

*Coyne*

TELEVISION SERVICING

**CYCLOPEDIA**



*Coyne* PUBLICATION

# COYNE TELEVISION SERVICING CYCLOPEDIA

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*A Practical Reference  
Book on Television  
Receivers and Reception*



***General Editors***  
BENNETT W. COOKE  
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**COYNE  
TELEVISION  
SERVICING  
CYCLOPEDIA**

# PREFACE

The COYNE TELEVISION CYCLOPEDIA is an entirely new book. It was planned especially as a guide and quick reference volume to provide instant, reliable information on Television receivers and reception.

**THE SCOPE OF THIS BOOK:** According to Webster's dictionary a Cyclopedia is: "a work giving a complete summary of some branch of knowledge." Any good Cyclopedia to be of maximum usefulness must be all of the following:

- |                  |                            |
|------------------|----------------------------|
| 1. CONCISE       | 4. EASY TO UNDERSTAND      |
| 2. AUTHORITATIVE | 5. SUBJECTS EASY TO LOCATE |
| 3. COMPLETE      | 6. ILLUSTRATED             |

The New COYNE TELEVISION CYCLOPEDIA meets all of these qualifications.

**CONCISE:** It is **CONCISE** because it expresses much technical material in brief, compact and "to the point" form. While every effort was made to present the material in a time saving way every subject was completely covered. The author and editors kept one thought foremost in mind—"make it complete but make it brief."

**AUTHORITATIVE:** Here is why this Cyclopedia is **AUTHORITATIVE**. Most of the information was selected from material contributed by over 70 companies engaged directly or indirectly in television set manufacture. The material was in the nature of special service and installation data which had previously been laboratory tested. This great quantity of tested data was then "screened" by Harold P. Manly (author of the original **CYCLOPEDIA OF RADIO**). This material for the **CYCLOPEDIA OF TELEVISION** was re-written to retain the best and most useful data. Mr. Manly's material was then carefully edited by members of the Coyne School staff. This procedure makes for the most accurate explanations possible.

**COMPLETE:** Any book with the above background and world of material could not help but be **COMPLETE**. The thought foremost in the minds of those who wrote and edited this book was—"make it completely cover each subject so there can be no doubt in the mind of the reader."

**EASY TO UNDERSTAND:** Throughout the book a "how-to-do-it" approach was used. Each article was written to meet the needs of the beginner but at the same time offer much that is **NEW** and **VALUABLE** to those with field experience in Radio or Television. An example of the procedure used can be seen in the material on **TELEVISION SERVICING**. In this section we included dozens of illustrated picture patterns (just as they were taken off the picture tube in operation). With these practical illustrations is complete information on the faults indicated and the remedy for correction of troubles. This is a **MODERN, PRACTICAL** approach to the servicing of Television receivers.

**SUBJECTS EASY TO LOCATE:** This book is written in the style of a dictionary. Subjects are arranged alphabetically making it amazingly simple for the reader to locate any subject he wishes. Hundreds of cross-references are used to assist toward easier and more complete understanding of every subject.

## PREFACE

**ILLUSTRATED:** The illustrations in the COYNE TELEVISION CYCLOPEDIA are many in number. They have all been planned with the same care as the text. Each photo or drawing has been prepared for this particular type of book and each vividly demonstrates some important point in the text.

Regardless of how many other Television books anyone may have this CYCLOPEDIA OF TELEVISION is a much needed book. Right now, great strides are being made in the industry. Television is destined to become one of the greatest forces in education and entertainment in the world. Special programs already conducted in surgical colleges (and pictured in this book) have vividly demonstrated the added advantage of this medium in teaching surgery. Where, only a few in a gallery could see an actual surgical operation in the past now hundreds can be as "close as the surgeon instructor" through the medium of Television.

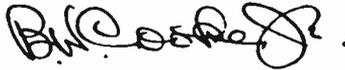
There will be many uses for Television in industry, the armed services, and schools of all types. Television today is still in its infancy but there is no doubt in anyone's mind of potential progress destined for this great new industry. This new medium can very easily become more important than the Press, Radio and Films for conveying information on National and World events.

The magnitude of the service industry is just beginning to be realized. With the tremendous growth of television the need for well trained technicians becomes more imminent. There is little doubt that the industry will experience a serious shortage of trained men for some time to come.

The Television industry will employ millions of men and women in the future. Anyone who is in the field or hopes to become associated with it should have and use a book of this type. A Cyclopedia is **ESSENTIAL** to a man already employed in Television—it is even more essential to men entering a new field like Television.

It is my belief that the COYNE TELEVISION CYCLOPEDIA provides a reliable reference source of information on Television. It can be a very important factor in the progress of anyone who enters the field of Television.

The Educational Book Publishing Division of the Coyne Electrical and Television-Radio School wishes to express its thanks to each of the contributing companies, editors, illustrators and others for their splendid cooperation and enthusiastic interest in every stage of the development of this book. The names are formally acknowledged on a following page.



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# CYCLOPEDIA OF TELEVISION

**ALIGNMENT.**—Alignment of a television receiver means the process of adjusting certain of its tuned circuits for suitable amplification or gain, or to specified band-pass characteristics throughout bands of carrier frequencies in the various channels, and at intermediate frequencies for which the receiver is designed. Alignment adjustments are found in the tuner section, in intermediate-frequency amplifiers for video and sound, and in the inputs to video detectors and sound detectors.

The shaded blocks of Fig. 1-1 indicate tuned circuits which may be aligned. Tubes are represented by circles. The transformer (coupling) between antenna and r-f amplifier may or

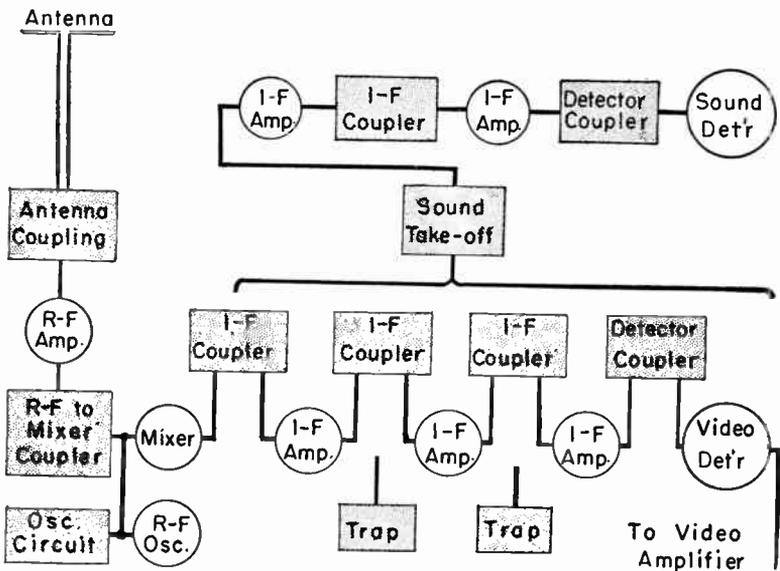


Fig. 1-1.—Parts or circuits of a television receiver in which there are adjustments for alignment.

may not be adjustable. The transformer between the r-f amplifier tube and the mixer tube usually is adjustable. The tuned circuit for the r-f oscillator nearly always is designed for alignment. Intermediate-frequency transformers between mixer,

i-f amplifiers, and video detector always are adjustable. Adjustably tuned interference traps are coupled to the video i-f amplifier stages of some receivers, and sometimes at the antenna transformer and preceding the sound takeoff. The sound takeoff may or may not be adjustably tuned. If there is more than one sound i-f amplifier each interstage transformer is adjustable, as is also the input transformer for the sound detector or demodulator.

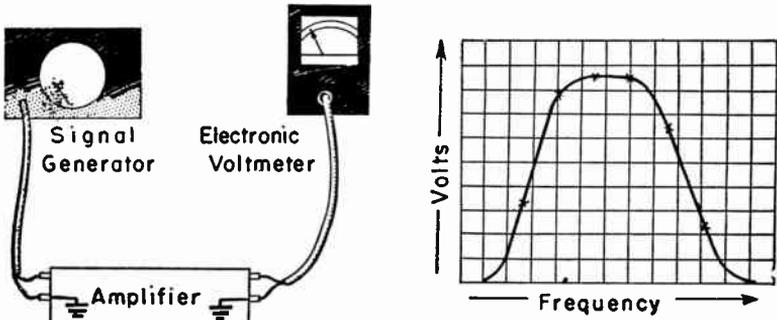
The order in which circuits or sections of the receiver are aligned depends on several factors. If trouble is indicated as being in some one section, because of observed symptoms or as the result of tests, and if the trouble appears due to misalignment, that is the section to be first aligned. If the entire receiver is to be realigned, or the adjustments checked, this work usually begins at the transformer preceding the video detector. Adjustments then are continued back to the transformer following the mixer. Traps in this portion of the receiver are aligned along with the transformers. Next would come the sound section, commencing at the transformer preceding the detector and following back to the sound takeoff. Final steps would include alignment of the r-f oscillator, the r-f to mixer transformer, and the antenna transformer.

**Instruments for Alignment.** — For aligning any section of the receiver it is necessary to have a signal source and an output indicator. The source is a generator which provides signal voltages at frequencies which may be varied throughout the band in which the section operates. For tuner alignment it is necessary that the generator provide carrier frequencies, for alignment of i-f amplifiers it is necessary to provide intermediate frequencies, and so on. The generator is coupled to the input of the section or circuit to be aligned.

To the output of the aligned section must be connected some instrument which will indicate response or relative voltage gain at all frequencies within the operating band. This output indicator may be an oscilloscope, an electronic voltmeter, or in some cases a high-resistance d-c voltmeter.

By using a signal generator for the source and an elec-

tronic voltmeter as the output indicator, as in Fig. 1-2, it is possible to take output voltage readings at a number of frequencies. These output voltages may be plotted against frequency on graph paper, as at the right, and a smooth curve drawn through them. Provided certain requirements are satisfactorily met, this curve represents the frequency response of the section or circuit being tested.



*Fig. 1-2.—Frequency response may be measured by using a signal generator and electronic voltmeter.*

The curve is a true frequency response only when the generator output is flat or of constant voltage throughout the range of frequencies, and when the indications of the voltmeter are unaffected by changes of frequency. Whether these requirements are reasonably satisfied may be checked by first connecting the output cable of the generator directly to the input of the voltmeter. Some generators are not capable of producing one volt of output. Note voltages obtained in Fig. 1-3. With the generator output adjusted to give a readable indication on the meter, the frequency is varied throughout the range to be used and the voltmeter readings are noted for various frequencies.

Fig. 1-3 shows variations of meter readings with frequency as found with two particular signal generators and electronic voltmeters for the range of intermediate frequencies between 20 and 27 mc. Obviously, an uncorrected response curve run with either of these combinations of instruments would not

show the true frequency response of an amplifier unless corrections were applied.

If the signal generator is equipped with a dependable calibrated attenuator the generator output may be adjusted to a uniform level at all frequencies. The voltmeter then must respond uniformly at all frequencies to be checked. Otherwise the non-uniform response voltages, such as those of Fig. 1-3, may be corrected to compensate for the variations. The ap-

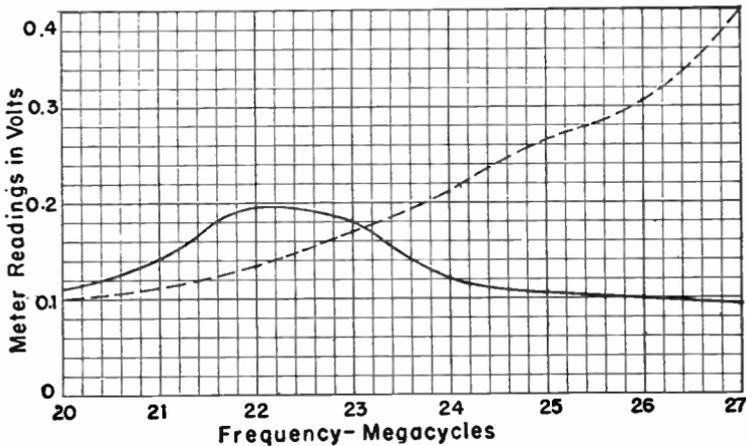


Fig. 1-3.—Correction curves for two combinations of signal generator and voltmeter.

proximate correction factor for each frequency may be determined from a curve plotted with the generator and meter connected directly together. The measured voltage at each frequency is divided by the voltage shown on the correction curve for that frequency. The corrected voltages then may be plotted on graph paper and a curve drawn through them.

Frequency response may be determined with an oscilloscope as the output indicator in a fraction of the time needed when using a voltmeter. When observing the output on an oscilloscope the input signal is provided by a sweep generator. A sweep generator furnishes a voltage whose frequency continually shifts back and forth throughout the range to be observed, usually at a rate of 60 times per second. The extent of frequency shift or sweep usually is adjustable from a fraction of a

megacycle to ten or more megacycles, depending on the design of the instrument.

If, for example, it is desired to observe the frequency response of an i-f amplifier between the limits of 21 and 29 mc, the sweep generator would be adjusted for a center frequency about midway between these limits, or to 25 mc, and the sweep width for about 10 mc to surely cover the extremes of response. Then, as shown at the top of Fig. 1-4, the frequency furnished by the generator would shift back and forth between 20 and 30 mc, increasing from 20 to 30 mc during  $1/120$  second, and decreasing from 30 to 20 mc in the following  $1/120$  second, assuming the sweep period to be  $1/60$  second.

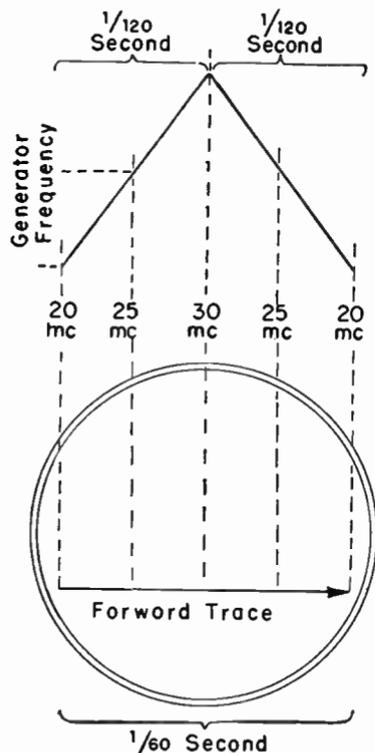


Fig. 1-4.—Generator frequency sweeps up and back again during each trace when using the internal sweep of the oscilloscope.

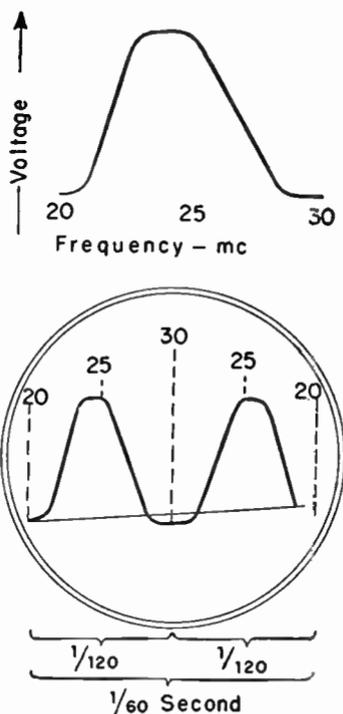


Fig. 1-5.—Appearance of a gain curve or frequency response on the oscilloscope when using the internal sweep.

The rate of horizontal sweep of the oscilloscope beam must match the rate of frequency sweep in the generator. This may be accomplished by using the internal sweep control of the oscilloscope or else by feeding to the horizontal input of the scope an external sweep voltage which is synchronized with the sweep timing of the generator.

If the generator sweep rate is 60 times per second, and the internal sweep of the oscilloscope is adjusted to 60 cycles per second, the generator frequency will increase and then decrease during each forward trace of the scope beam. This is illustrated at the bottom of Fig. 1-4.

We shall assume now that the generator output is coupled to the input of some amplifier, and that the amplifier output is connected to the vertical input of the oscilloscope. Relative voltage gain at swept frequencies might be as shown at the top of Fig. 1-5. This output voltage from the amplifier will cause two curves to be formed during each forward trace of the oscilloscope beam, as at the bottom of Fig. 1-5.

If a synchronized horizontal sweep voltage is supplied to the oscilloscope, instead of its own internal sweep voltage, and if the sweep rate still is assumed to be 60 times per second, the beam will travel from left to right during  $1/120$  second, and back from right to left during the following  $1/120$  second. A gain curve will be traced during each travel

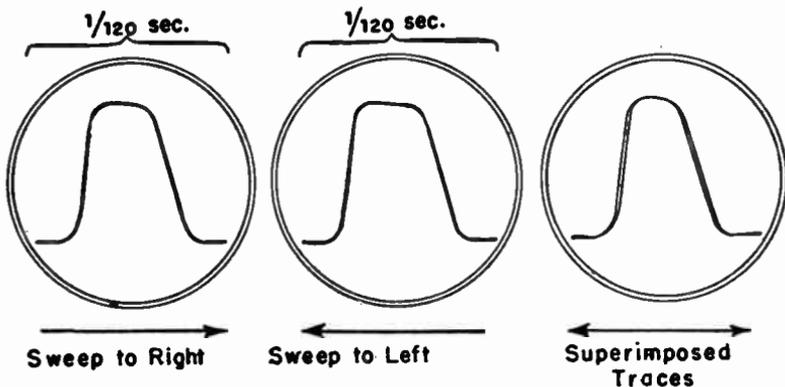
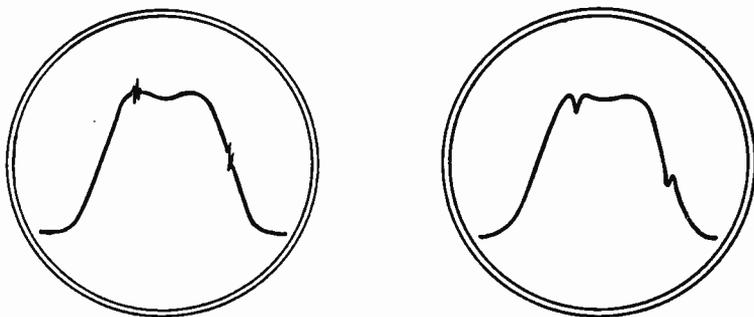


Fig. 1-6.—With synchronized sweep voltage for the horizontal input of the oscilloscope the two traces are made to appear as one.

of the beam. The forward and return curves will lie one over the other and will appear as a single curve such as illustrated by Fig. 1-6. Any change of frequency response brought about by alignment adjustment or otherwise will cause instant and corresponding change in the shape of the traced curve. The subject of synchronized sweep is treated more fully in the article on sweep generators.

Frequency calibration of a sweep generator seldom is precise enough to allow identifying an exact frequency by measuring horizontally across the response. Such identification requires an additional marker generator. The usual type of marker generator furnishes one or more steady frequencies at a time, like an ordinary frequency generator but of great accuracy. The marker frequency is adjustable, and may be set to any value within the range being observed.

When both the sweep generator and marker generator are coupled to the input of the amplifier their frequencies will beat together. At the instant in which the frequency from the sweep generator passes through a steady frequency of the marker generator the beat frequency goes through



*Fig. 1-7.—Frequency identification on traced responses by marker pips (left) and marker dips (right).*

zero, and there is a variation of gain. This variation is visible on the oscilloscope trace in the form of a marker pip as illustrated at two points on the trace at the left in Fig. 1-7.

Some marker generators are designed to absorb power at the frequency to which they are tuned. Absorption of power from the output of the sweep generator at this frequency reduces the gain where the two frequencies coincide on the trace. The result, shown at the right in Fig. 1-7, is a gap or dip in the trace at the point corresponding to the marker frequency.

Frequency at any point on a response curve may be identified by tuning the marker generator to produce a pip or a dip at that point. If a certain relative height or other characteristic is desired at some certain frequency on a response, the marker generator is tuned to that frequency and adjustments are made for the desired results where the marker appears. The functions of sweep generator and marker generator often are combined in a single instrument.

**Setups for Alignment.** — Nearly always it is necessary that the metal case of every test instrument, and the chassis of the receiver or amplifier worked upon, be securely connected to a common ground. This common ground is preferably a metal top on the work bench and a metal top on the instrument shelf, with these two well bonded together.

Instruments are provided with one or more ground posts or terminals in addition to their grounding connections made through shielded cables. These ground posts or terminals are to be connected to the metal shelf covering through flexible copper straps. It is possible that the chassis of the receiver or amplifier may be well grounded by its contact with the metal bench top, but it is better to make one or more connections with flexible copper grounding straps secured by screws or clamps.

The effectiveness of grounding may be checked by connecting all instruments to the receiver or amplifier, turning them on, and letting them warm up. Adjustments should be made to produce a reading on the voltmeter and a trace on the oscilloscope. Then the hand is touched to each instrument and to the receiver or amplifier chassis while observing the meter and scope. If there is any change in the meter reading

or in the form of the trace, the grounding is insufficient and additional connections should be made.

To duplicate the results of alignment as specified in manufacturers' instructions the a-c line should supply 117 volts. If actual line voltage is less than 115 volts or more than 120 volts a voltage adjusting transformer should be used between line and receiver.

When working with a receiver having series filaments or heaters, with one side of their circuit connected directly to the line and chassis, it is advisable to use an isolating transformer between power line and receiver. If a voltage adjusting transformer has its primary insulated from the secondary it will act also as an isolating transformer. This should be checked with an ohmmeter or circuit tester, since some voltage adjusting transformers are of the auto-transformer type.

If no isolating transformer is used, the line plug should be inserted in the receptacle in a manner to ground the chassis and thus avoid danger of shock from the hot side of the line. To check the plug position first connect between the chassis and a cold water pipe ground an a-c voltmeter capable of reading line voltage. Then insert the cord plug in the line receptacle. If the meter reads line voltage or nearly so the chassis is hot and the plug should be reversed. When the meter reads zero the chassis is connected to the grounded side of the power line. Some sets are made with the chassis "hot" regardless of the way the plug is inserted and therefore the test equipment must be connected to  $-B$ , and the chassis insulated from ground.

When aligning the front end or tuner, and also when aligning the video i-f amplifier, the automatic gain control should be overridden with a fixed bias as explained in the article on *Gain Control, Automatic*. When aligning the sound section its automatic volume control should be overridden in a similar manner. Otherwise these automatic controls may flatten the response curve unless the generator output is kept very low. Such flattening prevents showing true variations of gain with changes of frequency.

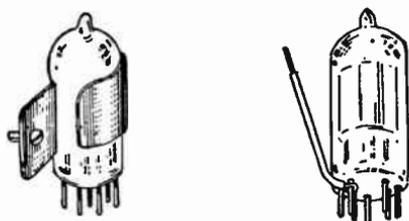
Alignment sometimes is carried out with the picture tube removed from the chassis, if the set is not of the type in which tube filaments are in series such as AC-DC sets. When a magnetic deflection type of picture tube is removed, the lead for its high-voltage anode must be well insulated and supported where it cannot make contact with chassis metal. Removal of an electrostatic deflection picture tube may so lessen the load on the high-potential voltage divider as to increase plate and screen voltages of amplifier tubes connected to this divider. In such a case the picture tube should remain connected.

All shielding should, if possible, remain in place during alignment. If tuning adjustments are altered while shielding is removed, the frequency response will change when the shielding is replaced. If many receivers of the same type are handled it may be advisable to make up dummy shields with cutouts for passage of tools. Most shielding has openings through which alignment tools may be inserted, and if not, no harm will be done by making the necessary small holes.

**Connections to Tubes.** — Connections from signal generators to tubes are best made to the tube base pins from instrument leads kept on top of the chassis, rather than to tube socket lugs from the bottom of the chassis. This is because signal radiation from exposed connections on the ends of generator leads then cannot reach the wiring in other circuits.

One of the simplest ways of injecting a signal into a glass-envelope tube is to slip a metal ring over the outside of the envelope and move this ring down to a position around the internal elements. A ring bent from a thin piece of aluminum is shown at the left in Fig. 1-8. With the generator lead clipped to the metal ring there is capacitive coupling through the glass envelope to the tube elements. This method may require a rather strong signal voltage from the generator in case the aligned circuits are far out of adjustment to begin with. Instead of the ring it is possible to use a close fitting tube shield with the generator lead connected to the shield. This shield must not touch chassis metal.

Another simple connection is made with a short length of insulated wire as at the right in Fig. 1-8. One end of the wire is bared just far enough to allow making a single tight turn around a tube pin, which usually is the pin for the control grid. This turn should be formed and forced onto the base pin while the tube is removed from its socket. The other end of of



*Fig. 1-8.—Connections for coupling signal generators to tubes.*

the short wire is bared to take the connection from a generator. Unless too much insulation is removed, the bare wire will not touch chassis metal when the tube is replaced. Solid wire, such as hookup wire, should be used in order that the tight turn will hold in place.

Some television chassis are built with cutouts or openings in the side to enable the serviceman to reach a socket lug or other component parts. In such cases the recommended procedure is to bare one end of wire, make a small hook on the end and pass it through the opening to reach the socket lug or other component part.

Test adapters are available for all types of tubes. Each adapter has socket openings mounted above base pins, with outwardly extending lugs connected to each pin. With a tube inserted in the adapter, and the adapter placed in the tube socket, it becomes possible to make test connections to any elements. Test adapters are not recommended when working at television frequencies since the removal after alignment will detune the circuit and possibly cause mis-alignment. Some receivers have special test jacks or terminals which allow making alignment connections to correct points for both signal input and output.

If a generator connection is made to the control grid of a tube it is necessary that a capacitor be in series with the lead in order that d-c grid bias may be maintained. Some generators have a capacitor built in. Otherwise an external capacitor must be on the generator lead. At the high frequencies used in television, a capacitance of 10 to 20 mmf should be ample to pass the required signal, although it is best to adhere to the manufacturer's recommended values. The series capacitor is not needed when introducing a signal through a metal ring or a shield placed over the tube.

After all test connections have been made and the grounding checked, the receiver or amplifier and all the test instruments should be turned on and allowed to warm up for 20 to 30 minutes before commencing alignment. This is required in order that all resistors, inductors, and capacitors may reach their normal working temperatures and values before being adjusted for frequency response.

For alignment of i-f amplifiers there is no objection to using a screw driver or wrench having a metal tip in an insulating handle, but for alignment of r-f and oscillator circuits the tool should be wholly of insulating material.

Every change of a tuning adjustment in any one circuit will affect not only that circuit but also others which are connected or coupled to it, or whose inductors or capacitors are close to those of the first circuit. As a consequence, the adjustments of all associated circuits should be gone over after any alignment has been altered.

**Tuning Wand.** — A tuning wand is a means for determining the probable effect of altering alignment adjustments before the adjustments actually are changed. The wand, illustrated by Fig. 1-9, usually consists of six to eight inches of insulating rod or tubing having on one end a small cylinder of non-

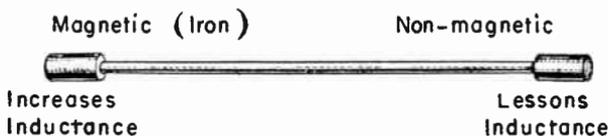


Fig. 1-9.—A tuning wand.

magnetic metal, such as copper, brass, or aluminum, and on the other end a small cylinder of iron, such as the powdered iron slugs used in some transformers. The metal ends should be covered with insulation, possibly cellophane tape, so they won't cause short circuits when touched to bare wires.

When the iron end of the wand is brought into the field of a coil it increases the inductance, and the effect is the same as though the alignment adjustment were altered in a way to increase inductance. When the non-magnetic end of the wand is brought into the field of a coil it reduces the effective inductance, and the effect is the same as though the alignment adjustment were changed to reduce inductance.

Frequency of resonance is decreased by adding inductance, and increased by reducing inductance. Resonant frequency is also decreased by adding capacitance, and increased by lessening capacitance in a tuned circuit. Capacitance acts like inductance in its effect on resonance. Consequently, using the iron end of the wand shows what will happen with an increase of inductance, an increase of capacitance, or both. Using the non-magnetic end shows what will happen when reducing inductance, reducing capacitance, or both.

While observing a response curve on the oscilloscope the ends of the wand may be brought near each coil or transformer to identify the portions of the response curve which will undergo maximum change when that coil or transformer is adjusted.

The wand may be used to determine what adjustment is needed to make a circuit resonant at any given frequency. With that frequency being applied to the circuit the ends of the wand are brought to the coil or inductor. If response or gain is increased by the iron end, the adjustment should be changed to increase either inductance or capacitance. If gain increases with the non-magnetic end of the wand, the circuit needs less inductance or capacitance to make it resonant. If gain decreases with both ends brought near the inductor, the circuit already is resonant at the applied frequency.

**Overall Check of Alignment.** — After a tuner and video i-f amplifier have been separately aligned, or after either one has

been aligned, they should be operated together while observing the overall response from antenna to video detector output.

With the regular antenna or transmission line disconnected from the receiver, connect the high sides of the sweep and marker generators to either antenna terminal. Connect the low sides of the generators to ground, as shown, or to a second antenna terminal if the receiver is designed for balanced input. Connect the vertical input of the oscilloscope to the high side of the video detector load resistor and the low side to ground. The horizontal input of the scope would be connected to the sweep output of the generator if synchronized sweep is employed.

Override the automatic gain control with a fixed bias. Set the contrast control to a usual operating position, say about one-third down from maximum. If there is a fine tuning or trimmer adjustment for the r-f oscillator set this control to its midposition. Turn on the instruments and the receiver, and allow about 20 minutes for warming up.

Then set the channel selector of the receiver to any channel, preferably one on which reception is known to be possible. Advance the vertical gain control of the oscilloscope and the output control or attenuator of the sweep generator. Adjust the sweep width for about 10 mc and tune the frequency adjustment of the sweep generator to bring a response curve onto

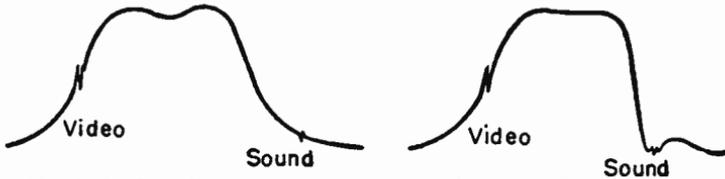


Fig. 1-10.—Overall response curves with markers for video and sound carrier frequencies.

the screen of the oscilloscope. Keeping the vertical gain of the scope well advanced, reduce the output of the sweep generator to the lowest value which produces a trace of readable height.

It is next in order to set the receiver channel selector to the highest channel and tune the sweep generator to bring a response curve onto the scope. Then tune the marker generator

to the video carrier frequency for the channel being checked, and next to the sound carrier frequency of the same channel. As shown by Fig. 1-10, the video marker should appear about half way down on one slope of the curve, and the sound marker far down or in a dip produced by a sound trap on the other slope.

If the marker generator does not provide carrier frequencies, but tunes only through intermediate frequencies, the setup may be modified as shown by Fig. 1-11. The sweep generator is connected to the antenna transformer and the oscilloscope to the video detector load resistor as previously explained. The

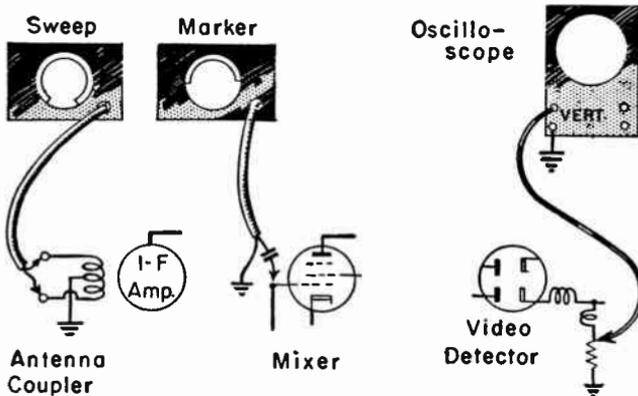


Fig. 1-11.—Applying marker signals to the mixer tube with the sweep generator connected to the antenna terminals.

marker generator is coupled to the mixer tube. All other preparations and steps for checking are the same as with the marker connected to the antenna terminals, except that the marker generator now is tuned first to the video intermediate frequency and then to the sound intermediate frequency of the receiver rather than to carrier frequencies.

This second method will show where the intermediate frequencies are located on the overall response curve. The markers will move with the curve as sweep frequency is changed. There will be nothing to identify either video or sound carrier frequencies on the response, since they are not being introduced by any of the test instruments.

**Uhf Alignment.** — When aligning ultra-high frequency receivers or converters use the sweep generator and oscilloscope to check frequency response throughout the entire range normally tuned. Obtain a response trace for the lowest channel, then vary together the tuning of sweep generator and receiver or converter to maintain a response through all higher channels. Shape and gain of the response may vary, but provided the uhf tuning operates properly there should be no channel frequencies at which gain drops to zero.

If uhf and vhf oscillators remain active at the same time there may be spurious response curves. The response to be observed while checking alignment is one that is shifted across the oscilloscope screen when sweep frequency is varied.

Alignment adjustments will be found in all or several of the following circuits.

**1. Channel selector.** Usually consists of two resonant circuits, one for the antenna and another for mixer input. In some units there is only a single adjustable coupling, from antenna to mixer input. Adjustments affect the pass band, the gain, and the tilt of response curves.

**2. Uhf oscillator tuning.** Adjustment affects tracking, the relation between channels actually tuned and channel numbers on the tuning dial. The position of a pointer on the tuning dial usually is readily adjustable, and may be set for best compromise between dial readings and channels actually tuned.

**3. Harmonic selector tuning.** This adjustment is found only in systems employing a harmonic of a vhf oscillator frequency instead of a separate uhf oscillator.

**4. I-f amplifier alignment.** Adjustments in the grid circuit and also in the plate circuit of amplifiers which follow the uhf mixer are similar to those for r-f and i-f amplifiers in vhf receivers, usually consisting of movable cores in tuning inductors.

Adjustments in uhf circuits are of various types. Most often there are variable capacitors with movable plates or movable cores. Also used are movable cores in inductors.

Sometimes the only adjustment is by spreading or squeezing the turns of inductors.

Adjustments in the grid-plate circuit of a uhf oscillator affect tuning on all channels. If oscillator capacitance and inductance are both adjustable, either may be designed to have chief effect on low channels and the other to have chief effect on high channels.

With single conversion uhf systems the uhf oscillator frequency is higher than the receiver carrier frequency, just as with vhf tuning. For double conversion the uhf oscillator frequencies are lower than those of received carriers.

For observing frequency responses the sweep generator is connected through a suitable matching pad to the uhf antenna terminals. The oscilloscope may be connected to any of the following points.

**1. Output of uhf mixer.** This connection allows observing the effect of adjustments in antenna, preselector, and mixer input couplings, but not the effect of adjustments between mixer and i-f amplifier. Since the signal has been demodulated at this point it is not necessary to use a detector probe on the scope input. Signals will be very weak, and the scope must have high vertical gain.

**2. Output of any i-f amplifier which follows the mixer.** This allows observing the effect of couplings between mixer and the i-f amplifier, as well as of all preceding couplings. Shape of the response curve is determined chiefly by i-f input or interstage couplings, but is affected also by all preceding couplings back to the uhf antenna input. Signals are at the intermediate frequency, consequently a detector probe must be used on the scope input.

**3. Video detector load.** This is the connection commonly used for i-f amplifier alignment in any receiver. Shape of the response curve is determined chiefly by the i-f amplifier section, but is affected also by adjustments of all preceding couplings. No detector probe is required on the scope input.

**AMPLIFIERS, BROAD BAND.**—A broad band amplifier is one having a large ratio of highest to lowest frequencies which

are amplified approximately equally. It is not the width of band pass in cycles or megacycles that makes an amplifier a broad band or narrow band type, it is the frequency ratio.

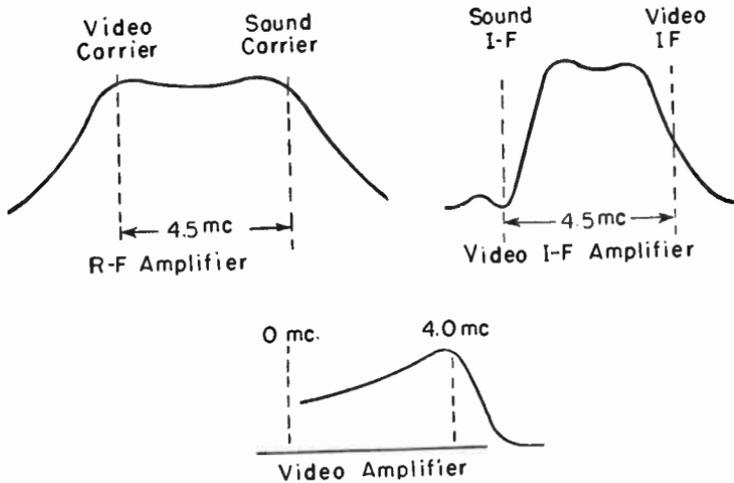


Fig. 2-1.—Typical frequency responses of broad band amplifiers.

Fig. 2-1 shows responses of narrow band r-f and i-f amplifiers. Although passbands are 4.5 mc, the high to low frequency ratio of a vhf r-f amplifier is no more than about 11 to 10, and of an i-f amplifier is no more than 12 to 10.

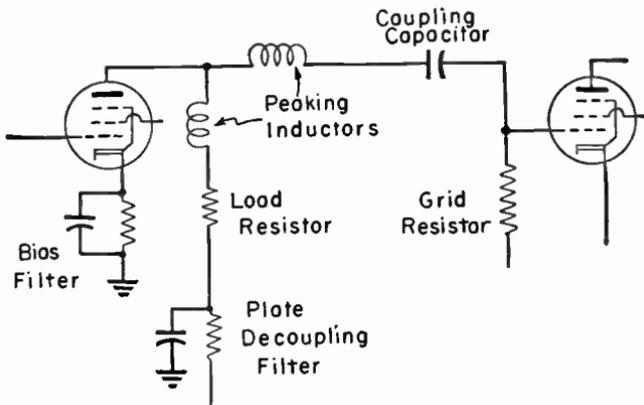


Fig. 2-2.—Parts which affect response at lowest and highest frequencies in video detector and video amplifier couplings.

On the other hand, frequency ratio for the video amplifier is very great, because highest frequencies to be uniformly amplified are around 4 mc and the lowest are near 50 cycles per second. This is a ratio of 80,000 to 1.

Fig. 2-2 shows elements used in video amplifier couplings to allow fairly uniform gain throughout a wide range of frequencies. How these elements perform is explained in the article on *Amplifiers, Video*.

**AMPLIFIERS, DIRECT-COUPLED.** — A direct-coupled amplifier is one in which the plate of one tube is conductively connected, without intervening capacitor or inductor, to the control grid of the following tube. The principle is shown by Fig. 3- , where the elements of both tubes are connected to various points along a voltage divider between B+ and B-. Any suitable voltages might be used, those shown on the diagram being merely illustrative.

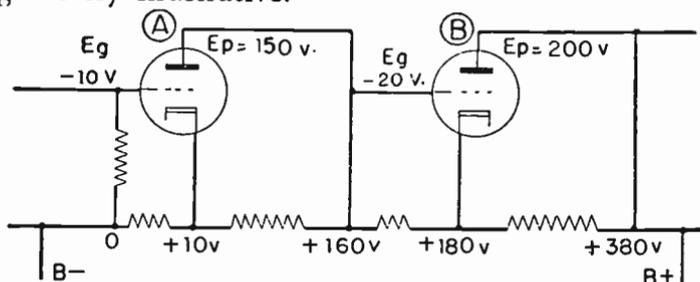


Fig. 3- .—Elementary principle of the direct-coupled amplifier.

A direct coupled amplifier will follow very low frequencies. The low limit is determined by reactances of capacitances in the tubes and the B-supply. High frequency amplification is limited by effects of tube and stray capacitances,

**AMPLIFIERS, PUSH-PULL.** — In a push-pull amplifier circuit the input signal is caused to form two similar voltages of opposite phase which are applied to grids of two push-pull tubes. Opposite-phase signal voltages from the two push-pull plates are fed to a transformer in such manner as to obtain at its input the sum of plate signal powers.

During one half-cycle of signal, plate voltage becomes less positive or effectively more negative at one push-pull tube and becomes more positive at the other push-pull tube. As a consequence, changes of signal voltage occur in the same direction across both halves of the primary in the output transformer. During opposite half-cycles of signal there is reversal of voltage across the output transformer primary.

Since only half the signal voltage from the preceding amplifier, or secondary of the input transformer, goes to each push-pull grid, total input signal must be twice that which would be needed for a single output tube if push-pull output is to be double that of the single similar output tube.

D-c currents between B+ and the two push-pull plates flow in opposite directions through the halves of the output transformer primary. When these d-c currents are equal, as with no signals applied, their magnetizing effects are equal but opposite, and can induce no currents or voltages in the output transformer secondary. Under such conditions there is cancellation of B-supply ripple or hum voltages, also of noise coming from the a-c power line. Hum or noise from preceding stages or from pickup by wiring and circuit elements is not cancelled.

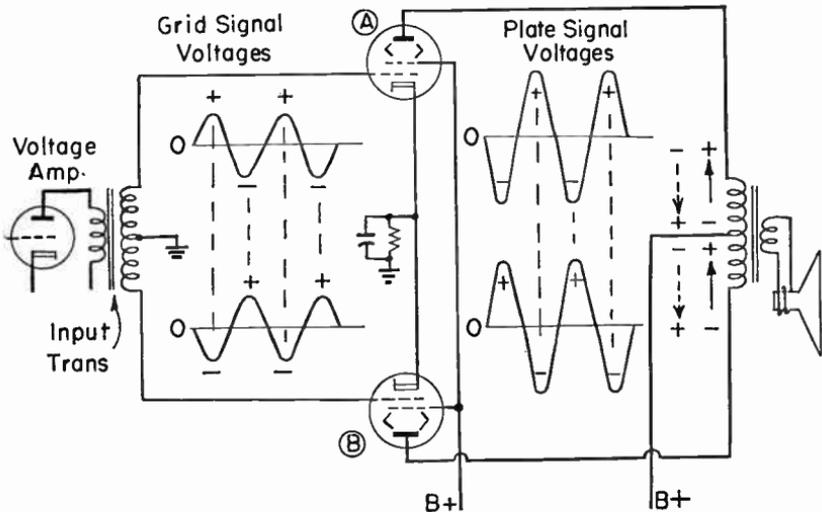


Fig. 3-1.—Two output tubes connected in push-pull, showing how signal voltages divide and recombine.

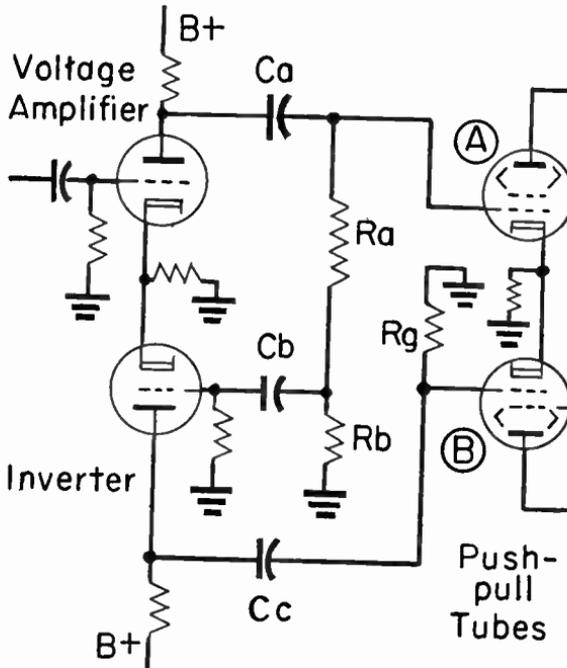


Fig. 3-2.—An inverter employing a resistance voltage divider.

The push-pull principle is illustrated by Fig. 3-1, where the two push-pull tubes are marked *A* and *B*. Opposite-phase grid signal voltages are obtained from opposite ends of a center-tapped transformer secondary, with the primary fed from a preceding voltage amplifier. Plate signal voltages go to opposite ends of a center-tapped primary in an output transformer.

There is cancellation of second and other even harmonic voltages which result from operating the push-pull tubes with signal voltage peaks in excess of negative bias voltage, or driving the tubes to plate current cutoff.

**Inverters For Push-pull.** — Push-pull amplifiers of moderate output power usually obtain equal and opposite grid signal voltages from a phase inverter rather than from a tapped transformer. A voltage divider type of inverter is shown by Fig. 3-2. Push-pull tubes are marked *A* and *B*, the preceding voltage amplifier is at the upper left, and the inverter tube

is at the lower left. The voltage amplifier and inverter usually are sections of a twin triode.

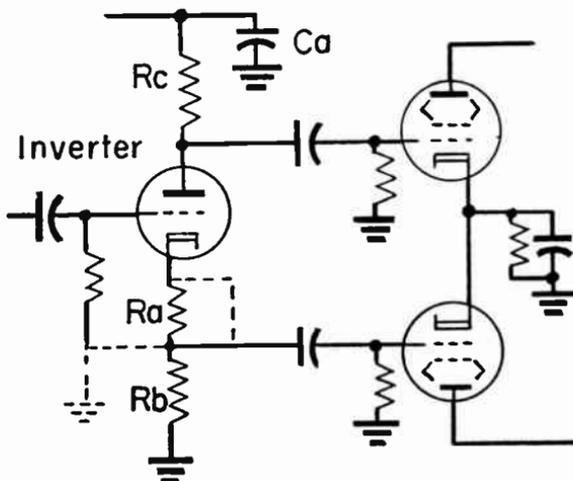


Fig. 3-3.—A push-pull inverter of the cathode follower type.

Signal voltage goes from the voltage amplifier plate through capacitor  $C_a$  to the grid of push-pull tube  $A$ . This same signal voltage is across resistors  $R_a$  and  $R_b$  in series, with the portion across  $R_b$  applied through capacitor  $C_b$  to the inverter grid. The inverter amplifies this portion of the signal voltage, and from the inverter plate the signal goes through capacitor  $C_c$  to the grid of push-pull tube  $B$ . Since there is polarity inversion between grid and plate of the inverter, the signal at the grid of tube  $B$  is opposite in polarity or phase to that at the grid of tube  $A$ .

Resistances at  $R_a$  and  $R_b$  are so proportioned that the ratio of signal voltage across  $R_b$  (at the inverter grid) to total voltage across the divider (at grid of tube  $A$ ) is inversely propor-

tional to amplification in the inverter. For example, were inverter gain to be 20, voltage across  $Rb$  would be  $1/20$  of that at the grid of tube  $A$ . Then amplified voltage from the inverter plate, going to the grid of tube  $B$ , would be made equal in amplitude to voltage at the grid of tube  $A$ .

Fig. 3-3 shows connections for a cathode follower type of inverter. Signals from a preceding amplifier come to the inverter grid. There is no polarity inversion between a grid and a cathode; consequently, signals from the inverter cathode to the grid of the lower push-pull tube are of the same phase as at the inverter grid. Polarity inversion between grid and plate allows the inverter plate to deliver to the grid of the upper push-pull tube signals whose phase is opposite to those at the inverter grid.

Resistors  $Ra$ ,  $Rb$ , and  $Rc$  are so proportioned as to allow equal signal amplitudes from cathode and plate of the inverter, and to properly bias the inverter grid. There is no gain from this type of inverter. Still other types of inverters are in use. All are used in connection with an output transformer having a center-tapped primary for push-pull plates and for the  $B+$  voltage supply.

**AMPLIFIERS, RADIO-FREQUENCY.** — The r-f amplifier is part of the tuner or front end of the television receiver, where are found also the r-f oscillator and the mixer. Signals from the antenna, at carrier frequencies, are applied to the input of the r-f tube, and from the output of this tube are taken to the control grid of the mixer. R-f amplifier tubes most often are sharp cutoff pentodes or else twin triodes, both of the miniature type.

Fig. 4-1 shows one of the simplest couplings used between antenna and r-f amplifier. The center-tapped coil  $La$  is assumed to provide a reasonably good match of its own impedance to that of the antenna and transmission line at all frequencies to be received. With such a simple coupling there

is little voltage gain from antenna input to r-f amplifier output.

The output circuit of the r-f amplifier always is tuned to frequencies in each channel to be received. This provides selectivity against signals in other channels, reduces the pos-

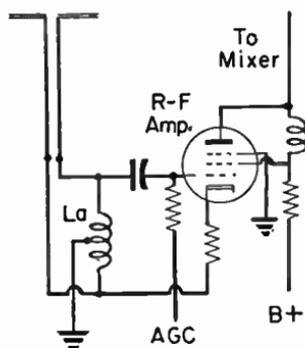


Fig. 4-1.—Untuned coupling between antenna and r-f amplifier.

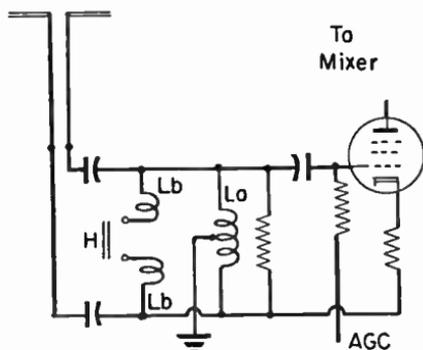


Fig. 4-2.—Input coupling whose impedance is changed for different channels.

sibility of image interference, and prevents radiation from the r-f oscillator through the antenna. It is for these reasons that the r-f amplifier is useful, even when it contributes negligible gain.

The more nearly the input impedance of the receiver is matched to the impedance of antenna and transmission line for each channel frequency the greater will be the gain. When a single antenna element is used for more than one channel there is change of antenna impedance with every change of frequency from channel to channel. In the receiver input coupling there is also a change of impedance with every change of frequency, as is true with any circuit containing inductance and capacitance.

Variation of input impedance between low band and high band channels is lessened in some receivers by altering inductance or capacitance of the input circuit. In Fig. 4-2 the inductance is lowered for high-band reception by closing switch  $H$  to place inductors  $L_b$  in series with each other and

in parallel with inductor  $L_a$ . In the article on tuners are shown more elaborate methods for tuning antenna coupling circuits for each channel, and methods for tuning the coupling between r-f amplifier and mixer tubes.

In ordinary triode amplifier circuits operated at very-high and ultra-high frequencies there is strong tendency to oscillate because of feedback through plate-grid capacitance, from the high r-f potential on the plate to the high side of the grid circuit. Oscillation may be prevented by using circuits such

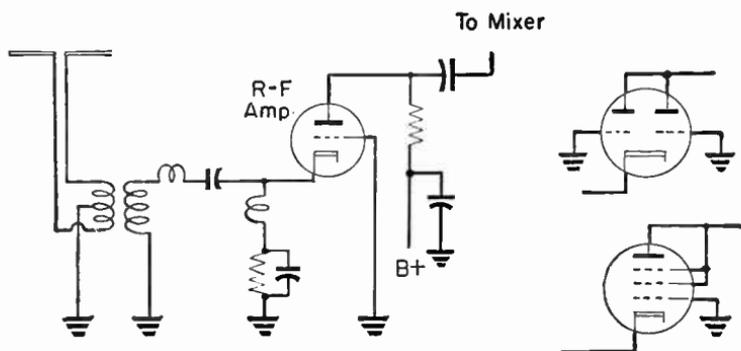


Fig. 4-3.—Connections used for grounded grid r-f amplifiers.

as that of Fig. 4-3. Here the grid is grounded. The antenna input signal is applied between the cathode and the grounded grid, which places the cathode at high r-f potential on the input side. The output signal is taken from between plate and grounded grid, with the plate at high r-f potential on the output side. The grounded grid acts as a shield between cathode and plate, or between the high sides of output and input circuits, and reduces feedback capacitance to a very low value.

**Cascode R-f Amplifiers.** — Many television tuners have what is called a cascode amplifier in the r-f stage. This is a combination of the familiar triode amplifier which has a grounded cathode and another triode amplifier of the grounded grid type. The two amplifiers are the two sections of a twin triode

tube designed especially for this and similar applications. Among such tubes are the 6BK7, the 6BQ7, and the 6BZ7, all of which perform satisfactorily at frequencies up to 250 or 300 megacycles and have transconductances on the order of 6,000 to 8,500 micromhos when operated with usual voltages and element currents.

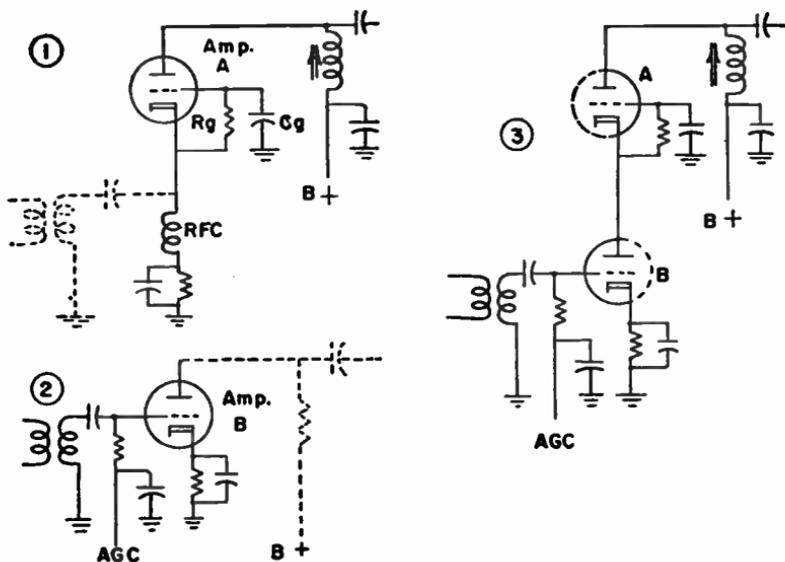


Fig. 137-19.—The principle of the cascode amplifier having the two triodes in series for d-c plate-cathode current.

The principle of the cascode circuit is illustrated by Fig. 137-19. Diagram 1 shows connections for a grounded grid amplifier. The grid is grounded for r-f signal currents and voltages through capacitor  $C_g$ , and is connected to the cathode through resistor  $R_g$  in order to stabilize the d-c potential on the grid. Signal input is between cathode and ground, with high impedance at signal frequencies in the cathode-to-ground path furnished by choke  $RFC$ . Signal output may be taken from the plate circuit in any way suitable for r-f amplification. For convenience during explanation we shall designate this grounded grid circuit as amplifier *A*.

Diagram 2 shows a triode circuit such as might be used as the r-f amplifier in a tuner. The cathode is grounded for r-f currents and voltages through the capacitor that bypasses the cathode-bias resistor. Signal input is between grid and ground. The output signal appears across a resistance or impedance load in the plate circuit. This grounded cathode amplifier will be called amplifier *B*.

Diagram 3 shows the cascode circuit. At the top is the grounded grid amplifier, *A*. The cathode circuit no longer contains only a simple impedance across which is applied the input signal, instead there is amplifier *B* between the cathode of amplifier *A* and ground. Signal voltage still is applied to the grounded grid amplifier, *A*, between its cathode and ground, but this is done by inserting amplifier *B* in this position. The load for the plate circuit of amplifier *B* is now the input impedance of amplifier *A*.

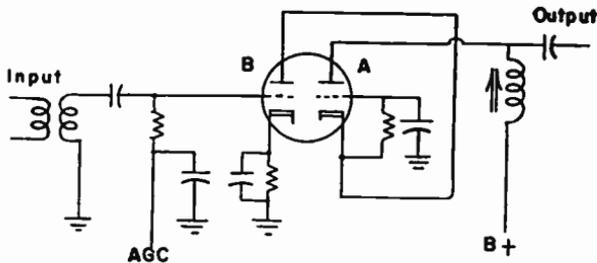


Fig. 137-20.—How the cascode r-f amplifier may be shown in television receiver diagrams.

The cascode r-f amplifier has overall gain comparable to that from an r-f pentode, and at the same time has the lower noise factor that is characteristic of triodes. Minimum noise is important in r-f amplifiers, because noise voltages originating here will be amplified by all following stages in the receiver. Feedback trouble, such as might occur with ordinary triode circuits, is practically eliminated because of input loading and rather limited gain in the first section, and by the grounded grid of the second section. In cascode circuits there may be various arrangements of resistors, capacitors and inductors



this section. Thus the automatic gain control affects both sections, although applied to only one of them. There is reduction of plate-cathode current in both sections.

There are many variations of the cascode amplifier circuit. One of the most widely used is shown in principle by Fig. 137-21. Section *A* still is a grounded grid amplifier, and section *B* is a grounded cathode amplifier. But the two amplifiers no longer are in series for d-c plate-cathode currents, although they are in series for r-f signal currents and voltages, due to the low impedance or reactance of capacitor  $C_c$  between the cathode of section *A* and the plate of section *B*. Total B-voltage from the d-c supply, except for the drops through resistors  $R_a$  and  $R_b$ , is applied to each of the plates. Each section has its own cathode bias, consisting of a biasing resistor bypassed by a capacitor to ground.

R-f signal voltages and currents are isolated from the d-c bias system of amplifier *A* by an r-f choke between the cathode and the bias resistor. Signal voltages and currents are isolated from the B-supply by another r-f choke in the plate circuit of amplifier *B*. Then the only low-impedance path from the plate of section *A* to the cathode of section *B* is through capacitor  $C_c$ .

R-f signal input is between the grid of section *B* and ground. Signal output is from the plate circuit of section *A*. The types of input and output couplings may be as shown, or of any other kinds suitable for r-f couplings. The types of input and output couplings have nothing in particular to do with the

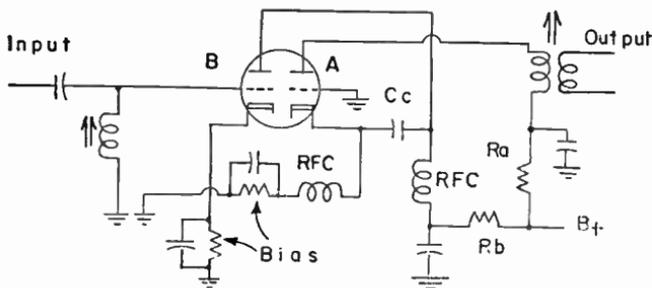


Fig. 137-22.—The second style of cascode r-f amplifier may be shown in this general manner as circuit diagrams.

principle of the cascode amplifier. This principle is the series connection for r-f signals of a grounded cathode amplifier and a grounded grid amplifier.

The cascode circuit of Fig. 137-21, shown with separate triodes for simplicity of explanation, ordinarily would be shown with a twin-triode tube symbol in some such manner as in Fig. 137-22. The same parts are similarly lettered in both diagrams.

**AMPLIFIERS, SWEEP.** — Vertical and horizontal sweep amplifiers provide sawtooth voltages which act through output transformers to cause sawtooth currents in the coils of yokes for picture tubes operating with magnetic deflection. Inputs to the grids of sweep amplifiers are sawtooth voltages coming from vertical and horizontal sweep oscillators. The electrical position of sweep amplifiers between the oscillators and the deflection system of television receivers is shown by Fig 5-1.

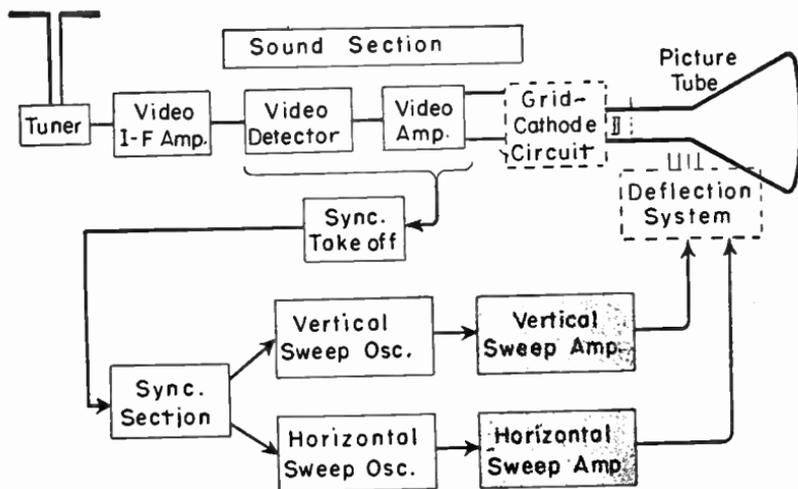


Fig. 5-1.—Electrical relations of sweep amplifiers to other parts of a television receiver.

**Magnetic Deflection Sweep Amplifiers.** — The deflection voltage developed by a sweep oscillator controlling charge and discharge of a capacitor is of sawtooth waveform. In the de-

deflection coils for the picture tube designed for magnetic deflection there must be a current of sawtooth waveform. When the original sawtooth voltage is applied to the control grid of a sweep amplifier there will be a sawtooth voltage developed across the load in the plate circuit, but only with amplifier tubes of high plate resistance will the plate current be of saw-

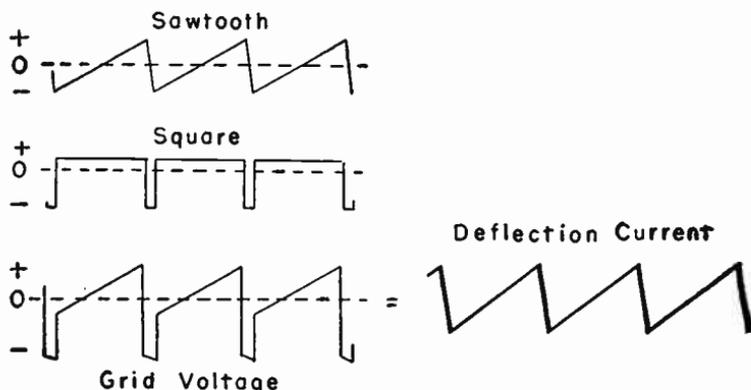


Fig. 5-9.—Characteristics of a grid voltage which produces a sawtooth plate current.

tooth waveform. Sweep amplifier tubes for vertical magnetic deflection systems ordinarily are power triodes, power pentodes, or beam power tetrodes, with the latter two types connected as triodes. All these tubes then have relatively low plate resistance and will not develop a sawtooth plate current when a sawtooth voltage wave is applied to their grids.

The reason for the difficulty is that the inductive reactance of the transformer windings and deflection coils in the amplifier plate circuit is large in comparison with the plate resistance of the tube. Were the plate load purely inductive, with only inductive reactance and no resistance, a sawtooth current would be produced in that load only by application of a square wave of voltage. Actually the plate circuit load is partly inductive and partly resistive, and to produce in it a sawtooth wave of current requires a voltage which is a combination of sawtooth and square waves. The general form of such a voltage is shown by Fig. 5-9. This is the form of voltage wave which

must be applied to the grid of the vertical sweep amplifier which has low plate resistance.

The combination sawtooth wave with a negative pulse or negative square wave may be produced by connecting a resistor  $R_s$  of Fig. 5-10 in series with the sawtooth capacitor  $C_s$ . When the oscillator or discharge tube becomes conductive for the retrace period the sawtooth capacitor has to discharge

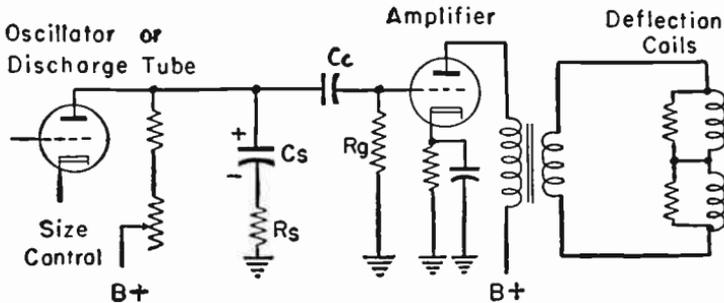


Fig. 5-10.—Circuit for production and use of a sawtooth voltage combined with a square pulse.

through the series resistor. The rate of discharge is slowed down by the resistor. The rate depends on the time constant of the sawtooth capacitor and its series resistor. The greater the capacitance and the greater the resistance, the slower is the discharge. By suitable choice of capacitance and resistance, or of their time constant, some of the original charge still will remain on the sawtooth capacitor when the oscillator or discharge tube again becomes non-conductive.

Charge and discharge of the sawtooth capacitor, and the resulting voltage for the amplifier grid, now are as shown by Fig. 5-11. Referring to the left-hand graph, the capacitor is charging and its voltage is rising until the instant represented by point  $a$ . At this instant the oscillator or discharge tube becomes conductive, and there is instant drop of voltage from  $a$  to  $b$  because the plate-cathode resistance of the tube becomes nearly zero when the grid is made positive. The capacitor discharges through the tube, and continues to discharge during the period from  $b$  to  $c$ . At the instant represented by  $c$  the

tube becomes non-conductive, its grid goes negative beyond the value for plate current cutoff.

Because the rate of capacitor discharge has been retarded by resistor  $R_s$ , there is still some charge remaining at instant  $c$ . Then the voltage rises to a value proportional to this remaining charge. The rise of voltage is from  $c$  to  $d$ . Then, with

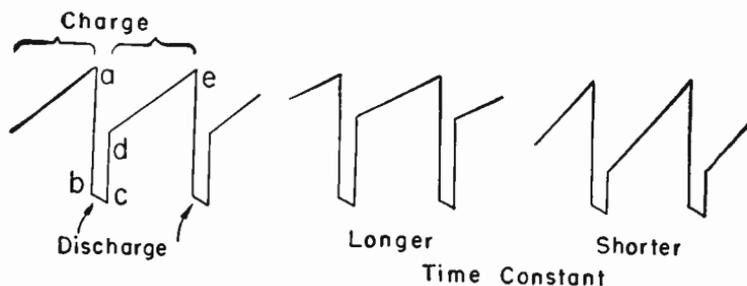


Fig. 5-11.—How the negative pulse is added to the sawtooth voltage.

the tube remaining non-conductive, there is the usual charging and rise of voltage along the line from  $d$  to  $e$ . Everything at  $e$  is the same as at  $a$ , and the performance repeats over and over.

If the resistance of  $R_s$  or the capacitance of  $C_s$  is increased, the time constant is lengthened, there is slower discharge, and a greater retained charge and voltage. Then the voltage wave changes as shown by the center graph of Fig. 5-11. If resistance or capacitance is lessened the time constant is shortened, there is more rapid discharge to a lower voltage, and the voltage wave becomes as shown at the right.

Such changes of voltage waveform at the amplifier grid cause changes in the waveform of current in the deflection coils and alter the distribution of picture elements between top and bottom of the screen. This means a change in vertical linearity. In some receivers the resistor  $R_s$  or part of it is made adjustable. This adjustment is a control for vertical linearity. Sometimes it is called a peaking control because it varies the amplitude of the negative peak added to the sawtooth voltage.

The vertical sweep circuit for magnetic deflection as shown by Fig. 5-12 is used in many receivers. In series with the sawtooth capacitor  $C_s$  is the discharge limiting resistor  $R_s$  whose function has been explained. But instead of the lower end of this resistor going to ground it connects to the high side of the cathode biasing resistors for the amplifier tube. Direct current for both charge and discharge of the sawtooth capacitor now flows through the biasing resistors in series with resistor  $R_s$ .

The adjustable cathode bias resistor forms a control for vertical linearity. Its principal effect is on the portion of the picture above the center, stretching or contracting this top portion as the adjustment is varied one way and the other.

Capacitance of the sawtooth capacitor  $C_s$  is usually 0.05 or 0.1 mf. Resistance in  $R_s$  is usually somewhere around 7,500 ohms, and in the cathode bias resistors is usually about 5,000 ohms or somewhat less, with most of this bias resistance in the adjustable unit. The full range of adjustment will vary the

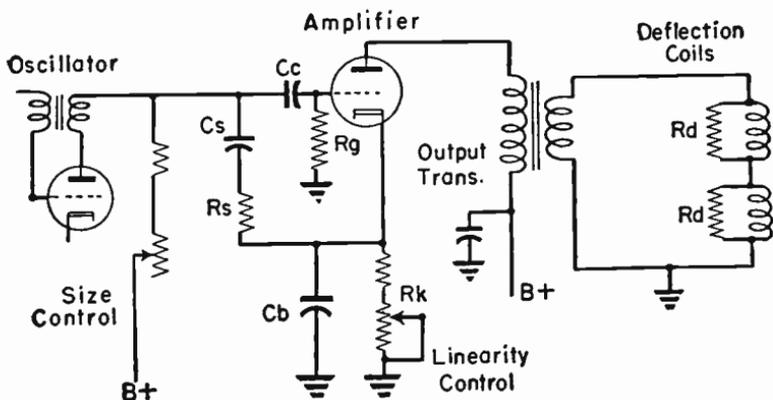


Fig. 5-12.—Typical vertical sweep circuit for magnetic deflection.

time constant of capacitance and resistance by about 25 per cent either way from its average value. At the same time, this adjustment varies the grid bias of the amplifier to make the tube operate along different portions of its grid-voltage plate-current characteristic. This latter effect changes the slope or

curvature of a curve showing plate current plotted against grid voltage, and consequently affects linearity. Capacitance of the bypass capacitor  $C_b$  always is large, common values ranging from 20 to more than 100 mf.

Oscilloscope traces taken from anywhere along the line from oscillator plate to amplifier grid will have the general form shown at the left in Fig. 5-13. The height and steepness of the sloped portion of the trace will be varied by adjustment of the vertical size control or height control. Traces taken

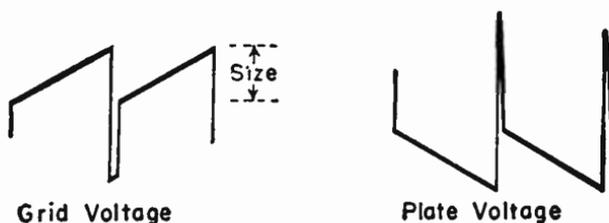


Fig. 5-13.—Oscilloscope traces from input and output sides of the vertical sweep amplifier tube.

from the amplifier plate will be of the same general form, but inverted as shown at the right. Adjustment of the vertical linearity control will make the sloped portion a practically straight line, as shown, or will make this portion curve either upward or downward.

**Horizontal Magnetic Deflection Amplifiers.** — In horizontal sweep systems for magnetic deflection are quite commonly found, in addition to the output amplifier circuits, some or all the features shown toward the right of the diagram in Fig. 5-14. Each is explained under its own heading in this book, but they will be mentioned here in relation to the amplifier circuit.

The high-voltage power supply shown at the top of the diagram, which is of the flyback type, furnishes as much as 20,000 volts for the high potential anode of the picture tube. High voltage applied to the plate of the diode rectifier is taken from the top of the primary winding on the output transformer. The tapped primary winding raises a pulse potential of about 5,000 volts to the higher voltage for the rectifier.

The damper tube suppresses oscillation which otherwise might continue in the deflection coil circuit after each pulse of retrace current. Such oscillation would continue at a frequency determined by inductance and capacitance in this circuit except for the fact that during the first cycle the damper is made conductive and places such a load on the circuit as to prevent additional cycles. Damping is not needed in circuits for vertical deflection, chiefly because of the lower operating frequency, although resistors usually are connected across the vertical coils to further reduce any tendency toward oscillation.

The width control of Fig. 5-14 is an adjustable inductor connected in parallel with a few turns of the secondary winding on the output transformer. This control alters the effective

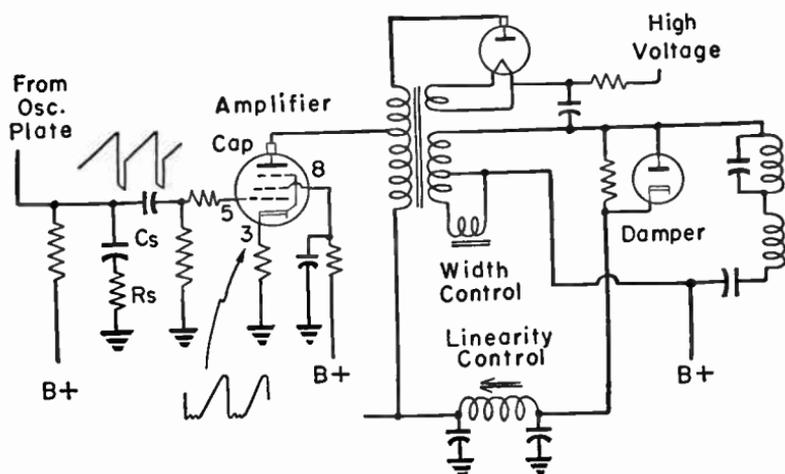


Fig. 5-14.—Typical horizontal sweep circuit for magnetic deflection.

inductance of the secondary winding, changes the strength of induced emf and deflection current, and thereby increases or decreases the extent of horizontal deflection and width of the picture. The width control here illustrated may be the only one provided for varying the horizontal size of the picture, or it may be used in addition to an adjustable resistor which varies the rate and amount of charge on the sawtooth capacitor.

The linearity control shown connected in series with the damper tube is another adjustable inductor. This inductor carries plate current for the amplifier tube. The current path may be assumed to start at B- or ground, go through the amplifier cathode to its plate, thence through the lower portion of the output transformer primary winding, the linearity control inductor, the cathode-to-plate path in the damper tube, the secondary winding of the output transformer, and to the B+ connection shown at the right. This linearity control is, in effect, a low-pass filter whose adjustment makes some change of phase in the amplifier plate voltage to alter the portion of its characteristic on which this tube operates.

All of the parts which have been mentioned are in the amplifier plate circuit or are coupled to that circuit. On the grid side of the sweep amplifier of Fig. 5-14 we find in series with sawtooth capacitor  $C_s$  a resistor  $R_s$  whose action adds a negative square pulse to the sawtooth wave going to the control grid of the amplifier. This action is the same as explained in connection with vertical amplifiers for magnetic deflection.

Oscilloscope traces taken between control grid and ground, and between cathode and ground, will be about as shown on the diagram. Without special voltage divider equipment it is not safe to try taking traces from the plate, because of the surge potentials which reach 4,000 to 5,000 volts.

Fig. 5-15 shows control grid and cathode connections for another general type of horizontal sweep amplifier circuit. Here the sawtooth capacitor  $C_s$  is connected to ground through a series fixed resistor  $R_s$ , an adjustable resistor marked *Drive Control*, and the cathode bias resistor  $R_k$ . Although the adjustable resistor usually is called a drive control, it may be called a linearity control or a peaking control. Actually it regulates the strength of the negative pulse added to the sawtooth voltage wave, by changing the time constant of the capacitor-resistor combination.

The fixed resistance at  $R_s$  may be anything between 2,000 and 10,000 ohms. Total resistance of the drive control usually is between 20,000 and 30,000 ohms. Cathode bias resistance

may be something between 50 and 150 ohms. The capacitance of the bypass,  $C_k$ , often is between 20 and 30 microfarads, although in some receivers it may be as small as 0.1 mf. The plate circuit of the amplifier, and associated parts, usually are

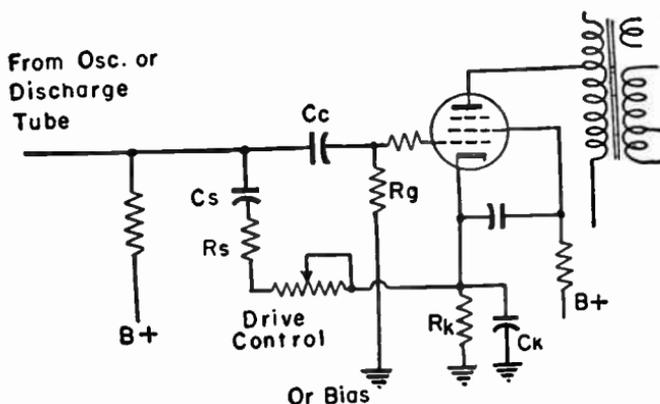


Fig. 5-15.—Sweep amplifier grid-cathode circuit with resistance type drive control.

about the same as shown by Fig. 5-14. Oscilloscope traces taken at the amplifier control grid and cathode are similar to those shown on Fig. 5-14.

In the amplifier grid-cathode circuit of Fig. 5-16 there is no means of adding a negative square pulse to the sawtooth voltage applied to the control grid. Sawtooth capacitor  $C_s$  is connected directly between the oscillator plate line and ground, being charged through the left-hand connection to B+ and discharged through the oscillator.

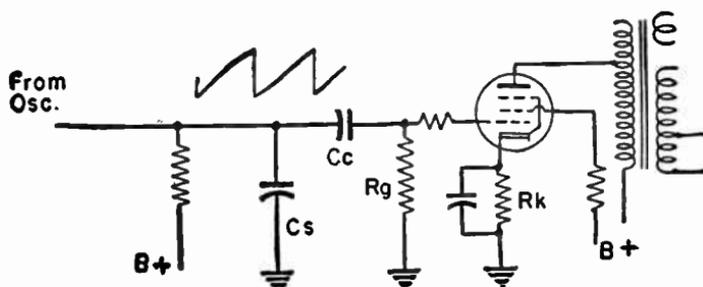


Fig. 5-16.—Sweep amplifier operated without a negative pulse on the sawtooth grid voltage.

It will be recalled that the reason for adding a negative peak to the sawtooth voltage wave is to overcome the excess of inductive reactance compared with resistance in the plate and deflection coil circuits. There is especial need for negative peaking with the low plate resistance of power triodes and other tubes connected as triodes. But beam power tubes have plate resistances in the neighborhood of 25,000 ohms, which is several times that of power triodes. By employing a beam power tube for the horizontal sweep amplifier and using the least practicable inductance in the horizontal deflection coils, the ratio of resistance to inductive reactance is considerably increased. With the tube working into a load which is more resistive than reactive, a plain sawtooth voltage applied to the control grid will produce a sawtooth current in the deflection coil circuit. This is the principle utilized for the amplifier circuit of Fig. 5-16.

In a number of receivers a negative peak is added to the sawtooth voltage for the amplifier grid by a feedback connection such as shown in Fig. 5-17. This feedback is taken through a high resistance or small capacitance from the line between output transformer secondary and deflection coils, on the side which connects to the plate of the damper tube. On this line

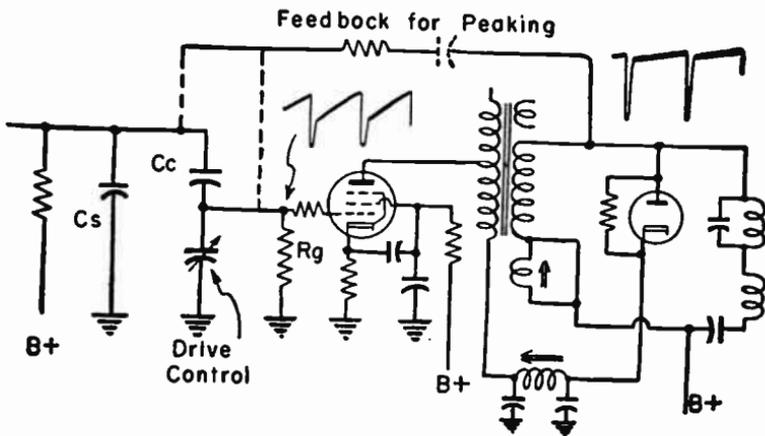


Fig. 5-17.—Feedback connection for adding a negative pulse to the sawtooth voltage for the amplifier grid.

the voltage consists of a series of sharp negative peaks. A trace taken with the oscilloscope will show peaks of the approximate form illustrated at the upper right in the figure. When a small part of the voltage from these pulses is added to the sawtooth wave developed by capacitor  $C_s$  the result is a negative peaked sawtooth of the general form illustrated above the amplifier tube. The feedback sometimes is brought to a point ahead of coupling capacitor  $C_c$  and again to a point following this capacitor.

Fig. 5-17 illustrates also a drive control consisting of an adjustable capacitor. This control capacitor and coupling capacitor  $C_c$  are in series with each other, and in parallel with sawtooth capacitor  $C_s$ . Thus the coupling capacitor and drive control capacitor form a capacitive voltage divider with voltage for the amplifier grid taken from between the capacitors. The fraction of total sawtooth voltage applied to the grid is equal to the ratio of capacitive reactance in the drive control to the sum of the reactances in the two divider capacitors. Adjusting the drive control for less capacitance increases its reactance, and increases the amplitude of voltage applied to the amplifier grid.

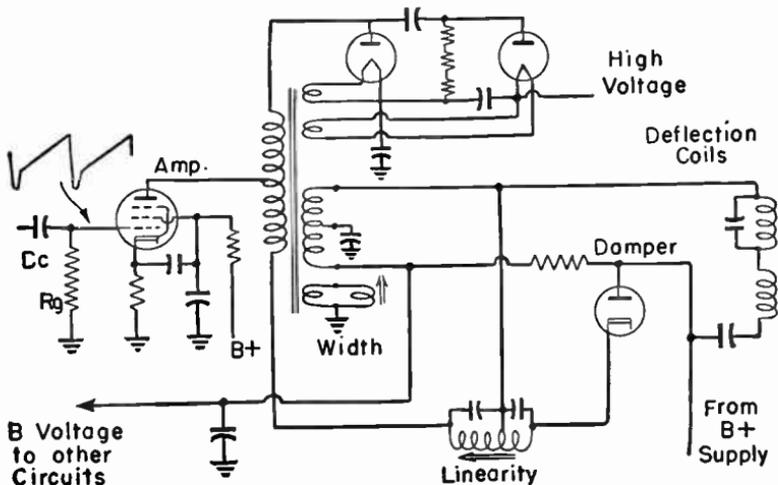
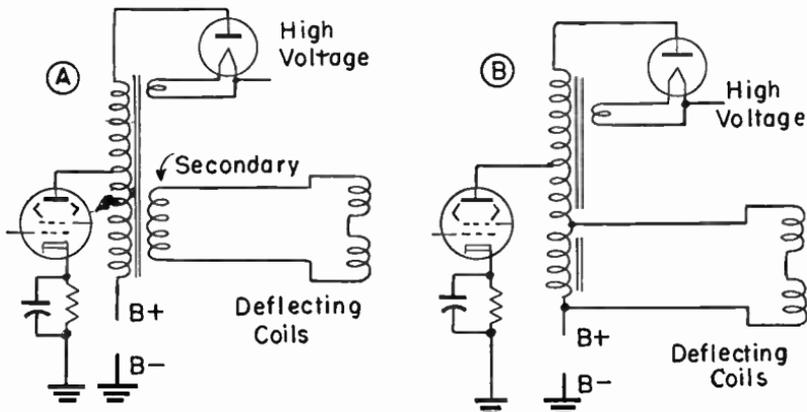


Fig. 5-18.—Modifications found in some of the circuits on the plate side of the horizontal sweep amplifier for magnetic deflection.

There are, of course, horizontal magnetic deflection amplifiers of types other than those for which diagrams have been shown. One of these other types is illustrated by Fig. 5-18. The control grid-cathode circuit of the amplifier tube is similar to types which have been described. In the plate circuit there is an output transformer with tapped primary, acting as a step-up auto-transformer for the high-voltage rectifier system. This high-voltage system for the high potential anode of the picture tube here employs the voltage doubling principle found also in other power supplies.

The width control consists of an adjustable inductor connected to one of the windings on the transformer, acting to vary the effective inductance and the extent of horizontal deflection of the picture tube beam. The horizontal linearity control is an adjustable inductor in the plate supply line to the sweep amplifier. The damper tube prevents continued oscillation in the deflection coil circuit.

There are many variations in circuit details and in connections of parts in the various sweep amplifier systems, but basic operating principles and methods of control are much the same in all of them.



**Fig. 5-19.**—Insulated winding transformer and autotransformer used in horizontal sweep sections.



plate instead of its cathode. An adjustable inductor for width control may be connected across various taps on the autotransformer, depending on circuit design.

The connection in Fig. 5-20 from damper heater to tap 4 on the autotransformer is required only where the damper

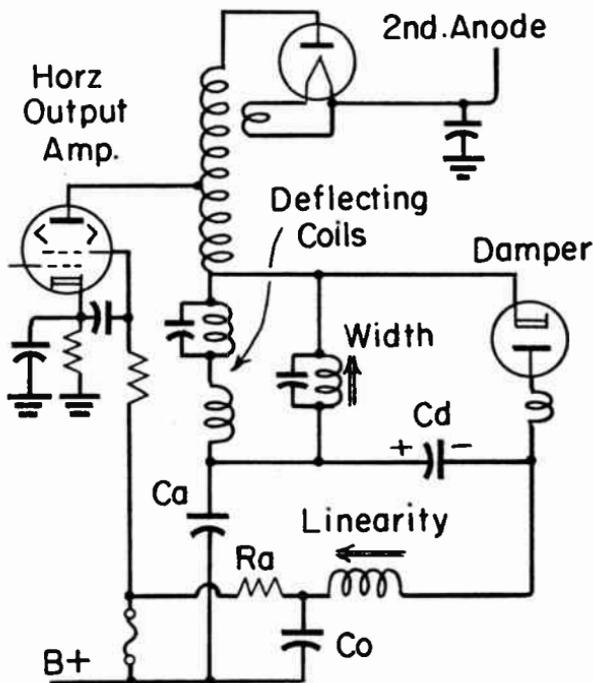


Fig. 5-21.—Deflecting coils of the yoke form a secondary for this autotransformer.

tube is not of a type which safely will withstand heater-cathode voltages much in excess of 2,000, the value of peak potential often found in horizontal yoke circuits.

In Fig. 5-21 the horizontal deflecting coils form the step-down secondary of a horizontal output autotransformer. Service diagrams usually show the yoke coils separated from the transformer, but electrically the connections are as shown here.

Autotransformers are used also for coupling vertical output amplifiers to vertical coils in deflecting yokes. In the typical circuit of Fig. 5-22 is a twin-triode blocking oscillator

and discharge tube preceding the output amplifier. The amplifier plate circuit includes the entire autotransformer winding, while the deflecting yoke is connected across only a portion. There is step-down of deflecting voltage, step-up of current, and matching of amplifier and yoke impedances.

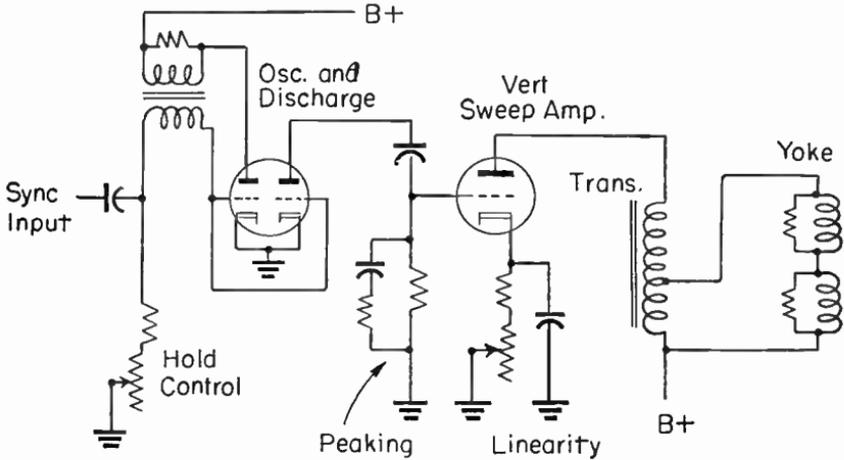


Fig. 5-22.—An autotransformer in a vertical sweep system.

**AMPLIFIERS, SYNC.** — In the sync section of the receiver, between the sync signal takeoff point and the sweep oscillators, are tubes operated as separators, clippers, limiters, and amplifiers. The chief purposes of these tubes, considered as a group, are (1) removal of picture signals from the composite signal to leave only sync pulses, (2) make all the sync pulses of uniform amplitude, and (3) bring the sync pulses to the sweep oscillators in correct polarity for triggering the oscillators. In carrying out these operations the sync pulses may be reduced below the needed amplitude. The sync amplifier tubes act to increase the pulse amplitude as may be required.

In addition to increasing the pulse amplitude, each sync amplifier inverts the polarity of the signal applied to its grid. Thus the amplifiers affect or determine the polarity of pulses reaching the sweep oscillators.

The sync amplifier tubes may be either triodes or pentodes.

They are operated with plate and screen voltages, and with control grid biases, which are normal for the particular type of tube when used as an amplifier. Other tubes in the sync section ordinarily are operated with very low plate and screen voltages, or with highly negative grid biases, to cause plate current cutoff, plate current saturation, or both, under some conditions.

Sync amplifiers may have coupling capacitor-resistor combinations acting in such manner as to decrease amplification of high frequencies, which would be picture signals, and to emphasize the low frequencies, which would be the sync pulses. The grid leak or grid rectification method may be used for all or part of the sync amplifier bias. This biasing often is designed in such a way that grid voltage becomes less negative on weak signals and more negative on strong ones. Thus the amplifier circuit tends to equalize incoming signals of varying strength, and to assist the action of sync limiter tubes.

**AMPLIFIERS, VIDEO.** — The video amplifier of a receiver consists of one or more tubes and interstage couplings between the video detector and the grid-cathode input circuit of the picture tube. The video amplifier receives composite television signals from the video detector output and delivers these signals, amplified, to the picture tube. Picture tube grid bias is such as to cut off the sync pulses while leaving the picture signals. The takeoff for pulses which go to the sync section of the receiver often is at some point along the video amplifier.

The composite signal in the video amplifier may be observed by connecting an oscilloscope to any control grid or plate between the detector output and the picture tube input. The vertical input of the scope might be connected first to the detector plate or cathode, whichever serves as output. Following connections might be to the control grid of the first video amplifier, then to the plate of this tube, then to the control grid of the second video amplifier and so on until the connection finally is made to either the control grid or the cathode of the picture tube, according to which of these elements acts as the input for this tube.

With the internal sweep of the scope synchronized for 60 cycles, 30 cycles, or other submultiple of 60, the vertical sync pulses and vertical blanking intervals will appear about as shown by Fig. 7-1. To observe a pulse and blanking interval on such a large scale as in the figure it is necessary to increase the horizontal gain of the scope while operating its horizontal centering control to bring a single blanking interval onto the screen.

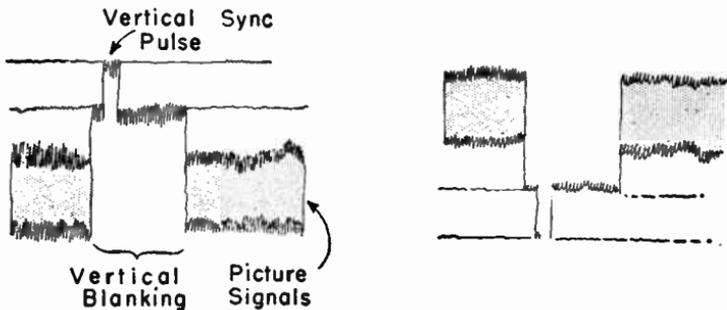


Fig. 7-1.—Vertical sync pulses and blanking intervals of the composite signal in the video amplifier.

The trace may be upright as at the left or inverted as at the right. If it is upright at the control grid of an amplifier the trace will be inverted at the plate of the same tube, and if the trace is inverted at the control grid it will be upright at the plate.

If the internal sweep of the oscilloscope is synchronized for 7,875 cycles per second, which is just half the line frequency

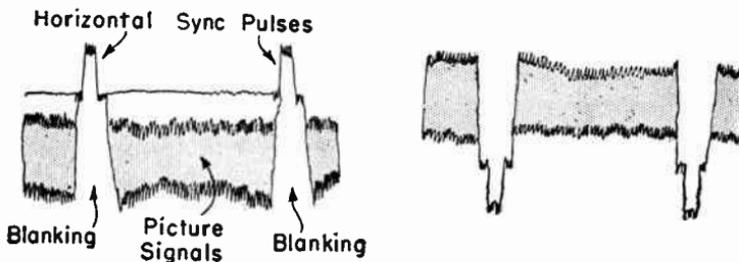


Fig. 7-2.—Horizontal sync pulses and blanking as shown by the oscilloscope.

of 15,750 cycles, it will be possible to observe the horizontal sync pulses and blanking intervals in about the form illustrated by Fig. 7-2. Again the signal may appear upright as at the left or inverted as at the right, depending on the point in the amplifier system to which the oscilloscope is connected. Because of the higher frequency it is more difficult to obtain clear traces of horizontal pulses than of vertical pulses. By connecting the oscilloscope successively to the inputs and outputs of the video amplifier tubes between the detector and picture tube it is possible to follow the composite signal all the way through this portion of the receiver. The effect of various controls is easily observed by watching for changes in the signal trace.

**Frequency response.**— Video amplifiers are resistance-capacitance coupled, with features which compensate in greater or less degree for the natural tendency of such amplifiers to drop off in gain at both very low and very high frequencies. If the gain of the amplifier decreases materially at low frequencies, any changes of picture tone or shading which occur only at relatively long intervals will not be well reproduced, instead of distinct changes there will be a tendency to merge all the shadings into a single tone or single degree of illumination. It is entirely possible for a certain shade to last throughout an entire frame of the picture, or for even more than a whole frame. This makes it necessary that the video amplifier have fairly good response at frequencies as low as 30 cycles per second.

The gain should extend with fair uniformity all the way to a high frequency limit which depends on how good the definition or picture detail is to be, and on the diameter of the picture tube or the size of the raster. With tubes of ten or twelve inch diameter it is assumed that there should be good definition of picture elements or details whose horizontal dimension is equal to the vertical distance between adjacent lines or traces. Since there are about 490 lines per frame it should be possible to resolve details whose width is as little as  $1/490$  of the height of the picture. To attain such resolution

there must be good response of the video amplifier through frequencies up to a high limit of four or more megacycles.

The larger the picture tube the greater must be the height of each horizontal trace line, and the poorer will be apparent definition as observed from distances close enough to allow distinguishing separate trace lines. On a 21-inch tube with pictures 14 inches high there will be about 35 horizontal lines per inch of height, and on a 27-inch tube with pictures 18½ inches high there will be only 25 or 26 horizontal lines per inch. Pictures appear coarse grained unless the observer is several feet from the face of the tube. With earlier 10-inch and 12-inch tubes there were 65 to 80 trace lines per inch of picture height, making for good apparent definition even when examined closely.

The video amplifier must have good high-frequency response in order to have sharply defined pictures. The low-frequency response must be good in order to reproduce slow changes in light and shade, to avoid trailers at the right-hand edges of large black objects or large white ones, and to prevent distortion of sync pulses. It is necessary also to avoid excessive differences between phase shifts at the two limits of frequency. Phase shift means a delay in the passage of signals through the amplifier. It is related to changes of voltage gain, also to time constants in the circuits. If some frequencies are delayed more than others the picture will become distorted. On top of all the other requirements it is desirable to have the highest practicable gain.

**Video Amplifier Tubes.** — The voltage amplitude of the picture signal as applied at the grid-cathode input of the picture tube must reach peak values which will produce white tones. That is, the picture tube input signal for white tones must overcome enough of the negative bias to cause maximum required intensity of the electron beam. This peak signal voltage depends on the type of picture tube and on the anode voltages applied to that tube, but in general the peak will be something between 40 and 80 volts.

The detector output voltage will depend on the maximum peak-to-peak video i-f input to the detector, which varies greatly in different receivers and with the strength of signal from the antenna. Maximum signal output voltage from the video detector usually will be at least 2 volts, but seldom more than 4 volts. To bring this 2- to 4-volt detector output up to the 40- to 80-volt input for the picture tube calls for an overall gain of 10 to 40 times. To provide this gain there may be one, two, or sometimes three video amplifier stages.

When there is only a single video amplifier it most often is one of the high-frequency broad-band pentodes designed especially for television service. When there is more than one stage the output amplifier most often is a power pentode or a beam power tube, and the preceding video amplifiers are voltage amplifying pentodes of types suited to high-frequency operation. In some receivers the video output amplifier is a triode.

The two features which are essential in video amplifier tubes are high transconductance or high mutual conductance, and small internal capacitances. The effectiveness of a video amplifier is very nearly proportional to the ratio of transconductance to its input and output capacitances, or, at least, this is the case at the high-frequency end of the band.

**Elements of the Amplifier.** — Fig. 7-3 shows fairly typical circuit connections for a two stage video amplifier. The inductors  $L_p$  in series with the plate load resistors  $R_o$ , and inductors  $L_s$  in series with coupling capacitors  $C_c$  are used only in wide-band high-frequency amplifiers such as video amplifiers. The functions of these inductors will be discussed a little later. All other elements in the circuit diagram might be found in any resistance-capacitance coupled amplifier designed to have reasonable gain at low audio frequencies.

Some of the features to be noted are as follows. There is a decoupling capacitor  $C_s$  connected from the screen to ground or B-. This capacitor bypasses variations of signal voltage which appear on the screen, and prevents interstage coupling which might otherwise occur because of the impedance of the



capacitance on the cathode will allow some degeneration at low frequencies, where the capacitive reactance becomes large, but at high frequencies this reactance becomes so small as to allow practically full gain.

**Low Frequency Compensation.** — Fig. 7-4 shows the parts of the interstage coupling which chiefly affect how low the frequency may be while still obtaining necessary gain. For any given transconductance in the tube, gain is directly proportional to impedance of the load in the plate circuit. With high resistance at  $R_g$ , and capacitance at  $C_c$  great enough to offer only small reactance, the plate load is approximately the resistance of  $R_o$  and  $R_g$  in parallel. It will be shown that, to have uniform gain over a wide band of frequencies, the resistance at  $R_o$  has to be small, only a few thousand ohms at most. Then the parallel resistance of  $R_o$  and  $R_g$  depends almost wholly on the resistance of  $R_o$ , the plate load resistor.

Whatever signal voltage is developed across load resistor  $R_o$  is applied through the reactance of coupling capacitor  $C_c$  to grid resistor  $R_g$ , and the signal voltage across the grid resistor is applied to the control grid of the second tube. The signal voltage divides between  $C_c$  and  $R_g$  proportionately to their impedances, or practically in proportion to the capacitive reactance of  $C_c$  and the resistance of  $R_g$ . The reactance of the coupling capacitor is inversely proportional to frequency. For example, the reactance of an 0.05 mf capacitor at 60 cycles is about 53,000 ohms, at 30 cycles it is about 106,000 ohms, at 20 cycles it is about 160,000 ohms, and so on.

If the resistance of  $R_g$  were 500,000 ohms the percentages of signal voltage appearing across this resistor when using the 0.05 mf coupling capacitor would be approximately 90.4 per cent at 60 cycles, 82.5 per cent at 30 cycles, and 75.9 per cent at 20 cycles. If the resistance of  $R_g$  were greater, the percentages of signal voltage across it and at the grid of the second tube would be greater.

In view of all this it is desirable to use the highest resistance at  $R_g$  that is permissible for the type of tube and the kind of grid biasing employed. The actual resistance usually is some-

thing between 0.5 and 1.0 megohm. At the same time the capacitance of the coupling capacitor should be as great as permissible, and its reactance low. Maximum capacitance usually is limited by the physical size of the capacitor, for the greater its size the larger is the capacitance to ground and the greater the bypassing effect at high frequencies. Capacitance of 0.1 mf is the usual high limit, although greater capacitances are used in some receivers.

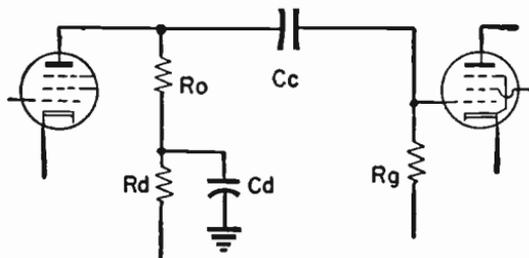


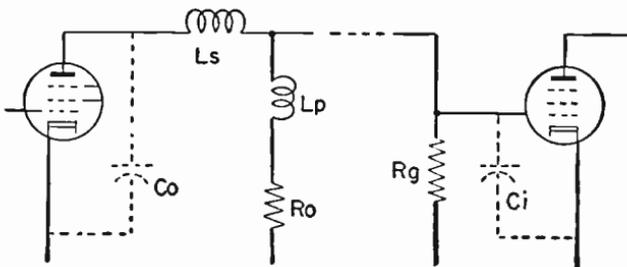
Fig. 7-4.—Parts of the video amplifier coupling which affect low-frequency response.

At frequencies for which the reactance of bypass capacitor  $C_d$  is small compared to the resistance of dropping resistor  $R_d$ , most of the a-c signal voltage in the plate circuit returns to the cathode of the first amplifier by way of the bypass. Then plate load impedance in the path of the signal includes load resistance in  $R_o$  and reactance in  $C_d$ . The reactance of capacitor  $C_d$  rises as signal frequency drops, and there is some increase of load impedance to retard the loss of gain at the lower frequencies. For this effect to be of much importance the capacitance of  $C_d$  would have to be smaller than ordinarily needed for decoupling, and the voltage dropping resistance would have to be greater than usually found in practice.

**High Frequency Compensation.** — Fig. 7-5 shows those portions of the interstage coupling which have greatest effect on how high the frequency may be at which the gain remains satisfactory. First to be noted are the capacitances  $C_o$  and  $C_i$ , which do not appear in usual circuit diagrams. At  $C_o$  is represented the output capacitance of the first amplifier tube, and at  $C_i$  is represented the input capacitance of the following tube.

Although these internal capacitances are of only a few micro-microfarads, their reactances drop to such low values at the higher frequencies as to have important effects on gain. It may be noted also that the coupling capacitor is not shown. This is because the reactance of this rather large capacitance becomes so very small at the high frequencies as to have no effect on circuit behavior or on gain.

The output capacitance of video amplifier tubes in general use ranges from 2 to 9 or 10 mmf. Input capacitances of these tubes range from about 5 to 12 or 13 mmf. There are also the capacitances of sockets and wiring, which usually are, at the



*Fig. 7-5.—Elements which chiefly affect the high-frequency response of the video amplifier.*

very least, from 3 to 5 mmf at each end of the circuit. All these capacitances are effectively in parallel with the load impedance, and are called shunting capacitances. A low value for total shunting capacitance on the plate side of the circuit might be 10 mmf, and the total on the grid side might be about 13 mmf. The total capacitance in parallel with the plate load then would be about 23 mmf if design and construction were first class in every respect.

The capacitive reactance of 23 mmf at a frequency of 4 mc is about 1,730 ohms, at 3 mc is about 2,310 ohms, and even at 2 mc still is only about 3,460 ohms. The impedance of elements in parallel always must be less than the impedance of any one of them alone. Therefore, the plate load impedance can be no greater than the capacitive reactances mentioned at the respective frequencies so long as the two shunting capacitances remain in parallel with each other.

The two shunting capacitances can be partially isolated from each other by inserting the series compensating inductor  $L_s$  anywhere in the line from plate to coupling capacitor. The inductance of  $L_s$  often is somewhere around 100 to 150 microhenrys. The inductive reactance of  $L_s$  increases with rise of frequency, while the reactances of the shunting capacitances become less. The inductive reactance of 100 microhenrys at 2 mc is 1,255 ohms, at 3 mc it is 1,885 ohms, and at 4 mc this reactance rises to 2,510 ohms. The reactance of 150 microhenrys would be just 50 per cent greater at each frequency.

With the shunting capacitances thus separated by  $L_s$  their total effective reactance is raised, and the plate load impedance is raised accordingly, for improvement of gain at the higher frequencies. The highest frequency at which the gain remains fairly good is that for which the capacitive reactance of the total effective shunting capacitance becomes equal to the impedance of the plate load. High frequency cutoff is extended by first doing everything economically possible to decrease the shunting capacitances. This is done by choosing tubes with small internal capacitances, using sockets whose insulation has a low dielectric constant, using short wires of small diameter in the plate and grid circuits, keeping these wires away from each other and from chassis metal, and by careful assembly in general.

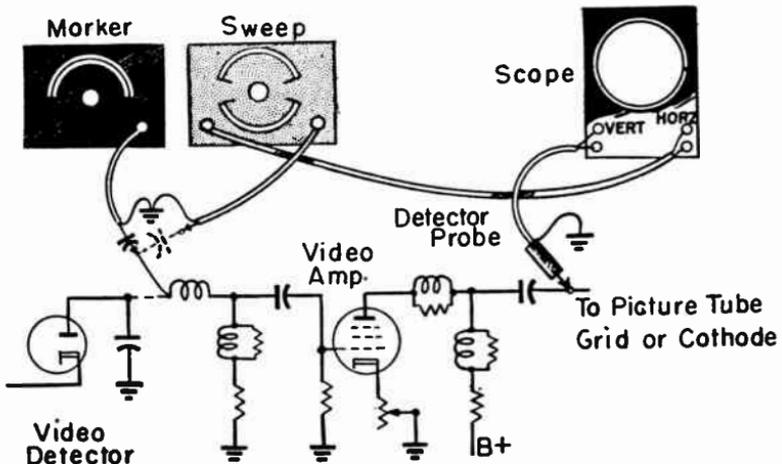


Fig. 7-6.—Instrument setup for observing frequency response of a video amplifier.

**Frequency Response Observation.** — Frequency response of the video amplifier section, from video detector to picture tube input, may be observed by using the test setup of Fig. 7-6. The sweep generator must be capable of tuning to a center frequency of two to three megacycles with sweep width of five to six megacycles. The marker generator may be any instrument such as would be suitable for i-f alignment. Synchronized sweep for the oscilloscope is desirable, but internal 60-cycle sweep may be used.

Connect the high side of the sweep generator to the video detector output through a paper capacitor of about 0.01 mf. If the video detector is a tube type, remove this tube from its socket and make the generator connection to the pin opening or lug for detector output, which may be either plate or cathode. With a crystal video detector not readily demountable, temporarily disconnect its output lead, as shown by a broken line on the diagram, and connect the sweep generator to this lead. Be sure to connect the generator ahead of any compensating inductor or peaker which is between the video detector and video amplifier.

Connect the marker generator through a capacitor as small as 10 to 20 mf to the same point as the sweep generator, or farther toward the picture tube in the video amplifier section. Use the least capacitance that allows clearly visible markers on the scope trace.

Connect the vertical input of the oscilloscope through a detector probe to the picture tube signal input. This input may be either the control grid or the cathode, depending on circuit design.

To prevent fuzzy response traces due to sync pulses, disable the horizontal sweep system by removing from its socket either the horizontal sweep oscillator or the horizontal sweep output tube. Disable the vertical sweep by removing the vertical sweep oscillator from its socket.

Set the receiver channel selector at a channel not locally allocated. Adjust the contrast control as for normal reception, and brightness at or near maximum. Adjust the sweep

generator for center frequency of two or three megacycles and sweep width of about five megacycles to begin with. Use fairly strong sweep output.

Turn on the receiver and instruments, and allow them to warm up. On the scope screen should appear a response trace

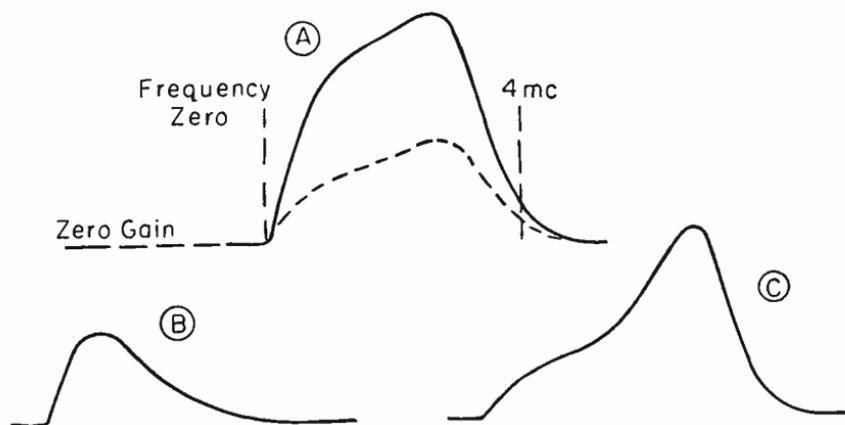


Fig. 7-7.—Frequency response curves of a video amplifier.

generally similar to those at *A* of Fig. 7-7. The full-line curve would be obtained with contrast advanced nearly to maximum, and the broken-line curve with contrast adjusted as for normal reception. Both ends of the response, where there is zero gain, may be brought onto the scope screen by adjustment of sweep center frequency and width.

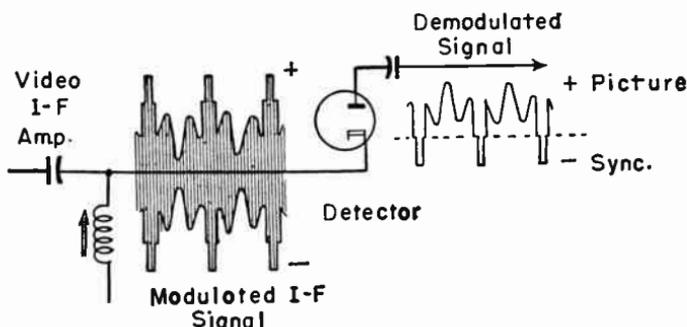
Since the total response extends over a frequency range of about four megacycles it is impossible to observe accurately any changes of gain at intervals of less than a few kilocycles, or to check relative gains at low frequencies around 50 to a few hundred cycles per second.

Effects of shunt capacitance and of high-frequency compensating inductors or peaks are shown plainly. Short circuiting a peaker which is in series with a plate load resistor, as  $L_p$  of Fig. 7-3, reduces high-frequency gain to produce a response such as at  $B$  of Fig. 7-7. Adjusting this peaker so that it resonates with shunt capacitances at a frequency of about  $3\frac{1}{4}$  megacycles produces at this frequency a high peak, as at  $C$  of Fig. 7-7. In some receivers there are adjustable peakers at one or more places, allowing a considerable degree of control over response shape and relative gains throughout the range.

Either series or shunt peakers may be adjusted, or different values may be substituted, and the results observed on the response trace. In this manner it is possible to shape the curve for fairly uniform gain. With the receiver in normal operation, video amplifier output will be affected by alignment and trap adjustments in the i-f amplifier section.

**Number of Video Amplifier Stages.** — There are definite relations between the number of stages or tubes in the video amplifier, the element from which detector output is taken, and the element of the picture tube at which there is signal input.

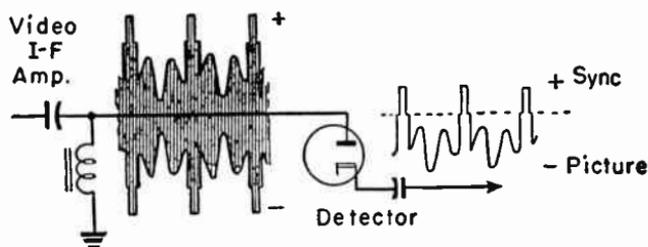
In Fig. 7-8 the modulated signal from the output of the video i-f amplifier is applied to the cathode element of the diode type video detector, and detector output is from the diode plate. For the detector to be conductive its cathode must



**Fig. 7-8.**—Picture signals are positive when the detector output is from the diode plate.

be negative with reference to its plate. With the connections shown here the cathode becomes negative only on the negative swings of the modulated signal. Consequently, only the negative side of the incoming signal envelope causes current in the detector output and this output reproduces the negative envelope. In the d-c output of the detector the picture portion of the signal is most positive and the sync pulses least positive, and when this output passes through any coupling capacitor to become an alternating voltage the picture signals are on the positive side and the sync pulses on the negative side of the wave.

In Fig. 7-9 we have the same modulated i-f signal as before, but now this signal is applied to the plate of the diode detector. For conduction to occur in the detector its plate must be made positive. The plate is made positive by the positive swings of the incoming signal, and the d-c output of the detector consists



*Fig. 7-9.—Sync pulses are positive in the output from the cathode of a diode detector.*

of the positive envelope of the i-f signal. Now the negative side of the modulated signal has been cut off by the detector, just as the positive side is cut off in Fig. 7-8. When the detector output passes through a capacitor and becomes alternating, the sync pulses are positive and the picture signals are negative.

The left-hand diagram of Fig. 7-10 shows the signal output from the video amplifier applied to the cathode of the picture tube. In order that picture variations in the signal which are to produce bright areas may cause increase of beam current and brighter traces on the picture tube screen the control grid of this tube must become more positive or less

negative with reference to the cathode. This is the same as saying that the cathode must become less positive or more negative, with reference to the grid, to have brighter traces. Then, for increasing the beam current, it is necessary that the picture signal make the cathode of the picture tube more negative. In order for this to happen, the picture side of the applied signal must be negative, and the sync pulses positive, just as shown on the left-hand diagram. For input to the picture tube cathode, the sync pulses must be positive and the picture signal negative.

At the right in Fig. 7-10 the signal from the video amplifier is applied to the control grid of the picture tube. In order that increases of picture signal amplitude may increase the beam current and brightness on the screen these increases must make the control grid more positive or less negative with reference to the cathode. For this to happen, the picture side of the applied signal must be positive, and the sync pulses negative, as shown on the right-hand diagram. For input to the picture tube grid, the picture signals must be positive and the sync pulses negative.

The detector connection shown in either Fig. 7-8 or in Fig. 7-9 may be used in a receiver where signal input is to either the control grid or the cathode of the picture tube. If the detector output is of such polarity that it must be inverted before reaching the picture tube input, a single video amplifier

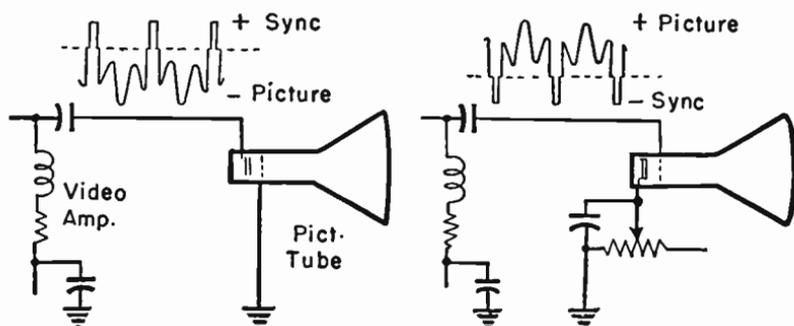


Fig. 7-10.—Polarities required in signal inputs to the cathode and to the control grid of the picture tube.

stage and tube will perform this inversion. Any odd number of video amplifiers would do the same thing, because any signal voltage applied to the control grid of a tube having a grid and a plate is inverted in the plate circuit of that tube.

If the detector output polarity is the same as the polarity required at the picture tube input this much of the problem could be handled by using no video amplifier, only a direct connection from detector to picture tube. But the detector output requires amplification. One video amplifier stage will invert the signal, and a second amplifier will bring the polarity back to its original form. Consequently, to have no inversion between detector and picture tube it is necessary to have at least two video amplifier stages. Any even number of video amplifiers would accomplish the same thing so far as maintaining signal polarity is concerned.

If output from one of the tubes in the video amplifier is from the cathode rather than from the plate of that tube there is no inversion by that one tube. Then any tube used as a cathode follower in the video amplifier system may be ignored when considering signal inversion or lack of inversion.

**Video Amplifier Troubles.** — The following are some of the more unusual faults which may be found in the video amplifier. If there is excessive peaking at some frequency it usually will show up as dark bands or smudges running horizontally across the vertical resolution wedges in a test pattern. The cause is likely to be too much inductance in a shunt compensation inductor or some construction or wiring fault which adds excessive stray capacitance, thus bringing a resonance peak down into the frequencies which should be uniformly amplified. Too much bypass capacitance across a cathode bias resistor may reduce degeneration too far and may allow enough regeneration to cause peaking. Any of these faults which might cause excessive or incorrectly placed peaking will affect the sync pulses when the sync takeoff is at some point in the video amplifier beyond where the trouble exists. The result may be white trailers or white bands on the right-hand edges of large black masses in the pictures, or sometimes

there may be fairly long white horizontal lines on the picture or pattern. Distortion of the sync pulse waveforms by the video amplifier may also make vertical lines appear crooked or skewed when they should be straight.

**AMPLIFIERS, VIDEO INTERMEDIATE-FREQUENCY.** — The video i-f amplifier of the television receiver extends from the output of the mixer tube in the tuner section to the input side of the video detector. Fig. 8-1 shows fairly typical circuit connections for one of the simpler types of video i-f amplifiers. To the control grid of the first amplifier tube come carrier and oscillator frequencies, also their sum and difference frequencies from the mixer. Tuned circuits select the difference frequencies and reject the others, just as in the i-f amplifier of a superheterodyne sound receiver. Whereas there may be only a single i-f amplifier stage in a sound receiver, the relatively low gains possible at high frequencies make it necessary to use two, three, or four such stages for television.

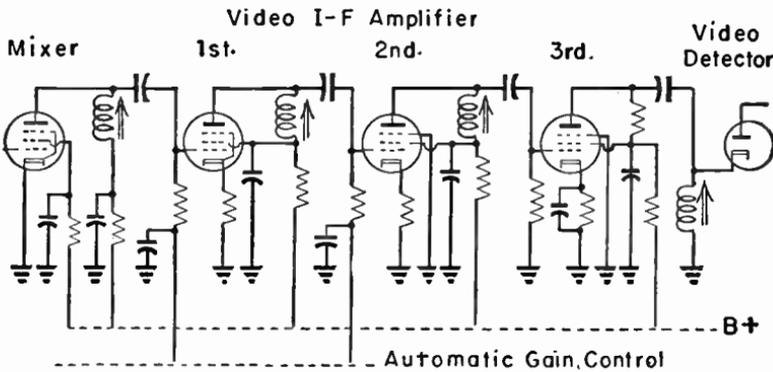


Fig. 8-1.—Connections in a three-stage video intermediate-frequency amplifier system.

Video i-f amplifier tubes are pentodes having high transconductance, usually of the miniature type. Tuning in Fig. 8-1 is by means of a single adjustable inductor in each interstage coupling. Tuning capacitance is furnished by the internal output and input capacitances of the tubes, by distributed capacitance of the inductors, and by stray capacitance

in sockets, wiring, and other parts. Each stage is tuned to resonance at some certain frequency which is close to or in between the video intermediate and sound intermediate frequencies. The tuning shown here is "staggered" at different frequencies so that the overall response of the whole amplifier is satisfactory for the band of frequencies to be handled.

In other receivers the necessary wide frequency response may be provided by transformers with two closely coupled windings, somewhat similar to i-f transformers for sound receivers. In still others, some of the couplings are with transformers giving a double-peaked response, and other couplings in the same amplifier are single tuned inductors or may be loosely coupled transformers providing a single peak in their response. In any of these designs the i-f amplifier must handle a frequency band three to four megacycles wide when it carries only picture and sync signals, or a band even wider in stages carrying the entire composite signal, including sound.

Some types of video i-f tubes have their suppressors internally connected to the cathodes. Suppressors provided with external connections or pins usually are connected directly to ground, as in Fig. 8-1, but sometimes are connected through an external resistor to the cathode. Automatic gain control commonly is provided for all except the last video i-f amplifier. Grid bias is from the agc system for some tubes and from a cathode bias resistor on each tube. Grid resistors usually are of 10,000 ohms or less. Coupling capacitors from plates to following control grids usually have values between 50 and 300 mmf.

In many video i-f amplifiers there are numerous features not shown by Fig. 8-1. These include traps for various interference frequencies and for the accompanying sound frequency, also sound takeoff connections or couplings when the receiver does not operate with intercarrier sound. In such receivers the sound takeoff may be anywhere between the mixer output coupling and the tuned circuits for the last video i-f amplifier. All stages between the mixer and the sound takeoff must carry signals for video, sync, and sound. Stages

following the sound takeoff carry only video and sync signals intentionally, with sound signals removed by means of traps.

In Fig. 8-1 the plate circuit load for the last i-f amplifier consists of a resistor between the plate and the B+ line. At the left in Fig. 8-2 this resistor is replaced by an untuned choke,  $L$ , in the plate circuit, with tuning still on the detector side of the coupling. In the right-hand diagram the tuned inductor is in the plate circuit of the last i-f amplifier, with

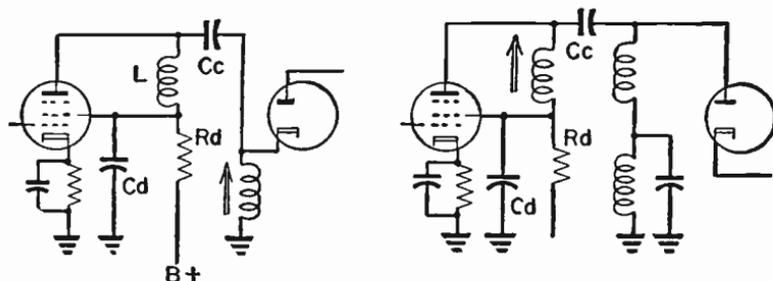


Fig. 8-2.—Tuning inductor locations in video i-f amplifiers.

an untuned choke on the detector side of the coupling capacitor. Not only in the coupling preceding the video detector, but in the coupling between any of the amplifier tubes, the tuned inductor may be on either the plate side or the grid side, either ahead of or following the coupling capacitor.

Note that the screen always is bypassed directly to ground through the capacitor marked  $Cd$  in Fig. 8-2. There is a single voltage dropping resistor,  $Rd$ , for both the screen and the plate. The bypass across this resistor carries to ground and the cathode return the high-frequency currents for both elements. If there is insufficient bypassing for the screen, high-frequency signal potentials from this element reduce signal voltage variations at the plate and drop the gain of the stage.

Parallel resonant types of traps often are found in the connection between the plate of one tube and the control grid of the following tube, in the line containing the coupling capacitor. Two such traps are illustrated by Fig. 8-3. The traps are tuned to the frequency which is to be reduced or

eliminated; they then offer high impedance at this frequency. The traps are in addition to the usual inductors or transformers which are tuned to resonance at the frequency to be amplified.

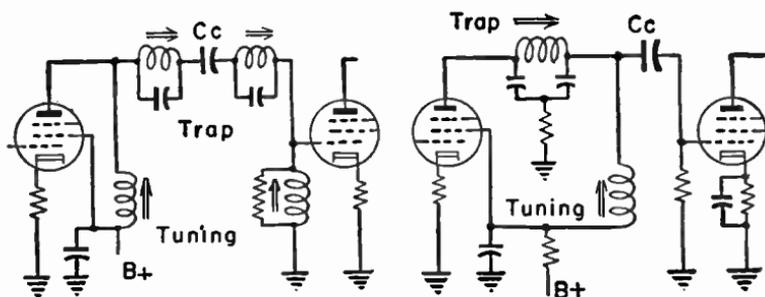


Fig. 8-3.—Parallel resonant traps in interstage couplings of video i-f amplifiers.

**Video and Sound Intermediate Frequencies.**— In the carrier frequencies for any television channel the sound carrier always is at a frequency 4.5 mc higher than the video carrier. For example, in channel 4 the video carrier is at 67.25 mc and the sound carrier is at 71.75, in channel 8 the video carrier is at 181.25 mc and the sound carrier at 185.75 mc, and so on.

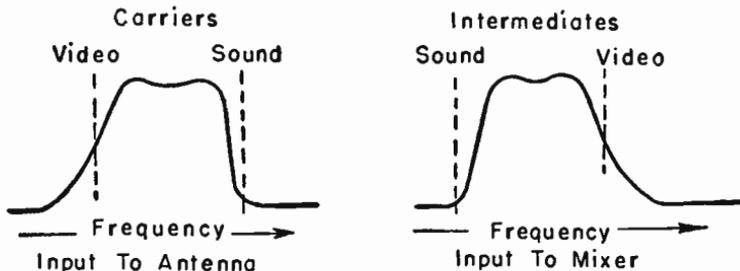
The frequency to which the r-f oscillator of the receiver is tuned for any channel is higher than the carrier frequencies in that channel. When the oscillator frequency and the video carrier frequency beat together in the mixer, the difference frequency becomes the video intermediate frequency which is amplified in the video i-f system. When the oscillator frequency and the sound carrier beat together the difference frequency becomes the sound intermediate frequency.

We may assume that a receiver tuned for channel 4, as an example, operates with an r-f oscillator frequency of 93.85 mc. The results are as follows:

<i>Oscillator frequency</i>	93.85	93.85
<i>Carrier frequencies</i>	67.25	71.75
<i>Difference (intermediate frequencies)</i>	<u>26.6</u>	<u>22.1</u>

Although the difference between the video and sound intermediate frequencies is 4.5 mc, the same as between the two carrier frequencies, the video intermediate frequency is higher than the sound intermediate frequency, while the video carrier frequency is lower than the sound carrier frequency. The same things are true in every channel; the video intermediate always is higher than the sound intermediate, and their difference always is precisely 4.5 mc.

If an oscilloscope is connected to the output of the video detector, across the detector load resistor, the frequency response represented by the trace will be shaped principally by the characteristics of the video i-f amplifier, and will be of the same general shape whether the input is from a sweep generator connected to the antenna terminals or from a generator connected or coupled to the mixer tube. With signal input to the antenna, and markers injected for video and



**Fig. 8-4.**—Where carrier frequencies and intermediate frequencies appear on the frequency response from the video detector.

sound carrier frequencies, these frequencies will lie on the response curve at about the positions shown at the left in Fig. 8-4. When signal input is to the mixer tube, with markers injected for the video and sound intermediate frequencies, these frequencies will appear on the response curve in the positions shown at the right.

The curves of Fig. 8-4 are based on the assumptions that sweep frequencies from the generator increase from left to right, and that increase of gain causes the curve to extend farther from the bottom toward the top, or that zero gain is

at the bottom and maximum gain at the top. If sweep frequencies decrease rather than increase from left to right, or if zero gain is at the top and maximum at the bottom, either of the curves may be shown with left and right sides reversed or may be shown upside down.

Fig. 8-5 shows examples of what may happen to the curve represented in its normal position at the right in Fig. 8-4. If sweep frequencies increase from right to left, with maximum gain at the top, the curve will be as at *A* in Fig. 8-5.

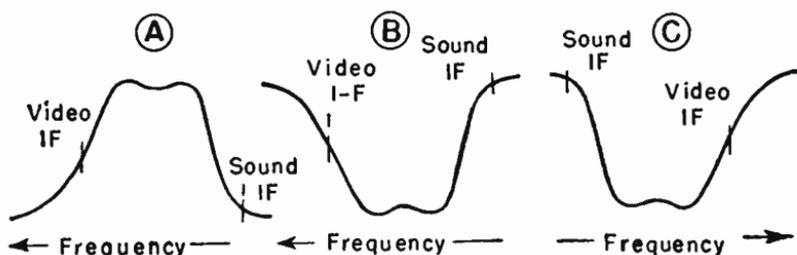


Fig. 8-5.—Inversions and reversals of frequency response curves.

With frequency the same, but maximum gain at the bottom the effect will be as at *B*. With frequency back to its first arrangement, increasing from left to right, but with maximum gain remaining at the bottom, the effect will be as at *C*. The curve, and the information it carries, are the same in every case. The differences are due solely to the characteristics of the sweep generator, the characteristics of the oscilloscope amplifiers, and the manner in which test connections are made.

The most commonly used video intermediate frequency is 45.75 mc, with which the corresponding sound intermediate frequency is 41.25 mc. Many receivers employ video intermediates of 25.75 mc up to 26.75 mc, and corresponding sound intermediates of 21.25 to 22.25 mc. Intermediate frequencies outside these ranges have been used, but in only a few cases. No matter what may be the video and sound intermediate frequencies, their difference always must be 4.5 mc, since this is the difference between video and sound carriers from which the intermediates are derived.

**Receiver Attenuation.**— With vestigial sideband transmission of television signals all frequencies from 0.75 mc below the video carrier to 0.75 mc above this carrier are transmitted in both the lower and upper sidebands. The strength of all these frequencies thus is doubled in the transmitted signal. Frequencies still lower than 0.75 mc below the video carrier are cut off. All frequencies still higher than 0.75 mc above the video carrier then are transmitted in only the upper sideband, and have only half the strength of those which are transmitted in both sidebands. The unequal strengths of various frequencies in the transmitted signal must be evened out or equalized in the receiver. The doubly transmitted frequencies are attenuated by the video i-f amplifier to make their strength equal to that of those singly transmitted. This process is known as receiver attenuation.

At the top of Fig. 8-6 is a curve representing relative strengths or amplitudes at various frequencies of the signal brought to the video i-f amplifier. This curve would corre-

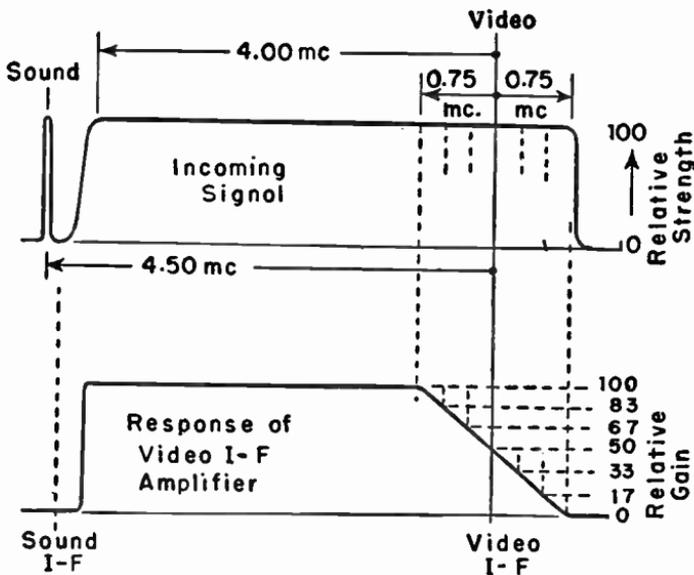


Fig. 8-6.—How the frequency response of the video i-f amplifier is shaped to equalize total amplification at all applied frequencies.

respond to the strength of the carrier signal at modulation frequencies measured both ways from the video carrier frequency as a reference point. After passing through the mixer, the video carrier frequency becomes the video intermediate frequency, so signal strengths shown by the upper curve are those measured above and below the video intermediate frequency at the input to the video i-f amplifier.

Modulation frequencies from zero up to 0.75 mc appear both above and below the video intermediate frequency. For example, with a video intermediate frequency of 26.25 mc these double-strength or doubly received modulation frequencies will be in the ranges from 25.50 mc (0.75 mc below the intermediate) up to 27.00 mc (0.75 mc above the intermediate). All greater modulation frequencies up to the limit of 4.00 mc appear on only one side of the video intermediate frequency.

In the video i-f amplifier there would be equal amplification or gain for all modulation frequencies were the amplifier response to be of the ideal form shown by the lower curve of Fig. 8-6. At the video i-f end of the response the gain is not uniform, it drops steadily from 100 at a frequency 0.75 mc below the intermediate down to zero at a frequency 0.75 mc above the intermediate, and has a value of 50, or 50 per cent of maximum, at the video intermediate frequency.

A modulation frequency of 0.25 mc will be applied to the video i-f amplifier on both sides of the intermediate frequency. On one side the gain will be 67 and on the other side the gain will be 33. The sum of the gains or the total gain at this modulation frequency then is 100 or 100 per cent. Another modulation frequency of 0.50 mc will be subjected in the video i-f amplifier to gains of 83 and 17, so that this frequency too is subjected to a total gain of 100. Every other modulation frequency between zero and 0.75 mc is similarly subjected to a total gain of 100. All greater modulation frequencies, from 0.75 through to 4.00 mc, act on the video i-f response in a region where its gain is 100. Thus all modulation frequencies receive the same total gain.

It would be theoretically possible to construct a video i-f amplifier system having a frequency response with the sharp changes illustrated at the bottom of Fig. 8-6, but it would be commercially impracticable and it is not necessary. A video i-f response of the general shape shown by Fig. 8-7 will give

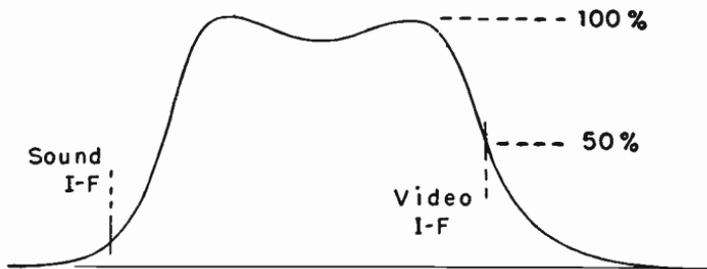


Fig. 8-7.—Typical frequency response of a video i-f amplifier system.

entirely acceptable results provided the video intermediate frequency is at or near the point of 50 per cent gain on the high-frequency side of the curve, and the sound intermediate frequency is far down on the low-frequency side.

**Positions of Intermediates on Amplifier Response.**—The shape of the frequency response curve for a video i-f amplifier is determined by adjustment of the tuned inductors or transformers in the interstage couplings. Altering any one adjustment will change the shape of the whole response curve to some extent, and usually will affect some one portion of the curve more than other portions. That is, the entire curve or some parts of it may be raised or lowered, or either side or both sides may be moved toward higher or lower frequencies. Adjustment ranges are limited, of course, but still it is possible to make great variations of response and always it is possible to produce a response of a shape satisfactory for required amplification.

We shall assume that the video i-f couplings have been adjusted or aligned to produce the frequency response shown at the top of Fig. 8-8. This response is well suited for video and sound intermediate frequencies used in an earlier example; 26.6 mc for video and 22.1 mc for sound.

Actual intermediate frequencies for video and sound depend on the frequency at which the r-f oscillator operates for any channel. This is because the intermediates are beat frequencies resulting from mixing of oscillator and carrier frequencies, and the carrier frequencies are fixed in any one

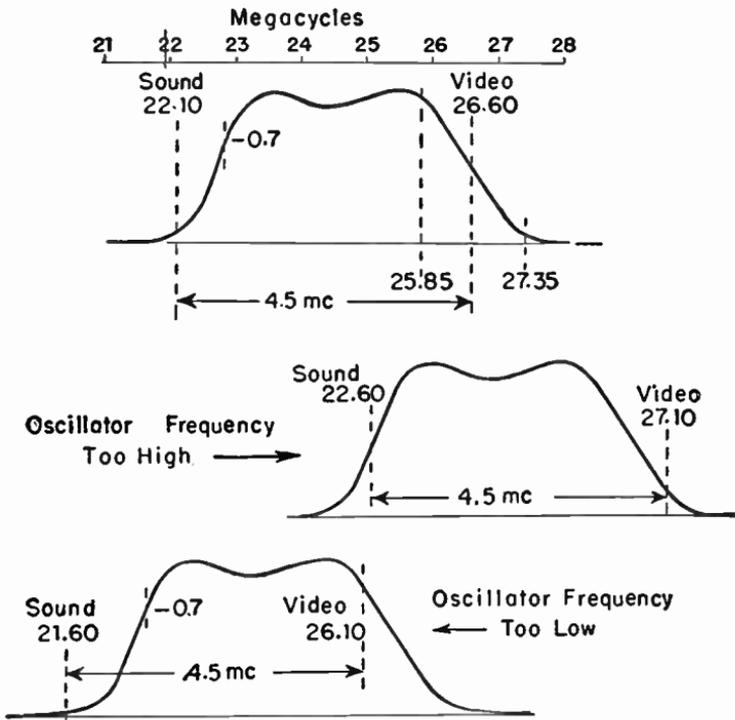


Fig. 8-8.—How video and sound intermediate frequencies are shifted on the i-f amplifier response curve by changes of r-f oscillator frequency.

channel. In the earlier example it was shown that an r-f oscillator frequency of 93.85 mc for channel 4 produces a video intermediate of 26.60 mc and a sound intermediate of 22.10 mc.

The next graph of Fig. 8-8 shows what happens when the r-f oscillator frequency is higher than it should be. For purposes of illustration the oscillator frequency is assumed to be 0.50 mc high, which makes it 94.35 mc instead of the

original 93.85 mc. Still using the carrier frequencies of channel 4, the intermediates now work out as follows.

<i>Oscillator frequency</i>	94.35	94.35
<i>Carrier frequencies</i>	<i>video</i> 67.25	<i>sound</i> 71.75
<i>Intermediate frequencies</i>	<i>video</i> $\overline{27.10}$	<i>sound</i> $\overline{22.60}$

It is apparent that the low frequencies of modulation, just above and below the video intermediate frequency, now are receiving far too little amplification—they are too far down on the gain curve. The resulting picture will be of generally dull appearance, and there are likely to be trailers or bands on the right-hand edges of any large or wide black objects. Sync pulses are of low frequency, and the lack of gain at low frequencies may make it difficult to maintain horizontal synchronization, and especially difficult to maintain vertical synchronization of the picture.

While the video intermediate has been moved too far down on the gain curve, the sound intermediate has been moved up by the excessively high oscillator frequency. Sound signals will be much too strong, and they are likely to cause horizontal bars or dark bands across the picture. These are called sound bars. If the sound section of the receiver is designed to operate from the intermediate frequency resulting from beating of carrier and oscillator frequencies, the actual sound intermediate will now have been moved far from the narrow frequency band in which the sound i-f amplifier has gain, and there will be no reproduction of sound from the loud speaker. If the receiver is designed to operate with intercarrier sound the result of the strong sound signal usually will be a loud buzz from the loud speaker, a buzz which cannot be removed by adjustment of the sound detector or demodulator circuit.

Next we shall assume that the r-f oscillator frequency has been made 0.50 mc lower than its original value, or brought down to 93.35 mc. The resulting intermediates are arrived at as follows:

<i>Oscillator frequency</i>	93.35	93.35
<i>Carrier frequencies</i>	<i>video</i> 67.25	<i>sound</i> 71.75
<i>Intermediate frequencies</i>	<i>video</i> $\overline{26.10}$	<i>sound</i> $\overline{21.60}$

These intermediate frequencies will appear on the video i-f amplifier gain curve as in the bottom graph of Fig. 8-8. Now the video intermediate is higher than originally, and the sound intermediate is much farther down. Note that the frequency response of the video i-f amplifier is not altered by these variations of oscillator frequency. The response of the amplifier is fixed by adjustments of its interstage couplings, and these are not being changed. Note also that the difference between video and sound intermediate frequencies always remains 4.5 mc. This is fixed by the difference between video and sound carrier frequencies, and never changes.

With the oscillator frequency too low the low modulation frequencies on either side of the video intermediate are receiving more than usual amplification. This may do no particular harm, in fact it may be desirable when signal strength from the antenna is very weak. There may, however, be trouble due to lack of gain at the high video frequencies or high modulation frequencies. In the top graph of Fig. 8-8 the band width from the video intermediate frequency to a point 0.7 down from maximum gain is about 3.75 mc. On the bottom graph this band width is reduced at about 3.25 mc. Frequencies farthest from the video intermediate, which are the highest modulation frequencies, are not so well amplified. This may cause poor definition or fuzzy appearance in the pictures.

With the oscillator frequency low the sound intermediate frequency is so far down on the gain curve as to be out of the band which is passed by a sound i-f amplifier designed to operate at the difference between carrier and oscillator frequencies. Then there will be no reproduction of sound. With an inter-carrier sound system the reproduction may be weak, but usually can be brought up to hearing level with the sound volume control.

Were it impossible to change the oscillator frequency to produce intermediates suited to the frequency-gain response of the video i-f amplifier, the amplifier might be realigned to make its gain curve suit the actual intermediate frequencies. With a sound system having an i-f amplifier and demodulator

designed to operate at the original sound intermediate frequency it would be necessary also to realign the sound i-f system and demodulator input for the new intermediate frequency. This might be impossible due to limited adjustment range. With intercarrier sound the sound modulation always is carried with the beat frequency of 4.5 mc which is the difference between video and sound carriers. This intercarrier beat, with its sound modulation, is not affected by changes of video and sound intermediates used in the video i-f amplifier.

**Band Width and Gain.** — It may be possible to improve reception of weak signals by increasing gain in the i-f amplifier section. Check first to see that plate and grid wires are short and direct, and have not been pushed close to and parallel with other wiring or metal. Gain may be raised by increasing B-voltage on amplifier plates and screens from the usual 125 or 135 volts to as much as 175 to 180 volts. Guard against excessive dissipation by measuring d-c voltage drops on cathode bias resistors and computing total screen and plate current from these drops and ohms of the bias resistors. Current must not exceed tube ratings.

Sharp definition in pictures may be sacrificed to obtain more gain by aligning for narrower i-f response and placing the video intermediate at 70 to 80 per cent of maximum gain instead of at the usual 50 to 60 per cent. A response such as at *A* of Fig. 8-9 might be changed to the form at *B*. Band-

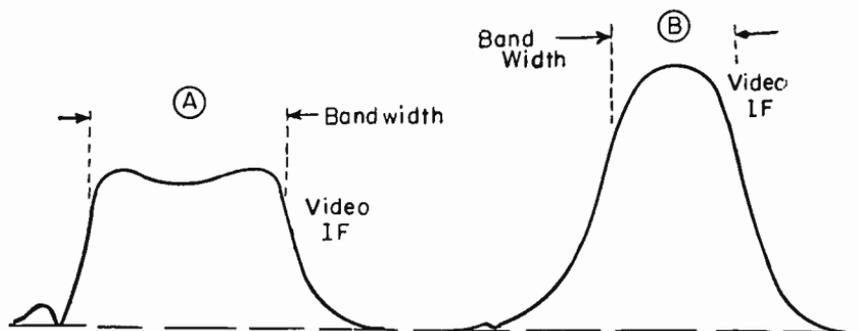


Fig. 8-9.—How i-f response curves are affected by increasing the gain.

width at *A* is about 3.75 mc at 70 per cent of maximum gain, and at *B* is less than 2.5 mc. The narrower band is accompanied by more overall gain and much stronger response at and near the video intermediate. It will be found possible to realign for high gain and a narrow band, or for a wide band and relatively low gain, but not for both at the same time.

**Stagger Tuning.** — No single interstage coupling and amplifier tube in an i-f section will provide enough gain and also a frequency response wide enough to cover the required band. To overcome this difficulty it is common practice to use two, three, or more i-f stages peaked at various frequencies between or close to the video and sound intermediates. This is called stagger tuning. Overall gain and band width meet the requirements.

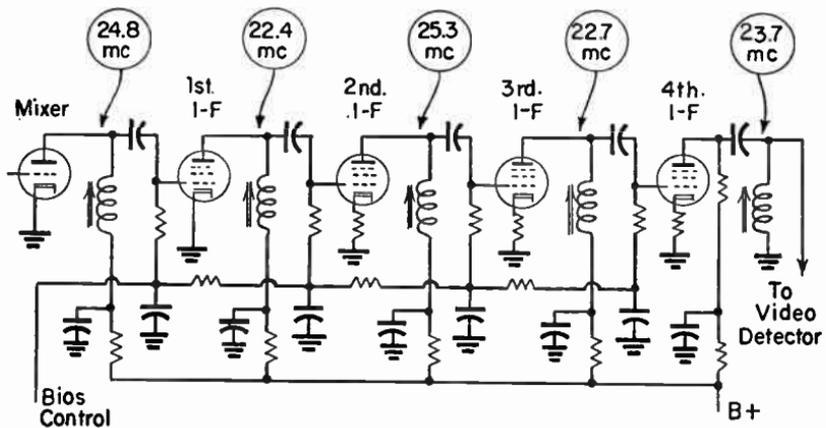


Fig. 8-10. An i-f section employing four amplifier tubes and five couplings tuned to different frequencies.

Fig. 8-10 shows circuit connections and peaking frequencies for a four-tube stagger tuned i-f amplifier operating with video intermediate of 26.1 mc and sound intermediate of 21.6 mc.

Each interstage coupling is peaked at a frequency between the two intermediates. In other cases one of the couplings may be peaked slightly outside this range.

In many receivers having four i-f couplings two are peaked at the same frequency and the other two at another frequency. Individual responses for stages in one such system are shown

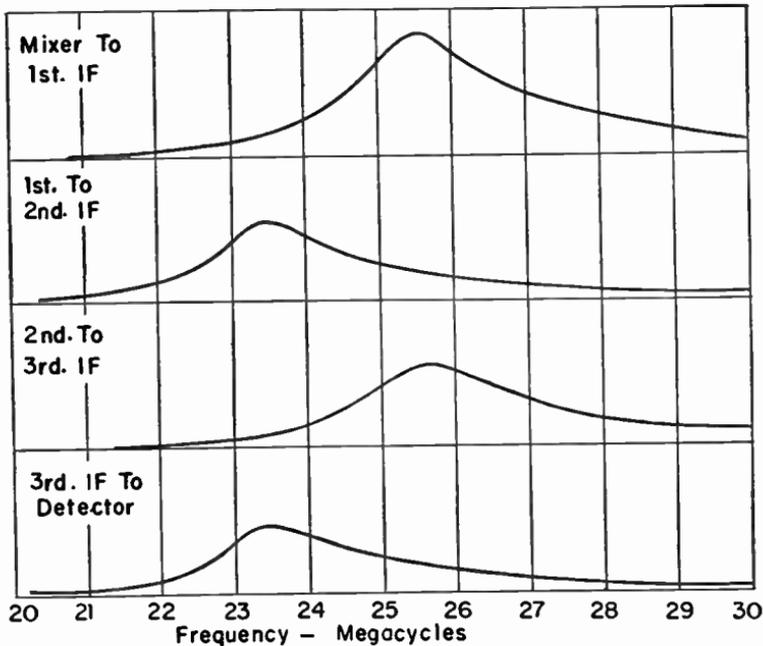


Fig. 8-11.—Individual frequency responses of four couplers used in a stagger tuned video i-f amplifier.

by Fig. 8-11. Reading from top to bottom the peak frequencies are 25.6 mc, 23.4 mc, 25.6 mc, and 23.4 mc.

The output of one i-f stage is the input for a following stage. Stage gains at each frequency multiply together. The overall result is shown by Fig. 8-12. This response is suited to a video intermediate of 25.7 mc and a sound intermediate of 21.2 mc. The shape of an overall frequency response is influenced largely by peak frequencies in the several stages, also by shapes of individual responses or by their amplifications at each frequency.

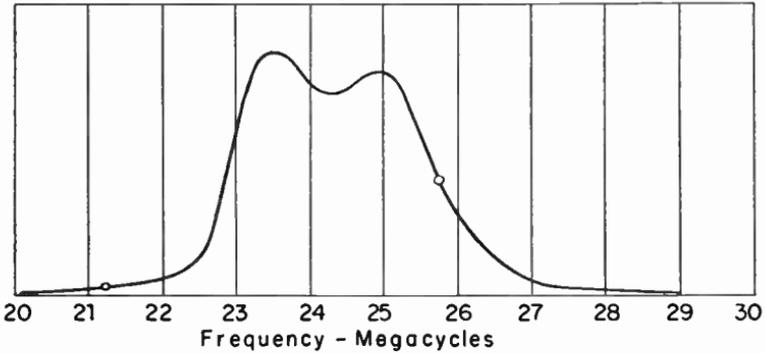


Fig. 8-12.—Overall frequency response of the four stagger tuned couplers.

Although i-f interstage couplings have been shown as simple tuned impedances in preceding diagrams, the principles of obtaining broad-band high-gain response by stagger tuning apply to all methods of coupling. Couplings from mixer to first i-f amplifier often differ from others in the section, as in the examples of Fig. 8-13. In diagram 1 inductor  $L_a$  is untuned

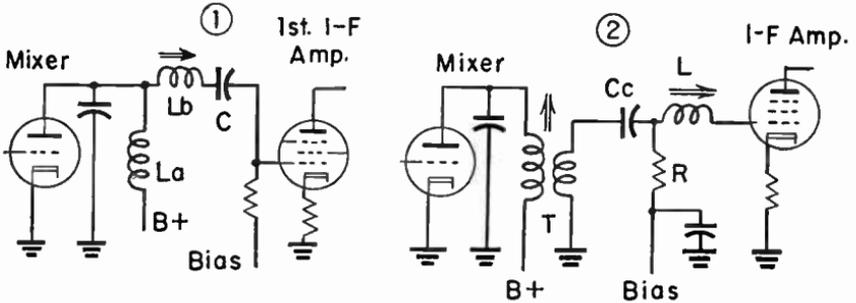


Fig. 8-13.—Couplings from mixers to i-f amplifiers.

but has high impedance at all frequencies within the i-f pass-band. Adjustable inductor  $L_b$  and capacitor  $C$  are made series resonant for peak response at a frequency near the video intermediate. In diagram 2 a tuned transformer trans-

fers signals from mixer to grid circuit of the first i-f amplifier. In the grid circuit is adjustable inductor  $L$  tuned to series resonance with capacitor  $Cc$ .

Close coupling and large signal transfer often are obtained by using bifilar interstage transformers. In a bifilar transformer the primary and secondary are insulated from each other, but their conductors are wound together, turn for turn, with wires for the two windings alongside each other throughout their length. The bifilar transformer is tuned by a single adjustable core within the winding form.

Although coupling in a bifilar transformer is close, it is not close enough to cause a flat-topped or a double peaked response, it is not close enough for overcoupling.

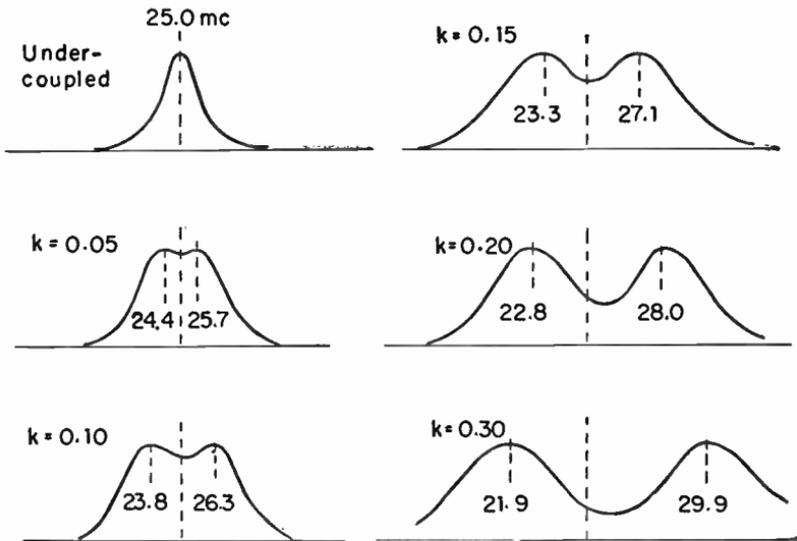


Fig. 8-14.—Effect on separation of resonance peaks when coupling is increased.

**Overcoupled Transformers.** — A transformer is said to be overcoupled when the coupling factor, whose symbol is  $k$ , is great enough or the coupling is close enough to cause resonance peaks at two frequencies. A transformer is undercoupled when the factor is small enough and coupling loose enough that only a single peak occurs. Fig. 8-14 shows what happens as the coupling factor is increased, beginning with undercoupling and a single peak at 25.0 mc.

Overcoupling is difficult or impossible with inductive coupling in an air-core transformer, as represented at the left in

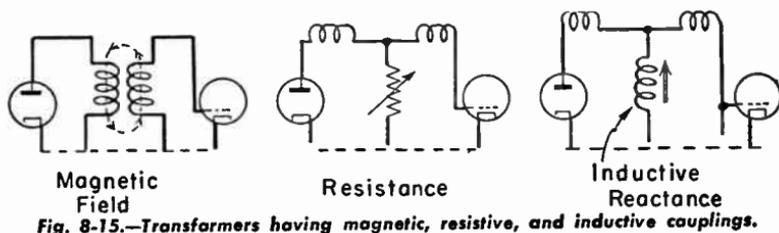
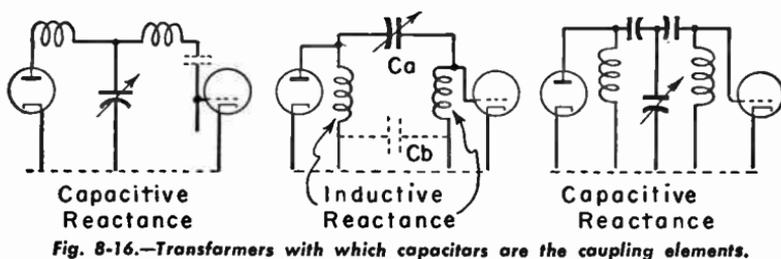


Fig. 8-15, but may be obtained when there is a resistance or an inductive reactance common to the two circuits which are to be coupled. Overcoupling in i-f amplifier circuits sometimes is secured as at the left in Fig. 8-16, where an adjustable



capacitor and its reactance are common to both circuits. Less capacitance and greater resulting reactance increase the coupling. In the center diagram, capacitor  $C_a$  determines the degree of coupling when capacitive reactance is enough to provide considerable opposition at intermediate frequencies. A capacitor at  $C_a$ , between plate and grid, is said to provide top coupling, while one at  $C_b$  provides what is called bottom coupling. At the right in Fig. 8-16 the adjustable capacitor is common to both the plate circuit and the grid circuit. With other elements of suitable values this capacitor may be adjusted for either undercoupling or overcoupling.

Combined inductive and capacitive couplings as used in the final stages of i-f amplifier sections are shown by Fig. 8-17.

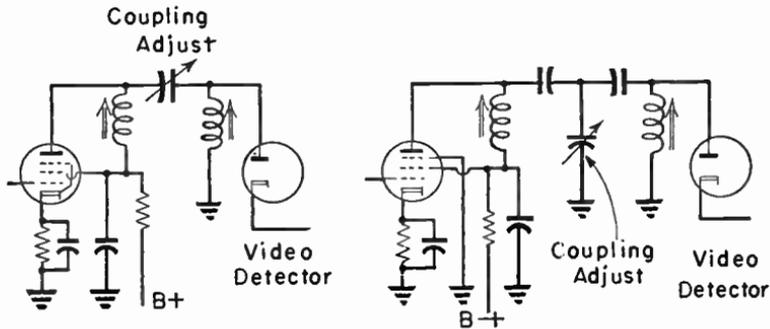


Fig. 8-17.—Adjustable capacitors used for varying the coupling and the band width at the output of video i-f amplifiers.

In the left-hand circuit the coupling is increased by increasing the adjustable coupling capacitance, while at the right the coupling is increased by decreasing the adjustable coupling capacitance and thereby increasing its reactance.

Fig. 8-18 illustrates frequency responses of single over-coupled stages. Video intermediate frequency is considered as zero on the graph, with higher and lower frequencies on lines at 1-mc intervals. The sound intermediate is, of course, 4.5 mc below the video intermediate. Couplings providing double peaks, as shown, often are used for the first or last i-f stage, with undercoupled single-peaking couplings in other stages for securing the desired overall response.

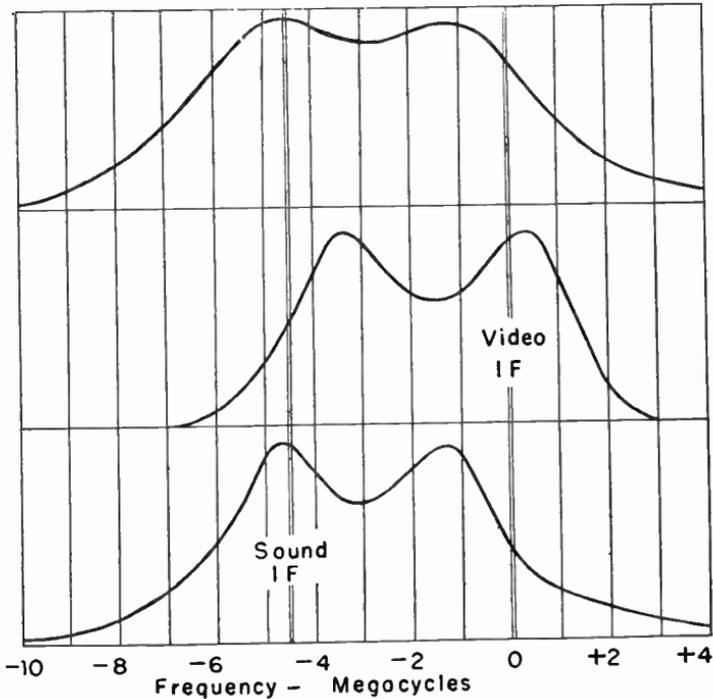
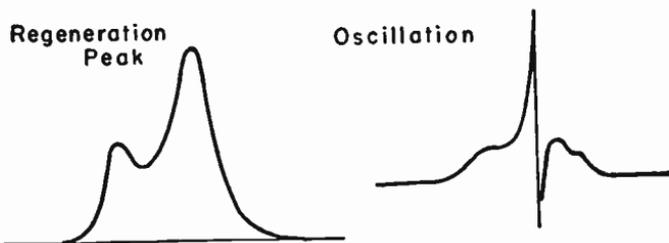


Fig. 8-18.—Typical response curves from stages having overcoupled transformers.

**Regeneration and Oscillation.** — Sometimes there is feedback from plate circuit to grid circuit of one tube, or from one stage to a preceding stage when both are tuned to the same frequency. The result may be to twist and skew lines which should be straight in pictures, or to partially cover the picture area with short, black, horizontal streaks, or to otherwise distort the reproduction. When such troubles are due to regeneration they often may be made to appear and then disappear by varying the setting of the contrast control, and sometimes by varying the setting of the fine tuning control which acts on the r-f oscillator.

While observing the overall response of a video i-f amplifier with a sweep generator and oscilloscope, regeneration will show up as an excessively high peak, on the order of the one at the left in Fig. 8-19. If regeneration becomes so great as to cause oscillation the oscilloscope trace will take on the appear-



*Fig. 8-19.—How regeneration and oscillation affect oscilloscope traces of response curves.*

ance shown at the right. There is violent oscillation where frequencies from the sweep generator pass through the one at which the trouble occurs.

If video i-f amplifier output is being observed with a voltmeter connected to the video detector load, oscillation will make this output voltage increase to a very high value and remain there when the input signal is at a single frequency, not at swept frequencies. Reducing the input to zero will not affect the output voltage, because the oscillation is self-sustaining. If a picture or test pattern is being observed when oscillation occurs, the screen of the picture tube will become brilliantly white all over. The quickest way to stop oscillation is to turn off the power to the receiver.

Regeneration and oscillation may be the result of tuning adjacent i-f amplifier stages to the same or nearly the same frequency. These troubles may be caused also by excessively long leads in plate and grid circuits, by these leads being too close to each other, by chokes or tuning inductors being close enough or in such relative positions as to allow feedback, or

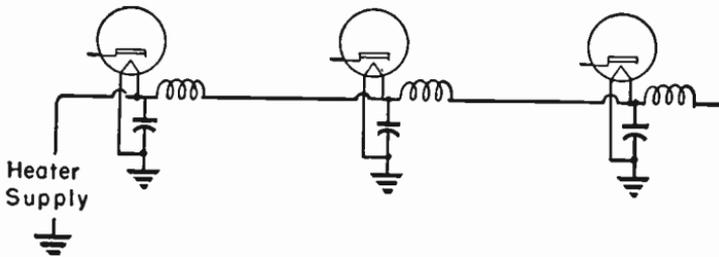


Fig. 8-20.—Decoupling capacitors and chokes in a heater circuit.

by insufficient decoupling capacitance from the bottom of the plate load to ground or from the screen to ground. These bypass capacitors must be large enough to prevent coupling through the voltage dropping resistors or through parts of the common B-power supply.

In video i-f amplifiers the tube heaters often are decoupled from one another by means of capacitors to ground, or series chokes, or both. Fig. 8-20 shows a heater circuit having one end of the heater in each tube connected to ground, which is one side of the a-c heater supply. The other side of the heater circuit is carried through insulated wire. From the end of each heater connected to this insulated line there is a bypass capacitor to ground, and between each pair of adjacent heaters there is an r-f choke. The bypasses and chokes, or either used alone, help prevent feedbacks from one stage to another and reduce tendency toward regeneration or possible oscillation.

Response curves often are badly distorted by connection of signal generator leads and oscilloscope or voltmeter leads to the receiver. The trouble is likely to be an input signal voltage which is too strong. There may be too much coupling from generator leads to tube grids. If this coupling is reduced by using a smaller capacitor in series with the lead, or by connecting a 300- or 400-ohm resistor from the grid to ground, the regeneration peaks or the oscillation often will disappear while allowing enough signal input to produce a good readable trace.

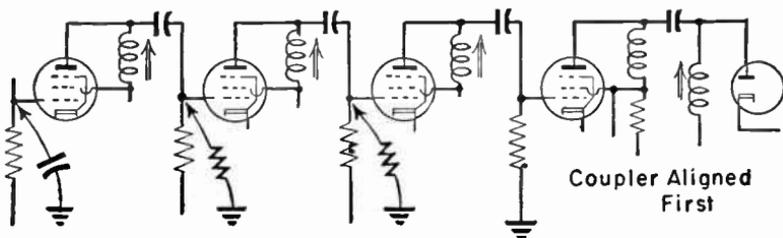


Fig. 8-21.—Grounding of control grids through capacitors or resistors to prevent oscillation during alignment.

The leads from generators and oscilloscope always should be of shielded cable. They should be no longer than necessary. The lengths exposed beyond the shielding should be as short as possible. Feedback from oscilloscope leads sometimes may be stopped by connecting a capacitor of about 0.002 mf from the vertical input lead to ground at the receiver end. It may be

possible to find certain positions of the contrast control where the trouble disappears.

If regeneration or oscillation cannot be otherwise prevented during alignment, the method of Fig. 8-21 may be used. The control grids of all tubes preceding the one for which the coupler is to be aligned are connected to ground through a capacitor or resistor. A capacitor, which should be of about 0.001 mf size, is shown on the first tube of the diagram. On the second and third tubes are shown resistors, which should be carbon types of about 300 ohms resistance. All grids are grounded through either capacitors or resistors, both being shown in the diagram only to illustrate their connections.

After aligning the coupler nearest the video detector, the grounding resistor or capacitor is removed from the next preceding tube on which it has been used, and the coupler second from the detector is aligned. The grounding elements are removed successively as alignment proceeds from stage to stage. Should regeneration or oscillation reappear after all stages are aligned, the trouble is due to some of the other causes mentioned earlier.

**Alignment of Video I-F Amplifiers.** — The subject of alignment as it applies to all parts of the television receiver is discussed in the article on *Alignment*, to which reference should be made. The present discussion relates specifically to alignment of video i-f amplifiers.

The shapes of the required overall response and of stage responses are affected by the type of sound system employed. With dual carrier systems the sound intermediate frequency is the beat frequency resulting from mixing of the sound carrier and r-f oscillator frequencies, and it will vary with change of oscillator frequency. The other general class of receivers employs intercarrier sound, where the center frequency for the sound system results from beating of sound and video intermediate frequencies in the video detector. Since these two frequencies always differ by 4.5 mc, the intercarrier sound frequency always is 4.5 mc, and is not affected by change of r-f oscillator frequency.

In receivers employing the sound intermediate frequency there is first the r-f response shown at the left in Fig. 8-22. This is changed by the r-f oscillator and mixer to a response of the general shape shown at the center. The sound takeoff may

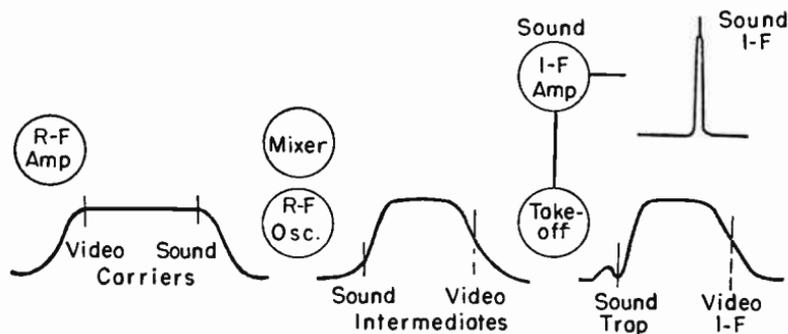


Fig. 8-22.—Frequency responses when the sound system operates at the sound intermediate frequency.

follow one or more i-f stages or may be located at the mixer output. The sound i-f signal goes from the takeoff to the sound i-f amplifier. In video i-f stages beyond the takeoff this sound i-f signal is trapped out, as shown.

The response of the sound i-f amplifier is centered at the sound intermediate frequency. This response is very narrow, seldom being more than 0.3 to 0.4 mc at points 50 per cent down on the curve. Consequently, the sound intermediate frequency, as fixed by r-f oscillator action, must be held within close limits in order to fall on this narrow response. The video intermediate frequency must, of course, be exactly 4.5 mc higher than the sound intermediate frequency. Since the sound intermediate must be exact, and the video intermediate must have an exact relation to sound, both frequencies are rigidly fixed in value. The shape of the video i-f response must be shaped to suit these intermediate frequencies. Even a small change of r-f oscillator frequency will prevent reproduction of sound, while having only a minor effect on the picture. With a correctly shaped video i-f response there will be the best picture when there is the best sound from the loud speaker.

What happens with an intercarrier sound system is shown by Fig. 8-23. There is the same r-f response as before. Both the sound and the video intermediate frequencies must pass through all video i-f stages and both must reach the video detector where they beat together to form a 4.5-mc center frequency, modulated with the sound signal. This modulated

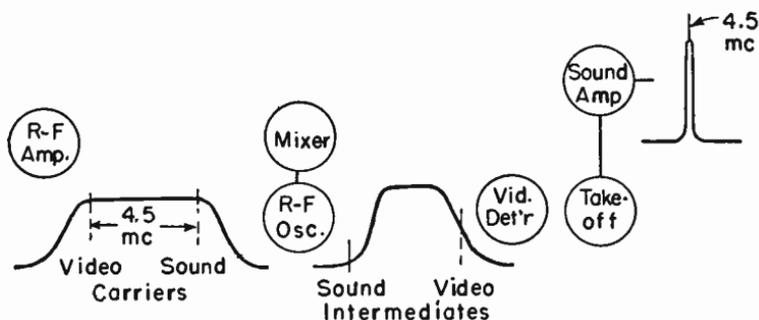


Fig. 8-23.—Frequency responses with on intercarrier sound system.

4.5-mc frequency is taken off at some point following the video detector and goes to the sound amplifier. The sound system has a very narrow response which is centered at 4.5 mc. This system always will receive a sound-modulated 4.5-mc signal unless either the sound or the video intermediate were to be completely dropped out in the video i-f amplifier.

The video i-f response for a receiver employing intercarrier sound must be so shaped that, at a frequency 4.5 mc lower than the 50 per cent point on the high-frequency slope, the response is down to something around 3 to 10 per cent with most designs. Now, provided there is sufficient range of adjustment, the frequency of the r-f oscillator may be brought to a value producing video and sound intermediate frequencies which fall on the specified points of the video i-f response. That is, shifting of the r-f oscillator frequency shifts both frequencies which result from beating with the two carrier frequencies, and these two frequencies always must remain 4.5 mc apart. At the output of the video detector there always will be a 4.5-mc modulated sound signal for any setting of the r-f oscil-

lator that does not move the sound intermediate frequency to zero on the video i-f response.

When a sweep generator is used for the signal source during alignment or test, and an oscilloscope is used as the output indicator, the presence or absence of traps in the video i-f amplifier will affect the shape of the responses. This is illustrated by Fig. 8-24, which shows successive video i-f stages.

On the coupler between first and second tubes there is no trap of any kind. A response taken at the input to the second tube would be of the general shape shown directly below, with sides or skirts sloping smoothly down to zero.

On the next interstage coupler is an accompanying sound trap, tuned to the frequency of the sound intermediate for the signal being received. As shown by the curve below this trap, the response at the accompanying sound frequency is reduced to zero or nearly so. The sound takeoff would precede this trap,

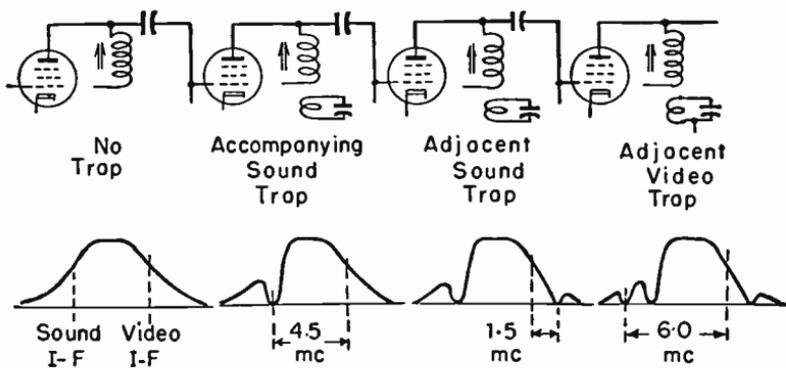


Fig. 8-24.—How various traps affect frequency responses in the video i-f amplifier.

for the sound signal will be very low or absent in following stages. The dip which appears in the response, for a correctly tuned accompanying sound trap, will be at a frequency 4.5 mc lower than the video intermediate frequency.

On the third coupler is an adjacent sound trap. The adjacent sound frequency is 6.0 mc higher than the accompanying sound, and 1.5 mc higher than the video intermediate

frequency. In the response will be a second dip at this adjacent sound frequency. The dip due to the preceding trap for accompanying sound will remain.

On the fourth coupler is a trap for the adjacent video frequency, which is 6.0 mc lower than the video intermediate frequency carrying the received signal. A third dip, corresponding to this adjacent video trap, will appear on the response which is the input for a following stage or the video detector. The other two dips still will appear.

Dips will appear in a video i-f response curve at the frequencies of any traps which are between the point of signal generator input and the point at which the output is observed. Just what dips appear on any particular response depends on the location of traps in the amplifier and the frequencies to which the traps are tuned.

It is recommended by some manufacturers that all traps be aligned before any of the interstage couplers are aligned. Others advise tuning the coupler and any trap on the same stage before proceeding to another stage or another coupler and trap. After aligning either a coupler or a trap in any one stage, the other of these two elements should be checked again for frequency response or effect. There is some change in the resonant frequency of any circuit coupled to another circuit when tuning of that other circuit is altered.

When aligning all the stages of a video i-f amplifier the test signal may be passed through one coupler or through more than one coupler between the generator and the output indicator. One method is illustrated by Fig. 8-25. The output indicator (voltmeter or oscilloscope) is connected across the load resistor in the output of the video detector. The signal generator output is coupled to the control grid of the video i-f amplifier preceding the detector, at 1, and alignment is made of the coupler between this amplifier and the detector. Then the generator input is moved back to the grid of the preceding amplifier, at 2, while aligning the coupler which precedes the one first adjusted. Next, with generator output moved ahead one more stage, 3, alignment is carried out on the next pre-

ceding coupler. Thus the work proceeds until the generator is coupled to the control grid of the mixer tube, 4, for alignment of the final coupler.

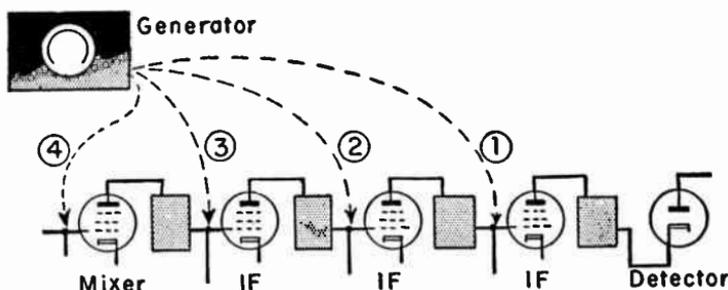


Fig. 8-25.—Connections of signal generator and output for video i-f alignment.

The method just described is commonly employed when interstage couplings are by means of overcoupled transformers, and for any type of coupling when one or more stages are thought to be far out of adjustment to begin with.

If it is probable that none of the stages are very far out of alignment, the signal generator may be coupled to the control grid of the mixer tube and kept there while aligning all couplers, with the output indicator remaining across the video detector load. This is common practice also for alignment of stagger tuned stages, where each stage is peaked at some one frequency.

The operation and use of sweep generators, single-frequency generators, marker generators, oscilloscopes, and electronic voltmeters are explained in articles devoted to these several instruments. Following are preliminaries to alignment of any video i-f amplifier.

a. Disconnect the antenna from the receiver, remove the transmission line from the antenna terminals on the receiver.

b. Set the channel selector for some high-frequency channel on which there is no transmission at the time of alignment.

c. Set the contrast control in its usual operating position, or about one-fourth down from maximum.

d. Override the automatic gain control voltage with a fixed bias voltage from dry cells, with battery positive to ground and negative to the agc bus. Use about 3 volts, or, if this lowers the response too much, try  $1\frac{1}{2}$  volts.

e. Disable the r-f oscillator. The entire signal will be furnished by the generator, and when this signal goes to the mixer control grid the results are confusing if the local oscillator remains active. If tube heaters are in parallel the oscillator tube may be removed from its socket during alignment. With series heaters, connect the control grid of the oscillator tube or section to ground through a capacitor of about 0.001 mf. This latter method may be used also when heaters are in parallel.

f. Allow at least 20 minutes for the receiver and instruments to warm up after all of them are turned on, then proceed to make adjustments.

**Alignment of Stagger Tuned I-F Stages.** — The only way of knowing the frequencies to which individual stagger tuned stages are supposed to be peaked is to have the manufacturer's service instructions or some reference book giving such information for all receivers.

In receivers having three video i-f amplifier tubes, and four interstage couplers between tubes, it is fairly common practice to tune two of these couplers to one frequency, and to tune the other two couplers to a different frequency—employing only the two peak frequencies for all four couplers. Such an arrangement is represented at the top of Fig. 8-26. The values of peak frequencies shown here are with reference to the video intermediate frequency, since they refer to no particular receiver, and might be used with any video intermediate. Furthermore, these peak frequencies are merely representative of what might be called average practice. In a given receiver, the lower of these two frequencies might be lower or higher than 2.6 mc below the video intermediate, and the higher peaks would not necessarily be 0.3 mc below the video intermediate.

In other receivers having four video i-f couplers for three amplifier tubes, each coupler is tuned to a frequency different from any of the others. Such practice is represented at the bottom of Fig. 8-26. Again the peak frequencies are with reference to the video intermediate frequency, and again they show only

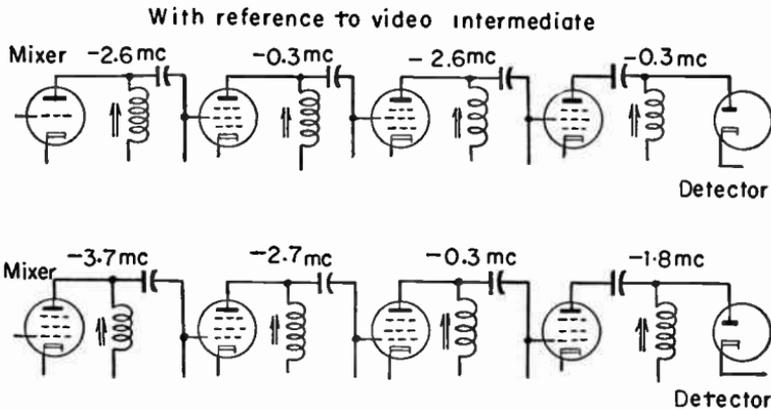


Fig. 8-26.—Typical peaking frequencies for four couplers in stagger tuned video i-f amplifiers.

average values. It is not necessary that the order of higher and lower frequencies be as shown by the diagram. Instead of the order used here, the peaks might be arranged as  $-3.7$  mc, then  $-1.8$  mc, then  $-2.7$  mc, and finally  $-0.3$  mc, or in some other order.

With four video i-f amplifier tubes, and five couplers, there is even less uniformity than with three tubes and four couplers. Sometimes two of the five couplers will be peaked at the same frequency, but more often there are five different frequencies. Fig. 8-27 shows some possible peak frequencies and orders in

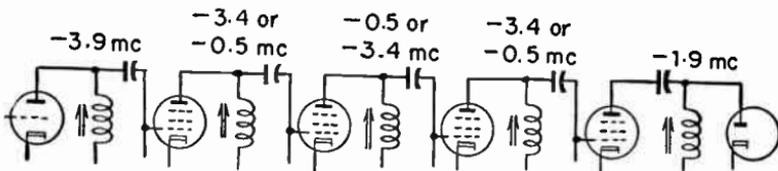


Fig. 8-27.—Typical peaking frequencies used with five couplers in stagger tuned video i-f amplifiers.

which they may follow one another. The first coupler is peaked somewhere around 3.9 mc below the video intermediate in a rather large number of receivers. Of the next three couplers, two sometimes are peaked up near the video intermediate with one peaked farther down toward the sound intermediate, or only one may be peaked up high and two of them lower down. The peak frequency for the last coupler often is around 1.9 mc below the video intermediate, or something between minus 1.4 mc and minus 2.4 mc.

Certain frequencies have been mentioned for peaking video i-f amplifiers only for use in an emergency when everything is completely out of adjustment, and when something must be tried in order to get the receiver on the road to recovery. These frequencies should not be used when there is any chance of employing those known to be correct. Some couplers have a range of frequency adjustment so limited that only those frequencies which are correct or nearly so may be tuned.

The usual method of aligning a stagger-tuned video i-f amplifier is as follows. Use a signal generator which furnishes a single steady frequency, not a sweep generator. This generator is used without modulation. Couple the generator output through a small capacitance to the mixer control grid, or connect without a capacitor to a metal ring or to an ungrounded tube shield placed around the mixer tube. When using a coupling capacitor the generator output may be applied to any point which is above ground potential in the mixer grid circuit.

Connect an electronic voltmeter across the video detector load resistor. Use the instrument as a d-c voltmeter. If the detector has a d-c bias, set the zero adjuster of the voltmeter to bring the pointer to zero or other easily identified marking on the scale when there is no signal from the generator.

Tune the generator precisely to a frequency at which one of the couplers is to be peaked, and vary the alignment adjustment of this coupler to produce maximum reading on the meter. Reduce the generator output voltage to allow using the lowest and most sensitive scale of the meter. If the pointer

tends to go off scale as alignment proceeds, bring it back by reducing the generator signal voltage, not by using a higher range of the meter.

Proceed to adjust any other couplers which are to peak at the same frequency, always aligning for maximum meter reading. Then tune the generator to another of the peak frequencies, and adjust other couplers which are to peak at this new frequency. Proceed thus until all couplers are aligned for their peaks. Always finish the alignment process by observing the overall response of the video i-f amplifier with a sweep generator feeding the mixer and an oscilloscope connected across the video detector load.

It is possible also to carry out the entire alignment by using a sweep generator and a marker generator, and by using an oscilloscope connected across the video detector load. The

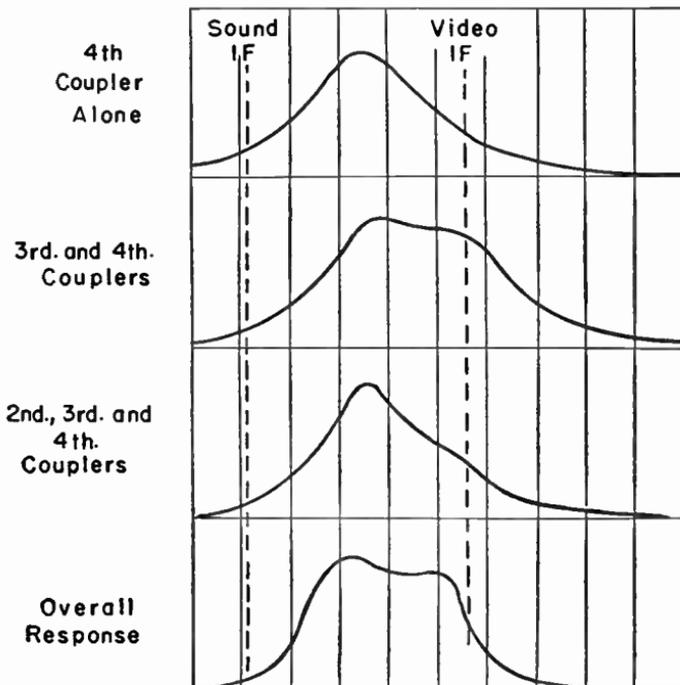


Fig. 8-28.—Frequency responses observed during alignment of a stagger tuned video i-f amplifier.

generators should be coupled first to the grid of the amplifier tube preceding the last coupler, as explained in connection with Fig. 8-25, then successively back to the grids of preceding tubes as alignment proceeds from the last coupler through to the coupler following the mixer.

To make a correct alignment by this method it is practically necessary to have available a set of response curves showing the results to be obtained in each step. The curves must, of course, apply specifically to the receiver being worked on. Fig. 8-28 shows fairly typical responses for a receiver having three i-f amplifier tubes and four couplers. At the top is the response with the generators to the grid of the last amplifier. Next is the response with the generators at the grid of the next preceding amplifier, so that the gain is affected by both the third and fourth couplers. The two lower responses show the effects of bringing in the second coupler, and of putting the signal through all four couplers for the overall response of the system.

**Alignment of Overcoupled I-F Stages.** — For accurate alignment of overcoupled transformers, without spending an excessive amount of time on the work, it is necessary to employ a sweep generator, a marker generator, and an oscilloscope. It is of utmost importance that the signal voltage from the sweep generator be held at the very lowest level which allows a readable trace on the scope. Otherwise the frequency response will be so distorted as to make correct alignment impossible. The signal must remain below a value with which a slight increase causes a change in the form of the trace as well as in its height or amplitude. The signal level must remain within the range where any small increase or decrease causes a proportional increase or decrease in the height of the trace, it must not be so high that further increase does not cause a proportional rise of the response curve.

If the attenuator of the generator will not drop the signal voltage low enough, try using a smaller coupling capacitor between the generator lead and the tube grid, or connect a resistor of a few hundred ohms from the grid to ground at a point beyond the coupling capacitor, not from the generator

lead to ground. A coupling ring may be moved higher up on the tube, or a tube shield used for coupling may be moved higher. Getting the ring or shield farther from tube elements reduces the coupling.

Marker signals must be kept weak enough to cause only the slightest drop of trace height when these signals are applied, and only the slightest rise when they are completely removed. Marker strength may be reduced by the same methods mentioned for the sweep signal. There may be enough coupling with the marker cable merely laid near the sweep generator cable or brought near the tube to which there is to be coupling. Another method is to couple the marker generator to the grid of some tube preceding the tube to which the sweep is coupled.

The oscilloscope usually is connected across the video detector load resistor. Sometimes the trace cannot be made high enough without using excessive signal input. To increase the height, the scope may be connected across the plate load resistor of either the first or second video amplifier tube. For still greater trace height the scope may be connected between ground and the picture tube control grid or cathode, whichever is used for signal input.

If video frequencies or high intermediate frequencies get into the lead to the oscilloscope they are likely to cause feedback to the early stages of the video i-f amplifier, which will result in distorted response. This may be prevented by connecting between the leads to the oscilloscope, at the receiver end, a capacitor of 0.002 mf or larger size, or by using a regular high-frequency filter probe on the oscilloscope vertical input.

When the trace tends to become too high for the screen as alignment proceeds, bring it down by reducing the signal voltage from the sweep generator. Always keep the vertical gain of the oscilloscope as high as possible; do not use it for reducing the height of the trace.

Overcoupled transformers usually are aligned while the oscilloscope is connected across the video detector load, and with signal generators coupled first to the video i-f amplifier

tube preceding the detector, then to amplifier tubes farther and farther from the detector until these instruments are coupled to the mixer tube.

Fig. 8-29 shows frequency responses obtained in the form of oscilloscope traces during successive steps of alignment on one type of receiver. With the generators coupled to the third i-f grid, which is the grid of the amplifier preceding the detector,

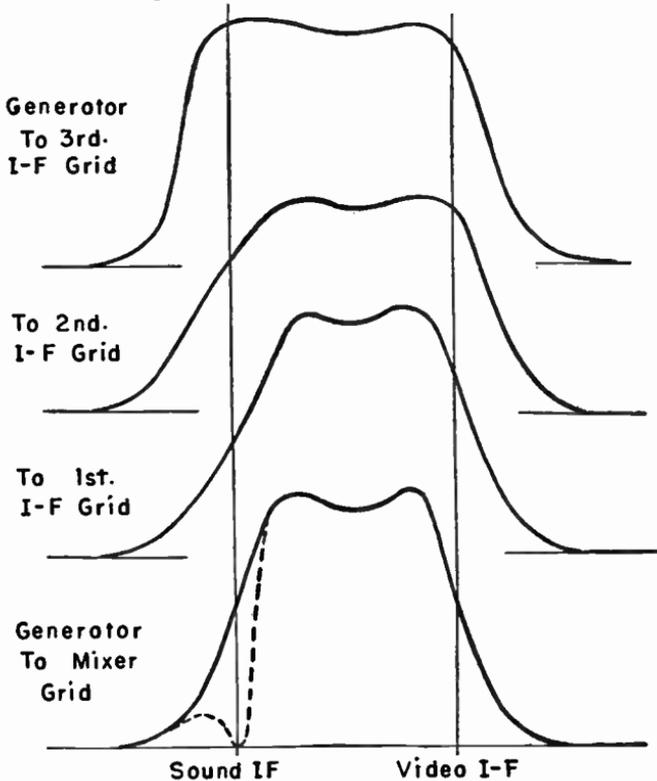


Fig. 8-29.—Frequency responses observed during alignment of overcoupled video i-f stages.

the response (at the top) is that of the single transformer between this last amplifier and the detector. With the generators moved back one stage the response, second from the top, is that of two coupling transformers working in cascade. Then, with the generators at the grid of the first i-f amplifier, comes the response for three coupling transformers working together.

Finally, with the generators coupled to the mixer tube, we have the overall video i-f response shown at the bottom of Fig. 9-29—with all four coupling transformers carrying the signal and affecting the shape of this response.

The response of a single stage is of the same general double-peaked form for any kind of overcoupled transformer. As more and more stages carry the signal from generators to oscilloscope the sides of the response curves tend to draw in, and the gain becomes much greater with each added stage. The two peaks of the overall response need not be of the same height, but at the deepest point of the valley between peaks the gain or amplitude should be no less than 70 per cent of the gain at the highest peak.

Any traps in the video amplifier will make decided change in the shape of the response from the coupling transformer with which the trap is associated, and in the shapes of responses from all stages between the traps and the video detector. As an example, a trap for accompanying sound might be coupled to the transformer between mixer and first i-f amplifier of the system to which Fig. 8-29 applies. All responses would be greatly altered from the forms shown, and in the overall response there would be an effect somewhat as shown by the broken line curve on the bottom (overall) response.

A careful analysis of the kinds of couplers used in various stages, and of the traps and trap frequencies associated with each coupler, will indicate the general form of response which should be obtained at each step in the alignment. But for best results, and least expenditure of time, the alignment always should be carried out in accordance with response curves furnished by the receiver manufacturer. Such curves show various marker frequencies in the positions they should occupy on the several responses when alignment is correct for each step in the process

Overcoupled transformers may be used in some stages of video i-f amplifiers, with single-peaked coupling coils in other stages. As an example, Fig. 8-30 represents a video i-f amplifier in which the first and last couplings are by means of overcoupled transformers, with the second and third couplings

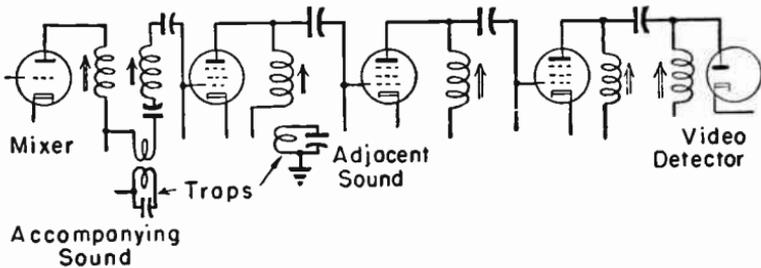


Fig. 8-30.—Overcoupled transformers and single-peaked coupling coils in the same video i-f amplifier.

by means of coils which are tuned to peak at a single frequency. On the transformer between mixer and first i-f amplifier is a trap for accompanying sound, and on the coupler in the following stage is a trap for adjacent sound.

**Overall Response Check.** — When a video i-f amplifier having overcoupled transformers has been aligned with the aid of an oscilloscope, the last step will have been observation of the overall response. When an amplifier having single-peaked couplers has been aligned for the several peak frequencies with the help of an electronic voltmeter it is highly advisable to make a final check on the overall response with sweep and marker generators coupled to the mixer tube, and with the oscilloscope connected across the video detector load.

The upper response curve of Fig. 8-31 shows frequencies whose positions on the trace should be checked by means of the marker generator. The video intermediate frequency on the high-frequency slope of the curve usually is required to be at 50 per cent of the height of the highest peak. This frequency should be no lower than 50 per cent, but on some receivers there is an advantage in having it somewhere between 50 and 60 per cent of the maximum gain.

At the frequency of each trap in the amplifier the gain should be zero or very nearly so. If there are any traps at all, one or more of them will be tuned to the accompanying sound frequency, which is the sound intermediate frequency. Adjacent sound traps will be tuned 1.5 mc higher than the video intermediate frequency, and adjacent video traps will be tuned 6.0 mc below the video intermediate frequency.

There should be no peak so high or so sharp as to indicate regeneration in the amplifier. The greatest dip between peaks should be no lower than 70 per cent of the highest peak. It

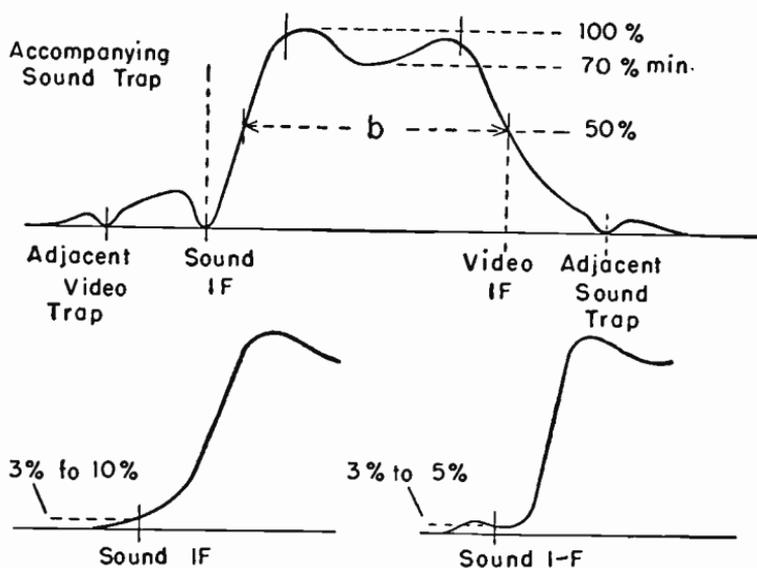


Fig. 8-31.—Frequencies at which the overall response curve should be checked for shape and height.

may be of interest to check the band width at a gain approximately 50 per cent of maximum, as indicated at *b* on the figure.

At the lower left in Fig. 8-31 is shown the low-frequency side of a response for an intercarrier sound system having no traps at all. At the sound intermediate frequency the gain should be between 3 and 10 per cent. Ordinarily the gain at this frequency must be held to 5 per cent or less in order to avoid "intercarrier buzz." If the video i-f amplifier for an

intercarrier sound system has some one stage containing a trap which forms a sound shelf or plateau, shown at the lower right, the gain at the sound intermediate frequency should be around 3 to 5 per cent of maximum.

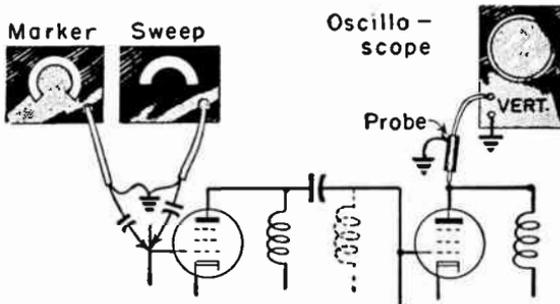
Should the shape of the overall response require correction it is desirable to change the adjustment of only the coupler or couplers which affect the portion of the curve which requires change. If peak frequencies for the various couplers are known, it is easy to determine which should be realigned, for frequency at the portion of the curve to be changed identifies the coupler tuned nearest to that frequency. If coupler peak frequencies are not known, the ends of the tuning wand may be brought near each of the coils. The portion of the response which then rises or falls will be chiefly affected by adjustment of the coil to which the wand is being presented. Otherwise touch a finger to the grid lug on each i-f socket. The portion of the curve chiefly affected will be altered by adjustment of the coupler preceding the tube whose grid is then being touched.

Quite often it will be found that some one coupler will have more effect than others on the entire shape of the curve and also on the overall band width. As a rule this coupler will be either the one just ahead of the video detector or else the one which immediately follows the mixer tube.

Sometimes an overall response is checked with the sweep generator connected to the antenna terminals. Then the response is affected by the r-f amplifier and tuner coils as well as by the video i-f amplifier. When this generator connection is being used it is inadvisable to attempt adjustments on the video i-f couplers. Adjustments made under this condition often turn out to be incorrect when again checking with the sweep generator at the mixer tube, and without the r-f amplifier carrying the signal.

**Individual Stage Responses.** — The method usually employed for observing the frequency response of a single stage or a single coupler is illustrated by Fig. 8-32. The sweep and marker generators are coupled very loosely to the control grid of the tube preceding the coupler to be checked. If the follow-

ing tube is an amplifier, whose signal output is at intermediate frequency, connect the vertical input of the scope through a detector probe to the amplifier plate. If the video detector follows the coupler, connect the oscilloscope across the video detector load, but do not use a detector probe. The same general method may be used whether the coupler being checked



**Fig. 8-32.**—Instrument connections for observing the frequency response of a single stage or single coupler.

is a single-peaked coil or a two-winding transformer. In the case of transformer couplings it may be necessary to short to ground the control grid of the coupler ahead of the one checked, as previously explained. The response trace may be satisfactory with the oscilloscope probe connected to the screen instead of to the plate of the second amplifier.

It is useless to connect the test instruments to the plate of an amplifier preceding the coupler being checked, or to the control grid of the amplifier following this coupler. It is useless also to make response observations with either of the tubes removed from its socket, since this removes the tube capacitances and changes the tuned frequencies.

With an entirely different method the generators may be coupled to the mixer tube and the oscilloscope connected across the video detector load for observation of responses in each of the individual stages. All the interstage couplers except the one whose response is to be observed are detuned by connecting across the ends of their coil or coils a fixed carbon

resistor of about 300 ohms. After observing the response of any one stage in this manner, a resistor is connected across the coupler of this stage and a resistor is removed from the coupler in another stage. Then the response of that second

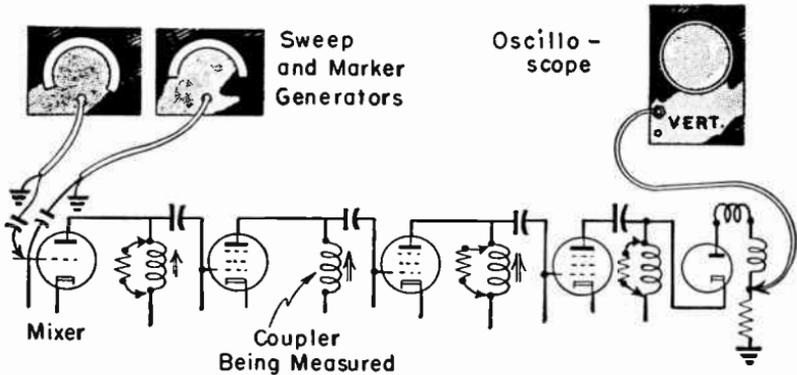


Fig. 8-33.—Observing individual stage response by shorting out the couplers in all other stages.

coupler is observed. So the work proceeds until all the couplers have been checked. The scheme of connections is illustrated by Fig. 8-33.

Instead of connecting resistors across all couplers except the one being measured, the control grids of all but one tube may be connected to ground through resistors of about 300 ohms. The results will be similar so far as the oscilloscope trace for response is concerned. The coupler remaining active is the one preceding the ungrounded grid, the coupler which is connected to and feeds to this grid.

The frequency at which any one stage is peaked may be determined with the couplers shorted with resistors or with the grids grounded through resistors on all stages except the one to be checked. The method is shown by Fig. 8-34, on which are drawn resistors from grids and detector input to ground, although the couplers might be shorted with resistors as in Fig. 8-33.

A constant frequency generator is coupled to the mixer tube. This generator may be a marker type, but it is not a

sweep generator. Across the video detector load is connected an electronic voltmeter or any high-resistance d-c voltmeter. The coupler which is not shorted or whose grid is not shorted to ground is peaked at the frequency to which the generator is tuned when meter reading is maximum. The generator is

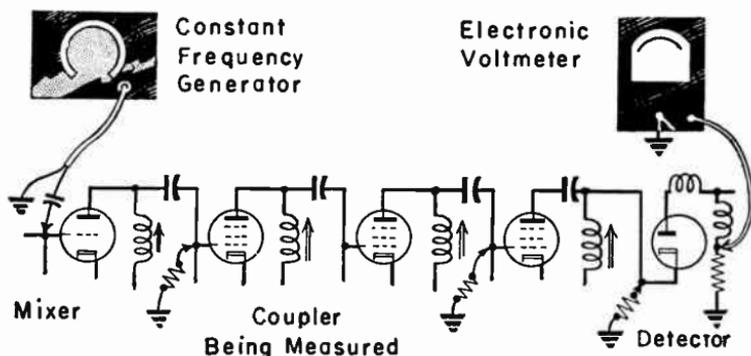


Fig. 8-34.—Determination of frequencies at which individual couplers are peaked.

tuned for maximum meter reading, and the frequency then read. An overcoupled transformer in the stage being checked will show two voltage peaks at its two peak frequencies if there is a dip or valley between peaks of the response. If there is no valley between peaks, the meter reading will remain nearly constant over a range of frequencies along the flat top of the response.

Individual couplers may be peaked at any desired frequency by using connections as described for Fig. 8-34. The generator is tuned to the desired peak frequency, then the unshorted coupler is aligned for maximum reading on the voltmeter.

If there is any doubt as to whether the meter reading is due to the coupler left unshorted or whose grid is unshorted, a finger may be touched to the grid pin or lead at the socket. The meter reading should drop nearly to zero. This finger test may be used also to make sure that an oscilloscope curve is due to the coupler being adjusted. When a grid connection is touched, the curve will flatten out if it represents response of the coupler ahead of the grid.

**I-f Amplifiers for Color Receivers.** — The intermediate - frequency amplifier section in a color television receiver is similar in most respects to the equivalent section in a black-and-white receiver, but performance requirements are much more stringent. In a fairly typical i-f amplifier for color there are four pentodes of medium high transconductance, 5,000 to 6,000 micromhos, having semi-remote cutoff characteristics. That is, for reduction of transconductance to a practical minimum of 50 micromhos the control grid must be made about 12.5 volts negative with reference to the cathode instead of only about six to eight volts negative as with most sharp cutoff pentodes. The semi-remote cutoff feature allows more satisfactory or more effective action of the automatic gain control system.

Interstage couplings, other than from mixer to first i-f amplifier, are by means of bifilar transformers in which plate and grid windings are insulated from each other, but with which there is very close coupling and large signal-power transfer because the two windings are run together around the supporting form.

In other designs there are only three i-f amplifier tubes, while in still others there are five tubes. Sharp cutoff pentodes of medium to high transconductance are used.

Intermediate frequencies nearly always are 41.25 mc for sound and 45.75 mc for video. This conforms to general practice in a large number of recently designed black-and-white receivers.

It is quite common to find more traps on the i-f amplifier section of a color receiver than on the black-and-white i-f amplifier. There may be one trap, but frequently there are two for the accompanying sound frequency, which would be 41.25 mc. In addition there may be another trap for the adjacent sound frequency of 47.25 mc, and third for the adjacent video frequency of 39.75 mc. In addition to rejecting or attenuating the interference frequencies, the accompanying sound traps help materially in providing cutoff at the high end of the video range while allowing high and uniform gain for all video frequencies up to the cutoff point.

Frequency response of the color i-f amplifier at the end carrying the video intermediate should be like that in any black-

and-white receiver. This means that gain at the video intermediate point should be 50 per cent of maximum, with a drop to zero or nearly zero at 0.75 mc higher and a rise to maximum at 0.75 mc lower frequency.

From the point at which the response rises to maximum gain the gain should be very nearly uniform to a frequency at least 3.9 mc and preferably to 4.0 or 4.1 mc from the video intermediate. This high and uniform gain over a range much wider than required in most black-and-white receivers is necessary because the chrominance signals in the i-f amplifier center at 3.58 mc, which is the frequency of the color subcarrier.

Chrominance signal sidebands, which represent coloring for pictures, extend to about 0.5 mc or slightly more, both above and below the subcarrier frequency of 3.58 mc. Adding 0.5 mc to the subcarrier frequency of 3.58 mc gives 4.08 mc or practically 4.1 mc as the range to which full response should extend above the video intermediate frequency.

In terms of intermediate frequencies observed on the response curve, the color subcarrier frequency is at 42.17 mc, because this is the difference between the video intermediate of 45.75 mc and the color subcarrier frequency of 3.58 mc. If the response is to extend 0.5 mc beyond the subcarrier frequency it must go to 41.67 mc on the i-f response curve. This is only 0.42 mc from the sound intermediate frequency. It is apparent that the low-frequency end of the i-f response must show a nearly square dropoff, and decrease very rapidly to zero.

If the i-f response does not maintain high gain almost as far as the sound intermediate, color sidebands or chrominance signal sidebands will not be properly amplified, or will be cut off on one side of the subcarrier frequency. Colors will be weak, or may disappear almost entirely, in spite of all possible readjustments of hue and saturation controls.

It is rather obvious that misadjustment of the fine tuning control on the tuner of a color receiver may have the same effects as poor alignment of the i-f amplifier section. Wrong adjustment of the fine tuning in one direction will shift the response curve in such manner as to leave little or no gain at the frequencies which carry color information. These are the frequencies from about 43.15 mc to 41.67 or 41.65 mc. Misadjustment of fine tuning in the opposite direction will have

the usual effect of losing sync control and of causing smears or trailers in pictures.

I-f alignment for color receivers is carried out with a sweep generator, marker generator, and oscilloscope in the same general manner as for black-and-white receivers. As a rule it is desirable to commence with the final i-f amplifier or coupling, by connecting the sweep and marker generators to the grid of the i-f amplifier tube just ahead of the final coupler, with the vertical input of the oscilloscope to the output of a video detector or a detector which feeds the luminance channel. Then the final coupler, also any trap on the final stage, may be adjusted.

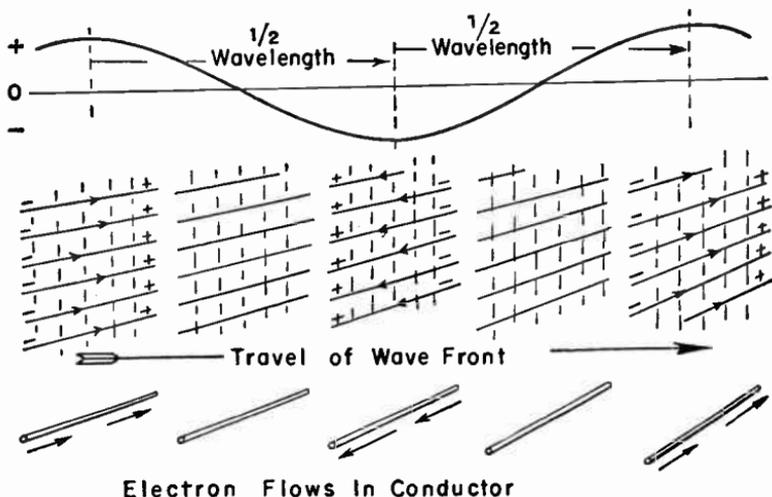
Preceding stages or couplers may be aligned by moving the generator connection back toward the tuner one tube at a time, but ordinarily it is satisfactory to go to the grid of the first i-f amplifier after adjusting the final coupling. Then remaining couplings between i-f amplifier tubes are adjusted. The final step in alignment is with the generators connected to the mixer grid. The oscilloscope remains at the output of a detector for all steps.

This stage-by-stage method of alignment requires that correct response curves be available for each step, such curves being found in service instructions for the receiver. Without such curves the alignment may be made for all couplings with the generators at the mixer grid. In this case the combined effects of so many couplers, traps, and tubes on the overall response is likely to make the process long and difficult.

So far as good color reproduction is concerned, the most important part of the alignment process is in obtaining high and uniform gain almost as far as the sound intermediate frequency, with a sharp drop just before this frequency is reached. Of course, alignment at the video frequency end of the response should be the best possible, provided the low-frequency end is not slighted.

During earlier days of television servicing it was felt that accompanying sound traps should not be between the tuner and the point of sound takeoff. With modern intercarrier sound systems it is desirable to attenuate the sound intermediate before the takeoff. Consequently, one accompanying sound trap usually is on the coupling between mixer and the first i-f amplifier.

**ANTENNAS.** — Every television antenna, no matter how elaborate, operates on the same basic principles as the simple half-wave straight dipole illustrated by Fig. 9-1. The antenna



*Fig. 9-1.—The electrostatic force of the radio wave causes alternating electron flows in the antenna conductors.*

consists of two conductors, metal rods or tubes, separated at the center and there connected through a transmission line to circuits in the receiver.

If overall length of the two antenna conductors is approximately half the wavelength of signals to be received, opposite ends of the antenna conductors are charged in opposite polarities at the same time. The charges cause signal currents to flow as shown by arrows on the diagram.

The half-wave dipole forms a series resonant circuit whose inductance is proportional to length of conductors and whose total capacitance is proportional to conductor diameter and to stray capacitances. Charges or emf's are induced in this resonant circuit when the electric fields of signal waves cut across the antenna conductors.

Induction results from the action illustrated by Fig. 9-2, where the antenna conductor is shown as continuous from end to end, although actually it would be divided for connection

## ANTENNAS

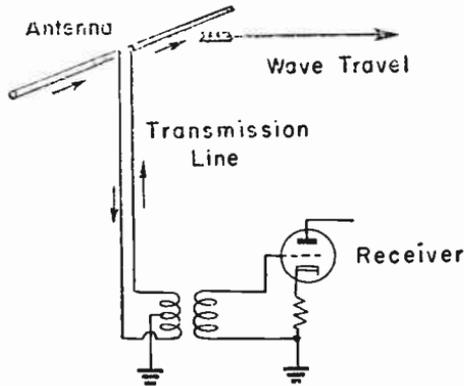


Fig. 9-2.—Connection of a dipole antenna through a transmission line to the receiver.

to the receiver. The electromagnetic signal waves consist of electric or electrostatic lines of force which are horizontal and of magnetic lines of force which are vertical. Electrons in the antenna conductor are caused to move in the same directions as the electrostatic forces.

In any conductor wherein electron charges are caused to move there is reflection and reversal of direction when the charges reach either end of the conductor. Electron velocity in the conductor approaches the speed of light, and after a charge is placed in motion it tends to reflect back and forth at intervals proportional to conductor length. If the antenna conductor is approximately half as long as a signal wave, every natural back and forth movement of charges is strengthened by reversing electric forces in passing waves, and maximum signal currents are induced in the antenna and connected circuits.

Based on actual velocity of charges in antenna conductors which are  $\frac{3}{8}$  to  $\frac{1}{2}$  inch in diameter, and on frequency of carrier signals or waves, satisfactory overall length in inches of a half-wave dipole may be found from this formula.

$$\text{Overall antenna length, inches} = \frac{5690}{\text{signal frequency, megacycles}}$$

Actual resonant frequency of an antenna varies to some extent with conductor diameter, shape, and material, also with

the type of supports, and is affected by surrounding objects. As a consequence, there are variations in formulas for antenna length. However, precise lengths are not particularly important in practice. Signals are received with maximum strength at a frequency for which the antenna actually is resonant, and with less strength at other frequencies.

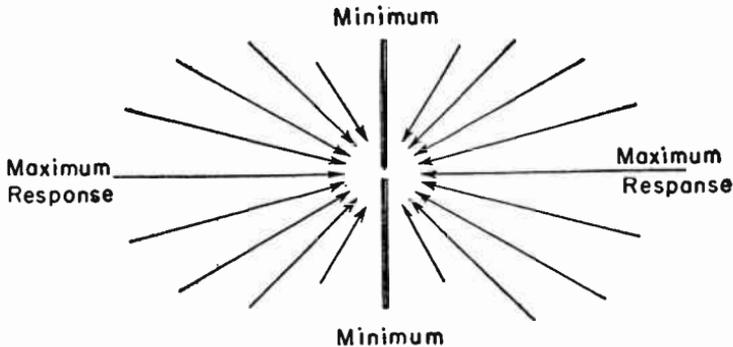


Fig. 9-3.—Directional properties of a simple half-wave dipole antenna.

**Directional Properties of Antennas.** — Fig. 9-3 shows directional properties of a simple half-wave dipole. Relative lengths of radial lines indicate relative response of the antenna to signals of equal strength arriving from the directions of the lines. There is maximum response to signals traveling on a line at right angles to the antenna conductors, and minimum response to signals traveling in line with the conductors.

For best reception of signals from any given direction the antenna conductors should be supported at right angles to that direction, although, as shown by the figure, response will remain almost as good when the antenna is rotated to some extent either way from the right-angle relation. With the simple dipole now being discussed there is maximum response to signals from two opposite directions, and minimum response from two other opposite directions.

When reception is desired in several channels the dipole may be supported at right angles to the direction of weakest signals. Often it is possible to greatly attenuate or even to

eliminate an undesired or interfering signal by rotating the antenna conductors to a position in line with the direction of that signal. The angle of minimum response is relatively narrow, and to cut out an unwanted signal the antenna conductors must be almost exactly in line with that signal. Desired signals may be well received with the conductors turned quite a ways either side of the maximum point. The directional pattern of Fig. 9-3 is based on having the antenna far enough from all other conductors, and from large bodies having poor dielectric properties, as to be unaffected by them.

Directional properties of antennas usually are shown by polar diagrams such as those of Fig. 9-4. Diagram *A* illustrates a response much the same as that shown by radial lines in Fig. 9-3, with the oval pattern of the polar diagram passing through what would be the outer ends of radial lines. Strength of signals as received from any direction is proportional to the distance in that direction from the center of the diagram to the polar pattern.

The outer circle of any polar diagram corresponds to maximum or 100 per cent signal strength, no matter what may be the actual strength as measured in microvolts or other unit.

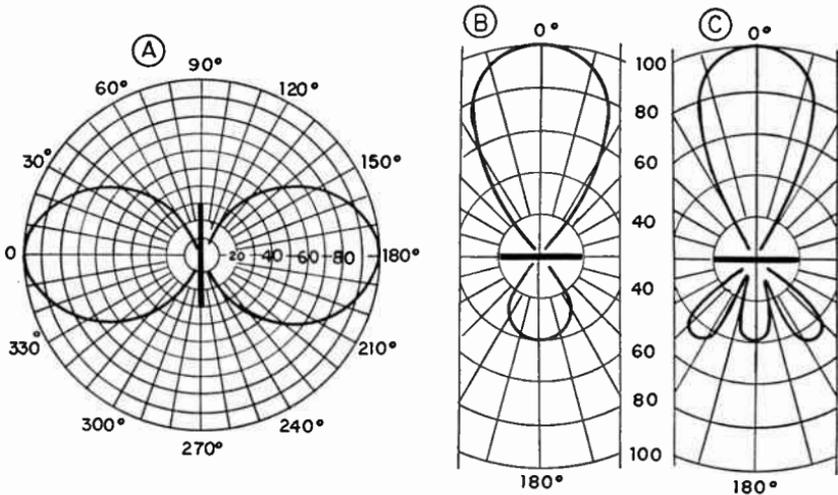


Fig. 9-4.—Polar diagrams showing directional responses of antennas.

Other circles are drawn at equal intervals for proportional drops of signal strength until reaching zero strength at the center. Radial lines on diagram *A* are at intervals of 10 degrees of geographical direction, commencing at zero degrees in the direction of best reception.

Polar diagrams as ordinarily shown are based on carrier waves traveling horizontally, or parallel to the surface of the earth. Actual wave travel may be at an angle with the horizontal, being inclined slightly upward or downward in some localities. This alters the directional response, especially in that signal pickup will not drop to zero for waves in the same direction as the line of antenna conductors.

Antenna assemblies usually include elements which increase response in one direction and reduce it in the opposite direction. Such unequal responses may show on a polar diagram as at *B*, where "back response" is only 40 per cent of "front response." The oval shaped outlines showing percentages of response in various directions are called lobes. The largest lobe or lobes are called major lobes, while others are called minor lobes. Many antennas having unequal responses in different directions have several minor lobes, as in diagram *C*.

**Broad Band Antennas.** — An important characteristic of any antenna is its bandwidth, which refers to the range of carrier frequencies or to the number of channels in which signal pickup is sufficient for satisfactory reception. A simple half-wave dipole tunes rather sharply over a rather narrow band of frequencies centering at the frequency for which length of the dipole makes it resonant. An antenna of this kind is satisfactory for either the high band or the low band, but not for both bands even where signals are strong.

One of the earliest antennas covering a broader band of frequencies, and a type still widely used, is the half-wave folded dipole of Fig. 9-5. As usually constructed the folded dipole is a long piece of tubing bent to form a flat loop with one side continuous and the other open at the center for connection of the transmission line. The loop is mounted with its plane vertical, with one side above the other. The

## ANTENNAS

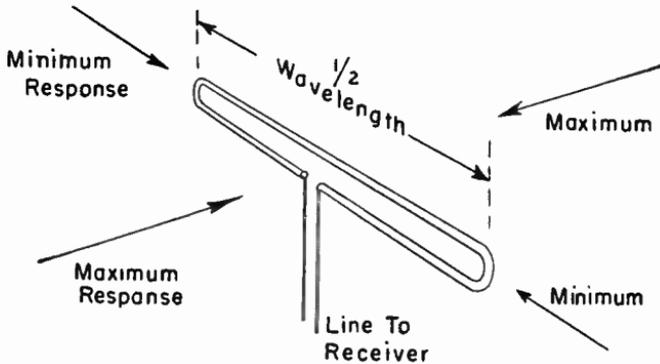


Fig. 9-5.—Half-wave folded dipole antenna and its directional properties.

transmission line gap may be either above or below. Distance between bends at the ends of the loop is made equal to a half wavelength for the frequency at which the antenna is to be resonant.

Like the straight dipole, the folded type has maximum response to signals at right angles to the length of antenna conductor, and minimum response to signals traveling in line with the conductor. The directional pattern of the folded dipole is like that for the straight half-wave dipole.

A popular style of broad band antenna is the conical, one type of which is illustrated at the left in Fig. 9-6. Three conductors radiate on each side from a common center. Were more and more conductors to be added they would have the approximate shape of two opposed cones, hence the name conical. The reason for using three conductors fanned out from the center on each side of the antenna is to pick up more

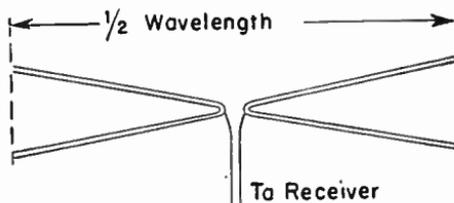


Fig. 9-6.—A double-V dipole antenna.

signal energy from carrier waves than would single opposite conductors of a straight half-wave dipole.

A conical antenna and other types having forwardly tilted conductors on opposite sides may be designed for good response in both low-band and high-band channels of the vhf television range. Attainment of wide frequency response depends on two facts. First, any half-wave dipole is series resonant not only at a frequency corresponding to overall length of its conductors but also to the third harmonic of this frequency. It will be found that all carrier frequencies in the vhf high band are third harmonics or are three times as great as some carrier frequency in the low band.

The second fact is this: Reception lobes may be shifted in direction by bringing the outer ends of an antenna ahead of the center to give a V-formation as viewed from above. A straight dipole will have directional response in the low band as shown at *A* of Fig. 9-4, but at third harmonic frequencies in the high band there will be six lobes, about as shown by Fig. 9-7. There are two strong major lobes and a minor lobe

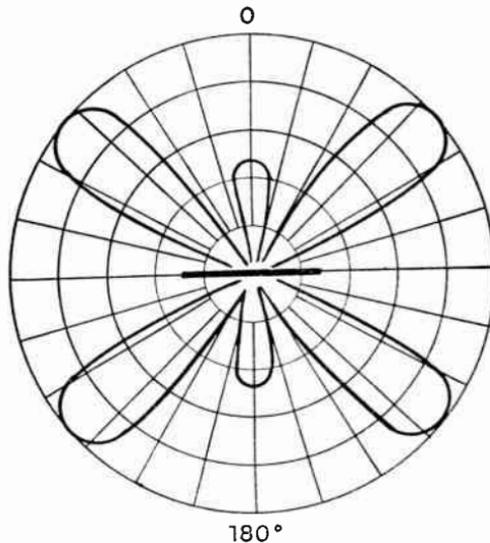
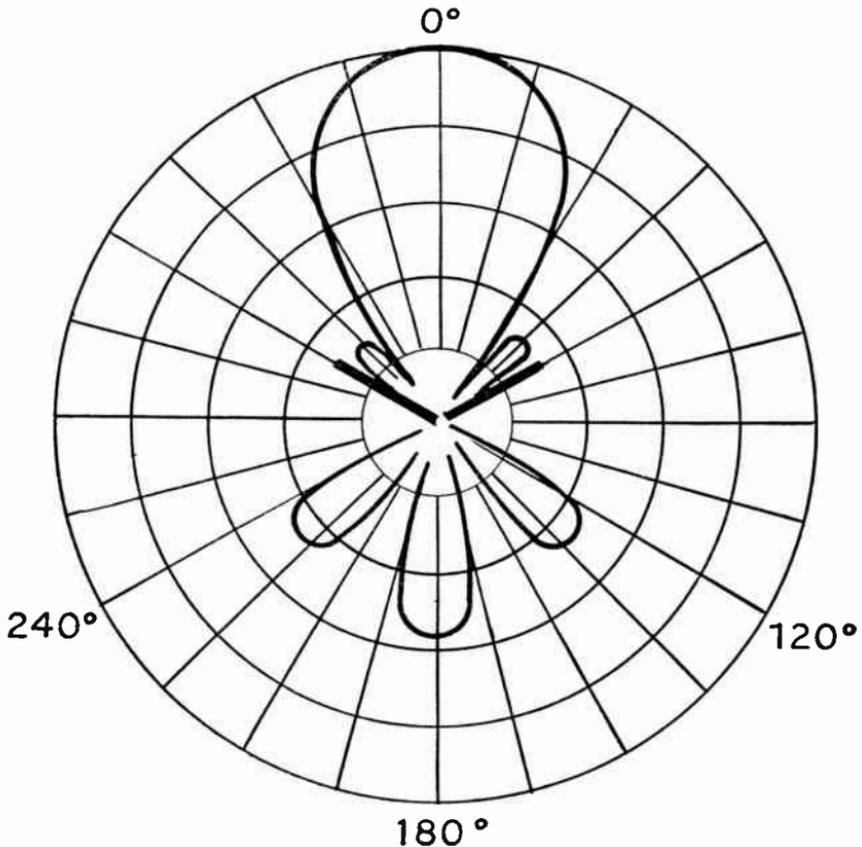


Fig. 9-7.—Antenna directional response at a third harmonic frequency of the fundamental for which cut.

in each direction. Although the major lobes are shown as extending to 100 per cent or to maximum pickup, this maximum actually is weaker than the maximum in low-band channels.

The two major lobes at the third harmonic frequency may be combined by inclining the sides of the antenna toward the direction of desired signal pickup. This is shown by Fig. 9-8.



*Fig. 9-8.—How response lobes are combined by forward inclination of antenna conductors.*

Principal response now is in one direction, both for high-band channels on third harmonics of antenna resonant length and for low-band channels on the fundamental antenna frequency. Back responses, between 120 and 240 degrees, become less in

relative strength. The principle of third harmonic response for broad band reception is employed also in V-type antennas of the general shape illustrated by Fig. 9-9.

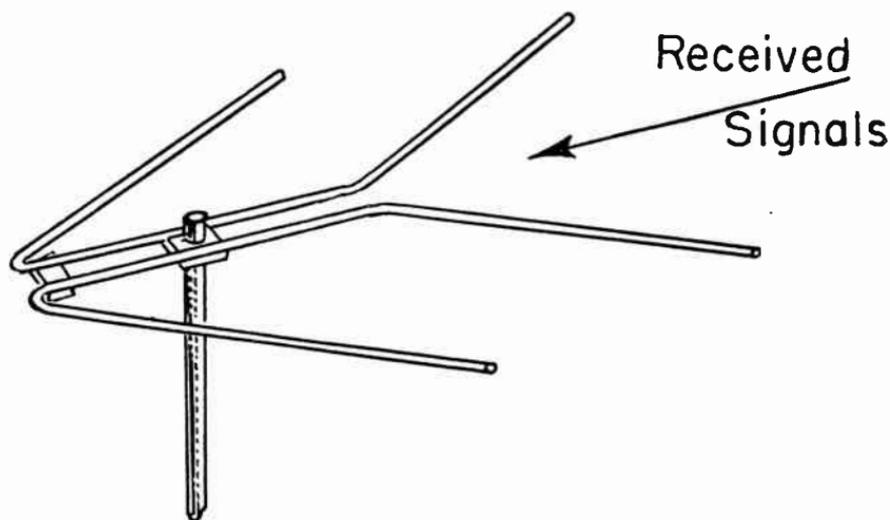


Fig. 9-9.—A V-type antenna for broad band reception.

**Gain of Antennas.** — The gain of an antenna is the ratio between its signal voltage delivered to a receiver and signal voltage delivered under the same conditions by some other antenna which is the standard of reference. The reference antenna commonly is a plain half-wave dipole or sometimes a folded dipole, of a length which makes it resonant at each frequency for which signal strengths are compared. Both antennas are subjected to the same or equal signals in space, and both feed into the same or equal impedances when comparisons are to be made on the basis of signal voltages.

Antenna gain is not amplification such as occurs in a tube, but is merely a comparison with performance of a reference antenna. Therefore, antenna gains are expressed in decibels, usually in decibels of voltage. Zero decibels or a one-to-one ratio corresponds to signal output from the reference antenna. Fairly typical gains of conical antennas throughout the low band and high band channels are shown by Fig. 9-10. The

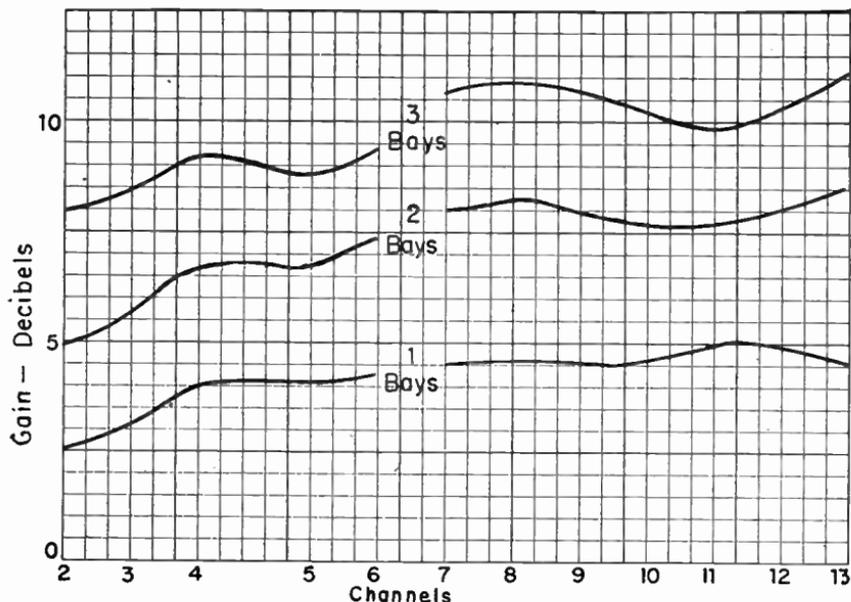


Fig. 9-10.—Relative gains in vhf channels of antennas having one, two, or three bays.

lower curves apply to single-bay antennas, which consist of one set of dipole conductors and a reflector. Two bays include two similar sets of elements which are stacked one above the other, connected together, and to the receiver. The three-bay antenna consists of three sets of similar elements, stacked.

**Reflectors and Directors.** — At the left in Fig. 9-11 an ele-

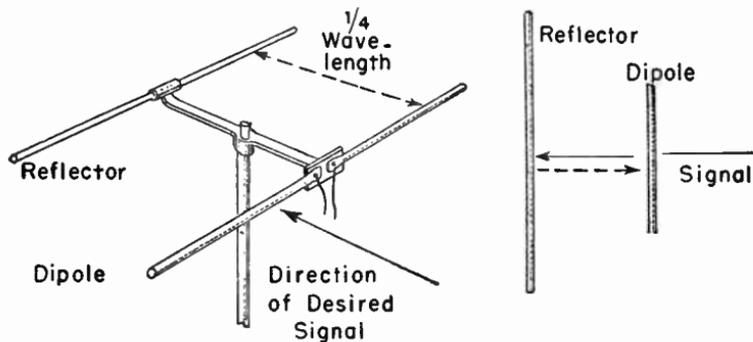


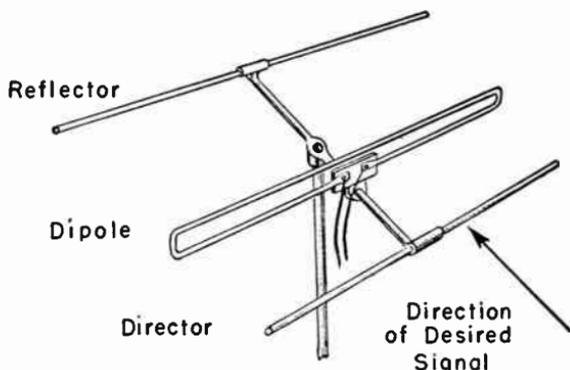
Fig. 9-11.—A half-wave dipole antenna with a reflector element back of it.

ment called a reflector is mounted back of the dipole antenna, or on the side opposite to that from which desired signals approach. The reflector may be a single piece of tubing or rod, or two pieces conductively joined together. The reflector is not electrically connected to the dipole, and may be either insulated or clamped conductively to the supporting boom and mast structure.

Signal waves induce emf's and currents in the reflector in the same manner as in the dipole. Induced signal currents cause reradiation of signal energy from the reflector. Part of this reradiated energy goes back to the dipole, reaching there in such phase relation to signal waves at the dipole as to reinforce received signals. The result is a decided increase of forward response and, relatively, the back response is weakened.

Reflectors usually are mounted back of the dipole at a distance equal approximately to half the overall length of the dipole itself, which is a quarter-wave spacing. Gain is increased by moving the reflector somewhat closer to the dipole and making it 5 to 10 per cent longer than the dipole, but this causes a considerable drop of antenna impedance. A reflector closer than a quarter-wavelength to the dipole reduces bandwidth, while greater spacing increases bandwidth.

In Fig. 9-12 a director element has been added in front of the dipole, while retaining the reflector on the opposite side.



**Fig. 9-12.**—A folded half-wave dipole with a reflector element back of it and a director element in front.

The dipole is shown as a folded type. Reflectors and directors are used with all forms and shapes of antennas. The director, like the reflector, is a conductively continuous rod or tube, either insulated from or conductively clamped to the boom and mast, but not electrically connected to the dipole.

Part of the carrier wave energy picked up by the director is reradiated to the dipole, and reaches the dipole in such phase as to reinforce received signals. Whether the added element acts as a director or as a reflector is largely a matter of spacing; a director being closer to the dipole than a reflector.

A director mounted close enough to the dipole to provide maximum gain causes a great reduction of antenna impedance, also a decrease of bandwidth. To avoid very low impedance the director may be moved slightly farther from the dipole, and made shorter than the dipole in order to retain as much gain as possible. A director increases the ratio of front to back response, thus adding to the similar effect of a reflector, and also produces a greatly narrowed front lobe or produces sharper directivity.

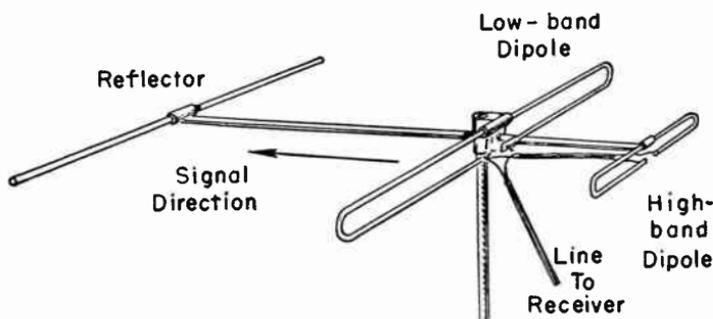


Fig. 9-13.—An "in-line" folded dipole antenna having elements for the high band and for the low band, and a single reflector element.

**Two-band Antennas.** — There are many antenna designs in which one element or set of elements is of length suited to reception in the vhf low band and another element or set of elements is "cut" for the high band. One such type is illustrated by Fig. 9-13. Toward the direction from which best reception is wanted is a folded dipole of overall length best

suited for high-band reception. Back of this high-band element is another folded dipole of length suited for low-band reception. Back of the low-band dipole is a reflector spaced to

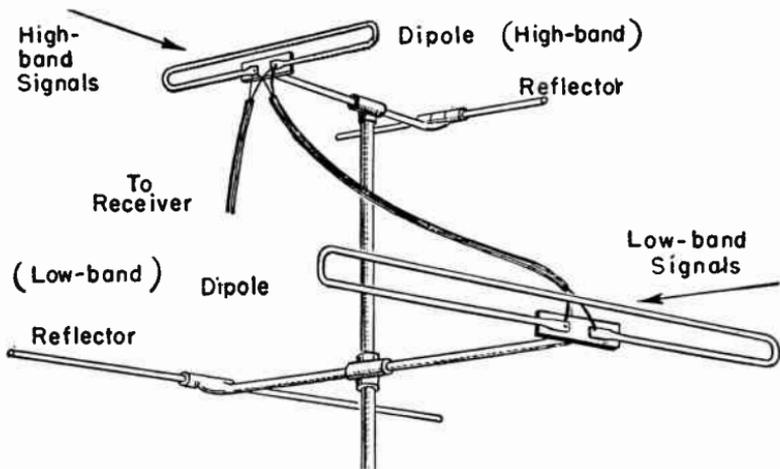
increase the gain in this band. Spacing between the two folded dipoles is such that the longer one acts as a reflector for the shorter one, thus increasing the gain in the high-band channels.

With the high- and low-band elements mounted in a fixed parallel relationship to each other, all signals to be received in full strength must come from the same general direction to the antenna location.

In localities where desired signals, or signals to be picked up with maximum strength, come from different directions in the high and low bands the dipoles suited to the two bands may be mounted one above the other as shown by Fig. 9-14. Either array may be rotated or orientated independently of the other, as may be found most advantageous for local conditions.

The high-band array usually is placed above the one for the low band, although this arrangement may be reversed. The figure shows the transmission line going from the high-band dipole to the receiver, with an extension or link going from the high-band dipole to the low-band dipole. This arrangement also may be reversed. Both dipoles may be provided with reflectors, as illustrated, or the reflectors may be omitted from either or both arrays—all depending on what is required for signal gain and reduction of interference from undesired signals. Not only the folded dipoles pictured in the figure may be used one above the other, but any other types of antennas may be similarly arranged.

When high-band and low-band elements are connected to a single transmission line, as in Fig. 9-14, or with any generally similar arrangement, both elements must be on the same mast. Were the elements even a few feet apart horizontally there might be enough phase difference between their received signals to cause blurring of pictures.



*Fig. 9-14.—Folded dipoles and reflector elements for the high band and the low band mounted for separate orientation in the two bands.*

**Yagi Antennas.** — A Yagi antenna, in the usual meaning of the name, is one having a single reflector and two or more directors. A Yagi with three directors is illustrated by Fig. 9-15. Types with as many as seven or eight directors sometimes are used. Most notable characteristics of the Yagi antenna are high gain, narrow bandwidth, sharp directivity in the forward direction, and small back response. Although most of these antennas have sufficient bandwidth for only one or two adjacent channels, some modified styles allow reception in either the entire low band or the entire high band of the vhf range, although with some sacrifice of gain.

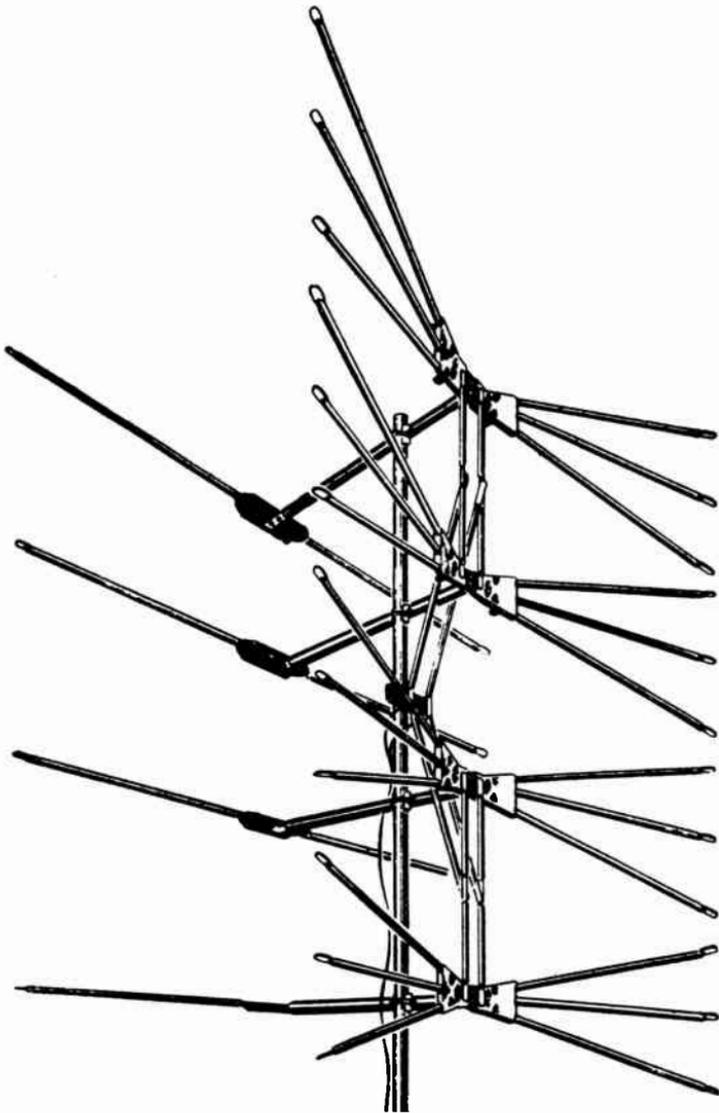
Because of its many director elements the Yagi antenna tends to have very low impedance. As a rule, impedance is maintained high enough to match transmission lines by such methods as using three-conductor folded dipoles, or by making the continuous conductor of a folded dipole of greater diameter than the divided element to which is connected the transmission line.



*Fig. 9-15.—A five-element Yagi antenna.*

**Stacked Antennas.** — Signal strength delivered to the receiver may be greatly increased by using two or more similar arrays (dipoles with reflectors and possibly directors) mounted one above another and suitably connected together and to the transmission line. With all the similar arrays designed for the same frequency band this arrangement is called a stacked antenna. A four-bay stacked antenna is illustrated by Fig. 9-16. The advantages in greater gain are clearly shown by the curves of Fig. 9-10.

Stacking may be carried out with any style of array, with straight dipoles, folded dipoles, V-type or conical dipoles, or anything else. The stacked elements must be of the same type. Each dipole may be used with reflector, director, both, or neither. Vertical spacing between centers of bays preferably is a half-wavelength at the frequency for which the antenna is to be resonant. Less spacing reduces gain, but may be necessary in order to limit overall size, especially with antenna elements cut for low-band or two-band reception.



*Fig. 9-16.—A fan antenna made with four stacked bays.*

The stacked elements to which connect the transmission line are connected together with phasing links which may be pieces of transmission line or else lengths of tubing or wire without insulation. With two stacked bays and transmission line takeoff from midway between them, as at the left in Fig.

9-17, conductors of the phasing link are joined to like ends of the dipoles. That is, the ends on the left of the gaps are connected together, and the two on the right are connected together. If transmission line takeoff is from one or the other of the bays, as from the bottom bay at the right in the figure, the phasing link conductors are transposed between upper and lower bays.

Were an additional bay added above or below or in both places on the stacked array at the left in Fig. 9-17, phasing

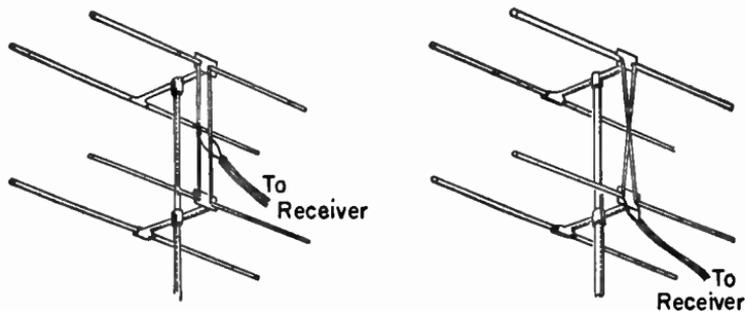


Fig. 9-17.—Whether or not the phasing link is transposed depends on the point of transmission line takeoff.

links to the added elements would be transposed. For example, with four bays the transmission line would connect midway between the two middle bays, to a link not transposed. Links to adjacent bays above or below would be transposed. Were line takeoffs midway between upper and lower pairs of bays, as in Fig. 9-16, there would be no transpositions.

Because of reduced wave velocity in insulated conductors a transmission line link will be electrically longer than the corresponding wavelength in air or in bare conductors. For example with vertical spacing of a half-wavelength between bays a straight link of transmission line would have electrical length greater than a half-wavelength.

**Impedance of Antennas.** — Since dipole antennas and all their modifications are tunable resonant circuits, inductive and capacitive reactances become equal and balance each other at the resonant frequency. This leaves high-frequency resistance as the only remaining factor in impedance, provided the received carrier frequency is the frequency of resonance and

there are no extraneous effects which prevent balancing of the reactances.

So far as reception is concerned, the antenna is the source of signal energy whose internal impedance depends on the type of antenna and conditions of operation. If the impedance of the antenna is matched by or is equalled by the impedance of the transmission line, half the energy picked up in the antenna will be transferred to the line. This represents maximum possible transfer of energy. If the impedance of the transmission line matches the impedance of the receiver input circuit, there will be maximum transfer of signal energy into the receiver. The subject of impedance matching is discussed in the article on *Lines, Transmission*.

The impedance at the center gap of a simple half-wave dipole (Fig. 9-2) and of other types which are essentially simple dipoles is about 73.2 ohms when the antenna is unaffected by any of the conditions to be mentioned in following paragraphs. The impedance at the center gap of a folded dipole, and of most other types which are developments of the folded dipole, is about 293 ohms, or approximately 300 ohms.

At all frequencies both higher and lower than the one for which the antenna is cut the impedance is greater than the values mentioned, because at all other frequencies there is an excess of either inductive or capacitive reactance. It follows that, if the antenna is longer than it should be for the received frequency, there is an excess of inductive reactance. And if the antenna is shorter there is an excess of capacitive reactance. An antenna used for reception in an entire band or in the whole range of very-high frequencies for television can have its working impedance matched by the transmission line and receiver at only one frequency, the one for which the antenna is cut, or the one with which the working impedance of the antenna becomes equal to the impedance of the line and receiver.

As an example, if a transmission line having an impedance of 300 ohms is used with a simple dipole antenna, there will be a match not at the frequency for which the antenna is cut, but

at some frequency enough higher or enough lower to bring the actual impedance of the antenna up to the impedance of the line. This explains why lines of 100-ohm, 150-ohm, or even 300-ohm impedance often seem to work very well with simple dipole antennas.

Antenna impedance is minimum at the center, and would become much greater were the takeoff point from either side of center. With takeoff from the center of a phasing link on stacked arrays, as at the left in Fig. 9-17, the impedance at the takeoff point is one-half that of the antenna elements, either element. With takeoff from the bottom, as at the right in that figure, or from the top, the impedance is the same as that of one of the elements.

A reflector tends to lessen the antenna impedance. The decrease is not great unless the reflector is mounted very close to the pickup element, or is of excessive length. Unless the antenna is supported at least a wavelength or farther from other conductors and from dielectrics of all kinds, the actual impedance will differ materially from theoretical values, usually being increased.

**Antenna Construction.** — Antenna conductors should be of weather-resistant metal, and preferably are of light weight to reduce stresses on mast and supports when subject to high winds, snow, and icing. Aluminum and aluminum alloys in tubular form best satisfy these requirements. Tubing of hard drawn copper, brass, bronze, and even of copper coated steel may be used where the weight is not objectionable. Soft-drawn copper tubing, obtainable from refrigeration supply houses, may be used for setting up experimental antennas. A dipole antenna may be made with two lengths of copper wire for the two conductors, held at the center gap with a radio antenna insulator, and similarly supported at the outer ends. The center gap of any antenna need be only great enough to allow for insulation, connection of the transmission line, and freedom from filling up with dirt.

Antenna insulation and supports must have high mechanical strength as well as good dielectric properties. These re-

quirements are satisfied by such substances as low-loss Bakelite and similar phenolic compounds, by steatite and various ceramics, by polyethylene, and under some circumstances by glazed porcelain. Experimental antennas may be set up by using wood which has been well impregnated with paraffin wax in a hot bath.

✦ **Locating and Orienting the Antenna.** — If the locality is a fringe area or if signals are known to be generally weak, it is worth while to mount the antenna at the highest practicable point. In many cases an extra five or six feet makes the difference between very poor and good reception. At the selected point it must be possible, of course, to erect a mast and needed guy wires without too much trouble. Always keep the antenna as far as possible from sources of electrical interference. Such sources include automobile traffic, public garages, buildings having many electrical machines, electric signs, X-ray and other kinds of medical laboratories. Keep away from large metal objects, such as metal roofs, gutters, and vent pipes.

When a tentative position for the antenna has been selected it is necessary to make a test of actual reception before final installation. With a transmission line of approximately correct length connected to the antenna and to the receiver, the antenna is moved about and rotated as to reception direction while observing the resulting signal strength and quality. This work is done most easily and quickly with one person moving the antenna while another watches the picture tube of the receiver. There must be some means of communication. The most popular means is a pair of self-energized phones, the kind which work without batteries or other external power, connected together by a cable running from receiver to the antenna location. Many manufacturers discourage the practice of connecting phones through the transmission line, although this sometimes is done with 75-ohm lines.

If the work must be done without the assistance of a helper it is possible to connect to the receiver, somewhere between the video detector load and the picture tube input, a high-resistance voltmeter or a microammeter which is taken to the

antenna location. Meters designed for this purpose are available. Any sensitive voltmeter or current meter may be used provided there is a rectifier and filter at the receiver end of the connecting line, so that the high-frequency signals produce a direct current or voltage in the long connecting line and the meter.

A detector probe such as used with an oscilloscope or an electronic voltmeter for high-frequency measurements is entirely satisfactory for use at the receiver. The scheme of connections is illustrated by Fig. 9-18. The high side of the probe may be connected to either the video detector load, the plate of one of the video amplifier tubes, or the control grid or

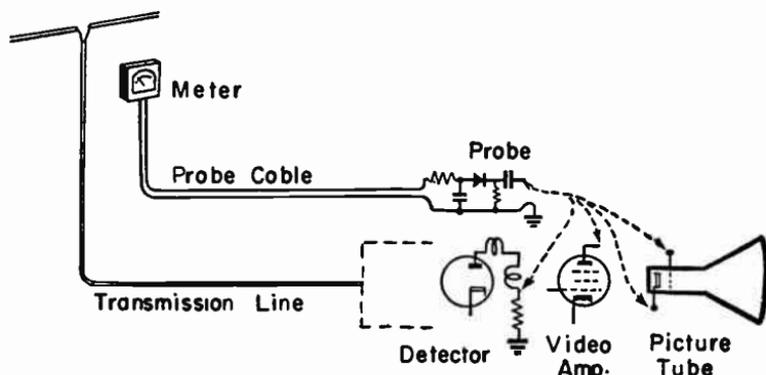


Fig. 9-18.—Connections for a meter which is carried to the antenna location during orientation.

cathode of the picture tube, whichever of these latter two elements is used for input to the picture tube in the receiver being handled. The other lead of the probe is grounded to the chassis. The cable from probe to meter may be 50 to 100 feet long, or even longer. Flexible lamp cord will do for this line, or any kind of transmission line may be used.

The receiver should be tuned to some channel, preferably one for which reception is known to be relatively weak. If a second person is watching the picture tube at the receiver the contrast control should be turned well down and all observations made on a test pattern, not a program picture. If the

meter is used, the contrast should be turned up to a usual operating position. The meter will indicate only changes in signal strength as the antenna is moved and rotated. Final inspection for picture quality must be made at the receiver with the antenna mounted in the position for greatest signal strength.

With the antenna at the position which appears to give maximum signal strength, it should be rotated one way and the other, while the conductors extend horizontally, until signal strength is at its peak. Then, if the construction allows, the antenna conductors may be inclined from the horizontal. It is possible that objects between transmitter and antenna may change the polarization of the signal away from the horizontal. If possible to make tests on more than one station, always favor the one which appears to furnish the weakest signal.

**Installation of Antennas.** — The actual mounting of the antenna mast and guy wires often calls for abilities such as possessed by a combination of carpenter, mason, plumber, electrician, and all around mechanic. It must be remembered that restrictions may be imposed on the height of masts, their location, or the overall size of arrays. Such restrictions may be in leases or in local ordinances. In addition, it is necessary to be familiar with rules of the National Electrical Code for antennas and transmission lines or "leadins." Elevated antennas and masts present lightning hazard. A lightning arrester of some type approved by the Underwriters' Laboratories should be used on the transmission line. With most types it is necessary only to clamp the arrester over the line, with or without removing some of the line insulation. If the arrester is grounded on a cold water pipe, keep the transmission line as nearly as possible at right angles to the pipe, never parallel for any distance at all. Often it is easier to drive a grounding rod down to permanently moist earth, solder a number 12 or larger copper wire to the top of the rod and run this wire to a location convenient for the arrester.

If the antenna mast is metal, as usually is the case, it should be connected through a heavy copper wire to a cold water pipe

or to the grounding rod. The charge accumulated on an ungrounded mast during dry summer weather, even when there is no lightning, is enough to give a shock that stings even though it is not dangerous.

If the antenna mast is 10 feet or more in height it should be supported by three or four equally spaced guy wires. A swaying antenna causes picture brightness to vary, and signal strength may drop enough to cause loss of synchronization. If guy wires have to come within one wavelength or less of the antenna conductors, one or more radio antenna insulators should be inserted in the guys somewhere within the first three or four feet from the mast.

**Built-in and Indoor Antennas.** — Many television receivers are equipped with antennas built into the cabinet or sup-

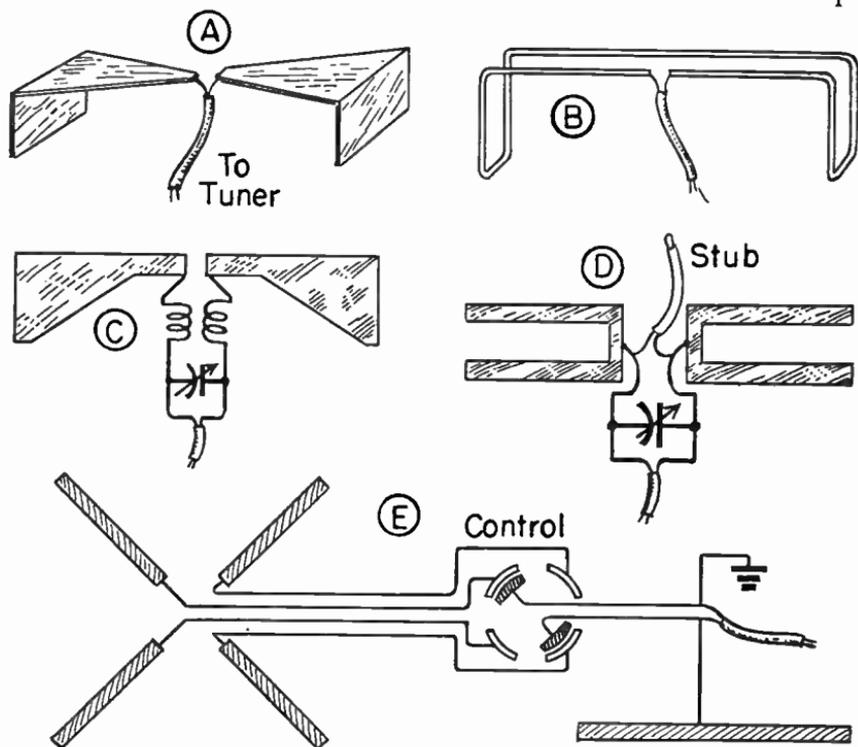


Fig. 9-19.—Various styles of built-in antennas.

ported from the chassis. Built-in antennas allow satisfactory reception only in localities where carrier signals are strong. A simple construction, shown at *A* of Fig. 19, consists of two sheets of metal foil, of triangular shape, attached inside the cabinet top and extending part way down on opposite sides. Folded dipoles may be used as at *B*, often being made of transmission line shorted at the outer ends and opened at the center of one side for transmission line connection. Elements of built-in antennas may be mounted on fibre which is tacked or stapled to the cabinet. Antennas mounted in top and bottom of a console may be stacked.

Some built-in antennas must be tuned to a desired channel or band by an adjustable capacitor and a small transformer as in diagram *C*. Resonant line stubs may be added, as at *D*. Inductors, stubs, and capacitors add to electrical length of antenna conductors for low-band reception, or shorten the length for high-band reception. Tuning knobs may be on the front, side, or rear of the cabinet.

Unless a built-in antenna is so designed that it may be rotated, either physically or electrically, for direction of best reception, it may be necessary to turn the cabinet to obtain satisfactory signal pickup. One of many methods of electrical rotation is illustrated at *E* in Fig. 9-19. There are four stationary antenna conductors in the form of a cross connected to four capacitor stators in the control. Two rotors in the control may be turned for capacitive coupling to pairs of opposite stationary conductors or to adjacent pairs. The control rotors connect to a counterpoise conductor and ground, and to the tuner of the receiver.

In some receivers a built-in antenna is used only for high-band reception, with low-band signals taken through blocking capacitors from the power cord where it enters the chassis. This method is found also in some radio broadcast receivers. Since the power cord acts as the low-band antenna, the cord must not be coiled, but laid on a fairly straight line and in position or direction for good reception.

Reception with built-in antennas is weakened by any metal

objects near the receiver, whether these objects are stationary, like a radiator, or movable like a lamp on the cabinet top. Reception varies with location of the receiver in a room, usually being best when near a window. Both leads from a built-in antenna must be disconnected from the tuner when the receiver is used with an outdoor antenna.

Indoor antennas may stand on or near the receiver cabinet, or be mounted on a wall or in an attic. Reception is satisfactory in areas of fairly high signal strength when the building is not of steel framework construction. Indoor antennas seldom are satisfactory under metal roofs or where there is metal foil heat insulation. Indoor antennas must be kept away from plumbing and vent pipes, from metal beams, and even from such things as metal curtain hangers.

Any indoor or built-in antenna is handicapped by lack of enough elevation to receive signals unaffected by all manner of obstructions in the wave path. From this standpoint an antenna located in an attic is likely to be more satisfactory than one placed near the receiver.

**Multiple Receiver Connections.** — What probably is the simplest method of connecting two receivers to a single antenna

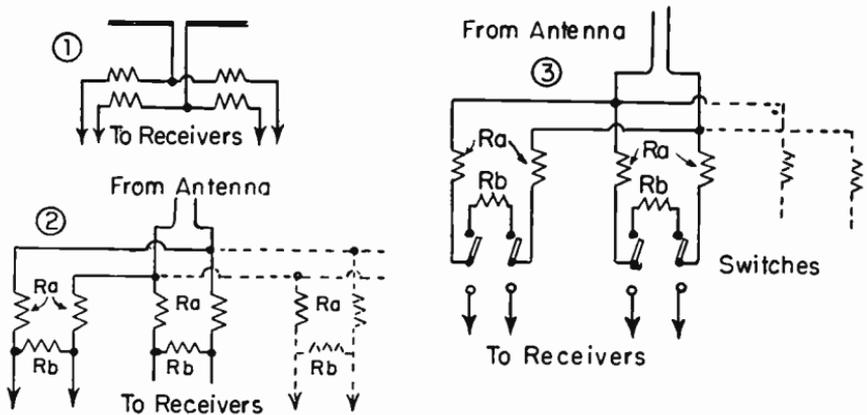


Fig. 9-20.—Connections for two or more receivers on a single antenna.

or transmission line is shown at 1 of Fig. 9-20. Both receivers remain permanently connected. With the usual 300-ohm impedance at receiver inputs, the resistors should be of 330 ohms

each. Either pair of resistors may be omitted and one receiver connected directly to the transmission line if this does not allow objectionable interaction between r-f oscillators in the two sets.

A method designed to preserve impedance relations between several receivers and the transmission line is shown by diagram 2. For two receivers of 300-ohm input impedance, resistors  $R_a$  would be 220 ohms each and  $R_b$  470 ohms. For three receivers each of the three resistors would be 390 ohms. For four receivers, resistors  $R_a$  would be 560 ohms each and  $R_b$  330 ohms.

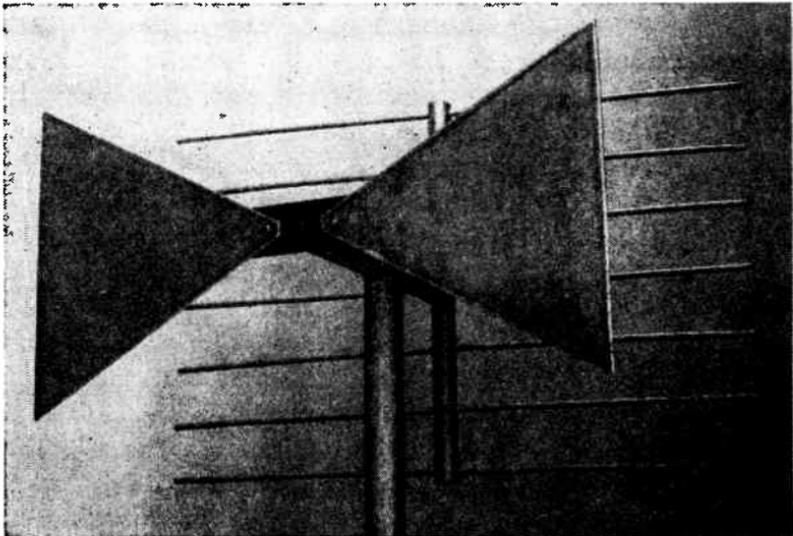
In diagram 3 two-pole double-throw switches allow removal of one or more receivers from the circuit. Assuming receivers with 300-ohm input impedance, resistors  $R_b$  always would be 300 ohms each. For two receivers, each  $R_a$  resistor would be 150 ohms, or 300 ohms for three, and 470 ohms for four receivers.

Resistors should be half-watt carbon or composition types of 10 per cent tolerance. The resistance pad for each set should be mounted close to the antenna input of that receiver. If any sets are transformerless types with hot chasses, use series mica or ceramic capacitors of at least 1,000 mmf in both signal leads to such receivers. Antenna output signal strength divides between connected receivers, causing less and less input as the number of sets and pads is increased. Inputs are still further reduced by passage of signals through resistors of the pads.

Multi-connectors are available in which coupling is by means of high-frequency transformers instead of with resistance pads. The transformers greatly lessen attenuation of signal strength, as compared with resistors, also reduce interference due to r-f oscillator frequencies passing from one to another of the connected receivers. Transformer connectors are to be preferred in localities where received signals are of only moderate strength, or are weak.

**Uhf Antennas.** — The short wavelengths in the uhf television band allow antennas less than two feet wide. Small size and light weights of uhf antennas allow convenient stacking for increased gains. Among the uhf types of design generally similar to earlier styles for the vhf band are conicals, plain V-styles, V-beams, folded dipoles, and yagis. An antenna having element dimensions suitable for the vhf band seldom allows satisfactory uhf reception unless received signals are strong and reflections few.

Signal reflections and resulting ghost images are more troublesome at ultra-high than at very-high frequencies, because the shorter waves are reflected more completely and with sharper directivity. This means that uhf antennas should have better directional properties and narrower reception lobes than vhf varieties. Location and orientation are important. A change of less than a foot horizontally or in elevation, or rotation through only a few degrees may greatly improve reception. It follows that supports or guying must be such as prevent swaying if reproduction is to remain steady.



*Fig. 138-20.—A uhf fan antenna with reflector.*

A number of antenna forms are better adapted to uhf than to vhf signals, chiefly because size would be excessive at the lower carrier frequencies. Among these are the fan, the rhombic, and the corner reflector types. A fan antenna, as pictured by Fig. 138-20 consists of two oppositely placed triangular sheets of metal, which may be solid or may be perforated to decrease wind resistance. The fan antenna, like most others for uhf signals, nearly always is used with a reflector. The reflector may be a number of parallel rods, as illustrated, or it may be a wire mesh screen supported in a vertical plane.

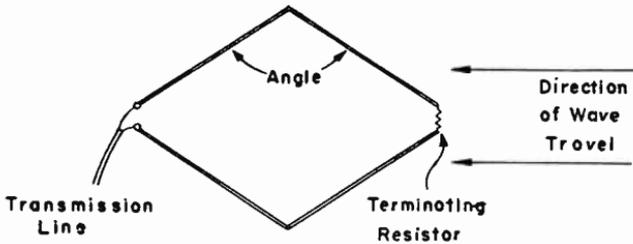


Fig. 138-21.—The form of a rhombic antenna, as seen from above or below.

A fan antenna without reflector provides moderate gain, a maximum of possibly 4 decibels over a plain dipole, but the gain is doubled or more than doubled by a reflector. Stacking two bays gives an additional 2 to 3 decibels of gain. The response may be broadly peaked around 700 to 800 mc, but does not fall off a great deal at lower and higher frequencies.

The rhombic antenna, looking from above, is of the form shown by Fig. 138-21. The four legs are of equal lengths, and each should be at least two wavelengths long at the lowest received frequency. Since one wavelength in air at the frequencies in uhf channel 14 is approximately 25 inches, each leg of the rhombic would be at least 50 inches long. The angle between opposite legs is something between 100 and 120 degrees, with the greater angles used when the legs are more than two wavelengths long.

The terminating resistor provides the effect of a reflector, making the antenna unidirectional in receiving characteristics, whereas omission of the resistor allows reception almost equally well from two opposite directions. This resistor must be of carbon or other non-inductive type. The value usually is between 400 and 500 ohms when a 300-ohm transmission line is used, although, in theory, the terminating resistance and line impedance should match.

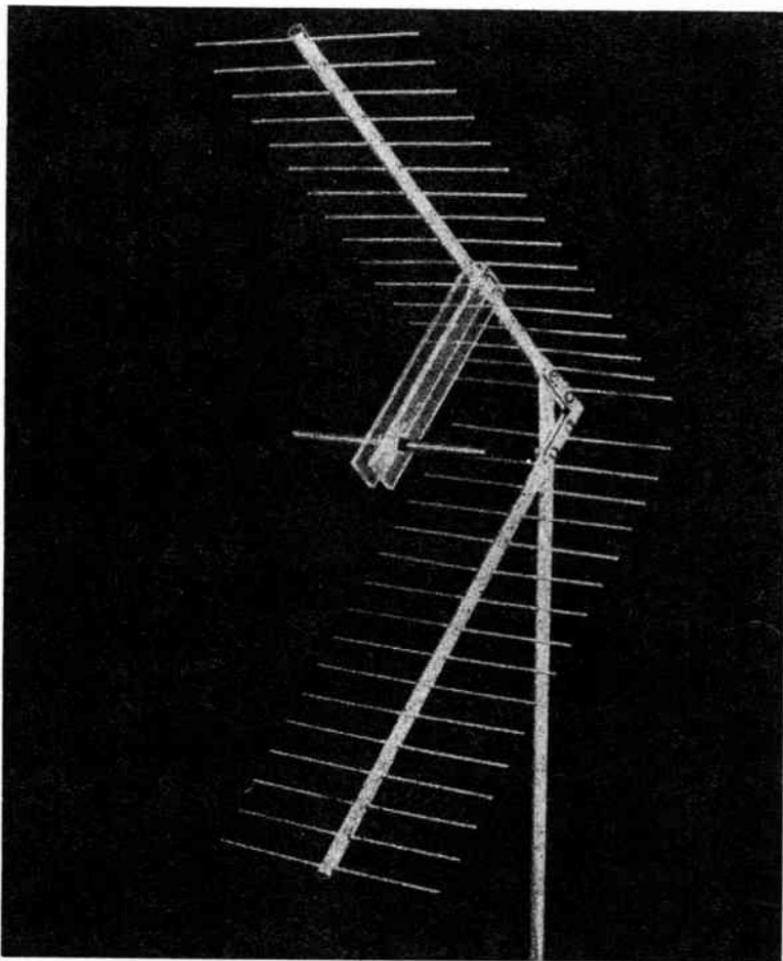


Fig. 138-22.—A corner reflector uhf antenna with a plain dipole as the active element.

Compared with a plain dipole, the rhombic antenna provides gain increasing from about 5 decibels at lowest uhf carrier frequencies to around 10 decibels at the high end of the band. Stacking of two bays, with half-wave vertical spacing, increases the gain by about two decibels.

A corner reflector uhf antenna is illustrated by Fig. 138-22. The reflector consists of a series of rods or spines arranged in two planes coming together at the back corner. The corner angle usually is approximately 90 degrees. Smaller angles allow somewhat greater gains, but only when the two sides of the reflector are lengthened at the same time.

The receiving element shown in the picture is a plain dipole. For more uniform response over the entire band of frequencies this element may be a small fan with the triangles bent at the same angle as the reflector corner, or it may be a folded dipole of dimensions suitable for the uhf band.

The corner reflector antenna is almost completely unidirectional. A single bay provides gain increasing from 7 or 8 decibels at the low end of the uhf band to about 13 decibels at 800 to 900 mc. This is greater gain than obtainable with other uhf antennas in general use. Two similar bays may be stacked for added gain, or two similar elements may be used with one large reflector.

## B

**BALLASTS.** — A ballast is a resistor which changes its value with variations of current or voltage in such manner as to partially compensate or reduce the effects of changes in current or voltage. Ballasts in television receivers most often are used in series with series-connected tube heaters. The purpose is to prevent excessive surges of current while heaters still are cold, and to reduce the effects of line voltage fluctuations.

These line voltage ballasts usually are of iron or iron alloy wire whose resistance increases sharply upon reaching a dull red heat. Excessive current, as may be caused by excessive voltage, results in an increase of ballast resistance and consequent limiting of current.

Some ballast resistors are enclosed within a glass bulb, like a tube, and have a tube base and pins. Others are enclosed within a perforated steel housing, usually cylindrical, fitted with an insulating base and pins like those of a tube. A single ballast unit may contain resistors for other than heater circuits, possibly for rectifiers and for plate circuits of power amplifiers.

**BIASING.** — Grid bias is the average potential difference, measured in volts, between control grid and cathode of a tube. With the grid negative in respect to the cathode there is negative bias. With the grid positive there is positive bias. With both elements at the same potential there is zero bias. A more negative grid bias allows less average plate current with given applied plate voltage. A grid sufficiently negative to reduce average plate current to zero is biased to plate current cutoff.

Variation or adjustment of grid bias serves two principal purposes; it determines whether or not there shall be grid current in the grid circuit of the tube, and it fixes the point of average operation on the curve showing relations between grid voltage and plate current. There will be no grid current

when the bias is appreciably more negative than zero. If the bias is very close to zero on the negative side, or is zero, or is positive, there will be grid current which becomes greater as the grid is made more positive. Grid current means that power is being taken from the source of grid signal voltage, which is undesirable in voltage amplifiers but may be allowed in some power amplifiers. To prevent flow of grid current the d-c value of negative grid bias must be equal to, and in practice must slightly exceed, the maximum peak potential of the signal applied to the grid. This peak is equal at least to 1.4 times the effective a-c value of the signal.

Fig. 10-1 illustrates effects of changing the grid bias to change the operating point on the grid-voltage plate-current curve. At the left the bias is sufficiently negative to hold the grid signal voltage on the straight portion of the curve. Plate

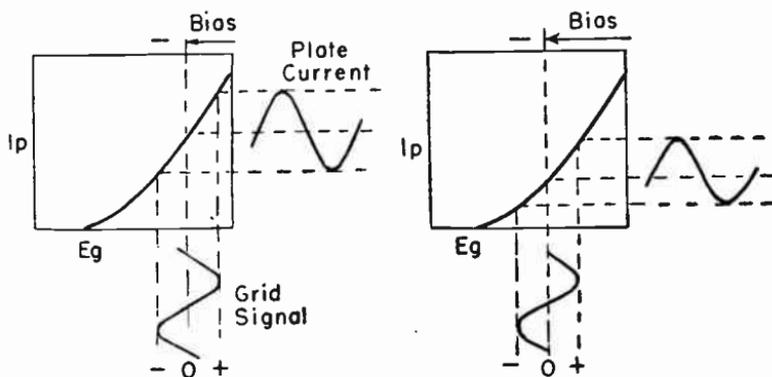


Fig. 10-1.—Varying the bias changes the operating point on the grid-voltage plate-current curve.

current changes have the same waveform as grid signal or grid voltage changes, there is no distortion of waveform. At the right the bias has been made more negative, and operation is on the lower bend of the curve. Now the plate current waveform is not like the waveform of grid signal voltage, there is greater increase than decrease of plate current with equal variations of grid voltage. Actual grid voltage for zero signal amplitude always is equal to the bias voltage. One example of such

biasing is found in some sweep amplifiers, where adjustable bias is used for linearity control.

At the left in Fig. 10-2 the grid bias is sufficiently negative to cause plate current cutoff when no signal voltage is applied to the grid. Pulses of plate current now flow only during positive swings of the grid signal. Plate current remains zero during negative swings of the signal. Were the bias slightly less negative than the cutoff value there would be some plate current during portions of the negative swings of grid signal.

If plate voltage is made very small there is limiting of maximum plate current due to saturation. That is, the low voltage is capable of drawing only a limited part of the electron emission from cathode to plate. The  $I_p$ - $E_g$  curve then

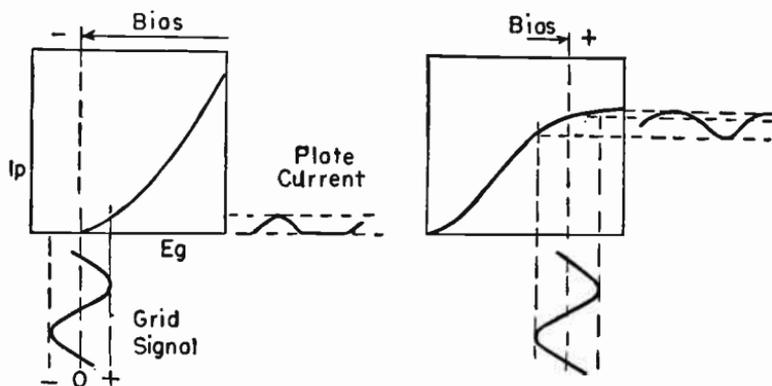


Fig. 10-2.—Biasing for plate current cutoff and for plate current saturation.

is about as shown at the right in Fig. 10-2. Even though the bias is made near zero or slightly positive, the plate current cannot be increased. Then the plate current waveform shows rather flat tops, the effect is somewhat similar to but is inverted as compared with what happens at the right in Fig. 10-1 and at left in Fig. 10-2. An example of distorted output as shown in Fig. 10-2 is found in some limiter and clipper tubes in the sync sections of television receivers.

Four different methods of biasing are found singly and in various combinations in television receivers. They are:

- a. Cathode bias, sometimes called self-bias.

b. Fixed bias secured from the B-power supply, or a partially fixed and partially variable bias sometimes called back-bias.

c. Bias by grid rectification or with grid leak and capacitor.

d. Bias furnished wholly or in part from the automatic gain control voltage.

**Cathode Bias.** — The principle of cathode bias is illustrated by Fig. 10-3. Between the cathode and B- or ground is a biasing resistor  $R_k$ . All d-c electron flow in the cathode must pass through this resistor. This electron flow will be the sum of the plate current, the screen current if the tube has a screen, and any grid current which may exist. The electron flow in  $R_k$  is in such direction that the grounded or B- end of this resistor is negative with reference to the cathode end. If the control grid is negative and there is no current in parts of the grid circuit such as  $L_g$  and  $R_g$  there will be no difference of potential between the grid and ground or B-. Since the grid is connected to ground or B- through the grid return, grid potential will be negative with reference to the cathode. This potential difference, which is the grid bias, is equal to the average voltage drop in  $R_k$ . Resistance for  $R_k$  is determined thus:

$$R_k, \text{ ohms} = \frac{1000 \times \text{required bias volts}}{\text{total cathode current, milliamperes}}$$

Because bias voltage depends on cathode current every variation of this current varies the bias. Every increase of plate current, screen current, or their sum increases the voltage

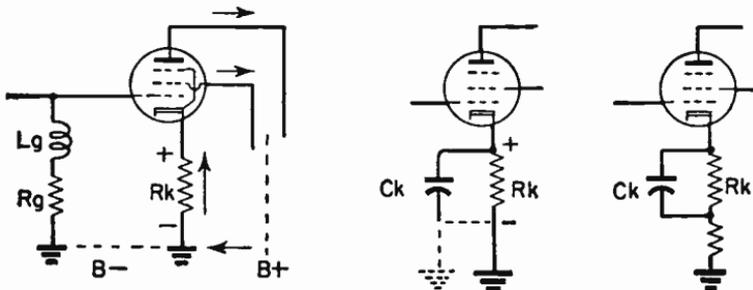


Fig. 10-3.—The principle of cathode bias.

drop in  $Rk$  and makes the bias more negative. Every decrease makes the bias less negative or more positive. Increases of plate and screen current accompany positive swings of the grid signal, and the positive swings thus make the bias more negative. Decreases of current accompany negative swings of signal, and the grid is made less negative. Consequently, the changes of bias always tend to counteract the changes of signal, and there is degeneration due to the negative feedback effect.

Cathode bias helps compensate for the differences between plate currents of original and replacement tubes, because more plate current makes the bias more negative and less plate current makes it less negative. There is small possibility of the grid ever becoming positive with this form of bias, and it is permissible to use greater values of resistance in the grid circuit than with fixed bias. Cathode bias is the kind nearly always used for power amplifiers giving high output voltages, and is used for voltage amplifiers as well.

To reduce the degenerative effect of cathode bias and allow more nearly the maximum possible gain from the tube, the biasing resistor  $Rk$  may be bypassed with a capacitor  $Ck$  as at the right in Fig. 10-3. When the chief purpose of the bypass capacitor is to lessen degeneration the capacitive reactance in ohms should be little if any more than 10 per cent of the ohms resistance in  $Rk$ , at the lowest frequency to be handled by the tube. In audio amplifiers and in cathode-biased sync and sweep amplifiers this calls for large capacitance. For example, with 300 ohms at  $Rk$  and 30 ohms reactance at  $Ck$ , the capacitance for low frequency of 60 cycles per second would have to be about 90 microfarads. For the same biasing resistance at intermediate frequencies it would be sufficient to use 250 to 300 mmf, and at carrier frequencies the capacitance would have to be only about 100 mmf. If a power supply is poorly filtered it would be necessary to bypass the 60-cycle or 120-cycle "hum" voltages even in high-frequency amplifiers. Sometimes only part of the biasing resistance is bypassed, as in the right-hand diagram.

Unless the cathode resistor is bypassed with capacitive reactance very small for the frequency considered, the cathode is not at r-f ground potential. Then it is desirable to have the least possible capacitance, stray and otherwise, between plate and cathode, also that the screen be bypassed to ground with capacitance whose reactance is low at the operating frequency.

**Fixed Bias.** — The principle of fixed bias is illustrated by Fig. 10-4, where the negative side of the B-power supply is shown at the left, with two amplifiers and their grid connections at the right. Between the center-tap plate connection of the power rectifier and ground are the two resistors  $R_a$  and  $R_b$  in series. All electron flow in this B-supply must pass through these resistors, in the direction shown by the arrow. The resistors form a voltage divider. The most negative point

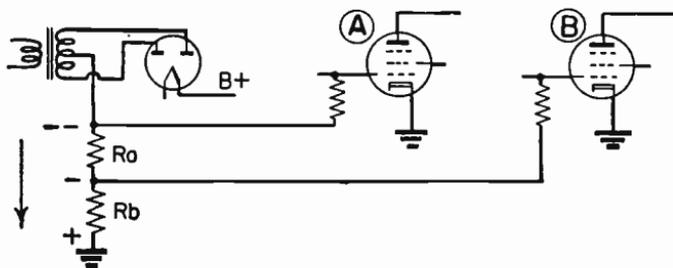


Fig. 10-4.—The principle of fixed bias.

is the top of resistor  $R_a$ , to which is connected the control grid of amplifier  $A$ . The cathode of this amplifier and the bottom of  $R_b$  are connected to the ground. Then the potential difference between grid and cathode of the tube is the same as the difference between the top of  $R_a$  and the bottom of  $R_b$  or ground. The grid bias for tube  $A$  is the total potential difference across the voltage divider. The grid of amplifier  $B$  is connected to a point between the two divider resistors, and the cathode to ground. Then the grid-cathode potential difference, or grid bias, on this tube is equal to the voltage drop across divider resistor  $R_b$  and is less than the bias for tube

A. By suitably proportioning the resistances at  $R_a$  and  $R_b$  the bias voltages may be made anything desired, and additional bias voltages may be provided by using more divider resistors.

Actual wiring for fixed bias usually will include some or all of the decoupling capacitors and resistors shown by Fig. 10-5. Across grid resistors  $R_g$  appear the r-f signal voltages for the control grids. Accordingly, these resistors are not bypassed. Near the grid resistors and tubes will be decou-

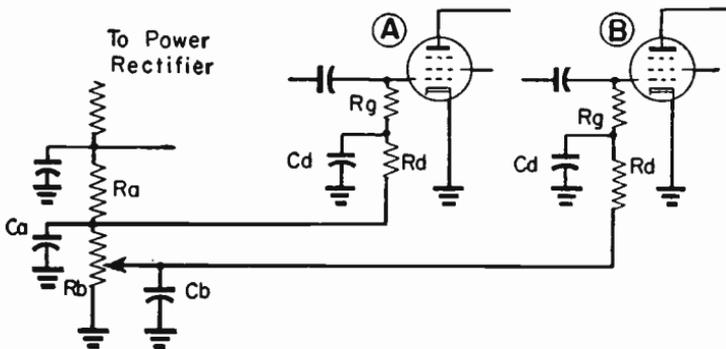


Fig. 10-5.—Fixed bias with decoupling capacitors and resistors.

pling capacitors  $C_d$ . Decoupling, or avoidance of interstage coupling and feedback, is made more certain by using decoupling resistors  $R_d$  near the tubes. To prevent interstage coupling in the d-c grid return lines, which may be rather long, additional decoupling capacitors  $C_a$  and  $C_b$  may be placed at the voltage divider, one from each return line to ground. The values of decoupling capacitances are based on the resistances and on the operating frequencies, as previously explained.

It must be noted that in Figs. 10-4 and 10-5 the current in the divider resistors includes not only bleeder current such as exists in any power-supply voltage divider, but also the cathode currents of the tubes. It is plain, from Fig. 10-4, that plate and screen circuits extend to B+, through the power rectifier, the voltage divider resistors, and ground back to

the cathodes. If most of the divider current is due to tube cathode current, variations of plate and screen currents which accompany variations of signal amplitude will vary the grid biases because they vary the divider current and voltage drops. Unless there is effective bypassing at the divider, the changes of grid bias or grid voltage due to signals in one tube may act on the grids of this tube and others too. This is a feedback. If feedback voltages to grid aid the signal voltages on the same tube there is regeneration and possible oscillation. If the feedback opposes the signal voltages there is degeneration and loss of gain.

If only a small portion of the divider current is tube cathode current, and most of it is bleeder current, changes of cathode current due to signals will have little effect on bias or grid voltage. With reasonably good bypassing there will be little danger of either regeneration or degeneration.

Fixed bias may be secured from a separate bias rectifier, one which is not part of the plate and screen power supply, and which does not carry cathode currents. Fig. 10-6 shows

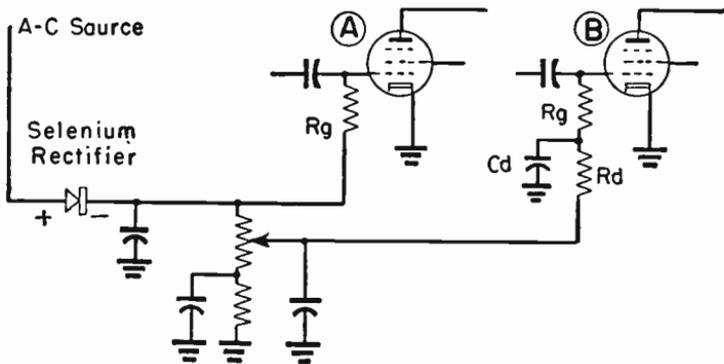


Fig. 10-6.—Fixed bias from a rectifier used only for biasing voltage.

such a system, where the rectifier is a selenium type. Any other kind of rectifier, either half-wave or full-wave, might be used provided it were capable of carrying the small current needed for biasing voltage drop. The a-c source may be anything providing a peak potential somewhat higher

than the maximum bias voltage to be provided. Heater transformers or circuits may act as this source. In Figs. 10-5 and 10-6 the biasing voltage for amplifiers *B* is adjustable, taken from the slider of a potentiometer in the divider string. With fixed bias on a tube the maximum resistance of the grid resistor usually may be only about half the maximum permitted with cathode bias.

**Grid Leak Bias.** — Fig. 10-7 illustrates the principle of grid leak bias, which is called also grid leak-capacitor bias or grid rectification bias. The coil at the left in the diagrams represents any source of signal voltage. Only the control grid and cathode of the tube are shown, because these are the only elements having anything to do with the biasing action. The tube might be a triode, a pentode, or a beam power tube. It might even be a diode, but then the diode plate would act like the grid of the other tubes.

In diagram *A* an alternating signal potential exists across the source and is applied through capacitor  $C_g$  between the grid and cathode of the tube. During each positive alternation of the signal voltage the grid is made positive with reference to the cathode, and there are pulses of electron flow as indicated by arrows. There can be no pulse during negative alternations of the signal, because these make the grid negative with reference to the cathode and there is no electron flow to a negative element.

As in diagram *B*, the continued pulses of electron flow add electrons to the side of the capacitor which is toward the grid. This side of the capacitor accumulates a negative charge, and because this negative side is directly connected to the grid, the grid becomes negative with reference to the cathode. The excess of electrons cannot escape from the negative side of the capacitor, for the electrons would have to flow from grid to cathode in the tube and they cannot flow from a negative to a positive element. The negative charge will build up until it equals the peak potential of the applied signal, whereupon the electron flows will cease and the grid will be highly negative.

In diagram *C* a grid resistor  $R_g$  has been connected between the grid and ground or between the grid and the cathode by way of ground. Some of the surplus electrons from the negative side of capacitor  $C_g$  now flow through the grid resistor as shown by long arrows. With this direction of electron flow the grid end of the resistor is negative and the

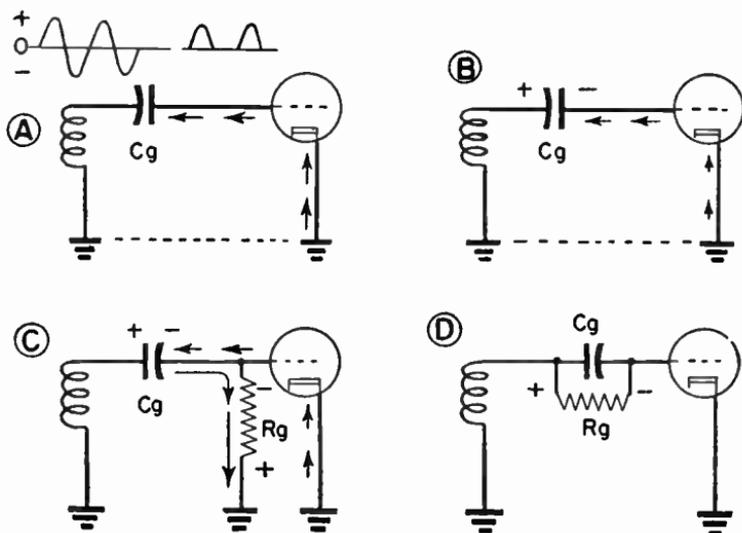


Fig. 10-7.—Principle of grid leak bias or of bias by grid rectification.

ground or cathode end is relatively positive. The potential difference between grid and cathode is the potential difference across  $R_g$ . This is a negative grid bias. Electrons, in pulses due to the signal, flow toward the capacitor while a steady electron flow away from the capacitor goes through the resistor.

The charge which remains on the capacitor varies with signal amplitude, being greater as the signal strength increases. The capacitor voltage varies likewise. The rate of electron flow through the resistor depends on capacitor voltage, or signal strength, and on the value of resistance in  $R_g$ . The greater the resistance the less will be the flow rate, the greater will be the charge held on the capacitor, and the

more negative will be the grid bias. Grid leak bias thus automatically adjusts itself to signal strength, and tends to maintain a constant output from the plate circuit of the tube when the signal varies. This feature is useful in sync amplifiers where it helps eliminate the effect of "noise" irregularities in the applied signal.

If the d-c grid return can be completed back to ground or cathode through the signal source, the biasing capacitor and leak may be in parallel with each other, as shown by diagram *D*. This simply connects the positive end of the resistor to ground or cathode through the conductors of the source.

In order that the grid leak-capacitor combination may provide bias voltage there must be a current in the resistor. This current must consist, on the average, of electron flow from the grid. This necessary grid current means that some power is taken from the signal source during positive alternations of the signal, but not during negative alternations. Thus the signal is distorted. This type of bias is not useful where the original waveform must be preserved without distortion, but it is useful where only the frequency rather than the waveform of pulses is of importance. This latter condition exists in most sync and sweep circuits. Once the bias is established by a capacitor charge which is added to and subtracted from equally, by signal pulses and leak current, the signal voltages pass through the capacitor and vary the grid voltage above and below the bias voltage.

With a grid leak-capacitor combination as the only source of bias voltage, the bias becomes zero in the absence of a signal. Then, unless there are other means of limiting plate current, this current may increase to a value harmful to the tube. Grid leak bias is used for limiters and clippers in the sync section, it may be used where detector action is wanted, it may be used with class C and other power amplifiers where grid current is permissible, and it is used with oscillators.

**BLANKING, RETRACE**

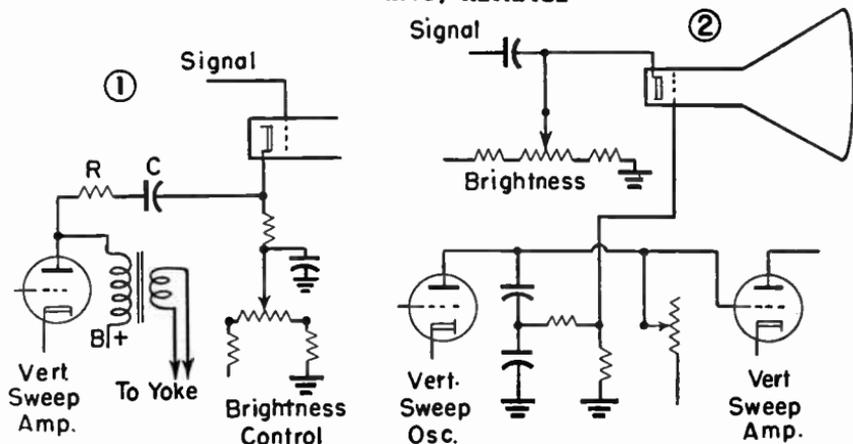


Fig. 10-8.—Connections for blanking of vertical retrace lines.

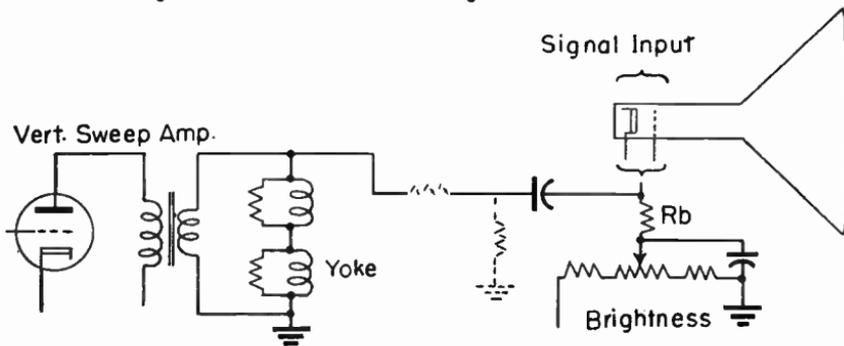


Fig. 10-9.—Blanking pulses from vertical deflecting coils in a yoke.

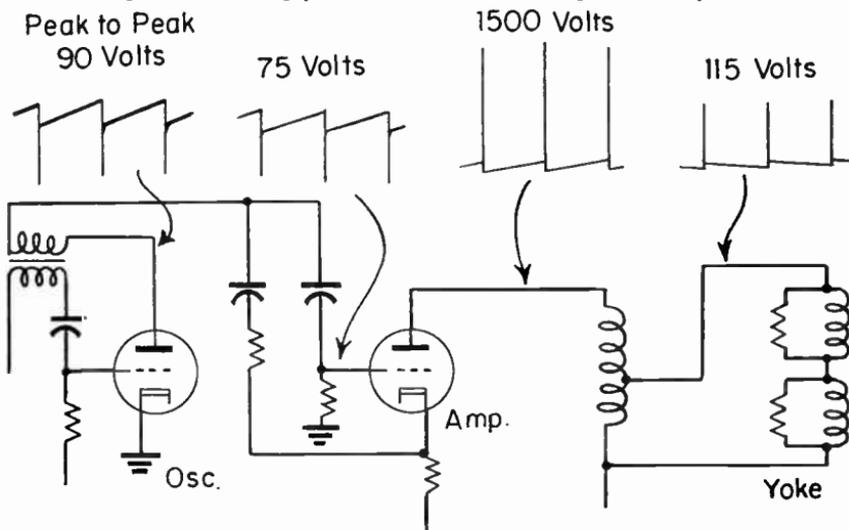


Fig. 10-10.—Waveforms in a vertical sweep section.

**BLANKING, RETRACE.** — Advancing the brightness control or retarding the contrast tends to cause narrow white lines sloping upward from left to right on pictures. These lines are produced as the electron beam sweeps upward during vertical retrace intervals. Usually the beam is cut off or blanked during vertical retrace by applying a pulse of negative voltage to the control grid, or of positive voltage to the cathode of the picture tube.

Diagram 1 of Fig. 10-8 shows how positive pulses from the plate of a vertical sweep amplifier may be applied through a dropping resistor  $R$  and blocking capacitor  $C$  to the picture tube cathode. In diagram 2 negative pulses from the plate of a vertical sweep oscillator are taken through a capacitor and resistance voltage divider to the control grid of the picture tube.

In Fig. 10-9 the blanking pulses are taken from the vertical coils of the deflecting yoke or from the output transformer. At the side of an output transformer not connected to ground or to B-plus, pulses may be positive or negative, depending on circuit design. An isolating resistor  $Rb$  is between the pulse line and the brightness control.

Fig. 10-10 shows negative and positive pulse waveforms at several points in a sweep section employing a blocking oscillator. Waveforms having a sawtooth portion may be reduced to sharp pulses by a differentiating filter (series capacitor and shunt resistor) in the lead to the picture tube. The filter time constant is such as to leave pulses wide enough for blanking, but not wide enough to cut off parts of pictures. Voltage dividers or dropping resistors reduce strong pulses to voltages suitable for trace elimination.

**BRIGHTNESS AND BRIGHTNESS CONTROLS.** — Average brightness of pictures corresponds to average current in the electron beam. Variations of beam current, resulting from changes of grid-cathode voltage at the picture tube, make some areas of pictures either lighter (more beam current) or darker (less beam current) than the average.

As may be seen from Fig. 11-1, curves showing the relation between control grid voltage and brightness on the screen of a picture tube have much the same shape as curves showing the relation between grid voltage and plate current of an amplifier tube. Just as average plate current depends on the bias-

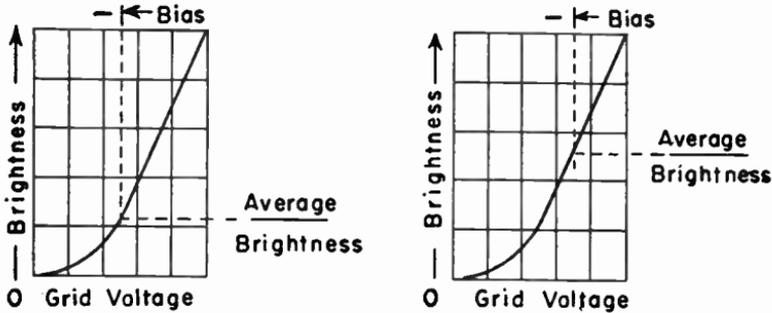


Fig. 11-1.—Relations between control grid voltage of a picture tube and average brightness of pictures or patterns.

ing of an amplifier, so average brightness of a picture or pattern depends on biasing of the picture tube with reference to the cathode. Making the picture tube grid bias less negative allows greater average brightness, which means greater beam current and also greater current in the high-voltage anode of the picture tube.

**Brightness Controls.** — Brightness controls, which usually are arranged for manual adjustment by the set operator, vary the control grid bias or the average potential difference between control grid and cathode of the picture tube. These controls are called also brilliancy controls, intensity controls, or background controls.

Fig. 11-2 shows connections for a typical brightness control where signal input is to the control grid of the picture tube.

Bias voltage is provided by potential drop in resistor  $R_a$ , the control potentiometer, and resistor  $R_b$ , which are in series between  $B+$  and ground or  $B-$ . The control grid of the picture tube connects to ground or  $B-$  through resistors  $R_c$  and  $R_d$ . Because of the direction of electron flow through the bright-

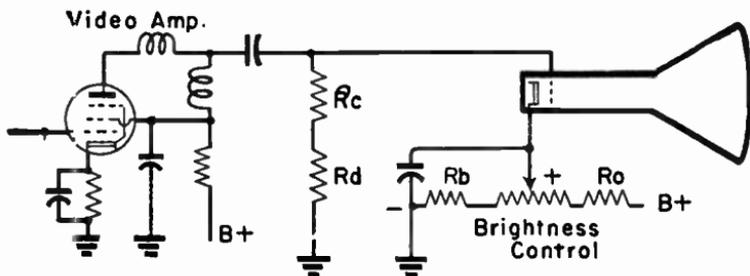


Fig. 11-2.—Brightness control used when signal input is to the picture tube control grid.

ness control resistors, the point connected to the cathode is positive in relation to the end which connects indirectly to the grid. Consequently the grid is negative with reference to the cathode, or has a negative bias whose value is varied by the control potentiometer.

Fig. 11-3 shows connections for a brightness control where signal input is to the cathode of the picture tube. Again the control potentiometer is in a resistor string between  $B+$  and ground, and again the point which connects to the cathode (through resistor  $R_b$ ) is positive with reference to the end which connects through ground to the control grid. Thus the

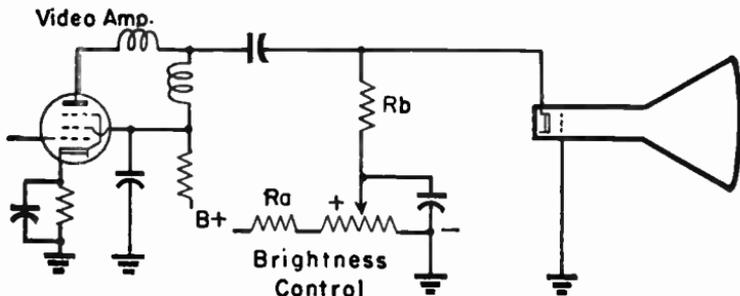


Fig. 11-3.—Brightness control when signal input is to the picture tube cathode.

grid is provided with negative bias voltage of a value which may be varied by movement of the slider on the control potentiometer.

Controls for brightness and for contrast always must be adjusted to suit each other. Too much or too little brightness for any given degree of contrast will cause poor pictures. Too much or too little contrast for any given setting of the brightness control also will cause poor pictures. It must be kept in mind that a contrast control is really a gain control, the higher the setting of the contrast control the greater is the amplification of the picture signals. A contrast control for television pictures is like a volume control for sound. It must be kept in mind too that a brightness control is a bias control, it fixes the operating point on the curve relating brightness and control grid voltage in the picture tube. Just as the bias of an amplifier tube must be suited to the amplitude or strength of the audio or radio-frequency signal, so must the bias (brightness) of a picture tube be suited to the strength (contrast) of the picture signals. The relations of these two controls are illustrated by Figs. 11-4 and 11-5.

At the left in Fig. 11-4 the brightness control is set too high for the strength (contrast) of the signal, the picture tube grid bias is not sufficiently negative. The entire signal acts high

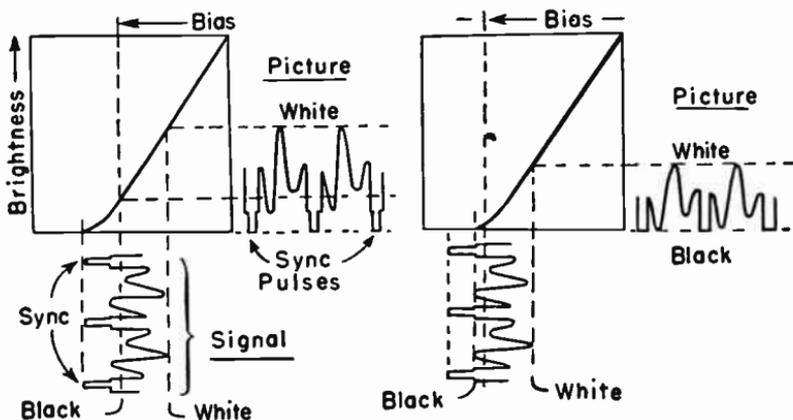


Fig. 11-4.—Left: Brightness control too high or contrast control too low. Right: Brightness and contrast correct for a weak signal.

up on the curve of brightness versus grid voltage. The intensity of the beam is varied not only by the picture elements of the signal, but also by the sync pulses. The black level of the signal is well above the cutoff point for beam current, so portions of the picture which should appear black actually are gray. The picture will appear washed out, and the blanking intervals and sync pulses will cause diagonal white lines.

These picture faults may be corrected, without changing the strength of the signal, by decreasing the brightness as shown at the right in Fig. 11-4. Decreasing the brightness makes the bias more negative. This brings the signal far enough down on the curve that the black level is at or near the point of beam cutoff. Since beam cutoff leaves the screen without illumination, the parts of the picture which should appear black actually will be black. The picture then has correct relative shading or relations of blacks, grays, and whites.

Average illumination of the picture with conditions as at the right is much lower than with them as at the left. With a signal of the strength indicated there would be no way of having a brighter picture while retaining correct shading. A signal which is weak at the input to the picture tube cannot be made to produce a very bright picture of good quality.

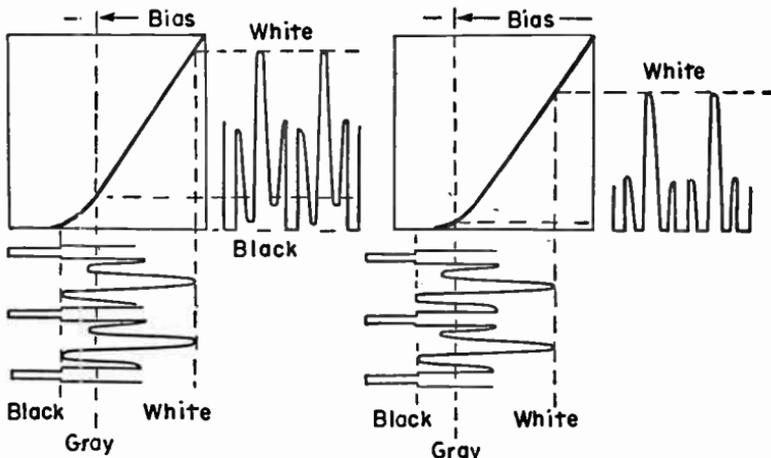


Fig. 11-5.—Left: Brightness and contrast correct for a strong signal.  
Right: Brightness control too low for a strong signal.

If we wish to retain the brightness setting at the left in Fig. 11-4, and have a good picture, the signal must be strengthened as at the left in Fig. 11-5. This can be done by advancing the contrast control, provided the contrast control is not already all the way advanced. With the stronger signal the black level is at or near the point of beam cutoff. All of the sync pulses are cut off, but in the picture there is a full range of shading from black through intermediate grays, to white. The picture is bright because the average level of illumination can be maintained high because the signal is strong.

At the right in Fig. 11-5 we are using the strong input signal with the brightness adjusted too low. The signal rides far down on the curve, due to the grid bias being too negative. Now it is not only the sync pulses which are cut off, but also a considerable portion of the signal lying above the black level. Parts of the signal which should produce gray tones in the picture are brought down to the point of beam cutoff, and these parts of the signal become black. Everything which should be black in the picture will, of course, appear black because it will be below the point of beam cutoff. The result is too much black, or too much contrast in the picture. The appearance will be mostly black and white without smooth gradation of intermediate grays.

## C

**CAPACITANCE AND CAPACITORS.** — A capacitor consists of two insulated conductors which may be charged with electrons through connections to a source of voltage and which may be discharged through any conductive circuit. Capacitance is a measure of the ability of a capacitor to receive a charge of electrons or electricity, it is the ratio of the number of coulombs of charge to the number of volts potential difference between the conductors of the capacitor. The two conductors are called the plates and the insulation between them is called the dielectric of the capacitor. A capacitor may be called an electrostatic condenser or simply a condenser. Ca-

capacitance may be called electrostatic capacity or simply capacity.

The fundamental unit of capacitance is the farad. If each capacitor plate will carry a charge of one coulomb when their potential difference is one volt the capacitance is one farad. Practical units are the microfarad, abbreviated *uf*, *mf*, or *mfd* and equal to one one-millionth of a farad, and the micro microfarad, abbreviated *uuf*, *mmf*, or *mmfd* and equal to one one-millionth of a microfarad. The electrostatic unit of capacitance sometimes is used. It is abbreviated *esu*. This unit is equal to 1.1126 micro-microfarads.

Rolled paper capacitors have plates of thin metal foil with two or more layers of thin paper as the dielectric. Usually there are only two plates. Adjustable trimmer capacitors may have two or more plates with air or thin sheets of mica for the dielectric, or they may be ceramic types mentioned later in this article. Capacitors used for tuning by the operator of television or radio apparatus usually have rigid metal plates and air dielectric. Alternate plates of multi-plate capacitors are conductively connected together to give the electrical effect of only two plates.

Rolled paper or rolled foil capacitors may be thought of as consisting of two long and relatively narrow plates separated by the dielectric as at the left in Fig. 12-1. All multi-plate capacitors may be represented in a general way as at the right.



Fig. 12-1.—Elementary forms of two-plate and multi-plate capacitors.

Capacitance depends on several factors. It increases with increase of active plate area, which is the area of plate surfaces in contact with the dielectric. Capacitance increases also with increase of dielectric constant of the dielectric material. The thicker the dielectric, or the greater the separation between plates, the less is the capacitance.

The approximate capacitance of a capacitor, in mmf, may be computed from the following formula when plate area is measured in square inches and dielectric thickness in inches.

$$Mmf = \frac{0.225 \times \text{dielectric constant} \times \text{area, one side of one plate} \times \text{number of plates, minus 1}}{\text{dielectric thickness}}$$

If plate area is in square centimeters and dielectric thickness in centimeters, the first term in the numerator is changed from 0.225 to 0.0885.

Actual capacitance will be affected by presence of supporting metal acting as additional plate surface, and by supporting dielectric or insulating material.

In multi-plate capacitors especially, and in two-plate types to some extent, additional capacitance results from edge effect, which is the curving of lines of force through the spaces around the edges of the plates. The approximate addition for edge effect is,

$$Mmf = \frac{\sqrt{\text{area of one side of one plate, sq. inches}}}{4}$$

In capacitors with long, narrow plates there is a considerable increase of capacitance due to elongation effect. Capacitance computed from the first preceding simple formula is multiplied by the elongation factor. The value of this factor is,

$$\text{Elongation factor} = 1 + \frac{0.015 \times \text{length}}{\text{width}}$$

The total capacitance of any number of capacitances connected together in parallel is equal to the sum of the separate capacitances. For example, with 20 mmf, 10 mmf, and 150 mmf in parallel the total capacitance is 180 mmf.

The combined capacitance of any number of equal capacitances connected together in series is the capacitance of one unit divided by the number of units. As an example, the

capacitance of three 60-mmf capacitors connected in series with one another is the quotient of 60 divided by 3, or is 20 mmf.

The combined capacitance of any two capacitances in series, regardless of whether the capacitances are equal or unequal, may be found from dividing the product by the sum. Assume capacitors of 12 mmf and 4 mmf in series. Then,

$$\frac{12 \times 4}{12 + 4} = \frac{48}{16} = 3 \text{ mmf, combined capacitance}$$

If more than two unequal capacitances are in series the combined capacitance may be found by first computing the capacitance of two of them according to the rule just explained, then considering this combined capacitance as in series with a third unit for a second similar computation, and so on. For example, assume that the capacitances of 12 mmf and 4 mmf are in series with a third unit whose capacitance is 60 mmf. The first part of the work would be to compute the combined capacitance of 12 and 4 mmf, which previously was found to be 3 mmf. The next step is to compute the capacitance of 3 mmf and 60 mmf in series, thus.

$$\frac{3 \times 60}{3 + 60} = \frac{180}{63} = 2 \frac{54}{63} = 2 \frac{6}{7} \text{ mmf}$$

The combined capacitance of any number of equal or unequal capacitances in series is equal to the reciprocal of the sum of the reciprocals of the separate units. Using the separate unit values of 12, 4, and 60 mmf, the computation would be as follows:

$$\frac{1}{12} + \frac{1}{4} + \frac{1}{60} \text{ equivalent to } \frac{5}{60} + \frac{15}{60} + \frac{1}{60} = \frac{21}{60}$$

The reciprocal of 21/60 is 60/21, which is equal to 2 6/7 mmf.

It is important to keep in mind that the total potential difference across any number of capacitors in series divides between the capacitors inversely as their capacitances. Continu-

ing with the example of 12, 4 and 60 mmf in series, any total voltage across all three would divide between them proportionately to 1/12, 1/4, and 1/60. This does not mean that 1/12 the total voltage would be across the 12 mmf unit, 1/4 the total across the 4 mmf unit, and so on, for all the fractions would not make up 100 per cent of the overall voltage.

The fractions may be changed to equivalent ones having the same denominator, as to 5/60, 15/60, and 1/60. Then the numerators, 5, 15, and 1, indicate the relative voltages on the capacitors. We might assume an overall potential difference of 315 volts. It would divide proportionately to 5, 15, and 1. The sum of these three numbers is 21. Dividing 315 volts by 21 gives 15 volts. Then voltages across the three capacitors are as follows:

$$\begin{array}{r}
 12 \text{ mmf unit. } 5 \times 15 \text{ volts} = 75 \text{ volts} \\
 4 \text{ mmf unit. } 15 \times 15 \text{ volts} = 225 \text{ volts} \\
 60 \text{ mmf unit. } 1 \times 15 \text{ volts} = 15 \text{ volts} \\
 \hline
 \text{Total voltage} \qquad \qquad \qquad 315 \text{ volts}
 \end{array}$$

The smallest capacitance in a series string always carries the greatest voltage, and the largest capacitance always carries the least voltage. Voltage ratings must allow for this fact if small capacitors are not to be punctured.

**Color Codes.** — Fig. 12-2 shows arrangements of dots and their significance in color coding systems which are and have been most generally used for mica capacitors. Meanings of the colors when used in the various dot positions are given by the accompanying table.

At the upper left in Fig. 12-2 are shown dot positions for the six-dot RMA (Radio Manufacturers' Association) color code. This system allows showing on the capacitor the first three numerals of the capacitance in mmf and a multiplier which shows the number of ciphers to be added in making up the complete capacitance value. In addition there is a dot space for showing capacitance tolerance in per cent, and another one for showing the working voltage of the capacitor. To indicate a capacitance of 2,500 mmf, tolerance of 5%,

and working voltage of 300, colors would be arranged thus:

*Red*      *Green*      *Black*  
*Orange*   *Green*      *Brown*

For 10 mmf capacitance, 10% tolerance, and 500 working volts, we would have,

*Brown*      *Black*      *Black*  
*Green*      *Silver*      *Gold*

## MICA CAPACITOR COLOR CODE

Color	Digit Numeral	Multiplier	Tolerance		Working Voltage (RMA)	Charac- teristic (JAN)
			(RMA)	(JAN)		
Black	0	1	20%	20%		A
Brown	1	10	1%		100	B
Red	2	100	2%	2%	200	C
Orange	3	1,000	3%		300	D
Yellow	4	10,000	4%		400	E
Green	5		5%		500	
Blue	6		6%		600	
Violet	7		7%		700	
Gray	8		8%		800	
White	9		9%		900	
Gold		0.1		5%	1,000	
Silver		0.01	10%	10%		
Characteristic		Temperature Coef. parts/million/deg. C		Capacitance Drift		
A (black)		Not specified		Not specified		
B (brown)		Not specified		Not specified		
C (red)		-200 to +200		+ or -0.5 per cent		
D (orange)		-100 to +100		+ or -0.3 per cent		
E (yellow)		- 20 to +100		+ or -0.1 per cent		

At the upper right in Fig. 12-2 are the positions for the JAN (Joint Army-Navy) capacitor color code. This code may be called ASA (American Standards Association) or AWS (American War Standard). Capacitors marked with this code are easily identified, because the upper left-hand dot always is black. To specify capacitance value there are dots for only two digits or numerals, and a multiplier. The tolerance dot is in the same position as for the RMA code, but the lower

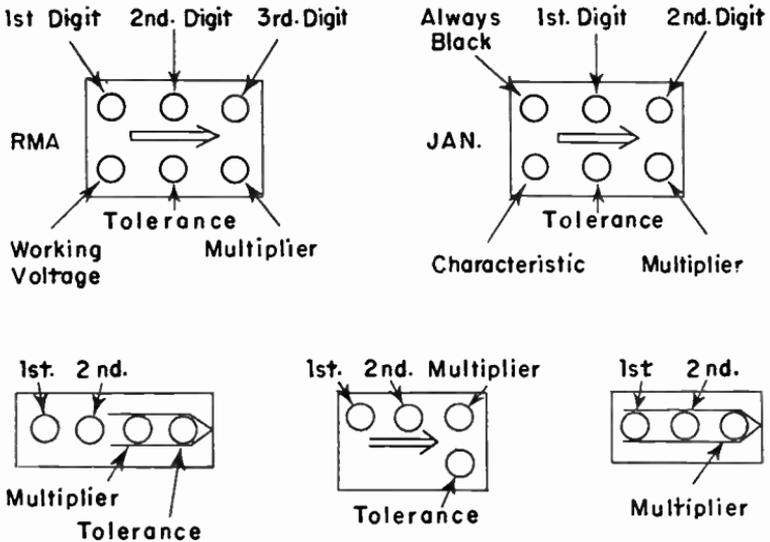


Fig. 12-2.—Meanings of color code dots in various positions on mica capacitors.

left-hand dot indicates temperature and drift characteristics as noted in the table.

For 2,500 mmf and 5% tolerance in the JAN code the color arrangement would be thus:

<i>Black</i>	<i>Red</i>	<i>Green</i>
<i>Black</i>	<i>Gold</i>	<i>Red</i>

For 10 mmf, 10% tolerance, in the JAN code the arrangement would be,

<i>Black</i>	<i>Brown</i>	<i>Black</i>
<i>Black</i>	<i>Silver</i>	<i>Black</i>

Some mica capacitors have a four-dot marking as at the lower left in Fig. 12-2. Capacitance digits, multiplier, and tolerance markings are as previously explained. With no voltage marking the working voltage of mica capacitors usually is 500. Another arrangement of a four-dot system is shown at the lower center of this figure. A three-dot marking is shown at the lower right. Here there are positions for only two capacitance digits and the multiplier. The tolerance ordinarily will be 20% and the working voltage 500 with this three-dot marking.

CERAMIC CAPACITOR COLOR CODE

Color	Digit Numerical	Multiplier	Tolerance		Temp. Coef.
			If C more than 10 mmf	If C 10 mmf or more	
Black	0	1	20%	2.0 mmf	0
Brown	1	10	1%		- 30
Red	2	100	2%		- 80
Orange	3	1,000			-150
Yellow	4				-220
Green	5		5%	0.5 mmf	-330
Blue	6				-470
Violet	7				-750
Gray	8	0.01		0.25 mmf	+ 30
White	9	0.1	10%	1.0 mmf	-330 to + or -500

Temperature coefficient of capacitance is the change in parts per million, per centigrade degree temperature change. Ceramic capacitors are color coded as in Fig. 12-3.

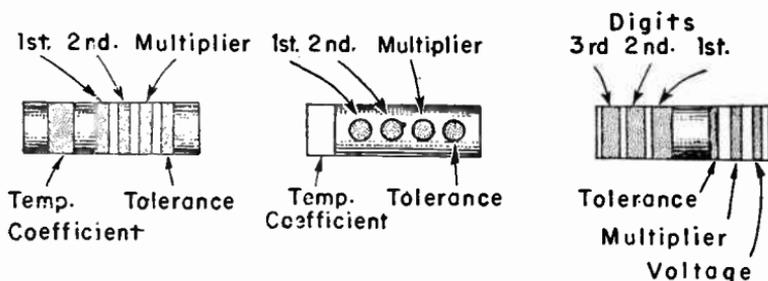


Fig. 12-3.—Color code markings for fixed ceramic and some other cylindrical or tubular capacitors.

Capacitors are held as follows while reading color coding. Tubular or disc ceramics have the pigtails extending upward. Mica capacitors (Fig. 12-2), hold the unit so any arrow or anything in the form of an arrow points toward the right, or so that the name of the manufacturer or any other wording is

right side up. The position of the coding bands or dots on ceramic units (Fig. 12-3) is identified by differences between widths of bands.

**Temperature Compensating Capacitors.** — When a temperature coefficient is specified in parts per million per degree change of centigrade temperature it is the same as saying in micro-microfarads per microfarad of capacitance per degree temperature change. It is the general tendency of tuned circuits to decrease their resonant frequency as temperature rises. This is most pronounced at very-high and ultra-high frequencies. The effect is as though there were an increase of inductance or of capacitance with rise of temperature. This tendency is opposed or compensated for by using capacitors having a negative temperature coefficient, indicated by the minus sign in the tables.

As an example, assume a capacitor having a nominal capacitance of 1000 mmf or 0.001 mf, and assume that the temperature coefficient of this capacitor  $-750$ . Then the capacitance will decrease in the ratio of 750 parts per million, or in the ratio of 750 mmf per microfarad of capacitance. Since the capacitance has been assumed to be 0.001 or  $1/1000$  microfarad, the actual decrease will be  $750/1000$  or 0.75 mmf per centigrade degree of temperature rise. With a rise of 10 degrees centigrade (18 degrees Fahrenheit) the capacitance will decrease by 10 times 0.75 mmf, or by 7.5 mmf, and instead of the original 1000 mmf it will become 992.5 mmf.

When a receiver operates in such manner that tuned circuits are subjected to high temperature, it is a rather common service operation to have to change capacitors of one temperature coefficient to others having a greater negative temperature coefficient.

**Mica Capacitors.** — Mica capacitors have high quality sheet mica as the dielectric, with plates of metal foil or of deposited silver films, and are completely enclosed within a housing of molded phenolic or plastic compound. Such capacitors have very small d-c leakage, and they have long service life with minimum danger of puncture or breakdown if the working

voltage rating is correct in the first place. The chief disadvantage of mica capacitors is their high cost in comparison with other types, but this cost is warranted when making replacements.

Silver mica capacitors are made with mica dielectric on which a coated compound is reduced to metallic silver by heat. The molded covering is of some material having small dielectric losses at very-high and ultra-high frequencies. Such capacitors retain their capacitance value and other operating characteristics for long periods. They are especially useful in high- $Q$  tuned circuits and wherever close tolerance is important.

**Ceramic Capacitors.**—The usual structure of a fixed ceramic capacitor is shown, in a general way, by Fig. 12-4. The capacitor element itself is a thin-walled tube of ceramic material, which is the dielectric, with the outside and inside surfaces of this tube coated with silver and sometimes other metals also,

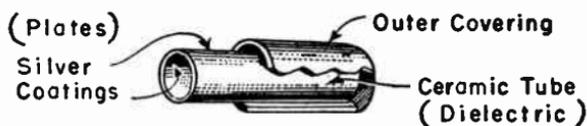


Fig. 12-4.—The principal parts of a fixed ceramic capacitor.

to form the two plates. This element is enclosed within a sealed outer covering, which may be of something such as steatite or a low-loss resin compound. The capacitor is wax impregnated under vacuum before sealing. The silver coatings are bonded to the ceramic tube by heating, and there may be over-coatings of copper or other metals.

The dielectric tube is given the desired dielectric constant by compounding the ceramic material with titanium dioxide which, by itself, has an extremely high dielectric constant. The compounding also determines the temperature coefficient of the capacitor. Ceramic capacitors originally were used almost exclusively where controlled temperature coefficients were required, but they now are used also for bypassing, coupling, and all other purposes for which moderate capacitances are needed.

Rotary ceramic capacitors are widely used as trimmers. The stationary base is of compounded ceramic material ground perfectly flat. On the under side is a silver deposit which forms one plate. The adjustable rotating member carries a semi-circular piece of metal which is held by spring pressure in close contact with the top surface of the ceramic base. There are other designs, some with the ceramic dielectric forming the rotating member. Usually the capacitance changes from minimum to maximum with a half-turn of the adjustment.

**Paper Capacitors.** — Capacitors with paper dielectric and metal foil plates, when used at frequencies lower than video and sound intermediate frequencies, usually are of the rolled type whose construction method is illustrated by Fig. 12-1. The cross section may be cylindrical or oval. If the sides of the long strips of foil do not protrude beyond the paper dielectric, and if terminal connections are made to the ends of the foils, there is not only capacitance but also a very considerable inductance because the foil strips are in the form of coils. The combination of capacitance and inductance makes the capacitor a complete resonant circuit at some frequency.

In a check on this feature, several 0.005-mf rolled paper capacitors were found to be resonant around 9 to 10 mc. One 0.001-mf unit was resonant at 65 mc. Where such capacitors are used for bypassing and decoupling, their high impedance at their resonant frequency makes it necessary to connect in parallel a smaller mica or ceramic capacitor which will have low reactance at this frequency provided it is to be bypassed.

Non-inductive paper capacitors are constructed with one foil protruding from one side of the paper, and with the other foil protruding from the other side. The protruding edges are pressed together to form the two terminals at which the pigtailed are attached.

The pigtail or terminal connected to the outside foil of paper capacitors may be marked with appropriate wording or with a narrow colored band. This end of the capacitor should be connected to the ground side of the circuit, whereupon the outside foil acts as a shield.

**Feed-through and Standoff Capacitors.**— A feed-through capacitor is one that mounts in an opening through chassis metal or through barriers or shielding partitions, with one capacitor terminal on one side of the metal support and the other terminal on the opposite side. Two of the many types of feed-throughs are illustrated by Fig. 12-5. One screws into a threaded hole. The other is soldered or sweated by means of a flange that fits over the opening. Mounting sometimes is by means of a clamp fastened to chassis metal. Feed-throughs are employed for coupling and blocking. Usually they are ceramic types.

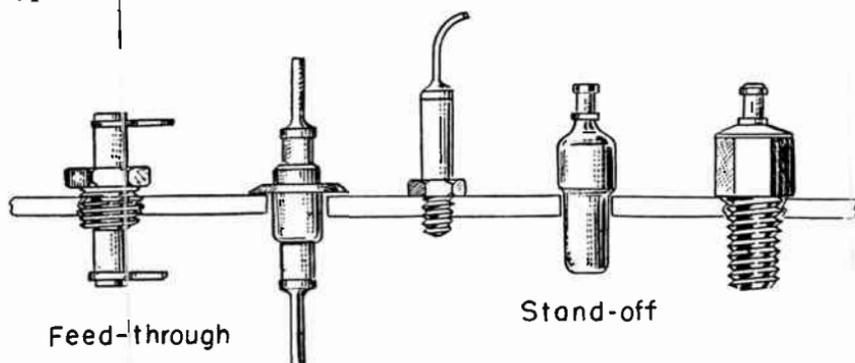
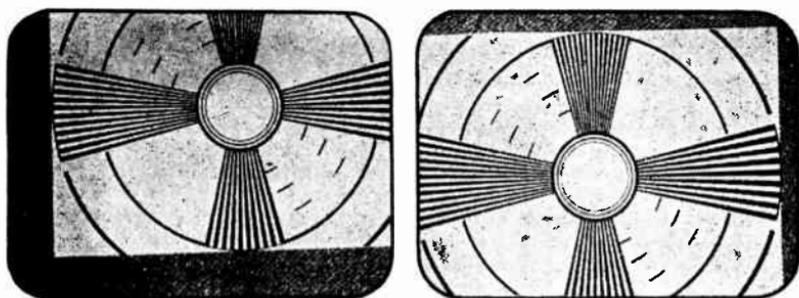


Fig. 12-5.—Capacitors of feed-through and standoff types.

A standoff capacitor, as shown by three examples in Fig. 12-5, mounts on grounded metal either by threading, by a tight push fit, or by soldering with a flange or ferrule. One terminal of the capacitor goes to ground through the mounting, the other is insulated and used for wiring connections. Standoffs are employed for bypassing and decoupling, and, like feed-throughs, usually are ceramic types.

**Gimmicks.**— A gimmick is a capacitor formed by winding several turns of bare wire around an insulated wire, or by twisting two insulated wires together. The metal conductors act as capacitor plates, with wire insulation forming the dielectric. Capacitance, only a few mmf at most, is varied by winding more or fewer turns.

**CENTERING AND CENTERING CONTROLS.**—The picture or pattern appearing on the screen of the picture tube may be too high and too far to the right, as shown at the left in Fig. 13-1, or too low and too far to the left, as in the right-hand illustration, or it might be off center in only one direction. Such faults are due to incorrect adjustment of centering controls. When no deflecting voltages or currents are affecting the electron beam in the picture tube the beam will not necessarily strike the exact center of the mask opening, and when there is deflection it may be greater in one direction than in the opposite direction.



*Fig. 13-1.—The pattern or picture may be off center in one or more directions.*

Centering adjustments should be made in such manner that pictures are centered in the cabinet mask opening and extend a quarter-inch or more beyond the opening on both sides and at top and bottom. Size and vertical linearity should be checked at the same time as centering. It may be necessary to adjust width or height, or both, to make pictures extend uniformly beyond the mask edges in all directions. If pictures are out of proportion vertically, the vertical linearity control requires adjustment. Although the horizontal drive control affects width, it seldom should be necessary to readjust the drive in connection with centering.

Before making centering adjustments observe programs on all available channels. Pictures on some channels may be smaller than on others. The smallest pictures should be used while centering. Also, the horizontal hold control should be at a setting midway between its positions at which pictures go out of sync either way. This is because the horizontal hold may shift pictures some distance to the left or right while still holding sync, and unless adjustment is made for an average or normal setting the centering later may be found incorrect.

**Centering with Magnetic Deflection.** — During normal operation of a picture tube designed for magnetic deflection the beam is moved up, down, and sideways by reversals of the magnetic fields from the deflection yoke or deflection coils. These magnetic fields are the result of, and are completely controlled by, currents in the deflection coils. Were a steady current passed through the deflection coils, in addition to the current causing the sweep, the beam would be deflected by magnetic fields resulting from the combination of steady and sweep currents.

Effects of a steady current are shown by Fig. 13-3. Electron flow in the left-hand diagram is in such direction around the coils as to produce magnetic polarities which deflect the electron beam to the left. Electron flow has been reversed in the right-hand diagram. This reverses the magnetic polarities and reverses the direction of beam deflection. These directions are based on the assumption that electrons in the beam are

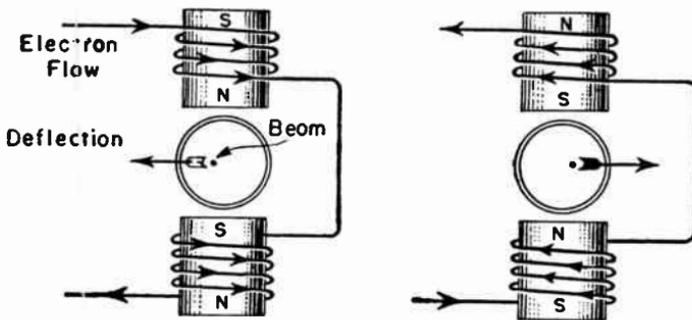


Fig. 13-3.—Effects of steady electron flow or direct currents in the yoke coils of a magnetic deflection system.

coming toward you, are coming out of the page. Variation of direction and strength of the steady electron flow will deflect the beam and hold it deflected at any distance toward the left or right. Variation of a steady current in the vertical deflection coils would deflect and hold the beam at any distance above or below center. Sweep distances will be approximately equal on both sides and above and below these centers. Thus the entire picture or pattern may be moved on the screen by varying the steady centering currents.

This principle of centering by variation of direct currents in the deflection coils is utilized with circuits such as shown by Fig. 13-4. Currents are taken from the B-power voltage

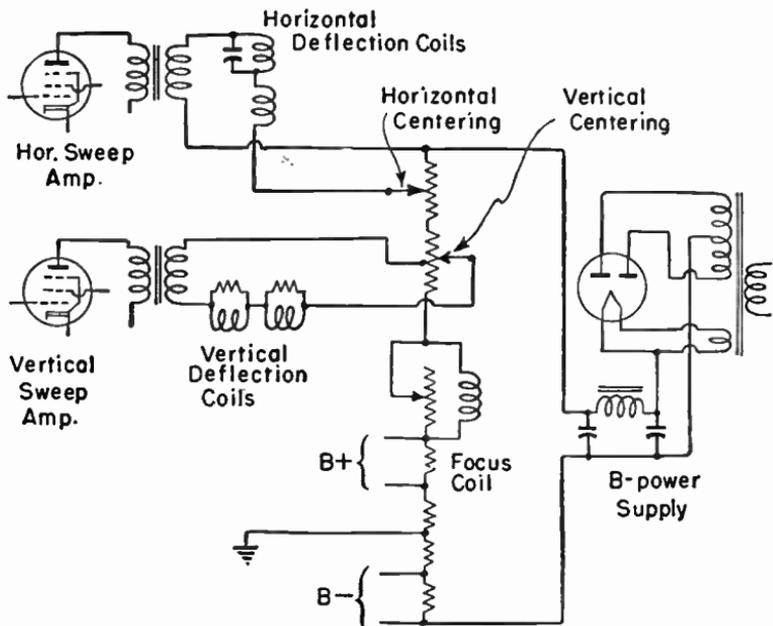


Fig. 13-4.—Electrical centering controls used with a magnetic deflection system.

divider system or other suitable source. The centering potentiometer for the horizontal deflection coils at the top of the diagram provides adjustable current, but this current flows in only one direction. It is used to counteract to a greater or

less extent another opposite current which flows in the horizontal coils during operation of the receiver. This other current tends to cause a steady deflection to one side, and the adjustable centering current or the voltage which would cause the centering current is adjusted as required to center the picture.

The centering potentiometer for the vertical deflection coils of Fig. 13-4 is a center tapped type whose adjustment allows current in either direction, and of any necessary value, through the coils. In this vertical deflection system there is no other steady direct current, as in the horizontal system up above. Therefore, it must be possible to have the centering current in either direction. This requires the center-tapped potentiometer for adjustment.

**Centering with Focus Coil.** — The focusing coil acts somewhat like a lens in collecting the beam electrons, which are spreading apart, and in focusing them to a small spot on the screen. Just as the position of a focused spot of light will change when the focusing lens is tilted with reference to its axis, so the beam spot is shifted on the picture tube screen when the focusing coil is tilted with reference to the axis of the picture tube.

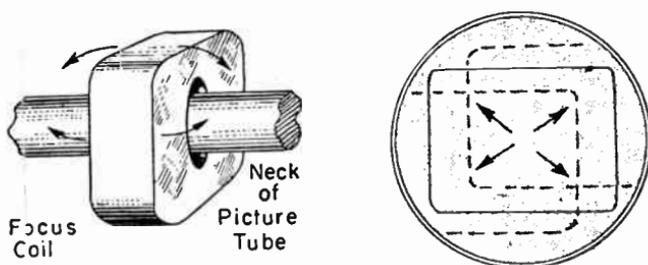


Fig. 13-5.—Tilting the focusing coil with reference to the axis of the picture tube neck will shift the center of a pattern or picture.

Focusing coils are mounted in such ways that the axis through the coil center may be shifted, as at the left in Fig. 13-5, with reference to the picture tube neck. Only a limited

degree of such shifting is possible without having the coil housing strike the tube neck, but it is enough to allow moving the picture or pattern on the screen as shown at the right. Moving the coil axis in either a vertical or horizontal plane causes diagonal shifting of the picture, so usually it is necessary to utilize both directions of coil movement for centering. In a large number of receivers this is the sole means of centering the picture.

If electrical centering controls such as those of Fig. 13-4 are employed, the initial adjustment may be of focus coil position while both electrical controls are set midway between their extreme adjustment positions. If an electrical control will not center the picture in the mask, that control should be placed at its mid-setting while the focusing coil is adjusted. Coil mounts may be of trunnion form, or may consist of three or more adjustable screws having tension springs, or of any other type which allows shifting the coil axis. The deflection yoke always should be as far toward the flare of the picture tube as possible while making any centering adjustments.

Control of centering for picture tubes employing electrostatic focusing may be with direct currents in the yoke, as shown by Fig. 13-4, or by means of ring-shaped permanent magnets rotatable around the tube neck, or by combination of both methods. Centering for such tubes may be adjusted also by means of permanent magnet focusing devices described in the article on *Focusing*.

There may be only a single ring-shaped centering magnet or there may be two, independently adjustable by rotation around the picture tube neck. At one point around each magnet is a gap, with magnet ends on both sides of the gap having the same polarity, either north or south. At a point directly opposite the gap is a consequent pole of opposite polarity. Magnetic fields between north and south poles extend through the tube neck and through the electron beam path to bend the beam as required. Adjustment is largely a matter of experiment in determining centering effects.

**CHANNELS, TELEVISION.** — The range of frequencies in which signals from a transmitter are radiated is called a channel. Each television broadcast channel includes a range of six megacycles. In the very-high frequency band there are twelve channels, numbered 2 through 13. Channel number 1, originally assigned for television broadcast, now is used for other services. In each channel the video carrier frequency is 1.25 mc higher than the low limit of the channel, and the sound carrier frequency is 0.25 mc lower than the high limit. Channel frequencies and carrier frequencies are listed in the accompanying table.

CHANNEL FREQUENCIES, MEGACYCLES			
Channel No.	Limits	Video Carrier	Sound Carrier
2	54 to 60	55.25	59.75
3	60 to 66	61.25	65.75
4	66 to 72	67.25	71.75
5	76 to 82	77.25	81.75
6	82 to 88	83.25	87.75
7	174 to 180	175.25	179.75
8	180 to 186	181.25	185.75
9	186 to 192	187.25	191.75
10	192 to 198	193.25	197.75
11	198 to 204	199.25	203.75
12	204 to 210	205.25	209.75
13	210 to 216	211.25	215.75

A band of ultra-high frequencies between 480 and 920 megacycles has been allocated for television broadcasting, but channel divisions have not been fixed or assigned in this band.

Under the original rules for assignment of channels, broadcast transmitters are not assigned to the same channel unless at least 150 miles apart, nor are transmitters assigned to adjacent channels unless separated by 75 miles or more. Note that channels numbered 4 and 5 are not adjacent, since there is a gap of four megacycles between them. Consequently, these two channels are assigned to stations in the same locality. These rules are modified to meet the public interest or necessity, and as may be required by conditions peculiar to certain geographical areas.

**Ultra-high Frequency Television Channels.** — The ultra-high frequency band of television broadcast frequencies extends from 470 to 890 megacycles. Each ultra-high frequency channel occupies 6 megacycles, which allows 70 channels in this band. Ultra-high frequency channels are numbered consecutively from 14 through 83, a continuation of the numbering for channels 2 through 13 in the very-high frequency television broadcast band. Channel frequency limits are listed by the accompanying table.

CHANNEL LIMITS IN THE ULTRA-HIGH FREQUENCY  
TELEVISION BAND

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

**CLIPPERS OR LIMITERS.** — In television receivers the name clipper may be applied to a tube whose purpose is to retain and sometimes to amplify the sync pulses of a signal while reducing or eliminating the picture variations of the signal. This would be one of the tubes in the sync section of the receiver. A tube performing this function may be called also a sync separator.

The name limiter may be applied to a tube whose chief purpose is to reduce all sync pulses of a signal to the same amplitude or same voltage strength, or to limit the amplitude to some certain maximum value which prevents passage of

extra strong pulses due to noise effects or to interference. Such a tube may be called also a sync clipper.

In television sound sections employing two or more intermediate-frequency stages the tube in the stage preceding the detector or demodulator may be operated as a noise limiter. The chief purpose of this limiter is to prevent passage through it of amplitude-modulated signals or impulses while delivering to the demodulator all the frequency-modulated signals for sound. The limiter may be operated with such combination of plate, screen, and grid voltages as to cause plate current cutoff, current saturation, or both effects when signal amplitude rises above a uniform level or when there are amplitude pulses which would cause noise after detection.

Diodes may be used as clippers, limiters, or levelers of sync pulses. One method is illustrated by Fig. 15-1. Sync pulses coming to the grid of the left-hand triode tube are of negative polarity, and in the output of this tube the pulses are positive due to signal inversion which always occurs in a triode amplifier. The positive pulses make the plate of the clipper diode positive and tend to make the diode conduct.

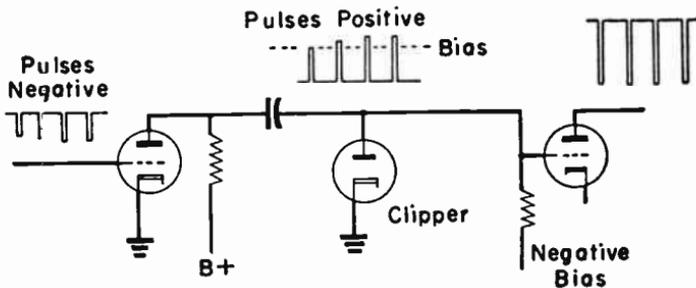


Fig. 15-1.—A diode used as a leveler or clipper for sync pulses.

But the plate of the diode is conductively connected to the negative grid bias voltage source for the right-hand triode. Only when the positive voltage of the sync pulses exceeds the negative bias voltage will the clipper diode actually conduct and reduce this positive voltage. Thus sync pulse volt-

ages applied to the grid of the right-hand diode are made fairly uniform and the output signal of this triode is likewise fairly uniform.

Another connection for a pulse clipper is shown by Fig. 15-2. Here the clipper diode plate is connected through capacitor  $C$  to the lead between the plate of the left-hand amplifier tube and the grid of the right-hand amplifier. In the circuits of video amplifiers the signal consists of both picture variations and sync pulses. It is assumed that in the signal

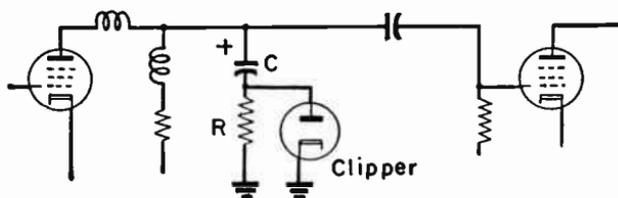


Fig. 15-2.—A clipper diode biased by a capacitor and a resistor.

applied to the diode plate the sync pulses are positive. Resulting current flowing in capacitor  $C$  and between plate and cathode of the clipper tube causes charging of the capacitor in the polarities marked. The capacitor discharges slowly through resistor  $R$ , and maintains the clipper plate negative with reference to its cathode. That is, the clipper plate is negatively biased at a potential proportional to the average or normal voltage of the sync pulses. Should interference or other external forces produce sync pulses of more than average voltage these higher voltages will overcome the negative bias and cause the clipper to conduct. The conduction loads the circuit sufficiently to level off most of the excess pulse voltage and to leave sync pulses of practically constant strength for the grid of the right-hand amplifier tube.

**COLOR TELEVISION.** — The elementary principles of color television are illustrated by Fig. 2. The upper diagram shows pickup of a televised scene and the manner in which color images are changed to signal voltages for transmission within the six-megacycle band of any television broadcast channel.

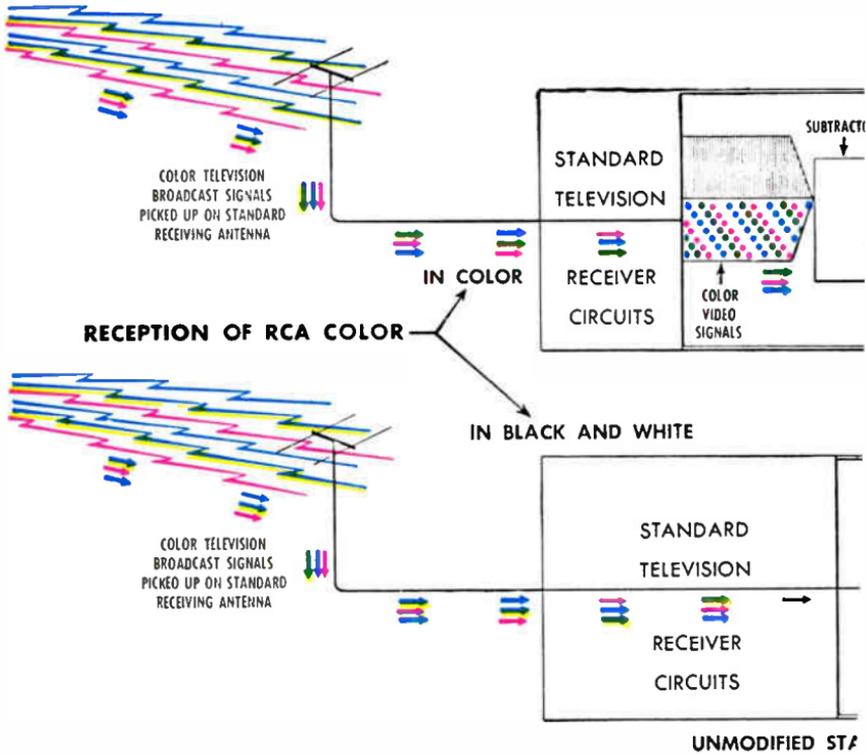
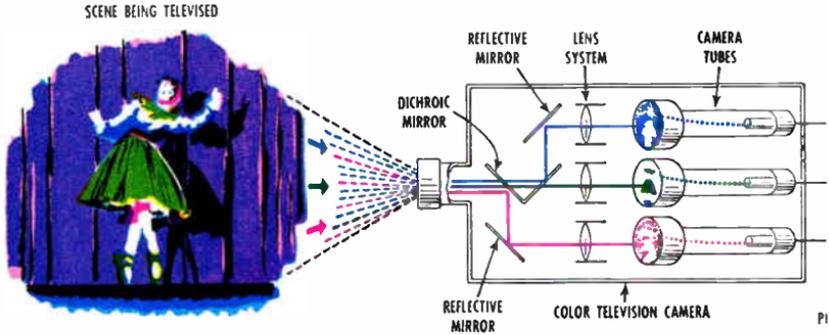
All colors reflected or radiated from objects in the televised scene enter the color television camera, but instead of going to a single camera tube as for black and white transmission the light rays pass through a system of mirrors and color filters which separate the original all-color image into three images for three camera tubes. One of the separate images consists of all colors from deepest violet to greenish-blue; this being called the blue image. Another, the green image, includes everything from blue-green through green and into yellowish-green. The red image consists of all colors from orange to deepest red. The blue, green, and red bands of color are called primary colors.

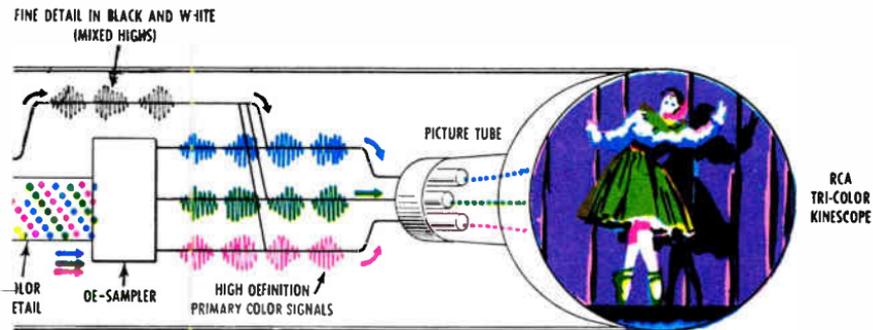
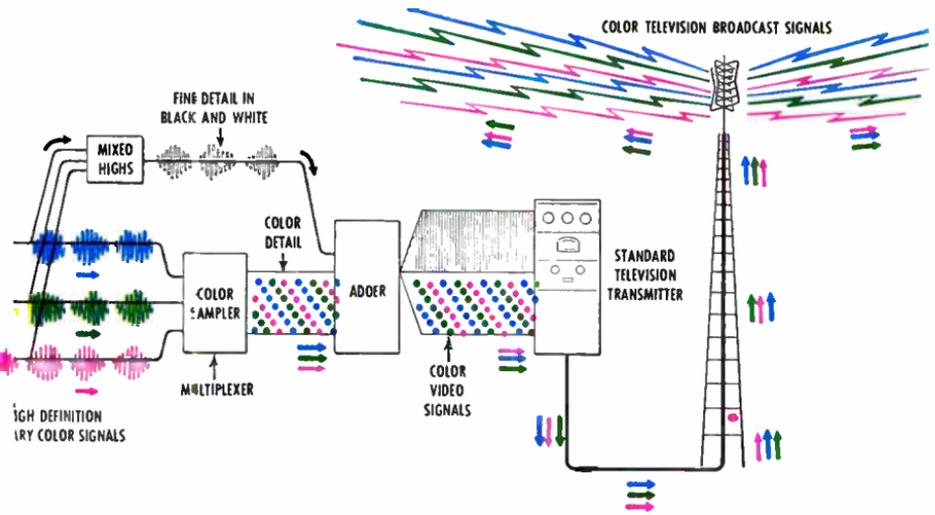
Each camera tube produces a signal voltage whose strength is proportional to intensities of colors in its portion of the complete image. From the blue tube is derived a blue-primary



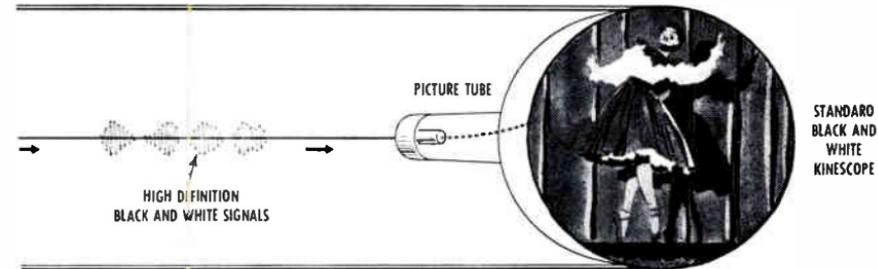
*RCA color camera in operation during a color broadcast from the WNBW studios in Washington, D. C.*

PICKUP AND TRANSMISSION OF RCA COLOR





**COLOR TELEVISION RECEIVER**



**STANDARD BLACK AND WHITE TELEVISION RECEIVER**

signal voltage, from the green tube a green-primary signal, and from the red tube a red-primary signal. These three primary color signals eventually cause to appear on the viewing screen of a color picture tube in a receiver all of the colors and shadings of the original scene.

Certain definite fractions of the three primary color signals are mixed together to form the "mixed highs" of the upper diagram. This mixture is a signal voltage similar to that from a single camera tube affected by all coloring in televised scenes, as in black-and-white television. The common name is luminance signal or Y-signal. The luminance signal represents picture elements and details such as carried by video frequencies in black-and-white television, but does not represent coloring.

Other fractions of the three primary signal voltages are mixed in color sampler or multiplexer circuits to form signals which represent coloring in images, as distinct from white, gray, and black details in the luminance signal. The color signal voltages are used to modulate a color subcarrier voltage whose frequency is approximately 3.58 megacycles. The resulting modulated signal is called the chrominance (color) signal. Now, at the transmitter the chrominance (color) signal and the luminance (black-and-white) signals modulate the video carrier of the television channel in which there is to be transmission.

In the center diagram of Fig. 2 the transmitted color television signal has reached the antenna of a color television receiver. The tuner, the i-f amplifier, possibly the video detector, and at least part of the video amplifier sections of the color receiver are practically the same as similar sections in a black-and-white receiver. At the output of the receiver video detector or detectors, or from a video amplifier, appear the combined luminance and chrominance signals, which are marked color video signals on the diagram. The luminance signal, carrying fine details and black-and-white shadings, is separated from the chrominance (color) signal, and after suitable amplification is applied to the color picture tube.

The chrominance signal passes through many circuits and tubes in the section marked de-sampler on the diagram, and emerges as blue, green, and red primary color signals like those which went into the color sampler of the upper diagram. Finally the primary color signals and the luminance signal go to grid-cathode circuits of the color picture tube.

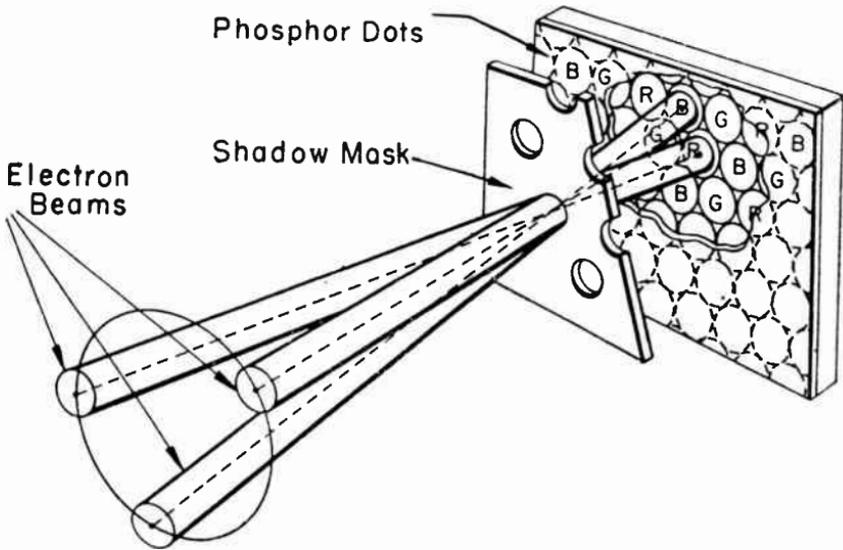
At the bottom of Fig. 2 the transmitted color television signal has reached the antenna of any receiver designed for ordinary black-and-white reception, including all receivers built before color television and all those presently made for only black-and-white reproduction. In these receivers are no circuits which respond to chrominance in the transmitted color signal. Consequently, the black-and-white set responds to only the luminance signal of the color broadcast. As mentioned before, the luminance signal is the equivalent of a black-and-white video signal, and causes reproduction of pictures in black, white, and shades of gray on the picture tube of the black-and-white set when receiving a color broadcast.

Not only will black-and-white sets receive color broadcasts, but a color receiver will reproduce black-and-white broadcasts in white, black, and tones of gray. This is possible because any black-and-white video signal is like the luminance portion of a color signal. Portions of the color receiver which reproduce coloring remain inactive while a black-and-white program is received. The fact that either color or black-and-white receivers will reproduce in one form or another all programs transmitted in either color or black-and-white, and the fact that color transmission is carried out in the same six-megacycle channels as black and white, is referred to as compatibility.

**Three-gun Picture Tubes.** — In the color picture tube which has been illustrated are three electron guns emitting three electron beams. Intensity of the beam from one gun is controlled by the blue signal voltage, beam intensity from a second gun is controlled by the green signal, and beam intensity from the third gun is controlled by the red signal.

On the viewing screen of the color picture tube are hundreds of thousands of tiny phosphor dots arranged in groups

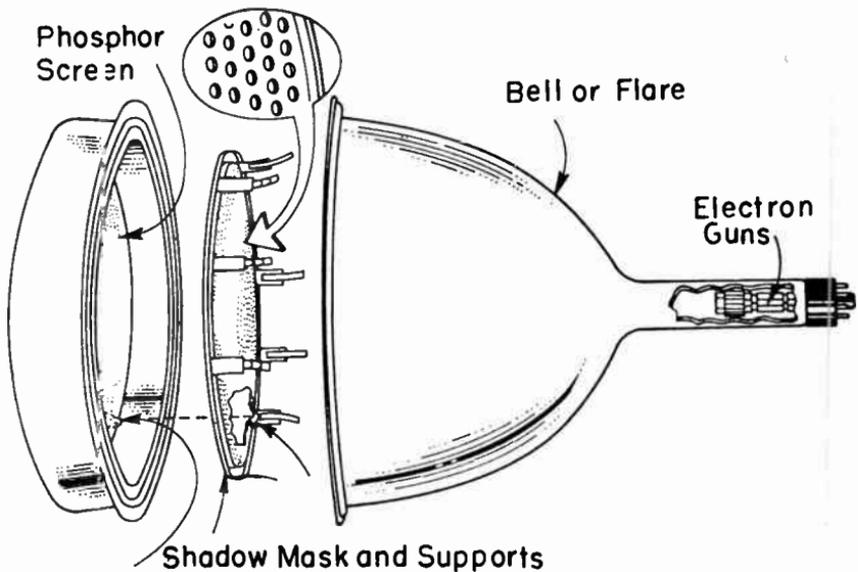
of three. In each group is one dot for each of the primary colors. A few of these color-phosphor dots are shown greatly enlarged by Fig. 3. Immediately back of the viewing screen is a shadow mask containing as many openings as there are groups of dots on the screen, one opening for each of the groups which consists of a red, a green, and a blue dot.



*Fig. 3.—The three electron beams pass through the shadow mask openings and strike phosphor dots which emit the three primary colors.*

The three electron beams come together or converge at openings in the shadow mask, then spread just enough so that the beam from the blue gun strikes only blue dots, and beams from the other two guns strike only dots of their respective colors. Phosphor dots for each primary color are excited to degrees of luminescence corresponding at every instant to intensities of the three primary colors at various places in the televised scene. The primary colors on the picture tube screen combine, in the eyes of viewers, to re-create all color tones and shadings, and all their changes, occurring at the scene.

The shadow mask of the CBS-Colortron picture tube is mounted between the face plate with its phosphor dots and the bell or flare of the tube, as shown by Fig. 4.



*Fig. 4.—The shadow mask is a thin plate supported just back of the face plate which carries phosphor dots.*

Elements and their connections to base pins in one three-gun picture tube are shown by Fig. 5. In each gun is a cathode, a control grid, a screen grid (second grid), a focus electrode (third grid), and part of the high-voltage anode. Cathodes, control grids, and screen grids are connected to individual base pins, while the three focus electrodes are connected together and to one base pin. The high-voltage anode terminal is a metal flange around the tube, just back of the face. The three electron guns are in relative positions illustrated, when looking at the rear or base of the tube when picture center lines are horizontal and vertical.

**Colors and Light.** — To understand the processes involved in producing television signals from variations of color, and in changing the signals back into color, it is helpful to examine some elementary principles of light and color. Light is radiant energy which travels through space as electromagnetic waves so short that there are from 36,000 to 63,000 per inch. Wavelengths usually are specified in a unit called the millimicron, equal approximately to one twenty-five millionth of

## COLOR TELEVISION

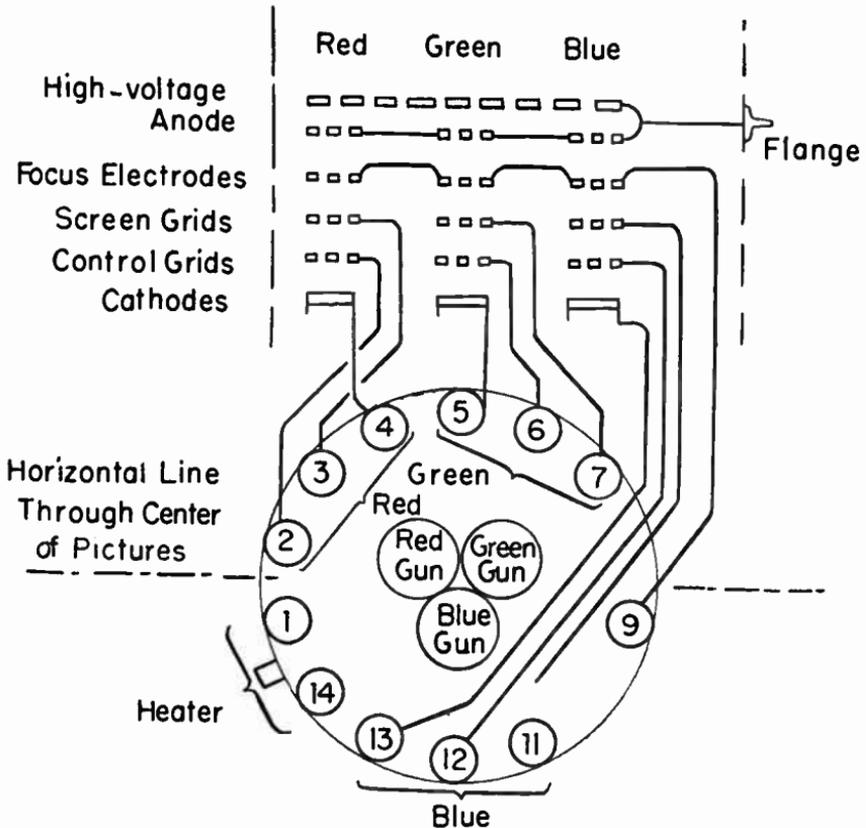


Fig. 5.—Base pin connections of a CBS-Colortron picture tube.

an inch. Visible light or the visible spectrum includes wavelengths from about 400 to 700 millimicrons, which cause color sensations shown at the left in Fig. 6. Actually there are no sharply defined boundaries between colors, each merges gradually into those on either side. When all these wavelengths reach the eye in suitable strengths the sensation is that of white.

As mentioned earlier, each of the three primary colors used in television includes a band of wavelengths from blue, green, and red portions of the spectrum, somewhat as at the right in Fig. 6. The blue primary includes wavelengths from 400 to about 525 millimicrons, with maximum strength around 445 millimicrons. The other primaries include the indicated

bands. When the three primaries are seen in suitable relative strengths at the same time the sensation is of white light.

Primary colors produced on the picture tube viewing screen may be made of such combinations and relative strengths as to give any and all color sensations. For example, certain proportions of green and red give the sensation of yellow, a mixture of blue and red appears as lilac, and so on. It is necessary only to excite one, two, or all three of the color phosphors to suitable degrees of luminescence to produce any color sensations.

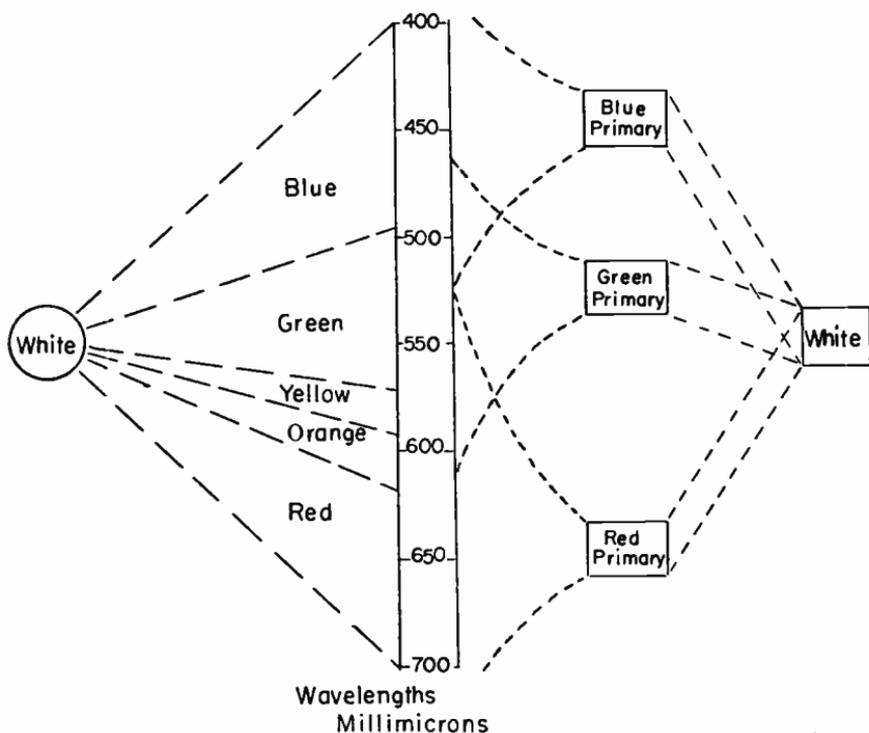


Fig. 6.—White light includes all wavelengths of the visible spectrum, while the color primaries include certain bands of the spectrum.

**Hue and Saturation.** — The color sensation which results from certain wavelengths of light is called hue. If wavelengths are predominantly those of the red portion of the spectrum the hue is red. If the sensation is what ordinarily is called pink, the hue still is red, because pink is merely light red. Hue

does not tell whether a color is bright or dark, faint or intense. Hue relates only to the color sensation, which depends on which wavelengths or mixtures of wavelengths are present.

Saturation refers to the proportion of white that is mixed with some certain hue. Pink or light red contains a considerable portion of white along with red, and is a red hue of relatively low saturation. The red alone would be of high saturation. Since the sensation of white light results from combining all three primary colors, any color of low saturation must contain all three primaries. The one which is strongest determines

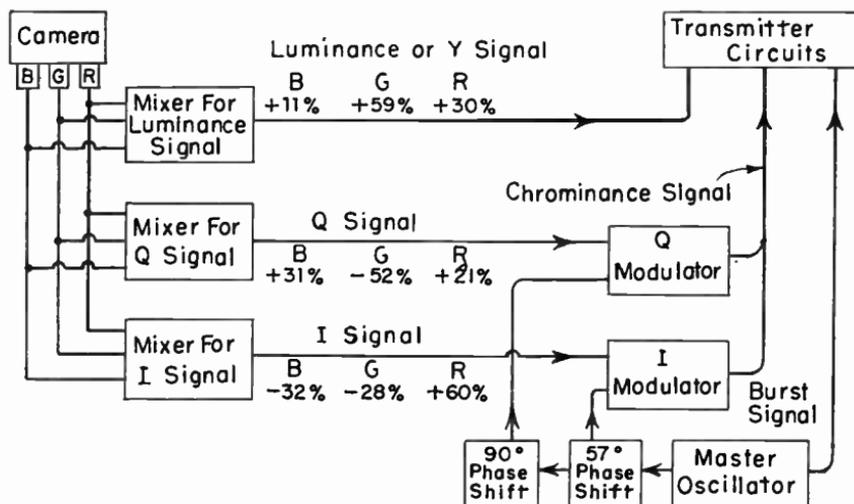


Fig. 7.—How color signals are formed at a transmitter.

the hue, while proportions of the other two determine the degree of saturation.

**Signal Formation.** — Camera tubes do not respond equally to all colors of equal radiant energies. Neither do picture tube phosphors emit equal apparent intensities of all three primary colors when excited by electron beams of equal strengths. Consequently, to produce the sensation of white light it is necessary to commence with only 11 per cent of blue signal voltage at the camera, with 59 per cent of green signal voltage, and with 30 per cent of red signal voltage. These percentages are mixed together to form the luminance or black-and-white sig-

nal, as shown at the top of Fig. 7. The luminance signal often is called the Y-signal.

The chrominance signal, which carries all the color information, is formed by combining two other signals, called Q and I, and these in turn are formed by mixing certain percentages of the three primary color voltages from the camera, all as shown by Fig. 7.

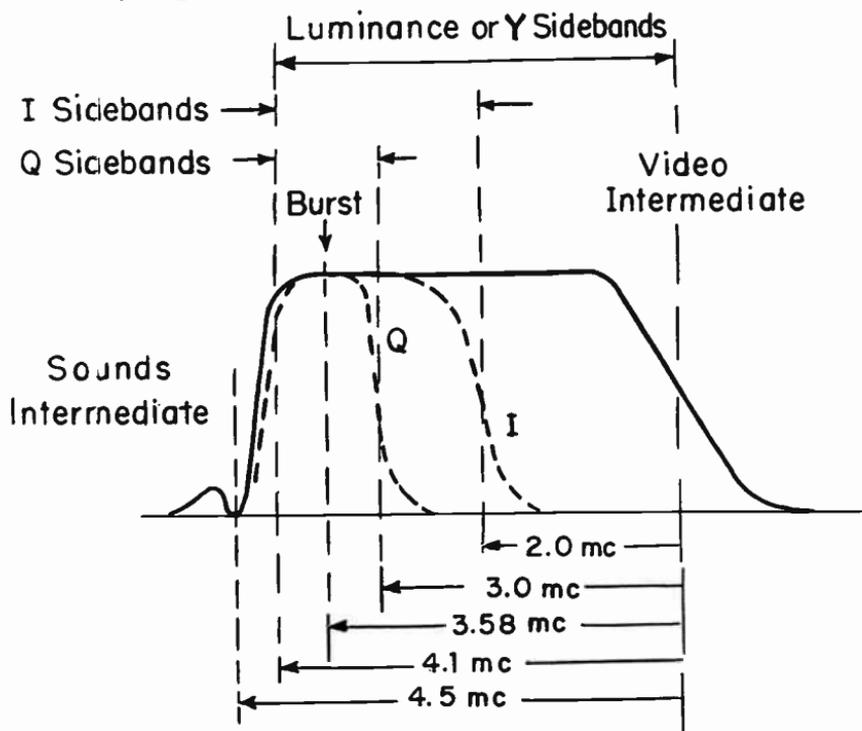


Fig. 3.—Ranges of video frequency covered by sidebands of Y-, I-, and Q-signals.

The Q-signal consists of 31 per cent of the blue camera voltage, 52 per cent of the green, and 21 per cent of the red. The blue and red components of the Q-signal are plus voltages, while the green is a minus voltage. A minus voltage or signal is one whose voltage changes are inverted or are in opposite phase with respect to the plus signal. Were a plus signal of any kind applied to the grid of a tube, the inverted voltage or waveform at the plate would be a minus signal of the same kind. The I-signal consists of minus blue and green, and of

plus red, in percentages marked. The Q-signal and I-signal go to modulators.

To the modulators are applied also voltages whose frequency is 3.58 mc, secured from a master oscillator. The master oscillator voltage is shifted  $57^\circ$  in phase and fed to the I-modulator, then shifted another  $90^\circ$  and applied to the Q-modulator. The I- and Q-signals become modulation on the 3.58-mc voltages, then are combined to form the complete chrominance signal as transmitted.

Also transmitted are intermittent bursts of 3.58-mc voltage from the master oscillator. Each burst consists of not less than eight cycles. The bursts are timed to occur immediately after each horizontal sync pulse, and appear on the back porch of the pedestal formed by horizontal blanking intervals.

Although I- and Q-signals are modulated onto the 3.58-mc voltage, the 3.58-mc frequency is suppressed at the transmitter and only the modulation sidebands go out in the transmitted

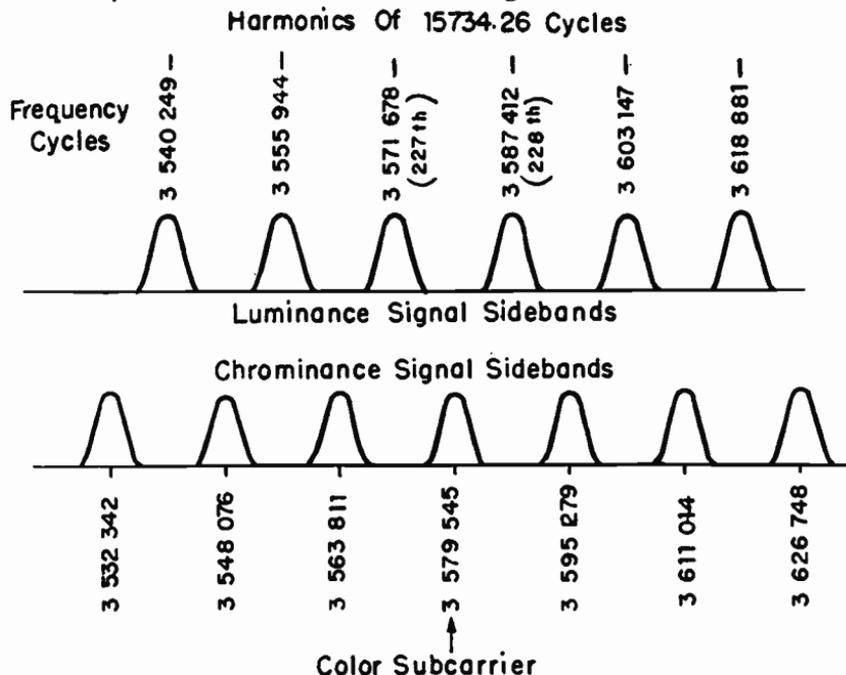


Fig. 9.—Luminance signals and chrominance signals are interleaved for transmission within the same range of video frequencies.

chrominance signal. It should be noted that the value of 3.58 megacycles for master oscillator and burst frequency is only approximate; the actual frequency is 3.579545 megacycles.

The composite color television signal as transmitted includes the Y-signal, the chrominance signal with I- and Q-sidebands, the burst signal, and also the usual horizontal, vertical, and equalizing sync pulses which are added in the transmitter circuits. The combination is called the NTSC color television signal. The letters NTSC stand for National Television System Committee.

**Interleaving.** — Fig. 8 shows relations of sidebands in the Y-signal and the I- and Q-components of the chrominance signal as they would appear after demodulation in a receiver. The video intermediate is considered to be zero frequency. The luminance or Y-signal is the equivalent of the video signal for black-and-white transmission. It includes frequencies from near 50 cycles per second to about 4.1 megacycles. The Q-component of the chrominance signal has sidebands extending both ways from 3.58 mc for 0.5 or 0.6 mc, while the I-component has sidebands to 0.5 or 0.6 mc on one side and to about 1.5 mc on the other side.

It is apparent that parts of the video frequency range used for the luminance signal are used also for the chrominance signal, yet these two signals must remain separate during transmission and reception. Separation is possible because neither kind of signal extends continuously throughout the video range, but consists of separated concentrations of amplitude around harmonics of the horizontal line frequency.

The horizontal line frequency for color television is 15,734.26 cycles per second instead of the 15,750 cycles used for black-and-white. Luminance signal sidebands concentrate around harmonics of this color line frequency, as shown at the top of Fig. 9. Chrominance signal sidebands would concentrate around harmonics of the color line frequency, but because the color subcarrier (3.579545 cycles per second) is an odd multiple of half the line frequency, concentrations of chrominance sidebands are shifted to half way between luminance concentrations. The color subcarrier frequency falls

midway between the 227th and 228th harmonic of 15,734.26 cycles. This "interleaving" of luminance and chrominance signals allows both to be transmitted within the same video-frequency range without interfering with each other to any noticeable extent.

**Color Receiver.** — To utilize the composite color signal for production of primary colors and their combinations at the picture tube the transmitter processes which produce the composite signal from primary colors at the camera tubes must be reversed in the receiver. Principal sections of a fairly typical color receiver are represented in Fig. 10. Many sections are similar to or may be exactly like corresponding sections of black-and-white sets, while other sections are decidedly different or are found only in color receivers.

Sets designed to reproduce pictures only in white, black, and shades of gray may be called either black-and-white receivers or monochrome receivers. As ordinarily used the two names have the same meaning.

The tuner, i-f amplifier, and automatic gain control (agc) sections in color receivers are essentially the same as for black-and-white or monochrome, with the exception that the i-f amplifier must provide nearly uniform gain to frequencies of 4.0 to 4.1 mc, as shown by the luminance response curve of Fig. 8.

Many color sets have a single video detector, as in monochrome sets, but in Fig. 10 there are two detectors, one for the luminance signal and the other for chrominance and sound signals. Following the luminance detector is a luminance or Y-amplifier quite similar to a video amplifier for black-and-white. The luminance amplifier feeds the luminance or Y-signal to the cathodes of the color picture tube. Between cathodes and control grids of the picture tube is a brightness control.

The output of the luminance amplifier goes also to the sync section, which is like similar sections in monochrome sets. The sync section feeds horizontal and vertical sweep and deflection sections, just as in monochrome practice. There is the usual horizontal automatic frequency control (afc) together with a horizontal oscillator, a drive control, and a horizontal

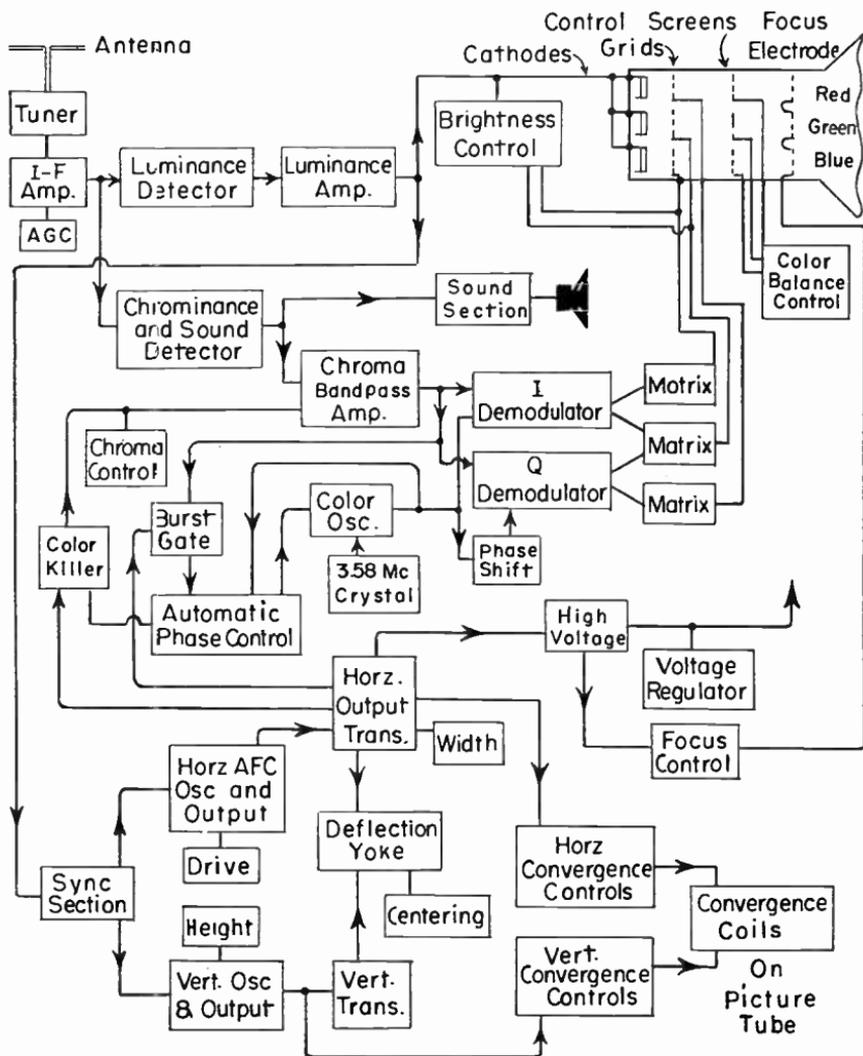


Fig. 10.—Principal sections in one type of color receiver, and signal flows between the sections.

sweep amplifier feeding to the horizontal output transformer. This transformer has more auxiliary windings and taps than needed in monochrome receivers, because of several additional pulse voltages needed for color reproduction. The width control on the output transformer is like width controls for monochrome.

Connected to the horizontal output transformer is a high-voltage rectifier in a flyback system. For three-gun picture tubes the high voltage for the ultor or second anode should not vary with fluctuations of load or of line voltage. Accordingly an automatic voltage regulator is on the high-voltage line. Three-gun picture tubes employ electrostatic focusing, for which the required high voltage is taken through a focus control from the high-voltage rectifier system.

The sync section feeds also to the vertical sweep oscillator, output amplifier, and output transformer. Any of the usual types of height control may be employed. Horizontal and vertical output transformers connect to the deflecting yoke, which is essentially like yokes for monochrome picture tubes. Centering ordinarily is by means of adjustable direct currents in the yoke coils.

To the horizontal output transformer and to the vertical output amplifier or transformer are connected convergence controls and coils used only with three-gun picture tubes. The purpose is to bring the three electron beams together at openings in the shadow mask of the picture tube.

Returning to the output of the i-f amplifier, the composite color signal is fed to the chrominance and sound detector as well as to the luminance detector. The sound section is of the intercarrier type, utilizing the same principles found in intercarrier sound sections of black-and-white receivers.

The demodulated output of the chrominance and sound detector goes to the chroma amplifier, which may be called the bandpass amplifier. This amplifier passes and strengthens the portion of the video signal that includes the I- and Q-components of the chrominance signal, as well as the burst signal, shown by Fig. 8. The chrominance signal is applied to the I- and Q-demodulators, which reverse the action of modulators at the transmitter to recover separately the I-signal and the Q-signal.

The I- and Q-signals originally were modulated onto a 3.58-mc color subcarrier voltage from the master oscillator in the transmitter, but the subcarrier was suppressed and the 3.58-mc voltage transmitted only in the form of intermittent bursts.

For demodulation of the chrominance signal in the receiver it is necessary to have a continual 3.58-mc voltage, which is produced by the color oscillator. The 3.58-mc voltage from the color oscillator, of suitable phase with respect to the bursts, is fed directly to the I-demodulator and through a  $90^\circ$  phase shift circuit to the Q-demodulator. When the 3.58-mc voltages and the chrominance signal voltages act together in the demodulators the demodulated outputs are the I-signal and the Q-signal.

It is the function of the burst signal to maintain color oscillator voltage synchronized in frequency and phase with master oscillator voltage at the transmitter. In Fig. 10 the color oscillator is held closely to its frequency by a quartz crystal which is resonant at 3.58 mc. In addition there is an automatic phase control system similar to automatic frequency controls for the horizontal sweep oscillator in monochrome sets, quite commonly employing a phase detector. The automatic phase control compares a signal taken from beyond the color oscillator with the burst signal, and applies a voltage for frequency and phase correction back to the oscillator.

The burst signal applied to the phase control comes from the chroma or bandpass amplifier through a burst gate. The gate tube is timed or gated by pulses from the horizontal output transformer, to permit passage of signals from the chroma amplifier only during intervals when bursts occur.

Connected to the chroma or bandpass amplifier is a color killer circuit whose purpose is to bias this amplifier to cutoff while there are no color transmissions, but only monochrome signals are received. This helps prevent appearance of color in black-and-white pictures. The color killer is actuated by one voltage derived from the horizontal output transformer and by another voltage from the automatic phase control; this latter voltage exists only during reception of color signals, which include bursts. In many receivers there is no color killer.

Plus I- and Q-signals, and minus (inverted) I- and Q-signals go from the two demodulators to color matrixes which mix the signals and thereby form primary color voltages. This reverses the action in mixer circuits at the transmitter. The red, green,

and blue signal voltages go to the respective control grids of the picture tube. Color signal voltages to the picture tube grids are of plus phase, while the luminance signal voltage applied to the cathodes is of minus phase. A minus or inverted signal at a cathode acts on the electron beam in the same way as a plus signal at a grid. Consequently, the minus-Y or luminance signal and the plus color signals act together to vary the intensity of the electron beams.

In the color balance system connected to the picture tube are adjustments for voltages on the screens in the red, green, and blue electron guns. The screen voltages determine relative average intensities of the three electron beams, and allow balancing the light emissions from the three kinds of phosphors to produce correct or desired color effects.

**Colors and Phase.** — The hue represented by any chrominance signal depends on the phase of this signal voltage with respect to master oscillator voltage. Saturation or intensity of color depends on amplitude of the chrominance signal. These facts may be illustrated by considering first the values of three voltages in the Q-signal of Fig. 7. These are relative values fed to the Q-modulator when white, without color, is being scanned.

Percentages or relative unit values in the Q-signal are  $+31$  and  $+21$ , for a total plus value of  $52$ , and there is also  $-52$ . Plus and minus cancel, and there is no Q-signal when no color

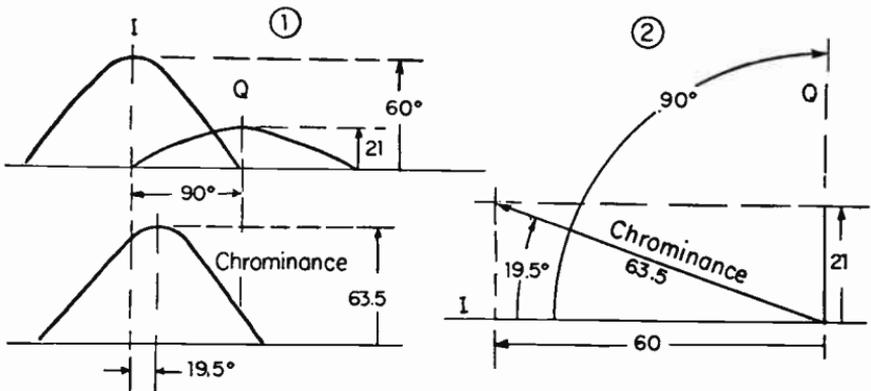


Fig. 11.—I- and Q-signals combine in this manner to form a chrominance signal corresponding to red.

is scanned. In the signal to the I-modulator there are unit minus values of  $-32$  and  $-28$ , for a total minus of  $60$ , and there also is  $+60$ . Accordingly, the I-signal cancels out when white is scanned.

Assume now that the color camera scans an area in which there is twice as much red as before, to double the red voltages in the Q- and I-signals, while blue and green remain unchanged. Combining plus and minus values of the two signals gives net relative voltages of  $21$  units for the Q-signal and  $60$  units for the I-signal, thus.

Blue	+ 31 Q	- 32 I
Green	- 52 Q	- 28 I
Red	+ 42 Q	+120 I
Totals	<u>+ 21 Q</u>	<u>+ 60 I</u>

These Q- and I-voltages will combine to form the chrominance signal, with the Q-component lagging the I-component by  $90^\circ$  in phase. Using this phase difference, and values of  $21$  units for the Q-component and  $60$  units for the I-component, the two modulating voltages combine as shown by sine-wave forms at 1 of Fig. 11, and by vectors at 2.

The chrominance signal now lags the original I-signal by about  $19.5^\circ$ , and leads the original Q-signal by about  $70.5^\circ$ . Since the I-signal already lags the master oscillator voltage (and transmitted burst) by  $57^\circ$ , total lag of the chrominance signal behind master oscillator and burst voltages is  $76.5^\circ$ . The chrominance phase angle of  $76.5^\circ$  results from a red hue in televised areas, and when suitably employed in a receiver will cause reproduction of red on the picture tube.

Unless specified otherwise, all phase angles of chrominance signals, of I and Q signals or R-Y and B-Y signals, and of  $3.58$ -mc voltages are with respect to the master oscillator and burst voltages considered as of zero phase. The master oscillator and burst phase is called the reference phase.

Returning to development of chrominance signals, assume that green instead of red is doubled in strength in an area scanned by the camera. Relative strength of the resulting Q-signal will total  $-52$ , and of the I-signal will total  $-28$

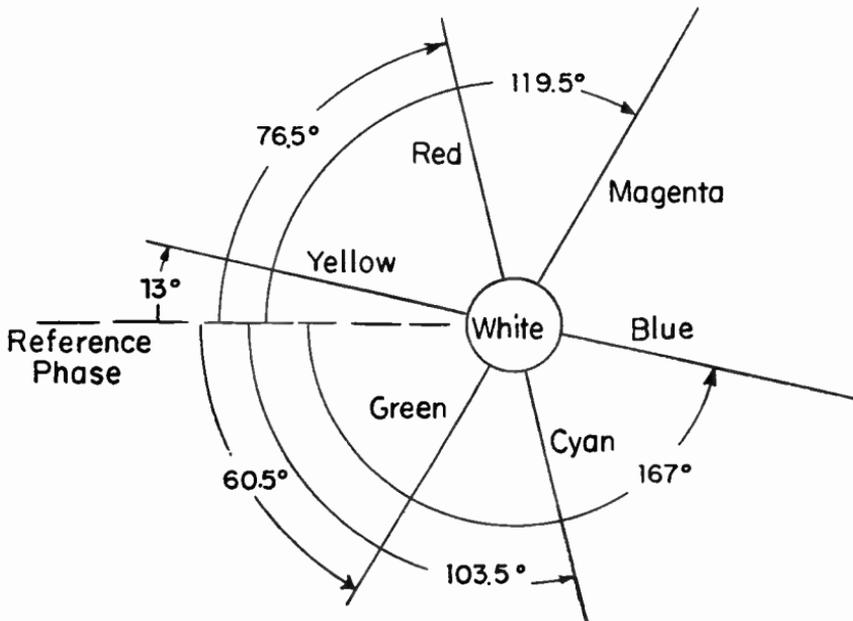


Fig. 12.—A color phase chart showing angles with respect to the reference phase for primaries and their complementary colors.

units. The chrominance signal will lead the reference phase by  $60.5^\circ$ , chrominance amplitude will be 59 units, and the hue will be green. Were blue to be doubled in the scanned area, the resulting chrominance signal would lead the reference phase by  $167^\circ$ , with amplitude of 44.5 units, and blue hue. As the color camera scans various hues, phase of the chrominance signal shifts to lag or lead the reference phase by angles up to  $180^\circ$  in either direction.

Fig. 12 shows by vector lines several hues corresponding to various phase angles of the chrominance signal with respect to the reference phase. All these chrominance phase angles result from certain combinations of color voltage values in the I- and Q-signals. Formation of red, green, and blue chrominance signals already has been discussed.

As another example, chrominance phase for yellow would result from doubling both green and red voltages going into the Q- and I-signals, with blue unchanged. The strength of the resulting Q-signal would be  $-32$  units, and of the I-signal

would be +32 units. Chrominance phase would now lag the reference phase by  $13^\circ$ .

On the color phase chart of Fig. 12 yellow is directly opposite blue, their phase difference is  $180^\circ$ . Were a chrominance signal for blue applied to the grid of a tube, phase inversion in the tube would shift plate voltage  $180^\circ$ , and at the plate would appear a chrominance signal corresponding to yellow.

Blue and yellow are complementary colors. Combining any two complementary colors in suitable proportions or strengths gives the visual sensation of white, as indicated at the center of the color phase chart. The chart shows that the complementary of red is cyan, a bluish-green hue. The complementary of green is magenta, a purplish hue. Any one complementary color phase may be changed to the other by  $180^\circ$  phase inversion in a tube.

**Receiver Demodulation.** — In the receiver are two demodulators that recover color signals, such as I- and Q-signals, by combining chrominance signals with 3.58-mc voltages originating at the color oscillator in the receiver.

Although phase and frequency of the receiver color oscillator are synchronized with phase and frequency of the transmitter master oscillator, by the burst signal, voltage which originates at the color oscillator may be shifted in phase before going to the demodulators. It is possible also to shift the phase of chrominance signals after they enter the receiver. Either or both such shifts would change phase relations between chrominance signals and color oscillator voltage.

If phases of two 3.58-mc voltages applied to the two demodulators in the receiver are the same, with respect to burst voltage, as phases of 3.58-mc voltages applied to transmitter modulators, the receiver demodulators will recover the I- and Q-signals shown on the color phase chart of Fig. 13. The positive I-signal (+I) lags the reference phase by  $57^\circ$  and the positive Q-signal (+Q) lags another  $90^\circ$ . Negative I- and Q-signals (−I and −Q) are inverted or differ by  $180^\circ$  from their positive counterparts.

On this chart is shown vectorially how a red color voltage may be secured by suitably combining 60 units of +I with

## RECEIVER DEMODULATION

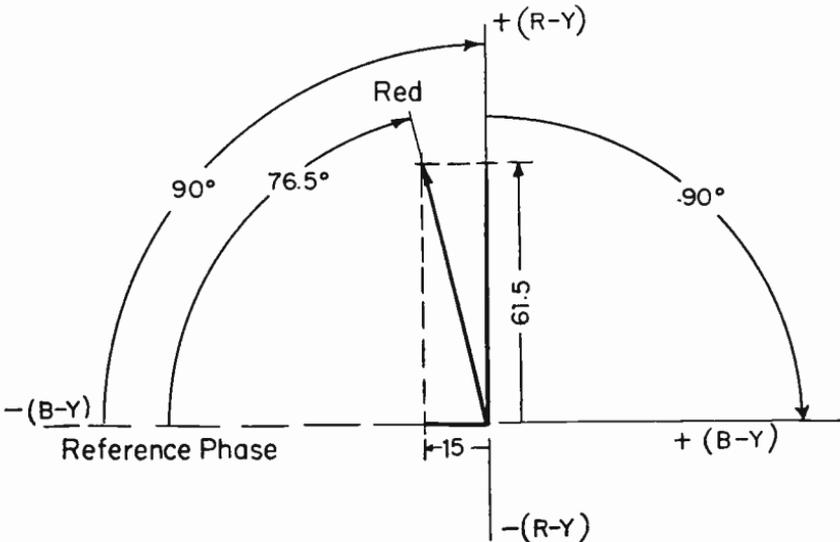


Fig. 14.—The red color phase may be secured by combining suitable proportions of R-Y and B-Y signals.

21 units of  $+Q$  signals. The red vector lags the reference phase by  $76.5^\circ$ , as in Fig. 12. A green color voltage would result from taking 28 units of the  $-I$  and 52 units of the  $-Q$  signals. Blue would result from 32 units of  $-I$  combined with 31 units of  $+Q$ .

In many color receivers demodulation is not for I- and Q-signals, it is for R-Y and B-Y signals. The results are illustrated by Fig. 14. One demodulator is supplied with a 3.58-mc voltage which produces an output lagging the reference or burst phase by  $90^\circ$ . This output is not the I-signal, instead it is a "red-minus-Y" signal, indicated by the symbol R-Y. Inversion produces a similar negative signal, so that there are positive and negative R-Y signals, usually shown by the symbols  $+(R-Y)$  and  $-(R-Y)$ .

The other demodulator no longer delivers positive or negative Q-signals, rather it delivers "blue-minus-Y" signals, marked B-Y. The positive B-Y signal is shown as  $+(B-Y)$  and the negative as  $-(B-Y)$ . Note that  $-(B-Y)$  is in phase with the burst or reference voltage.

The "Y" which is referred to in all these symbols is the

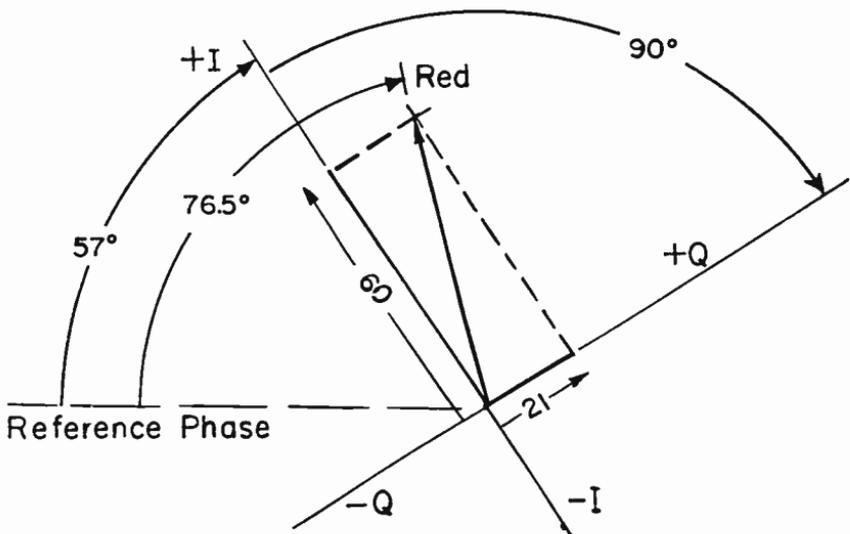


Fig. 13.—How the red color phase results from combining certain proportions of I- and Q-signals.

Y-signal or luminance signal. Then, as an example, the R-Y signal may be thought of as red with the Y-signal lacking. If the Y-signal is added to R-Y only the red color voltage will remain. The R-Y signal represents this combination of color voltages:

$$+(R-Y) = +70R - 59G - 11B$$

A positive Y-signal consists of the following:

$$+Y = +30R + 59G + 11B$$

Combining these two signals gives a total of +100 red, but positive and negative values cancel for green and blue, thus leaving only red. Similarly, combining +Y with +(B-Y) leaves only blue. Also developed in the color receiver is a green-minus-Y (G-Y) signal. Combining +Y with +(G-Y) leaves only a green color voltage. Combinations such as these are carried out in the matrixing system of the receiver.

When demodulator outputs are R-Y and B-Y it is possible to combine certain percentages or fractions of these outputs to produce the same color voltages as with I and Q outputs. Fig. 14 shows vectorially that combining 61.5 units of +(R-Y) with 15 units of -(B-Y) will produce a red color voltage lagging the reference phase by 76.5°, just as when red is produced

by fractions of I- and Q-signals. For green it would be necessary to take 51.5 units of  $-(R-Y)$  and 29 units of  $-(B-Y)$ . Blue would result from 10 units of  $-(R-Y)$  and 44 units of  $+(B-Y)$ . Phase angles for green and blue will be the same as when demodulating I- and Q-signals.

Either positive or negative R-Y and B-Y signals, or positive or negative I- and Q-signals, are secured by phase inversion in phase splitter tubes used for this particular purpose or in amplifiers used also for increasing signal strengths. Various percentages or fractions of the outputs from any kind of demodulators, phase splitters, or amplifiers are combined in matrix systems which follow the demodulators.

Demodulation is carried out by different circuit arrangements in various makes and models of receivers. One of the simplest methods employs two twin-diodes as in Fig. 15. The 3.58-mc voltage from the color oscillator goes to the plate and cathode tied together in each demodulator, with  $90^\circ$  phase shift for the B-Y demodulator.

With only the 3.58-mc voltage, and no chrominance signals, rectified currents flow through diode load resistors in such directions as to have positive voltages at the cathode ends and

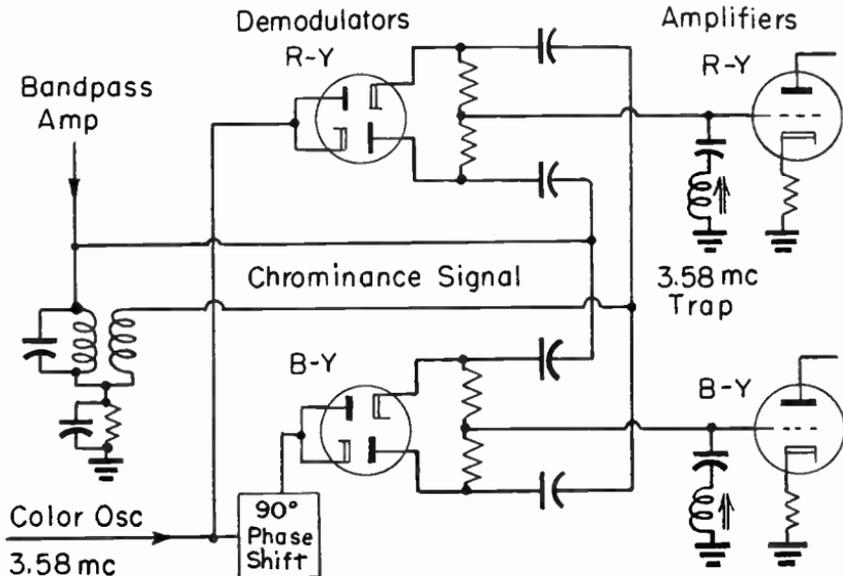


Fig. 15.—Twin-diodes used as R-Y and B-Y demodulators.

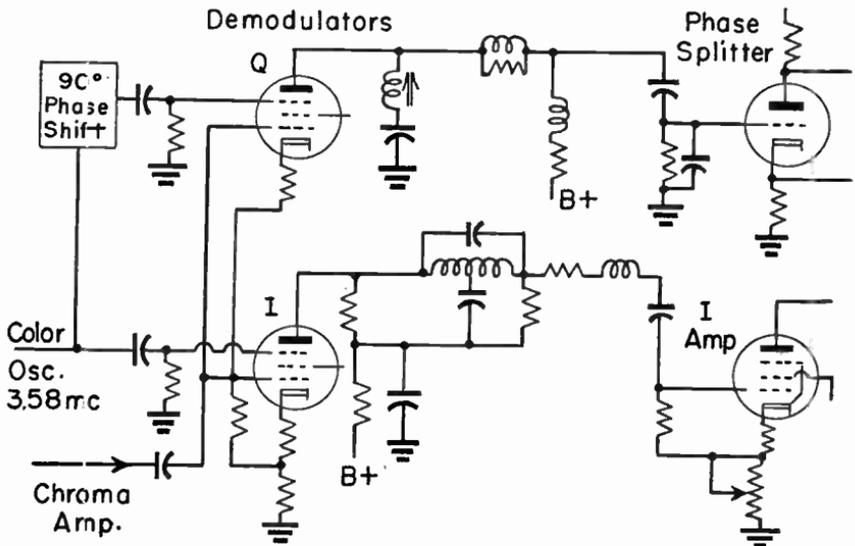


Fig. 16.—A method of demodulating to secure I- and Q-signals.

equal negative voltages at the plate ends of the resistors, with zero voltages at the center taps which connect to grid inputs of following amplifiers.

When chrominance signals are applied, the demodulator load voltages are unbalanced, and differences which are positive or negative and of varying amplitude go to the amplifier grids. A chrominance signal in phase with a 3.58-mc voltage causes maximum positive output from a demodulator. Opposite phase causes maximum negative output. A 90° lead or lag of chrominance phase with respect to the 3.58-mc voltage reduces the output to zero. Intermediate phase differences cause proportional changes of demodulator output voltage. Since there is 90° phase difference between the 3.58-mc voltages in the two demodulators, a chrominance signal of such phase as to cause maximum positive or negative output from one demodulator will at the same time cause zero output from the other.

A different demodulator circuit, shown by Fig. 16, employs two pentodes. The third grids, used as suppressors in a pentode amplifier, are here employed as additional control grids, to which are applied 3.58-mc voltages from the color oscillator

section. The 3.58-mc voltage to one demodulator is  $90^\circ$  out of phase with that to the other demodulator.

Chrominance signals are applied to the first grids of both demodulators, which are the control grids for pentode amplifier operation. Chrominance signal voltage and the 3.58-mc voltage act on the same electron stream which flows from cathode to plate. Plate output voltage varies in polarity and amplitude as phase of the chrominance signal varies with respect to that of the 3.58-mc voltage in each demodulator. Still other demodulators employ two triodes, usually in a twin tube. Chrominance and oscillator voltages are applied to the grids, with outputs taken from plates and cathodes.

The essential features of all demodulators are that changes of chrominance phase shall vary the output polarity, while changes of chrominance signal amplitude shall vary demodulator output voltage proportionately.

**Color Oscillator Section.** — The color oscillator in the receiver must be held precisely synchronized with the master oscillator at the transmitter, not only in frequency but also in phase. Master oscillator frequency and phase come to the receiver in the burst signals, which consist of eight or nine cycles at 3.58 mc frequency occurring only intermittently—only after each horizontal sync pulse. A crystal resonant at 3.58 mc is used in the oscillator grid circuit chiefly to maintain oscillation at practically constant frequency during intervals between bursts, this “flywheel” effect resulting from the extremely high Q-factor of the crystal.

It is the purpose of the automatic phase control system to hold oscillator voltage precisely in phase with burst voltage. All automatic phase controls (apc) operate essentially like horizontal automatic frequency controls (afc) used in all receivers. A frequency and phase taken from somewhere on the output side of the color oscillator is compared with frequency and phase of the burst voltage. Any difference results in a correction voltage which is applied directly or indirectly to the resonant circuit of the color oscillator.

Burst voltage comes to the apc system from the chroma or bandpass amplifier, or sometimes from a video detector. Video.

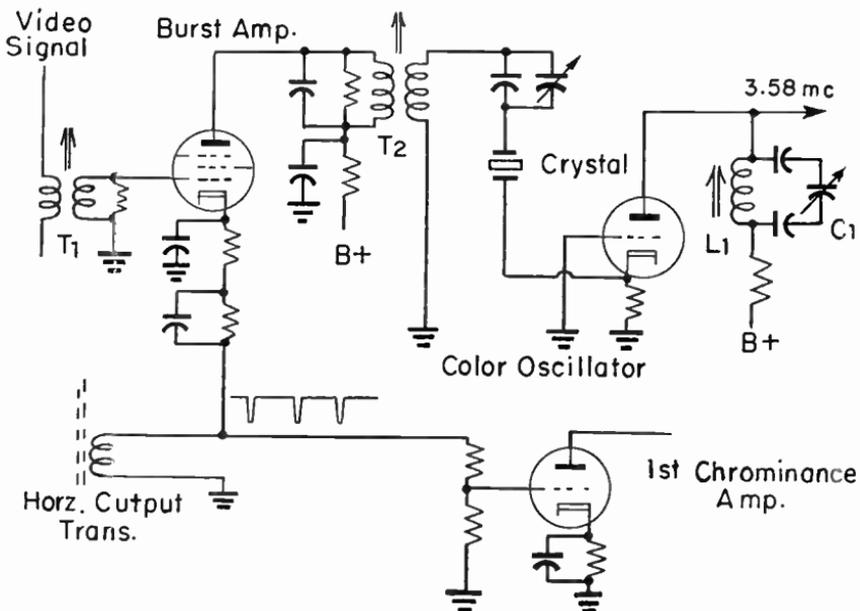


Fig. 17—The burst amplifier is gated to pass only the bursts, while excluding other signals.

luminance, and chrominance signals are kept from the apc system, and only bursts are passed, by gating the input to the apc system or by gating an amplifier tube which brings burst voltage to the crystal circuit of the color oscillator.

Fig. 17 illustrates one method of gating the burst input to a color oscillator. The video signal, carrying all frequencies, comes to the grid of the burst amplifier through transformer  $T_1$ , tuned sharply to 3.58 mc. The burst amplifier feeds through transformer  $T_2$ , also tuned to 3.58 mc, to the oscillator crystal circuit. The plate circuit of the oscillator is tuned to 3.58 mc by adjustable inductor  $L_1$  and capacitor  $C_1$ .

Cathode bias on the burst amplifier holds this tube at cutoff except while pulses of negative voltages from a winding on the horizontal output transformer are applied to the cathode circuit. These pulses occur during horizontal blanking periods, in which occur also the bursts. Thus the burst amplifier passes only burst signals to the color oscillator. The same pulses of negative voltage go also to the grid of the first chrominance amplifier to cut off the chrominance section during bursts and

thus prevent the strong burst voltages from accompanying the chrominance signals.

Many other gating systems are in use. Some provide a separate gating tube ahead of the burst amplifier. Others gate the burst amplifier by leaving its screen at zero voltage, for cutoff, except when pulses of positive voltage from a flyback transformer winding are applied to the screen. Instead of transformer *T2* of Fig. 17 there often is an automatic phase control system between the burst amplifier and color oscillator.

From the output of a color oscillator it always is necessary to have one 3.58-mc voltage differing in phase by  $90^\circ$  from another 3.58-mc voltage—for application to the two demodulators. Many commonly employed  $90^\circ$  phase shift circuits contain an adjustable inductor and a capacitor which may be made series resonant at 3.58-mc, and operate because of the fact that voltage across the inductor in a series resonant circuit leads the current by close to  $90^\circ$ . One of the required 3.58-mc voltages is taken directly from the oscillator output, with the other from the junction between the inductor and capacitor of the series resonant phase shift connection. Other circuits are based on the fact that voltage at resonance from the secondary of a doubly tuned transformer has phase displacement of  $90^\circ$  with respect to voltage in the primary of the same transformer.

**Color Controls.** — A color receiver has two controls different from anything found on monochrome sets. One is for hue and the other for saturation. Usually these color controls are accessible to the operator. The hue control may go by this name, or be called a phase control, a color shading control, or something equivalent. The saturation control may be called a color intensity control, a chroma control, or by other names.

Capacitor *C1* of Fig. 17 is a hue control which allows moderate changes of color oscillator phase with respect to phase of the burst signal. At *A* of Fig. 18 is a hue control operating on the apc system. At *B* is a hue control on the circuit feeding video signals to the burst amplifier. When a circuit which can be adjusted for resonance at an applied frequency actually is

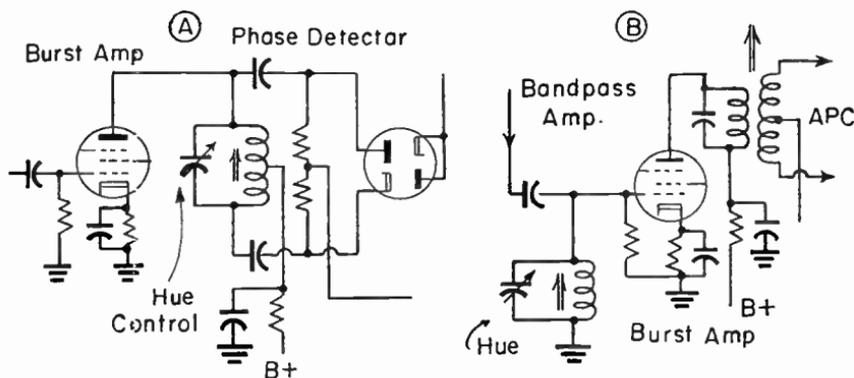


Fig. 18.—Examples of controls which change the hues of reproduced pictures.

tuned for a lower frequency, voltage is forced to lag the current, while if tuned for a higher frequency, voltage is forced to lead the current. This is the basic principle of the hue controls illustrated, and of many others.

A saturation control varies the gain applied to the chrominance signals, and thus varies their amplitude when reaching the demodulators. Since amplitude of the chrominance signal determines color saturation or color intensity, anything which varies chrominance gain varies saturation. Saturation controls act on one or more tubes in the bandpass or chroma amplifier section.

When the luminance or Y-signal remains of constant strength, reducing chrominance amplitude leaves more effect of the Y-signal, and since the Y-signal represents white this adds more white to all colors and thereby reduces saturation. More chrominance amplitude makes the white Y-signal less effective, and increases saturation.

Color receivers, like monochrome sets, have contrast controls which vary the strength of video signals to video amplifiers. Obviously, if varying the contrast is not to affect saturation, the contrast control must be arranged to act equally on both the luminance or Y-signal and on the chrominance signal. A contrast control may be ahead of the point at which luminance and chrominance signals are separated. Otherwise there may be two contrast controls, ganged together, with one

acting on the luminance signal and the other on the chrominance signal.

If a saturation control is sufficiently effective to drop chrominance gain practically to zero during reception of monochrome transmissions, this adjustment of the saturation control may allow omission of a color killer circuit.

**Matrixing.**— The final step in securing color signals for the grid-cathode circuits of a picture tube is mixing or matrixing the I- and Q-signals or the R-Y and B-Y signals from demodulators in such proportions as to strengthen some colors and weaken or eliminate others.

One method of matrixing demodulated I- and Q-signals is illustrated by Fig. 19. The Q-demodulator is here assumed to deliver a positive Q-signal while the I-demodulator delivers a negative I-signal. These signals go to the grids of phase splitters, at whose cathodes are signals of the same polarity as at their grids, and which furnish inverted signals from their plates. Relative values of red, green, and blue for all the I- and Q-signals are shown. From the luminance or Y-amplifier comes a positive Y-signal. The signals go to resistance mixer or matrix circuits in which combine the marked fractions or

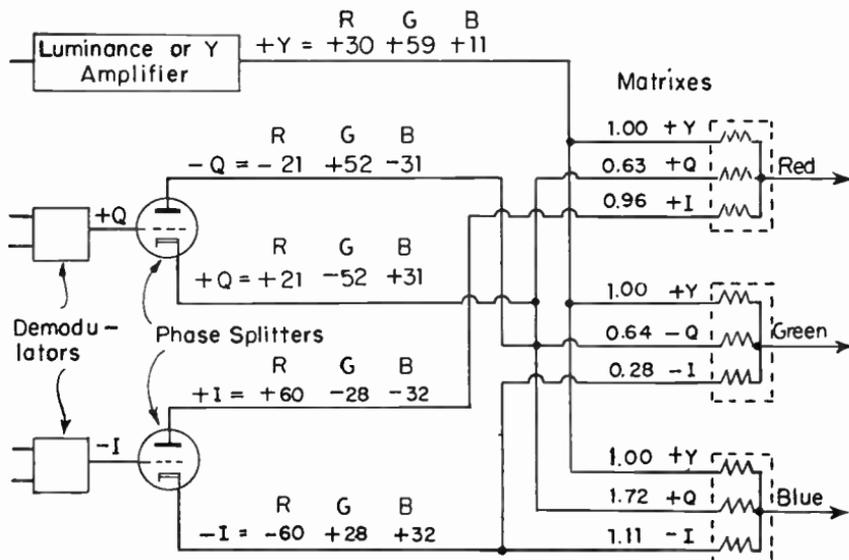


Fig. 19.—A matrix system producing primary color voltages from I- and Q-signals.

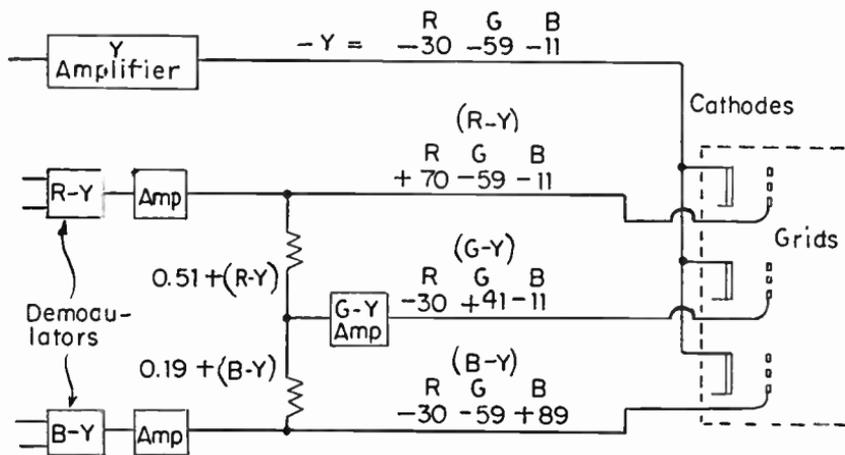


Fig. 20.—R-Y and B-Y signals are matrixed to form G-Y, while a -Y signal is matrixed with the others in the picture tube.

proportions of Y, I, and Q signals. The combinations are red, green, and blue color signal voltages.

An example of how the green color voltage is secured will illustrate the process in a general way. The green matrix accepts 100% of the +Y signal, 64% of the -Q signal, and 28% of the -I signal. Adding the percentages of red, green, and blue (algebraically) and rounding off decimal fractions shows that red and blue values cancel, while green becomes 100%.

+ Y x 1.00 =	+ 30 R	+ 59 G	+ 11 B
- Q x 0.64 =	- 13 R	+ 33 G	- 20 B
- I x 0.28 =	- 17 R	+ 8 G	+ 9 B
Sums	= Zero R	+ 100 G	Zero B

The combinations in the red and blue matrixes may be computed in similar fashion. Color voltages from matrix circuits may go to color amplifiers, then to the picture tube. Matrix systems sometimes are called adders.

Fig. 20 shows one method of deriving a G-Y signal by matrixing R-Y and B-Y signals from demodulators, and matrixing the Y-signal with the three color-difference signals between cathodes and control grids of the picture tube. The Y-amplifier or luminance amplifier furnishes a signal of

negative polarity. A negative signal applied to the cathodes of a picture tube, or of any other tube, has the same effect on electron flow or on the electron beam as would a positive signal applied to the control grid of the same tube. Both kinds of signals increase electron flow. Polarities of red, green and blue in the Y-signal should be thought of as inverted when considering their combination with color-difference signal polarities applied to the control grids.

The G-Y signal is formed as follows.

$$\begin{array}{rcl}
 \text{R-Y} \times 0.51 & = & +36 \text{ R} \quad - 30 \text{ G} \quad - 6 \text{ B} \\
 \text{B-Y} \times 0.19 & = & - 6 \text{ R} \quad - 11 \text{ G} \quad +17 \text{ B} \\
 \hline
 \text{Sums} & = & +30 \text{ R} \quad - 41 \text{ G} \quad +11 \text{ B}
 \end{array}$$

The polarities are inverted between grid and plate of the G-Y amplifier, and become.

$$\text{G-Y} \quad = \quad -30 \text{ R} \quad + 41 \text{ G} \quad - 11 \text{ B}$$

The green color voltage is formed in the grid-cathode circuit of the picture tube by combining the G-Y signal on the green control grid with the  $-Y$  signal on the green cathode, inverting the polarities of the Y-signal before adding them to G-Y. The combination becomes,

$$\begin{array}{rcl}
 \text{G-Y} & = & -30 \text{ R} \quad + 41 \text{ G} \quad - 11 \text{ B} \\
 -Y \text{ (inverted)} & = & +30 \text{ R} \quad + 59 \text{ G} \quad +11 \text{ B} \\
 \hline
 \text{Sums} & = & \text{Zero R} \quad +100 \text{ G} \quad \text{Zero B}
 \end{array}$$

Combining R-Y and inverted  $-Y$  (or  $+Y$ ) will yield 100% red, with green and blue cancelling. Combining B-Y and inverted  $-Y$  will yield 100% blue, with red and green cancelling.

There are numerous other matrixing methods, but all work out to recover separate pure primary colors, in each case cancelling all but one color voltage.

**Delay Lines.** — Somewhere in the luminance section, either before or after the amplifier, but always ahead of the matrix system, there is usually a delay line which actually does delay or hold back the luminance signal with respect to time so that chrominance and demodulated signals may catch up and reach the picture tube grid-cathode circuit at proper instants with respect to luminance signals. A delay line often looks

somewhat like a fairly short piece of large diameter coaxial cable. In the delay line is considerable inductance, usually in the form of a spiral conductor, associated with distributed capacitance.

Delay is necessary because the luminance signal passes through relatively few circuits containing inductance, capacitance, and resistance, as compared with the path followed by chrominance signals and demodulated signals. The chrominance signal may be slowed by as much as a microsecond or more, and the luminance signal must be equally delayed to have color effects in correct registration with luminance or brightness effects in pictures. Standard types of delay line provide delays of certain numbers of microseconds per foot of length, and must be cut to suit the needs of particular receivers.

**Convergence and Color Purity.** — When the three beams in the picture tube come together at a single opening in the shadow mask, and pass to the viewing screen, the beams

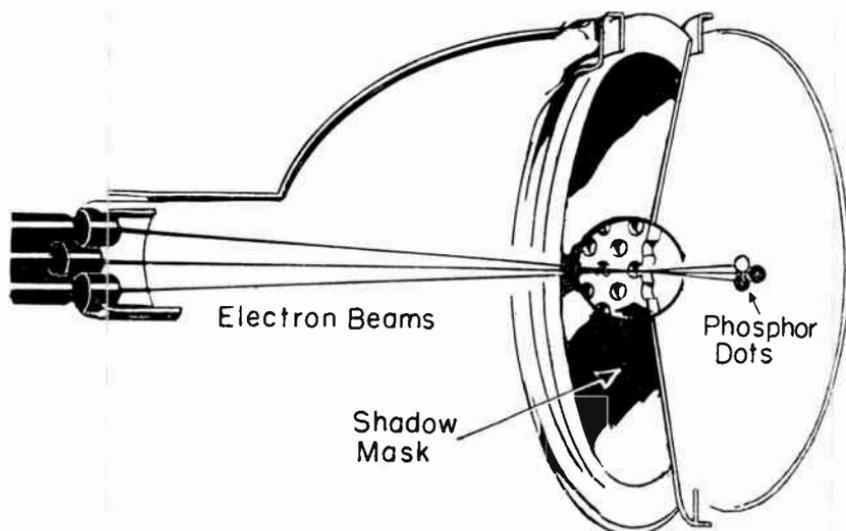
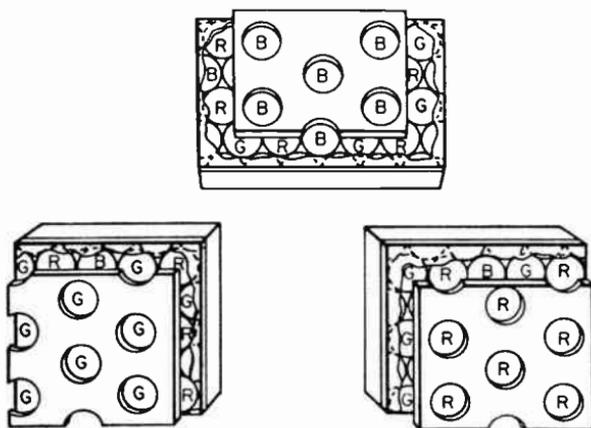


Fig. 21.—After convergence of the shadow mask, the three electron beams strike phosphor dots of three primary color emissions.

should strike three adjacent phosphor dots which form a triangle, as in Fig. 21. One dot emits red, another green, and the third emits blue light. If the beams have been made of

## CONVERGENCE AND COLOR PURITY



**Fig. 22.**—Each of the three electron beams must “see” phosphor dots of only one color.

proper relative strengths by adjusting color balance controls on the screen grids and biases on color amplifiers, and if only the luminance signal comes through, light emissions from the three phosphor dots will be such as to cause the sensation of white. The only way to produce white, or gray, on a color picture tube is to have a suitable combination of red, green, and blue light; there is no white phosphor such as in a monochrome tube.

Should one or more of the three phosphor dots be excited more or less strongly than others, the visual sensation will be of whatever color corresponds to relative intensities of red, green, and blue from the dots. The three electron beams are deflected together, horizontally and vertically, over the entire areas of the shadow mask and phosphor-dot screen. No matter where the beams are deflected, the beam from the blue gun must approach shadow mask openings at such an angle that it “sees” only blue phosphor dots, as at the top of Fig. 22. The green beam must see only green phosphor dots, and the red beam must see only red dots.

In case one or more electron beams approach the shadow mask at a wrong angle, each beam may strike dots of two or even three colors at the same time. Then there will be wrong coloring in pictures. Furthermore, it will be impossible to

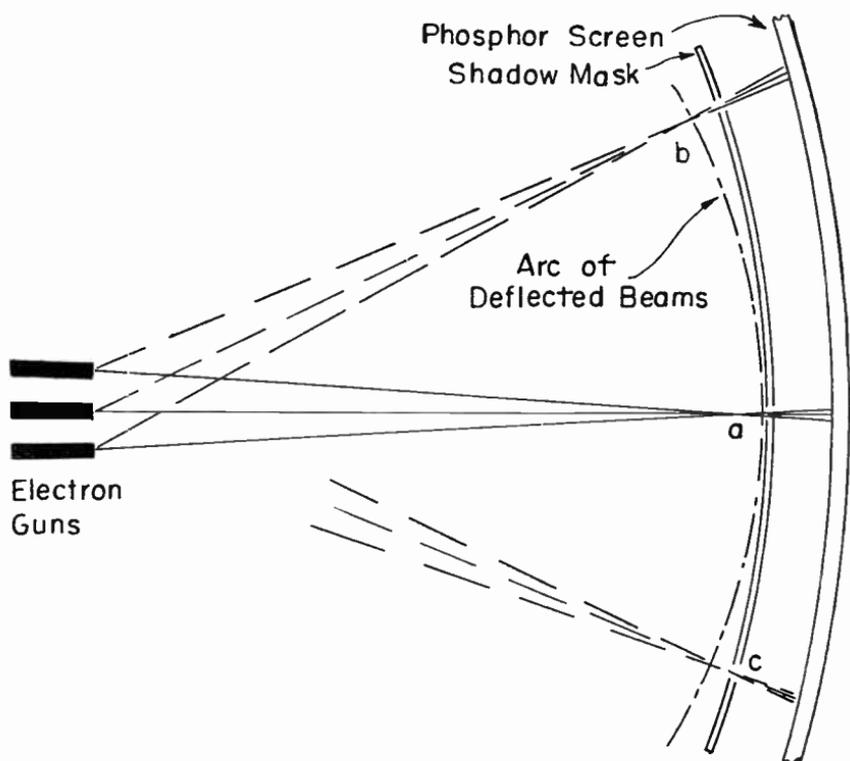


Fig. 23.—Points of beam convergence will not remain at the shadow mask when the beams are deflected, unless corrections are applied.

reproduce white or shadings of gray having no color. Monochrome pictures will be contaminated with color, and color pictures will have "impure" hues. These things may easily happen, because each phosphor dot is only about  $14/1000$  inch in diameter, shadow mask openings are about  $12/1000$  inch in diameter, and center to center separation between adjacent groups of dots is about  $25/1000$  inch.

**Convergence Adjustments.** — It is not too difficult to bring each of the three electron beams onto the appropriate phosphor dots at and near the center of the viewing screen, but many auxiliary devices and adjustments are needed to maintain this condition when the beams are deflected toward top, bottom, and sides of the viewing screen.

Fig. 23 illustrates a major obstacle in the way of keeping

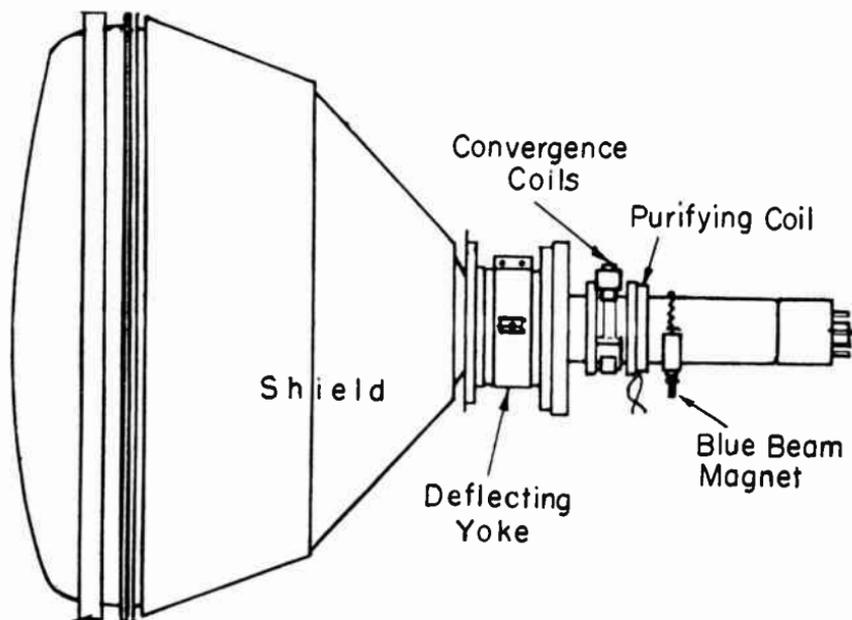


Fig. 24.—Several accessory devices for correction of convergence and purity are on the neck of a three-gun color tube.

the electron beams where they belong on phosphor dots. The shadow mask is just back of the phosphor screen, actually only about  $2/5$  inch away. Relative dimensions and shapes are exaggerated in the figure to better show what happens. Curvature of the shadow mask is not quite the same as that of the phosphor screen or face plate, but the arc followed by the deflected electron beams has less radius than the mask.

At *a* the three beams are properly converged at an opening near the center of the shadow mask, and pass through to properly excite the three phosphor dots of a group. Convergence is white similar to focusing, although concerned with three separate beams rather than with electrons in a

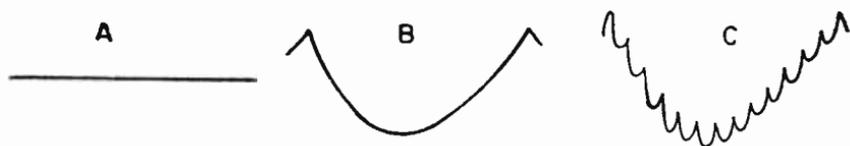


Fig. 25.—Voltages required for static and dynamic convergence.

single beam, and the three beams may be thought of as focused at point *a*.

When the electron beams are deflected away from the mask center, distance from the guns to a point of convergence at *b* does not change. But now the convergence point is behind the shadow mask, and even though all beams get through a mask opening, they will not strike phosphor dots in the proper way for good reproduction of color or monochrome. It would be possible to converge the electron beams at shadow mask openings away from the screen and mask centers, as at *c*, but then the point of center convergence would be ahead of the

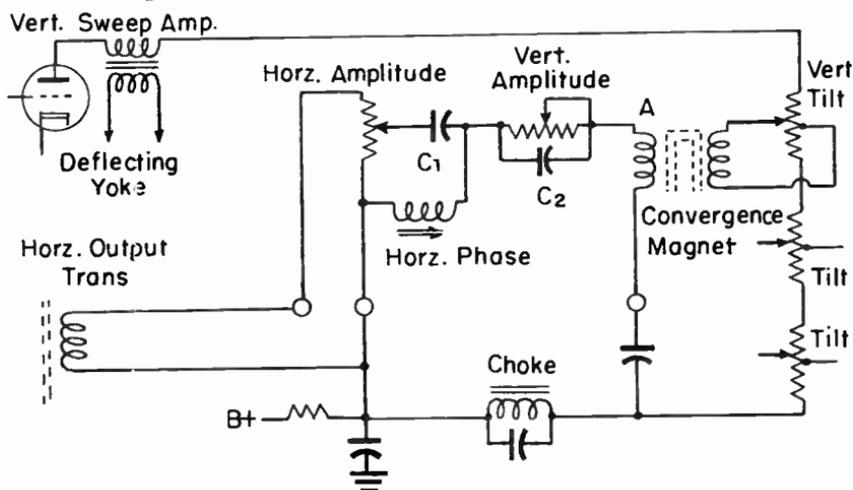


Fig. 26.—Controls for vertical and horizontal dynamic convergence voltages.

shadow mask at the center and there would be poor reproduction.

The electron beams are caused to converge by three electromagnets (convergence coils) mounted around the neck of the picture tube as shown by Fig. 24 for a CBS-Colortron tube. The three convergence coils are radially in line with the three electron guns. Direct current flowing in all three convergence coils may be adjusted for proper convergence at and near the center of the viewing screen. This usually is called static convergence adjustment, or d-c convergence adjustment. To maintain convergence as the beams are deflected, coil current

## CONVERGENCE ADJUSTMENTS

must be increased proportionately to deflection angles. To the d-c voltage required for center or static convergence are added voltages for convergence away from the center, which is called dynamic convergence. The added voltages are said to be of parabolic waveform. If static convergence voltage is represented as at *A* of Fig. 25, vertical dynamic voltage would be as at *B* for one vertical sweep. Adding the horizontal dynamic voltage, one cycle for each horizontal sweep, would make total static and dynamic convergence voltage appear somewhat as at *C*, but with more alternations.

After the beams have been converged at and near the center of the shadow mask, for static convergence, convergence in areas vertically above and below the center is secured by adjustment of the vertical parabolic voltage. This is vertical dynamic convergence. For areas horizontally to the left and right of center, convergence is secured by adjustment of the horizontal parabolic voltage. This is horizontal dynamic convergence.

Fig. 26 illustrates one kind of circuit for control of vertical and horizontal dynamic convergence, as affected by the electromagnet for only one of the electron beams. The other two

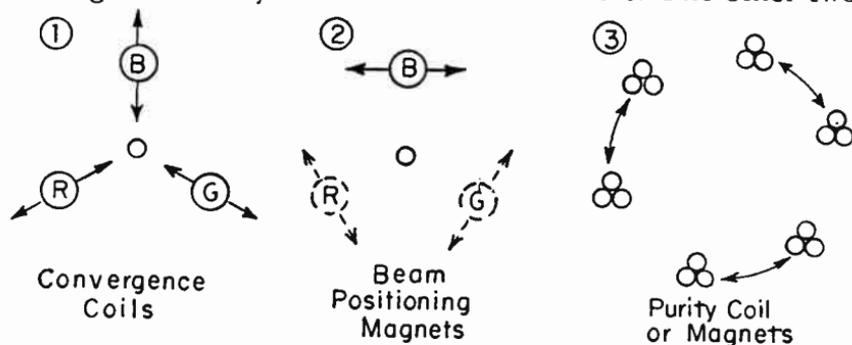


Fig. 27.—Effects of convergence and purity adjustments on the three electron beams.

beams are controlled by similar circuits for the other electromagnets, utilizing the same sources of vertical and horizontal voltages.

Each magnet consists of a single core on which are two windings. Vertical convergence voltage originates as the saw-tooth voltage in the plate circuit of the vertical sweep ampli-

fier. The plate circuit extends through the center-tapped potentiometer marked vertical tilt, in series with similar potentiometers on the other electromagnets, thence through a choke paralleled with a capacitor, and to B-plus. The choke and capacitor change the sawtooth waveform to a parabolic waveform which appears in the vertical tilt winding.

Horizontal convergence voltage originates as pulses from an auxiliary winding on the horizontal output transformer. Part of the pulse voltage is taken off by the horizontal amplitude control and put into a circuit consisting of the horizontal phase inductor and capacitor *C1*. Adjusting the inductance makes this circuit resonant at the horizontal line frequency, or at a frequency slightly higher or lower. This resonant circuit changes the pulsed voltage to a parabolic waveform, at or close to the horizontal line frequency, which acts in one of the electromagnet windings.

Vertical convergence currents flow in winding *A* because of high 60-cycle reactance in the paralleled path including the choke, and low 60-cycle reactance in the horizontal phase inductor. Amplitude of these vertical convergence currents is regulated by the vertical amplitude control, which is shunted by capacitor *C2*, having high 60-cycle reactance. The vertical tilt control changes the shape of the vertical parabolic waveform by "tilting" it one way with the slider on one side of the tap, and the opposite way with the slider on the other side of the tap.

Horizontal convergence currents, for winding *A*, are bypassed around the vertical amplitude control by low reactance at horizontal line frequency in capacitor *C2*, leaving horizontal convergence amplitude regulated by the horizontal amplitude control. Adjustment of the horizontal phase affects the horizontal convergence waveform in much the same way that the vertical waveform is affected by the vertical tilt control.

**Purity Adjustments.** — As at *1* of Fig. 27 the convergence electromagnets shift the three electron beams toward or away from what may be called the common central line or axis around which the beams are grouped. The picture tube is so

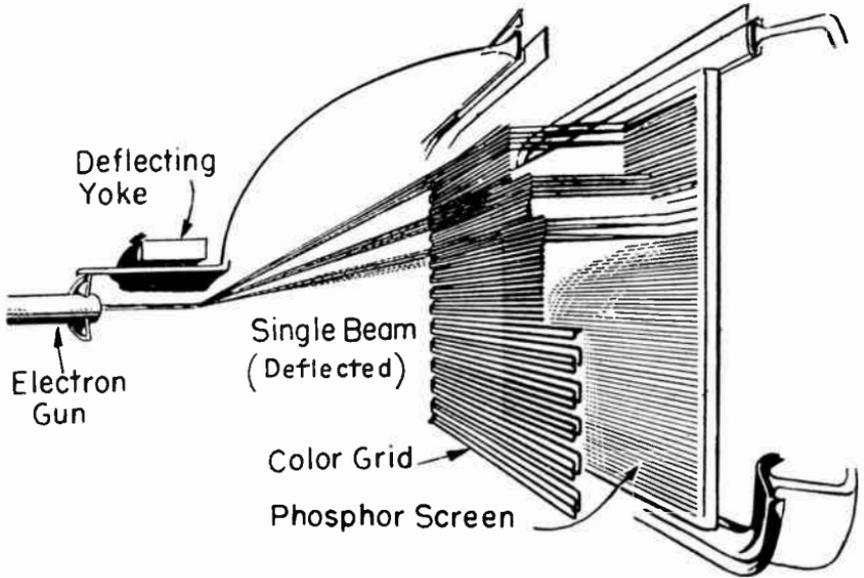


Fig. 28.—Operating principle of the Chromatron single-gun color picture tube.

positioned that the blue gun is at the top of bottom, consequently the convergence electromagnets move the blue beam straight up or down. Red and green beams always can be made to converge, because their diagonal movement lines

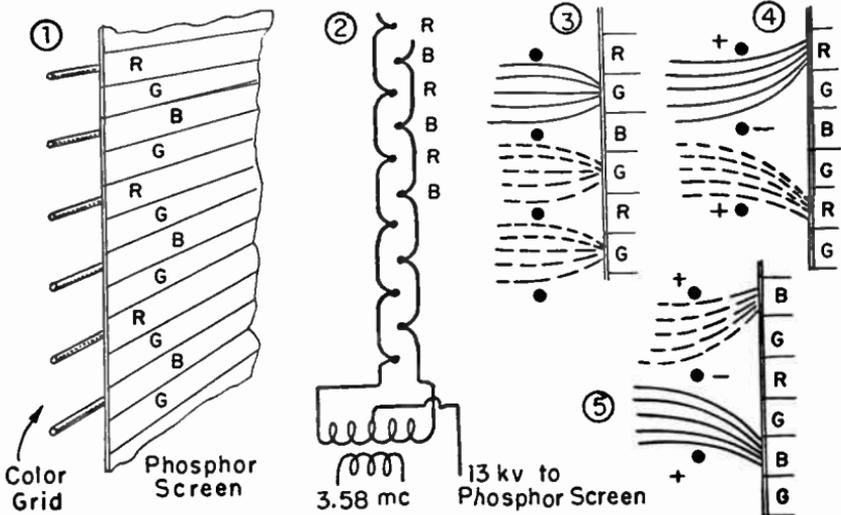


Fig. 29.—How the electron beam is focused, after deflection, onto the three color phosphors.

must meet at some point, but the path of the blue beam may not pass through this point when convergence is adjusted. To converge the blue beam it may be shifted sideways, as at 2, by means of a small permanent magnet marked blue-beam positioning magnet in Fig. 24. Sometimes, in addition to the blue beam magnet, there is a set of three more permanent magnets, supported ahead of the blue beam magnet, just back of the convergence coils.

Even though the three electron beams converge at openings in the shadow mask, their angle of approach may be such that the beams do not strike correctly on the three phosphor dots of each group. Resulting color impurity may be greatly lessened or eliminated by a purity coil or purifying coil mounted around the tube neck at the position shown by Fig. 24. This coil may be replaced by two concentric ring-shaped permanent magnets, somewhat similar to some types of permanent-magnet centering devices.

Either the coil or the permanent magnet type of purity device provides magnetic field lines which pass through the tube neck from side to side, or perpendicular to the tube axis. Rotating the coil or the magnets has the effect of moving all three electron beams, as a unit, about as shown at 3 of Fig. 27. Varying direct current in the coil, or changing relative positions of gaps in the ring magnets determines the extent to which the beams are shifted. Purity is affected also by moving the coil or the permanent magnet assembly forward or back along the tube neck.

With black-and-white picture tubes the deflecting yoke is placed as far forward on the neck and as close to the flare as possible. With three-gun color picture tubes the yoke may be as much as two inches back of the flare or bell. Moving the yoke slightly forward or back may make considerable improvement in color purity for areas near the edges of the viewing screen.

**Single-gun Color Tube.** — In the Chromatron color picture tube, often called the Lawrence tube, there is only a single electron gun. Gun elements are like those of electron guns in monochrome picture tubes, and base pin connections are

Phosphor Strips

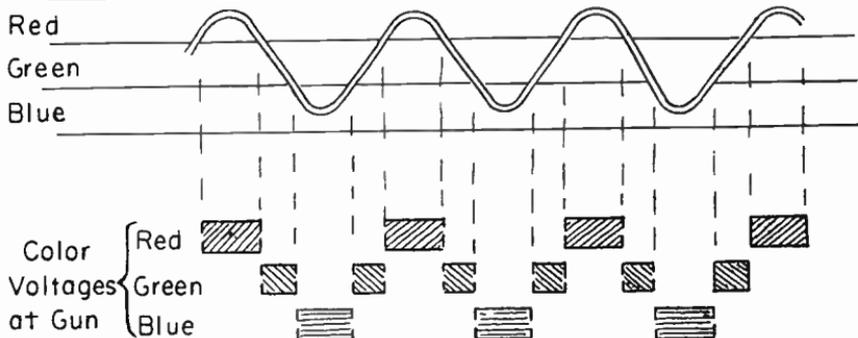


Fig. 30.—How the single electron beam traverses phosphor strips of three primary colors during horizontal traces.

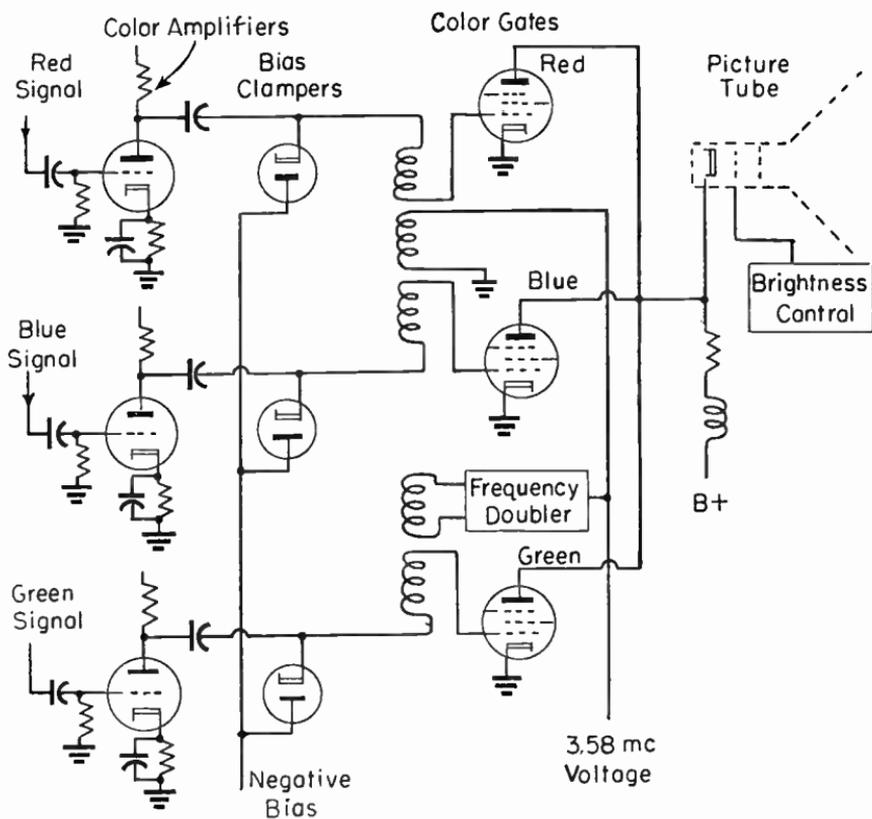


Fig. 31.—How simultaneous color signals are converted to sequential color signals for the single-gun tube.

the same. Before the single electron beam reaches red, green, and blue phosphors on a screen mounted internally immediately behind the face plate the beam passes through a color grid consisting of a great many thin parallel wires extending horizontally. This is shown by Fig. 28.

The color phosphors are in the form of horizontal strips arranged as at 1 in Fig. 29. Each strip measures about 10/1000 inch from top to bottom. Every second strip is a green phosphor, with red and blue strips alternating between the green ones. Directly back of each red phosphor strip and back of each blue strip is one wire of the color grid.

As shown at 2, all color grid wires back of red phosphor strips are joined together, at their outer ends, and all wires back of blue strips are joined together. The common lead for red wires goes to one end, and the common lead for blue wires to the other end of a center-tapped secondary on a transformer supplied with voltage at 3.58 mc. When wires back of red phosphor strips are at maximum positive potential with reference to the center tap, wires back of blue strips are at maximum negative potential. Positive and negative polarities reverse during each cycle. As the 3.58-mc voltage goes through zero, twice in each cycle, all grid wires are at equal potentials with reference to one another.

At all times there is a steady potential difference of about 13,000 volts between the entire color grid, considered as a unit, and an aluminized backing on the phosphor screen. This potential difference forms electron lenses between the color grid and phosphor screen. The result, illustrated at 3 of Fig. 29, is that while all grid wires are at equal potentials with respect to one another, the electron beam is focused onto green phosphor strips, which are midway between grid wires. While red wires are positive and blue ones negative in the color grid, as at 4, beam electrons are attracted toward the positive wire, repelled from the negative wire, and are focused on red phosphor strips. At 5, with grid wire polarities reversed, the beam electrons are focused onto blue phosphor strips.

Focusing the electron beam onto red, green, and blue phos-

phor strips as the electrons pass through the color grid is called post-deflection focusing, or focusing after the beam is deflected. There is also an initial focusing of the beam as it leaves the electron gun, by means of an electromagnetic focusing coil of the same kind used on monochrome tubes.

As may be seen at the top of Fig. 30, the 3.58-mc alternating voltage applied to color grid wires causes the electron beam to move up and down across a group of three phosphor strips while tracing each horizontal picture line. During every period in which the beam is on red phosphors the grid-cathode circuit of the electron gun must be actuated only by red color-signal voltages. While the beam is on green phosphors the gun must receive only green color voltages, and while on blue phosphors the gun signal voltage must be only blue. Required changes of color voltage fed to the electron gun are indicated at the bottom of the figure. Note that the electron beam is on the green phosphor twice as often as on either red or blue phosphors.

To deliver red, green, and blue color-signal voltages to the electron gun one after the other, the simultaneous or continual color voltages of the standard NTSC signal must be changed to sequential color voltages which occur in a definite order or sequence.

One method of converting simultaneous color signals to sequential signals for the single-gun tube is illustrated in principle by Fig. 31. Simultaneous color voltages for red, blue, and green enter the system at the left, and go to the grids of three color amplifiers. All circuits ahead of the color amplifiers may be the same as used in any type of color receiver for production of red, blue, and green color voltages such as would be applied to electron guns in a three-gun color picture tube.

The three simultaneous color-signal voltages from plates of color amplifiers go to grid circuits of three color gate tubes. The 3.58-mc voltage that switches the picture tube electron beam across the phosphor strips of three colors, as in Fig. 30, is coupled into the grid circuits of the red and blue gates in opposite polarities, from opposite ends of the 3.58-mc

winding. The red gate is held at cutoff except when the 3.58-mc voltage in its grid circuit is positive. Then the red gate passes a red color signal until change of the 3.58-mc voltage again allows cutoff. Similarly, the blue gate is cut off except while 3.58-mc voltage to its grid circuit is positive. This occurs midway between gating intervals for red, because of phase reversal between voltages coupled into the red and blue gates. Then the blue gate passes a blue color signal.

Since the picture tube electron beam falls on green phosphor strips twice as often as on red or blue strips, the gating voltage for the green gate goes through a frequency doubler, and the green gate passes green color signals during intervals between red and blue and between blue and red signals.

Color voltage signals from the three gates occur one after another in proper sequence, not simultaneously. Therefore, all three gated color voltages go through the same connection from gate plates to the cathode of the single electron gun. There are, of course, various adjustments for frequency and for phasing not shown on the simplified diagram of Fig. 31. In some applications the three gates are supplied with R-Y, B-Y, and G-Y color difference voltages, with gate outputs to either the cathode or grid of the electron gun. Then a  $-Y$  signal would go to the other gun element, the grid or cathode.

**CONSTANTS, TIME.** — There are two kinds of time constants, capacitive and inductive. A capacitive time constant is the length of time required for the charge and voltage of a capacitor to reach approximately 63 per cent of their final values when the capacitor is charged through a resistance. This same length of time, or same time constant, is required for the same capacitor to lose approximately 63 per cent of its charge and voltage when discharging through the same resistance.

An inductive time constant is the length of time required for direct or one-way current in a coil or inductor to reach 63 per cent of its final value when this current flows in resistance of the coil and its circuit.

Values of capacitive and inductive time constants may be determined from the following formulas.

$$\text{Time,} = \frac{\text{capacitance,}}{\text{resistance,}} \times \frac{\text{resistance,}}{\text{microfarads}} \times \frac{\text{ohms}}{\text{megohms}}$$

$$\text{Time,} = \frac{\text{inductance, henrys}}{\text{resistance, ohms}}$$

Applied voltage does not alter time constants, because time

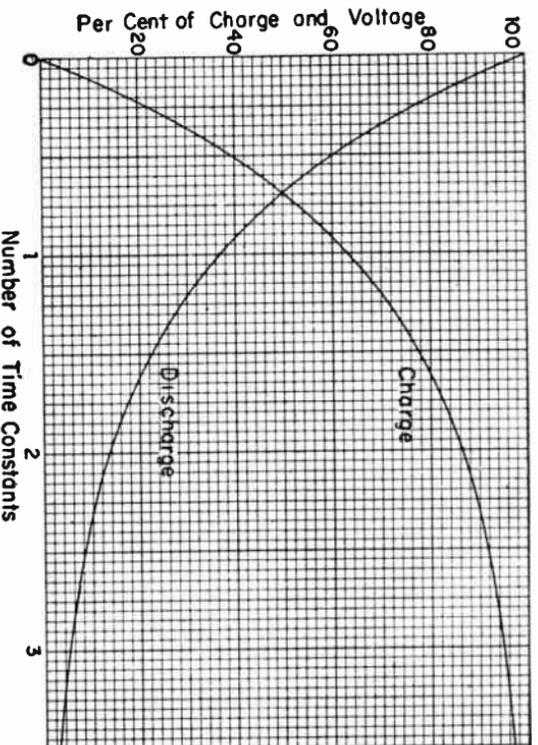


Fig. 17-1.—Rates of charge and discharge of a capacitor as related to the number of time constants.

for reaching full charge of a capacitor or final current in an inductor remains proportional to time for reaching any given fraction of full charge or of final current.

Fig. 17-1 shows percentages of charge and discharge during times measured in the number of time constants. Capacitor charge begins with a fully discharged unit; inductor charge begins with closing its circuit to a source. Capacitor discharge begins with a unit fully charged; inductor discharge begins with closing its circuit through which current flows from the inductor as a source. At the end of one time constant there is 63 per cent of full final charge, or loss of 63 per cent of full charge to leave 37 per cent. One time constant may be anything from a fraction of a microsecond to many seconds.

**CONTRAST AND CONTRAST CONTROLS.** — Contrast refers to differences between brightness of objects which appear at the same time in a television picture. There should be enough difference to make clear distinctions between adjacent or overlapping objects. There should be enough range of brightness to extend all the way from black to white, with all intermediate grays correctly rendered. Objects which should appear black should not be merely dark gray, and those which should appear white should not be merely light gray.

Contrast controls vary the gain or voltage amplification of the signal as it passes from the antenna to the picture tube. Contrast controls are gain controls, and, electrically, serve the same purpose as volume controls in sound radio.

Relations between picture contrast and gain, or contrast control, are illustrated by Fig. 18-1. The curves are brightness characteristics of a picture tube, showing relations between screen brightness and amplitude of applied signal voltage. Below the left-hand curve is represented the voltage changes of a weak signal, one resulting from a low setting of the contrast control. Corresponding changes of brightness on the

screen are shown at the right of the curve. Picture tube grid bias is assumed to be adjusted to a value such that blacks should be produced when the black level or blanking level of the signal is at the voltage for beam cutoff, and so that pure whites should be produced when the white level of the signal reaches the voltage corresponding to the line marked *White* on the screen brightness scale.

The black level of the weak signal actually is above the beam cutoff point, and the white level of this signal is below the value which produces white on the screen. As a consequence this weak signal, of too little contrast, will produce neither blacks nor whites, but only intermediate grays. Objects which should be black will be dark gray, and those which should be white will be lighter gray.

In the right-hand diagram the picture tube grid bias is unchanged, but the contrast control has been turned up to deliver at the picture tube a strong signal. Now the black level of the signal is beyond beam cutoff. Parts of the picture between cutoff and the signal black level will appear black, whereas they should be dark gray. The white level of the signal extends far beyond the value at which the screen is made

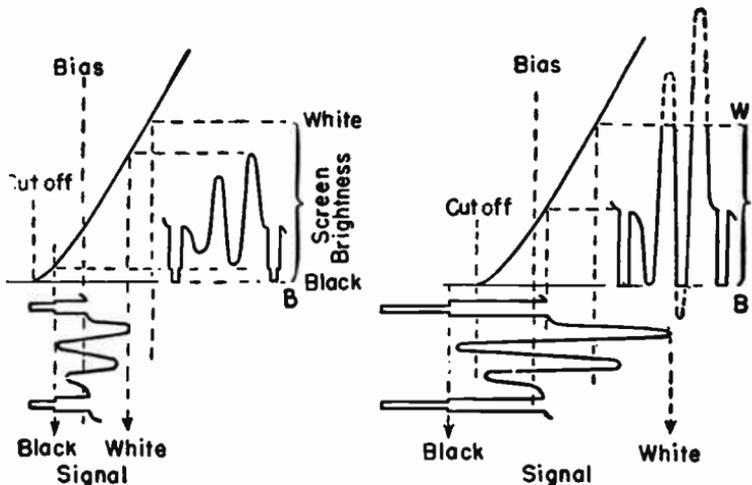


Fig. 18-1.—Relations between screen brightness and amplitude of the signal voltage applied to a picture tube.

pure white. Everything which should appear light gray now will be white on the screen, because the signal exceeds the voltage which produces white. The resulting picture will have too many objects completely black, and too many completely white. There will be too much contrast.

The conditions shown by Fig. 18-1 might be corrected by adjustment of the contrast control or by adjustment of picture tube grid bias. This bias is regulated by the brightness control of the receiver. Relations between settings of brightness and contrast controls are discussed in the article, *Brightness and Brightness Controls*.

**Contrast Controls.** — A simple but rather widely used type of contrast control is shown by Fig. 18-2. The video detector load resistor is in the form of a potentiometer, with the slider connected to the control grid of the following video amplifier tube. The percentage of signal voltage applied to the

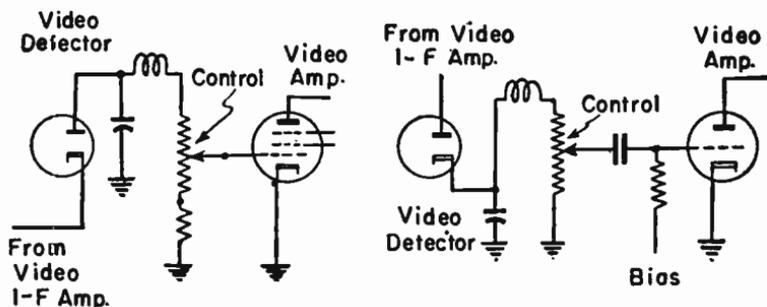


Fig. 18-2.—Contrast controls on grid input of video amplifier tubes.

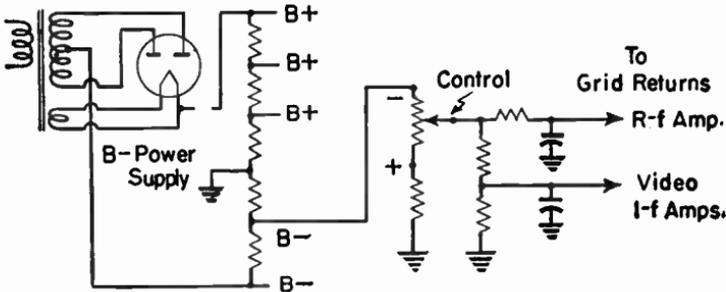
amplifier grid is that portion existing between the slider and the grounded end of the potentiometer. Contrast is increased by moving the slider toward the detector connection, and is decreased by moving the slider toward the grounded connection.

In many receivers the contrast control is an adjustable grid bias for certain of the amplifier tubes which precede the video detector. This bias control usually is applied to all the video i-f amplifiers except the last one, which operates with cathode bias in most receivers. The bias control is applied also to the

## CONTRAST AND CONTRAST CONTROLS

r-f amplifier tube or tubes. The grid returns for controlled tubes are connected to points shown in following diagrams.

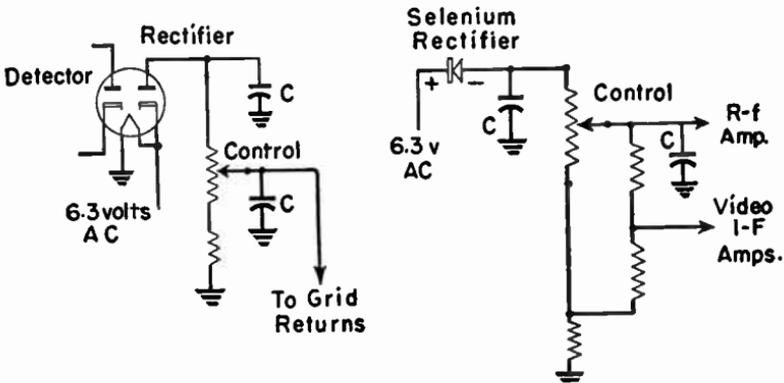
In Fig. 18-3 the negative biasing potential is obtained from a B- tap on the voltage divider system for the B-power supply. The contrast control unit is a potentiometer connected between the power supply tap and ground. Maximum negative bias voltage is applied to the r-f amplifier, and a less negative bias is applied to the video i-f amplifiers. Moving



**Fig. 18-3.**—Contrast control by means of adjustable negative voltage obtained through voltage dividers on the B-power supply.

the control slider toward the more negative end of the resistance makes the biases more negative, to reduce the gain and the contrast. Opposite movement increases the gain and contrast. This diagram shows only one application of this general method. Many variations are in use.

Fig. 18-4 shows two methods of obtaining the negative bias-



**Fig. 18-4.**—Adjustable negative bias voltage for contrast control obtained from the heater supply circuits of receivers.

ing potential from the 6.3-volt heater supply circuits of the receiver. In the left-hand diagram this a-c heater voltage is rectified by one section of the twin diode tube whose other section is the video detector. Rectified direct current flows in the resistance of the contrast control potentiometer and in the resistor which is between the potentiometer and ground. These resistances and capacitors *C* and *C* smooth the direct potential supplied from the control unit slider to the grid returns of whichever amplifiers are controlled for contrast variation.

In the right-hand diagram the a-c heater supply voltage is rectified by means of a selenium rectifier instead of by a tube rectifier. The direct current from the rectifier flows in the control potentiometer to produce a negative biasing potential applied to the grid returns of the r-f amplifier and some of the video i-f amplifier tubes through the voltage divider system connected between the slider and ground. Maximum negative bias is furnished for the r-f amplifier, and a less negative bias for the video i-f amplifiers. Both these bias potentials are adjusted together by movement of the control slider.

One of the most generally used contrast controls consists of an adjustable cathode bias resistor on the first or second video amplifier tube. Fig. 18-5 shows some of the connections employed for this general method of control. In any case, the adjustable resistor makes the tube cathode more or less posi-

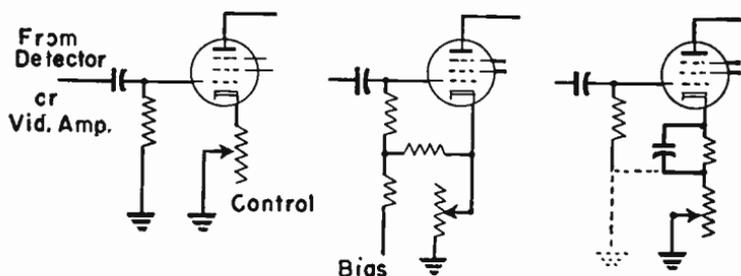


Fig. 18-5.—Contrast controls consisting of adjustable cathode bias on video amplifier tubes.

tive with reference to the control grid, thus varying the negative grid bias and the gain of this tube. Additional negative bias may be applied to the amplifier grid, as in the center diagram. Manual controls for contrast, such as shown here, ordinarily are used in connection with automatic gain control for video i-f amplifiers and a r-f amplifier. As explained in the article on *Gain Control, Automatic*, the adjustable cathode bias resistor may be connected to the agc tube from the end which here is shown unconnected.

Many other methods of gain control or contrast control may be used and are used in some receivers. Manual contrast controls most often are found somewhere in the video amplifier section of the receiver, between the video detector and the pic-

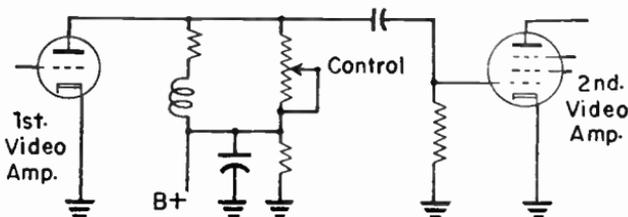


Fig. 18-6.—A contrast control in the plate circuit of a video amplifier tube.

ture tube signal input. In Fig. 18-6 the contrast control is between the first and second video amplifiers. This control consists of an adjustable resistor shunting the plate load of the first video amplifier. Reducing the control resistance and the plate load reduces the amplifier gain and lessens the contrast.

In Fig. 18-7 the manual contrast control is a potentiometer between cathode and ground of a cathode follower tube con-

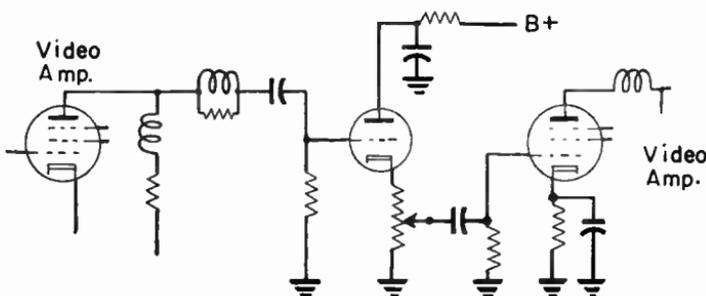


Fig. 18-7.—A contrast control on the cathode resistor of a cathode follower tube.

nected between the two video amplifiers. The signal from the first video amplifier plate is fed to the grid of the cathode follower, and appears in the control potentiometer. This potentiometer, of only 1,000 ohms resistance, provides a low impedance gain control. Moving the control slider toward the tube cathode increases the signal voltage applied to the right-hand video amplifier and increases picture contrast. There is no inversion of signal voltage between grid and cathode resistor of the follower tube.

**CONTROLS, TELEVISION.** — The controls or adjustments of a television receiver may be classified, first, as those operated by the user in obtaining the desired program and picture quality, and, second, as those employed only during servicing to place the receiver in operating condition or to correct certain faults. The first class may be called operating controls, the second may be called service adjustments or controls.

**Operating Controls.** — It is desirable, of course, to lessen the number and kind of controls which must be manipulated for satisfactory reception of a program. Controls made accessible to the operator in some receivers may be service adjustments in other receivers. In the hands of an experienced user additional operating controls allow better reception under adverse conditions, but an inexperienced or careless user is more likely to make misadjustments. The usual operating controls are as follows:

1. *Channel selector.* This is a knob, dial, or series of push buttons used to select the channel in which reception is desired. This control tunes the r-f amplifier, the r-f oscillator, and the mixer circuits to required frequencies.

2. *Fine tuning.* A fine tuning control makes slight changes of r-f oscillator frequency as may be required for best reception after the desired channel has been selected and after the receiver has had time to warm up. Not all receivers have a fine tuning control accessible to the operator.

3. *Contrast.* The contrast control varies the amplification of the signal to provide a picture containing a full range of shades from black to white, but without causing objects which

should appear either black or white actually to appear gray, and without causing those which should appear gray actually to become black or white. If contrast and brightness controls are both accessible to the operator they must be adjusted to suit each other.

4. *Brightness.* This control adjusts the bias of the picture tube control grid to bring the black level of the signal to the grid voltage for beam cutoff in the picture tube. Objects represented down near the black level of the signal then will appear black in the picture, and all sync pulses will be cut off so that they and the vertical blanking will not cause diagonal white streaks on the picture. Too much brightness makes the entire picture too light, with blacks lacking. Too little brightness makes the entire picture too dark. The brightness control sometimes is a service adjustment rather than an operating control.

5. *Vertical hold.* This control holds the picture stationary on the screen in a vertical direction, it prevents the picture from continually or intermittently shifting upward or downward. The adjustment varies the frequency of the vertical sweep oscillator to bring it into time with vertical sync pulses of the signal.

6. *Horizontal hold.* This control holds the picture stationary in a horizontal direction, preventing the picture from moving sideways either slowly or rapidly. The horizontal hold control usually will shift the picture slightly to the right or left while still keeping it from rapid or uncontrolled sidewise movement. The adjustment varies the frequency of the horizontal sweep oscillator to bring it into time or into synchronization with horizontal sync pulses of the signal.

7. *Volume.* The volume control alters the loudness of the sound program which accompanies the television picture, just as does the volume control of a receiver for sound radio. The volume control often is combined with the on-off switch. Some television receivers have controls for tone as well as for volume.

**Service Adjustments.** — Service adjustments usually are located on the rear of the receiver chassis, or sometimes they are concealed by movable portions of the front panel or a selector dial, and in some cases are found on the top of the chassis or even underneath.

1. *Focus.* The purpose of the focusing adjustment is to produce the smallest possible spot of light where the picture tube beam strikes the screen, thus allowing good definition or clear details in the objects pictured. Focusing with electrostatic picture tubes is adjusted by varying the voltage on one of the anodes in the tube. Focusing with magnetic picture tubes is adjusted by varying the direct current through a focusing coil or using permanent magnets.

2. *Centering.* The purpose of centering adjustments is, of course, to bring the center of the picture to the center of the mask opening which is in front of the picture tube screen. Centering, with electrostatic picture tubes, is accomplished by making one of a pair of deflection plates more positive and the other plate of the same pair more negative by means of a direct potential acting in addition to the alternating deflection voltages. Thus the picture, as a whole, is shifted in the direction of the plate which is made more positive. When this voltage change is applied to the vertical deflection plates the picture is shifted up or down, and when applied to the horizontal deflection plates the picture is shifted toward the right or left.

Electrical centering with picture tubes having magnetic deflection is accomplished by passing through the deflection coils a direct current in addition to the varying deflection currents. The polarity of the direct current determines the direction in which the picture is shifted, and the strength of this current determines the distance the picture is shifted. Centering currents are used in both the vertical and the horizontal deflection coils.

Mechanical centering with magnetic-deflection picture tubes is accomplished by tilting the axis of the focusing coil with reference to the axis of the picture tube which passes

through the lengthwise center of the neck. Both electrical and mechanical centering may be used on the same tube, or either method may be used alone.

3. *Size.* A vertical size control alters the height of the picture, and may be called a height control. A horizontal size control alters the width of the picture, and may be called a width control. Size is increased by increasing the amplitude or the maximum variation of sawtooth voltages or currents which act, respectively, in the deflection plates of electrostatic picture tubes or in the deflection coils on magnetic picture tubes.

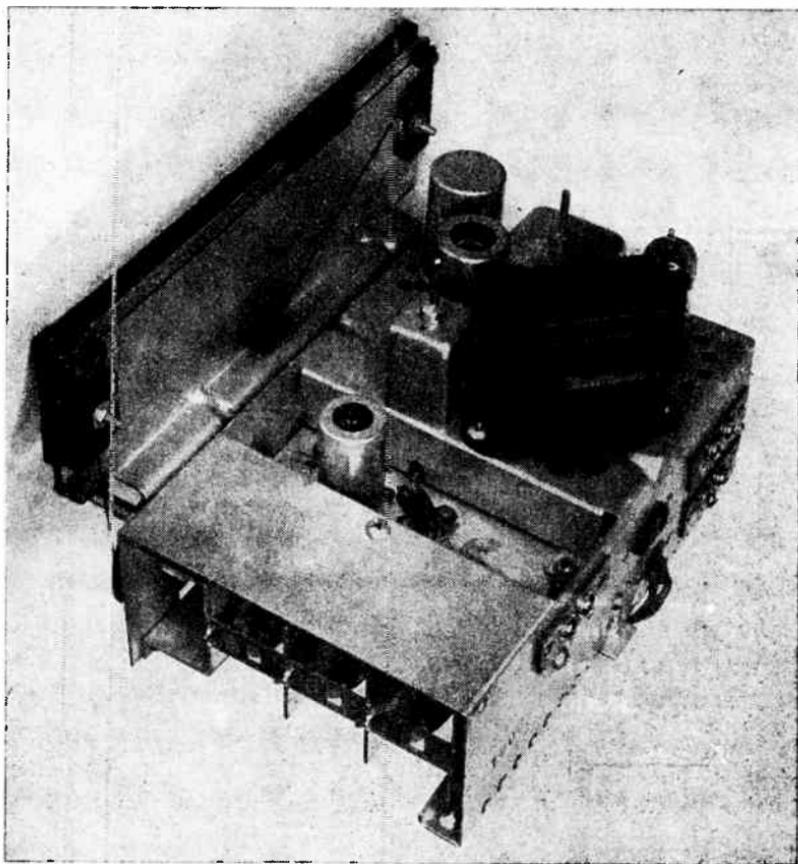
4. *Linearity.* The word linearity refers to shapes, sizes, and positions of objects and lines in the reproduced picture as these reproductions are related to actual shapes, sizes, and positions of objects and lines in the original televised scene or its image. When the outlines of objects in the reproduced picture are altered from their original forms the picture is non-linear. The purpose of linearity adjustments is to secure and maintain true proportions. Outlines may be distorted horizontally, vertically, or in both directions. Consequently there are horizontal linearity adjustments and also vertical linearity adjustments.

In general, the picture will be linear when the picture tube beam travels horizontally and vertically as uniform speeds throughout the entire deflection distances. This requires sawtooth deflection voltages or currents having a uniform rate of increase with respect to time. Since sawtooth voltages originally produced by the sweep oscillators are ordinarily not linear, corrections are made somewhere in the sweep amplifier sections by means of various kinds of linearity adjustments.

5. *Drive, peaking, etc.* These service adjustments vary the amplitude of sawtooth voltages applied to the grids of sweep amplifiers, or vary the waveform of these voltages. Drive controls affect linearity, and affect also the size of the picture because they alter the amplitude of sweep voltages.

**CONVERTERS AND CONVERSION, UHF.** — Parts of a television receiver which are different for uhf and vhf reception are those concerned with converting carrier signals to i-f signals. These are parts of the tuning system.

Uhf tuners may or may not have r-f amplifiers following channel selector circuits. Mixers sometimes are tubes, but commonly are crystal diodes. There may be a uhf oscillator tube, or oscillation frequencies may be harmonics from a vhf oscillator. Antenna signals may go to the mixer, to beat with uhf oscillator voltage. Gain which would be furnished by



*Fig. 128-1.—A converter which allows reception of uhf channels 14 through 83 on television receivers designed for channels 2 through 13.*

r-f amplifiers in vhf tuners then is added by amplifiers which follow the uhf mixer.

Uhf and vhf tuning based on this principle is shown in simplified form by Fig. 138-2. At the output of the uhf mixer appears a beat frequency equal to the difference between frequencies of the uhf carrier and uhf oscillator. This is the frequency at which the regular i-f amplifier of the receiver operates for vhf reception, and now also for uhf reception. The gain for ultra-high frequencies which would have been secured from an r-f amplifier, if present, is provided by the stage marked *UHF I-F Amp* inserted between the uhf mixer and the regular i-f amplifier section of the receiver.

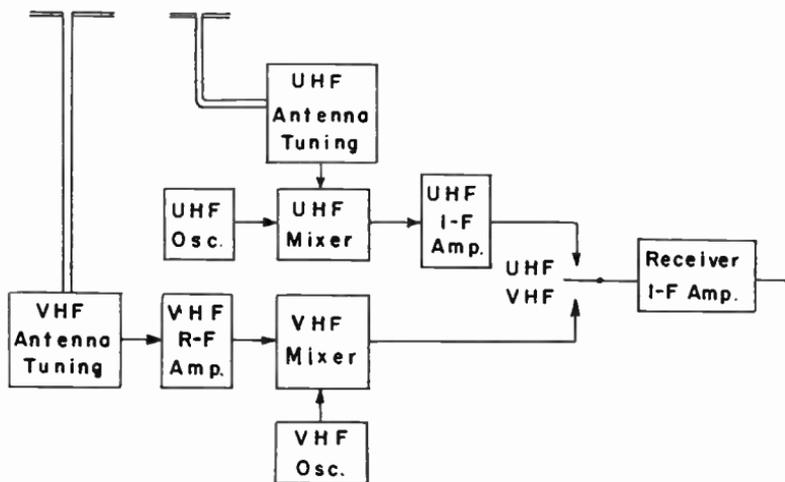


Fig. 138-2.—Tuner circuits in which there is no r-f amplifier for uhf signals.

The vhf tuner need be no different from that in any vhf receiver. It consists of the usual r-f amplifier, r-f oscillator, and mixer. A band switch connects the vhf mixer to the receiver i-f amplifier for vhf reception, and connects this i-f amplifier to the output of the uhf i-f amplifier for uhf reception. Additional contacts on the band switch would disconnect the B-voltage from the oscillator of whichever tuner is not in use.

The separate uhf i-f amplifier stage may be dispensed with when using the arrangement of Fig. 138-3. Here the vhf r-f

amplifier stage, in the vhf tuner, acts in its usual capacity and also as the amplifier required after the uhf mixer. Uhf antenna tuning, together with the uhf mixer and oscillator, are the same as before. But instead of being switched through a uhf i-f amplifier stage to the regular i-f amplifier of the receiver, the uhf mixer now is switched to the input for the r-f amplifier in the vhf tuner. During uhf reception the vhf antenna and its tuning circuits are disconnected from the vhf tuner.

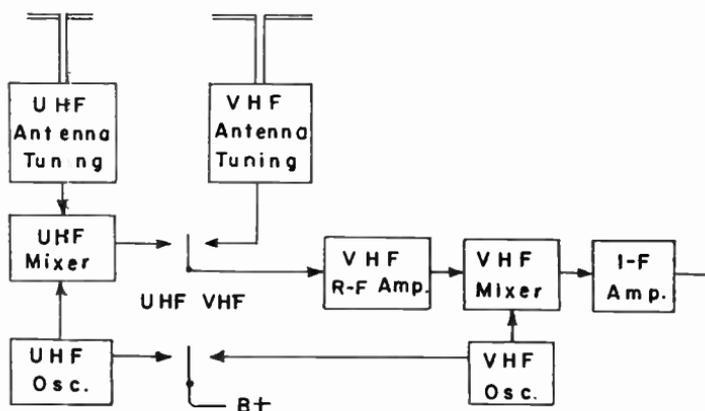


Fig. 138-3.—A system with which the r-f amplifier of the vhf tuner is used as an additional amplifier following the uhf mixer.

With this second arrangement the beat frequency from the uhf mixer may be anything for which the r-f amplifier and mixer coupling circuits of the vhf tuner may be adjusted. Oftentimes the tuner is of a type in which these circuits may be tuned to frequencies other than those of vhf channels. Then these other frequencies are used only for uhf reception. Tuning for uhf reception may be to some frequency between vhf channels 6 and 7, or to a frequency somewhat lower than for channel 2.

For uhf reception the vhf amplifier and mixer circuits are tuned to the frequency used only for uhf, which is the beat frequency from the uhf mixer. The band switch is placed in the uhf position, thus connecting the uhf mixer to the vhf r-f amplifier and at the same time applying B-voltage to the uhf

oscillator while cutting it off the vhf oscillator. When the band switch is changed back to the vhf position the uhf mixer and oscillator are cut off while the vhf antenna circuit and oscillator are reconnected as usual.

Figs. 138-2 and 138-3 illustrate the single conversion method of uhf reception, so called because frequency conversion from uhf carrier to receiver intermediate is accomplished at one step, in the uhf mixer. When employing a single conversion process the receiver intermediate frequency most often is in the 40-mc range, as with 45.75 mc for video and 41.25 mc for sound. This, however, is not always true; some sets use single conversion to intermediate frequencies in the 20-mc range.

**Double Conversion.** — Fig. 138-4 illustrates a method called double conversion. The uhf carrier frequencies first are converted in the uhf mixer to a lower frequency which is in or close to the range covered by vhf carriers. This frequency from the uhf mixer then is converted again, by the vhf mixer, to the intermediate frequency used in the receiver i-f amplifier. The vhf antenna is disconnected during uhf reception. For vhf reception this antenna is reconnected, while the uhf mixer is disconnected from the vhf tuner and B-voltage is cut off the uhf oscillator.

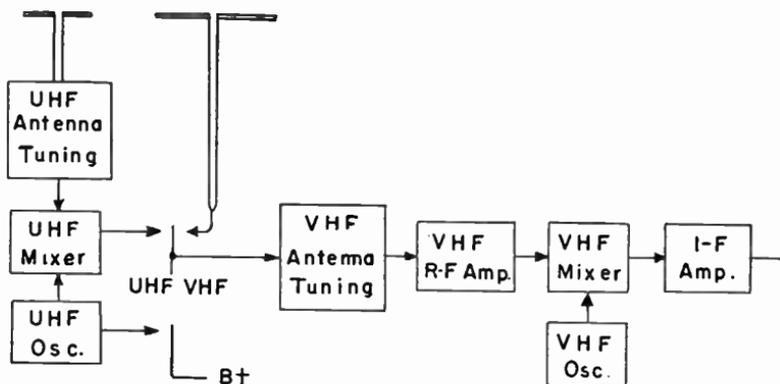


Fig. 138-4.—A double conversion system in which both the uhf mixer and vhf mixer remain in operation for uhf reception.

Double conversion is employed when separate converter units are used as attachments to allow uhf reception on any vhf television receiver. A typical arrangement is illustrated by Fig. 138-5. The converter unit contains the uhf antenna tuning circuits, mixer, and oscillator, also a stage of amplification operating at the beat frequency or uhf intermediate frequency from the mixer output. This is the frequency for which the tuner of the vhf receiver must be adjusted for reception of all uhf channels. Any particular uhf channel may be selected by tuning the uhf antenna and oscillator circuits in the converter.

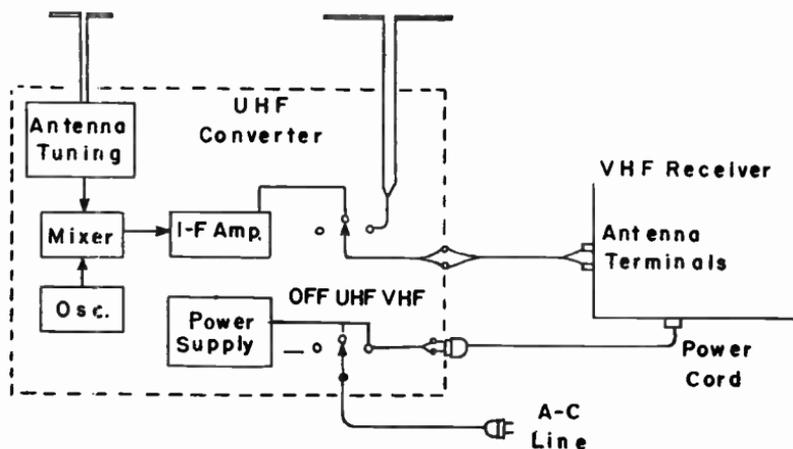


Fig. 138-5.—Circuit arrangements in a uhf converter used as an attachment to a vhf television receiver.

The a-c power line is connected into a band switch that is part of the converter. This switch has positions for *OFF*, *UHF*, and *VHF*. In either the *UHF* or the *VHF* position the a-c line is connected through the band switch to the power supply in the converter, and also to the regular power cord of the vhf receiver which is plugged into an output on the converter. With the band switch in its *UHF* position the uhf i-f amplifier stage within the converter is connected to the antenna terminals of the vhf receiver. In the *VHF* position the vhf antenna, attached to the converter, is connected through the band switch to the antenna terminals of the vhf receiver.

The converter may be designed to tune so broadly that its signal output extends over approximately the range for two adjacent vhf channels. Then the vhf tuner may be set for whichever of these two channels is not used for vhf reception in the locality.

As an example, the converter output might be centered at 82 mc, which is the high limit of vhf channel 5 and the low limit of channel 6. The vhf selector would be set for whichever of these two channels is not allocated to any nearby television station. In other cases the converter output may cover only the frequencies for one channel, and be pretuned for a vhf channel not locally allocated. The vhf tuner must be in correct alignment for the channel used for uhf reception, also for immediately adjacent channels in order to avoid pickup of unwanted vhf signals.

When employing double conversion the uhf oscillator frequency must be lower than the uhf carrier frequency to be received, this in order that video and sound intermediate frequencies for the receiver i-f amplifier may be correctly related to each other, with the sound intermediate below the video intermediate. What happens may be illustrated by an example in which the following conditions are assumed.

Uhf reception is to be from channel 26, in which the video carrier is 543.25 mc and the sound carrier 547.75 mc.

The vhf tuner is set for channel 5, in which the video carrier frequency is 77.25 mc and the sound carrier 81.75 mc.

Receiver intermediate frequencies are 25.75 for video and 21.25 for sound.

Accompanying tabulations show the resulting frequency conversions with uhf oscillator frequency below and above the uhf carrier frequencies. When oscillator frequency is higher than the uhf carriers, the final sound intermediate always will be higher than the video intermediate. This would be incorrect in accordance with the standard alignment of television i-f amplifiers.

## UHF FREQUENCY CONVERSIONS

Oscillator frequency lower than carrier frequencies	Video	Sound
Uhf carriers, channel 26	<u>543.25</u>	<u>547.75</u>
Uhf oscillator frequency	466.00	466.00
Resulting beat frequencies	<u>77.25</u>	<u>81.75</u>
Vhf oscillator, for channel 5	103.00	103.00
From uhf mixer, as above	<u>77.25</u>	<u>81.75</u>
Beats, to i-f amplifier	<u>25.75</u>	<u>21.25</u>
Oscillator frequency higher than carrier frequencies	Video	Sound
Uhf oscillator frequency	<u>620.50</u>	<u>620.50</u>
Uhf carriers, channel 26	543.25	547.75
Resulting beat frequencies	<u>77.25</u>	<u>72.75</u>
Vhf oscillator, for channel 5	103.00	103.00
From uhf mixer, as above	<u>77.25</u>	<u>72.75</u>
Beats, to i-f amplifier	<u>25.75</u>	<u>30.25</u>

**CRYSTALS, FREQUENCY CONTROL.** — Crystals for control of frequency are thin quartz plates, often called piezoelectric crystals. Crystals vibrate energetically at fundamental frequencies which depend on the manner in which the crystal was cut and on its thickness. When this crystal plate is in an electric field alternating at nearly the fundamental frequency of the crystal, the crystal is caused to vibrate at that frequency. The vibration produces, between opposite faces of the plate, alternating voltages at the fundamental frequency.

A vibrating or oscillating crystal is equivalent to a resonant circuit and may be used like such a circuit, usually between the grid and cathode of an oscillator tube. Connections for a simple crystal-controlled oscillator are shown by Fig. 20-1. Power for maintaining mechanical vibration of the crystal is fed back from the plate circuit of the oscillator tube. Since the crystal does not conduct direct current, a grid leak resistor is required between grid and cathode of the tube, in parallel with the crystal. A quartz crystal has a Q-factor of many thousands. It is more efficient than an ordinary resonant circuit

containing inductance and capacitance, and is less affected by the characteristics of any connected circuits.

The oscillating frequency varies inversely with thickness of any given type of crystal. A 2-megacycle crystal of the type commonly used for such a frequency is about 1/32 inch thick. A 6.5-megacycle crystal, again of an appropriate type, may be about 1/64 inch thick. This relation between frequency and

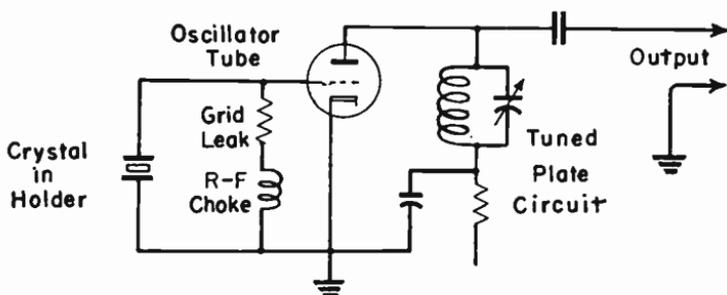


Fig. 20-1.—The circuit for a simple crystal-controlled oscillator.

thickness places practical limits on maximum fundamental frequencies, for crystals become more fragile as their thickness decreases. Fundamental frequencies seldom exceed 20 megacycles.

The maximum frequency limitation is not a serious disadvantage because, from a suitable oscillator or by the use of frequency multiplying circuits, it is possible to obtain usable harmonic frequencies of many times the fundamental. As an example, a 5-megacycle crystal will yield harmonic frequencies at every multiple of 5 megacycles all the way through to 200 megacycles, and even higher, with little difficulty.

Dual-frequency crystals will oscillate at either of two frequencies, depending on the tuning of the oscillator plate circuit. A dual-frequency crystal used in a number of signal generators will oscillate at either 0.1 or 1.0 megacycles, at 100 or 1,000 kilocycles. Harmonic frequencies are produced from both fundamental frequencies, usually as far as 20 megacycles from the 0.1 megacycle fundamental, and through 60 megacycles from the 1.0 megacycle fundamental.

The frequency accuracy of a crystal at some specified operating temperature may be expressed as a percentage, or as a maximum deviation of a certain number of kilocycles. For example, accuracy may be given as plus or minus something like 2, 5, or 30 kilocycles, with the greater deviations usually applying to crystals of relatively high frequencies. Percentage accuracies might be something like 0.002%, or 0.02%, or 0.05%, always plus or minus. These would correspond, respectively, to maximum deviations of 20, 200, and 500 cycles per megacycle of fundamental frequency.

Older types of crystals were subject to considerable frequency drift with change of operating temperature, and for maintained accuracy their temperature had to be held nearly constant by some kind of automatic heating device. Crystals of recent production undergo very small changes of frequency with moderate variation of temperature. Even in low cost units the frequency drift may be no more than 200 cycles per megacycle with the temperature varying between 14 and 140 degrees Fahrenheit. Some types of crystals have a positive temperature coefficient, their frequency increases with rise of temperature. Others have a negative coefficient, their frequency decreases with rise of temperature. The operating temperature of a crystal depends not alone on surrounding air temperature, ambient temperature, but also on the power dissipated by the crystal. Moderate oscillator plate voltage and moderate degrees of feedback help to reduce power dissipation and temperature.

Crystals are mounted in many different styles of holders. The body of the holder is of insulating material. Most often there are two extending pins which serve as mounting supports and as connections to internal metal plates which are on opposite sides of the crystal. Under the cover of the holder are insulating spacers and usually one or more pressure springs. Some of the two-pin types of crystal holders have the pins separated by  $\frac{1}{4}$  inch center to center, others have separation of  $\frac{1}{2}$  inch, and still others have  $\frac{3}{4}$  inch spacing, center to center. There are also holders with two banana plugs instead

of the solid pins, while still others are designed for mounting in standard 5-pin or else octal tube sockets. The two-pin styles mount in crystal sockets designed for the purpose. Holders having no pins may be provided with solder lugs, or may have screws for holding solder lugs, or may have wire pigtailed.

The two metal contact plates may both be in contact with the faces of the crystal, or there may be a small air gap between the top plate and the crystal. The spacing in the gap has some effect on oscillating frequency. Crystal holders used with some types of laboratory apparatus have adjustable air gaps for varying the frequency through a small fraction of one per cent. Crystals may have the metallic contacting elements plated directly onto the quartz surface. Then additional loose metal plates provide electrical connection between the crystal coating and the terminals.

Crystal controlled oscillator tubes usually are operated with plate and screen voltages below the maximum ratings for the tube. With excessive plate voltage or excessive feedback voltage the crystal may vibrate so violently as to crack, and become useless.

Crystals and holders are originally assembled with great care. The holder never should be opened unless the crystal cannot be made to oscillate in a suitable circuit. Specks of dust or an oily film due to touching the crystal surfaces may reduce oscillation or prevent it. Crystal surfaces may be cleaned with a soft brush, warm water, and mild soap, then rinsed with clear water and dried with lintless cloth or photographic lens paper. Then handle the crystal only by its edges while replacing in the holder. Do not alter the spring pressure on metal plates, since this may have some slight effect on oscillating frequency.

**CRYSTALS, RECTIFYING.** — Crystal rectifiers or crystal diodes depend for their action on unequal conductivity for currents flowing in opposite directions through the contact between a crystalline body and the tip of a fine wire which touches the crystal surface. The crystalline material in most common use is germanium, employed in germanium crystal diodes. The

contact point is the end of a fine wire of non-oxidizing metal such as tungsten or platinum. Germanium crystal diodes are used regularly at frequencies up to 100 megacycles, and in some applications up to 500 megacycles. Silicon crystal diodes are available for frequencies ranging from 3,000 to more than 30,000 megacycles.

Crystal diodes are used in television for video detectors, discriminators and ratio detectors, d-c restorers, noise limiters and pulse limiters, rectifiers for high-frequency meters, and probe detectors for electronic voltmeters, oscilloscopes, and signal tracers. There are a number of special types, including matched pairs for use as discriminators or in any full-wave rectifier circuits, also single units which have been dynamically tested for use as video detectors.

The diode units are cylindrical, ranging in diameter from about  $\frac{1}{8}$  to  $\frac{1}{2}$  inch and in length from about  $\frac{5}{8}$  to  $\frac{7}{8}$  inch for different makes. Bodies are of ceramic, glass, and other insulating materials. Some types have exposed metal end caps, others are completely insulated. Circuit connections are made, and the crystal diodes may be supported, by means of wire pigtailed built into the ends of the diodes, just as such pigtailed are used on many fixed resistors and capacitors. The negative ( - ) end of the crystal diode corresponds to the cathode of a tube rectifier or diode detector, and the positive ( + ) end corresponds to the anode or plate.

Shunt capacitance of crystal diodes, by themselves, is on the order of 1 mmf. This shunt capacitance is increased somewhat by the wiring connections, but with care in mounting it need be only about 3 mmf. Different types of diodes are designed to withstand continual maximum peak inverse voltages of from 50 to 200 volts. That is, they may be used in circuits where peak a-c voltages are from 50 to 200, or where effective sine wave voltages are from 35 to 140.

Compared with tube rectifiers the crystal diodes have advantages of requiring no heater power, of easier mounting, and sometimes of simpler wiring. The general purpose crystal diodes will carry more direct current without overheating

than will commonly used miniature tube diodes. The forward resistance of these crystal diodes is considerably less than that of otherwise equivalent tube diodes. This is an advantage where the output load resistance must be small, since performance improves with a smaller ratio of rectifier internal resistance to load resistance. A disadvantage of crystal diodes in some applications is their inability to withstand as high inverse voltages as may safely be applied to tube rectifiers. Crystal diodes, except in special types, tend to be somewhat less uniform in operating characteristics than do tube diodes.

General purpose crystal diodes, such as the 1N34 and equivalent types, may be tested with an ohmmeter capable of indicating resistances all the way from hundreds of ohms up to hundreds of thousands of ohms. Back resistance must be many times greater than forward resistance. Indications will depend on the kind of ohmmeter and on the scale used, since apparent diode resistance varies with applied voltage. The most practical meter test is comparison of readings from a doubtful unit with readings from one known to be satisfactory, using the same ohmmeter scale. Substitution of a new diode for a doubtful one is even better. When making a replacement watch the diode polarity markings, and make new connections the same as the originals. Do not solder onto pigtailed closer than  $\frac{1}{4}$  inch to the diode, and make the heating time brief.

## D

**DAMPERS AND DAMPING.** — In the vertical deflecting coils for a picture tube employing magnetic deflection there is required a sawtooth current whose frequency is 60 cycles per second. In the coils for horizontal deflection the frequency must be 15,750 cycles per second. There are, however, such values of inductance and distributed capacitance in the circuit containing the deflection coils and output transformer that any sudden change of voltage or current in this circuit could cause oscillation at higher frequencies. The higher frequencies would be the resonant frequencies corresponding to circuit inductances and capacitances. Such oscillation would

prevent uniform travel of the picture tube beam. Consequently the deflection circuits must be sufficiently loaded to stop any higher-frequency oscillations. This is called damping.

In a vertical deflection circuit, as shown at the left in Fig. 22-1, sufficient damping may be obtained by connecting across each deflecting coil a resistor  $R$  of a few hundred ohms. In a horizontal deflection circuit, as shown much simplified

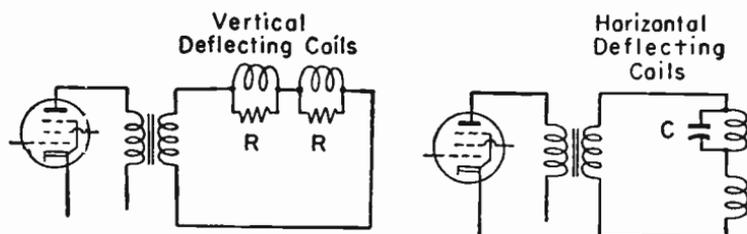


Fig. 22-1.—Simple circuits for magnetic deflection.

at the right, the tendency to oscillate is lessened by connecting across one of the coils a capacitor  $C$  of a few micro-microfarads. But because of the relatively high horizontal sweep frequency it is necessary to employ additional means for control and suppression of oscillation. These means will be described.

First it should be understood how oscillation is started and how it would continue were there no additional damping. At the top of Fig. 22-2 is shown waveform of voltage applied to the grid of the sweep amplifier tube. This is a sawtooth wave with each positive-going rise terminated by a sharp negative dip which precedes the next steady rise. The amplifier grid is negatively biased to a degree such that only about half of each sawtooth rise extends above plate current cutoff. Consequently, amplifier plate current flows only during separated peaked pulses as shown by the middle diagram. These pulses of current flow in the primary of the output transformer.

Each rise of amplifier plate current induces in the output transformer secondary and the deflecting coils a current increasing as from  $a$  to  $b$  in the bottom diagram. This is the current which causes the latter part of each horizontal trace. The rise of current produces a strong magnetic field around the

transformer and coils. When plate current is abruptly stopped by the amplifier grid going suddenly negative the magnetic field collapses. The magnetic lines contract during collapse.

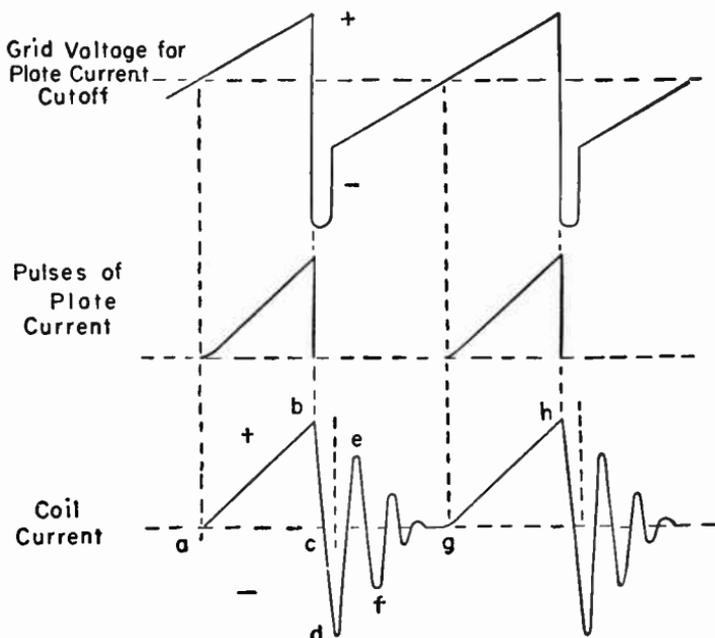


Fig. 22-2.—How oscillation may start and continue in a deflection coil circuit having no special means for damping.

whereas they expanded during growth of the field. The contraction induces in the coils and transformer secondary a current whose direction is opposite to that of the current which causes the magnetic field to expand. The opposite current is shown from *b* to *c* on the diagram: it is the start of the current which causes retrace of the picture tube beam.

Collapse of the magnetic field and decrease of current from *b* to *c* is at a rate corresponding to the resonant frequency of the coil circuit, as determined by inductance and capacitance of the coil circuit. This, of course, is the beginning of a cycle of oscillation. Since there now is little damping, the oscillatory current will continue from *c* to *d* and a new magnetic field

will be built up. Then this new field collapses and the current reverses. There are continued exchanges of energy between the magnetic fields and the circuit capacitance while the oscillatory current goes to *e*, then to *f*, and so on until it dies out because of circuit losses. Coil current would thereafter resume only with a new flow of plate current, at *g*.

During the period in which should be formed the first half of each horizontal trace, the beam actually would weave back and forth in following the current reversals from *d* to *g* of Fig. 22-2. Only during the second half of each horizontal trace period would the beam be deflected smoothly toward the right-hand side of the screen, by current changing from *a* to *b*

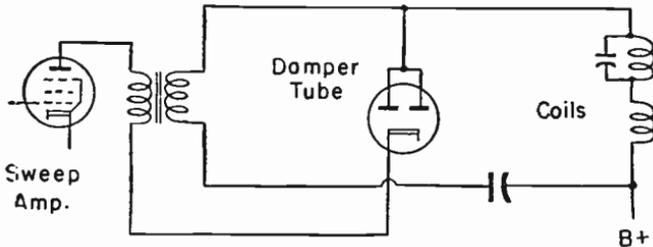


Fig. 22-3.—Connection of a damper tube in the coil circuit for magnetic deflection.

and from *g* to *h*. We need the first oscillatory reversal of current from *b* to *d*, for this is the retrace current, but to have a smooth forward trace from *d* to *g* there must be a uniform rise rather than oscillation.

To damp out continued oscillation the horizontal deflection circuit is changed as shown by Fig. 22-3. Here a diode damper tube is connected from one side of the coil circuit to the transformer primary. B-supply current flows in this path through the damper tube and the amplifier plate-cathode circuit. Because the damper tube must carry large currents at high voltage it usually is of a type such as commonly used for full-wave power rectifiers, but with its two plates connected in parallel.

Fig. 22-4 shows how oscillation is controlled by the damper tube. In the deflecting coil circuit the inductive reactance is

so much greater than the capacitive reactance that induced current lags the voltage by almost 90 degrees, and we may say that voltage leads the current by the same amount, which is almost a quarter-cycle. Consequently, by the time coil current has changed to point *b* the leading voltage has reached point

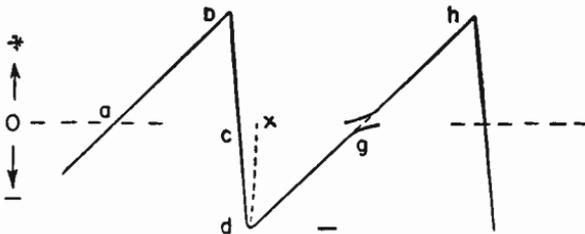


Fig. 22-4.—How oscillation is controlled and limited by the damper tube.

*c* and is making the damper plates negative. Then the damper tube is non-conductive. But by the time current has reached point *d* the leading voltage in this oscillating cycle has gone to point *x*, and is making the damper plates positive. Then the damper tube conducts and places a heavy current drain or heavy damping load on the coil circuit.

The loading or damping slows down the rate of magnetic field decay and the rate of oscillation. If the circuit is correctly designed and adjusted, the coil current which results from oscillation will decrease at a uniform rate from *d* to *g* and will die out on the zero line. Just as this current dies out there is the start of another current which is due to resumption of plate current in the sweep amplifier tube, as shown by the middle diagram of Fig. 22-2. This new current continues from *g* to *h* of Fig. 22-4. The two currents actually merge below and above the zero line to produce a smooth and uniform change all the way from *d* to *h*. Point *h* is the same as point *b* on the diagrams, and the whole process repeats for all following retraces and traces.

**Voltage Boosting by Damper.** — As shown by Fig. 22-3, the horizontal sweep amplifier receives its plate current and voltage through the damper tube. To the smooth B+ voltage from the receiver power supply are added pulses of voltage

and current produced by the quick changes of current in the deflecting coil circuit. The positive portions of these pulses are rectified by the damper tube; these are the current pulses which load the coil circuit. The voltage waveform at the damper cathode becomes as shown at the left in Fig. 22-5. In



Fig. 22-5.—Current irregularities are smoothed out by the inductor and capacitors of the linearity control.

the rising portion of this waveform are irregularities which must be removed to leave, at the lower end of the output transformer primary, a voltage waveform such as shown at the right.

Small irregularities of the voltage waveform are removed by adding a low-pass filter between the damper cathode and the transformer primary, as shown by Fig. 22-6. This filter is called a linearity control or linearity adjustment because it affects the deflection waveform through its influence on amplifier plate current. Pulses of voltage and current from the damper cathode charge capacitor  $C_a$  of the filter. This capaci-

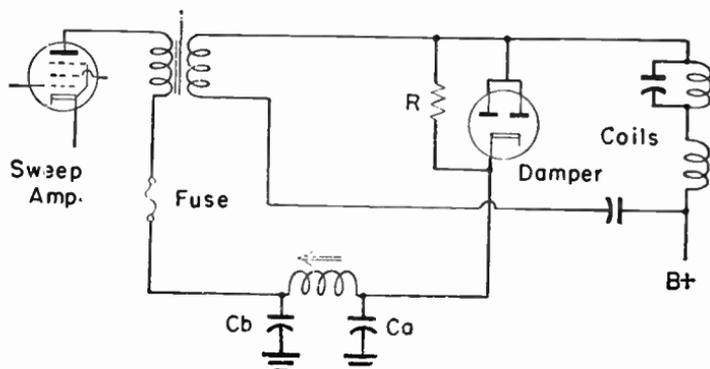


Fig. 22-6.—A horizontal linearity control connected between damper cathode and transformer primary.

tor discharges through the filter choke and the plate-cathode path of the sweep amplifier. The filter choke and capacitor  $C_b$  shift the phase of the pulses of voltage and current passing to the amplifier plate. The phasing or timing of these pulses is altered by adjustment of the core in the filter choke, so that the final result is a uniform or linear change of current in the deflecting coil circuit. Utilizing the deflection pulses or damping pulses in this manner increases the potential at the output transformer primary by 40 volts or more, in excess of the potential from the receiver B-power supply.

In the line between the filter choke and the transformer primary is a fuse, usually of  $\frac{1}{4}$ -ampere rating. This fuse blows should the sweep amplifier tube become gassy and allow plate current great enough to burn out the transformer primary if continued. Resistor  $R$ , between damper plates and cathode, provides damping effect during the periods in which the damper tube is non-conductive.

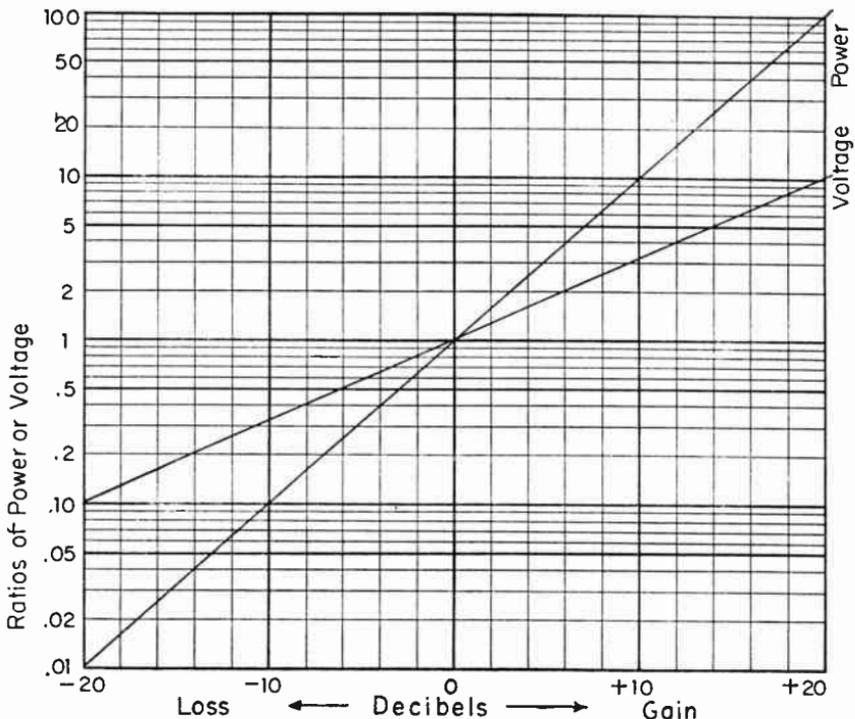


Fig. 1.—Relations between gains or losses in decibels and corresponding ratios of powers or voltages.

**DECIBELS.** — The decibel is a unit for expressing the ratio of one power to another, or of one voltage to another. Unless the value of one of the powers or one of the voltages is known, in watts or volts, a given number of decibels cannot be converted to equivalent watts or volts, because decibels indicate only how much a known power or voltage is multiplied, or how much it is divided or reduced. The abbreviation of decibel or decibels is *db*.

If two powers or voltages are equal their ratio is one-to-one, and it is expressed as zero db. Fig. 1 shows numbers of decibels from zero to 20 and the corresponding ratios of powers or of voltages for both gains and losses. When power is doubled, making the ratio 2-to-1, there is gain of +3.01 db. When power is halved, making a ratio of 1-to-2 or 0.5, there is loss of -3.01 db. Again doubling the power, for a 4-to-1 ratio, makes total gain expressed as +6.02 db. If power again is halved, to a ratio of 0.25, the loss is -6.02 db. Numbers of decibels are not directly proportional to ratios; the relation is logarithmic, as may be seen from the chart.

When expressing ratios of voltage instead of power, the number of db is twice that for the same power ratio. For example, doubling a voltage means voltage gain of +6.02 db, while halving a voltage means voltage loss of -6.02 db.

If two voltages are to be compared on the basis of db gains or losses, both voltages must appear or be measured across the same or equal impedances. Powers may be compared regardless of differences between impedances in which the two powers are developed.

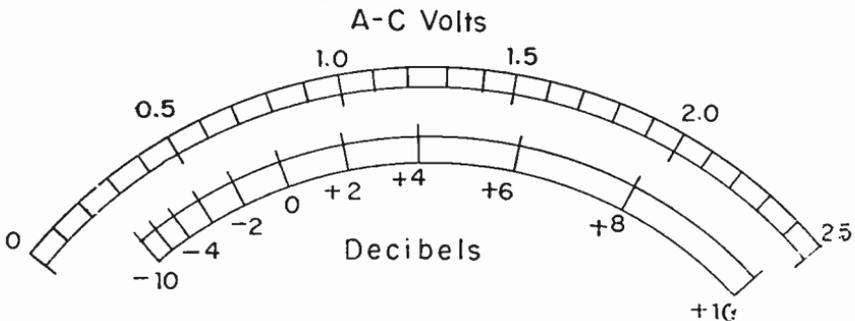


Fig. 2.—Relations between a-c volts and decibels on the scale of a meter.

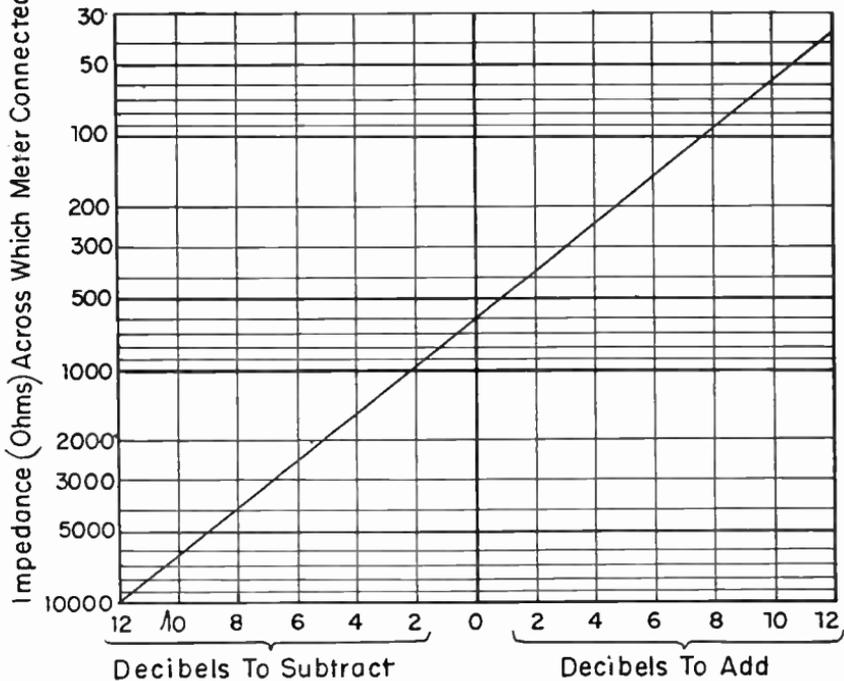
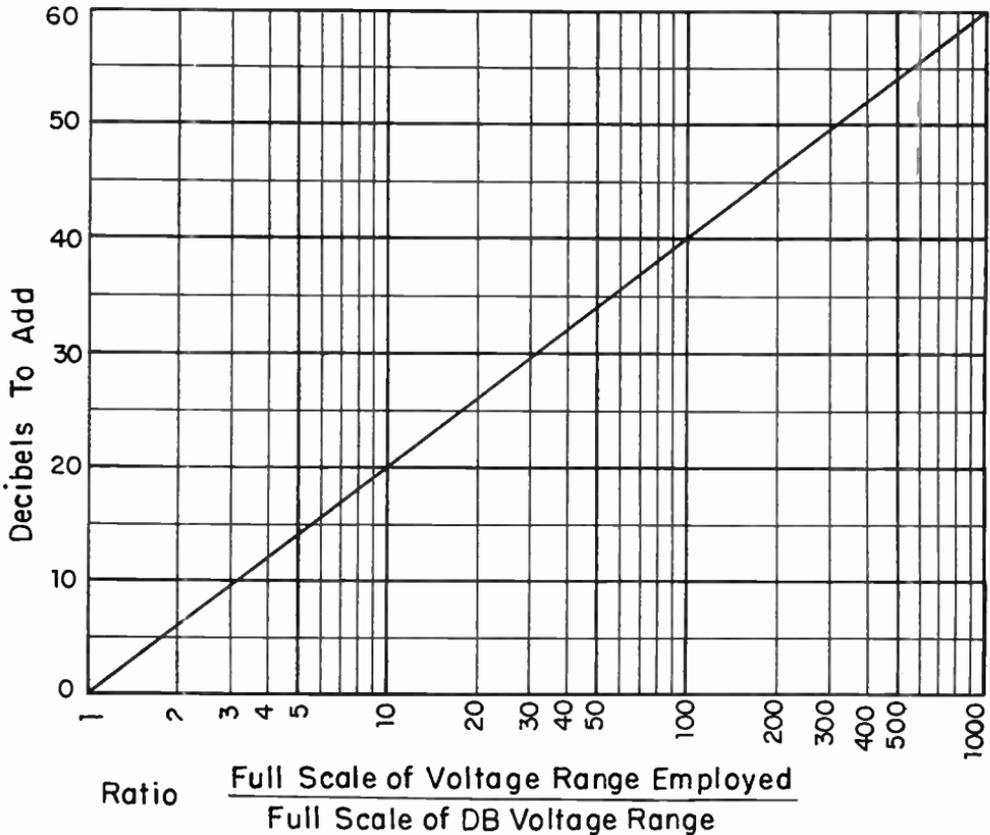


Fig. 3.—Corrections for decibel readings when a meter is across an impedance other than that for which calibrated.

A given number of decibels will express a definite power in watts provided some certain amount of power is assumed for zero db, this assumed power being called the reference level. In radio and television it is usual to assume a reference level of one milliwatt in an impedance of 600 ohms. An earlier reference level was six milliwatts in 500 ohms. The two reference levels are not equivalent.

Power gains or losses in decibels usually are measured on a db scale added to an a-c voltmeter, as in Fig. 2. Zero on the db scale is aligned with 0.775 on the a-c volts scale, because a reference power of one milliwatt in 600 ohms impedance is accompanied by a drop of 0.775 volt across the impedance. Various numbers of power decibels, both gains and losses are aligned on the scales with voltages appearing across 600 ohms of impedance.

Should a db meter calibrated for 600-ohm "load" be connected across more than 600 ohms, true power ratios in deci-



*Fig. 4.—Additions for decibel readings when a meter is used on a range other than that for which the db scale is calibrated.*

belts will be less than indicated, while with connection across less than 600 ohms true decibels are less than indicated. Corrections may be determined from Fig. 3. As may be seen, with the meter across 300 ohms, 3 db must be added to all readings, or with the meter across 4000 ohms it is necessary to subtract approximately 8 db.

A db scale reads true decibels directly only with the meter on a certain voltage range, usually the range closest to the db scale on the dial. For higher ranges certain numbers of decibels must be added to readings. Additions are shown by Fig. 4. First divide full-scale volts of the range actually used by full-scale volts for the range on which db readings would be direct,

thus finding a ratio listed on the chart. As an example, assume a db scale graduated for a 2.5-volt a-c scale, and a full-scale reading of 100 volts for the range actually used. The ratio of 100 to 2.5 is 40. For this ratio the chart shows that 32 db should be added to all indications.

**DECOUPLING.** — When alternating plate currents, screen currents, or grid currents of any two or more amplifying stages flow in the same resistance those stages are resistance coupled one to another. If such currents flow in a common inductance there is inductive coupling, and if they flow in a common capacitance there is capacitive coupling, provided the circuits may operate at the same frequency. In any of these cases the alternating voltages produced across the common resistance or reactance are applied to all the stages. Intentional couplings allow signal voltages to proceed in a forward direction from one amplifying stage to the next. Unintentional couplings allow voltages to feed back in a reverse direction, from the output end back toward the input end of the amplifying system.

Back couplings which are most troublesome are due to resistance or impedance of a plate and screen power supply connected to two or more amplifying stages. In Fig. 23-1 the plates and screens of both tubes are connected to a common power supply whose impedance causes back coupling. The back coupling may be lessened to some extent by connecting bypass capacitors  $C_s$  and  $C_s$  as shown. If the capacitive reactance of these bypasses is small in comparison with the impedance of the power supply, then most of the alternating signal will flow back to the tube cathodes through the bypasses and ground and will not flow in the power supply.

Back coupling through the power supply may be further lessened by inserting decoupling resistors  $R_d$  and  $R_d$  of Fig. 23-2. These resistances, in effect, add to the resistance or impedance of the power supply and make more opposition to signal currents flowing through the power supply. This forces more of the signal currents through the bypass capacitors to the tube cathodes.

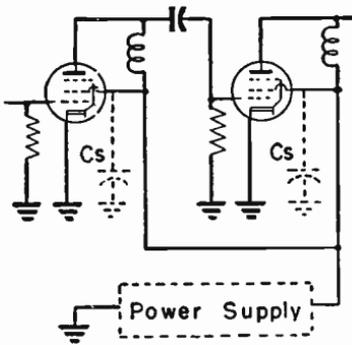


Fig. 23-1.—Feedback coupling through impedance of a common power supply.

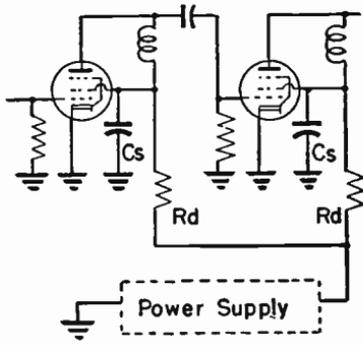


Fig. 23-2.—Partial decoupling by resistors in the plate-screen leads.

Fairly complete decoupling of one stage from all other stages is illustrated by Fig. 23-3. Plate circuit signal currents which would pass through the power supply now are opposed by decoupling resistor  $R_d$ , and are bypassed through ground to the tube cathode through decoupling capacitor  $C_d$ . Variations of signal voltage appear on the screen as well as on the plate of the amplifier. Accompanying variations of screen current are opposed in their passage to the power supply by resistors  $R_s$  and  $R_d$ , and are bypassed through decoupling capacitor  $C_s$  and ground to the tube cathode. Variations of screen current are prevented from affecting the plate circuit by the resistance of  $R_s$  and the bypassing effect of  $C_s$ .

The left-hand tube of Fig. 23-3 is shown as having a fixed bias or a bias provided by an automatic gain control system. A-c signal voltages in the grid circuit are opposed by decoupling resistor  $R_b$  in their effect on the common bias sup-

ply, and are bypassed to the tube cathode through decoupling capacitor  $C_b$  and ground. The right-hand tube is shown as having cathode bias. The bypass capacitor across a cathode bias resistor is not considered as a decoupling capacitor, its function is chiefly that of controlling degeneration and gain.

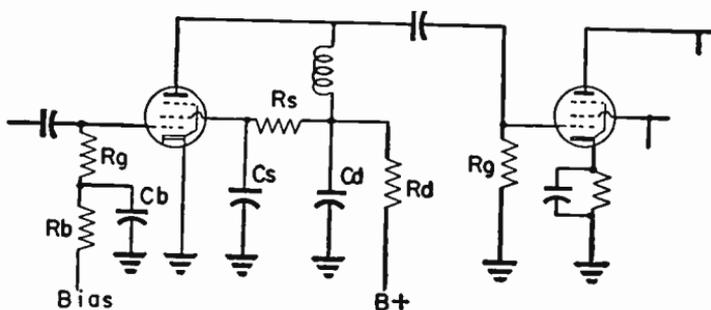


Fig. 23-3.—Separate decoupling for plate, screen, and grid circuits.

If the phase or instantaneous polarity of feedback voltage to the grid of an amplifier is opposite or nearly opposite to that of signal voltage for forward amplification the feedback is degenerative. This opposes forward signal voltage and reduces amplification. If phase or polarity of feedback is the same or nearly the same as that of forward signal voltage the feedback is regenerative. This aids the forward signal voltage and tends to increase amplification. Whether a feedback is degenerative or regenerative depends on the point from which this voltage originates in circuits following the affected grid circuit, and on whether or not there is phase shift between the origin and the amplifier grid input. Uncontrolled regenerative feedback may merely increase amplification, but usually it causes excessive peaking at some frequencies or outright oscillation. The purpose of decoupling ordinarily is to prevent uncontrolled regenerative feedbacks.

Decoupling capacitors and resistors are used in r-f amplifiers, in video i-f amplifiers, in video amplifiers, and also somewhat less completely in sound i-f amplifiers and in audio-frequency amplifiers. The subject of decoupling is closely related to that of shielding, since shielding lessens feedbacks

which are due to magnetic and electrostatic fields. Decoupling is assisted also by correct grounding, by using the shortest possible connecting leads, and by correct dressing of wires and parts.

**Decoupling Capacitance and Resistance.** — The effectiveness of decoupling is the ratio of the bypassed resistance to the capacitive reactance of the decoupling capacitor. If this ratio is 10 to 1, as an example, ten times as much signal voltage will be bypassed back to the cathode as goes through the power supply. Ratios between 10 to 1 and 5 to 1 are generally used. Although the bypassed resistance usually is considered as only that of the decoupling resistor, it really is the sum of this resistance plus the resistance or impedance of the power supply and associated circuits.

To retain a given effectiveness, the capacitance of a decoupling capacitor is inversely proportional to the lowest frequency at which the effectiveness is to be maintained. That is, at  $1/5$  the frequency it becomes necessary to use 5 times the capacitance to maintain the original capacitive reactance. A given degree of decoupling at a frame frequency of 30 cycles would call for a million times as much capacitance as at a video intermediate frequency of 30 megacycles.

The greater the decoupling resistance can be made the smaller need be the decoupling capacitance. A high resistance allows using a proportionately high capacitive reactance while retaining a given decoupling ratio. High capacitive reactance is found in relatively small capacitances. While increase of decoupling resistance is advantageous so far as capacitance is concerned, it causes greater voltage drop, requires more power supply voltage, and wastes more power in heating.

Electrolytic capacitors are used for decoupling at maximum frequencies around 5,000 cycles. They have too much internal resistance, and usually too much inductance and inductive reactance for use at higher frequencies. Ordinary rolled-foil paper dielectric capacitors have too much inductance and inductive reactance for use at frequencies even as high as standard broadcast radio frequencies. Non-inductive paper

capacitors are used for bypassing at radio frequencies. For television carrier frequencies, video intermediate frequencies, and the higher video frequencies it is necessary to use mica and ceramic capacitors which have negligible resistance and inductance.

When circuits, such as those in video amplifiers, have to carry a wide band of frequencies the large capacitance for low-frequency decoupling may be provided with electrolytic or large-capacitance paper capacitors having in parallel a small-capacitance mica or ceramic unit. The mica or ceramic capacitor bypasses the high inductive reactance of the electrolytic or paper unit at high frequencies.

**Plate and Screen Decoupling.** — Too little capacitance in a plate decoupling capacitor ( $C_d$  of Fig. 23-3) may allow frequency distortion or peaking at certain frequencies even though it does not allow oscillation. If there is to be no degeneration in a stage having cathode bias, the plate decoupling capacitor should be connected directly to the tube cathode. If this capacitor is connected to ground the plate signal currents must pass from ground through the biasing resistor unless the resistor is provided with a large capacitance bypass capacitor. It should be noted that, with a plate decoupling capacitor large enough, the decoupling resistor is not a part of the load in the plate circuit. Signal currents are kept out of this resistor by the decoupling capacitor.

The screen requires effective decoupling because variations of screen voltage may have almost as much effect on electron flow as would variations of grid signal voltage. Insufficient screen decoupling allows distortion and generally erratic performance. The small screen current allows using high de-

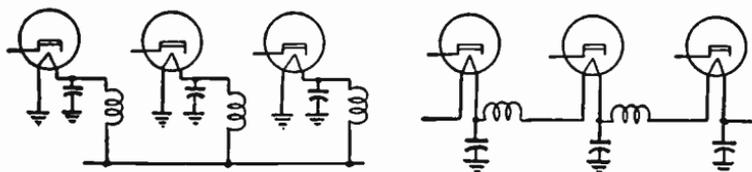


Fig. 23-4.—Decoupling chokes and capacitors for heater circuits.

coupling resistance without excessive voltage drop, and the high resistance permits fairly small capacitance for effective decoupling.

**Heater Decoupling.** — Back coupling may occur through leads for the cathode heaters in tubes for r-f and i-f amplifiers. Decoupling may be provided by r-f chokes and bypass capacitors to ground. A method used with heaters in parallel is shown at the left in Fig. 23-4, and a method for series heaters is shown at the right. The chokes may be made with 10 to 20 turns of enameled wire wound to a diameter of about  $\frac{1}{4}$  inch, closely spaced, and with air cores. The bypass capacitors may have each a capacitance of 100 mmf or more, often as large as 0.002 mf.

**Choke Decoupling.** — An inductor or choke coil might replace any or all decoupling resistors in preceding figures, to oppose alternating currents by inductive reactance. Advantages of a choke are negligible d-c voltage drop, better voltage regulation, and less power loss in heat. Disadvantages are greater cost and often greater bulk. Required reactance in ohms is the same as ohms of resistors.

**Tests For Decoupling.** — While operating at normal conditions a decoupling capacitor may be checked by connecting across it a high-impedance a-c voltmeter. The greater the voltage, the greater is opposition of the capacitor to voltages which should be freely bypassed, and the less is bypassing effect. Another method is to clip across the suspected capacitor an approximately similar unit in good condition. Marked improvement in performance indicates a defect in the original unit.

A decoupling resistor or choke may be checked with the high-impedance a-c voltmeter. The greater the voltage the poorer is the decoupling, since no alternating current should flow in the unit under ideal conditions. An oscilloscope may be used instead of a voltmeter. The scope usually has more impedance than an a-c meter, and in addition shows waveforms of voltages in decoupling units.

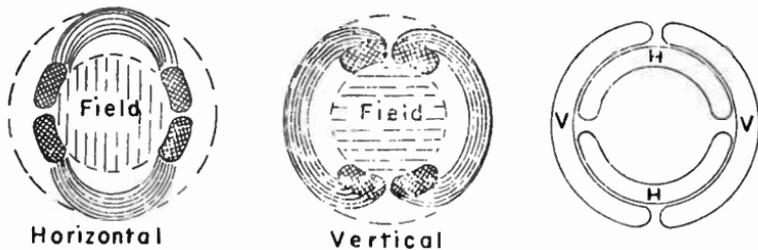


Fig. 25-1.—Arrangement of coils in a magnetic deflection yoke.

**DEFLECTION, MAGNETIC.** — The electron beam of a picture tube is deflected horizontally to form picture lines and vertically for fields by magnetic fields produced by currents of sawtooth waveform in coils of the deflecting yoke around the neck of the picture tube. An electron beam passing through a magnetic field is turned or deflected at a right angle to direction of field lines. Accordingly, as in Fig. 25-1, magnetic lines for horizontal deflection must be vertical, and lines for vertical deflection must be horizontal. Two coils for horizontal deflection and two for vertical deflection are placed in the yoke as at the right in Fig. 25-1, when looking length-

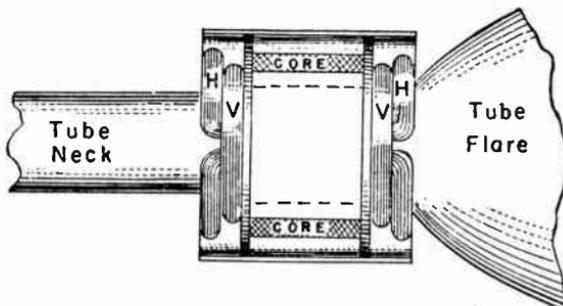


Fig. 25-2.—Positions of the yoke and deflecting coils on the neck of a picture tube.

wise, and as in Fig. 25-2 when looking toward one side of the tube.

Although voltage across the reflecting coils is of pulsed waveform, as at the left in Fig. 25-3, current in the coils is of sawtooth waveform, as at the right. The beam is deflected horizontally for picture lines and vertically for picture fields by the relatively slow increases of current. The sudden reversals cause retraces.

Greater face plate dimensions and shorter bells or flares in picture tubes call for greater total angles of beam deflection to cover viewing screens. Maximum angles in rectangular tubes up to 21-inch size are  $65^\circ$  horizontally and  $70^\circ$  diagonally, while in larger tubes the horizontal and diagonal deflection angles are respectively  $85^\circ$  and  $90^\circ$ .

**DEGENERATION.** — Degeneration occurs when there is fed back to the input or grid circuit of an amplifier a signal voltage which opposes the signal voltage being amplified. Degeneration requires a feedback which is in opposite or nearly opposite phase to the amplified signal. Degeneration is called also negative feedback or inverse feedback. A degenerative feedback reduces the gain of the amplifier. At the same time the amplifier is less likely to oscillate, the gain becomes more uniform at all frequencies, there is less harmonic distortion, less phase distortion, and less effect of tube noise. There is also less likelihood of trouble due to power line voltage fluctuations, to changes of load resistance, or to aging of amplifier tubes.

If a feedback voltage is in phase or nearly in phase with the amplified signal voltage the result is regeneration. This increases the gain because the input or grid signal voltage is increased, but there is more distortion and the amplifier may oscillate. If there is phase shift with variations of frequency in circuits furnishing the feedback, the result may be degeneration at some frequencies and regeneration at other frequencies.

There are two general methods of providing feedback. If feedback voltage results from changes of signal current in the amplifier output or in some following stage there is said to be current feedback. If feedback results from signal voltages anywhere in the output there is voltage feedback. Current feedback tends to maintain a stable output signal current, it acts like an increase of plate resistance in the amplifier tube when there is increase of feedback. Voltage feedback tends to maintain a stable output signal voltage, it acts like a decrease of plate resistance. Both kinds of feedback may be used in the same amplifier.

**Current Feedback.**—The most commonly employed method of obtaining current feedback is with a biasing resistor in the cathode circuit of the amplifier tube. Such a resistor is shown at  $R_k$  in Fig. 26-1. When the input signal goes positive the grid is made more positive, or less negative when there is a negative bias. There is a corresponding increase of plate current, a greater voltage drop in the plate load, and the voltage at the amplifier plate becomes more negative. The plate current now being considered is a variation of signal current. There must be the same change of current in the entire plate circuit, which includes not only the plate load but also the cathode resistor and capacitor  $C_d$  forming the plate return for the signal current. The increase of plate current in the cathode resistor causes increased electron flow toward the cathode, and the upper end of the resistor becomes more positive while the lower end becomes more negative. This

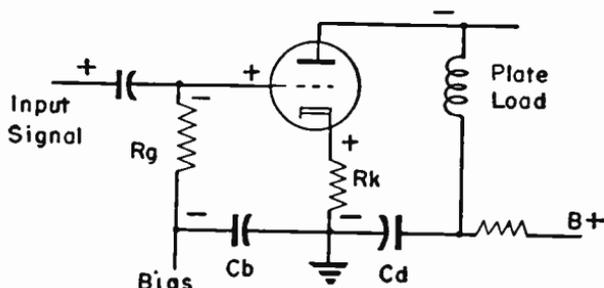


Fig. 26-1.—Current feedback obtained by means of a cathode resistor.

more negative signal voltage passes through capacitor  $C_b$  to the lower end of grid resistor  $R_g$ . Since, with negative bias, there is no current in  $R_g$  there is no change of potential in this resistor, and the increase of negative voltage is applied from the upper end of  $R_g$  to the grid. Thus a positive grid signal voltage causes a negative grid feedback voltage due to change of signal current in the cathode resistor.

The change of voltage actually applied to the tube grid is not the change of signal voltage, it is the signal voltage minus the opposing voltage developed across the cathode resistor and applied between grid and cathode. This negative feedback has the effect of flattening and straightening a curve showing the relation between grid voltage and plate current. Then harmonic distortion is reduced because such distortion results from curvature of this characteristic. Other advantages previously mentioned might similarly be traced to degenerative effects.

The greater the resistance of the cathode resistor  $R_k$  the greater will be the voltage drop or feedback voltage for a given variation of signal current, and the less will be the amplifier gain. If the cathode resistor is bypassed with a capacitor, part of the a-c signal current will act simply to charge and discharge the capacitor and less will remain across  $R_k$  to produce negative feedback. The greater the bypassing capacitance, and the less its capacitive reactance, the less will be the negative feedback. Capacitive reactance increases as frequency drops. Consequently, the bypassing will be less effective at low frequencies and there will be more degeneration at low frequencies than at higher ones.

Degeneration by means of a cathode resistor is commonly employed in video amplifiers. The degree of degeneration is determined by the resistance in series with the amplifier cathode and by the bypassing capacitance if any is used. Using a small bypass capacitance allows considerable degeneration at low video frequencies and automatically lessens degeneration and increases gain as frequency rises.

**Voltage Feedback.** — Fig. 26-2 shows one method of obtaining a degenerative voltage feedback. Between the plate of the right-hand amplifier and the grid input to the same tube are connected in series a capacitor  $C_f$  and a resistor  $R_f$ . Either the capacitor or the resistor often are used alone. Because change of signal voltage at the plate of an amplifier always is in opposite phase or opposite polarity to change of signal voltage at the grid of the same amplifier, the voltage fed back to the grid is degenerative. The degree of degeneration is fixed by the resistance of  $R_f$  or the capacitive reactance of  $C_f$  when only one of these elements is used. The greater the resistance or reactance the less is the degenerative voltage fed to the grid.

There is inversion of signal voltage in each stage of a cascade amplifier. Voltages are in phase, or approximately so, at the plate of a second amplifier and the grid of the first

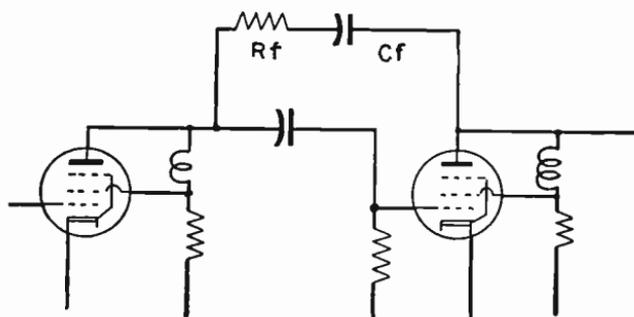


Fig. 26-2.—One method of obtaining voltage feedback.

amplifier in an adjacent pair. Were the feedback from the plate of the right-hand tube of Fig. 26-2 applied to the grid input circuit of the left hand tube the feedback voltage would be in phase with the grid voltage. The result would be regeneration instead of degeneration.

The phase or polarity of signal voltage at the cathode of an amplifier is the same as at the grid of the same tube. Therefore, a small capacitor connected between plate and cathode will apply a negative feedback to the grid.

There are numerous other methods of obtaining voltage feedback. A degenerative feedback may be had from any point after the output of an amplifier where the signal voltage is inverted or in opposite phase to the grid signal voltage of the amplifier. A regenerative feedback may be obtained from any point where the signal voltage is in phase with or of the same polarity as the grid signal voltage for the amplifier.

**DETECTOR, PHASE.** — A phase detector is a device which produces in its output a direct potential varying in accordance with changes of relative frequency in two voltages applied to the detector input. Fig. 27-1 shows circuit connections for a phase detector used in an automatic control for horizontal sweep frequency. The detector tube is a twin diode whose sections are marked *A* and *B*. Connected from the plate of *A* to the cathode of *B* are resistors *R<sub>a</sub>* and *R<sub>b</sub>*. To a point between these two resistors are applied the voltage pulses fed back from some point in the horizontal sweep output circuits. The frequency of these pulses is the same as the actual frequency of the horizontal sweep currents in the picture tube deflection coils. These sweep pulses appear without change of polarity at the upper end of resistor *R<sub>a</sub>* and at the lower end of resistor *R<sub>b</sub>*.

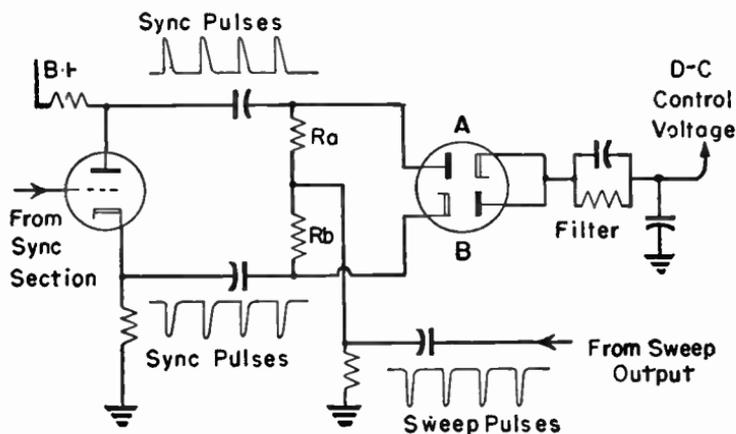


Fig. 27-1.—Connections in one type of phase detector circuit.

To the outer ends of the two resistors are applied voltage pulses coming from the sync section of the receiver. The frequency of these pulses is the same as that of the horizontal sync frequency in the received signal. The inverter tube, at the left, delivers the sync pulses in opposite polarity at the two resistors. The results of applying two voltages of the same or different frequencies to diodes connected as in Fig. 27-1 is explained at some length in the article on *Detector, Ratio*.

So long as the frequencies of sweep pulses and sync pulses are the same there is equal conduction in the two diodes, and zero potential with reference to ground at the detector output on the right. If the actual sweep pulse frequency should change with reference to the horizontal sync pulse frequency there will be unequal conduction in the diodes. Then there will appear at the output a potential which is either positive or negative with reference to ground, depending on whether sweep frequency increases or decreases. This output potential may be applied in various ways to the horizontal sweep oscillator to bring oscillator frequency and sweep frequency back into synchronization with the sync pulses. Any deviation of sweep frequency results in a change of output potential and amplitude with this phase detector, just as does deviation of signal frequency with the ratio detector for sound.

The filter of Fig. 27-1 prevents sudden changes of pulse amplitude from affecting the output potential. Such changes might result from electrical interference or other disturbances coming through the receiver to the sync section.

**DETECTOR, RATIO.** — A ratio detector is a device for obtaining from a frequency-modulated signal an audio voltage corresponding in frequency and amplitude to the original modulation. In a frequency-modulated signal the frequency varies or deviates above and below an average value which is the intermediate frequency of the receiver sound system. The number of times per second that the signal frequency deviates above and below the intermediate value is the same as the number of cycles per second of the audio-frequency modulation. How far each deviation goes above and below the inter-

mediate frequency corresponds to the amplitude or strength of the audio-frequency modulation.

In ratio detector circuits there are numerous modifications which do not alter the basic operating principles. These principles will be explained with reference to the circuit of Fig.

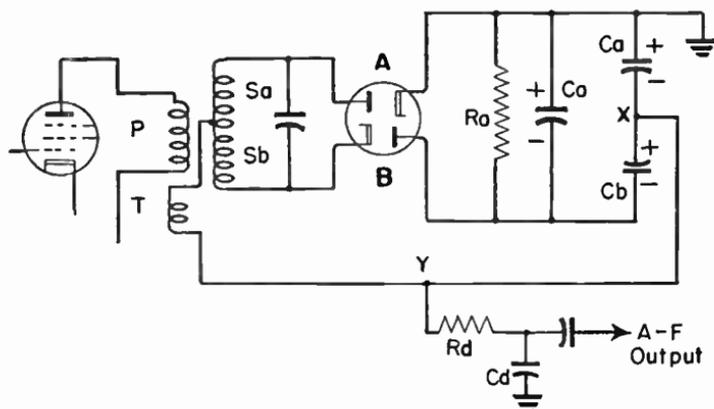


Fig. 28-1.—Typical circuit for a ratio detector used in f-m sound systems.

28-1. The detector consists of two diodes and a three-winding transformer which act together in obtaining from the f-m signal an audio voltage whose frequency is the same as the rate of deviation and whose amplitude varies as the extent of deviation.

The primary winding  $P$  of the detector transformer is in the plate circuit of the last sound i-f amplifier tube. The primary is coupled by mutual induction to a center-tapped secondary winding:  $Sa$  and  $Sb$ , whose outer ends connect to the plate of diode  $A$  and to the cathode of diode  $B$ . Primary and secondary windings have adjustable cores by means of which both windings are tuned to resonance at the sound intermediate frequency or center frequency. This is a service adjustment.

The third transformer winding  $T$  is connected from the center tap of the secondary to a point between two small capacitors,  $Ca$  and  $Cb$ , which are in series with each other between the cathode of diode  $A$  and the plate of diode  $B$ . This third winding is inductively coupled to the primary winding,

often being wound over the low end of the primary. Also connected between the cathode of diode *A* and the plate of diode *B* are the load resistor *R<sub>o</sub>* and a large capacitor *C<sub>o</sub>*.

In the third winding of the transformer is induced a voltage whose phase relation to the primary signal voltage is not altered by deviations of frequency. Because this fixed-phase voltage is put into the center tap of the secondary winding the same voltage appears at both ends of the secondary without change of either phase or polarity. However, the voltages which are induced in the secondary by means of its mutual inductive coupling to the primary will vary in their phase relations to the primary voltage when there is frequency deviation.

Changes of phase, and the results, are shown by Fig. 28-2. At the top are conditions when there is no frequency deviation. Voltage from the third winding of the transformer is at both ends of the secondary, being indicated by solid-line curves marked *T*. The phase is the same as in the primary. With no deviation the applied signal frequency is the resonant frequency to which the windings are tuned. Reactances balance at resonance, and only resistance remains to affect the secondary. Voltages induced in the secondary by inductive coupling to the primary are one-quarter cycle out of phase with voltage *T*. These induced voltages are shown by broken-line curves marked *A* and *B*. Their polarities are opposite at the two ends of the secondary, as occurs at the two ends of any transformer winding in which voltage is induced. The two pairs of voltages, *T-A* and *T-B*, at the top and bottom of the secondary combine as shown over at the right. These combination voltages are applied to the plate of diode *A* and to the cathode of diode *B*.

Diode *A* will conduct while its plate is positive. The brief conduction current will flow through the top half of the secondary and through winding *T*, will charge capacitor *C<sub>a</sub>* in the polarity marked in Fig. 28-1, and go to the cathode of diode *A*. Diode *B* will conduct while its cathode is negative. This conduction current will charge capacitor *C<sub>b</sub>* in the

polarity marked. The two diodes voltages are of equal amplitude, conduction currents will be equal, and the two small capacitors will be equally charged. Since capacitors  $C_a$  and  $C_b$

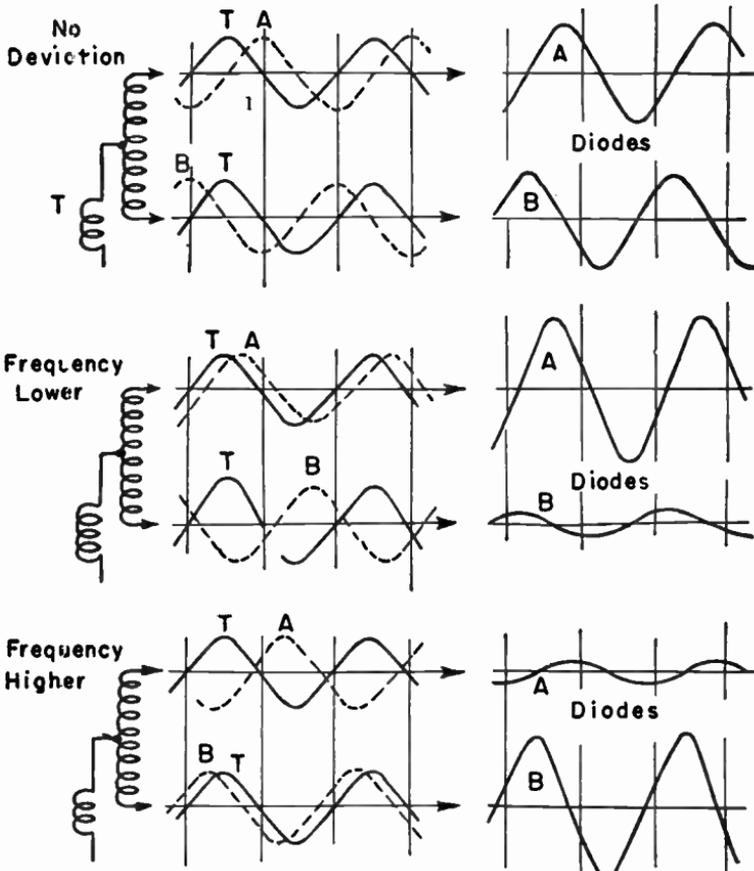


Fig. 28-2.—Changes of voltage phase in a ratio detector, and their results on diode voltage and current.

are across capacitor  $C_o$  this latter capacitor will be charged to a potential equal to the sum of the potentials across  $C_a$  and  $C_b$ . These charges gradually leak off through resistor  $R_o$ , but are continually renewed by the diode conduction currents.

The middle group of curves in Fig. 28-2 shows what happens when there is signal deviation to a lower frequency. When there is applied to a tuned circuit (the secondary wind-

ing) a frequency lower than that for which the circuit is resonant that circuit presents an excess of capacitive reactance, because capacitive reactance (in the secondary) increases while inductive reactance drops when frequency is lowered. Any capacitive reactance tends to make an induced a-c voltage lead the inducing (primary) voltage. This leading tendency has the effect of lessening the original lag of voltage  $A$  behind voltage  $T$  at the top of Fig. 28-2, and these voltages at the top of the secondary come more nearly into phase. The combination of these nearly phased voltages results in a much stronger voltage for diode  $A$ , as shown over at the right.

Since voltages at top and bottom of the secondary must remain in phase and of opposite polarity, we have at the bottom of the secondary an induced voltage  $B$  whose phase is nearly opposite to that of the fixed voltage  $T$ . Here the two voltages almost cancel and, as shown at the right, the combined voltage applied to diode  $B$  becomes very weak. Thus it appears that, when deviation is to a lower frequency, conduction in diode  $A$  will be greater than in diode  $B$ . Then capacitor  $C_a$  will receive more charge and will have greater voltage than capacitor  $C_b$ .

Conditions with deviation to a higher frequency are shown by the bottom group of curves in Fig. 28-2. Now the secondary winding offers an excess of inductive reactance, because inductive reactance increases while capacitive reactance drops when frequency goes up. Inductive reactance tends to make an induced a-c voltage lag the inducing voltage. This lagging tendency increases the original lag of voltage  $A$  behind voltage  $T$  in the top group of curves.

Now the two voltages at the top of the secondary are of nearly opposite phase, they almost cancel each other, and the combined voltage applied to diode  $A$  becomes very weak, as shown at the right. At the same time the combined voltage applied to diode  $B$  becomes much stronger. Conduction in diode  $B$  will be greater than in diode  $A$  when deviation is to a higher frequency, and capacitor  $C_b$  will receive more charge and will have a higher voltage than capacitor  $C_a$ .

It has been shown that deviation above and below the intermediate frequency causes changes of voltage on capacitors  $C_a$  and  $C_b$ . The number of deviations per second equals the audio frequency in cycles per second of the sound signal. Therefore, the changes of capacitor voltages will be at this audio frequency.

The greater the deviation or the greater the changes of frequency, the greater will be the differences between the phase of the fixed voltage  $T$  in Fig. 28-2 and the induced secondary voltages  $A$  and  $B$ . The greater the phase shift the stronger will be the combined voltages at one diode and the weaker they will be at the other diode. There will be corresponding differences in conduction currents and in charges and voltages on capacitors  $C_a$  and  $C_b$ . That is, greater deviation results in greater capacitor voltages, and since the extent of deviation represents strength of audio modulation the changes of capacitor voltages will follow the audio strength of signal modulation.

Capacitor voltages thus are found to vary at a rate corresponding to audio frequency, and to a degree corresponding to audio strength or "volume." This audio frequency voltage may be taken from across either of the capacitors  $C_a$  or  $C_b$  by suitable circuit arrangements. In Fig. 28-1 the audio frequency is taken from across capacitor  $C_a$ . The upper end of this capacitor is directly connected to ground. The lower end is connected to ground through points  $X$  and  $Y$ , and through resistor  $R_d$  and capacitor  $C_d$ . These latter two units form the de-emphasis filter required in all f-m audio amplifiers. The audio output is taken from the filter.

While a ratio detector translates frequency modulation into audio-frequency output, changes of input signal amplitude do not produce audio-frequency voltages in the output. This is due to the action of capacitor  $C_o$  and resistor  $R_o$  of Fig. 28-1. The capacitance of  $C_o$  is large, usually between 2 and 10 microfarads. The resistance of  $R_o$  is such that the discharge time constant of the combination is much longer than the period of the lowest audible frequency. Consequently, the

charge and the voltage of capacitor  $C_o$  cannot change at an audio rate.

Increases of voltage across capacitor  $C_a$  are matched by decreases of voltage across  $C_b$ , and vice versa. Although each of these separate voltages varies at audio frequency, their sum must remain practically constant because the sum always must equal the nearly constant voltage across  $C_o$ . Any change of signal amplitude (not frequency) tends to cause an increase or decrease of voltage across both small capacitors at the same time. This would change their sum voltage. But the sum voltage cannot change at an audible rate because it must remain equal to the voltage across the large capacitor  $C_o$ , and this voltage will not vary at an audible frequency.

**DETECTOR, VIDEO.** — The video detector recovers from the amplitude-modulated video intermediate frequency all the picture signal variations and all the sync pulses together with blanking level intervals of the television signal. This detector,

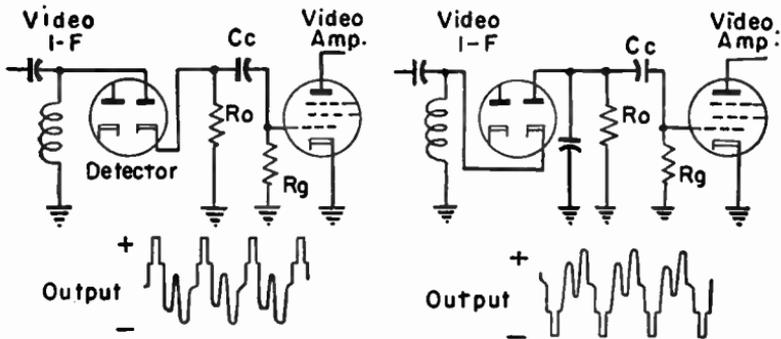


Fig. 29-1.—Video signal polarities at the outputs of diode video detectors.

as shown by Fig. 29-1, is connected between the last video i-f amplifier and the first video amplifier. In the majority of receivers the video detector is one section of a twin triode, whose other section may be used for automatic gain control, for d-c restoration, or any other purpose.

The i-f signal input may be to the plate of the detector with output from the cathode as at the left in Fig. 29-1, or the input may be to the cathode with output from the plate as at the right. With input to the detector plate, picture signals are negative and sync pulses are positive in the output. With input to detector cathode, picture signals are positive and sync pulses are negative in the output.

The detector must handle the same wide range of frequencies as the video amplifier, consequently is subject to the same difficulties due to wide band coverage as are encountered in the video amplifier. The detector load resistor,  $R_o$  in Fig. 29-1, is of only 2,000 to 5,000 ohms resistance in order to lessen the effects of shunting capacitances at high frequencies. Coupling capacitor  $C_c$  is of 0.05 to 0.10 mf capacitance in order to keep its capacitive reactance reasonably small at the lowest video frequencies.

The small load resistance in the detector output and the rather small internal resistance of a diode have the effect of

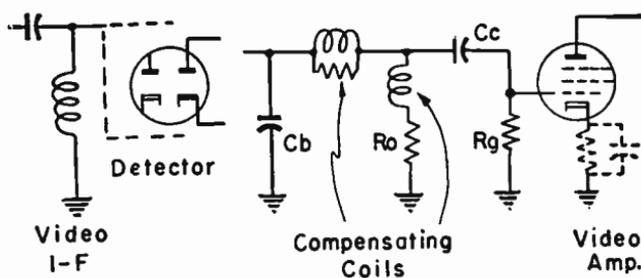


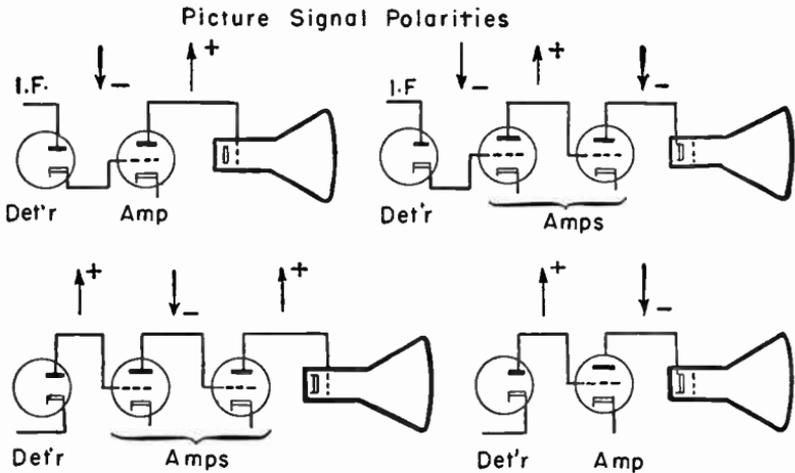
Fig. 29-2.—Features found in couplings between video detector and first video amplifier.

heavily loading the video i-f stage which precedes the detector. This broadens the frequency response while reducing the gain of that stage.

Fig. 29-2 illustrates some features which may be found in circuits between the video detector and video amplifier. The

series and shunt compensating coils are like those used in video amplifier couplings, and serve the same purpose of extending the response to higher frequencies. The small capacitor at  $C_b$ , usually 5 to 10 mmf, bypasses the high video intermediate frequencies to ground and back to the i-f coupler so that these frequencies do not go on into the video amplifier. This capacitor, in connection with the series compensating coil, form a low-pass filter which passes all frequencies up to the video limit of about 4.5 mc, while attenuating the much higher video intermediate frequencies.

**Detector Output Polarity.** — When video i-f input is to the detector plate, as at the left in Fig. 29-1, the positive side of the i-f carrier is rectified and in the detector output the sync pulses are positive and the picture signals are negative. Signal



**Fig. 29-3.**—Picture signal polarities with i-f input to detector plate or cathode, and with picture tube input to grid or cathode.

polarity will be inverted in passing through each video amplifier between detector and picture tube. Consequently, if the signal is to reach the picture tube with the picture signals positive there must be one inversion and one video amplifier

stage, or there might be any odd number. If the signal is to reach the picture tube with picture signals negative there must be two inversions and two video amplifier stages, or any even number. The signal must reach the picture tube with picture signals positive when input is to the grid of the picture tube, and the picture signals must be negative when input is to the cathode of the picture tube. These polarity relations are shown at the top of Fig. 29-3.

At the bottom of Fig. 29-3 are shown the stage polarities when video i-f input is to the cathode of the detector. Two, or some even number of video amplifiers are needed to make the picture signals positive at the picture tube grid. One or any odd number of stages are needed to make the picture signals negative at the picture tube cathode.

The same general rules for signal polarity apply when considering the sync pulses which go through the sync section of the receiver and are used to trigger the sweep oscillators. The relations are clearly shown by Fig. 29-3 if we assume the sweep oscillator sync inputs to be the same as the picture tube inputs illustrated, and assume the amplifiers to be tubes in the sync section. Most sweep oscillators require a positive triggering pulse, although some use negative pulses. It must be kept in mind that there is polarity inversion in any tube where input is to a grid and output from a plate, but there is no inversion in cathode followers where input is to the grid and output from the cathode.

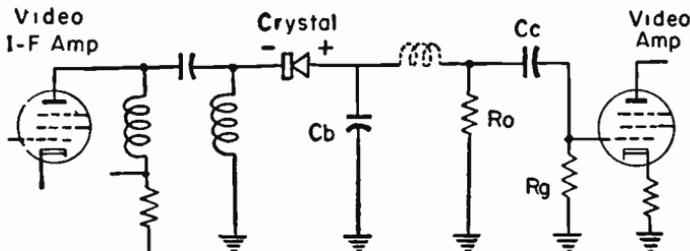


Fig. 29-4.—A crystal diode used as a video detector.

**Crystal Detectors.** — A number of television receivers have germanium crystal diodes for their video detectors. These diodes are described in the article on *Crystals, Rectifying*. A fairly typical crystal detector circuit is shown by Fig. 29-4. The negative end of the crystal is equivalent to the cathode of a tube diode, and the positive end is equivalent to the plate. Whether the i-f input is applied to the negative or positive end of the crystal determines the polarity of the output signal just as with tube diodes. Values of load resistance,  $R_o$ , and of coupling and bypass capacitors are like those for tube diode detectors. Series and shunt compensating coils may or may not be used, just as with tube detectors.

Care must be exercised in making service tests on crystal detectors, since strong a-c voltages from a signal generator may overload the crystal because of its low resistance to forward current. Neither end of the crystal should be connected to ground while making tests.

**DIELECTRICS.** — The accompanying table lists dielectric constants and power factors for materials commonly used in television capacitors and other circuit parts. The power factor is a measure of energy loss that occurs in the dielectric material. It is the fraction of the energy which would be dissipated as heat were all the current and applied voltage used to produce heating power, with this power in watts equal to  $EI$ . Were there no energy losses in a capacitor the current would lead the voltage by 90 degrees, and the power factor would be zero. When there are losses the current leads by less than 90 degrees and the power factor is more than zero.

High-frequency energy loss in a capacitor is assumed to be proportional to a loss factor which is equal to the product of dielectric constant and power factor. To provide a small loss factor both the dielectric constant and the power factor must be relatively small. This explains some of the advantage of using polystyrene, polyethylene, and some of the high grade dielectric waxes.

## PROPERTIES OF DIELECTRIC MATERIALS

Kind of Material	Dielectric Constant	Power Factor
Mica	5.4 to 8.0	0.2
Paper, plain	2.0 to 2.6	
Paper, wax impregnated	3.5	
Phenolic compounds, mica filled	5.0 to 6.0	0.8 to 1.2
Phenolic compounds, low loss	5.3	0.4 to 0.7
Polyethylene	2.3 to 2.4	0.02 to 0.05
Polystyrene	2.4 to 2.8	0.02
Quartz	4.7 to 5.1	0.02 to 0.03
Steatite	4.8 to 6.5	0.4
Steatite low loss	4.4	0.2
Titanium dioxide	90 to 170	0.1
Waxes	1.9 to 3.2	0.1 to 0.3

**DISCRIMINATORS.**— A discriminator is a device which, when supplied with voltages of varying frequency but constant amplitude, delivers from its output a voltage of varying polarity and varying amplitude. The polarity of the output voltage depends on whether the input frequency varies above or below an average frequency or center frequency. Amplitude of the output voltage depends on the extent of frequency shift at the input, or on how far the input frequency deviates from the average or center frequency.

Discriminators are used for various purposes in radio and television. One of the earliest uses was in automatic frequency controls or automatic tuning controls for radio receivers. The most widespread use is for sound detectors in f-m radio receivers and in the sound systems of television receivers, which operate from frequency-modulated sound signals. Here the discriminator recovers audio-frequency voltages from the frequency-modulated input signals. In many television receivers a discriminator is used as part of the automatic frequency control for sweep oscillators. Oscillator frequency is compared with sync pulse frequency in the discriminator, and a correction voltage is supplied for keeping the oscillator synchronized.

Fig. 31-1 is a typical circuit diagram for a discriminator used as an f-m sound detector in a television receiver. The discriminator consists of a special type of transformer and two diodes. The two diodes, marked *A* and *B*, may be the two sections of a twin diode tube, they may be matched germanium crystal diodes, or they may be parts of a tube having other sections serving as a-m diode detector, amplifier, inverter, or some other purpose.

In the transformer are a primary winding *P* and a center-tapped secondary winding *Sa-Sb*. The two windings are coupled by mutual induction. Both windings are adjustably tuned to resonance at the sound intermediate frequency or center frequency. These are service adjustments. To the

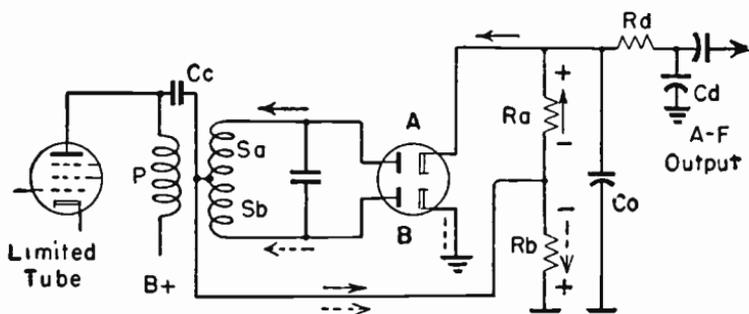


Fig. 31-1.—A typical circuit for a sound discriminator.

center tap of the secondary is applied primary voltage taken through capacitor *Cc* from the plate end of the primary winding. The outer ends of the secondary winding are connected to the plates of the two diodes. The cathode of one diode is grounded directly. The other cathode is connected to ground through two load resistors *Ra* and *Rb*. The center tap of the secondary winding is connected to a point between the load resistors. Across the load resistors is load capacitor *Co* which charges when there are unbalanced conduction currents in the two diodes and which discharges through the load resistors. Each of the resistances *Ra* and *Rb* may be of 100,000 ohms, with capacitance at *Co* of about 250 mmf. Audio output is

delivered from the top of load resistor  $R_a$  through a de-emphasis filter consisting of resistor  $R_d$  and capacitor  $C_d$ . Such a filter is required in all f-m sound systems.

A discriminator demodulates a frequency-modulated signal because, with deviation of input frequency, there is phase shift between the primary voltage and the voltage induced in the secondary winding by its inductive coupling to the primary. The phase shifting principle and its results on voltages applied to the two diodes are explained in the article on *Detector, Ratio*. In the discriminator circuit the voltage of fixed phase is that coming from the primary through capacitor  $C_c$  to the center tap of the secondary. This voltage appears without change at both top and bottom of the secondary.

Frequency deviation in one direction increases the voltage on diode  $A$  and at the same time decreases the voltage on diode  $B$ . Opposite deviation increases the voltage on diode  $B$  and decreases it on diode  $A$ . Electron flow in the circuit for diode  $A$  is indicated in Fig. 31-1 by full-line arrows, and in the circuit for diode  $B$  by broken-line arrows. Note that polarities are opposite in the two load resistors.

With deviation of frequency in one direction, the increase of voltage on diode  $A$  increases its conduction current and increases the voltage drop across resistor  $R_a$ . At the same time there is decrease of voltage on diode  $B$ , a decrease of its conduction current, and a decrease of voltage drop across resistor  $R_b$ . Then the overall voltage across both resistors, from the top of  $R_a$  to ground, becomes more positive at the top of  $R_a$  from where the audio output is taken.

When there is opposite deviation there is more voltage and more conduction current for diode  $B$  than for diode  $A$ . Consequently, there is more voltage drop across resistor  $R_b$  than across  $R_a$ . Then the overall voltage becomes more negative at the top of  $R_a$ , the audio output point.

Deviations of input frequency above and below the intermediate frequency have been shown to cause an alternating voltage at the audio output. This voltage is alternating because it goes first positive and then negative. The number of

alternations per second of this output voltage will be the same as the number of frequency deviations per second above and below the intermediate frequency. The number of deviations per second is the audio frequency of original modulation. The amplitude of the output voltage will be proportional to the extent of frequency deviation, because the greater the deviation the greater the phase shift in the transformer and the greater the difference between diode currents and voltage drops across the load resistors. This extent of frequency deviation corresponds to audio strength or "volume" of the original frequency modulation. There are many modifications of details in discriminator circuits but the basic operating principles remain as described.

While a discriminator for sound will produce output audio frequencies from input frequency modulation, it also will produce changes in the audio output whenever there is variation of input amplitude at the same time as frequency deviation. During any instant in which there is deviation of frequency there is phase shift of voltages at top and bottom of the secondary winding. Then any change of input amplitude will cause still greater difference between diode voltages, conduction currents, and voltage drops across the load resistors. This, of course, will affect the amplitude of the a-f output voltage.

Changes of amplitude in the received f-m sound signal must be removed by a limiter stage which precedes the discriminator. The limiter tube is operated with such voltages on plate, screen, and grid that alternations of signal voltage in the plate circuit cannot exceed a certain amplitude no matter how great the signal may become on the grid. So long as f-m signal voltages at the limiter input are strong enough to bring the output up to the limit of amplitude, the output must remain of constant amplitude. This frequency-modulated signal of constant amplitude is fed to the discriminator.

**DIVIDERS, VOLTAGE.**—A resistance voltage divider is shown at the left in Fig. 32-1. Voltage from *A* to *D* is the same as that of the source. If no current is being taken through taps *B* and *C* the voltages at these points will be proportional to the

resistances. When currents are drawn from the taps there is increased voltage drop in the resistances through which these currents flow, and tap voltages then depend on  $IR$  voltage drops rather than on resistance alone. Resistance voltage dividers have poor voltage regulation unless tap currents are very small, or voltage regulators are used between the taps and one side of the source, or a large bleeder current is allowed to flow from the source through the resistor string.

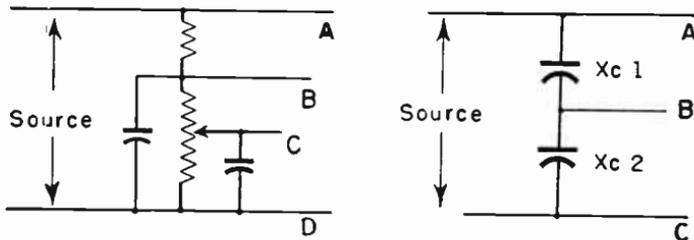


Fig. 32-1.—A resistance voltage divider (left) and a capacitance voltage divider (right).

A capacitance voltage divider is shown at the right in Fig. 32-1. Such a divider is used only with alternating voltages. Voltage at tap  $B$  depends on the relative capacitive reactances of capacitors above and below the tap. In making computations these reactances, in ohms, megohms, etc., are treated just as are resistances across the source with a resistance voltage divider. Voltage at a tap, with reference to one end of the divider, is proportional to the ratio of reactance between this tap and the end considered, to the total reactance in the entire divider.

For example, the voltage at tap  $B$  with reference to point  $A$  would be a fraction of the source voltage found by dividing reactance  $Xc1$  by the sum of  $Xc1$  plus  $Xc2$ . The voltage fraction with reference to point  $B$  would be found by dividing reactance  $Xc2$  by the sum of  $Xc1$  plus  $Xc2$ . Increasing a capacitance lessens the capacitive reactance and drops the voltage across the capacitance which is changed. Decreasing a capacitance raises the capacitive reactance and increases the voltage across that capacitance.

**DRESSING, WIRES AND PARTS.** — In all high-frequency circuits and in all wide-band amplifier circuits the layout of parts and the positions of connections are designed for small and constant stray capacitances and inductances. The higher the frequency or the greater the stage gain the more trouble may result from misplacement of conductors or circuit elements. All parts, including inductors, capacitors, and resistors, as well as wires and cables, should be replaced only with exact duplicates mounted in the same relative positions as the originals. When wires or cables are replaced the new ones should be of the same length, same gage size, should have the same kind of insulation as the originals, and should go in the same positions. Change of length will alter the inductance, while change of gage size or position will alter the circuit capacitances and may upset the tuning.

Grid leads and plate leads should be kept away from each other because of possible capacitive coupling and feedback. Grid leads should be kept clear of all other wiring. Plate leads often are dressed down close to chassis metal. These precautions apply also to plate and cathode leads for detector diodes. Screen leads should be dressed close to chassis metal. Grid returns and plate returns from transformers or other coupling units should be kept separated from each other until reaching the decoupling capacitors and resistors. Leads to and from r-f oscillator coils must be kept away from all other wiring and from each other. This applies to oscillators in the tuner and also in any r-f power supply. In general, all grid and plate leads should run as directly as possible between their terminals and should be kept as short as possible.

High-voltage wires and cables should be kept away from all low-voltage parts except where passing through suitable insulators or grommets. Leads to the high-voltage anode of outside-coated picture tubes should clear the coating.

Wires for the a-c power line, tube heaters, pilot lamps, and similar low-voltage a-c parts are either kept close to chassis metal or else are run well separated from all parts carrying signal voltages. These a-c leads must be kept well away from

audio detectors and amplifiers, and their plate and grid connections.

When wires remain longer than necessary between terminals, as may be the case with some transformer and choke connections, they should be kept close to chassis metal and dressed back toward the transformer or other part to which they are permanently connected.

Antenna connections or transmission line connections should be kept clear of metal parts, chassis and shielding, from their external terminals to where they enter the tuner.

Coupling capacitors should be kept clear of chassis metal and should be mounted as close as possible to the socket of the tube to whose grid the capacitors connect. Any extra length in the connections should be on the plate side of the capacitor. Changing the position of coupling capacitors may upset the alignment.

Peaking coils in wide band video amplifier and video detector circuits should be kept away from chassis metal and, so far as possible, away from other wiring. R-f chokes used for decoupling are mounted close to chassis metal unless the chokes have their own grounded shields.

It is especially desirable to keep all parts of oscillator circuits as far as possible from other parts and wiring, both in tuner circuits and sweep circuits.

Any added parts which are mounted close to other parts carrying radio or intermediate frequencies are likely to cause trouble because of altered capacitances in the original circuits.

Tuning capacitors, both fixed and adjustable, should be mounted as close as possible to the grid lugs or grid pins for the tube whose grid circuit is tuned.

The performance of high-gain audio amplifying tubes may be affected by strong magnetic fields from permanent magnets or field magnets of speakers, or from power transformers and filter chokes. New speakers or power supply parts should be mounted to avoid such possibilities.

**DRIVE CONTROLS.** — A drive control varies the voltage of the sawtooth wave which is taken from the output of a sweep

oscillator and applied to the grid circuit of the following sweep amplifier. It is this sawtooth voltage that drives the sweep amplifier. The sawtooth voltage is varied, in different receivers, by either of two distinctly different methods, both of which go by the name of drive control. One method utilizes an adjustable resistor, the other an adjustable capacitor. These controls are used in the horizontal sweep section of receivers having magnetic deflection picture tubes.

**Resistor Drive Control.** — A resistor type of drive control is shown at the left in Fig. 34-1. The control unit is an adjustable resistor in series with the sawtooth capacitor  $C_s$ . The effects of a resistor in this position are explained at some length in the article which is entitled *Amplifier, Sweep*. The resistor

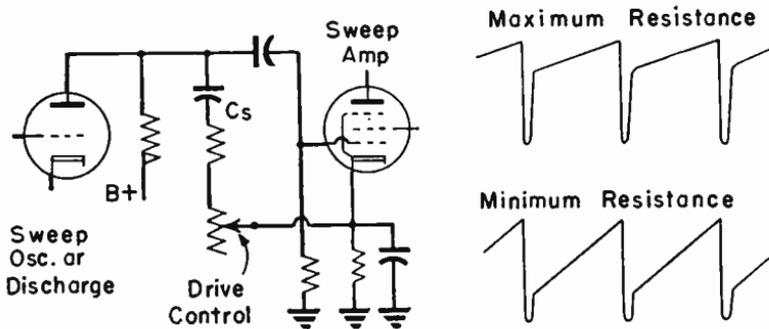


Fig. 34-1.—Resistor type drive control and its action.

causes retention of part of the charge in the sawtooth capacitor during the time in which the sweep oscillator or discharge tube is conductive. This retained charge accounts for the negative peak which precedes the sawtooth portion of the wave.

The greater the resistance in series with the sawtooth capacitor the more of the total voltage or peak-to-peak voltage of the wave appears in the negative peak, and the less remains for the sawtooth rise of voltage which follows the end of the conductive period in the oscillator or discharge tube. This is illustrated at the right in the figure. Decrease of resistance in the drive control lessens the negative peak and increases the sawtooth drive voltage applied to the sweep amplifier.

This type of drive control sometimes is called a peaking control. Again, because it affects the shape of the sawtooth wave, this adjustable resistor control may be called a linearity control.

**Capacitor Drive Control.**— Typical connections for a capacitor type drive control are shown at the left in Fig. 34-2. Sawtooth capacitor  $C_s$  is charged from the B-power supply and discharged through the oscillator or discharge tube in the usual manner. Across this capacitor is developed the sawtooth voltage for deflection. In parallel with capacitor  $C_s$ , and in series with each other, are coupling capacitor  $C_c$  and the adjustable drive control capacitor. The sawtooth voltage across  $C_s$  appears across this series pair,  $C_c$  and the drive capacitor, and divides between them proportionately to their capacitive reactances. Whatever portion of the voltage appears across the adjustable control capacitor is applied between grid and cathode of the sweep amplifier.

To increase the sawtooth voltage on the amplifier grid it is necessary to increase the capacitive reactance of the drive

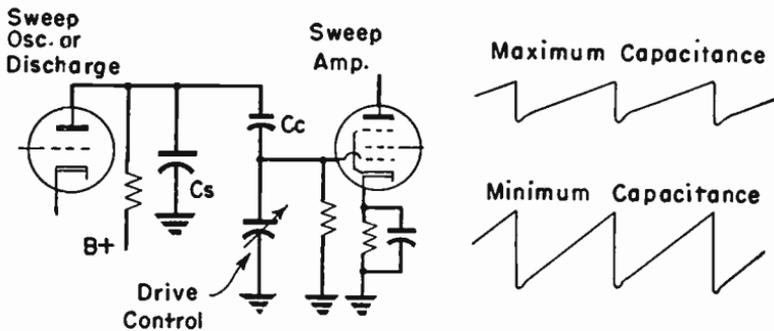


Fig. 34-2.—Capacitor type drive control and its action.

control, which requires a decrease of its capacitance. To decrease the sawtooth voltage on the amplifier grid there must be a decrease of reactance, which requires an increase of capacitance. These relations are shown at the right.

The sawtooth waves of Fig. 34-2 are shown without negative peaks because the capacitor type drive control does not pro-

duce such peaks. Negative peaks may be added, when necessary, by feedback of voltage pulses from some point on the output side of the sweep amplifier.

**Adjustments.** — Adjustment of the resistor and capacitor types of drive control are quite similar, since both controls alter the voltage of the sawtooth wave applied to the grid of the sweep amplifier.

If the drive control is adjusted for too little resistance or capacitance the center of a test pattern will be shifted toward the right-hand side of the screen, crowding the right-hand side and stretching the left-hand side. The width of the picture or pattern is affected by drive control adjustment because width, or extent of beam deflection, varies with voltage of the sawtooth portion of the wave. In some receivers there is no other control for width.

When either type of drive control is used in a receiver having a flyback style of high-voltage power supply, the voltage developed by this power supply and applied to the picture tube anode is affected by drive adjustment. Reducing the resistance or capacitance increases the sawtooth voltage and increases the anode voltage.

In some receivers it is possible to apply excessive anode voltage to the picture tube by wrong adjustment of the drive control, especially where this control is a capacitor type. Unless one is familiar with the effects of a capacitor drive control in the receiver being handled, it is advisable to keep a high-voltage voltmeter connected to the picture tube anode while the control is adjusted, and to avoid raising the anode voltage above the rating for the picture tube. When the drive control is so designed that anode voltage cannot be unduly increased, the control may be adjusted for maximum brightness with which there is satisfactory linearity.

Because the drive control affects both width and linearity, adjustment of any other width control or any other linearity control usually requires resetting of the drive control. One procedure is to first adjust the drive for equal distribution of the test pattern on the left and right, second to adjust any

other width control to make the pattern just fill the mask horizontally, and third to adjust any other linearity control to improve the center of the pattern and keep this point at the center of the mask opening.

With another method the drive control is initially adjusted for minimum sawtooth voltage, and any other width control is set for a picture or pattern slightly too narrow. Then any separate linearity control is adjusted for best pattern distribution right and left, and the drive control is increased just enough to make the picture or pattern fill the mask horizontally.

## E

**ELECTROSTATIC DEFLECTION SYSTEMS.**— Fig. 1 illustrates the principle of deflecting an electron beam by means of electrostatic charges on metal plates. With the beam directed between the plates it is deflected toward the one which is relatively positive and away from the one relatively negative. With two pairs of deflecting plates arranged as in Fig. 2 the beam may be deflected horizontally by alternating electric charges on one pair, and vertically by charges on the other pair.

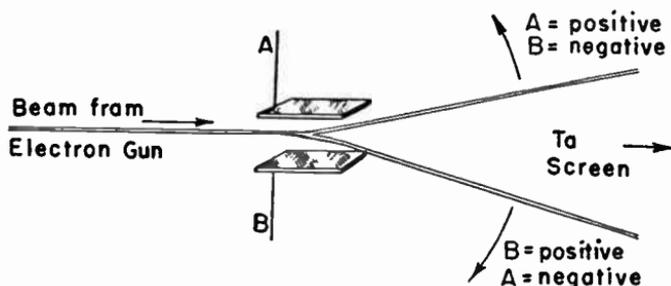


Fig. 1.—The electron beam is deflected by opposite charges on metal plates within the cathode-ray tube.

In Fig. 3 are shown waveforms of sawtooth voltages applied simultaneously to horizontal and vertical deflecting plates from amplifiers which follow horizontal and vertical oscillators. Sawtooths of opposite polarity for the two plates of each pair are obtained by using an inverter such as employed in push-pull amplifiers. Application of a resistance type inverter for electrostatic deflection is shown by Fig. 4. Tubes *A* and *B* are amplifier-inverters. Resistors *R<sub>a</sub>* and *R<sub>b</sub>* are inverter voltage dividers, *C<sub>g</sub>* is a blocking capacitor for the

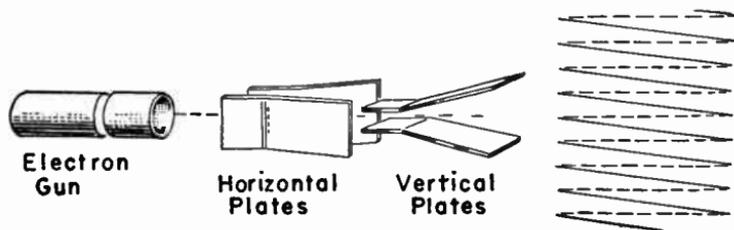


Fig. 2.—Relative positions of horizontal and vertical deflecting plates in an electrostatic picture tube.

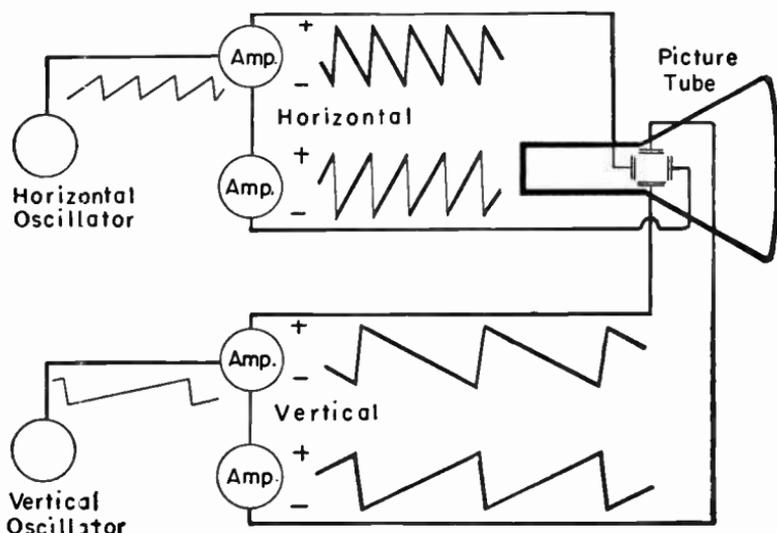


Fig. 3.—Sawtooth voltages for horizontal and vertical electrostatic deflection.

inverter grid, and at  $Ct$  are blocking capacitors for the deflecting plates.

Fig. 5 shows centering controls as used in conjunction with electrostatic deflection. Connections to deflecting plates from sweep amplifier-inverters are through the four blocking capacitors. Accelerating potentials are applied to the deflecting plates by connecting them to the rectifier through centering potentiometers. The rectifier circuit extends from the potentiometers to other receiver circuits. One deflecting plate of each pair connects to the center tap and the other plate to the slider of a control. Moving a slider one way or the other makes the connected deflecting plate positive or negative with reference to the other plate, to a degree depending on how far the slider is moved. Thus, while not being deflected, the beam may be shifted up, down, or to either side. The position to which the beam is shifted by the controls becomes the center for traces displayed on the screen of the cathode-ray tube.

The principle of electrostatic focusing employed with electrostatic deflection is illustrated by Fig. 6. Accelerating

## ELECTROSTATIC DEFLECTION SYSTEMS

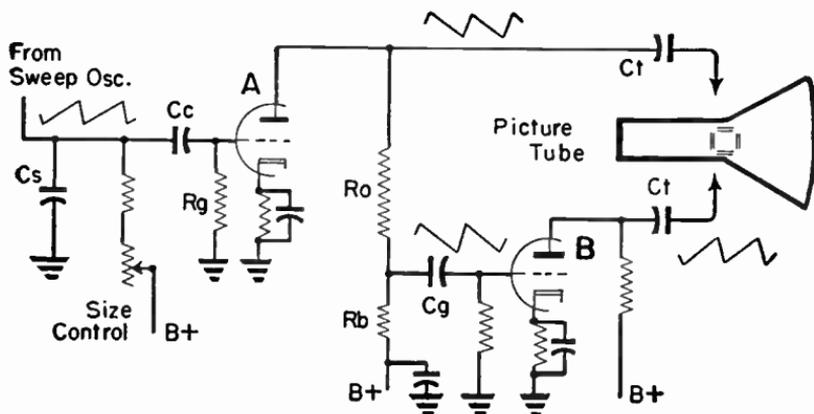


Fig. 4.—Connections for a phase inverter in an electrostatic sweep amplifier. Anodes are maintained at potentials higher than on the focusing anode, resulting in electron lenses in gaps between the anodes, with points of equal potentials along broken lines. Electrons tend to cross these points at right angles, and thus are turned to converge at a sharp focus. Adjusting the potential of the focusing anode affects the electron lenses to move the position of sharpest focus, and in this manner the beam may be focused at the screen of the tube.

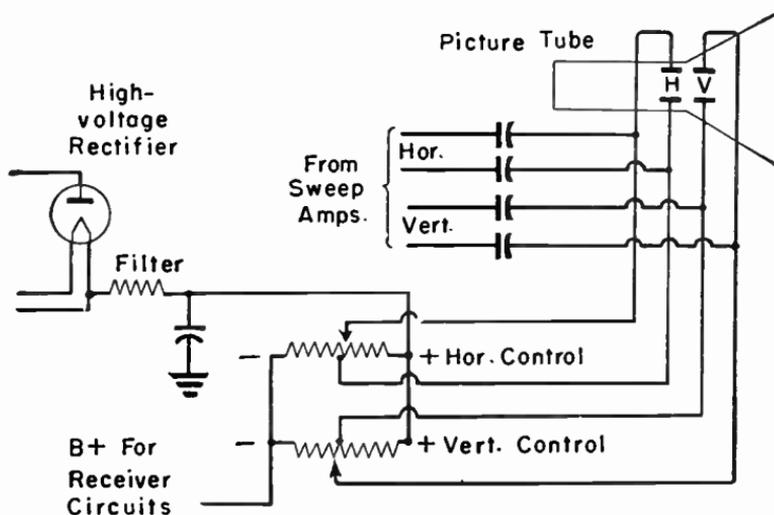


Fig. 5.—Centering controls used with electrostatic deflection.

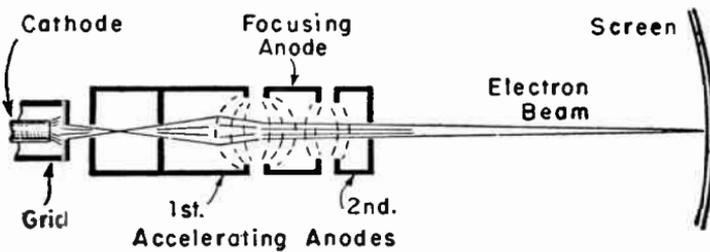


Fig. 6.—Electrostatic focusing is by means of electron lenses between electrodes

Electrostatic deflection is commonly used for oscilloscope cathode-ray tubes, and has been used for some television picture tubes. Fig. 7 shows base pin and element connections for a 7-inch diameter type 7JP4 picture tube.

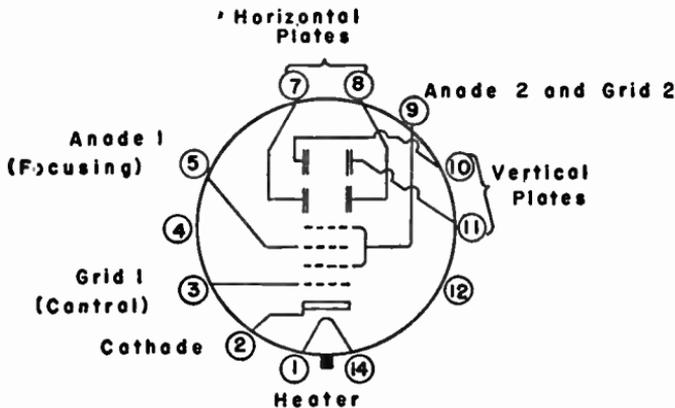
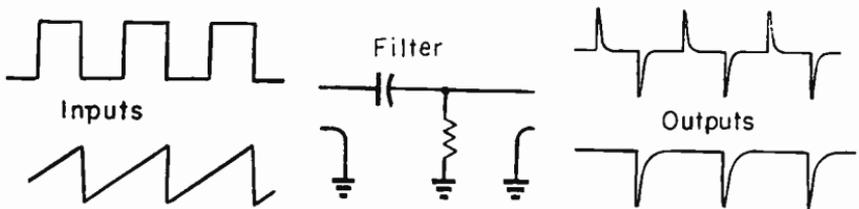


Fig. 7.—Basing connections for a type 7JP4 picture tube using electrostatic deflection.

## F

**FILTER, DIFFERENTIATING.**—As shown by Fig. 35-1, a single-section differentiating filter consists of a series capacitor followed by a shunt resistor between a voltage source and the filter output. In a multi-section filter there will be additional series capacitors, each followed by a shunt resistor.

When input is any voltage having sudden changes, the filter output consists of a series of pulses whose polarities are the same as those of sudden changes at the input. A square wave input, having sudden changes in both polarities, causes



**Fig. 35-1.**—A differentiating filter forms pulses from input voltages having sudden changes.

output pulses in both polarities. A sawtooth input produces output pulses only in the polarity of sudden reversals in the sawtooth. A sine wave or other form having only gradual changes produces no sharp output pulses.

The filter capacitor is charged by each sudden change of input voltage, and discharges through the resistor in a length of time proportional to capacitance-resistance time constant. A time constant short with respect to input frequency causes narrower and sharper pulses, but of less amplitude.

Were horizontal, equalizing, and vertical sync pulses applied to a differentiating filter, resulting output pulses would be as at the bottom of Fig. 35-2. A differentiating filter was used between the sync section and horizontal sweep oscillator of earlier receivers having no automatic frequency control for horizontal sweep oscillator.

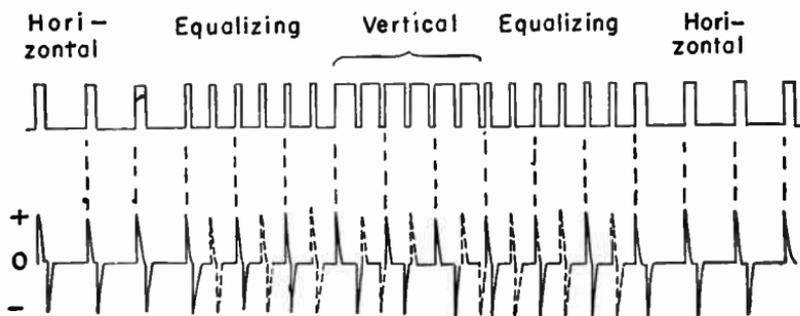


Fig. 35-2.—Action of the horizontal or differentiating filter.

**FILTER, INTEGRATING.** — An integrating filter section consists of a resistor in series between a voltage source and load, and a capacitor shunted across the filter output or load. Integrating filters, usually with two or more sections, are used between sync amplifier and inverter and the vertical sweep oscillator in receivers having no automatic frequency control for the vertical oscillator.

When input pulses are of long duration and close together in time, an integrating filter adds together the effects of successive pulses to build up a charge on the capacitor at the filter output. This is illustrated by Fig. 35-3, where capacitor charge shown along the bottom line increases quite steadily during the entire period of serrated vertical sync pulses. During these pulses the charge and voltage on the capacitor, also voltage to the connected sweep oscillator, reach a value which triggers the oscillator and causes it to

go through one vertical sweep cycle. The next triggering occurs at a similar point in the following vertical sync pulse period.

During widely spaced horizontal sync and equalizing pulses preceding each long vertical pulse the capacitor builds up negligible charges between pulses. During equalizing pulses after each vertical pulse the capacitor loses its accumulated charge, and remains practically without charge during following horizontal sync pulses.

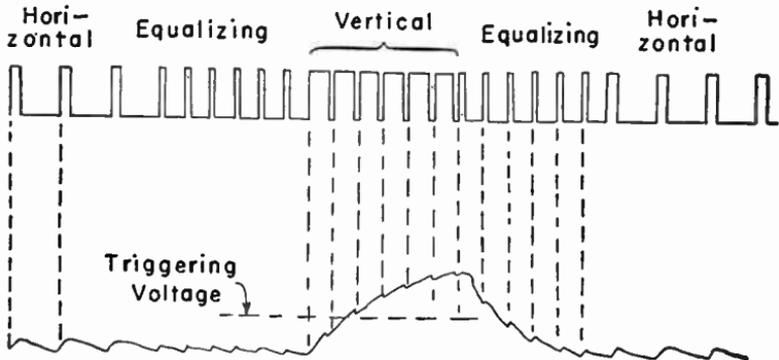


Fig. 35-3.—Action of the vertical or integrating filter.

**FINE TUNING CONTROL.** — A fine tuning control permits the operator of a television receiver to alter the frequency of the r-f oscillator while the receiver is in operation. The control unit usually is an adjustable capacitor, but may be an adjustment for inductance, in the tuned circuit of the oscillator. Fine tuning capacitors may have minimum capacitances of less than 1 mmf and maximums of as much as 10 mmf, but the tuning range ordinarily is something between 1 and 3 mmfd. A given change of fine tuning capacitance causes a much greater change of oscillator frequency on high-frequency channels than on low-frequency channels. The fine tuning control has more effect on tuning in the high-frequency channels.

With receivers employing intercarrier sound systems the most noticeable effect of a fine tuning control is on picture quality and brightness. In such receivers, an alteration of r-f oscillator frequency does not alter the frequency at which

the sound system operates, which always remains 4.5 mc, but does alter the point at which the video intermediate frequency falls on the frequency response curve of the video i-f amplifier and alters equally the point at which the sound intermediate frequency falls on this curve. This changes the amount of amplification applied to the picture signals, but has relatively little effect on sound reproduction unless the sound intermediate frequency is moved so high on the response curve as to introduce excessive amplitude modulation or so low as to drop the sound volume very low.

With receivers whose sound intermediate amplifiers and demodulators operate at the sound intermediate frequency corresponding to the sound carrier frequency the fine tuning control has its chief effect on sound reproduction. When the fine tuning control alters the frequency of the r-f oscillator it alters also the intermediate frequency applied to the sound section. The control must be adjusted to make this intermediate frequency match the frequency for which the sound i-f amplifiers and sound demodulator are aligned. Otherwise there will be no reproduction of sound or else excessive distortion. The video i-f amplifier must be so aligned that, when the sound intermediate frequency is correct for the sound section, the video intermediate frequency falls at the correct point on the frequency response curve.

**FOCUSING AND FOCUSING CONTROLS.** — The electrons in the picture tube beam tend to spread apart because they are negative charges, and like charges repel one another. Some means of focusing is required to direct beam electrons to a single sharply defined spot on the screen. Magnetic focusing utilizes the magnetic field of a focusing coil or of a permanent magnet. Electrostatic focusing employs the electric fields from an internal electrode.

**Magnetic Focusing.** — The principle of magnetic focusing is illustrated by Fig. 37-2. Around the outside of the neck of the picture tube is the focusing coil. In some receivers a permanent magnet is used instead of the current-carrying coil. Lines of the magnetic field produced in or around this coil or

magnet are parallel with and extend in the same direction as the axis of the tube where the lines are inside the neck.

If an electron coming from the cathode and the accelerating anode happens to enter this magnetic field exactly on the axis of the coil and the field, this electron will travel straight along the axis and proceed to the screen as indicated by line *a-a* of the sketch at the bottom of the figure. If the electron is moving

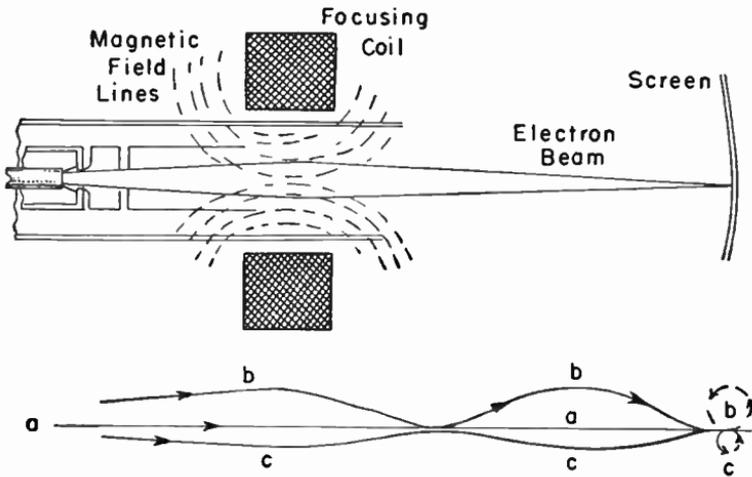


Fig. 37-2.—Principle of magnetic focusing.

outwardly from the axis when it enters the magnetic field, as most of the electrons are moving, this electron will be caused to move in a helical path as it continues on toward the screen. A helical path or a helix is the kind of curve followed by a thread on a straight screw, or around the outside of a cylinder.

The greater the divergence of the electron from the magnetic axis the larger is the diameter of the helix on which the electron is made to travel, and the less the initial divergence the smaller is the diameter. This is indicated by lines *b-b* and *c-c* of the bottom sketch. The circumferences of the helices being followed by all the electrons pass through the magnetic axis at every full turn. All the helices come in to the axis at the same points along the length of the axis. Consequently, all the beam electrons come to a sharply focused spot at each

of these points along the axis. By altering the current in the focusing coil the magnetic field strength may be made such that one of these focus points along the axis comes at the surface of the screen in the picture tube.

Various methods of altering the current in the focusing coil are shown by Fig. 37-3. At the upper left an adjustable control resistor is in series with the coil, between B+ and ground. At the upper right the adjustable control resistor is in parallel with the coil, so that more control resistance forces more cur-

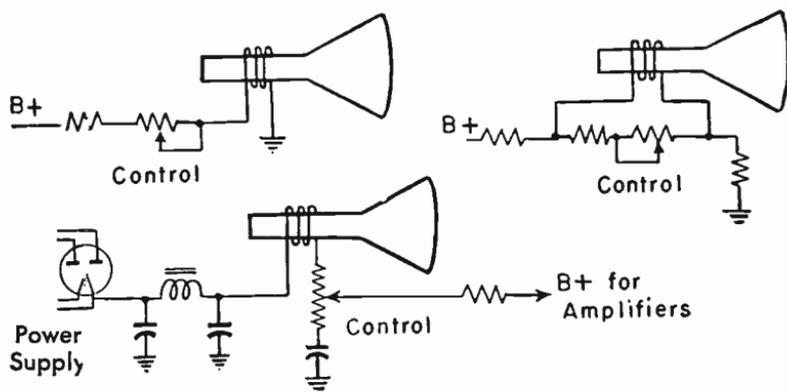


Fig. 37-3.—Control adjustments for current in magnetic focusing coils.

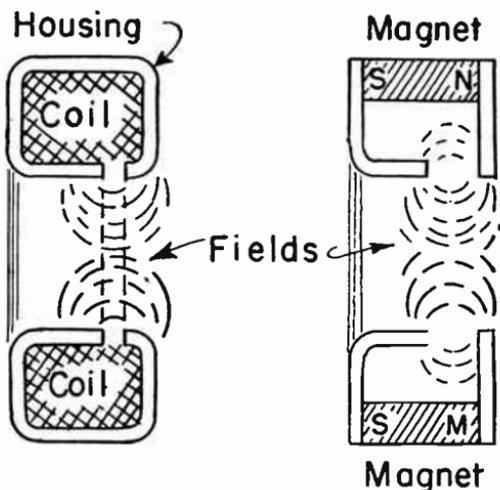
rent to flow in the coil. The lower diagram shows a method of using the focusing coil as a filter choke in the B-power supply system. Adjustment of the control potentiometer causes only small variations of B+ voltage to amplifier circuits, while causing large changes of current in the coil. Coil current may be almost anything between about 75 and 150 milliamperes; the greater currents being found with the larger picture tubes and with tubes operated at high anode voltages.

The focusing coil is mounted over the front end of the electron gun in the picture tube. Between the front side of the focusing coil and the rear of the deflection yoke is a space of about  $\frac{1}{4}$  to  $\frac{3}{8}$  inch to lessen interaction between the magnetic fields for focusing and for deflection of the beam. Moving the focusing coil forward or back along the tube neck may bring a greater screen area into good focus. The coil position may be

changed, while keeping the control adjusted for sharp focus, until there is the best compromise between sharp focusing at the center and the edges of the picture or pattern.

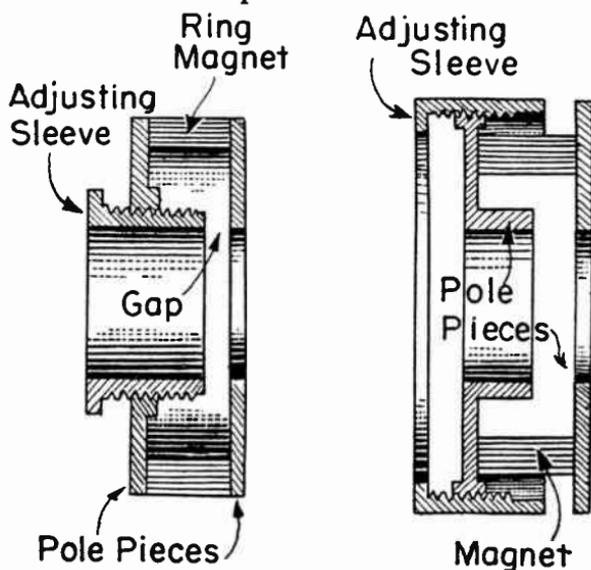
The axis of the focusing coil should coincide with the axis of the tube neck. If necessary, the coil should be moved vertically or horizontally in its mounting to bring the axes into line or very nearly so. Otherwise it may be difficult to avoid shadows at one or more corners of the picture or pattern.

Focusing is adjusted while there is a raster on the screen, but no picture or pattern. This is accomplished by turning the receiver on and setting the channel selector for a channel in which there is no transmission at the time of adjustment. The adjustment should be such that individual fine horizontal lines are clearly visible as separated lines with the contrast control set as for normal reception. Ordinarily it is impossible to have sharp focusing (well separated lines) at all points on the raster area. The adjustment should be set to provide sharp focusing at the center and as far toward the edges as possible. Once the focusing is correctly adjusted it seldom needs further alteration.



**Fig. 37-4.**—Magnetic fields are alike for electromagnetic and for permanent magnet focusing.

**Permanent Magnet (PM) Focusing.** — Magnetic fields for focusing are of similar form whether produced by an electromagnetic focusing coil, as at the left in Fig. 37-4, or by permanent magnets held between pole pieces as at the right. A PM focusing device is mounted to surround the neck of a picture tube in the same position as a focusing coil. Various mechanical arrangements involving screw adjustments allow altering the gap between pole pieces. Two methods are illustrated by Fig. 37-5. Changing the gap varies the strength of magnetic field lines in the region through which passes the electron beam, and thus the point of sharpest focus is shifted to fall at the screen of the picture tube.



*Fig. 37-5.—Some of the adjustments employed with permanent magnet focusing.*

Although centering of pictures often is brought about by tilting an electromagnetic focusing coil with reference to the picture tube axis, PM fusers ordinarily are not tilted for centering. Instead, the front pole piece or an additional small ring in close contact with that pole piece may be moved in relation to the tube axis. Movement of an added ring, here

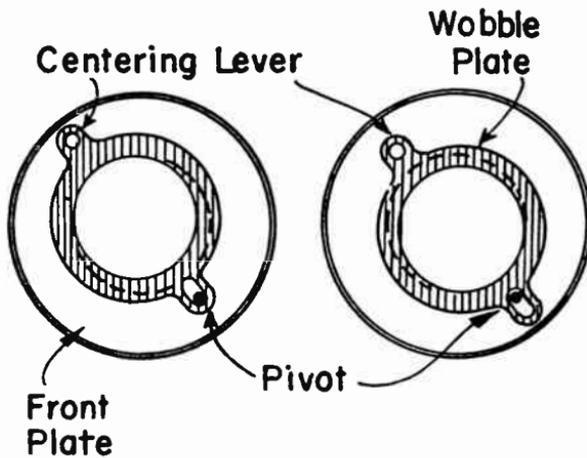


Fig. 37-6.—Movement of the wobble plate in relation to the picture tube axis controls centering.

called a wobble plate, is shown by Fig. 37-6. A slot on one side of the ring fits over a stationary pivot. The ring may be moved to various positions in relation to the tube axis by a centering lever that engages an opening on the opposite side of the wobble plate.

PM focusers should be adjusted with a non-magnetic screw driver or wrench. The limited adjustment range requires that the focuser or strength of its permanent magnets be suited for the size and length of picture tube with which used. The focusing device should not be close to steel supports, which

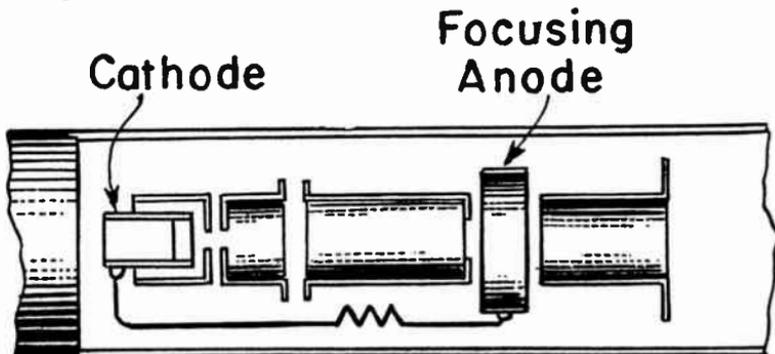


Fig. 37-7.—Position of an electrostatic focusing anode in the electron gun. The resistor is used only with automatic focusing.

would shunt the magnetic field, and should be carried on brackets of non-magnetic material to avoid similar trouble.

On some picture tubes are EM-PM focusers which consist of an electromagnetic focusing coil around the outside of which is a ring-shaped permanent magnet. Focusing current is less that required for a focusing coil having no added permanent magnet. Centering usually is accomplished by tilting the EM-PM focuser unless additional magnetic rings or d-c yoke currents are employed for focusing.

**Electrostatic Focusing.** — In picture tubes designed for electrostatic focusing the portion of the second anode or high-

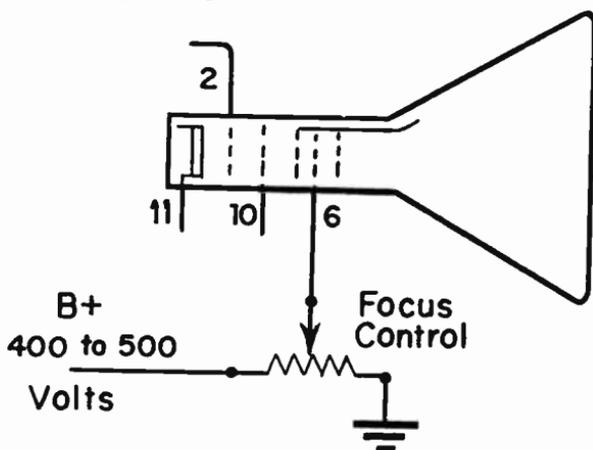


Fig. 37-8.—Control connections for a low-voltage electrostatic focusing anode.

voltage anode which is part of the electron gun is in two sections, one ahead of the other and with a gap between them as illustrated by Fig. 37-7. In this gap is a focusing anode. Potential differences between the focusing anode and the high-voltage anode on each side from electron lenses consisting of electrostatic lines of force, or electric fields. Electrons passing through an electric field tend to turn at right angles to field lines which are of equal potential, and thereby are caused to converge at a small point. The point of convergence or sharpest focus is shifted by adjusting the voltage on the focusing anode

Typical connections for electrostatic focusing are shown by Fig. 37-8. The focusing anode is connected to a base pin in position 6. The focus control potentiometer is connected to a B-plus voltage that allows 200 to 350 volts between the focusing anode and ground or B-minus. Sharpness and uniformity of focus usually improve with higher voltage on the second anode, and to a lesser extent with higher voltage on the second grid. In some receivers the focusing voltage is not adjustable.

Total resistance of the control potentiometer usually is between 1.0 and 2.5 megohms. Current to or from the focusing anode does not exceed 25 microamperes, and ordinarily is near zero. Voltage adjustment seldom is critical, although there will be one setting allowing best average focus over the entire viewing screen.

The system which has been described may be called low-voltage electrostatic focusing, because focusing anode voltage is low enough to avoid insulation difficulties. Earlier methods, designated as high-voltage electrostatic focusing, require 2,200 to 4,000 volts on the focusing anode. This calls for a high-voltage rectifier and filter for focusing voltage, and a control potentiometer with special insulation. Such high-voltage focusing systems commonly are mounted within the same protective housing that encloses the high-voltage power supply for the second anode of the picture tube.

Still another method of electrostatic focusing, called automatic focus or self-focus, has no base pin connection for the focusing anode. This anode is internally connected to the cathode through a fixed resistor, as in Fig. 37-7. There is, of course, no focusing adjustment. To maintain satisfactory focus it is necessary that voltage on the second grid be correctly proportioned to second anode voltage.

With picture tubes designed for electrostatic focusing, centering is accomplished by means of adjustable direct currents in deflecting yoke coils or else by two permanent magnets, usually ring-shaped, mounted concentrically around the tube neck just back of the deflecting yoke. On both sides

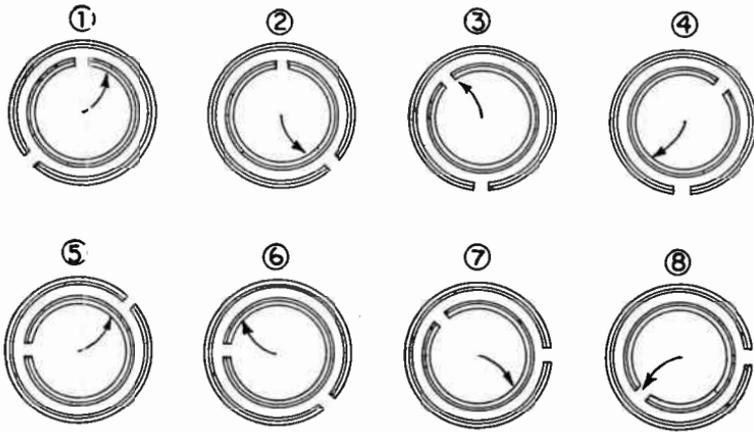


Fig. 37-9.—Permanent magnet centering when only one of the magnets is moved.

of the gap in each magnet are like poles, either north or south, and a consequent pole of opposite polarity is at a point directly opposite the gap. The two magnets may be rotated independently around the tube neck.

If the two gaps originally are above and below the neck axis, and if one gap remains in its original position while the other is rotated, the center of pictures will move as in

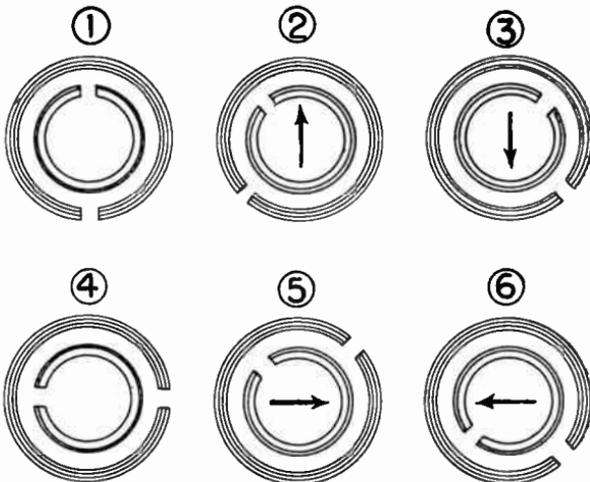


Fig. 37-10.—Effects on centering when both permanent magnets are moved during one operation.

diagrams 1 through 4 of Fig. 37-9. If both gaps originally are on opposite sides of the tube neck, and one remains there

while the other is rotated, effects on centering are as in diagrams 5 through 8. With the gaps initially above and below the neck, as at 1 of Fig. 37-10, shifting both gaps at one operation affects centering as in the other diagrams.

With adjustments indicated by Fig. 9 the picture as a whole is shifted along curved paths, while with adjustments indicated by Fig. 10 the picture may be shifted up or down, or to one side or the other. Distances through which pictures shift are increased as the two gaps are brought closer together, since then their magnetic fields aid, while with the gaps far apart the two fields oppose. This determines extent of shift, while relative positions of the gaps determine direction of shift.

**FREQUENCIES, BANDS OF.** — The accompanying table lists the bands or ranges of frequencies usually denoted by the various names which are in general use. The names which are followed by the word *radio* in parentheses are sub-classifications in the whole range of radio frequencies. A radio frequency is any frequency at which signals may be radiated through space.

Audio frequency	20 to 20,000 cycles
F-m broadcast frequency	88 to 108 megacycles
High (radio) frequency	3 to 30 megacycles
Hyper frequency	1000 to 100,000 megacycles
Longwave	30 to 300 kilocycles
Low (radio) frequency	30 to 300 kilocycles
Medium (radio) frequency	300 to 3000 kilocycles
Microwave	more than 600 megacycles
Power frequency	25 to 60 cycles
Radio frequency	more than 10 kilocycles
Shortwave	1.5 to 30 megacycles
Standard (a-m) broadcast	540 to 1600 kilocycles
Super-high (radio) frequency	over 3000 megacycles
Ultra-high (radio) frequency	300 to 3000 megacycles
Ultra-shortwave	30 to 300 megacycles
Very-high (radio) frequency	30 to 300 megacycles
Very-low (radio) frequency	10 to 30 kilocycles
Video frequency	30 cycles to 4.5 megacycles

**FREQUENCIES, BEAT.** — Along the top of Fig. 39-1 are short vertical lines which may represent the instants of maximum

amplitude in a wave of some certain frequency. On the middle line are vertical lines for the instants of maximum amplitude in a wave of somewhat lower frequency. If voltages at these two frequencies are combined in a single circuit their maximum amplitudes will occur as shown along the bottom line. At certain intervals the instant of maximum amplitude in one wave will coincide with the instant of maximum amplitude in the other wave. Then the two amplitudes will add together to produce strong potentials at these instants, as indicated above the arrows.

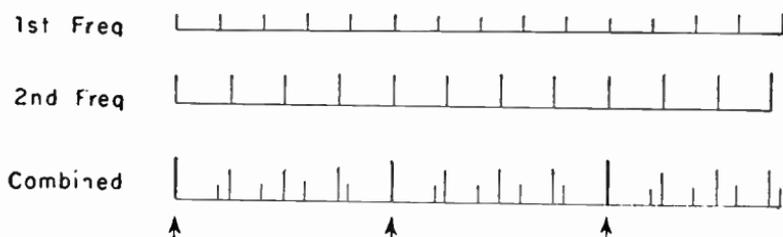


Fig. 35-1.—Constant amplitudes at different frequencies combine to form varying amplitudes and strong beats at the difference frequency.

The frequency at which these relatively strong potentials recur is equal to the difference between the two original applied frequencies. Were one frequency to be 1001 kilocycles and the other 1000 kilocycles the difference frequency would be 1 kilocycle. This difference frequency is called a beat frequency. If both frequencies are applied to a tube having the characteristics of a detector the output will contain the beat frequency, also another frequency equal to the sum of the original frequencies, and the two original frequencies.

The high potential or amplitude of the beat frequency occurs when the original waves are in phase with each other. There is a minimum amplitude when the original waves are

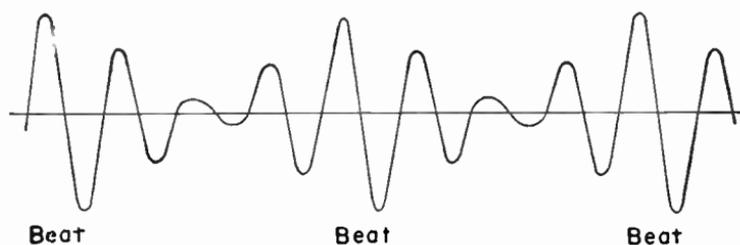


Fig. 39-2.—Output current or voltage showing beats.

in opposite phase. Changes of phase produce an output current and voltage of the general type shown by Fig. 39-2. The changes of amplitude in this combined wave are just like the changes in any wave which is amplitude modulated. To extract the beat frequency this combined wave must be applied to some kind of detector which cuts off either the top or bottom of the wave (a diode detector) or else displaces the wave toward the negative or positive side to give it a d-c component (a grid leak or a plate rectification detector). If either the original input voltages is modulated, that modulation will appear on the beat frequency in the output.

Frequencies produced by beating are employed in the intermediate amplifiers of television receivers and all superheterodyne receivers, where signal and local oscillator frequencies are applied to a mixer or converter tube. Beats are employed also for producing the 4.5 intercarrier sound signal in television receivers, where the video detector produces the intercarrier beat from the video and sound intermediate frequencies. Beat frequencies are utilized also in many types of signal generators and other testing instruments.

**FREQUENCIES, STANDARD.** — The most convenient and reliable source of standard frequencies which may be used for calibration of testing equipment is in transmissions from radio station WWV of the National Bureau of Standards at Washington, D. C. These frequencies are accurate to within one part in 50 million. They are broadcast continuously, day and night, with carrier frequencies of the following numbers of megacycles:

*2.5, 5, 10, 15, 20, 25, 30, and 35.*

The frequencies of 5, 10, 15, and 20 mc are transmitted with powers of 8 to 9 kilowatts, the 2.5 mc frequency with 700 watts, and the others with 100 watts. Carriers at 10, 15, 20, and 25 mc are modulated with audio frequencies of 440 and 4000 cycles per second. Those at 2.5, 5 and 30 mc are modulated with audio frequency of 440 cycles per second. There is a station announcement in voice at each half-hour and each hour.

**FREQUENCIES, TELEVISION.**—The following frequencies are used in or related to television reception.

**Accompanying Sound Frequency:** This is the sound intermediate frequency of the channel in which there is reception. It is a beat frequency produced by the r-f oscillator frequency and the sound carrier frequency in the received channel, and equal to the difference between those two frequencies.

**Adjacent Sound Frequency:** A beat frequency produced by and equal to the difference between the r-f oscillator frequency and the sound carrier frequency in a channel of lower number and immediately adjoining the one in which there is reception. This frequency is 6.0 mc higher than the sound intermediate frequency and 1.5 mc higher than the video intermediate frequency.

**Adjacent Video Frequency:** A beat frequency produced by and equal to the difference between the r-f oscillator frequency and the video carrier frequency in a channel of higher number and immediately adjoining the one being received. This frequency is 6.0 mc lower than the video intermediate frequency and 1.5 mc lower than the sound intermediate frequency.

**Associated Sound Frequency:** Same as *Accompanying Sound Frequency*, which see.

**Center Frequency:** An unmodulated carrier frequency or unmodulated intermediate frequency. This would be the carrier frequency before it is modulated for transmission of signals, or an intermediate frequency produced by beating of the r-f oscillator frequency with an unmodulated carrier frequency.

**Channel Frequency:** The carrier frequencies. In channels numbered 2 through 6, which may be called the low-band channels in the very-high frequency range, the carrier frequencies extend from 54 through 88 mc. In channels 7 through 13, the high-band channels in the very-high frequency range, the carrier frequencies extend from 175 through 216 mc. Television broadcast frequencies originally allo-

cated in the ultra-high frequency range extend from 480 through 920 mc. Television relay frequencies are still higher, as in bands such as 3500 to 4200 mc and 5850 to 8500 mc.

**Field Frequency:** The frequency with which successive picture fields recur; 60 cycles per second.

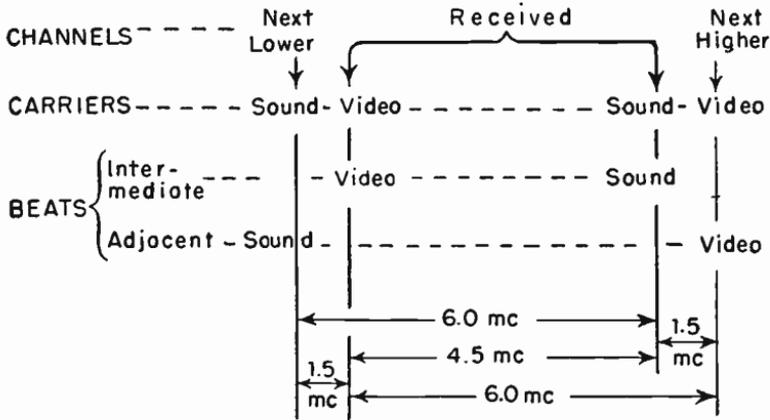


Fig. 41-1.—Relations between some of the television frequencies.

**Frame Frequency:** The frequency at which successive complete frames of the picture are produced; 30 cycles per second.

**Intercarrier Beat Frequency:** A beat frequency produced by the video and sound intermediate frequencies, and equal to the difference between those frequencies. This beat frequency always is 4.5 mc. The intercarrier beat frequency retains the modulation of sound signals and is used to feed the sound sections of receivers employing the intercarrier sound system.

**Line Frequency:** The frequency at which successive horizontal picture lines or traces recur. In the 525-line system the line frequency is 15,750 cycles per second; the product of 525 lines per frame and 30 frames per second.

**Sound Carrier Frequency:** The transmitted frequency which is frequency-modulated with signals for sound that is to accompany the visual television program. In each channel the sound carrier frequency for that channel is 4.5 mc higher than the video carrier frequency in the same channel.

**Sound Intermediate Frequency:** A beat frequency produced by the r-f oscillator frequency and the sound carrier frequency of the channel being received, and equal to the difference between those frequencies. The sound intermediate frequency is 4.5 mc lower than the video intermediate frequency when, as is usual, the r-f oscillator frequency is higher than the received carrier frequencies.

**Video Frequency:** Any frequency in the range required for reproduction of picture lights and shades. Video frequencies are employed for modulation in transmitting the picture portion of the television signal. This range extends from 30 cycles to about 4.5 megacycles, although not all receivers utilize or reproduce this entire range.

**Video Carrier Frequency:** The transmitted frequency which is amplitude-modulated with signals for picture lights and shadows, blanking level, and all synchronizing pulses required for reproduction of pictures. In each channel the video carrier frequency is 4.5 mc lower than the sound carrier frequency in the same channel.

**Video Intermediate Frequency:** A beat frequency produced by the r-f oscillator frequency and video carrier frequency of the channel being received, and equal to the difference between those two frequencies. The video intermediate frequency is 4.5 mc higher than the sound intermediate frequency in the same receiver when the r-f oscillator frequency is higher than the received carrier frequencies.

**FREQUENCY-WAVELENGTH RELATIONS.**—The accompanying formulas allow determination of equivalent frequencies and wavelengths.

$$\text{Wavelength, meters} = \frac{300,000}{\text{frequency, kilocycles}}$$

$$\text{Wavelength, meters} = \frac{300}{\text{frequency, megacycles}}$$

$$\text{Frequency, kilocycles} = \frac{300,000}{\text{wavelength, meters}}$$

$$\text{Frequency, megacycles} = \frac{300}{\text{wavelength, meters}}$$

## G

**GAIN CONTROL, AUTOMATIC.** — Automatic gain control circuits in a television receiver increase the amplification in r-f and i-f amplifiers when weak signals reach the antenna, and reduce the amplification for strong signals. Inputs to video and sound detectors then remain nearly constant with changes of antenna signal strength.

Gain control systems apply a variable negative bias voltage to the control grids of one or more i-f amplifiers, also in most cases to the r-f amplifier or amplifiers, and sometimes to one of the video amplifiers as well. This bias voltage becomes more negative when the received signal becomes stronger, and less negative on weaker signals.

An effective automatic gain control lessens the need for readjustment of contrast and brightness when changing from one station to another, also maintains uniform picture quality and sound volume when there are variations of antenna signal strength during a program. Many automatic gain controls reduce picture flicker such as caused by low flying airplanes, reduce the effects of outside electrical interference classed as "noise," and prevent pictures and sound from being greatly affected by power line voltage fluctuations, swaying antennas, and other faults.

Gain control bias voltage must vary with some characteristic of the received signal which changes only with change of signal strength, and not with changes of picture tone or shading. The characteristic which meets this requirement is the amplitude or voltage of the peaks of the sync pulses. This maximum pulse amplitude is maintained constant at the transmitter, and varies at the receiving antenna only when there is some variation in signal attenuation between transmitter and receiver. The sync pulse tips represent 100 per cent signal strength, while the black or blanking level represents 75 per cent, maximum white level represents 15 per cent, and picture modulation varies between 15 and 75 per cent.

As shown by Fig. 43-1, the sync pulse tips remain at the same amplitude with dark or light pictures and all intermediate shadings. Average amplitude varies, being high for dark pictures and low for light pictures. Were this average amplitude used for regulating the gain control, as it is used in automatic volume controls for sound receivers, the picture brightness would be maintained constant instead of showing changes which actually occur in brightness of televised scenes.

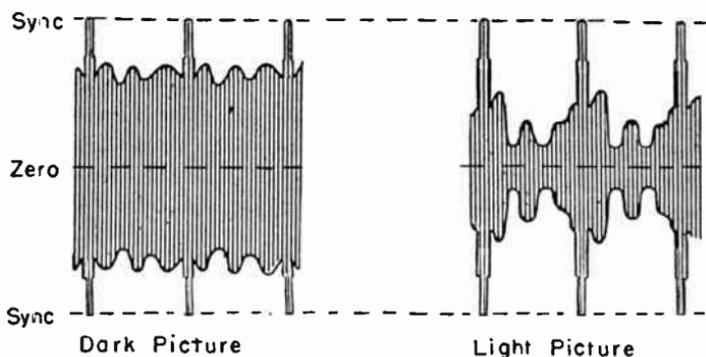


Fig. 43-1.—Sync pulse amplitude is not altered by changes of picture tone or shading.

**Gain Control Circuits.** — Circuits for one of the simplest automatic gain control systems are shown by Fig. 43-2. Connected to the output of the final i-f amplifier, through capacitor  $C_a$ , is the plate of the agc (automatic gain control) diode. The cathode of this diode is connected to ground. During each half-cycle of intermediate frequency in which the top of the coupling coil becomes positive with reference to the lower end and ground, the plate of the diode is made positive with reference to its cathode and ground. During this half-cycle there is electron flow in the direction of the arrows, and capacitor  $C_f$  is given a small charge in the polarity marked.

During intervening half-cycles the diode plate is made negative with reference to the cathode, and this tube becomes non-conductive. Then the charge of capacitor  $C_f$  may escape through resistors  $R_a$  and  $R_b$  and the ground path. But because the rather large capacitance of  $C_f$  and high resistance of  $R_a$  and  $R_b$  produce a long time constant for discharge, only a little of the charge escapes before more is added during the following half-cycle in which the diode again is conductive.

The charge on capacitor  $C_f$  increases until the accompanying voltage reaches and remains at practically the voltage of the peak amplitude of the sync pulses. This voltage, negative at the top of capacitor  $C_f$ , is applied as negative grid bias to the grid returns of whatever tubes are to have automatic gain control. One such tube is shown at the left in the diagram. The stronger the signal at the antenna and at the

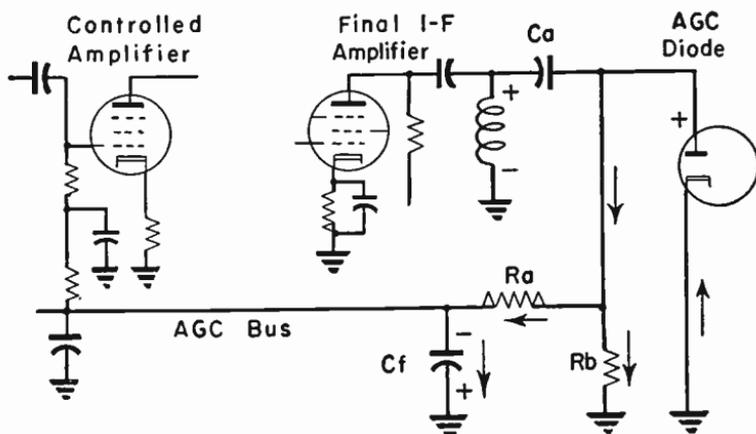


Fig. 43-2.—Circuits for a simple type of automatic gain control.

output of the i-f amplifier the greater becomes the peak amplitude of the sync pulses, and the greater the charge and voltage on capacitor  $C_f$ . The greater voltage then makes the controlled grids more negative and reduces amplification.

Weaker incoming signals reduce the amplitude of sync pulse peaks, allow reduction of charge and voltage of capacitor  $C_f$ , and make the controlled grids less negative for an increase of amplification.

The method of obtaining a gain control voltage which is proportional to peak voltage of sync pulses, as illustrated in Fig. 43-2, is utilized in some manner or other for the majority of automatic gain control systems. There is great variety in details of agc systems. Some are quite simple while others are complex. The principal variations will be examined.

**Delayed Automatic Gain Control.**—With the simple agc system of Fig. 43-2 the bias voltage, with no incoming signal, may be such as produces fairly high amplification. As soon as a signal is applied, or as soon as there is any increase in strength of an applied signal, there is an immediate increase of negative bias voltage and a reduction of gain. Often it is desirable that there be no such reduction of gain on very weak signals, which need all the amplification possible. Controls which do not reduce amplification until the received signal exceeds some certain strength are called delayed automatic gain controls.

A simple method of obtaining delay is illustrated by Fig. 43-3. The cathode of the agc diode is not connected to ground but to a tap on a voltage divider between  $B+$  and ground.

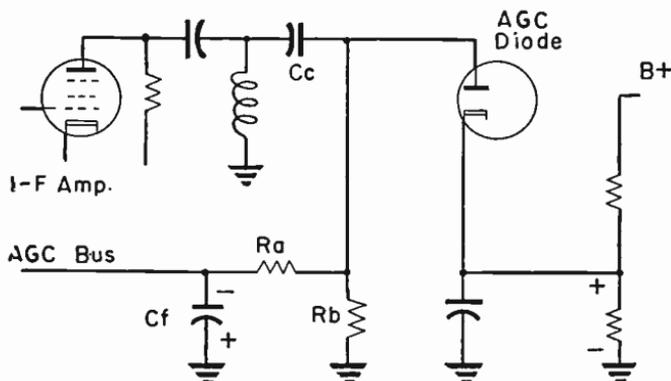


Fig. 43-3.—Delayed control resulting from a positive voltage on the cathode of the agc tube.

This tap usually is at a potential of 1 to 3 volts positive. Then, with no received signal, the cathode of the agc diode is 1 to 3 volts positive with reference to the plate, which is the same as making the plate 1 to 3 volts negative with reference to the cathode. Until i-f signal voltage applied to the agc diode exceeds the value of positive voltage on the cathode of this diode there can be no conduction in the tube. With no conduction there will be no negative biasing voltage built up across capacitor  $C_f$ , and there will be no reduction of amplification in the tubes connected to this agc system.

In some receivers the positive delay voltage is adjustable. The cathode of the agc diode is connected to the slider of a potentiometer in a voltage divider, instead of to a fixed point on the divider as in Fig. 43-3. Making the cathode less positive allows automatic reduction of gain on weaker signals, as might be required in a locality of high signal strength. Making the cathode more positive delays the agc action until stronger signals are received, as might be desirable in an area of low signal strength.

With the circuit connections of Fig. 43-4 the positive delay voltage is varied by the contrast control. This method of control is used in many receivers. The voltage divider between  $B+$  and ground includes resistor  $R_c$  and the portion of the contrast control potentiometer between its lower end and the

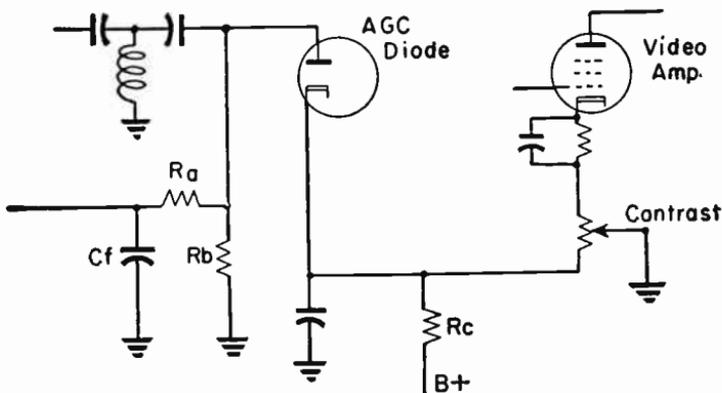


Fig. 43-4.—Agc delay voltage varies by adjustment of contrast control in the cathode circuit of a video amplifier.

grounded slider. Adjustment of the contrast control varies the grid bias and gain of the video amplifier tube in whose cathode circuit this control is located. At the same time, the adjustment varies the amount of positive delay voltage on the cathode of the agc diode.

When the contrast control is adjusted for maximum gain in the video amplifier, or for minimum bias in this tube, there is the greatest positive delay voltage. Thus, when the contrast control is set for maximum gain, the agc system is prevented from reducing the gain on weak received signals. When the contrast control is set for less gain or less contrast the agc system will act on weaker signals.

Fig. 43-5 shows another method of adjusting a positive delay voltage by means of a contrast control in the cathode circuit of a video amplifier. The variable positive delay voltage appears at the junction between resistor  $R_c$  and the lower end of the contrast control potentiometer. Instead of applying this delay voltage to the cathode of the agc diode it is fed

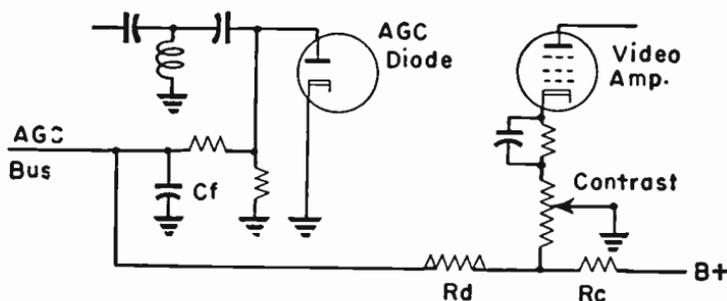


Fig. 43-5.—A positive delay voltage applied directly to the agc bus.

through about one megohm resistance at  $R_d$  to the agc bus, where the positive voltage opposes the negative voltage produced across capacitor  $C_f$ .

When the contrast control is adjusted for more contrast or more gain in the video amplifier there is increase of the positive delay voltage at the junction of  $R_c$  and the contrast control. This increased positive voltage reduces the effect of

negative biasing voltage from capacitor  $C_f$ , and tubes which are automatically controlled then operate with full gain until there is a stronger received signal.

**Amplifiers in Agc Systems.** — In many receivers there are agc systems which incorporate an amplifier tube in addition to the tube which furnishes a voltage proportionate to peak amplitude of the sync pulses. Some of the principles employed in circuits using an amplifier are illustrated by Fig. 43-6. As in agc circuits examined earlier, the video signal is rectified by a diode tube used for the agc takeoff. The cathode of this takeoff tube is connected to ground through a series of resistors in which is the potentiometer  $R_i$ , whose slider is connected to the grid of the amplifier tube. Increase of signal strength increases the electron flow in the diode. This flow will be upward in the resistance at  $R_i$ , tending to make the top of  $R_i$  more positive with reference to the ground end, and making the amplifier grid less negative with reference to the amplifier cathode. Thus an increase of signal strength causes an increase of amplifier plate current. Time constants in the diode cathode circuit are such that capacitor  $C_i$  remains charged to a voltage proportional to the peak amplitude of the sync pulses.

The entire amplifier system of Fig. 43-6 is connected between B- and ground. Voltage divider action in the various resistances, combined with potential drops which accompany the electron flows, may be assumed to place the elements of the amplifier tube at the voltages marked, which are with reference to ground. The grid (-95v) is more negative than the cathode (-90v), so is negatively biased. The plate (-6v) is much less negative than the cathode and, in effect, is 84 volts positive with reference to the cathode. All the negative potentials shown on the diagram are purely for illustration of principles. Actual receiver voltages might differ widely from these and still have the same general relations at the amplifier.

When a stronger incoming signal increases plate current in the amplifier there is increased flow in resistor  $R_o$  and the top of  $R_o$  becomes more negative with reference to ground.

The potential at the top of  $R_o$ , and at the amplifier plate, is that for the agc bus which connects to the grid returns of all tubes whose gain is automatically controlled. Thus the automatic bias voltage is made more negative by increased signal strength and amplification of the controlled tubes is reduced.

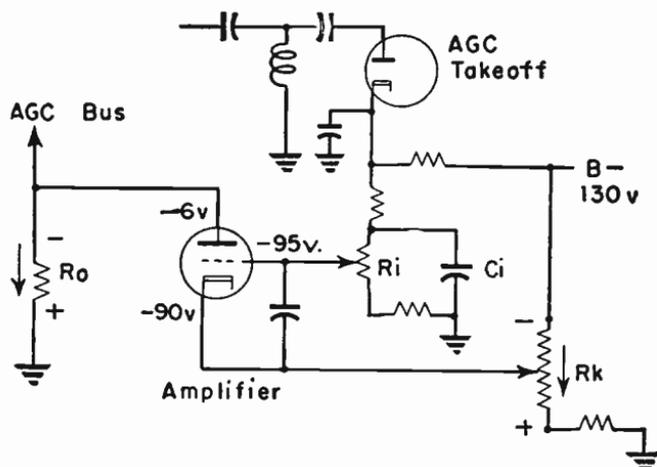


Fig. 43-6.—An amplifier for increasing the automatic biasing voltage and inverting its polarity.

Shown in Fig. 43-6 are two adjustments which affect the strength of signal necessary to cause any given automatic bias voltage at the amplifier plate. The adjustment at  $R_i$  alters the steady negative grid voltage which exists in the absence of current in the agc takeoff diode. This is a grid bias voltage for the amplifier. The adjustment at  $R_k$  alters the steady negative potential of the cathode with reference to ground. Change of cathode potential alters both the grid-cathode and the plate-cathode potential differences in the amplifier tube, alters the amplifier plate current, and the automatic bias voltage for any given signal amplitude or signal strength. Ordinarily only one or the other of these adjustments will be found in the same receiver.

**Limited Control for R-f Stages.**—The prime purpose of automatic gain control is to provide constant signal strength

at the video detector and thus reduce the need for readjustment of brightness and contrast controls. This purpose is satisfied by making the automatic negative bias increase uniformly with signal strength. It is found, however, that an improved signal to noise ratio may be secured by allowing the r-f amplifier or amplifiers to operate at full gain on all signals which won't cause overloading of the following i-f stages. With full automatic control on r-f amplifiers the highly negative bias may cause cross modulation, which allows strong signals to ride through with weak ones and produce an effect of poor selectivity.

In some receivers only some definite fraction of the automatic negative bias is connected to grid returns of r-f amplifiers while the full agc voltage or a greater fraction of it is connected to grid returns of i-f amplifiers. Voltage divider circuits such as those of Fig. 43-7 are used for this division. In other receivers there is a switch for cutting off the agc action

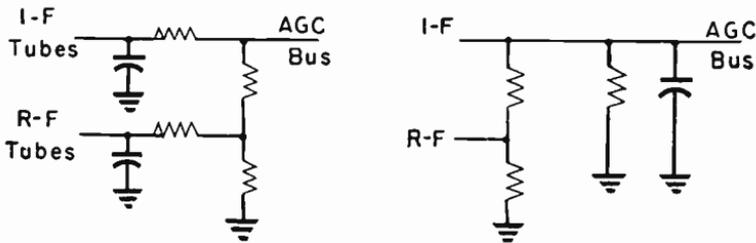


Fig. 43-7.—Methods of lessening the automatic gain control action on r-f amplifiers.

when receiving weak signals. If agc action is cut off when receiving strong signals the resulting overloading of i-f stages may cause excessive control on these stages, and a picture which appears weak or all over gray.

By using two separate diode takeoff tubes, one for i-f biasing and the other for r-f biasing, it is possible to use different delay voltages on the two tubes. Then the greater delay is applied to the r-f amplifier or amplifiers and the lesser delay to the i-f amplifiers. The same general method may be used

with two separate amplifiers, whose steady grid or cathode voltages may be adjusted for different delays, or for different signal strengths required for production of equally negative bias voltages. The same general idea is utilized by having a single takeoff diode feeding a single amplifier, with the agc voltage output of this amplifier applied to two separate diodes. One of these diodes may be cathode-biased for a delay suited to i-f amplifiers and the other diode may be cathode-biased for a delay suited to r-f amplifiers.

**Time Constants and Noise.** — In order that the automatic negative bias voltage may remain proportional to input signal strength the time constant of the gain control system must be longer than the period of one horizontal line. This time constant is the discharge time constant of the capacitor which has been marked  $C_f$  and the resistors marked  $R_a$  and  $R_b$  in some preceding diagrams. This time constant must be long enough to prevent any great loss of charge and bias voltage between successive horizontal sync pulses.

Should any external electrical interference be of a type causing brief voltage peaks to come through the r-f and i-f amplifiers these peaks may act on the agc system in the same manner as the sync pulses. These interference voltages, usually called noise, may establish a negative biasing voltage of their own if the interference is stronger than the sync pulses. This false agc voltage will reduce amplification for desired signals, possibly to the extent that program reception is impossible when or where the noise interference level is high.

Many expedients are employed for making the agc system more or less independent of noise interference. Referring to Figs. 43-2 and 43-3, the time constant of capacitor  $C_a$  and resistor  $R_b$  may be made relatively short. Then, even though  $C_a$  is charged by noise to a voltage higher than the sync pulse peaks, this capacitor discharges so quickly as to add but little to the biasing voltage maintained on capacitor  $C_f$ .

Fig. 43-8 illustrates the principles of two methods of using diodes as noise limiters in connection with an agc amplifier.

At *A* the cathode of the diode limiter is connected to the negative agc bus on the plate side of the amplifier. The plate of the diode limiter is connected to a negative voltage equal approximately to the maximum automatic bias which is to be allowed. That is, this limiting negative voltage would correspond to maximum peak amplitude of sync pulses. Noise peaks whose voltage exceeds that of the signal sync pulses

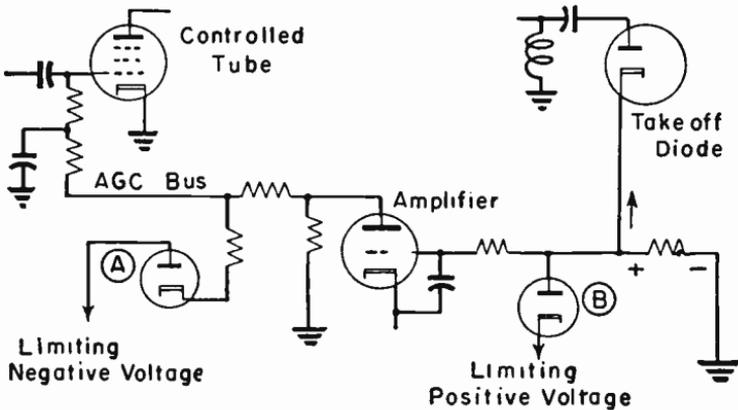


Fig. 43-8.—Two ways in which diodes may be used as noise limiters.

will make the amplifier plate and diode limiter cathode more negative. Then the cathode of the diode will be more negative than its plate, which is equivalent to making the plate positive with reference to the cathode. This will cause the diode to conduct and load the circuit sufficiently to prevent excessively negative voltage on the agc bus.

Diode *B* of Fig. 43-8 is connected to the input or grid side of the agc amplifier, with the plate of this limiter diode to the line between the takeoff diode and the amplifier. In this line the potential becomes more positive with reference to ground when there is increase of signal strength. This positive potential acts on the plate of the limiter diode. The cathode of this limiter diode is connected to a positive voltage which prevents conduction in the diode so long as voltage on the diode plate does not exceed that corresponding to peak amplitude of the sync pulses.

Noise peaks which would cause more current in the take-off diode and a more positive voltage at the amplifier grid also make the plate of the limiter diode more positive. Then this diode conducts and loads the amplifier grid circuit to keep the grid from becoming sufficiently positive to make a material increase of negative bias voltage in the plate circuit of the amplifier. Only one or the other of the diode limiters ordinarily would be used in an agc system.

**Keyed or Gated Automatic Gain Control.** — In keyed or gated methods of automatic gain control the control tube is conductive during only the periods of horizontal sync pulses. Unless noise peaks occur simultaneously with the sync pulses, the noise voltages can neither add to nor subtract from the automatic biasing voltage which is developed by direct or

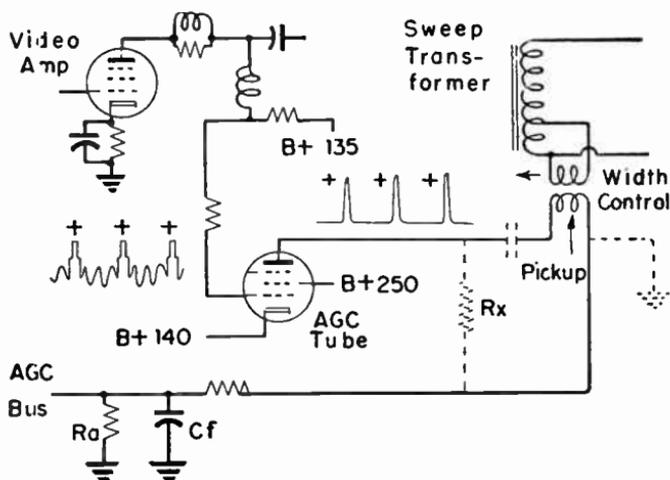


Fig. 43-9.—A keyed agc system in which keying pulses are taken from the horizontal sweep output transformer.

indirect action of the horizontal sync pulses. In between successive horizontal sync pulses the control tube is made non-conductive either by holding its plate at a voltage less positive than necessary for conduction of else by holding its grid more negative than required for plate current cutoff.

Principles employed in one method of keyed automatic gain control are illustrated by Fig. 43-9. To the grid of the agc tube is applied the entire composite video signal as taken from the output of a video amplifier. Average voltage on the agc grid is highly positive, shown here for purposes of explanation as 135 volts. The cathode is connected to a voltage slightly more positive, which makes the grid negative with reference to the cathode. Screen voltage is relatively high.

In the system illustrated the pickup coil in the plate circuit is coupled to the width control inductor on the secondary of the sweep transformer in a flyback type of high-voltage power supply. In this secondary winding there is sudden reversal of current during each horizontal retrace period. The change of magnetic field due to current reversal induces a strong voltage pulse in the coil and a brief highly positive voltage on the plate of the agc tube. This happens while the grid of the agc tube is being driven more positive by a sync pulse.

There are various ways of introducing positive voltage peaks from the sweep circuit into the plate circuit of the agc tube. With one method the conductor from the tube plate extends unbroken through the pickup coil and to the agc bus. With another method, indicated by broken lines in Fig. 43-9, the pickup coil is connected to the plate circuit through a capacitor and ground, with a conductive connection from the plate through resistor  $R_x$  to the agc bus. In any case, the average potential of the plate in the agc tube is practically ground potential except when the positive voltage peaks occur, and is so highly negative with reference to the cathode as to prevent conduction except during those voltage peaks.

While the plate of the agc tube is positive, the positive sync pulses on the grid cause conduction. The plate current charges capacitor  $C_f$  to build up a negative biasing voltage at the agc bus in the usual way. Since the tube is conductive only while sync pulses act on the grid, any voltage pulses due to noise interference occurring between sync pulses cannot affect the negative agc voltage. Strength of negative biasing

voltage is proportional to plate current of the agc tube. Plate current is proportional to strength of sync pulses on the grid, and sync pulse amplitude is proportional to strength of the received signal. Consequently, the agc bias is proportional to signal strength.

In the gated agc system of Fig. 43-10 the video signal with negative sync pulses is taken from the output of the video detector and applied to the grid of the first agc amplifier. Stronger incoming signals make the grid more negative and reduce the plate current. This plate current of the first agc amplifier flows in cathode resistor  $Rk$ , which is also in the cathode circuit of the second agc amplifier. When stronger signals

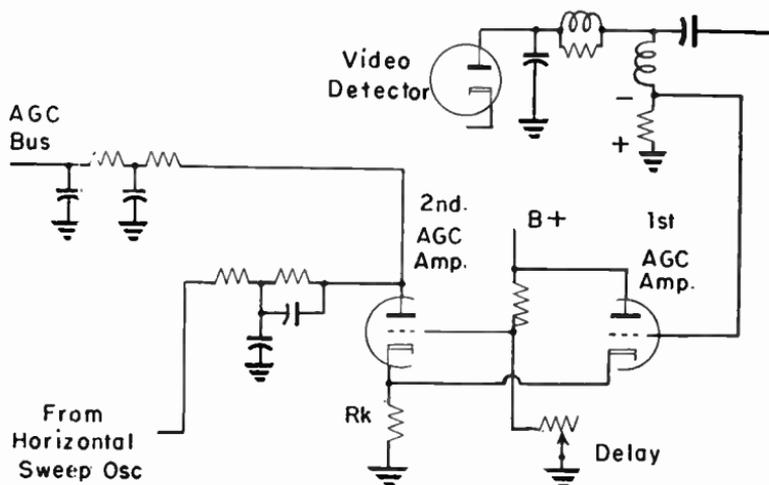


Fig. 43-10.—Gated agc system in which voltage peaks for making an amplifier conductive are taken from the horizontal oscillator circuit.

reduce plate current in resistor  $Rk$  there is less voltage drop across this resistor, which makes the grid bias of the second amplifier less negative and increases plate current in the second amplifier. Increase of plate current in the second amplifier makes its plate more negative, due to inversion of grid and plate voltages which occurs in any amplifier. Thus the automatic biasing voltage from the plate of the second amplifier becomes more negative when there is increase of signal strength.

## GAIN CONTROL, AUTOMATIC

The gated effect in the system of Fig. 43-10 results from the fact that plate current can flow in the second agc amplifier only when its plate is made positive with reference to its cathode by a sine wave voltage taken from the circuit of the horizontal sweep oscillator. This voltage is positive at the plate of the second amplifier only while a horizontal sync pulse is reaching the grid of the first amplifier. Noise voltages occurring at other times have no effect on the automatic biasing voltage.

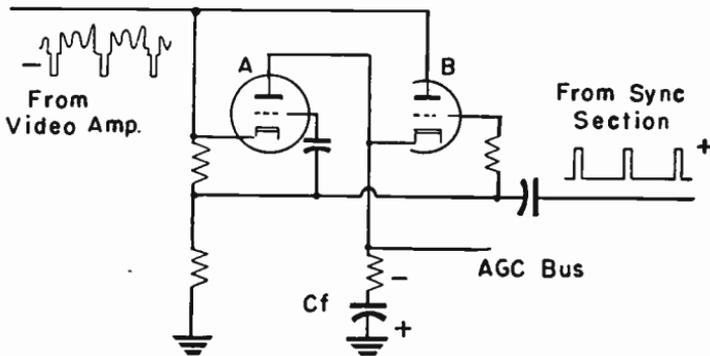


Fig. 43-11.—Keyed agc system in which positive keying pulses for control grids are taken from the sync section of the receiver.

The elementary principle of another keyed agc system is shown by Fig. 43-11. The video signal with sync pulses negative is taken from the output of a video amplifier and applied to the cathode of agc tube *A*, also to the plate of agc tube *B*. A stronger signal tends to increase conduction in tube *A*, because the cathode is made more negative, and at the same time tends to decrease or stop conduction in tube *B*, because its plate is made more negative.

To the grids of both agc tubes are applied positive sync pulses taken from a point in the sync section where picture signals have been removed and only the uniform sync pulses remain. The agc tubes can conduct only while these positive pulses from the sync section are acting on the agc grids. This happens only while negative sync pulses from the video

amplifier are acting on the cathode of tube *A* and on the plate of tube *B*. Noise voltages occurring between the sync pulses find the tubes nonconductive, and have no effect on agc voltage.

In the system of 43-11 the automatic negative bias is maintained proportional to strength of received signals by the following actions. Conduction in agc tube *A* charges capacitor *C<sub>f</sub>* to make its upper end, connected to the agc bus, of negative polarity with reference to ground. Since conduction in tube *A* increases with a stronger signal, the stronger signal acts to increase the charge on capacitor *C<sub>f</sub>*, which makes the automatic bias more negative and lessens amplification in the controlled amplifiers.

The plate of tube *B* is being maintained negative with reference to the tube cathode by the negative peaks of the video signal which are applied to this plate. At the same time, since the cathode of tube *B* is connected to the top of capacitor *C<sub>f</sub>*, increase of negative biasing voltage makes this cathode more and more negative. Finally, the increase of biasing voltage makes the cathode negative with reference to the plate, and the tube becomes conductive. Then current which would cause additional charging of *C<sub>f</sub>* flows instead through tube *B*. Biasing voltage increases no further, but is held at a value proportional to strength of the received signal.

Should strength of the received signal decrease, the plate of tube *B* becomes less negative, and through this tube there is discharge of capacitor *C<sub>f</sub>* to reduce the biasing voltage. If there is increase of signal strength the plate of tube *B* is made more negative, stopping discharge of capacitor *C<sub>f</sub>* until its biasing voltage builds up to a value corresponding to the stronger signal.

**Gain Control During Alignment.** — When frequency response is being observed during alignment of r-f and i-f amplifiers the apparent response will not conform to actual characteristics of the amplifiers if there is any automatic gain control action. For true response characteristics the grid bias of all amplifiers must remain of constant value, unaffected by

signal strength. This might be accomplished in some receivers by keeping the output voltage of the signal generator below the value which causes automatic control of amplifier bias. In receivers having delayed automatic gain control the delay voltage may be enough to permit moderately high input signal voltage without causing automatic change of bias, or it might be possible to temporarily adjust the delay voltage to a value high enough to prevent automatic control.

As a general rule it is quicker, easier, and more certain to substitute for the automatic biasing voltage a temporary fixed bias obtained from one or more dry cells. The negative terminal of the biasing battery should be connected to the agc bus, and the positive terminal connected to ground or to the positive side of the agc capacitor which has been designated as  $C_f$  in several diagrams. A battery thus connected usually will override all effect of the automatic gain control tubes and maintain a steady bias. If a gain control tube is a separate unit, not in the same envelope with the video detector or some other tube, the gain control tube may be removed from its socket while the biasing battery is connected.

If the fixed battery bias is sufficient to actually prevent all agc action there will be no change of shape in the frequency response curve when output voltage of the signal generator is increased or decreased by small amounts. Such changes of generator output should cause the response curve to become higher or lower, but not to change its shape. If there is change of curve shape the bias must be made more negative, or a gain control tube must be made inoperative.

A 3-volt bias from two dry cells in series usually is satisfactory. Adjustable bias may be had with a potentiometer of 5,000 to 10,000 ohms resistance connected across a  $4\frac{1}{2}$ -volt battery. The positive side of the battery, with one end of the potentiometer attached, is connected to chassis ground, and the slider of the potentiometer is connected to the agc bus. Adjustment of the slider then will provide any negative bias between zero and full battery voltage.

**GAIN, STAGE.**—The gain of one stage of amplification usually refers to voltage gain unless power gain is specified. Voltage gain is the number of times a signal voltage is increased between the grid input to one amplifier tube and the grid input to the following amplifier tube, or between the first grid input and the output of the same stage into something like a speaker coupling or a picture tube.

Following are usual formulas for voltage gain of a stage.

$$G_{cin} = \frac{Gm \times Rp \times Ro}{Rp + Ro} \quad \text{Gain} = \frac{\mu \times Ro}{Rp + Ro}$$

*Gm.* Mutual conductance or transconductance, micromhos.

*Rp.* Plate resistance of tube, megohms.

*Ro.* Plate load resistance, megohms.

*mu.* Amplification factor.

All values used in the formulas must be those actually existing with the operating conditions such as plate, screen, and grid voltages, and whatever frequency is being considered. The two formulas are equivalent because *mu*, the first term on the right in the second formula, is equal to  $Gm \times Rp$  in the first formula.

Actual stage gain may be determined as illustrated in Fig. 44-1 from a load line drawn for the actual plate load resistance

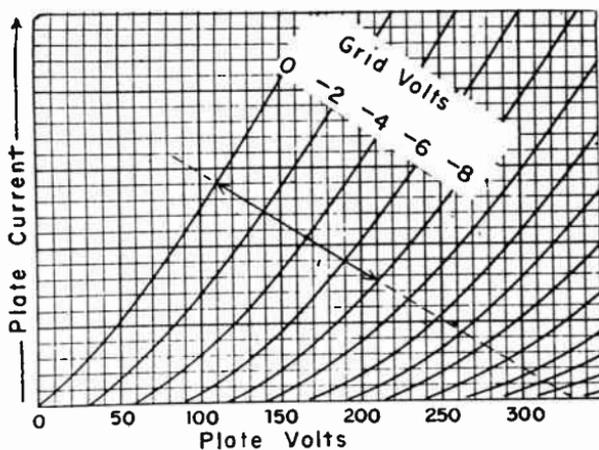


Fig. 44-1.—On the plate and plate load there is change of 100 volts for 8 volts change on the grid, indicating a gain of 12½ times.

on a family of plate characteristics for the tube in use. Points are marked on the load line at the high and low limits of grid voltage or input signal swing above and below the point corresponding to grid bias. Plate voltages corresponding to these limits of grid voltage then are read from the graph. Change of voltage across the plate load may be assumed equal to change of plate voltage. Dividing the number of volts change in plate voltage by the number of volts change in grid voltage gives the voltage amplification of the stage.

Stage gain may be measured with a calibrated signal generator and an electronic voltmeter. The generator must have accurate calibration for its output, usually in numbers of microvolts, and should deliver minimum stray signal voltage at low settings of the attenuator. A check should be made with the generator connected directly to the electronic voltmeter while determining the relation between changes of attenuator settings and resulting meter readings, or while calibrating one of the instruments in relation to the other.

If measurements are to be made in r-f or i-f stages the automatic gain control should be overridden with a fixed bias from a battery. When working on i-f stages the r-f oscillator should be disabled in the same manner as for alignment. The remainder of the process is illustrated by Fig. 44-2.

1. Connect the electronic voltmeter across the output of the last stage in the amplifier system. If connection is to other

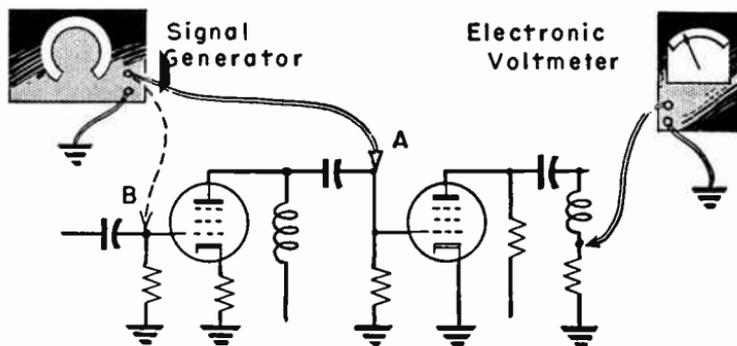


Fig. 44-2.—Connections for measuring voltage gain with signal generator and electronic voltmeter.

than the c-c output of a detector it will be necessary to use a detector probe on the voltmeter input.

2. Use the signal generator unmodulated. For measurements in r-f or i-f stages tune the generator to the amplifier frequency or to the frequency at which gain is to be measured.

3. Connect the output of the generator to the grid of the final amplifier tube, just as when making alignment adjustments. This is the connection at point *A*.

4. Adjust the generator attenuator to produce from the amplifier an output which may be read accurately on the voltmeter, such as 0.5 or 1.0 volt.

The gain, or loss, of the final stage is the quotient of dividing the amplifier output by the generator output.

5. To measure gain of the preceding amplifier stage connect the output of the signal generator to the grid of the tube in the preceding stage, as at *B*.

6. Adjust the generator attenuator to produce from the final stage of the amplifier the same output as in step 4.

7. The gain of this preceding stage is the quotient of dividing the generator output used in step 4 by the generator output used in step 6

**GENERATORS, AUDIO-FREQUENCY.**—Generators which furnish variable and adjustable frequencies in the audio range usually employ the principle of beat frequencies, or of resistance-capacitance tuning, or of relaxation oscillation as found in multivibrators and blocking oscillators. For fixed audio frequencies such as used for modulation in r-f signal generators it is common practice to use some form of feedback oscillator, such as tuned grid or tuned plate with tickler feedback or else a Hartley circuit. Feedback oscillators do not lend themselves to adjustment over such wide ranges of audio frequency as do the other types mentioned.

Service types of audio-frequency generators usually provide for outputs anywhere between about 20 cycles per second and a high limit of 12,000 to 20,000 cycles per second. Beat frequency oscillators contain one r-f oscillator operating at a fixed frequency and another r-f oscillator whose frequency is ad-

justable. The outputs are fed to a mixer or converter from which is taken the difference beat frequency in the audio range. Remaining high frequencies are removed by a low pass filter. The filter often is followed by a degenerative amplifier tube.

Fig. 45-1 illustrates the principle of resistance-capacitance tuning employed in many audio-frequency generators. The operating frequency is selected or adjusted by means of a Wein bridge in which the four arms consist of  $Ca-Ra$  in series,

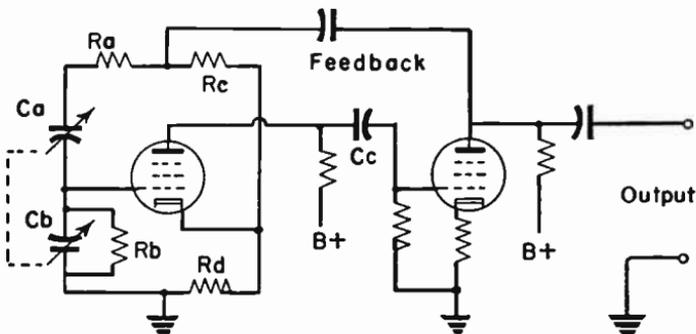


Fig. 45-1.—A Wein bridge frequency-selective circuit as used in audio-frequency generators.

of  $Cb-Rb$  in parallel, of  $Rc$ , and of  $Rd$ . Adjustment of capacitors  $Ca$  and  $Cb$  will vary the frequency through a range of about 10 to 1. Additional ranges are provided by switching in different resistances at  $Ra$  and  $Rb$ .

Plate output of the left-hand tube is fed through capacitor  $Cc$  to the grid of the right-hand tube. Part of the output from the right-hand tube goes to the generator output terminals and part is a feedback. The feedback is applied to the upper input point of the bridge, between  $Ra$  and  $Rc$ , through the feedback capacitor, and is applied to the bottom input point of the bridge through ground. The output of the bridge is applied to the grid of the left-hand tube from between  $Ca$  and  $Cb$ , and is applied to the cathode of this tube from between  $Rc$  and  $Rd$ . At the frequency for which the bridge is balanced or

adjusted the feedback is regenerative, and causes oscillation. At all other frequencies the phase of the feedback is such as to make it degenerative.

**GENERATORS, CRYSTAL CONTROLLED.** — The characteristics of piezo-electric quartz crystals employed for maintaining the frequency of r-f oscillators are explained in the article on *Crystals, Frequency Control*. Fig. 46-1 shows the circuit

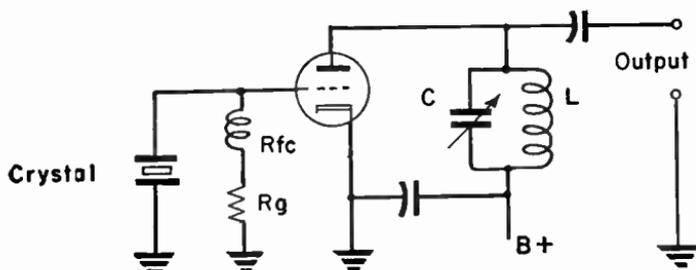


Fig. 46-1.—A simple crystal-controlled oscillator.

for a simple crystal controlled oscillator using a triode voltage amplifier tube. The radio-frequency choke  $R_{fc}$  has inductance of 1 to 3 millihenrys. The grid resistor is of 1 to 5 megohms resistance. Inductance and capacitance at  $L$  and  $C$  are of any values which will tune to the frequency of the crystal. When

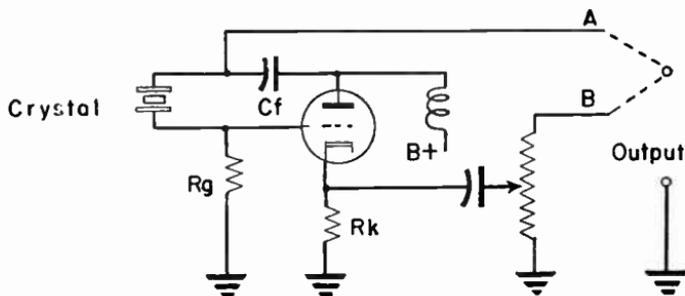


Fig. 46-2.—An untuned crystal-controlled r-f oscillator.

the plate circuit is tuned to a frequency slightly higher than that of the crystal the reactance of the plate circuit is inductive and there is feedback through the plate-grid capacitance of the tube. The feedback sustains oscillation at the crystal frequency.

Fig. 46-2 shows an untuned oscillator circuit which will,

however, sustain oscillation at the frequency of any crystal connected between plate and grid of the tube. This is one variety of the Pierce oscillator, a type used in many r-f generators where the crystal frequency is employed for markers. The feedback capacitance  $C_f$  may be about 500 mmf. The grid leak resistance  $R_g$  may be 0.1 megohm or more. The cathode resistor  $R_k$  is chosen to maintain a suitable grid bias for the tube used.

Output from the oscillator of Fig. 46-2 may be taken from the plate side of the circuit at  $A$  or from the cathode side at  $B$ . An output attenuator is shown for the cathode connection. Some generally similar attenuator would be used for plate output if employed. The oscillator tube may be a pentode instead of the triode illustrated. The only addition would be the screen supply dropping resistor and a screen bypass capacitor to ground.

**GENERATORS, MARKER.** — A marker generator provides means for identifying the frequency at any point along a curve of frequency response being traced on the screen of an oscilloscope with the help of a sweep generator. The marker generator is essentially an oscillator furnishing radio-frequency voltage of fixed frequency which, when introduced into the same receiver circuit with the output of the sweep generator, forms a strong beat voltage every time the frequency of the sweep generator goes through the fixed frequency for which the marker generator is adjusted.

At points on the frequency response curve where marker and sweep frequencies are equal the beat voltage produces an effect usually called a marker "pip," as illustrated at the left in Fig. 47-1. Some marker generators have provision for turning off their oscillator tube. Then, with the tuned circuits of the marker generator remaining connected to the receiver circuit in which are the sweep voltages, there is absorption of power every time the sweep frequency passes through the marker frequency. The result is a break or a "dip" in the frequency response curve, as at the right in Fig. 47-1. Either pips or dips may be used for frequency identification.

Frequency calibration of the marker generator must be highly accurate and stable. Many service operations require frequency settings to within a small fraction of one per cent of a specified value. For any frequencies which are likely to be used many times, as those for receivers most often serviced, it

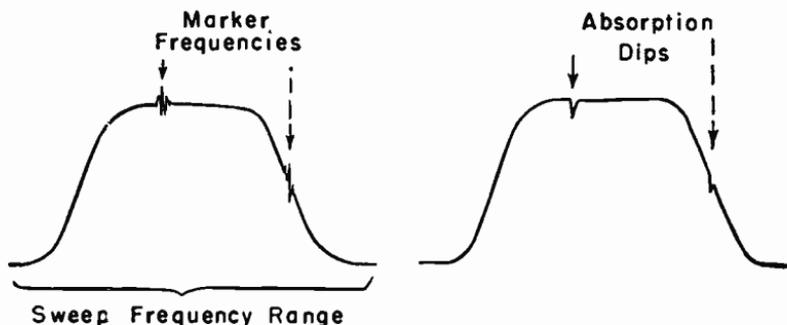


Fig. 47-1.—Marker pips resulting from oscillator action (left) and marker dips resulting from absorption (right).

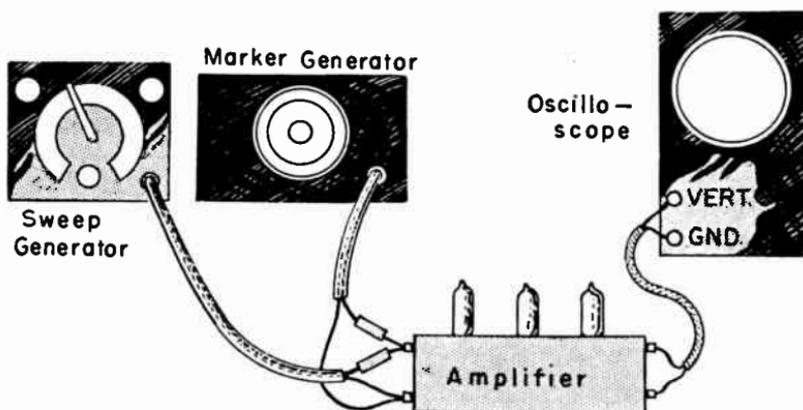
is desirable to have crystal control. It is necessary also that the tuning dial drive and scale of the marker generator be such as will allow resetting to a given frequency with a high degree of accuracy after the generator has been used for some other frequency between times.

Frequency ranges of the marker generator must include all the intermediate frequencies, from 20 to 50 mc. It is necessary also to provide the 4.5 mc intercarrier beat frequency used for intercarrier sound systems. Marker frequencies are not always provided for video and sound carriers and for r-f oscillator frequencies. If carrier and oscillator frequencies are provided there will be a range from 54 through about 125 mc for the low band and from 174 through about 250 mc for the high band. Fundamental frequencies are preferable to harmonic frequencies in all ranges, although harmonics often are used for carrier and oscillator frequencies.

The marker generator nearly always is used also as a general purpose signal generator where an adjustable constant fre-

quency is needed, as for alignment of stagger tuned i-f stages and many other purposes. Such uses require that the r-f output may be modulated with audio frequency, although no modulation is needed for the marker function. It is desirable also that the output attenuator be accurately calibrated in microvolts.

The marker generator and sweep generator may be constructed as a single instrument called a television generator. In such a combination instrument the marker and sweep outputs are internally connected together so that only a single



*Fig. 47-2.—Connections of marker generator, sweep generator, and oscilloscope to amplifier or receiver.*

external cable is used for connection to the receiver. Through this cable are applied the sweep frequency and the adjustable marker frequency, with each frequency regulated by its own controls on the generator panel.

If the marker generator is an instrument separate from the sweep generator, the sweep generator is coupled to the receiver circuits in the usual way. Usual procedure then is to connect the high side of the marker generator through a fixed capacitor of something between 2 and 20 mmf capacitance to the same point as the high side of the sweep generator. This is illus-

trated by Fig. 47-2. The low sides of both generators are connected to chassis ground or to B- of the receiver. Output cables must be shielded, with the shield grounded.

Enough coupling for marker frequencies may be secured with the cable from the marker generator laid near the cable from the sweep generator, or the marker cable may be given two or three turns around the sweep cable with the end of the marker cable left unconnected. The marker need not be connected to the same point as the sweep, but may be coupled to the grid of a tube in a stage either preceding or following that to which the sweep is coupled. In some generators there is an internal fixed capacitor in series with the output. Otherwise an external series capacitor must be used to prevent a d-c short circuit of grid bias through the generator when the sweep output is connected to the grid of an amplifier tube. If a d-c ohmmeter gives a steady resistance reading when connected to the output of the generator there is no series capacitor inside the generator, and an external capacitor should be used. The 2- to 20-mmf coupling capacitor mentioned in the preceding paragraph serves this purpose.

The marker generator employed for producing marker pips or dips is used without modulation or is used with pure r-f output. To identify the frequency at any point on a curve produced by the sweep generator, tune the marker generator to bring the pip or dip to the point in question and read the frequency from the marker dial. When it is desired to alter the response at some certain frequency, tune the marker generator to that frequency and then make changes of alignment or other adjustments as required to obtain the desired response where the marker pip or dip appears on the curve. These methods are used when working with frequency responses of r-f and i-f amplifiers, also when working with S-curves of sound discriminators or ratio detectors. Markers may be used to identify high and low limits of amplification, to identify peak frequencies, trap frequencies, and for other purposes.

Some generators provide two different marker frequencies at the same time. This is an advantage in many cases. For example, when working with the response of the television i-f amplifier section one marker may be set for the video intermediate and the other marker set for the sound intermediate frequency. When adjusting for a desired band pass the two markers may be set for the high and low limits of the band.

In order to avoid distortion of the response curve the output of the marker generator must be kept at the lowest level which gives a recognizable pip. If the marker generator is tuned through the range of sweep frequency the pip will move up on one side of the response curve, across the top, then down on the other side. The height of the pip will be maximum at the top of the curve, where there is maximum amplification, and will decrease to zero at the bottom of the curve where there is no amplification. Marker output must be adjusted according to the part of the curve being checked if the curve is not to be distorted.

It is difficult to see a marker on a steeply sloped portion of a curve. The marker may be located by widening the curve with the horizontal gain control of the scope. Weak markers may be seen more easily if the phasing control of the generator or scope is adjusted to separate the forward and return traces, or the return trace may be blanked if there is provision for blanking on the scope.

If all marker pips appear wide and fuzzy they may be sharpened by connecting across the vertical input of the scope a capacitor of something between 100 and 2,000 mmf capacitance, using the least capacitance that gives satisfactory pips. Resistance of about 10,000 to 20,000 ohms in series with the vertical input lead may help to sharpen the pips. A regular filter probe for the scope, containing a shunting capacitor and series resistance, will sharpen the pips.

Harmonic frequencies from crystal oscillators in a generator or from a separate crystal controlled generator sometimes are used to identify frequency points on r-f response curves, with the sweep generator operating at carrier frequencies and

connected to the receiver antenna terminals. Various crystal harmonics may beat with the r-f oscillator frequency and produce a variety of pips on the response curve. The spurious beats due to the r-f oscillator may be identified by shifting a fine tuning control. The oscillator beat markers will move with the curve, holding their same positions on the curve. Pips not due to oscillator beats will remain stationary on the screen of the scope or on a graph scale used in front of the screen. These pips not due to r-f oscillator beats will not move along with the response curve but will appear at different parts of the curve as the fine tuning control is adjusted.

If the r-f oscillator is not made inoperative during alignment of i-f stages the i-f response curve is likely to carry several pips, only one of which corresponds to the frequency for which the marker generator is tuned. The other pips are due to beating with the r-f oscillator voltage. These spurious pips usually are smaller than the one for the marker frequency. If the generator provides for a dip marker, and is switched from pip to dip, the dip marker will appear at the frequency to which the marker generator is tuned.

**GENERATORS, PATTERN.** — A pattern generator produces on the picture tube screen a number of regularly spaced bars, lines or dots. These patterns are particularly useful for adjustment of horizontal and vertical linearity in the absence of transmitted test patterns, and useful also for checking size, hold, and drive controls, as well as general performance of the receiver. Patterns of white dots or rectangular spots on a dark raster are used during service adjustments of convergence and purity controls of color receivers.

A pattern generator may produce horizontal lines as at *A* of Fig. 48-1, or vertical lines as at *B*. Some produce simultaneous horizontal and vertical lines to form the crosshatch pattern at *C*. Patterns commonly are dark on a light raster, but some instruments produce white on a dark raster. Dots or rectangles for color servicing, as at *D* and *E*, must be white on a dark raster.

## GENERATORS, PATTERN

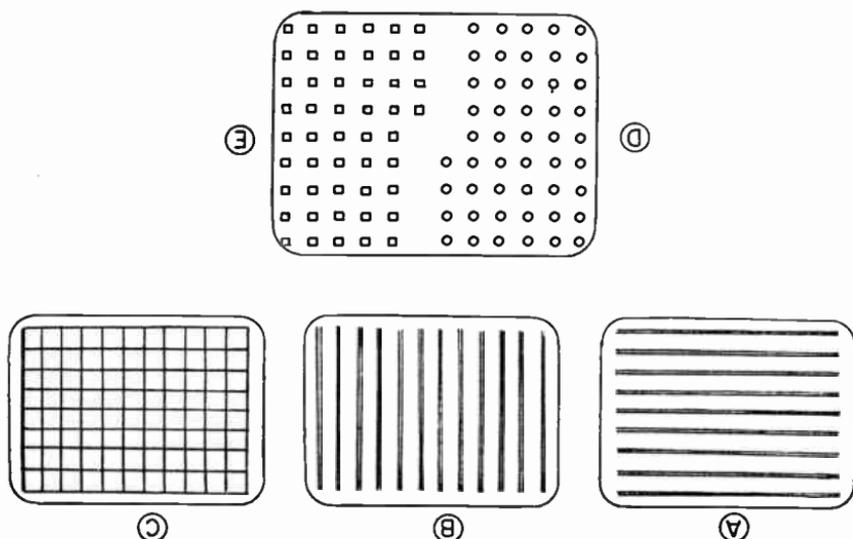


Fig. 48-1.—Patterns produced on the viewing screen of a picture tube by a pattern generator.

Pattern generators contain oscillators operating at multiples of vertical field frequency or of horizontal line frequency, which are respectively 60 cycles and 15.75 kc for monochrome transmission. Were oscillator frequency a multiple of 60 cycles, but less than 15.75 kc, there would be horizontal lines in number equal to oscillator frequency divided by 60. A frequency of 600 cycles would produce 10 horizontal lines. However, some are lost during vertical retrace periods and fewer than this number would be visible. Were oscillator frequency a multiple of 15.75 kc there would be vertical lines in number equal to that frequency divided by 15.75, except for the fact that some of the pattern is lost during horizontal retrace periods.

Most generators have controls for changing, over a limited range, the numbers of horizontal and vertical lines or rows of dots or rectangles. It is customary to employ a horizontal to vertical ratio the same as the standard aspect ratio, which is 4 to 3. For instance, with 12 lines or rows from side to side there would be 9 from top to bottom.

Instruments may be connected either to antenna terminals or to the input of the video amplifier. When connected to antenna terminals, both the generator and the receiver must be tuned for the same channel. Generators usually tune only to low-band channels. It is necessary to adjust receiver hold controls or generator frequency controls to secure a stationary pattern. Contrast and brightness of the receiver are adjusted for satisfactory contrast between pattern and raster background.

**GENERATORS, SQUARE WAVE.** — A square wave generator furnishes an alternating voltage which changes very suddenly between its maximum positive and negative values and remains at each of these values during approximately a half-cycle. Most instruments produce the square wave by generating a sine wave which is shaped or distorted into the square form.

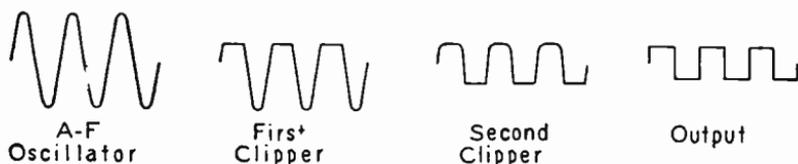


Fig. 50-1.—Steps in changing a sine wave to a square wave.

Typical steps in the formation of a square wave in successive circuits of a generator are illustrated by Fig. 50-1.

Either diodes or triodes may be used for clipping the tops and bottoms of a sine wave. Fig. 50-2 shows connections for one type of twin-diode clipper circuit. The cathode of diode *A* is positively biased from one of the adjustable potentiometers. This diode will conduct only while potential of the positive alternations of the sine wave exceeds the bias voltage. The conduction current flows in resistor *R* and is accompanied by enough voltage loss in this resistor to prevent further rise of current in the diode and, consequently, to flatten off the positive tops of the sine wave. The plate of diode *B* is negatively biased from the other potentiometer. This diode conducts only while potential of the negative peaks of the sine

wave exceeds the bias voltage. Then the conduction current through resistor  $R$  is accompanied by enough voltage drop to flatten the negative peaks of the sine wave. The greater the resistance at  $R$  the more squarely the wave will be clipped, or

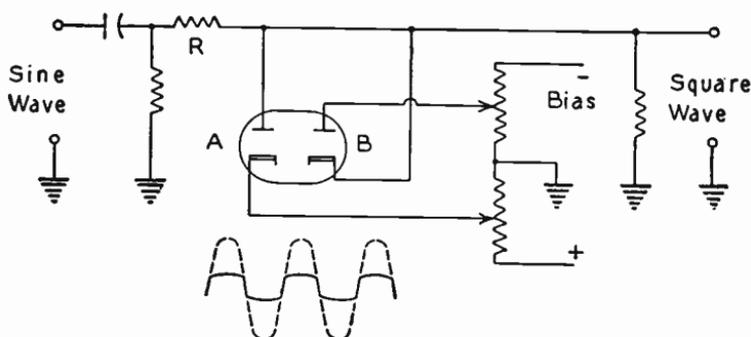


Fig. 50-2.—Square wave clipping circuit with diodes in parallel.

the flatter will be the tops of the squared wave. Squaring is improved also by using small resistance in the biasing potentiometers.

The twin-diode clipping circuit of Fig. 50-3 gives sharper squaring of the output wave and operates well at higher frequencies, but is somewhat more difficult to adjust for satisfactory performance than the circuit of Fig. 50-2. The two diodes are in series with each other and are opposed, cathode to cathode, between the sine wave input and the square wave output. Both cathodes are negatively biased to maintain conduction in both diodes until the bias voltage is overcome by input potentials during positive and negative alternations of the sine wave.

Positive alternations of the sine wave make the plate of diode  $A$  and the cathode of diode  $B$  more positive. This allows continued conduction in  $A$ , but stops it in  $B$  during the entire period in which positive sine wave potential exceeds the bias voltage. Negative alternations of the sine wave make the plate of diode  $A$  and the cathode of  $B$  more negative. This allows

conduction to continue through *B*, but cuts it off in *A* during the period in which negative sine wave potential exceeds the bias voltage.

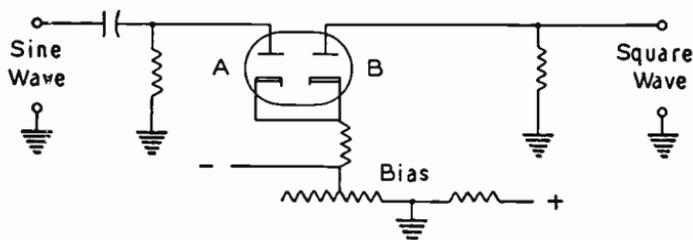


Fig. 50-3.—Square wave clipping circuit with diodes in series.

The chief difference between the two diode clipping circuits is that conduction is limited but not stopped by voltage drop in resistor *R* of Fig. 50-2, while conduction is stopped completely by interaction of sine wave and bias potentials in Fig. 50-3. Naturally, with complete stoppage of current and of output voltage the tops of the square wave will be straight across.

Triode clipping circuits utilize either of two general methods. With one method the negative peaks of the input sine wave are clipped squarely by biasing the triode grid so far negative as to cause plate current cutoff shortly after the sine wave commences to go negative. The output from the plate circuit of this first clipper is applied as grid input to a second similarly biased triode. Because of signal polarity inversion which occurs between grid and plate of any tube, the second clipper squares off the sine wave peaks which originally were on the positive side but have been made negative at the grid of the second clipper. Other triode square wave circuits employ the general method of using such resistance-capacitance constants in grid circuits as will cause plate currents and voltages to appear as sharp pulses which, by further filtering and amplification, become square waves.

**Uses of Square Waves.** — Square wave voltages are used as input signals for audio-frequency and wide-band amplifiers which are to be tested for frequency response and for distor-

tion. The output of the amplifier is connected to the vertical input of an oscilloscope. Using the square wave input allows rapid testing because a square wave really consists of a great number of harmonically related frequencies acting together.

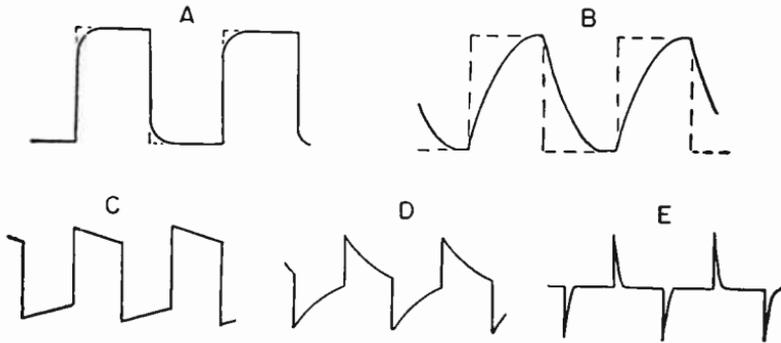
The sharp corners at the leading edges of the square wave are due to frequencies 10 to 100 and more times the fundamental frequency. The flat top and flat bottom of the square wave represent no voltage change, or zero frequency, during the intervals of time occupied by these portions of the wave. Applying a square wave as a signal input is equivalent to making separate tests with a great number of sine wave frequencies. If an amplifier reproduces a square wave without serious distortion, the response of the amplifier usually will be satisfactory at all frequencies from that of the square wave fundamental up to frequencies at least ten times as high.

Any amplifier should be tested with at least two fundamental square wave frequencies. One should be the lowest frequency at which the amplifier is expected to operate satisfactorily. The other should be a frequency of about one-tenth the maximum at which the amplifier is expected to have good performance.

A preliminary check should be made with the output of the square wave generator connected directly to the vertical input of the oscilloscope. The generator should be tuned successively to each fundamental frequency which will be used later for testing the amplifier. At each of these frequencies the internal sweep of the scope should be adjusted to bring two or three cycles onto the screen. If the trace is not a true square wave the fault may be in the generator, in the scope, or in both places. Often it will be possible to secure a good square wave by trying different sweep frequency ranges on the scope, even though this changes the number of cycles traced, and by trying different settings of the vertical gain control. Often there will be improvement of the square wave within some limited range of settings for the generator output attenuator. Unless a square wave with very little distortion can be secured with this direct connection there is little use of checking the

amplifier. When a good square wave is obtained with certain settings of the controls, the amplifier tests should be carried out with the same settings.

If the output trace shown by the oscilloscope has rounded corners at the leading edges of the square wave, as shown at *A* in Fig. 50-4, the amplifier or other circuit being tested is not passing the highest frequencies in the square wave. This usually would indicate poor response at frequencies not much



**Fig. 50-4.**—Faults indicated by output traces with square wave input. *A*, lack of high frequencies. *B*, extreme high-frequency loss. *C*, moderate low-frequency loss. *D*, serious low-frequency loss. *E*, no low-frequency gain.

higher than 10 times the square wave fundamental. The trace at *B* corresponds to an extreme case of high frequency loss or attenuation. Excessive capacitance shunted across the tested unit will cause such a trace, as will also excessive inductance in series with the signal-carrying paths.

The lower row of traces in Fig. 50-4 show the effects of poor response, lack of gain, or excessive attenuation of low frequencies or of frequencies down toward the fundamental of the square wave. A moderate degree of low-frequency distortion is indicated at *C*, a very serious loss is indicated at *D*, and at *E* the lows are wholly lacking and only the highest frequencies are being reproduced or transmitted by the circuit on test. This general class of square wave distortion indicates too much series capacitance or else too much shunted inductance somewhere between the square wave input and the output connected to the oscilloscope.

At the top of Fig. 50-5 are shown traces illustrating the effects of phase shift in the amplifier or other circuit between the square wave input and the output to the oscilloscope. At *A* the phase is leading, with current leading the voltage or voltage lagging the current, as when a circuit has excessive capacitive reactance. In an amplifier this, naturally, gives an indication somewhat similar to that from poor response at low frequencies, as at *C* and *D* of Fig. 50-4. It might result from too small coupling capacitance in a resistance-coupled amplifier used at low frequencies. At *B* is shown the opposite effect,

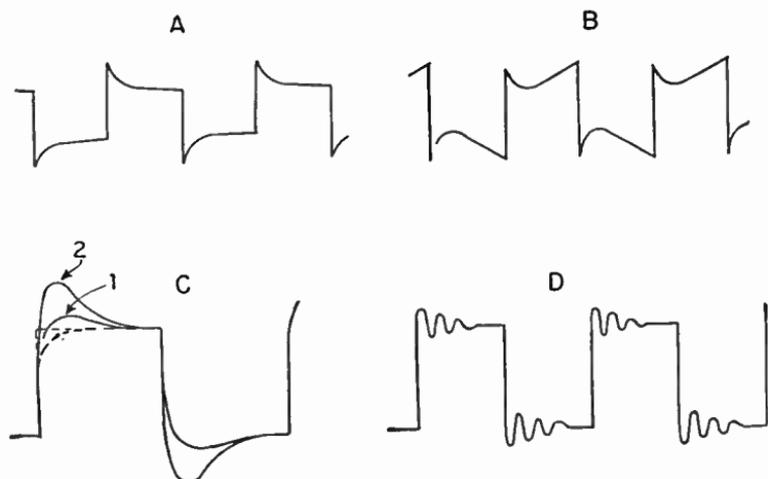


Fig. 50-5.—Faults indicated by traces from square waves. *A*, phase distortion with current leading. *B*, phase distortion with current lagging. *C*, effects of frequency compensation. *D*, oscillation in amplifier.

which results from lagging phase, with current lagging the voltage. Except for the sharp peak on the leading edges this phase shift trace bears considerable resemblance to the one indicating poor response at high frequencies as shown at *A* in Fig. 50-4.

The trace marked *C* in Fig. 50-5 shows the effects of adding inductance in the plate circuit of an amplifier to maintain or increase the gain at the high frequencies. Such inductance is supplied by peaking coils used for compensation in broad band video amplifiers. The broken line curve at the top corner

of the first leading edge would correspond to the trace at *A* in Fig. 50-4, indicating lack of high-frequency response. Adding plate circuit inductance will bring the curve up to 1, and adding still more inductance will bring it up to 2.

At *D* in Fig. 50-5 is a trace which indicates oscillation in the amplifier being tested. Following each leading edge is a train of damped oscillations. These result from circuit inductance and distributed or stray capacitance which are resonant at some frequency in the square wave. It is possible to determine the approximate frequency of oscillation by counting the number of oscillatory cycles in a half-cycle period or in some other fraction of the square wave. The square wave period is known from the generator tuned frequency. The oscillation frequency may give some indication as to the part of the amplifier circuit causing the trouble.

**GENERATORS, SWEEP.** — A sweep generator furnishes a signal voltage whose frequency varies continually and at a constant rate above and below some center frequency for which the generator is adjusted or tuned. The generator may be adjusted to operate above and below the center frequency of any television channel, or the center of an intermediate-frequency band of any television receiver. When correctly adjusted, the generator output frequency will continually sweep back and forth through the entire range of frequencies included in the television channel or through the entire *if* range of the receiver.

A sweep generator is used in conjunction with an oscilloscope for observing simultaneously the relative gains of an amplifier at all frequencies in the range or band being covered or swept by the generator. The method of use is illustrated by Fig. 51-1. The output of the generator is applied to the input of the amplifier whose gain is to be observed. The output of the amplifier is connected to the vertical input of the oscilloscope.

The oscilloscope beam is deflected horizontally at a rate which is exactly the same as the rate at which generator frequency is shifted through its sweep range. Consequently, the

oscilloscope beam always arrives at the same horizontal position for any given frequency during each sweep. At each horizontal position, and at the frequency corresponding to that position, vertical deflection or height of the beam and the

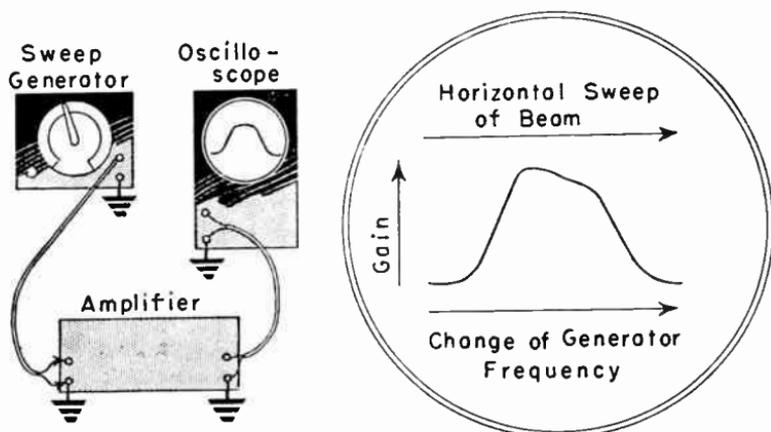


Fig. 51-1.—Connections for a sweep generator and features of the oscilloscope trace.

trace is proportional to the voltage gain of the amplifier at that particular frequency. As the beam moves horizontally across the oscilloscope screen it traces a curve of relative gains at all the frequencies within the swept range. This curve is the frequency response of the amplifier.

Sweep generators are used during alignment of antenna couplers and r-f amplifier stages in television receivers, also during alignment of video and video-sound i-f amplifiers, and in the sound section for alignment of i-f amplifiers, drivers, sound detectors, and discriminators. A television sweep generator is generally similar to f-m generators used for alignment of f-m sound receivers, except that the television type furnishes a wider range of swept frequencies and tunes in the higher carrier and intermediate frequencies of the television bands.

**Methods of Frequency Sweep.** — Frequency is shifted back and forth by varying the tuning capacitance or tuning inductance of an oscillator which is part of the sweep generator. One method makes use of a variable capacitor whose shaft

carrying the rotor plates is continually revolved by a small electric motor built into the sweep generator. Capacitance varies between maximum and minimum as the rotor plates are turned into and out of mesh with the stator plates of the capacitor. This rotary capacitor is connected in parallel with the main tuning capacitor or with the tuning inductance of the oscillator. Change of capacitance increases and decreases the oscillator frequency at a rate depending on the speed of the driving motor. Motor-driven capacitors were used in most of the early types of sweep generators, but less costly methods are now more generally employed.

Many sweep generators secure their frequency shift by means of a reactance tube operating quite similarly to such tubes used in some kinds of automatic frequency control for television sweep oscillators. In the sweep generator connections shown by Fig. 51-2 the plate of the reactance tube is connected to one end of the oscillator coil. High-frequency currents in the plate circuit of the reactance tube is out of phase with oscillator voltage. Connections to the reactance tube may be such that reactance plate current either lags or

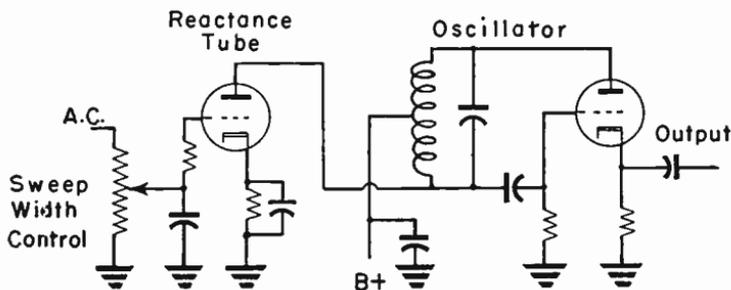


Fig. 51-2.—A reactance tube employed for frequency sweeping.

leads the oscillator voltage. With a lagging reactance current applied to the tuned circuit of the oscillator the effect is equivalent to adding inductance in the oscillator circuit. This lowers the oscillator frequency. If reactance current leads the oscillator voltage the effect is equivalent to adding capacitance in the oscillator tuned circuit, which also decreases the oscillator frequency.

How much the oscillator frequency is decreased is proportional to reactance plate current, the greater the current the greater the shift of frequency. Oscillator frequency may be thus decreased, and allowed to again increase, at a rate of any number of times per second by increasing and decreasing the reactance plate current at that desired rate of frequency sweep.

Both the extent of frequency sweep and the number of sweeps per second are controlled by suitably varying the grid voltage of the reactance tube. In Fig. 51-2 the reactance tube grid is connected through a potentiometer either to the a-c power line or else to a tube heater circuit. Grid voltage then will vary at power line frequency, usually 60 times per second, and oscillator frequency will vary or sweep at the same rate. The extent of the alternating swings of reactance grid voltage depends on adjustment of the slider on the sweep width control. This control accordingly regulates the extent of change of oscillator frequency or the width of frequency sweep.

Another method of frequency sweep, illustrated in principle by Fig. 51-3, makes use of the fact that moving a piece of metal toward and away from the inductor of a tuned circuit alters the effective inductance of that circuit, and there are

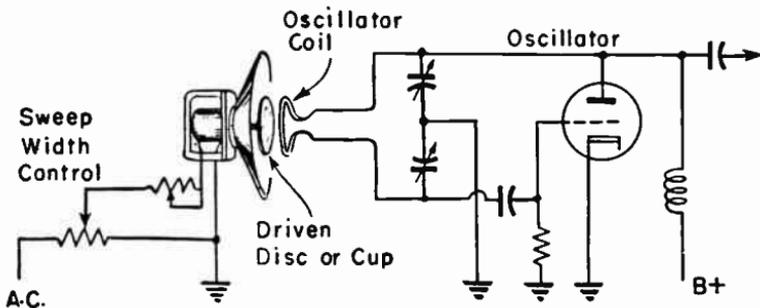


Fig. 51-3.—Electromechanical or electrodynamic method of frequency sweeping.

accompanying changes in resonant frequency of the circuit. The coil or inductor for the oscillator is supported in a fixed position. A fraction of an inch from the oscillator inductor is a metallic disc, plate, or cup which is attached to the moving

coil and cone of a small permanent-magnet speaker. The speaker is driven at power line frequency by taking current for its voice coil from an a-c heater circuit or other equivalent source.

The metal which is near the oscillator inductor is moved toward and away from the inductor at a rate corresponding to power line frequency. Consequently, the oscillator frequency is caused to increase and decrease at power line frequency, and this is the rate of frequency sweep for the generator.

The extent of frequency change is proportional to the distance through which the piece of metal is moved as it approaches and recedes from the oscillator inductor. The greater the movement of the metal the greater is the sweep. This movement, and the resulting range through which oscillator frequency is shifted, are varied by altering the current in the voice coil of the speaker. The current and the width of sweep are controlled by means of the potentiometer or rheostat connected to the voice coil.

**Width of Sweep.** — For television r-f and i-f alignment it should be possible to obtain maximum sweep width of at least 10 megacycles and preferably as much as 15 to 20 megacycles. This refers to the overall sweep, from minimum to maximum frequency, with part of this sweep range above the center frequency and part below. In order that a sweep generator may be used also for alignment of television sound sections and for alignment of f-m sound receivers it is desirable that the minimum sweep width be no more than two or three megacycles.

When a generator is in use, the sweep width control should be adjusted to bring onto the screen of the oscilloscope the entire gain curve or frequency response curve, plus a small extension of the zero gain or base line on both sides. The result of such adjustment is illustrated at the center of Fig. 51-4. The amplifier being tested has more or less gain over a total range of about 8 mc, as shown by the fact that the curve rises from zero and falls back to zero within this range. By making the sweep width approximately 10 mc it is possible to view the

entire curve, and to be sure of seeing all the curve by displaying a small amount of zero base line at both ends of the curve.

At the left in Fig. 51-4 is shown the result of using too much sweep width. The response curve then is cramped into a small part of the entire trace, with too much of the trace used for

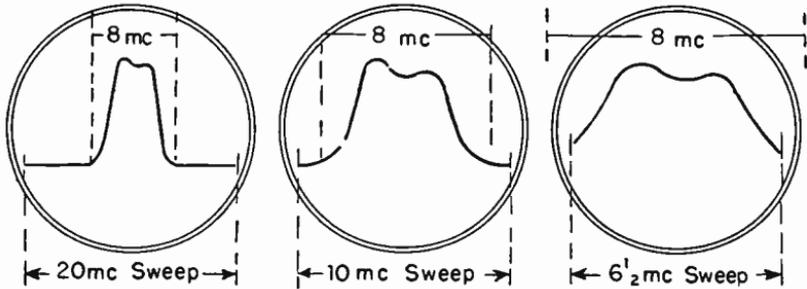


Fig. 51-4.—Effects of varying a sweep width control.

the meaningless base line or line of zero gain. The gain curve is too narrow to allow critical inspection of its shape and characteristics. At the right is shown the result of too little sweep width. Only a portion of the gain curve or frequency response can be examined. The rule is to use the least width of sweep which will display both ends of the frequency response all the way down to the zero line. Sweep width should not be altered during adjustment of any one piece of apparatus, since a change in curve width alters the appearance enough to cause some confusion.

**Positioning of Response Curve.** — The same frequency response or same relative gains at various frequencies may be shown by the same curve in any of four different positions on the screen of the oscilloscope. The positions are shown by Fig. 51-5. At *A* there is increase of frequency from left to right and increase of gain from the base line upward. At *B* the frequency still increases from left to right but the increase of gain is downward from the base line. At *C* there is increase of frequency from right to left, which is a reversal of the frequency direction, and the gain increases upward. At *D* the frequency increase remains from right to left, but increase of gain is downward.

If the oscilloscope beam is deflected horizontally by the internal sweep of the scope, the direction of frequency increase depends on the design of the internal sweep circuit, on the direction in which there is the relatively slow rise of sawtooth voltage which causes the useful trace, not the retrace.

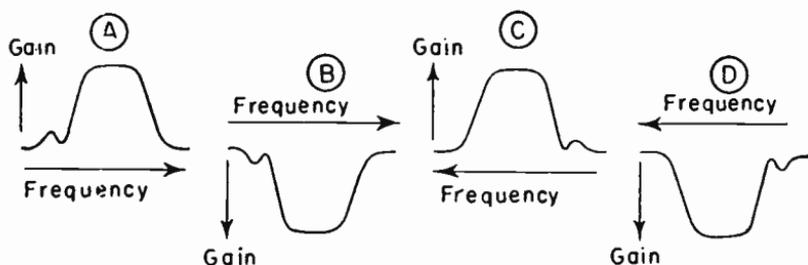


Fig. 51-5.—The same curve of frequency response may appear in different positions on the oscilloscope screen.

In practically all service oscilloscopes the trace is from left to right and there will be increase of frequency from left to right.

Usually the horizontal deflection of the oscilloscope beam is caused by a synchronized sweep voltage supplied from the sweep generator and fed to the horizontal input of the scope. Then the direction of frequency increase depends on the phase relation of this synchronized sweep voltage to the voltage which varies the frequency of the oscillator in the sweep generator. Synchronized sweep voltages from sweep generators usually are in such phase relation as to cause frequency to increase from left to right on the oscilloscope screen.

Whether there is increase of voltage gain upward or downward depends, for any given oscilloscope, on the polarity of the signal at the amplifier output or at the point connected to the vertical input of the scope. Each amplifier stage inverts the signal polarity, there are differences in numbers of stages in various receivers, and the response curve is just about as likely to show increase of gain in one direction as the other.

Any response curve may be moved bodily in a horizontal direction across the screen of the oscilloscope by changing the center frequency to which the sweep generator is tuned. At

the left in Fig. 51-6 the sweep generator is tuned to a center frequency which is too low to bring the entire response curve onto the screen. Increasing the center frequency will bring the entire curve onto the screen, as in the middle diagram.

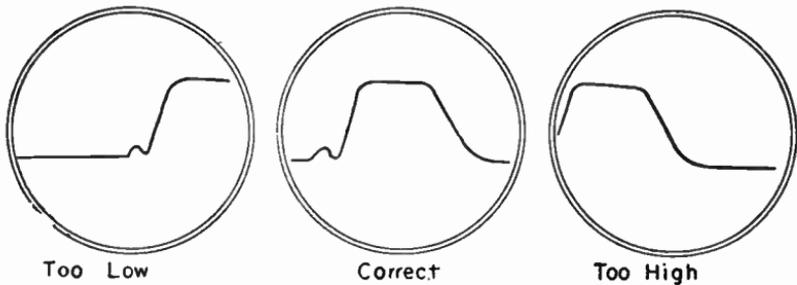
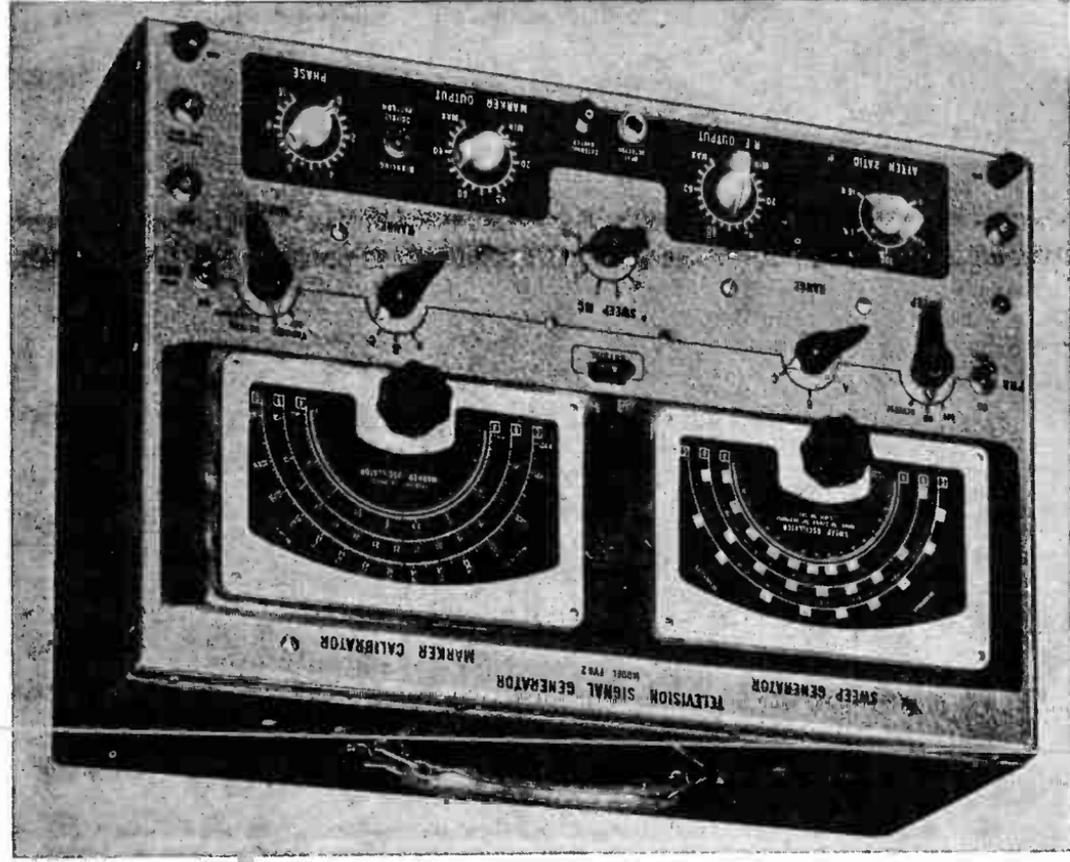


Fig. 51-6.—Effects of altering the center frequency of the sweep generator.

Further increase of center frequency will move the response curve off the screen in the opposite direction, as shown at the right.

**Synchronizing the Sweep.**— In order that a given point measured horizontally across the response curve may always represent the same frequency it is necessary that the horizontal sweep of the oscilloscope beam occur precisely in step with the sweep or the change of frequency of the generator. The beam may be deflected horizontally either by the internal sweep voltage of the oscilloscope or else by a synchronized sweep voltage supplied from the sweep generator. While the beam is being deflected, the frequency furnished by the sweep generator will be changing. If frequency sweep is at the rate of 60 times per second the frequency will go from minimum to maximum and back to minimum during  $1/60$  second. Another way of saying the same thing is to state that frequency will go from minimum to maximum during  $1/120$  second, then from maximum back to minimum in the following  $1/120$  second, and will continue to change at this rate.

In order to deal with some concrete values we may assume that the generator frequency sweeps back and forth between 20 and 30 megacycles, going from 20 to 30 mc during  $1/120$



A modern type TV Signal Generator

second and from 30 back to 20 mc in the following 1/120 second. If the internal sweep of the oscilloscope is adjusted for a 60 cycle sweep, the beam will travel all the way across the screen in 1/60 second and the trace will show two response curves as at *A* in Fig. 51-7. One curve will show the change of gain as frequency increases from 20 mc to 30 mc, and the other curve will show the change of gain as frequency decreases

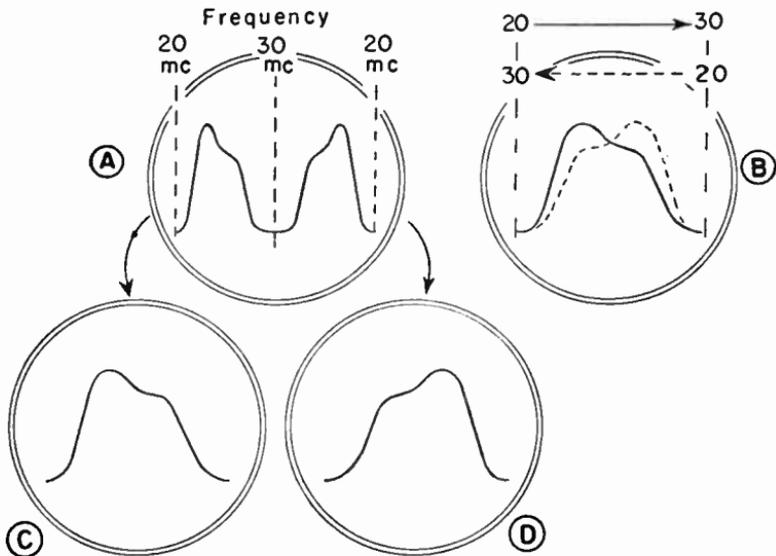


Fig. 51-7.—Response traces produced by using the internal horizontal sweep of the oscilloscope.

from 30 mc to 20 mc. The two curves will be symmetrical, and reversed left to right.

If the internal sweep of the oscilloscope is adjusted to 120 cycles per second the frequency trace will be split and the two parts will be superimposed as shown at *B*. Although this permits widening the response by using the horizontal gain control of the scope, the curves usually will be difficult to identify or to view with enough separation to make an analysis of the frequency characteristic.

A better way is to increase the horizontal gain of the oscilloscope until one of the two curves will nearly fill the screen, and

to operate the horizontal centering control of the scope to center that curve on the screen. Then it is possible to view the curve of increasing frequency as at *C* in Fig. 51-7 or to view the curve of decreasing frequency as at *D*. These separated curves would be secured with the internal sweep of the scope adjusted for 60 cycles, or for whatever may be the sweep rate of the generator.

Instead of using the internal horizontal sweep of the oscilloscope it is more common practice to sweep the beam horizontally by means of the same alternating voltage that shifts the generator frequency. The alternating voltage at power

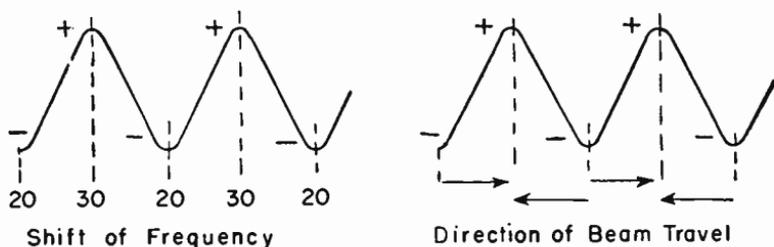


Fig. 51-8.—Relations between changes of frequency and directions of sweep of the oscilloscope beam when using synchronized sweep voltage.

line frequency used for frequency shift in the generator, as in Figs. 51-2 and 51-3, is fed also to the horizontal input of the oscilloscope. Then the oscilloscope beam is swept horizontally at a rate precisely the same as the rate of frequency shift. This principle is illustrated by Fig. 51-8. As the alternating voltage changes from negative to positive it shifts the frequency from 20 to 30 mc in the generator and at the same time deflects the oscilloscope beam to the right. While the control voltage changes from positive back to negative it shifts the generator frequency from 30 to 20 mc and deflects the beam to the left.

Resulting traces on the oscilloscope screen are shown by Fig. 51-9. At *A* is the trace formed while frequency shifts from 20 to 30 mc with the beam moving from left to right. At *B* is the trace formed while frequency shifts from 30 back to 20 mc with the beam moving from right to left. Because the impression of each luminous trace remains visible while the other

one is formed, both traces appear together as in diagram *C*. The two traces now are formed in the same direction with respect to frequency, not in reversed directions as at *B* in Fig. 51-7.

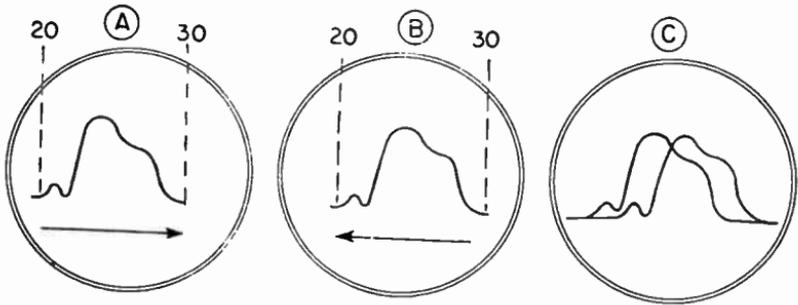


Fig. 51-9.—Traces formed by travel of the oscilloscope beam toward the left and toward the right with synchronized sweep voltage.

Because of the effects of various inductances and capacitances in the frequency shifting circuits of the generator the alternating power line voltage fed to the horizontal input of the oscilloscope will not be exactly in phase with the frequency shift. This is the reason for the displacement of the two traces at *C* in Fig. 51-9. It is desirable that the two traces be brought into register or superimposed to make them appear as a single trace. This requires a phase shifting circuit for the horizontal sweep voltage being fed to the horizontal input of the oscilloscope. \*

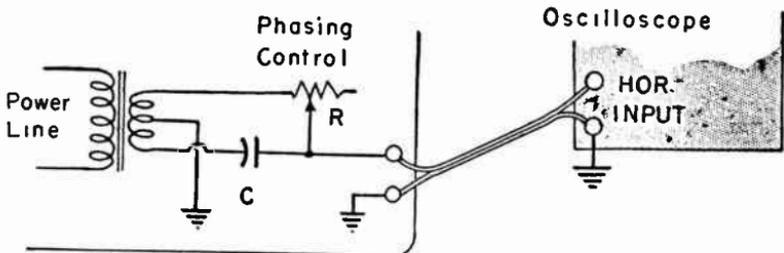


Fig. 51-10.—A phase shifting circuit for synchronized sweep voltage.

A typical method of phase shift or phase control is illustrated by Fig. 51-10. This control circuit may be in either the

sweep generator or oscilloscope, but most often is in the generator. The connection for the horizontal input of the oscilloscope is taken from between capacitor  $C$  and adjustable resistor  $R$  which are in series with each other across the ends of a center tapped transformer secondary winding. Moving the slider of the phasing control causes one trace to move toward the left while the other trace moves toward the right. There is one position of the control with which the two traces come together and appear as a single trace. The secondary winding of the control transformer usually furnishes about 6 volts from end to end. Resistance at  $R$  may be almost anything between 0.1 and 0.5 megohm. Capacitance at  $C$  may be something between 0.01 and 0.25 mfd.

**Frequency Ranges.** — The center frequency of the sweep generator should be adjustable throughout the video and sound intermediate frequency ranges used in television receivers. This requires continuously variable tuning from about 20 through 40 megacycles, and for the intercarrier beat frequency employed in many sound sections there must be sweep centered at 4.5 megacycles. Tuning through the carriers in the low band of the very-high frequency television channels requires continuous tuning between 54 and 88 megacycles, and the high band requires tuning from 174 to 216 megacycles.

Sweep frequency dial calibration need not be very accurate because, as explained in connection with Fig. 51-6, the tuning is varied to center the response curve regardless of what frequency may be shown on the tuning dial.

Fig. 51-11 illustrates principles commonly employed for producing sweep frequencies in the various ranges. At the left in this diagram is the sweep oscillator whose frequency is caused to continually shift back and forth through the sweep width by methods illustrated in Figs. 51-2 and 51-3, or anything equivalent in results. The center frequency of this oscillator remains fixed for all ranges or bands. Next toward the right is a variable oscillator whose frequency may be varied throughout the change required in any one range. The fre-

quency of this oscillator is adjusted by variable capacitor  $C$  whose control or dial is on the generator front panel.

Outputs of the two oscillators are fed to the mixer tube at the right where the two oscillator frequencies beat to produce difference frequencies and add to produce sum frequencies in

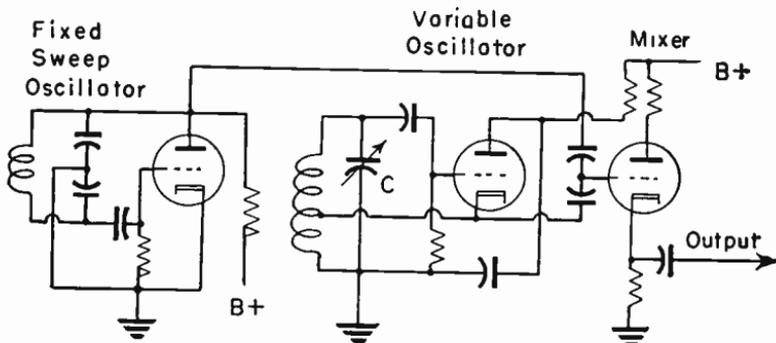


Fig. 51-11.—Oscillators for producing sweep frequencies which are centered at various values.

the mixer output. The mixer output goes to the attenuator of the generator, possibly with some amplifier or coupler tube between the mixer and the attenuator. Some sweep generators are designed to cover all required frequencies by utilizing the difference frequencies, the sum frequencies, and also harmonics of either or both these frequencies. Other instruments provide separate tuning coils for the variable oscillator in such ranges of output frequency as will avoid the use of harmonics which fall within one band while another band is being used for alignment. Still other designs use a beat frequency or difference frequency for only the intermediate-frequency ranges where relatively low sweep frequencies must be provided, and vary the tuning of the sweep oscillator to provide outputs in the carrier-frequency ranges. For this latter method the two capacitors in the tuned circuit of the sweep oscillator in Fig. 51-11 would be variable, operated by a panel control.

**Output Voltage.**—Maximum output voltage from the sweep generator must be enough to produce on the oscilloscope screen a response curve of useful height even when the

amplifier being tested has but small voltage gain. The fraction of a volt which will satisfy this requirement depends largely on the sensitivity of the oscilloscope, but at least 0.1 volt from the generator usually will be needed.

Output voltage should remain within a few per cent of a constant value throughout any sweep of frequency. For example, if the sweep is from 20 to 30 mc there should be practically no variation of voltage during this sweep. Any humps or valleys will be amplified when observing a response curve, and the curve will be so distorted as to be useless for alignment work. Provided the oscilloscope has high vertical sensitivity, the flatness of the generator output voltage may be checked by connecting the generator output to the vertical input of the scope and observing whether the trace is straight or wavy. To make this test it is necessary to use a detector probe on the vertical input of the scope, since the signal from the generator will be at high frequency. The trace should be observed in different frequency bands, with different settings of the attenuator, and different widths of sweep. The generator should be used for alignment work only at settings and adjustments with which the trace is a nearly straight line.

**Coupling Sweep Generator To Receiver.** — For alignment of tuner circuits the output of the sweep generator is connected to the receiver or tuner antenna terminals. This connection is used also for observing overall frequency response, from antenna through the video detector. The high side of the generator output cable may be directly connected to either antenna terminal and the low side or ground side to the other antenna terminal. Otherwise the low side of the generator cable may be connected to chassis ground or to B-minus, and the high side to either one of the antenna terminals.

With the sweep connected to receiver or tuner antenna terminals the generator must sweep frequencies of the channel to which the receiver is tuned during alignment. At these very-high carrier frequencies it is highly desirable that the generator output cable be terminated, at the receiver end,

with a matching pad whose resistance is close to the characteristic impedance of coaxial cable used for generator output. Such matching pads are described in this book.

For alignment of i-f amplifier stages the high side of the generator output cable must be connected through capacitance to the mixer tube or to one of the i-f amplifier tubes, depending on what stages are to be aligned. Capacitive connection is necessary to prevent shorting the d-c grid bias of the tube. A fixed capacitor of 10 to 50 mmf usually is satisfactory, or the coupling may be with a ring or shield over the glass of the tube, as described in the article on *Alignment*.

Sweep voltage to receiver circuits must be low enough to avoid distortion of response curves. Should small changes of generator output alter the shape of the response, in addition to merely changing the height or amplitude, sweep input is too strong. If sweep voltage cannot be sufficiently reduced by the generator attenuator, a smaller coupling capacitor may be used, or a coupling ring moved higher on the tube envelope.

Many sweep generators have special features adding to convenience of operation. There may be provision in the generator for blanking return traces on the oscilloscope screen by cutting off either the positive or negative swing of each cycle of sweep voltage. Combination sweep and marker generators allow reducing the number of cable connections.

**GROUNDING.** — In receivers operating at frequencies no higher than those in the standard broadcast band it is assumed that all points in the chassis metal are at the same r-f potential, and that all parts directly connected to chassis metal will be at the same r-f potential even though these parts are some distance from one another. In apparatus operating at television carrier and intermediate frequencies it cannot be assumed that there is no r-f potential difference between various points on the chassis metal when high-frequency circuits are completed through this metal. The effects of capacitance, inductance, and high-frequency losses in general are equivalent to enough resistance to cause regenerative and degenerative feedbacks from plate circuits to grid circuits whose returns are through

any considerable length of chassis metal or other metallis supports. Such troubles could be lessened by using such low resistance metals as copper and aluminum alloys, but this construction is seldom economically possible in commercial products.

R-f bypass and decoupling capacitors should be mounted as close as possible to socket lugs or base pins of tubes to which these capacitors are connected. The leads from capacitors to ground should be as short as possible. When decoupling capacitors are used for the plate, screen, and grid of one tube all these capacitors should be connected to the same point on the chassis metal. This point should be the one to which the suppressor and a directly grounded cathode are connected. Then none of the element returns follow a common path through chassis metal. Ground connections for rotors of variable tuning capacitors should go to the same point at which a grid decoupling or bypass capacitor is connected.

When using paper capacitors it is essential that connections from chassis ground or from B-minus be made to the pigtail or other terminal that connects to the outside foil. This terminal is identified by words or by a printed band at its end of the capacitor.

Where center terminals of miniature tube sockets are grounded, and have various ground connections made to them, this arrangement should not be altered during servicing operations. When one heater pin of a single-ended tube is grounded to complete the a-c heater circuit, this grounded pin should be the one nearest the grid pin of that tube.

It is inadvisable to change the grounding points on a chassis, or to change the length or arrangement of wires coming to these grounds. Grounding of any part by merely bolting or screwing its frame tightly to chassis metal is not dependable in high-frequency circuits. A grounding wire or flexible metallic braid should be soldered to both the frame and the chassis metal to insure a permanent and noise-free connection. It is easy to make soldered connections directly to the metal of a steel chassis which is coated with cadmium, zinc, or copper.

Several grounding wires may be soldered to the same point on the chassis when all the wires come from nearby parts. Rather than using long grounding wires from widely separated parts it is better to make additional soldered connections to the chassis with short wires to each connection.

It must be kept in mind that chassis metal or chassis ground is not necessarily the most negative point in the B-power circuits. The most negative point is that to which is connected the plate or the plate return of the power rectifier and high-voltage winding of the power transformer. Chassis ground may be at a potential much more positive. In receivers having series heaters the chassis ordinarily is not at B-minus potential nor is it a ground so far as the B-power circuits are concerned.

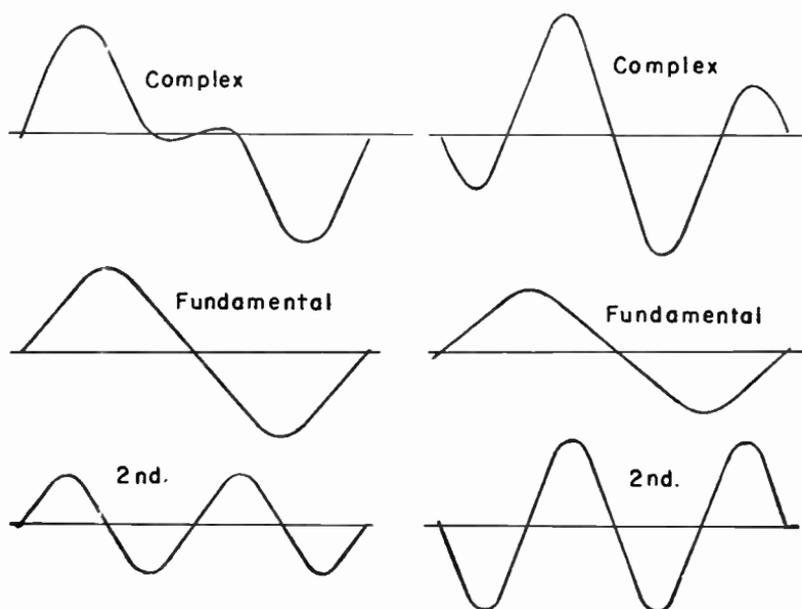
## H

**HARMONICS.** — A harmonic is a frequency which is an even or odd number of times another frequency which is called the fundamental. For example, with a fundamental frequency of 5 mc, the harmonics will be 10 mc, 15 mc, 20 mc, 25 mc and so on for still higher multiples of 5. The frequency which is twice the fundamental is the second harmonic, the one which is three times the fundamental is the third harmonic, and the fourth, fifth, and higher harmonics are respectively four, five, and higher numbers of times the fundamental. The fundamental frequency sometimes is called the first harmonic.

At the top of Fig. 54-1 are two complex waves, or two waves which are not of the same form in the first and second alternations or on positive and negative sides. Each wave is produced by combining in one circuit the fundamental and the second harmonic shown directly below. Both the fundamental and the harmonic are sine waves. Differences between the two complex waves result from different phase relations and different relative amplitudes in the fundamentals and harmonics which are combined with each other.

Any wave which recurs with the same form in succeeding cycles, and is not of sine wave form, consists of a combination

of sine waves which are harmonics of the lowest frequency in the complex wave. In highly complex waves there are large numbers of harmonics. In square waves there are, theoretically, an infinite number of harmonics with extremely high fre-



**Fig. 54-1.**—Complex waves formed by fundamental and harmonic frequencies.

quencies making up the straight leading edges and square corners.

Harmonic frequencies are produced by detectors and by any tube which rectifies, and by any tube with such a combination of plate voltage and grid bias that the output in the plate circuit does not follow the input to the grid, which means non-linear operation of the tube. All oscillators tend to produce harmonic frequencies. Inductive feedback types usually produce more harmonics than do those operating with capacitive feedback. All relaxation oscillators, including the multivibrators, produce many harmonic frequencies along with their

fundamentals. Crystal controlled oscillators will deliver many harmonics, the number and their strength depending on the type of circuit employed. With oscillators which produce beat frequencies there are harmonics of the beats. Harmonics are produced by frequency multipliers, which are essentially amplifiers whose plate circuits are tuned to some harmonic of the frequency applied to their grid circuits.

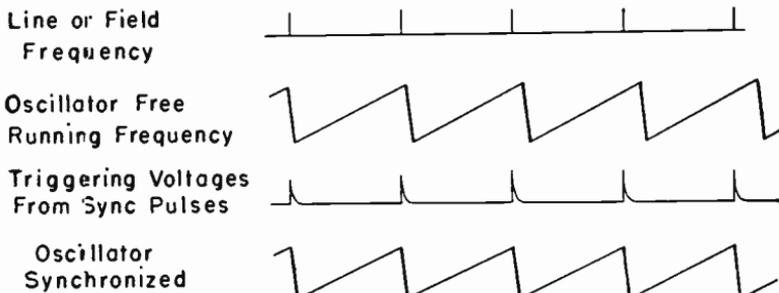
Harmonics are useful for calibration of testing apparatus. Following is an example of how harmonics may be used for calibrating or checking the calibration of a test oscillator or signal generator with the help of a receiver and any received signal of known frequency and accuracy. First the receiver is carefully tuned to the signal. Then the oscillator or generator is tuned to produce an audible beat note from the receiver and is adjusted for zero beat as nearly as this can be done. A lead from the generator output may be laid near the receiver antenna terminals if necessary. This procedure adjusts the generator to the frequency of the received signal. Keeping the receiver tuned to the original signal frequency, the generator now is tuned to lower and lower frequencies. There will be zero beat from the receiver each time the generator is tuned to a frequency which divides without a fractional remainder into the frequency to which the receiver is tuned. At each of these points the generator is tuned to its own fundamental, and the receiver is beating with a harmonic of that oscillator fundamental.

To check frequencies higher than that of the original received signal the generator is tuned to the received signal frequency by beating as previously described. The generator then will furnish harmonics of this signal frequency. The receiver may be tuned for zero beat with successively higher generator harmonics, thus calibrating the receiver to each of these harmonic frequencies. Finally, with the receiver tuned to any harmonic, the oscillator may be tuned to the same harmonic by listening for zero beat. The whole process is one of transferring calibration points to and from the receiver and

generator, commencing with the received signal and working with harmonics and zero beat points.

**HOLD CONTROLS.** — A hold control is an adjustable resistor which regulates the operating frequency of a horizontal or vertical television sweep oscillator to allow timing of oscillator action by sync pulses of the received signal. A hold control is connected to the grid of the sweep oscillator or discharge tube. The control is so adjusted that, in the absence of sync pulses, the sweep oscillator would operate a little slower or at a “free running” frequency slightly lower than the horizontal line frequency in the case of a horizontal oscillator, or the vertical field frequency in the case of a vertical oscillator. This is shown at the top of Fig. 55-1.

Just before the oscillator would become conductive in its free running state, and discharge the sawtooth capacitor for a retrace, a triggering voltage derived from a sync pulse acts on



*Fig. 55-1.—Sweep oscillator free running frequency is changed to sync frequency by the triggering pulses.*

the grid of the oscillator or discharge tube. This causes the tube to become conductive at the instant of the triggering voltage, thus starting the sawtooth discharges and retraces in time with or in synchronization with the sync pulses and triggering voltages rather than at the slightly slower timing of the free running frequency.

The hold control is made adjustable in order that the free

running frequency of the oscillator may be brought so close to sync pulse frequency that triggering voltages surely cause starting of the retraces. The hold control allows the grid voltage of the oscillator or discharge tube to become so nearly positive just before the instant of each triggering voltage that the application of even a small triggering voltage makes the grid sufficiently positive to allow discharge of the sawtooth capacitor, thus forming the retrace. Were the free running frequency of the oscillator not kept very close to synchronization by the hold control, the triggering voltages would not be enough to overcome remaining negative potential of the oscillator grid, and the grid would not be made positive to start the retrace at the correct time.

Fig. 55-2 shows how an adjustable hold control is connected to the grid of a blocking oscillator. The control resistance determines how fast the blocking charges leak off grid capacitor  $C_g$ , and thus regulates the intervals at which the grid

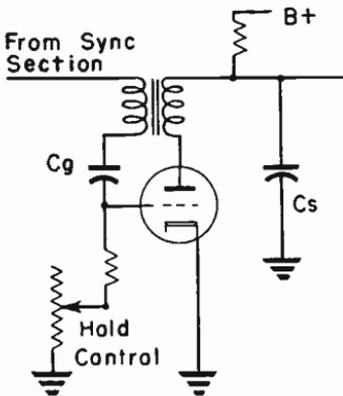


Fig. 55-2.—Hold control connected to a blocking oscillator.

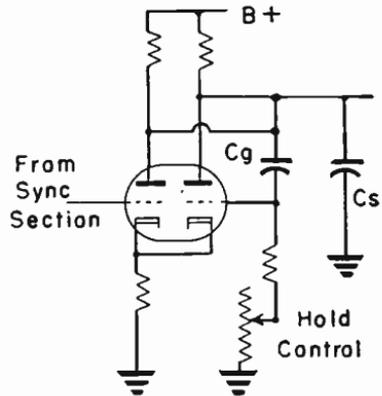
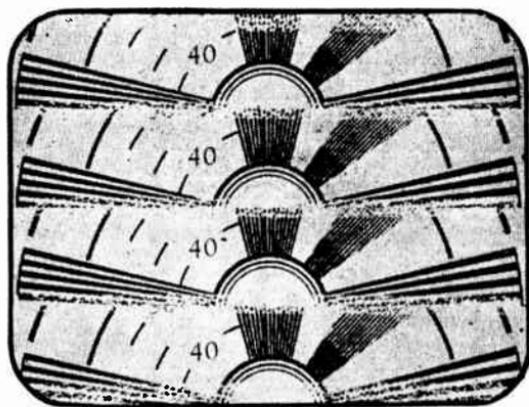


Fig. 55-3.—Hold control applied to a multivibrator.

becomes positive to allow discharge of sawtooth capacitor  $C_s$ . Fig. 55-3 shows an adjustable hold control connected to the grid of the second section or output section of a multivibrator sweep oscillator. Here too the control resistance determines

the time required for charges to leak off capacitor  $C_g$ , and determines the intervals between discharges of the sawtooth capacitor  $C_s$ . There are many modifications in details of sweep oscillator circuits and connections, but in all of them the hold control is effectively connected as illustrated or in some equivalent manner to vary the discharge rate of the grid capacitor which maintains the grid negative between the instants of sync pulses and triggering voltages.

Hold controls sometimes are called frequency controls and again may be called speed controls. In some receivers the horizontal and vertical hold controls are adjustable by the operator, and in others these controls are service adjustments



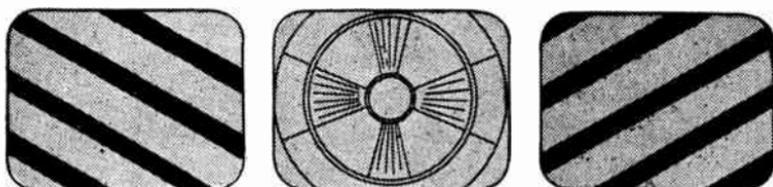
*Fig. 55-4.—Misadjustment or faults in the vertical hold control allow the pattern to move upward or downward on the picture tube screen.*

accessible from inside the cabinet or else upon removal of a portion of the front panel.

When a vertical hold control is incorrectly adjusted, the picture or pattern will move upward or downward on the screen of the picture tube. The greater the misadjustment the more rapid will be this movement. The picture appears to be rolling, with an effect somewhat as shown by Fig. 55-4. Incorrect adjustment of a horizontal hold control may cause com-

plete breakup of the picture or pattern to leave on the screen only a mass of horizontal streaks in rapid motion. As the hold control is slowly moved toward a correct adjustment the streaks will resolve themselves into sloping bars as at the sides of Fig. 55-5. With still closer adjustment the bars become more nearly vertical and finally the pattern drops into synchronization as at the center diagram.

Adjustment of hold controls is carried out as follows: With the receiver operating at normal contrast and brightness, and with a pattern or picture tuned in, turn the hold control adjustment as far as possible in one direction. If the pattern or picture drops out of synchronization, turn the control slowly



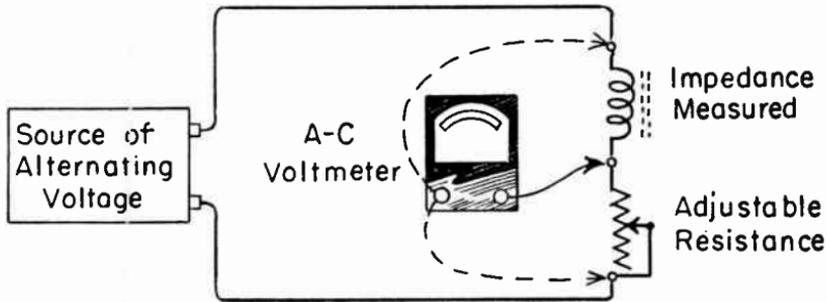
*Fig. 55-5.—Effects on the pattern of adjusting a horizontal hold control either way from synchronization.*

back until reaching the position at which the pattern or picture locks in. Note this position. Then turn the control as far as possible in the opposite direction and again back to the position at which the pattern or picture just locks in. Leave the control at a point midway between the two positions at which there is lock-in.

If turning the hold control all the way in one direction does not cause drop-out, leave the control in its extreme position while momentarily switching to another channel and back again, or while turning the receiver off and on again. This usually will cause drop-out of the pattern or picture. Then turn the control slowly back to the position where the pattern or picture pulls into synchronization. Note this position. Turn the control to the limit of its travel in the opposite direction,

again drop the pattern or picture out of synchronization, and slowly bring the control back to the position where there is pull-in. Leave the control midway between the two pull-in positions.

**IMPEDANCES.** — Impedance is effective opposition, in ohms, offered to alternating current by any combination of inductive reactance, capacitive reactance, and resistance.



*Fig. 56-1.—Measuring impedance with a voltmeter and adjustable resistance.*

Fig. 56-1 illustrates a method of measuring impedances at frequencies up to a few kilocycles. The unit whose impedance is to be measured is connected in series with an adjustable resistance, across a source of alternating voltage whose frequency is that for which impedance is to be measured. One terminal of an a-c voltmeter is connected to the junction between impedance and resistance. The other terminal is connected first to the outer end of the impedance, then to the outer end of the resistance. Resistance is adjusted for the same voltage as measured across the impedance, whereupon impedance is equal to the series resistance. If the resistance unit is not calibrated, its adjusted value may be measured with an ohmmeter.

Applied a-c voltage should be no greater than normal for the measured unit. Frequency of 60 cycles may be had from a tube heater circuit, or from the a-c power line if voltage is suitably limited. An a-f generator may be used for other frequencies, although there may be difficulty in obtaining enough signal voltage for measuring small impedances.

With reactance and resistance in series, impedance is shown by this formula.

$$Z = \sqrt{R^2 + (X_L - X_C)^2} \text{ or } Z = \sqrt{R^2 + X^2}$$

$Z$  Impedance, ohms

$X_L$  Inductive reactance, ohms

$X_C$  Capacitive reactance, ohms

$X$  Net reactance, ohms

$R$  Resistance, ohms

With reactance and resistance in parallel the impedance is shown by either of the following formulas, one for inductive reactance and the other for capacitive reactance. Meanings of the letter symbols are the same as before.

$$Z = \frac{R \times X_L}{\sqrt{R^2 + X_L^2}} \qquad Z = \frac{R + X_C}{\sqrt{R^2 + X_C^2}}$$

**INDUCTANCE.** — The approximate self-inductance of a single-layer air-core coil may be found from this formula. The result of the computation will be fairly accurate when the ratio of winding length to winding diameter is between 0.3 and 1.0.

$$L = \frac{D \times N^2}{40 \left( \frac{B}{D} + 0.45 \right)}$$

$L$  Inductance, microhenrys

$B$  Winding length, inches

$D$  Winding diameter, inches

$N$  Number of turns

The approximate number of turns required for a single-layer air-core coil of specified inductance may be computed from the following formula. Meanings of the letter symbols are the same as before.

$$N = 6.3 \sqrt{\frac{L \left( \frac{B}{D} + 0.45 \right)}{D}} \qquad \sqrt{\frac{\quad}{D}}$$

Approximate self-inductance of a coil or circuit may be measured by a resonance method. The equipment includes a signal generator, either a high-resistance r-f voltmeter or

else a low-resistance r-f current meter, and a fixed capacitor of some non-inductive type having capacitance on the order of 0.01 to 0.02 mfd. Precision of measurement is directly proportional to the accuracy with capacitance of the capacitor is known.

The unknown inductance may be connected in parallel with the known capacitance, with the generator connected to operate the combination at parallel resonance while voltage is indicated. An r-f electronic voltmeter or a d-c type with a detector probe makes a good resonance indicator. With the alternative arrangement, the inductance, the capacitance, the r-f current meter, and the signal generator are connected in series to operate the inductance-capacitance combination at series resonance.

With either connection the signal generator is tuned for maximum meter reading. Care must be exercised to tune the generator to a fundamental frequency, not a harmonic. The fundamental will be the highest frequency at which it is possible to obtain peaking of voltage or current which indicates resonance. Unless the fundamental is used, the computed inductance will be four or more times the actual value. Inductance is computed from the following formula.

$$L = \frac{25330}{C \times f^2}$$

*L* Inductance, microhenrys

*C* Parallel or series capacitance, micro-microfarads

*f* Fundamental tuned frequency, megacycles

Computed inductance will be the apparent value, which is affected to a greater or less extent by distributed capacitance of the coil and other circuit parts. Distributed and stray capacitances make the apparent inductance, as computed, greater than the actual inductance. The simplest way to minimize this effect is to use the largest parallel or series capacitance which allows distinct peaking of the resonance indicating meter. Capacitances of 0.01 to 0.02 mfd are not too large provided they are non-inductive. This large tuning capacitance makes the much smaller distributed and stray

capacitances relatively unimportant in determining the resonant frequency.

**INDUCTORS OR COILS.** — Choice of the type of winding and support for high-frequency tuning inductors often depends largely on how much distributed capacitance can be tolerated. When a coil is in series with a source of emf, and the coil possesses distributed capacitance, the inductance and capacitance of the coil are parallel resonant at some frequency. That is, the distributed capacitance of the coil has the same effect as a capacitor connected across the ends of the coil.

There is minimum distributed capacitance in a self-supporting single-layer coil with spaced turns of solid wire. Distributed capacitance is increased by winding the turns close together and by using more than a single layer. When multi-layer coils must be used in order to have the required inductance in a limited space, small distributed capacitance is secured by using duolateral or honeycomb windings having the turns of successive layers crossing each other at an angle rather than lying parallel. Capacitance is reduced also in pie windings which consist of a number of sections, each of very short length and relatively large diameter, mounted with their axes in line and connected in series to form the entire coil.

Self-supporting coils have minimum distributed capacitance because the dielectric constant of air is less than that of any supporting material which may be used. If winding wire is too small to be self-supporting, rigidity may be had with a few narrow strips of thin insulation laid lengthwise and cemented to the coil turns. When coils are wound on solid supporting forms the distributed capacitance is reduced by using material of small dielectric constant. This also reduces energy losses because it lowers the loss factor which is proportional to dielectric constant and power factor of the material.

If coil cement is used for mounting or additional support the cement should be of some low-loss kind such as made from polystyrene.

Self-supporting or strip supported windings may be made with bare copper or tinned copper wire when the turns are

spaced. Enamel covered copper wire is in general use for both spaced and close windings. To obtain worth-while reduction of capacitance by spacing the turns the distance between adjacent turns should be not less than one-third the diameter of the wire. There is little advantage in using spacing much greater than two-thirds the wire diameter. There is little if any advantage in using Litz wire for windings to be employed at frequencies higher than about three megacycles. At very high frequencies there is more loss with Litz or other stranded wire than with solid round wire.

Self-inductances of coils in series add together like series resistances, provided the coils are so far apart or are so positioned in relation to each other that there is negligible coupling and consequently negligible mutual inductance. Otherwise the total inductance is the sum of self- and mutual-inductances. Mutual inductances adds to self-inductance when the mutuals are positive, which occurs when directions of currents in the coils are such that the magnetic fields are of the same polarity and aid each other. Mutual inductances are subtracted from the total self-inductance when the mutuals are negative, when currents are in such directions that the magnetic fields oppose each other.

If coils are connected in parallel with each other, and are so positioned or spaced as to have no coupling and no mutual inductance, their total self-inductance is equal to the product of the separate inductances divided by the sum of these inductances, just as in the case of paralleled resistors. If there is coupling, and resulting mutual inductance, the total inductance is found thus.

$$L = \frac{La \times Lb - M^2}{La + Lb - 2M}$$

Where  $L$  is total inductance,  $La$  and  $Lb$  are the two separate inductances, and  $M$  is the mutual inductance, all measured in the same unit. The formula as written applies when mutual inductance is positive. If mutual inductance is negative the last term in the denominator becomes  $+2M$  instead of  $-2M$ .

**INSTALLATION, RECEIVER.** — The following should be considered when installing a television receiver in a room. The location should, if possible, be convenient to a power outlet and close enough to the point at which the transmission line enters the room that running of this line will not make an unsightly appearance nor bring the line so close to electrical machines, appliances, and switches as to allow pickup of interference. If there are ventilating openings in the sides or back of the cabinet do not place the receiver so close to a wall that these openings are shut off; leave two to three inches of clear space. See that ventilating openings in the cabinet bottom are not obstructed.

The receiver should be so placed that it may be viewed by the largest number of people in the room, allowing some viewers to be at least ten feet from the screen even though the picture tube is of only ten-inch diameter. The front of the receiver and the mask opening should be at such an angle with windows and lamps that reflections from these light sources will not make it difficult to view pictures.

When a receiver is first installed, make sure that all tubes are firmly seated in their sockets, that the face of the picture tube is against a mask cushion if such a cushion is provided, and that the front of the flare of a magnetic deflection tube is against the cushion in the neck bracket.

The chassis usually is held in its cabinet with bolts or screws through the bottom of the cabinet. These fastenings may be fitted with cushion washers. Front panel knobs usually pull straight off their shafts, although it is possible that set screws may be used and they should be looked for. Before attempting to take the chassis out of a cabinet make certain that cords or cable to the speaker are disconnected, that a cable to the picture tube socket or base is disconnected, and that the power cord is out of the wall receptacle and free from the cabinet.

**INSTRUMENTS, SERVICE.** — The two instruments which are essential for any kind of television servicing are a high-frequency signal generator which will furnish input voltages, and some form of indicating meter which will measure result-

ing outputs of amplifiers and other circuits. The signal generator must meet the requirements outlined for a marker generator in the article on *Generators, Marker*.

The meter for indicating output may be a moving coil voltmeter which usually will be part of a volt-ohmmeter or of a volt-ohm-milliammeter. In order to avoid such loading of television circuits as would destroy the usefulness of many indications the sensitivity of the meter should be at least 10,000 ohms per volt, and greater sensitivity is much to be preferred. For tests and measurements in circuits working at radio, intermediate, and audio frequencies an electronic voltmeter is much more useful than the moving coil type because the very high input resistance or impedance of the electronic type causes almost negligible loading of most television circuits. With either the moving coil voltmeter or the electronic voltmeter there will be needed a high-voltage probe allowing measurements to a maximum of about 30,000 volts in the anode circuits of picture tubes. With the electronic voltmeter should be provided a detector probe for measurements in circuits operating at audio and higher frequencies.

Although many service tests and adjustments may be carried out with only the signal generator and voltmeter, an oscilloscope allows much faster work and a more complete analysis of faults and of the results of adjustments. To make full use of the oscilloscope during alignment and trouble shooting in all circuits between the antenna and the video and sound detectors it is necessary to employ a sweep generator and a marker generator. These two generators may be combined in a single television generator. With the oscilloscope it is convenient to have a high-frequency detector probe which will allow observation of frequency responses directly from circuits preceding the video and sound detectors.

A variable audio-frequency generator is necessary only for checking performance in the audio amplifier and speaker of the sound section and for checking low-frequency performance of the video amplifiers. A square-wave generator allows more rapid checking of these sections than does a sine-wave

generator, and if the square-wave generator will furnish fundamentals up to 200 or 300 kilocycles it is useful for checking performance of video amplifiers throughout their usual frequency range.

It is assumed that an ohmmeter function is included in either the moving coil voltmeter or in the electronic voltmeter. In addition to this instrument for measuring resistances, it is highly desirable if not absolutely necessary to have a capacitor tester which allows direct reading of capacitances and which makes tests for shorts, opens, and leakage in capacitors. An inductance tester or bridge is convenient, but no so essential as the capacitor tester.

⇒ Many kinds of trouble may be quite quickly located by means of a signal tracer which will pick up signals at fre-

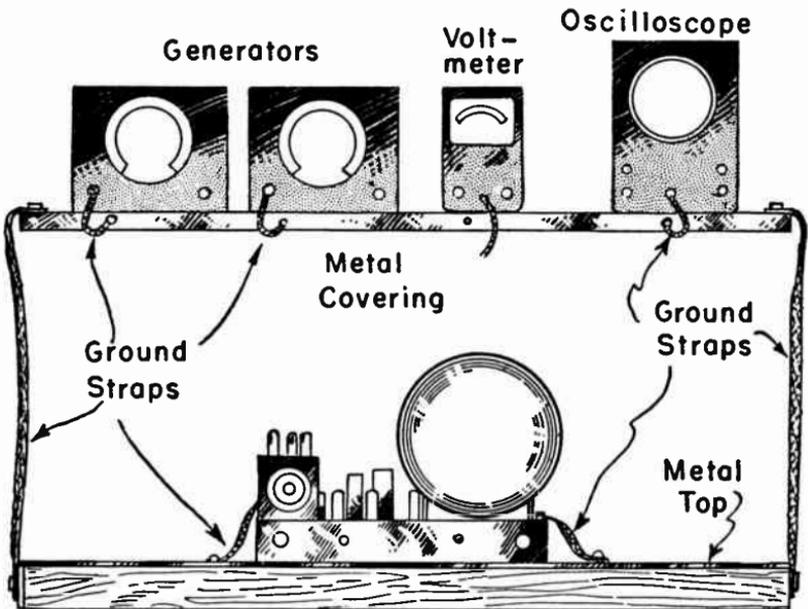


Fig. 60-1.—Service instruments and a receiver or amplifier being worked upon should be connected to a common ground.

quencies at least as high as those in television channels 2 through 6. A grid dip meter or oscillator allows measuring frequency of resonance and allows preliminary adjustment of all couplers and wave traps. Pattern generators of some types

are useful for locating troubles and making adjustments in the sync and sweep sections of the television receiver. A tube tester is convenient where a large amount of service work is being handled, although a tube which tests good with almost any type of tube tester may fail to operate in the high-frequency television circuits.

All testing instruments used during any one service operation are preferably supported on a sheet-metal covered shelf above and somewhat back of the service bench, as shown by Fig. 60-1. All instrument grounds should be made to the metal shelf cover. This cover should be bonded through large copper straps to a sheet metal covering on the test bench. The receiver being worked upon should have its chassis grounded to the bench top at one or more points. It is not necessary that the bench and shelf covers be connected to a cold water pipe or other conductor leading into moist earth. Such a ground connection makes a conductive connection to one side of the building power line. When this kind of ground connection is used, any receivers having series heaters should be fed only through an isolating transformer if shocks are to be avoided when the line cord is plugged in to make the chassis "hot."

**INSULATION.** — The three characteristics which chiefly determine the usefulness of a material as insulation are dielectric strength, volume resistivity, and surface leakage. Dielectric strength is measured in volts. It is the potential difference, in volts, at which the insulation punctures and an electric current in the form of a spark or arc passes through the opening. Dielectric strength usually is specified as the number of volts per mil (1/1000 inch) of thickness. This has little practical meaning unless the total thickness of the test specimen is mentioned, because dielectric strength in volts per mil decreases as the total thickness increases. As an example, in the case of polystyrene 0.005 inch thick the dielectric strength is about 3500 volts per mil and the total strength is 17,500 volts. With twenty-five times the total thickness (0.125 inch) the dielectric strength drops to about 600 volts per mil and the total strength is up to 75,000 volts.

The dielectric strength of air is somewhat less than 10,000 volts per  $\frac{1}{8}$  inch between flat or gently curved surfaces, and this value holds with little change regardless of the separation. Between two sharp points the dielectric strength drops to about 1250 volts per  $\frac{1}{8}$  inch. Dielectric strength of any substance varies with frequency, waveform, temperature, humidity or moisture, and length of time the voltage is applied.

Volume resistivity is the actual ohmic resistance which determines the rate of current flow through the body of insulation when a potential difference is applied on opposite sides. Volume resistivity of all commonly used insulation materials is so great as to make little practical difference in comparison with dielectric strength. Among the poorest of common insulators is red fibre, with which resistance of a cube about 0.4 inch on each side is around 5,000 megohms.

Surface leakage ordinarily results from films of moisture, oil, and dirt in general. Rough surfaces tend to collect these things and to increase the surface leakage. Some insulators absorb moisture and thus increase their leakage. Actual insulation resistance is that of volume resistance and surface resistance acting in parallel.

**INTERCARRIER SOUND.** — In a television receiver employing an intercarrier sound system the center frequency of the frequency-modulated signal delivered to the demodulator in the sound section always is 4.5 mc, produced by beating of the video and sound intermediate frequencies in the video detector and taken to the sound section from some point following the video detector. In receivers which do not employ the intercarrier sound system the center frequency for sound i-f amplifiers and sound demodulator is the sound intermediate frequency produced by beating of the video and sound carrier frequencies in the mixer.

The path of the sound signal in an intercarrier system is shown by Fig. 62-1. The received carriers enter the tuner as usual, and video and sound intermediate frequencies are delivered from the tuner to the i-f amplifier. The sound intermediate frequency is not separated from the video interme-

diate anywhere in the i-f amplifier, but both intermediate frequencies go to the video detector where they beat to produce the difference frequency of 4.5 mc. Regardless of the channel received, or of any other conditions, the video and sound

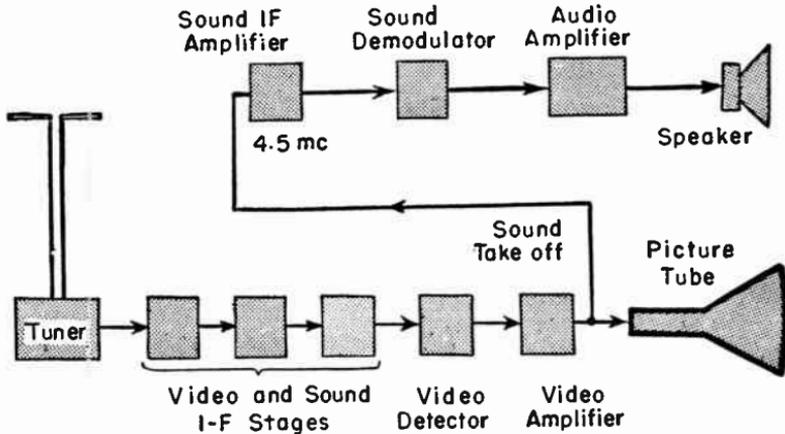


Fig. 62-1.—Signal paths in a receiver employing intercarrier sound.

carriers always are separated by 4.5 mc, the resulting intermediate frequencies always are separated by 4.5 mc, and the beat frequency from the intermediates must be 4.5 mc.

This 4.5 mc beat frequency is frequency-modulated with the sound signal, which has the standard television deviation of 25 kc. This f-m sound signal may be taken off in the output of the video detector, but usually is further amplified in one or two video amplifier stages and then taken to the sound section. It should be noted that drift or misadjustment of oscillator frequency can have no effect on the frequency difference between carriers or intermediates, nor on the center frequency fed to the sound section, which always remains 4.5 mc.

Fig. 62-2 shows connections for intercarrier sound takeoffs in which are employed double tuned transformers with their primaries in the plate or screen circuit of a video amplifier and their secondaries in the grid circuit of the sound i-f ampli-

fier or driver tube for the sound demodulator. Other takeoff couplings are shown by Fig. 62-3. The one at the left consists of an auto-transformer between the video amplifier plate and ground, with a tap connection for the sound i-f amplifier grid.

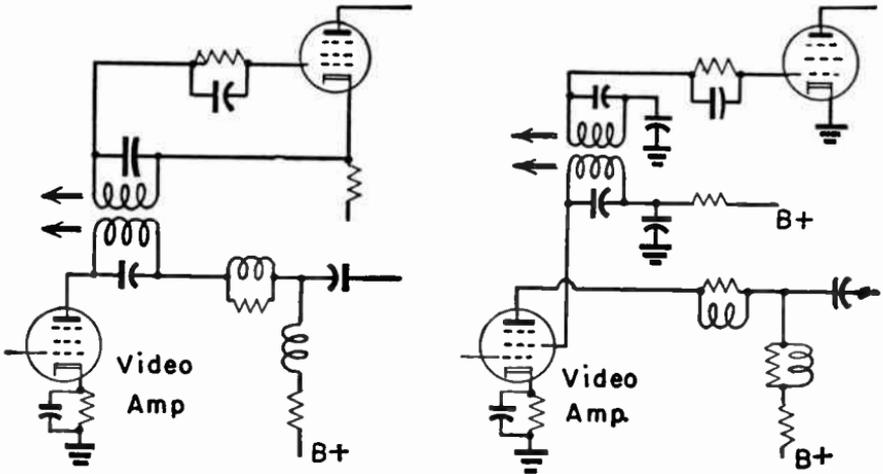


Fig. 62-2.—Intercarrier sound takeoffs consisting of double-tuned transformers.

At the right the takeoff coupling is a tuned impedance between the video amplifier plate and the sound i-f amplifier grid.

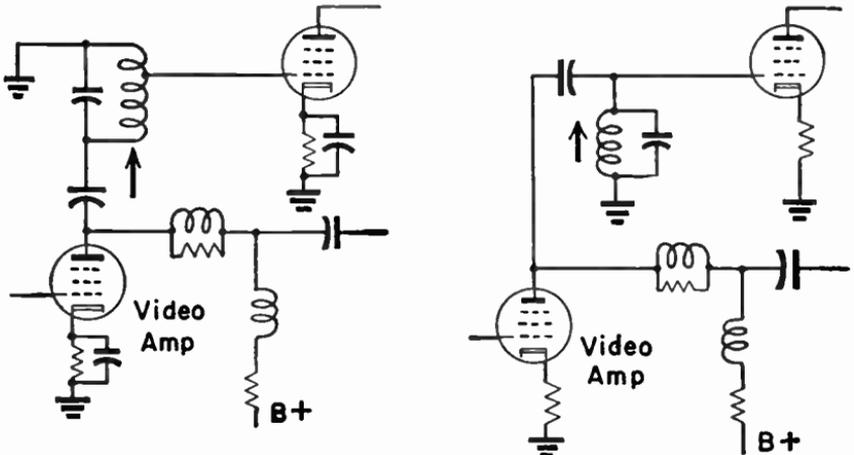


Fig. 62-3.—Intercarrier sound takeoffs through an autotransformer and a tuned impedance coupling.

All these takeoff couplings act also as traps to greatly attenuate the 4.5 mc intercarrier beat voltage which otherwise would reach the picture tube input. Following the sound takeoffs in the plate circuits of video amplifiers are the usual shunt and series compensating or peaking coils.

Frequency response of the r-f amplifier and antenna coupler of a receiver having intercarrier sound is no different from that in a receiver having any other sound system and otherwise similar. As shown at the left in Fig. 62-4 the video and sound carrier frequencies should be at or close to the peak of the response.

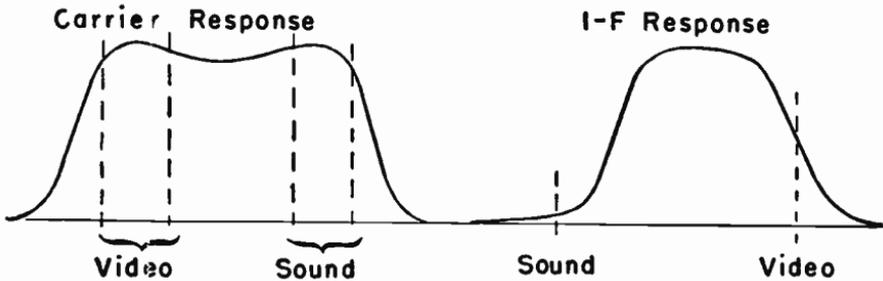


Fig. 62-4.—Typical frequency responses for receivers having intercarrier sound systems.

The intermediate-frequency amplifier of a receiver employing intercarrier sound must be aligned for a response about as shown at the right in Fig. 62-4. The video intermediate frequency should be in its usual location, about half way down on the high-frequency side of the curve. The sound intermediate frequency response or gain should be down 26 to 30 decibels or at a value of 3 to 5 per cent of the peak gain. This very low gain at the sound intermediate frequency is necessary in order that sound signals may not reach the picture tube and in order that a trouble called intercarrier buzz may be minimized.

The chief aim of the low response at the sound intermediate frequency is to keep the f-m sound signal on a relatively flat portion of the gain curve, where changes in deviation and degree of modulation, and in the center frequency itself, will not cause amplitude modulation. Were the f-m

sound signal to get onto a steeply sloped portion of the gain curve, there would be more gain with deviation toward higher frequencies than with it toward lower frequencies. Such a change of gain would constitute amplitude modulation, which would pass through the video detector along with amplitude modulation for picture and sync signals, and would place sound bars or wavy horizontal edges on the pictures.

Were the sound intermediate frequency to reach a sloped portion of the gain curve, or were the gain at this frequency to be too great even on a fairly flat part of the curve, the resulting amplitude modulation would be more than could be removed by the ratio detector which nearly always is the sound demodulator with intercarrier systems. This would

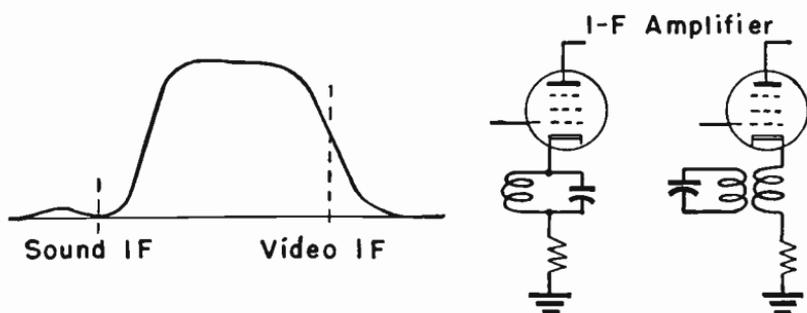


Fig. 62-5.—The shelf or plateau response sometimes used for sound intermediate frequency with intercarrier sound.

cause frequencies evident as a sharp buzzing sound. This buzz may result also from other causes, such as enough overloading of the video amplifier to bring vertical sync pulses into the audio system.

Fig. 62-5 illustrates one method of keeping the sound intermediate frequency on a low-gain and relatively flat portion of the response curve. The curve is formed with a sort of extended shelf or plateau at the point of sound intermediate frequency. This flattened extension is brought about by placing an attenuation trap on any one of the i-f amplifier tubes. Usually the trap is in a cathode circuit, as shown by diagrams at the right. The trap circuit may use a tuning

capacitor or a self-resonant coil tuning with distributed capacitance. In any case the tuning is rather broad, and, of course, is at the sound intermediate frequency.

**Alignment.** — With receivers which do not use intercarrier sound it is rather common practice to align the r-f oscillator so that maximum signal at the sound intermediate frequency passes through the sound i-f amplifier and demodulator, relying on the sharply peaked and narrow response of the sound i-f amplifier. This cannot be done with an intercarrier sound system, because no matter how the frequency of the r-f oscillator may be changed the result always is shifting of both video and sound intermediate frequencies in the same direction and by the same amount. The 4.5 mc center frequency or beat frequency, with more or less sound signal modulation, always remains.

A satisfactory method of oscillator alignment is as follows:

1. Check the frequency response of the i-f amplifier to see that it conforms to frequency requirements mentioned in connection with Figs. 62-4 and 62-5. This response must be reasonably correct. The check is made with a sweep generator and marker generator feeding into the mixer tube, and with the oscilloscope connected across the video detector load.

2. Adjust a fine tuning control to the center of its range and leave it there. Override the automatic gain control, set the contrast control at a normal operating position, and take other preliminary steps outlined in the article on *Alignment*.

3. Connect the sweep generator to the antenna terminals of the receiver. Couple the marker generator very loosely to the same antenna terminals. Connect the oscilloscope across the video detector load as is done for most alignment work.

4. Tune the receiver and the sweep generator for the channel to be first aligned. Tune the marker to the sound carrier frequency of this channel. Align the r-f oscillator by adjusting the trimmer for this channel to bring the sound frequency far enough down and onto a relatively flat part of the response.

5. Tune the marker generator to the video carrier of the same channel. The marker for this frequency should be approximately half way down on the high-frequency slope of the response. If the video marker is not reasonably close to this position the i-f amplifier section needs to be realigned.

For alignment of the sound takeoff in the intercarrier system (Figs. 62-2 and 62-3) a highly accurate signal at 4.5 megacycles should be applied to the high side of the video detector load. This assumes that the takeoff follows one of the video amplifiers, not the video detector. If the takeoff is from the detector output the 4.5 mc alignment signal must be coupled to the grid of the i-f amplifier which precedes the video detector. This signal should be accurate to within a small fraction of one per cent. It is best secured from a 4.5 mc crystal in the marker generator or with the marker calibrated from a crystal controlled oscillator working at this frequency. Alignment of the demodulator transformer then is carried out in the same manner as for any other sound system, the only difference being in the center frequency of 4.5 mc instead of the sound intermediate frequency of the receiver.

**INTERFERENCE.** — The effects of any electrical waves or fields, other than desired signals, which cause confused pictures or sound are known as interference. Kinds of interference may be broadly classified in two groups. In one class the electrical waves or fields have no definite frequency or band of frequencies, but cover very wide ranges. Such interference is caused by electrical machines and apparatus in which there are rapid changes of current, switching, sparking, and discharges. All these may be called, for convenience, spark interference. In the second class of interference are all radio transmissions, including standard broadcast, amateur, international short wave broadcast, f-m broadcast, aviation signals and communication, all governmental and commercial radio services, and television signals other than the one which it is desired to receive. All interferences in this second class are confined to definite frequencies or limited bands of frequency.

Electrical faults within the television receiver which affect picture and sound reproduction are not classed as interference. Interference from adjacent channel transmissions in the same television band, and picture interference due to accompanying sound, are considered in the article on *Traps, Interference*.

**Spark Interference.**—The troubles classed generally as spark interference cause steady or intermittent bright flashes, horizontal white streaks, as in Fig. 63-1, or short black lines. The trouble usually will appear on all channels. This interference will come and go if it results from passing motor vehicles, street cars, elevated trains, elevators, or other sources

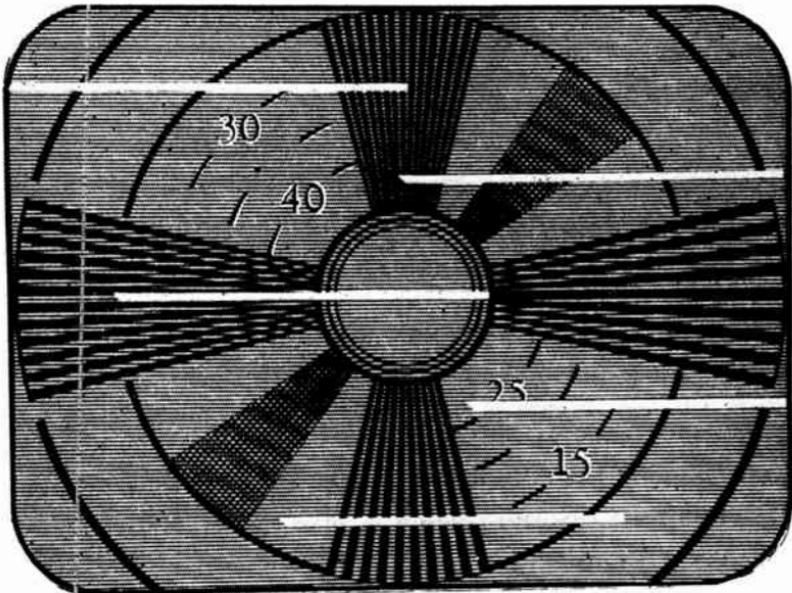


Fig. 63-1.—Effects of spark interference as seen at the picture tube.

in motion. Very strong interference may cause tearing of the picture due to momentary loss of synchronization.

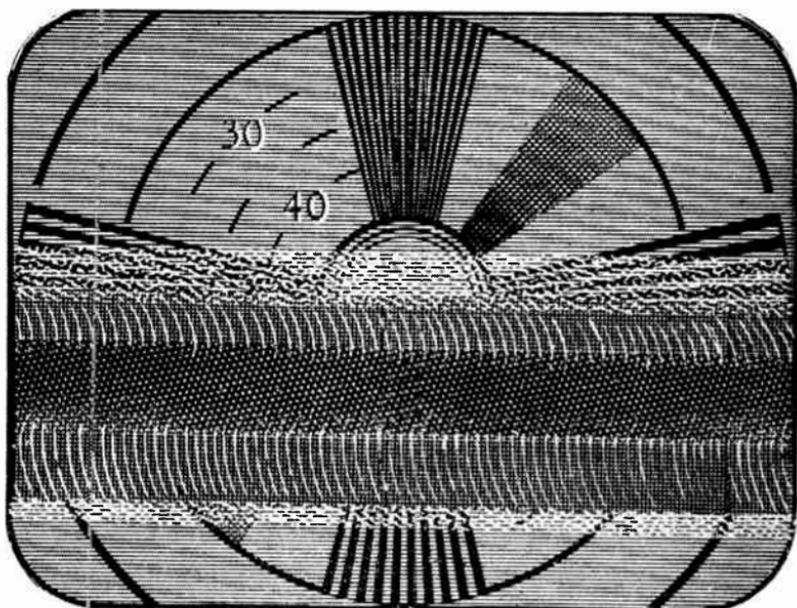
Principal sources of spark interference include the following. Automobiles, airplanes, trucks, buses, and everything else which uses internal combustion engines with electric

ignition. Street cars, elevated cars, electric trains, elevators, and everything else containing direct-current motors or other motors having brushes and commutators. Flashing electric signs and any electrical display mechanism having intermittent switching. Household and store appliances operated with series a-c motors or universal motors. It should be noted that symptoms similar to those from external spark interference may result from corona or flash-over in the high-voltage power supply and wiring of the television receiver.

Remedies for spark interference are most effective if they can be applied at the electrical machine or device which is causing the trouble. Sparking contacts and brushes should be cleaned, adjusted, or replaced if necessary. Contacts may be bypassed with a quarter-mfd or larger fixed capacitor. Filters such as used for sound radio may be used in the power line connection of the offending device. Grounding of motor frames and machine frames to cold water pipes will help. If such grounding already is in use, the connections should be checked for conductivity. If nothing can be done at the source of interference, and if the electrical disturbance is being transmitted by way of the power lines, it may help to connect the television receiver to its power supply through an isolation transformer of one-to-one voltage ratio. There should be a shield between transformer windings. This shield and the transformer core, usually connected together, should be grounded to a cold water pipe.

Interference from medical apparatus such as diathermy machines and some X-ray apparatus emits high frequency radiations, but is in the general class of spark interference because these radiations extend over a very broad band of frequencies, and because remedies are generally the same as for other spark interference. Power line filtering and grounded shielding at the troublesome apparatus usually will give relief. This general class of interference, where high frequencies are radiated, causes herringbone patterns either in broad horizontal bands or all over the picture tube screen. The effect is illustrated by Fig. 63-2.

Trouble due to interference radiated through space may be lessened by using shielded transmission line. Antennas of the H-type and some other stacked arrays tend to reject signals from below and above the antenna level, hence may help in some cases of radiated interference. Any antenna may be partially shielded from interference at a lower level by con-



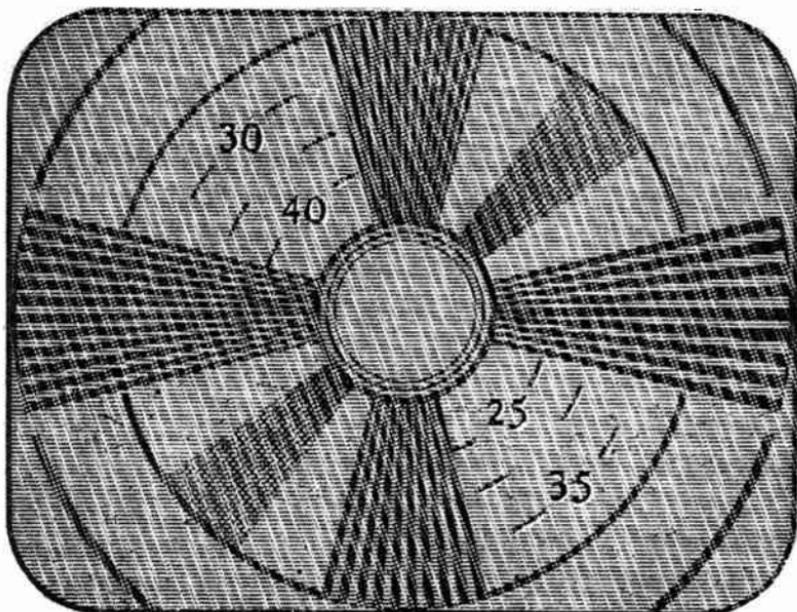
*Fig. 63-2.—Effect at the picture tube of interference from diathermy machines and other medical apparatus.*

ductively attaching to the metal antenna mast a dipole and reflector, or reflector and director, a few feet below the regular antenna conductors. Orienting the antenna in a direction away from the interference seldom helps, because desired signal strength on weak stations will be reduced more than the interference. Radiated interference will be reduced by an antenna with greater gain or better signal to noise ratio. This is true partly for the reason that such antennas usually are quite directional and tend to pick up more signal while excluding more interference. A tuned preamplifier at the receiver will bring up the strength of signals in and near the

desired channel more than it will increase the level of broad band interference.

**Radio-frequency Interference.**— Interference from f-m broadcast stations appears on the picture tube screen as narrow diagonal lines which constantly change their slope and pattern to give a weaving or rippling effect. Very strong f-m interference may produce an all-over herringbone pattern made up of fine curved lines.

Interference from transmitters using amplitude modulation in standard broadcast, short wave, amateur, or other bands may cause narrow shifting lines which appear similar to those from f-m interference. More often these transmissions produce on the picture tube screen a very fine grained all-over plain or crosshatch pattern which may change to a



*Fig. 63-3.—Radio-frequency interference causes beats which produce narrow lines on the screen pattern or picture.*

fine-lined herringbone pattern with modulation of the transmission. Only when a-m interference is very strong will horizontal sound bars appear on the picture tube of the

television receiver. The usual effect is as shown by Fig. 63-3.

It is characteristic of r-f interference that the pattern will be stationary or nearly so when there is no modulation of the interfering signal. The pattern will change and weave, or may develop bars, spots, and rather strong herringbone forms when there is modulation.

Remedies for r-f interference include the following. The use of directional forms of antenna, with reflectors and possibly directors also, which are oriented for maximum desired signal and minimum interference. There is advantage in any antenna of relatively high gain or better signal to noise ratio when the interfering frequencies are not in the television bands where reception is wanted. Tuned preamplifiers are helpful when interference is not at frequencies desired. Shielded transmission line allows much improved reception in localities where there is interference extending throughout a wide range of radio frequencies.

The most direct approach to reduction of r-f interference occurring at only one frequency or in a narrow band of frequencies is use of tuned trap circuits at the antenna terminals or in the antenna coupler of the television receiver. The traps may be series or parallel resonant circuits made with coils and capacitors or they may be short sections of transmission line, all as described later in this article.

**Frequency of R-f Interference.** — It is possible for strong r-f interference to be picked up by parts of the i-f amplifier when the frequency of the interference is within the band pass of the i-f amplifier. Such interference does not come through the antenna and tuner, but acts directly on the i-f circuits. To occur within this range of frequencies the interference would come from government services or from police, various transportation services, commercial controls, or from one of the amateur radio bands.

If interference which affects pictures or sound is picked up by the antenna, the signals must come from the mixer at frequencies within the band pass of the i-f amplifier. The total band width of this amplifier may extend from about  $1\frac{1}{4}$  mc

below the sound intermediate frequency to about  $21\frac{1}{4}$  mc above the video intermediate frequency as extreme limits. In order to cover this i-f band pass, signals reaching the antenna and the r-f amplifier of the receiver must extend from about 1 mc below to 1 mc above the 6-mc limits of the channel to which the receiver is tuned when affected by the interference. This consideration fixes the limits of interfering frequencies for the channel to which the receiver is tuned. As an example, with the receiver tuned to channel 5, whose frequencies extend from 76 to 82 mc, the frequency of interference at the antenna would be between about 75 and 83 mc in order that the interference might be accepted and amplified.

There are not many signal frequencies which will fall directly in the channel limits as just mentioned. But other signal frequencies are brought within this range by second harmonics of these signals, by second harmonics of the r-f oscillator in the receiver, by signals which are at image frequencies, and by conversion action in the r-f amplifier when this amplifier is so biased as to have the properties of a detector, mixer, or converter.

To identify frequencies which may be brought within the interference range by means mentioned in the preceding paragraph we shall use letters for the several classes.

*A. A frequency within the tuned channel or 1 mc either side.* Here we may find radiation from the r-f oscillator of a nearby television receiver tuned to the third or fourth channel below the one in which the first receiver experiences interference. Example: Interference on channel 5. Channel frequencies are 76 to 82 mc, and assumed interference limits are 75 to 83 mc. A nearby receiver may be tuned to channel 2, where the video carrier is at 55.25 mc. If the video intermediate of that receiver is 26.4 mc its r-f oscillator frequency will be 81.65 mc, which is within the interference range of the first receiver.

*B. A frequency whose second harmonic is within the tuned channel or 1 mc either side.* This would be any fre-

quency whose fundamental is equal to one-half the interference range for the channel to which the receiver is tuned. When tuned to channel 5, where the interference range is 75 to 83 mc, the fundamental frequency of the interference would be between 37.5 and 41.5 mc. It would be different for other channels.

*C. Any signal which is an image frequency for the channel in which there is interference.* Television oscillators nearly always operate at frequencies higher than carrier frequencies. Consequently, an image frequency must be higher than the oscillator frequency, and must be equal to the sum of the oscillator frequency and the intermediate frequency of the receiver.

Example: Interference is on channel 5. This channel is being used for all examples in order to avoid confusing changes of frequencies and limits, but every channel would require computations on the basis of its own frequencies. The video carrier is at 77.25 mc. Assume the video intermediate is 26.4 mc. Then the frequency of the r-f oscillator is 103.65 mc. Interference limits for this channel are taken as 75 to 83 mc. Adding these limit frequencies to the r-f oscillator frequency gives frequency range of this interference as between 178.65 and 186.65 mc. These are within the ranges of modulated carriers in channels 7 and 8.

Correctly designed r-f amplifier stages have ample selectivity to prevent image interference, but poor design and construction or lack of a tuned r-f stage may allow such interference.

*D. Any signal whose second harmonic is an image frequency for the channel in which there is interference.* This would be a signal whose fundamental is within a range of frequencies equal to one-half of those in the preceding class of interference. Continuing with the example there worked out, the interference range would be one-half of 178.65 to 186.65, or would be in a range of fundamentals between 89.33 and 93.33 mc. These harmonic image frequencies of interference are not so very far above the channel frequencies

for the tuned channel, and would call for more selectivity in the r-f amplifier.

*E. Any signal which may beat with the second harmonic of the receiver r-f oscillator to produce a difference frequency within the interference range for the tuned channel.* In earlier examples for interference in channel 5 the frequency of the r-f oscillator was computed as 103.65 mc. The second harmonic is 207.30 mc. Frequency limits for interference signals would be arrived at by subtracting the channel limits from this second harmonic frequency. Subtracting 75 mc and 83 mc from 207.3 mc gives a range of 132.3 mc down to 124.3 mc for the interference. Some aviation radio services are in this range. It must be kept in mind that other channels and other intermediate frequencies would yield entirely different results.

*F. Any signal whose frequency becomes an image frequency for the received channel when that signal beats with the second harmonic of the receiver r-f oscillator.* In the example for preceding class *E* the second harmonic of the r-f oscillator was computed as 207.30 mc. In the example for preceding class *C* the image frequency range was computed as 178.65 to 186.65 mc. The interfering signal frequency range would have to be the difference between the image range and the oscillator second harmonic. The difference would be from 28.65 mc down to 20.65 mc. In this range there are amateur and other a-m transmissions.

*G. This classification includes all interference which is made possible by conversion or mixing in the r-f amplifier of the receiver.* When some types of automatic gain control or of manual gain or contrast control are applied to an r-f amplifier the amplifier tube may operate on the lower bend of its mutual characteristic. Then the tube does not act as a linear amplifier, with changes of plate current and voltage proportional to changes of grid voltage, but rather acts as a partial rectifier or mixer or converter.

It is quite possible for a small voltage at the frequency of the r-f oscillator to reach the grid of the r-f amplifier along

with signal frequencies from the antenna. If the amplifier is acting as a converter, there is mixing of the signal and oscillator voltages in the amplifier. If the difference frequency or beat frequency is within the band pass of the tuned coupling between r-f amplifier and regular mixer, this difference frequency will again be converted by the regular mixer and will be passed through the i-f amplifier.

In order that this action of r-f conversion may cause interference, the received signal frequency must be such that beating with the r-f oscillator frequency will produce a frequency within the interference range of the tuned channel. For channel 5 this interference range is assumed to be from 75 to 83 mc. In previous examples the r-f oscillator frequency for this channel has been computed at 103.65 mc. Then the interfering frequency must be in a range equal to the sum of the oscillator frequency and the channel range. Adding 103.65 mc to the range of 75 to 83 mc gives the range of the interfering signal as 178.65 mc to 186.65 mc. Signals within this frequency range normally would be rejected due to selectivity of a r-f amplifier stage tuned for channel 5, but conversion in the r-f amplifier produces a beat frequency which is accepted.

The class of trouble just explained may be called interchannel interference because it permits signals of transmissions in a higher channel to interfere with reception in a lower channel. The difficulty usually is interference by signals from channel 7 with reception in channel 5, and interference by signals from channel 10 with reception in channel 6.

Interchannel interference usually may be eliminated or greatly reduced by a trap circuit or circuits on the grid of the r-f amplifier or on the antenna coupling which connects to the r-f grid. This trap is tuned to the frequency of the r-f oscillator when operating for the channel in which there is interference. Thus the trap removes the oscillator frequency from the r-f grid and prevents conversion action.

Another method of preventing conversion interference is operation of the r-f amplifier with fixed bias rather than automatic or manually controlled bias. The tube then is worked on a straight or nearly straight portion of its mutual characteristic, with full gain, and so provides linear amplification without acting as a converter or mixer.

**Identifying Interference Frequency.**— Interference from f-m broadcasting stations sometimes is strong enough that station announcements, including call letters, will be heard from the television receiver loud speaker. When interference is evident on the television screen, changing a fine tuning control or temporarily changing the r-f oscillator alignment may bring in voice announcements which allow identifying a station and its operating frequency. A signal tracer containing a detector for amplitude modulation sometimes will make it possible to hear voice or code identifications of a-m broadcasters, short-wave stations, and amateur stations.

While interference is present, the r-f oscillator of the receiver may be removed from its socket or made inoperative by connecting the oscillator grid to its cathode or B-minus. If interference continues, it is being picked up on the i-f amplifier and interference frequency is within the band pass of this amplifier.

Interference frequency often may be determined by means of a constant frequency (marker type) signal generator having reasonably accurate calibration in the range of television carrier frequencies. The generator output is loosely coupled to the antenna terminals of the receiver. Merely laying the output cable near the antenna terminals usually gives enough coupling. While interference is evident on the picture tube screen, tuning the signal generator through the interference frequency will cause distinct changes of the interference pattern, or movement of the pattern, or a new and different pattern. If there is no transmission in the channel at the time of testing, but only the interference, audible beat notes may be produced by the generator and interference frequencies.

Generators usually produce harmonic frequencies. Con-

sequently, the interference may be at a harmonic of the frequency to which the generator is tuned. A check should be made by tuning the generator to twice the frequency at which interference effects first appear, and this doubling of generator frequency should continue until reaching the highest frequency at which the effects appear. Then the interference is at the generator fundamental frequency or tuned frequency.

This test with a signal generator identifies the frequency of interference as it reaches the television antenna. This may be a second or higher harmonic of the frequency at which the interference source is operating.

**Transmission Line.** — An unshielded transmission line may pick up any kind of r-f interference, and will do so where interference is strong and line layout is conducive to pickup. It is advisable to use the shortest possible length of line, to avoid horizontal and nearly horizontal runs so far as is possible, and to twist a ribbon type line about one turn per foot of length.

Shielded line is of great help where there is strong interference. This may be coaxial line with its outer conductor grounded or may be shielded two-conductor line with the grounded outer shield entirely independent from the two signal-carrying conductors. Unshielded 300-ohm line may be replaced with shielded 300-ohm two-conductor line to avoid upsetting impedance match between antenna, line, and receiver.

A balanced 300-ohm line which is open or has a high resistance joint in one side becomes unbalanced and will pick up much interference while reducing the strength of desired signal. This will happen also if there is an open in one side of the antenna coupling circuits in the receiver.

**R-f Inductor-capacitor Traps.** — Fig. 63-4 shows typical series-resonant traps consisting of inductor coils and capacitors connected from antenna and r-f grids to ground. The coils may be adjustably tuned with slugs or the capacitors may be of the adjustable trimmer type. In either case the construction is of high-Q design, allowing quite sharp tuning to the interference

frequency or r-f oscillator frequency which is to be bypassed to ground by the trap. When only a single r-f amplifier is used instead of a twin type, as at the right, the trap elements on the ground side of the antenna and cathode side of the tube sometimes are omitted.

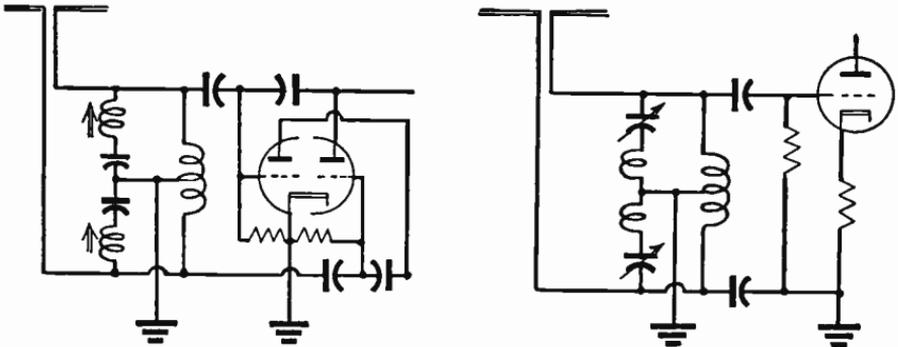


Fig. 63-4.—Traps in r-f grid circuits and antenna couplings for 300-ohm balanced transmission line.

Fig. 63-5 shows connections for traps used with unbalanced coaxial transmission lines. At the left a parallel resonant rejection trap is between the high side of the transmission line and the lead to the r-f amplifier. At the right a series resonant trap is connected from the high side of the line to ground. Either kind of trap may be tuned with adjustable inductance or capacitance. Traps of types shown in the two figures are used for reduction or elimination of every type of r-f interference which has been described.

These interference traps are adjusted by tuning the receiver to the channel in which interference is evident, or to the lowest channel of a band when interference extends throughout the entire band. If there is a fine tuning control it should be adjusted for correct oscillator frequency on the tuned channel, usually by setting this control for best sound or picture. The contrast control should be set at the highest point which does not distort the picture or pattern. The trap then is adjusted for minimum interference. When there are two tuned inductors or capacitors, as in Fig. 63-4, both should be kept approxi-

mately together in position until interference is reduced as far as possible. Then first one and then the other of the two adjustments should be changed one direction and the opposite direction to obtain any further improvement possible.

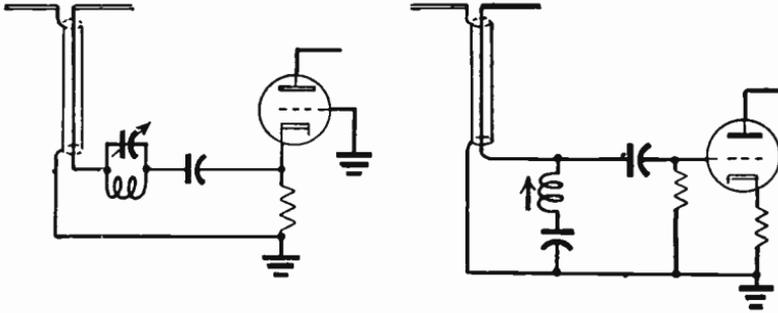


Fig. 63-5.—Parallel and series resonant traps shown connected to inputs from coaxial transmission line.

Trap adjustment may be made with a constant frequency or marker type generator coupled to the receiver antenna terminals with matched impedance, and tuned to the frequency of the interference. The generator thus simulates the interference. An electronic voltmeter or high-resistance d-c voltmeter is connected across the video detector load. Channel tuning, fine tuning, and contrast control are adjusted as in the preceding paragraph. The trap then is adjusted for minimum meter reading.

**Line Stubs for Interference Traps.**— Pieces of transmission line whose length is approximately a half-wavelength or quarter-wavelength corresponding to the interference frequency may be connected to the receiver antenna terminals to act as traps. Such transmission line stubs are illustrated by Fig. 63-6. A stub whose physical length is close to a half-wavelength of the interference signal, when shorted at its far end, acts like a series resonant circuit connected across the antenna terminals. A half-wave shorted stub is shown at the left. The resonance peak of the stub is quite sharp and there is correspondingly great attenuation of the interference signal. If a broader response is wanted, with some sacrifice of inter-

ference attenuation, the ends of the stub may be connected together through a fixed composition resistor of 50 ohms or less.

If the stub is cut to approximately a quarter-wavelength of the interference signal, and the ends are left open, there is

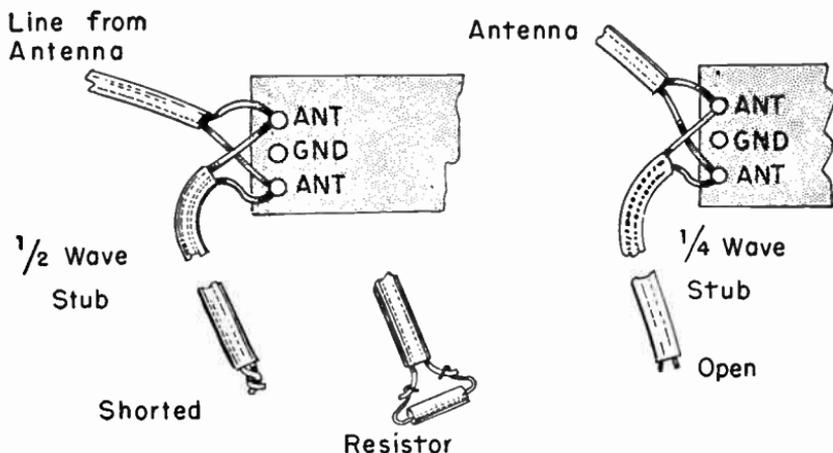


Fig. 63-6.—Transmission line stubs used as interference traps.

again the equivalent of a series resonant circuit across the antenna terminals. That is, either the half-wave shorted stub or the quarter-wave open stub is the equivalent in impedance characteristics to the series inductor and capacitor at the right in Fig. 63-5.

The interference stubs illustrated are made from 300-ohm twin-conductor transmission line, and are used where the transmission line from the antenna is of this type. If the transmission line is of the 75-ohm or 52-ohm coaxial type the stub will be made of the same kind of coaxial line with its central conductor connected to the receiver antenna terminal and the outer shielding conductor connected to the ground terminal.

Lengths of interference stubs may be computed by dividing the following factors by the interference frequency in megacycles.

Twin-line  $\frac{1}{2}$ -wave stub, inches = 4900/megacycles

Twin-line  $\frac{1}{4}$ -wave stub, inches = 2450/megacycles

Coaxial  $\frac{1}{2}$ -wave stub, inches = 3880/megacycles

Coaxial  $\frac{1}{4}$ -wave stub, inches = 1940/megacycles

The stub should be cut an inch or two longer than the computed length, then reduced after connection to the receiver until there is maximum reduction of the interference. The best length may turn out to be either more or less than the computed length. Shorting a half-wave stub with a resistor instead of a direct connection makes it necessary to cut the stub shorter in order that it may be resonant at a desired frequency. Stubs should be kept clear of chassis metal. If brought close, the stub will have to be shortened in order to be resonant at the desired frequency.

**INVERTERS AND INVERSIONS.** — With any tube having a grid and a plate, signal voltage at the plate is in opposite phase to signal voltage at the grid. At any one instant of time the voltage at the plate is changing in an opposite direction to change of voltage at the grid of the same tube.

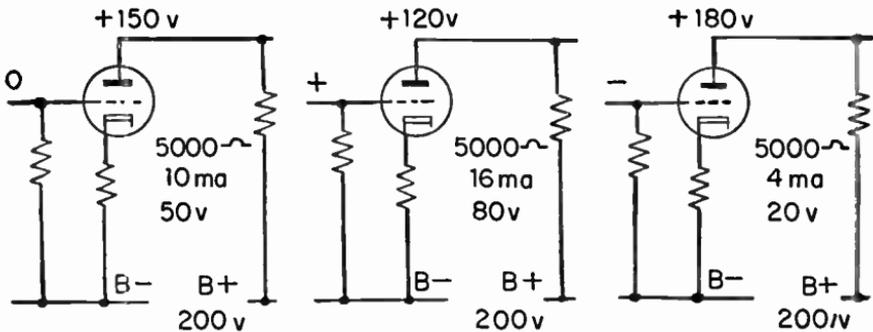


Fig. 64-1.—Inversion or phase reversal of voltages changes at grid and plate.

The reason is illustrated by Fig. 64-1, where plate load resistance is 5,000 ohms and plate supply potential is 200 volts, which is assumed to remain constant or nearly so when there are changes of plate current. At the left there is zero voltage on the grid, an assumed plate current of 10 milliamperes, and

a resulting drop of 50 volts in the plate load. This leaves 150 volts at the plate of the tube.

In the center diagram the grid voltage has been made positive, resulting in an assumed increase of plate current to 16 milliamperes, a load drop of 80 volts, and 120 volts remaining at the tube plate. Increase of grid voltage in a positive direction has made the plate voltage less positive, and output voltage from the plate and the top of the load will become less positive or negative.

In the right-hand diagram the grid voltage has been made negative. Plate current decreases to an assumed 4 milliamperes, with a consequent 20-volt drop in the load and 180 volts remaining at the tube plate. Increase of grid voltage in a negative direction has made the plate voltage more positive, and output voltage from the plate circuit will become more positive.

Signal inversion is employed in video amplifiers to make picture signals positive and sync pulses negative at the picture tube grid, or to make picture signals negative and sync pulses positive at the picture tube cathode. Inversion is used also in the sync section to provide correct or required polarity of triggering voltages for sweep oscillators. Inversion is utilized in one of a pair of deflection or sweep amplifiers for picture tubes having electrostatic deflection, and in somewhat similar fashion for one tube or one section of a tube in push-pull audio amplifiers. There is inversion also wherever signal input is to a grid and signal output from the plate of the same tube, although such inversion may not be desired and may require an additional inverter tube to bring the signal back to its original phase or polarity relation.

At the left in Fig. 64-2 the signal input is between grid and cathode, with signal output taken from across the cathode resistor. This is the arrangement usually called a cathode follower. There is no phase reversal or inversion of the signal. When the input signal makes the grid more positive or less negative there is increase of plate current. Plate current flows not only in the plate circuit as far as the B-supply, but also in

the cathode resistor. With electron flow toward the cathode, the more positive grid and increased plate current make the cathode more positive, and signal output voltage becomes more positive. An input signal making the grid more negative

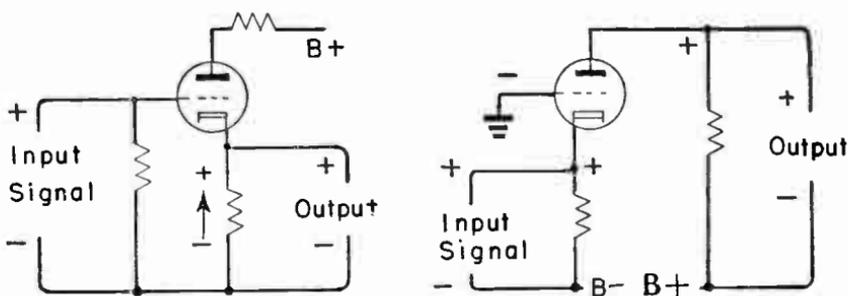


Fig. 64-2.—Tube circuits with which there is no inversion between input and output voltages.

reduces plate and cathode current, making the output voltage less positive or negative.

At the right in Fig. 64-2 the input signal is applied across the cathode resistor, as with a grounded grid amplifier. There is no phase reversal or signal inversion when the output signal voltage is taken from between plate and cathode. When the input signal makes the cathode more positive the effect is the same as making the grid more negative, and, as in Fig. 64-1, this makes the plate signal more positive. When the input signal makes the cathode less positive or effectively more negative, the result is the same as when making the grid less negative or more positive. Then the plate voltage and output voltage becomes less positive or more negative, because there is inversion between grid and plate voltages.

## L

**LIGHT AND VISION.** — The velocity of light in empty space or a vacuum is 300 million meters per second or approximately 186,000 miles per second. Wavelength of light usually is measured in a unit called the angstrom, which is equal to one ten-billionth of a meter or about 1/250 millionth of one

inch. Another commonly used unit of wavelength is the millimicron, which is equal to 10 angstroms.

Wavelength in angstroms at the centers of the various color ranges in the visible spectrum are approximately as follows:

Violet	4150	Yellow	5800
Blue	4700	Orange	6100
Green	5200	Red	6700

Wavelengths shorter than violet and having no effect on vision are classed as ultra-violet. Those longer than red and having no effect on vision are classed as infra-red. Wavelengths of all visible frequencies reaching the eye together form white light. Absence of all visible radiation leaves the sensation of black. The average human eye is most sensitive to a wavelength of about 5560 angstroms, which is between the green and yellow bands.

**LIMITERS, F-M SOUND.**—A limiter stage as used in f-m amplifiers for sound consists of a tube operated in such manner that plate signal voltages cannot greatly exceed a predetermined amplitude, even when amplitude of input signal voltage increases. Limiter stages precede f-m demodulators, and prevent amplitude modulation in the form of variations of signal strength from reaching the demodulator. Limiters are more common with discriminator types of demodulator than with ratio detector types, because a discriminator accepts and passes amplitude modulation more than does a ratio detector.

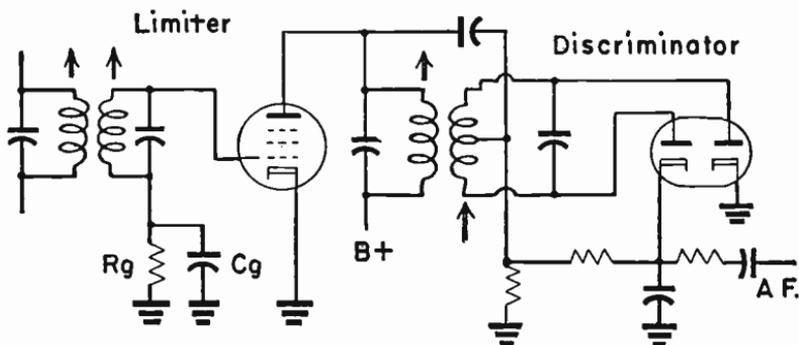


Fig. 66-1.—A limiter with its biasing resistor and capacitor in the grid return connection to ground.

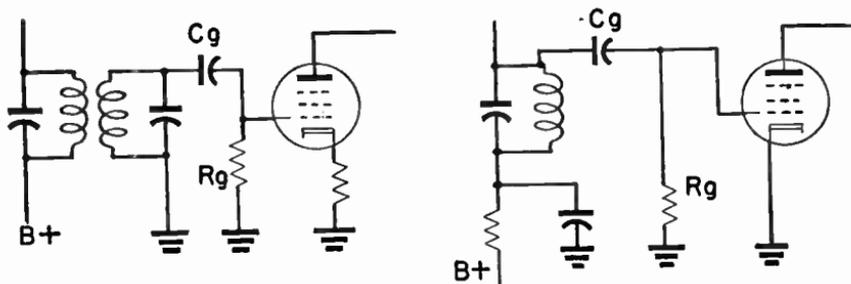


Fig. 66-2.—Limiters having biasing resistors and capacitors connected directly to the grid.

A limiter stage preceding a discriminator is shown by Fig. 66-1. The limiter tube is operated with grid-leak bias furnished by resistor  $R_g$  and capacitor  $C_g$ , and usually with screen voltage lower than for amplifier tubes. Bias elements sometimes are in positions shown by Fig. 66-2.

Limiting action is shown by Fig. 66-3. At the left there is slight cutoff at  $B-B$  in the output of positive input peaks  $A-A$ . This is due to grid current for charging the bias capacitor. Highly negative grid bias causes cutoff of negative out-peaks  $C-C$ , at the right, while positive peaks at  $B-B$  are cut off by plate current saturation due to low screen voltage. Thus

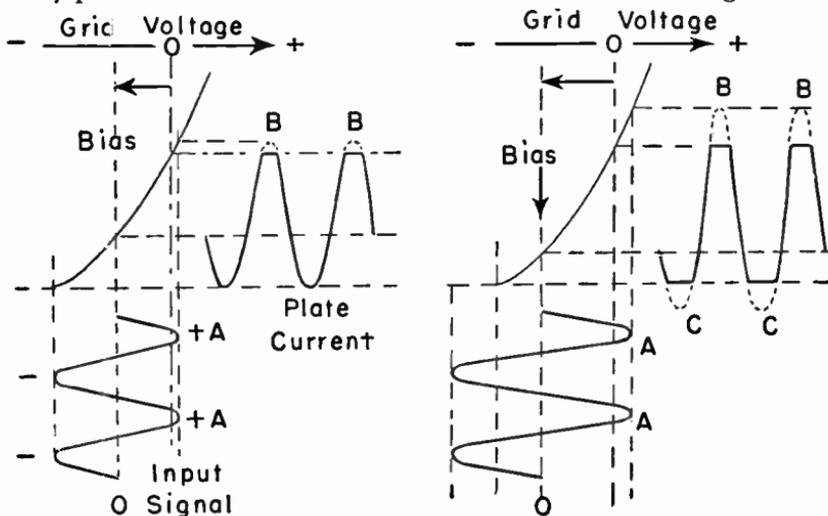


Fig. 66-3.—How limiting is caused by plate current cutoff and by positive grid voltage.

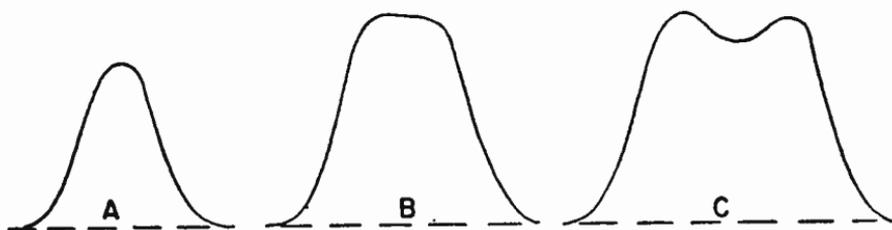


Fig. 66-4.—Frequency response of a limiter with various input voltage amplitudes.

output signals are held closely within certain amplitude limits when input voltage increases.

Limiting can occur only when input signals are strong enough to cause plate current saturation and cutoff with element voltages and grid bias employed. A weaker input leaves frequency response peaked and of proportionately low amplitude as at *A* of Fig. 66-4. Limiting of a stronger input signal is shown by the flattened response at *B*. Still stronger input distorts the response as at *C*, but still there is effective limiting. Such distortion is not harmful, since it is only variations of frequency, not of amplitude, that are to be passed on to the demodulator.

Limiter tubes are sharp cutoff pentodes, which allow plate current cutoff at moderately negative grid biases. Low screen voltage contributes to quick cutoff with any increase of bias, also prevents excessive screen and plate currents at the small bias existing when input signals are of low amplitude. Stronger input signals allow higher screen and plate voltages on the limiter, and the stronger will be signals passed on to the demodulator. Relatively weak inputs, of small average amplitudes, require lower screen and plate voltages in order to have limiting, and weaker signals go to the demodulator.

**LINES, RESONANT.** — A resonant line consists of a pair of insulated parallel conductors, usually a section of transmission line, of such length that when connected to a source of radio-frequency voltage the line will behave like a parallel or series resonant circuit, like an inductive or capacitive reactance, or like an open or short circuit.

Resonant lines sometimes are called line stubs. Ordinarily their length is a half-wavelength or less at the frequency where resonant effects are wanted. Resonant lines are not used for transmission of power, rather they are used for matching of impedances between antennas, transmission lines, and receivers, also for filters, for step-up and step-down transformers, and, in general, for circuit elements at frequencies with which coils and capacitors are inefficient because of small size.

**Wave Reflections.** — The characteristics of resonant lines depend on and may be explained with reference to wave reflections and resulting standing waves of voltage and current

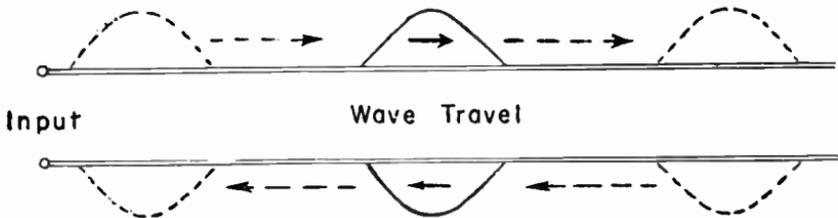


Fig. 67-1.—Electromagnetic waves and pulses of potential and current move along the line conductors.

which are formed along the lines. Reflections are caused as follows. If, in Fig. 67-1, a potential difference of particular polarity is applied across the input ends of the two conductors forming the line, an electromagnetic wave will move away from the source along one conductor and a similar wave, in opposite phase will move toward the source along the other conductor. The waves are accompanied by movements of electrons, or by currents, and by changes of potential.

Were the conductors of the line infinitely long, the waves with their currents and potentials would travel on until all

energy has been dissipated in line losses or attenuation. If the line were not infinitely long, but ended at a pure capacitance as at *A* in Fig. 67-2, the electric energy of the wave would charge the capacitance. Then the capacitance would discharge and return all the original energy to the line, since a pure capacitance absorbs no energy. This discharge would start an

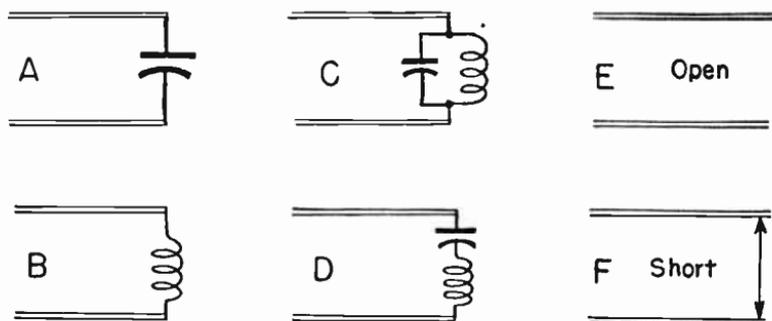


Fig. 67-2.—Line terminations which cause wave reflections.

electric wave in the opposite direction; back toward the input along the line conductors. This would be a reflected wave.

Were the line to end at a pure inductance, as at *B*, the magnetic energy of the wave would build up a magnetic field around the inductance. Then the field would collapse and return all the original energy to the line, for a pure inductance absorbs no energy. This energy would start a magnetic wave in the opposite direction, back toward the inlet. The returning wave would be a reflected wave.

There would be reflection also were the line to end at a parallel resonant circuit as at *C* in Fig. 67-2, provided the frequency from the source connected to the input were the resonant frequency of this parallel tuned circuit. Reflection would occur again were the line to end at a series resonant circuit as at *D*, provided the input frequency were the resonant frequency of this terminating circuit.

At the two terminals of an ideal parallel resonant circuit containing no resistance there would be infinitely high impedance to an applied voltage at the resonant frequency of the

tuned circuit, and there would be zero current from one circuit terminal through to the other, although there might be large circulating current within the tuned circuit. At the parallel resonant termination of the line the condition of maximum voltage and zero current is equivalent to an open circuit, with which there is maximum voltage but zero current. Consequently, there will be total wave reflection when the end of the line is open, as at *E*.

Between the two terminals of an ideal series resonant circuit there is maximum current when applied voltage is at the resonant frequency of the tuned circuit, this because a series resonant circuit containing no resistance has zero impedance at resonance. At the same time the voltage across the ideal series resonant circuit is zero, because there is no potential difference across zero impedance. Maximum current and zero voltage exist also at any short circuit. Consequently, there is total reflection from the end of the line when the end is shorted, as at *F*.

**Standing Waves.** — Direct waves originating at the input to the resonant line are continually traveling toward the far end of the line, while waves reflected from the end are continually traveling back to the input. The waves in opposite directions cross each other. At some points the two waves will be in phase and will add their strengths to produce a strong maximum. At other points the waves will be in opposite phase and they will cancel to leave zero strength. At intermediate points there will be such phase relations as to make the combined strength something between maximum and zero.

If the electrical length of the line is some exact multiple of a quarter-wavelength corresponding to the input frequency, the in-phase peaks and opposite-phase peaks always will occur at the same points along the line and the peaks will be of maximum strength. The waves which act together to produce the maximum and zero values are in continual motion away from and toward the input, but the points where they add and cancel stand still. These stationary points where appear the maximum and minimum voltages and currents form what are called standing waves on the line.

Standing waves of current are displaced from standing waves of voltage by one-quarter wavelength along the line. This is true because in every case where there is reflection, as illustrated by Fig. 67-2, there is maximum current accompanied by minimum voltage or else minimum current accompanied

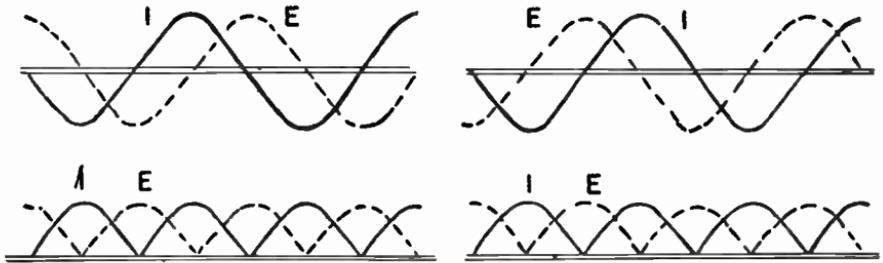


Fig. 67-3.—How standing waves may be represented in diagrams.

by maximum voltage. These phase relations of voltage and current are illustrated at the top of Fig. 67-3. Were the currents and voltages to be measured by some form of a-c meter, the readings would be the same whether the instantaneous alternating voltages and currents were positive or negative. Then the waves may be more simply and conveniently represented as at the bottom of the figure, with all the alternations above the zero line.

**Open and Shorted Lines.** — The relations between current and voltage in standing waves due to reflections cause certain lengths of open and shorted lines to act like particular combinations of inductance and capacitance. The relations for open lines are illustrated by Fig. 67-4. At the top are standing waves of current and voltage. At every point where there is maximum voltage and minimum current these relations are the same as at an open circuit or at the terminals of a parallel resonant circuit. Where there is minimum voltage and maximum current the relations are the same as at the terminals of a series resonant circuit.

The open ends of the lines in the figure are placed directly below a point having conditions like those of an open circuit. If the input to this line is  $\frac{1}{4}$  wavelength from the open end,

this input will be at a point where voltage and current relations are like those at the terminals of a series resonant circuit. Then at the input to this  $\frac{1}{4}$ -wave open line will be the same effect as though the line were a series resonant circuit, and the line may be used instead of such a circuit. If the input to the

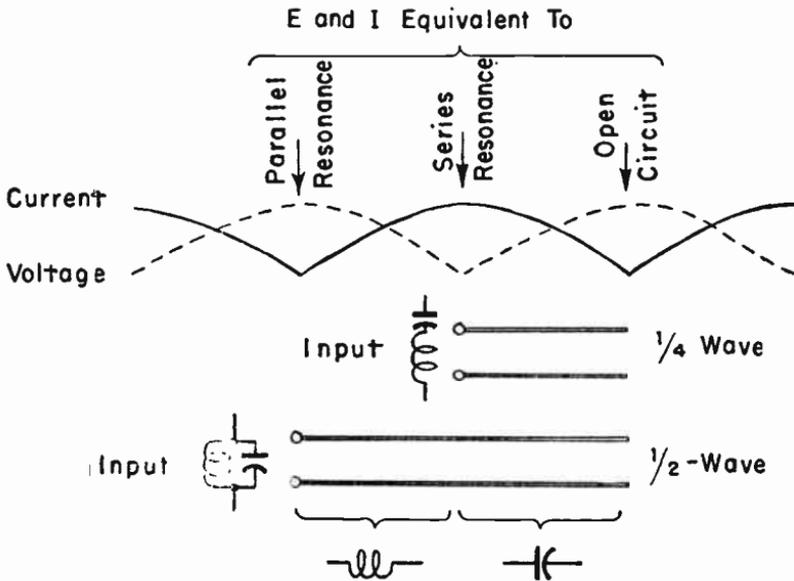


Fig. 67-4.—Relations between lengths of open lines and the kinds of circuit elements to which the lines are equivalent.

line is  $\frac{1}{2}$  wavelength from its open end, the voltage and current relations at the input are like those at the terminals of a parallel resonant circuit, and this  $\frac{1}{2}$ -wave open line may be used like such a circuit.

At the bottom of Fig. 67-4 it is shown that an open line whose length is less than  $\frac{1}{4}$  wavelength at the applied frequency acts like a capacitance or a capacitive reactance. If the length of the open line is between  $\frac{1}{4}$  and  $\frac{1}{2}$  wavelength, the line acts like an inductance or an inductive reactance.

Fig. 67-5 illustrates the effect of length of shorted lines on how the lines behave toward anything connected to their input. Here the shorted ends of the lines have been placed

directly below a point where voltage and current relations are like those at a short circuit. When the input to the shorted line is  $\frac{1}{4}$  wavelength from the shorted end, this line acts at the input as though it were a parallel resonant circuit or an open circuit. This  $\frac{1}{4}$ -wave shorted line may be used like a parallel resonant circuit. If the input to the shorted line is  $\frac{1}{2}$  wavelength from the shorted end, the line acts at the input as though it were a series resonant circuit or a short circuit. This  $\frac{1}{2}$ -wave shorted line may be used like a series resonant circuit.

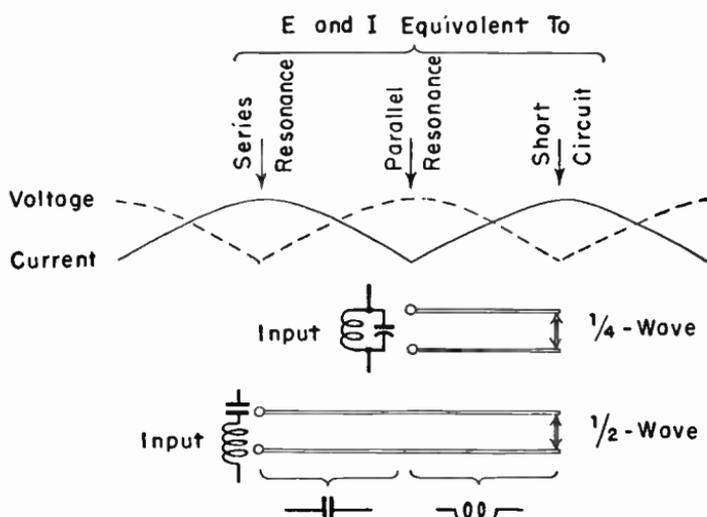


Fig. 67-5.—Relations between lengths of shorted lines and the kinds of circuit elements to which the lines are equivalent.

At the bottom of Fig. 67-5 it is shown that a shorted line less than  $\frac{1}{4}$  wave in length behaves at its input like an inductance or an inductive reactance. A shorted line whose length is between  $\frac{1}{4}$  and  $\frac{1}{2}$  wavelength behaves at its input like a capacitance or a capacitive reactance. The behavior of shorted lines is not altered by cutting off the portion of the line which extends beyond the short.

**Effects of Half-wavelengths.** — Examination of Figs. 67-4 and 67-5 shows that relations of voltage and current repeat at every half-wavelength along the line. It was shown also that

the electrical effect at the input to a  $\frac{1}{2}$ -wave open line is that of an open circuit, and at the input to a  $\frac{1}{2}$ -wave shorted line the effect is that of a short circuit. Adding or subtracting half-wavelengths to a resonant line has no effect on the electrical behavior or characteristics of the line as seen from the input end.

Characteristics have been described for lines of four lengths; less than  $\frac{1}{4}$  wave, exactly  $\frac{1}{4}$  wave, something between  $\frac{1}{4}$  wave and  $\frac{1}{2}$  wave, and exactly  $\frac{1}{2}$  wave. From any longer line it would be possible to remove exact half-wavelengths until that line was brought down to one of the lengths first mentioned. It is not necessary to remove  $\frac{1}{2}$ -wave sections physically, the line may be considered as shortened by any number of half-wavelengths, and the electrical length of the remainder will indicate the properties of the whole line.

**LINES, TRANSMISSION.** — A transmission line consists of two conductors which connect the two halves or sections of a television or f-m antenna to the receiver circuits. In types made for reception the two conductors of a transmission line are molded or embedded in low-loss insulation, such as polyethylene. In some types of line there is additional outer covering for protection against weather and mechanical damage. High quality transmission line avoids serious reduction of signal strength due to attenuation in the line, conserves signal strength by allowing correct matching of impedances, and reduces pickup of electrical interference.

Transmission lines may be either balanced to ground as at the left in Fig. 68-1 or unbalanced as at the right. Each conductor of a balanced line is connected to the inner end of one-half the antenna and, in the receiver, to one end of a coil in the receiver antenna circuit. There is a ground connection from the center tap of the coil or of a resistance or capacitance network. The two sides of the lines and everything to which they connect are symmetrical. Radiations reaching the balanced line cause voltages and currents which act in the same direction at the same time in both conductors, either upward or downward on both sides. These forces oppose each other

in the two halves of the center tapped coil or other balanced antenna coupling in the receiver, thus canceling their effects and leaving only the signal from the antenna to act in the coupling.

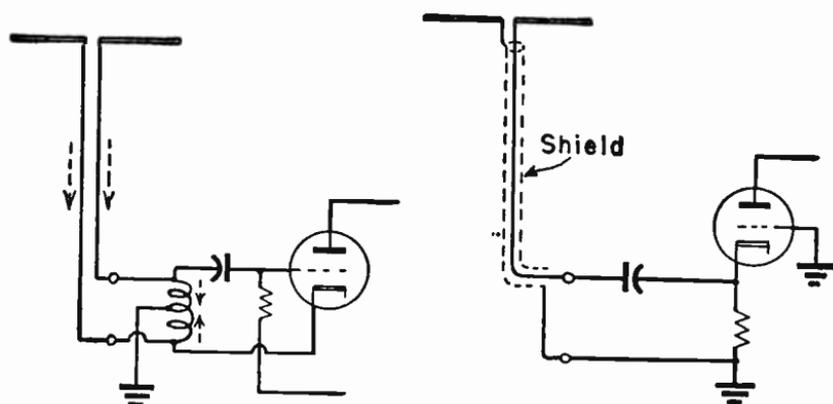


Fig. 68-1.—Connections to a balanced transmission line (left) and to an unbalanced line (right).

Cancellation of voltages picked up by a balanced line makes this type fairly immune to interference even when there is no shielding, as is the case with most balanced lines. However, where there is strong interference and electrical noise, the two signal-carrying conductors within their insulation may be enclosed by a metallic braid shield protected with other external insulation. The braid shield is connected to chassis ground and sometimes to a cold water pipe or earth ground.

A transmission line used for unbalanced connections nearly always is of coaxial construction, in which a single central conductor is embedded in insulation around which is the other conductor in the form of metallic braid. There is additional protective insulating covering outside the braid conductor. The central conductor is connected to one side of the antenna and, in the receiver, to the high side of the coupling in the r-f grid circuit. The braid of the coaxial line is connected to the other side of the antenna, to the low side of the input coupling in the receiver, and usually to ground so that

the braid acts as a shield. A coaxial line with its braid conductor grounded is not affected by external electrical fields, nor does it have any external field of its own.

**Impedance of Line.** — Any transmission line may be thought of as consisting of a very great number of very short sections as shown at the top of Fig. 68-2. On both sides of every section there is the inductance of the conductors in that section, with all the inductances in series with the line. Across each section is the capacitance resulting from the two conductors as plates, separated by insulation which is the dielectric. All the capacitances are in parallel with one another. A somewhat more exact representation is shown by the bottom diagram, where are included the series resistances of the line conductors and the shunt resistances across the conductors.

Any one section of the line offers a certain opposition to alternating current because of the series inductances and their inductive reactances. At the same time this section aids flow of alternating current because of the shunted capacitance and

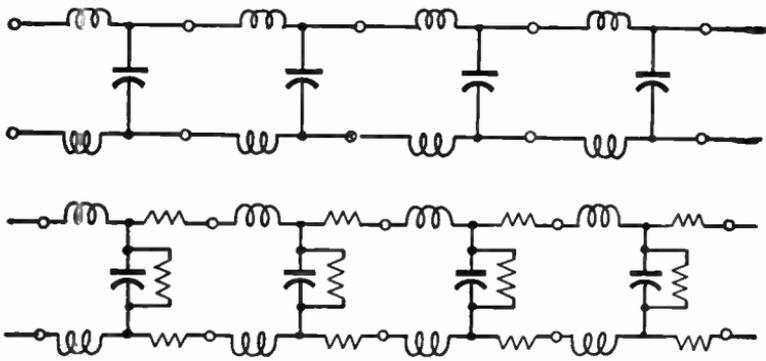


Fig. 68-2.—A transmission line consists, in effect, of many sections in which there is inductance, capacitance, and resistance.

capacitive reactance. Adding another section of line offers more opposition in the series reactance, but less in the shunt reactance, and the result is no change of impedance or in the net amount of opposition to alternating current flow. Any

number of sections may be added, and still there will be no change of impedance.

This unchanging impedance usually is called characteristic impedance. Other names are surge impedance, image impedance, and iterative impedance. This impedance depends on inductance, capacitance, and resistance in any length of line. These things depend on the size and separation of the two conductors and on the dielectric constant of the insulating material in the line. The characteristic impedance remains the same regardless of actual length of line, and it is the same regardless of frequency because inductive and capacitive reactances change oppositely with variation of frequency.

Characteristic impedance is the impedance which would exist at the input end of a line infinitely long. This impedance, in ohms, would be equal to the quotient of dividing the input alternating volts by the input alternating amperes. Line impedance could be computed from the following formula.

$$Z = \sqrt{\frac{L}{C}} \times \frac{1}{\sqrt{k}}$$

Where  $Z$  is the impedance in ohms,  $L$  is the inductance in henrys of any unit length of line,  $C$  is the capacitance in farads of the same length of line, and  $k$  is the dielectric constant of the insulating material in the field between conductors. Any change in unit length of line changes  $L$  and  $C$  together, and their ratio remains constant.

Capacitance increases with conductors closer together and of larger diameter. This lessens the impedance. Wide spacing and small conductors reduce the capacitance and increase the impedance. The greater the dielectric constant of the insulation the less becomes the line impedance.

With parallel conductors such as used for a balanced transmission line the formula for characteristic impedance is,

$$Z = 276 \times \log \frac{2 \times S}{D} \times \frac{1}{\sqrt{k}}$$

Where  $Z$  is the impedance in ohms,  $S$  is center-to-center separation of conductors in inches,  $D$  is the conductor outside

diameter in inches, and  $k$  is the dielectric constant of the insulation. The common logarithm is used.

When the line is of the coaxial type with concentric conductors, as commonly used for unbalanced line connections, the formula for characteristic impedance is,

$$Z = 138 \times \log \frac{D}{d} \times \frac{1}{\sqrt{k}}$$

Where  $Z$  is impedance in ohms,  $D$  is the inside diameter of the outer or shielding conductor in inches,  $d$  is the outside diameter of the inner conductor in inches, and  $k$  is the dielectric constant of the insulating material. The common logarithm is used.

Television receivers having balanced antenna inputs nearly always are designed with approximately 300 ohms input impedance at their antenna terminals. The transmission line for such receivers usually is of the type having insulated twin conductors, either unshielded or shielded, and having 300-ohm characteristic impedance. Unshielded line of this type most often is in the form of a flat ribbon about  $\frac{3}{8}$  inch wide and  $\frac{1}{16}$  inch thick.

Receivers with unbalanced antenna input nearly always are designed with approximately 75 ohms impedance at their antenna terminals. The transmission line should have impedance of 75 ohms. Unshielded 75-ohm twin-conductor line may be used, but the choice ordinarily is coaxial line of 75-ohm impedance which is used with the outer conductor grounded to act as a shield. There are other coaxial lines or cables having impedance of approximately 95 ohms. There are also unshielded lines having impedances of 150 ohms and of 100 ohms. Lines with impedances other than 300 ohms and 75 ohms are used chiefly for experimental work and some special applications.

**Velocity Constant.** — Radio-frequency waves traveling along transmission lines move more slowly than waves of the same frequency traveling in empty space. Consequently, the wavelength in a transmission line is shorter than in space. This means that a line whose length is specified as a certain number

of wavelengths or as quarter-wavelength, half-wavelength, and other fractions, will not have a physical length in inches or feet as long as the same number of wavelengths or fractions in space.

The fraction of the space speed at which a wave of given frequency travels in a transmission line is the velocity constant or velocity factor of the line. This constant or factor is also the fraction by which wavelength in space is multiplied to determine equivalent wavelength of the line. Some velocity constants for Amphenol cables and lines are as follows.

Twin-lead, 300-ohm impedance (unshielded)	0.82
Twin-lead, 150-ohm impedance (unshielded)	.77
Twin-lead, 75-ohm impedance (unshielded)	.69
Coaxial, 73 or 75-ohm impedance	.66
Coaxial, 53-ohm impedance	.66

For air-insulated twin-conductor lines supported with usual insulators the velocity constant will be something between 0.85 and 0.95 in most cases.

Actual line lengths corresponding to wavelengths and fractions are determined thus.

One wavelength, feet	=	983	×	factor/megacycles
inches	=	11800	×	factor/megacycles
1/2 wavelength, feet	=	492	×	factor/megacycles
inches	=	5900	×	factor/megacycles
1/4 wavelength, feet	=	246	×	factor/megacycles
inches	=	2950	×	factor/megacycles

**Matching of Impedances.** — If the impedance of a load to which a transmission line connects is not the same as the characteristic impedance of the line there will be wave reflections in the line. Signal energy which should pass into the load and be used there will be reflected back into the line. The load is the receiver, so signal energy which should go into the receiver is reflected back to the antenna and is wasted. If impedances of line and load are alike there are no reflections, regardless of the length of line, and all power or energy enters the load. When impedances are equal the line and load are said to be matched.

The transmission line is the load for the antenna, and line impedance should match antenna impedance in order to have maximum useful transfer of signal energy from antenna to line and thence to the receiver. Lack of impedance matching is the cause for weak signals at the receiver, for poor definition and apparent poor focusing, and for excessive interference or noise effects. Mismatching occurs not only where line impedance is incorrect for antenna or receiver, but wherever there are electrical changes such as high-resistance splices or couplings or material changes of spacing between line conductors.

**Line Stubs for Matching.** — When the impedance of a load, such as the receiver input, is an exact match for the impedance of the line the effect at the antenna end of the line is as though the receiver input were a pure resistance equal to line impedance. For instance, if receiver input impedance actually is 300 ohms, and if line characteristic impedance is 300 ohms, the antenna will effectively be connected to a pure resistance of 300 ohms. If antenna impedance is 300 ohms, as with connection at the center of a folded dipole, all signal energy picked up by the antenna will go through the line and into the receiver input without loss.

The receiver input may be considered as equivalent to a series resonant circuit in which inductance and capacitance cancel each other to leave only resistance. This resistance is the impedance of 300 ohms. If the receiver input is not effectively resonant it will have either excess capacitance or else excess inductance. In either case the impedance will increase, just as it would increase in a circuit having inductance and capacitance in series when the applied frequency is either higher or lower than that for resonance. If there is excess capacitance it may be balanced out by adding inductance to resonate with the capacitance, and excess inductance may be balanced with capacitance.

Inductance or capacitance may be added by connecting to the junction of line and receiver input a resonant line or a line stub. As shown in the article on *Lines, Resonant*, inductance

will be provided with either an open stub between  $\frac{1}{4}$  and  $\frac{1}{2}$  wavelength in electrical length or else a shorted stub less than  $\frac{1}{4}$  wavelength in electrical length. Capacitance will be provided by an open stub less than  $\frac{1}{4}$ -wave in length or by a shorted stub whose length is between  $\frac{1}{4}$  and  $\frac{1}{2}$  wavelength.

Ordinarily there is no apparatus available for determining the kind of mismatch, and even were measurements possible it would be quicker and more certain to try stubs of different length, with their free ends open and shorted, and to use whatever gives the most improvement. The stub is connected to the receiver antenna terminals along with the transmission line, as at the left in Fig. 68-3. It is easiest to experiment with an open stub cut initially to about half-wavelength in the channel where there is greatest evidence of faults which may be due to mismatching. While the stub is connected to the receiver and a test pattern is tuned in, pieces about one-quarter inch long may be cut from the free end of the stub until there is improvement and until reception again commences to become poor. Then a permanent stub may be cut to the length at which best results were obtained. Experiments with a shorted stub may

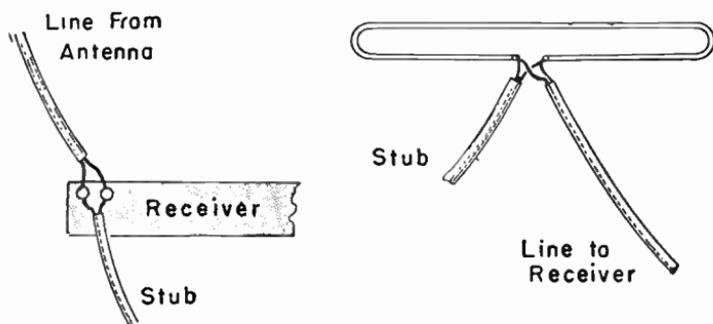


Fig. 68-3.—Matching stubs connected at the receiver and at the antenna.

be carried out by arranging pin points or needle points which can be pushed through the insulation into the conductors at various points.

Matching stubs may be used at the antenna end of the line, connected to the points at which the line conductors are soldered or screwed to the ends of the antenna rods or to the

link between upper and lower sections of a stacked array or between sections of a two-band antenna. Connection at an antenna is shown at the right in Fig. 68-3.

**Line Sections for Matching.**—Unlike impedances of antenna and transmission line or of transmission line and receiver may be matched with a section of line  $\frac{1}{4}$  wavelength long at the frequency for which the best match is desired. The impedance of the  $\frac{1}{4}$ -wavelength matching section is made equal to the square root of the product of the unlike impedances which, in the following formula, are designated as  $Z_a$  and  $Z_b$ . All impedances are in ohms.

$$\text{Matching impedance} = \sqrt{Z_a \times Z_b}$$

The most common matching problems occur with simple dipole 73-ohm antennas, folded dipole 300-ohm antennas, and receivers having inputs designed for 75 ohms or for 300 ohms.

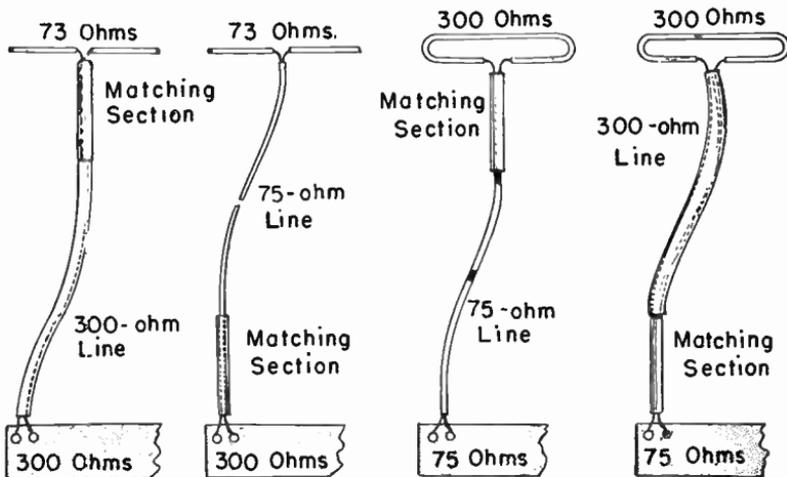


Fig. 68-4.—Connections of matching sections at the antenna end and at the receiver end of the transmission line.

The formula will show that a matching section for joining 73 or 75 ohms to 300 ohms should have impedance of approximately 150 ohms, which is the impedance of a generally available twin-conductor line. Fig. 68-4 shows several applications of such a matching section.

**Installing Transmission Lines.** — Loss of signal strength in the transmission line is directly proportional to line length, and for this reason the length should be as short as possible. However, the antenna should be located for best signal pickup and least interference even though such a position calls for a relatively long transmission line. Much of the loss in a long line may be compensated for by good impedance matching and by careful attention to installation methods.

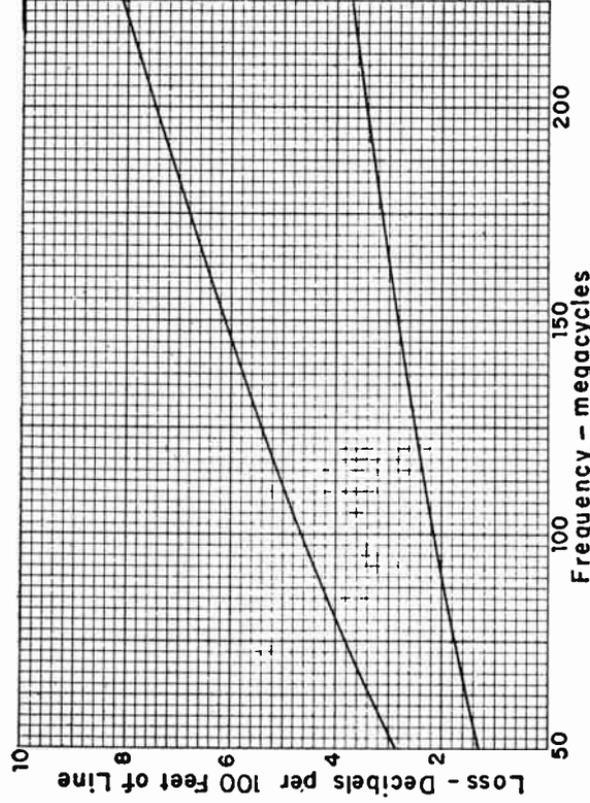


Fig. 68-5.—Average losses of signal energy in transmission lines of 75-ohm impedance (top curve) and of 300-ohm impedance (bottom curve).

Fig. 68-5 shows signal loss in decibels per 100 feet of unshielded twin-conductor transmission line. The upper curve, showing the greater loss, applies to line having characteristic impedance of 75 ohms. The lower curve applies to line of 300-ohm impedance. The loss in coaxial lines of approximately 53 and 75-ohm impedance varies widely with construction and size of the cable, but generally is within the limits of the two curves shown in Fig. 68-5.

Transmission line of any type should be fastened or supported every 6 to 8 feet in such a manner that it will not sway in the wind. Enough slack should be left in a short length at the antenna end to prevent pulling or breaking the line in case the antenna sways. Any excess length should be cut off, not folded or rolled out of the way. There should be no very sharp bends which might cause changes in conductor spacing and allow reflections. The line conductors are preferably soldered to the antenna terminals.

Coaxial and other shielded transmission line may be run almost anywhere that is convenient. It may be clamped or taped to a metal antenna mast, run inside of metal conduit or piping, in wall spaces, or in or on top of earth. Excessive heat will damage the insulation. Connections are made to coaxial line by removing the outer insulation far enough to permit exposing the inner conductor and some of the braid. The braid may be unraveled, then twisted and tinned with solder for connection to the ground terminal, or the braid may be left in place while tinned with a hot iron quickly enough to prevent damage to the insulation. Then a ground wire may be soldered to the tinned braid. The least possible length of inner conductor should be exposed beyond the shielding braid.

Unshielded transmission line should be kept from direct contact with building surfaces, even though the surfaces are nonconductors. This kind of line must be kept well separated from power lines and all electrical machinery, also from all metal. Metal includes gutters and downspouts, roof vents, conduits, water and gas piping, and the antenna mast. Unshielded line must not be run inside of conduit or metal pipe of any kind. Supports should be standoffs made entirely of insulation or with insulation held by metal. If standoffs having a metal loop around a slotted insulator are used on a line running vertically or nearly so, the loops may be squeezed down onto the insulation and line after everything is in final position.

Unshielded lines should be fastened at a point near the antenna, then carried away from the antenna mast at a slight angle. Interference pickup is lessened by twisting the line

about one full turn per foot of length. Horizontal runs should be as short as possible. One reason is that a horizontal run is subjected to the same kind of signal pickup as the horizontal antenna conductors. Another reason is that a horizontal line, unless protected by some building projection, tends to collect rain, snow, condensed moisture, and dirt. These things then are in the field space between the conductors and cause temporary or permanent changes in line impedance.

Shielded line can be spliced satisfactorily, from the transmission standpoint, only with connectors made especially for this purpose. Unshielded twin-conductor line may be spliced with soldered joints provided the conductor spacing is not altered, the least possible solder is used, and excess conductor is carefully removed. A soldered splice should be protected with wrapping of plastic tape. Most connectors made for unshielded twin-conductor cable are not weather-resistant and are not satisfactory for outdoor use. It is far better to use continuous runs rather than any kind of splice.

A line should be brought indoors through a hole sloping downward toward the outside. Unshielded line should be protected with insulating tubing large enough not to change the shape or conductor spacing of the line. Just outside the building there should be a downward drip loop in the line. Space remaining in the opening may be filled with mastic or caulking compound.

Unshielded line indoors must not be run near radiators, pipes, or other large metal objects, whether these objects are exposed or are concealed in walls and floors. A line may be run across basement joists and up through a floor near the receiver. The line should be kept out of sight so far as possible by running it behind moldings and in similar recesses, but not by placing the line underneath carpets or rugs. Twin conductor line may be fastened with a few very small tacks or brads driven through the insulation midway between the two conductors.

**Uhf Transmission Lines.** — Energy losses from flat twin-conductor transmission line increase materially when the line is wet or dirty, and may become serious with weak uhf signals.

Moisture or dirt is in the field space which surrounds the flat conductor. The energy loss may be decidedly reduced by using tubular transmission line of 300-ohm impedance. This style has the two conductors molded into opposite sides of a tube of dielectric material about  $\frac{3}{8}$  inch in outside diameter. The tubular form maintains an internal open air space between conductors and a longer external surface between conductor positions. Energy loss per 100 feet of the flat and tubular lines is about equal when both are dry and clean.

Energy loss only a small fraction of that in lines having conductors molded in dielectric material is obtained with open wire line having the two conductors supported only by insulating or dielectric spacers at intervals of about six inches. Commonly available line of this kind is about 1 inch wide, between conductors, and has characteristic impedance of 450 ohms. For best results the impedance should be matched at the antenna and at the receiver.

Energy loss in shielded twin-conductor line and in coaxial cables is much greater per foot than in any of the unshielded lines mentioned in preceding paragraphs. In spite of this it may be necessary to use some form of shielded line where there are likely to be many and strong signal reflections.

It is even more important in uhf than in vhf installations to have the transmission line as short as possible, and, unless of shielded type, to keep the line well away from all large metal objects. Insulating supports should be provided close enough together and in such positions that the line remains taut throughout its length. The usual precautions must be observed in making clean, tight terminal connections and in protecting them from weather effects. When tubular transmission line is used, both ends must be tightly sealed against entrance of moisture. If a drip loop is used near the receiver end of such line the tubing should be pierced with a small opening at the bottom of the loop to allow escape of any internal moisture which may condense there.

**LINEARITY AND LINEARITY CONTROLS.** — Linearity refers to positions and shapes of lines and objects in reproduced pictures, in relation to positions of lines and objects in the original image. If proportions are the same in the reproduction as in the original, horizontally and vertically, there is good linearity, otherwise it is poor. With poor linearity there may be displacements at the center or elsewhere, circles may be distorted, and crowding or stretching may be evident at various places.

Usually there are adjustable controls for correction of non-linearity. There may be one or more than one such control in either the horizontal or vertical sweep section, or in both sections. Some of these controls may affect chiefly the right side and others the left side, while some may affect chiefly the bottom and others the top of the pattern or picture. There

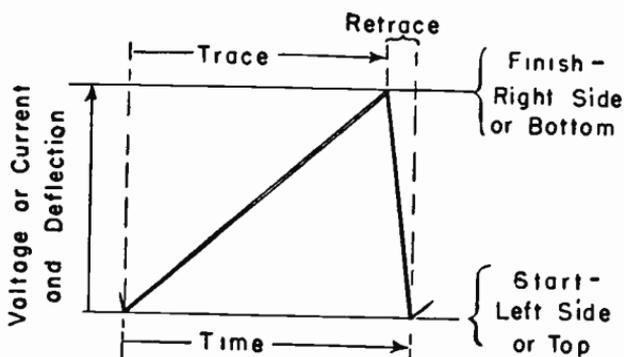


Fig. 69-1.—Relations of sawtooth wave slopes to deflections of the electron beam.

may be no controls marked specifically as linearity adjustments, the same purpose being served by drive controls or by peaking controls in some receivers. Linearity may be affected by adjustment of controls for width or height, for centering, and sometimes by adjustment of hold controls. These should be checked before changing the linearity adjustments. Always there is the possibility that the pattern from one station may be temporarily non-linear, so more than one pattern should be observed before changing adjustments.

Linearity or lack of it depends largely on the form of the sawtooth wave of voltage or current that deflects the picture tube beam. If the slope of the wave is a straight line, as in Fig. 69-1, the reproduction should be linear. During the time that sawtooth voltage or current increases, the beam is deflected from left to right or from top to bottom of the picture tube screen for a horizontal or a vertical trace. The rate of change of voltage or current should be constant in relation to time.

At the left in Fig. 69-2 the sawtooth is shown as rising too rapidly at the beginning. When half the time period for a horizontal trace or a vertical trace has elapsed, the deflection voltage or current will have gone through more than half its total increase and the beam will have been moved more than half the distance from left to right or from top to bottom of the mask area. At the left, the sawtooth rises too slowly at the beginning and too rapidly at the end. Now, by the end of half the time period for a horizontal or vertical trace, the

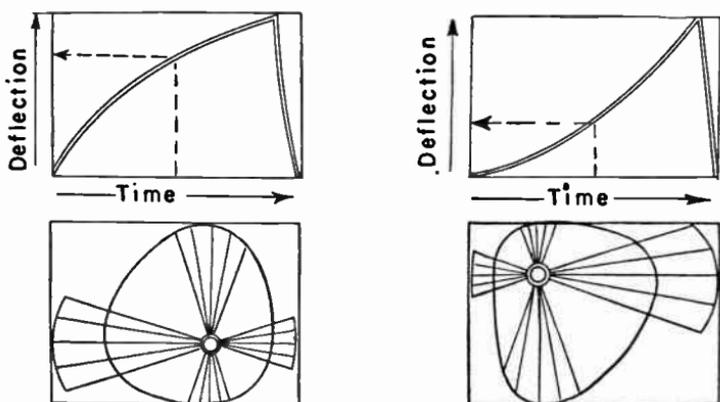


Fig. 69-2.—How curvatures of the sawtooth slope affect linearity of a pattern.

deflection voltage or current will not have gone through half its total change. The picture tube beam will not have been deflected far enough, and whatever should appear at the center of the screen actually will be too far up or too far to

the left, depending on whether the fault is in the vertical or horizontal sweep sections.

The sawtooth curve at the left in Fig. 69-2 is of nearly the same form as the charging curve for a capacitor subjected to a constant charging voltage. The sawtooth curve at the right is like the mutual characteristic for an amplifier tube having an excessively negative bias and working down near the lower bend of its characteristic. A common style of linearity control is based on balancing these opposite curvatures against each other to obtain a sawtooth curve with a straight slope. This control varies the resistance in series with the cathode of the sweep amplifier or deflection amplifier to change the grid bias and vary the points on the characteristic curve between which the tube operates.

Adjustment of amplifier bias to change the point of operation on the characteristic curve not only compensates for opposite curvature of a capacitor charge, but also for the similar opposite curvature due to inductance. A curve similar to that at the left in Fig. 69-2 would show the manner in which current increases in an inductance which is subjected to a constant voltage. Such effects on currents may occur in deflection transformers and deflection coils.

**Horizontal Linearity Controls.** — Fig. 69-3 shows two types of horizontal linearity controls used in connection with horizontal sweep transformers having insulated primary and secondary windings. Either or both controls may be found in the same receiver. The more common type is the adjustable inductor on the line from the damper cathode. Varying this inductance alters the time at which the damper becomes conductive, to prevent continued oscillation in deflecting coils, and alters the waveform of deflecting current in these coils. The control inductor and capacitors at its ends cause some phase shift in voltage applied through the output transformer to the plate of the sweep amplifier. Hence an adjustment alters the plate characteristic, also curvature of the mutual characteristic which relates plate current to grid voltage in the amplifier.

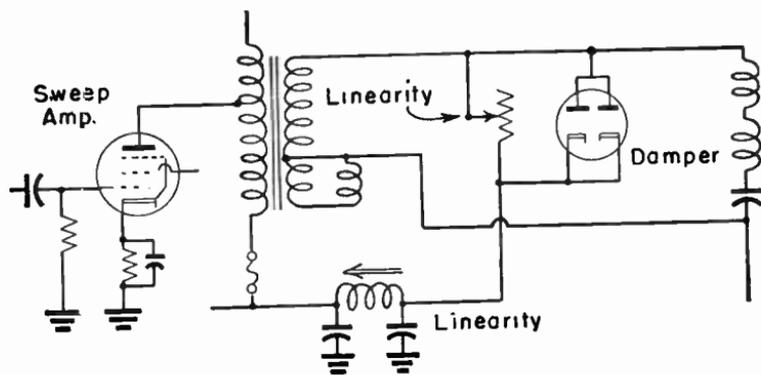


Fig. 69-3.—Horizontal linearity controls in the plate and cathode circuits of a damper tube.

The adjustable resistance linearity control of Fig. 69-3 shunts the damper tube. It changes the rate of damper conduction or the period during which there is conduction, and thus alters the time during which there is decay of oscillatory current in the yoke coils.

Fig. 69-4 shows connections for a horizontal linearity

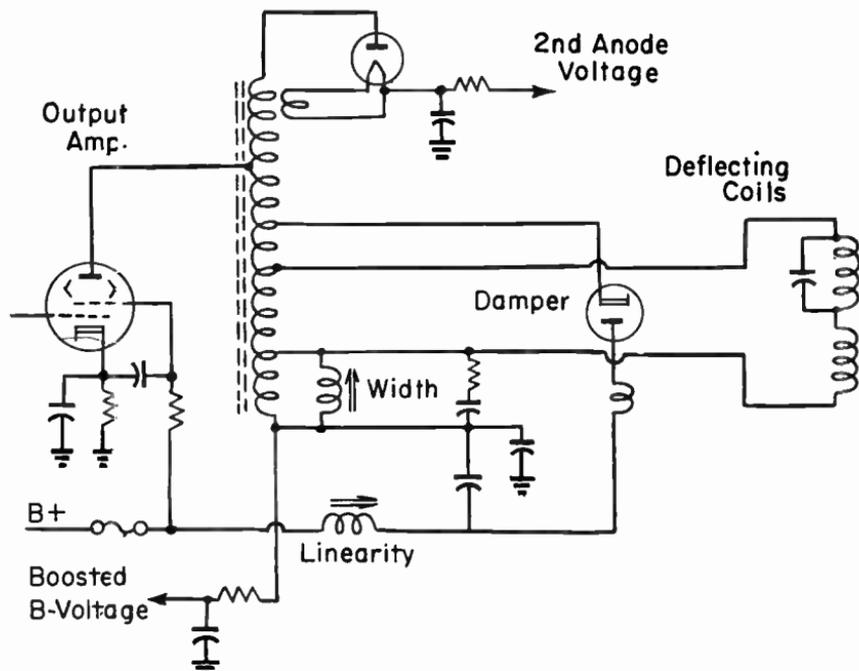


Fig. 69-4.—A linearity control in a horizontal sweep system employing an autotransformer.

control inductor used with an autotransformer for output coupling between amplifier and yoke coils.

A horizontal linearity control should be adjusted at the same time as a horizontal drive control and a width or horizontal size control. All three controls alter picture width. While the horizontal linearity control has maximum influence on correct proportions, the drive and width controls may affect linearity to a greater or less extent.

**Vertical Linearity Controls.** — Vertical linearity commonly is adjusted by a variable cathode bias on the vertical sweep amplifier. Two such controls are shown in Fig. 69-5. These adjustments alter the shape of the grid-voltage plate-current characteristic of the amplifier. They will correct either undue compression or expansion in portions of the picture above the center, and will make the upper part of pictures proportional to or linear with respect to the bottom part.

A vertical linearity control always is adjusted in conjunction with a height or vertical size control. Both controls alter picture height, and they must be adjusted together for correct height combined with good vertical linearity.

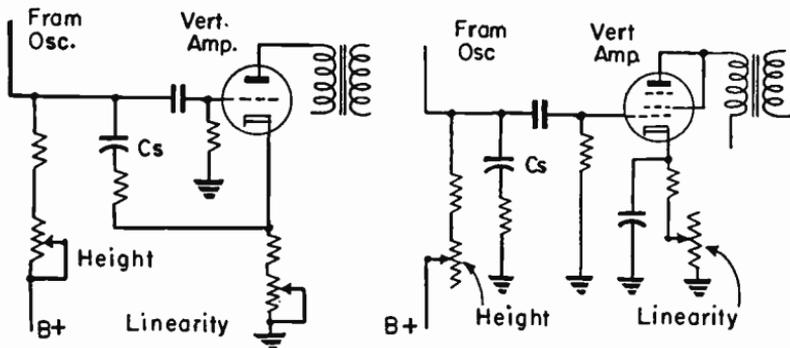


Fig. 69-5.—Linearity controls on output amplifiers in a vertical deflection system.

**Adjusting Linearity.** — Should pictures appear non-linear, proceed with adjustment as follows:

Make sure that the contrast control is not advanced so far as to distort pictures. Adjust vertical and horizontal hold controls midway between positions at which pictures drop out of synchronization. Check the effects of horizontal drive and size controls, and of a vertical size control, before altering linearity controls. A misadjusted horizontal hold control may cause poor linearity at the sides of pictures. Adjust linearity controls while observing pictures having minimum motion, such as some commercials. Better, use a pattern generator furnishing vertical lines for checking horizontal linearity and horizontal lines for vertical linearity, or a cross-hatch pattern for simultaneous checks.

**LOSSES, HIGH FREQUENCY.** — In many types of circuits the energy losses are greater with alternating current than with direct current, and increase with frequency. The additional energy is considered a loss because it produces heat rather than doing anything useful. An equal loss of energy and production of heat might result from addition of resistance to a circuit. Consequently, the actual loss may be called high-frequency resistance or effective resistance or a-c resistance. This a-c resistance would not be the same as impedance, because impedance is the combined effect of reactance and d-c resistance in a circuit. The principal causes for high-frequency loss are mentioned in following paragraphs.

*Dielectric loss* or dielectric hysteresis occurs in dielectric materials subjected to an alternating electrostatic field. The electrons and the nucleus of atoms are displaced in opposite directions within the atoms. The displacements or strains reverse with alternations of the field. The result is dielectric heating.

Energy loss results from distributed capacitance in coils and circuit conductors when this capacitance resonates with coil or circuit inductance to cause oscillating currents at the resonant frequency of this capacitance and inductance, which ordinarily would not be the frequency for which the circuit is intentionally tuned. Heating and energy loss result from flow of the spurious oscillating currents in the conductor

resistance. There is no loss of energy in capacitance or inductance because all energy used for charging a capacitance and all energy used for producing a magnetic field around an inductance are returned to the circuit and source during discharge of the capacitance or collapse of the magnetic field.

Alternating current in any conductor causes self-induction or counter-emf in the conductor. This action is due to magnetic lines of force spreading outwardly from and contracting back into the conductor with alternations of current. The effect is least at the surface of the conductor, for metal at the surface is subjected only to magnetic lines from the interior, while metal inside is subjected to cutting by lines from both the center and outside. Because counter-emf is least at the surface of the conductor, the current tends to concentrate at the surface rather than flowing uniformly in all parts. The greater current density near the conductor surface causes increased heating and energy loss. This action is called *skin effect*.

Skin effect is worse in coils than in straight conductors. Loss due to this effect varies approximately with the square root of the frequency, and is serious at all frequencies in excess of about 3 megacycles. The effect is more pronounced in conductors of large diameter than in those of smaller diameter, because with larger diameter there is less surface area in proportion to the volume or the cross sectional area. Since the inner portion of a conductor is of little use at high frequencies, thin walled tubing may be used, or some highly conductive metal like silver may be plated on a core of other metal.

When alternating or varying lines of magnetic force cut through any conductive metal the cutting induces emf's and circulating currents in the metal. The currents are called *eddy currents* because they circle or eddy around the magnetic field lines. Flow of eddy currents in the resistance of the conductor wherein they are formed causes heating and energy loss. This loss increases with conductor resistance, so is greater in iron and steel than in metals such as copper and aluminum.

When magnetic cores such as used in coils and transformers are subjected to alternating magnetic fields the magnetic polarity of the atoms undergoes continual reversal. Energy is required for this magnetization first one way and then the other, and to destroy the magnetization in one polarity as the opposite polarity is established. Magnetic reversals lag behind the reversals of field lines. This lag is called magnetic hysteresis. The accompanying loss of energy is called a *hysteresis loss* or a hysteric loss.

Energy loss does not always occur exclusively in the circuit wherein is the alternating current, it may occur in any nearby circuit or conductors or dielectrics. Magnetic fields around the current-carrying conductor may cause eddy currents in other conductors, or they may displace the currents in those other conductors due to an effect similar to skin effect. Energy wasted in the nearby parts must be furnished by the first current-carrying conductor and its circuit. Consequently the effect of high-frequency loss appears in this first circuit even though most of the heating appears somewhere else.

## M

**MAGNETIC CIRCUITS.** — A magnetic circuit is a complete or closed path in which may act the magnetic force or lines of force produced by a permanent magnet or by a current-carrying winding. A magnetic circuit is like an electric circuit in several ways. Although nothing actually moves, the magnetic flux or magnetic lines of force are thought of as moving in the magnetic circuit, as current moves in the electric circuit. The permanent magnet or current-carrying winding furnishes magnetomotive force, which is analogous to electromotive force. Magnetic flux starts from one of these sources, passes through all the circuit, and returns to the source. Magnetic flux is opposed by reluctance, as electric current is opposed by resistance. Calculations for magnetic circuits cannot be so accurate as for electric circuits because flux cannot

be closely confined and because there is wide variation of reluctance with change in quantity of flux.

Except for leakage, flux is the same in all parts of a series magnetic circuit. Total reluctance is the sum of the reluctances in all the series parts. Flux is proportional to the ratio of magnetomotive force to reluctance, or to magnetomotive force divided by reluctance, just as electric current is proportional to electromotive force divided by resistance. The unit of flux is the line or the maxwell. One maxwell is equal to one line.

The unit of magnetomotive force from a current-carrying winding is the ampere-turn. Another unit is the gilbert, equal to 0.796 ampere-turn. The number of ampere-turns is equal to the product of number of winding turns and the number of amperes flowing in the winding. Magnetomotive forces add, like emf's, when there is more than one magnetic source in a circuit. There is drop of magnetomotive force in a magnetic circuit, just as there is drop of voltage or potential in an electric circuit.

Reluctance, like resistance, increases with the length of the circuit and decreases when there is greater cross sectional area. Reluctance is the same or has the same value in air, other gases, all liquids, all dielectrics and insulators, and in all metals and other solids except those classed as magnetic materials. Magnetic materials include iron, steel, alloys containing iron, and a few other special alloys which have the magnetic properties of attraction and repulsion. Nickel, cobalt, and gadolinium are weakly magnetic. In magnetic materials, especially in iron, steel, and iron alloys, the reluctance is only a small fraction of that in non-magnetic materials. If there is a gap in the iron or steel of a magnetic circuit the reluctance of the gap is so great compared with that of the iron or steel that the length of iron or steel in the circuit has practically no effect on quantity of flux.

The property of a material which determines the number of lines per square inch cross section for any given magnetomotive force per inch of length is called the permeability of

the material. The number of ampere-turns per inch is determined by dividing the total ampere-turns by the length of magnetic path in inches. This ratio is called magnetizing force. Its symbol is the capital letter **H**. The number of lines per square inch of cross section is called magnetic flux density. Its symbol is the capital letter **B**. The ratio  $B/H$  is the permeability of the material at the assumed working conditions.

**MEASUREMENTS, HIGH-FREQUENCY.** — Measurements of capacitance, inductance, reactance, effective resistance, and other circuit quantities existing at high frequencies are conveniently measured by resonance methods. A common type of setup is shown by Fig. 72-1. A signal generator coupled to the tested circuit through a loop or a small coil furnishes energy in the ranges of frequency where performance is of interest. A satisfactory resonance indicator is an electronic voltmeter equipped with a detector probe. With the meter connected across the tested circuit, as shown, resonance is indicated by maximum voltage. Another type of indicator may be made from a 20,000 ohms per volt meter, or any 50 to 100 microampere meter, with a crystal diode rectifier and resistance of about 5 megohms in series. See under *Meters, Rectifier*.

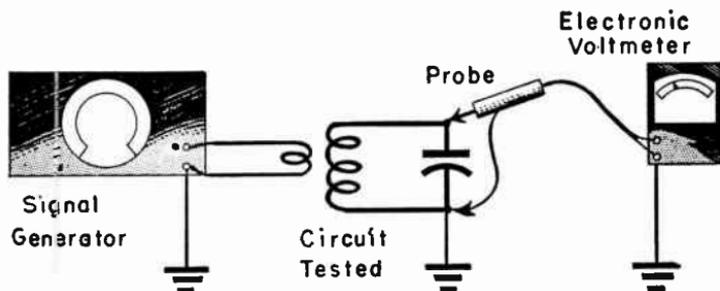


Fig. 72-1.—Determining resonance from maximum voltage developed in a tuned circuit.

The three quantities which enter into measurements are frequency, capacitance, and inductance. With any two of these known, the other may be computed. This means that an unknown inductance may be measured with a known capacitance

and generator frequency, or an unknown capacitance may be measured with a known inductance and frequency, or the resonant frequency may be determined for any combination of capacitance and inductance.

If the resonant circuit cannot be made up with inductance and capacitance in parallel, as illustrated, they may be connected in series with each other and with a fixed carbon or composition resistor of 5 to 10 ohms or more. The electronic voltmeter then is connected across the resistor. There will be maximum series current at resonance, and the voltmeter will give maximum reading.

Formulas for the various computations are as follows:

$$\text{Inductance, } \frac{25\ 530}{\text{microhenrys}} = \frac{25\ 530}{(\text{frequency, megacycles})^2 \times \text{micro-microfarads}}$$

$$\text{Capacitance, } \frac{25\ 530}{\text{micro-microfarads}} = \frac{25\ 530}{(\text{frequency, megacycles})^2 \times \text{microhenrys}}$$

$$\text{Frequency, } \frac{25\ 530}{\text{megacycles}} = \sqrt{\frac{25\ 530}{\text{micro-microfarads} \times \text{microhenrys}}}$$

In any of these formulas the frequency may be changed to kilocycles provided the capacitance is changed to microfarads in the same formula. The formulas apply alike to parallel resonance and series resonance.

Measured inductors or coils must be placed where there is a minimum of either other conductors or dielectrics in their fields. If the capacitor has a shield, ground the shield. Otherwise ground the rotor plates of an adjustable unit. If possible to use large capacitance do so, for then this known capacitance will largely overcome the effects of unknown stray and distributed capacitances. All leads and connections have appreciable capacitance and inductance, so all of them should be short and their positions and separations should not be altered during tests.

Use the weakest coupling from the signal generator which allows readable indications on the meter. Always retune to resonance after making any change whatever in any of the

parts or connections. Be sure to tune the generator either for fundamentals or harmonics in all parts of the same test, not to a fundamental one time and a harmonic the next time. If retuning the generator to some approximately integral multiple of the original frequency gives a meter reading, the origi-

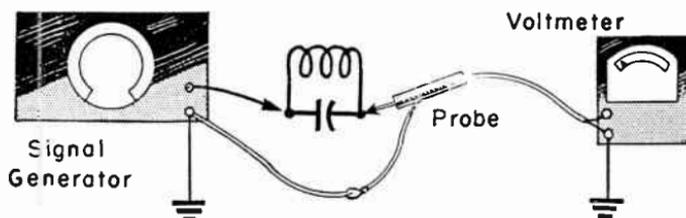


Fig. 72-2.—Determining resonance from impedance of a tuned circuit.

nal frequency is a fraction of the resonant frequency of the inductor-capacitor combination, and the tuned circuit is responding to a harmonic frequency from the generator. Always use the highest generator frequency at which there is a response from the tuned circuit and the resonance indicator.

Tests may be made without a coupling loop or coil by using the connections shown in Fig. 72-2. The high side of the signal generator is connected to one side of the resonant circuit. The opposite side of the resonant circuit is connected to the high side of the electronic voltmeter. The low sides of both generator and voltmeter are connected together and to ground, but not to the resonant circuit. With these connections the resonant circuit offers maximum impedance to current between generator and voltmeter at the resonant frequency. The generator is adjusted to the required frequency and either the capacitance or inductance of the resonant circuit is adjusted for minimum reading on the meter, or with any existing adjustment of capacitance and inductance the generator is tuned for minimum reading on the meter. The minimum reading indicates resonance.

**METERS, ELECTRONIC VOLT (VTVM).** — Usual types of electronic voltmeters (vacuum tube voltmeters) employ a bridge circuit in which two arms are triodes. Definite fractions of measured voltages are applied to one grid circuit. This unbalances the bridge. Current due to unbalance flows in an indicating meter calibrated for direct readings of measured voltages. Divider circuits for voltage ranges are of high resistance, and the negatively biased bridge tube offers high impedance. Uniform input resistance of the instrument is so many megohms as to cause almost negligible loading of circuits across which connected.

Electronic voltmeters have utilized a great variety of principles and many different methods of applying direct, alternating or rectified potential differences to the grid circuit of a

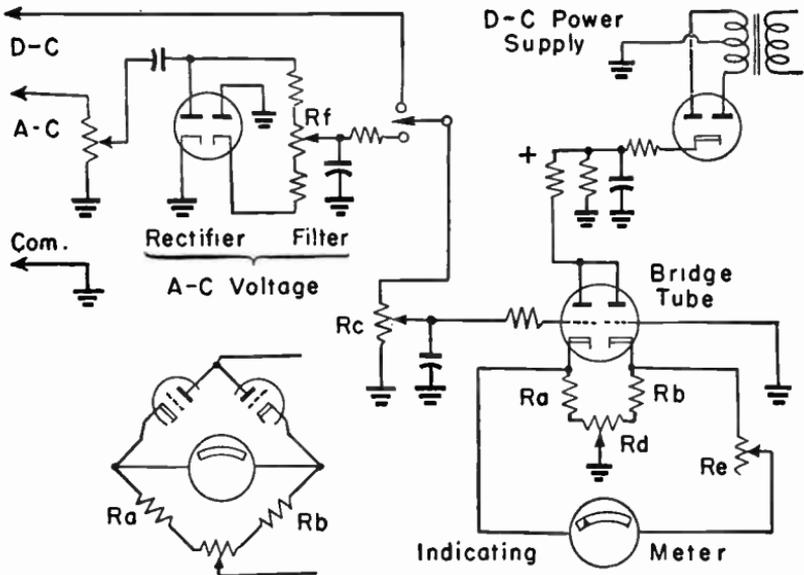


Fig. 74-1.—Typical circuits used in electronic voltmeters or vacuum tube voltmeters.

tube, and of measuring resulting changes of plate current. Fig. 74-1 illustrates features found in many present designs. The moving coil indicating meter is connected across a balanced bridge circuit whose electrical equivalent is shown at

the lower left in the diagram. Two of the bridge arms consist of resistors  $R_a$  and  $R_b$ . The other two arms are the two sections of the twin-triode bridge tube. The positive side of the d-c power supply is connected directly to the plates of the bridge tube, while the negative side connects through ground to resistors  $R_a$  and  $R_b$ .

A direct potential difference to be measured is applied between the left-hand grid of the bridge tube and ground through any conventional type of range control resistors, represented in simplified form at  $R_c$  in the diagram. The right-hand grid is grounded. With no potential difference applied to the grid-cathode circuit of the left-hand section, the zero adjustment potentiometer  $R_d$  is set for equal plate-cathode currents in the two sections of the bridge tube. Then there is no potential difference across the meter, and the meter reads zero. Resistor  $R_e$  is a calibration adjustment, of which there may be several in the complete instrument.

When potential difference to be measured is applied to the grid circuit of the left-hand section of the bridge tube, the bridge becomes unbalanced because this section then carries more or less current than the other section. Current equal to the difference of those in the two sections then flows in the indicating meter. Not shown in the diagram is a reversing switch usually installed for allowing the meter to read up scale whether d-c input to the instrument is positive or negative with reference to ground.

For measurement of alternating potentials in the range of power and audio frequencies, the a-c input to the instrument is rectified by the twin-diode rectifier at the upper left in Fig. 74-1. The d-c filtered output potential from this rectifier system is applied to the grid circuit of the bridge tube. The a-c balancing potentiometer  $R_f$  is for maintaining the pointer of the indicating meter at zero when changing from one voltage range to another. Radio-frequency voltages to be measured are rectified by a tube or a crystal diode mounted at the end of a cable long enough to bring the rectifier close to the point where measurement is desired, with the rectified voltage

brought through the cable to the bridge circuit of the electronic voltmeter. These accessories are described in the article on *Probes, Detector*.

The input resistance of an electronic voltmeter as used for d-c measurements consists of the total resistance in the voltage divider system which provides the several voltage ranges, plus a fixed resistance incorporated in the outer end of the test prod. This fixed resistance usually is something between 1 and 15 megohms, and the instrument resistance most often is on the order of 10 to 12 megohms.

Input impedance for alternating voltage measurements would vary with frequency. Instead of giving this impedance as a certain number of ohms it is specified as some certain number of megohms resistance, shunted with capacitance of some specified number of micro-microfarads. From the two values it is possible to compute approximate effective impedance for any frequency. The shunt capacitance lowers the input impedance very materially at high frequencies. The ratio between a power frequency and a frequency of 100 megacycles often is on the order of 500 to 1. A-c input resistances commonly are between 1 and 10 to 15 megohms, with shunting capacitances of 10 to 200 mmf. The input capacitance may be enough to detune any resonant circuit to which the instrument is connected. For precise measurement, the measured circuit sometimes is retuned to resonance while the meter is connected, then tuned back to the original settings after tests are completed.

**Using the Electronic Voltmeter.** — Although most electronic voltmeters are designed so that application of excessive voltage will not overload the indicating meter, it is well to avoid excessive voltage, just as with any other type of voltmeter. The instrument should be allowed to warm up for 5 minutes or more after it is turned on and before measurements are made. Shifting of zero during tests usually indicates too little preliminary warmup time. As a general rule, all voltages should be measured with reference to chassis ground or B-minus, with the ground terminal of the meter connected to chassis metal

or B-minus, the high side to the point whose potential is to be measured. To determine potential differences between two points when neither is ground or B-minus, make separate measurements to each point and then subtract voltages which are of the same sign or add those of opposite sign. That is, subtract when both voltages are positive or both negative, and add when one is positive and the other negative.

When measuring an alternating potential where a higher direct potential is present it is advisable to protect the meter with a series capacitor as shown at C in Fig. 74-2. Here the measurement is of a-f voltage in a plate circuit where the direct B+ voltage is relatively high. The series capacitor must have great enough capacitance that its reactance is reasonably low, say 5,000 to 10,000 ohms at most, and its d-c working voltage rating must be well above the maximum direct voltage in the tested circuit.

In an electronic voltmeter of usual design the basic measuring circuit operates from a direct potential whose value would be the equivalent of the peak potential of an alternating volt-

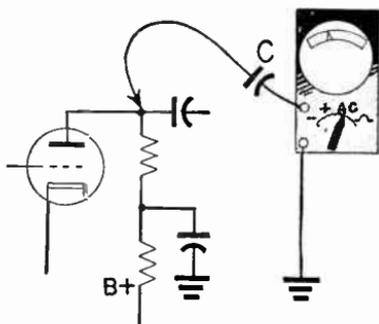


Fig. 74-2.—Protecting the voltmeter against high direct potential.

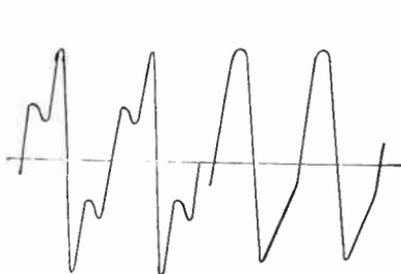


Fig. 74-3.—Waveforms which may cause errors in indications of electronic voltmeters.

age. When alternating potential actually is measured it is applied first to a rectifier, as in Fig. 74-1, whose direct voltage output goes to the measuring circuit. Although the measuring circuit is responding to peak potential, the dial of the indicating meter is calibrated and marked for effective or r-m-s

alternating voltage of sine wave form. If the measured voltage is not a sine wave, but of some other form such as that at the left in Fig. 74-3, the indicated voltage will not be the effective value of the actual wave. The wave at the left in Fig. 74-3 is symmetrical on its positive and negative alternations, and voltage indications will be the same no matter which way the meter leads are connected to the circuit in which is the measured voltage. The wave at the right is not symmetrical, and reversing the input connections will change the indicated voltage with most of the electronic voltmeters in common use.

The high input resistance and impedance of the electronic voltmeter suit this instrument for many measurements and tests which are difficult or impossible with other kinds of meters. D-c measurements may be made of grid bias voltage, of automatic volume control and automatic gain control voltages, and of r-f oscillator action. Audio-frequency measurements may be made of grid signal voltages, plate signal voltages, stage gains, and the instrument may be used as an output meter. With a detector probe it is possible to measure grid signal voltages, plate signal voltages, and stage gains in r-f, video i-f, and video amplifier sections, also to make resonance tests such as described in the section on *Measurements, High-frequency*. Other important uses include alignment of r-f oscillators, peaked couplers in i-f amplifiers, and sound discriminators or ratio detectors.

The value of a-f or r-f current in a circuit may be determined with fair accuracy from measurement of voltage drop across known resistances or reactances. The electronic voltmeter may be used to measure the voltage across any resistor which is in series with the circuit whose current is to be measured. Multiplying the indicated number of volts by 1,000 and dividing by the number of ohms resistance in the resistor then gives the number of milliamperes of current. The resistor should be a carbon or composition type, free from inductive reactance.

High-frequency current may be determined also by measuring the voltage drop across a capacitor of known capacitance

which is in series with the circuit. It is necessary to know the frequency in this circuit, and to compute the capacitive reactance in ohms. Current in milliamperes then is equal to the quotient of dividing 1,000 times the indicated voltage by the number of ohms capacitive reactance. It is assumed that the capacitor has no appreciable resistance or inductance, which is reasonable in the case of air dielectric and mica dielectric capacitors.

**METERS, FIELD STRENGTH.** — Field strength of a television signal is measured as the number of signal microvolts induced in a standard antenna one meter long, correctly erected and oriented at the point where field strength is to be measured. This constitutes absolute field strength, in microvolts per meter.

Measurements during service operations usually are of relative field strengths of signals from various transmitters or channels, in terms of signal microvolts at the antenna terminals of a receiver or at the input of a field strength meter designed for such measurements. The number of signal microvolts required at the antenna terminals of a receiver for satisfactory reception depends on receiver sensitivity, on the ratio of internally generated noise to gain, and on how much external interference and noise may be picked up by the receiving antenna system. Assuming no interference and negligible external noise, few receivers provide passable pictures with less than 50 to 100 microvolts of input signal. Good pictures may result from input of 200 to 300 microvolts, and excellent pictures from 500 to 1,000 microvolts of signal at the antenna terminals.

Essential elements of a field strength meter are shown by Fig. 74-4. There is a television tuner covering all vhf and uhf channels in which measurements are desired. There are one or more stages of intermediate-frequency amplification, usually peaked at the intermediate frequency of the tuner. There follows a video detector, to whose load may be connected a d-c voltmeter calibrated to read signal microvolts. Additional video amplifications may be used, in which case

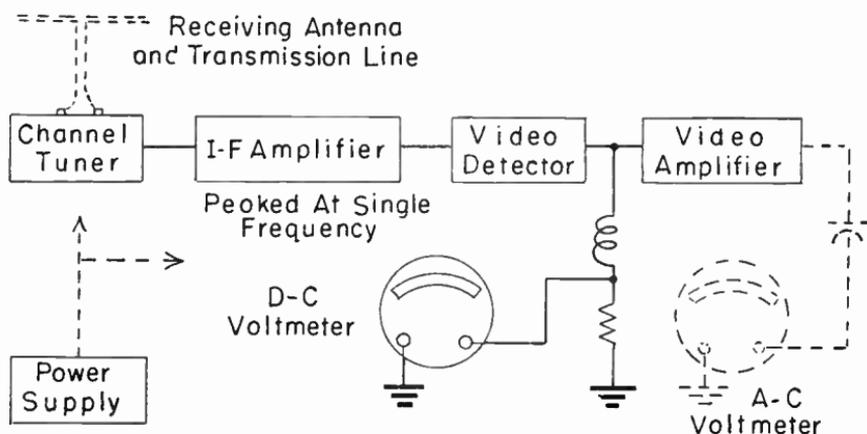


Fig. 74-4.—Elements of a field strength meter. Either meter, but not both, would be used.

the indicating meter may be an a-c type if amplified video signals are alternating. The instrument may be battery powered, or may have an a-c operated power supply for B-voltages and heater voltages, with a long external power cord.

Field strength meters are used chiefly for orienting antennas for best average reception in all channels. These instruments are used also for checking signal pickup of various types of antennas, and for checking noise and interference if the instrument includes a picture tube. For antenna adjustments the meter is connected to the regular transmission line and antenna. Since frequency response of the field strength meter may not be like that of a receiver to be used, it may not be possible to determine relative strengths of signals in different channels as they will affect the receiver. However, an antenna may be located and oriented for best reception on any one or on several channels.

Obsolete television receivers may be altered to serve as field strength meters for relative measurements. Light-weight transformerless sets are preferred. The sound system and speaker are removed. Fixed or cathode bias is substituted for automatic gain control. If the picture tube is not wanted it is usual practice to remove the high-voltage power supply, the damper, the sync section, and sweep oscillators. This assumes, as usually is true, that boosted B-voltage is not

needed for remaining circuits. A high-resistance low-range d-c voltmeter may be connected across the video detector load, or a rectifier a-c voltmeter across the output of any video amplifier stage. The entire tuning and amplifying system of the reconstructed receiver must be well shielded. The altered receiver is connected and used in the same manner as any other field strength meter, and will indicate relative field strengths.

**METERS, FREQUENCY.** — The name frequency meter as used in the radio and television fields commonly refers to a device measuring radio frequencies, or sometimes for audio frequencies. The simplest frequency meters, usually called wavemeters, consist of a tunable capacitance-inductance circuit across which is connected some form of detector (such as a crystal diode) and an indicator which usually is a d-c current meter or headphones. The meter is held near the circuit whose frequency is to be measured and is tuned for maximum indication, whereupon the frequency is read from the tuning dial of the frequency meter. These instruments obtain all power for their operation from the measured circuit, hence are suitable for use only with transmitters or high-power oscillator circuits, not for receiver circuits.

Very-high and ultra-high frequencies often are measured by noting the distance between successive nodes or anti-nodes in standing waves which are excited in a length of open or short circuited air-insulated transmission line. This distance is equal to a half-wavelength. Frequency in megacycles is determined by dividing the number 5,905 by the half-wavelength distance in inches.

**Heterodyne Frequency Meter.** — Frequencies in low-powered circuits are conveniently measured with some form of heterodyne frequency meter. A frequency to be measured is fed to some form of detector along with an adjustable or variable frequency produced by an oscillator in the meter. The meter oscillator is tuned for zero beat from the detector, which acts as a mixer. The tuned frequency then is the same as the frequency being measured, and its value is read from the tun-

ing dial of the frequency meter. The action depends on heterodyning or beating of the known and unknown frequencies.

Fig. 75-1 shows in simplified form the essential parts of a heterodyne frequency meter. The oscillator may be of any type. Here it is an electron coupled Hartley type with tuning in the grid-plate circuit at the left and also in the isolated plate circuit at the right. Output coupling or load has little effect on frequency in the grid-plate circuit, because of electron

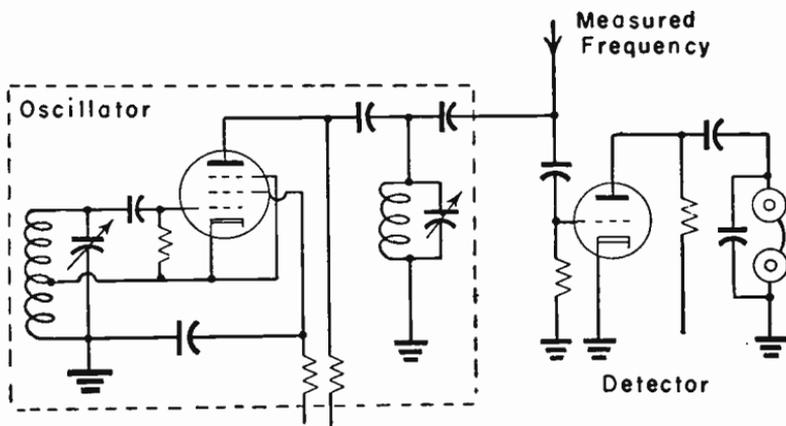


Fig. 75-1.—Essential parts of a heterodyne frequency meter or heterodyne detector.

coupling, and harmonic frequencies may be had from the plate circuit. The output of the meter oscillator and also the frequency to be measured are fed to the input of the grid-leak detector or mixer at the right-hand side of the diagram. The beat notes may be heard from headphones in the detector plate circuit.

The oscillator must have precise frequency calibration, and must be of such design and construction as to maintain its accuracy. The variable oscillator may be calibrated from a crystal controlled oscillator whose output is fed to the detector along with the output of the meter oscillator. A crystal oscillator supplying frequencies at various check points may be built into the instrument. Instead of connecting headphones into

the detector plate circuit there may be additional amplification following the detector. Then approximate zero beat is determined with headphones or a loud speaker, and the final precise setting is made with some visual indicator such as an a-f voltmeter.

**METERS, GRID DIP.** — The grid dip meter or grid dip oscillator is an instrument whose primary purpose is the measurement of resonant frequencies in tuned circuits and in all circuits and parts containing combinations of inductance and capacitance which may produce resonance. Fig. 76-1 shows connections such as generally used in grid dip meters employed for experimentation and service operations.

The meter consists of an oscillator in whose grid circuit is an indicating current meter. The tuned circuit of the oscillator includes an adjustable or variable capacitor attached to a dial calibrated in frequencies, and a series of plug-in coils for the several frequency ranges in which the meter is to operate. The oscillator coil of the meter is coupled to the circuit or part whose resonant frequency is to be measured. When the oscillator of the meter is tuned to the resonant

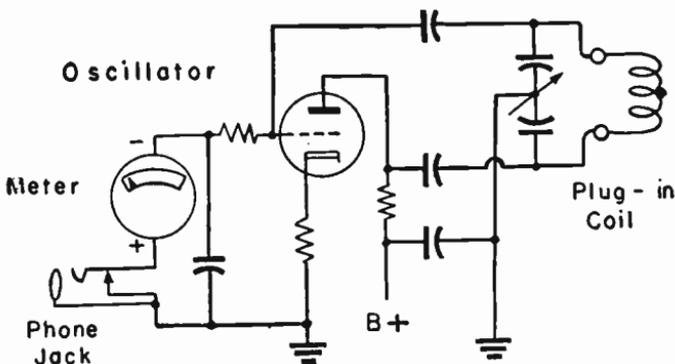


Fig. 76-1.—The parts and connections in a grid dip meter or grid dip oscillator.

frequency of the external circuit or part, or when the external circuit is tuned to the oscillator frequency, a greatly increased amount of energy goes from the meter oscillator circuit into the external circuit. The result is a reduction of feedback

energy in the meter oscillator circuit and a decrease of grid current. This decrease of grid current is indicated by the grid meter as a decrease or "dip." As either the meter oscillator or the measured circuit is tuned through resonance, the pointer of the meter will fall and then rise again. Approximate equality of resonant frequencies in oscillator and coupled circuit occurs at the minimum meter reading.

The pointer of the indicating meter will rise or fall to some extent as oscillator frequency is changed by rotation of the tuning dial, but this pointer movement will be steady or smooth. The resonance dip is a sharp drop and quick rise as the dial is turned through the point of resonant frequency match.

When the grid dip meter is of such size and construction as to make it possible, coupling usually is secured by bringing the oscillator coil close to some part of the measured circuit around which there is a magnetic field. Closest coupling and maximum meter indications are secured when field lines of the oscillator coil and of the measured circuit are parallel, or when the axes of the oscillator coil and a coupled coil are

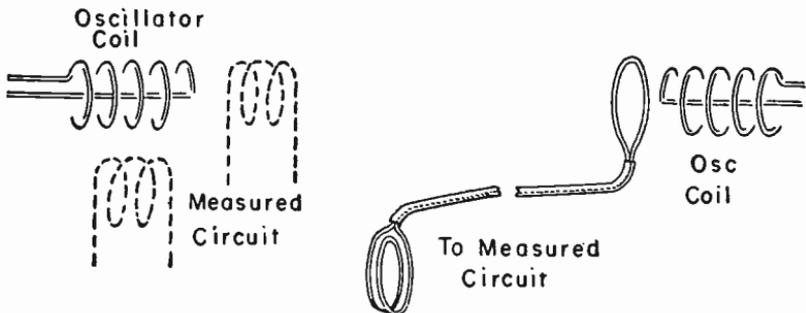


Fig. 76-2.—Methods of coupling between a grid dip meter and measured circuits.

parallel. These relations are shown at the left in Fig. 76-2. A link coupling may be used, as shown at the right. Transmission line of any kind may be used between the single turn coil at the oscillator end and a coil of two or three turns at the pickup end.

Coupling to the measured circuit should be as loose as allows readable dips on the indicating meter. Coupling which is too tight or close causes the meter pointer to snap back upward from its minimum reading. Excessively close coupling causes the meter to read too high. For example, when resonant frequency of a measured circuit actually is 20 mc the meter may indicate it as 21 or 22 mc with close coupling.

**Using the Grid Dip Meter.** — The measured circuit or part need not be carrying current, nor connected to any source of voltage or current when its resonant frequency is measured. This allows pretuning of transformers, couplers, takeoffs, traps, and other elements. When a pretuned unit is connected into its working circuit some final adjustment will be required because of stray capacitances and inductances, and tube capacitances, which add their effects.

If the measured circuit is operating and is carrying r-f voltages and currents, the grid dip meter is used similarly to a heterodyne frequency meter by connecting headphones into the jack of Fig. 76-1. There will be audible beat notes or whistles as the meter is tuned to the frequency of the measured circuit or as that circuit is tuned to the meter frequency, and there will be zero beat at matching of frequencies.

It is possible to measure the self-resonant frequencies of coils and chokes having distributed capacitance, or resistors, and of capacitors which are of inductive construction. The meter may be used also for locating parts which cause parasitic oscillation due to their own capacitance and inductance.

The grid dip meter may be used for measuring the resonant frequency and adjusting the lengths of half-wave dipole antennas by making a temporary wire connection across the center gap of the antenna and coupling the meter to this center connection. Similar measurements and length adjustments may be made on transmission lines and on resonant line stubs. With the meter coupled to the shorted end of a line which is open at the other end, as at *A* in Fig. 76-3, the resonant frequency corresponds to a quarter-wavelength. That is, the line is effectively a quarter-wavelength long at the frequency in-

indicated by the meter. If both ends of the line are shorted, as at *B*, the resonant frequency corresponds to a half-wavelength. For an open line, as at *C*, the meter is coupled to the center of the line. The resonant frequency corresponds to a half-wavelength.

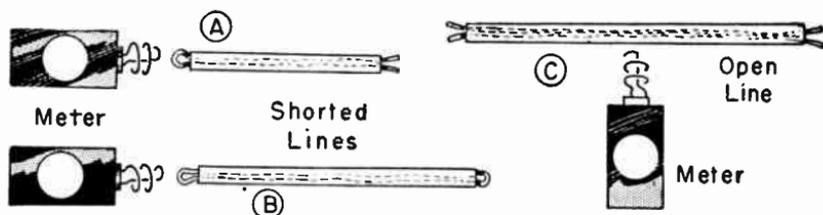


Fig. 76-3.—Coupling of grid dip meter to transmission line sections or stubs.

In making line measurements it is necessary to consider the velocity factor for the kind of line in use. The meter must be tuned to the lowest frequency at which there is a dip in order that this frequency may indicate effective wavelengths. There will be resonance and additional dips at odd multiples of the fundamental frequency, at 3 times, 5 times, and other odd multiples. These precautions apply also to measurements and adjustments of length with dipole antennas.

The grid dip meter may be used as the calibrated signal generator for measurements of capacitance and inductance as described in the article on *Measurements, High-frequency*. The meter may be used also as a signal generator for many other purposes, but since the meter oscillator coil is not shielded the signals will be radiated to everything within many feet of the instrument, and there will be no means for attenuating or otherwise regulating the output.

Amplitude modulation at audio frequencies or other frequencies may be applied to a signal from the grid dip meter by connecting the modulating voltage into the grid circuit by means of the phone jack of Fig. 76-1. For instance, the output of any type of audio generator or audio oscillator may be connected to this jack, and the audio note picked up by a receiver or any detector.

Among other uses, the grid dip meter may be used as a diode detector by cutting off plate voltage from the oscillator tube, but leaving the cathode heated. Signals picked up on the oscillator coil may be heard in the headphones. Some signal tracing may be done with the meter used thus. With no plate voltage, but with the cathode heated, the grid dip meter may be used as an absorption wavemeter by bringing the oscillator coil or coupling near an energized circuit. There will be a slight rise of the pointer on the indicating meter when the grid dip tuning is varied through the frequency of the coupled circuit.

**METERS, RECTIFIER.** — Voltmeters having useful readings at frequencies as high as 100 megacycles employ moving coil d-c microammeters in connection with crystal diode rectifiers. Half-wave rectifying circuits are shown at *A* and *B* of Fig. 77-1. A full-wave circuit is shown at *C*. Indicating meters are preferably of 50 to 100 micro-ampere range at full-scale. Calibra-

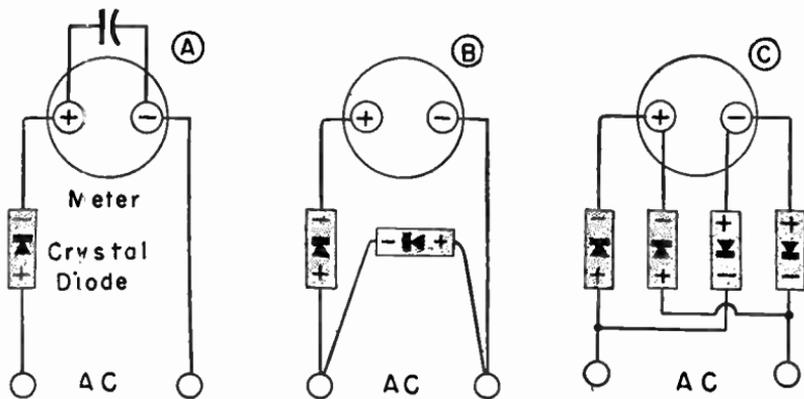


Fig. 77-1.—Connections of crystal diodes to d-c microammeters for measurements at high frequencies.

tion varies with the meter and the circuit. Scales are not linear with respect to current or voltage. Multiplier resistors for voltage measurement must be connected so that all current for all rectifiers flows in these resistors. With connections of Fig. 77-1 a multiplier resistor could be between the negative

terminal of the meter and one of the a-c terminals at *A*, but in the other diagrams the multiplier would have to be in series with one of the a-c terminals to protect all the rectifiers.

**METERS, THERMOCOUPLE.** — Two dissimilar metals in contact form a thermocouple which, when heated by flow of alternating current, produces a direct emf and current which may be used to actuate a moving coil meter. The meter is calibrated in terms of alternating voltage applied to the heater or of alternating current flowing in the heater. Because the thermocouple meter utilizes only the heating effect of applied alternating current these instruments may be made of high accuracy up to frequencies somewhat less than 100 mc in full-scale ranges of 1 to 5 milliamperes and up to several hundred megacycles in full scale ranges of 100 to 200 milliamperes or more. Indications are independent of waveform, always showing effective values.

Thermocouple voltmeters are quite insensitive, having resistances of 125 to 500 ohms per volt in order to permit enough current for heating. Because developed heat and indicated current or voltage are proportional to the square of applied current in the thermocouple, the scale is crowded at the low end.

**METRIC CONVERSIONS.** — Multiply the following metric quantities by the listed factors to determine equivalent quantities in English units.

Centimeters	×	0.3937	=	inches
Cubic centimeters	×	0.06102	=	cubic inches
Grams	×	0.03527	=	ounces
Liters	×	61.02	=	cubic inches
Liters	×	1.057	=	quarts
Meters	×	3.2808	=	feet
Meters	×	39.37	=	inches
Millimeters	×	0.03937	=	inches
Square centimeters	×	0.1550	=	square inches
Square millimeters	×	0.00155	=	square inches

To convert centigrade temperature to equivalent Fahrenheit temperature proceed thus:

$(1.8 \times \text{centigrade degrees}) + 32^\circ = \text{Fahr. degs.}$

As an example, convert  $-10^\circ \text{ C.}$  to equivalent  $^\circ \text{ F.}$

$(1.8 \times -10) = -18^\circ$

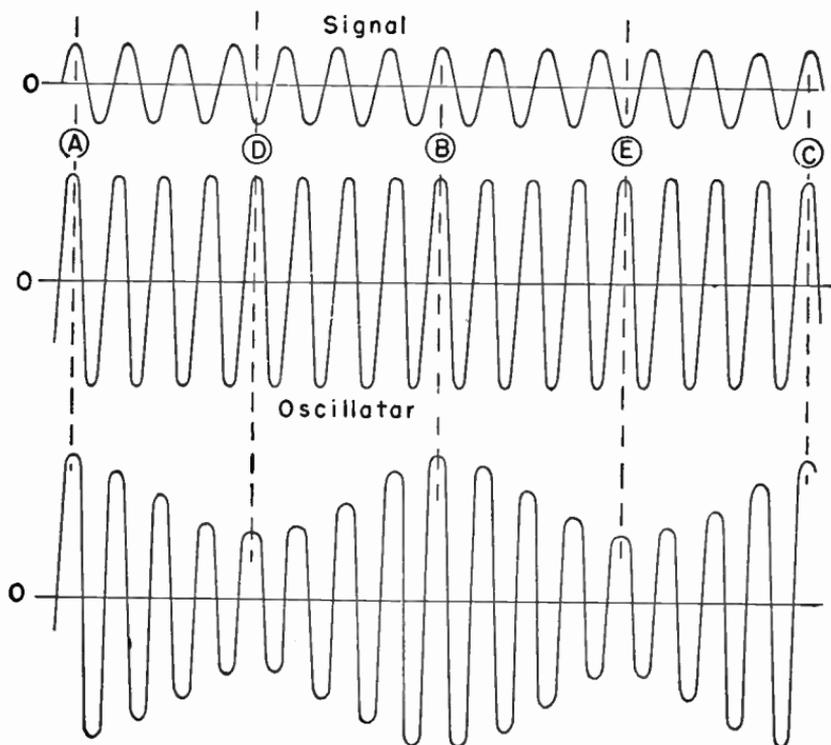
$-18^\circ + 32^\circ = +14^\circ \text{ F.}$

**MIXERS.** — A mixer is a tube, or sometimes a crystal diode, in which the frequencies of the carrier signals and the r-f oscillator combine to produce frequencies among which there is the beat frequency that becomes the intermediate frequency for the receiver. The mixer in broadcast sound receivers usually is combined with the r-f oscillator in the converter tube, but in television receivers most often is a separate tube or a separate section of a twin tube. The mixer sometimes is called the first detector, because it acts to rectify or partially rectify the frequencies applied to it.

In the article on *Frequencies, beat* it is shown that two voltages of different frequencies put into the same circuit combine in such manner as to produce peak amplitudes of resultant voltage or current when the two voltages are in phase, and to produce zero amplitude when they are of opposite phase. These peak and zero amplitudes occur when the two original voltages have equal amplitudes. Ordinarily the two input amplitudes are different and the result is not zero when the voltages are of opposite phase.

Fig. 80-1 illustrates the heterodyne action with incoming signal voltage of lower frequency and also of less amplitude than the voltage from the local r-f oscillator. These are the conditions usually obtaining in television receivers. The combined voltage is shown at the bottom of the diagram. In this combined voltage there are instants of maximum amplitude at *A*, *B*, and *C* when signal and oscillator voltages are in phase. There are instants of minimum amplitude at *D* and *E* when signal and oscillator voltages are of opposite phase. Maximum amplitudes are equal to the sum of the amplitudes of signal and oscillator voltages, while minimum amplitudes are equal to the difference. Oscillator voltage is modulated by signal voltage.

The resulting combination voltage varies in amplitude, but still is at a frequency in the carrier signal or oscillator range. This combination voltage acts on the mixer grid. The mixer is negatively biased to a degree which makes it operate as a



*Fig. 80-1.—How signal and oscillator frequencies combine with amplitude variations which furnish a beat frequency when rectified.*

detector, with the grid signal voltages working down on the lower curve of the mutual characteristic. The bias would be almost that for plate current cutoff. Oscillator voltage is held to a value such that the sum of oscillator and signal voltages, at peak amplitudes, will not exceed the bias and cause grid current to flow. Plate current flows during positive alternations of oscillator voltage, but during very little if any of the negative alternations. This is rectification or detection. The

average value of plate current then varies according to the average of alternations above (or below) the zero line at the bottom of Fig. 80-1.

One cycle of the frequency of average change in plate current would extend from *A* to *B* or from *B* to *C* in Fig. 80-1. During the time represented by the diagram there are two such cycles, and were the time to be one second the beat frequency in the mixer plate circuit would be two cycles per second. In the signal wave at the top are 14 cycles, and in the oscillator wave are 16 cycles. The difference is the beat frequency of 2 cycles. The difference frequency is selected as the intermediate frequency by tuning the i-f transformers or couplers to this frequency. In the mixer output there are also the oscillator frequency, the signal frequency, a frequency equal to their sum, and still other frequencies formed by harmonics of the oscillator frequency.

Conversion transconductance of a mixer is the ratio of i-f signal voltage in the plate circuit to carrier frequency signal voltage applied to the grid circuit. The value is specified in micromhos, which are equivalent to microamperes of i-f plate current per volt of r-f signal on the grid. Conversion transconductance of a triode or pentode used as a mixer is usually about one-third or less of the grid-plate transconductance of the same tube used as an amplifier.

Conversion gain is the voltage gain between signal voltage at the grid of the mixer tube and signal voltage delivered to the i-f amplifier, usually measured at the grid input of the first i-f amplifier. The gain is the ratio of i-f output voltage to r-f signal input voltage.

**Uhf Crystal Mixers.** — Any element which is capable of acting as a detector may be used also as a mixer for frequency conversion, since the prime requirement is non-linear rectification or unequal current flow in opposite directions when alternating voltage is applied to the element. A number of germanium crystal diodes have been designed especially for mixer service at the television ultra-high frequencies, including the 1N72, CK710, G7, and others. The type number

1N72 is also a designation of uhf crystal diodes as a general class.

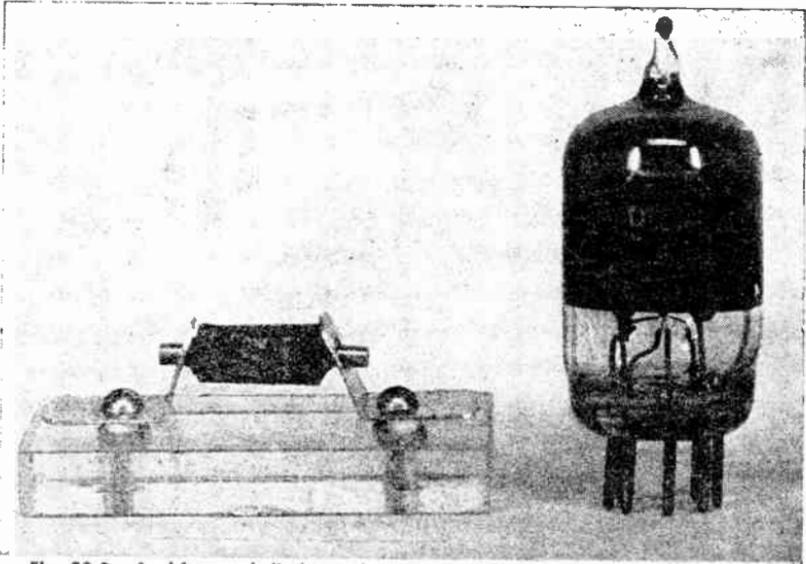


Fig. 80-2.—A uhf crystal diode used as a mixer, and a miniature diode tube.

At the left in Fig. 80-2 is a 1N72 germanium crystal diode and at the right is a type 9006 uhf diode tube, of about the same size as a 6AL5 or a 6AK5 miniature tube. The 1N72 will operate satisfactorily at input frequencies of 100 mc to 1,000 mc, with a noise figure of approximately 15 decibels.

Fig. 80-3 is a simplified circuit diagram for a uhf tuner containing a crystal diode mixer, a triode tube oscillator, and a triode amplifier for the beat frequency or intermediate frequency from the carrier and oscillator voltages that combine in the mixer circuit. The i-f amplifier is needed because no r-f amplifier is used. Furthermore, there is not only no gain in the crystal mixer, but there is a loss of 60 per cent or more of the signal strength in this kind of mixer. The crystal mixer requires less voltage from the oscillator or less injection voltage than a tube used as a mixer, and the crystal has a noise factor lower than that of tubes not designed for uhf mixer service.

Measurement of direct current in a crystal diode mixer provides a check of crystal condition and oscillator operation.

Crystal current varies with changes of oscillator injection

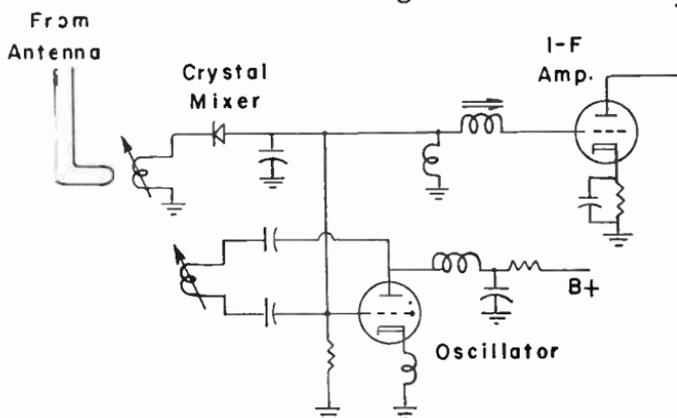


Fig. 80-3.—Connections between a crystal diode mixer, uhf antenna, and uhf oscillator tube.

voltage, with frequency, with the type of crystal, and with its condition. Many receivers have provision for measuring mixer current by means of a d-c milliammeter temporarily connected in series with the crystal. Mixer current will vary widely with receiver or converter tuning through the range of uhf carriers, but always there will be some current provided the oscillator furnishes injection voltage and the crystal is in good condition. Maximum current may be 2 ma or more, depending on design of the circuits.

The surest test for a defective mixer crystal is replacement with one known to be good, preferably of the same type. Ohm-meter measurements of forward and back resistance seldom give reliable indications of crystal condition. Replacement usually calls for realignment of antenna and mixer input circuits. Crystal mixers sometimes have positive bias of a fraction of a volt on their anode.

**MODULATION, AMPLITUDE.** — Modulation is a variation of some characteristic of a carrier wave by means of a signal which represents or is derived from sounds, pictures, or other intelligence to be transmitted. Present methods of modulation include variation of carrier amplitude, frequency, or phase. The principle of amplitude modulation is shown by Fig. 81-1. At the top of the diagram is represented the modulating signal in the form of an approximate sine wave of voltage. Any other

simple or complex signal waveform might be used. At the lower left is represented the unmodulated carrier voltage, which is of constant frequency and constant amplitude. The signal voltage is combined with the carrier voltage to increase and decrease the amplitude of the carrier without altering its frequency.

Not clearly evident from the conventional diagram of Fig. 81-1 is the fact that in the modulated carrier there are at any instant a total of three frequencies. One is the frequency of

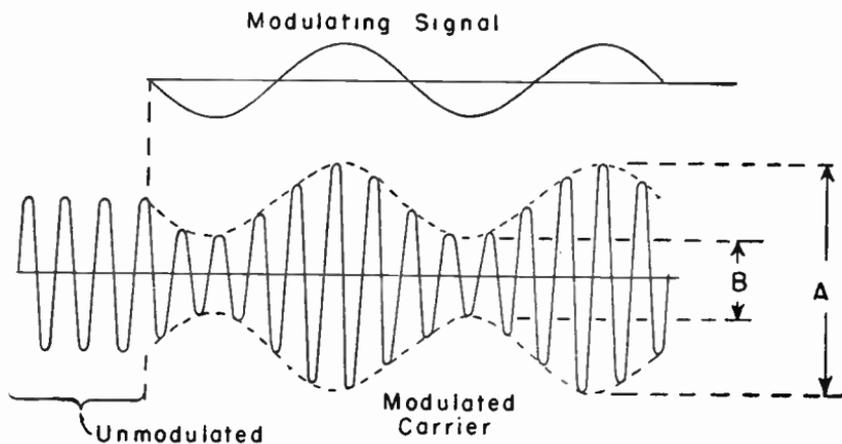


Fig. 81-1.—The principle of amplitude modulation.

the unmodulated carrier, a second is this frequency plus the instantaneous modulating frequency, and the third is the unmodulated frequency minus the instantaneous modulating frequency. Thus the transmission becomes a band of frequencies extending below and above the unmodulated carrier frequency by amounts equal to the maximum modulating frequency being used. Amplifiers must be capable of handling this entire band of frequencies.

Percentage of modulation may be determined from values of maximum and minimum voltages in the modulated carrier, these values being marked respectively  $A$  and  $B$  in the diagram.

$$\text{Percentage modulation} = 100 \times \frac{A - B}{A + B}$$

Amplitude modulation is used for transmission of picture and synchronizing signals in television, for standard broadcast and international shortwave broadcast, and for most other purposes. An amplitude-modulated carrier may be demodulated for recovery of the signal by any detector which acts as a rectifier. This class includes diode detectors of both electronic (tube) and crystal types, also grid-leak detectors and plate rectification detectors.

**MODULATION, FREQUENCY.** — The principle of frequency modulation of a carrier wave or carrier voltage is illustrated by Fig. 82-1. The unmodulated carrier, shown at the left, is of constant frequency and constant amplitude. The modulating signal, here shown as a sine wave voltage, may be of any waveform. This modulating signal is used to cause variations of reactance which change the frequency of the carrier in accordance with amplitude and phase or polarity of the signal. Frequency of the unmodulated carrier may be called the center frequency. At each change between maximum and minimum frequencies the modulated carrier frequency passes through this center value, at C-C-C in the diagram. The modulated carrier remains of constant amplitude so far as effects of the modulating signal are concerned, although the amplitude may be varied by effects of interference after leaving the transmitter.

The change of frequency, from the center frequency to either the maximum or minimum frequency, is called *frequency deviation* or just deviation. Deviation is not the total change as measured from maximum to minimum. Deviation is proportional to amplitude of the modulating signal. If that signal is for sound or is an audio-frequency signal, the extent of deviation is proportional to strength or volume of sound. The greater the sound volume, or the stronger the modulating signal as measured in any manner, the greater is the deviation or the greater is the change of frequency away from the center frequency.

Frequency modulation is used for television sound, for sound programs accompanying the pictures. Maximum devia-

tion for television sound is 25 kilocycles above or below the center frequency. Frequency modulation is used also for f-m broadcasting of sound in carrier channels between 88 and 108 megacycles. For this class of service the maximum deviation is 75 kilocycles. These maximum values of deviation correspond to 100 per cent modulation. There are no electrical or

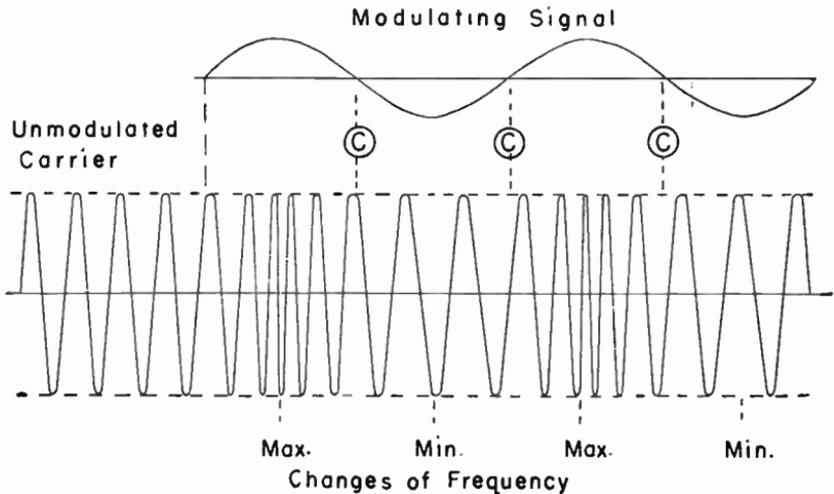


Fig. 82-1.—The principle of frequency modulation.

physical reasons why deviations might not be greater, or less. The limits mentioned have been agreed upon for the respective classes of service.

*Modulation frequency* is the number of times per second that the frequency goes through a complete change, as from a maximum through a minimum and back to the maximum. Each such complete change occurs during one complete cycle of the modulating frequency, as is apparent in Fig. 82-1. Consequently, the modulation frequency of the frequency-modulated signal is the same as the frequency of the modulating signal. With a sound signal the modulation frequency would correspond to frequency or pitch of the sound. The higher the sound frequency the higher will be the modulation frequency. Modulation frequency and deviation are entirely independent of each other.

Frequency-modulated signals are demodulated or detected by discriminators and ratio detectors. Limiters are used to remove amplitude modulation which may have been unintentionally added to the frequency modulation during transmission. There are articles describing *Discriminators*, *Detector*, *ratio*, and *Limiters, f-m sound*. The chief advantage of using frequency modulation for sound transmission and reception is its insensitivity to static and electrical interference. These, and most other forms of interference with sound transmission, tend to produce amplitude modulation rather than frequency changes.

**MODULATION, PHASE.** — With phase modulation of a carrier voltage or current the modulating signal is used to advance or retard the phase of the carrier with reference to the unmodulated carrier. The idea is illustrated in an elementary way by Fig. 83-1 where the unmodulated carrier in its relation to time is represented by the broken-line curve, with portions advanced and retarded in phase represented by full-line curves.

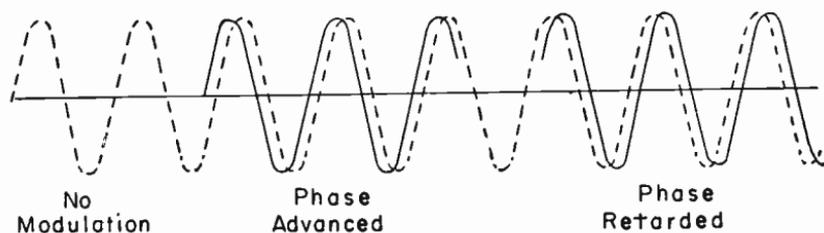


Fig. 83-1.—Phase changes for phase modulation.

The phase is shifted at the transmitter by circuits causing instantaneous lead and lag. Change of phase is proportional to amplitude of the modulating signal.

Inspection of the diagram will show that there must be an increase of frequency during the period in which phase is advanced, because the waves are compressed in time, and there must be a decrease of frequency while phase is being retarded. However, with completion of a phase change, the frequency returns to its former value. Change of phase in relation to the unmodulated carrier is measured in degrees of

the cycle, and is called phase deviation. Frequency of the modulating signal determines the number of phase changes per second. Phase modulation is at present used chiefly for experimental transmission and reception.

**MULTIPLIERS, FREQUENCY.** — A frequency multiplier consists of one or more amplifying stages with tuned plate circuits operated in such manner as to have distortion and strong harmonic frequencies in the plate circuit, one of these harmonics being taken as the output.

A frequency multiplier tube is biased nearly to plate current cutoff or sometimes just beyond cutoff when no signal is on the grid. The plate circuit then is tuned to the second or third harmonic of the frequency applied to the grid circuit. Such tuning of the plate circuit provides high impedance and high voltage drop (output voltage) for the tuned harmonic frequency. The fundamental and unwanted harmonics are bypassed to ground. With grid bias made somewhat more negative than for plate current cutoff the third harmonic is fairly strong, although much weaker than the second. Still higher harmonic frequencies may be obtained from the plate circuit, but they provide little power. Push-pull connections sometimes are used for frequency tripling, since the push-pull amplifier suppresses the second harmonic and provides fairly efficient operation on the third.

Frequency doublers and triplers often are used to follow a crystal controlled oscillator where the final output frequency must be higher than can be directly controlled by crystals of practicable thickness. Frequency doublers often are used in cascade; two stages then giving four times the original frequency, and three stages giving eight times the frequency.

**MULTIPLIERS, VOLTAGE.** — Transformerless receivers often employ voltage doubler rectifier systems which supply for plate and screen circuits a d-c voltage approximately double the effective a-c voltage from the power line. Voltage doubler circuits may employ electronic tube rectifiers, often twin tubes with two plates and two cathodes in one envelope, or they

may use pairs of selenium rectifiers. The high output voltage is obtained by allowing two or more capacitors to charge alternately during opposite half-cycles of line voltage to potential differences equal very nearly to peaks of the a-c line voltage. These capacitors then are discharged in series to the load.

A circuit commonly used for full-wave voltage doubling is shown by Fig. 85-1. Rectifiers *A* and *B* may be the two sections of a rectifier-doubler tube, or two electronic rectifiers, or two selenium rectifiers. The two capacitors which take part in the doubling action are marked *A* and *B*. These two capacitors,

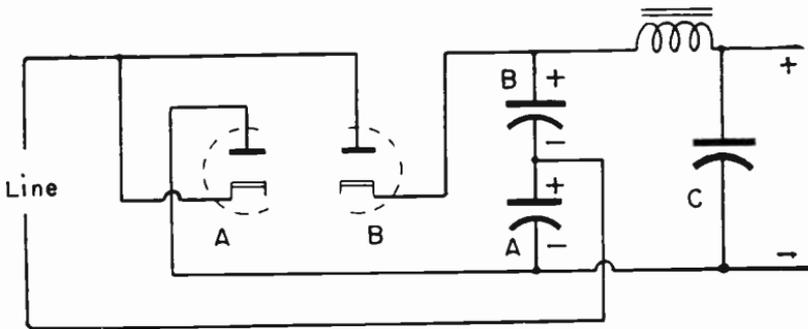


Fig. 85-1.—Connections for full-wave voltage doubler.

in combination with the choke and capacitor *C*, form a power supply filter to which the plate, screen, and biasing loads are connected at the positive and negative output terminals on the right. One side of the a-c line is connected to the cathode of rectifier *A* and to the plate or anode of rectifier *B*. The other side of the line is connected to a point between the two doubler capacitors.

When the upper end of the line is negative, electron flow through rectifier *A* charges capacitor *A* in the polarity shown, while electron flow from this capacitor goes to the lower positive side of the line. When line polarity reversed, electron flow through rectifier *B* charges capacitor *B* in the marked polarity. These two capacitors are in series with each other

across the output. Their polarities are in such relation that the voltage across the output is the sum of the capacitor voltages.

If no current were being drawn from this rectifier-filter system, the capacitors *A* and *B* would charge to the peaks of line voltage or to about 1.4 times the effective line voltage. Then the sum of the capacitor voltages would be about 2.8 times the effective line voltage. Actual output voltage becomes less when current is being supplied to external loads at the same time the capacitors are being charged from the line.

Both alternations of each a-c cycle are rectified, so ripple frequency from the filter output is twice the line frequency; for example, is 120 cycles per second on a 60-cycle line. To provide fairly high d-c output voltage and satisfactory voltage regulation, the capacitances at *A* and *B* usually are something between 16 and 32 mf, with still greater capacitance at *C*. When the circuit of Fig. 85-1 is used with high-voltage filament-cathode rectifier tubes, each of the filaments must be heated from a separately insulated secondary winding on the power transformer.

No direct or conductive ground connections may be made to either the positive or the negative output sides of the voltage multiplier system. This is because one of the line wires always is grounded in the building circuits, and this line ground extends either to the cathode of rectifier *A* and the plate or anode of rectifier *B*, or else to the connection between capacitors *A* and *B*. A ground on either side of the output would short circuit either a capacitor or a rectifier. Radio-frequency grounds are made through capacitors of small capacitance, which then are in shunt with a doubler capacitor or a rectifier.

At the top of Fig. 85-2 are shown connections for a half-wave voltage doubler with which one side of the a-c power line and B-minus of the doubler circuit are common, allowing this to be grounded. The resistor in series between the power line and capacitor *A* reduces current surges during charging

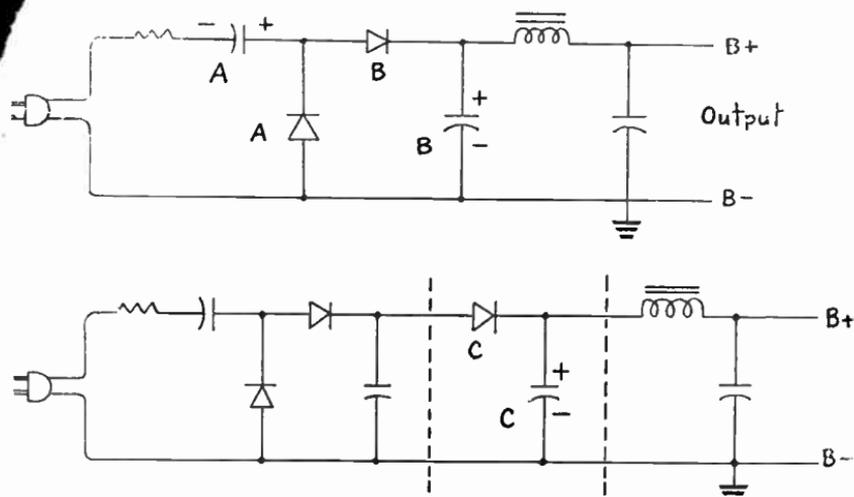


Fig. 85-2.—Half-wave voltage doubler (above) and half-wave voltage tripler (below).

of the capacitors and limits the current in case of accidental short circuits. This resistor may be on either side of the line.

When the lower side of the a-c line is positive and the upper side negative, electron flow through rectifier *A* charges capacitor *A* in the polarity marked. When line polarity reverses, electron flow goes to B-minus of the output, also through capacitor *B*, rectifier *B*, and capacitor *A* to the upper side of the line, now positive. Capacitor *A* is discharging, and its voltage acquired during the preceding half-cycle of line voltage adds to line voltage of the half-cycle now considered. Capacitors *A* and *B* are in series across the output, on which there is approximately twice the line voltage. Capacitor *A* must withstand peak line voltage, and usually is of 150 d-c working volts rating. Capacitor *B* must withstand nearly double the peak line voltage, and would have a rating of 250 or more d-c working volts.

A voltage tripler circuit is shown at the bottom of Fig. 85-2. The portion of this circuit on the left of the broken lines is the same as the voltage doubler circuit shown above. Between the broken lines is added a half-wave rectifier circuit in which capacitor *C* is charged to line voltage by electron flow through rectifier *C*. To this voltage on capacitor *C* is added twice the

line voltage developed in the doubler circuit, giving a total output of approximately three times line voltage. Capacitor must withstand almost three times peak line voltage, and ordinarily would have a rating of 450 or more d-c working volts.

Voltage multipliers here described are used for low-voltage B-supplies. These circuits are not the same as those of voltage multipliers used in high-voltage supply systems for second anodes or ultors in picture tubes, because second-anode voltage supplies operate on pulsating voltage instead of on the approximately sine-wave voltage from a-c power lines.

## N

**NOISE, RECEIVER.** — Noise may result not only from external interference but also from many causes within the receiver itself. Many of the internal causes for noise are quite easily corrected. These include looseness, dirt, or corrosion at terminal connections, at contacts for band switches, in tuning capacitors or trimmers, and especially in adjustable potentiometers or rheostats. Where it is necessary for an adjustable resistor to carry only alternating current, and where it is also carrying direct current, noisy operation sometimes is prevented by using a series capacitor to block out the direct current.

Internal noise results also from leaky insulation or leaky capacitors, from tubes which are microphonic or which have loose elements, and from loose fastenings for shields or any other parts in which there may be electrostatic or magnetic fields. Noise results from tubes of poor quality, due to internal leakage, secondary emission, and gas effects.

An effect called thermal noise results from motions of free electrons within resistors and all conductors, even when there is no current. The random movements of electrons balance to leave no net flow or current in either direction, but there are local potential differences which cause noise when greatly amplified. The thermal noise increases with temperature and with resistance of the parts in which it originates. The effect becomes evident as a hissing sound.

Another effect, called tube noise, results from slight irregularities of the electron stream in even the best of tubes. Tube noise is lessened in tubes capable of providing large emission from the cathode in comparison with actual plate current, so that there is no emission limiting but only space charge limiting of plate current. Tube noise is lessened also by high values of transconductance in comparison with the plate current. The most severe kind of tube noise is called shot effect, which may cause popping or sputtering noises. In tubes designed especially for television all precautions

have been taken to lessen the tube noise. This is necessary because the small values of stage gain in wide band amplifiers limit the signal strength and leave tube noise more prominent.

Thermal noise and tube noise ordinarily extend throughout the entire band of frequencies which are amplified. Highly selective circuits having good amplification tend to lessen noise because they react only to the portion of noise voltage included within the range of amplified frequencies, whereas non-selective circuits would bring in a greater total noise voltage while still amplifying only the same signal. Noise voltages always are amplified by any amplifier which will amplify a desired signal. When tuning a sound receiver from one station to another, there is a great increase of noise (hiss) if there is automatic volume control allowing increased gain as signal strength drops off.

It is most important to reduce tube and circuit noise in the first stage of any amplifier. Here the desired signal is relatively weak, and may be almost completely obliterated by strong noise. If, however, the noise level in the first stage is low in relation to signal strength, equal amplification of both in following stages will not permit overemphasis of the noise.

In any receiver or amplifier there is generation of noise voltages, and these voltages will be amplified from stage to stage. The best way of obtaining the highest possible ratio of signal to noise at the output is to increase signal strength at the input. Signal input is improved by careful attention to selection and installation of the antenna and transmission line. Unless the input signal is strong enough to override the internal noise generated in the first couplings and amplifying stages it will be impossible to have good reproduction of either pictures or sound. All noise effects which would mar the reproduction of voice and music from a sound receiver will have equally bad effects on reproduction of pictures in a television receiver.

**NUMBERS, PREFERRED.** — Values of fixed resistors and of small fixed capacitors usually are of such numbers of ohms or megohms for resistors, and of such numbers of micro-microfarads for capacitors, as to greatly reduce the variety of these units which must be available, while still allowing choice of 20%, 10%, or 5% plus or minus tolerances as may be required for various applications.

PREFERRED NUMBERS					
20% Tolerance	10% Tolerance	5% Tolerance	20% Tolerance	10% Tolerance	5% Tolerance
10	10	10	33	33	33
		11			36
	12	12		39	39
		13			43
15	15	15	47	47	47
		16			51
	18	18		56	56
		20			62
22	22	22	68	68	68
		24			75
	27	27		82	82
		30			91

The accompanying list gives preferred numbers for tolerances of 20% in one column, for 10% in another column, and for 5% in a third column. The list includes only preferred numbers from 10 to 100, but these may be increased or reduced by multiples or submultiples of 10 to find preferred values any amount larger or smaller.

Tolerances work out as follows: Consider three resistors of 22 ohms, 27 ohms, and 33 ohms, all with tolerance of 10% plus or minus. The range between 10% low and 10% high for the resistor rated at 22 ohms would be from 19.8 ohms to 24.2 ohms. For the unit rated at 27 ohms the low and high limits would be 24.3 and 29.7 ohms, and for the 33-ohm unit would extend from 29.7 to 36.3 ohms. Provided 10% tolerance is satisfactory, these three resistors give continuous coverage of all possible values from 19.8 ohms through 36.3 ohms. Consequently, there would be no object in having inter-

mediate values when 10% tolerance is satisfactory. The same reasoning applies to fixed capacitors whose values are in preferred numbers.

Similar computations for any adjacent values for 20% tolerance, and for any having 5% tolerance, will show continuous coverage for all values intermediate between those listed. Similar continuous coverage extends throughout all preferred numbers in the list. If these numbers are divided by 10 or any multiple of 10, or are multiplied by 10 or any multiple of 10, there still will be continuous coverage throughout the extended list for each tolerance.



**OHMMETERS.** — At the left in Fig. 87-1 is shown the elementary circuit for an ohmmeter designed for connection of the measured resistance in series with the indicating meter of the instrument. The unknown resistance is connected across terminals  $R_x$ . The current source may be a battery or any rectified d-c supply. Change of source voltage is compensated for by an adjustable shunt resistance across the meter. With the  $R_x$  terminals shorted together, the meter shunt is adjusted to bring the pointer to zero. With this style of ohmmeter, zero measured resistance allows full current, and deflection of the meter pointer all the way to the right. Maximum or infinite measured resistance allows no current, and the meter pointer remains at the left.

Instead of an adjustable resistance shunt for zero setting, the indicating meter may be equipped with a magnetic shunt, or the calibrating resistance in series with meter and source

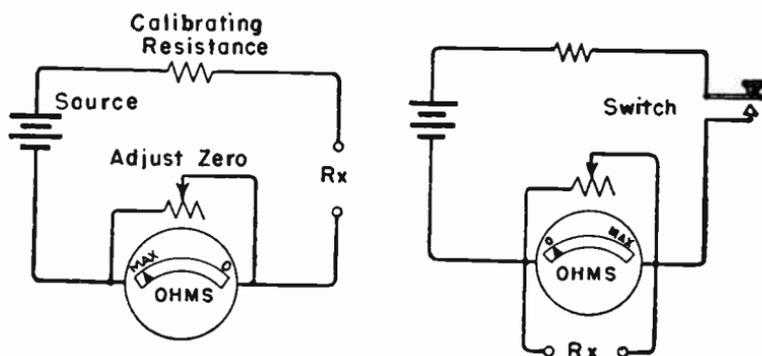


Fig. 87-1.—Series resistance and shunt resistance types of ohmmeters.

may be adjustable. The total internal resistance of the ohmmeter is the sum of the resistances of the meter as shunted, of the calibrating resistance, and of the source. Relations of internal resistance, the source voltage, and the full-scale current of the indicating meter is as follows:

$$\text{Source volts} = \frac{\text{full-scale milliamperes} \times \text{internal resistance, ohms}}{1000}$$

$$\text{Full-scale milliamperes} = \frac{1000 \times \text{source volts}}{\text{internal resistance, ohms}}$$

There is half-scale or mid-scale deflection of the meter pointer when the measured resistance connected across  $R_x$  terminals is equal to the internal resistance of the instrument.

The diagram at the right in Fig. 87-1 shows the elementary form of connections for an ohmmeter with which the measured resistance is connected across the terminals of the indicating meter, thereby acting as a shunt on the meter. Zero measured resistance leaves the meter pointer at the left, since all source current then goes through the measured resistance. With infinite measured resistance, all current flows through the indicating meter and its pointer moves to the right. The instrument circuit is closed by a switch only while measurements are being made, since otherwise there would be continual current drain on the source.

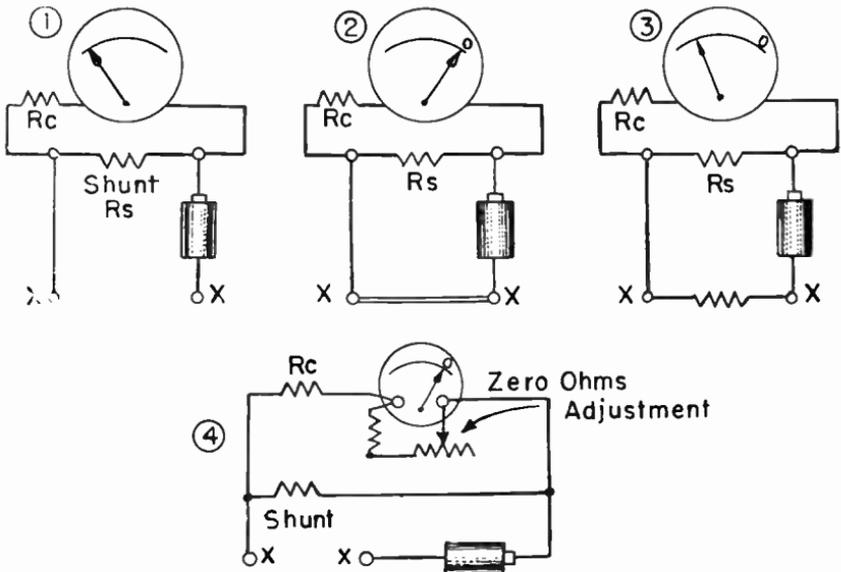


Fig. 87-2.—Ohmmeter circuit in which a meter shunt allows changing the measurement range.

The principle of an ohmmeter circuit commonly used in

service instruments is illustrated by Fig. 87-2. Resistor  $R_c$  is for calibration. A shunt resistor,  $R_s$ , is across the meter. In series with one lead to the external terminals for connection of measured resistances,  $X-X$ , is the ohmmeter battery. When terminals  $X-X$  are open no battery current can flow in the meter, and its pointer stands at the position for infinite measured resistance. If test terminals are short circuited, or if resistance measured is practically zero (2) battery current through the short circuit and meter swings the meter pointer to the position for zero resistance. When a measurable resistance is connected at  $X-X$  (diagram 3) part of the battery current flows through that resistance and the shunt, while part goes through the measured resistance and the meter circuit. The pointer deflects to a position corresponding to the measured resistance.

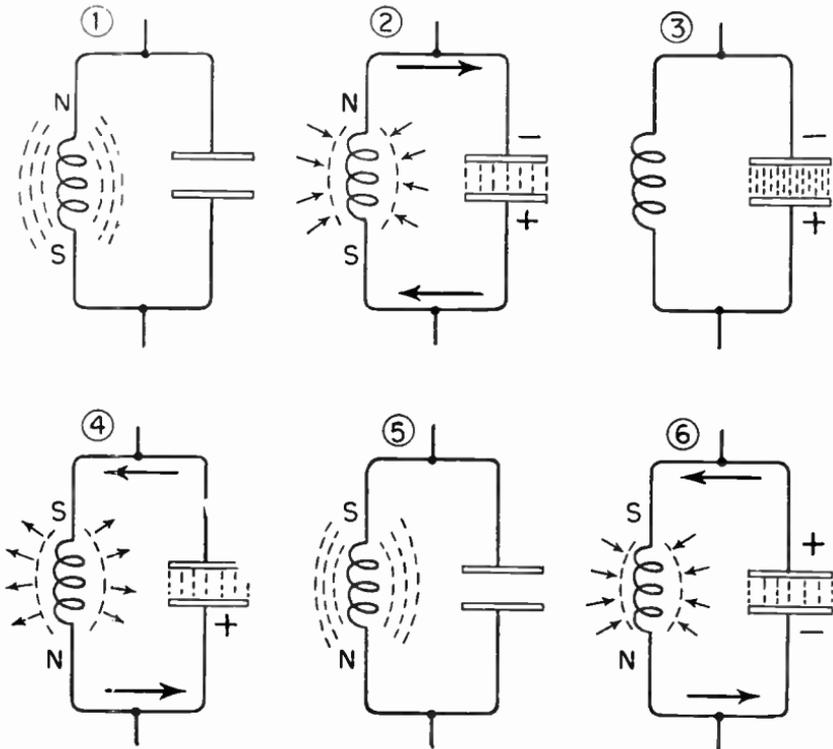


Fig. 88-1.—During oscillation there is interchange of energy between the magnetic field of an inductor and the electrostatic field of a capacitor.

In practice, the shunt resistance  $R_s$  consists of several resistor sections which may be switched into the circuit to provide a number of ranges for measurement. Weakening of the battery may be compensated for with an adjustable resistor across the meter, as shown by diagram 4, or in other ways. This is the zero ohms adjustment.

**OSCILLATION.** — Oscillation may be defined as back-and-forth movement of electric charges between a capacitance and an inductance, as illustrated for one type of connection by Fig. 88-1. Assuming that energy originally exists in a magnetic field around the inductor (1), collapsing magnetic lines cut the conductors and induce emf which sends energy in the form of electron flow to the capacitor. The capacitor commences to charge (2) and with continued electron flow becomes fully charged (3) as the magnetic field disappears. Energy now is in the electrostatic field between capacitor plates. This capacitor charge causes a reversed electron flow (4) in the inductor, building up a new magnetic field (5). When this magnetic field collapses (6) there is electron flow



*Fig. 88-2.—Barkhausen oscillation causes a dark vertical line at the left on the picture tube screen.*

and transfer of energy back to the capacitor, and so the action continues. Frequency of oscillation or energy transfer is the resonant frequency of inductance and capacitance.

Energy is dissipated by circuit resistance and other losses.

If energy from an external source replaces losses there is sustained or continual oscillation. Otherwise the oscillating or circulating currents die away, and there is said to be damped oscillation.

**Barkhausen oscillation** at high frequencies occurs in horizontal output amplifiers and their circuits, producing a form of interference picked up by r-f and i-f circuits in the receiver to cause a ragged dark vertical line at the left of the picture tube screen, as in Fig. 88-2. As output amplifier grid voltage goes suddenly less negative, plate voltage drops momentarily below screen voltage and electrons between plate and screen move toward the screen. But before all electrons reach the screen the plate becomes more positive, and electrons between the elements move back toward the plate. This back-and-forth movement of electrons or charges constitutes oscillation, within the tube and in capacitances and inductances of connected wiring.

**Parasitic oscillation** occurs in inductances and capacitances not intentionally tuned to resonance at the oscillation frequency. For example, inductance of any coil, or even of lengths of wiring, may act with stray, distributed, or tube capacitances to form an oscillatory circuit in which oscillation occurs at the resonant frequency of these inductances and capacitances. Parasitic oscillation may occur at frequencies from a few hundred kilocycles up to hundreds of megacycles, but the name usually is applied to oscillation at frequencies much higher than those for which affected circuits are intentionally tuned.

**Ringings** is a form of oscillation, usually damped, occurring in inductance and capacitance of a circuit not intentionally tuned to the oscillation frequency, hence is a variety of parasitic oscillation. The name ringing usually is applied to oscillation occurring at frequencies lower than those considered to exist with parasitic oscillation. Ringing occurs in circuits having inductance large in comparison with capacitance; for example, in circuits of deflecting yoke coils, or in feedback transformers of blocking oscillators. The name ringing circuit may be applied to a parallel resonant inductor and capacitor in series with one of the plate leads to a multi-

vibrator sweep oscillator. The purpose is to improve horizontal synchronization.

**Oscillation constant** is the product of inductance and capacitance which are resonant at a certain frequency. The formula is as follows.

$$\begin{array}{l} \text{Inductance,} \\ \text{microhenrys} \end{array} \times \begin{array}{l} \text{Capacitance,} \\ \text{mmf.} \end{array} = \frac{25330.2}{(\text{megacycles})^2}$$

It is evident that resonance at any given frequency may be had with an almost infinite variety of inductances and capacitances, provided one of these factors is increased as the other is increased, to preserve the same product. Also, any combination of inductance and capacitance will be resonant at some frequency which satisfies the formula. The constant is the product forming the first term of the formula.

**OSCILLATORS.** — An oscillator, as used in radio and television receivers, consists of an electronic tube in a circuit which produces alternating currents from direct-current power. Most oscillator circuits include capacitance and inductance whose resonant frequency is the frequency of the alternating current. Energy oscillates back and forth between the capacitance and inductance as the current reverses. Some oscillator circuits include capacitance and resistance, rather than capacitance and inductance. Then the oscillating or alternating frequency is inversely proportional to the capacitive time constant.

To have sustained oscillation it is necessary to compensate for energy losses in the oscillatory circuits. In what are called feedback oscillators, losses in the grid circuit are replaced by energy fed back from the plate circuit. In this group are inductive feedback oscillators and capacitive feedback oscillators, wherein feedback is respectively through inductive coupling and through capacitance. Less common are negative resistance oscillators in which negative resistance de-

veloped between plate and cathode of a tetrode tube balances the effective resistance of the oscillatory plate circuit.

So long as circuit losses are balanced by feedback or negative resistance these losses no longer are effective in limiting the oscillating current. Limiting usually results from using plate, screen, and bias voltages with which there is plate current saturation before undesirable values are reached during the half-cycle in which the grid becomes positive. There is plate current cutoff during the negative half-cycle.

At frequencies up to a few megacycles both triodes and pentodes are used as oscillator tubes, with pentodes favored when it is desired to limit the feedback which may occur through the grid-plate capacitance of the tube. Triodes are favored for r-f oscillators operating in the carrier frequency range. The grid-plate capacitance of a triode often is part of the resonating capacitance for the oscillator circuit. With any type of tube, the same characteristics which are desirable for amplification are desirable also for oscillation.

**OSCILLATORS, BEAT-FREQUENCY.** — A beat-frequency oscillator consists of two separate oscillators whose frequencies beat together to produce an output frequency equal to the difference. Beat-frequency oscillators are used in testing instruments producing audio frequencies in a range usually extending from 20 to 20,000 cycles, also in types producing radio frequencies up to about 5 megacycles.

The general scheme of connections is shown by Fig. 89-1. The two oscillators at the left are represented as tuned plate types with tickler feedback, but any of various types may be used. Frequency of the upper variable oscillator is tuned by capacitor  $Ct$  in obtaining desired beat frequencies. An initial setting for the frequency of the lower fixed oscillator is made with capacitor  $Ca$ . This setting is used to bring both oscillators to the same frequency, and to zero beat, when the instrument dial shows zero output frequency.

Outputs of the two oscillators may be fed together to any type of mixer tube, but to lessen the tendency of one oscillator to pull into step with the other when approaching zero

beat, it is common practice to use a pentagrid mixer as shown in the diagram. Each of the oscillators is connected to one of the control grids of the mixer tube.

The output of the mixer contains not only the desired beat frequency or difference frequency, but also the two original frequencies, the sum frequency, and many harmonics. For a reasonably pure output from the instrument it is

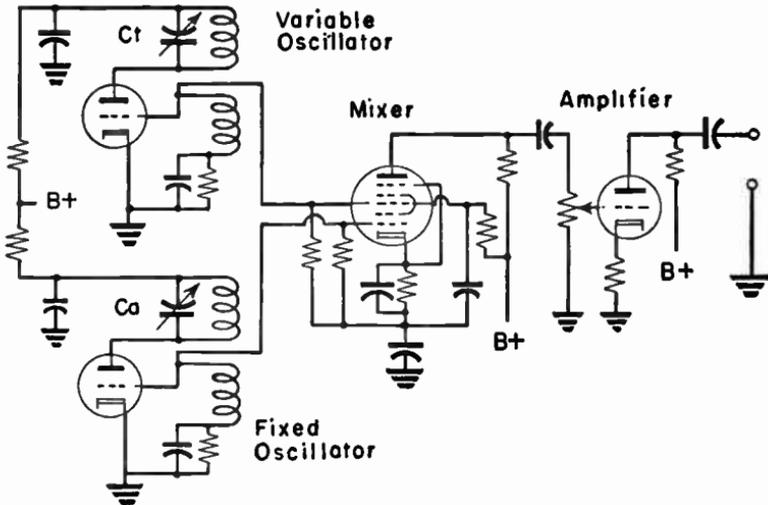


Fig. 89-1.—Circuits in one type of beat-frequency oscillator.

necessary to filter the mixer output. Usually the mixer is followed by an amplifier stage which serves as a filter and at the same time strengthens the rather weak signal from the mixer.

The variable and fixed oscillators must be well shielded from each other, and care used to eliminate common resistances, inductances, and capacitances in the two oscillator circuits. It is important that the only coupling be at the mixer. The variable oscillator may be tuned either higher or lower than the fixed oscillator to provide a desired beat frequency. The fixed oscillator sometimes is crystal controlled. Adjustment for zero beat frequency then must be provided in the circuit of the variable oscillator.

A small percentage change in frequency of the variable oscillator, or either oscillator, causes a relatively large percentage change in beat frequency and instrument output frequency. For example, with one oscillator at 100,000 cycles and the other at 105,000 cycles for an output frequency of 5,000 cycles, a change of 1,000 cycles or 1 per cent in the first oscillator frequency will change the output by 1,000 cycles, which is 20 per cent of the original output frequency. All precautions are taken to prevent greater or less temperature drift in the frequency of one oscillator than in the other one. If both oscillators drift higher or lower together there will be little or no change in output frequency.

**OSCILLATOR, ELECTRON COUPLED.** — An electron coupled oscillator is any radio-frequency oscillator with which the only intentional coupling between the tuned oscillating circuit and the output or load is by means of variations in the electron stream within the oscillator tube. The purpose is to prevent the capacitance and inductance of the load from reacting on the oscillator circuit to alter its tuned frequency.

Fig. 90-1 shows a Hartley oscillator arranged for electron coupling. Other types of oscillators may be used in a similar

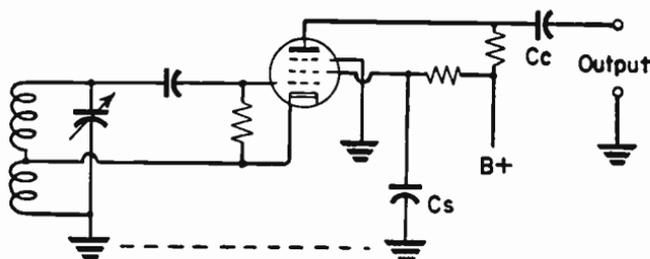


Fig. 90-1.—An electron coupled Hartley oscillator.

way. The tube is a pentode with its suppressor grounded to act as an r-f shield between the screen and the plate. The portion of the tube used as an oscillator consists of the screen, the control grid, and the cathode, with the screen acting as the oscillator plate. The screen (oscillator plate) circuit is completed through capacitor  $C_s$ , the ground connections,

the lower part of the tuned coil at the left, and to the cathode of the tube.

The plate is operated at a higher voltage than the screen. Electron flow from cathode to plate, and through the output and load, must pass through the screen. The screen-to-cathode potential is varying at the oscillating frequency and this varying screen potential modulates the plate current or output current at the oscillating frequency. The load has little effect on oscillation frequency because, in any pentode, changes of plate voltage have little effect on plate current, which is controlled almost entirely by screen voltage.

Electron coupled oscillators often are used in signal generators. The r-f output of such a generator may be modulated with audio frequency by putting the a-f voltage into the suppressor of the oscillator tube. That is, instead of connecting the suppressor to ground, as in the diagram, it is connected through some source of a-f modulating voltage and thence to ground.

**OSCILLATORS, FEEDBACK.** — Feedback oscillators include many types, but all are of the general class in which the tube acts as an amplifier of alternating voltages caused to appear in its grid circuit, and in which energy or power dissipated in the grid circuit is replaced by power fed back from the plate circuit of the oscillator tube. With circuit losses thus compensated for, oscillation will continue at the resonant frequency of capacitance and inductance in either the plate circuit or the grid circuit. Oscillation is begun in the first place by any change, however small, in plate current or voltage. Even the changes due to normal instantaneous variations of emission are enough to cause feedback and introduce a change of voltage into the grid circuit. Then oscillations build up to maximum value in a brief fraction of a second.

Feedback from plate circuit to grid circuit may occur in any of the three principal ways illustrated in Fig. 91-1; by means of a common magnetic field, by means of a common inductance, or through tube capacitance or other capacitance between the plate and grid circuits.

When considering any type of feedback it will be helpful to keep in mind the phase relations of grid voltage, plate current, and plate voltage. When a grid becomes more positive or less negative there is increase of plate current, but because of increased voltage drop in the plate load there is a simultaneous decrease of plate voltage at the tube. When the grid becomes less positive or more negative there is decrease of plate current, but an accompanying increase of plate voltage at the tube. Grid voltage and plate current are of like phase, but grid voltage and plate voltage are in opposite phase.

At *A* in Fig. 91-1 there is feedback by means of inductive coupling through a single magnetic field common to one coil in the plate circuit and to another coil in the grid circuit. Relative directions of currents in the two coils must be such

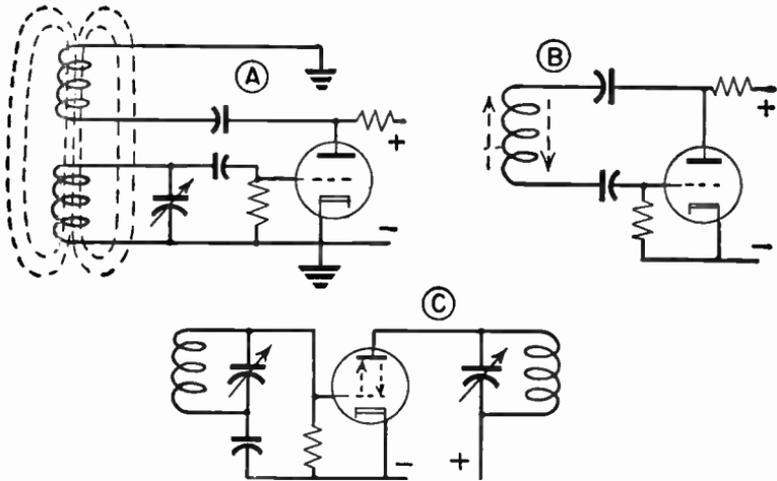


Fig. 91-1.—The three principal ways in which feedback occurs in oscillators.

that the magnetic field of the plate coil aids the field of the grid coil, or so that north and south magnetic poles of both coils point the same direction at the same time. The phase relationship of voltages and currents in grid and plate circuits determines the relative directions of the magnetic fields,

which must be in the same direction or polarity to have feedback such as will sustain oscillation.

If directions of currents in the coils happen to be such as produce opposite magnetic fields, and oppose rather than sustain oscillation, it is necessary only to turn either coil end for end, or else to interchange the circuit connections to the ends of either coil. Feedback voltage or current must be in correct phase no matter how it is accomplished.

In diagram *B* of Fig. 91-1 the same coil is in both the plate circuit and the grid circuit for oscillatory currents. The opposite phase of voltages at plate and grid insures feedback which sustains oscillation. That is, with the coil a continuous winding in the same direction throughout its length, the positive voltage polarity of the grid at one end acts to cause electron flow in the same direction as does negative voltage polarity of the plate at the other end, or vice versa.

In diagram *C* of Fig. 91-1 the feedback from plate to grid acts through the grid-plate capacitance of the oscillator tube. In order that feedback may reach the grid in such phase as to aid changes of grid voltage, the plate circuit must act as an inductive reactance rather than as a pure resistance. This is accomplished by tuning the plate circuit to a frequency slightly higher than the oscillating frequency, which is controlled by resonant frequency of the grid circuit.

With feedbacks such as have been described, every increase or decrease of plate voltage acts through the feedback to cause still further change of grid voltage in the direction that is responsible for that increase or decrease of plate voltage. This further change of grid voltage causes an even greater increase or decrease of plate voltage, and so the interaction continues until plate current reaches saturation value if it is increasing, or drops to zero at plate current cutoff if it is decreasing.

At either saturation or cutoff, the feedback can have no more increase, and the momentary loss of energy causes grid voltage to commence a reversal. This grid voltage acts immediately to start a reversal of the previous change of plate current and voltage, thus commencing a feedback in the opposite

direction or alternation. This opposite feedback continues until plate current again must cease its changing as it comes to either saturation or cutoff. Here the grid voltage is again caused to reverse, and the whole process repeats over and over.

The result of this repeated process is continual rise and fall of plate current between cutoff and saturation, and corresponding changes of voltage across the load in the plate circuit. These changes are the alternating components of plate current and voltage, having the frequency of oscillation or of resonance in the tuned circuit which controls the oscillating system.

As will appear in following circuit diagrams, oscillator tubes nearly always are operated with grid-leak bias, a method which supplies negative biasing voltage proportional to grid current.

**Hartley Oscillators.** — The Hartley oscillator circuit, shown in principle by Fig. 91-2 is used generally in both radio and television receivers for generation of low and intermediate radio frequencies, also in r-f signal generators for service work and in some audio-frequency generators and audio modulating circuits.

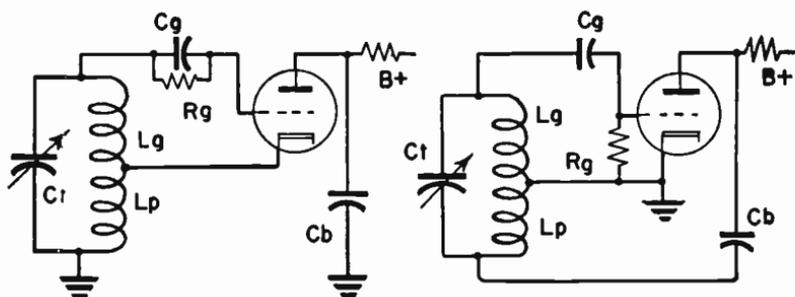


Fig. 91-2.—Hartley oscillators with grounded tuning capacitor (left) and with grounded cathode (right).

Two coils,  $L_g$  and  $L_p$ , or a single tapped coil, are tuned to resonance at the desired oscillator frequency by capacitor  $C_t$ . The two coils or two parts of a coil usually are so mounted as to have inductive coupling, although such coupling is not essential for the reason that the same oscillating currents always must flow in both parts. Feedback is from coil  $L_p$ , in the

plate circuit, to coil  $L_g$  in the grid circuit. The two coils ordinarily are about alike, or a single coil is tapped near its center. Output amplitude is increased by more inductive reactance in the plate coil.

With connections shown by the left-hand diagram of Fig. 91-2 one side of the tuning capacitor and one end of the tuned coil are grounded, an arrangement desirable for prevention of body capacitance effects if tuning is to be manually altered while the oscillator is operating. In the right-hand diagram the cathode of the oscillator tube is grounded, with both ends of the tuned circuit at potentials above ground. The oscillatory

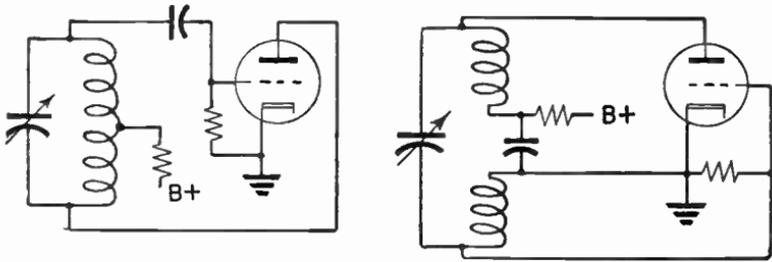


Fig. 91-3.—Modifications of the Hartley oscillator circuit.

circuit is connected to the tube plate through capacitor  $C_b$ . Capacitance at  $C_b$ , also at the grid capacitor  $C_g$ , should be great enough to have small reactance at oscillator frequencies.

Hartley oscillator circuits are found with many modifications, two of which are shown by Fig. 91-3. The essential and easily recognized feature of this circuit is connection of the oscillator cathode to a tap on a single coil or to a point between two coils which are tuned to resonance by a single capacitor or by a principal capacitor paralleled by a trimmer or a fine tuning capacitor.

**Colpitts Oscillators.** — The Colpitts oscillator circuit is shown by Fig. 91-4. This is a type in general use for r-f oscillators in the tuners of television receivers, also for the very-high frequency oscillators in television types of signal generators. The connection method at the left is the one almost al-

ways used. There is grounding of one side of each of the tuning capacitors  $C_a$  and  $C_b$ , also of the tube cathode. In the much older method of connection shown at the right both sides of both capacitors are above ground potential.

The single coil  $L_t$  is tuned to resonance by varying both tuning capacitors in unison, for more capacitance in both or else for less capacitance in both. Oscillating voltage in the plate circuit is across capacitor  $C_b$ . Capacitor  $C_a$  is in the grid circuit. Thus part of the energy put into the tuned circuit from the plate is fed back to the grid. Feedback is increased by increasing the ratio of capacitive reactance at  $C_b$  to capacitive reactance at  $C_a$ .

In the circuit connections at the left in Fig. 91-4, feedback is accomplished through the magnetic field of coil  $L_t$  which is common to both the plate and grid circuits. Oscillating energy

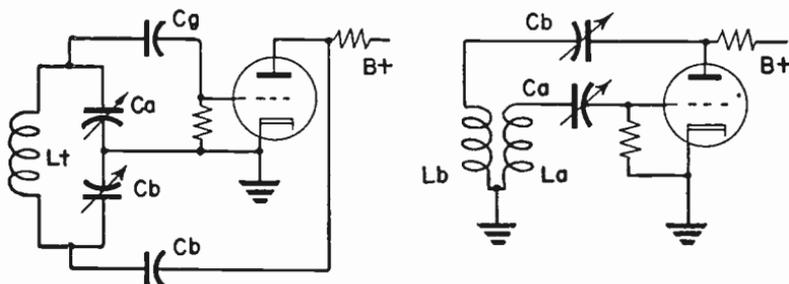


Fig. 91-4.—Colpitts oscillator circuits.

which is in this coil during one half-cycle is divided between the tuning capacitors on the opposite half-cycle. With connections at the right, feedback is by means of inductive coupling from plate coil  $L_b$  to grid coil  $L_a$ . Colpitts oscillators have less tendency than most other simple oscillators to produce harmonics of the tuned frequency.

**Tickler Feedbacks.** — With the oscillator circuits of Fig. 91-5, energy from the plate circuit is fed back to the tuned grid circuit by inductive coupling between the “tickler” coil  $L_p$  in the plate circuit and coil  $L_g$  in the grid circuit. Feedback is varied by changing the position or the angle of coil  $L_p$  in relation to coil  $L_g$ . The tuned-grid tickler-feedback circuit is used in

radio-frequency oscillators operating at frequencies up to 10 or 15 megacycles, and occasionally in audio-frequency oscillators. This circuit, operated with feedback just below that for oscillation, was used in most of the regenerative detectors for early radio broadcast receivers.

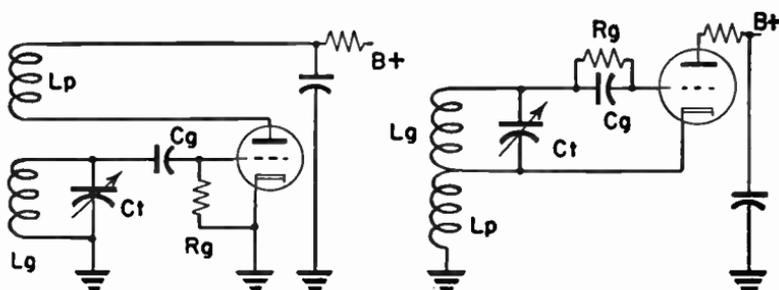


Fig. 91-5.—Tuned grid oscillators with tickler feedbacks.

In Fig. 91-6 the plate circuit rather than the grid circuit is tuned to resonance, and to the coil in this tuned plate circuit is inductively coupled a tickler coil which is connected into the grid circuit. Such tuned-plate tickler-feedback circuits are used in r-f signal generators operating at frequencies as high as 15

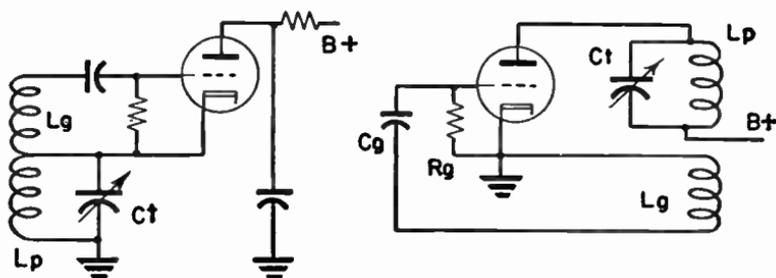


Fig. 91-6.—Tuned plate oscillators with tickler feedbacks.

to 20 megacycles and sometimes in audio-frequency generators. With the connection scheme shown at the left, one side of the tuning capacitor  $C_t$  is grounded. Although neither side of the tuning capacitor is grounded in the right-hand circuit, the one which connects to  $B+$  may be grounded for radio frequencies through a bypass capacitor.

**Other Feedback Oscillators.** — At *C* in Fig. 91-1 are shown the essential features of what is called a tuned-grid tuned-plate oscillator. There is no inductive coupling between the tuned coils in plate and grid circuits. Feedback current flows in the capacitance between plate and grid of the oscillator tube, which is a triode because triodes have greater grid-plate capacitance than pentodes.

In Fig. 91-7 is illustrated the usual form of ultraudion oscillator circuit. Feedback is through the magnetic field of tuned coil *L* which is common to both the plate circuit and the grid

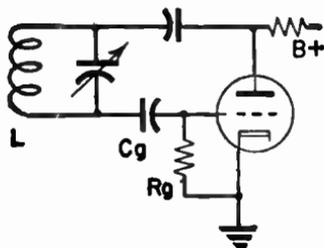


Fig. 91-7.—Ultraudion oscillator circuit.

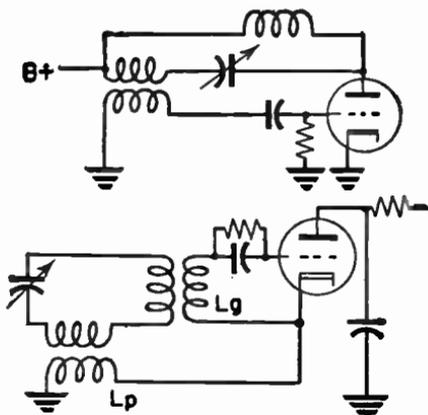


Fig. 91-8.—Two variations of the Meissner oscillator circuit.

circuit. The amount of feedback may be controlled by connecting a variable bypass capacitor from plate to cathode or plate to ground, through which part of the feedback energy is shunted.

Fig. 91-8 shows at the top an Arco-Meissner oscillator circuit in which there is a tuned plate circuit consisting of a series connected variable capacitor and coil which are paralleled by another coil. To the series plate coil is inductively coupled an untuned coil in the grid circuit to which is transferred the feedback energy.

Down below is the oscillator circuit usually associated with the name Meissner. Neither the plate circuit nor the grid cir-

cuit is tuned, but coils in both these circuits are inductively coupled to coils which are in a tuned link circuit. Feedback is from coil  $Lp$  in the plate circuit through the link to coil  $Lg$  in the grid. The Meissner oscillators are not in common use for present radio and television receivers, but sometimes are referred to.

**OSCILLATORS, FREQUENCY DRIFT IN.**—Frequency stability and constancy are necessary in any oscillator which is calibrated, as are the oscillators of signal generators, and is highly desirable where oscillator output is combined with other frequencies, as in superheterodyne tuners and controls for television sweep frequency. Frequency stability in test instruments is best secured by crystal control where crystal frequencies or their harmonics can be made to cover enough points in the operating band. Automatic frequency control is generally used for horizontal sweep oscillators, sometimes for vertical sweep oscillators, and occasionally for r-f oscillators in tuners.

Slow drift of oscillator frequency during possibly the first half hour of operation usually is due to temperature effects which alter the positions and dimensions of parts to change the inductance of coils and the capacitance of capacitors and tubes. Naturally, anything that lessens the production of heat will help. This will include the use of tubes requiring relatively small plate current, and of circuits and parts having least possible resistance. High-wattage resistors may allow improvement by reducing their own temperature, although this will not lessen the total production of heat. Temperature rise is lessened by good ventilation, absence of crowding, and by keeping parts free from dust and other dirt. Heat may be carried away and also distributed more uniformly by chassis metal and shielding of heavy gauge and of good thermal conductivity, as found in copper and aluminum.

When all practicable changes have been made in design and construction, the remaining temperature drift may be balanced out by connecting across the tuned oscillator circuits temperature compensating ceramic capacitors having

negative temperature coefficients. These compensating capacitors must, of course, become active parts of the circuits and replace all or part of non-compensating types of capacitance which would be connected into the same positions. As temperature rises there is decrease of the compensating capacitance with a tendency to raise the oscillator frequency to the same extent it would be lowered by heating of the various other parts.

More rapid changes of oscillator frequency may result from changes of load unless the tuned oscillating circuits are well isolated from the output. Such isolation usually is provided in test instruments by using electron coupled oscillators or by using untuned buffer amplifiers between the oscillator itself and the output to the load.

Fluctuations of plate and screen voltage will alter the frequency of oscillation by changing plate resistance and transconductance of the oscillator tube. The best preventive is use of standard voltage regulating tubes or voltage regulating transformers or both. Effects of voltage variation are minimized by using oscillator tubes of fairly high plate resistance and by working with strongly negative grid bias and the required high plate supply voltage. The use of insulation and other parts having minimum high-frequency resistance and losses helps maintain frequency stability.

Very rapid changes of oscillator frequency may be due to parts which are loose or which vibrate, or to loose electrical connections. Very slow changes which do not correct themselves usually are due to capacitors which have developed some leakage or which have changed their capacitance, or to resistors which have changed their resistance with aging. Gradual aging of tubes will alter the oscillating frequency. If no parts prove actually defective it is best not to make replacements but to make readjustments or recalibrations after the first one or two hundred hours of use. Apparatus then should operate for long periods without enough additional change to cause any difficulty.

**OSCILLATORS, NEGATIVE RESISTANCE.**— In negative resistance oscillators there is no feedback from the plate circuit to compensate for loss of energy in the grid circuit; rather there is utilized an effect called negative resistance to compensate for losses in a tuned circuit connected between plate and screen and cathode of the oscillator tube.

Connections for a dynatron type negative resistance oscillator are shown by Fig. 93-1. The tube is represented as a tetrode or screen grid tube, but might be a pentode connected as a tetrode or might be a triode. The essential feature for dynatron operation is that the element between plate and cathode, which may be either a screen or a control grid, shall be at a potential higher than that on the plate.

Fig. 93-2 shows variations of dynatron plate current, screen current, and total or cathode current when the screen is maintained at constant voltage while there are increases and decreases of plate current and voltage at the oscillating frequency, as occur during operation of any oscillator. With zero plate voltage all emission goes to the screen, and screen current is total current. With increasing plate voltage, more and more electrons are drawn through the screen to strike the plate. These primary electrons cause emission of second-

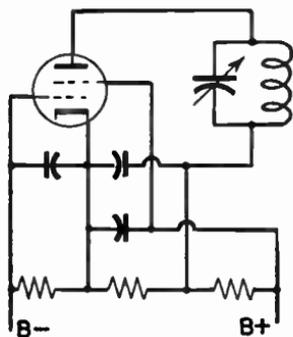


Fig. 93-1.—Circuit connections for a dynatron oscillator.

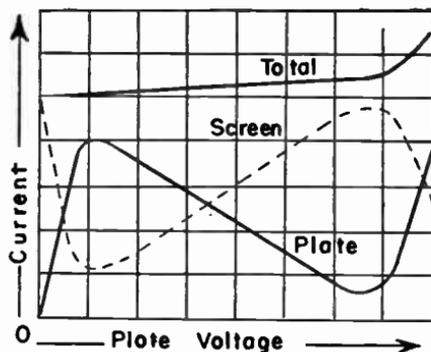


Fig. 93-2.—Currents in a dynatron oscillator tube.

creases of plate current and voltage at the oscillating frequency, as occur during operation of any oscillator. With zero plate voltage all emission goes to the screen, and screen current is total current. With increasing plate voltage, more and more electrons are drawn through the screen to strike the plate. These primary electrons cause emission of second-

ary electrons from the plate. The secondary electrons then are drawn to the more positive screen and add themselves to the screen current.

Continued rise of plate voltage increases the velocity of electrons striking the plate, which increases the rate of secondary emission. Soon there are more secondary electrons leaving the plate than primary electrons coming to the plate from the cathode. The result is decrease of electrons remaining in the plate and flowing out to the connected circuit in the form of plate current. If plate voltage rises above screen voltage there is no further dynatron action, and, as at the right in Fig. 93-2, there are rapid rises of plate current and total current, with rapid drop of screen current. The grid of the tube is maintained at such negative voltage as limits emission to a suitable rate and helps maintain a constant emission rate.

With a tube operated in the ordinary manner (plate voltage higher than screen voltage) every increase of plate current in the tube and the plate load means more voltage drop between plate and cathode. Consequently, plate current soon is limited by lack of voltage applied from the B-supply through the load. But when the tube is operated as a dynatron the increase of plate current means less voltage drop between plate and cathode, and more B-supply voltage becomes available to the tuned load circuit connected between plate and cathode. Any increase of current through the tuned circuit now is accompanied by an increase of voltage applied across it, while decrease of current is accompanied by decrease of applied voltage. Consequently, once the load current commences to either increase or decrease, there is a change of voltage across the load (plate to cathode) tending to accelerate the increase or decrease. This effect is equivalent to that obtained with any form of feedback.

Dynatron oscillators are used chiefly in test equipment, where they provide fair frequency stability and good waveform, but usually require considerable recalibration when the oscillator tube is replaced. Among other negative resistance oscillator circuits is the negative transconductance

type employing a pentode tube with the tuned oscillatory circuit connected between screen and cathode. A modulating voltage may be applied between grid and cathode of a negative resistance oscillator.

**OSCILLATORS, RELAXATION.** — A relaxation oscillator is a type in which control grid voltage is changed by the charging and discharging of a capacitor or inductor through a resistance rather than as a result of oscillations in capacitance and inductance. That is, the operating frequency of the relaxation oscillator depends on capacitive or inductive time constants rather than on resonant frequency in a capacitance and inductance. The output of a relaxation oscillator is not naturally a sine wave, but may be a pulsed wave, a square wave, a triangular wave, a sawtooth wave, or other forms. The output may be changed to a sine wave by suitable filtering.

The frequency of a relaxation oscillator may be quite easily synchronized with that of a periodic voltage. The oscillator is adjusted to operate naturally at a frequency slightly lower than the desired rate, and the synchronizing voltage is applied in a manner to hasten the capacitor charge or the change of grid voltage and thus start each cycle of oscillation in time with synchronizing pulses.

**Multivibrator Oscillator.** — One of the earliest relaxation oscillators is the multivibrator, of which a typical circuit is shown by Fig. 94-1. It should be noted that this style of multivibrator is used for testing and frequency-measuring equipment, but is not the type employed as a sweep oscillator or deflection oscillator in modern television receivers. The oscillator shown here is essentially a two-stage resistance coupled amplifier with the output of one stage fed back as input to the other stage.

Feedback is in such polarity as to reinforce the changes of grid voltage in the first stage, this being due to the fact that voltage is inverted in polarity between grid and plate of any tube. For example, when the grid of the first tube is going positive its plate is going negative. This causes the grid of the second tube to be driven negative, and the plate of that tube

goes positive. The positive-going plate voltage from the second tube is fed back and added to the positive-going grid voltage of the first tube, with which the action commenced.

The feedback voltages which act in the same way as existing grid voltages drive plate current to saturation in one tube and to cutoff in the other tube. Then the actions can continue

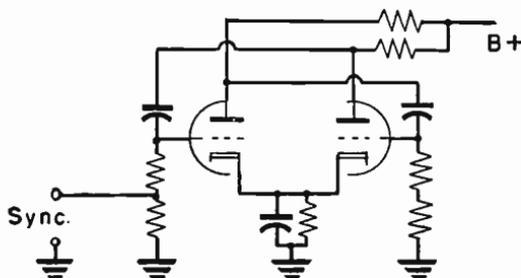


Fig. 94-1.—Connections for a multivibrator oscillator.

no further in the original directions, but reverse and again cause saturation and cutoff in the two tubes. Operating frequency depends on time constant of capacitors feeding the grids and the resistances between grids and ground. Frequency may be altered by adjustment of either capacitance or resistance, usually the latter. A synchronizing voltage may be put into either grid circuit.

**Gaseous Tube Oscillator.** — A relaxation oscillator shown in principle at the left in Fig. 94-2 uses a gaseous triode (thyatron) whose frequency of breakdown, along with output frequency of the system, is determined by the time constant of a capacitance-resistance combination. This type, and modifications of it, are used as sweep oscillators in many oscilloscopes.

Capacitor  $C$  is connected between plate and cathode of the gas-filled tube. The maximum voltage to which the capacitor may charge is the breakdown or ionization voltage of the tube. The capacitor charges from a d-c source through adjustable resistor  $R_a$ . The time required for capacitor voltage to rise to the breakdown value of the tube depends on the time constant of  $C$  and  $R_a$ . During the charging time there is gradual in-

crease of capacitor voltage, which also is the output voltage from terminals at the right. When the tube breaks down or ionizes, its internal resistance suddenly drops to a very low value, and the capacitor discharges through the tube from cathode to plate until voltage drops to a value at which the

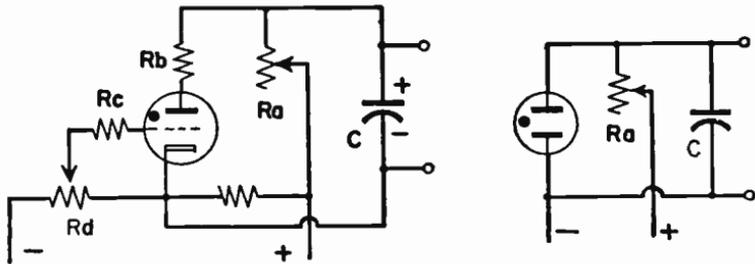


Fig. 94-2.—A gaseous tube relaxation oscillator (left) and a neon oscillator (right).

tube deionizes. Tube resistance then returns to a high value, the capacitor re-charges, and the process repeats. The output voltage is of sawtooth waveform.

Resistors  $R_b$  and  $R_c$  respectively limit the plate current after breakdown and limit the grid current. Grid voltage may be made more or less negative by adjustment of  $R_d$ . Grid voltage or bias voltage determines the plate-cathode and capacitor voltage at which the tube will break down, and so determines maximum output voltage. Frequency of oscillation is inversely proportional to the time constant of  $C$  and  $R_a$ . Various ranges of frequency may be covered by switching different values of capacitance into the circuit at  $C$ .

At the right in Fig. 94-2 is a simple "neon oscillator," which is a relaxation type. The tube is a neon lamp. D-c supply voltage is higher than breakdown voltage of the lamp. Capacitor  $C$  charges through resistor  $R_a$ . Oscillation frequency depends inversely on the time constant of  $R_a$  and  $C$ . Output voltage is of sawtooth waveform.

**OSCILLATORS, TELEVISION, R-F.**—A television radio-frequency oscillator produces very-high frequencies which beat with received carrier frequencies to produce intermediate fre-

quencies which are amplified in the intermediate frequency amplifier. The oscillator and carrier frequencies are combined in the mixer tube, whose output goes to the intermediate-frequency amplifier. This oscillator, which is part of the tuner, is called the r-f oscillator or the local oscillator to distinguish it from the television sweep oscillators.

In the television r-f oscillator circuit there is practically always a separate tube or separate section of a twin tube serving this one function, rather than having oscillator and mixer functions combined in a converter tube having a single electron stream, as is general practice in receivers for sound radio. Often the television r-f oscillator is one section of a twin tube whose other section is the mixer, but there are separate electron streams for the two functions. The great majority of r-f oscillator tubes are miniature triodes of high transconductance.

The output of the r-f oscillator consists of a single sharply tuned frequency for each channel. This oscillator frequency beats with the amplitude-modulated video carrier frequency to produce a video intermediate frequency having the same signal modulation and extending over the same band width in megacycles as covered by the video carrier. The same oscillator frequency beats with the frequency-modulated sound carrier in the same channel to produce a sound intermediate frequency having the same signal frequency modulation and covering the same band width as the sound carrier.

It is a general rule that the r-f oscillator frequency shall be higher than the carrier frequencies in the same channel. There are a few television receivers in which the oscillator frequency is lower than the carrier frequencies, but they are the exceptions. The difference between oscillator frequency and video carrier frequency is the video intermediate frequency of the receiver, and the difference between oscillator frequency and sound carrier frequency is the sound intermediate frequency of the receiver. The following tabulation shows these relations in a few channels for a receiver having

video intermediate frequency of 26.60 mc and sound intermediate frequency of 22.10 mc. All the listed frequencies are in megacycles.

Channel	Carriers		Intermediates		Oscillator
	Video	Sound	Video	Sound	
2	55.25	59.75	26.60	22.10	81.85
6	83.25	87.75	26.60	22.10	109.85
7	175.25	179.75	26.60	22.10	201.85
13	211.25	215.75	26.60	22.10	237.85

**Oscillator Circuits.**—The Colpitts circuit is used more than any other for television r-f oscillators, but it often appears in such modifications as to make identification rather difficult. The basic Colpitts circuit is shown at the left in Fig.

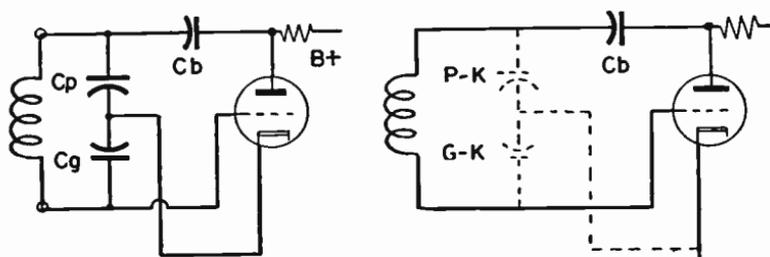


Fig. 95-1.—Colpitts oscillator with external tuning capacitances (left) and with tuning by tube interelectrode capacitances (right).

95-1. Tuning to resonance is by means of capacitor  $C_p$  in the plate circuit and capacitor  $C_g$  in the grid circuit. One side of both these capacitors connects to the cathode, either directly or through ground. The blocking capacitor at  $C_b$  keeps d-c plate current and voltage out of the grid circuit.

At the very-high frequencies existing in television r-f oscillator circuits the interelectrode capacitances of the tube are ample for tuning to resonance with only a moderate amount of inductance in the tuned circuit. Consequently, in television circuit diagrams the Colpitts oscillator may appear as shown by the full lines at the right in Fig. 95-1. Tuning of the oscillator coil is accomplished by plate-to-cathode capacitance of the tube, represented by broken lines at  $P-K$ , and by grid-to-cathode capacitance represented at  $G-K$ . These tube

capacitances take the place of capacitors  $C_p$  and  $C_g$  in the left-hand diagram.

Other Colpitts oscillator circuits are shown by Fig. 92-2. The left-hand diagram contains fine tuning capacitors  $C_m$  and  $C_n$ , of which either one might be used, but not both in the same oscillator circuit. Capacitor  $C_m$  is connected across the entire tuning coil and also is in parallel with the grid-plate interelectrode capacitance of the tube. Capacitor  $C_n$  is in parallel with the plate-cathode capacitance of the tube, which is one of the interelectrode capacitances that tune the coil to resonance. With the fine tuning capacitors removed, the oscillator circuit would be essentially the same as at the right in Fig. 95-1. With fine tuning capacitor  $C_m$  appearing in a circuit diagram the oscillator could be mistaken for an ultradion unless values of capacitance, inductance, and frequency were checked to show that  $C_m$  is not the principal tuning element.

At the right in Fig. 95-2 is illustrated still another modification of the Colpitts oscillator. Capacitor  $C_t$  is for fine tuning.

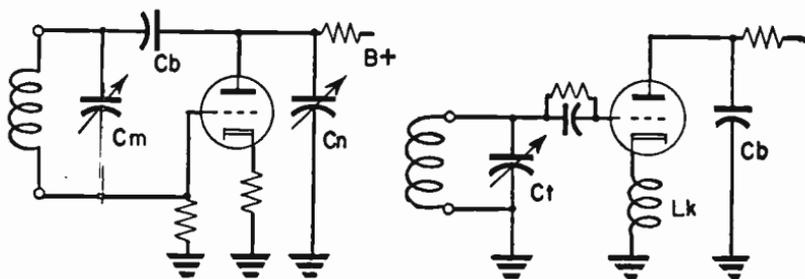


Fig. 95-2.—Circuits connections for modified Colpitts oscillators.

The r-f plate circuit with its plate-to-cathode capacitance is connected to the tuning coil circuit through capacitor  $C_b$  and ground. The interelectrode capacitances of the tube are made effective for tuning by isolating the cathode from ground by means of r-f choke  $L_k$ , so far as r-f currents are concerned, while maintaining the cathode-to-ground conductive path for direct current. There are many modifications of the funda-

mental Colpitts oscillator circuit in addition to those which have been illustrated.

Fig. 95-3 shows circuit connections for a push-pull r-f oscillator employing a twin triode tube and an oscillatory circuit consisting of capacitor  $C$  and coil  $L$  used as a parallel resonant circuit between the two plates. A different coil  $L$ , of suitable inductance for tuning, is switched into position for each channel.

This oscillator operates so far as feedback is concerned like two resistance coupled amplifiers with the output of each plate fed as the input to the opposite grid through coupling capaci-

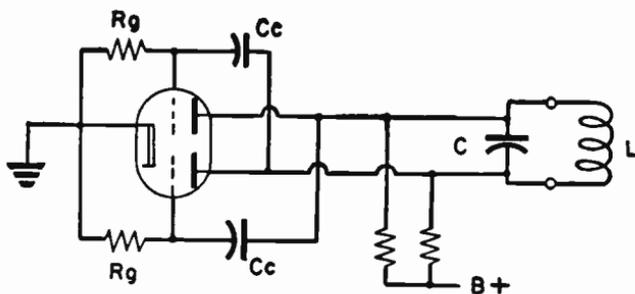


Fig. 95-3.—Push-pull r-f oscillator with parallel resonant tuned circuit.

tors  $C_c$  and  $C_c$ . The reactance of these capacitors is very small in comparison with resistance of grid resistors  $R_g$  and  $R_g$ . The oscillator output is of good frequency stability, and there is little tendency to produce harmonic frequencies.

Fig. 95-4 shows connections for a resonant line oscillator in which the tuned resonant circuit consisting of lumped capacitance and inductance elements is replaced with a quarter-wave shorted resonant line, which is the equivalent of a parallel resonant circuit so far as tuning is concerned. The left-hand end of this diagram is practically the same as the one in Fig. 95-3. Dual capacitor  $C_t$  is a variable fine tuning unit.

• Opposite sides of the resonant line, shown toward the right in the diagram, consist of coils of a few turns, or small loops, or lengths of straight wire between the successive terminal

points, depending on the amount which the resonant frequency must change between tuning positions. The movable shorting bar shown between opposite line terminals is shifted one way and the other to vary the effective length of line and provide resonance in the various channels. The shorting bar

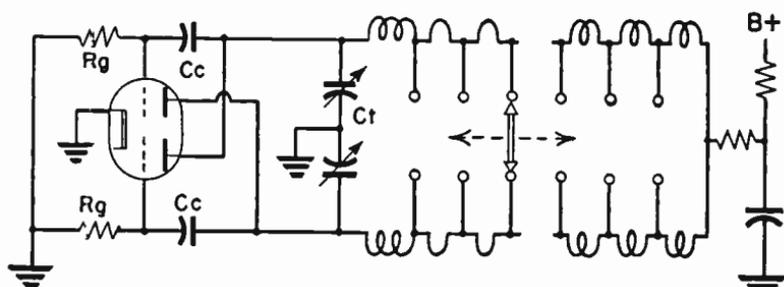


Fig. 95-4.—Resonant line oscillator with a shorted quarter-wave resonant line as the tuned circuit.

or connection usually is a conductor between movable contacts of a rotary switch, whose stationary contacts connect to or support the line inductors. The diagram would represent the circumference of the rotary switch spread out as a flat surface from left to right.

**Effects of Oscillator Frequency.** — Fig. 95-5 illustrates the effects of oscillator tuning or alignment in changing the intermediate frequencies which are produced in the mixer and applied to the input of the i-f amplifier. The curve, which is the same at left and right, represents the frequency response or gain versus frequency in the i-f amplifier. It is assumed that the sound intermediate frequency applied to this amplifier should come low down on the left-hand or low-frequency skirt of the response, and that the video intermediate frequency should come at the point of 50 per cent gain on the right-hand side of the response curve. The two intermediate frequencies always will be separated by 4.5 megacycles, as marked on the diagrams.

If the oscillator is mistuned or misaligned, both the sound and the video intermediate frequencies will be made too low

or too high. Their separation, however, will remain 4.5 mc because separation of sound and video carriers in any channel always is 4.5 mc, and both intermediate frequencies are produced by beating of a single oscillator frequency with the two carriers.

With the oscillator tuned or aligned for a frequency which is too low, both intermediate frequencies will be too low by the same number of megacycles or fraction of a megacycle. This is shown at the left in Fig. 95-5. If the receiver employs the intercarrier sound system the gain or response at the actual sound intermediate frequency will be so low as to allow practically no sound reproduction, while if the sound system is of the dual carrier type the center frequency fed to the sound amplifier will be nearly or completely outside the band pass of this amplifier, and there will be distorted sound or no sound at all. The video intermediate frequency will move up on the

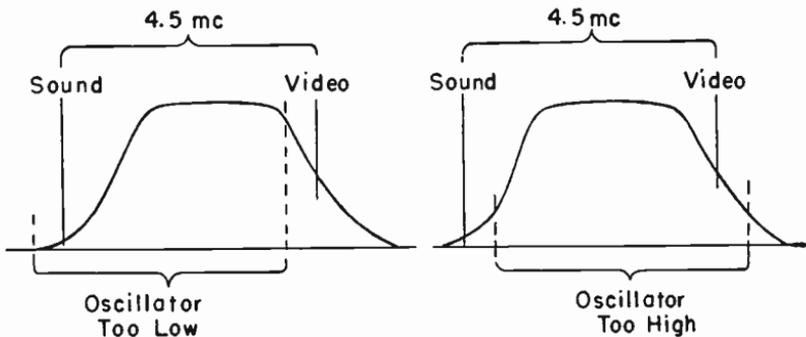


Fig. 95-5.—How shifting of r-f oscillator frequency alters both the intermediate frequencies.

response curve. There will be too much gain at low video frequencies, while the highest video frequencies will be moved down on the opposite side of the curve, to give them too little amplification. This will prevent reproduction of fine details in the pictures.

With the oscillator tuned to a frequency which is too high, the results will be as at the right in Fig. 95-5. The sound intermediate frequency now moves up on the gain curve while

the video intermediate frequency moves down. With an inter-carrier sound system the amplification for sound will be so great as to cause possible overloading of the sound amplifiers, usually also the effect called intercarrier buzz, and it is quite probable that sound bars will appear on the picture or pattern. With a dual carrier sound system the intermediate frequency passed on to the sound amplifier again will be nearly or wholly outside the band pass of this amplifier, and there will be distorted sound or none at all. The video carrier will move down on the gain curve. Low video frequencies will receive too little amplification, and there is probability of trailers and black "speckling" in pictures and patterns. Excessive reduction of low-frequency gain also may make it difficult to obtain satisfactory adjustment of the hold controls.

If oscillator frequency is not too far out of alignment, and if there is a fine tuning control accessible to the operator, this control may have enough range to make corrections in the various channels. With receivers having dual channel sound systems the fine tuning control may bring either satisfactory pictures or satisfactory sound, but not both at the same setting.

**Alignment of R-f Oscillators.** — General instructions for alignment in all parts of the receiver are given in the article on *Alignment*. The following precautions should be observed when aligning the r-f oscillator.

When test signals will pass through the video or sound intermediate-frequency amplifiers to be picked up by meters or oscilloscope at the detectors, these amplifiers should be in reasonably good alignment before working on the oscillator.

If a signal generator is used, the antenna should be disconnected from the receiver.

It is advisable to override any automatic gain control, even though this control does not act directly on the r-f oscillator.

A fine tuning control, if there is one, should be set at the approximate center of its adjustment range, and left there throughout the alignment process. This control sometimes is a service adjustment, as with types consisting of a cam with screw settings for each channel.

Only non-metallic alignment or adjustment tools should be used where frequencies are so high as in r-f oscillators. It is essential that the receiver and all test instruments be allowed to warm up for at least 15 minutes before any adjustments are altered.

In some tuners there is an overall adjustment for the oscillator, an adjustment that alters the tuning in all channels at once. This adjustment, if used, most often is associated with channels 2 or 13.

R-f oscillators in receivers having intercarrier sound systems may be aligned in any of several ways. One method requires use of a sweep generator, a marker generator, and an oscilloscope. The sweep and marker generators are connected or coupled to the antenna terminals of the receiver, with correct impedance matching for the sweep input. The oscilloscope is connected across the output load of the video detector. Sweep frequency and width are adjusted to produce a video i-f response curve on the oscilloscope. Then markers are applied at video and sound carrier frequencies while the r-f oscillator is aligned for correct marker positions as shown by Fig. 95-5.

Approximate alignment may be made with an electronic voltmeter connected across the video detector load and using any television signal for the input. The oscillator is aligned for a meter reading about half way between minimum and maximum shown on the meter as the tuning adjustment is changed throughout its range. Some readjustment usually is required to obtain the combination of sound fully regulated by the volume control and a picture that is clear. With correct alignment the picture will not be of maximum brightness for that would mean a low oscillator frequency and the conditions shown at the left in Fig. 95-5.

If no marker frequencies are available in the carrier ranges it is possible to feed to the mixer input a marker at the video intermediate frequency and one at the sound intermediate frequency. With an oscilloscope connected across the video detector load, these i-f markers will appear on the curve pro-

duced by a sweep generator connected to the antenna terminals and operating through the carrier frequencies for the channel in which adjustments are made. This is little more than a check on the fact that i-f markers will appear at certain positions on the i-f response curve. It will show gross misalignment, but little else.

R-f oscillators in receivers having dual carrier sound systems nearly always are aligned by putting a single-frequency signal (not a sweep frequency signal) through the sound amplifier and demodulator. The band pass of the sound i-f amplifier and demodulator hardly ever is more than a half megacycle and usually is much less. If the r-f oscillator is so aligned that a sound signal comes through this narrow band pass and is well amplified without distortion it may be assumed that the oscillator alignment is sufficiently accurate for all practical purposes. If the sound intermediate frequency is thus found to be correct, the video intermediate frequency must also be correct, and if the video i-f amplifier is properly aligned there will be good reproduction of both pictures and sound.

Some of the many methods of aligning the r-f oscillator through the sound system are as follows. The simplest method, but not the most accurate, is to tune in a signal consisting of a test pattern with accompanying audio note as sound modulation, then to adjust the oscillator alignment for maximum sound volume. An ordinary output meter may be used instead of or to supplement the speaker indications. A method of no greater accuracy, with equipment usually available, substitutes a constant frequency tone modulated signal from a generator of the marker type instead of the television signal. The r-f oscillator is aligned for maximum volume from the speaker or maximum reading of an output meter.

Satisfactory alignment may be made by using a marker generator, unmodulated, connected to the receiver antenna terminals, with an electronic voltmeter connected to the output of the sound discriminator or ratio detector. If the voltmeter is connected to a point of unbalanced demodulator output, not to the output going to the a-f amplifier, alignment

is made for maximum meter reading. Alignment is for maximum meter reading also if the meter is connected across the grid resistor of an audio limiter tube. If the voltmeter is connected to a point of balanced demodulator output, which would be just ahead of the de-emphasis filter in most cases, alignment is made for zero reading on the meter. The marker generator is tuned to the sound carrier frequency for the channel in which alignment is being carried out.

Another satisfactory method employs a sweep generator and marker generator connected to the antenna terminals of the receiver, and an oscilloscope connected to the balanced a-f output of the sound discriminator or ratio detector. The sweep frequency and width are adjusted to produce a demodulator S-curve on the oscilloscope. The marker generator is tuned accurately to the sound carrier frequency of the channel in which alignment is being carried out, and is used without modulation. The oscillator tuning is adjusted to bring the marker to the center of the S-curve.

**OSCILLATOR, TELEVISION SWEEP.** — The purpose of sweep oscillators in television receivers is to control charging and discharging times of a capacitor which furnishes sawtooth voltages for directly or indirectly deflecting the picture tube beam vertically or horizontally. As shown by Fig. 96-1, a sweep oscillator is essentially an electronic switch shunted across the sawtooth capacitor. When the oscillator tube is made non-conductive by highly negative grid voltage, at the left, it allows the capacitor to charge from the B-power supply through a resistor. Direction of electron flow is shown by arrows. When the oscillator tube is made conductive by positive grid voltage, at the right, the capacitor discharges through the tube.

Charging of the capacitor is at a relatively slow rate, which produces the gradual rise of sawtooth voltage for causing a horizontal active trace or a vertical downward travel of the beam in the picture tube. Charging continues while the grid of the tube is held negative. When the grid is made positive there is rapid discharge of the capacitor, producing the sharp

drop of sawtooth voltage which causes horizontal or vertical retrace in the picture tube. The discharge period lasts only while the grid of the tube is held positive. Capacitor discharge and retrace are immediately followed by another charge and active deflection of the beam.

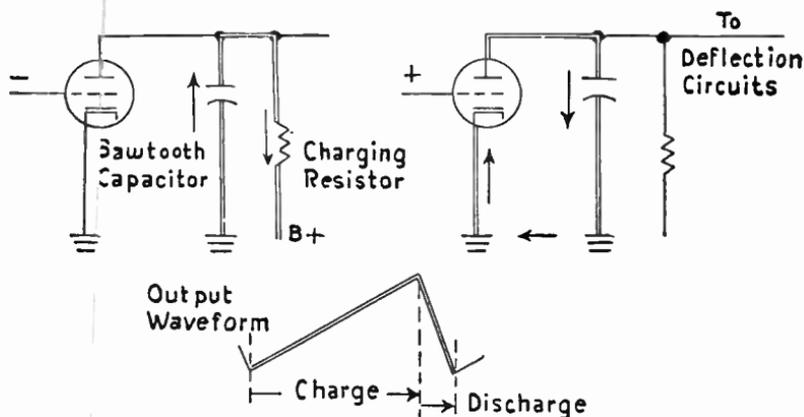


Fig. 96-1.—How a sweep oscillator controls charge and discharge of the capacitor.

The time constant of the sawtooth capacitor and charging resistor is too long to permit charging the capacitor to full voltage of the B-supply during the time in which the tube grid remains negative. The charge is thus limited in order to use only the first portion of the charging curve, which is fairly straight or linear. This helps maintain satisfactory linearity of deflection.

The frequency at which the capacitor is charged and discharged must be the same as the deflection frequency, which is 60 cycles per second for vertical deflection or is 15,750 cycles per second for horizontal deflection. In circuits having no automatic control for sweep frequency, and often in circuits which do have such control, the sweep oscillators are either of the blocking type or else are multivibrators. In both these types the operating frequency is fixed by combined effects of resistance-capacitance time constants in the oscillator grid circuits and the vertical or horizontal sync pulses coming to these circuits.

Frequency is regulated by adjustment of a grid circuit resistor which, in connection with a grid circuit capacitor, determines the time period during which the grid remains so negative as to keep the tube non-conductive. The adjustable resistor is a hold control, shown in Fig. 96-2. Frequency is adjusted to a rate just a little slower than that of the sync pulses.

Changes of grid voltage which would occur in the absence of sync pulses are shown at the center in Fig. 96-2. The principal differences between various blocking oscillator circuits and various multivibrator circuits are in the manner of producing

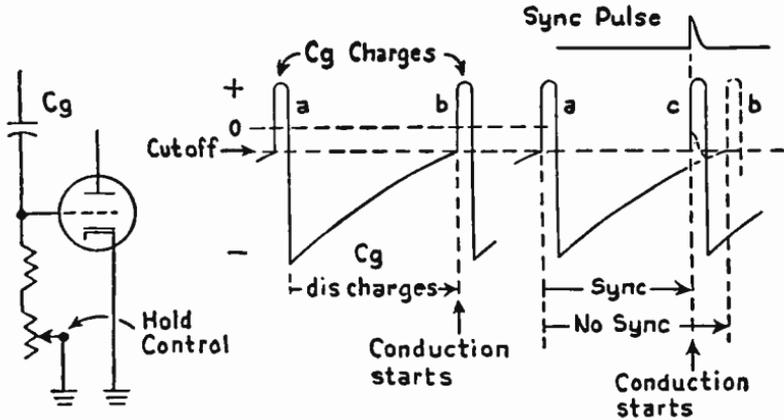


Fig. 96-2.—How sweep oscillator frequency is controlled by sync pulses.

these grid voltage changes. There is first a short interval,  $a$ , during which the grid is made positive and in which flow of grid current charges grid capacitor  $C_g$ . At the end of this interval the grid is suddenly made negative to a value well below plate current cutoff. Then the grid capacitor discharges rather slowly through the hold control resistor until grid voltage rises to the cutoff value, whereupon conduction in the tube starts a train of events which cause the grid again to become positive, at  $b$ , and the entire grid voltage cycle repeats over and over.

The effect of sync pulses is shown at the right. These pulses, of positive polarity, add their potential to the grid voltage just before the instant at which conduction would have started due to the action previously described. The result of adding the sync pulse potential to the grid voltage is to bring grid voltage up past the cutoff value at the instant of the pulse. This starts conduction in the tube, and there is a charging interval at *c* instead of at *b* as would have occurred with no sync pulse. The time period from *a* to *c*, marked *Sync*, corresponds to the correct deflection frequency as determined by sync pulse frequency. This time is shorter, or the frequency is somewhat higher, than *a* to *b* which is marked *No Sync*.

The sweep oscillator will continue to act even when there are no sync pulses, as when no television signal comes into the receiver circuits. Action will be at a frequency somewhat lower than signal sync frequency, as at the center in Fig. 96-2, but there will be continual deflection of the beam in the picture tube.

**Blocking Oscillator.** — Fig. 96-3 shows a typical circuit for a blocking oscillator. Grid capacitor  $C_g$ , the hold control resistor, and the sawtooth capacitor have been shown in

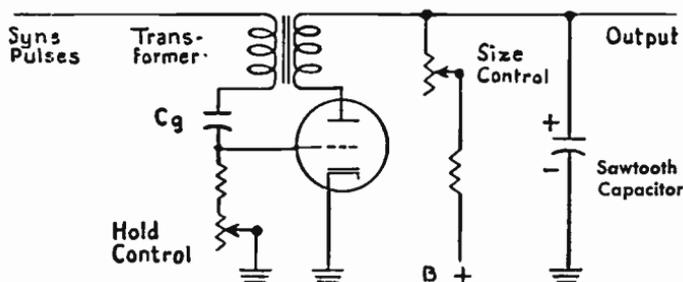


Fig. 96-3.—Blocking oscillator connections.

preceding figures. The size control resistor of Fig. 96-3 is equivalent to the charging resistor of Fig. 96-1. There is inductive feedback from plate to grid of the oscillator tube through the transformer. Transformer connections are such

that feedback is of polarity which increases the grid voltage in whichever way it is already changing. That is, the feedback is positive or is regenerative. Sync pulses are applied to the grid circuit through the grid winding of the transformer.

There are continual changes of plate current in the tube, but to explain the action we shall commence with a period during which plate current is increasing in the tube and in the plate winding of the transformer. At this time the grid voltage is becoming less negative or more positive, which accounts for increase of plate current. Regenerative feedback makes the grid still more positive. As the grid becomes positive there is flow of grid current, which charges capacitor  $C_g$  to make the side toward the grid of negative polarity. The charge voltage will reach approximately the peak value of feedback voltage. This action is like that occurring with grid-leak bias.

Feedback from the plate circuit continues because plate current continues to increase. The increase of plate current is not instantaneous because it is acting through inductance

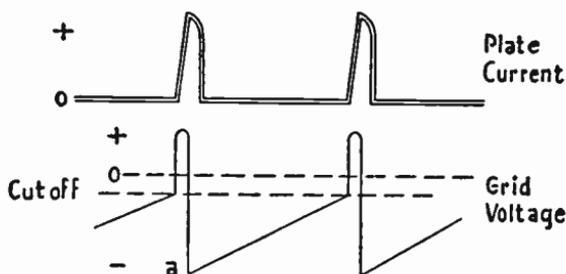


Fig. 96-4.—Changes of plate current and grid voltage in a blocking oscillator.

of the transformer plate winding wherein counter-emf slows down the rate of increase. Consequently, the grid is driven more and more positive, which causes plate current to increase to saturation at whatever value is proportional to applied plate voltage. This increase of plate current is shown at the top of Fig. 96-4, with the simultaneous changes of grid voltage shown below.

The length of time for plate current to reach saturation increases with more inductance in the transformer. It is desirable to have small inductance and small distributed capacitance to lessen this time and keep the plate current pulses narrow or of short duration, for retrace time cannot be short unless these pulses are short. When plate current reaches saturation there is no further change of current, therefore no further induction through the transformer and no further feedback. Grid voltage then commences to drop toward zero.

The negative-going grid voltage decreases the plate current. This decrease is a change which again causes feedback, but now the feedback polarity is such as to drive the grid more and more negative which, in turn, causes further decrease of plate current until this current drops to zero. Now the charge which was placed on grid capacitor  $C_g$  forces the grid voltage to a negative value far below cutoff, at  $a$  in Fig. 96-4. The capacitor discharges slowly through the hold control resistance, allowing the grid to become less and less negative until reaching cutoff voltage or until a sync pulse brings the grid voltage to or above cutoff. When this happens there is a new pulse of plate current, and the entire action repeats.

**Discharge Tube.** — In some receivers the charge and discharge of the sawtooth capacitor are controlled by a discharge tube which is made alternately conductive and non-conductive by a blocking oscillator. Typical connections are shown by Fig. 96-5. The oscillation and discharge functions usually are performed by the two sections of a twin triode tube.

Oscillator connections are essentially the same as in Fig. 96-3, except that the oscillator plate is connected through the transformer plate winding to  $B+$  voltage for this section of the tube rather than to the sawtooth capacitor and the output. The two cathodes are connected together and to ground. The two grids are connected together. Then changes of grid voltage on the discharge section must be exactly the same, and must occur at the same times, as on the oscillator.

The plate of the discharge section connects to its own B+ voltage through the size control, also to the sawtooth capacitor and the output, just as does the plate of the oscillator in Fig. 96-3.

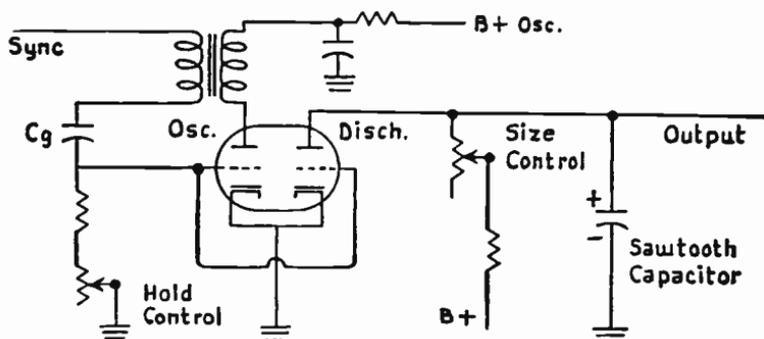


Fig. 96-5.—Discharge tube used with a blocking oscillator.

While the oscillator grid voltage is so negative as to prevent conduction, there likewise can be no conduction in the discharge section, and there is charging of the sawtooth capacitor from the B-supply. When the oscillator grid voltage becomes positive the grid of the discharge section is made positive, and the sawtooth capacitor discharges through the discharge section.

With a discharge tube in the circuit the B+ voltage applied to this tube and to the sawtooth capacitor is made higher than B+ voltage to the oscillator, thus allowing higher charge voltage on the sawtooth capacitor and a stronger sawtooth signal voltage for following deflection circuits. The relatively low B+ voltage on the oscillator allows smaller current at saturation, a quicker change of plate current, and faster action.

**Multivibrator Oscillator.** — Multivibrator sweep oscillators employ two triodes or two sections of a twin triode, as in Fig. 96-6, so connected that changes of plate voltage or cathode voltage from each section are applied to the grid of the other section to make the sections alternately and oppositely conductive and non-conductive. The two sections are coupled through the common cathode resistor  $R_k$ . Section B is shunted across the sawtooth capacitor to control charge and discharge

of that capacitor as this section is made alternately non-conductive and briefly conductive.

The action may be presumed to begin with electron flows as shown by arrows, and as occur when B+ voltage is applied to the circuit. One electron path is through cathode resistor  $R_k$ , the hold control resistor, capacitor  $C_g$ , and resistor  $R_o$  to

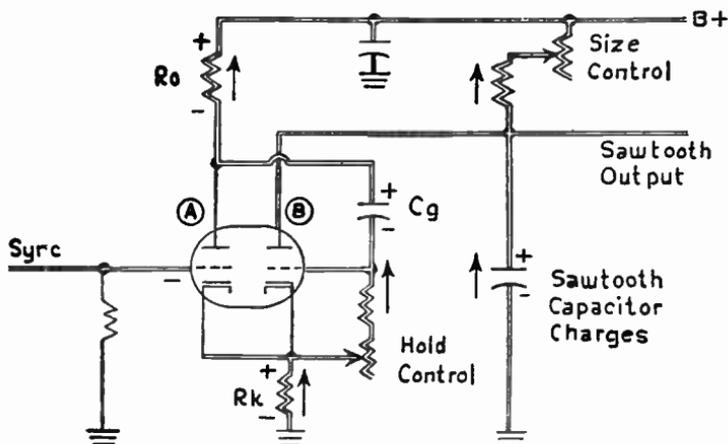


Fig. 96-6.—A cathode-coupled multivibrator sweep oscillator.

B+. Capacitor  $C_g$  is charged in the polarity marked. Capacitor voltage in relation to ground makes the grid of section  $B$  so negative with reference to its cathode that there is plate current cutoff and the section remains non-conductive.

While section  $B$  is non-conductive there is electron flow from ground (B-minus) through the sawtooth capacitor and through the size control resistors to B+. This flow charges the sawtooth capacitor in the marked polarity.

Electron flow toward the cathodes through resistor  $R_k$  means that the polarity of  $R_k$  is such as to make the grid of section  $A$  more negative. At the same time the direction of electron flow through resistor  $R_o$  indicates that the plate end of this resistor is more negative or less positive than the B+ end, an effect which reduces the plate voltage on section  $A$ . These two actions combine to cut off plate current in section  $A$  of the tube.

As capacitor  $C_g$  nears full charge there is lessening of current through resistors  $R_k$  and  $R_o$ , which makes the bias on section  $A$  less negative while increasing the plate voltage, and this section becomes conductive. Capacitor  $C_g$  discharges through section  $A$  as at the left in Fig. 96-7. This discharge through the hold control resistors is in a direction with which the grid of section  $B$  is made less negative with reference to the cathode, but still there is plate current cutoff in this section. As the discharge ceases, the grid voltage of section  $B$  becomes zero and this section is made conductive. Now there may be a discharge of the sawtooth capacitor through section  $B$  of the tube, as at the right in Fig. 96-7. Discharge current through cathode resistor  $R_k$  makes the grid of section  $A$  so negative as to maintain plate current cutoff in that section. As the end of discharge from the sawtooth capacitor we return to the starting condition of Fig. 96-6, and the events repeat as before.

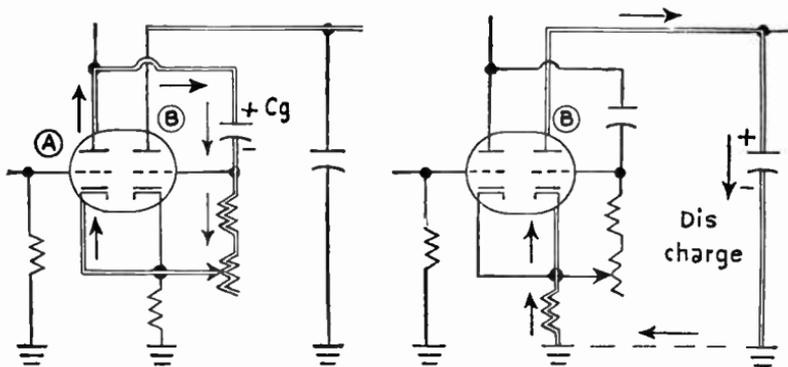


Fig. 96-7.—Actions in grid and plate circuits of the discharge section of the multivibrator.

Operating frequency of the oscillator is determined in the absence of sync pulses by the charge and discharge time of capacitor  $C_g$ , whose potential fixes the time at which section  $B$  of the tube is made conductive to allow discharge of the sawtooth capacitor. Discharge time for capacitor  $C_g$  is varied by adjustment of the hold control resistance.

Sync pulse voltages of negative polarity come to the grid of section *A* while this section is conductive, as at the left in Fig. 96-7. The pulse voltages are amplified and inverted by this section, and are applied through capacitor  $C_g$  to the grid of section *B*. The amplified pulses, now positive, make the grid of section *B* positive. Thus this section becomes conductive at the instant of each sync pulse voltage, and allows discharge of the sawtooth capacitor. The condition represented at the right in Fig. 96-7 lasts during only the very brief period of discharge for the sawtooth capacitor.

Fig. 96-8 shows the circuit of a multivibrator sweep oscillator with which coupling between the tube sections is not through a common cathode resistor but is through capacitors  $C_c$  and  $C_g$  connected between the plate of each section and the grid of the other section. Section *B* acts as an electronic switch across sawtooth capacitor  $C_s$  to control charge and discharge of this capacitor as this section is made alternately

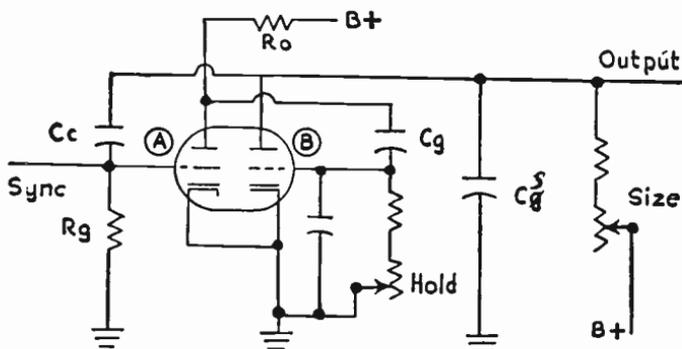


Fig. 96-8.—A capacitor-coupled multivibrator sweep oscillator.

non-conductive and conductive by charge and discharge of grid capacitor  $C_g$ . The sawtooth capacitor is charged through the size control resistance.

Increase of plate current in section *A* lessens its plate voltage because of more voltage drop in resistor  $R_o$ . The less positive plate voltage, applied through capacitor  $C_g$  to the grid of section *B*, makes that grid less positive or more nega-

tive. This change of grid voltage decreases plate current in section *B* and there is an increase of plate voltage on this section because of less drop in the size control resistance. The increased plate voltage or more positive plate voltage is applied through capacitor *C<sub>c</sub>* to the grid of section *A*, making that grid still more positive and causing further increase of plate current and drop of plate voltage in that section until plate current reaches saturation.

Capacitor *C<sub>g</sub>* now is charged in a polarity and to an extent making the grid of section *B* negative beyond the value for plate current cutoff. This condition continues until capacitor *C<sub>g</sub>* discharges through the hold control resistance sufficiently to allow resumption of plate current in section *B*. Then increase of plate current in section *B* drops the plate voltage on this section, and the effect is carried through capacitor *C<sub>c</sub>* to make the grid of section *A* become negative. Currents and voltages now are changing in polarities opposite to those first assumed. This action continues until there is plate current cutoff in section *A* while capacitor *C<sub>c</sub>* discharges through grid resistance *R<sub>g</sub>*. At the completion of this discharge there will be increase of plate current in section *A*, and the whole process repeats.

Sync pulses of negative polarity are applied to the grid of section *A*, are amplified, inverted, and applied to the grid of section *B* to make *B* conductive and allow discharge of the sawtooth capacitor at the instant of the pulse voltages. The small capacitor from the grid of section *B* to ground bypasses high-frequency interference voltages which otherwise might act like sync pulse voltages to trigger the tube.

The multivibrator of Fig. 96-8 may be called a capacitor-coupled type. It is used not only for sweep oscillators but also for generating various kinds of pulses and peaked waveforms for service instruments. The circuit shown by Figs. 96-6 and 96-7 usually is called a cathode-coupled multivibrator, but may be called a Potter multivibrator. It is the most common multivibrator for sweep oscillators, but is used also in many service instruments.

**OSCILLATOR, TELEVISION SWEEP, AFC FOR.** — Automatic frequency control (afc) is applied to all horizontal sweep oscillators and sometimes to vertical sweep oscillators. The purpose is to keep oscillator frequency properly synchronized or timed with frequency of sync pulses of received signals. Especially in horizontal sweep systems it is possible for various interference and noise pulses to act similarly to sync pulses in “triggering” an oscillator, but at wrong instants, unless an afc system is employed.

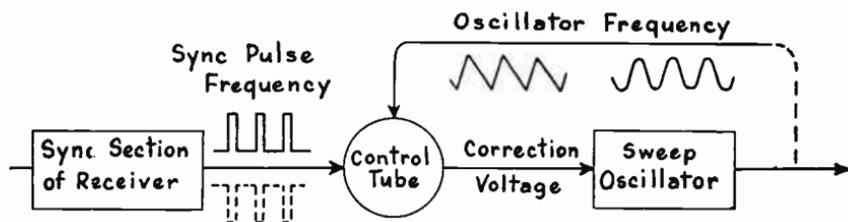


Fig. 97-1.—The principle of automatic frequency control for a sweep oscillator.

The basic principle of automatic frequency control is illustrated by Fig. 97-1. To the control tube are applied sync pulses coming through the sync section of the receiver from received signals. To the control tube is applied also a voltage consisting of sawtooth waves, sine waves, or some generally similar waveform derived from some circuit on the output side of the sweep oscillator. Frequency of this latter voltage is, of course, the actual operating frequency of the sweep oscillator. Unless this operating frequency is the same as that of received sync pulses, the control tube and associated circuits produce a correction voltage. This correction voltage is applied to frequency-determining circuits of the sweep oscillator in such manner that oscillator frequency is made equal to sync pulse frequency.

A principal object of automatic frequency control is to produce a correction voltage proportional to the constant frequency of sync pulses and to the uniform intervals of time between these pulses. If this is done, momentary appearance of interference or electrical noise pulses between regularly

spaced sync pulses will not appreciably affect the correction voltage. That is, interference pulses at instants between regular sync pulses cannot alter a correction voltage which depends for its value on regularity of spacing between sync pulses of received signals.

Afc systems are of three principal classes: 1. Sync pulse voltages are added in the control system to sawtooth waves coming back from the oscillator output. 2. Sync pulses are added to sine waves derived from oscillator output. 3. Sawtooth waves obtained from the oscillator output are utilized to vary the time duration or width of pulses from received signals.

Required correction voltages depend on the type of sweep oscillator. Frequency of a multivibrator oscillator is increased when correction voltage makes the frequency-control grid more negative, and frequency is lowered by making this grid less negative. With a blocking oscillator, frequency is increased by making its grid less negative, and is lowered when correction voltage makes the oscillator grid more negative. In feedback oscillators, tuned by inductance-capacitance circuits, frequency is increased by reducing either inductance or capacitance, and is lowered by using more inductance or capacitance.

Some examples of afc systems will be described. Circuit variations which appear from time to time are aimed at making synchronization more certain, at simplifying the control circuits, or both.

**Phase Detector and Multivibrator.** — Fig. 97-2 is a circuit diagram for an afc system with which correction voltage developed by a phase detector changes the grid voltage on a multivibrator sweep oscillator when oscillator frequency tends to vary from sync pulse frequency. Positive sync pulses are applied to one plate and negative sync pulses to one cathode of the phase detector, from plate and cathode of an inverter which is part of the sync section in the receiver. To the other plate and cathode of the phase detector, tied together, is applied a sawtooth voltage which originates as pulses from

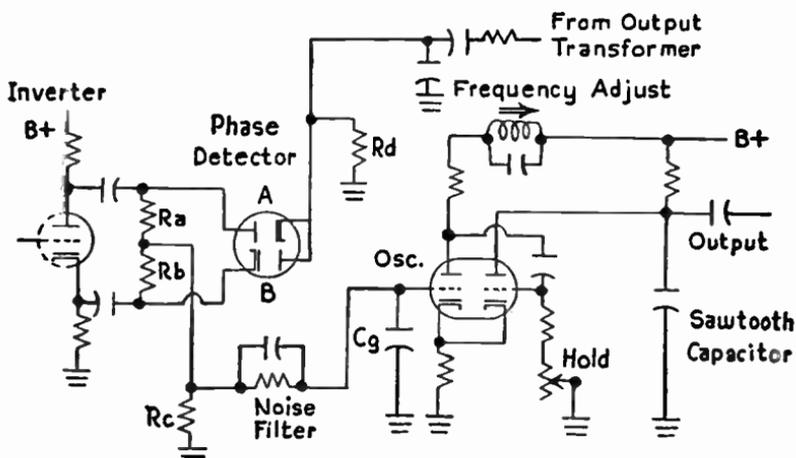


Fig. 97-2.—Phase detector for direct frequency control of a multivibrator oscillator.

the sweep output transformer and is changed to sawtooth form by a resistor and capacitors on the line going to the phase detector.

Pulses and sawtooth voltages on the phase detector are shown by Fig. 97-3. It must be realized to begin with that neither diode can conduct unless its plate and cathode are relatively positive and negative, this being true of any tube

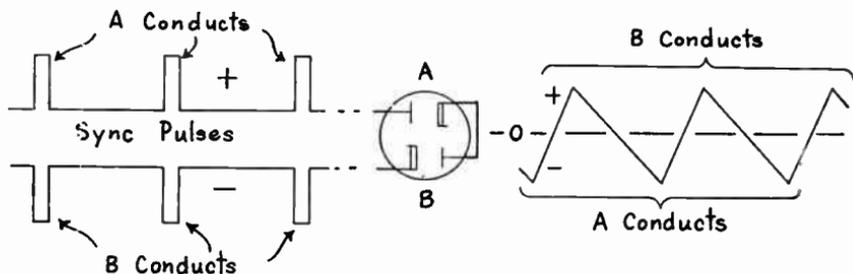


Fig. 97-3.—Sync pulses and a sawtooth voltage as applied to opposite sides of a phase detector.

containing cathode and plate. Accordingly, the diodes can conduct only during sync pulses, because only then are there

positive pulses on the plate of diode *A* and negative pulses on the cathode of diode *B*.

Consider next the effects of sawtooth voltage which is applied simultaneously to the cathode of diode *A* and to the plate of diode *B*. Diode *A* can conduct only while the sawtooth is on a negative half-cycle, for only then is the cathode of *A* negative. Diode *B* can conduct only while the sawtooth is positive, for only then is the plate of *B* positive. Under the combined effects of sync pulses and sawtooth voltage neither diode can conduct except during sync pulses, and then only if sawtooth voltage is of certain polarity.

In Fig. 97-4 the pulse periods are marked on sawtooth voltages as the sawtooths affect diode *A* and diode *B*. Half-cycles which make each diode conductive are in full lines, and half-cycles which make each diode non-conductive are in broken lines.

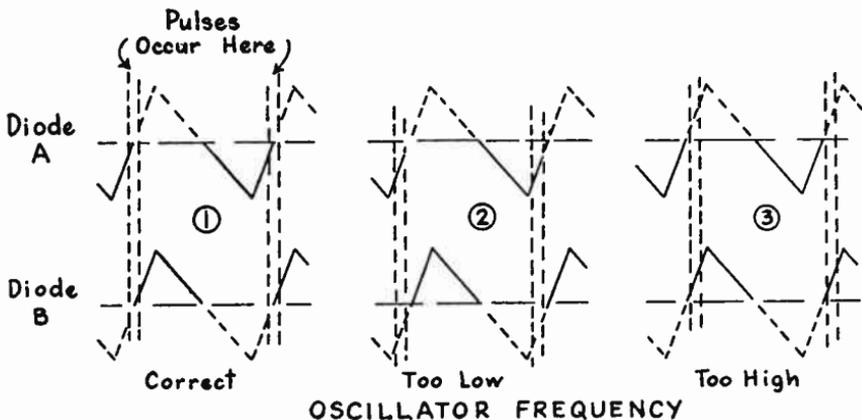


Fig. 97-4.—Conductions in the phase detector diodes.

with non-conductive half-cycles in broken lines. At 1 the oscillator frequency is correct in relation to received sync pulses. Conduction occurs only while the sawtooth voltage is at and near zero. There is but little conduction in either diode, and this little is equal in the two diodes.

At 2 the oscillator frequency has become too low, and its cycles occur slightly later in relation to sync pulses. Then there is conduction in diode *A* because the sawtooth is on a half-cycle making *A* conductive. But there is no conduction in diode *B* because the same sawtooth half-cycle makes *B* non-conductive. At 3 the oscillator frequency has increased, and cycles of sawtooth voltage occur slightly earlier in relation to sync pulses. Now there is no conduction in diode *A*, but there is conduction in diode *B*.

Diode currents are rectified direct current which can flow only in conductive paths, not through any of the capacitors. Direct current for diode *A* follows the path (on Fig. 97-2) upward through resistor *Rd*, from cathode to plate in the diode, and downward in resistors *Ra* and *Rc*. For diode *B* the direct current flow is upward through resistor *Rc*, down through *Rb*, from cathode to plate in the diode, and downward in resistor *Rd*. Resistor *Rc* is between grid and cathode of the oscillator, by way of the noise filter and ground connections. Therefore, any voltage developed across *Rc* becomes part of the oscillator grid voltage.

Diode currents flow oppositely and may partially balance in resistor *Rc*. When there is greater conduction in diode *A*, the top of *Rc* and the oscillator grid become more negative, to increase oscillator frequency. Fig. 97-4 shows that such conduction occurs when oscillator frequency commences to become lower—so there is correction of frequency. Greater conduction in diode *B*, resulting from increase of oscillator frequency, causes the top of resistor *Rc* and the oscillator grid to go more positive, or actually less negative. This lowers the oscillator frequency to provide correction.

The only service adjustment for the afc system of Fig. 97-2 is marked *Frequency Adjust* on the diagram. In the inductor and paralleled capacitor are produced a voltage of approximate sine-wave form, at the frequency of oscillation. This sine-wave form is added to voltage at the grid of the second section in the multivibrator. When this added voltage is suitably phased it raises effective amplitude of sync pulses

which should trigger the oscillator, while reducing effective amplitude of all other pulses, both sync and noise. This stabilizes oscillator frequency in relation to sync pulses. Adjustment consists of setting the inductor core so that, with the hold control at its mid-position, pictures lock into synchronization. The adjustment may be varied slightly to maintain synchronization over the widest possible range of hold control settings.

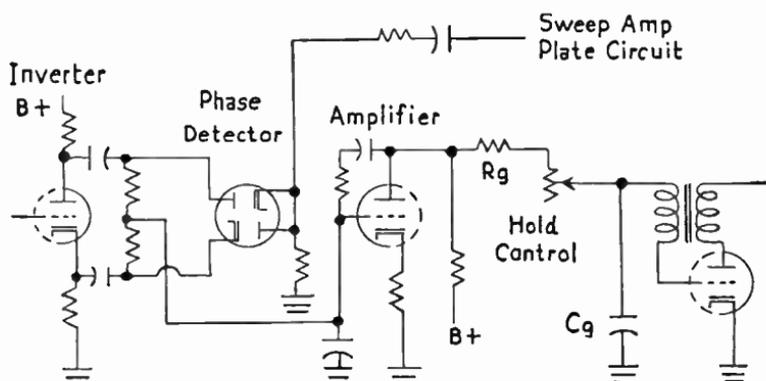


Fig. 97-5.—Sweep afc system employing an amplifier between a phase detector and a blocking oscillator.

There are many modifications of the phase detector method of automatic frequency control, of which Fig. 97-5 is one example. Here the correction voltage is strengthened and its polarity is inverted in an amplifier tube between the control tube and a blocking oscillator. In this case inversion of correction voltage is necessary because at the grid of a blocking oscillator the effects of positive and negative correction voltages are just the reverse of their effects at the synchronizing grid of a multivibrator.

An inverting amplifier may be used also between the phase detector and the grid of the second section of a multivibrator, where polarity of correction voltage must be opposite to that at the first grid in order to have the same effects on oscillator frequency.

**Variable Pulse Width For Afc.** — Fig. 97-6 is a circuit diagram of an afc system with which grid voltage and frequency of a blocking oscillator are regulated by varying the time

duration or width of sync pulses. The control tube is a triode in whose cathode circuit is resistor  $R_g$ . This resistor is also in the grid-to-ground circuit of the oscillator. Voltage across  $R_g$  is altered by changes of conduction in the control tube, which are caused, in turn, by differences between oscillator output frequency and frequency of received sync pulses. The control voltage becomes part of the oscillator grid voltage, and increases or decreases oscillator frequency as may be required for synchronization.

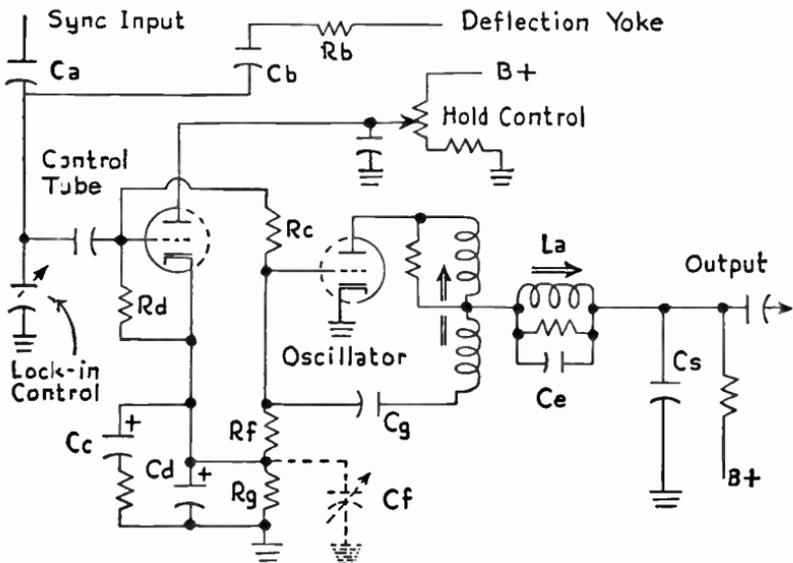


Fig. 97-6.—The afc system varies the width of sync pulses to alter oscillator bias and control frequency.

Capacitor  $C_c$  and a series resistor, also capacitor  $C_d$ , are in parallel with resistor  $R_g$ . These capacitors are charged by pulses of conduction current through the control tube, and discharge rather slowly. Holding the charges for a fraction of a second prevents control voltage across resistor  $R_g$  from being affected by momentary pulses of interference or noise.

To the grid of the control tube are applied positive sync pulses coming from the sync section of the receiver through capacitor  $C_a$  and the capacitor in series with the grid. To the control tube grid is applied also a sawtooth voltage derived from the deflection yoke circuit. Sync pulses and the

sawtooth voltage combine at the control grid as in Fig. 97-7.

At the top of Fig. 97-7 are shown conditions when the sweep is in time, which means that frequency of sawtooth voltage from the oscillator output is correctly in synchronization with received sync pulses. About half of each pulse rides on the sawtooth voltage peak and the remainder is down in the trough. The portion of the pulse on top of the peak

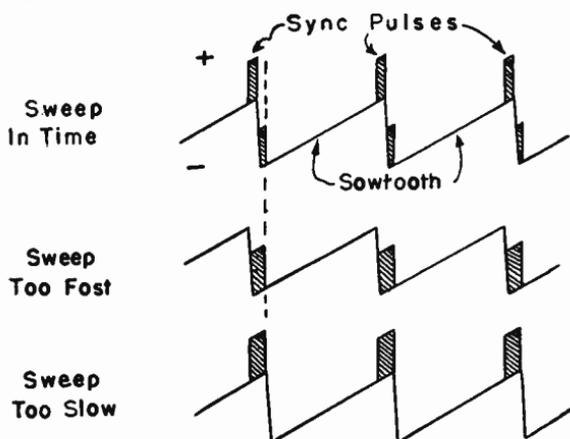


Fig. 97-7.—How effective width or time duration of sync pulses is varied by combining them with a sawtooth voltage.

makes the combined positive voltages great enough to overcome part of the negative bias at the control tube grid and make this tube conductive.

Should the sweep be too fast, meaning that oscillator frequency is too high, a greater part or possibly all of each sync pulse drops into the trough, as shown by the center diagram. This reduces or completely prevents conduction in the control tube. Should oscillator frequency become too low, as shown by the bottom diagram, a greater part or possibly all of each sync pulse rides on top of the sawtooth peak. This increases the time of conduction in the control tube.

Each period of conduction in the control tube charges capacitors  $C_c$  and  $C_d$ , which are in the cathode circuit, making the upper plates of these capacitors positive. The greater the total duration of conduction periods in the control tube the greater is the charge and the voltage placed on the capacitors.

Since these capacitors are connected across resistor  $R_g$ , positive voltage of the capacitors counteracts more or less of that part of the oscillator negative grid bias which is across resistor  $R_g$ .

When oscillator frequency becomes too high there is less charging of capacitors  $C_c$  and  $C_d$ , and oscillator grid bias is allowed to become more negative. This will lower the frequency of a blocking oscillator. Should oscillator frequency become too low there will be increase of capacitor charge and voltage. Then the oscillator bias is made less negative and oscillator frequency is increased.

Feedback from plate to grid of the oscillator is through the auto-transformer shown at the right of the tube in Fig. 97-6. An adjustable core in this transformer allows varying the free running frequency of the oscillator. Output to the sweep amplifier which follows the oscillator is from the transformer tap. On the output line is adjustable inductor  $L_a$ , sometimes called a stabilizing coil or ringing coil. In the inductor and paralleled capacitor is produced a sine-wave voltage which adds itself to the voltage waveform at the oscillator grid to increase effective amplitude of sync pulses which should trigger the oscillator, while reducing effective amplitude of other pulses, including those due to interference and noise.

In some receivers there is an adjustable capacitor  $C_f$  in parallel with capacitors  $C_c$  and  $C_d$ . Adjustment of  $C_f$  varies total capacitance which is charged from the control tube, thus varying the charge voltage and its effect on oscillator bias. The adjustable lock-in capacitor on the grid circuit of the control tube acts as a variable bypass to allow greater or less amplitude of combined control voltages to act on the grid of this tube. Instead of an adjustable capacitor a fixed unit may be used in this position. All adjustments mentioned so far are service adjustments. The one front panel adjustment is the hold control in series with the plate of the control tube. This adjustment varies plate voltage on the control tube and thus alters the amount of charging current for capacitors  $C_c$  and  $C_d$ .



in which a triode afc tube is used as a phase detector. Positive sync pulses (1) are taken from the inverter plate and applied to the afc grid. To this grid is applied also a sawtooth voltage (2) obtained from the plate of the second section of the multivibrator oscillator. The pulses and sawtooth voltage combine at the afc grid about as shown by the waveform at 3. Positive peaks of this composite waveform cause grid current to flow in the path of broken-line arrows, through resistors  $R_a$  and  $R_b$ .

Negative sync pulses (4) from the inverter cathode are applied to the cathode of the afc tube. To the plate of this tube is applied the sawtooth voltage 5, obtained from the deflection system. Voltages having the waveforms at 5 and 4 combine at the cathode of the afc tube to produce the plate-current waveform at 6. Plate current flows, as shown by full-line arrows, from the plate through resistor  $R_d$  to ground and from ground through resistor  $R_c$  back to the cathode.

The d-c return circuit for the first grid of the oscillator extends through the noise filter resistor and to the top of resistor  $R_b$ , thence through  $R_b$  and  $R_c$  to ground, and back to the oscillator cathode. Voltage due to afc grid current in  $R_b$  tends to make the oscillator grid negative, because the negative end of  $R_b$  is toward the oscillator grid. Voltage due to afc plate current in resistor  $R_c$  tends to make the oscillator grid positive, because the positive end of  $R_c$  is toward the oscillator grid.

Voltage on the oscillator grid will be the difference between opposing voltages across  $R_b$  and  $R_c$ , and will be of polarity depending on which of these resistors carries greater current and has greater voltage drop. When sawtooth (oscillator) voltage and received sync pulses are in correct synchronization, grid current and plate current in the afc tube are such as to produce equal and opposite voltages across resistors  $R_b$  and  $R_c$ . These voltages cancel, and no correction voltage is applied to the oscillator grid.

Should the oscillator tend to run slow, the sawtooth voltages shift in relation to sync pulses to make afc grid current greater

than afc plate current. Then the oscillator grid is made more negative, to increase oscillator frequency. Should the oscillator tend to run fast, the phase shift of oscillator output and of sawtooth voltages is such as to increase afc plate current more than afc grid current. This makes the oscillator grid less negative and decreases oscillator frequency.

The only service adjustment is a movable slug in the inductor marked *Horz Freq.* This inductor and the paralleled capacitor have the same function as similar parts on the multivibrator plate connection in Fig. 97-2, and adjustment

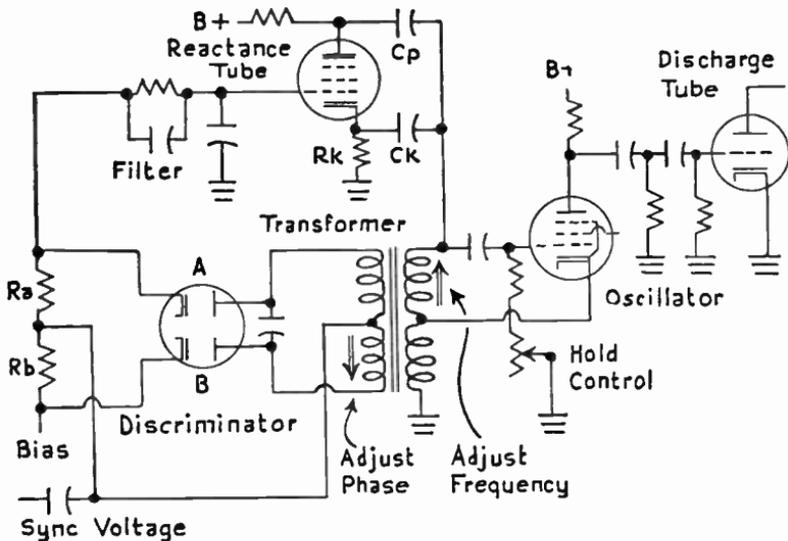


Fig. 97-9.—The discriminator furnishes correction voltage to the reactance tube grid, and reactance tube plate current varies the oscillator frequency

is carried out in the same manner. That is, the slug is adjusted so that pictures remain synchronized with the hold control turned as far as possible in either direction.

**Reactance Tube And Discriminator For Afc.** — Fig. 97-9 is a circuit diagram for an afc system employing a reactance tube and a frequency or phase discriminator for synchronizing a Hartley oscillator in a horizontal sweep section. Action of a reactance tube is explained elsewhere in this book. Capacitor *Ck* and resistor *Rk* are the units which apply voltage at oscillator frequency in the tank circuit to the reactance

tube cathode. Capacitor  $C_p$  is the unit which allows reactance tube plate current to flow in the oscillator tank, while excluding  $B+$  voltage and direct current. The reactance tube grid circuit extends through a noise filter and through discriminator load resistors  $R_a$  and  $R_b$  to a source of negative grid bias for the reactance tube. Correction voltage is produced in these resistors.

The oscillator tank inductor forms the primary of a transformer whose center-tapped secondary connects to the discriminator plates. A sine-wave voltage induced in the secondary makes the plate of diode  $A$  positive with reference to the center tap while the plate of diode  $B$  is made negative, and vice versa. Each diode conducts while its plate is positive.

Conduction current of diode  $A$  flows upward through resistor  $R_a$ , and of diode  $B$  flows downward through resistor  $R_b$ . When oscillator frequency is the same as resonant frequency of the transformer secondary, as tuned by the capacitor between diode plates, there will be equal conduction currents and equal but opposite currents in resistors  $R_a$  and  $R_b$  so far as these currents are influenced only by the sine-wave voltage. Voltages accompanying the currents in resistors  $R_a$  and  $R_b$  cancel, and no correction voltage is applied to the reactance tube grid.

If oscillator frequency is not the same as resonant frequency of the transformer secondary, diode conductions and voltages across  $R_a$  and  $R_b$  will not be equal. Net voltages or difference voltages across the two resistors will be positive or negative, and will alter grid voltage and plate current amplitude of the reactance tube. This would make oscillator frequency that of resonance in the transformer secondary.

Both positive and negative sync pulses come to the transformer secondary center tap and to the junction between resistors  $R_a$  and  $R_b$ , but only the positive pulses are rectified to cause conduction currents in the diodes. Because the sync pulses are applied at center points they make the plates of both diodes positive at the same time. Resulting pulses of current are equal and opposite in resistors  $R_a$  and  $R_b$ , and

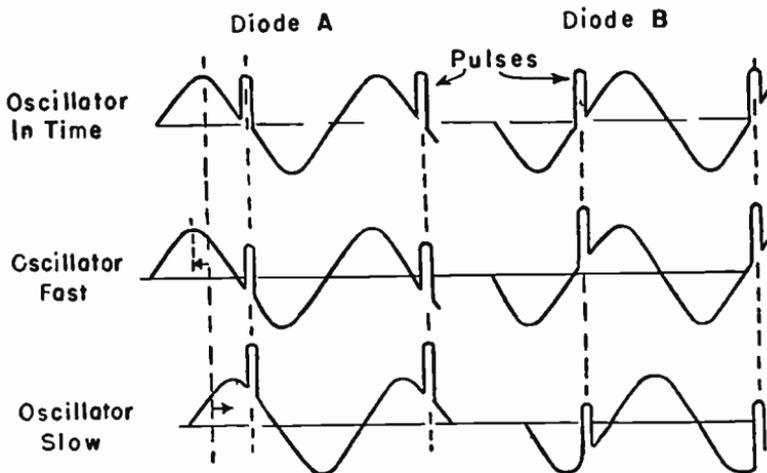


Fig. 97-10.—How sync pulses and a sine wave voltage from the oscillator circuit combine at the discriminator diodes.

accompanying voltages would cancel when due only to action of sync pulses.

Actually the sine waves and sync pulses combine at the two diodes as shown by Fig. 97-10. If the oscillator is in time with or is correctly synchronized with received signals, pulses in both diodes will occur while sine waves go through zero, as shown by the top diagram. These pulses cause equal but opposite voltages which cancel in resistors  $R_a$  and  $R_b$  of Fig. 97-9. Then no correction voltage is applied in the grid circuit of the reactance tube.

As oscillator frequency tends to become fast, or higher than sync pulse frequency, the sine waves will occur earlier. Then sync pulses will appear while diode  $A$  is non-conductive, but  $B$  is conductive. More current will flow in diode  $B$  and in resistor  $R_b$ , and net voltage across the load resistors will become negative toward the grid of the reactance tube. This reduces oscillator frequency, as is required to restore synchronization.

Should oscillator frequency become slow, conditions would be as at the bottom of Fig. 97-10. Sine waves would occur later, and during pulse periods diode  $A$  would be conductive, with diode  $B$  non-conductive. Then more current would flow

in resistor  $R_a$  to make net voltage across the load resistors positive toward the reactance tube grid. This would increase oscillator frequency, as required.

The purpose of the filter in Fig. 97-9 is primarily to furnish a slight time delay for instantaneous changes of correction voltage developed in the discriminator load resistors. Then correction voltage at the reactance tube grid is of an average value resulting from several sync pulses, and the control system does not respond to single intermittent pulses of interference or noise. Otherwise these undesired pulses might cause pictures to jump sideways.

Service adjustments for this afc system are in the discriminator transformer. A movable slug in the primary alters resonant frequency of the oscillator tank circuit. This adjustment should be set to hold pictures synced on all channels, and for prompt pull-in when changing channels. The adjustment may be varied slightly so that pictures remain synced with the hold control turned as far as possible in each direction.

Another slug in the transformer secondary alters resonant frequency of the secondary circuit. This adjustment should be made with the contrast control retarded, brightness turned up, and the hold control at its mid-position. Should a vertical dark bar (horizontal blanking bar) appear in pictures, the phasing adjustment is altered to move this bar just outside the picture area.

**OSCILLATORS, ULTRA-HIGH FREQUENCY.** — Operation of oscillators at frequencies higher than about 300 mc introduces many difficulties which are unimportant or absent at lower frequencies. Among the principal difficulties are excessive interelectrode capacitance effects in the oscillator tubes, also the inductance and capacitance of internal leads for the tube elements. Distributed and stray capacitances throughout the circuits have more effect on tuning as frequency increases, and it becomes difficult to construct circuits of high inductance-capacitance ratio or of high Q-factors. Excess capacitance tends to improve frequency stability, but the lack of inductance makes the circuits less efficient in use of applied power.

Electron transit time from the cathode to other elements in a tube is the factor which limits the operating frequency in many constructions. When the distance between elements is any considerable fraction of a wavelength corresponding to operating frequency an electron may progress only part way from one element to another before the electric fields reverse. Even when electrons have time to complete their travel the variations of field strength during a cycle have an effect equivalent to reduction of resistance between cathode and negative grid, with corresponding decrease of Q-factor and increase of power taken from the grid circuit.

A number of miniature tubes are especially designed for use as r-f oscillators in ultra-high frequency tuners and converters. The construction of these tubes is such as to overcome difficulties inherent in operation at high frequencies, and they give dependable performance throughout the uhf carrier range and to 1,000 megacycles.

**Oscillator Drift.** — Gradual change or drift of oscillator frequency until circuit parts reach stable operating temperatures causes the same difficulties in uhf reception as at lower frequencies. Since oscillator frequency combines with carrier frequency to produce the intermediate beat frequency, any great change of oscillator frequency may so change the intermediate as to throw it outside the range within which the i-f amplifier has correct frequency response. The receiver or

converter, or a fine tuning control, have to be retuned several times during the first few minutes of operation.

The most common cause for oscillator drift is change of temperature in parts of the resonant circuit during the warm-up period, usually with a decrease of resonant frequency as temperature rises. This effect may be compensated for with temperature compensating ceramic capacitors, as is done with vhf oscillators and others. Another common cause is change of plate voltage applied to the oscillator tube. This varies the oscillator frequency and also the output which is the injection voltage to the mixer. Variations of oscillator plate voltage also change the relative strengths of harmonic frequencies, which sometimes are used instead of the oscillator fundamental for formation of intermediate beat frequencies. Plate voltage may be held nearly constant by any of various types of voltage regulation in the d-c power supply system, but this seldom is done in receivers.

Oscillator frequency drift in well designed uhf tuners is not enough to cause much difficulty when the sound system of the receiver is of the intercarrier type. With this system the modulated sound signal, at 4.5 mc, results from the difference between video and sound carriers and the difference between video and sound intermediate frequencies, and it is not altered by variations of oscillator frequency. When the receiver sound system is of the dual or split type an oscillator frequency drift of  $1/20$  of one per cent at the middle of the uhf band would throw the sound intermediate completely outside the usual pass band of the sound i-f amplifier, and much smaller drifts would make the sound decidedly unsatisfactory without retuning of the channel selector or a fine tuning control.

**Using Harmonics from Vhf Oscillator.** — The separate uhf oscillator may be dispensed with by using harmonic frequencies from the vhf oscillator during uhf reception. This method is chiefly useful in the design of channel strips for turret tuners, wherein all principal frequency conversion elements for a single uhf channel may be mounted on one set of strips.

With no uhf oscillator needed, and with the uhf mixer a crystal diode type, there is ample space at a single turret position for all circuit elements required for reception of any one uhf channel.

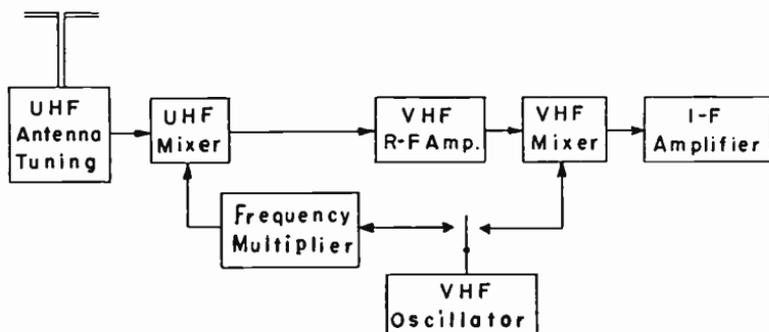


Fig. 138-17.—To the uhf mixer is applied a harmonic of the fundamental frequency at which the vhf oscillator operates.

The general scheme of connections is illustrated by Fig. 138-17. Uhf antenna tuning and crystal mixer circuits may be any of the usual types. The beat frequency in the output of the uhf mixer is the intermediate frequency regularly used in the vhf section of the receiver, practically always in the 40-mc range of intermediates. During uhf reception the r-f amplifier and also the r-f mixer tuning are made resonant at the intermediate frequency, and act as additional i-f amplifier tubes, with the vhf oscillator cut off from the vhf mixer. The r-f amplifier is a cascode type which provides the high gain and low noise factor required for an amplifier following a uhf crystal mixer. Cascode amplifiers are described in the section on *Amplifiers, Radio-frequency*.

During uhf reception the vhf oscillator is connected through a frequency multiplier circuit to the uhf mixer circuit, taking the place of a uhf oscillator tube otherwise employed. The vhf oscillator is resonated at a frequency of which the second, third, or fourth harmonic is the oscillation frequency required in the uhf mixer circuit for the channel received. In the system illustrated there is only one conversion between uhf carrier

and receiver intermediate frequencies, this occurring in the uhf mixer. Consequently, the oscillation frequency for this mixer is higher than the carrier frequencies of the received channel.

A frequency multiplier circuit is shown by heavy lines in Fig. 138-18. The principle is the same as in multiplier circuits using electronic tubes. Such tubes are operated with negative grid biases so strong as to cut off or nearly cut off plate current during negative alternations of grid voltage, and to leave only a series of current pulses for positive alternations. Such action is essentially that of a rectifier. Consequently, a crystal diode rectifier may be used. The crystal diode is biased to insure operation at the sharpest portion of the bend on its characteristic.

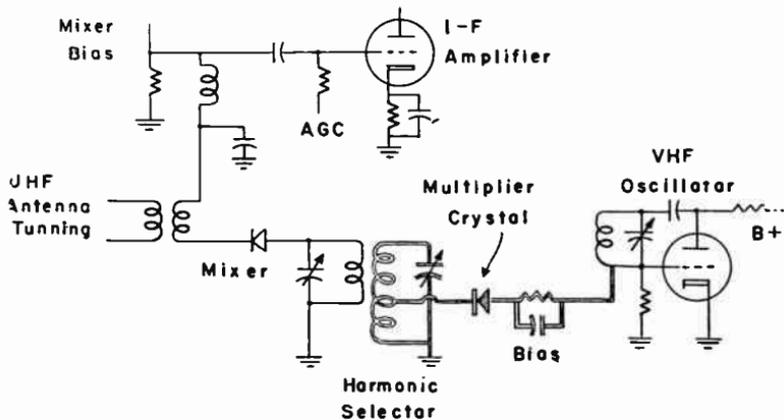


Fig. 138-18.—A frequency multiplier circuit used between the vhf oscillator and the uhf mixer.

In the output of the crystal there are many harmonics of the fundamental frequency at which the vhf oscillator is operating. These harmonics, along with the fundamental, go to the harmonic selector, which is a parallel resonant circuit tuned for the harmonic frequency to be applied to the uhf mixer. The selector circuit responds strongly at the tuned harmonic frequency, while the fundamental and other harmonics pass quite freely to ground.

The harmonic selector circuit may be coupled to the mixer circuit in any of various ways. In Fig. 138-18 there is inductive coupling, with the mixer side tuned to the desired harmonic frequency. Uhf carrier frequencies are coupled into the mixer circuit from the antenna tuning circuits. The resulting beat frequency is applied to the grid of the i-f amplifier, which is the r-f amplifier tube for vhf reception, here retuned to the intermediate frequency. The mixer crystal may be biased for operation on the most favorable point of its characteristic.

How this system works out in practice is illustrated by the accompanying tabulation. It is assumed that the vhf oscillator, during uhf reception, may be tuned for any fundamental frequencies between 172.33 mc and 232.75 mc, all of which are within the normal range for any vhf oscillator. Third harmonic frequencies could be used for reception in uhf channels 14 through 44, and fourth harmonics for channels 45 through 83. The table shows frequency relations for channels 14 and 83, at the bottom and top of the uhf band.

USING HARMONICS FROM VHF OSCILLATOR

	Channel 14		Channel 83	
	Video	Sound	Video	Sound
Vhf osc. fundamentals, mc	172.33	172.33	232.75	232.75
Harmonics — 3rd	517.00	517.00		
4th			931.00	931.00
Uhf carrier frequencies, mc	471.25	475.75	885.25	889.75
Intermediates (beats), mc	45.75	41.25	45.75	41.25

There are several tuners or converters having uhf oscillator tubes not in any way related to the vhf oscillators, and employing second or third harmonics of the oscillator frequency for injection into the mixer circuit. This permits use of tubes well suited for operation as oscillators in the vhf band instead of the types especially designed for ultra-high frequency operation.

**OSCILLOSCOPES.** — In the oscilloscope or oscillograph is a cathode-ray tube on whose screen are produced luminous traces showing waveforms of any voltage which varies with respect to time. The oscilloscope is used also for display of frequency responses of r-f, i-f, and audio amplifiers. Service oscilloscopes employ electrostatic deflection cathode-ray tubes, whose performance is explained under *Electrostatic Deflection Systems*.

The electron beam in the cathode-ray tube is deflected vertically by the voltage whose waveform or amplitude variations are observed. At the same time the beam is deflected horizontally by a separate deflection voltage which may be produced within the oscilloscope or else taken from various external sources.

Some certain fraction of a second is required for the electron beam to travel horizontally across the screen. During this time the beam is moved up and down by increases and decreases in amplitude of the observed voltage. When horizontal deflection or sweep time is equal to or is a simple fraction of the time for one cycle of observed voltage, the electron beam will trace an extended curve showing changes of amplitude during the period of one horizontal trace. With the horizontal sweep rate properly related to frequency of observed voltage, curves traced during all sweeps will be alike, will follow the same path across the screen of the cathode-ray tube, and will appear as a single continuing trace. The observed waveform may be seen to change during service adjustments in any circuits to which the vertical input of the oscilloscope is connected.

Fig. 99-1 shows principal parts of a service oscilloscope. Arrows indicate application of control voltages. Voltage for vertical deflection goes from vertical input terminals through a gain control and amplifier to vertical deflection plates. Horizontal deflection voltage is fed from the horizontal sweep oscillator or from horizontal input terminal through a sweep selector switch, a gain control, and an amplifier to horizontal deflection plates.

To the plates and other electrodes of the cathode-ray tube are connected vertical and horizontal centering controls, a focus control, and an intensity control that varies the brightness of the luminous trace on the tube face. These three controls are in the circuit of the high-voltage power supply system. A low-voltage power supply furnishes voltages and currents

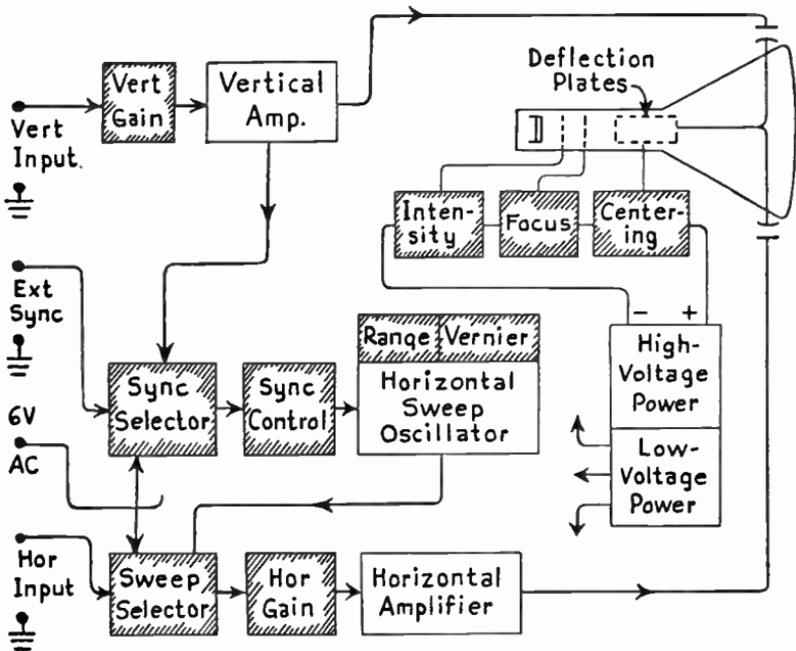


Fig. 99-1.—Principal parts used in service oscilloscopes.

for plates, screens, grid biases, and heaters in amplifier and oscillator tubes. Controls which are shaded in the diagram, and the functions of these controls, will be described in more detail in pages following.

Centering controls and focusing controls such as commonly employed with electrostatic deflection cathode-ray tubes are explained under *Electrostatic Deflection Systems*. There are, of course, numerous modifications of all such operating controls. The horizontal sweep oscillator may use a gaseous tube

as shown in the article on *Oscillators, Relaxation*, or it may use a multivibrator oscillator described under *Oscillators, Television Sweep*.

Cathode ray-tubes of the electrostatic deflection type are used in practically all oscilloscopes because this deflection method allows satisfactory operation throughout a wide range of deflection frequencies, whereas with magnetic deflection circuits the changes of inductive reactances with frequency would limit the useful frequency range.

Screens usually are of phosphor number 1, which produces a bright green trace. Phosphor number 4, with the white trace found in television picture tubes, sometimes is used. Where a principal object is photography of traces, the phosphor may be number 5 or number 11, whose color strongly affects photographic emulsions.

When working on the high-voltage circuits of the cathode-ray tube and high-voltage power supply it is necessary to observe the same precautions as when working with high-voltage circuits of television picture tubes.

Shortcomings found in some of the lower cost oscilloscopes include lack of uniform frequency response in the vertical amplifier, insufficient gain in this amplifier, and a non-linear horizontal sweep which may be due to faults of the internal sweep oscillator but more often to poor frequency response in the horizontal amplifier. Horizontal sync signals from television receivers may be much distorted if the oscilloscope vertical amplifier does not provide nearly uniform response at frequencies up to at least 200 or 250 kilocycles. Frequency response curves taken from receiver amplifier circuits will not indicate true relative gains when using an oscilloscope with internal sweep which is non-linear. This latter difficulty is avoided when using a synchronized sweep from a signal generator or sweep generator.

**Controls and Terminals.** — The arrangement of controls and terminal connections on the panels of service oscilloscopes is shown in a general way by Fig. 99-2. Names used on this

figure are the same as on Fig. 99-1. Exact positions of controls, and names applied to them, will vary with the make of instrument and between models of the same make.

At the left side of the cathode-ray tube is an *Intensity* control which changes the bias on the control grid of this tube to vary the brightness of the trace formed on the screen. This corresponds to the brightness control of a television receiver. In order to prevent damaging the screen material of the cathode-ray tube, brightness or intensity always should be kept as low as produces satisfactory visibility of the trace. Keeping strong external light away from the face of the tube allows using low intensity. Many oscilloscopes have a hood extending out around the tube face or have the face recessed in the housing to provide shielding against external light.

Should there be temporarily neither vertical nor horizontal deflection of the beam in the cathode-ray tube, the intensity must be kept so low as to make the spot of light on the screen barely visible or completely extinguished. A stationary bright spot will burn the screen material. The on-off switch for the oscilloscope may be combined with the intensity control, just as it is combined with the sound volume control in receivers.

On the right-hand side of the cathode-ray tube is the *Focus* control, which performs exactly the same function as a focusing control in a television receiver—keeping the trace line of minimum width and maximum clarity. Turning this control either way from a correct setting will widen the trace. The focus control will require readjustment when the intensity control is varied.

On the left side of the panel of Fig. 99-2 is a vertical *Centering* control, sometimes called a positioning control, which will move the entire trace up or down to center it between top and bottom of the tube face or to bring the trace to any desired position. On the right is a horizontal *Centering* or position control which will move the entire trace sideways to center it between right and left sides of the tube screen or to bring the trace wherever desired.

Down below the centering controls are vertical and horizontal *Gain* controls. These controls vary the gain of vertical and horizontal amplifiers which are part of the oscilloscope.

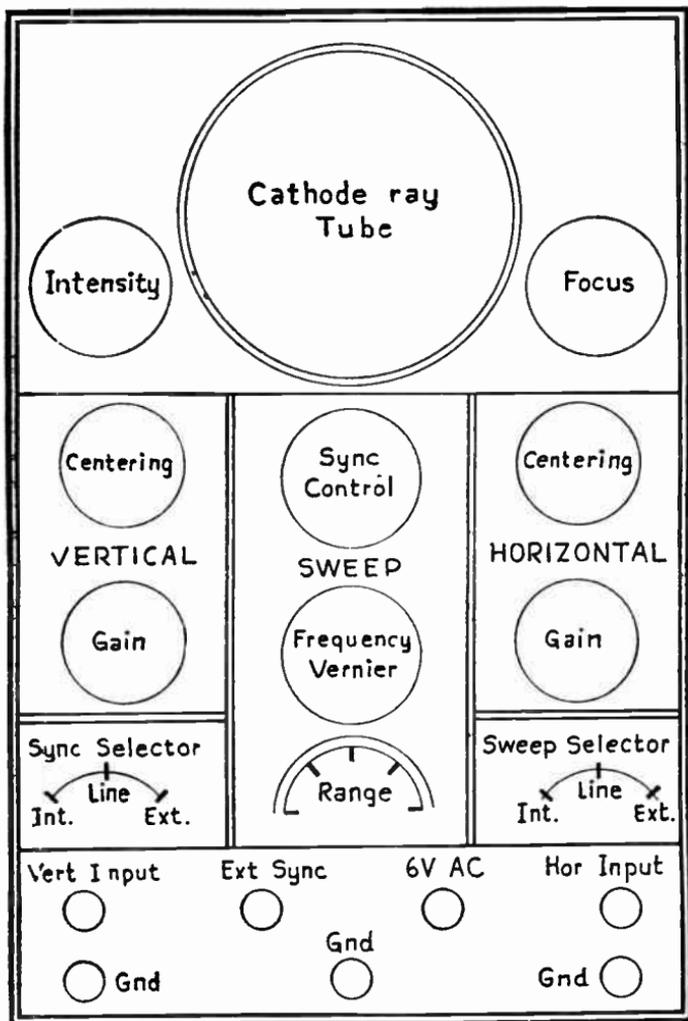


Fig. 99-2.—Controls and terminals on the panel of an oscilloscope.

Input voltage for the vertical amplifier is applied to the oscilloscope through the terminal marked *Vert Input* and a ground

terminal at the lower left-hand corner of the panel. To these terminals may be brought any voltage which is to cause vertical deflection of the beam or whose waveform is to be observed. Changing the gain of this amplifier will either extend or contract the trace in a vertical direction without altering the width.

Input to the horizontal amplifier of the oscilloscope may be from the horizontal sweep oscillator built into the instrument or from any external voltage applied between the panel terminal marked *Hor Input* and a ground terminal at the lower right-hand corner of the panel. Changing the gain of the horizontal amplifier makes the trace occupy more or less space from side to side, but does not alter the height.

When a voltage at power line frequency is to be applied to either the vertical or horizontal amplifiers this voltage may be obtained from the terminal marked *6V AC*. This terminal usually is connected inside the instrument to one of the low-voltage heater circuits. An external connection from *6V AC* to the *Vert Input* terminal allows applying power line frequency to the vertical amplifier, while a connection from *6V AC* to the *Hor Input* terminal allows applying line frequency to the horizontal amplifier.

The gain controls normally are used to provide traces of desired size from input voltages of various strengths. These controls serve the same purpose as size controls or height and width controls in television receivers. The gain controls may be used also to enlarge the trace for close examination of any one portion. This particular part of the trace then may be centered on the screen by manipulating the vertical and horizontal centering controls. Other parts of the enlarged trace then will be outside the screen area and will be invisible.

Underneath the horizontal gain control of Fig. 99-2 is a *Sweep Selector* switch that selects the source of voltage input to the horizontal sweep amplifier. When this switch is turned to the position marked *Int*, for internal, the amplifier input is connected to the output of the sweep oscillator which is within the oscilloscope. This switch position sometimes is

marked *Sweep*. With the switch set at the *Line* position the horizontal amplifier is fed voltage at power line frequency, as secured from one of the tube heater circuits inside the instrument. With the switch turned to the position marked *Ext*, for external, the input of the amplifier is connected to the terminal immediately below the switch, marked *Hor Input*. To this terminal and the ground terminal underneath may be connected any source of voltage to be used for horizontal deflection of the oscilloscope beam.

In the center of the panel of Fig. 99-2 are three controls which alter the operating frequency of the horizontal sweep oscillator built into the oscilloscope. The lowermost of these controls is a range switch for connecting sawtooth capacitors of different capacitances to the oscillator. This switch may be marked *Coarse Frequency* or with some other equivalent name. Since greater sawtooth capacitance causes increase of the time constant for charging through any given resistance, the greater capacitance decreases the oscillator frequency. Conversely, when a smaller sawtooth capacitance is switched into the oscillator circuit there is increase of oscillator frequency. The several positions of the range switch provide sweep frequency bands extending from about 15 cycles per second to about 30 kilocycles with gaseous tube oscillators, or to 70 or more kilocycles per second with vacuum tube oscillators.

Immediately above the range switch is a *Frequency Vernier* control for varying the resistance through which the sawtooth capacitor is charged. This control may be marked *Frequency*, *Fine Frequency*, or with some other equivalent name. Increasing the charging resistance lengthens the time constant for charging and lowers the oscillator frequency, while decreasing the resistance shortens the time constant and raises oscillator frequency. This control is capable of varying the horizontal sweep frequency throughout any one of the bands covered by the range switch.

The range switch and frequency vernier control may be operated to bring one or more complete cycles of vertical

input voltage onto the oscilloscope screen. If these controls are adjusted to make the horizontal sweep frequency the same as frequency of the vertical input voltage there will be one complete cycle of vertical voltage on the screen. If sweep frequency is lowered to exactly half the vertical frequency there will be two cycles, if made one-third the vertical frequency there will be three cycles, and so on.

In order that a trace to be observed on the oscilloscope screen may remain stationary it is necessary that the horizontal sweep frequency remain in time with or synchronized with frequency of the voltage whose waveform or amplitude is to be observed, and which is causing vertical deflection. If the internal sweep oscillator is being used for horizontal deflection, this oscillator must be synchronized with the vertical deflection voltage. Such synchronization may be effected by applying part of the vertical deflecting voltage to the horizontal sweep oscillator, just as sync pulses are applied to sweep oscillators in television receivers. This synchronizing voltage for the sweep oscillator is taken from some point in the vertical amplifier.

The frequency controls are used to make the free running frequency of the internal sweep oscillator just a little lower or slower than the vertical input frequency. Then a synchronizing voltage applied to the sweep oscillator will bring the oscillator frequency into time with vertical input frequency. The synchronizing voltage has the effect of locking the trace in a stationary position on the screen of the oscilloscope tube.

When the internal sweep oscillator is not being used for horizontal deflection it still is necessary to synchronize the horizontal deflection frequency with the vertical input frequency. Necessary synchronizing voltage then must be secured from some external source whose frequency is equal to or is some simple fraction (sub-multiple) of the vertical input frequency.

The source of synchronizing voltage is selected by the *Sync Selector* switch located at the left of the range switch in Fig. 99-2. With this switch turned to its *Int* (internal) position the

synchronizing voltage is taken from the vertical amplifier as mentioned before. When the sync selector switch is turned to its *Line* position the synchronizing voltage is being taken from one of the tube heater circuits within the oscilloscope, and is at power line frequency. If the switch is placed on the *Ext* (external) position it becomes possible to use for synchronizing any voltage applied between the *Ext Sync* terminal and one of the ground terminals along the bottom of the panel.

At the top center of the panel, just below the cathode-ray tube, is a *Sync Control* consisting of a potentiometer which varies the strength or amplitude of whatever voltage is being used for synchronization. This control sometimes is marked *Sync Amplitude*, *Sync Adjust*, *Sweep Sync*, or with some other equivalent name.

Synchronizing voltage must be of the least strength which allows holding the trace stationary. Otherwise the trace will be distorted and will not show true waveform or response. The sync control should first be turned nearly all the way off while the frequency vernier is used to make the trace as nearly stationary as possible. Then sync control should be increased just enough to hold the trace stationary.

Some oscilloscopes have provision for intensity modulation, which is a periodic blanking of the beam or reduction of intensity by means of some external voltage which makes the grid of the cathode-ray tube periodically more negative. A suitable modulating voltage is applied to intensity modulation terminals of the instrument, and from there passes, either with or without amplification, to the grid-cathode circuit of the tube. This voltage usually consists of short pulses at a frequency higher than that of vertical input voltage, which will break up the trace at the modulation intervals or points. The modulating voltage otherwise may be at some lower frequency which will blank certain portions of the trace.

**Operating the Oscilloscope.** — Any voltages applied from external sources to the vertical input of the oscilloscope must be brought from their sources through shielded cable with the shield connected to ground on the oscilloscope. Although

shielded cable may not be a necessity for other inputs it is preferable also for bringing external voltages to the horizontal input and to the external sync input terminals.

Voltage brought to the vertical input often is accompanied at its source by higher frequencies than are to be observed. A case in point is when a signal is taken from the load resistor of a video detector. The higher frequencies should be bypassed by using on the source end of the input cable a filter probe. Otherwise the traces will be indistinct or have fuzzy outlines and backgrounds.

If vertical input voltage for the oscilloscope is to be the low-frequency modulation of a radio- or video-intermediate-frequency signal taken from one of the high-frequency amplifiers, the demodulation may be accomplished by using a detector probe on the source end of the input cable. Filter probes and detector probes are described in the articles on *Probes*.

When necessary input connections have been arranged, the next step is to make preliminary settings of the various controls on the oscilloscope. When the internal horizontal sweep oscillator is to be used for horizontal deflection, place the sweep selector switch at *Int* and the sync selector switch to *Int*, thus taking a synchronizing voltage from the vertical amplifier which is carrying the vertical input voltage. There may be cases where the internal sweep is to be synchronized by some voltage other than that of the vertical input, whereupon the sync selector would be set at *Ext* and the synchronizing voltage connected to the *Ext Sync* terminal.

In most television alignment work synchronization will be by means of a synchronized sweep voltage supplied from a television signal generator or sweep generator. Then the sweep output from the generator is connected to the *Hor Input* terminal and a ground terminal. The sweep selector switch is set at *Ext*. The sweep voltage already is synchronized in the test generator, so the sync selector switch of the oscilloscope may be set at *Ext* with no connection made to the *Ext Sync* terminal.

When some other external synchronizing voltage is to be used, this voltage is connected to the *Ext Sync* terminal and a ground terminal on the oscilloscope. The sync selector switch is set at *Ext*. Setting of the sweep selector switch depends on the source to be used for horizontal sweep voltage. If this sweep voltage is to come from the internal sweep oscillator the sweep selector is set to *Int*. If an externally supplied voltage is to cause horizontal sweep the sweep selector switch is set to *Ext*, and the synchronizing voltage is connected to the *Ext Sync* terminal and a ground terminal.

When preliminary settings and connections are completed the oscilloscope is placed in operation and allowed to warm up. Final adjustments then may be made as follows:

1. Set the range switch to include the frequency of the voltage applied to the vertical input.

2. Turn the sync control to a low setting, for minimum synchronization voltage.

3. Adjust the intensity control to produce a visible trace or pattern.

4. Adjust the frequency vernier or fine frequency to produce a trace with the desired number of cycles of vertical input voltage on the screen. The range switch may have to be set for a lower frequency band to obtain the desired number of cycles. Continue adjusting the frequency vernier to make the trace stationary or until there is the slowest possible travel to the right or left.

5. Advance the sync control just enough to hold the trace stationary.

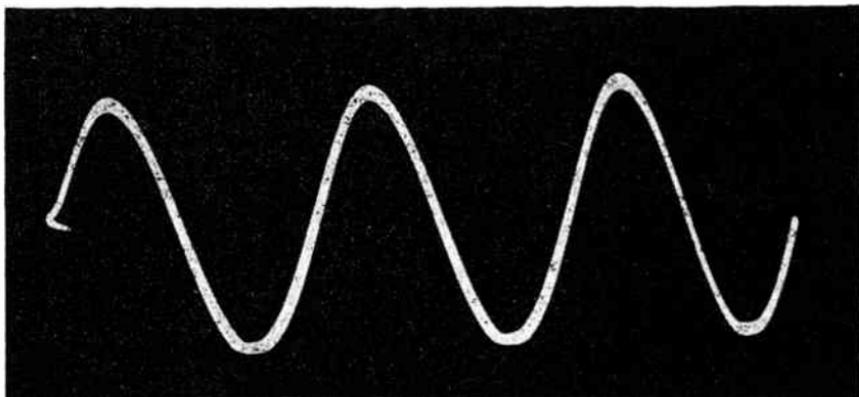
6. Adjust the focus control to make the trace line as thin as possible.

7. Use the centering controls, if necessary, to bring the trace to the desired position on the screen.

8. Adjust the gain controls to make the trace of the desired height and width.

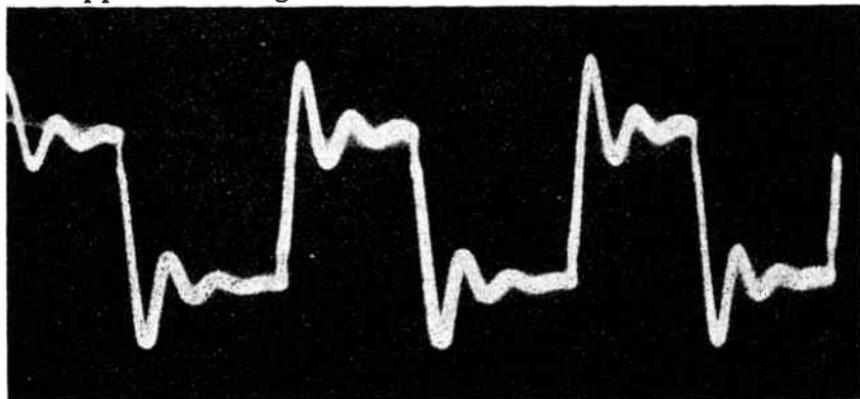
If the oscilloscope is to be left turned on between observations, to avoid need for a reheating time, always reduce the intensity to prevent formation of a trace.

A voltage of approximate sine-wave form applied to the vertical input of the oscilloscope, with horizontal sweep frequency adjusted for one-third the sine-wave frequency, will



*Fig. 99-3.—Three cycles of a sine-wave voltage are shown on the screen of the oscilloscope.*

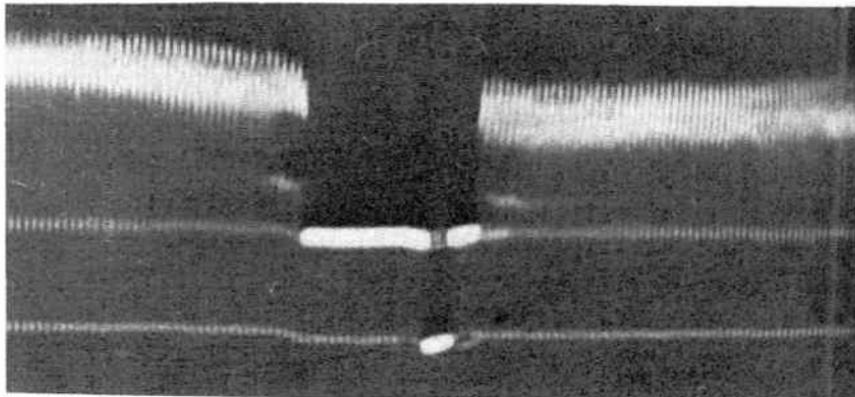
appear on the oscilloscope screen as in Fig. 99-3. A square wave voltage accompanied by oscillation at higher frequency will appear as in Fig. 99-4.



*Fig. 99-4.—A square wave voltage is seen to be accompanied by damped oscillation.*

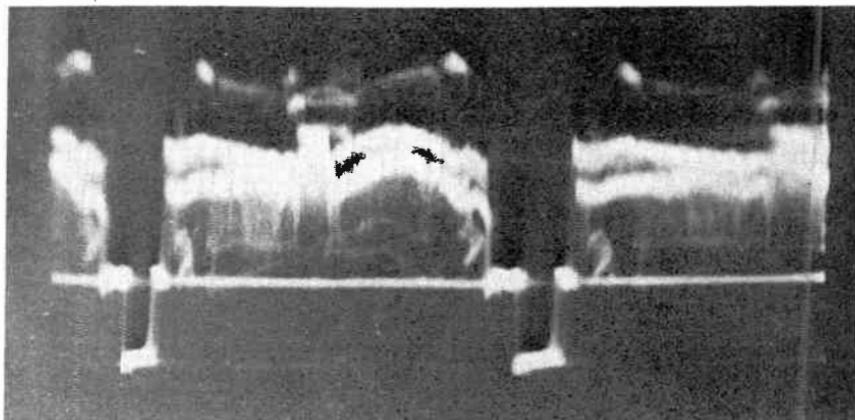
With the oscilloscope vertical input connected to a video detector load, and with horizontal sweep of the oscilloscope adjusted for 60 cycles per second, or some submultiple such as 30 cycles, vertical blanking intervals together with their sync pulses will appear as in Fig. 99-5. With the vertical

input still connected to the same video detector load, but with horizontal sweep of the oscilloscope adjusted for horizontal line frequency or a submultiple of that frequency,



*Fig. 99-5.—A trace showing a vertical sync pulse with preceding and following equalizing and horizontal sync pulses.*

horizontal blanking intervals, with sync pulses and intervening picture signal voltages, will show on the screen of the oscilloscope as in Fig. 99-6.



*Fig. 99-6.—Two horizontal sync pulses with front and back porches of the horizontal blanking pedestal.*

It is necessary only to change the horizontal sweep or frequency controls of the oscilloscope to observe waveforms either at the field frequency or at the horizontal line frequency. Connections to the vertical input remain unchanged.

To adjust horizontal sweep frequency of the oscilloscope to some desired field or line frequency, or to a submultiple of such a frequency, it is necessary only to adjust horizontal sweep controls to obtain on the screen a stationary trace showing some number of cycles of observed voltage. Sweep switches and dials are marked only with approximate frequencies, if at all. There is correct sweep frequency when waveform traces remain stationary.

**Sensitivity and Input Impedance.** — Sensitivity of an oscilloscope is specified as the r-m-s or effective alternating voltage at the input terminals which will cause a trace deflection of one inch when attenuator and gain controls are at their maximum settings. Vertical sensitivity is measured with voltage applied to the vertical input, and horizontal sensitivity with voltage applied to the horizontal input. Sensitivity usually is given as the number of millivolts per inch of deflection. The fewer the millivolts per inch, the greater is vertical or horizontal sensitivity.

For general television servicing it is desirable that vertical sensitivity be no less than 300 millivolts per inch, while sensitivity of 50 to 30 millivolts is better. Practically any oscilloscope will have sufficient horizontal sensitivity for all usual service operations.

If sensitivity is specified in peak-to-peak volts per inch, rather than in r-m-s volts, the peak-to-peak sensitivity may be divided by 2.8 to find equivalent r-m-s sensitivity. For example, peak-to-peak sensitivity of 100 millivolts per inch, divided by 2.8, shows equivalent r-m-s sensitivity to be about 35.7 millivolts per inch.

Any circuit across which the vertical input of the oscilloscope is connected is shunted by impedance due to combination of internal resistance and capacitance between the vertical input terminals. Vertical input resistance often is between a half-megohm and five or more megohms, while input capacitance often is between 25 and 50 mmf.

Reactance of input capacitances is many megohms at low frequencies, and causes no serious loading of measured cir-

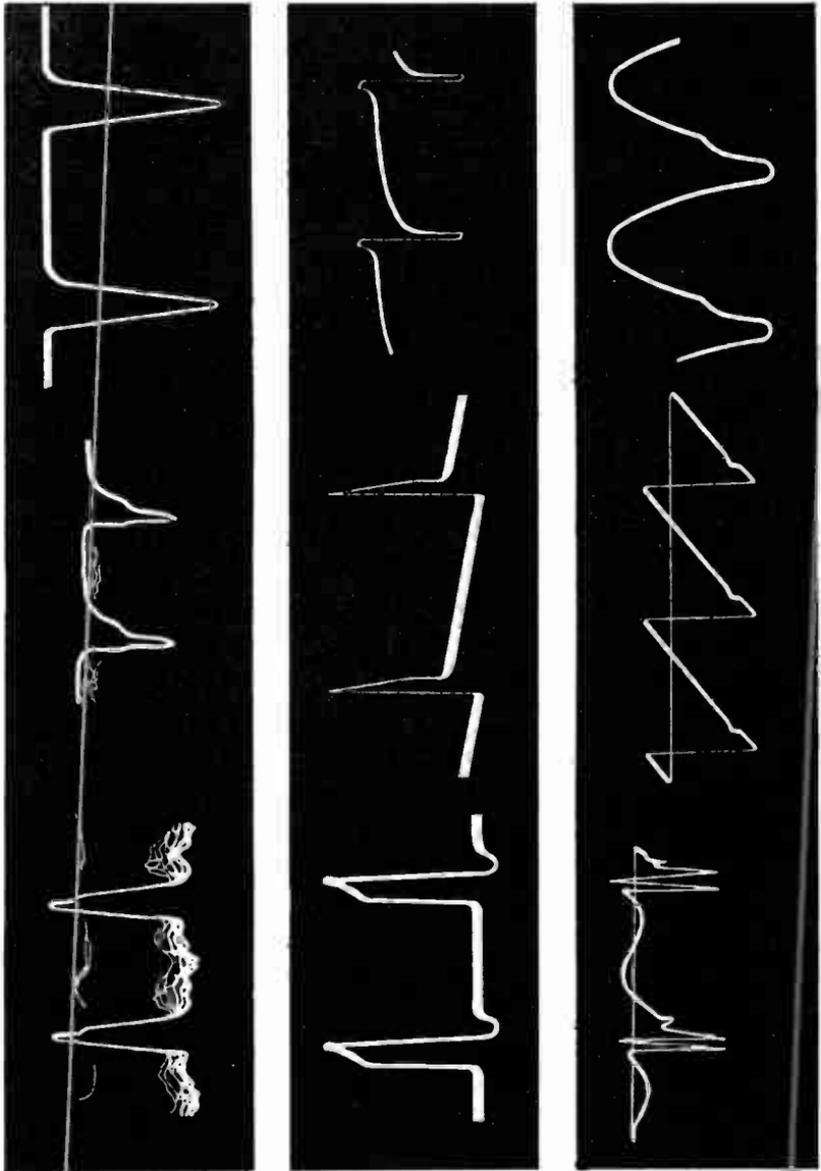


Fig. 99-7.—Some of the waveforms which accompany correct performance of circuits between a video detector and deflecting coils.

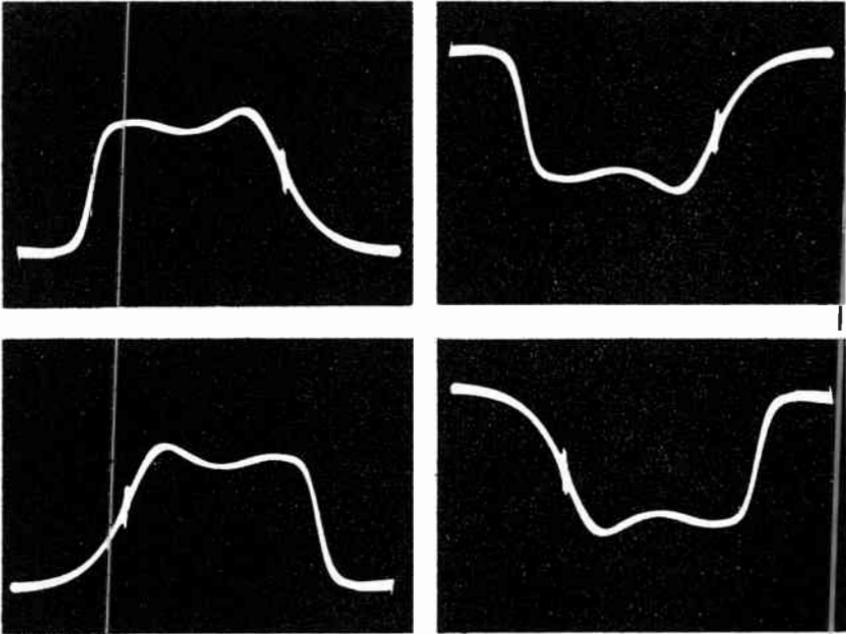
cuits. But should frequency of observed voltage be as high as a half-megacycle, the capacitive reactance would drop to between 5,000 and 10,000 ohms. Since impedance always is less than either of the factors, capacitance and resistance, the low impedance at high frequencies may cause serious loading of measured circuits.

Frequency response of amplifier sections in oscilloscopes may be defined as the range of observed frequencies in which gains remains practically flat or constant, or it may be specified as the number of decibels by which gain drops below maximum at certain frequencies. For general television servicing a vertical response flat to 500 kilocycles should be satisfactory. For servicing color receivers the vertical response should have only slight drop at 3.58 megacycles, the frequency at which the color oscillator section and related circuits operate.

**Receiver Waveforms and Frequency Responses.** — Service operations performed with the oscilloscope are of two general kinds; observations of waveforms and observations of frequency responses. Waveform traces, such as those of Figs. 99-3 to 99-6, show how an observed voltage varies during one or more of its cycles. Certain waveforms must be present at various points in receiver circuits in order that necessary functions may be performed. Fig. 99-7 shows a number of waveforms which should appear in circuits between the video detector and coils of the deflecting yoke in a particular receiver. Decided departures from these waveforms would indicate trouble in that receiver, and its approximate location.

An oscilloscope trace of frequency response shows relative voltage gains of an amplifier or of several cascaded amplifiers at all frequencies within some selected range. Fig. 99-8 shows a typical frequency response observed at the load of a video detector. All four traces indicate the same response, but there are reversals left and right and inversions of top and bottom. This comes about because sweep frequencies may increase from left to right or from right to left, and because voltage gains of amplifiers may be either positive or negative with

reference to potential of chassis ground or B-minus in the receiver.

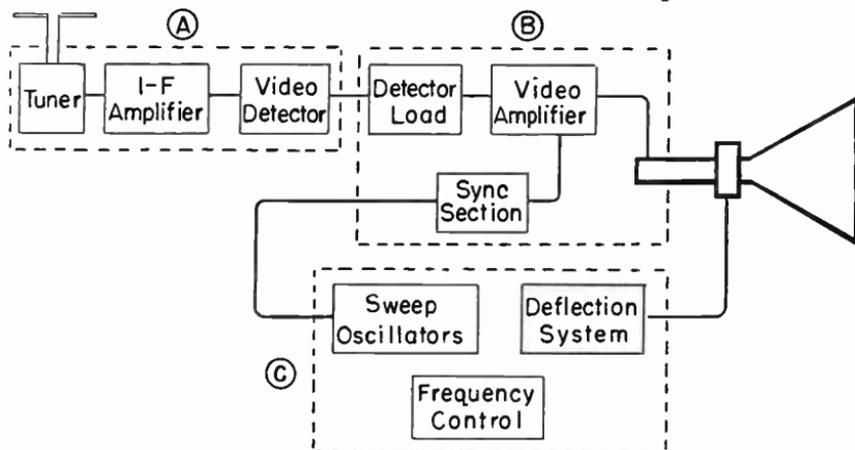


*Fig. 99-8.—The same frequency response may appear in any of these positions on the oscilloscope screen.*

For waveform observation the vertical input of the oscilloscope is connected across the high side and the low side or ground at the output of the amplifier in which the varying voltage exists. Voltage applied to the input of the amplifier or receiver may be a television signal, or the tested circuit may produce voltage waveforms without help from any external signal. In only a few cases would it be necessary to have a signal generator of any kind furnishing an input voltage.

For observation of frequency response the vertical input of the oscilloscope is connected across the output of the amplifier or receiver section whose relative gains in a range of frequencies are to be observed. Input voltage for amplifier or receiver is supplied by a sweep generator adjusted for the range of frequencies within which there should be voltage gains. The greater the gain at any frequency, and at the cor-

responding point along the response trace, the higher will be the trace at that point. Sweep range should be wide enough that gain drops to zero on both sides of the response trace.



*Fig. 99-9.—Waveforms, frequency responses, or both, may be observed from these sections of a television receiver.*

Fig. 99-9 shows receiver sections in which waveforms and frequency responses may be observed. Frequency responses may be observed for any amplifier enclosed at *A*. Except at output of the mixer in the tuner section it is necessary that the vertical input cable of the oscilloscope be terminated in a detector probe which demodulates signal voltages. Waveforms cannot be observed in sections at *A* unless the oscilloscope vertical amplifier has good gain at carrier and intermediate frequencies, and unless horizontal sweep frequency can be made high enough to clearly separate cycles of voltage at these frequencies. Such abilities would be unusual in service oscilloscopes.

Waveforms may be observed in any sections enclosed at *B* and *C*. A detector probe is not needed, because i-f signals have been demodulated by the video detector. Television broadcast signals must reach the receiver antenna terminals in order that waveforms useful for service work may appear in sections enclosed at *B*.

At and beyond the sweep oscillators, in sections enclosed

at C, there are deflection sawtooth voltages and other deflection voltages produced by the sweep oscillators whether or not a received signal or voltage from any signal generator is applied anywhere in the receiver. Therefore, waveforms may be observed in these sections with the oscilloscope as the only service instrument.

Frequency response may be taken from the video detector load, and will indicate combined relative gain of all amplifier stages from the detector load back to the point at which the sweep generator is connected. Frequency response of the video amplifier section may be observed provided the sweep generator will operate in a range of frequencies from near zero up to five megacycles. Frequency responses could be observed in the sync section and in following sections enclosed at C of Fig. 99-9, but these responses ordinarily are not made use of during service work, and would have no particular significance.

**Frequency Compensation.** — Frequencies of picture signals and of voltages in color oscillator and chrominance sections extend throughout the video-frequency range up to about four megacycles. Sync pulses and many deflection voltages have characteristics of square waves, meaning that they are combinations of frequencies from about one-tenth up to about ten times the frequencies of the waves themselves. For these reasons it is desirable that vertical amplifier sections of service oscilloscopes have almost uniform gain from about 10 cycles per second to at least 200 kilocycles, and to about four megacycles for color servicing.

Even though the vertical amplifier is compensated to provide broad frequency response, traces will not show true waveforms unless the vertical attenuator or gain control provides uniform attenuation over the entire frequency range for all settings of this control. In a gain control not designed for frequency compensation and uniform attenuation, the capacitive reactances of stray capacitances vary with frequency and alter the attenuation. With such gain controls there is absence

of frequency discrimination only when the adjustment is for maximum gain.

Circuits of a fairly typical frequency-compensating attenuator are shown by Fig. 99-10. With the attenuator switch at *1*, the vertical input is connected directly to the grid of the first amplifier. There is maximum gain. Provided the amplifiers

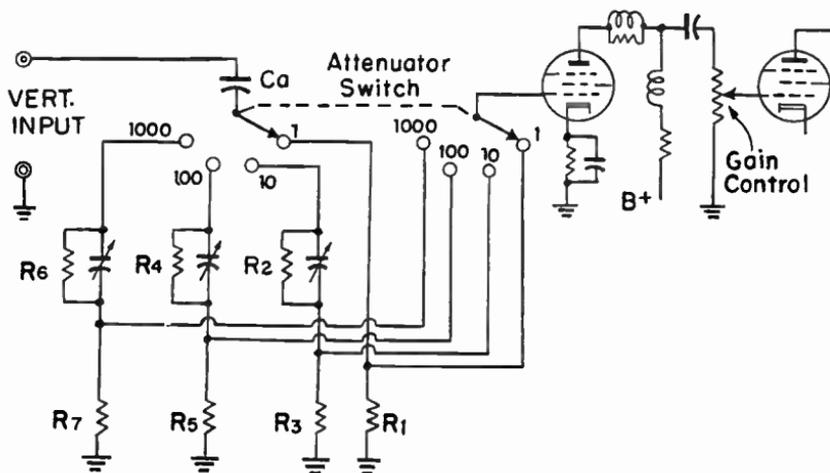


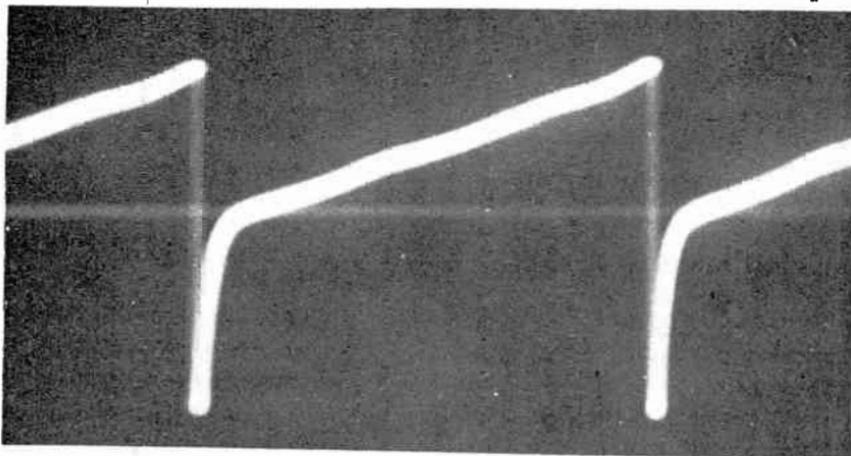
Fig. 99-10.—A frequency-compensating attenuator between the vertical input and first vertical amplifier.

themselves are frequency compensated, waveforms will be shown correctly.

Moving the attenuator switch to *10* reduces gain to  $1/10$  of maximum, by increasing attenuation. This is because resistors *R2* and *R3* form a voltage divider to reduce voltage going from their junction to the amplifier grid. To prevent frequency discrimination the small capacitor across resistor *R2* is adjusted to compensate for effects of stray capacitances, and is left so. With the switch at *100*, effective gain is reduced to  $1/100$  of maximum, or the input voltage must be 100 times as strong as before in order to cause traces of the same height. With the switch at *1000* the effective gain is  $1/1000$  of maximum, due to the ratio of voltage divider resistors. Capacitors across resistors *R4* and *R6* are adjusted to compensate for stray capacitance effects at each switch position.

The attenuator switch varies effective gain only in large

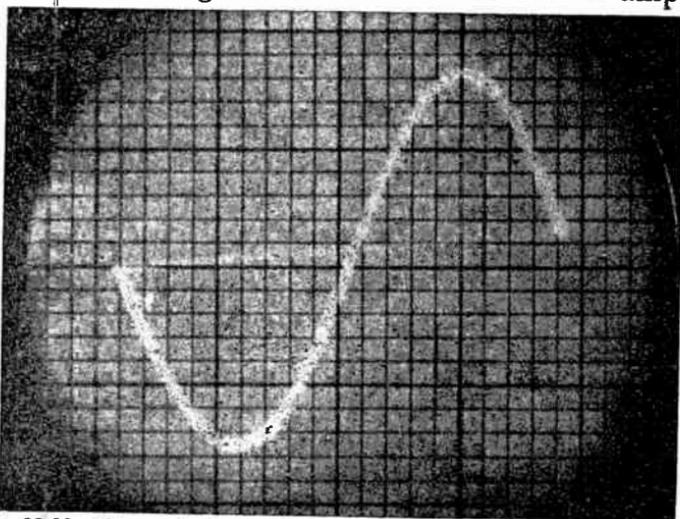
steps. For intermediate adjustments there is a separate gain control between first and second vertical amplifier tubes in Fig. 99-1C. To have the advantages of a frequency-compensated attenuator it is necessary to use on the vertical input



*Fig. 99-11.—Peak-to-peak voltages usually are shown on oscilloscope traces in service manuals.*

cable of the oscilloscope a frequency-compensating probe such as described elsewhere in this book.

**Peak-To-Peak Voltages.** — Voltages of irregular waveform, such as those of Fig. 99-7, should be of overall amplitudes



*Fig. 99-12.—This graph sheet allows convenient measurement of trace heights.*

within about ten per cent of values listed in receiver service manuals. These overall amplitudes are specified as peak-to-peak voltages, or potential differences between extreme top and extreme bottom of the waveforms. Typical examples are illustrated by Fig. 99-11.

A peak-to-peak voltage is measured by comparing the height of its trace on the oscilloscope screen with height of a trace produced by a known or measurable voltage applied to the vertical input. For comparisons of heights it is necessary to have in front of the cathode-ray tube a transparent sheet with graduations as shown by Fig. 99-12 or something equivalent.

The comparison voltage may be alternating, of sine-wave form, such as can be accurately measured with an a-c voltmeter incorporated as one of the functions in vacuum tube voltmeters and volt-ohm-milliammeters. Procedure is as follows: On the oscilloscope screen set up a trace of the voltage waveform to be measured. Adjust vertical attenuation and gain to make this trace of some easily recognized height, such as one inch or two inches from peak to peak. Without altering vertical gain, transfer the vertical input of the oscilloscope to the source of comparison voltage. Frequency of the comparison voltage need not be the same as that of the voltage to be measured, since comparison will be only of trace heights or of overall heights of screen illuminations produced by the two voltages.

If the comparison voltage is adjustable, make its value such as to extend over the same screen height as the measured voltage. The r-m-s or effective value of the sine-wave comparison voltage then is measured on an a-c voltmeter and is multiplied by 2.8 to determine peak-to-peak voltage of the irregular waveform being measured.

If the comparison voltage is not adjustable, height of the waveform being measured may be adjusted by the attenuator and gain control of the oscilloscope to equal the height of the trace or the illumination from the comparison voltage. Then r-m-s value of the comparison voltage, as measured, is multi-

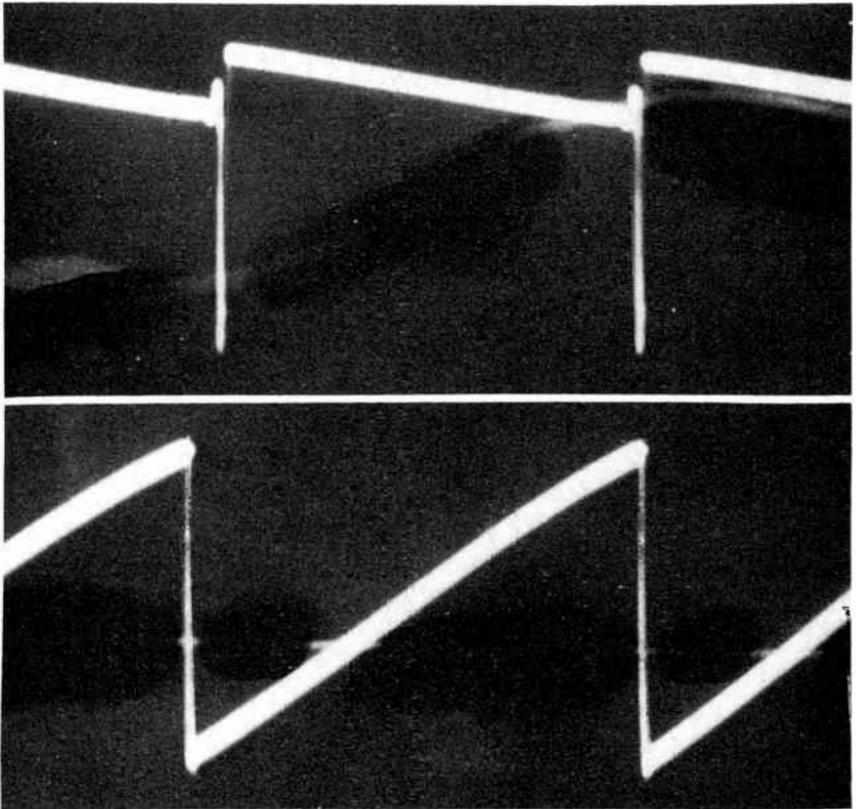
plied by 2.8 to determine peak-to-peak voltage of the waveform being observed.

The ratio of any two alternating voltages applied to the vertical input of the oscilloscope is equal to the ratio of trace heights produced by the two voltages, provided both frequencies are within the range over which the oscilloscope attenuator and vertical amplifier provide uniform or flat gain. If a sine-wave comparison voltage is measured with an a-c voltmeter and the measurer r-m-s value multiplied by 2.8, the product is peak-to-peak voltage of the sine wave. Multiplying this peak-to-peak voltage by the ratio of trace heights gives peak-to-peak value of the irregular waveform in question.

Comparison voltages are obtained most conveniently from a voltage calibrator, which may be a separate instrument or may be built into the oscilloscope. The calibrator usually furnishes a square wave or a clipped sine wave, both of which are flat across the top and bottom. The flat portions produce horizontal lines whose separation or height is easily noted whether or not the horizontal sweep of the oscilloscope is adjusted to hold a stationary trace of the comparison voltage. Calibrator voltage is indicated directly in peak-to-peak values, either by a built-in voltmeter or else by graduated dials on which pointers are rotated to adjust calibrator voltage for a height equaling that of any waveform to be measured.

**Oscilloscope As A-c Voltmeter.** — An oscilloscope may be used for measuring r-m-s or effective value of any alternating voltage of symmetrical waveform, provided the frequency of measured voltage is within the range for which the oscilloscope has flat response. The advantage of this method is that input impedance of an oscilloscope usually is much greater than that of any commonly available a-c voltmeter, including the a-c voltage function of most electronic voltmeters. Consequently, the oscilloscope does not so heavily load a circuit in which voltage is to be measured, and indicated voltages are more nearly those existing during normal operation.

Procedure is to compare the height of a trace produced by the unknown voltage with height of a trace from another alternating voltage which may be measured on an ordinary voltmeter. When traces are of equal height the two voltages are equal. If traces are of unequal heights the ratio of heights is equal to the ratio of voltages compared. It is not necessary that traces be held stationary on the oscilloscope screen by adjustment of horizontal sweep controls, but only that overall heights may be measured on a graduated scale or otherwise.



*Fig. 99-13.—The voltage waveform at A is quite different from the current waveform at B, although both are taken from the same circuit components.*

**Current Waveforms.** — The oscilloscope is primarily designed to display traces of voltage waveforms, not of current waveforms. However, it often is desirable to observe a current waveform, since only in a pure resistance is it certain

that waveforms of a voltage and its resulting current are alike. Waveform of current in an inductance or capacitance may be decidedly different from waveform of voltage across the same unit. At *A* of Fig. 99-13 is the waveform of voltage across vertical coils in a deflecting yoke, while at *B* is the waveform of current in the same coils.

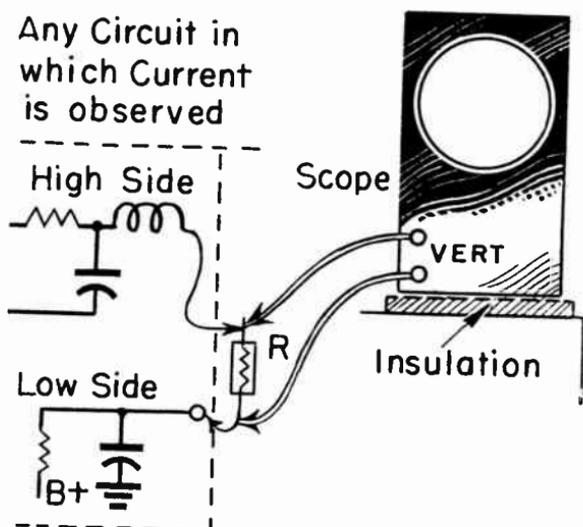
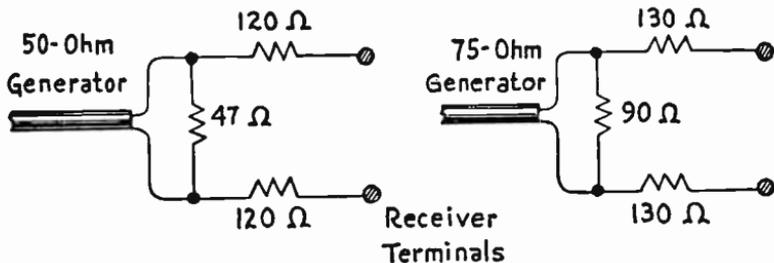


Fig. 99-14.—Connections for observing a current waveform on the oscilloscope.

To observe a current waveform the circuit carrying the current is opened, as in Fig. 99-14, and between the opened ends is connected a non-inductive resistor. This series resistance must be small enough not to disturb normal operation of the circuit; a value of one to five ohms usually being satisfactory. Vertical input of the oscilloscope is connected across the inserted resistor. Since current and voltage are in phase in a resistance, the observed waveform of voltage across the resistor will also be the waveform of current in the resistor and in the series circuit.

## P

**PADS, MATCHING.** — A matching pad is an arrangement of resistors used to provide desired impedance relations between a source and a load. When a signal generator is connected to the antenna terminals of a receiver the resistance load on the generator should approximately equal the internal output impedance of the generator. At the same time the resistance



*Fig. 100-1.—Matching pads used between signal generators and receiver antenna terminals.*

placed across the receiver terminals should approximately equal the receiver input impedance. If this is not done there will be wave reflection losses at the junctions and standing waves at certain frequencies when generator output and receiver input impedance are not alike. Effects are similar to those with an antenna, a transmission line, and a receiver that are not matched.

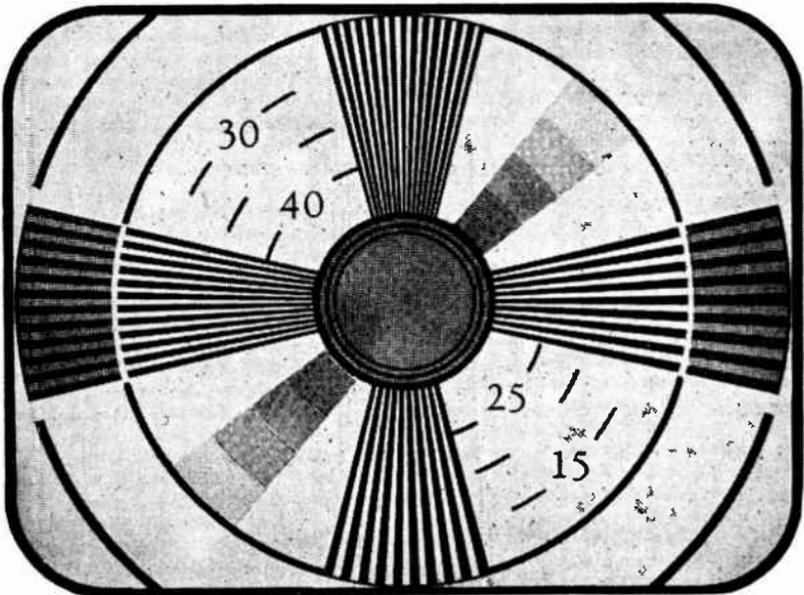
Signal generators usually have output impedances of either 50 ohms or 75 ohms. The output of such generators may be directly connected to antenna terminals of a receiver designed for unbalanced transmission lines of similar impedance, such as coaxial transmission lines. But a matching pad should be used between a 50-ohm or 75-ohm generator output and the antenna terminals of a receiver designed for a 300-ohm balanced transmission line. Fig. 100-1 shows connections and resistances used in two pads arranged for this purpose. The pads usually are placed at the receiver end of the cable between generator and receiver.

**PATTERNS, TEST.** — A test pattern is a stationary design transmitted in the picture channel of a television station between the time when the station goes on the air and the time when regular programs are transmitted. The test pattern is accompanied by a steady audible tone transmission in the sound portion of the channel. The chief purpose of test pattern and tone transmission is to furnish types of signals which are of assistance in alignment and other adjustments on television receiver installations.

Test patterns from different stations vary in design, but many features common to most patterns are illustrated by Fig. 101-1. There are parts of a large outer circle whose edges at the left and right indicate the correct width of picture area in the mask, and there is a smaller inner circle or parts of such a circle whose top and bottom indicate the correct height of picture area. These circles are useful in adjustment of size controls for width and height, also in adjustment of centering controls. The circles provide a check on linearity and provide a guide for adjustments, since all parts of each circle should be equidistant from the center of the pattern, or the circles should be truly round.

Resolution wedges extend vertically and horizontally from the center of the pattern. These wedges consist of lines and spaces which are relatively wide and far apart at the outer ends, and narrow and close together near the center. The farther toward the center the wedge lines remain distinctly separated the better is the resolution or the greater is the detail which will be rendered in pictures. The approximately vertical lines provide a check on horizontal resolution, since the electron beam there travels alternately from dark to light shadings. The approximately horizontal lines provide a check on vertical resolution, since they extend across vertical lines of pictures. On some test patterns there are reference numbers showing resolution in number of lines at various points along the wedges. The numbers on Fig. 101-1 would each be multiplied by 10 to give corresponding numbers of lines which are separated at the respective positions.

The resolution wedges provide an additional check of linearity, since their lines and spaces should be straight at all points if there is linear reproduction. The vertical lines of the wedges for horizontal resolution will show certain kinds of peaking or excessive response at certain frequencies in ampli-



*Fig. 101-1.—Features of a test pattern which are of assistance when making adjustments.*

fiers. When such trouble is present there will be a dark smudge from left to right across the wedge lines.

With the pattern here illustrated there are two tone wedges whose shading varies along the length of the wedges. These wedges provide a check on contrast and brightness, for with correct adjustments the change of shading or tone will be fairly gradual. Parts of a pattern other than tone wedges often have graduated shading to allow checking contrast and brightness relations.

The lines on test patterns make evident the presence of ghosts or signal reflections. The large clear areas allow easy identification of the formations which are peculiar to and

which indicate the existence of many kinds of interference. Test pattern observation and the interpretation of various effects is the basis for a system of trouble location described in this book.

**POTENTIAL, CONTACT.** — Contact potential in a tube is a small negative potential, with reference to the cathode, which is produced on elements close to the cathode when the cathode is heated. This negative contact potential is the result of electrons collected by the element near the cathode when the cathode is heated to cause emission and a negative space charge within the tube. The higher the cathode temperature or the greater the emission, the greater becomes the contact potential. In tubes having heater cathodes the value usually is between 0.3 and 0.7 volt, but it may be as high as 1.4 to 1.5 volt. Contact potential is maximum on the element closest to the cathode, which is the grid of a triode, pentode or beam power tube, or is the plate of a diode.

Contact potential will cause current of a few microamperes in a resistance connected between the cathode and the nearest other element provided no other opposing potentials or voltages are present. Contact potential will affect a grid bias. It causes incorrect indications when using detector probes containing a tube unless the effect is compensated for, usually by another similar tube in the same circuit but with connections reversed between cathode and the nearest element. Contact potential is the chief cause of erratic control of plate current in a tube with a free grid or a disconnected grid. The value of negative contact potential tends to drop as a tube ages in use. Contact potential will be indicated by a voltmeter having sensitivity of 20,000 ohms per volt, or better by an electronic voltmeter, when no external voltages are applied to any of the tube elements while the cathode is heated.

**POWER SUPPLY, HIGH-VOLTAGE.** — The high-voltage power supply for a television receiver furnishes direct current obtained by rectification of alternating voltage, and by filtering of rectified current. D-c potentials are 25,000 or more volts for the high-voltage anode of direct view magnetic

deflection tubes. Two principal types of high-voltage supply are in use. In one of them, called a pulse-operated or flyback type, there is rectification of high-potential pulses caused by auto-transformer action in an extension of the primary winding on the horizontal sweep output transformer. The rectified pulse voltage may be raised in a voltage multiplier before being filtered and passed to the picture tube. Flyback or pulse-operated power supplies are used with magnetic deflection picture tubes.

The other type of high-voltage power supply employs an oscillator operating at radio frequency to produce from the d-c voltages on its elements an alternating voltage which is stepped up by a transformer. This stepped-up voltage is filtered and delivered to the picture tube circuit. This type is used with electrostatic deflection picture tubes and sometimes with magnetic deflection tubes.

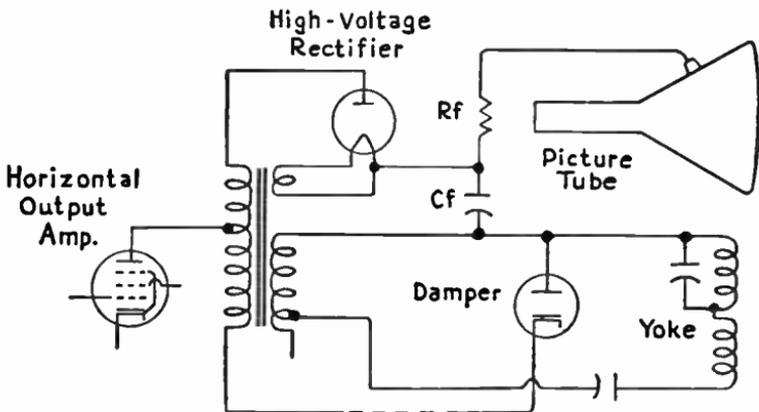


Fig. 103-1.—High-voltage power supply of the flyback or pulse-operated type.

**Flyback Power Supply.** — Circuit connections for a typical flyback or pulse-operated power supply are shown by Fig. 103-1. This style of power supply, shown at the top of the diagram, may be connected to any of the deflection yoke circuits in general use. One such circuit is shown in the lower part of the diagram.

At the instant of plate current cutoff in the horizontal output amplifier there is collapse of magnetic fields in the deflection yoke circuit, which includes the secondary winding of the output transformer. It is this collapse that starts retraces. The sudden change of magnetic field induces a pulse of negative potential in the secondary. Induction between secondary and primary induces a positive pulse in the primary winding. In the portion of the primary connected between the amplifier plate and damper cathode this pulse potential reaches 4,000 volts or more. The upward extension of the primary makes this winding an auto-transformer, and between top and bottom of the entire winding the pulse potential is stepped up to something on the order of 8,000 to 10,000 volts. This potential varies with operating conditions in the sweep amplifier and yoke circuits and on the strength of the negative pulse initially produced in the yoke circuit.

The high-voltage pulses are applied to the high-voltage rectifier. Rectified current charges capacitor  $C_f$ . The time constant of  $C_f$  and resistor  $R_f$  is longer than intervals between pulses, so the capacitor retains a charge and delivers fairly constant voltage to the picture tube anode through  $R_f$ . Filtering effect is increased by capacitance of conductive coatings and the glass of the envelope in some types of picture tubes.

Power supply voltage is affected to some extent by adjustment of width controls, drive controls, and peaking controls, since these adjustments alter the plate current of the output amplifier and currents in the yoke circuit. Any failure which prevents operation of the output amplifier stops the high-voltage supply to the picture tube. Damper failure stops the high-voltage supply when output amplifier plate voltage is furnished through the damper circuit.

**Voltage Multipliers.** — A voltage doubler circuit for a fly-back power supply is shown by Fig. 103-2. Because input to the doubler consists entirely of positive voltage pulses instead of alternating voltage the action differs from that in doublers used on line voltage in low-voltage B-supplies of receivers. In the doubler circuit are two high-voltage rectifier tubes, 1 and

2, also capacitors  $C_a$ ,  $C_b$ , and  $C_c$ , and resistor  $R$ . Input is from the top of the auto-transformer type of primary winding on the output transformer. Output to the picture tube anode is through resistor  $R_f$ .

During each voltage pulse from the transformer the plate of rectifier 1 is made positive, and there is conduction through this tube to charge capacitor  $C_a$ . Capacitor  $C_b$  previously will have been charged in the marked polarity by action to be

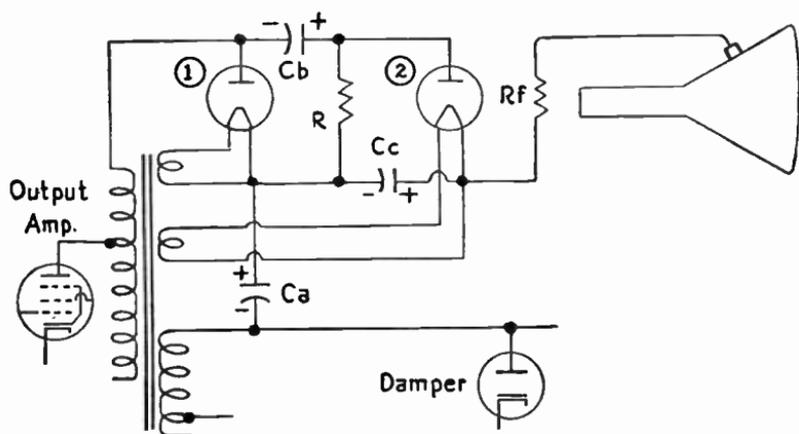


Fig. 103-2.—Voltage doubler circuit used with a flyback high-voltage supply.

explained. The potential on the side of  $C_b$  toward the plate of rectifier 2 is positive, so this rectifier conducts to charge capacitor  $C_c$  as marked. Capacitors  $C_a$  and  $C_c$  are thus charged simultaneously to approximately peak pulse voltage. These two capacitors are in series with each other and the connection through resistor  $R_f$  to the picture tube anode. Consequently there is applied to the anode a voltage approximately equal to twice the pulse voltage.

During intervals between pulses the plate of rectifier 1 is negative and there is current cutoff in this tube. Discharge from capacitor  $C_a$  flows through resistor  $R$  to charge capacitor  $C_b$  in the marked polarity. It is this charge on capacitor  $C_b$  which remains long enough to make the plate of rectifier 2

positive at the beginning of the next voltage pulse. Capacitors  $C_a$  and  $C_c$  are charged during each pulse period, and capacitor  $C_b$  is charged during intervals between pulses.

**R-f Power Supply.** — Connections for a typical r-f type high-voltage power supply are shown by Fig. 103-3. The oscillator tube is furnished with plate and screen voltages from the low-voltage B-power supply through a filter system which prevents escape of r-f voltages and currents from the oscillator circuit to other parts of the receiver. The entire high-voltage power supply system is enclosed within a shield to prevent r-f radiation to the remainder of the set.

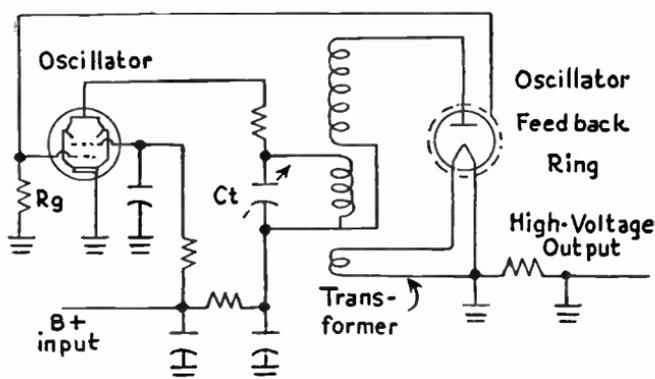


Fig. 103-3.—The r-f oscillator type of high-voltage power supply.

The oscillator tube is biased by grid resistor  $R_g$ . The tuned plate circuit of the oscillator consists of tuning capacitor  $C_t$  and the portion of the transformer winding across which this capacitor is connected. Capacitive feedback to the oscillator grid is here from a ring placed around the outside of the glass envelope of the high-voltage rectifier tube. This feedback ring usually is a coiled spring long enough to go around the envelope, with the ends of the spring joined so that it may be slipped onto the tube. Other styles of metallic clamps or rings may be used. There is capacitance between the metal ring and the tube elements, with the glass envelope and vacuum space as dielectric. In some power supplies of this general type the feedback is from a tickler coil coupled to the high end of the

transformer winding. Oscillator frequency most often is in the neighborhood of 200 kilocycles.

The oscillator tube usually is a beam power type, although it may be a triode, or a pentode connected as a triode. The oscillation transformer is an air-core type of high-Q construction. The turns ratio, secondary to primary, may be as much as 35 to 1. The transformer secondary winding is tuned by its inductance and the sum of distributed capacitance in the winding, stray capacitance in connections, and internal capacitance of the rectifier tube.

Output voltage from the rectifier will be maximum when the oscillator plate winding is tuned to the natural frequency of the secondary winding, but to provide stable oscillation the plate winding is tuned to a frequency slightly higher or lower. When capacitor  $C_t$  is adjustable it is used for varying the output voltage. The method is to connect to the output a suitable voltmeter, adjust the capacitor first for maximum voltage, and then detune it to allow normal output voltage specified for the receiver. The only other adjustment is for feedback. A feedback ring or clamp should be placed, to begin with, so that it surrounds the lower edge of the tubular shield which extends up toward the top cap inside the rectifier tube, then moved slightly if necessary to have oscillation and power output. Incorrect positioning will reduce the maximum output voltage or may prevent any output at all.

**POWER SUPPLY, LOW-VOLTAGE.** — Low-voltage B-power supplies for television receivers employ the same general principles as found in power supplies for sound radio receivers. In most of the larger television receivers the low-voltage supply system includes a power transformer, one or more full-wave tube rectifiers, and capacitor-choke filters which sometimes are supplemented with capacitor-resistor filters for circuits taking small currents.

Many of the smaller television receivers, and some large ones, utilize transformerless or ac-dc power supplies, often with selenium rectifiers or combinations of tube and selenium rectifiers. High voltages for plates and screens may be had from

voltage doublers and sometimes from voltage triplers. Another method makes use of two line-voltage rectifier systems in series, with the high positive voltage fed to plates and screens while the low negative voltage goes to cathodes of the same tubes.

Heaters and filaments are operated in parallel from a transformer secondary in some receivers, and in other receivers there are various series and series-parallel heater systems operated from line voltage.

**Series Cathode Currents.** — Amplifiers and other tubes may be connected to a B-power supply in such a manner that electron flow from B-minus of the supply goes to the cathodes of only part of the tubes. When this electron flow leaves the plates and screens of this first group of tubes it does not return directly to the positive side of the power supply but goes to the cathodes of a second group of tubes. From the plates and screens of this second group the electrons flow returns to B-plus of the power supply. This general principle of connecting tubes in series to utilize the same current for plates and screens of more than one group is subject to almost innumerable modifications of the particular tubes included in the different groups.

One application is shown by the simplified diagram of Fig. 104-1. All electron flow from B-minus of the power supply that goes through ground to the cathodes of three video i-f amplifiers passes from the plates and screens of these amplifiers to the cathodes of the audio output and driver tubes. All plate current from the audio output tube then returns to the power supply through the B+ 375-volt line. Screen current from the audio output tube, and both plate current and screen current from the audio driver tube return to the power supply through the 350-volt B+ line. Part of the electron flow which has passed from B- through the video amplifier goes from the plate and screen of this tube along with flows from the i-f amplifiers to the audio tubes. The remainder flows through a resistor to the B+ 350-volt line and the power supply.

Between the lower and upper groups of tubes is a resistance-capacitance filter, which is necessary to prevent signal currents from one group reaching the other one. Power supply output voltage divides between the groups of tubes. In the arrangement illustrated, plate voltage on the audio out-

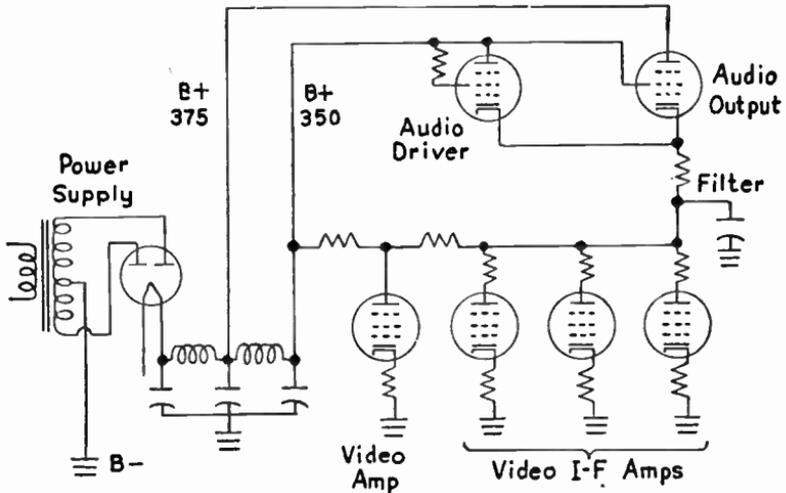


Fig. 104-1.—Groups of tubes so connected as to be in series for their cathode currents.

put tube would be 375 volts and on the plate of the audio driver would be 350 volts. The cathode voltage of both these tubes may be 140 volts positive with reference to ground. There is some drop in the filter, which may leave about 130 to 135 volts at the plates of the video i-f amplifiers. Because of its connection to the B+ 350 volt line the plate voltage of the video amplifier may be on the order of 160 to 175 volts. If currents could not conveniently be equalized in series connected groups of tubes, additional current from one group may be carried around the other group through a resistor paralleling that other group. With any arrangement which may be used there still is the advantage that total current from the power supply is a minimum for the total number of tubes in the circuits.

The cathodes of one group of tubes will be at rather high positive voltage above ground. In order to maintain suitable bias voltages between grids and cathodes the grids of this group also will be highly positive. With some series circuits the B-minus lines are insulated and ground connection is made from between groups of tubes. Then cathodes of one group will be highly negative with reference to ground while their plates and screens will be at nearly ground potential.

When tubes are connected for series cathode currents the tubes in the high-voltage group provide a certain amount of voltage regulation for plates of the other tubes which are fed from the first cathodes. As an example, assume that a change of signal strength at the lower tubes of Fig. 104-1 alters the grid voltage and internal resistance of these tubes to raise their plate voltages above the 130 or 135 volts previously mentioned. There will be similar rise of voltage at the cathodes of the upper tubes, or the cathodes will become more positive. This is equivalent to making the grids of the upper tubes more negative, and their plate-screen currents or cathode currents will decrease. Then there will be less voltage drop across all the tubes in the lower group, which means decreased voltage between cathodes and plates or decreased plate voltage for these lower tubes. Since the action commenced with a rise of this plate voltage, the final effect is compensation for the change.

**PREAMPLIFIERS.** — A television preamplifier or booster is a radio-frequency amplifier constructed as a separate self-powered unit for connection between a television receiver and the transmission line coming from the antenna. The principal purpose of a preamplifier is to provide increased signal strength at the receiver input in localities classed as fringe areas, or when only an indoor antenna may be used, under which conditions the gain in many receivers is insufficient. There is also an increase in selectivity, as with added r-f stages in any receiver.

Signal booster units use miniature r-f pentodes for amplifiers, the most common type being the 6AK5. The amplifier

circuit most often is a tuned-grid tuned-plate type, for maximum gain, although sometimes only the grid is tuned and again only the plate is tuned. The method of tuning is about equally divided between movable coil cores, or permeability tuning, and variable capacitors. Tuning ordinarily is continuous through each television band, with switching for manually selecting either the low-band channels 2 through 6 or the high-band channels 7 through 13. A few units provide continuous tuning through both the television bands and the f-m broadcast band between them. Always there is provision for connecting the transmission line directly through the booster unit to the receiver antenna terminals, bypassing the amplifier.

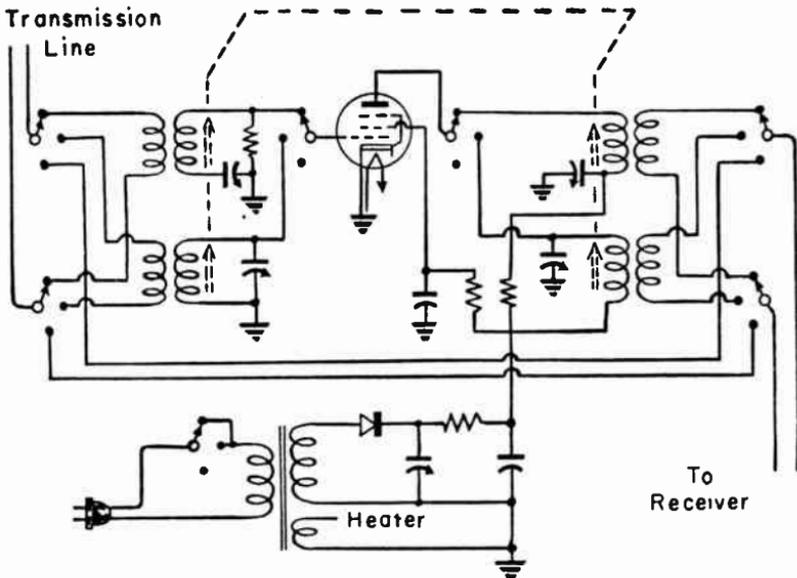


Fig. 105-1.—Preamplifier circuit employing a single r-f stage with band switching.

In boosters or preamplifiers of one general class a single tube is switched into the tuned coil circuits for whichever band of channels is to provide reception. A fairly typical circuit and switching connections are shown by Fig. 105-1. Switching for the two bands and for direct connection from

line to receiver is handled by a three-position seven-circuit ganged switch. Two switch circuits connect to the transmission line, one each to the grid and plate of the tube, two more to the output for the receiver, and one goes to the power line. With switch elements in their upper positions, as on the diagram, connections are completed for high-band reception. The next lower position makes connections for low-band reception. The bottom position connects the line input directly to the receiver output of the booster.

Tuning in Fig. 105-1 is by means of movable cores in the grid windings and plate windings of the input and output coupling transformers. Capacitor trimmers are shown on each of the four windings. In some units the tuning is with capacitors which are variable, with adjustable cores or capacitors for trimmers.

In another general class of preamplifiers there are separate tubes for each band as well as separate sets of tuning coils. The generally equivalent connections would be made upon placing a second tube between the tops of the grid and plate windings of Fig. 105-1, and omitting the switch elements there connected to the grid and plate of the single tube.

Power supplies for plates and screens of boosters nearly always have a selenium half-wave rectifier, as shown in Fig. 105-1. In this diagram is shown a power transformer which completely insulates the line from the receiver parts. The three-position switch on the primary energizes this winding for high-band and low-band positions, but turns it off on the lowermost position for a direct connection from transmission line to receiver. There are many boosters in which one side of the power line is grounded to the chassis, and few in which the entire B-minus side of the circuit is run with insulated wire and in which the chassis connection is through a small capacitor, as in many transformerless radio receivers. In any case there is required a step-down transformer for heater voltage and current.

**PROBE, DETECTOR.** — A detector probe consists of a detector or demodulator built into a small metallic shield which is on

the end of a flexible shielded cable. From the probe shield extends an insulated contact point which may be touched to parts of r-f, i-f, and oscillator circuits which are carrying the signal voltages to be demodulated and observed. Signal modulation or a rectified signal from the probe detector is carried through the cable to an indicator. The indicator may be only a sensitive d-c meter or headphones, but usually is an oscilloscope, an electronic voltmeter, or the a-f amplifier of a signal tracer or any other a-f amplifier feeding a loud speaker or output meter.

The detector probe and indicator permit observation or measurement of signal modulation without change of connections or excessive loading of the high-frequency circuit from which the signal is taken. The detector used in a probe most often is a germanium crystal diode or other crystal diode such as described in the article on *Crystals, Rectifying*. A diode or triode tube detector may be used, with voltages for plate and filament or heater supplied through the connecting cable.

One of the simplest detector probe circuits is shown at the left in Fig. 106-1. The insulated probe tip or contact connects through capacitor  $C_a$  and resistor  $R_a$  to a spring clip. The clip

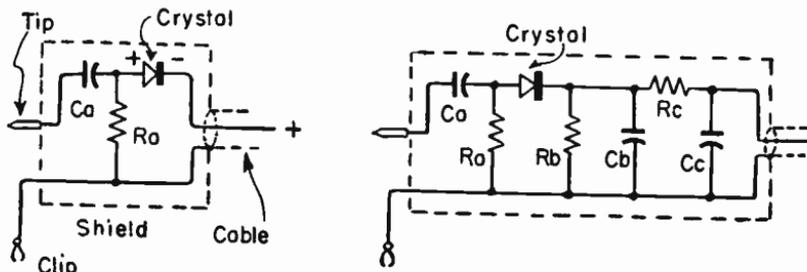


Fig. 106-1.—Detector probes with crystal diode in series between input and output.

is attached to chassis ground or B-minus or other low side point of the circuit from which a signal is to be taken, and the tip is touched to any point on the high side. The high side point might be the grid or plate of a tube. The crystal diode is connected between the junction of  $C_a$  and  $R_a$ , and the central conductor of the shielded cable. With crystal polarity

as marked on the diagram, potential to the cable conductor will be positive with reference to the shield. Reversing the crystal will reverse the output polarity, making the cable conductor negative. The crystal may be reversed to change the polarity of the trace on an oscilloscope indicator, or to make the trace "right side up."

Some or all of the additional capacitors and resistors shown by the right-hand diagram may be used in a probe. Values suitable for use when the indicator is either an oscilloscope or an electronic voltmeter are as follows:

- |                              |                                |
|------------------------------|--------------------------------|
| <i>Ca.</i> 0.001 mf          | <i>Ra.</i> 3,300 ohms          |
| <i>Cb.</i> 0.0001 mf         | <i>Rb.</i> 3,300 ohms          |
| <i>Cc.</i> 0.001 mf or less. | <i>Rc.</i> 1,000 ohms or more. |

Fig. 106-2 is a detector probe circuit often recommended for use with electronic voltmeters. Suitable values for capacitors and resistors are as follows:

- |                      |                        |
|----------------------|------------------------|
| <i>Ca.</i> 0.0005 mf | <i>Ra.</i> 1 megohm    |
| <i>Cb.</i> 0.01 mf   | <i>Rb.</i> 10,000 ohms |

All probe capacitors should be of the ceramic or mica type. The input capacitor *Ca* is subjected to whatever d-c potential

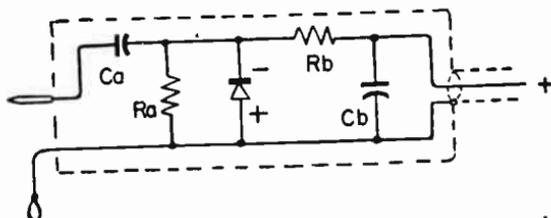


Fig. 106-2.—A detector probe with the crystal shunted across the line.

may be in the measured circuit. This may be very high when plate circuits are measured. Consequently, the d-c working voltage of this capacitor must be enough to withstand the highest plate voltage with which the probe may be used. Capacitances listed as suitable at this position are amply large for measured frequencies as low as 100 kilocycles. For voltmeter measurement of audio frequencies down to 60 cycles this capacitance should be at least 0.25 mf. Large capacitance sometimes is placed in parallel with capacitor *Ca* for low-frequency work.

All probe resistors should be non-inductive carbon types, never wire-wound types. Increasing the resistance at  $R_a$  increases the voltage output from the probe, but will distort the waveform observed on an oscilloscope.

Filter capacitors,  $C_b$  and  $C_c$  in the diagrams, bypass radio frequencies and allow oscilloscope traces free from fuzziness. The filter capacitor shown connected directly from cable conductor to ground usually is unnecessary, its purpose being served by capacitance of the shielded cable. Filter resistor  $R_b$  in Fig. 106-1 must be used in order to have well defined oscilloscope traces. Increasing this resistance raises the detector output for use with an electronic voltmeter.

**Probe Performance.** — Capacitor  $C_a$  of the diagrams is charged through the detector crystal and discharges through resistor  $R_a$ . The longer is the time constant of  $C_a$  and  $R_a$  the more nearly the average capacitor charge will approach the peak value of applied signal voltage. Periods between charging pulses from the signal voltage increase as signal frequency drops, so this effect alone would cause higher charge voltage at higher frequencies. At frequencies higher than about 10 mc there are likely to be resonances of capacitances and lead inductances of the probe, which will cause peaking of probe output voltage at the resonant frequencies. These effects which tend to raise the output voltage are more or less counteracted by all the usual energy losses which occur at high frequencies.

Probe construction must be such as to minimize stray capacitances between parts, and to reduce inductances by using the shortest possible connections between parts if performance is to be satisfactory at high frequencies. Connection to B-minus or ground at the measured circuit must be with a clip connected at the probe, not through a separate conductor leading to the indicator. Such a conductor extends the high-frequency circuit all the way to the indicator.

A probe is connected across any measured high-frequency circuit as in Fig. 106-3. Across the measured circuit in the left-hand diagram is the capacitance at  $C_a$  and a parallel com-

bination of resistor  $R_a$  in one leg and the crystal and resistor  $R_b$  in the other leg. Across the measured circuit at the right is the capacitance at  $C_a$  and a parallel combination of the crystal and resistor  $R_a$ . Also across the measured circuits are all the stray capacitances of the probe, represented by a capacitor in broken lines.

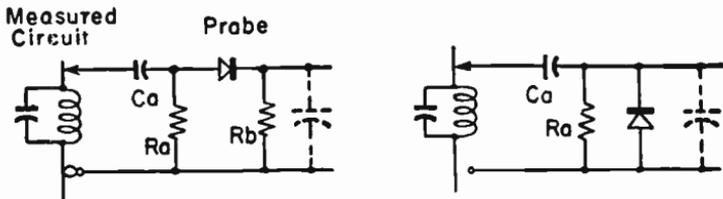


Fig. 106-3.—Detector probes may both load and detune a measured circuit.

The measured circuit is loaded by the impedance resulting from capacitances and resistances of the probe. The smaller the capacitances, including strays, and the greater the resistances, the less will be the loading and the more nearly the measured circuit will maintain normal behavior.

The measured circuit, if a tuned type, will be detuned by probe capacitances. This detuning may completely alter the performance in high-frequency amplifiers and will change the frequency or prevent oscillation in high-frequency oscillators. Often it is necessary to retune the measured circuit so that it may function with the probe connected.

Calibration of electronic voltmeters is difficult when used with a detector probe. The probe diode is a half-wave rectifier, so readings will change if connections are reversed to a measured circuit in which is an unsymmetrical waveform. Voltage readings will vary with frequency except through a rather narrow range. Harmonics are measured along with fundamental frequency voltages, and cause change of indications. Substituting a different detector or capacitor will require recalibration.

When using a probe its shield housing should be kept as far as possible from parts of measured circuits. The clip for B-minus or ground should be connected as close as possible

to the active portion of the measured circuit. If there is a choice, the operating frequency of the measured circuit should be as low as possible. It is better to work with low television channels than with high ones, and better to make measurements in i-f circuits rather than r-f circuits if the desired information can thus be obtained. Measurements are made preferably on low-impedance circuits, since then the impedance of the probe will form a smaller load and will have relatively little effect on circuit performance.

**PROBE, FILTER.** — A filter probe is a capacitor-resistor filter enclosed within a small metallic shield and connected through a shielded cable to the vertical input of an oscilloscope as shown by Fig. 107-1. A filter probe is used when making observations on circuits wherein there are frequencies higher

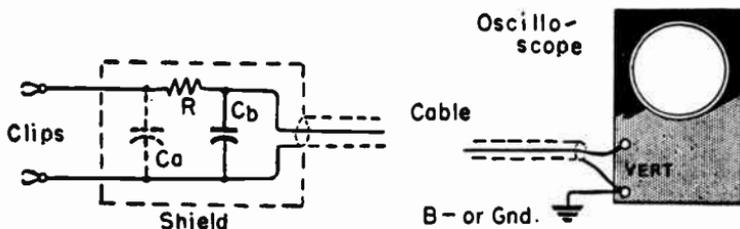


Fig. 107-1.—Filter probe for connection to an oscilloscope.

than those to be shown on the oscilloscope. Otherwise the high frequencies will cause fuzziness of the trace. Resistance at  $R$  usually is 10,000 ohms, but may be between 5,000 and 20,000 ohms.

The capacitor at  $C_a$  usually is omitted and capacitance at  $C_b$  is made 0.001 mf. If both capacitors are used, each may be 0.001 to 0.002 mf. Two capacitors provide more than twice the effective filtering action, and are useful where unwanted high-frequency voltages are strong. A filter probe is a virtual necessity when taking traces from the load of a television video detector.

**PROBE, FREQUENCY COMPENSATING.** — A frequency compensating probe, called also a low-capacitance probe, is used on the vertical input of an oscilloscope to isolate tested circuits from capacitances of the connecting cable and oscilloscope. A shielded cable ordinarily has capacitance of 50 to 75 mmf between inner conductor and shield. This capacitance, plus input capacitance of the oscilloscope, causes frequency discrimination because reactance varies with frequency.

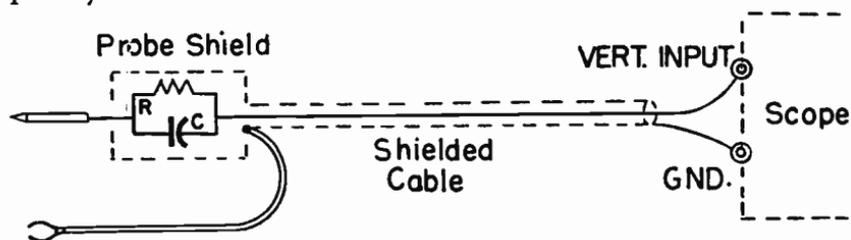


Fig. 107-2.—A low-capacitance probe for lessening frequency discrimination.

As shown by Fig. 107-2, a compensating probe contains a paralleled resistance and capacitance. Capacitance  $C$  may be about 5 mmf, and resistance  $R$  several megohms. The product of probe resistance and capacitance should equal the product of oscilloscope input resistance and combined capacitances of shielded cable and oscilloscope. Attenuation in the probe then will be nearly uniform over a wide frequency range, while impedance shunted across measured circuits is so high as to cause but little loading. A compensating probe is of practical benefit only with an oscilloscope having a frequency compensated vertical attenuator.

Resistance and capacitive reactance in the probe lessen vertical sensitivity, commonly to about 1/10 of the oscilloscope sensitivity without the probe. In order to allow matching of resistances and capacitances the small capacitor in the probe may be adjustable.

**PROBE, HIGH VOLTAGE.** — A high voltage probe extends the range of an electronic voltmeter or a sensitive moving coil voltmeter to 30,000 or more d-c volts. The probe is used chiefly for measuring voltage at the second anodes of pic-

ture tubes. A high voltage probe contains resistance of several hundred megohms which, in series with internal d-c resistance of the meter, acts as a multiplier resistance.

Resistance required in the probe is determined thus: Divide the desired maximum volts on the extended range by full-scale volts on the meter range to be used for high-voltage measurements. Multiply this quotient by internal resistance in megohms of the meter when used on the range to be employed. From this product subtract the internal resistance of the meter. The difference is required probe resistance in megohms.

Example: To extend the 1000-volt range to 30,000 volts with a meter whose internal resistance is 11 megohms on the 1000-volt range. Computation is as follows:

$$30000 \div 1000 = 30 \qquad 30 \times 11 = 330$$

$$330 - 11 = 319 \text{ megohms, probe resistance.}$$

All readings on the 1000-volt range of the meter then would be multiplied by 30.

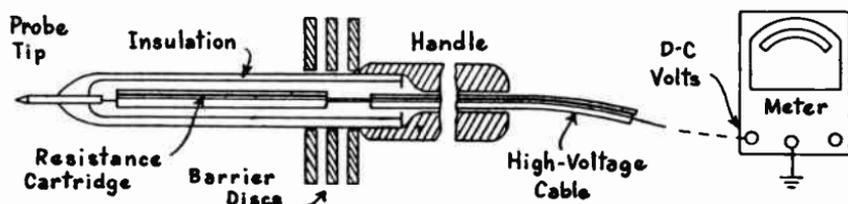


Fig. 107-3.—Principal parts of a high voltage probe.

Principal parts of a high voltage probe are illustrated by Fig. 107-3. The resistance cartridge is removable, when part of the insulating head is unscrewed, in order that cartridges of various values may be used in the same probe for meters of different internal resistances. Barrier discs extend leakage path length from probe tip to handle. The cable from probe to meter is insulated for voltage in excess of the highest measured. This cable may be shielded, with the shield connected to ground.

**PROBE, HIGH VOLTAGE WAVEFORM.** — A special type of probe is required on the vertical input of an oscilloscope for observation of waveforms whose peak-to-peak amplitudes are

in excess of a few hundred volts. This is because a capacitor internally in series with the vertical input might puncture, or other internal circuit components might be damaged when attempting to observe waveforms at horizontal output amplifiers and transformers, at dampers, and possibly at either horizontal or vertical deflecting yoke circuits.

It is usual practice to employ a capacitance voltage divider which applies only a fraction of receiver circuit voltage to the oscilloscope. The principle is shown at *A* of Fig. 107-4. The small capacitance has high reactance, and the large capacitance has small reactance. Most of the source voltage is dropped across the high reactance, and only the fraction across the small reactance goes to the oscilloscope.

One construction is shown at *B*. The small capacitance is plate-cathode capacitance, about two mmf, in a miniature half-wave high-voltage rectifier tube used without heating of its cathode. To lessen frequency discrimination, due to

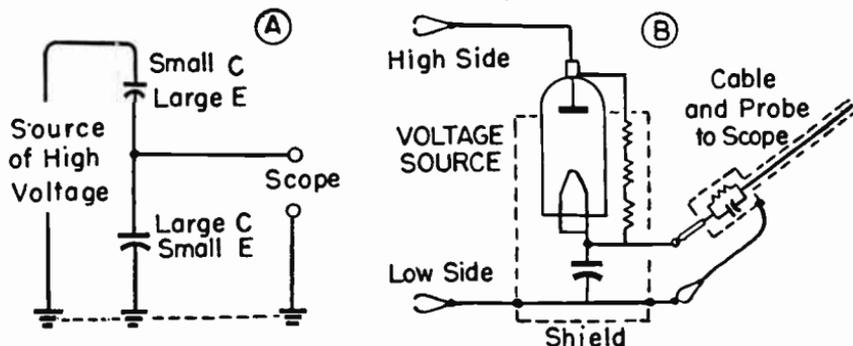


Fig. 107-4.—The principle of a capacitance voltage divider, and its application in a probe for waveform observation at high voltages.

variation of capacitive reactance with frequency, a compensating resistance may be connected across the rectifier tube. Waveform distortion in observed traces is further lessened by use of a frequency-compensating probe on the vertical input of the oscilloscope.

**PROPAGATION, WAVE.** — An electromagnetic wave, as employed for television and f-m radio transmission, consists of electric force acting alternately one way and the opposite in

a horizontal direction, and of magnetic force acting alternately up and down or vertically. Such a wave is represented by Fig. 108-1. The waves travel away from the transmitter and carry their energy in a direction at right angles to both electric and magnetic forces. Velocity through space is the same as that of light, 300,000,000 meters per second or about 186,000 miles per second.

Polarization of a wave refers to the direction of its electric lines of force. Horizontal polarization is used for television and f-m radio, and is shown by Fig. 108-1. Horizontal polarization requires the use of a horizontal antenna conductor at the receiver. Horizontal polarization, compared with vertical, allows a better ratio of signal to noise, since most electrical

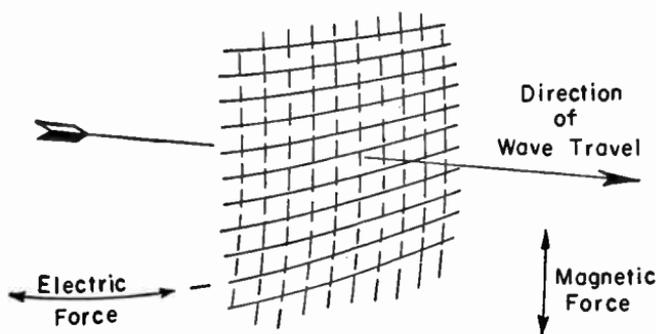


Fig. 108-1.—Forces in an electromagnetic radiation wave.

interference travels with vertical polarization. Horizontal polarization provides decided directional properties at the receiving antenna, whereas an antenna for vertical polarization is not directional. Waves horizontally polarized when leaving the transmitting antenna may become tilted slightly one way or the other, but this effect is slight in the short distances for television reception.

Electromagnetic waves at frequencies used for standard broadcast and medium short wave transmission are reflected back and forth between a layer of ionized gases high in the atmosphere, or in the ionosphere, and the moist earth and bodies of water. These successive reflections carry such waves

for long distances. Very-high and ultra-high frequency waves are not reflected from the ionosphere, and can be received in practice only so far as points which are in an unobstructed straight line or a "line of sight" from the transmitting antenna.

The maximum line of sight distance in which there is reliable reception is the sum of distances  $D_1$  and  $D_2$  of Fig. 108-2, these being the distances from transmitting and receiving antennas to their common horizon. Assuming uniform curvature of the earth's surface, and no high intercepting objects near the horizon, each of the distances may be computed from this formula.

$$\text{Miles} = \sqrt{2 \times \text{antenna height in feet}}$$

Any additional height which ordinarily is possible at the receiving antenna is too little to make much difference in the horizon distance. The chief purpose of added height of the

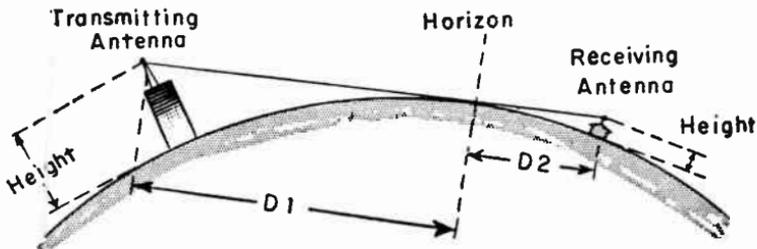


Fig. 108-2.—Factors which affect television and f-m radio reception distance.

receiving antenna is to get it above sources of interference and above nearby objects which might interrupt or reflect the radiation.

Occasionally there may be satisfactory reception at distances somewhat greater than the sum of those shown by Fig. 108-2, because of some refraction or bending of the electromagnetic waves at and near the horizon. With usual conditions and usual antenna heights there will be satisfactory reception over a range of no more than 20 to 30 miles. Unusually high transmitting antennas allow reception at great distances. With the approximately 6,000 foot elevation of transmitters on Mount

Wilson, near Los Angeles, the theoretical or computed horizon distance is nearly 110 miles, and as a rule there is good reception as far as San Diego, 116 miles away.

**Reflections and Ghosts.** — Radiation waves at very-high and ultra-high frequencies are reflected quite effectively from metallic objects and those consisting largely of metal and other good conductors. The strength or completeness of such reflections increases at higher frequencies or shorter wavelengths, because then the lines of force may more nearly complete their action toward one side or the other within the distance spanned by the reflecting object.

Both the direct and the reflected waves may reach a receiving antenna, as shown by Fig. 108-3. Although the reflected wave at the receiving antenna normally is weaker than the direct wave, the reflection still may act as a complete picture signal and produce on the screen of the picture tube an image of its own, in addition to the image formed by the direct wave.

The reflected wave, having traveled farther through space, will arrive at the receiving antenna slightly later than the direct wave. During the interval between arrivals of the two waves the beam in the picture tube will have traveled toward

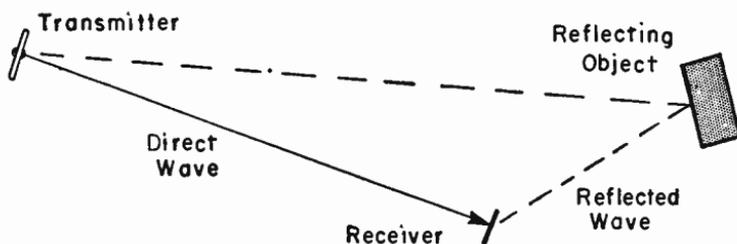


Fig. 108-3.—Wave reflection which may cause ghost images.

the right on the screen. Then the image formed by the reflected wave will appear slightly to the right of the image formed by the direct wave. The extra image due to the reflected wave usually is called a ghost. If the reflected wave has traveled 500 feet farther than the direct wave the ghost image will be displaced slightly more than  $\frac{1}{8}$  inch from the regular image on the face of a 16-inch picture tube. Actual separation

may be greater with longer travel of the reflected wave, or with short extra travel the two images may be so close as to cause only a blurring effect.

Wave reflections and resulting ghost images ordinarily are more troublesome where there are large buildings, bridges, large smoke stacks, gas tanks, and such objects. Usually there is little of this trouble in districts of single homes and in suburban areas generally.

If reflected and direct waves reach the receiving antenna at a considerable angle one to the other it usually is possible to rotate or orient the antenna to strengthen the direct signal and weaken the reflected one. If the two waves arrive from nearly opposite directions, as in Fig. 108-3, a reflector back of the antenna usually helps. Strength of the reflected wave may vary greatly within short distances, and moving the antenna may so weaken reception from the reflected wave as to remedy matters. An antenna director may so sharpen the directional effect of the antenna as to make the direct wave much stronger than the reflected one.

There are cases where the reflection is stronger than the direct wave, which may encounter obstructions not in the paths of the reflection. Then the antenna may be oriented for regular reception from the reflected wave. What otherwise would be a direct wave may be completely blocked by objects between transmitting and receiving antennas. Naturally, any object which reflects a wave does not allow the wave to pass through it, unless possibly with greatly reduced strength. Thus there often are shadow effects, where the receiving antenna is the electrical shadow of some interfering object.

**Q-FACTOR.** — The Q-factor or  $Q$  of an inductive or capacitive element or circuit is the ratio of its reactance to its high-frequency resistance, or is equal to reactance divided by high-frequency resistance. Inductive reactance and accompanying high-frequency resistance may be those of an inductor or of an entire circuit. Capacitive reactance and accompanying high-frequency resistance may be those of a capacitor or of an entire circuit. Energy losses which cause high-frequency resistance are treated in this book under *Losses, High-frequency*. Any construction which reduces such losses increases the  $Q$  of an element or circuit.

The letter  $Q$  refers to electrical Quality of an element or circuit. Where reactance is large in comparison with energy losses there is *high Q*, while large losses compared to reactance cause *low Q*. So-called low-loss construction means *high Q* construction.

Since most energy losses in a tuned circuit are in inductors rather than capacitors, the  $Q$  of such a circuit is affected chiefly by  $Q$  of its inductors. Inductive Q-factor undergoes little change with frequency, because both inductive reactance and high-frequency losses increase at about the same rate as frequency rises. Capacitive Q-factor decreases with rise of frequency, since capacitive reactance drops, but capacitors have inherently higher Q-factors and less energy loss than inductors.

The  $Q$  of good capacitors is high enough that moderate decrease with higher frequency will not greatly lower the  $Q$  of a resonant circuit. The  $Q$  of an inductor does not change a great deal with variation of frequency. Accordingly, to maintain fairly constant  $Q$  in tuned circuits, it is desirable to produce resonance by using large inductance and small capacitance rather than with small inductance and large capacitance.

**RASTER.** — A raster consists only of the luminous traces caused by horizontal and vertical deflection of the electron beam across the picture tube screen, without pictures. There will be a raster provided sweep oscillators and amplifiers operate, even though no picture, sync, or sound signals are present. The oscillators then sweep the beam at their free running frequencies, without synchronization, and produce a raster.

**REACTANCES.** — Capacitive reactances in ohms may be found from any of the following expressions, according to whether frequency is measured in cycles, kilocycles, or megacycles, and capacitance in microfarads (mf) or in micro-microfarads (mmf).

$$\frac{160\ 000}{\text{cycles} \times \text{mf}}$$

$$\frac{160}{\text{kilocycles} \times \text{mf}}$$

$$\frac{0.16}{\text{megacycles} \times \text{mf}}$$

$$6.28 \times \text{megacycles} \times \text{mf}$$

$$\frac{160\ 000\ 000\ 000}{\text{cycles} \times \text{mmf}}$$

$$\frac{160\ 000\ 000}{\text{kilocycles} \times \text{mmf}}$$

$$\frac{160\ 000}{\text{megacycles} \times \text{mmf}}$$

Inductive reactances in ohms may be found from one of the following expressions, in which there are various combinations of frequency measured in cycles, kilocycles, or megacycles, and of inductance measured in henrys, millihenrys, or microhenrys.

$$6.28 \times \text{cycles} \times \text{henrys}$$

$$\frac{\text{cycles} \times \text{millihenrys}}{160}$$

$$\frac{\text{cycles} \times \text{microhenrys}}{160\ 000}$$

$$6280 \times \text{kilocycles} \times \text{henrys}$$

$$6.28 \times \text{kilocycles} \times \text{millihenrys}$$

$$\frac{\text{kilocycles} \times \text{microhenrys}}{160}$$

$$6\ 280\ 000 \times \text{megacycles} \times \text{henrys}$$

$$6280 \times \text{megacycles} \times \text{millihenrys}$$

$$6.28 \times \text{megacycles} \times \text{microhenrys}$$

**RECTIFIERS.** — A rectifier delivers one-way current or direct current when supplied with alternating voltage. A diode vacuum tube in a half-wave rectifier circuit is shown by Fig. 110-1. Applied voltage makes the rectifier plate alternately positive and negative, and the cathode of opposite

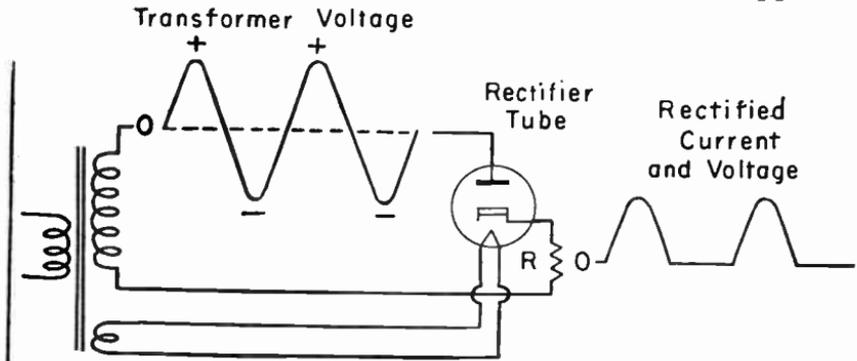


Fig. 110-1.—How alternating voltage causes one-way current in a half-wave rectifier employing a vacuum tube.

polarity. There is conduction only while plate and cathode are respectively positive and negative, not while the plate is negative and the cathode relatively positive. The result is that only the positive alternations or only half of each applied voltage wave causes pulses of current. Rectified current is one-way or direct, but is pulsating or intermittent and requires filtering to form continuous direct current.

A full-wave rectifier delivers one-way current pulses from both positive and negative alternations of applied voltage, in

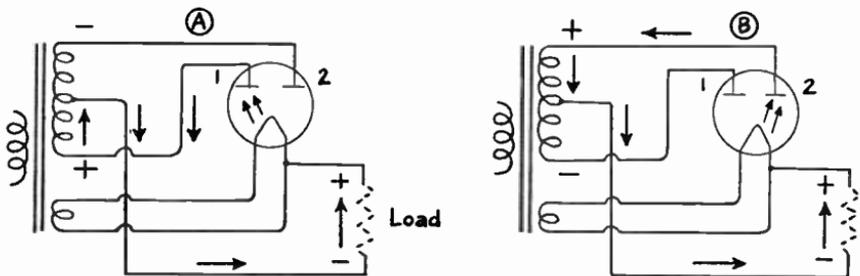


Fig. 110-2.—Current flow in a full-wave rectifier during opposite alternations of applied voltage.

the manner illustrated by Fig. 110-2. The rectifier tube has two plates. The single cathode may be either a filament type, as shown, or an indirectly heated type. The plates are connected to outer ends of a center-tapped secondary on the supply transformer. These ends become alternately positive and negative with reference to the center tap.

While transformer polarity is such as to make plate 1 positive, electron flow is in the path shown by arrows on diagram *A*, being upward through the load. When transformer secondary polarity reverses, as at *B*, plate 2 is made positive and electron flow follows the arrows. This flow is in the same direction through the load during both alternations of supply voltage, or during the full wave.

**Selenium Rectifiers.** — A single element of a selenium rectifier consists of an aluminum plate coated on one side with a thin layer of selenium, over which is another layer of some metal which distributes electron flow over the selenium. When alternating voltage is applied across the rectifier element there is fairly free flow of electrons from the outer coating to the selenium, but strong opposition to opposite flow. There is one-way conductivity, and current will flow in pulses of a single polarity.

In a complete rectifier unit there usually are several selenium elements or cells mounted between extended metal flanges which help carry away heat. A selenium rectifier has no heater, and requires no heater current or wiring.

A single selenium element will operate without overheating at a maximum of about 26 r-m-s or effective applied a-c volts. Then, for example, five elements side by side in a series connection will operate on 135 a-c volts and may be connected to a 117-volt a-c power line. How much current may be rectified depends on surface area of the elements. Ratings for assemblies of various sizes range from 65 to 450 ma.

A half-wave selenium rectifier circuit is shown by Fig. 110-3. The standard rectifier symbol is an arrowhead and bar, with the arrowhead representing the anode or "plate" and the

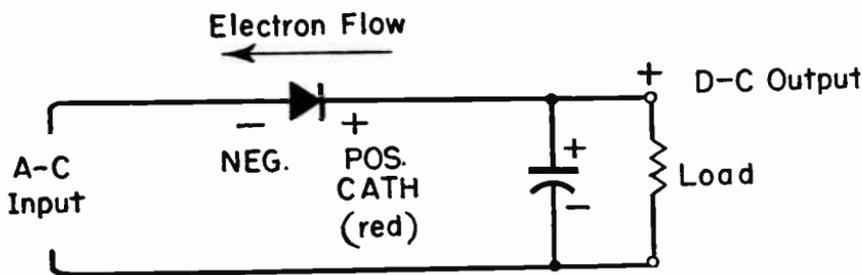


Fig. 110-3.—A selenium rectifier in a half-wave circuit.

bar representing the cathode. Electron flow is from cathode to anode or plate, as in a vacuum tube, consequently is opposite in direction to the direction in which the arrowhead points.

Two selenium rectifiers might be used with a transformer in a full-wave circuit such as shown by Fig. 110-2. However, it is more common practice, when using two rectifiers, to omit the transformer and use a voltage multiplier connection to provide either full-wave or half-wave rectification with approximate doubling of voltage from the alternating supply.

**REGULATION, VOLTAGE.** — There is decrease of terminal voltage at a source when there is increase of current in the source and a connected load. This is because more of the source emf is used to force increased current through internal resistance of the source, and less of the emf remains as voltage at the output. Automatic voltage regulation refers to various methods for maintaining nearly constant voltage across a connected load when there are changes of voltage at the source or changes of current in the load.

The basic principle of one general method for automatic voltage regulation is maintaining constant total current in and from the source, thereby maintaining constant voltage drop within the source and constant terminal voltage.

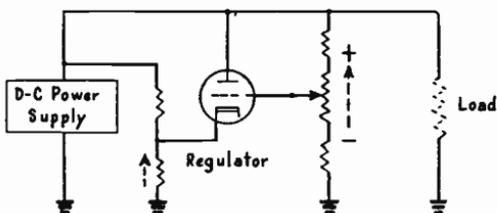


Fig. 111-1.—With one method of automatic voltage regulation there is constant current in the source while load current varies.

Fig. 111-1 illustrates one application of this principle. The regulator is a vacuum triode in parallel with a d-c power supply (the source) and a load represented as simple resistance. The regulator grid connects to the slider of a potentiometer in a resistor string across the power supply and load. The grid is at a potential positive with respect to ground. The cathode is connected to a positive potential of such value that the regulator grid may be biased negatively or slightly positive by moving the slider of the potentiometer. Change of grid bias alters internal plate-cathode resistance of the regulator tube.

Should there be increase of source and load voltage, more current would flow in the resistor string, there would be greater voltage across the resistors, and the regulator grid would become more positive with respect to the cathode. This would reduce internal resistance of the regulator, which then would take more current from the source and thereby bring source and load voltage down to the former value. Decrease of source and load voltage would have opposite effects; less current would be taken through the regulator and source, allowing source voltage to rise to its former value.

Another method of voltage regulation with vacuum tubes is illustrated by Fig. 111-2. The principle is that of increasing a resistance and thus interposing more voltage drop between power supply and load when supply voltages rises, as

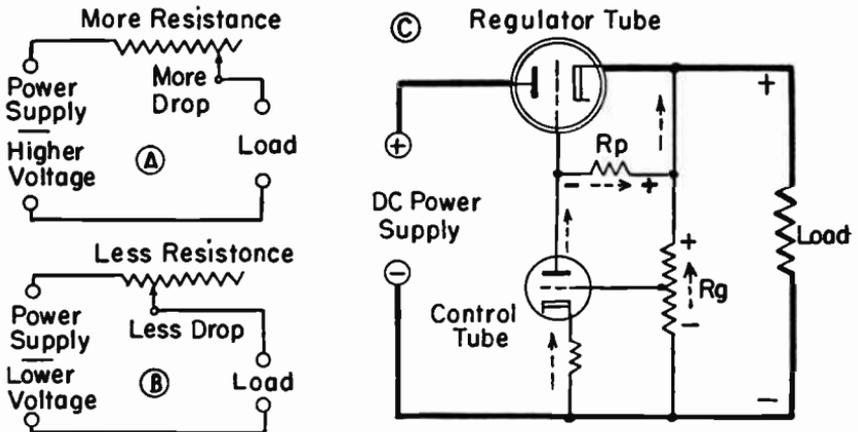


Fig. 111-2.—A method of automatic voltage regulation by variation of resistance in series between source and load.

in diagram *A*. Series resistance is decreased when supply voltage drops, as at *B*.

Circuit connections for such voltage regulation are shown by diagram *C*. The regulator tube acts as the variable series resistance between power supply and load, carrying the entire load current. Plate-to-cathode resistance of this tube is changed by altering its grid voltage with respect to its cathode. This grid voltage on the regulator is altered by changes of voltage across resistor  $R_p$ , which is in the plate circuit of a control tube. Plate current in the control tube, and in  $R_p$ , is altered by varying its grid voltage. The control tube grid is connected to potentiometer  $R_g$  which is across the load whose voltage is to be regulated. Negative grid bias is provided by a cathode resistor on the control tube.

A slight increase of voltage across the load will increase current and voltage drop through  $R_g$ , thus making the control grid more positive, or actually less negative. Then there is more plate current in the control tube and resistor  $R_p$ . The resulting greater voltage across  $R_p$  makes the end of this

resistor connected to the regulator grid more negative than before. The more negative regulator grid increases internal resistance of the regulator, there is greater voltage drop through this tube, and less of the supply voltage remains at the load. This counteracts the increase of load voltage which started the regulating action. A slight decrease of voltage across the load reverses the entire process. Then internal resistance of the regulator tube is decreased, it interposes less resistance and less voltage drop, and there is correction of load voltage.

**Glow Tube Voltage Regulators.**—A tube having only a cathode and anode as elements, with a small amount of argon or neon gas in an otherwise evacuated bulb, has the property of greatly changing its internal resistance with small changes of applied direct voltage. This occurs when there is ionization of gas in the tube. During ionization there is visible glow in the tube, hence the name glow tube. Other names are gaseous voltage regulator and VR tube.

Each type of glow tube regulator is designed to operate at some specified average voltage, with which the tube has certain effective internal resistance; for example, 10,000 ohms. Were potential difference across a fixed resistor of this value increased by two volts, current in the resistance would increase only 0.2 ma. But the same two-volt increase on the regulator tube causes such drop of effective internal resistance that the tube current increases by 14.0 ma. Effective resistance

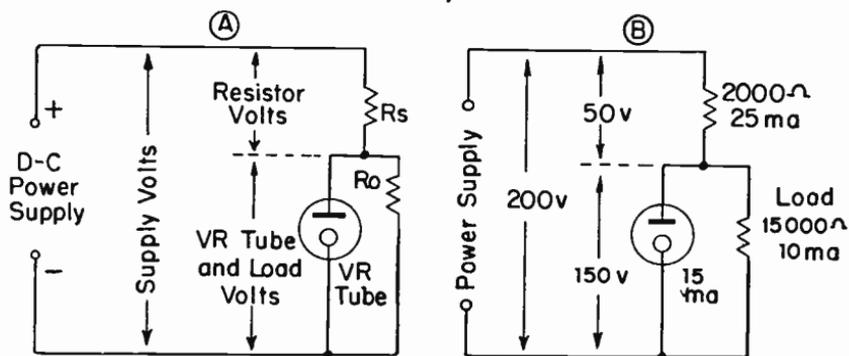


Fig. 111-3.—Automatic voltage regulation with a glow tube as the regulator

of the regulator drops from about 30,000 ohms to only little more than 5,000 ohms.

Gaseous voltage regulator (VR) tubes commonly are used in a circuit such as shown at *A* of Fig. 111-3. Resistor  $R_s$  limits maximum current in the VR tube and load  $R_o$  when power first is applied. Diagram *B* shows currents and voltages with a 15,000-ohm load and a VR tube designed to maintain 150 volts across itself and the load. A drop of 50 volts across the series resistor reduces the 200 volts at the power supply to 150 volts at the load and regulator tube.

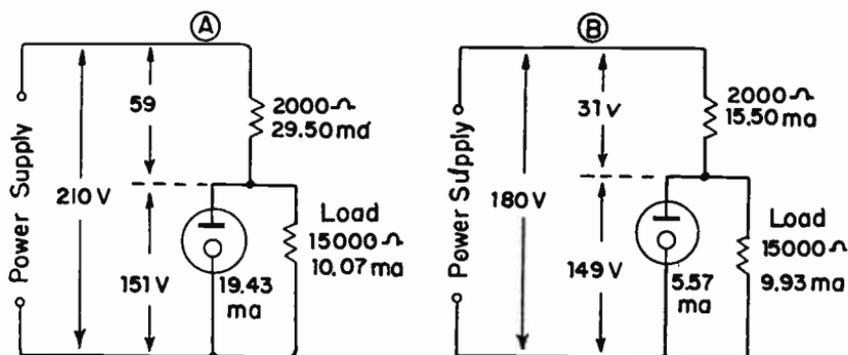


Fig. 111-4.—Performance of a glow tube regulator when there are fluctuations of power supply voltage.

Should power supply voltage increase to 210, as at *A* of Fig. 111-4, the regulator tube would draw more current through itself and the series resistor. By the time voltage across the regulator and load rises only one volt, voltage drop across the series resistor increases almost as much as supply voltage. Should supply voltage drop, as in diagram *B*, the regulator would take less current, causing less voltage drop across the series resistor. By the time voltage across regulator and load decreases one volt from the normal 150 volts, currents and voltages stabilize at values on the diagram.

Were power supply voltage to remain constant while load resistance dropped, more current through the load and series resistor would cause more voltage drop across that resistor

and leave less across regulator and load. Then the regulator would take much less current, compensating for increased load current, and allowing but little decrease of voltage on load and regulator. More load resistance would have opposite effects, with the load taking less current and the regulator more, thus limiting the change of voltage across regulator and load.

**RESISTORS, HIGH-FREQUENCY.** — Wire-wound resistors, because of their inductive effects, are not suited for use in circuits operating at frequencies above about 10 megacycles. All resistors, both wire-wound and carbon or composition, have a certain amount of capacitance between their terminals. Because of this the carbon types may act as capacitive reactances to allow signal leakages when frequencies go up around 100 megacycles. A unit whose resistance measures several megohms at direct current and low frequencies may have effective resistance of only a few thousand ohms at very-high frequencies. Troublesome effects sometimes are avoided by using several units in series, with total resistance of the desired value. Then the capacitances are in series to increase the capacitive reactance. With several resistors there is, however, some danger of greatly increasing the capacitance to ground.

**RESONANCE.** — The accompanying formulas relate to frequencies of resonance in megacycles and kilocycles, and to combinations of inductance in microhenrys and capacitance in micro-microfarads (mmf) for such frequencies. The formulas apply to circuits which are either series resonant or parallel resonant.

$$\text{Megacycles} = \frac{160}{\sqrt{\text{mmf} \times \text{microhenrys}}}$$

$$\text{Microhenrys} = \frac{25\ 330}{\text{mc}^2 \times \text{mmf}}$$

$$M\text{mf} = \frac{25\ 330}{\text{mc}^2 \times \text{microhenrys}}$$

$$\text{Kilocycles} = \frac{160\ 000}{\sqrt{\text{mmf} \times \text{microhenrys}}}$$

$$\text{Microhenrys} = \frac{25\ 330\ 000\ 000}{\text{kc}^2 \times \text{mmf}}$$

$$M\text{mf} = \frac{25\ 330\ 000\ 000}{\text{kc}^2 \times \text{microhenrys}}$$

A series resonant circuit subjected to frequencies lower than the resonant frequency acts like a capacitive reactance, or a capacitance and resistance in series. At frequencies above resonance the series resonant circuit acts like an inductive reactance, or an inductance and a resistance in series. Alternating potential differences across the inductance and/or the capacitance may be many times as great as the voltage applied in series with the resonant circuit when the frequency is that of resonance. The capacitors may have to withstand potential differences much greater than the applied voltage.

A parallel resonant circuit subjected to frequencies lower than the resonant frequency acts like an inductive reactance, while at frequencies above resonance this circuit acts like a capacitive reactance. There is the same potential difference across both capacitance and inductance, and it is no greater than voltage applied in series with the resonant circuit.

**RESTORATION, D-C.** — D-c restoration or d-c reinsertion is a means for automatically maintaining correct values of tone or shading in a television picture when there are changes of average brightness at the scene being televised. Correct tone values can exist only when the black level of the television signal is at the grid voltage for beam cutoff in the picture tube, and when control grid bias in the picture tube is such as to maintain this condition.

The term d-c restoration means restoration of the d-c component of the video signal when this component disappears as the signal passes through any coupling capacitors between the video detector and the picture tube. The original d-c component is identified in Fig. 113-1. At *A* is represented a video i-f signal modulated with sync pulses and with picture signal variations for a light toned picture. At *B* is represented a modulated signal for a dark toned picture. So long as carrier signal strength and video i-f signal strength remain constant, the tips of the sync pulses will remain at the same amplitude with reference to the zero voltage line whether the picture is light toned or dark.

When the i-f signals are demodulated by the video detector only the positive or the negative modulation envelope will remain. If the detector is so connected as to pass only the positive envelope, the detector output will consist of a direct voltage whose variations are the sync pulses and picture signals shown at *C* and *D*. The average potential of this detector

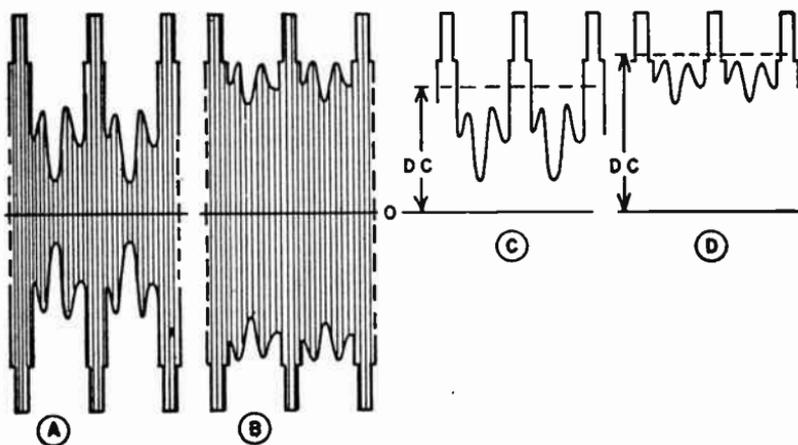


Fig. 113-1.—Modulated i-f signals and their d-c components for light toned and dark toned pictures.

output is the d-c component of the signal. For the light toned picture the d-c component is less than for the dark toned picture. In the final reproduction at the picture tube screen the amount of d-c component, or some equivalent effect, will determine whether the picture appears generally light or generally dark.

The signal voltages at *C* and *D* of Fig. 113-1 usually go through at least one coupling capacitor between the detector and a video amplifier, or between two video amplifiers, or between a video amplifier and the picture tube. Any capacitor will pass only the alternating component of the signal voltage, and on the output side of the capacitor there will be only an a-c signal voltage with no d-c component. This is shown by

Fig. 113-2. The alternating signal voltages for both light and dark toned pictures will be centered on their zero potentials, or will extend equally above and below zero.

The two alternating signal voltages which come through a capacitor are shown by Fig. 113-3 applied to the characteristic curve for control grid voltage and beam current of a picture



Fig. 113-2.—Picture signals and sync pulses represented as alternating potentials without d-c components.

tube. Here it is assumed that the brightness control of the receiver has been adjusted to correctly render the light toned picture, from signal *C*. The alternating signal voltage centers at the grid bias voltage, just as in an amplifier tube. The bias is such that sync pulses of the signal make the control grid more negative than for beam cutoff, and consequently these pulse voltages have no effect in the picture.

Signal *D*, for the dark toned picture, also centers on the bias voltage which is correctly adjusted for the light toned picture. In actual picture reproduction such changes of tone or average shading may come very rapidly. Nearly the entire sync pulse voltage for the dark toned signal now is above beam cutoff, whereas all of each sync pulse should be below the point of beam cutoff. The black level for this dark toned picture really should be at beam current cutoff, but it is far above cutoff. All the picture variations likewise are far above cutoff. The supposedly dark picture will be made of generally light tone.

Were picture tube grid bias made more negative, to bring the black level of the dark toned picture down to beam current cutoff, and were the light toned signal applied on this

more negative bias, there would be another fault. The darker portions of the picture would become wholly black, because they would drop below beam current cutoff. Also, the remaining portions of the picture would come down too low on the brightness scale, and the supposedly light toned picture would become dark. The result of losing the d-c component, and not

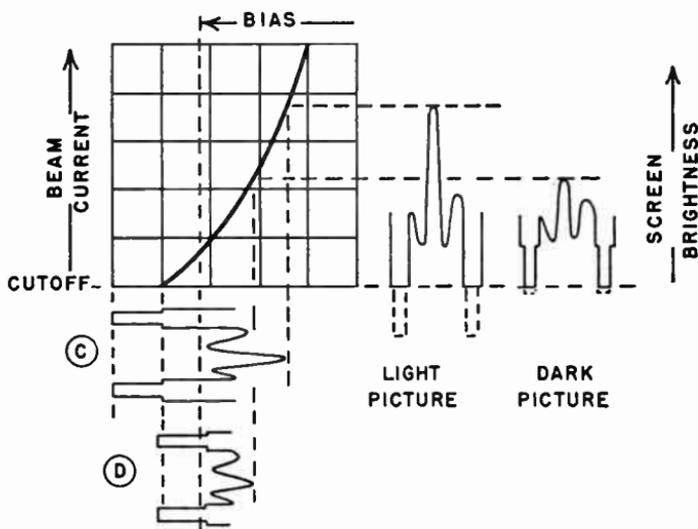


Fig. 113-3.—Signals for light and dark pictures reproduced with the same grid bias in the picture tube.

restoring it before the signal reaches the picture tube, is to make dark pictures too light and light pictures too dark. Contrast is greatly reduced.

For correct reproduction of pictures the black level of the signal voltage must be at the picture tube grid voltage which causes cutoff of the beam. Then sync pulses in the signal will make the grid still more negative, and these pulse voltages will be cut off. The black levels of the light and dark toned signals were at the same potential in the detector output, at the right in Fig. 113-1. They became of different potentials when losing their d-c components, as in Fig. 113-2. It is change of black level that causes difficulty.

In Fig. 113-4 the black levels for light and dark toned pictures are kept at equal potential by avoiding the use of coupling capacitors. The video detector is conductively connected to the control grid of the video amplifier tube. The positive tops of all sync pulses bring the control grid to the same voltage, and the black level in both signals likewise brings the

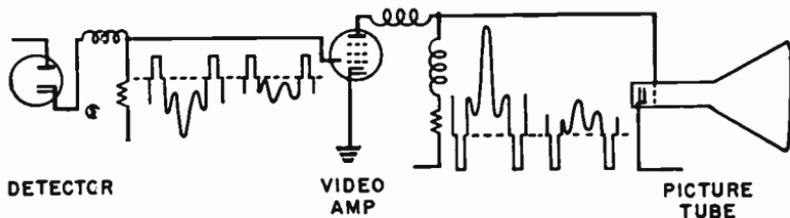


Fig. 113-4.—Direct coupling between detector and picture tube retains the d-c component of the signal.

control grid to equal voltage. The signal polarity is inverted by the amplifier. Variations of d-c plate voltage will be just like the variations of d-c detector voltage applied to the amplifier control grid, and in these plate voltage variations the two black levels will be at the same potential.

The plate of the video amplifier is conductively connected to the control grid of the picture tube. The variations of plate voltage, with their constant potential for the black levels, become the input signal voltage for the picture tube. When picture tube grid bias is adjusted to bring the black level of either signal to the point of beam current cutoff, the black level of the other signal will also be at cutoff.

Returning now to the unequal black level potentials shown in Fig. 113-3, it is apparent that both light and dark toned pictures could be correctly reproduced only by altering the control grid bias of the picture tube to compensate for every change of tone. For dark toned pictures the bias would have to be made more negative than shown. When tone again became lighter, the bias would have to be made less negative.

Picture tube grid bias can be changed by the manually operated brightness control, but not rapidly enough to com-

pensate for every change of picture tone and signal amplitude. Consequently, the bias must be changed automatically. This automatic change of bias must be brought about by the very change of signal amplitude which makes correction necessary.

Bias voltage is the difference of potential between grid and cathode of the picture tube. Bias may be altered in any way required by varying either the grid voltage or the cathode voltage, while average voltage of the other of these two elements remains fixed. Some d-c restoration systems act to alter the average grid voltage and others alter the cathode voltage. Whatever method is used to bring all black levels to the same potential it must follow the last coupling capacitor in the video detector and amplifier circuits. Black levels may be held at equal values as the signal passes from tube to tube, as in Fig. 113-4, but not when the signals go through a coupling capacitor.

**Amplifier Bias for Restoration.** — In Fig. 113-5 the d-c component of the signal which is present at the detector output is lost in the coupling capacitor between detector and first video amplifier grid. The resulting alternating signal voltage

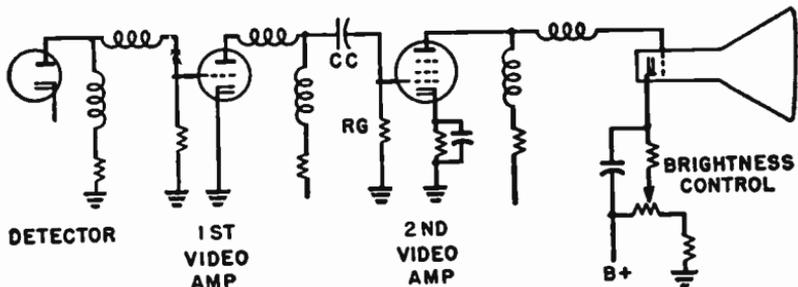


Fig. 113-5.—D-c restoration by means of grid-leak bias for a video amplifier tube.

is amplified and goes through another coupling capacitor,  $C_c$ , to the grid of the second video amplifier tube. Had the d-c component not been lost in the earlier capacitor it would disappear here.

By employing grid leak-capacitor bias for the second video amplifier the d-c component will be restored and will appear

In the plate circuit of this amplifier. The restored d-c component will not again be lost between the plate of the second amplifier and the control grid of the picture tube, because there is direct coupling between the amplifier and the picture tube.

Restoration is accomplished as illustrated in Fig. 113-6, where the alternating signal voltages for the light toned and dark toned pictures are shown applied to the mutual characteristic of an amplifier tube. With grid-leak bias the positive

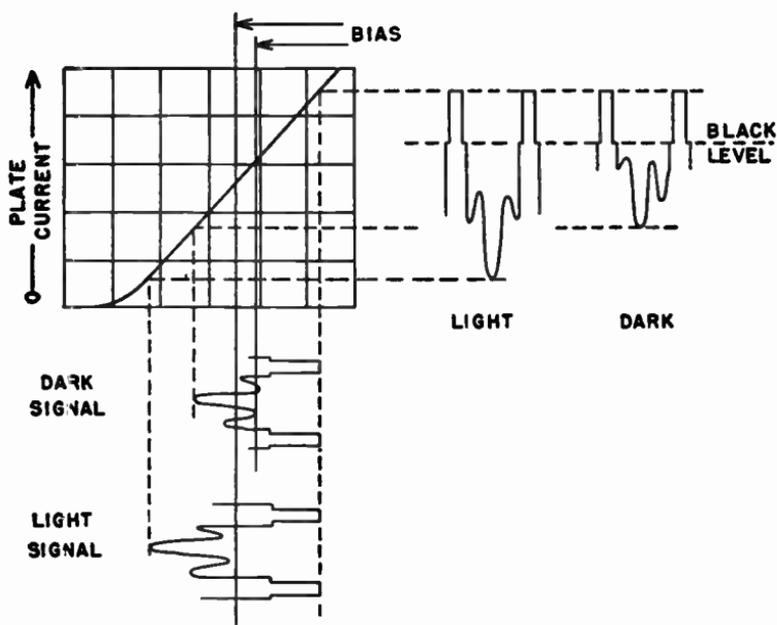


Fig. 113-6.—Action of grid-leak bias in restoring a d-c component.

alternations of the signals will add to the negative charge on the grid side of capacitor  $C_c$ . Grid bias will be made more negative because the charge leaks away only slowly through grid resistor  $R_g$ . The greater the signal amplitude the greater will be the charge added to the capacitor, and the more negative the bias will become.

The signal for the light toned picture has greater amplitude than the signal for the dark toned picture. Consequently, the light signal will make the amplifier grid bias more negative than will the dark signal, and the light signal will be applied lower than the dark signal on the characteristic curve. In the resulting amplifier plate currents, and in accompanying plate voltages, the tips of the sync pulses and also the black levels will be made equal for both kinds of pictures.

Amplifier plate signal voltages will be inverted in polarity with respect to amplifier grid signal voltages, making the sync pulses negative and the picture variations positive in the plate circuit. This inverted signal voltage will be applied to the control grid of the picture tube. When the receiver brightness control is adjusted to suit either the light or dark toned picture it will be correct for the other one, because both black levels have been brought to the same potential.

**Diodes for Restoration.**—When the last video amplifier tube is coupled to the picture tube through a capacitor, the d-c component of the signal must be restored in the picture tube grid-cathode circuit which follows the capacitor. The method frequently used employs a diode in any of several circuit arrangements. Always the picture tube control grid is made less negative with reference to its cathode when there are light toned pictures, with their greater signal amplitude.

One of the simplest diode restoration circuits is shown by Fig. 113-7. Picture tube grid return to the cathode is through resistors  $R_a$  and  $R_b$ , ground, and the brightness control resistors. This or some equivalent grid return would be necessary even were there no restoration. Potential drop in the brightness control resistors makes the cathode of the picture tube positive with reference to ground, and since the grid is connected to ground through  $R_a$  and  $R_b$  the grid will be negative with reference to the cathode. With the grid thus negative there would be no current and no potential difference in  $R_a$  and  $R_b$  were it not for action of the diode, as will be explained.

The diode cathode is connected to the grid side of the circuit. Consequently, the cathode is made negative with reference to the plate by the negative sync pulses of the grid signal, and the diode will conduct during these negative alternations of the signal. The diode will not conduct during positive

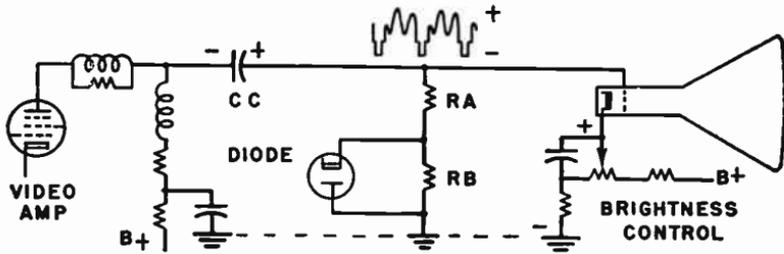


Fig. 113-7.—D-c restoration by means of a diode when signal input is to the picture tube grid.

alternations, because these make the diode cathode positive, or its plate negative. When the diode conducts during negative alternations it forms what amounts to a short circuit for resistor  $Rb$ .

Resistance at  $Rb$  is far greater than at  $Ra$ , usually 100 times or more. Then, during negative signal alternations when  $Rb$  is shorted by diode conduction, capacitor  $Cc$  charges quickly through the small resistance of  $Ra$ . Charge potential on the capacitor becomes nearly equal to the potential or amplitude of positive signal alternations, which are the sync pulses.

During positive alternations of the signal, when  $Rb$  is not shorted by the diode, capacitor  $Cc$  can discharge only slowly through the high resistance of  $Rb$  added to that of  $Ra$ . As a result, there appears on the diode side of capacitor  $Cc$  and at the top of resistor  $Ra$  a positive potential equal approximately to sync pulse amplitude. This positive potential on  $Cc$  and at the top of resistor  $Ra$  is applied to the picture tube grid. It counteracts more or less of the negative grid bias supplied from the brightness control.

Strength of this opposing voltage from the capacitor will be proportional to sync pulse amplitude and signal amplitude.

Since signal amplitude is greater for light toned pictures than for those of darker tone, picture tube grid bias is made less negative for light toned pictures. This is the action required for d-c restoration.

The restorer diode has been shown as a tube. Sometimes it is a germanium crystal diode. Restoration action is the same with one kind of diode as with the other.

The diode of Fig. 13-7 may be replaced with a triode which then is used as a combination restorer and sync takeoff tube. The grid of the triode is grounded and acts as a diode plate so far as restoration action is concerned, or the grid acts like the grounded diode plate in the diagram. The entire triode is an effective sync separator because it conducts only during the sync pulse alternations of the signal. Pulse voltage is taken from the triode plate to remaining tubes in the sync section of the receiver.

The principle of restoration just explained may be employed when signal input is to the cathode of the picture tube, as in Fig. 113-8. Here the diode plate is connected to the side

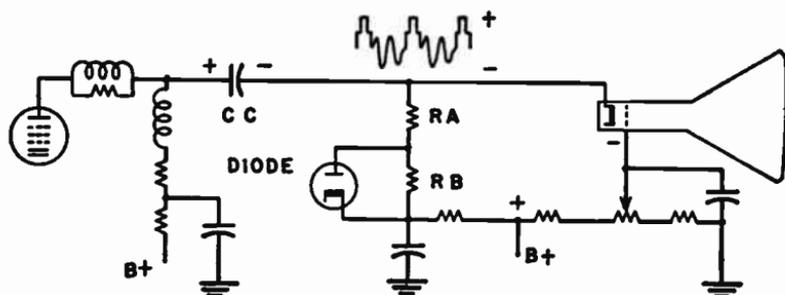


Fig. 113-8.—Diode method of d-c restoration when signal input is to the picture tube cathode.

of the circuit on which is the picture tube cathode. Sync pulses of the signal are positive in this side of the circuit, and by making the diode plate positive the pulse alternations cause conduction in the diode.

Now resistor  $R_b$  is short circuited by diode conduction during positive alternations which rapidly charge capacitor  $C_c$ .

The diode is non-conductive during negative alternations of the signal, and but little of the capacitor charge escapes during these alternations. Therefore, a negative charge is held on the side of capacitor  $C_c$  which is toward the picture tube cathode. This negative charge and its potential are proportional to signal amplitude. The greater amplitude of light toned picture signals increases the charge and the negative potential applied to the picture tube cathode.

This negative potential counteracts more or less of the positive cathode potential supplied from the B+ connection below resistor  $R_b$ . Signals for light toned pictures make the cathode more negative with reference to the grid of the picture tube, which is equivalent in effect to making the picture tube grid less negative with reference to the cathode. Thus there is the action required for restoration.

The accumulated charges on coupling capacitor  $C_c$  in Figs. 113-7 and 113-8 do not interfere with transfer of signal voltages to the grid or cathode of the picture tube. The charges which cause restoration form the average potential difference across the capacitor, this potential difference being the required d-c component for the signal. Signal variations which go to the picture tube grid or cathode are merely the instantaneous changes above and below the d-c component which exists at the capacitor.

In another diode restorer circuit, shown by Fig. 113-9, the capacitor charge which represents the d-c component is built up on a separate capacitor,  $C_r$ , instead of on the signal coupling capacitor  $C_c$ . Signal voltage for the restorer circuit is taken from across the amplifier load resistor  $R_o$ . Sync pulses are negative, just as in Fig. 113-7. This polarity is indicated by the fact that the amplifier plate is coupled to the picture tube control grid. Negative sync pulses or negative alternations of the signal act on the diode cathode to make the diode conductive. Electron flow then is as shown by broken-line arrows. This flow charges capacitor  $C_r$  in the polarity marked, then goes on through the temporarily conductive diode and resistor  $R_c$ .

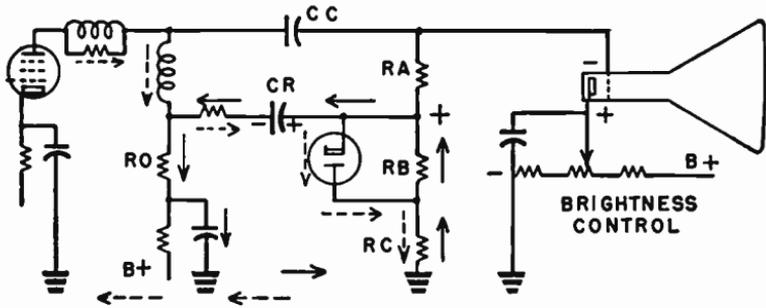


Fig. 113-9.—D-c restoration by means of a diode which allows charging of a separate capacitor for restoration potential.

During opposite alternations of signal voltage the polarity on the diode elements is reversed, and the diode is made non-conductive. Then capacitor  $C_r$  discharges slowly through the path shown by full-line arrows. Discharge is slow because it is forced by the non-conductive diode to go through the high resistance at  $R_b$ . The charge retained on capacitor  $C_r$ , and the potential at the top of  $R_b$ , become nearly equal to the peak amplitude of the signal which is intermittently charging the capacitor. This is the amplitude of the sync pulses.

Potential drop across the resistors in the brightness control makes the picture tube grid negative with reference to the cathode, because the grid is connected through resistors  $R_a$ ,  $R_b$  and  $R_c$ , and through ground to the negative end of the brightness control. The positive potential developed at the top of resistor  $R_b$  by diode action counteracts more or less of the negative grid bias on the picture tube. The positive potential from resistor  $R_b$  increases with greater signal amplitude, which means a picture of lighter tone. Thus the picture tube grid is made less negative for reproduction of light toned pictures. The only purpose of resistor  $R_a$  in this system is to transfer the biasing and restoration potentials to the picture tube grid. This resistor carries no currents.

**Black Level Controls.**— Diode systems of d-c restoration which have been discussed derive the potentials for their d-c components from the peak amplitude of sync pulses in the video signal. So long as signal strength remains constant at

the output of the last video amplifier, and at the input to the picture tube, the black level of the signals will hold a constant relation to sync tip potential. Then, with d-c restoration voltage derived from sync tip amplitudes, the black levels will be correctly maintained during reproduction of pictures which have varying tone or shading.

But if signal strength does vary at the output of the last video amplifier there will be proportional variation between potentials at the sync tips and the black levels. This is shown by Fig. 113-10. At *A* we have the same modulated i-f signal

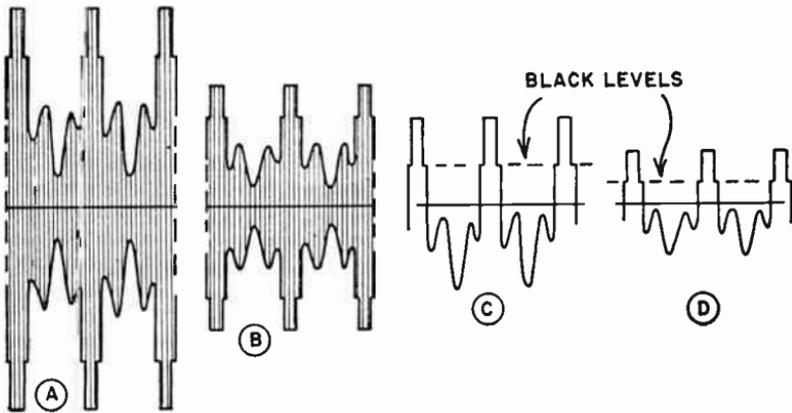


Fig. 113-10.—Black level potential or amplitude with reference to sync pulse tips is altered by change of signal strength.

as at *A* in Fig. 113-1. At *B* in Fig. 113-10 is shown a signal of exactly the same light tone, but of less strength. At *C* is the stronger signal after being demodulated by the video detector and passed through a coupling capacitor, which removes the d-c component. At *D* is the weaker signal after demodulation and removal of its d-c component.

Sync pulse tips now are at different potentials or amplitudes for the stronger and weaker signals. Because sync pulses are higher for the strong signal than for the weaker one, the two black levels would be at different potentials even with the pulse tips brought to the same potential by d-c restoration. In Fig. 113-11 the pulse tips of the stronger and weaker sig-

nals have been brought to the same potential, and the two signals are applied to the grid-voltage beam-current characteristic of a picture tube, just as the signals for light and dark toned pictures were applied in Fig. 113-3.

Resulting beam currents and picture effects are shown at the right-hand side of the picture tube characteristic. By means of the brightness control the picture tube grid bias has been

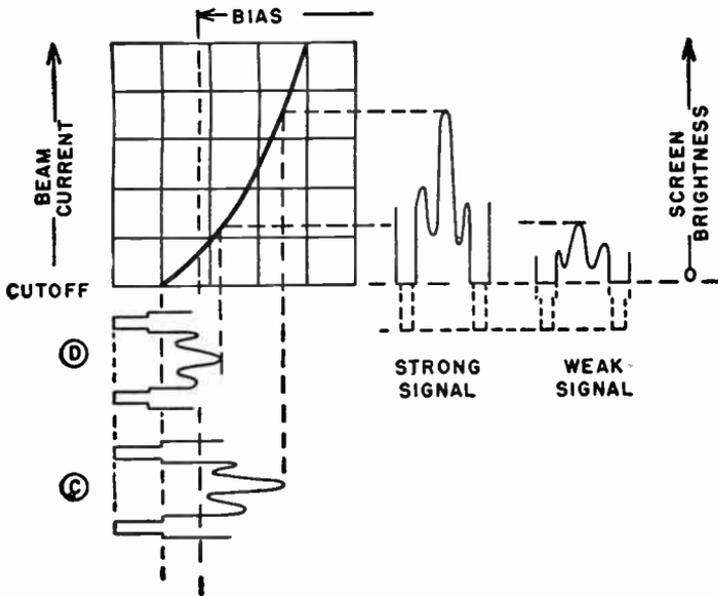


Fig. 113-11.—Effect on picture reproduction of equalizing the sync tip voltages for strong and weak signals.

made correct for the stronger signal, so that the black level of this signal is brought to beam cutoff. This adjustment brings the black level of the weaker signal beyond cutoff, and in the picture itself the average brightness is lower than it should be, while any portions of the picture which should have been dark gray will be made wholly black. Were brightness control of grid bias made correct for the weaker signal, the stronger one would produce a picture too bright.

The faulty reproduction illustrated by Fig. 113-11 is not too serious, because signal strength from a given transmitter seldom changes to any great extent during one program, and much of the change is compensated for by automatic gain controls. When switching from one channel to another the change of signal strength usually is enough to require readjustment of the receiver contrast (manual gain) control. If such changes are to be minimized or made unnecessary, the d-c restoration system should operate automatically from the black level potential rather than from sync tip potential. Systems which do this may be called black level controls, or by some equivalent name. There is further advantage in that adjustment of the contrast control to make any given picture more or less contrasty will not necessarily make it necessary to readjust the brightness control at the same time.

Black level controls of one general type act to remove the sync pulse voltages from the signal that charges the capacitor to the correct voltage of d-c component. The principle is illustrated in Fig. 113-12. The restorer diode and its connections may be generally similar to those shown in Figs. 113-7 or

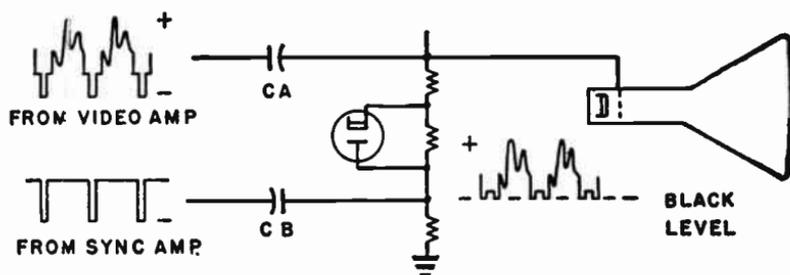


Fig. 113-12.—Black level control system in which sync pulses are removed to leave only the pedestals of the signal.

113-9. The complete signal from the video amplifier comes to the cathode of the diode through capacitor  $C_a$ . Since the sync pulse alternations of this signal are negative, the diode will be made conductive during each complete blanking interval within which a pulse occurs.

Through capacitor  $C_b$  there are applied simultaneously to the plate of the diode a succession of negative sync pulses obtained from some point in the sync section of the receiver. These negative sync pulses which act on the diode plate are stronger than pulses in the signal which is acting on the cathode. As a result, the diode plate is more negative than its cathode during each pulse period, and during these periods the diode is non-conductive.

In effect, the pulses are removed from the signal, as shown toward the right in the diagram. The negative peaks of the remaining portion of the signal are the porches or the tops of the pedestals. Potential at the tops of the pedestals is the black level of the signal. It is this signal, with sync pulses removed, which charges capacitor  $C_a$  to provide the d-c component.

Capacitor  $C_a$ , on which is produced the potential that partially counteracts negative grid bias, is charged only to a potential corresponding to the remaining black level of the signal. Thus the correction voltage which affects picture tube grid bias is based on the black level of the signal rather than on the sync pulse tips. Peak amplitude of the sync tips no longer determines the value of the d-c component, because the effect of these peaks has been removed.

**Restoration from Sync Tube.** — Fig. 113-13 illustrates a d-c restoration method with which the d-c component is obtained from a cathode bias resistor on a sync amplifier, separator, or limiter tube. The signal from the video amplifier is fed to the cathode of the picture tube, which requires that sync pulses be positive and picture variations negative.

A portion of the complete video signal is taken through capacitor  $C_r$  to the grid of the sync tube. This tube usually is a pentode, but it may be a triode. The sync tube is operated with cathode bias so negative and with voltages on the plate or on the screen and plate so low that the tube conducts only during the more positive portions of the sync pulses.

Increase of signal amplitude, as with a light toned picture or any generally stronger signal, increases the plate-cathode

current of the sync tube. This is accompanied by greater voltage drop across resistor  $Rk$  and greater charge voltage on capacitor  $Ck$ . Positive potential from the tops of  $Rk$  and  $Ck$  is applied directly to the picture tube grid, while the negative ends of this resistor and capacitor are connected through ground and the brightness control resistors to the cathode of the picture tube.

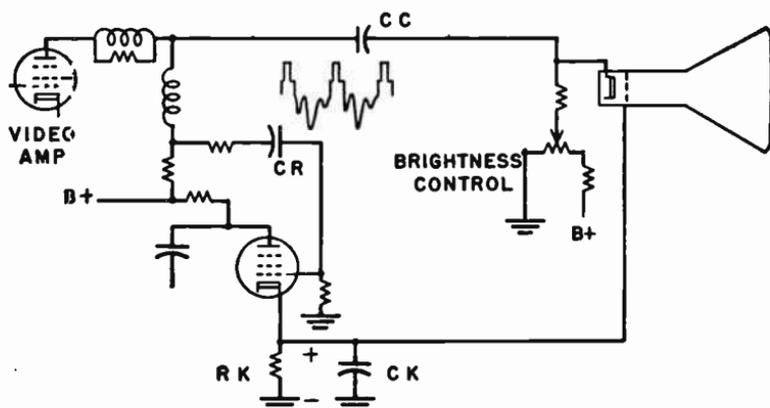


Fig. 113-13.—D-c restoration from cathode voltage of a tube in the sync section of the receiver.

Potential drop in the brightness control makes the picture tube grid negative with reference to its cathode. Voltage from across the cathode resistor and capacitor of the sync tube opposes the negative bias of the picture tube, as furnished by the brightness control. A light toned picture signal or any stronger signal increases the voltage from  $Rk$  and  $Ck$  to make the grid of the picture tube less negative with reference to its cathode. This is the action required for d-c restoration.

**Time Constant for Restoration.** — In any d-c restoration system the capacitor whose charge voltage fixes the value of d-c component is charged by sync pulse or pedestal top voltages which occur once for each line of the picture. There must be only negligible discharge of this capacitor between charges, or during one line period. On the other hand, the capacitor must not lose charge so slowly nor gain extra charge so slowly

that the d-c component cannot follow ordinary changes of picture brightness or of signal strength. The time required for charge voltage to change is proportional to the time constant of the capacitor and resistance in series with it.

The time constant usually is about 0.05 second, which would be the time for discharge to 36.8 per cent of the initial maximum voltage. During 0.05 second there are about 780 line periods, so discharge during one period would be entirely negligible. With this value of time constant the capacitor charge and voltage could drop to about half their initial value during one complete frame period of  $1/30$  second. Such discharge, or similar increase of charge, would allow the d-c component to follow any ordinary changes of picture tone or signal strength.

## S

**SCANNING.** — Scanning is the process of dividing the lights and shadows of the image of a televised scene into parts so small as individually to consist of only single tone values. These tone or shading values are transmitted one after another and reproduced one after another to reassemble the complete image at a receiver. Although scanning is primarily a separation of the image in the television camera tube, the word is used also to describe the building up of the reproduced image in the picture tube of a receiver.

Television is practicable because, during any one instant of time, the television camera is allowed to view, and the receiver picture tube to reproduce, only an image area so small that it can consist of only a single value of light or shade. The entire image or picture, whatever its overall size, is effectively divided into approximately 300,000 of these tiny areas.

All these areas making up a complete scene are viewed and reproduced within  $1/30$  second, a time so brief that during it there can be only a little relative movement between objects in the televised scene. Movement becomes apparent in the reproduced picture due to completed images following one another at intervals of only  $1/30$  second. Because of persis-

tence of vision, the human eye retains the impression of each completed image until the next one takes its place. Changes of position of objects in successive images appear as continual smooth movement. This is the principle employed also for motion pictures in theatres and the home.

Output voltage or signal voltage from the television camera tube is proportional to the intensity of light in only one small image area at a time. This is accomplished by forming the image in the camera tube on a surface which transforms light energy into electrical potential proportional to light intensity, but does so only where the surface is made light-sensitive by a stream of electrons striking it. An electron stream or electron beam in the camera tube is focused to fall on only the very small image area which has been mentioned. Consequently, signal voltage at any one instant of time is proportional to intensity of light on the small image area then being covered by the focused electron beam.

In the first step of the scanning process the electron beam in the camera tube is swept from left to right across the uppermost one of the narrow horizontal lines on the picture image as shown by Fig. 114-1. The picture in this illustration is divided into only a few of the several hundred horizontal lines of a television picture, in order to make the scanning process clearer.

Next the electron beam is dropped the width of two lines, whereupon it starts again from the left and sweeps the horizontal line which is third from the top. This continues until, within  $1/60$  second, the beam has swept over every alternate horizontal line from top to bottom of the image. Then the beam is returned to the top of the image and similarly sweeps all the intervening lines which were not covered during the first downward travel. The entire image, or all its lines, are scanned in two periods of  $1/60$  second each, or in a total of  $1/30$  second.

Images are reproduced on the screen of a picture tube in a receiver. The screen, on the inside of the rather flat face of the picture tube, contains a phosphor compound

which becomes momentarily luminous wherever struck by a beam of electrons moving at high velocity. Such an electron beam is swept from left to right across the picture tube screen, is dropped down and again swept horizontally, and continued in such motion until this picture tube electron



*Fig. 114-1.—A picture image is divided into many horizontal lines by scanning.*

beam has covered the entire screen surface on which pictures are formed. This happens while the electron beam in the camera tube is sweeping over the entire light-sensitive surface on which is the image of the televised scene.

The rate of electron flow in the picture tube beam is altered by the signal voltages which originate at the camera tube, and are transmitted. While the camera tube beam is traversing a light toned area of the original image, the transmitted signal voltage increases electron flow in the picture tube beam, and a proportionately light toned area is caused to appear on the screen. While the camera tube beam is on a dark area of the image, the resulting signal causes reduction of electron flow in the picture tube beam, and a proportionately dark area appears on the screen.

Travel of the camera tube beam is controlled by pulses of voltage produced and accurately timed at the transmitter. These synchronizing pulses are added to the transmitted signal. They maintain travel of the picture tube beam in precise timing or in synchronization with travel of the camera tube beam. Therefore, light and dark areas formed on the picture tube screen appear in the same relative positions as light and dark areas of the image in the camera tube. Thus the image on the picture tube screen becomes a duplicate of the image existing at the same time in the camera tube.

The manner in which the electron beam sweeps over the screen of the picture tube is shown by Fig. 114-2. Only enough of the horizontal lines are drawn to illustrate the principles involved. The beam commences its travel in the upper left-hand corner of diagram 1, at point *A*. The beam is swept to the left along the solid line path to point *B*, and along this line there appears a trace of successive light and dark areas in accordance with variations of picture signal voltage and of variations in light in the image at the camera tube. This completes one active line or one horizontal trace.

When the beam completes the luminous trace as far as point *B* the electron flow is stopped. This stoppage of electron flow is called blanking. The electric or magnetic forces which caused the beam to sweep from *A* to *B* continue acting, and were the beam not blanked, these forces would move it along the broken-line path to point *C*. The time during which the beam would be returned to the left side of the screen is the horizontal retrace period. The retrace occurs while the beam is blanked. Horizontal traces and retrace periods continue until the beam arrives at the lower right-hand corner of the picture space, at point *D*.

When the electron beam reaches the bottom of the picture space the beam again is blanked. It remains blanked while the deflecting forces would cause travel as shown by diagram 2. Were the beam not blanked it would follow a zig-zag path from point *D* upward through the picture space until reach-

ing point *E*. This point is only half way from left to right across the picture space. The time interval during which the beam would have been moved upward to the top of the picture space is a vertical blanking period. It may be called also a vertical retrace period.

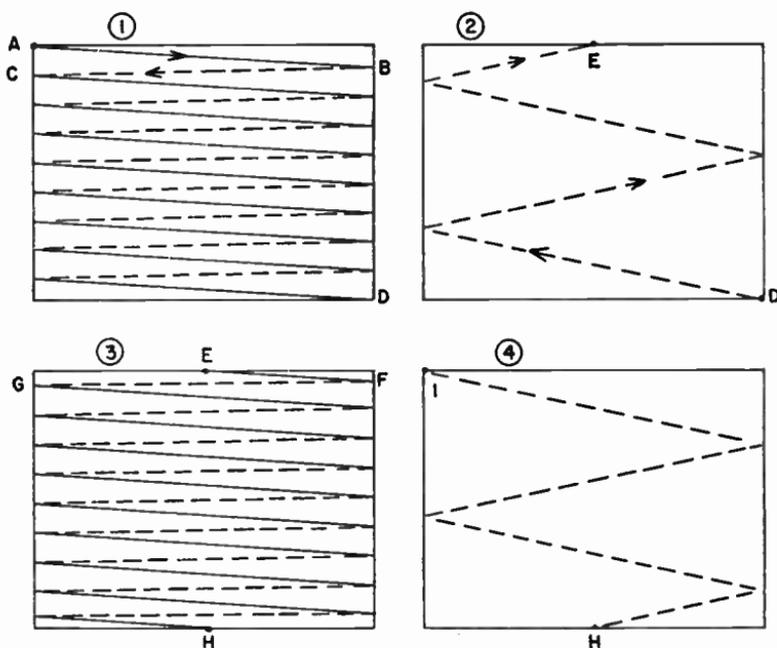


Fig. 114-2.—Travel of the electron beam over the screen in the picture tube.

The beam now is re-established or electron flow is allowed to resume at point *E* of diagram 3, which is the same as *E* on diagram 2. The beam now traces a half line from *E* to *F*. Next follows a horizontal retrace from *F* to *G*. This is followed by active traces or lines and intervening blanked retraces until the beam comes down to point *H* at the middle of a horizontal active line. Note that each active line or trace in diagram 3 is midway between traces of diagram 1.

At point *H* the beam once more is blanked, and the forces which would cause sweeping or deflection act as shown on diagram 4. During this second vertical retrace period the

beam would have arrived at point *I*. This is the same as point *A* in diagram 1, from where the entire sweeping action continues over and over again.

**Fields and Frames.** — One-half the total lines which form a complete picture are scanned during the action shown by diagram 1 of Fig. 114-2. This constitutes one field. The remaining half of the necessary lines are scanned during the action shown by diagram 3. This is the second field. Active lines of one field are midway between active lines of the other field. The two fields together are called one frame. A frame includes the scanning for one complete picture.

One field and the following vertical retrace are completed during  $1/60$  second. The field frequency thus becomes 60 cycles per second. One frame, consisting of two fields and two vertical retraces, is completed during  $1/30$  second. The frame frequency is 30 cycles per second.

During one complete frame period of  $1/30$  second there would be time for 525 horizontal active traces and the accompanying 525 horizontal retraces. The time required for one horizontal line and one horizontal retrace is a line period. With 525 line periods per frame and 30 frames per second there are 15,750 line periods per second, and the line frequency is 15,750 cycles per second.

Of the 525 line periods per frame 35 will be used during the two vertical retraces. It may be seen from Fig. 114-2 that each vertical retrace period will occupy the time for some number of whole lines plus a half line. There remain for the two fields a total of 490 line periods. In one complete picture there will be 490 luminous horizontal lines. This is true regardless of the height in inches of the picture. A large picture has the same number of luminous traces as a small one, but the traces or lines are farther apart in the large picture.

The image area actually scanned at the camera tube has a width to height ratio of 4 to 3. If a reproduced picture maintains proportions like those of the original image this picture will have the same ratio of 4 to 3 for width and

height. This is called the aspect ratio. The complete picture might be 4 inches wide and 3 inches high, or 8 inches wide and 6 inches high, or of any other dimensions with which width and height are in this ratio.

The system of scanning alternate lines of the complete picture during each field or each half-frame is called interlaced scanning. There are two principal reasons for using interlaced scanning rather than line by line scanning. First, there is less danger of noticeable flickering of the pictures. Alternate lines still are so close together that the impression on the eye is that of a complete picture in each field—this because of persistence of vision. Then with interlaced scanning there is the visual effect of 60 pictures per second instead of the actual 30 completely filled-in pictures per second. Even with high levels of screen illumination there is no visible flicker at a repetition rate of 60 per second, although there might be at 30 per second.

The second advantage of interlaced scanning is reduction of line frequency, and consequently of the frequency width of the band required for transmission. If 60 completely filled-in pictures were to be formed during each second, and were there still to be 525 line periods per complete picture, the line frequency would be 60 times 525, or would be 31,500 cycles per second instead of 15,750 cycles per second as with interlaced scanning.

**SENSITIVITY.**—Sensitivity of a television receiver usually is specified as the number of microvolts of r-f signal at the antenna terminals which will cause a one-volt signal strength at the video detector load. The fewer the required microvolts of r-f input the better is the sensitivity of the receiver. Sensitivity is checked with an accurately calibrated attenuator of a signal generator. Measurement is made while any automatic gain control is overridden or otherwise made inoperative, and with the manual contrast control set for maximum contrast or gain.

With receivers designed and constructed for good performance in localities of low signal field strength the sensitivity

for satisfactory reception from low-band channels 2 through 6 may be between 25 and 200 microvolts, and for high-band channels 7 through 13 it may be between 100 and 400 microvolts.

**SEPARATION, SYNC.** — The purpose of sync separation is to provide for the sweep oscillator circuits the necessary sync pulses without any of the picture variations which accompany the pulses in the complete television signal. The complete signal from which pulses are to be obtained may be taken from any of several points. It may come from the output of the last video i-f amplifier, from the output of the video detector, or from the output of any video amplifier.

Fig. 116-1 is a circuit diagram showing signal takeoff from the output of the last video i-f stage, through a diode. The takeoff diode ordinarily would be one section of a twin diode whose other section is the video detector. The takeoff diode acts as a rectifier or detector connected to deliver a demodulated signal in which sync pulses are positive and picture variations negative.

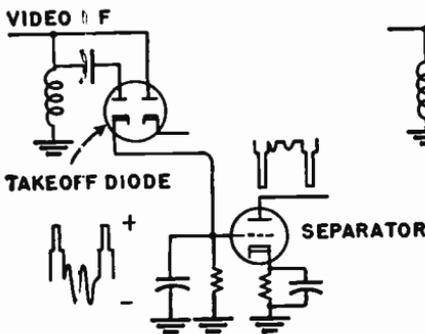


Fig. 116-1.—Sync takeoff from the final amplifier in the video i-f section.

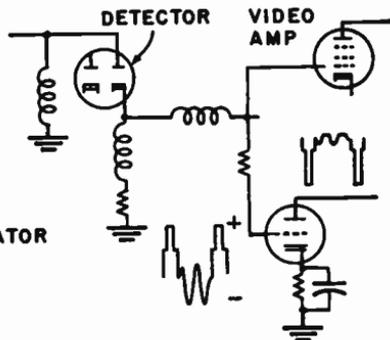


Fig. 116-2.—Sync takeoff from the output of the video detector.

The grid of the separator triode is negatively biased by the cathode resistor to a degree that cuts off most of the picture signal variations, which are negative at the grid. In some designs an additional negative biasing voltage is applied to the separator grid for more complete cutoff. In any case the tube passes positive sync pulses, and may amplify them to a greater or less extent, before the pulses go to following sync tubes or to an oscillator circuit.

In Fig. 116-2 the complete television signal is taken from the output of the video detector and applied through a resistor to the grid of a separator triode. Again the separator grid is biased sufficiently negative to partially cut off the picture variations of the applied signal, while passing and possibly amplifying the sync pulses. In this circuit and also in Fig. 116-1 the separator tube might be a pentode instead of the diode shown by the diagrams.

The outputs of the separator tubes in Figs. 116-1 and 116-2 retain some of the picture variations, which were not completely removed by cutoff action. One or more additional tubes will be used to completely remove all traces of picture variations and to amplify the sync pulses. The separator output usually would go to a sync amplifier which would strengthen the signal and make the sync pulses positive. The output of this amplifier would then go to a second separator with highly negative grid bias. This tube would pass only the sync pulses.

In the circuit illustrated by Fig. 116-3 the complete television signal is taken from the output of a video amplifier. In

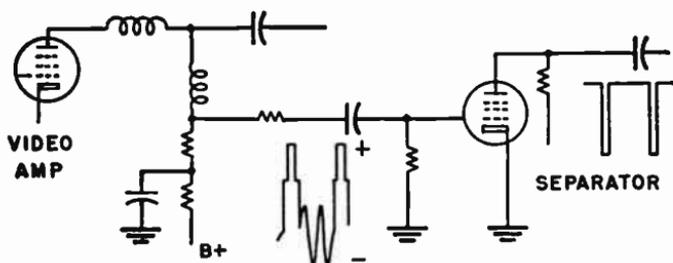


Fig. 116-3.—Sync separator connected to output of a video amplifier.

this output signal the sync pulses are positive and the picture variations negative. This might be the last video amplifier of a receiver wherein signal input is to the grid of the picture tube, or it might be the preceding video amplifier when signal input to the picture tube is to its cathode.

The signal from just above the load resistor of the video amplifier is taken through a resistor and capacitor to the grid

of the separator tube. The complete signal as taken from the output of a video amplifier has considerable overall amplitude and strong sync pulses. The grid bias of the separator may be so highly negative as to cut off all the picture variations and still have ample pulse voltage in its own output.

The process of sync pulse separation by means of a negatively biased grid is illustrated by Fig. 116-4. The bias is shown as being sufficiently negative to bring the point of plate current cutoff slightly above the black level voltage of the input signal. Bias is secured by the grid leak-capacitor method, which automatically makes the grid more negative upon increase of signal amplitude, and thus maintains the desired cutoff point even when there are decided changes of signal strength.

There may be supplementary fixed bias, with the grid resistor returned to some fixed negative potential, or there may be supplementary cathode bias. Supplementary bias which is

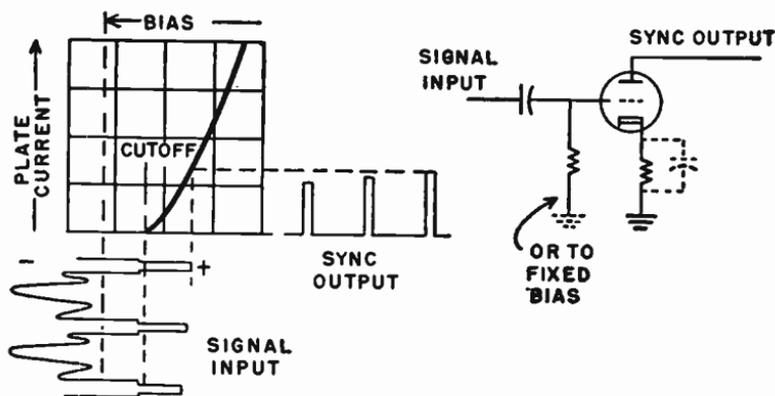


Fig. 116-4.—How sync pulses are retained and picture signals discarded by suitable biasing of a separator grid.

not wholly dependent on signal strength helps prevent formation of voltage pulses from noise interference effects accompanying the signal. Such pulses in the separator output might trigger the sweep oscillators and cause loss of synchronization.

The time constant of the grid-leak bias may be adjusted, by suitable choice of capacitance and resistance, to allow some

discharge of the capacitor between sync pulses. Then average bias voltage will be somewhat less than sync peak voltage, and cutoff may be maintained just above the black level, as in Fig. 116-4. The separator tube may be a voltage amplifying triode or one section of a twin triode, or it may be a sharp cutoff or semi-remote cutoff pentode. This tube often is operated with plate voltage or plate and screen voltages high enough to cause amplification of the sync pulses.

The function of sync separation often is combined with that of d-c restoration in a single tube. Fig. 116-5 shows a diode type restorer tube used also as a sync takeoff and partial separator. At the plate of the restorer diode the signal con-

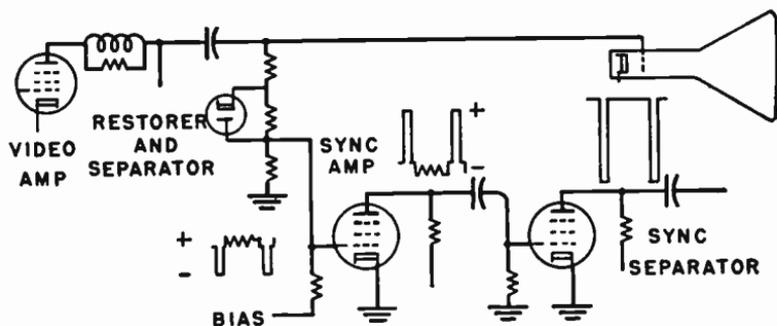


Fig. 116-5.—Sync takeoff from the plate of a d-c restoration diode.

sists of negative going sync pulses and of picture signal variations having relatively little strength. This signal from the restorer is applied to the grid of a sync amplifier. The amplifier strengthens and inverts the signal, making the sync pulses positive.

The output of the sync amplifier is applied to the grid of the sync separator tube. The grid of the separator is negatively biased to a degree which cuts off the remaining picture variations by the process illustrated in Fig. 116-4. Plate and screen voltages on the separator tube are high enough to make it also an amplifier for the sync pulses. Following tubes in the sync section may provide any necessary clipping or limiting, also inversion and amplification.

Fig. 116-6 shows how a triode used as a d-c restorer may serve also as a sync separator. Restoration action is carried out by the cathode & grounded grid of the triode acting as a diode restorer. Sync pulses at the cathode result in electron flow to the plate, where appear only the sync pulses of the signal in negative going polarity. These sync voltage pulses

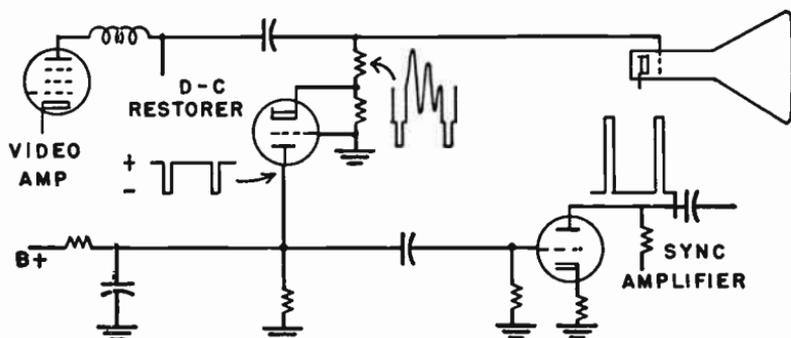


Fig. 116-6.—A triode used for d-c restoration and sync separation.

are applied to the grid of a sync amplifier, which may be a triode or a pentode. Following tubes may provide clipping, inversion, or additional amplification as required. In many receivers the output of the sync amplifier shown in the diagram goes to the sweep oscillators or to automatic sweep frequency controls through the usual vertical and horizontal filters.

**SHIELDING.** — Shielding consists of metal partitions or enclosures whose purpose is to protect circuits from electrostatic or magnetic fields produced by other circuits or parts. A shield is most effective when it surrounds or is near the part or circuit producing the field lines which are to be kept away from other parts. Well designed shielding is, however, effective when used at the part or circuit to be protected.

Electrostatic shielding is provided by metal even so thin as a coating or plating on a non-conductor. Field lines coming to the shield induce on the surface of the metal an electrostatic charge whose polarity is opposite to that of the body from which the field originates. This induced charge distributes

itself over the metal surface so that all electrostatic lines reaching the surface end on an opposite charge. Consequently, the field lines do not penetrate or pass through the shield.

Magnetic shielding at radio frequencies depends on induction by the magnetic field of eddy currents within the shield metal. These eddy currents produce magnetic fields of their own which oppose the field that induces the currents. The greater the conductivity of the shielding metal the stronger are the eddy currents and their fields, and the better is the shielding. Moderately thick shields of copper or aluminum are highly effective.

R-f magnetic shielding is preferably free from open spaces or joints which interfere with flow of the eddy currents, or else joints may be soldered to form a continuous conductor. Shields which are not formed by the main body of the chassis metal should be connected to this metal by soldered copper wires or straps, not only by contact pressure at screwed or bolted supports.

Magnetic shielding at audio and power frequencies depends on drawing the magnetic field lines into a shield of magnetic material and thus deflecting the lines away from or around parts to be protected. Iron and alloys of steel which have high permeability and are of considerable thickness are the most effective magnetic shields at low frequencies.

Any shield metal close enough to intercept the magnetic field of a coil will lessen the self-inductance of the coil and increase the tuned frequency in a circuit of which the coil is a part. The Q-factor of the tuned circuit is reduced by the decrease of inductance in relation to high-frequency resistance, also by increase of high-frequency resistance because of energy dissipated in the shield metal. Shielding, especially when close to current-carrying conductors, may cause a material increase of capacitance to ground or stray capacitance. This lowers the ratio of inductance to capacitance in tuned circuits, and may seriously reduce the gain or response at very-high-frequencies.

Metal tubes are shielded by their envelopes when a pin connected with the envelope is connected to ground. Some

glass-envelope tubes have internal shields connected to one of the base pins. This pin may be grounded to shield the tube. Tubes which are not self-shielded may be provided with a close fitting external shield which, when in position on the tube, is grounded by clips that hold the shield on chassis metal or other grounded metal. External shields are commonly used on r-f oscillator tubes in television receivers. When an oscillator has been aligned with a tube shield in place, removal of the shield may detune the circuit to a greater extent than can be compensated for with a fine tuning control.

Shielded wires and cables should have the shield soldered to the chassis or other grounding metal at both ends of the shield, and at any intermediate points where the wire needs support.

Television power transformers usually have a static or electrostatic shield between the primary and secondary windings. This shield consists of a band of thin copper or aluminum completely surrounding the primary but not joined together at its overlapping ends, in order that the band may not form a shorted single-turn conductor. The static shield usually is connected to the core, so that grounding the core grounds the shield. There may be an external lead coming from the shield, which allows making better connection to an external ground.

**SIGNAL, TELEVISION.** — The television signal consists of voltage variations which are utilized in the receiver for three purposes. First is the reproduction on the picture tube screen of lights and shadows forming the images. Second is blanking of the picture tube electron beam between picture lines, fields, and frames. Third is correct synchronization or timing of the horizontal and vertical sweep oscillators which control deflection of the beam in the picture tube. The television signal is carried between transmitter and receiver as amplitude modulation of video carrier waves whose frequencies are between 54 and 216 megacycles per second in the very-high frequency bands. Maximum frequency of signal modulation is in the neighborhood of 4 megacycles per second.

The television signal, as considered in this article, is the modulation of the video carrier wave. This signal modulation consists of voltage variations which remain after demodulation by the video detector, or consists of the output voltage of the video detector.

Sound which accompanies the television picture is transmitted as frequency modulation of an entirely separate sound carrier wave whose center frequency is  $4\frac{1}{2}$  megacycles higher than that of the video carrier for the same program.

Carrier frequencies and modulations for both picture signals and accompanying sound signals on the same program are transmitted within a 6-megacycle range called a television channel. As an example, channel number 10 includes the 6-megacycle range between 192 and 198 megacycles. Within this channel the video carrier modulation utilizes frequencies from 192.50 to 197.25 megacycles, and the sound carrier modulation utilizes frequencies from 197.725 to 197.775 megacycles. Additional information is given in the article on *Transmission*.

As explained in the article on *Scanning*, a picture or pattern on the screen of the picture tube in a receiver is made up of about 490 closely spaced luminous horizontal lines. Every second line, or every alternate line, is formed during a time period of  $1/60$  second, called one field. The intervening lines are formed during the following similar field period. There are 245 lines in each field. The two fields required for completing all the lines in one picture occur within a time period of  $1/30$  second, called one frame. The picture actually is transmitted, received, and reproduced one line at a time.

The interval between the start of one line and the start of the next line in a field averages 63.5 microseconds. Signal voltage during one line period may be represented as at the left in Fig. 118-1. This would be the voltage applied to the control grid of the picture tube with the cathode maintained at a constant potential. Polarity is such that increase of signal voltage in the positive direction makes the grid less negative, allows increase of current in the electron beam, and produces

a brighter portion of the luminous trace or a portion more nearly white. Signal voltage less positive allows the grid to become more negative with reference to the cathode, reduces beam current, and decreases the brightness. Minimum signal voltage shown here allows grid voltage to reach the value for

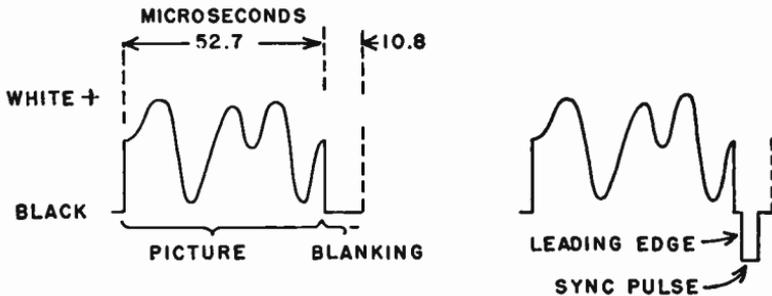


Fig. 118-1.—Television signal voltages during one line period.

cutoff of the electron beam in the picture tube. Then there is no luminous trace, or there is black in the picture.

One line of the picture or pattern is traced as the electron beam produces variations of brightness which are to exist along that particular line of the image. At the end of the line period the signal voltage goes to the value for beam cutoff, and there is blackness or blanking until time to start the next line.

Horizontal travel or sweep of the electron beam in the picture tube is caused by the horizontal sweep oscillator of the receiver. Action of the oscillator must be timed or synchronized to start each picture line at the same instant a corresponding line is started on the image in the camera tube. Oscillator action is timed in this manner by a pulse of voltage added to the signal during the blanking period between lines. This voltage pulse, shown at the right in Fig. 118-1, is called a horizontal sync pulse. The sweep oscillator is caused to commence one oscillation cycle by the sudden change of voltage at the beginning of this pulse, this being called the leading edge of the pulse. The horizontal sync pulse itself lasts for

between 5 and 6 microseconds, starting a little more than 1 microsecond after the beginning of the horizontal blanking period.

If the signal is to be applied to the cathode of the picture tube, with the control grid maintained at constant potential, the signal polarity is inverted as shown by Fig. 118-2. The signal may be obtained in either polarity from the video detector, depending on how the detector is connected to the preceding i-f amplifier. Changes of relative voltage between grid and cathode of the picture tube are the same with either polarity of the signal. There is the same reproduction of lights and shadows for a picture line, and the same blanking period at the end of a line.

The transmitted signal is of such form that sync tips of the modulation are at maximum amplitude on both sides of zero, or are at 100 per cent amplitude. Were the negative side of

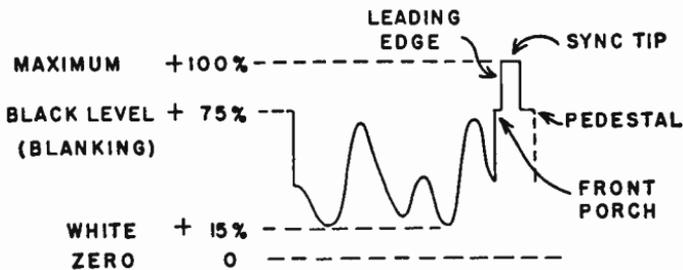


Fig. 118-2.—Voltage levels in a television signal.

the modulation recovered by the video detector, the resulting signal would appear as at the right in Fig. 118-1. Were the positive side recovered, the signal would appear as in Fig. 118-2, which is the usual way of showing the signal in diagrams.

With the tips of the sync pulses considered as 100 per cent signal voltage, blanking of the picture tube beam should occur with voltage equal to 75 per cent of maximum, plus or minus  $2\frac{1}{2}$  per cent. This 75 per cent of maximum signal voltage is called the black level of the signal. Voltage from top to bottom of the sync pulse is 25 per cent of total signal

voltage. This entire range of sync pulse voltage is beyond the black level, and may be called "blacker than black."

Purest white tones of the picture correspond to 15 per cent of maximum signal voltage. Various gray tones in the picture result from signal voltages between 15 and 75 per cent of maximum. The portion of the blanking voltage occurring before the sync pulse often is referred to as the front porch of the signal. The part of the voltage waveform on which the sync pulse appears to stand may be called the pedestal. Pedestal voltage is the same as black level voltage.

During reproduction of one picture field the signal will consist of 245 picture traces and horizontal blanking periods such as shown by Figs. 118-1 and 118-2. In each horizontal blanking period will be a horizontal sync pulse. At the end of the field there will commence a vertical blanking period during which occurs the vertical retrace action, as explained in the article on *Scanning*.

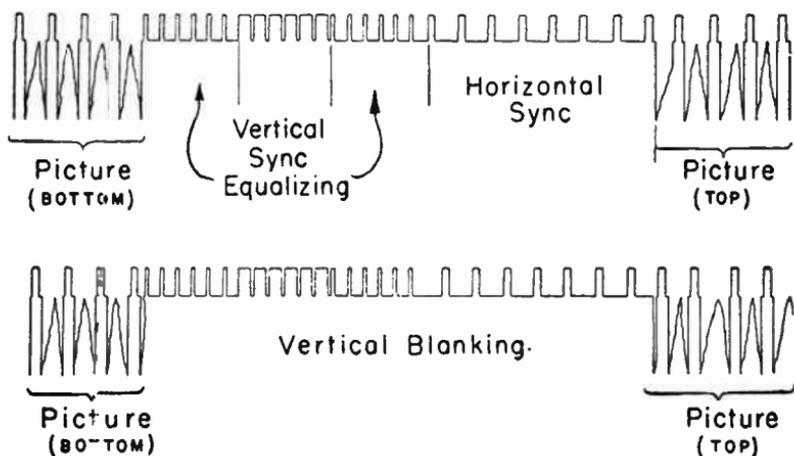


Fig. 118-3.—The television signal during the two vertical blanking periods of one frame.

Fig. 118-3 shows the signal during vertical blanking periods following each of the two fields which make up one frame. The vertical blanking period shown at the top is considered to begin with the picture tube beam at the bottom of the scanned

area, and shows at the left the last four picture signals, horizontal blanking periods, and horizontal sync pulses. Then occurs a vertical blanking period, followed at the right by the beginning of the next field, which will be at the top of the picture area. This second field ends with the horizontal lines, blanking, and pulses at the bottom of the picture area, as represented by the signal at the left on the lower part of the diagram. Then comes the second vertical blanking period, at the end of which the lines for another field are started as shown at the right in the diagram.

Note that during the entire vertical blanking period the signal voltage remains in the blacker than black region, so the electron beam remains cut off or blanked and no light appears on the screen of the picture tube. Obviously, if an increase of signal voltage as far as the black level cuts off the beam, any greater voltage in the same polarity will keep the beam cut off. Each vertical blanking period lasts for the equivalent of  $19\frac{1}{2}$  line periods, which is about 1,240 microseconds.

During each vertical blanking period there occur three kinds of synchronizing pulses. First there are six equalizing pulses spaced only half as far apart in time as the horizontal sync pulses, and each lasting only half as long as a horizontal sync pulse. Then come six parts of a long vertical synchronizing pulse in which each part lasts about five times as long as a horizontal sync pulse. Each separation or serration between the parts last somewhat less than the period of a horizontal sync pulse. The entire serrated vertical sync pulse lasts about 185 microseconds. Following the serrated vertical pulse are six more equalizing pulses like the earlier ones. The remainder of the vertical blanking period contains horizontal sync pulses like those between picture lines, and spaced at one line intervals.

The sync pulses are shown by themselves in Fig. 118-4. Examination of the time spacing between leading edges of pulses of all kinds shows that there is a leading edge of one pulse or another at every line interval all through the vertical blanking period. That is, the timing between leading edges of

horizontal sync pulses occurring between lines is continued in one way or another all through the vertical blanking. Thus the horizontal sweep oscillator is kept in synchronization throughout the vertical blanking period. Otherwise the horizontal sweep oscillator might get so far out of time that sync pulses which follow the vertical blanking and equalizing pulses could not pull this oscillator back into its original synchronization.

Equalizing pulses are spaced at half-line intervals in order that alternate picture fields may begin and end on either full lines or half lines, as shown in the article on *Scanning* to be necessary for interlaced scanning. The half-line spacing also allows synchronization of both horizontal and vertical sweep oscillators to remain uniform no matter which way a field begins or ends.

In Fig. 118-4 it is easy to see that the regular horizontal pulses in the top diagram occur at times half way between those in the bottom diagram. Then picture lines in one field

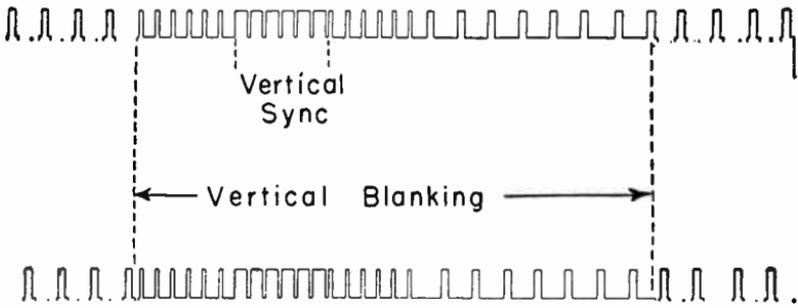


Fig. 118-4.—Sync pulses of the several kinds occurring during vertical blanking periods.

must likewise be half way between lines of the other field in a frame, as is necessary for interlaced scanning. Yet the beginning of the serrated vertical sync pulse is at the same point in both diagrams. This is necessary in order to have uniform timing of vertical retraces following both kinds of fields.

Were the regularly spaced horizontal sync pulses continued from the end of each field until the beginning of a vertical sync pulse, there would be a full-line interval preceding the vertical pulse at the top of Fig. 118-4, and only a half-line interval preceding the vertical pulse at the bottom. Then the capacitors which are to be charged by the vertical pulse would have more time for preliminary discharge with conditions at the top of the diagram than as down below. Timing of the vertical sweep oscillator would be slightly later in one case than in the other, and there would be defective interlacing. But with uniformly spaced equalizing pulses preceding both the vertical pulses, the capacitors will be equally discharged before charging commences, and timing will be uniform. These are the capacitors in the vertical integrating filter.

Half-line spacing of equalizing pulses also allows maintaining synchronization of the horizontal sweep oscillator regardless of whether a field ends with a full line or a half line. The explanation is as follows. In the upper diagram of Fig. 118-3 the final picture signal occupies a full line, while in the lower diagram it occupies only a half line. But either the first or second equalizing pulse always occurs at exactly one line period after the final horizontal sync pulse. Then timing of the horizontal sweep oscillator is picked up by an equalizing pulse after exactly one line period, whether the preceding field ends with a full line or with a half line.

The horizontal sweep oscillator is caused to commence an oscillation cycle only by an equalizing pulse occurring at full line intervals after preceding horizontal sync pulses. At the half-line points, where occur the alternate equalizing pulses, the oscillator is so far from being able to commence a cycle that it is insensitive to these alternate pulses.

The reason for breaking the total vertical sync pulse with serrations at half-line intervals is to provide leading edge voltage changes which maintain synchronization of the horizontal sweep oscillator throughout the vertical pulse periods. These leading edges after each serration in the vertical pulse act just like leading edges of equalizing pulses and horizontal sync

pulses. So far as controlling the vertical sweep oscillator is concerned, a long unbroken vertical pulse would be even better, but synchronization of the horizontal sweep oscillator might be lost.

**SIZE CONTROLS.** — In a television receiver there is a vertical size control or height control which alters the height of the scanned area on the picture tube screen, also a horizontal size control or width control which alters the width of the scanned area. To increase the size in either direction requires greater deflection of the picture tube electron beam in that direction. Consequently, all size controls act directly or indirectly to alter deflection voltages or fields in electrostatic deflection cathode-ray tubes, or alter deflection currents or magnetic fields for magnetic deflection tubes.

A size control used for either height or width, and found with both styles of picture tubes, is shown in elementary form by Fig. 119-1. The amount of control resistance brought into the circuit regulates the rate of electron flow for charging the

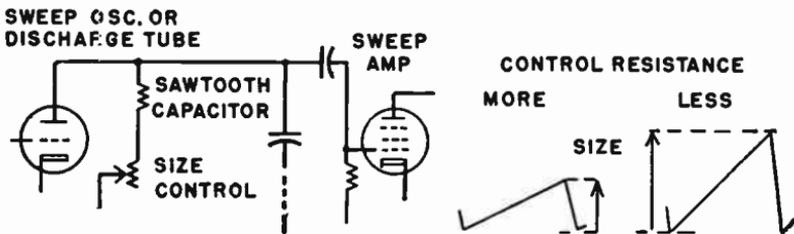


Fig. 119-1.—A size control which varies the charge and maximum voltage of the sawtooth capacitor.

sawtooth capacitor, consequently determines the voltage to which this capacitor is charged during the time in which the discharge tube or sweep oscillator remains non-conductive in the oscillation cycle. More control resistance decreases the rate of charging and the maximum charge voltage, while less resistance increases the charge rate and maximum voltage.

Changes of voltage on the sawtooth capacitor are applied to the control grid of the sweep amplifier. The greater the amplitude of grid voltage, the greater will be the deflection voltage

or current in sweep circuits connected to the amplifier output, and the greater will be the size of the picture. A smaller amplitude of grid voltage reduces the size of the picture.

In a few receivers the size control consists of an adjustable cathode-to-ground resistor on the sweep amplifier tube. Altering this resistance changes the grid bias and amplification of the sweep amplifier, and thus changes the deflection voltage or current in the output circuits. More bias resistance decreases picture size, while less resistance increases the size.

Another size control which alters the gain of the sweep amplifier tube acts by changing the screen voltage on this tube. The control unit is an adjustable resistor in series with the screen, between the screen and the B+ supply. Less resistance allows increase of screen voltage, which increases the transconductance of the tube, increases the deflection voltage or current in the output circuits, and increases the picture size. More control resistance has opposite effects. There also are size controls which vary the plate voltage on the sweep amplifier tube.

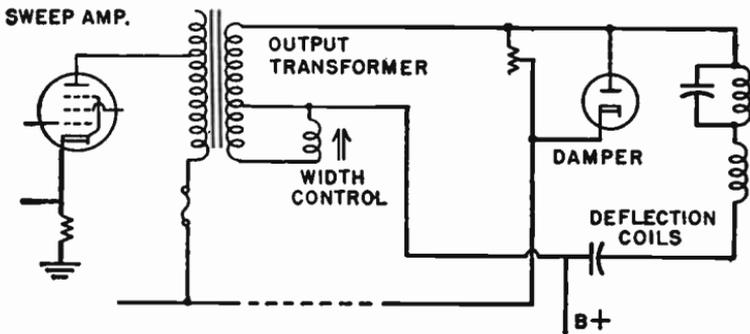


Fig. 119-2.—Width control inductor in parallel with part of the secondary on the horizontal output transformer.

Fig. 119-2 shows a method of width control used in connection with magnetic deflection picture tubes where the high-voltage supply is of the flyback or pulse operated type. The control unit is an adjustable inductor connected in parallel with part of the secondary winding on the output transformer.

Adjustment is by means of a movable core. Turning the core farther into the control unit winding increases the inductance of the unit and of the secondary winding. This increases the deflection current and makes the picture wider. Turning the core farther out of the inductor winding decreases picture width.

Still another method of width control for magnetic deflection picture tubes employs an aluminum sleeve that slides over the neck of the picture tube. One end of the sleeve goes through the focus coil and into the deflection yoke. Moving the sleeve farther into the yoke lessens the horizontal sweep and makes the picture narrower. Withdrawing the sleeve allows a wider picture. The aluminum sleeve acquires an electrostatic charge while the receiver is in operation, and should be discharged to ground before adjustment is made.

Controls for width and height should be adjusted to make the picture or pattern extend out in all directions to a small fraction of an inch beyond the edges of the mask, so that limits of the scanned area on the picture tube screen are not visible. There may be changes in size when receiving different stations, also when changing the settings of contrast and/or brightness controls. The size controls should be adjusted while the received picture or pattern is of smallest dimensions likely to result in actual reception.

Picture size is altered by adjustment of linearity controls, drive controls, and some peaking controls. When any of these other controls are changed it will be necessary to readjust the size controls at the same time.

**Size Magnifiers.** — Some television receivers are provided with a switch for increasing the height and width of the picture beyond the normal dimensions with which the entire picture is visible. The enlarged dimensions move the outer parts of the picture off the screen, while leaving the central portion magnified.

The magnifier switch is a ganged multi-pole single- or double-throw type. This switch may be mounted on the front panel of the receiver, or it may be constructed with a relay

mechanism whose magnet coil circuit extends through an external flexible cord to a push button which allows remote control of picture size.

**SOUND, TELEVISION.** — Voice, music, and sound effects which are to accompany the television picture are transmitted as a frequency-modulation signal on a sound carrier whose frequency is 4.5 megacycles higher than that of the video carrier in the same channel. In all television receivers the frequency-modulated sound carrier passes from the antenna and transmission line through the r-f amplifier and to the mixer. In the mixer tube the frequency modulation of the sound carrier beats with a fixed frequency introduced from the r-f oscillator. Among other frequencies in the output of the mixer there appears an intermediate frequency equal to the difference between the frequencies in the modulated sound carrier and the r-f oscillator. This sound intermediate frequency is frequency-modulated with the sound signals.

Between the output of the mixer tube and the audio frequency amplifier which feeds the speaker of the television set the sound signals may be handled by either of two general methods. One method is called the intercarrier sound system, which is described in the article, *Intercarrier Sound*. The other method may be called the dual-channel or split sound system, which will be described in this article.

There are articles in this book on *Detector*, *Ratio* and on *Discriminators*, both of which are used as sound signal demodulators. There is also an article on *Modulation, Frequency*. Principles of frequency modulation for television sound are the same as for f-m radio broadcasting and reception. There are some differences in application of principles. As an example, maximum deviation for television sound is 25 kilocycles instead of the 75-kilocycle limit in f-m broadcasting.

**Intercarrier Sound.** — With intercarrier sound the f-m sound signal modulation is taken all the way through the i-f amplifier section that handles also the amplitude-modulated video intermediate frequencies. Frequency difference between the i-f center frequency for sound and the video i-f carrier remains

at exactly 4.5 megacycles all the way to the video detector. This is the same difference existing between the r-f or space wave carriers. There are no traps for accompanying sound in the i-f amplifier section of a receiver employing intercarrier sound.

Since the output of the video detector is not linear, this detector acts as a mixer for the sound and video intermediate frequencies which have the 4.5 megacycle difference. In the detector output there is then a beat frequency which is the frequency-modulated sound signal with a center frequency of 4.5 megacycles. This f-m sound signal centered on 4.5 megacycles is taken from the output of the video detector, or from the output of a following video amplifier, and is carried to the i-f amplifier for sound. From this amplifier the f-m signal goes to the demodulator and thence to the audio amplifier.

**Dual-channel Sound.** — With dual-channel or separate channel sound systems the frequency-modulated i-f sound signal and the amplitude-modulated video i-f signal are not carried

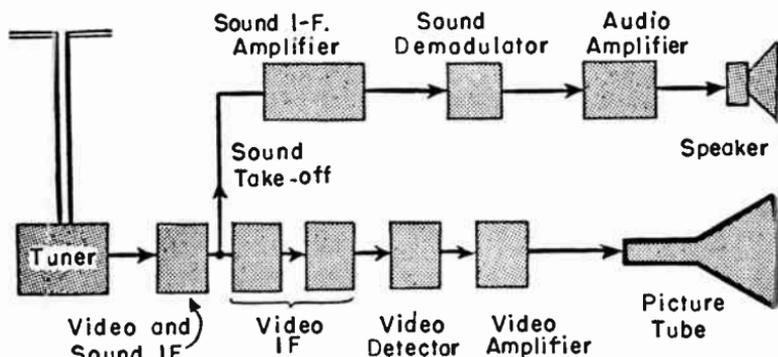


Fig. 120-1.—Scheme of connections for a dual-channel sound system.

together all the way through the i-f amplifier from mixer to video detector. Rather the sound i-f signal is separated from the video i-f signal either immediately following the mixer or at the output of one of the i-f amplifiers which precede the video detector. The principle is illustrated by Fig. 120-1. The

sound takeoff coupling is shown here at the output of the first i-f amplifier, but it might be at the output of the mixer in the tuner, or at the output of one of the following video i-f amplifiers. In a receiver employing a dual-channel sound system there may be traps for accompanying sound in one or more of the video i-f stages that follow the sound takeoff.

The intermediate frequency for sound is 4.5 megacycles lower than the video intermediate frequency for the same channel, whereas the sound carrier frequency is 4.5 megacycles higher than the video carrier frequency. This change occurs in the beating process at the mixer, because the two carrier frequencies are subtracted from the same r-f oscillator frequency.

The f-m sound signal from the takeoff coupling goes to an i-f amplifier designed to handle such a signal. Usually there are two or more sound i-f amplifier stages. If the sound demodulator is a discriminator, the last tube in the sound i-f section is operated as a limiter, as described in the article on *Limiters, F-m Sound*. Following the demodulator there is the usual audio amplifier and speaker.

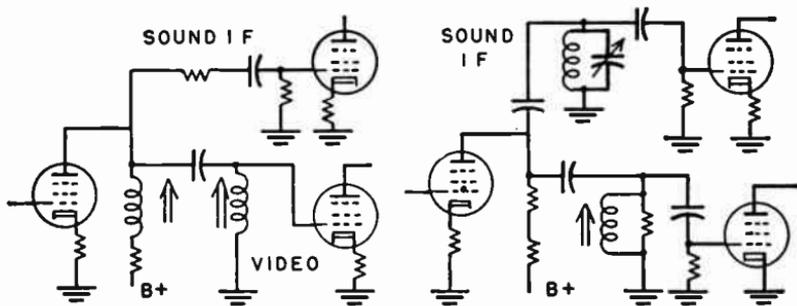


Fig. 120-2.—Sound takeoff connections for dual-channel sound systems.

Fig. 120-2 illustrates two of the several types of sound takeoffs found with dual-channel sound systems. The takeoff connection at the left is untuned and both intermediate frequencies would be carried to the first sound i-f amplifier, which is shown in the diagram. The plate circuit of this

amplifier, and following interstage couplings in the sound system, would be rather sharply tuned to the sound intermediate frequency and would amplify this frequency while rejecting the video intermediate frequency.

In the right-hand diagram there is a tuned impedance coupling in the grid circuit of the first sound i-f amplifier. Frequency response in the sound system need be no more than 300 kilocycles or 0.3 megacycle wide, and often is somewhat narrower. This provides ample coverage for the total deviation of 50 kilocycles above and below the center frequency, with allowance for some mistuning while retaining plenty of gain. Consequently, the sound takeoff couplings and the following interstage couplings in the sound system are tuned much more sharply than in the video i-f stages which handle the wide frequency range for television signals.

Fig. 120-3 shows sound takeoffs in which there are coupling transformers. In the left-hand diagram the primary of the takeoff transformer is the tuned plate coil for the mixer or the

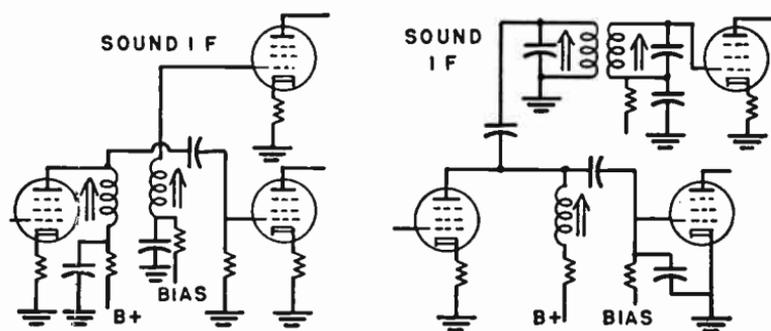


Fig. 120-3.—Sound takeoffs of the transformer type.

amplifier for both sound and video. The secondary is in the grid circuit of the first sound i-f amplifier. The plate winding is tuned for suitable response in the video intermediate frequency range, while the winding for the sound amplifier is tuned sharply to the sound intermediate frequency.

In the right-hand diagram there is a takeoff transformer whose primary winding is coupled to the plate circuit of the

mixer or first video-sound i-f amplifier, but there is a separate plate circuit coupling coil which is tuned for the video intermediate frequency range. Both the primary and the secondary of the takeoff transformer are tuned to the sound intermediate frequency.

Any tuned takeoff for the sound intermediate frequency acts also as a fairly effective trap for accompanying sound, so far as the remainder of the video i-f amplifier section is concerned. A takeoff sharply tuned to the sound intermediate frequency removes most of the energy at this frequency from the i-f amplifier circuit that is carrying both the video and the sound intermediate frequencies, leaving nearly all the video i-f energy but relatively little sound i-f energy to pass on through the remainder of the i-f amplifiers leading to the video detector.

**Alignment of Sound Section.** — The television sound section may be aligned in either of two general ways. One method employs an oscilloscope as output indicator, with input signals furnished by sweep and marker generators or a television generator combining both these functions. The other method employs as indicator a d-c voltmeter, either an electronic type or a moving coil type having sensitivity not less than 20,000 ohms per volt. Input signals then are furnished from an adjustable or a crystal controlled constant-frequency generator.

Either method may be used for dual-channel sound systems or for intercarrier sound systems, and either may be used where sound demodulators are of the discriminator type, ratio detector type, or other types. Instructions and precautions of a general nature are given in the article on *Alignment*.

With any method of alignment it is usual practice to connect the signal generator to the grid of the tube preceding the sound takeoff. With dual-channel sound systems this tube may be either the mixer or one of the video i-f amplifiers. With intercarrier sound systems this tube usually is the video amplifier ahead of the sound takeoff, although it would be the i-f amplifier preceding the video detector when intercarrier sound takeoff is from the output of the video detector.

If the output indicator is either an electronic voltmeter or a sensitive moving coil voltmeter the constant-frequency signal generator must be accurately tuned to the sound intermediate frequency used in a receiver having a dual-channel sound system, or to 4.5 megacycles for any intercarrier sound system. Crystal control of generator frequency or calibration against a crystal oscillator is desirable to insure accuracy. The generator is used without modulation, with pure r-f output.

If the output indicator is an oscilloscope the sweep generator is tuned for a center frequency which is approximately the sound intermediate frequency for a dual-channel sound system, or approximately 4.5 megacycles for intercarrier sound. The marker generator will be used to precisely identify this center frequency and other frequencies in the response curves.

**Connections for Alignment.** — General principles underlying all methods of alignment in the sound section are explained in the following paragraphs.

The output indicator, either voltmeter or oscilloscope, is connected to the load circuit of the sound demodulator. In the two sides of the demodulator load circuit are produced voltage pulses occurring at an audio frequency which is the same as the rate of frequency deviation of the sound signal during normal reception. When using the ordinary type of sweep generator this rate remains constant at 60 cycles per second.

The strength of voltage pulses on one side of the demodulator load circuit increases while pulse strength on the other side decreases when there is deviation from the center frequency. The result is a combined or net pulse voltage which is alternately stronger on one side and then on the other side of the load circuit as frequency deviates below and above the center frequency. The combined pulse voltages during deviations below the center frequency may be represented as at *A* in Fig. 120-4, and during deviations above the center frequency as at *B*.

Connections from the two sides of any demodulator load circuit are made to the following audio-frequency amplifier

in such manner that deviations below the center frequency cause negative alternations of voltage in the output to the audio amplifier. When deviations are above the center frequency the result is positive alternations in the output to the audio amplifier.

In any f-m sound signal, also in the signal from a sweep generator, are frequency deviations alternately below and above the center frequency. These deviations produce in the

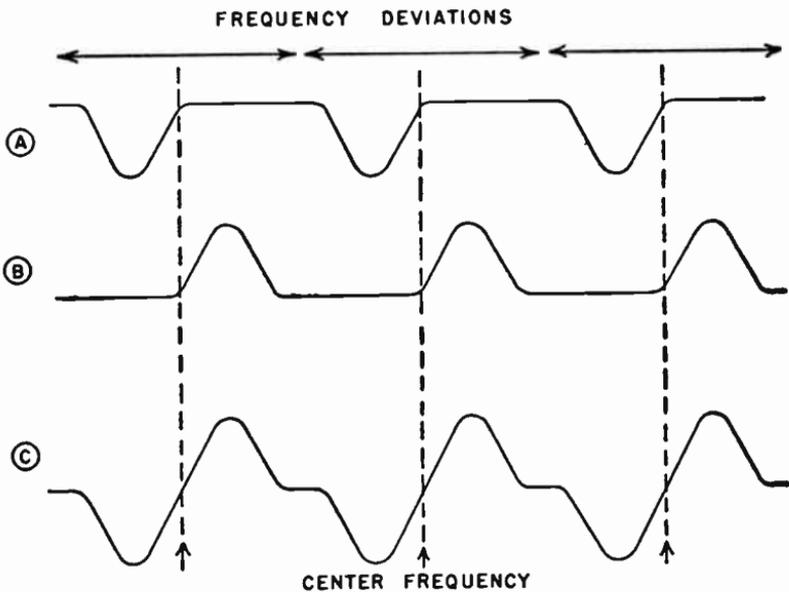


Fig. 120-4.—Audio-frequency voltage pulses produced by frequency deviations.

demodulator output to the audio amplifier a voltage wave such as shown at C in Fig. 120-4. This is a voltage alternating at audio frequency.

If we connect a d-c voltmeter across one side or the other of the demodulator load circuit the meter is subjected to voltage pulses which always are of one polarity, either positive or negative depending on the side of the demodulator load to which connection is made. Furthermore, the strength of these one-way pulses will vary with the strength of sound signal

from a transmitter or a sweep generator, and will not be affected by opposite or compensating pulses in the other side of the load circuit. It is the strength of these one-way or d-c voltage pulses that usually serves as an indication of gain or response during alignment of sound take-offs, sound i-f transformers, and the primary of the demodulator transformer. This is true whether the output indicator is a voltmeter or an oscilloscope.

With a d-c voltmeter connected to the demodulator output going to the audio amplifier the meter will be subjected to the alternating voltage at *C* in the diagram of Fig. 120-4. If the demodulator is correctly aligned the amplitudes will be equal on both sides of zero. Then the average voltage will be zero, and the meter will read zero. If the demodulator or stages preceding it are not correctly aligned the amplitudes will not be equal, and there will be a d-c component which may be either negative or positive. The d-c meter will indicate this d-c component so long as alignment is incorrect. Zero voltage on the meter is the guide for correct alignment of the secondary of the demodulator transformer.

If an oscilloscope is connected across either side of the demodulator output, and if input signal is from a sweep generator, the oscilloscope traces will show pulses like those at *A* or *B* in Fig. 120-4. The stronger the applied signal and the greater the gain the higher the traces will become.

Whenever an oscilloscope is used it is advisable to have on the receiver end of the vertical input cable a resistor whose value is between 100,000 ohms and one megohm. This resistor helps isolate the aligned circuits from capacitance of the cable and of any capacitor in the vertical gain control of the oscilloscope.

When the oscilloscope is connected to the demodulator output going to the audio amplifier, and the input signal is from a sweep generator, the trace will be as shown at *C* in Fig. 120-4. The two peaks which are on either side of a center frequency, and the portion of the trace joining the peaks, make up what is called the S-curve of the demodulator.

There are a number of alignment adjustments in any sound section whose object is to produce maximum readings of a voltmeter or maximum height of oscilloscope traces for any given strength of input signal. These adjustments include those of the takeoff coupler or transformer, of any sound i-f transformers, and of the primary in the demodulator transformer.

For these alignments in a sound section which includes a discriminator type of demodulator the meter or oscilloscope usually is connected across only one side of the demodulator output. This connection would be across resistor  $R_a$  or else

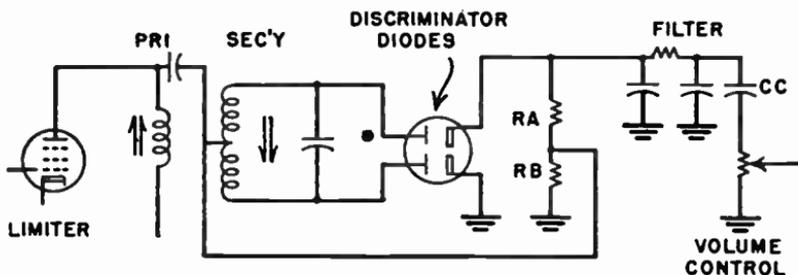


Fig. 120-5.—Typical circuit of a discriminator type of sound demodulator.

across resistor  $R_b$  of Fig. 120-5, which shows a typical discriminator circuit. Most often it is convenient to connect the meter or oscilloscope from ground or B-minus to the junction between these two resistors.

When making the adjustments mentioned in a sound section having a ratio detector for demodulator, and using a volt-

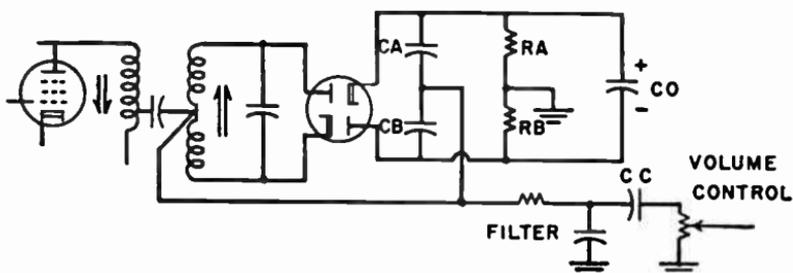


Fig. 120-6.—One type of circuit for a ratio detector type of sound demodulator.

meter for output indicator, the voltmeter is connected across the two ends of the large capacitor marked  $C_o$  in Fig. 120-6. The charge and voltage of this capacitor vary with strength of incoming signal and with gain. In some cases it may be practicable to connect the voltmeter across either of the capacitors  $C_a$  or  $C_b$ , or across either of the resistors  $R_a$  or  $R_b$  if such capacitors or resistors are in the circuit. Voltages across these elements ordinarily are so weak that this method may not be satisfactory.

Many ratio detectors employ a circuit such as that of Fig. 120-7, where a single resistor  $R_o$  is across the large capacitor  $C_o$  and the detector diodes. In this case the testing voltmeter may be connected only across capacitor  $C_o$  or resistor  $R_o$  while making the adjustments requiring maximum meter readings.

If an oscilloscope is used as output indicator for a ratio detector the vertical input may be connected across either of the divided capacitors or resistors shown in Fig. 120-6 while

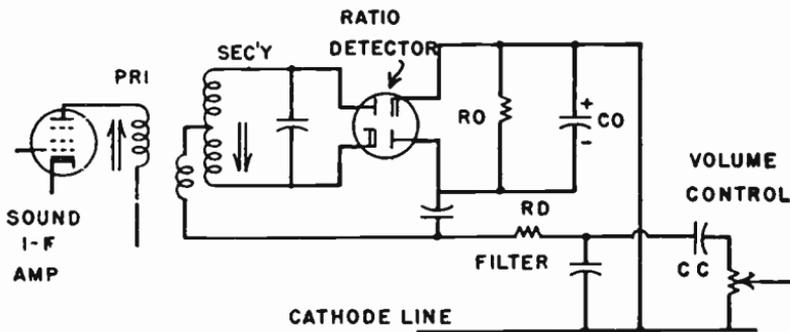


Fig. 120-7.—Ratio detector in a sound system whose cathode line supplies plate and screen voltages to other tubes.

making the adjustments requiring maximum height of trace. Traces secured with these connections often are of too little height to allow satisfactory adjustments.

When the ratio detector load circuit is arranged as in Fig. 120-7, with no divided output, the oscilloscope cannot be connected directly to the detector load for alignment of sound

takeoff, i-f transformers, and primary of the demodulator transformer. These adjustments may be made with a voltmeter connected across the large capacitor.

It is possible to align the sound takeoff, i-f transformers, and primary of the demodulator transformer with the voltmeter or oscilloscope connected to the demodulator output which goes to the audio amplifier, rather than to any of the points previously mentioned. This method is applicable with discriminator or ratio detector circuits such as shown in Figs. 120-5 and 120-6. The points for output indicator connection are from ground or B-minus to any point in the de-emphasis filter line which is on the demodulator side of coupling capacitor  $C_c$  leading to the volume control. With this connection of the output indicator it is necessary to detune the secondary of the demodulator transformer by giving its adjustment one or two turns in either direction.

While the secondary is detuned and a voltmeter connected as output indicator, all adjustments from sound takeoff to demodulator transformer primary, inclusive, are adjusted for maximum voltage. With an oscilloscope for output indicator all these adjustments are aligned for maximum height of trace. After these adjustments are completed, the meter or oscilloscope may remain connected to the same points during alignment of the demodulator transformer secondary, according to methods which will be described.

**Limiter Input for Alignment Indications.** — If the sound section includes a limiter tube preceding a discriminator the direct voltage developed across the limiter grid resistor may be used to indicate correct alignment of the sound takeoff and all sound i-f transformers. Voltage across this grid resistor varies proportionately to input signal strength and gain in all parts from sound takeoff through the i-f transformer just ahead of the limiter tube. Circuits are described in the article on *Limiters, F-m Sound*.

One method of using the limiter grid resistor utilizes a constant-frequency generator connected and operated as previously described, and an electronic voltmeter or highly sensi-

tive moving coil voltmeter connected across the resistor. Alignment adjustments are made for maximum reading on the meter.

Another method employs an oscilloscope connected across the limiter grid resistor, with sweep and marker generators furnishing input signals. In series with the vertical input to the oscilloscope must be the highest resistance which allows readable traces. Alignment adjustments are made for a trace of maximum height, centered at the sound intermediate frequency as identified by a marker, and reasonably symmetrical in shape on both sides of the center frequency. This method is especially useful for alignment of sound i-f transformers of the overcoupled type whose frequency response is a double-peaked curve.

When using the oscilloscope on the limiter grid resistor the two sides of the trace should be at least 70 per cent of the peak height for frequencies which are 75 to 150 kilocycles below and above the center frequency. These points on the response may be identified with markers. This response below and above the center frequency should be obtained with either single-peaked or double-peaked curves.

In case some or all the sound i-f transformers are far out of alignment, the generator may be connected first to the grid of whatever tube precedes the limiter, while adjusting the transformer between that tube and the limiter. Then the generator connection may be moved to the grid of the second preceding tube while adjusting the second preceding transformer. Thus the adjustments are carried out stage by stage until reaching the sound takeoff, which is aligned last.

**Demodulator Transformer Secondary Alignment.** — For alignment of the secondary in the demodulator transformer the testing meter or oscilloscope must be connected across points furnishing the voltage wave shown at *C* in Fig. 120-4. With a discriminator as shown by Fig. 120-5 this requires connection between the top of resistor *Ra* and the bottom of *Rb*. Usually the output indicator may be connected between ground or B-minus and any point along the de-emphasis fil-

ter line which is on the discriminator side of coupling capacitor  $C_c$ . The output indicator will be thus connected if preceding adjustments have been made with the secondary intentionally detuned. A voltmeter will read zero when the secondary is correctly aligned, and an oscilloscope will show a symmetrical S-curve with correct alignment.

Transformer secondary alignment for a ratio detector requires consideration of whether the detector output is practically at the potential of ground or B-minus, or is highly positive with reference to ground. Fig. 120-6 shows an output which is near ground potential, as is evident from the ground (or B-minus) connection between resistors  $R_a$  and  $R_b$ , also from the ground connections on the filter and volume control. With this general type of ratio detector circuit either the voltmeter or the oscilloscope may be connected from ground or B-minus to any point along the filter line on the detector side of coupling capacitor  $C_c$ . The meter will read zero or the oscilloscope will show a symmetrical S-curve when alignment is correct.

When using a meter to obtain zero reading upon correct alignment of the secondary it may be possible to obtain as many as three zero readings if the alignment adjustment has a very great range. The correct zero point is at the center frequency, as shown at  $C$  in Fig. 120-4. The other two zero points are at the outer ends of the complete S-curve, at frequencies in between the center frequencies of that diagram. The correct zero reading is the one occurring between two peak readings, one positive and the other negative. Adjustment for this zero point is critical. The least change of adjustment either way will produce either a positive or negative reading. At either of the two incorrect zero points the adjustment is not critical.

Fig. 120-7 applies specifically to a sound system in which the cathode line for all the audio tubes supplies plate and screen currents to other tubes in the receiver. This cathode line usually is 130 to 150 volts positive with reference to ground or B-minus. In this circuit there are no points between

which a voltmeter may be connected to indicate zero voltage when the transformer secondary is correctly aligned.

With a circuit of the general type shown by Fig. 120-7 an oscilloscope may be connected from any point on the detector side of the filter line to either the cathode line, B-minus, or ground, and will display a symmetrical S-curve when the secondary is correctly aligned.

If a ground terminal of the oscilloscope is connected to B-minus or chassis ground with a high-voltage audio system it is advisable to make this connection through a capacitor of about  $\frac{1}{2}$  mf to prevent high voltage from going through a direct ground connection to the sweep or marker generator. If no oscilloscope is available the secondary may be aligned merely for good sound quality and volume and for freedom from intercarrier buzz while receiving a signal with sound from a television station. Final alignment with intercarrier sound may be for minimum buzz with the contrast control at its maximum setting.

**S-curve Requirements.**— Fig. 120-8 shows features which are desirable in the S-curve of a sound demodulator, as observed on an oscilloscope with signal from a sweep generator. Peak-to-peak voltage, *A*, should be maximum obtainable while

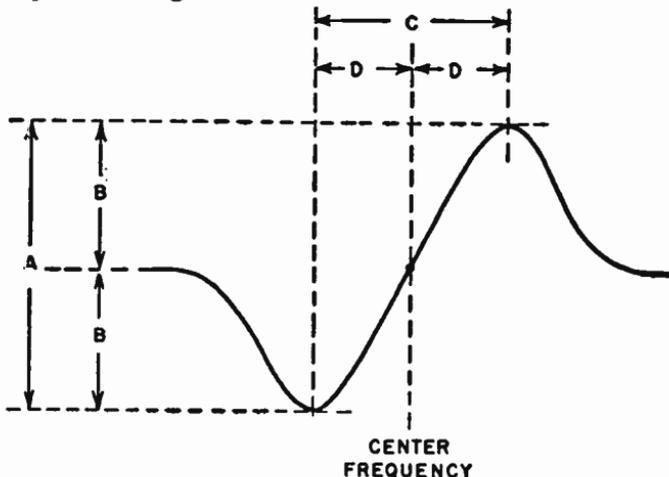


Fig. 120-8.—A demodulator S-curve showing features which are checked and adjusted during alignment.

observing other requirements. Amplitudes  $B$  and  $B$  must be equal above and below the zero point. Maximum voltage depends on correct alignment of the demodulator transformer primary and other transformers and couplers back through the sound takeoff. Equal amplitudes depend largely on adjustment of the demodulator transformer secondary.

Peak separation,  $C$ , should be not less than 300 kilocycles and not more than 500 or 600 kilocycles, this latter limit being for prevention of sound interference. The two peaks should be equally spaced from the center frequency, as at  $D$  and  $D$ . Separation and spacing of the peaks depend chiefly on adjustment of the transformer primary, but are affected also by secondary adjustment. These features are checked with markers.

The center of the curve must be at the sound intermediate frequency for a dual-channel sound system or at 4.5 megacycles for an intercarrier sound system. This point is checked with an accurate marker. It is affected by adjustment of both primary and secondary in the demodulator transformer.

The curve should be symmetrical or of similar shapes from the center out to the peaks on each side. This is chiefly a matter of primary adjustment. The sloping line between the peaks, and through the center, should be practically straight for at least 50 to 125 kilocycles on both sides of the zero or center point. The line is straightened largely by correct adjustment of the transformer primary.

**Steps in Alignment.** — Following is a brief summary of the steps ordinarily taken during alignment of the sound section when using an oscilloscope with sweep and marker generators.

1. Connect the oscilloscope and the sweep and marker generators as directed. Turn on all the instruments and the receiver, then allow at least a 15-minute warmup period for operating conditions to stabilize.

2. Set the sweep width for something between 1 and 3 megacycles.

3. Adjust the sweep frequency or center frequency to bring an S-curve to the center of the oscilloscope screen.

4. Readjust the sweep width to leave only small horizontal extensions on opposite sides of the S-curve, about as shown by Fig. 120-8. Readjust the sweep frequency as may be necessary to keep the S-curve in the center of the screen.

5. Adjust the oscilloscope horizontal gain so that the S-curve and horizontal lines nearly fill the screen. Adjust the vertical gain to make the curve of the general proportions shown by Fig. 120-8.

6. Adjust the transformer primary for maximum peak-to-peak height of the curve. Unless adjustments of preceding transformers and the sound takeoff are known to be correct it may be well to check them at this point for possible improvement of the S-curve.

7. Adjust the transformer secondary for equal amplitudes of peaks, for peaks equally spaced from the center, and for the straightest sloping line between peaks.

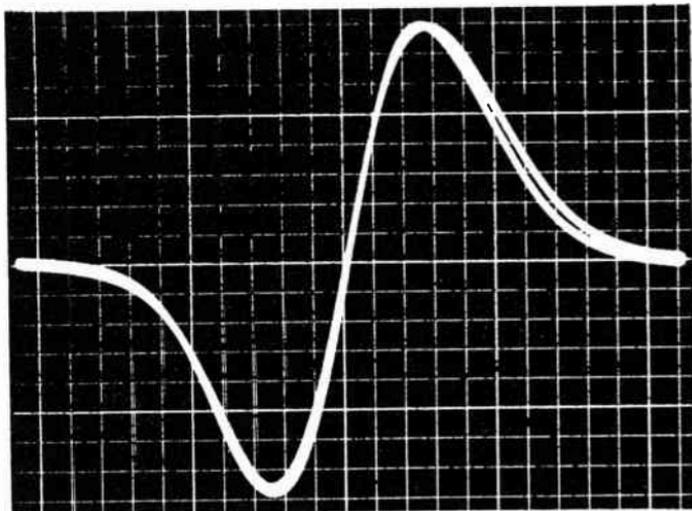
8. Work back and forth between primary and secondary adjustments to get the best balance between requirements outlined in steps 6 and 7.

9. Use frequency markers to check the frequency at the center of the curve, the separation between peaks, the frequency difference from the center to each peak, and the frequency range in which the sloping line is fairly straight.

10. To move the entire curve with reference to a marker frequency it will be necessary to adjust both the secondary and primary of the demodulator transformer. The secondary alone will shift the curve so that the marker moves up or down on the central slope, but to preserve symmetry it will be necessary to readjust the primary.

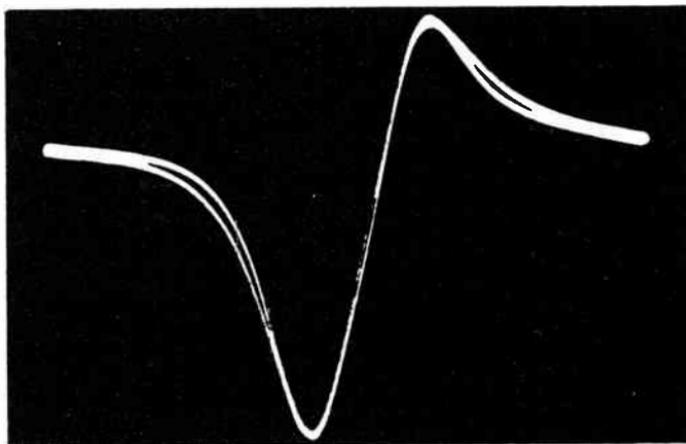
A marker strong enough to be clearly visible at the center of the S-curve may so distort the curve and alter the heights of peaks as to make adjustments difficult. When the center of the curve is at the marker frequency any change of marker strength will affect the amplitudes of both top and bottom, or both peaks, equally. As a last resort the marker may be made strong enough to be distinct, regardless of its effect on the curve. Then the sweep frequency or center frequency of the

generator may be adjusted to bring this marker exactly on the center line of the graph scale for the oscilloscope screen. Finally, the marker may be turned off and the curve adjusted to bring its center point exactly on the center line of the graph scale.

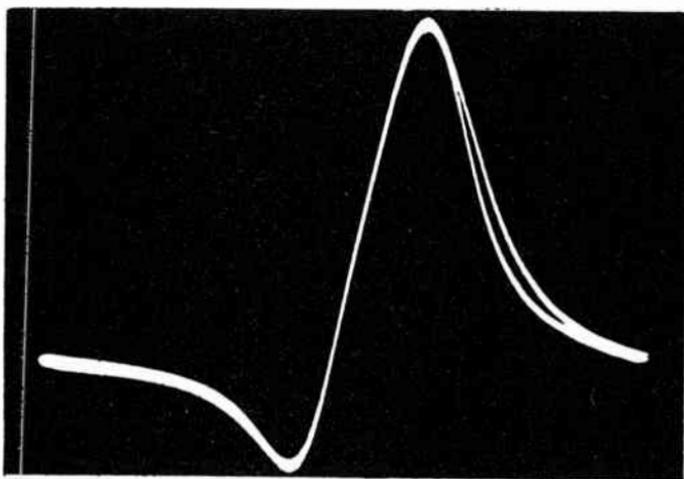


*Fig. 120-9.—A satisfactory S-curve as it appears on the screen of an oscilloscope.*

Fig. 120-9 shows a correctly adjusted S-curve as it appears with a graph scale used in front of the oscilloscope screen. The effects of incorrect adjustments are illustrated by the

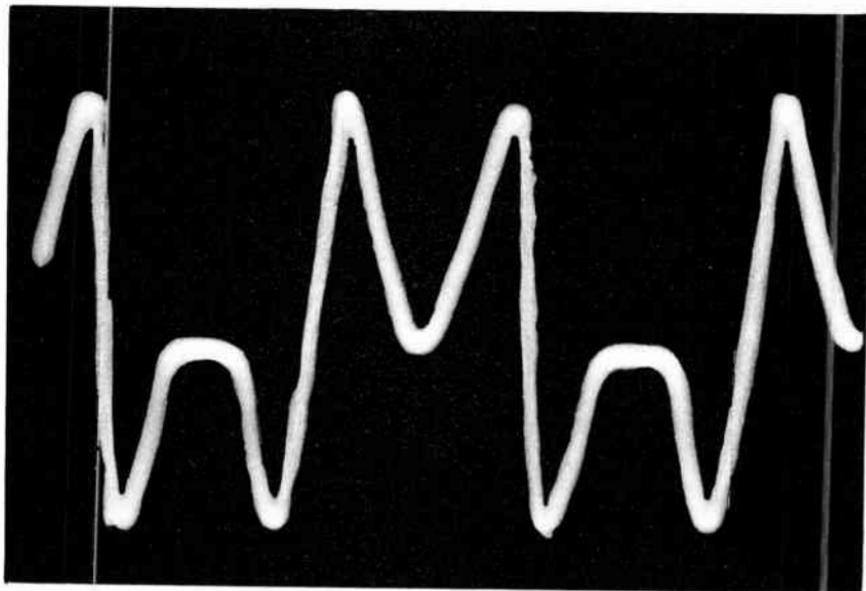


*Fig. 120-10.—The positive peak has less amplitude than the negative peak, due to misadjustments.*

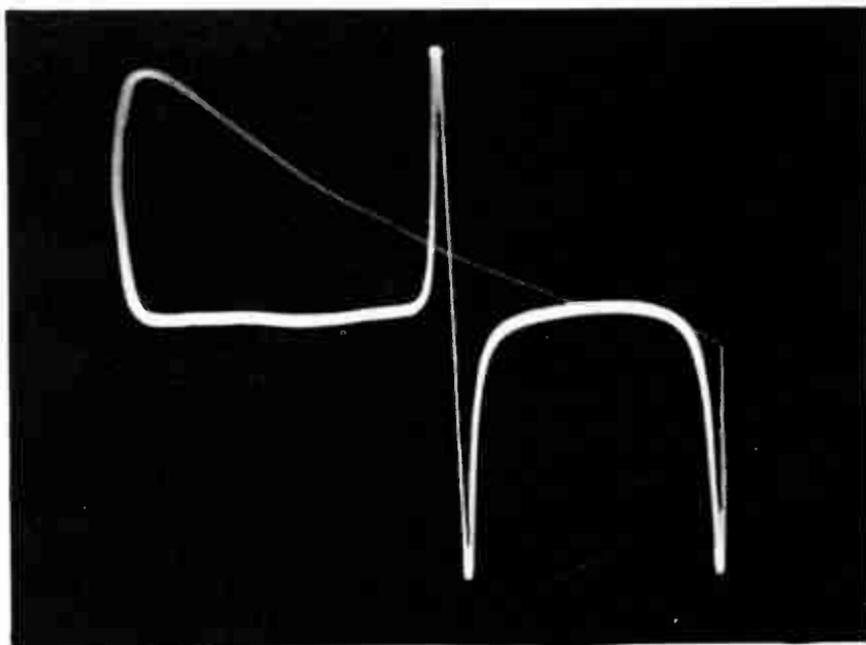


*Fig. 120-11.—When peaks are of unequal amplitudes the center frequency of the S-curve will not be at the center frequency of f-m sound.*

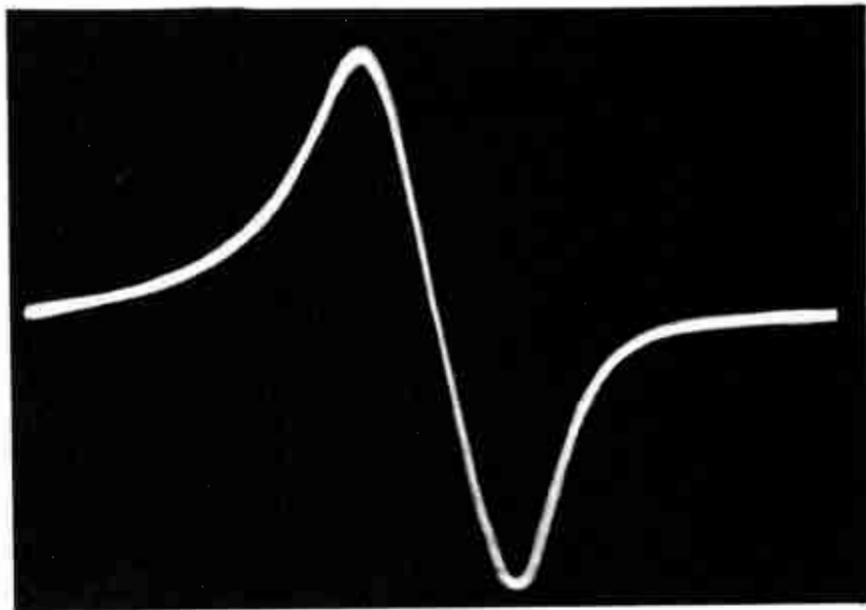
oscilloscope traces of Figs. 120-10 and 120-11. Amplitudes of positive and negative peaks are not equal with reference to zero, which is indicated by the relatively flat or horizontal extensions at the left and right of the S-curves. It



*Fig. 120-12.—When using internal horizontal sweep of the oscilloscope the trace will show a number of S-curves.*



*Fig. 120-13.—Increasing the horizontal sweep frequency will reduce the number of S-curves shown by the trace.*



*Fig. 120-14.—Adjustments of centering and of width will bring a single S-curve onto the oscilloscope screen.*

would be found also that such misadjustments of amplitude are accompanied by shifting of the center of the S-curve to frequencies higher or lower than the desired center frequency or sound i-f frequency.

It is possible to bring an S-curve onto the oscilloscope when using the internal horizontal sweep rather than synchronized sweep, although the sweep generator must be employed in the usual manner to sweep through frequencies on both sides of 4.5 mc for intercarrier sound or on both sides of the sound intermediate for dual sound systems.

Internal sweep frequency of the oscilloscope is adjusted for 60 cycles or for 30 cycles per second which, with the center frequency of the sweep generator suitably adjusted, will bring onto the oscilloscope screen a number of S-curves joined as a single trace. A trace such as shown by Fig. 120-12 will appear when internal sweep frequency is 30 cycles per second, and a trace as shown by Fig. 120-13 will appear with horizontal sweep at 60 cycles per second, where the S-curve is plainly visible although very narrow. Increasing horizontal gain of the oscilloscope will widen the S-curve, while simultaneous adjustment of horizontal centering will keep the curve centered on the screen. The result of gain and centering adjustments is shown by Fig. 120-14. The single S-curve will show frequency increasing from left to right or else from right to left, depending on internal sweep frequency and on which way the horizontal centering is adjusted. With synchronized sweep there are two superimposed S-curves produced by forward and back frequency sweep. The curves of Figs. 120-9 to 120-11 are with synchronized sweep.

The secondary of the demodulator transformer may be adjusted quite easily to bring the center of an S-curve to the center frequency for f-m sound if the signal from the marker generator can be modulated at any audio or tone frequency. So long as the secondary is not correctly adjusted, the sound modulation of the marker signal will come through to the audio output at which the oscilloscope is connected. The audio frequency will show up as strong rippling and increased

vertical height of the left-hand and right-hand ends of the S-curve trace. The central portion of the trace may be affected, but to a much lesser degree.

Correct adjustment of the transformer secondary will cause the waviness to disappear or to decrease to small amplitude. It may be necessary to make some readjustment of the transformer primary or of couplers in the sound system ahead of the demodulator transformer. Such readjustments would be called for no matter in what manner the center frequency of the S-curve is determined. Waviness disappears because there is zero gain at the f-m center frequency to which the marker generator is precisely tuned.

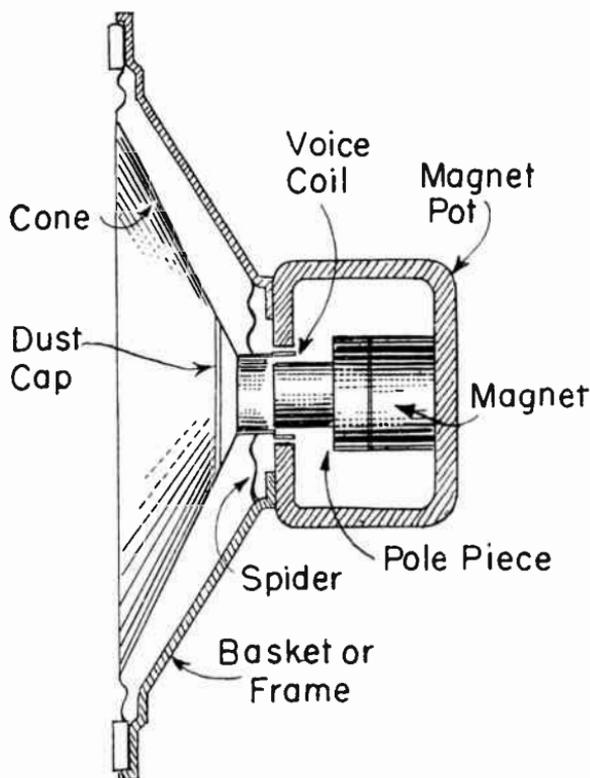


Fig. 1.—Structural elements of a permanent magnet (PM) speaker.

**SPEAKERS.** — Most television speakers are of the permanent magnet (PM) type, whose construction is shown by Fig. 1. A cone of paper or fibre is flexibly supported at its outer edge between the frame and a spacer ring. The cone is flexibly supported at its center by a spider of stiffened cloth that fastens to the frame. A voice coil consisting of a few turns of insulated wire is around the outside of a tube extending back from the center of the cone.

A strong permanent magnet and soft iron pole piece are mounted in a "pot" whose forward end has a circular opening around the circular pole piece. In the small gap between pot and pole piece is a concentrated magnetic field. Flexible leads from the voice coil go to terminals of the frame, thence to the secondary of a step-down speaker coupling transformer. The transformer primary is in the plate circuit of the audio output tube or tubes.

Alternating currents at audio frequencies in the speaker voice coil produce their own reversing magnetic fields, which react with the field of the permagnet magnet to move the voice coil and attached cone forward and back at the audio frequencies. This movement of the cone sets up sound waves at audio frequencies in surrounding air.

Principal catalog specifications of a speaker include cone diameter, power capacity in watts, voice coil impedance, and permanent magnet weight or magnetic field strength. Cone diameter may be that of the entire cone, including the part that rests in the frame, or of only the portion exposed within the spacer ring.

Power capacities range from two to 20 watts for frequencies of 400 to 1000 cycles per second. To avoid severe distortion of sound and possible damage to the speaker, its rating should be in excess of the maximum undistorted output power at which the audio output tube or tubes are rated.

Voice coil impedance at 400 to 1000 cycles per second most commonly is 3.2 ohms or 3 to 4 ohms, but may be 6 to 8 ohms, 16 ohms, or of higher values. Impedance rises gradually at higher frequencies, and usually shows a resonant peak at

something lower than 100 cycles. These low impedances result from the limited number of voice coil turns which may be used without weight so great as to make the cone move sluggishly. Impedance of the secondary in the speaker transformer should be no greater than voice coil impedance. Transformer primary impedance should be that recommended in tube ratings for the audio output tube or tubes used.

Ability of a speaker to handle audio power without distortion depends largely on weight and consequent strength of its magnet. Magnet weights are from less than an ounce in the smallest speakers up to 8 or 10 ounces or more for power ratings up around 20 watts. The stronger the magnet the greater will be cone movement for a given audio power, and the more uniform will be the magnetic field and movement of the cone throughout its travel.

Sound output power of a speaker is not directly proportional to audio input power, but varies with frequency in the general manner shown by Fig. 2 for a speaker of ordinary design. An actual sensitivity curve is not so smooth as shown,

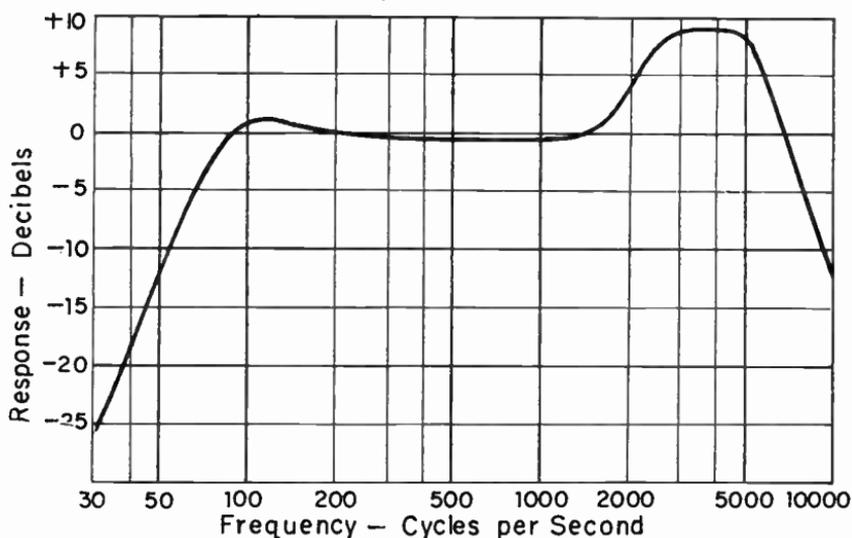


Fig. 2.—Sound output power of speakers usually varies with frequency when electrical input power remains constant.

but will have peaks and valleys all the way across. This may be observed when operating a speaker from an audio-

frequency generator. Sound output will become irregularly weaker and stronger as frequency gradually is increased, even with constant audio input power.

Some speakers have an electromagnet rather than a permanent magnet furnishing the steady magnetic field in the voice coil gap. Connections are shown by Fig. 3. The field magnet winding is around a core which extends forward inside the

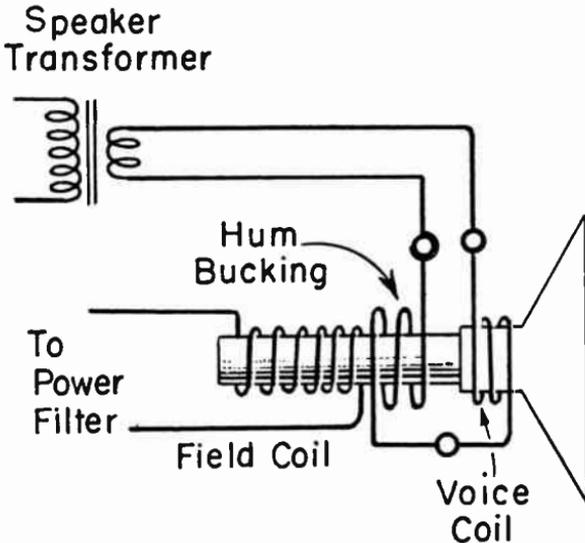


Fig. 3.—Internal wiring and external connections for an electro-magnetic speaker having a wound field coil.

voice coil. This winding nearly always is used as a choke in the filter of the low-voltage B-power supply system, thus being supplied with direct current. The small ripple voltage is counteracted to a greater or less extent by a hum bucking winding in series with the voice coil and transformer secondary. Field coil resistances range from less than 100 ohms to as much as 3000 ohms. Speakers of the electromagnetic type may be recognized by their four terminal leads, as compared with two leads from a PM speaker.

**Speaker Trouble and Repair.** — Troubles which are not too difficult for repair or for replacement of parts include punctured or deformed cones, loose spiders, and voice coil tubes which are loose, deformed, or not centered around the mag-

net. Other than new cones and spiders when needed, repairs require only cone cement and centering shims. A damaged cone may be replaced with a new one as follows.

Do all work on a thoroughly clean surface, to avoid possibility of getting dirt or metal chips into the magnetic gap. Loosen the outer edge of the cone and attached spacer ring from the frame by carefully prying with a knife blade or similar tool. The joint may be loosened with cement thinner. Disconnect the flexible voice coil leads from terminals on the frame. Pull the cone outward to reach the cloth spider and free the edges of the spider from the frame with a thin screw driver. Withdraw the cone, voice coil, spider, and spacer ring from the speaker frame. Carefully pull the dust cap away from the cone. This cap will be cemented back onto the replacement cone.

For selection of a new cone and spider make measurements on the original parts as shown by Fig. 4. Cone diameter is that of the cone alone, not of the spacer ring. Voice coil tube

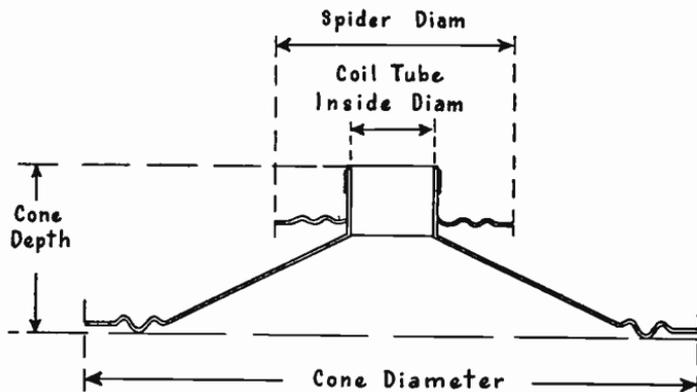


Fig. 4.—Dimensions which must be known for selection of a replacement cone and spider.

inside diameter must be accurately measured, since variations between standard sizes may be as little as  $1/32$  inch where cones are less than eight inches in diameter. Note whether the spider is flat or cupped and obtain a replacement of similar shape.

Examine the magnet gap. Should particles of foreign mat-

ter be present, work them out with paper, thin brass, or thin aluminum. A jet of compressed air is better if it is available. If the unit is to be laid aside, cover the magnet gap with adhesive tape to keep out dirt. Otherwise proceed to clean the frame where the new cone and spider will be cemented. Cement the cone where the spider will attach to it, work the spider over the voice coil tube, and make sure the cement is evenly distributed.

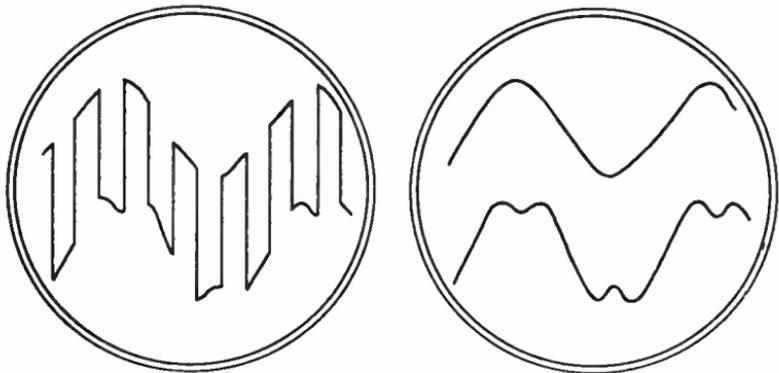
Before the cone and spider are cemented to the frame the voice coil tube must be centered around the magnet pole by using shims. Centering shims are available in thicknesses from five to fifteen thousandths inch, with four of each size in a set. Otherwise the magnet end may be wrapped with paper of suitable thickness or with suitable thickness of brass shim stock obtainable from auto supply stores. A wrapped shim should extend about  $\frac{3}{4}$  inch beyond the magnet, so the voice coil tube may be worked onto the outside of the shim and over the magnet pole. Strip shims are inserted after the cone is in place. With any kind of shim there must be a snug fit to insure correct centering.

When the shim or shims are ready, cement the frame where the edge of the cone is to rest, also where the edge of the spider will rest. Position the voice coil leads as with the original cone, put the cone with spider in place, and press the cement joints together. Next cement the cardboard or fibre spacer ring in place. Remove shims only after the work dries for several hours. Finally, cement the felt dust cover over the center opening of the cone.

**SWITCH, ELECTRONIC.** — An electronic switch is an instrument that allows observing at the same time on the screen of an ordinary oscilloscope the traces for two independent voltages or signals. The switch applies each of the observed voltages to the oscilloscope during alternate very brief intervals of time. The switching rate is so rapid that traces for both voltages remain visible, due to persistence of illumination on the oscilloscope screen and persistence of vision in the eye.

The principle is illustrated at the left in Fig. 121-1. The oscilloscope beam is caused to trace first a short length of one curve, then a short length of the other curve, and so on all across the horizontal travel. By suitable adjustments it is possible to make both traces appear continuous or practically so, as at the right.

The two observed voltages are applied through separate input terminals on the switch to separate amplifiers. The amplifiers are alternately biased to cutoff, and allowed to amplify the voltages applied to their grids. The outputs or plates of the two amplifier tubes are connected together and to the output terminal of the switch. This output terminal is con-



*Fig. 121-1.—How two simultaneous oscilloscope traces are formed by the electronic switch.*

nected to the vertical input of the oscilloscope. On each amplifier is a gain control allowing the height of its portion of the trace to be adjusted independently of the height of the trace from the other amplifier. The switching rate is adjustable, usually from 10 or 20 times per second up to 2,000 times.

The switching or cutoff biasing of the amplifiers is accomplished by a multi-vibrator oscillator. The rate of switching is altered by varying the operating frequency of the multi-vibrator. In most electronic switches the alternating positive and negative pulses from the multivibrator are applied to the grids of two blocking tubes. In the cathode resistors of these blocking tubes are produced corresponding pulses of current and voltage. These two resistors are also the cathode resistors for the two amplifier tubes. When the voltage pulses make the amplifier cathodes highly positive it is equivalent to making the amplifier grids highly negative, and during these intervals the amplifiers are cut off.

The biasing voltages have the form of square waves. The multivibrator may be designed to produce the square wave voltages when no separate blocking tubes are used, or these separate tubes may do most of the squaring of the voltage waves.

The relative vertical positions of the two traces may be varied as desired. One may be above the other, as in Fig. 121-1, or their positions may be reversed as at the left in Fig.



Fig. 121-2.—The two traces may be moved vertically with reference to each other.

121-2. The two traces may be superimposed for close comparison, as at the right. This is accomplished by a position control or balance control which will increase the gain of one amplifier while decreasing the gain of the other for raising one

## SWITCH, ELECTRONIC

trace above the other, or which will make the gains equal when the two traces are to have the same zero or base line, as at the right in Fig. 121-2. The gain may be altered by change of plate voltage, screen voltage, or grid bias.

**Operating the Electronic Switch.** — On the panel of an electronic switch are two input terminals, one for each observed voltage, also a single output terminal for the oscilloscope lead, and one or more ground terminals. Connections for observing simultaneously the input and output voltages of an amplifier are shown by Fig. 121-3. The signal source may be a generator. The place of the amplifier might be taken by any other device being tested.

All ground terminals are connected together or all are connected to a grounding metal top of a test bench. The high side of the signal source is connected to the input of the device being tested and also to either of the input terminals on the switch. The output of the device being tested is connected to the other input terminal of the switch. The output terminal

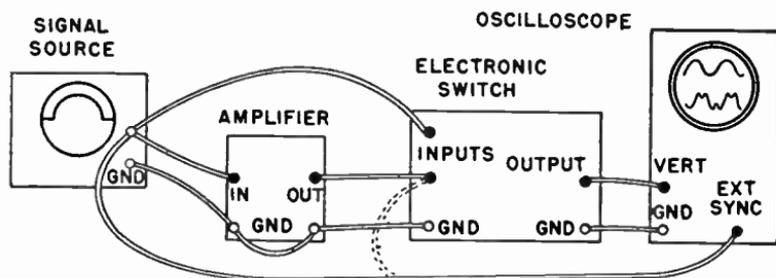


Fig. 121-3.—Electronic switch connected for observing input and output voltages of an amplifier.

of the switch is connected to the vertical input of the oscilloscope. In order to obtain reasonably continuous traces, as at the right in Fig. 121-1, it is necessary to employ the external synchronizing connection of the oscilloscope, and turn off the internal synchronization. The external sync terminal of the oscilloscope may be connected to the signal source, as shown, or to the amplifier output or to either of the input terminals of the switch.

To begin with, the two input gain controls of the switch may be set at zero while the position control or balance control is adjusted to produce two separate horizontal lines on the oscilloscope screen. Then the gain controls of the switch may be increased enough to produce two voltage wave traces. The height of either trace may be changed by its own gain control. The heights of both traces may be changed together by varying the output of the signal source. The internal sweep frequency and the sync control of the oscilloscope are adjusted in the usual manner to hold the traces steady on the screen.

The switching rate, varied by the frequency controls of the electronic switch, need have no particular relation to signal frequency or oscilloscope sweep frequency. Switching frequency should be made whatever allows the clearest and most nearly continuous traces on the oscilloscope screen. In general, it is best to employ high-speed switching when observing low-frequency signals, and to employ relatively low-speed switching for high frequency signals. This usually will allow the least flickering and greatest continuity of traces.

To avoid distorted traces it is necessary that the frequency response of the oscilloscope vertical input attenuator and vertical amplifier be practically flat up to at least ten times the switching rate. This is because voltages being applied to the vertical input are of square waveform, as shown at the left in Fig. 121-1, and such waveforms contain frequencies many times as high as the fundamental or the switch rate frequency.

If either of the gain controls on the electronic switch is set at zero the corresponding trace becomes a straight horizontal line. This line may be used as a reference level or zero level by operating the position or balance control to bring the line into any desired relationship with the remaining trace.

**Uses for the Electronic Switch.** — The principal uses for the electronic switch include comparison of two waveforms, checking phase relations and phase shifts, comparison of amplitudes or signal strengths, and comparison of frequency responses.

Waveforms may be compared at the input and output of any device or apparatus. It is possible also to compare the inputs to two devices or their output waveforms. Relative changes in any two such waveforms may be observed while input signals or other operating conditions are varied. As one example, the input and output waveforms of all types of filters may be compared.

One of the more important uses of the electronic switch is in checking and measuring the phase shift occurring in all types of amplifiers and amplifying stages at various frequencies. With input and output of an amplifier connected as in Fig. 121-3, and with the two traces superimposed, the direction and extent of phase shift becomes clearly apparent. Phase relations are shown because timing always remains the same in both traces. Any given instant of time is the same on a vertical line passing through both traces. The graph screen of the oscilloscope may be used to establish corresponding instants of time along both traces. Polarity inversions may be examined. Various distortions and corrections of distortion may be studied.

The electronic switch is useful for comparing waveforms and amplitudes in the outputs from the two sides of any inverter tube or cathode follower, for comparing the signals at various points in push-pull amplifier circuits, for comparing the sawtooth voltages applied to opposite plates of a pair in an electrostatic deflection picture tube, and many similar purposes.

The electronic switch allows observation of direct potentials and their changes on an oscilloscope whose input is designed for only alternating potentials. The direct potential to be observed is connected between one of the input terminals and ground on the electronic switch. The gain control for the other input terminal is set at zero, to provide a zero reference trace on the oscilloscope screen. Usually it is best to adjust the switching rate to some rather high frequency. Then the switching action breaks the direct potential into pulses occurring at the switching rate. This is the equivalent of a

square wave alternating potential, which will pass through the input attenuator and vertical amplifier of the oscilloscope to produce on the screen a second trace whose separation from the zero trace is proportional to the observed direct potential. Cathode biases, power supply bleeder voltages, and other direct voltages may be examined with this method.

Any electronic switch will serve as a square wave generator for frequencies up to about one-fourth the maximum switching rate or frequency. At higher frequencies the waveform will have considerable distortion. No connections are made to the input of the switch. Both gain controls are set at zero. The position control or balance control is used to vary the amplitude or height of the square wave. The square wave signal is taken from the output terminal and ground terminal of the switch. Frequency of the output signal is varied with the frequency controls of the switch.

**SYNC SECTION.** — The sync section of the television receiver extends from the video detector or a video amplifier through to the sweep oscillators. The input to the sync section consists of the composite television signal containing picture signals, vertical and horizontal and equalizing pulses, and blanking intervals. The sync section delivers to the vertical and horizontal sweep oscillators the pulses of voltage which keep oscilla-

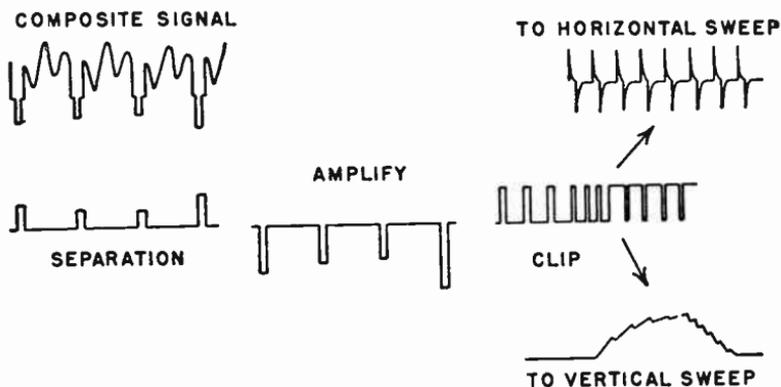


Fig. 122-1.—Principal steps in producing oscillator synchronizing voltages from the composite television signal.

tor frequencies correctly timed or synchronized with the vertical and horizontal sync pulses of the received signal.

The following operations are performed by the sync section, as illustrated by Fig. 122-1.

1. Remove all picture signals from the incoming composite signal, also remove the pedestals which are at the black level and leave only the sync pulses. This function is described in detail by the article on *Separation, Sync*.

2. Amplify the remaining pulse voltages as may be necessary to allow other functions to be carried out satisfactorily. Such amplification is covered by the article, *Amplifier, Sync*.

3. Cut down the strength of sync pulses to a uniform value, and remove so far as is practicable all abnormally strong voltage pulses due to noise types of interference. This action is explained in the article on *Clippers or Limiters*.

4. Produce from the horizontal sync pulses sharp pips of voltage which will correctly trigger or control the horizontal sweep oscillator, either directly or through some form of automatic sweep frequency control.

5. Produce from the vertical sync pulses the relatively long voltage pulses which will trigger the vertical sweep oscillator. Voltages for triggering of the vertical and horizontal sweep oscillators are produced from the sync pulses by filters described in the article on *Filters, Differentiating and Integrating*.

6. Make the polarity of the triggering voltages whatever is required by the type of sweep oscillator or automatic frequency control used in the receiver. This is done by providing a suitable number of signal or polarity inversions as explained in the article on *Inverters and Inversions*. In Fig. 122-1 inversions of polarity are indicated by the directions in which sync pulses are drawn in the composite signal and in steps of the diagram.

A blocking oscillator is triggered by positive voltage pips or pulses applied to its grid. This is true also of a blocking oscillator used with a discharge tube. Therefore, with sweep oscillators of this type the voltage pips produced by leading edges of horizontal sync pulses must be positive.

Triggering voltages applied to the first section of a multivibrator sweep oscillator must be negative. They are inverted between the first and second sections of the multivibrator, and become positive at the grid of the second section for controlling discharge of the sawtooth capacitor. If triggering voltages are applied to the cathode of the first section in a multivibrator these voltages must be positive, because they will be of this same polarity at the plate of the first section and at the grid of the second section.

When a sweep oscillator is preceded by an automatic frequency control for sweep frequency, the polarity of the synchronizing voltage pips or pulses will depend on the type of automatic frequency control as well as on the type of oscillator. In any event, the triggering voltages arriving at the grid of the oscillator itself must be as previously described.

It is not necessary that separation, amplification, and clipping or limiting be performed in the order shown by Fig. 122-1. Separation often is provided in a d-c restorer tube, with no other tube needed for this purpose in the sync section. A single tube in the sync section may partially or completely remove picture signals while at the same time amplifying sync pulses, or may both amplify and clip or limit the pulses. Any tube providing necessary inversion of polarity nearly always acts also as an amplifier, a separator, or a clipper.

Instead of combining two or more functions in a single tube there may be more than one tube for the same function. It is rather common practice to provide amplification in more than one tube of the sync section. Often there will be one clipper acting on the bottom of the pulses and a second clipper acting on the top, with inversion of polarity between the two clippers, so that what was the bottom of the pulses in the first operation becomes the top of the pulses in the second operation.

A few of the many arrangements of tubes in sync sections are shown by Fig. 122-2. In the upper left-hand diagram are shown, in order, a separator, an amplifier, and a clipper. At the upper right there is first a separator, then a clipper, fol-

lowed by two amplifiers. The top amplifier is fed from the plate of the clipper and the bottom one from the cathode of the clipper. The signal to the top amplifier will be much

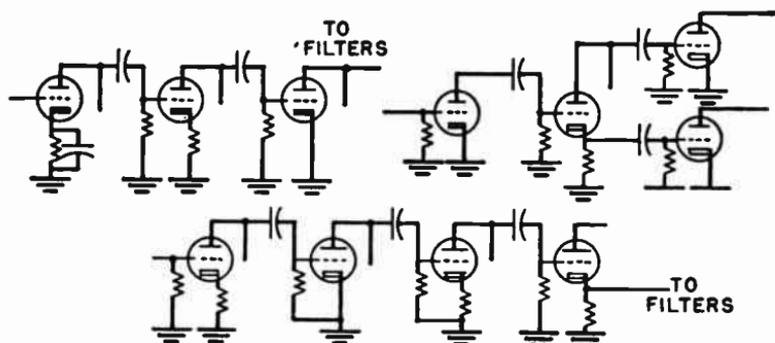


Fig. 122-2.—Connections of tubes in sync sections of receivers.

stronger than to the bottom amplifier, because there is amplification at the plate output but not at the cathode. Signals to both amplifiers would be of the same polarity. The same or a similar diagram might represent first an amplifier, then a separator, and finally two clippers. How a tube performs and what it accomplishes depend more on plate, screen, and grid voltages or biases than on circuit connections.

The lower diagram of Fig. 122-2 shows first a separator, then a clipper, then an amplifier, and at the output end a cathode follower feeding the vertical and horizontal filters. Any of the tubes in these diagrams might be pentodes instead of triodes.

At the upper left in Fig. 122-3 is first an amplifier, then a clipper, and last a combined inverter and cathode follower. From the plate and cathode of this last tube, acting as an inverter, are taken signals for an automatic frequency control system. From the cathode, with the tube acting as a cathode follower, are taken signals for the vertical sweep oscillator. The upper right-hand diagram shows a separator feeding a cathode follower for the vertical oscillator and feeding an amplifier for the horizontal oscillator. The lower diagram

shows a separator feeding a clipper, with between them a diode limiter or leveler. The clipper acts also as an inverter for its plate signal.

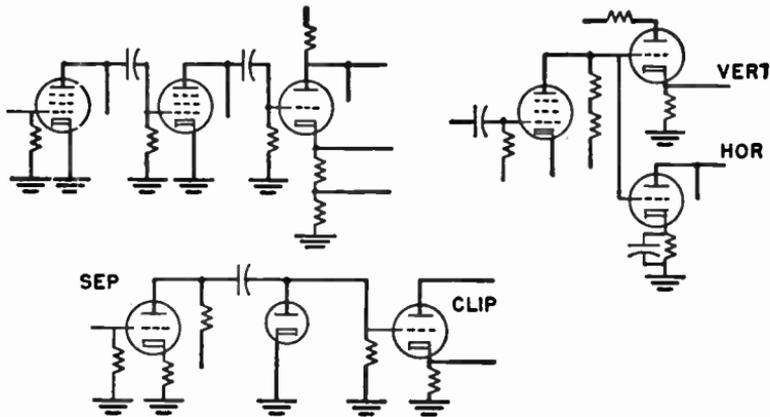


Fig. 122-3.—Connections in sync sections which provide two or more separate outputs to filters.

With input to the sync section taken at or near the video detector it becomes necessary to provide more following amplification than when a stronger signal is taken from a first or second video amplifier stage. In some receivers there are amplifiers following one or both filters. Greater amplification may be provided for horizontal synchronizing voltages than for the vertical pulses, since it is more difficult to maintain synchronization at the relatively high horizontal frequency than at the low vertical frequency.

Cathode followers may be used not only to provide low impedance couplings but also to secure or retain correct polarity of voltages going to the sweep oscillators. A cathode follower furnishes no amplification, rather there is some loss of signal strength. Therefore, an additional amplifier may be used in a branch containing a cathode follower. Also, as at the upper right in Fig. 122-2, a cathode follower might be in the line to the vertical sweep oscillator and an amplifier in the line to the horizontal oscillator.

Because a single tube may perform two or more functions, and because two or more tubes may perform similar functions, there is no standardization of names applied to tubes in the sync section. It is the intention to use names denoting the principal purpose, but the same tube may be called either a separator, an amplifier, or a clipper, and may be called by different names in different receivers.

When examining oscilloscope traces of waveforms in the sync section it usually is necessary to receive a transmitted picture or test pattern signal in order to have synchronizing pulses. For examination of vertical pulses and related actions the internal sweep of the oscilloscope should be adjusted for 30 cycles per second, which will display two vertical pulse intervals. If it is difficult to synchronize the oscilloscope for holding a vertical blanking period near the center of the screen, the sweep frequency may be changed to 20 cycles to bring three vertical pulse intervals onto the screen. Any part of the center interval then is easily examined. For examination of horizontal sync pulses and related actions the internal sweep of the oscilloscope may be adjusted to 7,875 cycles per second to display two pulse intervals, or at 5,250 cycles to display three pulse intervals.

**TELEVISION RECEIVERS.** — The major divisions or sections of a television receiver are as shown by Fig. 124-1, where directions of signal travel are indicated by arrows.

Each television program is carried through space from transmitter to receiver by two independent radio waves at very-high frequencies in the range between 54 and 216 megacycles per second or at ultra-high frequencies between 470 and 890 megacycles per second. These frequencies of transmission are called carrier frequencies.

One of the waves is varied or modulated by the signal corresponding to the moving pictures of television and providing also the means for controlling reproduction of pictures at the receiver. This composite picture and control signal is the video signal. The radio wave on which this composite signal is transmitted is called the video carrier.

The other wave is varied or modulated by the signal corresponding to speech, music, and other sounds which are to accompany the television picture. The radio wave on which the sound signal is transmitted is called the sound carrier.

The modulation or variation corresponding to the video signal is of the same kind employed for sound radio in the standard broadcast and short-wave bands. This is called amplitude modulation. The sound signal modulation is of the kind employed for frequency-modulation or f-m sound radio, it is a frequency-modulation signal.

The video carrier and its signal, also the sound carrier and its signal, are collected by the receiving antenna from radio wave energy being propagated through space from television stations or transmitters. The amplitude-modulated video carrier and the frequency-modulated sound carrier pass together from the antenna into the radio-frequency section of the television receiver.

In the r-f (radio-frequency) section the very-high carrier frequencies on which the signals have been transmitted are changed to lower intermediate frequencies which still carry the original signals for pictures, picture control, and sound. These intermediate frequencies are in the range between 20

and 50 million cycles (megacycles) per second. Then the video intermediate frequency carrying picture and control signals, and the sound intermediate frequency carrying sound signals, are passed on to the video section and the sound section of the receiver.

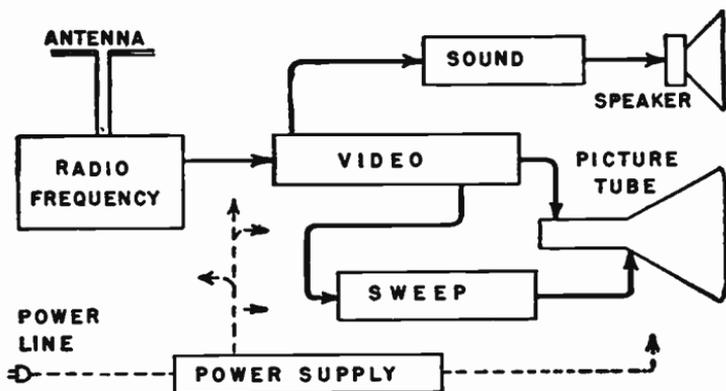


Fig. 124-1.—The principal sections of a television receiver.

In the video section the picture and control signals at video intermediate frequencies are strengthened and caused to control voltages at still lower frequencies that act in the picture tube to form lights and shadows viewed as moving pictures on the screen. These lower picture frequencies are in the range from 30 cycles per second up to about 4 megacycles per second. These are the frequencies of the picture signal itself, which have been carried as modulation on very-high and intermediate frequency waves and voltages, and which finally have been separated for use in the picture tube.

In some receivers the intermediate frequencies carrying sound signals go from the radio-frequency or r-f section directly to the sound section. More often these sound intermediate-frequency signals pass first through part or all of the video section that is handling the video intermediate-frequency signals, and then the sound signals go to the sound section.

The sound section strengthens or amplifies the signals coming to it and removes the relatively high intermediate fre-

quencies. This leaves only the signals or voltages which correspond to sounds in the television program. The sound section of the television receiver is much like that portion of an f-m sound receiver that follows the r-f section of the sound receiver.

While the video section of the television receiver is strengthening or amplifying the picture-forming signal it is also amplifying the portions of the video signal which will control reproduction of the picture. This control ensures that every light or shadow produced on the screen of the picture tube occupies the same relative position as the original corresponding light or shadow in the televised scene, so that the reproduced picture will not be scrambled.

Control of reproduction depends on correctly timing the movement or sweeping in the picture tube of the electron beam that illuminates the screen, so that this sweep will be precisely the same as that of the electron beam in the camera tube which is viewing the original scene. When the electron beams in the picture tube and camera tube are moving in the same directions over the same picture areas at the same instants of time the two beams are said to be synchronized.

Travel of the electron beam in the receiving picture tube, and also in the camera tube at the point of picture pickup, is called scanning. The sweep section of the receiver takes from the video section those control portions of the video signal which regulate travel of the electron beam in the picture tube. These portions are the synchronizing voltage pulses or the sync pulses of the signal. The sync pulses are used in the sweep section to control voltages or currents which move the electron beam in the picture tube.

The power supply section of the receiver takes energy from the electric power and lighting lines in the building, and furnishes to all other parts of the receiver whatever values of voltage and current are required for correct operation.

**Radio-frequency Section.** — Fig. 124-2 shows the principal parts of the r-f section of the television receiver as these parts are related to one another by signal travel through them.

Carrier-frequency signals from the antenna are applied to the grid circuit of a tube in the r-f amplifier and are strengthened in this amplifier. The carrier-frequency signals then pass to the grid of the mixer tube. The r-f oscillator tube and its circuit produce frequencies which usually are higher than the carrier frequencies, but may be lower in some receivers. These oscillator frequencies are applied to the grid of the mixer tube

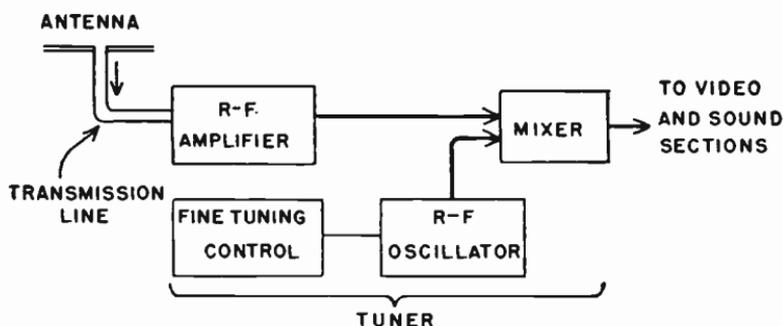


Fig. 124-2.—Principal parts of the radio-frequency section of a television receiver.

along with the carrier frequencies. The difference frequency between carrier and oscillator frequencies appears in the output of the mixer. This difference frequency is the intermediate frequency.

In the r-f section are inductances and capacitances which are tuned together for reception of programs in any of the various television channels. Selection or adjustment of inductances and capacitances for each channel is accomplished by switching or other means in the mechanism called the tuner of the receiver. With the channel selector knob or dial set for any one channel, small changes of r-f oscillator frequency may be made in some receivers by the fine tuning control. This allows improved reception in case the oscillator frequency does not at first act to produce the correct intermediate frequency by beating in the mixer tube with the carrier frequency.

**Video Section.** — The principal parts of the video section of the television receiver are shown in relation to signal travel by

Fig. 124-3. To this section comes the video intermediate-frequency signal from the mixer tube of the r-f section. R-f oscillator frequencies are so adjusted for each received channel by action of the tuner as to produce the same intermediate frequency from all channels or all carrier frequencies. Consequently, the video i-f (intermediate-frequency) amplifier operates in the same limited range of intermediate frequencies at all times and does not require any change of its tuning for reception in different channels.

The video i-f amplifier usually consists of three or four stages and tubes. All stages carry the video i-f signals. In receivers having intercarrier sound systems all stages of the i-f amplifier carry also the sound i-f signals. With dual channel sound systems the sound i-f signals may be carried through one or more, but not all the i-f stages, and then taken to the sound section.

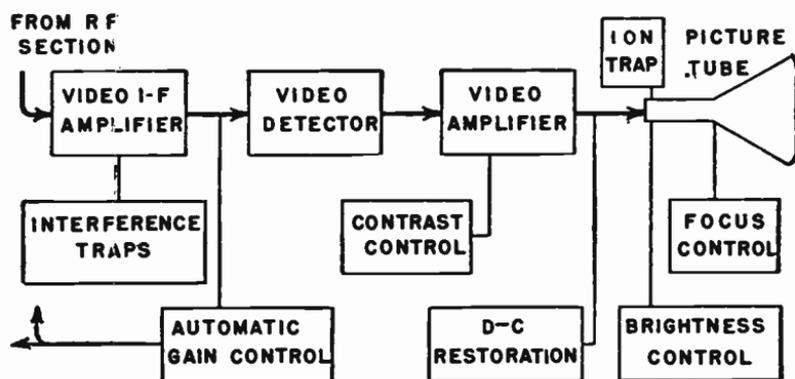


Fig. 124-3.—The video section with its accessories and controls.

Some receivers have interference traps connected or coupled into the video i-f amplifier. These traps absorb or reject signal energy at various signal frequencies which are not to pass all the way through the i-f amplifier.

The output of the i-f amplifier goes to the video detector. This detector recovers from the modulated video i-f signal the

variations which correspond to picture lights and shadows, also the synchronizing pulses. The relatively high intermediate frequencies are gotten rid of in the detector circuit, while the signal modulations are passed on to the video amplifier.

A portion of the output of the i-f amplifier may act in an automatic gain control circuit. This control regulates the amplification or gain in all or part of the i-f amplifier and possibly also in the r-f section to compensate in great measure for variations of strength in the received signal. Picture brightness and contrast thus are maintained at reasonably constant levels.

The video amplifier strengthens the composite signal coming from the video detector and applies this signal to the picture tube. The picture tube is operated with such combinations of voltages that sync pulses of the composite signal have no effect in the pictures, only the picture variations being used to control the electron beam for lights and shadows.

The contrast control varies the gain of the video amplifier to provide a picture having a full range of shades from black to white while preserving the correct relations of various gray tones. Sometimes the contrast control acts also on the video i-f amplifier, or it may be interconnected with the automatic gain control. The contrast control is adjusted during reception of programs.

With some types of signal transfer or coupling between video detector and picture tube it is necessary to employ what is called d-c restoration. This is a means for automatically maintaining correct values of tone or shading in the reproduced picture when there are changes of average brightness at the scene being televised by the camera at the point of pickup.

Inside the large end of the picture tube is a screen which is illuminated to form the reproduced pictures. The screen becomes illuminated wherever it is struck by electrons coming toward the screen in a very small beam from an internal part of the tube called the electron gun. The density of electrons in the beam is varied by the picture portion of the video signal. The greater the density the brighter becomes the illuminated spot or area on the screen.

The electron beam in the picture tube is rapidly deflected from left to right and also more gradually downward while its density is varying. Thus the lights and shadows are distributed over the screen in their correct positions. The electron beam strikes each tiny spot on the screen during only the briefest fraction of a second while being deflected, but due to persistence of illumination on the screen and persistence of vision in the human eye the impression of each illuminated spot or area remains while the entire picture is being traced by the beam.

Some types of picture tubes require a device called an ion trap. Relatively heavy ions which tend to travel to the screen along with the electrons are drawn out of the beam by this trap, thus preventing damage to the screen which would result from its continual bombardment by these heavy particles.

The brightness control acts to maintain such average density of electrons in the picture tube beam as will preserve the correct average brightness or background illumination in the picture. In some receivers the brightness control may be adjusted by the operator during reception of programs, while in other receivers this control is a service adjustment.

The focus control acts on the electron beam in the picture tube to produce the smallest possible spot of light or the narrowest possible horizontal line of light on the screen. This allows clarity of detail or good definition of lines, points, and small objects in the picture. The focus control is a service adjustment.

**Sweep Section.** — Electrical relations between parts of the television sweep section are illustrated by the block diagram of Fig. 124-4. The composite video signal coming from the video section goes to one or more tubes and circuits in a group which may be considered as a sub-section for handling the synchronizing voltages or pulses. The functions of these sync section tubes and circuits are as follows. Separate the synchronizing voltage pulses from the picture signals, and discard the picture signals. Amplify the synchronizing pulses to whatever extent may be necessary. Clip the pulse voltages, which means

making them of uniform strength even when there are variations of strength in the incoming composite signal. Invert the polarity or reverse the positive and negative directions of the pulse voltages as may be necessary for operation of the following sweep oscillators. The order in which these functions are performed will vary in different receivers.

The output of the synchronizing or sync section goes to the vertical sweep oscillator through a filter that selects pulses which are to control vertical sweeping of the beam. Output of the sync section goes also to what might be considered a filter, but in practice is an automatic frequency control for the horizontal sweep oscillator. This control keeps the oscillator precisely in time with horizontal sync pulses of received signals. The sweep oscillators consist of tubes and associated circuits which produce voltage waves of such forms, or waves which vary in such manner, as will cause correct sweeping rates and times of the electron beam.

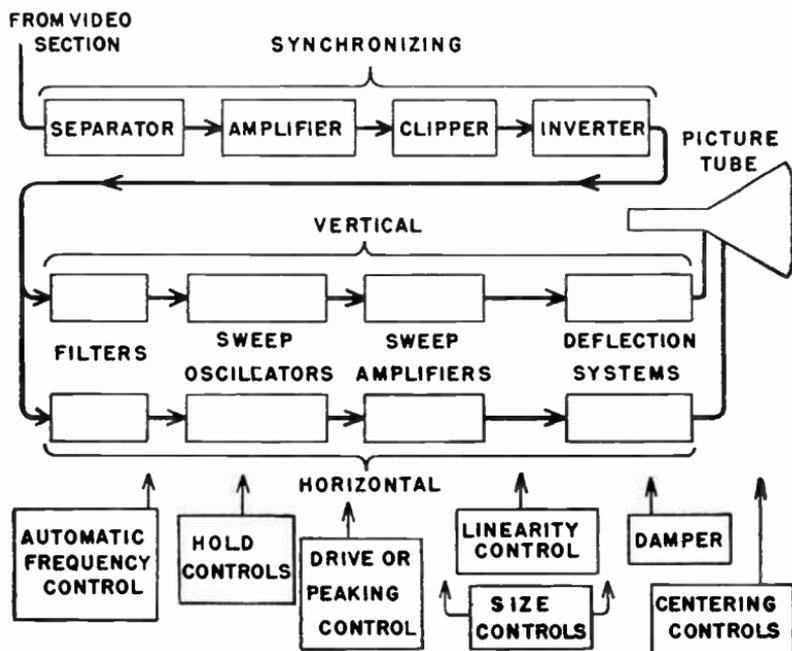


Fig. 124-4.—Principal parts in the sweep section of a typical television receiver.

In some receivers there is automatic frequency control for the vertical sweep oscillator, in addition to such a control for the horizontal sweep oscillator. Hold controls on each sweep oscillator permit adjustment of free-running oscillation rates to values which may be easily changed to the rates or frequencies of sync pulses coming from a filter or an automatic frequency control.

Output voltages from the sweep oscillators are applied to the vertical and horizontal sweep amplifiers. These amplifiers deliver from their own outputs, with the help of transformers, currents which sweep the electron beam horizontally and vertically across the face of the picture tube. Drive controls regulate strengths of voltages delivered from oscillators to sweep amplifiers, and in some cases make necessary corrections in the waveform of these voltages.

Linearity controls prevent or correct distortion of shape, size, and relative positions of objects and lines in reproduced pictures, to ensure that proportions are like those in original images or scenes being televised. Size controls allow adjustment of height and width of reproduced pictures for correct proportioning and for filling the opening of a mask in front of the picture tube face. Drive, peaking, and linearity controls are service adjustments.

Currents from sweep amplifiers and their transformers pass to the vertical and horizontal deflection system for the picture tube. The deflection system consists principally of four electromagnetic coils and cores in a yoke that fits around the neck of the picture tube. In the horizontal yoke circuit is a damper tube whose function is to prevent undesirable fluctuations or oscillations of current in the deflecting coils.

Centering controls bring the reproduced pictures within the mask opening in front of the picture tube. All receiver parts and sections which have been here mentioned and shown by diagrams are fully described by articles under their respective headings elsewhere in this book.

**TONE, CONTROL FOR.**—Tone controls in television receivers allow varying the sound reproduction for apparent emphasis of high audio frequencies, called treble tones, or apparent emphasis of low frequencies, called bass tones. With simple tone control circuits such as commonly employed the effect of bass emphasis is secured by attenuating middle and high frequencies, while treble emphasis is secured by attenuating or weakening middle and low frequencies.

Tone controls consist of one or more fixed capacitors and usually at least one adjustable resistor connected between an audio voltage amplifier and output amplifier as in Fig 1,

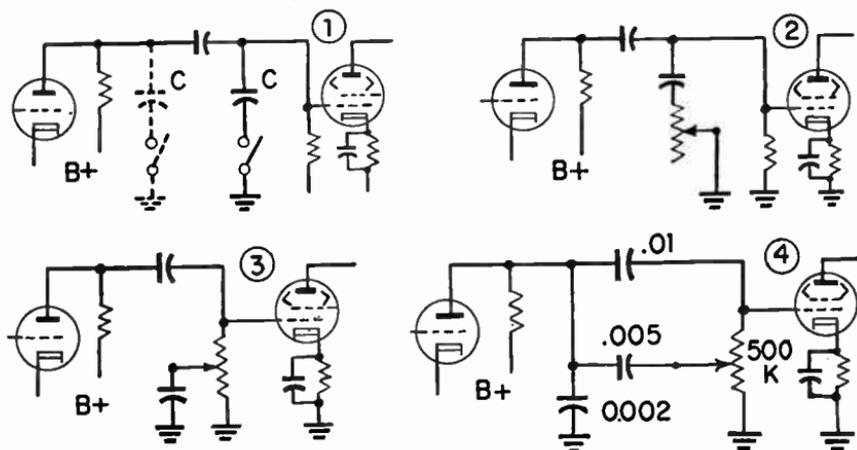


Fig. 1.—Simple tone control circuits such as found in many television receivers.

or between a volume control and the grid of the output amplifier. A capacitor connected as in diagram 1, from the high side of the audio amplifier circuit to ground, attenuates high frequencies more than low frequencies and thus emphasizes the bass. High frequency attenuation results from the fact that reactance of a capacitor decreases as frequency increases.

In diagram 2 the effect of the bypass capacitor may be varied by an adjustable series resistance. Decreasing the resistance increases high frequency attenuation and leaves low frequencies relatively stronger. In diagrams 3 and 4 the bypass capacitor is connected to an adjustable resistor in the grid

return of the output amplifier. Moving the control slider toward the grid connection increases high frequency attenuation and thus emphasizes low frequencies.

With simple volume controls of the general style shown at 1 in Fig 2 there is apparent weakening of bass tones, also some apparent loss of treble, when the control is adjusted for low volume. This effect is due to the human ear having less sensitivity for both bass and treble tones at low volume levels than at higher volume.

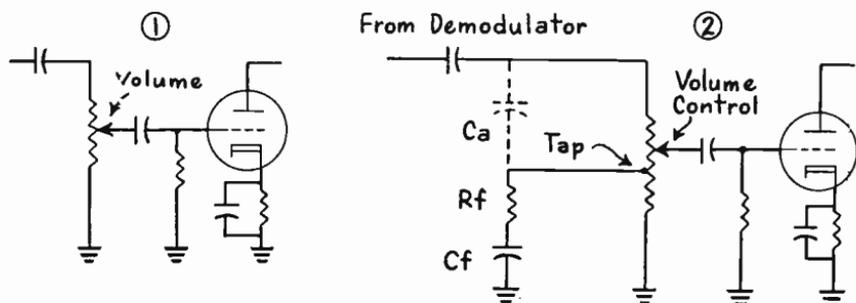


Fig. 2.—A compensated volume control helps maintain equality of low, middle, and high frequency sound at low volume.

A compensated volume control, such as the one of diagram 2, increases attenuation of middle audio frequencies in relation to lows and highs as volume is reduced, and maintains more uniform reproduction at all sound frequencies. The volume control potentiometer is tapped at one-fifth to one-half its total resistance, measured from ground. Resistor  $R_f$  and capacitor  $C_f$  act as a tone filter. Capacitor  $C_a$  may or may not be used. There is maximum compensation with the control slider at the tap position, and least compensation with the slider adjusted for maximum volume.

**TRACING, TV SIGNAL.** — Signal paths from antenna terminals to picture tube video input may be checked for continuity, and amplifiers for gain, by using a modulated r-f signal generator and an audio amplifier as in Fig. 125-1. The audio amplifier usually is a signal tracer of the type commonly used for following audio signals through sound radio receivers. The signal generator may be connected through a matching pad to the antenna terminals of the receiver and

## TRACING, TV SIGNAL

tuned on fundamentals or harmonics within the range of carrier frequencies of any channel to which the receiver is tuned. Otherwise the generator may be connected through a small capacitor to the mixer grid or to the grid of any i-f amplifier, and tuned within the intermediate-frequency range of the receiver without regard to channels.

Input to the signal tracer or amplifier is through a detector probe on the end of a shielded cable. The probe must not seriously load or detune circuits to which connected. A suitable low-capacitance probe is shown by Fig. 125-2. Capacitor

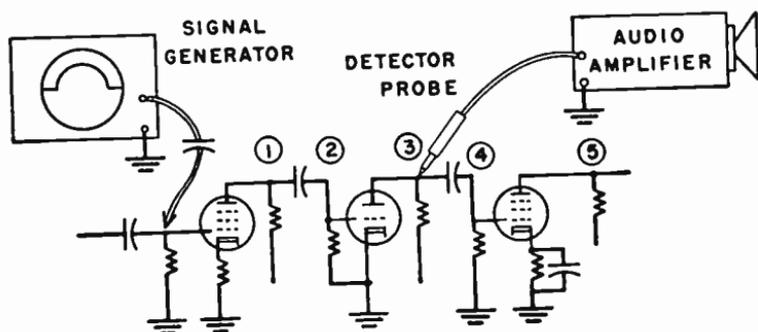


Fig. 125-1.—Connections for testing with a signal tracer.

$CA$  is of one to three mmf value, resistor  $RS$  is of about five megohms, and resistor  $RB$  of 0.1 to 0.2 megohm. The crystal diode may be a general-purpose type, such as the 1N34.

The vertical sweep oscillator should be removed or disabled, to avoid pickup by the probe tip of confusing 60-cycle buzz. The horizontal oscillator causes only a slight hissing sound, which should not be troublesome. The signal generator may be modulated with any audible frequency. Before commencing a series of tests the probe tip should be touched to the high side of the generator output to make sure a tone signal can be heard.

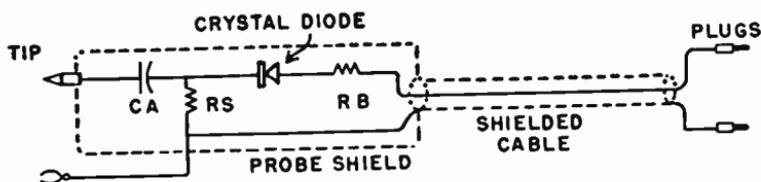


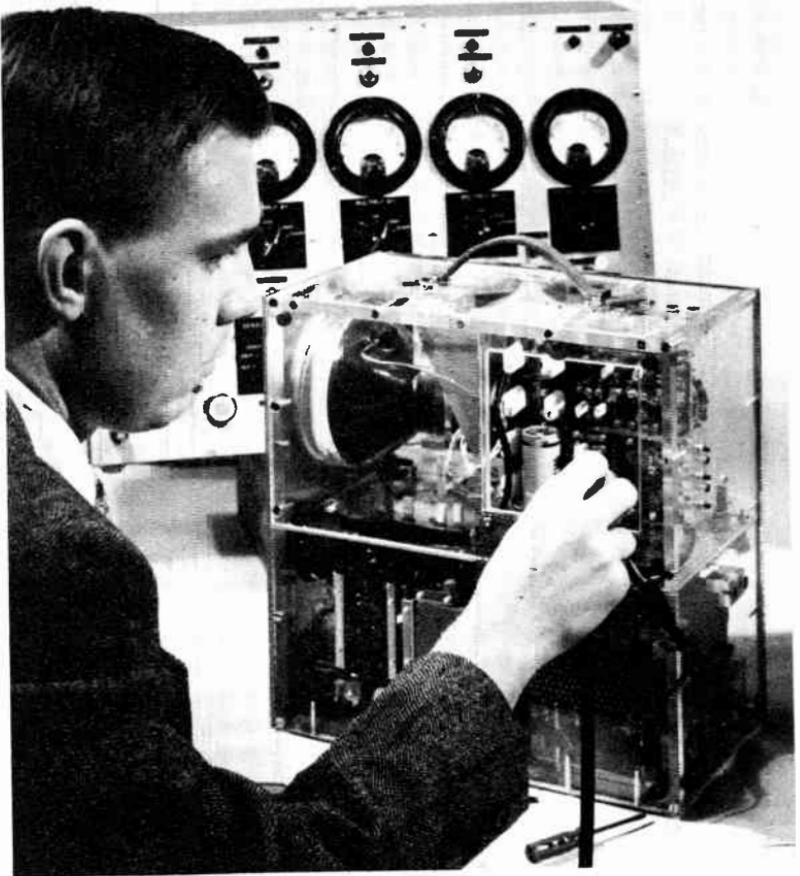
Fig. 125-2.—Detector probe suitable for use with a signal tracer.

With the probe at the output of the first stage following the signal generator connection the generator is tuned for loudest sound from the tracer or amplifier, assuming that the stage passes a signal. Then proceed to following grids and plates as in Fig. 125-1. Due to detuning effect of the probe, sound may be weaker at a grid than at a preceding plate. Generator output should be low enough to prevent hearing the modulation tone with the probe held an inch or so from amplifier circuit wiring and elements.

This method of testing shows only whether a signal passes through tested stages, not whether regularly received signals would be satisfactorily reproduced. A fault is indicated between a point at which one modulation is audible and a following point at which sound disappears.

Sound and audio sections cannot be satisfactorily tested because the generator does not furnish a frequency-modulated signal. Tests in a sync section may be satisfactory if generator modulation can be made of 60-cycle frequency. Tests in the horizontal afc and sweep sections would be useful only were modulation to be at the line frequency of 15,750 cycles per second, but this frequency would be inaudible. Signal tracing in sync, sweep, and deflection sections is best carried out with an oscilloscope which shows signal waveforms. Methods of making such observations are explained in the article on *Oscilloscopes*. Typical waveform traces as obtained at various points throughout a receiver are illustrated in the article on *Waveforms*.

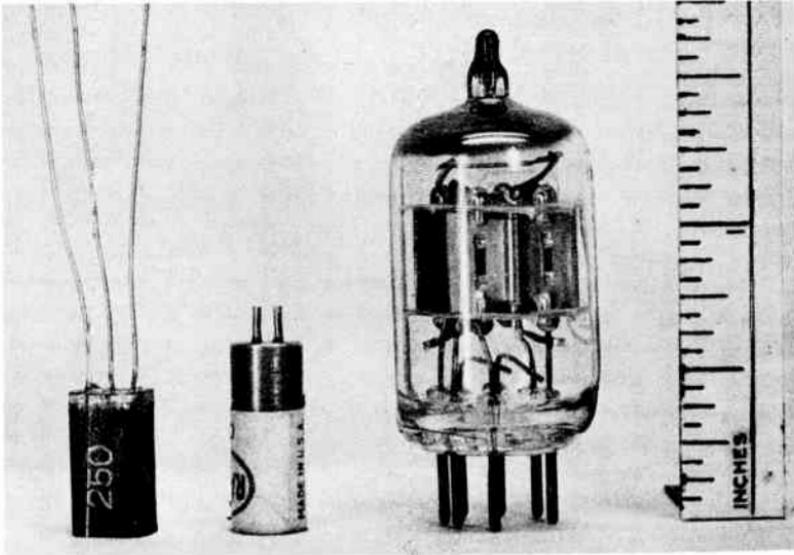
**TRANSISTORS.** — A transistor consists essentially of a tiny crystal of the element germanium to which have been added minute quantities of certain other elements. Attached to the



*Experimental Portable TV Receiver with only one tube—the Kinescope Picture tube.*

crystal are terminals or leads for convenience in making connections to external circuits. The crystal itself is molded in plastic, glass, or both, for protection and mechanical stability, and there may be an outer shell or base of metal. Some of these

assemblies weigh about one-sixth as much as the smallest tube used in television receivers and have about one-fifteenth the bulk. The transistor may be used as an amplifier, as an oscillator, and for many other purposes served by tubes.



*Fig. 1.—At the left is a junction transistor, with next to it a point-contact transistor. The miniature tube is a 6AL5 twin-diode type.*

In addition to being of diminutive size and weight the transistor does not operate with a hot cathode and consumes no heater or filament power. A transistor can withstand mechanical abuses which would ruin a tube, and unless mistreated electrically its life could be about 40 years when used five hours a day, every day of every year.

Transistors have shortcomings which partially offset their advantages. Maximum operating frequencies of types commonly available are limited to a few megacycles or, at best, to a few tens of megacycles. Transistors have limited power handling ability; it is sufficient for many applications, but prohibits the use of transistors instead of tubes for many of the important circuits in television and radio receivers. These,

and other less serious disadvantages, are being gradually overcome as development proceeds.

**Crystals for Transistors.** — The germanium crystal used in a transistor differs from that for a germanium crystal diode in being composed of two varieties of the material, whereas the diode crystal is of the same composition throughout. The two modifications of transistor crystal material are produced by adding to the germanium either of two general classes of other elements. How these additions affect the crystal may be explained as follows.

First to be considered is the fact that the outermost shell of each atom of germanium has, normally, four electrons. It is these outer electrons that link together in the crystal structure. When the crystal is pure germanium, with four outer electrons per atom, all of these electrons link together with none left over.

Certain added elements have five electrons in the outer shell of each atom. At every place in the crystal where such an atom replaces a germanium atom, four outer electrons of the intruder link with outer electrons on neighboring germanium atoms. The fifth is left over and becomes a free electron, wandering about or traveling one way or another when influenced by electric fields or potentials applied across the crystal.

Among the elements having five outer electrons per atom are antimony, arsenic, and phosphorous. When these are added to the germanium the resulting crystal is called an N-type, because it possesses excess negative charges, the free electrons.

Other elements which may be added to the germanium, such as aluminum, have only three electrons in the outer shell of each atom. At each point where an atom of such an element takes the place of a germanium atom there are only three outer electrons, while all around it are germanium atoms having four outer electrons. At every one of these points there is, consequently, a deficiency of one negative electron. Such a

deficiency constitutes a positive charge, so the crystal has positive charges wherever there is one of the intruder atoms.

The absence of an electron is called a "hole." A hole is a positive charge. This positive charge attracts negative electrons which are in surrounding atoms, and the hole may be filled by a negative electron to complete the group of four outer electrons which normally would be at every atom position.

When a negative electron fills one hole it leaves another hole at the point from whence it came. Thus, although atoms of the solid body which is the crystal cannot move, the holes do move from place to place. It is as though you were to fill one hole in a lawn with a spadeful of earth taken from some nearby point. The hole, in effect, would have moved from the point filled to the place from which you took the earth.

Electrons which are caused to move within the transistor crystal may flow in external circuits. But since the conductors of such circuits lack the peculiar properties of a P-type crystal, the holes do not leave the transistor. They do, however, influence the resistance and impedance of the transistor during the application of direct and alternating voltages.

**Transistor Elements.**—In the transistor crystal are three sections called the emitter, the collector, and the base. They may be employed similarly to the three elements of a triode tube, as illustrated by Fig. 2. The emitter of the transistor may be connected into various circuits in much the same way as the cathode of a tube. The collector section or element of the transistor is employed or connected similarly to the plate of a

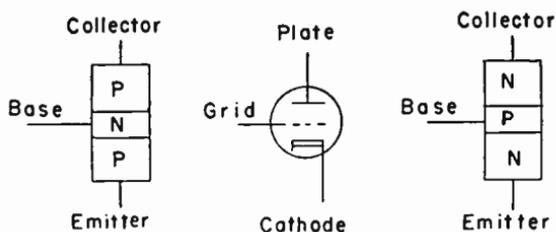


Fig. 2.—Transistor elements function similarly to the elements of a triode tube.

tube. The base element of the transistor acts somewhat similarly to the grid of a tube.

Since the transistor has three sections or elements, but there are only two basic kinds of crystal material, two of the transistor elements must be of the same basic type. That is, two may be of P-type and one of N-type, or else two may be N-type and one P-type. But when two elements, such as emitter and collector, are of the same basic crystal material, they are not exactly alike nor interchangeable. The part which is to be the emitter has been treated during manufacture to perform best as an emitter, while the part designated for the collector has been prepared especially for such service and would not make a satisfactory emitter.

**Junction and Point-contact Transistors.** — Considered from the structural standpoint there are two distinctly different methods of building transistors, one resulting in a junction transistor and the other in a point-contact transistor.

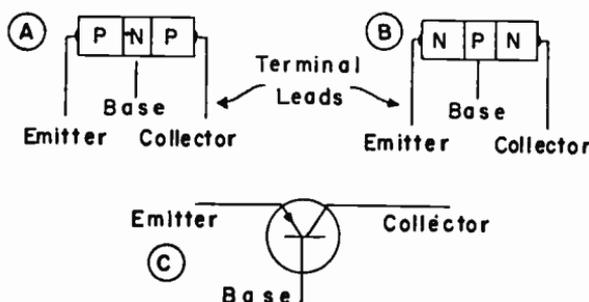


Fig. 3.—Element arrangements for junction transistors, and the symbol for this general type.

The structure of a junction transistor is shown in principle by Fig. 3. Opposite sides of the single piece of crystal are of the compositions which are to act as emitter and collector. In between is a section, possibly one-thousandth inch thick, which will act as the base. If, as at *A*, emitter and collector are of P-type material, with the base N-type, we have what is called a PNP junction transistor. When, as at *B*, the emitter and

collector are of N-type material, with the base P-type, we have an NPN junction transistor.

The usual symbol for either kind of junction transistor is shown at *C*. The base is represented by a straight line, with emitter and collector represented by lines at angles to the base. The slanting line for the emitter is marked with a small arrowhead, which sometimes looks like a dot on diagrams. The collector line has no such identification.

With the second method of transistor construction the internal connections to emitter and collector are made through the ends of very small wires in contact with the surface of the crystal at points very close together. This makes what is called a point-contact resistor, illustrated in principle by Fig. 4. As

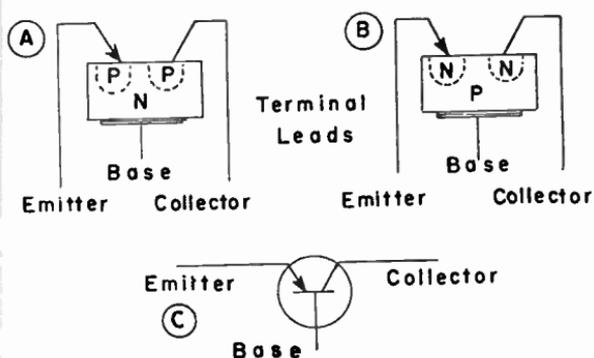


Fig. 4.—Construction and element arrangement for point-contact transistors, and the usual symbol.

actually constructed the crystal of the point-contact transistor may be only about 1/20 inch diameter or square, and about 1/50 inch thick. Contacts of the two wires on the crystal surface are separated by only a few thousandths of an inch.

The design represented at *A* of Fig. 4 employs an N-type crystal; the main body of the crystal being of such material. During manufacture there is a forming process which produces at each contact position a small area of P-type material. Therefore, the contact points for emitter and collector actually connect to P-type material, between which is the N-type body

of the crystal. This makes the N-type point-contact transistor equivalent in many respects to the PNP junction type. That is, emitter and collector leads go to P-type material, while the base lead goes to N-type material.

In diagram *B* the body of the crystal is of P-type material. There are small areas of N-type material formed at the emitter and collector contact points. This gives, in many respects, the equivalent of an NPN junction transistor, at least so far as crystal materials and terminal connections are concerned.

At *C* of Fig. 4 is the usual symbol for any kind of point-contact transistor. The base is represented by a straight line, the emitter and collector by two slanting lines. The emitter is identified by an arrowhead placed right at the contact on the base. The collector is represented by a plain slanting line.

**Transistor Circuits.**—Inasmuch as the three elements of a transistor correspond roughly to the three elements of a triode tube, it is natural to find the transistor used in the same general types of circuits as the triode. The most familiar triode tube circuit is shown at the left in Fig. 5. The input signal is applied between grid and cathode, while the output signal appears between plate and cathode. Average grid potential is determined by a grid biasing voltage, and average plate potential with reference to the cathode is fixed by a d-c plate voltage. The cathode is common to both input and output, and may be connected to ground.

The equivalent transistor circuit is shown at the right. Input is applied between base and emitter. Average d-c base poten-

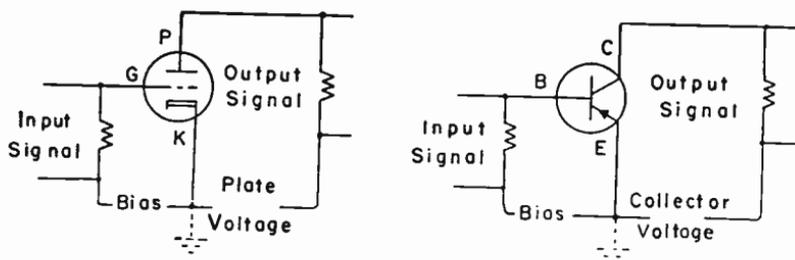


Fig. 5.—Grounded cathode circuit for a triode tube (left) and grounded emitter circuit for a transistor (right).

tial, with reference to the emitter, is determined by bias voltage. Output is taken from between collector and emitter, with average collector potential fixed by d-c collector voltage. This is called a grounded emitter circuit, because the emitter is common to both input and output, and may be grounded. When specifying a general type of transistor circuit, and referring to a certain element as being grounded, the reference is to the element that is common to both the input and the output signal circuits. In practice, this element need not actually be grounded.

Another fairly common triode circuit employs a grounded grid, as shown by the simplified diagram at the left in Fig. 6. Input is between cathode and grid, by way of ground, with bias voltage applied between cathode and ground, which is equivalent to applying the bias between cathode and grid. Output is from between plate and grid, again by way of ground. Average plate potential is determined by d-c plate voltage.

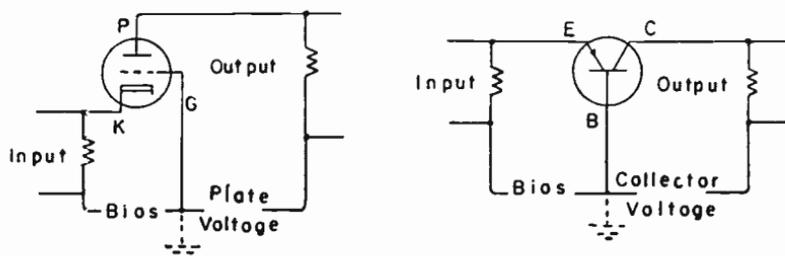


Fig. 5.—Grounded grid triode tube connections (left) and grounded base connections for a transistor (right).

The transistor equivalent for the grounded grid tube circuit is shown at the right. Input is applied between emitter, corresponding to the tube cathode, and the base, comparable to the tube grid. Average emitter potential, with reference to the base, is determined by a bias voltage in the input circuit. Output is taken from between collector and base, with average collector potential fixed by the d-c collector voltage. This is

called a grounded base circuit, because the base is common to both input and output, and may be connected to ground.

A third familiar circuit used with triode tubes is called the cathode follower. The tube version is shown at the left in Fig. 7, and the comparable transistor circuit at the right. When using a triode tube the signal input is between grid and the grounded side of the plate circuit. Average grid potential is determined by a biasing voltage in the input circuit. Signal output is taken from between the cathode and the grounded side of the plate circuit. Plate potential is maintained by a source of d-c plate voltage. It is the plate that is common to both input and output.

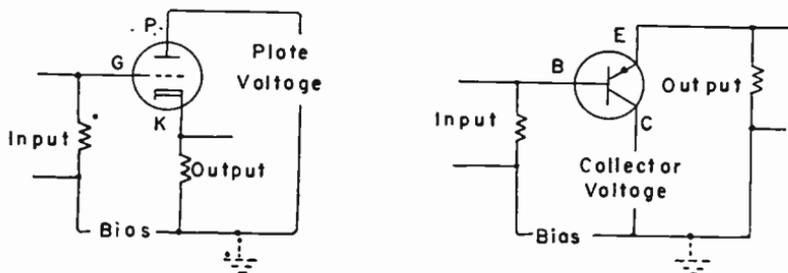


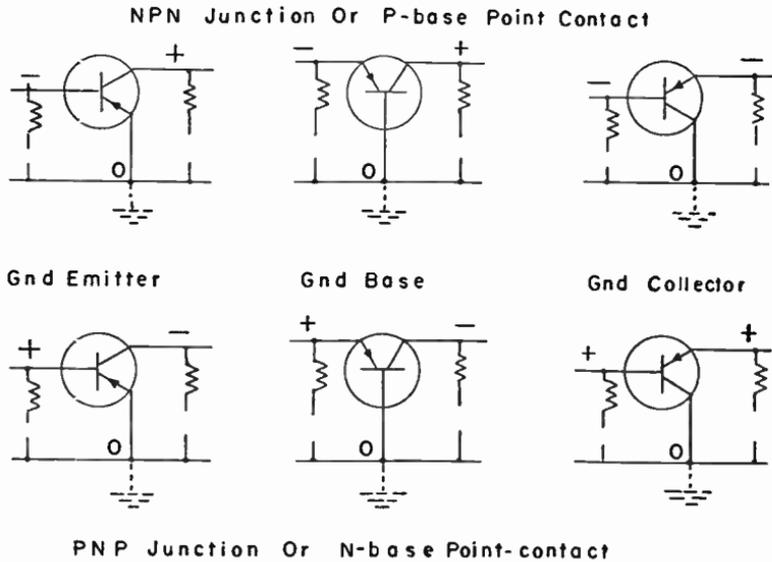
Fig. 7.—Cathode follower circuit for a triode tube (left) and the grounded collector circuit for a transistor (right).

Transistor connections equivalent to the cathode follower provide what is called a grounded collector circuit. Signal input is applied between base and collector, comparable to grid and plate of a tube. A biasing voltage maintains average base voltage, with reference to the collector. Average collector potential is fixed by a source of collector voltage.

**Element Polarities.**—The polarities of the high-side elements for signal input and output of the transistor, with reference to the common or grounded element, depend on the type of transistor and on the kind of circuit in which it is used. The relative polarities for types of transistors and circuits are shown by Fig. 8. These polarities are indicated as positive or negative when measured at the high-side elements themselves,

not at the sources of d-c voltages of bias for the input side or of d-c voltages for the output element.

Along the upper row of diagrams are shown polarities for NPN junction type transistors or for point-contact types with P-material in the body of the base. The relative polarities are the same as for tubes used in comparable circuits. For example, at the right the grounded emitter circuit is comparable to the grounded or common cathode circuit for a triode tube. The transistor base, or tube grid, is negative with reference to the transistor emitter, or tube cathode. The transistor collector (tube plate) is positive with reference to the emitter (tube cathode).



*Fig. 8.—Transistor element polarities with reference to the common or grounded element.*

The lower row of diagrams shows relative polarities for PNP junction transistors or for point-contact types with N-type base. Polarities of the high-side input and output elements, with reference to the common or grounded element, are exactly opposite to those for opposite kinds of transistors

shown in the upper row of diagrams and, accordingly, are opposite to those used with tubes in comparable circuits.

Polarities of the high-side output elements, either collector or emitter, always are as shown with reference to the common or grounded element. Polarities of the high-side input elements sometimes are reversed in practice. This may occur when the biasing voltage, at the element itself, is 0.2 volt or less. Gain may be greater or distortion lessened or eliminated by using a bias voltage of polarity opposite to that shown by the diagrams of Fig. 8.

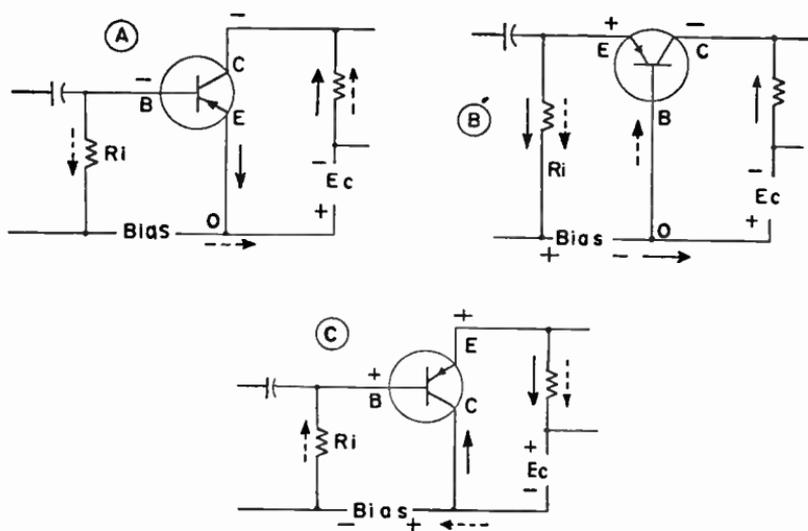


Fig. 9.—Directions of electron flows in transistor circuits.

**Electron Flows.** — Fig. 9 shows directions of electron flows in typical circuits having PNP junction transistors or N-base point-contact types. There are d-c electron flows in all three elements of the transistor, including the base element which is comparable to the grid of a triode tube. There is no electron flow in the grid of a triode when bias is more negative than a very small fraction of a volt.

It is convenient to consider the transistor circuits as having two paths for electron flow, with these paths separate in some

conductors and joining in other conductors. In the paths indicated by full line arrows on the diagrams there is relatively large electron flow, and in the paths indicated by broken-line arrows there is smaller electron flow. The larger flows never are more than a few milliamperes, and the smaller flows may be on the order of 10 to a few hundred microamperes. These are the d-c electron flows, they are not alternating signal currents.

In the grounded emitter circuit at *A* the larger flow is from collector to emitter within the transistor, and in the external circuit it is from emitter through the source of collector voltage,  $E_c$ , and the output load resistor back to the collector. The smaller flow is from collector to base within the transistor, and in the external circuit this flow is from base through the input resistor,  $R_i$ , the source of bias voltage, and thence through the external output circuit to the collector.

The direction of electron flow through resistor  $R_i$  is such as to make the base negative with reference to the emitter. Under some conditions of operation the voltage drop across  $R_i$  provides suitable bias for the base without the need for any external source of bias voltage.

In the grounded base circuit of diagram *B* the larger electron flow is from collector to emitter within the transistor, and in the external circuit this larger flow passes through input and output resistors and the voltage sources from emitter to collector. The smaller flow is from base to emitter within the transistor, and in the external circuit is through only the input resistor and source of bias voltage.

In this grounded base circuit the input resistor,  $R_i$ , is carrying the total of both electron flows in such a direction as would tend to make the emitter negative with reference to the base. Since it is desired to have the emitter positive with reference to the base, the bias voltage must be great enough to counteract all of the drop in  $R_i$  and leave enough positive potential at the emitter for the desired bias of this element.

With the grounded collector circuit at *C* of Fig. 9 the larger

and smaller electron flows pass through the same transistor elements and the same portions of the external circuits as in diagram *A* for a grounded emitter. But with the grounded collector all directions of electron flows are reversed and all polarities are reversed in comparison with the grounded emitter arrangement. When the collector is grounded, electron flow through resistor  $R_i$  is in the direction that makes the base positive with reference to the collector. When voltage drop across  $R_i$  is more than the desired bias at the base, part of this voltage is counteracted by the external source of bias voltage.

Because there always is electron flow in the high-side element of the transistor connected to the input, there always is a self-biasing effect when this flow passes through any appreciable external resistance in the input circuit. Any source of external biasing voltage may subtract from or add to the self-bias. Consequently, the polarity in which this external source is connected to the input circuit may be the opposite of that which, apparently, will provide the desired bias potential at the high-side input element. This occurs when the external bias voltage must counteract part of the self-bias voltage.

**Impedance Relations.** — For maximum transfer of a-c power or signal power from a source to a load the impedances of source and load must be equal. There is rapid decrease of power transfer as the load impedance is made less than the source impedance, and a less rapid decrease as impedance of the load is made greater than that of the source.

In a single amplifier stage there are two sources and two loads requiring matching of impedances. There is first the signal source connected to the transistor input, for which the input impedance of the transistor forms the load. The transistor itself is the second source, at its output terminals. Impedance at the transistor output requires matching to that of the load connected across the output terminals.

Transistor input and output impedances depend on the kind of transistor, whether PNP or NPN junction styles, or P- or N-base point-contact styles. Impedances depend also on

the circuit in which the unit is connected, whether grounded emitter, grounded base, or grounded collector, and impedances at both input and output vary with the make and model of transistor. Consequently, each application becomes a problem in itself, to be solved by reference to specified characteristics or by adjustment of impedances in the signal source and the signal load.

With resistance couplings having blocking and bypass capacitors large enough to maintain low reactances at all applied frequencies there is but little variation of impedances connected to transistor input and output. With transformer coupling there may be large variations of winding impedance unless the transformers are designed for fairly flat inductive reactances and impedances over the range of applied frequencies.

Transistor output impedances are greater than their input impedances when using grounded emitter or grounded base circuits. With the grounded collector circuit the output impedance is less than the input impedance, just as with a cathode follower circuit using a tube.

With grounded emitter and grounded base circuits the ratio of output to input impedances may be only moderately large or it may be very large. In general, the ratio of output to input impedances is much greater for junction than for point-contact transistors, and it is greater when using a grounded base circuit rather than a grounded emitter with either kind of transistor. Input impedances range from less than 100 ohms up to about 2,000 ohms. Output impedances may be as low as 4,000 to 5,000 ohms, or as high as several megohms.

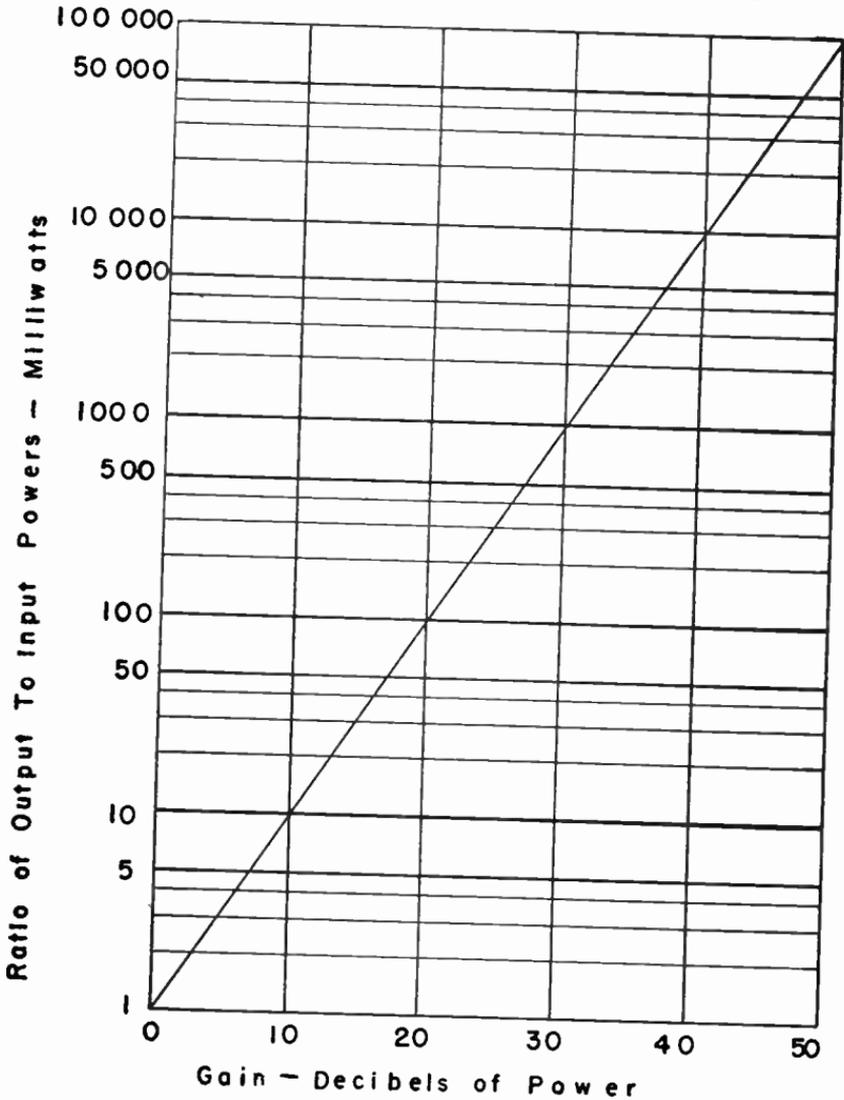
**Gains of Power and Voltage.** — Gains of transistor circuits usually are specified in decibels of power, instead of as multiplications of voltage such as specified for most tube circuits other than those for audio power and output amplifiers.

Fig. 10 gives relations between power gains in decibels and gains measured as the ratio of output to input powers as measured in milliwatts, or in any other common unit of power.

## TRANSISTORS

Milliwatts are shown on the chart because power outputs and permissible power dissipations for transistors usually are measured in milliwatts.

Power applied to and used in the input circuit is directly proportional to the input impedance when input signal cur-



rent remains constant and when it is assumed that there is reasonably good matching of source and transistor impedances. Then only small power will be needed at the input when input impedance is small. If output signal current is constant, output power is directly proportional to output impedance, and with high output impedance there will be high output power for any given current. Were input and output currents to be alike, the ratio of gain would be equal to the ratio of output to input impedances.

With input and output signal currents which are not alike, the input and output powers are proportional to the squares of the currents, as well as being directly proportional to the impedances. Actual power gains depend not only on characteristic impedances of the transistor, but to a great extent on how well these impedances are matched by the source and the load, on the type of circuit employed, and on the relative potentials at the high-side input and output elements with respect to the common or grounded element. Gain which may be realized in practice is a matter of correct circuit design, as would be the case when employing tubes.

Junction transistors are capable of giving more power gain than point-contact types, when both kinds are in the same types of circuits, possibly 150 to 200 per cent more. There is little difference between the possible gains of grounded emitter and grounded base circuits when using the same type of transistor in both. Power gain in grounded collector circuits is decidedly less than in either of the other circuit types, decreasing to two-thirds with point-contact transistors and to around 40 per cent with junction types, as compared with gains of the same transistors used in grounded emitter or grounded base circuits.

There so many variables that it is impossible to give any definite numbers of decibels as representing typical power gains. However, the average of gains for a number of well designed circuits of the grounded emitter and grounded base type is about 40 decibels when using junction transistors and

about 25 decibels when using point-contact types. The average power gain for a few grounded collector circuits is on the order of 15 decibels when using either type of transistor.

**Power Dissipations and Outputs.** — Transistors, like tubes, are rated for certain maximum permissible power dissipations—usually in watts for tubes and in milliwatts for transistors. Dissipation in milliwatts is equal to the product of voltage measured from the element in question to some other element, and the current in milliamperes which flows in the element considered. If there is current of more than a few microamperes in other elements it may be necessary to consider the dissipation at these other elements in arriving at the total dissipation, which is the number of milliwatts of power producing heat in the transistor.

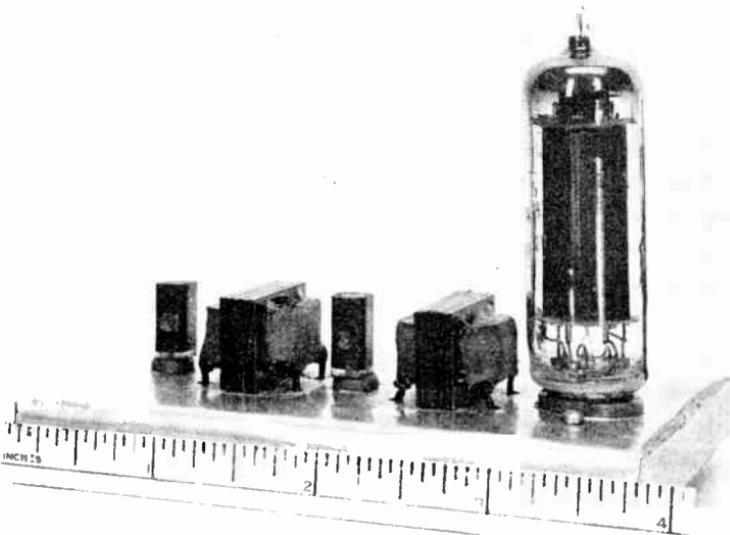
Permissible element dissipations vary between wide limits in various makes and types of transistors. For example, maximum collector dissipations may be almost anything between 20 and 200 milliwatts in units commonly employed, or still greater in other kinds. Maximum safe dissipation depends chiefly on cross section of the transistor crystals and on concentration of currents in the crystal. In general, the larger the unit the greater is the safe dissipation.

In addition to having ratings for maximum power dissipations, transistors are rated also for maximum voltages and for maximum currents at the various elements. Although power dissipation is the product of voltage and current, neither of the latter values may be exceeded regardless of the resulting power. As an example, consider a transistor rated for 40 maximum milliwatts, maximum current of 5 ma, and maximum voltage of 20 at some element. Using maximum current and maximum voltage would result in 100 milliwatts dissipation, and would destroy the unit. To remain within the 40-mw limit while using 20 volts the current could not exceed 2 ma, and while using 5 ma the voltage could not exceed 8, since the products of these currents and voltages equal the maximum milliwatts of dissipation.

In applications where signal distortion is an important consideration the operating voltages, currents, and dissipations must be kept well below the rated maximums. In some oscillator circuits, where the output need not be of sine-wave form, the d-c voltages and currents may approach maximum rated values.

**Frequency Response.** — As operating frequency is increased, the gain of transistor circuits decreases because of two actions in the transistor itself. The first is the time required for electrons to move through the crystal material and for holes to shift from place to place. These movements are much slower than through the vacuums between elements of an electronic tube. This "transit time" through the crystal depends to some extent on the exact composition of the P- or N-type materials.

With any kind of crystal material, transit time in point-contact transistors is lessened by reducing the spacing between contacts. In junction types the time is lessened by smaller dimensions between end contacts and by a thinner base layer.



**Fig. 11.**—A two-stage amplifier in which transistors feed a 6AQ5 beam power audio amplifier tube, at the right.

The constructions which decrease transit time increase the internal capacitances of the transistor. The resulting small capacitive reactances become smaller and smaller with increase of frequency, and thus help limit the response.

For maximum or near maximum gain the commonly available junction transistors are limited to the audio range of frequencies, although gain still great enough to be useful may extend to three megacycles or even slightly higher.

Point-contact transistors have satisfactory response through higher frequencies than junction types, being useful at standard broadcast intermediate and carrier frequencies, and, with some kinds, for television intermediate frequencies. At these relatively high frequencies the gain of transistor circuits drops very rapidly, but this is true also with electronic tubes in television i-f amplifiers.

**Amplifiers.** — Fig. 11 is a picture of a two-stage transformer-coupled amplifier using two junction transistors to feed an output beam-power amplifier of the type often used in television receivers. In the circuit, Fig. 12, the transistors are connected with emitters grounded. To match the higher output impedance of the first transistor to the lower input impedance of the second transistor it is necessary to use an interstage coupling transformer with step-down turns and impedance ratio.

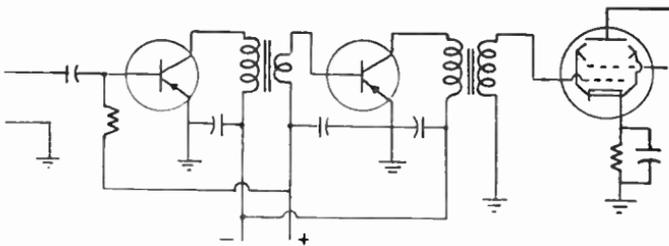


Fig. 12.—The two transistors are used in grounded emitter circuits. The first coupling transformer has a step-down turns ratio for matching of transistor impedances.

Positive biasing voltage is provided for the transistor base circuits. This external bias may be made zero or negative by opposing voltage drops across the resistor and across the resistance of the transformer secondary in the base circuits. Negative voltage is provided for the collectors of the two transistors. The bypass capacitors on output and input circuits of the transistor may be in the d-c power supply instead of in the amplifier section.

At frequencies above the audio range, iron-core coupling transformers is usual construction cause excessive energy losses. For higher frequencies it may be possible to use double impedance couplings somewhat similar to those employed for wide-band amplification in some television i-f amplifiers, with one inductor providing high impedance for the collector output and the other providing lower impedance for the base input, with energy transfer through a blocking and coupling capacitor.

Resistance couplings and impedance couplings present difficulties in matching output to input impedances of the transistors. When using such couplings, as illustrated by Fig. 13, it is assumed that the capacitance of blocking capacitor  $C_b$  is so great, and its capacitive reactance so small at signal frequencies, as to provide what amounts to a direct connection

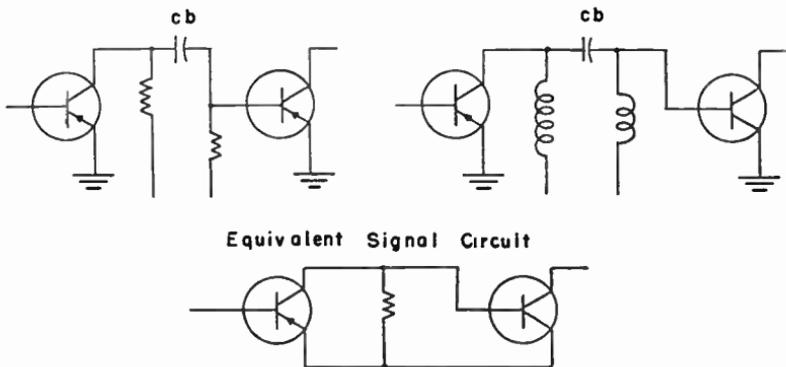


Fig. 13.—Resistance and impedance couplings have equal output and input impedances for signal voltages and currents.

for signal currents and voltages. This leaves the output load resistance or impedance and the following input resistance or impedance in parallel with each other so far as signals are concerned, and the effective resistance or impedance must be less than the value in the smaller unit.

This combined or parallel resistance or impedance is in the output circuit of the first transistor and in the input circuit of the following transistor. Thus there is the same impedance for both output and input, and a match on both sides cannot be had when output and input impedances of the transistors are different.

In resistance-coupled tube circuits this difficulty is not present or, at least, is not serious. The input impedance of tubes with negatively biased grids always is high, while plate resistance or output impedance in the tube is relatively low. Then a combination of coupling resistors may be selected to give a parallel resistance which is a fair match for the output impedance of the first tube. This rather low parallel resistance will be in the grid circuit of the following tube, but energy transfer from a lower to a higher impedance is quite good unless there is a very great difference. With transistors we would have transfer from a higher to a lower impedance, which

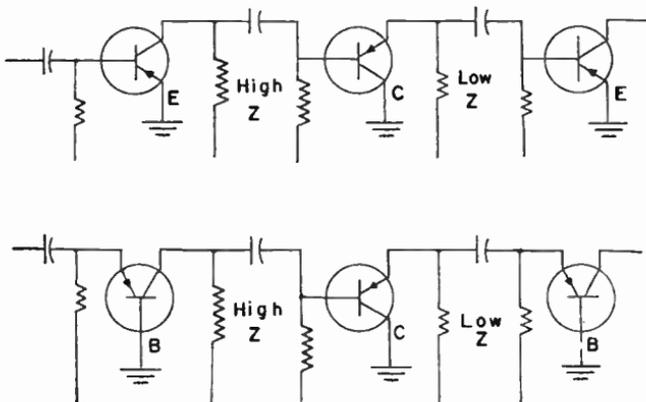


Fig. 14.—Impedances may be matched fairly well by using a grounded collector circuit between two circuits of other types.

means that much more power would be dissipated in the first unit than passed along to the second.

Because input and output impedances are not so different in the grounded emitter circuit as in the grounded base circuit, resistance or impedance coupling allows better matching with the grounded emitter circuit.

Still another method of obtaining reasonably good impedance matching utilizes the grounded collector circuit as a sort of step-down transformer, much as it is used with cathode follower tubes in radio and television apparatus. In the upper diagram of Fig. 14 a grounded collector circuit is between two grounded emitter circuits. There is high coupling impedance for the output of the first transistor and the input to the grounded collector transistor. There is relatively low coupling impedance between output of the grounded collector circuit and input to the following grounded emitter circuit. In the lower diagram there is a grounded collector circuit between two grounded base circuits.

**Signal Polarity Inversion.** — In television video amplifiers, between the video detector and picture tube grid-cathode circuit, the matter of polarity inversion is important. Inversion or lack of it must be such as brings picture signals in positive polarity to the grid or in negative polarity to the cathode of the picture tube. Polarity is important also in television sync

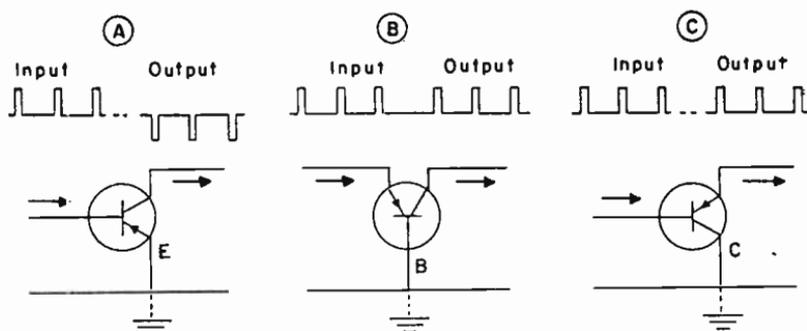


Fig. 15.—Signal polarity is inverted by a grounded emitter circuit, but not by a grounded base or by a grounded collector circuit.

amplifiers, in order that sync signals to sweep oscillators may be positive or negative in accordance with the type of oscillator.

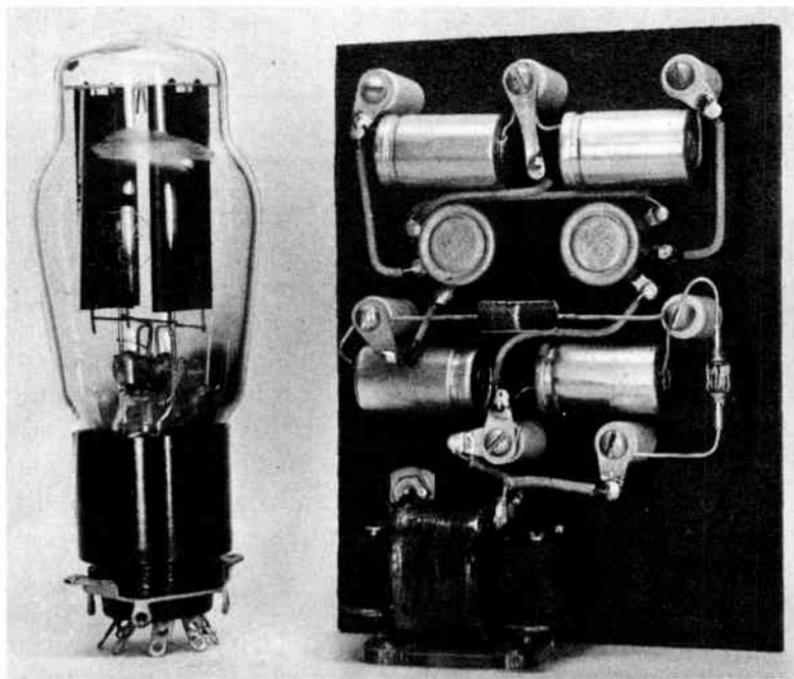
Were transistors to be used in video amplifier or sync section circuits, or anywhere else that polarity inversion affects performance, it would be found that there is inversion between input and output of a grounded emitter circuit, as illustrated at *A* in Fig. 15. There is no inversion of signal polarity between input and output of the grounded base circuit, at *B*, nor with the grounded collector circuit at *C*.

**Biasing for Transistors.** — With most applications and circuits the bias voltage is quite critical in value for the high-side element of the input circuit. In general, the bias voltage on the input side must be of value suited to voltage on the high-side element for the output side of the circuit, both of these voltages with reference to the common or grounded element. If voltage on the high-side output element, the collector for example, is increased, then it is necessary also to increase the bias on the high-side input element, the base or emitter in most circuits.

If bias is incorrect in relation to voltage on the output side of the transistor, the output waveform will be distorted on either the positive or negative alternations. This usually occurs with excessive bias voltage. Making the bias voltage less than the maximum value which prevents distortion may simply reduce the output power or voltage without distortion of waveform. Then it becomes possible to control the gain by varying the bias voltage up to the limit at which there is distortion.

It must be remembered, as mentioned earlier, that potential at the high-side input or output element is affected by voltage drop in any resistance in the input or output circuits as well as by voltage applied from external sources. This is true also of tube plate circuits, where voltage at the plate is less than voltage applied to the plate circuit from the B-supply. It is not true of grid circuits in which there is no current with the grid negatively biased.

**D-c Power Supplies.** — At the left in Fig. 16 is a 5U4-G power rectifier tube such as found in a great number of television receivers. At the right is a complete d-c power supply of an experimental type used with transistor amplifier circuits. At the top of the transistor power unit are two bypass capacitors for input and output circuits of a connected amplifier. Immediately below are two potentiometers for adjusting the bias voltage and the voltage to the high-side output elements. Near the center of the unit are the filter capacitors and filter resistor. At the lower left is the power transformer, which might be smaller to handle the small transistor currents. At the right, opposite one of the filter capacitors, is the "power rectifier," which is a high-back voltage type of germanium crystal diode.



*Fig. 16.—The power rectifier tube at the left is almost as large as the entire d-c power supply for transistors at the right.*

The circuit for this power supply unit is shown by Fig. 17. The rectifier is connected to provide, at the bias terminal, a potential positive with reference to ground, and at the output terminal a potential negative with reference to ground. Bypass and filter capacitors, of the electrolytic type, must be connected to suit the circuit polarities. Were the d-c output polarities to be reversed, to provide a negative bias and a positive potential

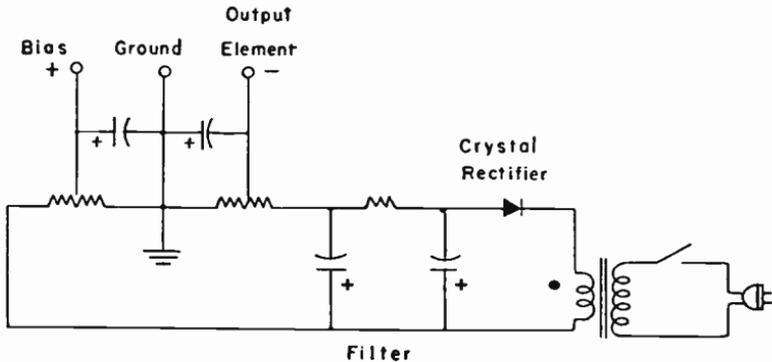


Fig. 17.—The circuit of the d-c power supply used with transistors.

for the amplifier output element, it would be necessary to reverse the rectifier and all the electrolytic capacitors. The experimental unit is made up with screw terminals to permit convenient reversal of the parts mentioned.

From power supplies designed for electronic tube circuits it is desirable to have as nearly as practicable a constant voltage. Then variations of signal currents in amplifiers and other connected circuits will not cause excessive changes of plate, screen, and grid bias voltages. A d-c supply unit designed for use with junction transistors in amplifier circuits may be of the constant voltage type.

With many circuit applications of point-contact transistors the applied potentials are of such polarity as might cause excessive current through the low impedance or low resistance side were these potentials to remain constant or nearly so regardless of current. That is, the transistor acts like a rectifier

to which voltage is applied in the forward or low-resistance direction, with possible destruction of the unit due to excessive current unless the voltage drops. Such conditions require a d-c power supply furnishing a nearly constant current, or, at least, a limited current. This is accomplished, in practice, by making power unit resistances and other resistances in series with the transistor element much greater than the element resistance. Current tending to become excessive then causes enough voltage drop to effectively increase the resistance, and limit the current.

Principal applications of transistors in the field of television servicing include oscillators and amplifiers for test instruments and accessory devices in which space and weight are to be made the least possible. Some of the more recently developed transistors are capable of handling enough power, at least one watt, to furnish adequate output without need for added amplification by means of tubes. There also are a number of applications in which transistors and miniature tubes are used together for reduction of total weight and of total power for cathode heating.

**TRANSMISSION, NEGATIVE.** — Negative transmission refers to the method of television carrier modulation with which a decrease of brightness in the televised image causes an increase of transmitted power. This method, which is standard in the United States and some other countries, is illustrated at the left in Fig. 126-1. The signal modulation displayed here is explained in the article, *Signal, Television*.

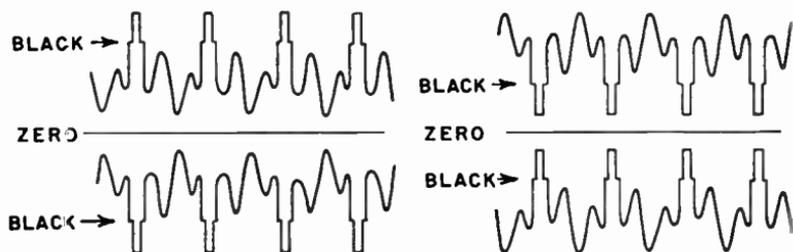


Fig. 126-1.—Negative transmission at the left and positive transmission at the right.

With negative transmission there is maximum or 100 per cent amplitude for the tips of sync pulses, approximately 75 per cent of maximum for the black level, and about 15 per cent of maximum for pure white in the pictures. With positive transmission, shown at the right, maximum amplitude corresponds to the lightest or brightest tones and minimum amplitude corresponds to the tips of sync pulses.

Negative transmission allows synchronization to be more strongly maintained, since the strongest portions of the signal are the sync pulses. Severe noise interference causes amplitude to increase beyond the black level, and thus puts dark spots rather than bright flashes in the reproduced picture. Either negative or positive transmission signals may be inverted as may be necessary for their application to picture tube grids or cathodes, and to synchronization of sweep oscillators.

**TRANSMISSION, VESTIGIAL SIDEBAND.** — Vestigial sideband transmission consists of transmission at full strength of all video modulation frequencies extending to a limit of 4 megacycles above the video carrier frequency, while transmitting at full strength the modulation frequencies below the video carrier to a limit of only  $\frac{3}{4}$  megacycle. That is, all modulation frequencies in the upper sideband are transmitted, but the lower sideband is partially cut off. The name vestigial sideband refers to the portion or vestige of the lower sideband that remains.

This method is intermediate between double sideband transmission with which all modulation frequencies above and below the carrier are radiated, and the single sideband method in which there is radiation of modulation frequencies only above or below the carrier frequency. Vestigial sideband transmission is standard for television.

Cutting off a large part of the lower sideband still allows transmission of video signals to a high limit of 4 megacycles, also a modulated sound carrier, within a frequency channel of 6 megacycles while providing enough spare band width above and below all modulation frequencies to avoid overlapping adjacent channels.

Fig. 127-1 shows distribution of transmitted frequencies within a 6-megacycle channel. The video carrier frequency is  $1\frac{1}{4}$  megacycles above the low limit for the channel. The sound carrier frequency is exactly  $4\frac{1}{2}$  megacycles above the video carrier, and  $\frac{1}{4}$  megacycle below the high limit for the channel. The height of the curve corresponds to the relative values of

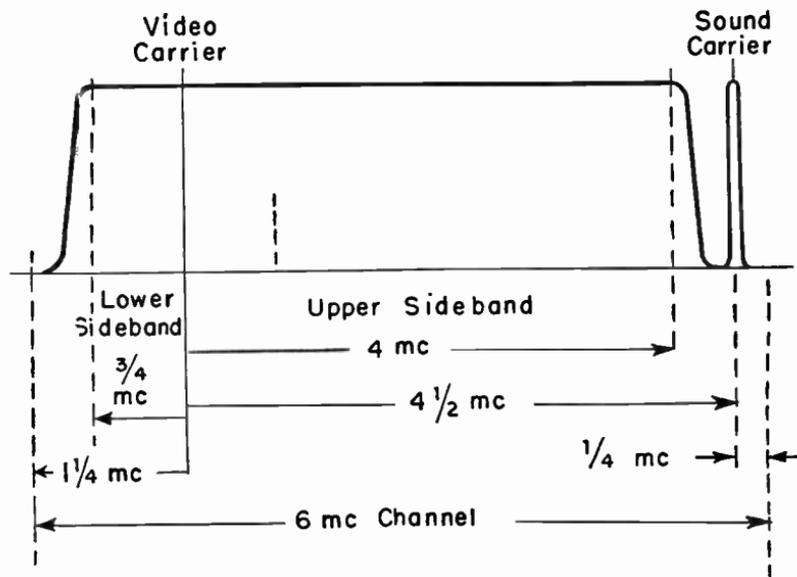


Fig. 127-1.—Distribution of transmitted frequencies within a six megacycle channel.

radiated field strength at the transmitter. Maximum field strength extends  $\frac{3}{4}$  megacycle below the video carrier frequency, and to 4 megacycles above this frequency. There must be practically complete attenuation between these limits and the low side of the channel and the sound carrier with its modulation.

When any one video signal frequency is applied as modulation to a carrier it will produce two sideband frequencies of transmission. The lower sideband frequency is equal to the difference between carrier and modulation frequencies, and

the upper sideband frequency is equal to the sum of carrier and modulation frequencies.

If frequency of video signal modulation is no more than  $\frac{3}{4}$  megacycle it will produce transmitted frequencies at full strength in both sidebands, as is evident from Fig. 127-1. This means that all video signal frequencies up to  $\frac{3}{4}$  megacycle will be transmitted at full strength in both sidebands, while all higher video signal frequencies will be transmitted at full strength in only the upper sideband. Consequently, all video signal frequencies up to  $\frac{3}{4}$  megacycle are effectively transmitted twice, and actually are transmitted in double strength. All higher video signal frequencies are transmitted in only single strength.

Were the television receiver to provide uniform amplification throughout the whole range of low and high transmitted sidebands, all video signals at frequencies up to  $\frac{3}{4}$  megacycle

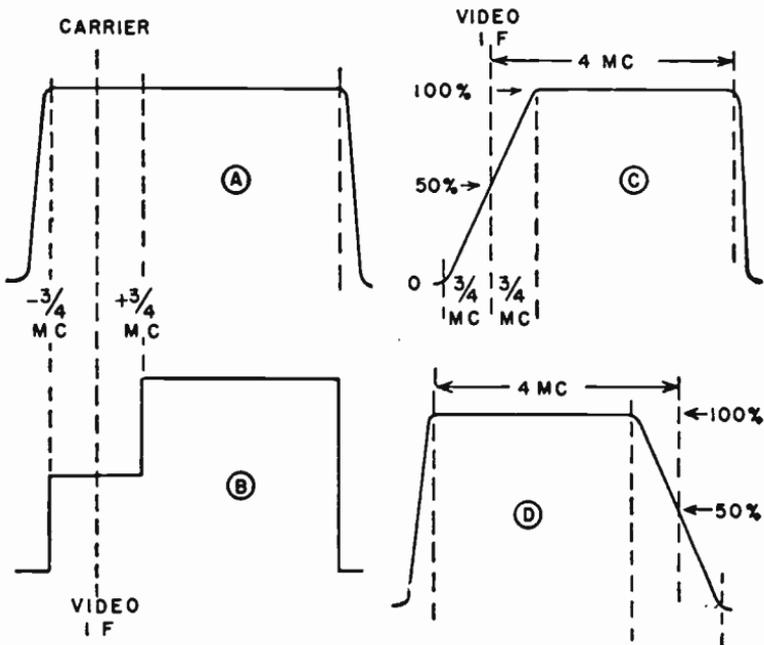


Fig. 127-2.—How receiver attenuation compensates for double transmission of some signal modulation frequencies.

which are received in double strength would be reproduced with twice the output of signals at higher video frequencies. This undesirable result is prevented by what is called receiver attenuation.

Receiver attenuation provides overall frequency response giving the effect shown by Fig. 127-2. At *A* is the curve for radiated field strength. At *B* is shown a receiver response providing 50 per cent of maximum gain for all the doubly transmitted frequencies, and 100 per cent gain for all singly transmitted frequencies. The result would be uniform output for the whole range of video signal frequencies after detection. The video intermediate frequency in circuits following the mixer corresponds to the carrier frequency in circuits preceding the mixer.

It would be difficult to construct receiver video i-f amplifiers having such sharp changes of gain at certain frequencies as shown at *B*. The same effect may be attained much more easily with a video i-f amplifier frequency response such as at *C*. Here the gain is zero at  $\frac{3}{4}$  megacycle below the video intermediate frequency, increases uniformly to 50 per cent of maximum at

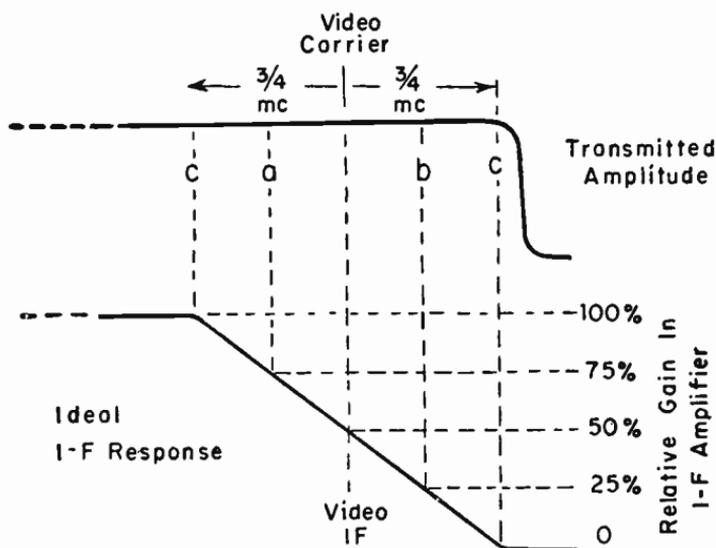


Fig. 127-3.—Relative gain percentages with an ideal form of receiver attenuation.

the video intermediate frequency, and continues uniformly to 100 per cent at a frequency  $\frac{3}{4}$  megacycle higher than the video intermediate.

On this sloping portion of the receiver response any two frequencies which are equally below and above the video intermediate frequency are modulated with the same video signal. For instance, there is the same signal modulation on a frequency  $\frac{1}{4}$  megacycle below the video intermediate as on another frequency  $\frac{1}{4}$  megacycle above the video intermediate.

If we add together the percentage gains for any two frequencies equally below and above the video intermediate, the sum of the two gains always will be 100 per cent. This is the same as the 100 per cent gain for higher frequencies along the flat top of the curve. Therefore, with this ideal form of receiver response curve, all signal modulation frequencies will receive the same total amplification, and amplifier output will be uniform for all frequencies extending from the video intermediate frequency to 4 megacycles above this frequency.

Video carrier frequencies are lower than sound carrier frequencies as transmitted. When the video carriers and their modulation sidebands are subtracted from r-f oscillator frequencies in the mixer this relation is reversed, and the video intermediate frequency becomes lower than the sound intermediate frequency. For this reason the video i-f response curve more often is shown as at *D* in Fig. 127-2 than as at *C*.

Fig. 127-3 shows some relative gain percentages illustrating the effect of receiver attenuation. It may be seen that gains at *a* and *b* are respectively 75 and 25 per cent, making a total of 100 per cent. Also, gains at *c* and *c* are 100 per cent and zero, again making a total of 100 per cent. Curves which have been shown are ideal frequency responses for the video i-f amplifier of a receiver. Actual responses as seen on an oscilloscope approximate this ideal to a greater or less extent when alignment adjustments are correctly made.

**TRAPS, INTERFERENCE.** — An interference trap or wave trap is a device for preventing the appearance in an amplifier output of voltages at undesired frequencies, or at frequencies

which might cause interference with the desired signals. Interference traps described in this article are those associated with the video i-f amplifier, video detector, and video amplifier of the television receiver. Traps such as used in antenna and tuner circuits are described in the article on *Interference*.

Almost any type of circuit capable of being sharply tuned to resonance may be utilized for a wave trap. Parallel resonant trap circuits are connected in series with the signal path and tuned to offer high impedance at the unwanted frequency, and may be used to dissipate energy at unwanted frequencies by means of the large circulating currents in the resistance of the parallel resonant trap circuit. Series resonant traps may be connected from the signal circuit to ground and tuned to allow escape of unwanted frequencies while having high impedance at frequencies of the desired signals.

Fig. 128-1 shows several traps of types which absorb and dissipate the unwanted signal energy, or bypass this energy to ground. At *A* a parallel resonant absorption trap is inductively coupled to the tuning coil in the plate circuit of the left-hand tube. This trap is shown as being tuned or aligned by means of adjustable inductance, but an adjustable capacitor may be used instead. Most trap circuits are adaptable for either method of tuning. A generally similar trap is shown by broken lines.

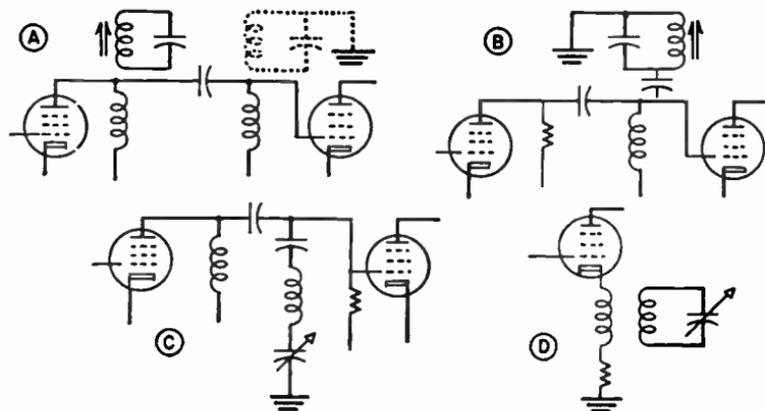


Fig. 128-1.—Interference traps of absorption types.

inductively coupled to the grid circuit of the second tube. One side of this trap circuit is grounded. Other parallel resonant traps may be similarly grounded to increase their energy dissipation.

At *B* in Fig. 128-1 a parallel resonant trap circuit is coupled through a small capacitance to the grid circuit of an amplifier tube. Diagram *C* shows a series resonant trap circuit connected through a small capacitance to an amplifier grid circuit. This trap is tuned by an adjustable capacitor. Adjustable inductance might be used instead. At *D* a parallel resonant trap circuit is inductively coupled to a coil which is in series with the cathode of an amplifier tube. Any undesired frequencies may be removed at cathode circuits, grid circuits, or plate circuits, or may be removed at two or more of these points in the same amplifying stage.

Traps offering high impedance to undesired frequencies are illustrated by Fig. 128-2. At *A* there is a parallel resonant trap circuit in series between the plate of the first tube and the grid of the second tube. A similar trap might be used on the plate side of the circuit, or there might be one of these traps on the plate side and another on the grid side. Two traps ordinarily

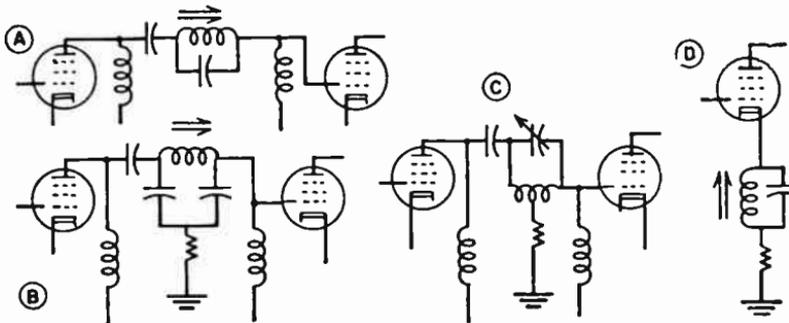


Fig. 128-2.—Interference traps of impedance types.

would be tuned to two different interference frequencies. Diagram *B* shows another series connected parallel resonant trap, this one having two capacitors in the trap circuit with a ground connection from a point between them.

Diagram *C* of Fig. 128-2 shows a series connected parallel resonant trap with a center tapped coil. From the center tap there is a connection through a resistor to ground. Any of the traps which are shown on the grid side of the interstage couplings might be on the plate side provided blocking capacitors are used to isolate the B+ lines from conductive connections with ground or B-minus. At *D* is shown a parallel resonant trap connected in series with the cathode of a tube. A trap of this general type may be in the r-f return connection of a video detector to prevent demodulation of signals for which the trap is tuned.

**Trap Frequencies.** — Fig. 128-3 illustrates the manner of determining the frequencies to which traps should be tuned in a receiver operating with certain video and sound intermediate frequencies. It has been assumed, for purposes of illustration, that the particular receiver here considered is to operate with a video intermediate frequency of 25.75 megacycles and with a sound intermediate frequency of 21.25 megacycles. It is assumed also that the receiver is tuned for channel 8, although trap frequencies would be the same for tuning to any channel having other channels immediately adjoining its upper and lower limits.

For reception in channel 8 the r-f oscillator of the receiver would be tuned to a frequency equal to the sum of the video intermediate frequency and the video carrier frequency for that channel, or to the sum of the sound intermediate and sound carrier, which would be the same thing. With the assumed intermediate frequencies this oscillator frequency would be 207.00 megacycles. If video or sound carriers from the adjacent channels reach the mixer these other carriers will beat with the frequency of the receiver r-f oscillator to produce beat frequencies. Along the top of Fig. 128-3 are shown these other carriers and also the carriers for the tuned channel, number 8. Each carrier frequency is subtracted from the r-f oscillator frequency to show the resulting beat frequency. All these beat frequencies would be applied to the video i-f amplifier of the receiver.

The frequency response of the video i-f amplifier might be as shown by the curve. There is the required 50 per cent of maximum gain at the video intermediate frequency of 25.75 megacycles. There is almost full gain through all frequencies extending 4 megacycles from the video intermediate, which is

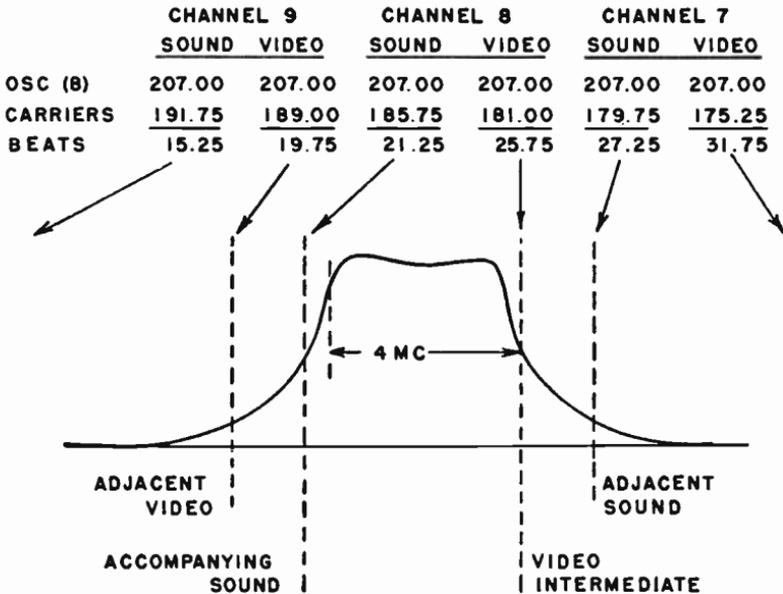


Fig. 128-3.—How the required trap frequencies may be determined in accordance with video or sound intermediate frequencies used in a receiver.

a full range of high quality picture reproduction. There is very nearly 50 per cent gain at the sound intermediate frequency which, when considering traps, is called the accompanying sound frequency or the associated sound frequency. Were such highly amplified sound signals to go through the i-f amplifier, the video detector, the video amplifier, and to the picture tube there would be severe sound interference with pictures. This would become apparent as sound bars on the screen. Consequently, the accompanying sound frequency must be trapped out before it reaches the picture tube.

With the frequency response as illustrated there is considerable gain at the adjacent channel video frequency of 19.75 megacycles. Unless this adjacent video frequency is trapped

out before reaching the picture tube it will cause beat interference patterns consisting of narrow vertical and diagonal lines which weave and sway one way and another.

There is considerable gain also at the adjacent channel sound frequency of 27.25 megacycles. This adjacent sound frequency must be trapped out in order to prevent interference patterns and possibly sound bars on the screen of the picture tube.

If the video i-f response or band pass is not so wide as shown by Fig. 128-3 some or all the traps may not be required. A relatively narrow band pass, usually not much more than 3 megacycles from the video intermediate, is found in receivers having only limited amplification in intermediate-frequency stages. In receivers of this type there may be none of the traps mentioned; neither accompanying sound, adjacent sound, nor adjacent video traps.

Traps for accompanying sound are not used in the i-f amplifiers of receivers employing the intercarrier sound system, because both the sound and the video intermediate frequencies must reach the video detector. The band pass in many of these receivers is narrow enough to exclude adjacent channel interference, so there may be omission also of traps for adjacent video and adjacent sound signals.

Trap frequencies determined as in Fig. 128-3 are correct for only the intermediate frequencies mentioned. Similar computations may be made for any other intermediate frequencies as used in particular receivers. It is not necessary to consider adjacent channel interference unless the frequency limit of the other channel is the same as that of the received channel. There is separation of 4 megacycles between channels 4 and 5. This makes it practically impossible for sound in channel 4 to affect reception in channel 5, or for video in channel 5 to affect reception in channel 4. Because there is no channel immediately below number 7 this channel would suffer no adjacent sound interference, and because there is no channel immediately above number 13 there would here be no adjacent video interference.

Were sharply tuned traps applied to the video i-f amplifier whose original response is shown by Fig. 128-3 the response as modified by the traps might appear about as in Fig. 128-4. There should be attenuation nearly to zero at each of the trap frequencies.

A trap for completely removing accompanying sound must never precede the point at which the sound i-f signal is taken from the mixer or i-f amplifier and carried to the sound section

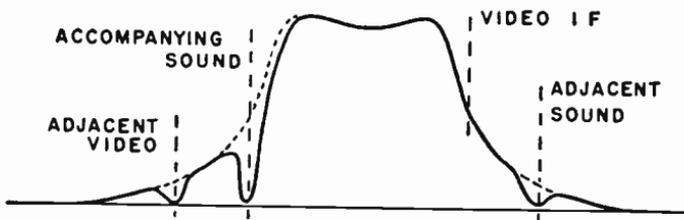


Fig. 128-4.—Video i-f frequency response as modified by three types of traps.

of the receiver. Otherwise the sound signal would not reach the point of takeoff. Other traps, when used, may be associated with any of the interstage couplers or may be on the cathodes of i-f amplifiers or video detector.

**Alignment of Video i-f Traps.** — Traps which are connected to or coupled to the video i-f amplifier are aligned at the same time as this amplifier, because adjustment of the traps will affect alignment of the transformers or couplers, and adjustment of the transformers or couplers will affect the trap frequencies. When working through the i-f amplifier it usually is most satisfactory to adjust the trap coupled to each transformer just before that transformer is aligned. A sound takeoff transformer should be treated like a trap, and aligned in its regular order.

Methods of alignment are the same regardless of the types of traps, the object always being to have minimum response at the trapped frequencies. Traps should be adjusted only for this minimum response at their respective frequencies, and no attempt should be made to use the traps for shaping the response curve around the video intermediate frequency. Traps

designed for certain interference frequencies usually have a total adjustment range of no more than 2 or 3 megacycles. Consequently, a trap designed for one kind of interference seldom can be used satisfactorily for any other kind.

Traps usually are aligned by means of a signal from a constant frequency generator, such as a marker type or any accurately calibrated signal generator, while attenuation is measured with an electronic voltmeter. The first steps include disconnecting the transmission line or antenna from the receiver, setting the channel selector on a channel where there is no transmission, and placing the contrast control at the maximum setting which does not cause regeneration or oscillation in the video i-f amplifier.

The signal generator is coupled to the mixer tube as for video i-f alignment, or it may be coupled to a tube preceding each trap aligned. The generator is used without modulation. The electronic voltmeter is connected to the video detector load resistor or, for greater sensitivity in adjustments, to the load resistor of a video amplifier. This latter connection will require using a detector probe on the meter.

The generator should be tuned accurately to the frequency at which each trap is to be aligned while adjusting that trap for minimum reading on the meter. If there is more than one trap for the same frequency, first adjust the one farthest from the generator. A trap closer to the generator connection may have to be temporarily detuned to pass enough signal for correct adjustment of a trap farther away in the signal path. Generator output may be increased as alignment progresses. With all traps aligned the generator should be tuned slowly through the whole range of trap frequencies while watching the meter. At each minimum voltage reading the generator frequency should be noted, and if it is not at one of the trap frequencies it will be necessary to readjust the corresponding trap.

Traps may be aligned and their effects on frequency response observed by means of a sweep generator for signal input and an oscilloscope as the indicator. Fig. 128-5 shows interstage couplings and traps of an i-f amplifier from which were ob-

served the oscilloscope responses of Fig. 128-6. Between the mixer and first i-f amplifier tube is an overcoupled transformer to which is inductively coupled a trap for accompanying sound. Between first and second i-f amplifier tubes is a single tuned coupling coil with an inductively coupled trap for adjacent sound. The next coupler is a single tuned coil and the last one is an overcoupled transformer. There are no traps on the latter two couplers.

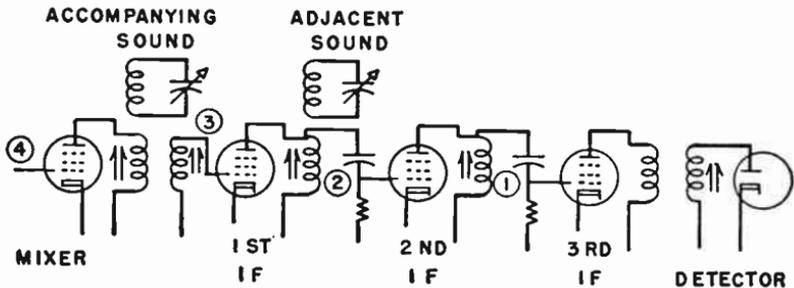


Fig. 128-5.—Traps for accompanying sound and adjacent sound coupled to a video i-f amplifier section.

The top curve of Fig. 128-6 is taken with the sweep generator coupled to the grid of the third i-f amplifier and the oscilloscope connected across the detector load resistor. The oscilloscope remains connected here for the other curves. This top curve shows a typical frequency response for an overcoupled transformer. It covers a wide range of frequencies and shows no dips due to traps.

The second curve from the top is taken with the sweep generator moved back to the grid of the second i-f amplifier. This is the frequency response of the single-coil coupler and the overcoupled transformer working together. The high peak on this response is at the frequency to which the single-coil coupler is tuned.

The third curve from the top is taken with the sweep generator at the grid of the first i-f amplifier tube. The response now shows one peak due to the single-coil coupler between first and second amplifiers, and the other peak due to the similar coupler between second and third amplifiers. There now

appears a dip in the response caused by action of the adjacent sound trap coupled to the single-coil coupler last brought into the amplification system between generator and oscilloscope.

The fourth or bottom curve is taken with the sweep genera-

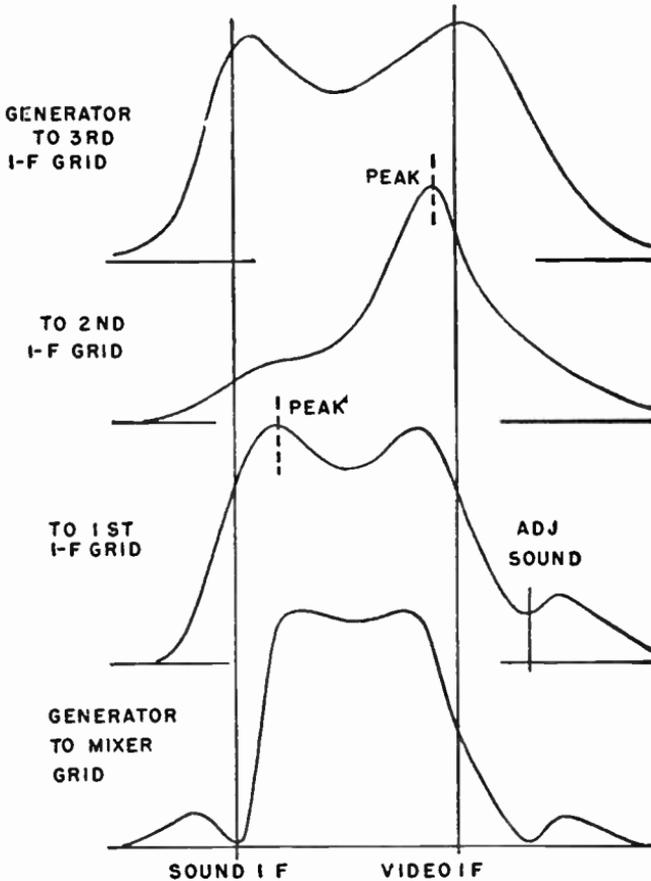


Fig. 128-6.—Frequency responses at various points in the i-f amplifier having traps for accompanying and adjacent sound.

tor coupled to the mixer grid. The effect of the first over-coupled transformer is chiefly a narrowing of the frequency response. At the sound intermediate frequency on this final curve appears a dip due to action of the accompanying sound trap which is coupled to the first transformer.

For trap alignment with the sweep generator and oscilloscope the generator should sweep the entire video i-f range and through the adjacent sound and video frequencies. A sweep of about 10 megacycles should be sufficient. The oscilloscope may be connected across the video detector load resistor for all observations, or for greater sensitivity may be connected to the load resistor of a video amplifier tube. A marker must be accurately adjusted for the frequency of each trap to be aligned. The trap is adjusted to eliminate the center of the marker pip so far as is possible while gradually increasing the output level from the marker generator.

**Intercarrier Sound Traps.** — The intermediate frequencies for video and for sound beat together in the video detector to produce their difference frequency, which always is 4.5 megacycles. This is called the intercarrier beat frequency. If this frequency reaches the picture tube it causes a pattern similar to that from any other high-frequency interference, closely spaced thin vertical or sloping lines which shift one way and the other.

If the sound intermediate frequency is thoroughly trapped out of the video i-f amplifier it will not reach the video detector and there will be no intercarrier beat frequency in the detector output nor in following circuits leading to the picture tube. If some of the sound intermediate frequency does reach the video detector the resulting intercarrier beat must be trapped before reaching the picture tube.

In receivers employing the intercarrier sound system the intercarrier beat frequency with its sound modulation is taken through the sound takeoff transformer or other form of coupling to the sound section of the receiver. The sound takeoff is sharply tuned, and removes practically all the 4.5-megacycle energy from circuits extending from the takeoff to the picture tube. Thus the sound takeoff acts, or should act, as an effective trap for the intercarrier beat so far as picture tube input is concerned.

In receivers having dual channel sound systems it often is necessary to place an extra trap for the intercarrier beat fre-

quency in the output circuit of the video detector or in any of the grid or plate circuits of the video amplifier. This trap nearly always is a parallel resonant type in series with a signal circuit, as shown at *A* in Fig. 128-2. High quality picture reproduction is not impaired by the trap cutoff at 4.5 megacycles, since this is well above the maximum video frequency needed for good definition.

A trap for intercarrier beat may be aligned with a 4.5-megacycle signal from an adjustable or crystal controlled constant-frequency generator and either an electronic voltmeter or an oscilloscope as response indicator. The generator is coupled to the grid of the tube preceding the trap, and is tuned accurately to 4.5 megacycles.

An electronic voltmeter with a detector probe on the input cable is applied to the plate of the tube which follows the trap or to the input of the picture tube. The generator is used without modulation. The trap should be aligned for minimum reading on the meter. This reading will not drop to zero because of other frequencies which are present in the receiver circuits, but will become minimum as the trap is adjusted.

The vertical input of an oscilloscope, fitted with a filter probe, is applied to the plate of the tube following the trap or to the picture tube input. The generator is used with tone modulation, usually at 400 cycles. The internal sweep of the oscilloscope is adjusted for the tone modulation frequency of the generator and the trap is adjusted for minimum height of the trace on the oscilloscope screen.

When using either the electronic voltmeter or the oscilloscope the output of the generator should be increased as the trap adjustment proceeds. The contrast control of the receiver may be at its maximum setting for final adjustment of the trap.

**TRAPS, ION.** — An ion trap consists of a specially designed electron gun acting in conjunction with the fields of one or two externally mounted magnets to separate ions from the electron beam in certain types of picture tubes which are operated with magnetic deflection.

The ions are atoms of gases which have acquired a negative charge by temporary attachment to them of one or more negative electrons. An ion has the weight or mass of an atom, nearly two thousand times that of an electron. The negative ions are accelerated toward the screen of the picture tube by the same positive potentials which accelerate the negative electrons. Because of their great mass, the ions are deflected only slightly by the magnetic fields of the deflection coils which deflect the electrons of the beam horizontally and vertically during scanning. As a result, all the ions tend to strike near the center of the picture tube screen, within an area about the size of a half dollar. Such concentrated bombardment by the heavy ions would produce on the screen a brown spot called an ion burn or ion spot.

Tubes operating with electrostatic deflection do not permit concentrated ion bombardment because the electrostatic fields which deflect the electron beam also deflect the ions to distribute them quite uniformly over the screen surface. In some magnetic deflection tubes the inside surface of the screen is protected by a metallic coating of atomic thickness which prevents ill effects from ion bombardment. Most other tubes operating with magnetic deflection require ion traps.

Separation of ions from the electron stream is made possible by two facts. First, both ions and electrons are strongly deflected by electrostatic fields or electric fields. Second, only electrons are strongly deflected by magnetic fields. Although a magnetic field acts on the negative charge of an ion just as it acts on the negative charge of an electron, the relatively great mass or weight of the ion keeps it traveling along its original course in spite of the deflection force, whereas the negligible mass of an electron does not prevent its deflection.

**Double Magnet Traps.** — Fig. 129-1 illustrates the construction of one style of ion trap. Looking at the side of the picture tube neck which is on the right when facing forward from the base end, there may be seen a diagonal gap between the first and second anodes of the electron gun or between the second grid and the first anode, according to which names are applied to

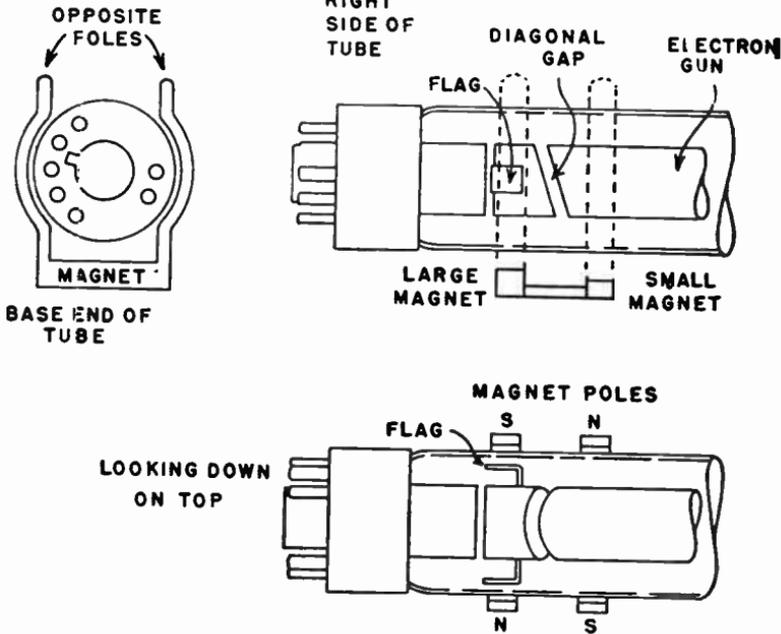


Fig. 129-1.—Ion trap of the type requiring a double magnet.

these elements. Extending through the gap is a strong electric field, due to the second element being much more highly positive than the first one. The lines in this electric field are almost at right angles to the direction of the gap, or the lines slope upward toward the front of the tube. The result is that both electrons and ions are deflected at this upward angle by the electric field.

On the outside of the tube neck are the poles of two permanent magnets. One magnet, the stronger one, is mounted a short ways back of the diagonal gap. The other magnet is ahead of the gap. The four poles of these two magnets are arranged as shown by the top view of the tube in Fig. 129-1. The magnetic field between poles of the rear magnet is in such direction as to deflect electrons downward, while the field of the front magnet is in the opposite direction, and deflects electrons upward. These two magnetic fields have hardly any effect on the ions.

The action is illustrated in a general way by Fig. 129-2 where are shown the paths followed by one ion and one electron. Direction of ion travel is practically unaffected by the field of the rear magnet, but ions are deflected upward almost into line with the electric field in the diagonal gap. The nega-

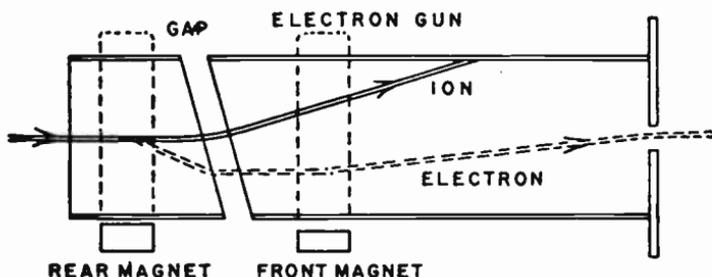


Fig. 129-2.—Paths of ions and electrons in the double magnet ion trap.

tive ions continue on their new course until striking the inside of the positively charged forward anode, where they are collected. Electrons are deflected downward as they pass through the field of the rear magnet, are straightened out by the electric field in the gap, then are deflected upward by the field of the front magnet just enough so that the electrons pass through the opening at the front end of the anode to form the electron beam.

The double magnet structure whose action has been described is attached to the neck of the picture tube as shown at the upper left in Fig. 129-3. This is a view looking down on top of the tube neck. Passing through the pole piece extensions of the two magnets are long screws fitted with knurled nuts. With these screws loosened the magnet frame will slide over the tube base and forward to its working position. Then the screws are tightened just enough to prevent slipping.

At the upper right in Fig. 129-3 is shown an arrangement consisting of two ring-shaped permanent magnets having open air gaps forming their poles. The magnets are carried by a supporting sleeve of fibre or rubber. Tension springs between the sleeve and tube neck hold the magnets wherever they are

placed, but allow sliding and rotating for adjustment. The larger and stronger ring magnet is placed toward the base of the picture tube. The smaller front ring may be rotated independently of the large one, which sometimes is desirable when making an adjustment. Normally the gaps in the two rings are in line lengthwise of the tube axis.

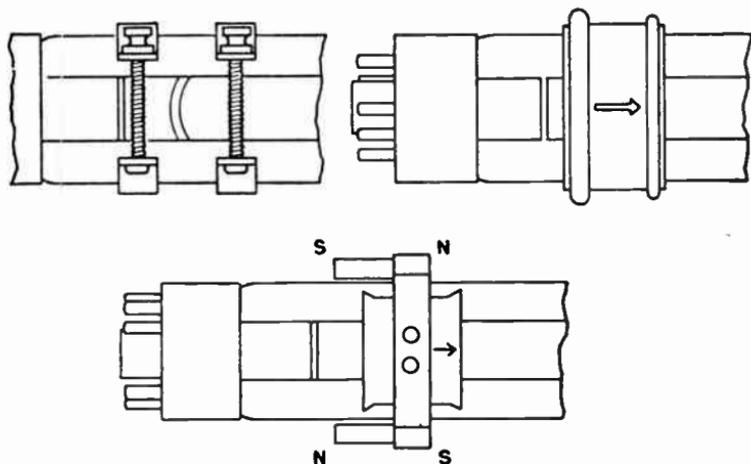


Fig 129-3.—Methods of mounting double magnet ion traps on the necks of picture tubes.

The larger or stronger magnet of a double unit goes toward the tube base. Poles of this larger magnet should be initially in line with two small metal “flags” which extend outward from the electron gun near the gap between first and second grids, as in Fig. 129-1. Some magnet structures are marked with arrows which should point toward the front end or viewing screen of the picture tube.

**Single Magnet Traps.** — Fig. 129-4 illustrates the principle of the bent-gun ion trap which requires only a single external magnet for deflection of electrons. The cathode, the pre-accelerator anode, and the rear end of the second anode are at an angle with the axis of the electron tube neck. The remainder of the second anode is in line with the neck axis. Ions and electrons enter the rear end of the second anode while traveling

toward a point on the inside of this anode. The ions continue along this original direction until striking the anode surface, where they are collected. The field of the single external magnet deflects the electrons so that they are directed toward the opening at the front end of the second anode, where the electron beam is formed and continues through the focus coil and deflection coils to the tube screen.

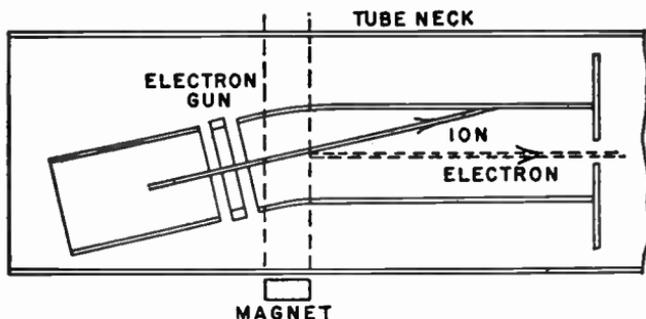


Fig. 129-4.—Operating principle of the bent-gun ion trap employing a single magnet.

Another electron gun requiring only a single trap magnet is shown by Fig. 129-5. The gun itself is straight, but is inclined with respect to the neck axis. The opening through

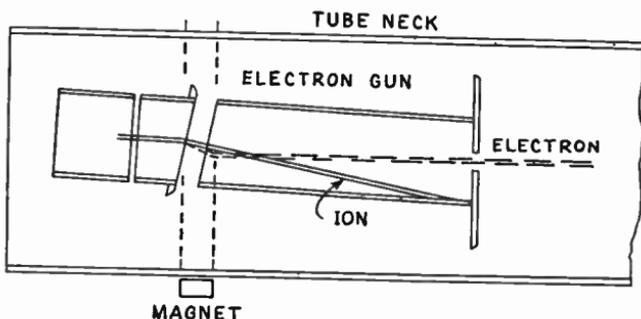


Fig. 129-5.—A tilted electron gun for removal of ions from the electron beam.

the front of the gun is centered in the neck, while the rear end is offset from the neck axis. Tilt of the gun directs both electrons and ions slightly downward. In the large gap is a

strong electric field which turns electrons and ions still farther downward. Outside this gap is the single magnet, whose strength and field polarity are such as to turn electrons upward in spite of the electric field. Electrons pass through the opening at the front of the gun. Ions continue their downward path and strike inside the positively charged second anode.

A single trap magnet, unless plainly marked for correct position, may go on the tube neck with either pole on either side. Should the first position prove wrong, rotating the magnet a half-turn around the neck will correct it.

**Trap Magnet Adjustment.** — The contrast control should be set as for normal reception during trap magnet adjustment. The brightness control must not be advanced more than allows barely visible illumination of the picture tube screen when magnet adjustment begins; otherwise the electron gun may be permanently damaged. The first step in adjustment is slow rotation of the magnet, with simultaneous forward and back movement along the tube neck to locate the position for maximum screen brightness or brightest raster.

The next step is to slowly advance the brightness control while rotating and sliding the magnet structure on the tube neck to obtain the brightest possible raster with minimum setting of the brightness control. Should the tube screen remain nearly dark with brightness advanced, the position of the magnet should be checked; it may be necessary to rotate the structure a half-turn.

Should a bright raster be obtained with the magnet in either of two positions, one farther back than the other, final adjustment should be with the magnet closer to the tube base. If shadows appear around the edges of the picture tube screen they are not to be eliminated by magnet adjustment, but by correct positioning of the deflecting yoke and of focusing and centering devices. The ion trap magnet is to be adjusted only for the brightest raster while the brightness control is adjusted as for normal viewing.

**TROUBLE, LOCATION OF.** — For rapid location of troubles it is necessary first to consider which sections of the receiver are most likely at fault and in which it is improbable that the trouble exists. Fig. 130-1 illustrates typical "sectionalizing" for trouble shooting.

1. *Power supply.* This section extends from the plug inserted in the building power line receptacle through the voltage dividers of both low-voltage and high-voltage power supply systems.

2. *Sound section.* This section extends from the sound takeoff through to the speaker.

3. *Sweep section.* This section extends from the vertical and horizontal sweep oscillators through the deflection coils or plates and includes the beam deflection functions in the picture tube.

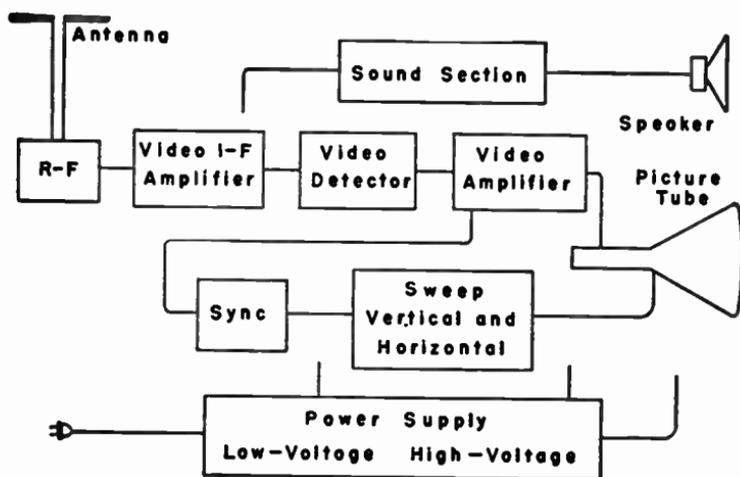


Fig. 130-1.—Sections of receiver to be considered during location of troubles.

4. *Sync section.* This section begins at the sync takeoff from the video amplifier or detector and extends to the inputs of the sweep oscillators.

5. *Video amplifier.* This portion of the receiver extends from the video detector through the grid-cathode signal input of the picture tube.

6. *Video detector.* Located between the video i-f amplifier and video amplifier.

7. *Video i-f amplifier.* This amplifier system extends from the output of the mixer or converter tube to the signal input for the video detector.

8. *R-f section.* This section includes the channel selector or tuner, the r-f amplifier, r-f oscillator, and mixer.

9. *Antenna system.* Including the outdoor, indoor, or built-in antenna, and also the transmission line.

Preliminary examination for determining in which section the trouble probably exists may be carried out by observing the screen of the picture tube, listening to the speaker, using a signal generator, and sometimes an oscilloscope or other testing instruments.

The exact procedure followed during trouble location depends largely on the kind of equipment available and on the experience and preferences of the technician. If experience has shown that the receiver being worked upon is likely to develop certain kinds of trouble, and if evidences of such trouble are present, it is that trouble that should be looked for without delay. Methods described in this article assume that no such information on probable troubles exists.

When preparing to locate trouble the receiver switch should be turned on and the channel selector set for a channel in which a station should be transmitting at the time of testing. Analysis then may be carried out in the following steps, which list parts that should be examined for each class of symptoms.

- A. No picture or pattern, no raster, no sound. Glass tubes do not light nor metal tubes become warm. Examine the following parts.
  1. Line plug and its contacts in the outlet receptacle.
  2. Line cord or extension cord from plug to receiver.
  3. On-off switch in the receiver.
  4. Any interlock switch that may have been opened by removal of a cover or panel on chassis or cabinet.
- B. No picture or pattern, and no sound. There is a raster, either uniform or mottled.

1. Change the selector to another channel or to several other channels where stations are known to operate. The first station may be off the air. If the trouble exists only on some channels, but not on all, the fault must be in the tuner or channel selector, in the tuned couplings of the r-f section, or in the r-f oscillator. If the same symptoms exist on all channels the fault may be in any of the following.
    2. Antenna or transmission line.
    3. R-f section, including its tubes.
    4. Video i-f amplifier stages which carry signals for both picture and sound.
    5. Video detector, if the receiver has intercarrier sound.
    6. Sync section. Since there is a raster, the vertical and horizontal sweep sections must be operating.
- C.** No sound. There is a picture or pattern.
1. Sound section.
  2. R-f oscillator out of alignment or fine tuning incorrectly adjusted in receivers using dual or split sound system.
- D.** No distinct picture or pattern. There is a raster, either uniform or mottled. There is sound.
1. Sync section. If neither of the hold controls have material effect on the appearance of picture, pattern, or raster it is probable that trouble exists in the sync section.
  2. Video amplifier.
  3. Video detector.
  4. Video i-f amplifier, beyond the point of sound take-off.
- E.** No picture, pattern, or raster, but only a horizontal or vertical line on the picture tube screen. There is sound.
1. Sweep section. If there is only a vertical line, trouble is in the horizontal sweep system, if only a horizontal line the trouble is in the vertical sweep.

One of the quickest ways of locating a section or part in which trouble probably exists is by observation and analysis of test patterns appearing in the picture tube screen. This method is explained in the article *Trouble Shooting, Pattern Analysis for*.

An oscilloscope instead of picture tube indications may be used for locating the section or circuit in which trouble exists. With an oscilloscope it is possible to observe the frequency response of all stages or combinations of stages between the antenna terminals and the signal input to the picture tube. This is done with the help of sweep and marker generators or a television signal generator.

With an oscilloscope it is possible to observe also the signal waveforms as well as other waveforms existing at points between the video detector and the beam deflection circuits and elements at the picture tube. Use of the oscilloscope for trouble location and signal tracing is explained in articles on *Oscilloscopes*, on *Trouble Shooting, Waveforms and Frequency Responses*, and on *Waveforms*. Under *Tracing, TV Signals* are described methods of following from antenna terminals to the video input of the picture tube a tone modulated signal furnished by an r-f generator.

After some one section has been identified as the probable location of a fault there are three commonly employed methods for tracing the trouble to a single part or circuit in that section. These methods include tube substitutions, voltage measurements, and resistance measurements.

New tubes or tubes known to be in good condition for television reception should be substituted, one at a time, for the original tubes in the section being checked. Checking by means of a tube tester may or may not indicate a defective tube. A tube which tests bad doubtless will fail to operate, but a tube may test good and yet fail to operate satisfactorily in the television circuits. Since it would be possible for more than one tube to burn out in a series heater system, an ohmmeter may be used to check heater continuity in single tubes or in groups of tubes.

Further tests require that the chassis be removed from its cabinet. If the picture tube in a series heater system is disconnected, the heater leads for this tube must be completed with a resistor of the same resistance and wattage dissipation as the heater itself.

Examine wiring, resistors, capacitors, and small inductors for evidences of overheating. Look for charring, blistering, melting of wax and other insulation, and note any burnt odor. These things indicate severe overloading, or possible short circuits or accidental grounds. When a part has failed because of overloading there is little use of replacing the part until the cause of overloading has been located and corrected.

Look for loose or dirty circuit and terminal connections while the set is turned on. At all suspected connections make checks by using light pressure, tapping, gentle pulling, or careful bending while watching for any changes in operation.

Voltage at tube socket terminals should be measured and compared with correct voltages listed in service instructions for the receiver if such instructions are available. Otherwise comparison should be made with voltages which would be suitable for each type of tube as listed in typical ratings for the tube. Voltages at plates and screens of tubes in the sync section may be much lower than any typical ratings, but there should be some voltage at these elements.

When working from service instructions it is necessary to follow testing conditions as there specified. Such conditions may include the type of voltmeter to be used, whether measurements are to be from tube sockets to ground or to B-minus, and the positions of contrast controls and other controls. Measurements usually are to be made with 117 line volts. Plus and minus voltage limits for satisfactory performance usually are given.

Resistance measurements from tube sockets to chassis ground or B-minus should be made if there is available a list or diagram showing correct resistances and resistance limits. Most faults which cause incorrect resistances will also cause incorrect voltages, and vice versa. The chief advantage of

resistance measurements is that they may be made on a completely dead receiver, where there are no voltages to measure.

It may be difficult to make measurements on individual resistors or capacitors thought to be open circuited, since, with these units connected in place, there are many paralleled paths whose effects can cause misleading readings. The suspected unit may be bridged or jumped momentarily with a good resistor or capacitor while observing results. It is important to make jumper connections with shortest possible leads to avoid introducing so much extra capacitance or inductance as will prevent normal action of the tested circuit.

**Sound Section Troubles.** — Causes for faulty reproduction of sound are, in general, the same as found in parts of f-m sound receivers which follow the mixer or converter tube in such receivers. In a television set these faults will be in parts which follow the sound takeoff. Articles which treat of these parts are under the following headings.

*Detector, ratio.*

*Discriminators.*

*Intercarrier sound.*

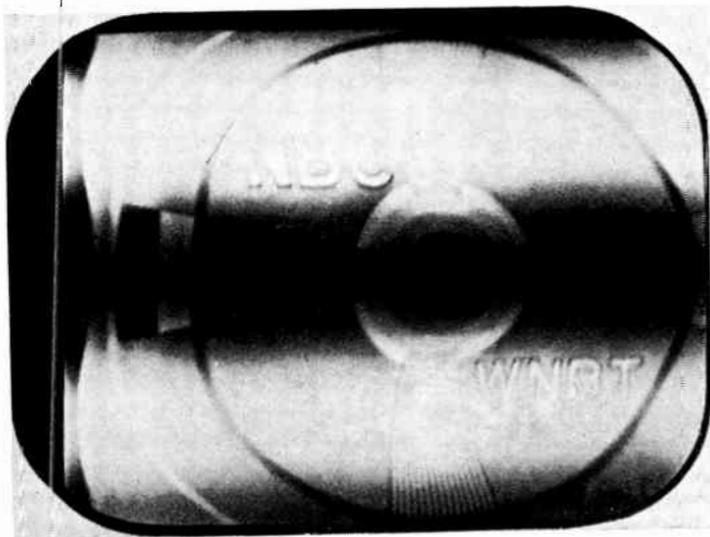
*Limiters, f-m sound.*

*Modulation, frequency.*

*Sound, television.*

**TROUBLE SHOOTING, PATTERN, ANALYSIS FOR.** — Differences between abnormal and normal patterns or pictures often allow determining what troubles cause faulty reproduction. Features of test patterns are described under *Patterns, Test*. Succeeding pages in the present section show examples of abnormal patterns, and include notes on causes and corrections for defects listed below. Many of the photographs are reproduced through courtesy of Allen B. Du Mont Laboratories, Admiral Corporation, General Electric, Philco Corporation, Sentinel Radio Corp., and Radio-Electronics magazine.

Bars, horizontal, wide	Line, dark vertical
Bars, horizontal, narrow	Lines, narrow, allover
Bars, sloping	Lines, white, curved
Bars, Vertical	Movement, horizontal
Bending	Movement, vertical
Brightness excessive	Multiple images
Brightness lacking	Pincushion effect
Centering incorrect	Raster only, no pattern
Contrast excessive	Reversed pictures
Definition poor	Shadows
Focus poor	Snow
Folded pattern, horizontal	Speckles or streaks
Folded pattern, vertical	Split pattern, left-right
Ghosts	Split pattern, top-bottom
Herringbone effect	Tear out
Jitter or jumping	Tilting in mask
Linearity poor, horizontal	Trailers or smears
Linearity poor, vertical	Wavy edges
Line only, no raster	Wedge shape
Line, bright vertical	



**Bars, Horizontal, Wide.**

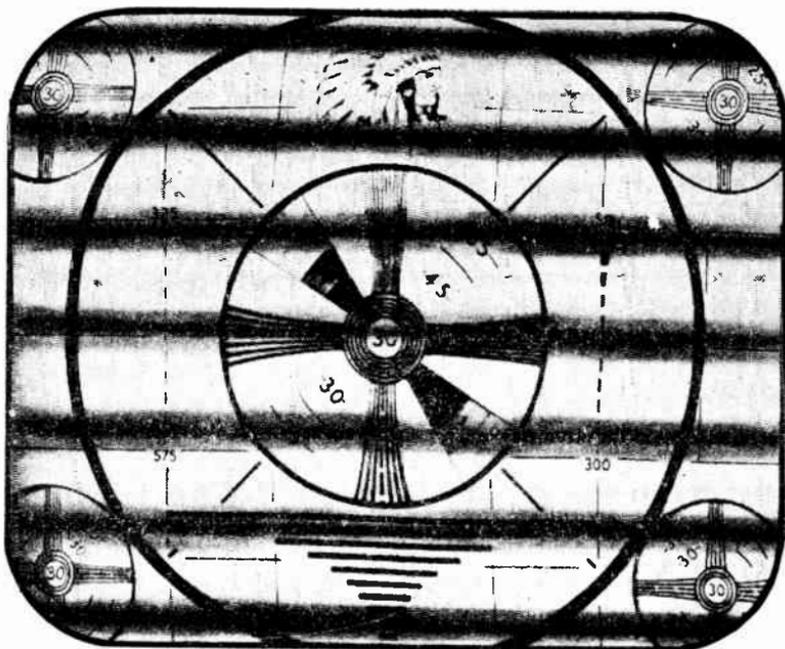
One or more wide black or dark toned bars separated by light toned spaces of about equal size. The bars remain stationary or nearly so. The condition may be called "hum," because the bars result from faults which would cause hum in a sound receiver.

*Causes for trouble.*

Insufficient filtering in low-voltage B-power supply. Open filter capacitors, etc.

Open or disconnected decoupling capacitors in sound or video circuits.

Locate points at which audible hum is present by using a signal tracer, or use oscilloscope for visual tracing of hum voltages.



**Bars, Horizontal, Narrow. "Sound Bars."**

A number of dark toned horizontal bars which result from audio-frequency voltages getting into the video i-f amplifier, video detector, or video amplifier circuits. If the audio frequency is constant, the bars remain stationary or move up or down at a uniform rate. If the audio frequency is modulated, the bars weave and cross with the modulation. The number of cycles per second of the audio frequency causing the bars is approximately equal to 60 times the number of bars.

*Causes for trouble.*

Fine tuning control misadjusted.

Sound volume control set too high.

R-f oscillator incorrectly aligned. Change alignment by least amount which gets rid of the bars.

Microphonic tube in r-f section, or possibly in i-f amplifiers, sync, or sweep sections. Tap tubes lightly to identify the one causing trouble. If bars are most noticeable with the sound volume control at maximum the fault probably is in the sound amplifier. If bars are most noticeable with volume control at minimum, check the other sections mentioned.

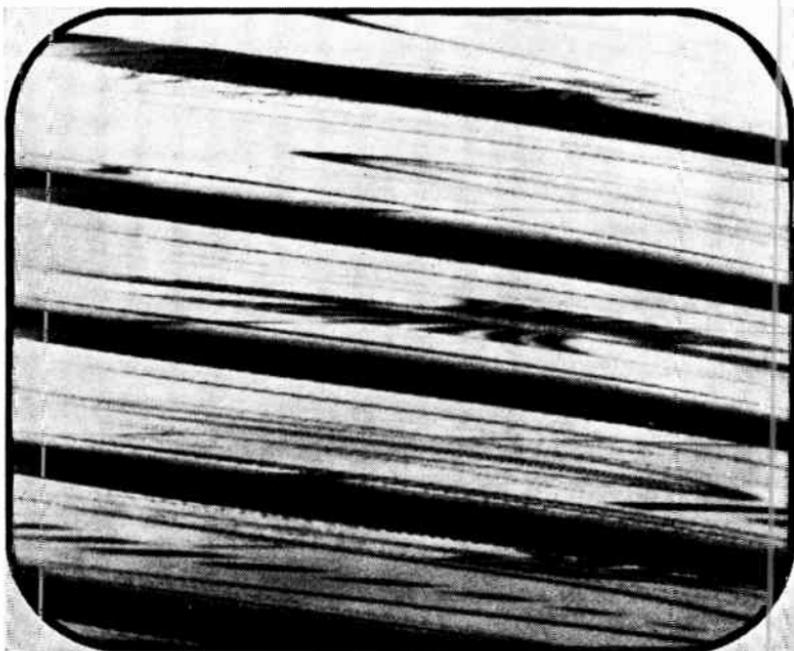
Traps for accompanying sound or adjacent sound incorrectly aligned.

Sound takeoff coupling incorrectly aligned.

Video i-f alignment wrong.

Regeneration in video i-f amplifier.

Check also for troubles listed under *Lines, Horizontal.*



***Bars, Sloping.***

The bars change their degree of slope, their number, and their positions as the horizontal hold control is altered.

*Causes for trouble.*

Horizontal hold control incorrectly adjusted.

Faulty connections, resistors, or capacitors in circuits for horizontal hold control or for horizontal automatic control of sweep frequency.

See also troubles listed under *Movement, Horizontal.*

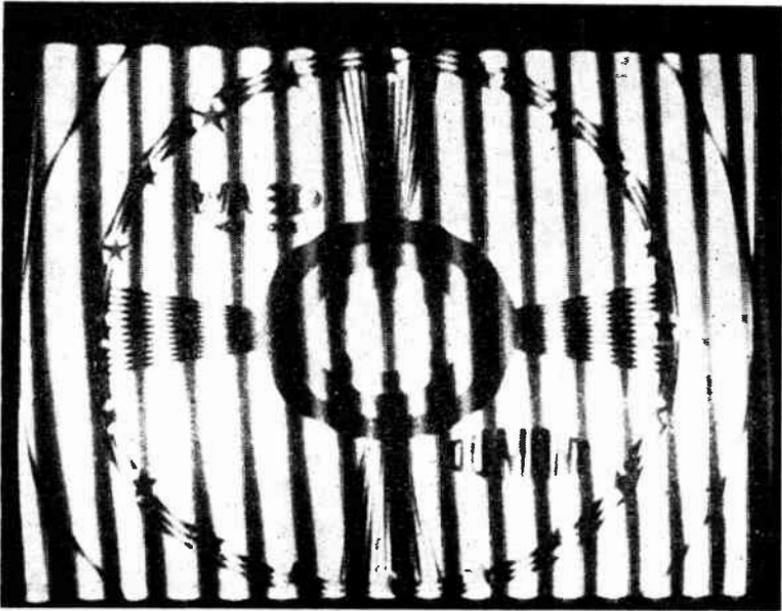
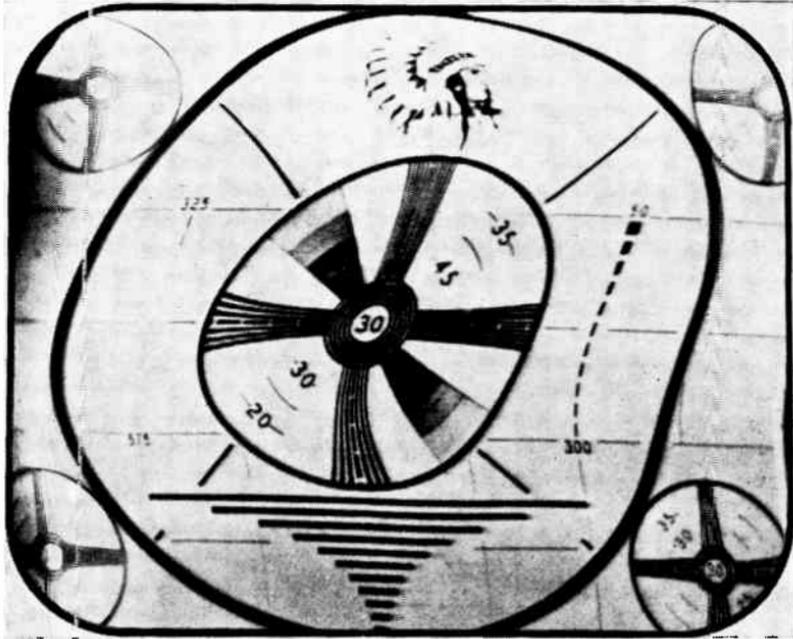


Fig. 131-4.—Bars, Vertical.

*Causes for trouble.*

Usually due to beat interference at frequencies lower than standard a-m broadcast carriers but higher than audio frequencies. The interference may originate inside or outside the receiver. The bars may remain stationary, may move to the right or left, or may change their slope.

The interference frequency is equal approximately to the number of dark or light bars multiplied by 15,750. Presence of the interference voltage at various points may be identified by means of an oscilloscope whose internal sweep is adjusted to maintain a steady trace pattern with the vertical input connected to various receiver circuits.



***Bending and Pulling.***

Lines which should be vertical at the top of pictures bend to the right or left. The degree of bend varies with movement of objects in pictures and with adjustment of the horizontal hold control. The underlying cause is weak horizontal sync pulses.

***Causes for trouble.***

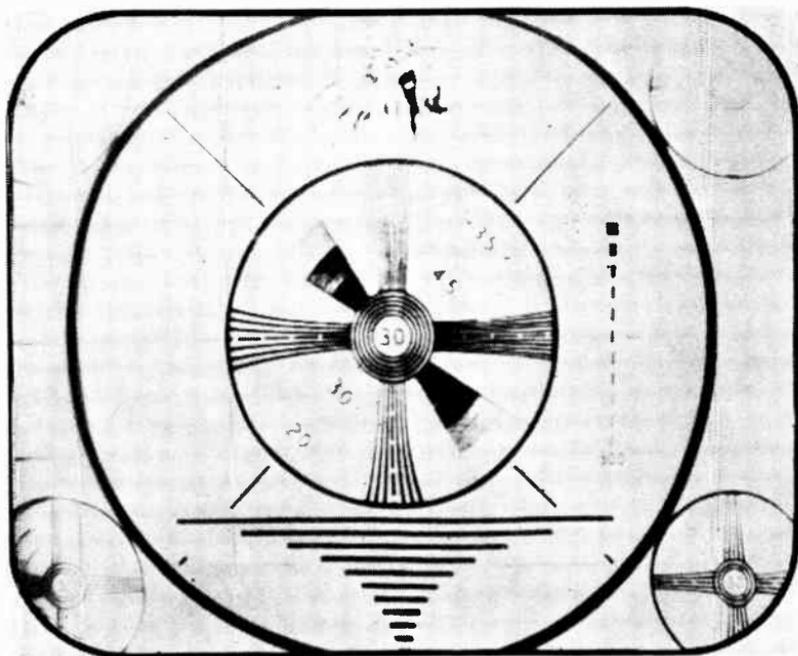
Horizontal afc system adjustments incorrect. Control tube weak. Defective capacitor or resistor. Ripple or audio voltage getting into afc circuits because decoupling resistor or capacitor faulty.

Composite video signal to sync section is too weak. Incorrect voltage on some sync tube, or defective tube. Defective coupling capacitor or plate load resistor in sync section.

Video amplifier tube defective or operating with wrong plate, screen, or grid voltage. Faulty coupling capacitor or plate load resistor in vidac amplifier section.

I-f amplifier or tuner aligned for video carrier of intermediate too low on frequency response.

Agc system not maintaining grid voltages sufficiently negative, to allow overloading of tubes. Agc delay or threshold misadjusted.



**Brightness Excessive or Varying.**

*Causes for trouble.*

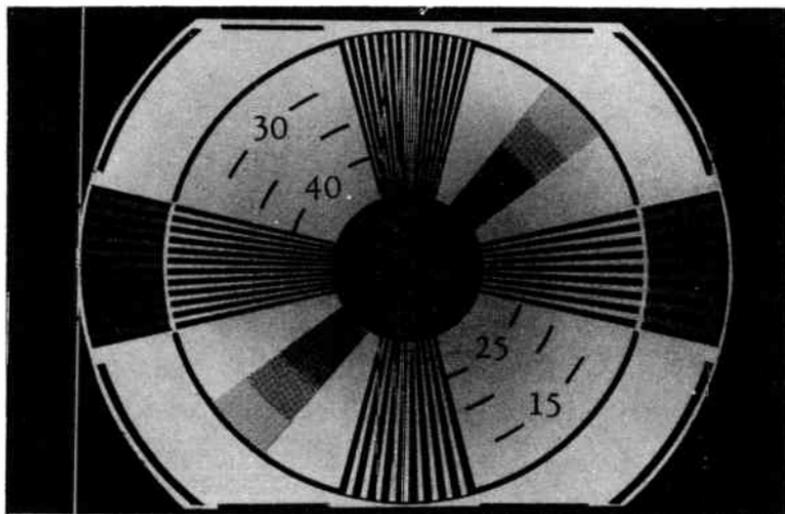
Brightness control turned too high. Vertical retrace lines appear.

Brightness control defective or open circuited.

Open circuit in picture tube grid-cathode circuit.

Trouble in d-c restorer circuits. Brightness varies irregularly.

Oscillation in video i-f amplifier. Picture tube screen becomes brilliant white, with no pattern or picture. Receiver usually must be turned off to stop the oscillation.



***Brightness Lacking.***

It is assumed that details are well rendered, but that pictures are too dark all over.

*Cause: for trouble.*

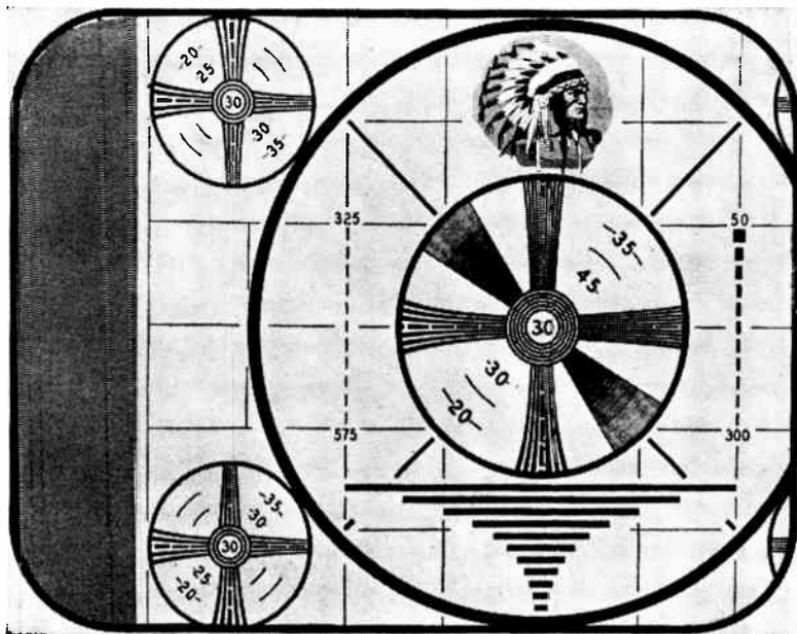
A-c power line voltage, or low B-voltage from the d-c power supply system.

Ion trap magnet misadjusted, or magnet weak.

I-f amplifier or tuner tubes weak, or operated with plate and screen voltages too low or with grid bias too negative. Check agc system. Dirty or bent contacts in tuner switch.

I-f amplifier misaligned, with video intermediate too low on the frequency response.

Picture tube grid too negative with reference to its cathode. Check voltages and connections in the brightness control circuits.



**Centering Incorrect.**

Picture or pattern may be too high or too low, incorrect vertical centering, or it may be too far to the right or left, incorrect horizontal centering, or there may be incorrect centering in both directions at once, as in the photograph.

*Causes for trouble. Magnetic deflection.*

Focusing control wrongly adjusted.

Ion trap magnet in wrong position on picture tube neck, or weak.

Horizontal hold control misadjusted.

Focusing coil axis direction requires adjustment. Should be in line with picture tube axis.

Focusing coil too far forward or back. Usually should be  $\frac{1}{4}$  to  $\frac{3}{8}$  inch from the deflection yoke.

Focusing coil short circuited.

Deflection yoke too far back on neck of picture tube, or not centered around neck.

Defective bypass capacitor on focusing control.

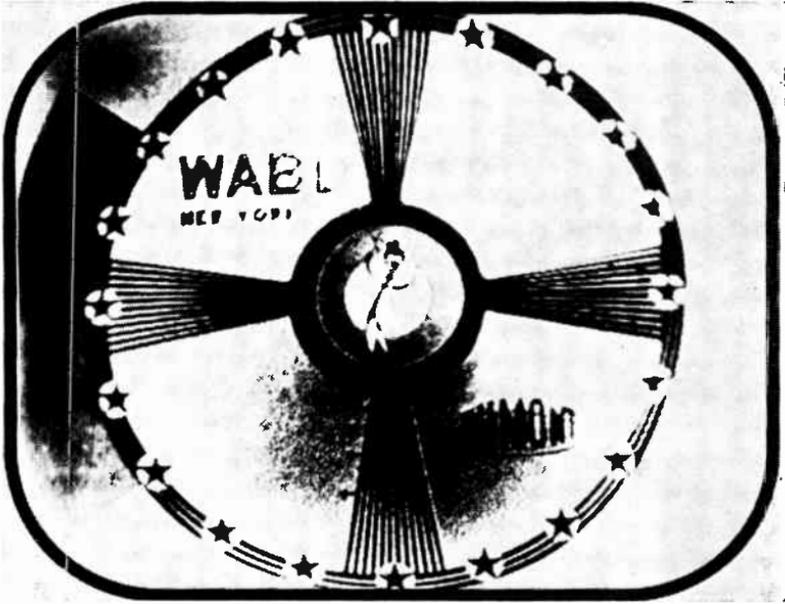
*Causes for trouble. Electrostatic deflection.*

Centering control or controls wrongly adjusted.

Horizontal hold control misadjusted.

Picture tube shield magnetized.

Leaky capacitor or capacitors between outputs of deflection amplifiers or oscillators and the picture tube deflection plates.

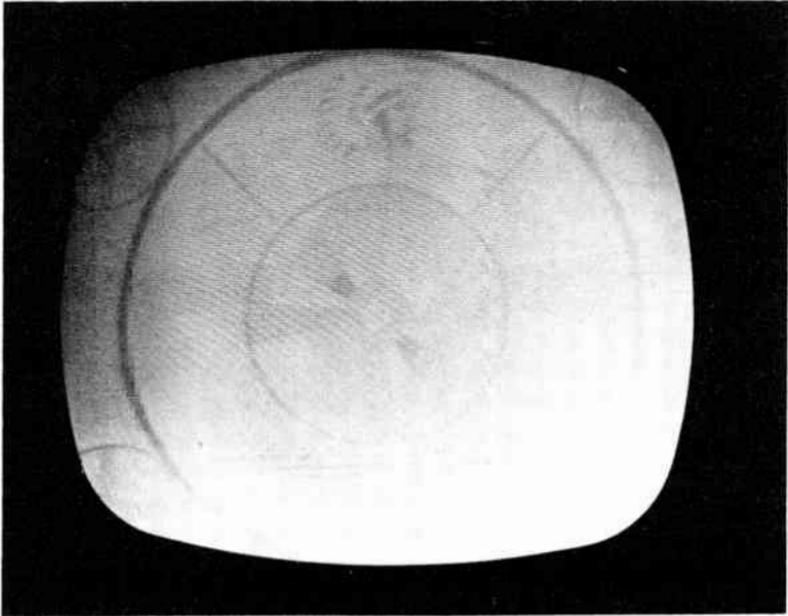


**Contrast Excessive.**

*Causes for trouble.*

Contrast control adjusted too high.

Faults in automatic gain control circuits or tubes.



*Contrast Insufficient.*

*Causes for trouble.*

Contrast control adjusted too low.

Brightness control too high. Causes vertical retrace lines as in photograph.

Weak signal reaching the antenna coupling input at the tuner. Check the transmission line, the antenna, and their connections. Built-in or indoor antenna may not deliver enough signal where the receiver is located.

Video i-f amplifier wrongly aligned, or has defective tube.

R-f amplifier or antenna coupling, or both, misaligned. Defective tube in r-f amplifier position.

R-f oscillator wrongly aligned, or weak tube.

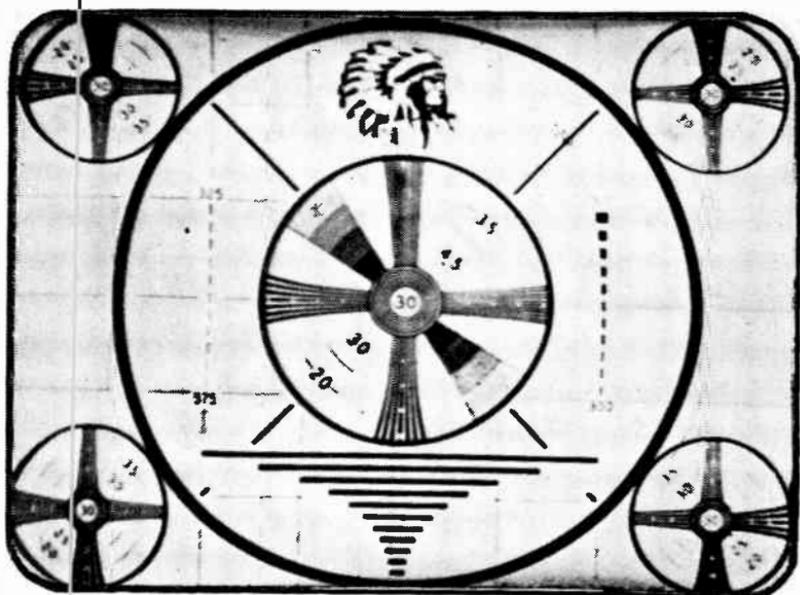
Defective tube or crystal for video detector.

Automatic gain control providing excessive negative bias.

Video amplifier tube defective.

Fault in d-c restorer circuit, or defective tube.

Picture tube defective, or so old as to have poor emission.



*Definition Poor, Detail Lacking.*

Poor definition which is not due to faults in focusing nearly always results from lack of response or lack of sufficient gain at the high video frequencies. Such lack of high frequency response is evident when lines and spaces of the vertically extending resolution wedges blur or run together a short distance inside the upper and lower sides of the pattern.

*Causes for trouble.*

Contrast control too high.

Fine tuning control misadjusted.

Weak received signal.

Ghost images. See troubles listed under *Ghosts*.

Video i-f amplifier incorrectly aligned. Not enough gain or response toward the sound intermediate frequency side of the response curve.

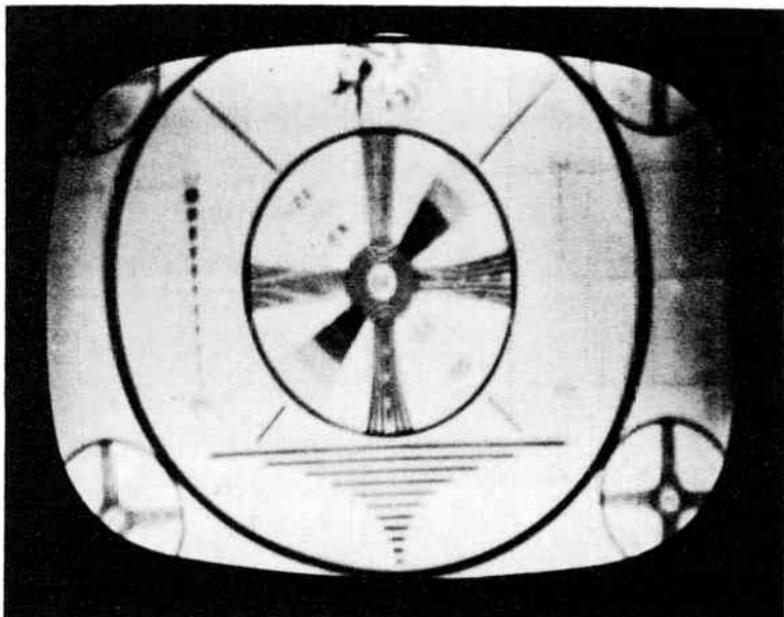
R-f amplifier or antenna coupling not aligned to provide sufficiently wide band pass.

Voltage too low on high-voltage anode or anodes of picture tube. Check the high-voltage power supply.

Video detector and/or video amplifier have poor response at the higher video frequencies. Coupling capacitors may be leaky or shorted. Peaking coils may be shorted. Try temporarily shorting each peaker. If this causes no change in definition, the shorted coil probably is defective.

Interlacing defective. Trouble in the sync section.

Focusing poor. See troubles listed under *Focus Poor*.



**Focus Poor.**

Focus is poor when horizontal traces or lines on the picture tube screen are not clearly separated when looking closely at the tube. Focus will vary in sharpness at different areas, even with best possible operation.

*Causes for trouble.*

Focusing control incorrectly adjusted.

Ghost images causing only slight displacement in the pattern. See *Ghosts*.

Focusing coil in wrong position, not centered around picture tube neck.

Ion trap magnet not in correct position, or weak.

Deflection yoke in wrong position on picture tube neck. Causes poor focusing at corners or outer edges of screen.

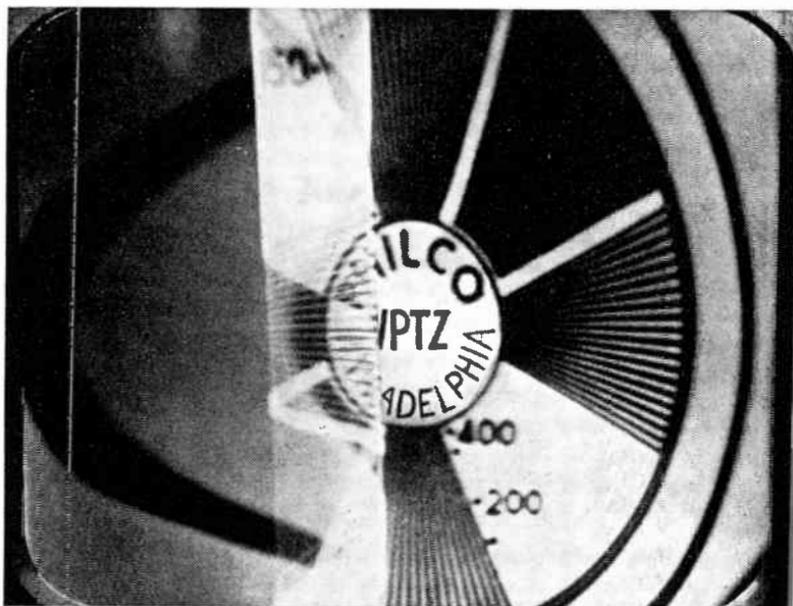
Voltage too low on high-voltage anode or anodes of picture tube. Check the circuits and tube or tubes in the high-voltage power supply.

Voltage and current too low or too high in focusing coil of magnetic focusing picture tube, or voltage too low or too high at focusing electrode of electrostatic picture tube. Focusing will be best with control moved to either end of its range.

Picture tube defective. Lack of emission due to age. Possibly gassy.

Weak signals may cause appearance similar to that due to poor focusing. See troubles listed under *Snow*.

Poor definition or resolution may appear like poor focusing. See troubles listed under *Definition Poor, Detail Lacking*.



*Folded Pattern, Horizontally.*

The picture or pattern appears as though it has been pushed together from left to right.

*Causes for trouble.*

Incorrect adjustment of horizontal drive control.

Insufficient sawtooth voltage from horizontal sweep oscillator or amplifier.

Trouble in damper circuit of magnetic deflection system, or defective damper tube.

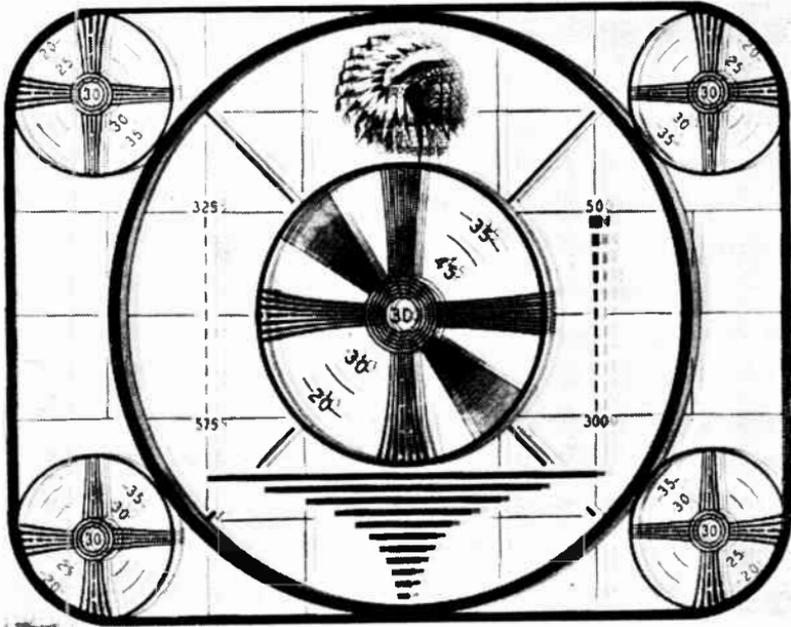


*Folded Pattern, Vertically.*

*Causes for trouble.*

Vertical hold control incorrectly adjusted.

Faults in vertical hold control circuit causing vertical deflection frequency which is too high.

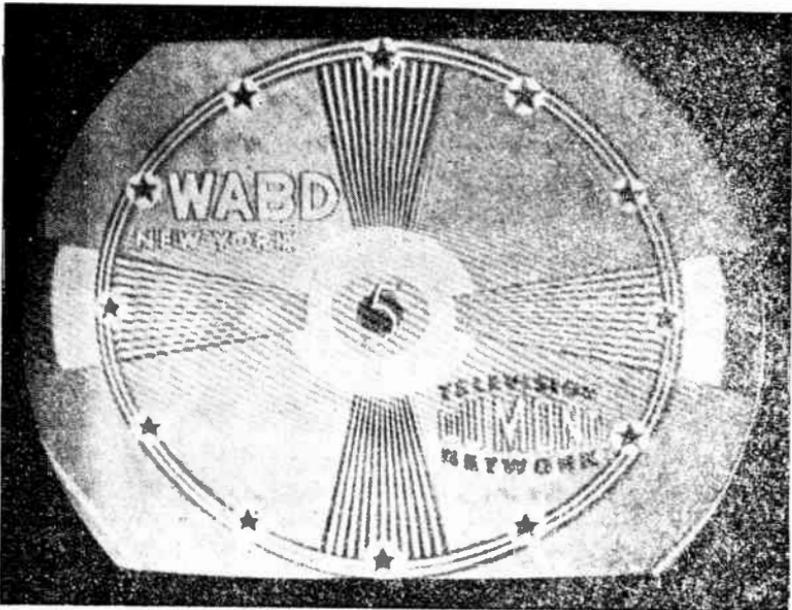
**Obsta.**

There are multiple images in the test pattern or picture. The displaced images, of which there may be one or more, may be so close to the principal image or may be so faint as to cause only a blurring effect. In other cases the displaced images may be at a considerable fraction of inch from the principal image, and may be distinct.

*Causes for trouble.*

Part of the transmitted signal is being reflected from large conductive or semi-conductive objects, such as buildings, bridges, tanks, or steep hills, and the reflected portion is reaching the receiver antenna a fraction of a second later than the direct signal. Try rotating the receiving antenna to reject the reflected signal without too much loss of direct signal. Fit a reflector, and possibly also a director, on the antenna. Try the antenna in various locations.

Incorrect matching of impedances between antenna and transmission line, or between transmission line and receiver input. There are standing waves on the line. Use antenna and transmission line whose impedances match that of the receiver and of each other.



*Herringbone Effect.*

A series of fine or coarse irregular lines and curves extending horizontally near the top, center, or bottom of the picture area. The photograph shows a rather faint effect across the center of the pattern.

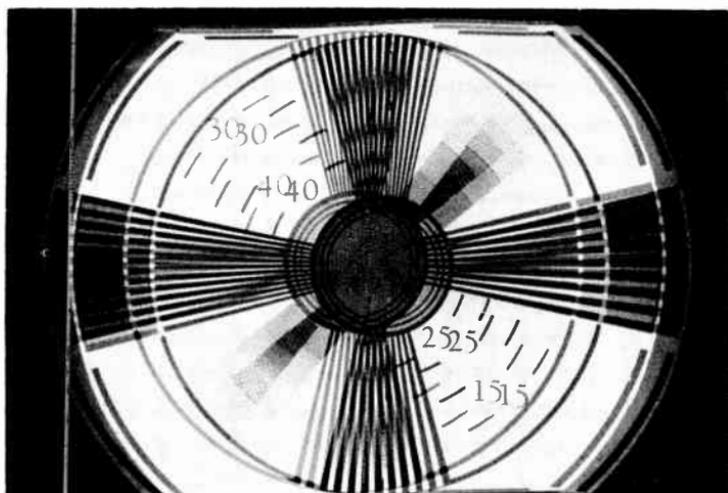
*Causes for trouble.*

Interference from medical apparatus, such as diathermy equipment, or from any similar devices being used in the neighborhood.

Signals from f-m radio stations operating in the area where the receiver is located. Other high-frequency transmitters could cause a similar effect. Turning the antenna or changing its location may be of help. An antenna trap may be tuned to the interfering frequency.

Regeneration in the video i-f amplifier.

Regeneration in the sound i-f amplifier.



*Jitter or Jumping.*

Pictures shift rapidly back and forth from side to side through small distances.

*Causes for trouble.*

Pickup of strong noise pulses through the antenna, or from defects within the receiver.

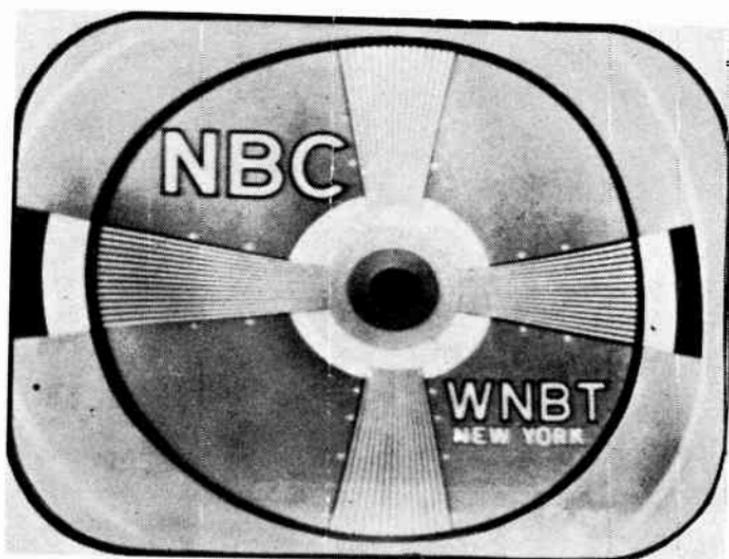
Check mechanical condition of switch contacts in units such as the tuner, a sensitivity switch, a TV-Phono-Radio switch, and others. Check contacts of tube base pins in sockets.

Defective tubes. Tap lightly to check for microphonic condition or internal short circuits.

Horizontal afc system may vary a frequency correction voltage too much or too rapidly. Check this system, especially its noise filter capacitors and resistors.

Agc system may contain loose contacts or defective capacitors.

Lack of decoupling in horizontal sweep circuits beyond the horizontal oscillator. Check B-plus decoupling capacitors and resistors.



*Linearity Poor, Horizontal.*

Linearity is the name given to distortion of outlines or of proportions in television patterns and pictures. With poor horizontal linearity there may be stretching at the right and crowding at the left, as in the photograph, or there may be stretching at the left with crowding at the right. Circles become deformed.

*Causes for trouble.*

Horizontal linearity control or controls misadjusted.

Horizontal drive control or peaking control misadjusted.

Defective inductors, capacitors, or resistors, fixed or adjustable, in horizontal linearity control circuits.

Horizontal sweep oscillator tube defective, or supplied with incorrect voltages on its elements.

Horizontal sweep amplifier tube defective, or supplied with incorrect voltages.

Defective damper tube in magnetic deflection system.

Trouble in any circuits or parts which follow the horizontal sweep oscillator, and in which are sawtooth voltages and currents.

Horizontal sweep output transformer defective.

Shorted turns in horizontal coil or coils of the deflection yoke.

Poor filtering of low-voltage B-power supply.



*Linearity Poor, Vertical.*

With poor vertical linearity the pattern or picture is compressed or flattened from above, below, or from both directions.

*Causes for trouble.*

Vertical linearity control wrongly adjusted.

Defective capacitors or resistors, fixed or adjustable, in vertical linearity control circuits.

Vertical sweep oscillator tube defective, or supplied with wrong voltages.

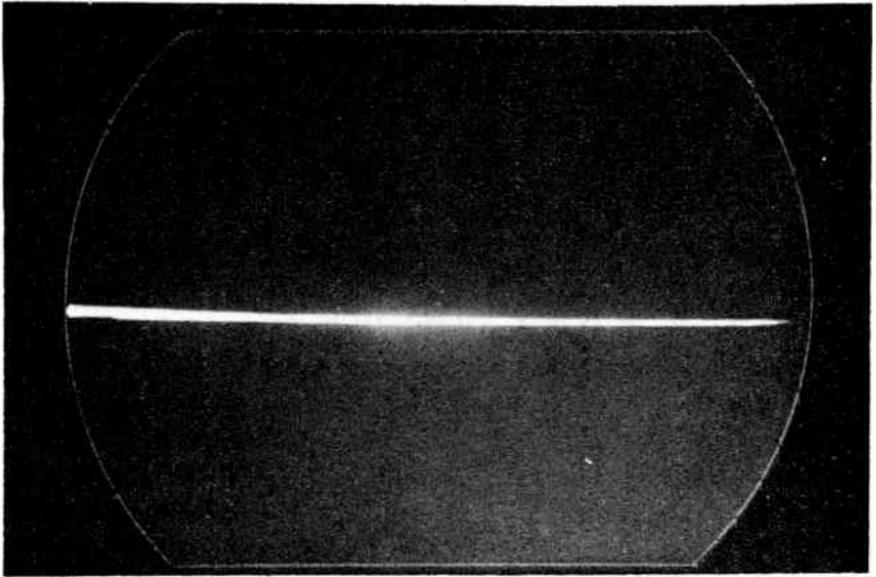
Vertical sweep amplifier tube defective, or supplied with incorrect voltages.

Trouble in any parts which follow the vertical sweep oscillator, and which carry sawtooth voltages and currents.

Vertical sweep output transformer defective.

Shorted turns in a vertical deflection coil.

Poor filtering of low-voltage B-power supply.



*Line Only, No Raster or Pattern.*

There may be a single trace line running horizontally, as in the photograph, or a similar single line running vertically on the screen.

*Causes for trouble.*

A horizontal line indicates that there is no vertical deflection, only horizontal deflection remains. A vertical line indicates that there is no horizontal deflection, only vertical deflection remains.

With no vertical deflection the trouble may be anywhere from the vertical sweep oscillator through to the deflection yoke of a magnetic picture tube or to the deflection plates of an electrostatic tube. Conversely, with no horizontal deflection the trouble may exist anywhere from the horizontal sweep oscillator through to the deflection elements at the picture tube.



*Line, Bright Vertical.*

The basic fault is in horizontal deflection of the electron beam at a rate not constant all the way across, but delayed where one or more bright vertical lines appear.

*Causes for trouble.*

Horizontal sweep oscillator tube weak or operated with wrong element voltages. Check also for defective capacitor in grid circuit of this tube, for sawtooth capacitor on the plate side defective or of wrong value, and for defective resistors in the plate circuit.

Horizontal output amplifier tube weak. Check also for leaky coupling capacitor and for otherwise defective capacitors or resistors between this amplifier and the horizontal oscillator.

Damper tube may be defective, or have faulty capacitors or resistors in its circuit leading to the boosted B-voltage connection.

A width control inductor on the horizontal output transformer may be connected to taps between which there are too many turns.



*Line, Dark Vertical.*

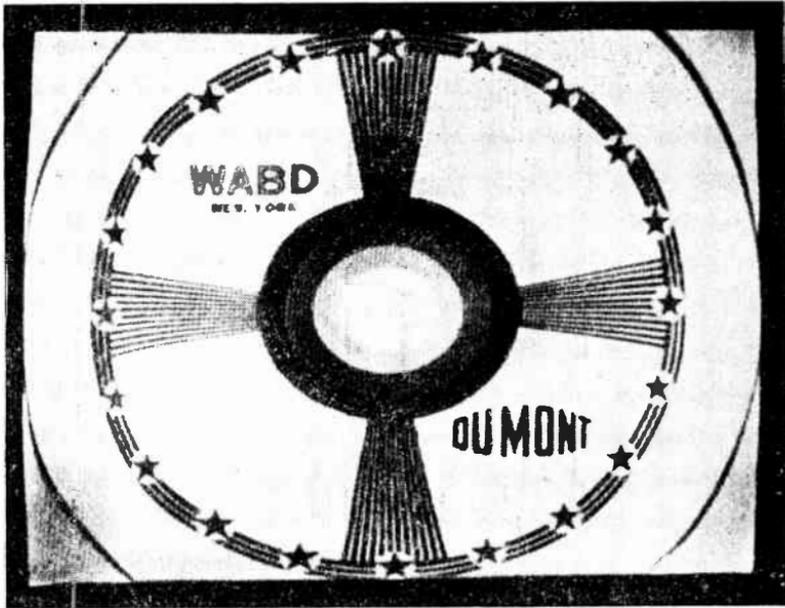
This usually is a case of Barkhausen oscillation explained under *Oscillation*. The line may weave slightly. Ordinarily it shows only on pictures, not on a raster.

*Causes for trouble.*

Misadjustment of horizontal drive, linearity, or width controls.

Horizontal output amplifier tube defective, or operated with excessive screen voltage and plate current.

A common remedy is a device similar to a single ion trap magnet placed on the horizontal output amplifier glass envelope near the top. A fixed non-inductive resistor of about 50 to 100 ohms in series with the amplifier grid may prevent oscillation.



***Lines, Narrow, Allover Pattern.***

Due to beat interference from radio frequency and television frequency signals or voltages originating from outside or within the receiver. The number of cycles per second of the interfering frequency is equal approximately to the number of lines, either light or dark, but not both, multiplied by 15,750. The lines may lie vertically or diagonally on the picture tube screen. They weave or ripple and change their direction.

*Causes for trouble.*

Interference from f-m radio broadcasting stations operating in the area where the receiver is located. Change the direction of the receiving antenna. Tune an antenna trap to the interfering frequency. Check the transmission line for possible signal pickup.

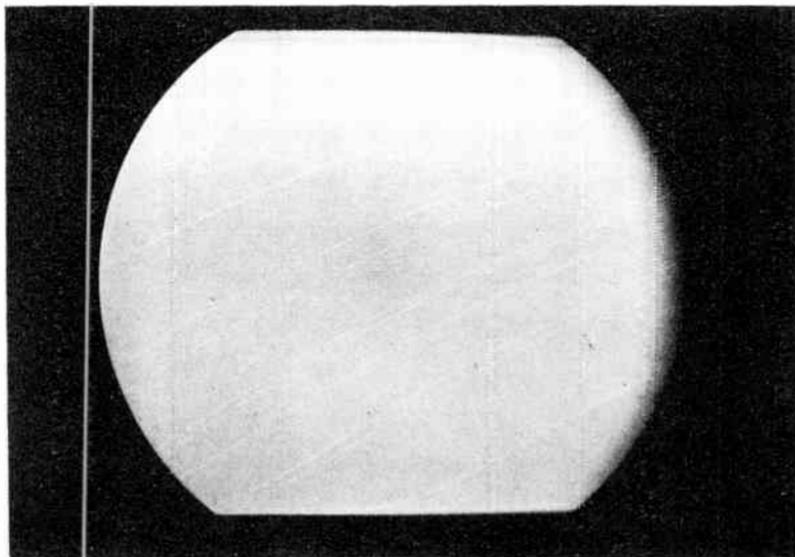
Interference from nearby short-wave transmitters. Same remedies as for f-m interference.

Interference from television channels other than the one to which the receiver is tuned. Try adjusting the fine tuning control.

Beating frequency of 4.5 megacycles from sound section of a receiver having intercarrier sound system, or getting past the sound takeoff and reaching the picture tube grid cathode circuit through all or part of the video amplifier. Check dressing of all grid and plate leads following the takeoff.

Incorrect dressing of capacitors and resistors in grid and plate lines of the video amplifier, or dressing picture tube grid or cathode leads too close to the chassis.

Interference from r-f type high-voltage power supply in the receiver. Not shielded, shield not connected to chassis or B-minus, poor filtering at output or input, defective or shorted isolating chokes or capacitors. Incorrect dressing of r-f oscillator plate and grid leads.



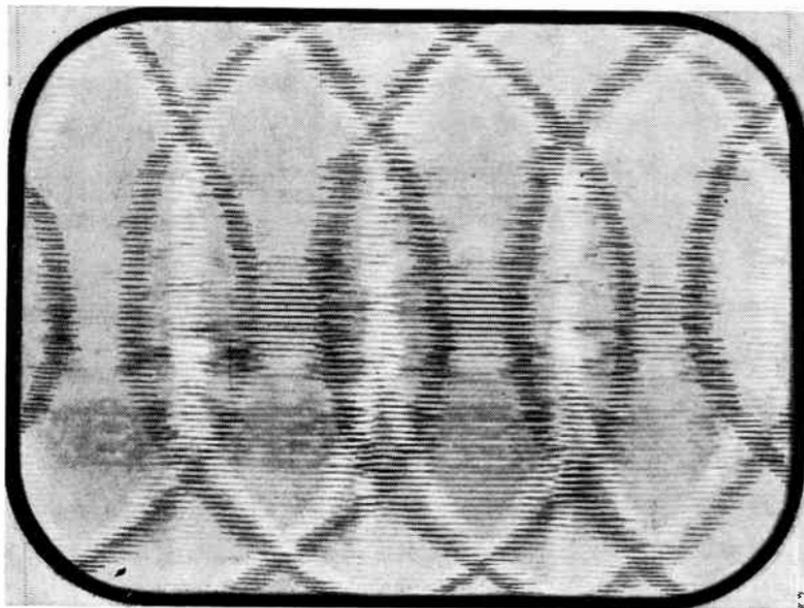
*Lines, White, Curved and Sloping.*

These are called vertical retrace lines. The photograph shows such lines on a plain raster. They appear similarly on patterns and pictures.

*Causes for trouble.*

Brightness control too high, contrast too low, or both together. Adjust one or both controls.

Vertical hold control slightly out of adjustment. The retrace lines will appear only while the picture or pattern moves slowly up or down. Adjust the control.



*Movement, Horizontal.*

The parts of the pattern or picture move slowly or rapidly toward the right or left, sometimes with complete scattering and breakup of all elements, and again with the appearance of strong diagonal bars at changing angles. These things are due to faulty horizontal synchronization.

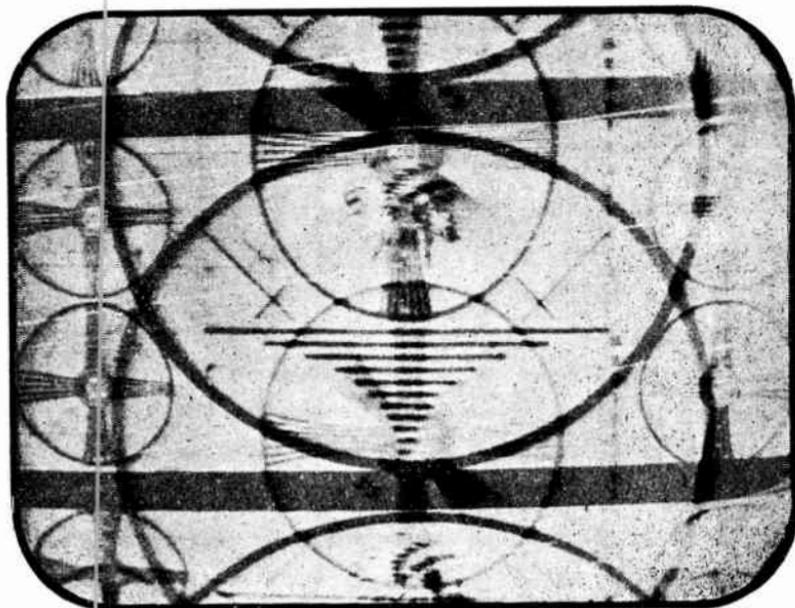
*Causes for trouble.*

Strong reflected signals prevent synchronizing action of the horizontal hold control instead of having their more usual effect of producing ghost images.

Leaky, shorted, or open capacitors or resistors in the horizontal hold control circuit that is connected to the grid of the horizontal oscillator.

Defective horizontal oscillator tube, or incorrect tube voltages.

Trouble in the horizontal automatic frequency control for synchronizing. Check tubes, tube voltages, transformer adjustments, coupling capacitors, and all resistors.



*Movement, Vertical.*

The pattern or picture moves more or less rapidly up or down on the screen of the picture tube, or appears to "roll." Vertical blanking bars appear crosswise of the screen, or, with very slow movement, there may be only one such bar at a time.

*Causes for trouble.*

Vertical hold control misadjusted.

Strong reflected signals may interfere with action of the vertical hold system.

Corona or arcing in high-voltage power supply may interfere with vertical hold.

Vertical sweep oscillator tube defective or supplied with incorrect voltages. Opens, leakages, or shorts of capacitors and resistors connected to the grid of the vertical sweep oscillator tube.

Leaky capacitors, or resistors open or shorted, in the integrating filter between the sync section and the vertical sweep oscillator.



*Movement, Vertical and Horizontal.*

*Causes for trouble.*

Both the horizontal and the vertical hold are out of adjustment.

Contrast control adjusted too low.

Weak signal reaching the antenna or tuner terminals of the receiver.

Strong electrical interference of the spark type.

Corona or arcing in the high-voltage power supply system of the receiver.

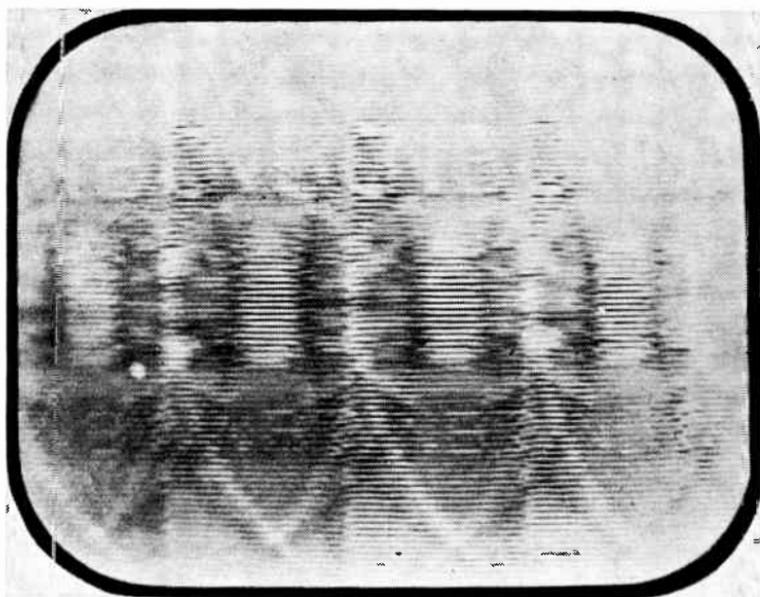
Strong reflected signals reaching the antenna. Such signals ordinarily cause the appearance of ghost images.

Faults in the sync section, anywhere from the takeoff through to the integrating and differentiating filters. Check tubes, tube voltages, coupling capacitors, resistors, and dressing of grid and plate leads.

Video i-f amplifier incorrectly aligned, so that the video intermediate frequency is too far down on the response curve. This reduces low-frequency response.

R-f amplifier or antenna coupling incorrectly aligned, reducing the response at low video frequencies.

D-c restorer inoperative. This trouble may cause loss of synchronization chiefly on dark toned pictures. Check the tube and all parts of its circuit.



***Multiple Images.***

Multiple images consist of alternate light and dark lines closely following at the right of any sharp changes of shading. The successive images or lines become less and less distinct, because they result from damped oscillation in the video amplifier or closely associated circuits. Multiple images and ghosts or reflections may be distinguished from each other because almost always there will be but a single ghost image, while close inspection will show two or more sets of equally spaced lines in multiple images.

*Causes for trouble.*

Inductance may be too great in one or more peakers of the video amplifier section. On peakers through which pass video signals from video detector through amplifier and to picture tube the shunt resistors may be open or, rarely, of too great resistance.

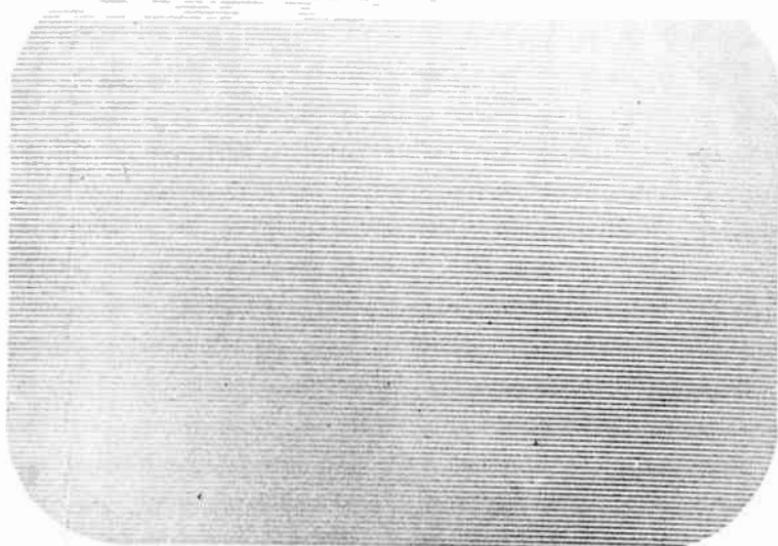
Plate and grid leads between video detector and picture tube signal input may be dressed close together, allowing regenerative feedbacks.

Mismatch of impedances in horizontal output transformer and horizontal coils of the deflecting yoke.

*Pincushion Effect.*

Corners of pictures extend outward, with top, bottom, and sides bowed inward. The fault may occur with picture tubes having cylindrical face plates.

Correction is by altering the positions of small permanent magnets supported on opposite sides of the picture tube close to the point at which the neck joins the flare. After loosening any locking device that holds the magnets, they may be moved toward or away from the tube, upward, downward, or to a different angle as may be required for straightening the sides of pictures.



***Raster Only, No Pattern or Picture.***

If there is no reproduction of sound, and neither pattern nor picture, the trouble may be any of those listed. If there is reproduction of sound the trouble can be only in parts of the receiver which follow the sound takeoff.

*Causes for trouble.*

Antenna or transmission line open circuited, shorted, have high resistance connections, or for any reason fail to deliver a signal to the receiver.

Dirt, looseness, open connections, or other faults in the tuner.

R-f oscillator or r-f amplifier out of alignment, tube defective, wrong voltages, or other troubles preventing operation.

Video i-f amplifier far out of alignment, defective tube, no voltage on one or more tubes, open or shorted coupling capacitors, etc.

Defective video detector tube or crystal.

Contrast control open circuited or otherwise defective.

Automatic gain control system supplying excessively negative voltage.

Faults in the video amplifier. Check tubes and their voltages, coupling capacitors, resistors, wiring connections, and other causes for no amplification.

Examine all circuits through to the picture tube signal input.



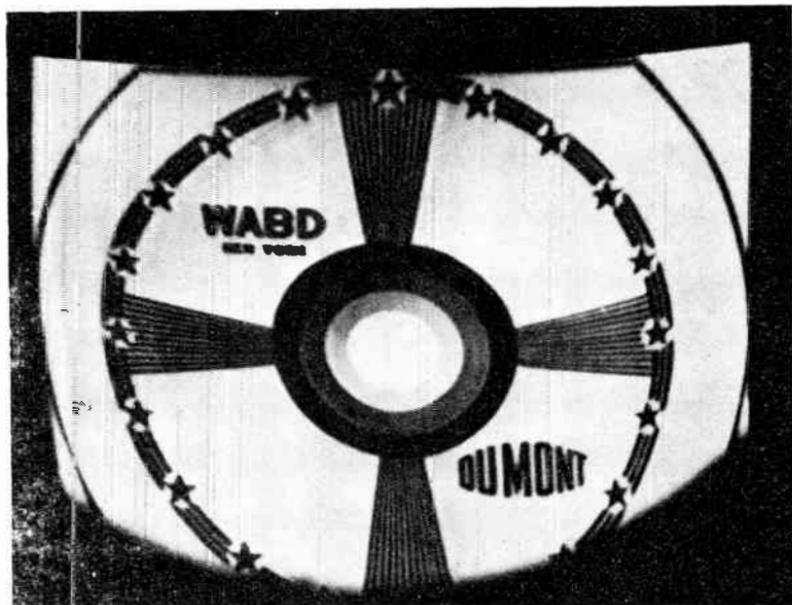
*Reversed Pictures.*

Reversal between left and right, inversion between top and bottom, or both effects combined result from interchanging of leads to horizontal coils, vertical coils, or to both sets of coils in the deflecting yoke. Wrong connections may be anywhere between the sweep output transformers and the yoke.

Pictures right side up but reversed left and right means wrong connections only to the horizontal coils. This is shown by the photograph.

Pictures upside down but not reversed left and right means wrong connections only to vertical coils.

Pictures reversed left and right, also upside down, means wrong connections to both sets of coils in the yoke.



**Shadows.**

*Cause: for trouble.*

Ion trap magnet incorrectly adjusted, in wrong position.

Focus coil or magnet not centered around neck of picture tube.

Focus coil or magnet too far back on picture tube neck, too close to base.

Deflection yoke too far back from flared portion of picture tube. It should be as close as possible to the flare.

Frame which carries the yoke and focusing coil or magnet out of position.

Picture tube may require slight rotation around its axis.

Reversing the connections to the focusing coil may help in some cases.



*Size Incorrect.*

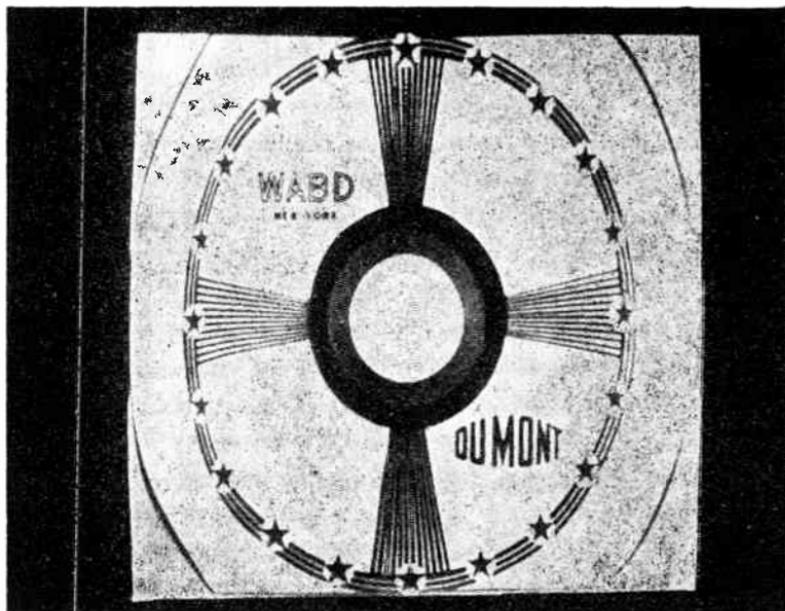
The size of the picture, with reference to the mask opening or screen size, may be too small in a vertical direction as shown by Fig. 131-24 or too small in a horizontal direction as shown by Fig. 131-25. In some cases the size may be too great in both directions. In considering these faults it is assumed that both the vertical and horizontal size controls, or the height and width control, have been adjusted so far as is possible to make the size correct in both directions. Troubles listed below are to be looked for in the vertical sweep section when height is incorrect, or in the horizontal sweep section when width is incorrect, or in both these sections when size is incorrect both ways.

*Causes for trouble.*

Re-check adjustments of size control or controls.

Size may be affected by incorrect adjustment of linearity controls, either vertical or horizontal, and of horizontal drive or peaking controls.

*(Continued on following page)*



*Size Incorrect.*

Centering control adjustments may have reacted on size.

Sweep oscillator tube defective or supplied with wrong voltages.

Examine circuits which carry sawtooth voltages and currents, and check sawtooth waveforms with oscilloscope.

Trouble in circuits between sweep oscillator and amplifier. Examine coupling capacitors and voltage divider capacitors or resistors.

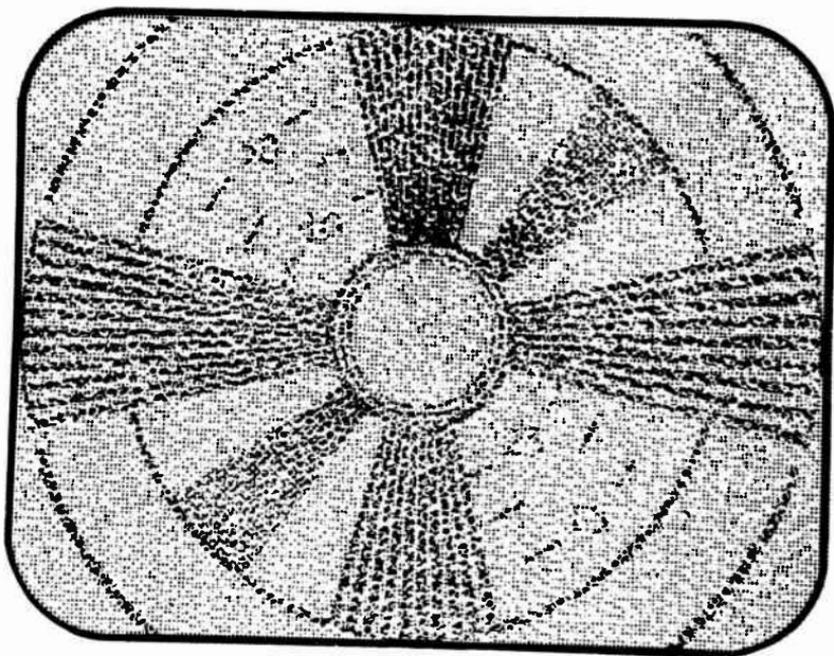
Defective sweep output amplifier tube, or tube supplied with wrong voltages.

Defective sweep output transformer.

Defective damper tube or incorrect damper adjustments. Only in horizontal magnetic deflection systems.

Picture tube may be gassy.

Size which remains excessive after all adjustments may indicate voltage too low on the high voltage anode or anodes of the picture tube.



Snow.

The pattern or picture appears covered with small white specks or flashes. The primary cause is a weak signal at the antenna input terminals of the receiver, or there is a low ratio of signal to noise. With the contrast control turned up for maximum gain there is high amplification of noise pulses, which cause the white flashes.

*Causes for trouble.*

Antenna trouble. Incorrectly oriented. Poor location. Shorted, open circuited, dirty or corroded terminal connections. The receiver may be located so far from transmitters as to require an outdoor antenna instead of a built-in or indoor type, or may be too far away for good reception under any conditions.

Transmission line open circuited, shorted, dirty connections, or may be picking up excessive interference.

Antenna coupling or r-f amplifier incorrectly aligned.

Weak or noisy tubes for r-f amplifier, mixer, or r-f oscillator.

Dirty contacts or loose connections in selector switch or tuner.

Video i-f amplifier incorrectly aligned or contains weak tubes.

Weak tube or crystal in video detector.

Video amplifier has weak tubes, or there are defective parts such as coupling capacitors.

Corona or arcing in parts or wiring of the high-voltage power supply may cause interference flashes.

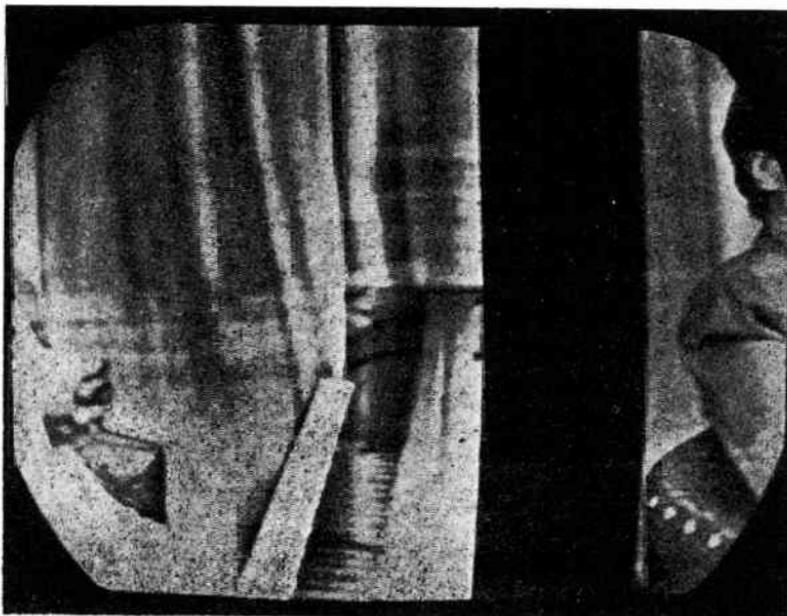


*Speckles or Streaks.*

Short, dark, irregularly spaced streaks somewhat longer from left to right than their height result from electrical impulse voltages commonly called "noise". Troublesome voltages come through the antenna or be produced within the receiver.

External causes for streaking are discussed in the article on *Interference* under the subheading of *Spark Interference*.

Among the more common internal causes for streaking are loose connections anywhere from tuner to picture tube, through the i-f amplifier, video detector, and video amplifier sections. This trouble may result also from internal shorts in tubes of the sections mentioned. Corona or arcing anywhere in the high-voltage power supply is another possible cause.



*Split Pattern, Left and Right.*

A horizontal blanking bar runs from top to bottom of the picture area, remaining stationary or moving slowly. The portion of the picture or pattern which should appear at the left is on the right side of this bar, and the portion which should be at the right is on the left side of the bar.

*Causes for trouble.*

Horizontal hold control wrongly adjusted.

Faults in horizontal afc system for sweep oscillator. Defective tubes, incorrect voltages. Wrong adjustments, as of phasing or lock-in controls. Coupling capacitors open or shorted.

Faults in horizontal hold control circuit, making frequency too low or of half the correct value. Check capacitors and resistors, incorrect values, opens, shorts, and leakage.

Excessive undamped oscillations in horizontal sweep oscillator circuit. Check resistors and decoupling capacitors.

Coupling capacitors for a sync inverter tube leaky or connected wrong.

Defective damper tube.

Oscillation in horizontal deflection coil. Check the bypass capacitor.



*Split Pattern, Top and Bottom.*

A vertical blanking bar runs from side to side of the picture area. The bar may be stationary or slowly moving. The top part of the picture will appear below the bar, and the bottom part will be above the bar. Sometimes there are two complete pictures, each occupying about half the total height in the mask.

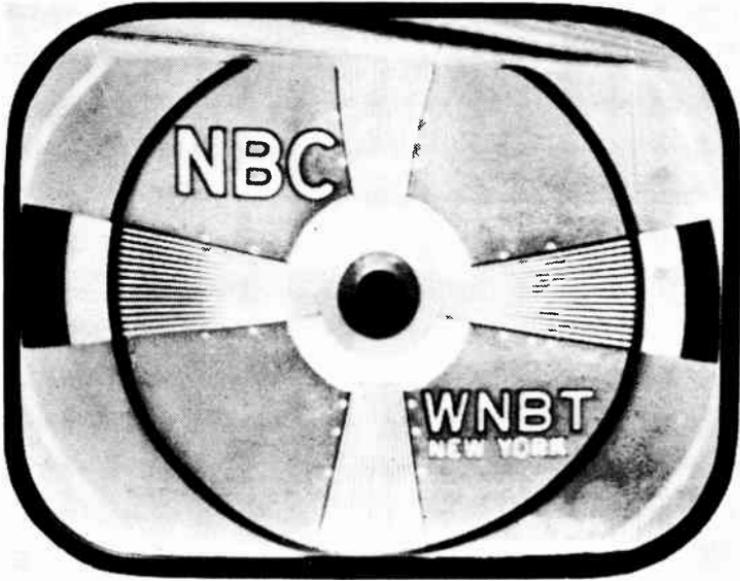
*Causes for trouble.*

Vertical hold control wrongly adjusted.

Faults in the vertical hold control circuit, making the vertical sync frequency too low or half its correct value. Check for leaky or shorted capacitors. Check for shorted resistors. Values of capacitors or resistors may be incorrect.

If there is a vertical afc system for the sweep oscillator, check it as explained for a horizontal afc system in the preceding section.

Sync inverter coupling capacitors leaky or connected wrong.



**Tear Out.**

*Causes for trouble.*

Horizontal hold control wrongly adjusted.

Contrast control set too high.

Excessively strong signal in channel where trouble appears.

Strong but irregular electrical interference temporarily affecting the horizontal hold control system.

Microphonic tubes anywhere in r-f, video, or sync circuits. Tear out will occur when the receiver is jarred.

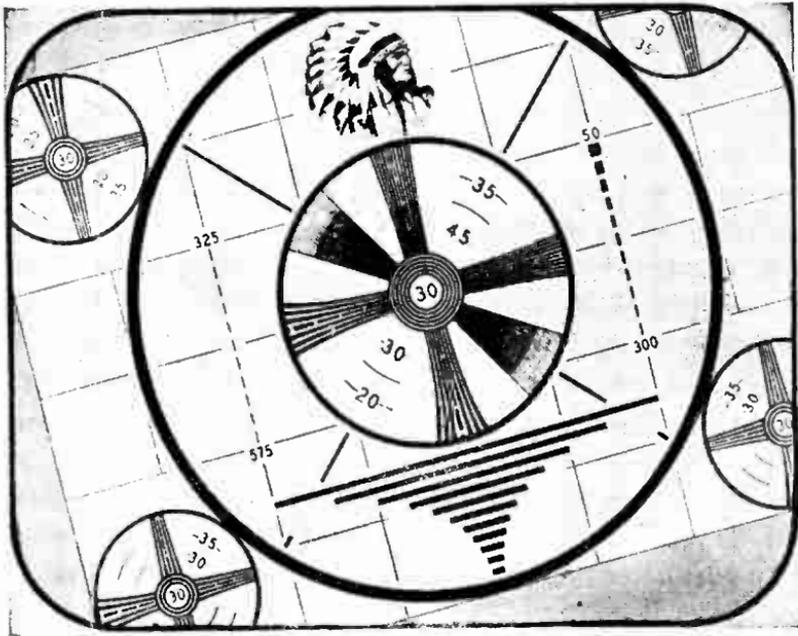
Poor filtering in low-voltage B-supply.

Alignment of video i-f amplifier incorrect, bringing video i-f carrier too low down on the response curve.

Misalignment of traps for accompanying or adjacent sound, allowing audio interference to reach the picture tube input.

Poor low-frequency response in the video detector or video amplifier circuits. Check the tubes, and their plate and grid bias voltages. Look for defective units and incorrect values in coupling capacitors, decoupling capacitors, cathode bypass capacitors, decoupling resistors, and grid resistors.

Incorrect adjustments in the horizontal afc system for the sweep oscillator.



*Tilting In Mask.*

*Causes for trouble.*

Incorrect position of magnetic deflection yoke. Loosen the yoke fastening while rotating the yoke around the picture tube neck to straighten the pattern.



*Trailers or Smears.*

Dark objects or lines in the pattern or picture are followed on their right by light toned or white areas or lines. The trouble is due primarily to poor response or lack of gain at the low video frequencies.

*Causes for trouble.*

Contrast control too high, overloading the amplifiers.

Video i-f amplifier wrongly aligned, with the video i-f carrier too low on the side of the response curve.

Faults in video detector and/or video amplifier circuits. Check coupling capacitors, decoupling capacitors, and bypass capacitors for open circuits. Check plate load resistors and grid resistors for incorrect values or shorting. Amplifier grid biases may be insufficiently negative, allowing grid current to flow.

Ghost images due to signal reflections may give an appearance similar to trailing. See troubles listed under *Ghosts*.



*Wavy Edges on Pattern.*

Usually the right-hand edge of the pattern or picture is waved. Dark and light bands may appear. The entire picture may appear to sway sideways.

*Causes for trouble.*

Contrast control adjusted too high.

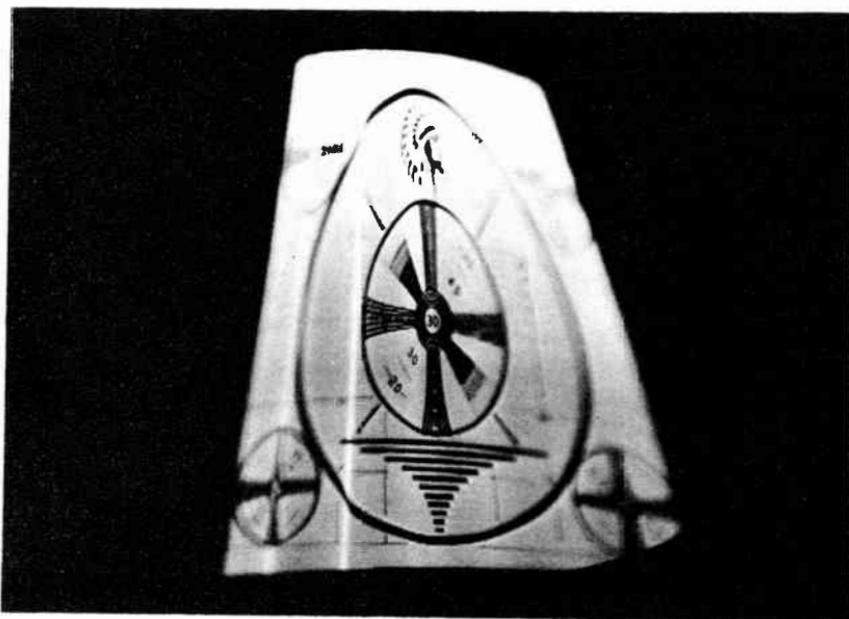
Horizontal drive control wrongly adjusted.

Insufficient filtering or defective filter capacitors in low-voltage B-power supply. The ripple voltage may be getting into the focusing coil or an electromagnetic centering circuit.

Defective horizontal sweep amplifier tube. Try changing the tube.

Vertical deflection voltages or fields reacting on horizontal deflection circuits. There may be coupling between the circuits, possibly in the deflection yoke. Check bypass capacitors and resistors which are across the deflection coils.

Defective capacitor or resistor in noise filter of a horizontal afc system for the sweep oscillator.



*Wedge Shape.*

Pictures are not of the same width at top and bottom, as in the photograph, or may not be of the same height at left and right.

*Causes for trouble.*

Slight tapering may result from the following: Deflecting yoke not concentric with tube axis or tilted with respect to axis. Focus coil or PM focuser in wrong position or not concentric with tube axis. Ion trap magnet in wrong position on tube neck, not adjusted for maximum brightness.

Severe tapering, as in the photograph, results from trouble in coils of deflecting yoke. Unequal widths at top and bottom indicate trouble in horizontal coils. Unequal heights left and right indicate trouble in vertical coils. A coil or its connections may be open or short circuited. A capacitor or resistor across a coil may be shorted or very leaky. Coils may have become displaced in the yoke structure.

**TROUBLE SHOOTING, WAVEFORMS AND FREQUENCY RESPONSES.** — The oscilloscope is used for observing voltage waveforms and frequency responses throughout the television receiver during location of trouble.

A trace of frequency response on the oscilloscope screen shows relative gains at all frequencies being fed into an amplifying section or any single stage of the receiver. Traces taken from the mixer tube circuits show response of the r-f amplifier or tuner to the carrier frequencies in a channel. Traces taken at the output of the video detector or from the video amplifier show performance of the video i-f amplifier, of this amplifier and the r-f section combined, or of any single stage or combination of stages. Response traces taken from the sound discriminator or ratio detector show performance of the sound section or of parts in this section.

Observation of frequency responses requires a signal from a sweep generator, and identification of frequencies by means of a marker generator, or requires use of a combination television generator. Instructions for checking frequency response are given in the following articles.

*Alignment.*

*Amplifiers, video intermediate-frequency.*

*Oscillators, television r-f.*

*Sound, television.* Also related subjects mentioned in that article.

*Tuners.*

Waveform traces show signal voltages and other voltages as they actually exist at various points between the video detector and the picture tube. Signal voltage and voltages resulting from or controlled by the signal may be compared with forms known to be correct for the receiver or for the type of circuit involved. Material differences indicate trouble at or preceding the point where the faulty waveform appears. Instructions for taking waveform traces, also typical examples of such traces, are given in the article on *Waveforms*.

**TUBE, ELECTRON-RAY.** — The electron-ray tube is a device for visually indicating increase or decrease of an applied voltage. Visible through the end or top of the glass envelope is a circular “target” which is made luminous over all of its area except a pie-shaped sector or shadow which becomes smaller as an applied voltage is made more negative, and larger as the applied voltage is made less negative. The electron-ray tube was used originally as a tuning indicator in sound radios, being operated by changes of negative a-c voltage with the receiver tuned on or off resonance. The tube is quite widely used instead of a d-c voltmeter in service instruments such as resistance bridges, capacitor testers, electronic voltmeters, and heterodyne frequency meters.

The type of electron-ray tube commonly used in service instruments contains the circular cup-shaped target, a deflector, and also an amplifying triode, with these elements arranged as in Fig. 1. The cathode extends through the triode and up inside the target. A circular light shield hides the

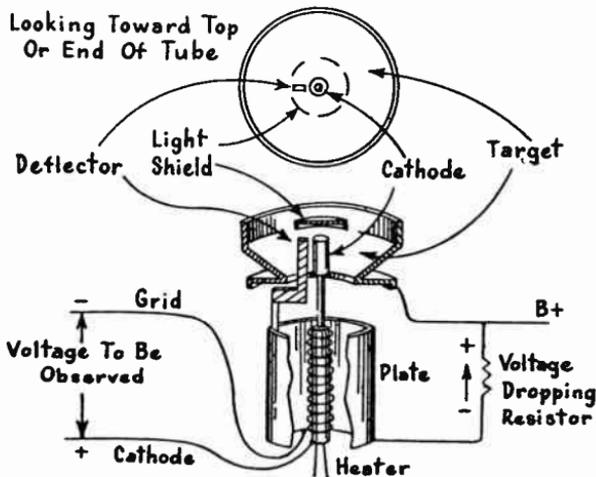


Fig. 1.—The elements in an electron-ray tube.

end of the cathode but leaves the target visible through the glass envelope. The inner surface of the target is coated with fluorescent material which emits a green glow when struck by electrons from the cathode. Electrons are drawn from

cathode to target because the target is directly connected to B-plus potential and is highly positive with respect to the cathode. At one side of the cathode is a narrow deflector attached to the triode plate.

Electrons flowing from cathode to one side of the target must pass the deflector. If the deflector is at the same positive potential as the target there is little interference with the electron flow. Then electrons reach all of the target and the entire ring-shaped surface becomes uniformly luminous. If the deflector is less positive than the target, which means relatively negative, negative electrons are prevented from reaching the portion of the target surface which is in the "electron shadow" of the deflector. Then the shadowed area remains dark. The less positive the deflector, or the more negative with respect to the target, the greater will be the deflecting action and the larger will become the dark shadow on the target.

Deflector potential and shadow area are controlled by voltage between the triode grid and cathode, in the following

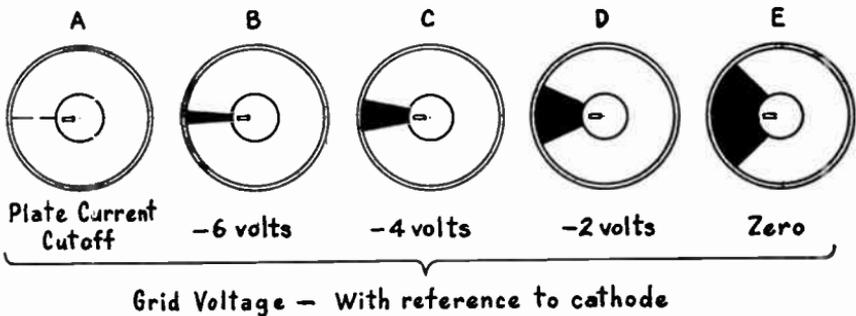


Fig. 2.—Changes of triode grid voltage alter the width of shadow area on the target.

manner. With grid voltage so negative as to cause plate current cutoff, there will be no plate current in the dropping resistor, no voltage across this resistor, and the plate with attached deflector will be at the same positive potential as the target. Then there will be no deflection of electrons and the entire target will be luminous, as at A of Fig. 2.

As triode grid voltage becomes less negative than for cut-

off there will be flow of plate current and some voltage across the dropping resistor. This makes the plate and attached deflector negative with respect to the target. Electron deflection will cause a narrow shadow on the target, as at *B*. With the grid still less negative there will be increase of plate current and of voltage across the dropping resistor. The plate and deflector become more negative with respect to the target, and the shadow widens as at *C* and *D* of Fig. 2.

When grid voltage becomes zero with respect to the cathode there will be maximum plate current and maximum voltage drop across the resistor. Plate and deflector potential will become maximum negative with respect to the target. As at *E*, the shadow will become of greatest width allowed by design of the tube, this width usually being from 90 to 100 degrees of the target circle.

The grid of the electron-ray tube is connected to a smooth direct voltage through a low-pass resistor-capacitor filter as

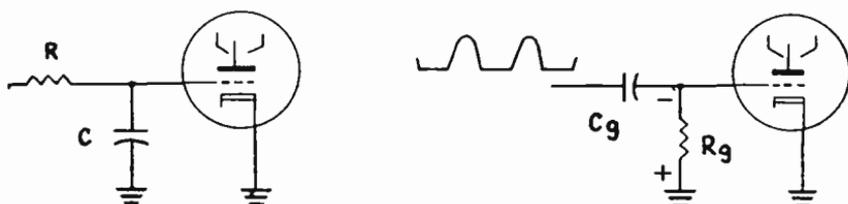


Fig. 3.—Connections for the triode grid of an electron-ray tube.

at the left in Fig. 3. If the d-c voltage is pulsating, as at the right, connection is through what amounts to a grid-leak bias system. Capacitor voltage becomes nearly equal to peak potential of the pulses. Negative capacitor voltage is held on the grid by slow leakage of the charge through the resistor to ground.

**TUBE, GATED BEAM.** — The gated beam tube has five active elements, but does not act like a pentode. The elements are, in order, a cathode, a control grid, a screen grid, a quadrature grid, and a plate. The most notable characteristic of this tube is the manner in which plate current is varied by changes of control grid voltage.

When the control grid is three to four volts negative with reference to the cathode there is plate current cutoff. As the control grid is made less negative there comes a point at which plate current commences to increase rapidly. Then, with a change of only about one volt on the control grid, plate current rises to its maximum value. Varying the control grid voltage back and forth through this range of approximately one volt will cause plate current to vary between zero and maximum. Even though the control grid is made positive by as much as 20 to 30 volts, plate current increases little if any above the value reached with the control grid still slightly negative.

Maximum plate current is determined chiefly by plate voltage or by B-voltage applied to the plate load resistor. The maximum increases almost directly with this B-voltage so long as screen voltage is held constant. The effect of screen voltage on the gated beam tube is much like that of plate voltage on a pentode. That is, plate current increases with screen voltage up to a certain value, after which further increase of screen voltage causes hardly any additional increase of plate current.

Maximum plate current is determined also by voltage on the quadrature grid. With this grid a few volts negative there is plate current cutoff. There is gradual increase of maximum plate current as quadrature grid voltage changes from negative through zero and to positive.

The gated beam tube was designed originally for use as a combined demodulator and limiter for f-m sound, and is thus employed in both sound and television receivers. A typical circuit is shown by Fig. 1. There is limiting because plate current cannot exceed a predetermined maximum on positive alternations of input grid voltage. For demodulator action, pulses of electron flow pass through the screen, acting as an accelerator, and reach the quadrature grid. Connected to the quadrature grid is an inductor-capacitor circuit tuned to the f-m center frequency. Resonant voltage in this tuned circuit lags signal voltage on the control grid by about 90 de-

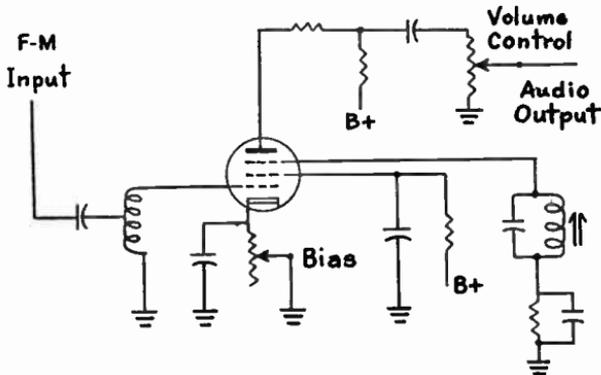


Fig. 1.—The gated beam tube in an f-m sound demodulator circuit.

gress. The combined effect of this lagging voltage and frequency-modulated voltage on the control grid vary the time duration of current pulses to the plate. Resulting variations of average plate current are proportional to frequency deviation of the f-m signal. Consequently, plate current varies at audio frequency.

The gated beam tube is used also as a sync separator in circuits such as that of Fig. 2. Each positive sync pulse ap-

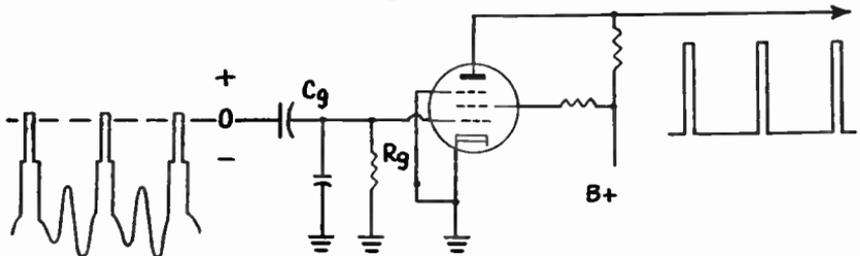


Fig. 2.—Action of the gated beam tube when used as a sync separator.

plied to the control grid causes plate current to reach its maximum. Portions of the video signal more negative than pedestal voltage keep the plate current cut off. The cathode and quadrature grid are grounded. There is grid-leak bias by means of capacitor  $C_g$  and resistor  $R_g$ . Noise voltages whose positive amplitude exceeds that of sync pulses are cut down, since plate current cannot exceed a certain maximum no matter how positive the control grid may become.

Other applications of the gated beam tube are in square wave generators and pattern generators. If a sine-wave voltage from any type of oscillator is applied to the control grid of the gated beam tube, and grid bias is adjusted to suitable negative value, plate output from the gated beam tube will be a sine wave clipped or flattened at top and bottom, as at *A* of Fig. 3. This is an approximate square wave. A greater

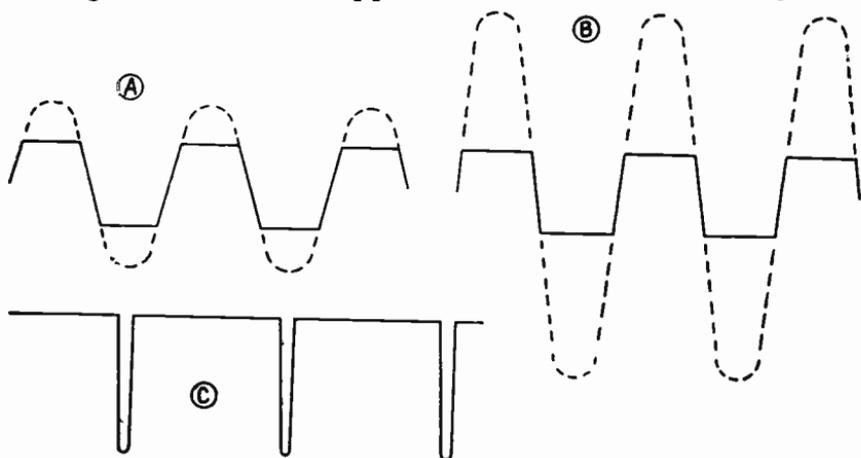


Fig. 3.—With sine-wave voltage applied to a gated beam tube, the output may be made square waves, pulses, or other forms.

amplitude of sine-wave input voltage, as at *B* makes the output more nearly approach a true square wave.

Other adjustments of control grid bias change the gated beam output to narrow negative peaks, as at *C*. There is cut-off of all positive alternations and part of the negative alternations. With the gated beam tube in a pattern generator these narrow voltage peaks produce narrow lines on the screen of a picture tube.

**TUBES, HIGH-FREQUENCY.** — Tubes for use at very-high and ultra-high frequencies as amplifiers, oscillators, and mixers should have small internal capacitances, small inductance in their internal leads, small dielectric losses, and, so far as is compatible with other requirements, should have high transconductance. Also, at ultra-high frequencies, the matter of transit time for electrons traveling between the elements may be of importance.

Standard types of tubes which are entirely satisfactory at the lower radio frequencies may cause losses instead of gains when used as amplifiers, or may fail to oscillate when required to do so, if these tubes are used at television frequencies. Most low-frequency tubes become unsatisfactory at frequencies above 50 to 60 megacycles. When used in amplifier circuits at higher frequencies these tubes may oscillate because their internal capacitances resonate with the inductance of even short straight connecting wires. Tube noise tends to become excessive. The high-frequency limit for satisfactory operation as oscillators usually is lower than the limit for amplification.

R-f amplifiers in circuits where signal input is to the grid usually are pentodes rather than triodes because an ordinary triode circuit tends to allow oscillation when grid and plate are tuned to the same or nearly the same frequency. Triodes may be used as grounded grid high-frequency amplifiers as explained later in this article.

High transconductance is desirable or necessary to compensate for the low effective impedances of plate circuits at the higher frequencies. This low impedance is due chiefly to input and output capacitances of the tubes and to distributed and stray capacitances of inductors and circuit connections.

Transconductance is increased in tubes having closely spaced elements and by having adjacent turns of the grid close together. These structural practices tend to increase the internal capacitances at the same time as the transconductance. A figure of merit for a tube intended for high-frequency operation sometimes is taken as the quotient of dividing the transconductance in micromhos by the total of input and

output capacitances in mmf. Such figures of merit will range between 500 and 1,000 for tubes in general use as television r-f amplifiers, video i-f amplifiers, and video amplifiers.

Triodes, as a class, have less transconductance than pentodes, but the triodes also have less total input and output capacitance. The two factors tend to balance, so that figures of merit for triodes are as high as for pentodes. The grid-plate capacitance of pentodes is, of course, far less than for triodes.

Many miniature tubes are well suited for high-frequency operation. This is due in part to absence of plastics employed in bases of other tubes and to the use of very short internal leads and small base pins. Several of the lock-in types are well suited for high-frequency operation. Lock-in tubes have only glass and metal in their construction, have small base pins and relatively short internal leads. A number of miniature and lock-in types give satisfactory performance at frequencies from 400 to 600 megacycles.

Some of the tubes especially designed for high-frequency operation have two or even three pins and internal leads for a single element. Such construction allows circuit connections which reduce the effects of lead inductance, because several inductances (leads) connected in parallel with one another have less effective inductance than any single one of the group. When external circuit connections are made to base pins which are at opposite ends of an internal lead there is division of internal tube capacitance between the connected circuits. It is the shunting effect of internal capacitances that is largely responsible for low impedances in plate and grid circuits.

When there are two pins and two internal leads for the cathode of a tube it becomes possible to maintain separation between input and output circuits right through to the cathode itself. All returns for the grid circuit should be made to one cathode pin. Plate, screen, and heater returns are made to the other pin. Then there is very little cathode lead inductance common to the output and input circuits, there is lessened coupling through common inductive reactance, and

less degeneration. This, of course, allows realizing a considerably greater gain.

**Grounded Grid Amplifiers.** — Fig. 133-1 shows typical circuits for grounded grid high-frequency amplifiers. The tube is a triode, usually of special design and construction for this kind of operation. The grounded grid acts as an r-f shield between plate and cathode to permit amplification without the feedback which could cause oscillation at very-high and ultra-high frequencies.

The high side of the signal input circuit is connected to the cathode of the tube, with the low side connected through ground to the grid. The output is connected from plate to

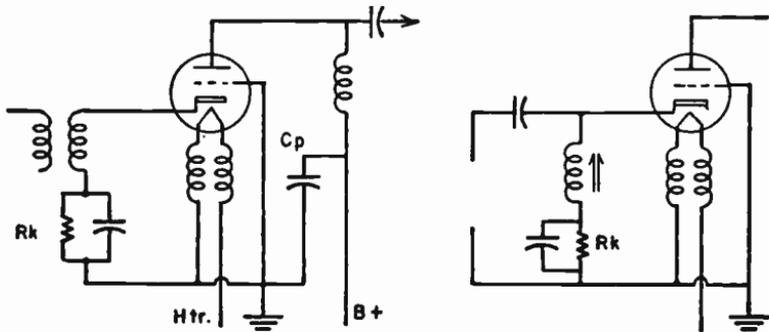


Fig. 133-1.—Circuits for grounded grid r-f amplifiers.

ground and grid through capacitor  $C_p$ . There is a certain amount of negative feedback from plate to grid, which adds to the effect of the grounded grid in preventing oscillation. Resistor  $R_k$ , which always is bypassed, provides cathode bias for the tube. R-f chokes in both heater leads maintain the r-f potential of the heater the same as that of the cathode and thus prevent adding the effect of cathode-heater capacitance to the input circuit.

**Transit Time.** — Transit time is the length of time required for an electron to travel from cathode to plate under the accelerating force of plate voltage. Average electron velocity in inches per microsecond is equal approximately to the prod-

uct of 11.7 and the square root of the plate voltage, assuming the electron to start from rest at the cathode. For example, with 130 plate volts the average velocity is about 133 inches per microsecond.

To illustrate the effects of transit time we may assume a cathode to plate distance of  $\frac{1}{4}$  inch. Transit time then will be a little less than 0.002 microsecond. Were signal frequency to be 200 megacycles per second, each complete cycle would require a time of 0.005 microsecond. With the assumed plate voltage and cathode-plate separation an electron would travel from cathode to plate during about 135 degrees of a signal cycle.

Because of the time required for the electron to travel from cathode to plate there will be a very considerable difference between the point in a signal cycle at which the electron leaves the cathode and the point in the same cycle where the electron reaches the plate. Consequently, signal variations of plate current will not be in phase with signal variations of grid voltage. When current and voltage are out of phase there is a lowered power factor and loss or waste of energy. This reduces the output power otherwise available from the tube. There is also a reduction of effective transconductance.

Transit time is lessened by closer spacing between cathode and plate, also by use of higher plate voltage. Both these measures, however, increase the heating while reducing the ability of the elements to lose heat by conduction and radiation.

**TUBES, PICTURE.** — The earliest picture tubes employed the now obsolete method of electrostatic deflection, which has been replaced by magnetic deflection. Round tubes of nominal diameters from 10 to 19 inches formerly were popular, but all recent types have rectangular face plates and screens. Both round and rectangular tubes may be of all-glass construction or may have metal cones or flares. Electron guns in earlier tubes were designed for double-magnet ion traps, but later types employ single-magnet traps. The nominal size of a rectangular tube is the approximate external diagonal dimension in inches. Popular sizes have included 14, 17, 21, 24, 27, and 30 inch nominal diagonal dimensions. Useful screen sizes in all types of tubes are one to two inches less than corresponding outside dimensions.

**Picture Tube Construction.** — Electrons which form the beam in the picture tube are emitted, controlled in density, and accelerated in a part of the tube called the electron gun. The essential parts of the electron gun in a typical tube are shown by Fig. 134-1. An oxide coating on the forward end of the cathode sleeve is heated dull red by action of the heater

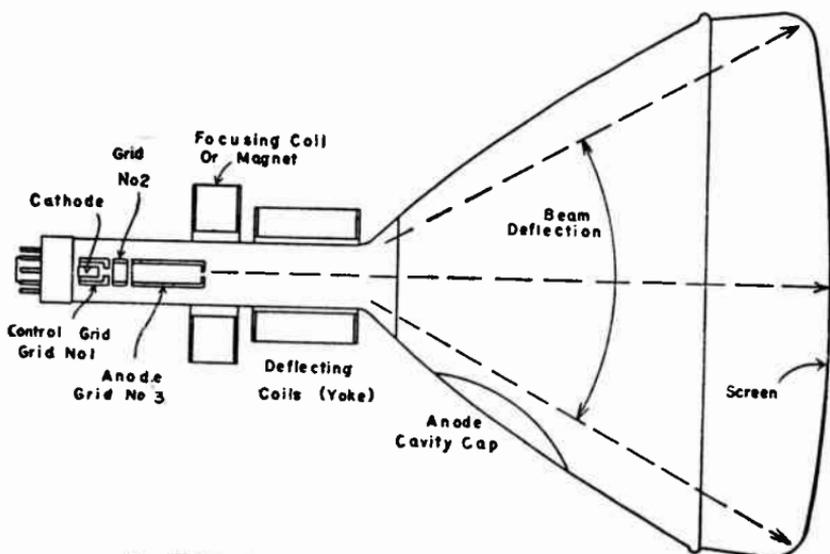


Fig. 134-1.—Location of the electron gun in a picture tube.

which is enclosed within the sleeve. Electrons emitted from the cathode are drawn into a narrow stream at the opening through the control grid or grid number 1.

The electron stream then spreads to some extent as it is drawn on through grid number 2 which is operated at a fixed potential, usually about 250 to 300 volts positive with reference to the cathode. The electrons pass next through the anode, called also the accelerating electrode or grid number 3, where they are accelerated to maximum required velocity on their way to the screen.

The inside of the flared portion of tubes of all-glass construction is covered with a conductive coating of finely divided graphite which is electrically connected to grid number 3. This internal coating is part of the anode of the tube. Sometimes the internal coating and grid number 3, considered as a single electrical element, are called the anode. The internal coating collects electrons which leave the screen as a result of secondary emission at the point where primary electrons from the gun strike the screen.

Grid number 2, due to its constant potential, insures that control of electron density of the beam by control grid voltages will be practically unaffected by different potentials applied to the anode in different receivers.

The electron stream still is spreading to some extent as it reaches the space in which is a magnetic field produced by the focusing coil or focusing magnet. This field draws the electrons together to form a narrow beam that strikes the screen of the tube on a spot of small diameter. This is the action of focusing, which is explained in the article on *Focusing and Focusing Controls*.

Approximately half the types of all-glass picture tubes which are in general use have an external conductive coating over the flared portion, in addition to the internal coating. The external conductive coating is connected to ground through a contact spring attached to the chassis, and acts as a shield against external electric fields.

The internal and external conductive coatings act as the plates of a capacitor, with the glass of the tube envelope as dielectric. The capacitance of the coatings, which may be anything between 500 and 3,000 mmf, is used as a high-voltage filter capacitor with connections shown at the left in Fig. 134-2. When the picture tube does not have an external coating a separate filter capacitor of 500 to 2,000 mmf is connected between the high-voltage lead and ground, as at the right.

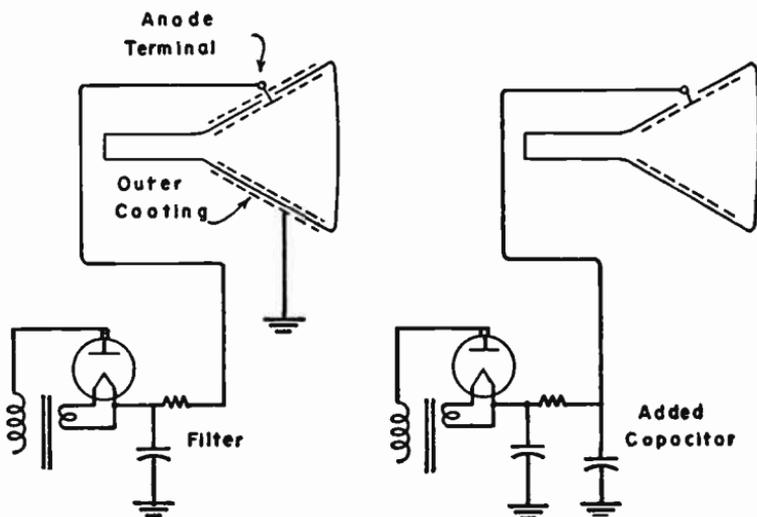


Fig. 134-2.—Using picture tube coatings as a high-voltage filter capacitor (left), and connections for an added capacitor (right).

The metallic conical portion of the envelope of metal picture tubes is an electrical part of the anode. Consequently, this cone is at a very high potential and is dangerous to touch while the receiver is turned on. The high-voltage lead for the anode of metal-cone tubes is fitted with a clip connector which attaches to the front lip of the cone. On most all-glass tubes the anode terminal is a cap recessed in a cavity on the side of the flare which ordinarily is placed at or near the top when the tube is mounted. This recess is indicated in Fig. 134-1. Other all-glass tubes have for their anode terminal a ball which is not recessed.

**Tube Voltages.** — All magnetic deflection picture tubes in general use are designed for 6.3 volts and 0.6 ampere in their heaters. Negative grid potential for complete cutoff of illumination on the screen of various types of picture tubes ranges from 35 to 77 volts. The higher the anode voltage and the higher the voltage on grid number 2 the more negative must be the control grid voltage to attain cutoff with any given tube.

Fig. 134-3 shows typical relations between control grid voltage and brightness of the screen. One curve is drawn for 250 volts and the other for 400 volts on grid number 2. Curves showing relations between control grid voltage and anode current would be of the same general form. This would be expected, inasmuch as illumination is roughly proportional to electron density in the beam reaching the screen.

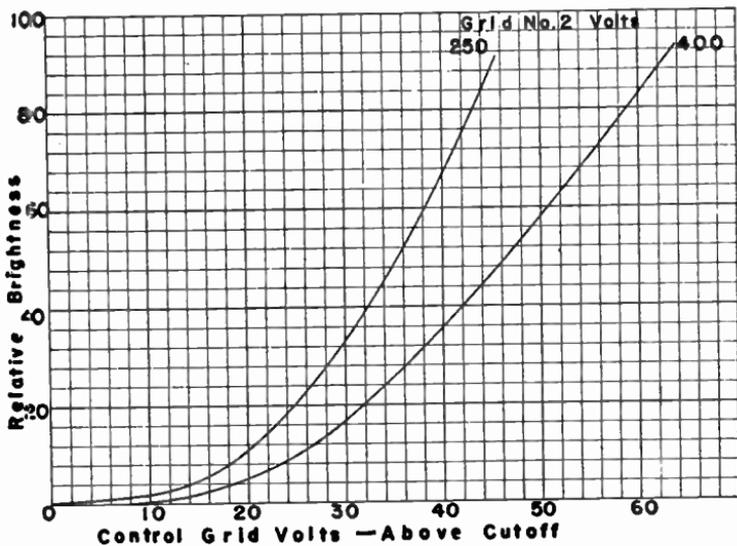


Fig. 134-3.—Relations between control grid voltage and brightness at the screen of a picture tube.

Note that voltage shown on the graph is voltage above cutoff, it is not grid-cathode voltage. As an example, were cutoff potential to be 60 volts negative, the graph voltages related to the curves would be those subtracted from 60 volts,

or would be the number of volts by which the control grid is made less negative than for cutoff in causing the degrees of brightness indicated by the curves.

All curves showing the effect of control grid voltage on brightness and on anode current are quite similar to curves showing relations between control grid voltage and plate current in triode amplifier tubes. These mutual characteristics or transfer characteristics of picture tubes may be used to illustrate relations between picture signal voltages and picture brightness just as they are used to illustrate relations between sound signal voltages and sound output of amplifiers.

Maximum permissible anode potentials for 10-inch, 12-inch, and 14-inch picture tubes are between 10,000 and 12,000 volts, with typical operating voltages ranging from 7,000 to 11,000. For 15-inch and larger tubes the maximum anode voltages are between 14,000 and 20,000 volts, depending on the type of tube, while typical operating voltages are between 9,000 and 15,000. Higher voltage on the anode tends to give better definition and greater brightness in the pictures.

**Deflection of Beam.** — Factors affecting the distance the electron beam is deflected either way from the center of the screen are illustrated by Fig. 134-4. Deflection distance is directly proportional to strength or flux density in the magnetic field of the deflecting coil. Flux density, in turn, is approximately proportional to deflecting current in milliamperes, but depends also on permeability of any iron core used in the coil. Permeability of an iron core varies with changes of current and of flux. Deflection distance increases with increase of deflecting current in the coil, but not proportionately when the coil has an iron core.

Deflection distance on the screen increases directly with length of the deflecting coil in line with the tube axis, or rather with length of the magnetic field when the field is uniform. The deflection is increased also, and increases directly, with increase of distance from the center of the deflecting coil to the center of the screen. This is because a longer electron beam deflected over a certain angle moves farther

at the screen end than does a shorter beam deflected over the same angle.

Deflection distance is inversely proportional to the square root of anode voltage. The greater the anode voltage the less will be the deflection distance with all other factors remaining unchanged, but decrease of deflection will be only in the ratio of the square roots of the anode voltages. Anode voltage or electron accelerating voltage tends to pull the beam along a

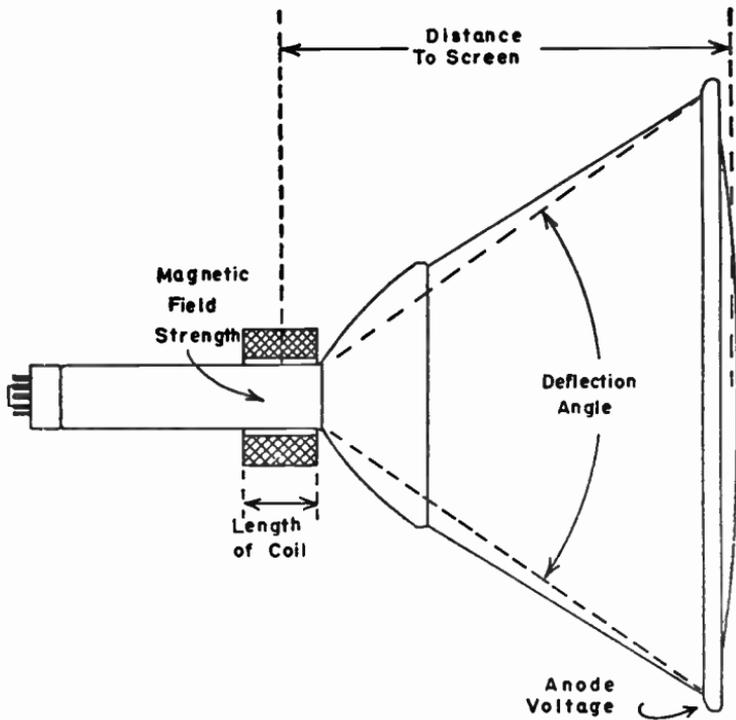


Fig. 134-4.—Factors which affect distance of deflection in a magnetic picture tube.

straight line, and opposes deflection or bending. It turns out that distance of deflection is increased by more deflecting current in the coil, by less anode voltage, or by both these changes. Deflection is decreased by less deflecting current, more anode voltage, or both.

The tube illustrated by Fig. 134-4 is designed for a maxi-

imum deflection angle of 70 degrees, while in Fig 134-1 the angle is 56 degrees. Rectangular tubes in sizes to and including 21 inches are designed for diagonal deflection angles of 70 degrees and horizontal angles of 65 or 66 degrees. Larger rectangular tubes are designed for diagonal deflection of 90 degrees and horizontal deflection of 85 degrees.

Although there is no exact relation between tube size and second anode voltage, the larger the tube the greater is the anode voltage commonly employed. The combination of higher anode voltage and greater deflection angle requires stronger magnetic fields and greater swings of field strength than needed for smaller tubes. Consequently, tubes having deflection angles of 70 or 90 degrees require deflecting yokes and sweep output transformers designed and operated to meet the requirements of the larger tubes.

**Tube Types and Base Connections.**— Type designations consist of numerals and letters such as 24CP4-A. The first number, 24 in the example, is the greatest outside dimension at the face end, measured to the nearest whole inch. The following letter indicates the order in which a particular design was registered with the Radio-Electronic-Television Manufacturers Association. *A* indicates the first tube of a general kind, *B* the second, and so on.

The next letter and numeral, *P4* in the example, show the kind of phosphor in the screen. Phosphor number 4, indicated by *P4*, produces a white luminance and is used in all picture tubes for black-and-white or monochrome receivers. A three-color phosphor is number 22, which would be indicated in the type number 19VP22 for a color picture tube.

A letter following the phosphor number indicates a modification of a tube whose original type designation does not include the final letter. For example, the 21WP4-A has an aluminized screen, while the 21WP4 is an otherwise identical tube except for having a plain screen. A given final letter does not refer to any certain modification, it is merely assigned in alphabetical order as variations are developed.

Fig. 134-5 shows base pin positions and connections for

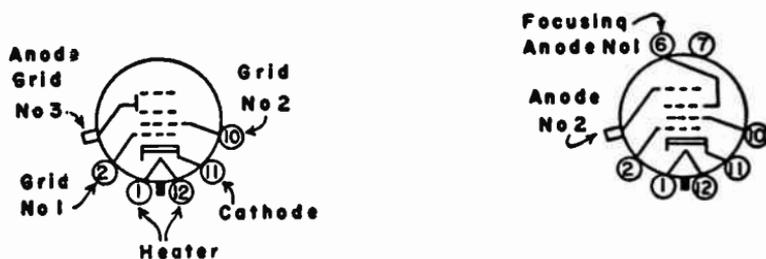


Fig. 134-5.—Base pin positions and element connections for magnetic deflection picture tubes.

picture tubes, other than three-gun color tubes, which employ magnetic deflection. Pin positions are as seen from the outside of the tube base. Tubes designed for magnetic focusing have basing arrangements at the left. Tubes designed for electrostatic focusing have an additional pin in position 6, as shown at the right. Only in a few early types is there a pin in position 7. Pins in positions 6 and 7, with no internal connections, have been used to help locate the socket on some tubes.

Electrostatic focusing may be employed for both rectangular and round picture tubes of all sizes, and with either glass or metal envelopes. No particular type numbers or letters indicate whether a tube is designed for electrostatic or for magnetic focusing, although the registration letter which immediately follows the size number will be different for the two methods of focusing. For instance, a 24DP4 tube employs electrostatic focusing, while a 24VP4 is designed for magnetic focusing.

Whether a tube is of all-glass construction or has a metal flare or cone is not shown by any certain letters or numerals in the type designation. Here again, however, the registration letters always are different for the two constructions. A 27MP4 tube is a rectangular type with a metal shell, while a 27LP4 is a rectangular all-glass tube.

Sockets for the magnetic deflection tubes are specified as the duodecal type, in which there are openings and lug connections for all 12 pin positions. Picture tube sockets are supported only by being pressed onto the base pins of the tube, with circuit connections made through flexible wires leading to the socket lugs.

**Screens for Picture Tubes.** — In the screen which covers the inside of the exposed face of the picture tube are solid materials which become luminous when their particles are struck by the electron beam. These materials are called phosphors. Different kinds of phosphorescent substances emit light of various colors when excited by the electron beam. Mixtures of these substances will produce intermediate hues and will produce an approximation of white.

The phosphors are identified by numbers. Phosphor number 4, used in all monochrome picture tubes, gives the effect of

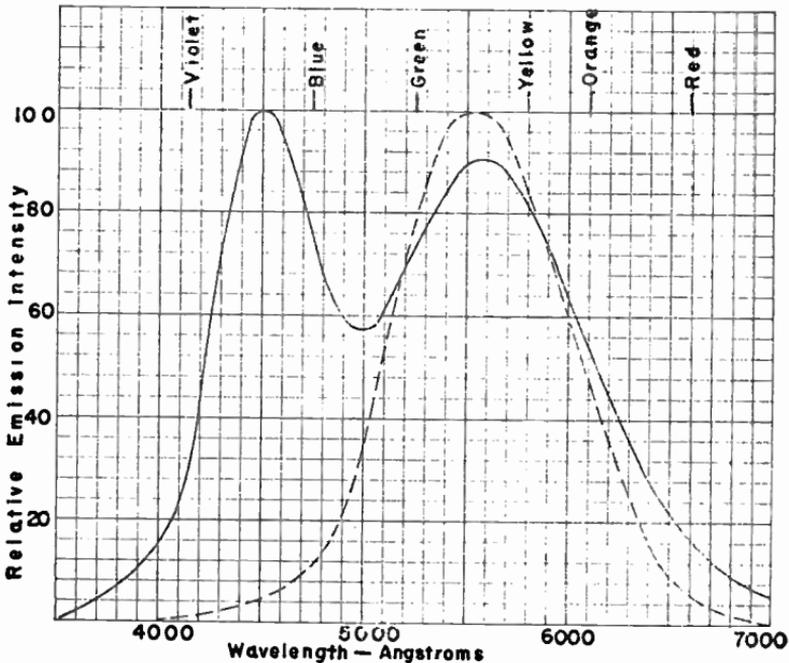


Fig. 134-6.—Luminous radiation at various wavelengths for phosphor number 4 (full-line curve) and luminous sensitivity of the average human eye (broken-line curve).

white light by delivering radiations from blue through green and yellow. The full-line curve of Fig. 134-6 shows relative intensities of emission at various color wavelengths in the visible spectrum for phosphor number 4.

Wavelengths are in Angstroms or Angstrom units. One Angstrom is a wavelength of one hundred-millionth of a centimeter or about one 250-millionth of an inch. A wavelength of 5,000 Angstroms, at the approximate center of the visible spectrum, corresponds to a frequency of 600 million megacycles per second. In the emission from phosphor number 4 there are peaks between violet and blue, and between green and yellow, with lesser intensities between blue and green, and at the violet and red ends of the curve. These emissions combine to produce the light seen on the picture tube screen.

Among other phosphors in general use, number 1 is found in the screens of most oscilloscope tubes. The color of its trace is green. Phosphors 5 and 11 produce a blue trace suitable for photography of oscilloscope traces. Numbers 7, 12, and 14 produce combinations of blue, yellow, and orange traces used for radar observation. Number 15, with a blue-green trace, is used for flying spot scanning in film reproduction.

The broken-line curve of Fig. 134-6 shows average human eye response to the various color wavelengths when radiation intensity is uniform for all wavelengths. Maximum eye sensitivity, for normal vision, is to a greenish-yellow hue of about 5,550 Angstroms in wavelength.

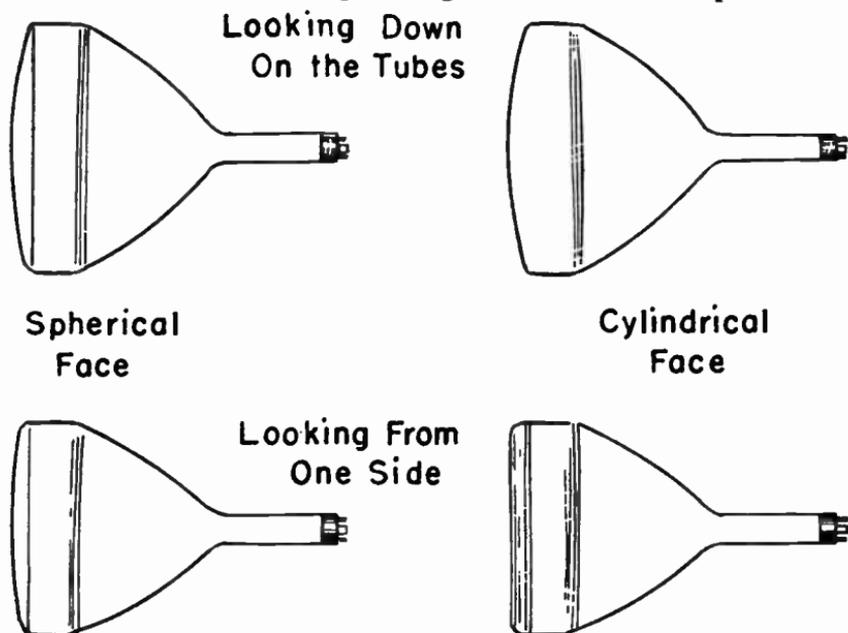
In the light or the luminescence appearing on the picture tube screen there are two effects. One is called fluorescence, the light from which ceases instantly when the electron beam moves on. The other is phosphorescence, which causes emission of light from the phosphor particles for an appreciable time after the beam has left a spot on the screen. Phosphorescence accounts for what is called persistence of the particular phosphor considered.

Persistence of phosphor number 4 is long enough to help prevent the appearance of flicker, but is short enough that

luminescence remaining at the end of a field period is only a few per cent of the initial value.

**Metal-backed Screens.** — In a picture tube having an aluminized or metal-backed screen there is deposited over the inner surface of the phosphor coating a very thin layer of aluminum. This layer is so thin that beam electrons penetrate it and excite the phosphor as usual. The principal advantage is increased brightness for any given drive voltage and beam intensity. With two tubes identical except for an aluminized screen in one of them, with both operated at the same element voltages, the one with the metallized screen will show 25 to 40 per cent greater brightness.

The thin layer of aluminum acts somewhat like a mirror in reflecting outward through the face much of the light which otherwise would go back inside the tube. This improves contrast. The metal backing has good electrical conductivity and allows electrons from the beam to more easily flow off the screen. The backing also gives considerable protection



**Fig. 134-7.**—A spherical face plate curves in all directions, but a cylindrical face plate curves only from side to side.

should ions reach the screen. The earliest tubes having aluminized screens were used without ion trap magnets, although the electron guns were straight and concentric.

**Face Plates and Reflections.**— Most picture tubes have spherical face plates, meaning that the outside of the glass face is curved in all directions like the surface of a sphere or ball of large diameter or radius. This construction is shown at the left in Fig. 134-7. In a typical 27-inch tube, face plate radius may be about 40 inches. When objects anywhere in front of the tube are illuminated or self-luminous, they may be reflected in the curved face plate surface.

Some tubes have cylindrical face plates, as at the right in Fig. 134-7. The outside of the face is curved from left to right like a section of a cylinder standing vertically, but there is no curvature from top to bottom. There may be reflections from objects at either side, but there is less chance of reflections from above or below the viewing level. Still other tubes have frosted face plates which change sharply defined (specular) reflections to diffused images which do not greatly interfere with pictures.

Another method of lessening reflections, while at the same time increasing the range of tones on the contrast, consists of using phosphor and screen materials which are of rather dark gray when not luminous, instead of being nearly white. Less external light is reflected back outward from the gray phosphor than from a white one.

Glass used for face plates commonly is tinted to act as a neutral density filter, often being called gray filter glass or identified by trade names such as *Filterglass* and *Teleglass*. Only 65 to 75 per cent of light reaching this glass passes through it. Light from pictures passes outward through the face plate only once, but external light reflected back from the phosphor has to pass through the glass twice, and suffers twice as much absorption.

**Masks.**— The picture image originally formed in camera tubes is of standard aspect ratio, four units of width to three units of height, as shown by broken lines in Fig. 134-8. Full

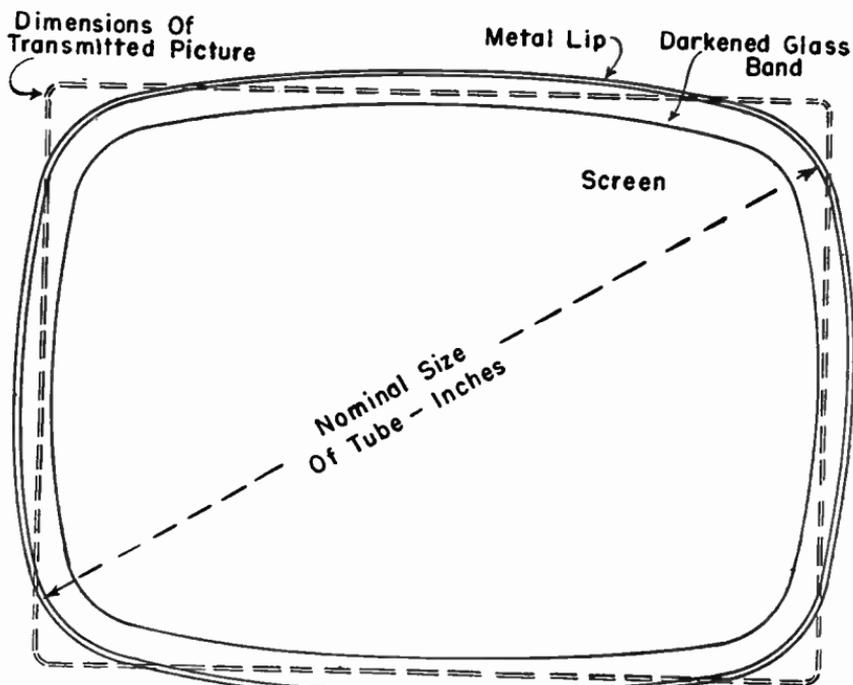


Fig. 134-8.—Useful screen area of a picture tube is of nearly the same proportions as the standard aspect ratio.

lines show typical proportions of the outside, also of the useful screen area of rectangular picture tubes. Since any transmitted picture is of the same proportions as the original image, while the reproduction must be of width and height to fill the screen or the mask opening, the extreme corners of pictures do not appear. Useful width and height of picture tube screens always are very nearly in the ratio of four units to three units respectively.

**Handling of Picture Tubes.**— All types of picture tubes must be handled with great care to avoid the possibility of breakage. On the outer surface of even a 17-inch tube, the total atmospheric pressure exceeds internal vacuum pressure by more than three and one-half tons. A break allows an implosion, with particles of glass driven inward and then rebounding outward with great force, and possibility of serious injury unless suitable precautions are observed.

A picture tube should not be handled by anyone not wearing shatter-proof goggles and heavy gloves. Everyone in the near vicinity should be similarly protected, or should move to a safe distance. A tube should not be held near the body. The tube should be held near the large end to support its weight, with only enough support on the neck to guide or steady the tube. Especial care should be used to place no undue stresses on the neck nor at any place where the glass is rather sharply curved.

While a tube is removed from a receiver it should be rested face down on some soft surface, with precautions against tipping over. The safest procedure is to place the tube in a regular tube carton. A picture tube must not be forced into its mountings on the receiver. Coils, magnets, sockets, and any other parts which go over the tube neck must be a free fit. A picture tube must not be struck by nor allowed to strike against any hard objects. The glass surfaces must not be scratched. It is advisable to keep all metal tools away from the tube, and to keep the tube away from metal objects so far as is possible.

**Removal and Replacement of Tubes.** — In the majority of receivers the picture tube is mounted on and supported by the chassis as shown in a general way by Fig. 134-9. The chassis with the tube in place may be removed from the cabinet as a unit. First it is necessary to take off control knobs whose shafts protrude through small openings in the front of the cabinet, and usually it is necessary to disconnect the leads for the speaker. Of course, any screws, bolts, or clamps holding the chassis in the cabinet must be removed or loosened. The tube may be taken off the chassis as follows.

1. Take the socket off the base of the tube, leaving the socket attached to the flexible leads that run to the chassis.

2. If there is an ion trap magnet slide it back off the base end of the tube, first loosening the clamping screws if the magnet is held in this manner.

3. Remove the high-voltage anode lead from the cap or ball on the flare of an all-glass tube or from the front rim of a

metal-cone tube. With the lead removed, and before touching its clip or cap and before touching the anode terminal of the tube, discharge the high-voltage filter capacitors by touching the clip or cap to chassis metal or a B-minus point. Then discharge any capacitance of the tube itself by touching the bared ends of a piece of insulated wire to the chassis metal or B-minus and to the cap, ball, or rim which is the anode terminal of the tube. This discharging of the tube is especially important with all-glass types which have an external conductive coating, but the same procedure should be followed for all kinds of picture tubes. It must be kept in mind that both the cone and the glass face of metal-cone picture tubes operate at anode voltage, and these parts should be treated accordingly.

4. For ease of later replacement of the socket and anode leads it is advisable to note the position of the tube with reference to its angular position around the axis. The reference point may be the anode terminal of all-glass tubes, the base key, or certain parts of the electron gun.

5. In order that the tube neck may slide forward through the yoke and focus coil or magnet it may be necessary to loosen their mountings or adjustments.

6. Loosen or remove the strap and any other fastenings that hold the large end of the picture tube on the chassis. Now the tube may be carefully moved forward out of the yoke and focus coil or magnet.

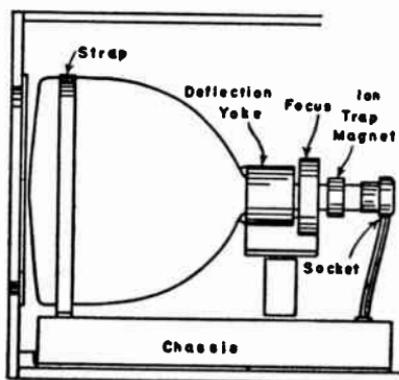


Fig. 134-9.—Picture tube mounted on chassis.

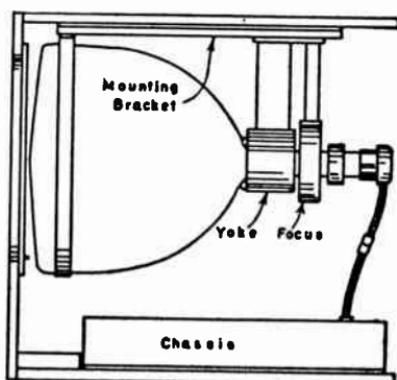


Fig. 134-10.—Picture tube mounted in top of cabinet.

Fig. 134-10 shows in a simplified manner a method of mounting the picture tube and its accessory parts in the top of the cabinet, independently of the chassis except for any necessary connecting leads. These leads must be disconnected before removal of either the tube or the chassis. Always there will be leads to the tube socket and the deflecting yoke, and leads will run also to electromagnetic focusing coils mounted on the tube. With construction of the type illustrated a single bracket or frame ordinarily carries the picture tube, the yoke and focusing magnet. The bracket and all these parts may be removed from the cabinet as a unit. Coils and other parts then may be removed from the tube neck, and the tube taken off the bracket as described earlier.

A bracket or separate frame mounting for the picture tube and accessory parts may be mounted on one side of the cabinet instead of from the top. In some sets the support for the picture tube mounts in the cabinet bottom alongside the chassis, and is independent of the chassis except for connecting leads.

Fig. 134-11 illustrates in a general way one method of mounting the picture tube on the front of the cabinet. The large end of the tube is carried by cushions or pads which are on the cabinet. The back of the flare or cone of the tube is supported by cushions or pads carried by the deflection yoke cover or by the yoke mounting.

When the picture tube is to be removed the first steps are to take off the socket and the ion trap magnet as earlier described. Then the focusing coil or magnet and the deflection yoke are loosened from their support or supports, their leads are disconnected if necessary, and these two parts are taken off the base end of the tube. While carefully supporting the tube with one hand, its fastenings to the front of the cabinet now are loosened or removed, whereupon the tube may be taken out from the rear of the cabinet.

Some receivers and their cabinets are designed to permit removal of the picture tube through the front of the cabinet, after opening a front panel inside of which are supports for

## TUBES, PICTURE

the large end of the tube. Such construction is illustrated in a simplified way by Fig. 134-12. To take out the tube the socket and ion trap are removed as usual, and the high-voltage anode lead is disconnected from the tube. The high-voltage filter capacitor and the tube capacitance are discharged as described earlier. Next the deflecting yoke and focus coil or magnet are loosened on their support and moved as far toward the tube base as they will go. The front panel now is opened or removed, while carefully supporting the picture tube, and the tube is taken out through the front of the cabinet as its neck slides through the yoke and focus coil or magnet.

Picture tube replacement calls for reversal of the steps carried out during removal. There are, however, a few special precautions to be observed. Make certain that the deflecting yoke and focusing coil or magnet are in such positions, or are sufficiently free, that the tube neck will pass freely through

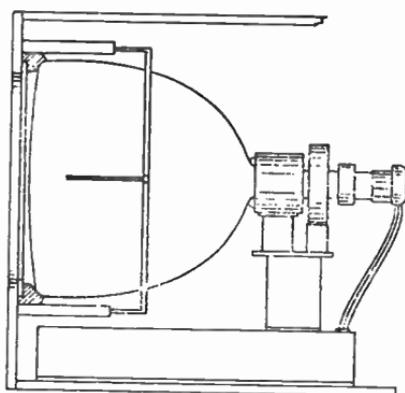


Fig. 134-11.—Picture tube mounted between cabinet front and a cushion on the yoke assembly.

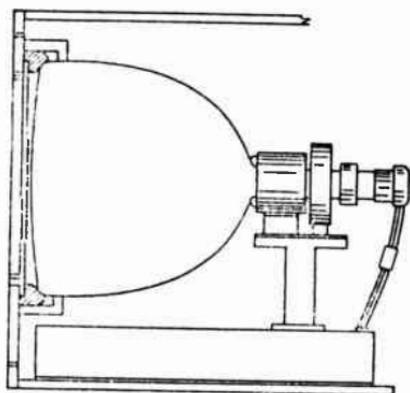


Fig. 134-12.—Picture tube removable from the front of the cabinet.

these parts. The tube neck should be centered in the openings through yoke and coil or magnet, or these parts should be adjusted to center around the neck.

The back end or small end of the tube flare or cone insulator must fit snugly against the cushion or pad carried by the deflecting yoke or its support. Either the yoke mounting or

the picture tube may be moved to accomplish this. If the tube has an outer conductive coating the grounding spring attached to chassis metal must make good contact with the coating. Since the front rim or lip of metal shell tubes is at anode potential it must be well insulated from all metal parts except the clip of the high-voltage cable.

**TUBES, REACTANCE.** — A reactance tube is a triode or pentode operated in such manner that it acts like an inductance or a capacitance which may be varied in value by changes of control grid voltage. Reactance tubes are used in automatic frequency controls for sweep oscillators and for r-f oscillators, for varying the frequency of oscillators in sweep generators, for frequency modulation, and otherwise where automatic variation of inductive or capacitive reactance is required.

With one method of connection, plate current in the reactance tube is made to lag voltage on its plate by approximately 90 degrees, and since alternating current in any inductor lags applied voltage, the tube then acts like inductance. With another method of connection, plate current is made to lead voltage on the plate, and since alternating current leads applied voltage in a capacitor, the tube acts like a capacitance.

Fig. 1 is a connection diagram for a reactance tube furnishing variable inductance for an oscillator, as in certain automatic frequency control systems. The oscillator is shown

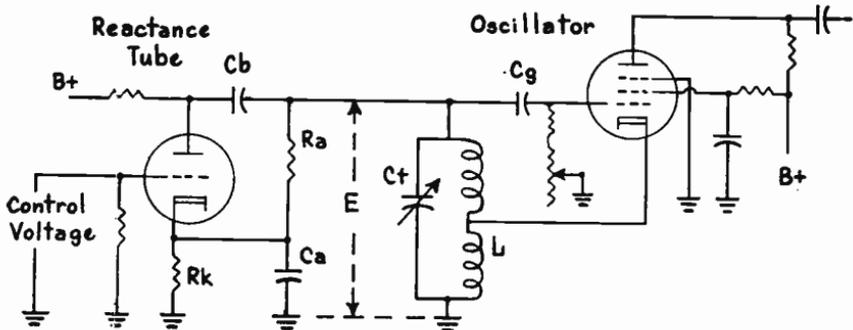


Fig. 1.—A reactance tube controlling oscillator frequency by means of a variable inductive effect.

as an electron coupled Hartley type, but might be of any other kind employing a tuning inductor in its resonant circuit or tank circuit. The tank circuit here consists of inductor  $L$ , capacitor  $C_a$ , and resistor  $R_a$ . Grid biasing capacitor  $C_g$  and the hold control are such as might be found with any oscillator, and take no part in reactance tube action. At  $C_b$  is a capacitor of small reactance which keeps B-plus voltage from the oscillator tanks circuit but allows alternating plate current from the reactance tube to flow in that circuit. It should be noted that the reactance tube is in parallel with the oscillator tank inductor.

Resistor  $R_a$ , usually of only about 10 ohms, is part of the oscillator tank circuit and also is in series with the reactance tube cathode. Any alternating voltage produced in  $R_a$ , as part of the oscillator tank circuit, appears also in the grid-cathode circuit of the reactance tube.

Oscillator voltage, marked  $E$  on the diagram, is applied across  $C_a$  and  $R_a$  in series. As in any capacitance, alternating current in  $C_a$  leads applied voltage  $E$  by very nearly 90 degrees. Capacitive reactance at  $C_a$  is made so great in comparison with resistance at  $R_a$  that current in  $R_a$  is practically in phase with current in  $C_a$ , and leads voltage  $E$  by about 90 degrees. Because voltage across any resistor is in phase with its current, voltage across  $R_a$  leads voltage  $E$  by approximately 90 degrees. This leading voltage across  $R_a$  is applied to the cathode of the reactance tube.

A voltage applied to a cathode is equivalent, in effect on plate current, to a similar voltage 180 degrees out of phase applied to the grid. This shift of 180 degrees makes equivalent alternating grid voltage at the reactance tube lead voltage  $E$  by about 90 degrees. Plate current in any tube is in phase with its grid voltage. This means that plate current in the reactance tube lags voltage  $E$  by about 90 degrees, because grid voltage has this relation to voltage  $E$ .

Reactance tube plate current, lagging voltage  $E$ , flows in the oscillator tank circuit. A current which lags applied voltage is characteristic of inductance, so adding the lagging plate-

current from the reactance tube has the same effect as adding inductance across the oscillator tank circuit. It must be kept in mind that the reactance tube is in parallel with inductor  $L$  of the oscillator tank circuit. As a consequence, inductive effect of the reactance tube is in parallel with inductance at  $L$ . When inductances are paralleled the combined inductance is decreased, so the reactance tube decreases effective inductance in the tank circuit. Less inductance in any tuned circuit raises the resonant frequency. Thus the effect of reactance tube plate current is to increase the oscillator frequency.

How much the oscillator frequency is increased depends on amplitude of reactance tube plate current. The greater the amplitude the greater is the parallel inductance effect, and the higher becomes oscillator frequency. Plate current amplitude may be varied by changing reactance tube grid voltage, as would be true of any tube.

In practice the grid of the reactance tube is negatively biased to a degree which prevents flow of grid current when grid voltage is varied to provide a variable inductive effect. If the grid then is made less negative there will be increase of plate current and consequent increase of oscillator frequency. Making the grid more negative reduces plate current amplitude, and lowers oscillator frequency.

By interchanging the electrical positions of capacitor  $C_a$

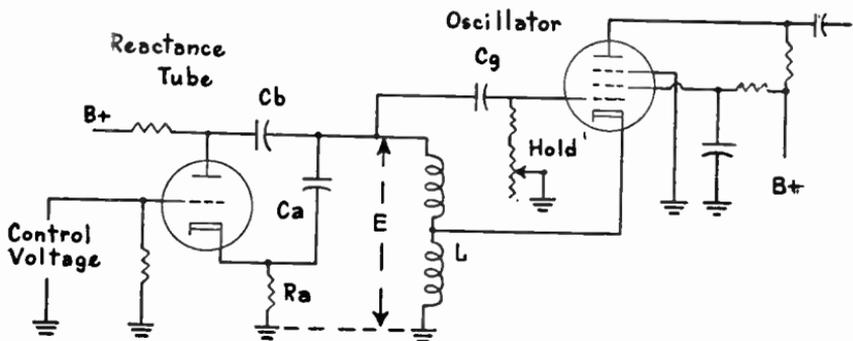


Fig. 2.—The reactance tube controls oscillator frequency by means of a variable capacitive effect.

and resistor  $R_a$ , as in Fig. 2, the reactance tube is made to act like a variable capacitance instead of an inductance. Resistance at  $R_a$  is large, possibly several hundred thousand ohms, while capacitive reactance at  $C_a$  is only a few ohms at the oscillator frequency. Resistor  $R_k$  is needed only to provide a conductive path from the reactance tube cathode to ground, its value in ohms being suitable for grid bias, but several times the value of reactance at  $C_a$ .

Current in  $R_a$  and  $C_a$  is in phase with voltage  $E$ , but voltage across  $C_a$  lags the current by about 90 degrees. This lagging voltage is applied across  $R_k$  and to the reactance tube cathode. Equivalent voltage at the reactance tube grid is 180 degrees out of phase, as explained in connection with Fig. 1, and alternating grid voltage accordingly leads voltage  $E$ . Since reactance tube plate current is in phase with its grid voltage, this current leads voltage  $E$  by about 90 degrees. A leading current with respect to applied alternating voltage is characteristic of capacitance, so now the reactance tube is acting like a capacitance.

The reactance tube is in parallel with the oscillator tank circuit and with tuning capacitance  $C_t$  in that circuit. Consequently, reactive capacitance is added in parallel with  $C_t$ , to increase the combined capacitance. Increase of tuning capacitance lowers the oscillator frequency, and since reactive capacitance is increased by more amplitude of reactance tube plate current, the total effective capacitance is increased. Thus oscillator frequency is lowered proportionately to amplitude of reactance tube plate current. Plate current amplitude, and oscillator frequency, are varied by making the grid of the reactance tube more or less negative.

There are numerous variations of reactance tube circuits for providing either inductive or capacitive effects, but the basic principles are as explained.

**TUBES, ULTRA-HIGH FREQUENCY.** — A variety of tubes perform satisfactorily at uhf television carrier frequencies, among them being oscillators, mixers, and amplifiers. In all the uhf tubes interelectrode capacitances are small. Lengths and inductances of internal connections to elements have been reduced, and parallel leads to a single element often are used for further reduction of inductance in the tube and in connected circuit conductors.

Internal elements themselves are of small dimensions, even when the glass envelope is of the size used for ordinary miniature tubes. A comparison is shown by Fig. 1. At the left are the elements of a miniature pentode of the general purpose type. At the center are the elements of a 6J6, often used as a combined oscillator-mixer in vhf tuners. At the right is a 6AF4 uhf oscillator built with a regular miniature envelope and 7-pin base, but with the very small plate, grid, and cathode assembly which may be seen inside the bulb.

Among other uhf receiver tubes is the 6AN4 designed for use either as a mixer or as an amplifier. This is a miniature 7-pin type. With operating voltages ordinarily applied, the 6AN4 as an amplifier has transconductance of 10,000 micromhos, and as a uhf mixer has conversion transconductance of 2,900 micromhos. This, and other tubes designed especially for uhf application, have noise factors considerably lower than found with any type designed for general purposes.

Uhf tuners may have circuits basically like those for vhf reception and frequency conversion, as illustrated by the example of Fig. 2. There is a tuned antenna circuit coupled to a tuned mixer input circuit, but there is no r-f amplifier and the mixer is used in a grounded grid circuit. The oscillator is a modified Colpitts type, with all necessary inductance in leads and connections within and outside the tube. Each of the three tuned circuits is resonated by one section of a three-gang capacitor, with trimmer capacitors on each section.

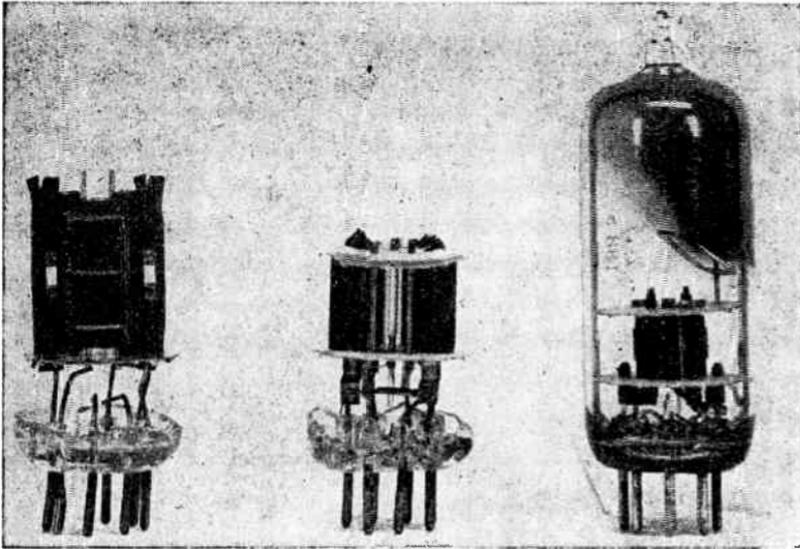


Fig. 1.—The tube at the right, designed for uhf application, has short internal leads which reduce the internal inductances.

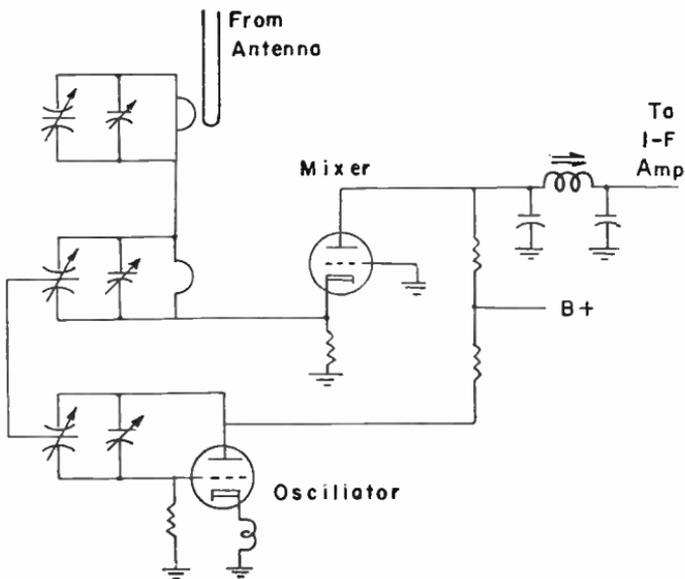


Fig. 2.—A circuit diagram, simplified, for a uhf tuner.

**TUNERS.** — It is the function of the television receiver tuner to convert all carrier frequencies in which signals are transmitted in different channels to a single intermediate frequency, while retaining the modulations which represent pictures and sound. Such conversion of many different modulated frequencies to a single frequency carrying the same modulation is the principle of the superheterodyne.

Essential parts of a tuner are represented in Fig. 1. Signals transmitted on carrier frequencies are picked up by the

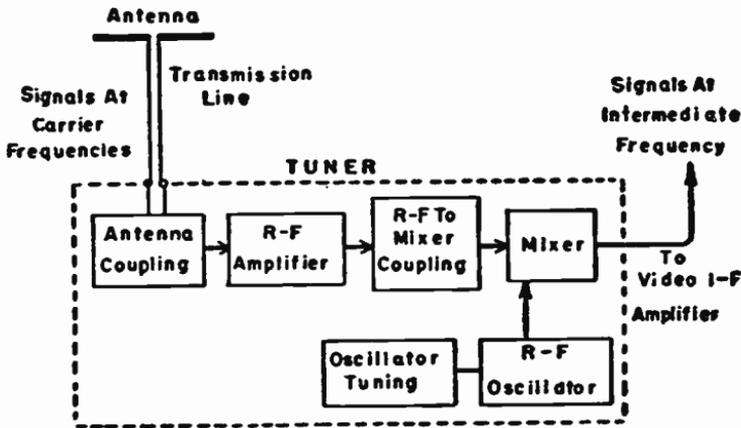


Fig. 1.—Principal parts of the tuner.

antenna, pass through the transmission line to the antenna coupling in the tuner, and are applied to an r-f amplifier consisting of one or more tubes. Amplified modulated carrier frequencies pass from r-f amplifier to mixer through a coupling adjustably tuned to carrier frequencies of the channel in which reception is desired.

The r-f oscillator is adjustably tuned to produce a frequency a certain number of megacycles higher than that of carrier frequencies of the channel in which there is to be reception. This requires that oscillator tuning be varied simultaneously with tuning of couplings for the antenna and between r-f amplifier and mixer. Thus oscillator frequency is maintained the same number of megacycles higher than carrier frequencies no matter what channel is received.

The oscillator frequency and carrier frequencies beat to-

gether in the mixer. The result, in the mixer output, is a beat frequency which is the difference between oscillator frequency and the carrier frequencies in any channel to which the receiver is tuned. Frequencies in the mixer output are equal, in megacycles, to the constant differences between carrier and oscillator frequencies. These are the intermediate frequencies at which amplification is carried out in sections of the receiver which follow the tuner.

*Example:* The received channel is number 5, in which carrier frequencies extend from 76 to 82 megacycles. Couplings between antenna and mixer are tuned to pass this range of frequencies. For reception in channel 5 the r-f oscillator may be tuned to 123 mc. Differences between this 123 mc and the low and high limit frequencies of channel 5, which are 76 and 82 mc, are 47 and 41 mc. Therefore, intermediate frequencies from the mixer will be in the range from 47 to 41 mc.

Were the receiver tuned to channel 7, in which carriers extend from 174 to 180 mc, the r-f oscillator tuning would be changed to 221 mc. Differences between oscillator and carrier frequencies again become 47 mc and 41 mc at the channel limits. By tuning together the r-f oscillator circuit and the carrier frequency couplings for each channel, the difference frequencies or intermediate frequencies will remain 47 to 41 mc regardless of the channel.

Tuning for oscillator and carrier frequencies may be by varying the inductance or by varying the capacitance of resonant circuits. With the very-high and ultra-high frequencies at which tuners operate there ordinarily is enough capacitance in tubes, wiring, and components for tuning in all frequency ranges merely by changing inductors or by varying the effective value of a single inductor. Accordingly, channel selection in many tuners is by switching of suitable inductors or else by adjusting the effective value of inductors. Small variable capacitors then are employed only for service adjustments. Other tuners provide channel selection by variable capacitors or sometimes by switching of a number of capacitors. Here again there may be additional small variable capacitors for service adjustments.

**Turret Tuners.** — The turret type tuner, of which an ex-

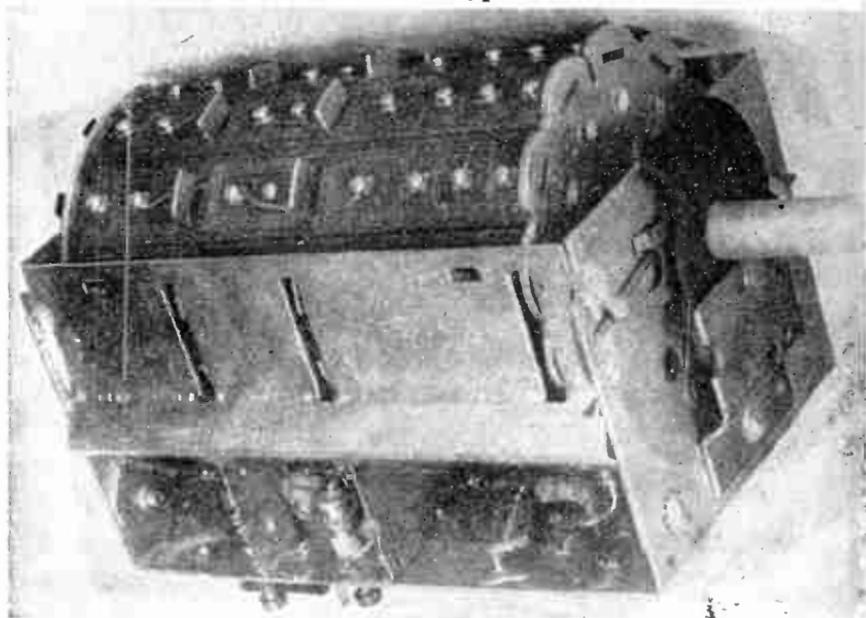


Fig. 2.

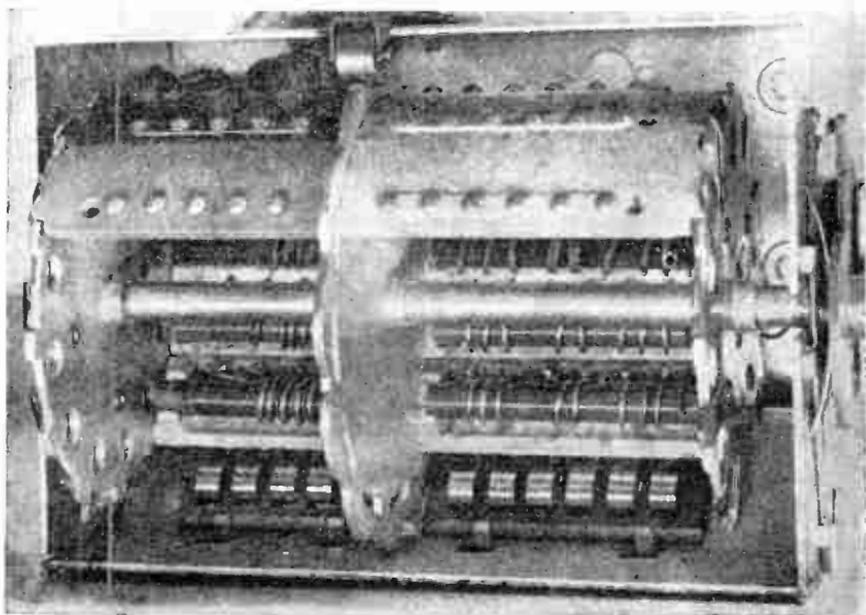


Fig. 3.

ample is pictured by Fig. 2, carries on a revolving drum a separate set of tuning inductors for each channel to be received. Inductors are on the inside of strips of insulation, and connect to small metal buttons on the outside of the strips. As the drum is revolved from one position to another by the channel selector knob the buttons engage stationary spring contacts from which connections lead to tube sockets and to the various circuit components in the tuner. Fig. 3 is a picture of a turret drum from which a few channel strips have been removed to expose inductors or remaining strips. The contact

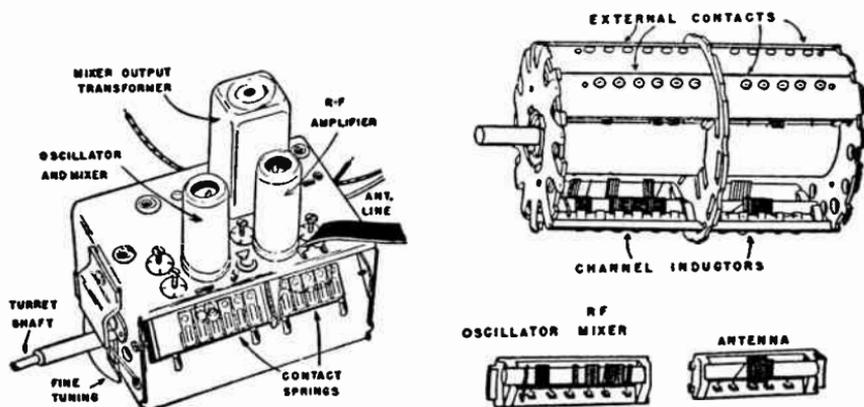
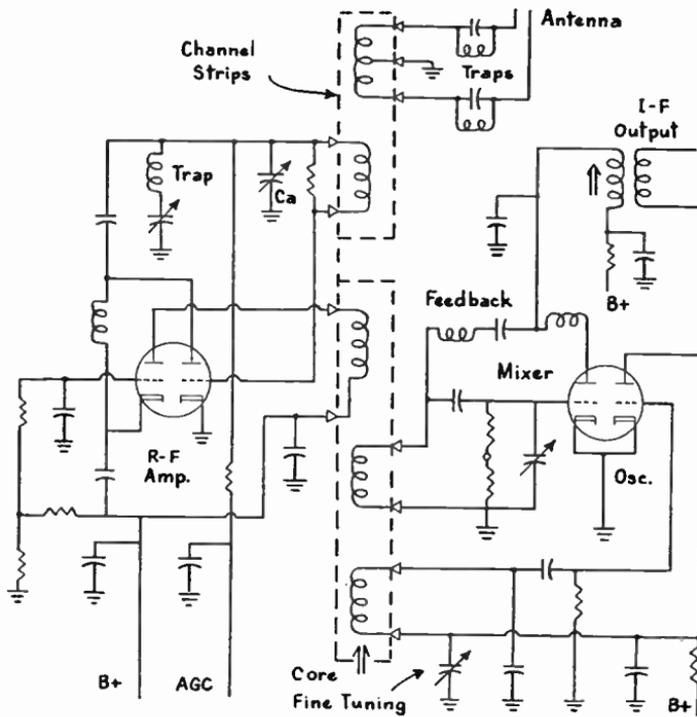


Fig. 4.—A turret tuner.

springs may be seen in the upper part of the frame. As shown by Fig. 4, the drum is removable from the frame after removing parts of the fine tuning control and retaining springs. Strips may be removed with the drum still mounted in the frame.

Fig. 5 is a circuit diagram for one style of turret tuner. Inductors for one channel are shown within broken lines at the center of the diagram. Inductors for other channels are brought into this broken-line position by rotating the drum. Contact buttons and springs are indicated by arrowheads to which connect the various circuits in stationary parts of the tuner.

From top to bottom of the diagram the inductors are as fol-



**Fig. 5.**—Circuit connections in a turret tuner employing a cascade r-f amplifier with another twin-triode as oscillator and mixer.

lows: First, a center-tapped coil whose ends connect to the antenna through traps which help prevent interference in the range of intermediate frequencies. Second the grid coil for the input section of the cascade r-f amplifier. On this coil is a trimmer capacitor  $C_a$ , also a series resonant trap adjustable to frequencies of local interference. Third, the plate coil for the output section of the r-f amplifier. Fourth, the grid coil for the mixer. From mixer plate to grid there is degenerative feedback through a capacitor and inductor. Fifth, the coil in the tuned circuit of the Colpitts type r-f oscillator. Inside this oscillator coil is an adjustable core used for alignment. Connected to one side of the oscillator circuit is a fine tuning capacitor.

Fig. 6 is a circuit diagram for a turret tuner in which the

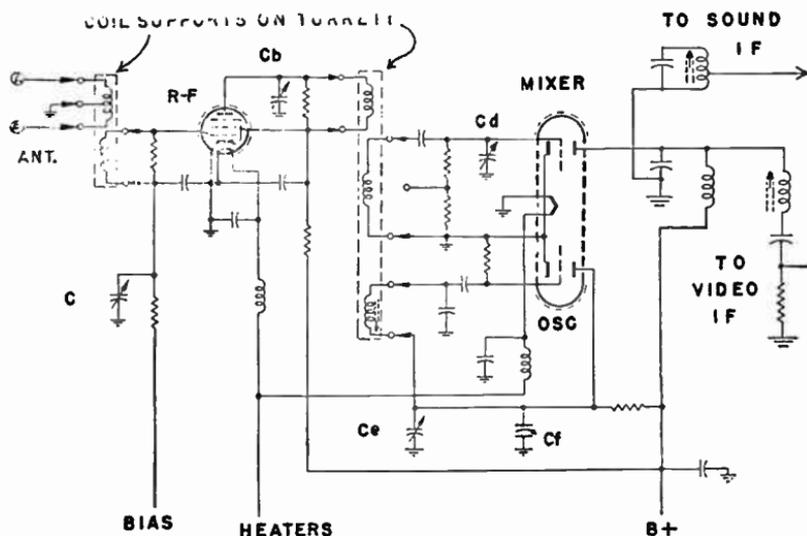


Fig. 6.—Circuit connections in a Standard Coil turret tuner.

r-f amplifier is a pentode. The channel strip carrying the antenna coil and r-f grid coil is shown at the left of the r-f amplifier, while the strip carrying the other three coils is shown in the center of the diagram. Alignment trimmer capacitors are as follows:  $C$  for the r-f amplifier grid.  $C_b$  for the r-f amplifier plate.  $C_d$  for the mixed grid.  $C_e$  for the oscillator. The fine tuning capacitor is marked  $C_f$ . The output of the mixer goes through one coupler to the video i-f amplifier section of the receiver and through another coupler to a dual or split sound section.

**Incremental Tuners.**—In a general kind of tuner often called an incremental inductance type the grid and plate circuits of r-f amplifiers, mixers, and oscillators are tuned for channel selection by a number of inductors connected in series. Between adjacent inductors are switch connections that allow selecting a portion of the total inductance suitable for tuning each channel. These connections are contacted one after another by rotor segments or rotor tongues of a multi-

section rotary selector switch. The switch and inductor compartment of a tuner of this type may be seen in Fig. 7.

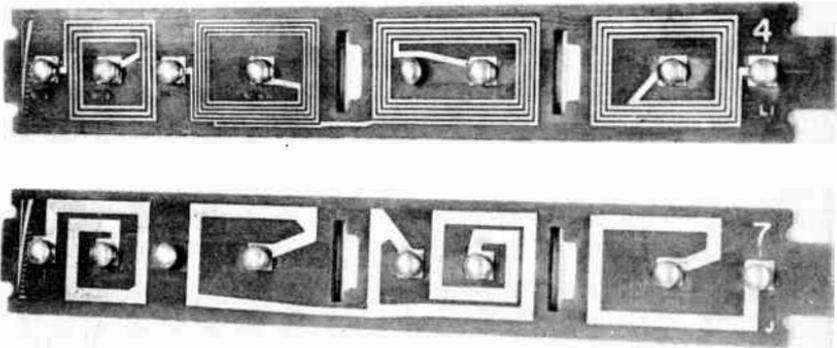


Fig. 7.—An incremental tuner with which channel selection is by means of rotary switch sections.

The principle of channel tuning by means of incremental inductance is illustrated by Fig. 8. At the top are shown in-

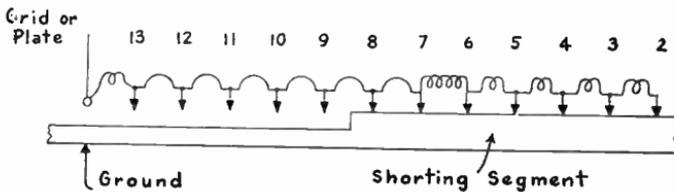
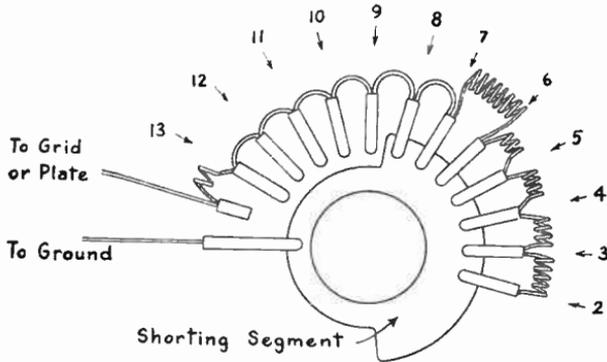


Fig. 8.

ductor sections on a switch wafer that has a shorting segment

on its rotor. Down below is represented the same switch as it may be drawn on some service diagrams. On other diagrams the switch sections are shown by symbols which look much like the wafers themselves. The switch in the figure has its shorting segment in position for reception of channel 8. All of the inductors used for reception of channels 2 through 7

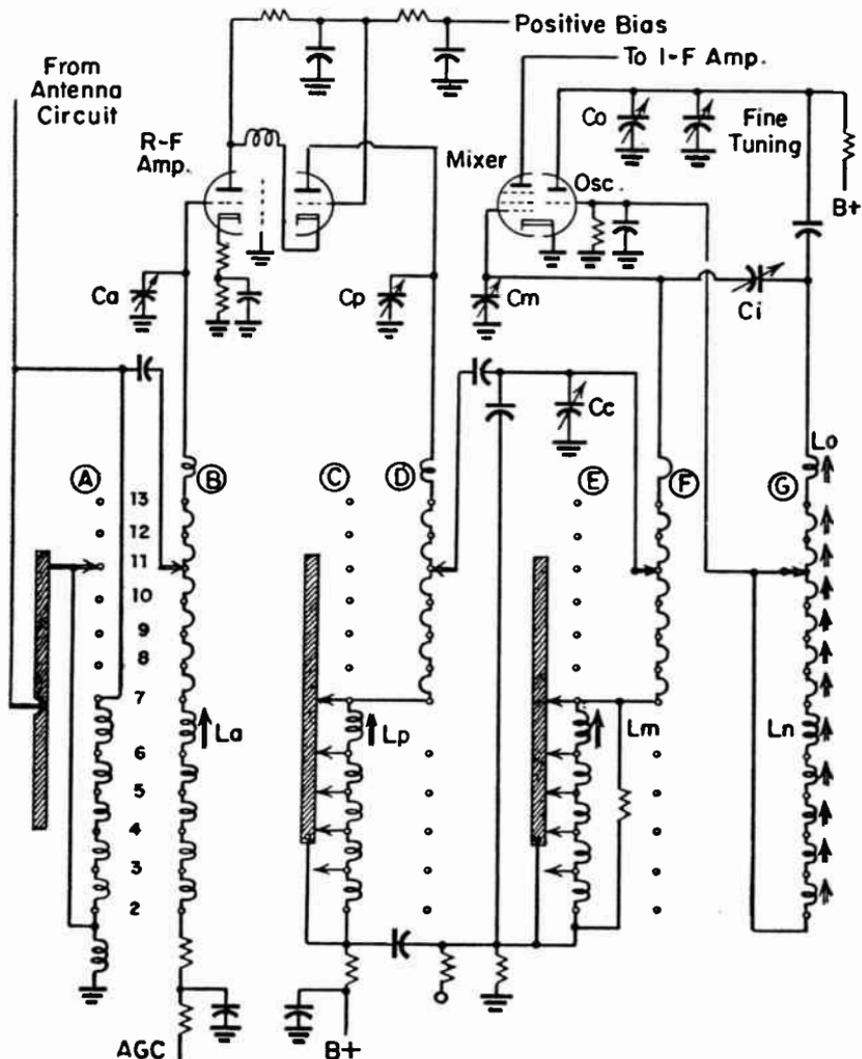


Fig. 9.—Circuit connections for an incremental tuner in which the r-f amplifier is a cascade type, with a pentode-triode for mixer and oscillator.

are shorted on one another and to ground. All inductors which are not thus shorted remain in series between a grid or plate and ground.

As the shorting segment is rotated clockwise it brings into the grid or plate circuit additional inductance in small steps or small increments. Inductors for high-band channels 7 through 13 consist of short, nearly straight pieces of wire between switch points. Inductance between the switch point for channel 13 and the grid or plate connection is sufficient for tuning this channel.

Were total circuit capacitance to be 15 mmf, total inductance for tuning to the center of channel 13 would be about 0.0372 microhenry, which would be the value of inductance between switch point 13 and the grid or plate connection. Then, to tune all the way from channel 13 to channel 7 would require only 0.0167 microhenry additional inductance, which would be the combined value of all the small inductors between positions 13 and 7 on the switch.

Tuning to channel 6, still with 15 mmf of circuit capacitance, requires total inductance of 0.2337 microhenry. There-

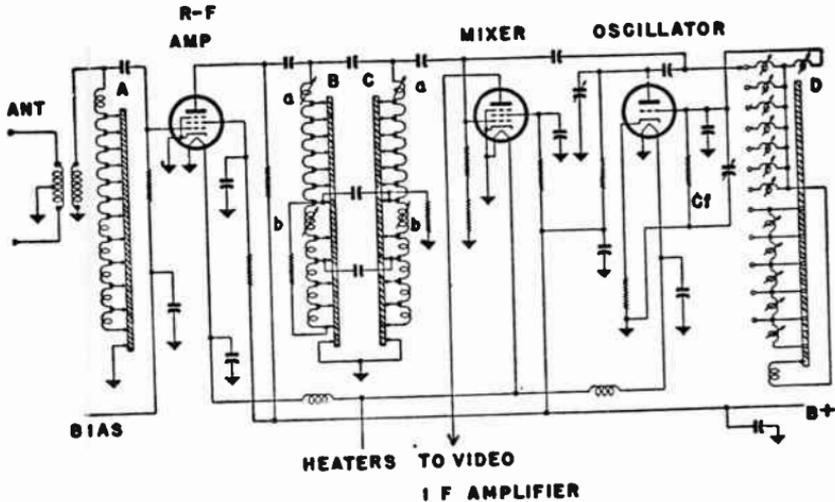


Fig 10.—Addition of antenna tuning with shorting connections made through a rotary selector switch.

fore, almost 0.1800 microhenry must be added by the large coiled inductor between positions 7 and 6. The remaining smaller coiled inductors bring total inductance to about 0.5200 microhenry for tuning channel 2, all when total circuit capacitance for tuning is 15 mmf.

In Fig. 9 are circuit connections for an incremental tuner having a twin-triode cascode r-f amplifier, and a pentode mixer and triode oscillator. Switch sections *A* and *B*, for antenna and r-f amplifier grid, are on opposite sides of the same

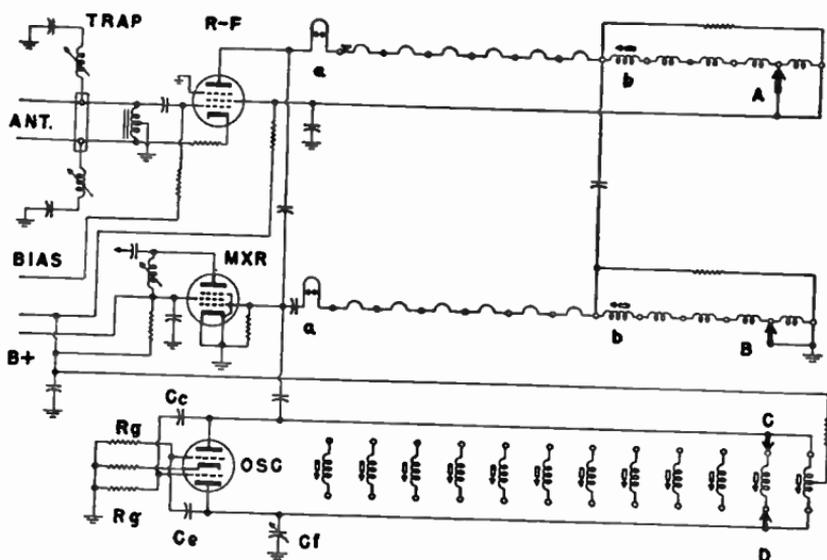


Fig. 11.—Tuner employing a push-pull oscillator with a twin triode tube.

wafer. This is true also of sections *C* and *D* for the r-f amplifier plate, and of sections *E* and *F* for the mixer grid. Shorting segments and tongues on opposite sides of each wafer move together. The switch is shown in position for reception of channel 11.

Inductors at *La*, *Lp*, and *Lm* are adjustments for alignment of the low-band. High-band alignment is made with adjustable trimmer capacitors *Ca*, *Cp*, and *Cm*. Capacitor *Cc*, also an alignment adjustment, varies the degree of coupling between r-f amplifier plate and mixer grid. At *Ci* is an adjustable

alignment capacitor for varying injection voltage from oscillator to mixer grid.

The r-f oscillator is a Colpitts type tuned by switch section *G*. Each small inductor on this section is individually adjustable for alignment of each channel. Trimmer capacitor *C<sub>0</sub>* is for overall alignment of the oscillator, being in parallel with the fine tuning capacitor.

Fig. 10 is a circuit diagram for an incremental tuner having a pentode r-f amplifier, a pentode mixer, and a triode r-f oscillator. Switch section *A* tunes the r-f amplifier grid circuit, *B* tunes the r-f amplifier plate, *C* the mixer grid,

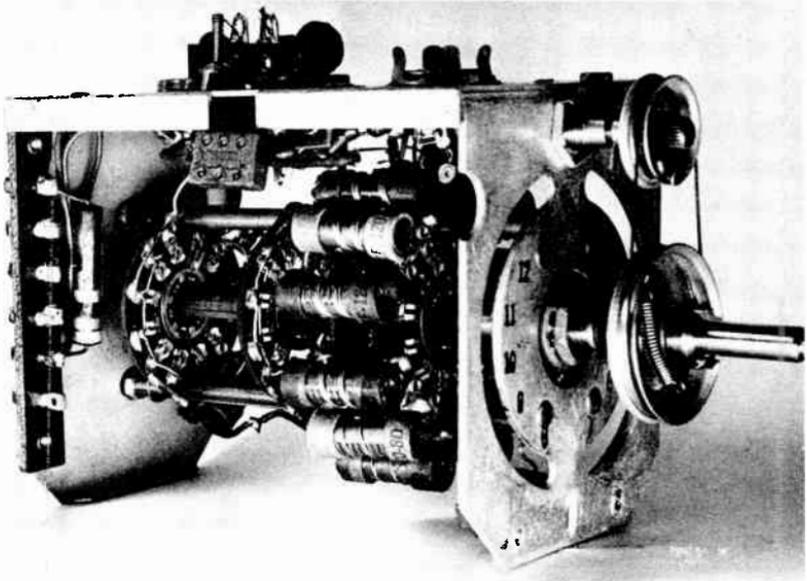


Fig. 12.—Inductors and switches for incremental tuning in the left-hand sections, with separate inductors at the right.

and *D* the oscillator. The upper inductors between r-f amplifier and mixer, marked *a-a*, are adjustable for alignment of high-band channels. Adjustment is by spreading or squeezing the coil turns. Inductors *b-b* are adjustable in the same manner for alignment of low-band channels. Other than these inductors, and a trimmer capacitor on the oscillator plate, there are no service adjustments. Capacitor *C<sub>f</sub>* is for fine tuning.

Fig. 11 shows circuits for a tuner employing incremental inductances for the r-f amplifier plate and the mixer grid, with shorting of unused inductor sections by means of rotor tongues instead of segments on the selector switch. The oscillator is tuned by separate inductors for each channel, with connections to the two ends of each inductor through contact tongues on rotors of two wafers in the selector switch. The switch and inductor compartment of a tuner of this general style is pictured by Fig. 12. The r-f oscillator used in tuners of Figs. 11 and 12 is a twin-triode push-pull type.

**Resonant Line Tuner.** — Fig. 13 shows circuits of a formerly popular resonant line tuner employing rotary selector

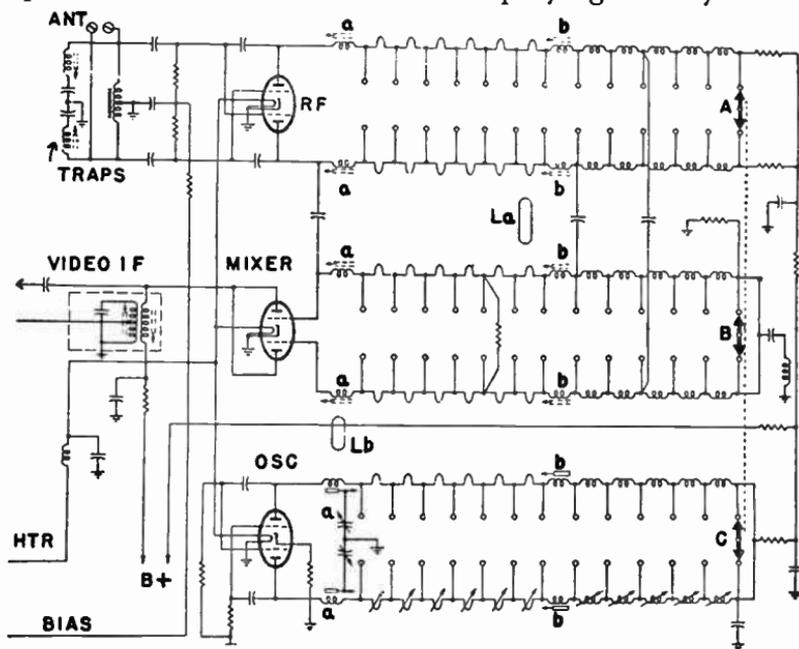


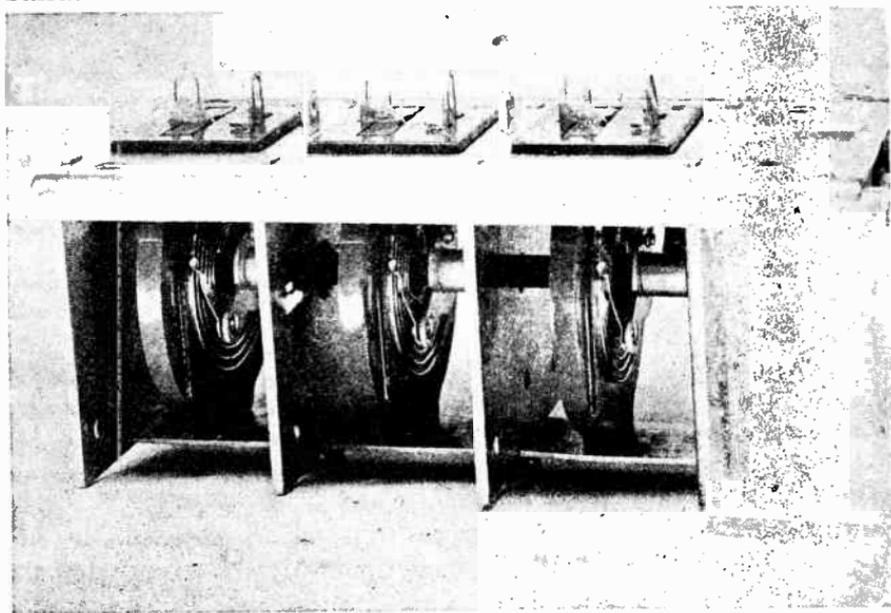
Fig. 13.—Shorted quarter-wave resonant lines in a tuner.

switches and appearing externally somewhat like some incremental inductance types. Each of three tuned circuits consists of a shorted quarter-wave resonant line. The two sides of each line are shorted together at points suited for tuning by movement of contact members *A*, *B*, and *C*. This alters effec-

tive or electrical lengths of the lines, and changes their resonant frequencies to suit each channel.

A shorted quarter-wave line is equivalent to a parallel resonant circuit tuned to a frequency depending on electrical length of the line, which is the length between two grids or two plates in the tubes and the shorting members operated by the selector switch. Link circuit *La*, and capacitors on either side, couple the r-f amplifier plate to the mixer grid. Link *Lb* couples the r-f oscillator to the mixer grid.

**Continuous Tuning With Inductors.** — Many tuners are designed for changing their resonant frequencies only in steps. They tune to channel frequencies, but not to any frequencies between channels or between bands. Other designs, which may be called continuous tuners, may be made resonant at any frequencies in either the low or high bands of the very-high frequency carrier range, and sometimes to frequencies between the top of the low band and the bottom of the high band.



**Fig. 14.**—A three-circuit Inductuner providing continuously variable inductance for channel tuning.

A continuous variable inductance unit called an Inductuner is pictured by Fig. 14. There are three inductors, each consisting of a spiral of metal ribbon over which is rotated a contact arm operated through gearing from the channel selector knob. Similar units are made with four adjustable inductors.

One method of using three spiral inductors in a complete tuner is shown by Fig. 15. The variable inductors are enclosed within the broken line. Between the r-f amplifier plate

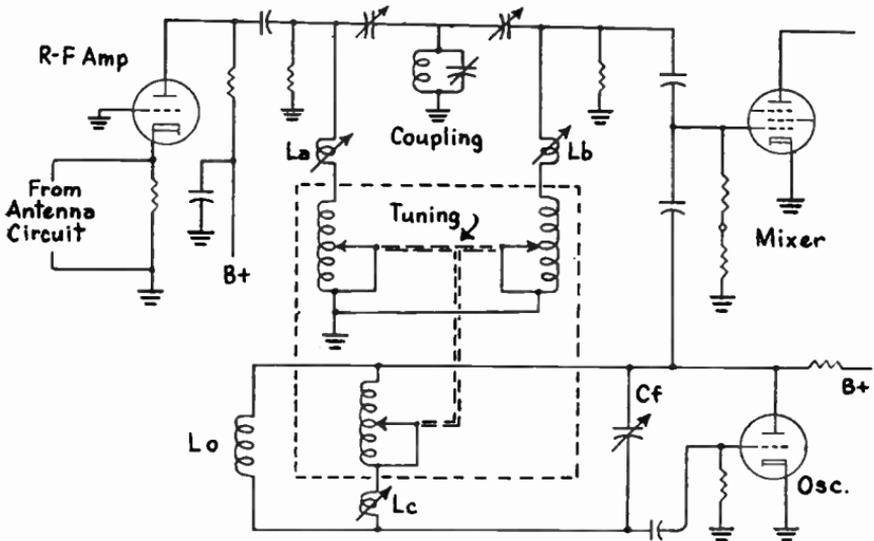
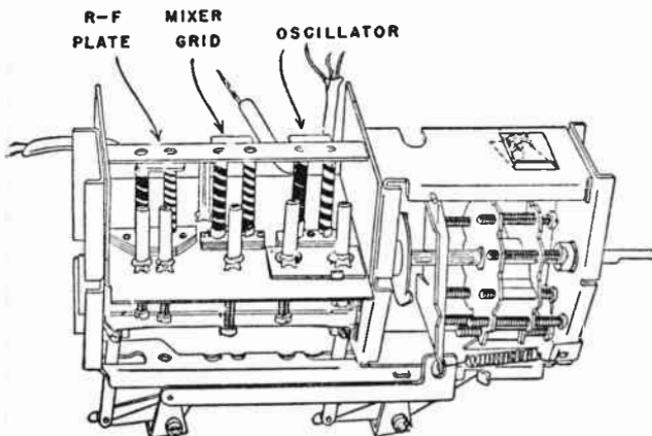


Fig. 15.—Connections for three spiral inductors in a tuner.

and mixer grid are three adjustable capacitors used for low-band alignment. In series with each variable inductor is a loop or small coil used for high-band alignment by spreading or squeezing the turns or by moving the loop or coil toward or away from adjacent metal.

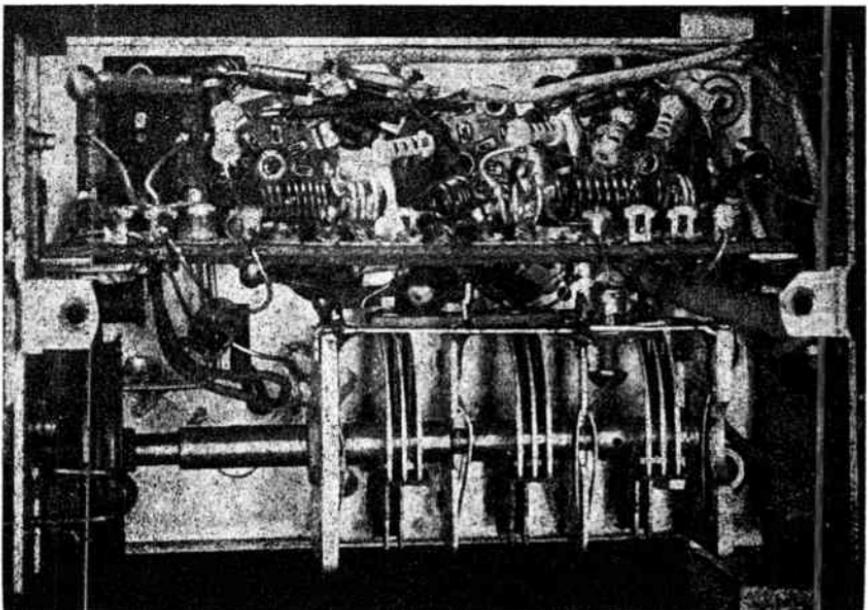
External inductor  $L_o$ , connected in parallel with the oscillator variable inductor, reduces effective oscillator tuning inductance to allow frequencies higher than received carrier frequencies. Capacitor  $C_f$  is an alignment adjustment for the oscillator. If the tuner is provided with fixed stops for each channel this capacitor may be used for fine tuning.

Fig. 16 illustrates an earlier type of tuner whose plate



**Fig. 16.**—A variable inductance tuner in which inductor cores are moved within the windings by the channel selector shaft.

and grid inductors are windings on tubular forms, with adjustment by movable cores operated from the channel selector knob. There may be continuous tuning through the low band and through the high band, with switching between bands.



**Fig. 17.**—Channel selection is by means of three continuously variable air-dielectric capacitors.

Ordinarily, however, there are stops or detents for each channel position, and for each position there is an alignment adjustment for varying the positions of the cores in the inductors.

**Capacitance Tuning.** — In Fig. 17 may be seen the interior of a continuously variable tuner with which channel selection is by means of three air-dielectric capacitors at the lower right in the picture. A circuit diagram for such a tuner is

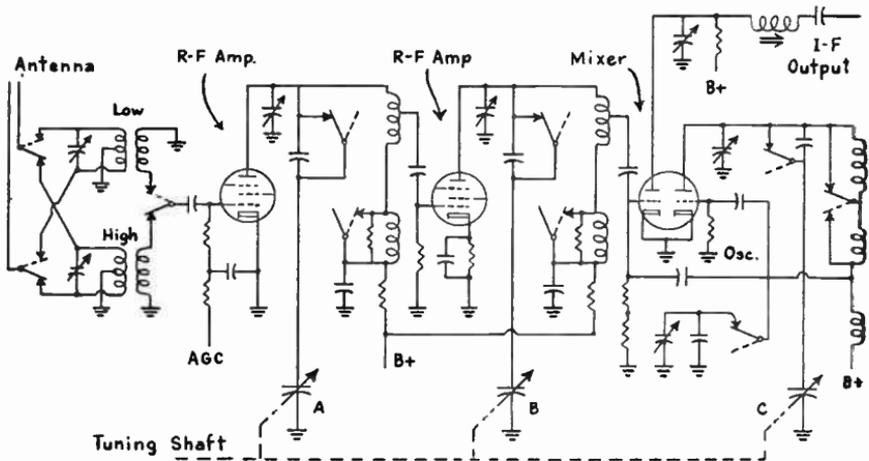


Fig. 18.—Circuit connections for continuous tuning with three variable capacitors on one shaft.

shown by Fig. 18. The three variable capacitors are operated by the single tuning shaft. There are two pentode r-f amplifiers in cascade. The first amplifier peaks at approximately the video carrier frequency of each received channel, while the second peaks at approximately the sound carrier frequency.

All inductors are brought in and out of the resonant circuits, for the low and high bands, by slide switches operated by a control knob concentric with the channel selector knob. For simplicity in circuit tracing these switches are represented by single-pole double-throw symbols, with low-band positions in full lines and high-band positions in broken lines.

Band switching is carried out as follows: The antenna transmission line is connected through two switch sections to

the primary of either a low-band or high-band coupling transformer. Another switch section simultaneously connects the corresponding transformer secondary to the grid of the first r-f amplifier. The coupling between first and second r-f amplifiers, also between the second r-f amplifier and mixer, includes two windings in series. Inductance is reduced for high-band tuning by a switch section that shorts one winding.

Small fixed capacitors are in series with the variable tuning capacitors, to decrease capacitance for high-band reception. These fixed capacitors are shorted for low-band reception. A small fixed capacitor in series with the oscillator variable tuning capacitor is similarly switched. There are two oscillator tuning inductors, one shorted out for high-band reception and the other for low-band reception.

Alignment trimmer capacitors are on each antenna transformer primary. There is a trimmer capacitor on the plate of each r-f amplifier, and another on the oscillator plate. With switches in the low-band position an additional trimmer capacitor is connected to the oscillator grid circuit. Since there is continuous tuning through each band, the variable capacitors may be adjusted to any frequencies within the band. Consequently there is no need for a fine tuning capacitor.

**Tuner Alignment.** — Alignment of tuners, as here considered, relates to adjustment of couplings between antenna and r-f amplifier and of couplings between r-f amplifier and mixer. Alignment of r-f oscillators is described elsewhere in this book. General instructions for all alignment operations are given under the heading of *Alignment*.

The preferred method for tuner alignment is with sweep and marker generators connected to the antenna input terminals, with a matching pad on the sweep output cable. The vertical input of an oscilloscope is connected through a fixed resistor of 5,000 to 15,000 ohms to the mixer grid return. Grid return resistors or mixers often are in two parts, with an intermediate test point to which the oscilloscope may be con-

nected. Such test points may be seen in Figs. 5, 6, 15, and 18. Where there is no grid return test point, the vertical input of the oscilloscope may be connected through a fixed resistor of 5,000 ohms or more to the mixer grid.

It is essential that any automatic gain control voltage on the r-f amplifier or amplifiers be temporarily overridden with a fixed bias. An override of  $1\frac{1}{2}$  volts should be satisfactory for pentode r-f amplifiers, and of 3 volts for cascode amplifiers. The contrast control may be set at or below a position for normal reception, since this control should not affect tuner frequency response.

Examination of tuner circuit diagrams will show that mixers are biased by series capacitors and grid-leak resistors. This bias becomes so highly negative during reception that the tube operates on the lower bend of its grid-voltage plate-current characteristic, behaving like a detector to partially rectify carrier-frequency voltages coming to the grid. For this reason it is not necessary to use a detector probe on the vertical input of the oscilloscope during tuner alignment.

Output voltage of the sweep generator must be kept within limits which only increase or decrease the height of the response curve without altering its shape or form. That is, sweep generator output must remain below the value at which the response curve commences additional flattening at the top, or develops new peaks, or otherwise changes its form. This precaution applies also to output of the marker generator.

With only limited output from the sweep generator, and with only the gain of the r-f amplifier or amplifiers between generator and oscilloscope, the signal at the oscilloscope vertical input will be weak. Then, to produce a trace of useful height, vertical sensitivity of the oscilloscope should be no less than about 0.01 r-m-s volt per inch of deflection, except for tuners having high r-f gain.

If a tuner lacks r-f gain, or if an oscilloscope lacks in vertical gain, it might seem possible to take a tuner frequency response from the mixer plate, thus utilizing the conversion gain. This would require disconnecting the mixer plate from

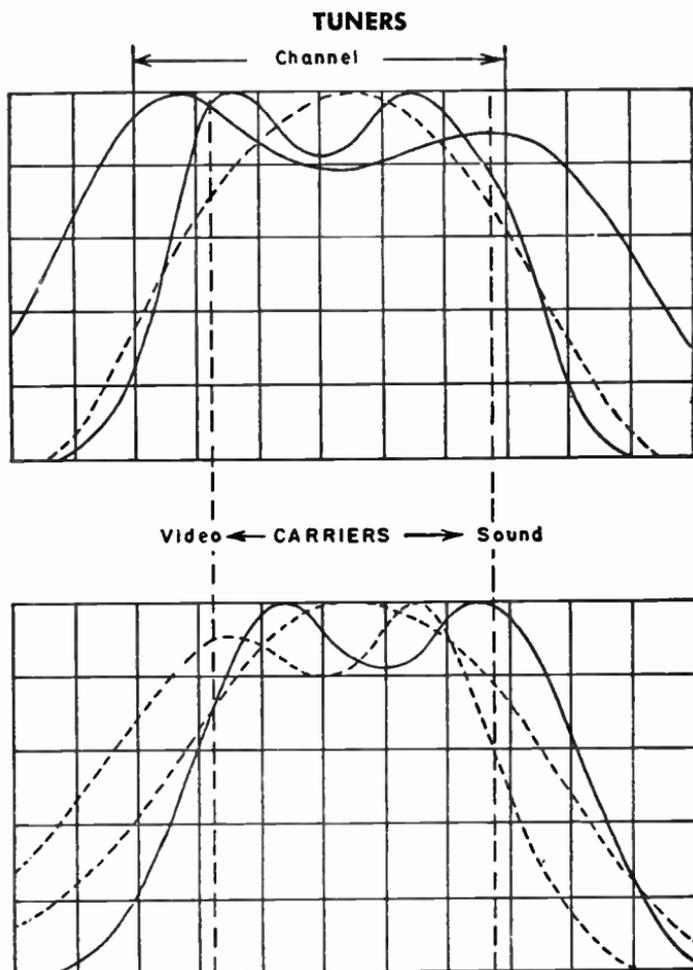


Fig. 19.—Typical response curves for tuners when input is to the antenna terminals and output at one of the mixer circuits.

any tuned coupling between mixer and first i-f amplifier, and usually would require providing temporary B-voltage for the mixer plate through a fixed resistor. Otherwise the following tuned coupling will prevent the trace from showing true frequency response of tuner circuits.

Furthermore, when taking a response from the mixer plate, it would be necessary to use a detector probe on the vertical input of the oscilloscope. Capacitance of the probe usually

would so alter the frequency response that it would not be that of tuner circuits between antenna and mixer grid.

Full-line curves of Fig. 19 show tuner frequency responses which are entirely satisfactory, while broken-line curves show responses not quite so satisfactory, but still acceptable. Video carrier frequencies should be between 70 and 100 per cent of peak gain, while sound carriers should be at least 50 per cent of the maximum. The top of the curve is preferably flat. A valley between peaks should not dip below 80 per cent of peak gain. Two peaks need not be of equal heights, but should not differ by much more than 10 per cent. Responses need not be exactly alike in shape for all channels.

Overall bandwidth or frequency separation between peaks of the tuner response, also the maximum gain, are affected chiefly by adjustment of coupling elements between r-f amplifier plate and mixer grid. Relative heights of peaks or tilt

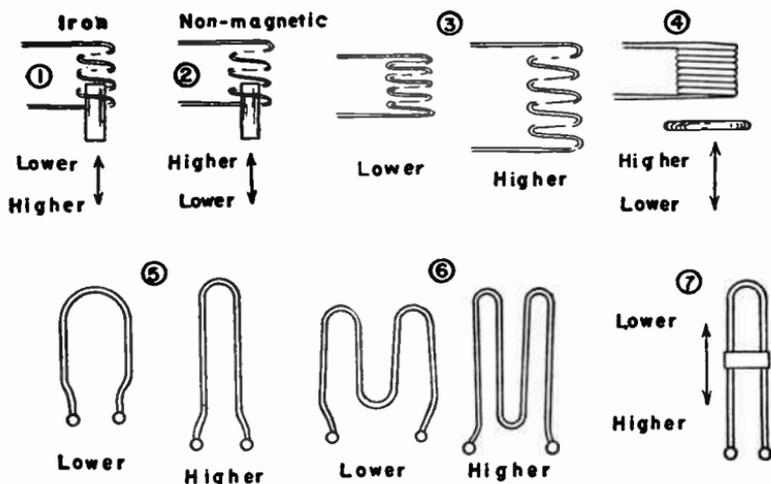


Fig. 20.—How resonant frequency is affected by various methods of adjusting inductors used in tuners.

across the top of the response is affected chiefly by adjustment of any coupling between antenna and r-f amplifier grid. Any adjustment which broadens a response will reduce maximum gain, while a narrower response will allow greater gain.

Adjustment should be for greatest possible gain or greatest height of the response trace while maintaining video and sound carrier markers within limits shown by Fig. 19.

Most earlier tuners, also many more recent types, provide r-f amplifier and mixer alignments only by means of adjustable inductors for the high band and for the low band. This practice is illustrated by Figs. 10, 11, and 13. Adjustments may be any of the types shown by Fig. 20, where changes for making the resonant frequency lower or higher are marked on the diagrams. Methods are as follows: 1 and 2. Metal cores screw in or out of the coils. Effects on frequency of a brass, copper, or aluminum non-magnetic core are opposite to those of a powdered iron core. 3. Coil turns may be squeezed together or spread apart. 4. A shorted turn may be moved with respect to the coil. 5 and 6. A hairpin loop or several flat loops may be made wider or narrower. 7. A slider may be moved on a hairpin loop while solder holding the bar is kept hot.

Movable cores often are held in position by a coating of wax over the ends. The wax may be scraped away or melted by bringing a hot soldering iron close to it. Where adjustable elements are held by cement, the fastening may be dissolved with lacquer thinner or some solvent made especially for the purpose. Movable cores should be adjusted only with alignment tools containing no metal at all, unless the core itself is grounded through the mounting. Tools used for bending or shaping loops and coils while the tuner is in operation should contain no metal.

Tuners such as shown by Figs. 5, 6, 9, and 18 have trimmer capacitors for r-f grid, r-f plate, and mixer grid. Capacitors on r-f plate and mixer grid are adjusted for bandwidth or separation of response peaks, while a capacitor on the r-f grid is adjusted for tilt or for relative heights of two peaks. Adjustments are made for best compromise of responses in channels affected by each trimmer. In addition to trimmer capacitors there may be adjustable inductors as, for example, at *La*, *Lp*, and *Lm* of Fig. 9.

**R-f Alignment.** — Tuner alignment consists of two inde-

pendent operations, one called r-f alignment and the other called oscillator alignment. R-f alignment is concerned with adjustments on couplings between antenna and r-f amplifier grid, and on couplings between r-f plate and mixer grid. Oscillator alignment has little or no effect on bandwidth or

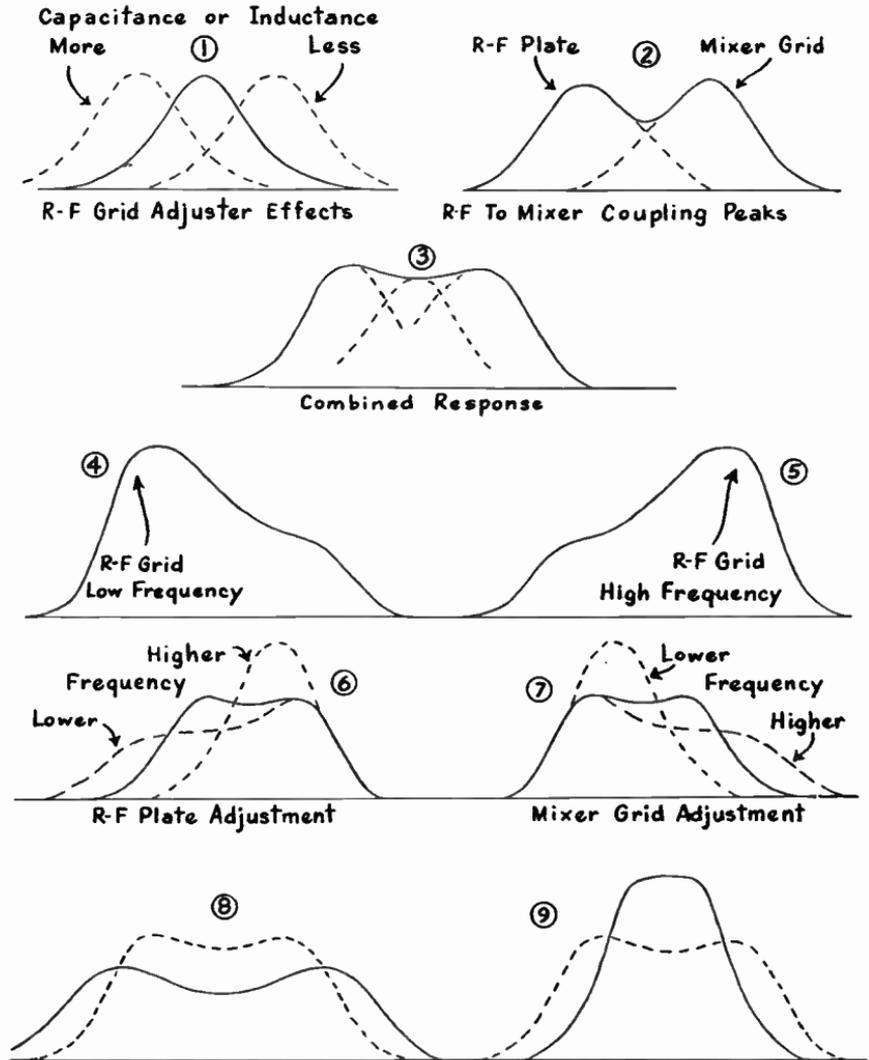


Fig. 21.—Frequency responses which may be observed during r-f alignment

gain, but does determine the intermediate frequencies which will appear at the mixer output.

R-f alignment adjusters are found on resonant circuits connected to the r-f amplifier grid, to the r-f amplifier plate, and to the mixer grid. There may be only adjustable trimmer capacitors, or only adjustable inductors, or there may be both capacitors and inductors on any or all of the three resonant circuits. Adjusters directly connected to tube grids or plates usually alter the alignment for all channels, in both low and high bands. Other adjusters may affect only high-band channels and have negligible effect in the low band. Still others may affect only low-band channels and have no effect in the high band.

No matter whether an adjuster is a capacitor or an inductor, it is capable of lowering the resonant frequency (with more capacitance or inductance) or of raising the resonant frequency (with less capacitance or inductance). To simplify explanations, capacitors or inductors will be referred to as (1) the r-f grid adjuster, (2) the r-f plate adjuster, and (3) the mixer grid adjuster.

General principles of r-f alignment are illustrated by Fig. 21. Diagram 1 shows frequency response of the coupling between antenna and r-f amplifier, considered by itself. If the original peak is at the full-line curve, the r-f grid adjuster will move the peak to a lower or higher frequency, as shown by broken-line curves.

Diagram 2 shows two peaks on the response of the coupling between r-f amplifier plate and mixer grid, considered by itself. One of these peaks may be shifted by any r-f plate adjuster and the other by any mixer grid adjuster. The r-f plate adjuster is here indicated as affecting the low-frequency side of the response, while the mixer grid adjuster affects the high-frequency side. This relation of adjusters and frequencies may be reversed, with the r-f plate adjuster affecting chiefly the high-frequency side of the response and with the mixer grid adjuster affecting the low-frequency side. Such reversal would not affect general principles of alignment now being discussed.

At 3 is shown the combined response due to r-f grid coupling and to coupling between r-f plate and mixer grid, as these two couplings affect carrier signal voltages at the mixer grid.

Diagram 4 shows what happens when the r-f grid adjuster lowers the frequency to bring the peak of diagram 1 more nearly into line with the r-f plate response of diagram 2. The result is a high peak on the low-frequency side of the combined frequency response, and a drop on the high-frequency side.

At 5 is shown the effect of changing the r-f grid adjuster for higher frequency, to move the peak of diagram 1 more nearly into line with the mixer grid peak of diagram 2. Now there is a drop on the low-frequency side of the combined response and a high peak on the high-frequency side. Were an observed frequency response to have excessive tilt in either direction, as at 4 or 5, the tilt would be removed by appropriate adjustment at the r-f grid.

In diagram 6 the full-line curve represents a combined response similar to that of diagram 3. Were this response to be extended on the low-frequency side it would be done by adjusting for lower frequency at the r-f plate, assuming that the r-f plate adjuster has its principal effect on the low-frequency side of the response. Were the original response to be narrowed or drawn in on the low-frequency side, adjustment would be made for higher frequency at the r-f plate.

Diagram 7 shows how an original response might be either widened or narrowed on the high-frequency side by adjustment at the mixer grid, assuming that the mixer grid adjuster has its principal effect on the high-frequency side.

Any adjustments at the r-f plate or mixer grid, as illustrated by diagrams 6 and 7, will produce tilting or unequal peaks at the top of the response. This must be corrected by suitable adjustment at the r-f grid. That is, the peak controlled by the r-f grid adjuster must be shifted toward the side of the response which drops, or away from the side that acquires a higher peak.

Should it be necessary to make the combined response wider at both sides, as in diagram 8, the r-f plate adjuster would be altered for lower frequency and the mixer grid adjuster for higher frequency. With resulting increase of bandwidth there will be decrease of gain. It will be necessary to vary the r-f grid adjuster only should tilt become excessive.

To make the response narrower on both sides, as in diagram 9, the r-f plate adjuster would be varied for higher frequency and the mixer grid adjuster for lower frequency. Maximum gain will increase with the narrower bandwidth. Again there will be need for r-f grid adjustment only should there be excessive tilt of the narrowed response.

It becomes fairly obvious that an entire response may be shifted to lower frequencies, without altering bandwidth, by varying all the adjusters for lower frequencies. Similarly, movement of the entire response to higher frequencies requires varying all the adjusters to higher frequency settings.

Although diagrams of Fig. 21 show certain adjusters as affecting only certain parts of the response, every adjustment actually affects every part of the response to greater or less extent. For this reason, each adjuster should be altered only a very little at a time, and all of them must be thus altered, one after another, until the desired result is attained. Too much change of an adjuster on any one circuit, without accompanying changes of other adjusters, is almost sure to throw the response so far out of alignment as to make recovery quite difficult.

**TUNERS, ULTRA-HIGH FREQUENCY.**— While uhf tuners commonly are combined with vhf tuners in receivers, similar types are found in separate uhf converter units. Although the fundamental principles of uhf circuits employed in uhf tuners may be the same as in tuners for very-high and other radio frequencies, the physical constructions are entirely different in many respects. This is due to such factors as the large inductive reactances of even the shortest straight conductors, and the small capacitive reactances between even the smallest conductors which are separated by insulation or dielectric materials.

For resonance at the lower limit, the middle, and the high limit of the uhf television band the products of capacitance and inductance must be approximately the following oscillation constants.

$$470 \text{ mc. Micromicrofarads} \times \text{microhenrys} = 0.115$$

$$680 \text{ mc. Micromicrofarads} \times \text{microhenrys} = 0.056$$

$$470 \text{ mc. Micromicrofarads} \times \text{microhenrys} = 0.032$$

Dividing the oscillation constant by any given capacitance will give the inductance required for resonance at the corresponding frequency, and dividing by a given inductance will give the required capacitance at the same frequency. For example, if total circuit capacitance is only 5 mmf and the frequency is 680 mc, dividing the oscillation constant, 0.056, by 5 gives approximately 0.011 microhenry as the inductance for resonance.

The inductance of one inch of straight wire of number 20 gage is about 0.019 microhenry, and of one-half inch of this wire it is about 0.008 microhenry. Therefore, a resonant circuit for the preceding example, with 5 mmf capacitance, would be completed with between one-half and one inch of number 20 straight wire.

Inductance of straight conductors is not directly proportional to length. Doubling the length more than doubles the inductance. Doubling a length which originally is less than about two inches increases the inductance about  $2\frac{1}{2}$  times.

The diameter, or cross sectional and surfaces areas of con-

ductors, have important effects on inductances. A small diameter conductor of any given length has more, not less, inductance than one of greater diameter having equal length. Inductance is increased by using smaller diameters, and is decreased by using larger diameters. This may be illustrated by comparing the inductances of one-inch lengths of conductors having different diameters, as follows.

No. 30 gage. Diameter 0.010 inch. Inductance 0.0254 microhenry.

No. 20 gage. Diameter 0.032 inch. Inductance 0.0190 microhenry.

No. 10 gage. Diameter 0.102 inch. Inductance 0.0136 microhenry.

When inductance must be reduced to an extent greater than allowed by a practicable increase of conductor diameter, two or more conductors may be arranged in parallel. Provided the paralleled conductors are not too close together the ordinary rules for parallel inductances hold good.

Fixed stray capacitances in uhf tuned circuits are large enough that addition of a variable tuning capacitor of ordinary design, with maximum to minimum capacitance ratio large enough for tuning through the uhf band, would result in total capacitance too great for use with inductance ordinarily attainable. It is for this reason that uhf tuning usually is with variable inductances used in combination with circuit and tube capacitances. Adjustable capacitances are, however, employed for tracking and other service adjustments which require only a relatively small change of capacitance.

The effective values of capacitances which appear unavoidably in tuned circuits may be reduced by arranging as many capacitances as possible in series with one another, thus making good use of the laws which apply to all capacitances in series. Trimming capacitances may be in series with fixed capacitances. While this reduces the range of the adjustable unit, the range for trimming purposes ordinarily remains amply great.

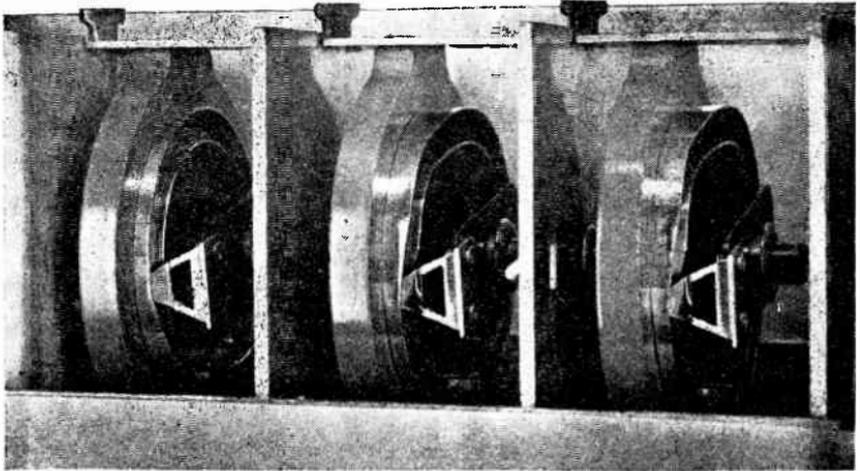


Fig. 138-10.—A uhf tuning unit having adjustable quarter-wave shorted lines for oscillator, mixer, and antenna circuits.

Many uhf receivers and converters have quarter-wave shorted resonant lines for tuning elements. Elementary principles are explained in the article on *Lines, Resonant*. A quarter-wave shorted line acts like a parallel resonant circuit when length of the line is adjusted for the applied frequency. The effective length is adjusted by sliding contacts which bridge across the two sides of the line, and may be moved by a tuning control to the position at which the line becomes resonant at a carrier frequency or at an oscillator frequency.

A tuning unit of this kind is pictured by Fig. 138-10. There are three sections, one for tuning the uhf oscillator, a second for tuning the antenna-mixer coupling, and a third for the antenna input circuit. The two sides of each line consist of concentric circular strips of metal embedded in low-loss insulating material. The sliders which short the lines at any required point around the circles are attached to a common shaft that is rotated by the control for selecting uhf channels.

Fairly typical circuits containing quarter-wave shorted lines are shown by Fig. 138-11, with the three lines represented as pairs of straight parallel conductors. Portions of the lines

above the shorting bars have no effect on resonant frequencies, the effective length of line being determined by the portions between the shorting bars and connections for antenna, mixer, and oscillator. The mixer is shown as a crystal diode, but could, of course, be a diode or triode tube. The center tuning element would be in the grid circuit of a mixer tube.

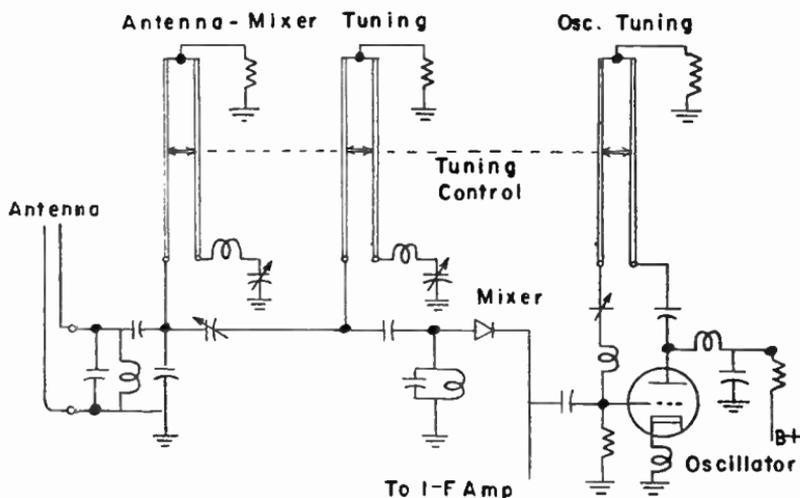


Fig. 138-11.—Quarter-wave resonant lines with adjustable shorting for uhf tuning.

Resonant frequencies of quarter-wave shorted lines are inversely proportion to the effective length of line. In Fig. 138-11 the lowest frequencies would be obtained with the bars or sliders moved all the way up, and the highest frequencies of resonance would result from moving the sliders downward to the limit of their travel. The upper ends of the two conductors of each line are permanently connected together and through resistors to ground as a means for completing return circuits for tubes and crystals as may be required by circuit design.

The electrical length of a resonant line is considerably greater than the actual measured or physical length, which is the same as saying that the physical length of the conductors from the shorting bars to the open ends is much less than the length of a quarter-wave in air at the resonant frequency. This

is because inductances and capacitances are added by external circuit conductors, by various trimmer elements, and by internal inductances and capacitances of connected tubes. There is also a reduction of velocity constant because the line conductors are supported by dielectric material rather than in air, and this tends to make the electrical length greater than the physical length.

Correct tracking of antenna, mixer, and oscillator circuits may be provided by making the line conductors of different dimensions, widths, thicknesses, or both, at various points along their lengths. Variations of this kind may be seen in Fig. 138-10. Tracking at various points in the frequency range may be accomplished in whole or in part by trimmer capacitances on the line conductors. Such trimmers may be seen at the open ends of the lines in Fig. 138-12. Additional trimming capacitances might be placed at points intermediate between the open and the permanently shorted ends of the lines.

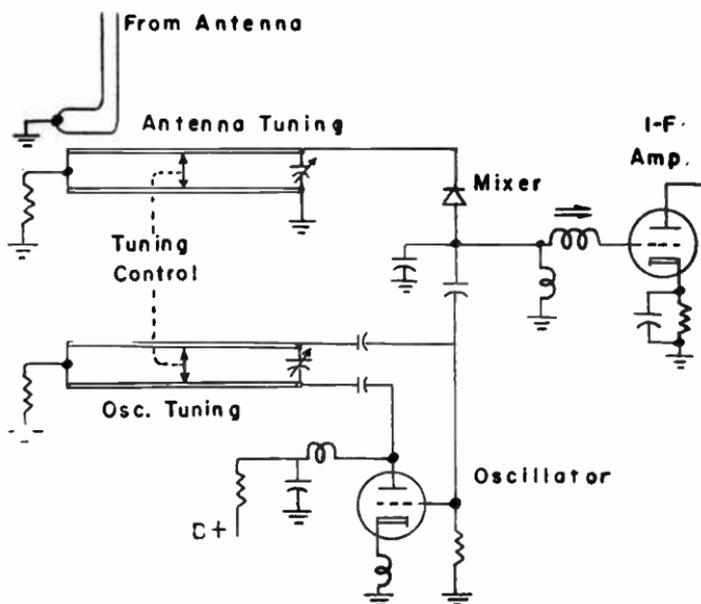


Fig. 138-12.—Two shorted quarter-wave resonant lines for tuning the antenna, mixer, and oscillator circuits.

In the design of Fig. 138-12 only two resonant lines are used for tuning. The line at the top of the diagram tunes the antenna input and the coupling to the mixer, with antenna at one end and mixer at the opposite end of this line. The lower resonant line is for tuning the uhf oscillator.

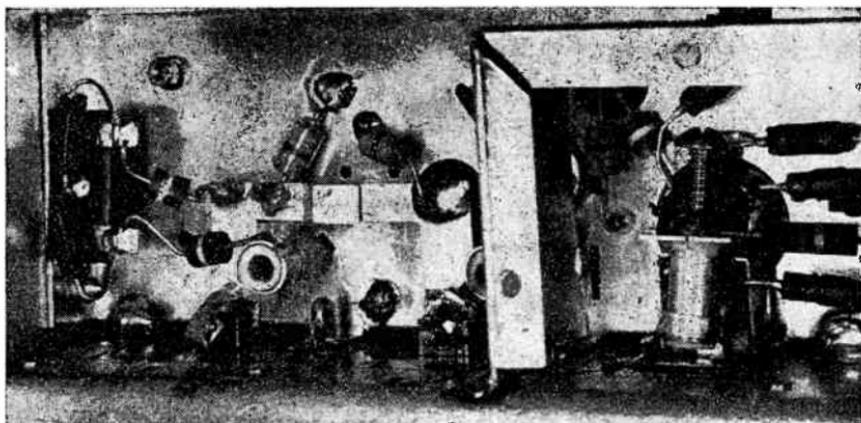
Spacing between the two conductors of a resonant line affects the characteristic impedance of the line, and by appropriate choice of spacing the line impedance may be matched to impedances of connected circuits and tubes. Wider spacing increases the line impedance. The characteristic impedance of the line is not altered as the effective length is changed by moving the shorting bars, this because the ratio of line inductance to capacitance is not affected by length.

The mechanical design of shorting contacts on the conductors of a quarter-wave line is important in relation to continued good performance and servicing problems. At the point of shorting there is maximum current, which calls for contacts of ample area. But voltage at this point is minimum, and may be so small that it cannot force the large current through even the slightest trace of corrosion or dirt.

At carrier and oscillator frequencies of the uhf television range there is strong tendency toward radiation of signal energy from resonant lines. This requires that the adjustable lines be enclosed by effective shielding, well grounded to the chassis. The outside shielding enclosure was removed when making the picture of Fig. 138-10. As may be seen, there are shielding partitions between the adjacent sections in this tuning unit.

**ULTRA-HIGH FREQUENCIES.** — Difficulties in uhf reception arise from the exceedingly small inductances and capacitances which produce resonance in this frequency range.

All resonant circuits must be isolated from other circuits by voltage dropping resistors or by r-f chokes where fairly high resistance cannot be used. High-frequency circuits must be completed through capacitances bypassing the currents around other circuits. Bypass capacitors to ground must be close to r-f chokes used for circuit isolation. Decoupling chokes ordinarily are found in all plate and screen circuits, also in both sides of leads to tube heaters. Grounding of plate, grid, and cathode returns for any one tube should be to the same point for avoidance of inductive couplings in metal between separated grounds.



*Fig. 138-19.—Circuit elements and connections in a uhf tuning system.*

Fig. 138-19 shows some of the elements and connections for antenna, mixer, and oscillator circuits in a typical unit. Parts associated with a tube are grouped close to the socket, with supports and interconnections provided by terminals and pigtail leads of the units themselves, to eliminate wire connections. Sockets must be made with low-loss dielectric materials. Tubes should be held rigidly by their sockets. Oscillator tubes always are provided with close fitting shields held securely to

the chassis by grounding clamps. This helps to stabilize the oscillator frequencies.

Circuit diagrams usually represent inductors and capacitors by the usual symbols for such elements, but structurally the actual elements may bear little resemblance to those in circuits operating at lower frequencies. Capacitances may be furnished by any adjacent pieces of metal, however small, with any kind of dielectric material between them. Inductive couplings may be those between two straight conductors of small size, or between a screw or a metal pin and other parts. Coils, when used, are of diminutive size when in resonant circuits, but may be much larger for r-f chokes. In one case the coupling from oscillator to mixer is through the cathode-heater capacitance of the oscillator tube.

In each ultra-high frequency channel the video carrier frequency is 1.25 mc above the low limit. There is the standard separation of 4.5 mc between video and sound carriers, bringing the sound carrier frequency 0.25 mc below the high limit of the channel. These are the same frequency distributions as in each very-high frequency television channel, with vestigial sideband transmission in both cases. Usual abbreviations for ultra-high frequency are *uhf* or *u-h-f*, corresponding to the abbreviation *vhf* or *v-h-f* for very-high frequency.

Uhf carrier waves behave like light rays to an even greater extent than vhf carrier waves. There is less bending at and around objects in the path of the waves, and transmission more nearly approaches a line-of-sight direction between transmitting and receiving antennas. Intervening structures, rises of land, trees, and other objects cut off the uhf waves somewhat more than vhf waves. With transmitting power and conditions roughly equivalent between transmitters and receivers, the maximum reception distance for uhf signals may be about the same or slightly less than for vhf signals.

A major advantage in uhf television broadcasting is the reduction or elimination of electrical interference of the spark type, and from sources operating at such moderately high.

frequencies as used in medical and some industrial apparatus. The chief disadvantage, so far as reception is concerned, is the increase of signal energy losses at ultra-high as compared with very-high carrier frequencies.

Detailed information on ultra-high frequency receiver components will be found under the following heading in this book.

*Antennas.*

*Channels.*

*Converters and Conversion,  
Uhf.*

*Lines, Transmission.*

*Mixers.*

*Oscillators.*

*Tubes, Ultra-high Frequency.*

*Tuners, Ultra-high Fre-  
quency.*

## W

**WAVEFORMS.**— A waveform trace on the screen of an oscilloscope exhibits the voltage variations or the "shape" of the composite television signal and of other voltages derived from or controlled by the signal.

Waveforms may be observed at the output of the video detector and at all points from there to the picture tube. The signal may be followed from the video detector through the video amplifier to the signal input at the picture tube grid or cathode. The signal may be followed also from the video detector through the sync and sweep sections to the deflection coils or the deflection plate connections for the picture tube.

For checking waveforms at all points through the video amplifier and also through the sync section as far as inputs to the sweep oscillators it is necessary to employ an input signal from a regular television transmission or one from a television pattern generator. When using a signal from a television transmission it is preferable to work with a test pattern rather than a picture in which there is movement. From the outputs of the sweep oscillators through the picture tube deflection systems it is not necessary to have an input signal, since the oscillators themselves will control sweep of the electron beam in the picture tube by providing necessary voltages.

For observation of waveforms the vertical input of the oscilloscope is connected between the point where the signal or voltage waveform is to be observed and chassis ground or B-minus. Voltage at the point of test must be no greater than the rated maximum vertical input voltage for the oscilloscope. To help isolate the tested circuit from capacitance of the oscilloscope input it is advisable to use a series resistor of about 100,000 ohms on the receiver end of the vertical input cable.

To observe waveforms related to vertical deflection the internal sweep of the oscilloscope is set for 60 cycles per second to cover a period of one field, for 30 cycles to cover two fields, or for 20 cycles to cover three fields on the same trace. To observe horizontal deflection waveforms the internal sweep of the oscilloscope is set for 15,750 cycles per second to observe one line period, for 7,875 cycles to observe two lines, or for 5,250 cycles to observe three lines on a single trace. Usual practice is to employ a trace covering two fields or two lines.

The composite television signals from different transmitters may show variations in apparent form. The shapes of all traces and the details which may be seen are affected by frequency response characteristics and such features as linearity in the oscilloscope. There may be slight differences between waveforms taken from different receivers of the same model. Waveforms taken from sync and sweep sections will conform to the operating principles employed in the receiver being tested. The height of traces taken from points between video detector and inputs to the sweep oscillators is altered by adjustment of the contrast control, while adjustment of size controls varies the height of traces taken from sweep circuits which follow the oscillators.

Examples of typical waveforms and descriptions of what they indicate are shown in articles listed below.

<i>Amplifiers, sweep</i>	<i>Restoration, d-c</i>
<i>Clippers or limiters</i>	<i>Scanning</i>
<i>Deflection, electrostatic</i>	<i>Separation, sync</i>
<i>Deflection, magnetic</i>	<i>Signal, television</i>
<i>Drive controls</i>	<i>Sync section</i>
<i>Filters</i>	

Figs. 139-1 to 139-24 are waveform traces observed at various points between video detector and picture tube of a typical receiver having a magnetic deflection picture tube and em-

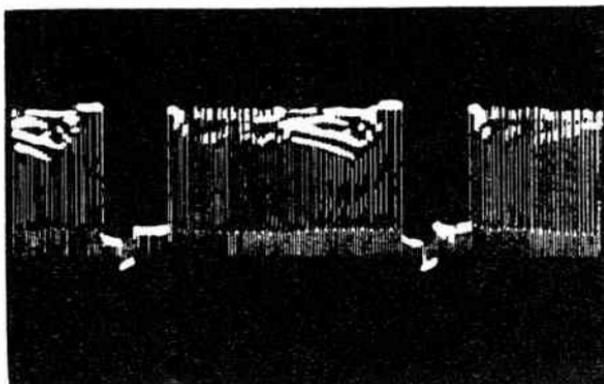
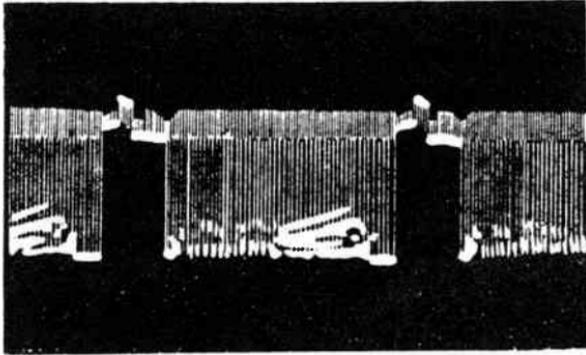


Fig. 139-1.

ploying a triode afc control tube for the horizontal sweep oscillator. All traces in this series are reproduced by permission of the Admiral Corporation.

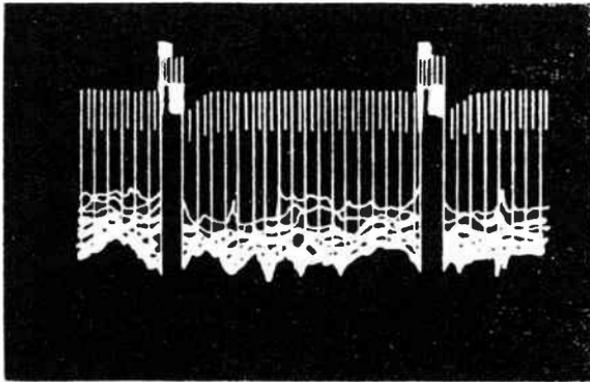
Traces shown by Figs. 139-1 to 139-10 inclusive are taken with the internal sweep of the oscilloscope adjusted for 30 cycles per second, to show two field periods. Consequently all these traces show waveforms associated with complete fields, with vertical blanking intervals, vertical sync pulses, equalizing pulses, and vertical deflection voltages.

*Fig. 139-1:* Taken at the top of the video detector load resistor. This is the output of the video detector and the input to the grid circuit of the video amplifier. Here appears the entire composite television signal with picture variations, positive, at the top and with sync pulses, negative, at the bottom. Two vertical blanking intervals are plainly visible between the fields. During each blanking interval there appear in order, from left to right, the equalizing pulses which follow one field, then the vertical sync pulses at the bottom-most points along the trace, and finally the remaining equalizing and horizontal sync pulses which precede the next field.



*Fig. 139-2.*

*Fig. 139-2:* Taken at the plate of the video amplifier tube. Here again is the complete composite signal, but now the polarity has been inverted to make sync pulses positive and picture variations negative. This waveform is applied to the cathode of the picture tube, which is the point of signal input to the picture tube of this particular receiver.



*Fig. 139-3.*

*Fig. 139-3:* Taken at the grid of the first tube in the sync section, which is a sync amplifier. The signal shown here comes from the output of the video amplifier, and accordingly is of the same polarity and has the same general characteristics as shown in Fig. 139-2.

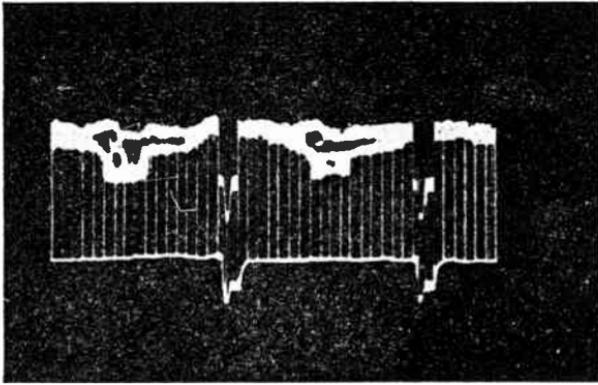


Fig. 139-4.

*Fig. 139-4:* Taken at the plate of the sync amplifier tube. The polarity has been inverted with respect to polarity in Fig. 139-3. The peak-to-peak voltage of this amplifier output waveform actually is about four times as great as voltage at the input to the tube.

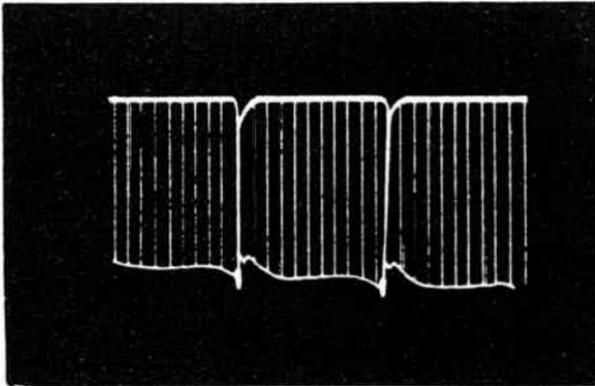
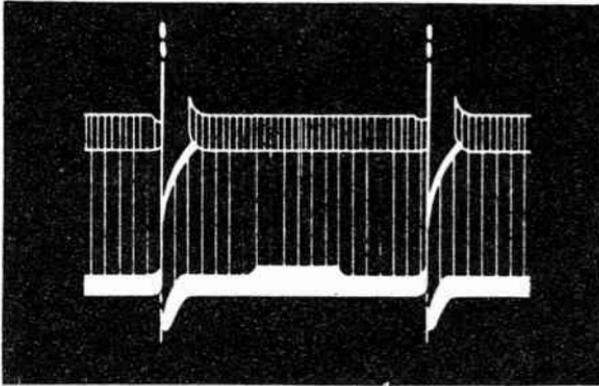


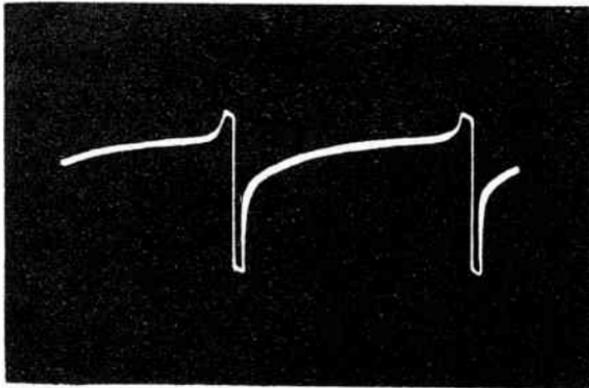
Fig. 139-5.

*Fig. 139-5:* Taken at the plate of the second tube in the sync section, which is operated as a separator. Picture variations have all but disappeared from the signal, while the vertical sync pulses have been retained. Polarity has not been inverted, because signal input is to the cathode rather than the grid of this separator.



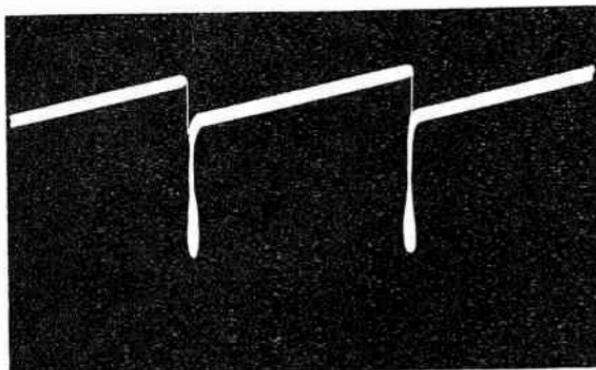
*Fig. 139-6.*

*Fig. 139-6:* Taken at the plate of the third tube in the sync section, which is operated as a clipper. Polarity has been inverted with respect to that of Fig. 139-5. Vertical sync pulse voltage peaks have become very pronounced. This is the signal which goes to the integrating filter located between the sync clipper and the input for the vertical sweep oscillator.



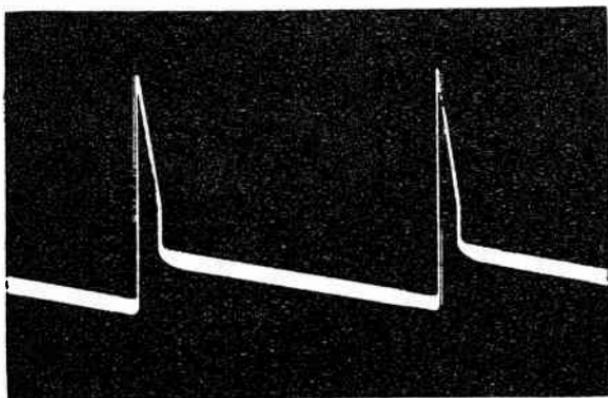
*Fig. 139-7.*

*Fig. 139-7:* Taken at the grid of the vertical sweep oscillator, which is a blocking type. Note the sudden changes of potential in the negative direction, downward on the trace, as the oscillator blocks. Then comes the quick partial recovery in the positive direction and the more gradual change preceding the positive peak that triggers this oscillator.



*Fig. 139-8.*

*Fig. 139-8:* Taken at the grid of the vertical sweep amplifier which follows the vertical oscillator. This is the sawtooth voltage combined with negative (downward) peaks as required for magnetic deflection.



*Fig. 139-9.*

*Fig. 139-9:* Taken at the plate of the vertical sweep amplifier. Polarity has been inverted with respect to the previous trace, taken at the grid of the same tube. Peak-to-peak voltage here is about 18 times as great as at the grid.

*Fig. 139-10:* This final trace for the vertical deflection system is taken from the circuit which includes the secondary winding of the vertical output transformer and the two ver-

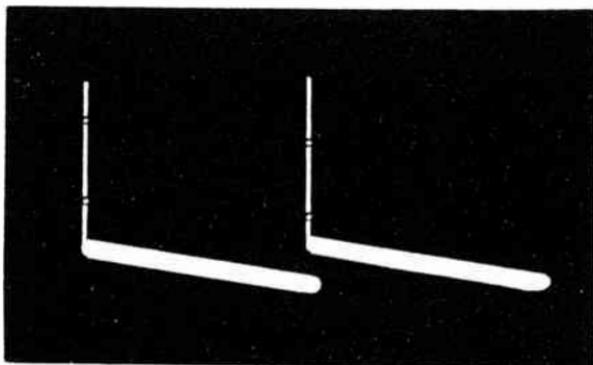


Fig. 139-10.

tical deflection coils of the yoke on the picture tube. Peak-to-peak voltage is between one-ninth and one-tenth of that at the plate of the vertical sweep amplifier, which connects to the primary of the output transformer.

Traces which are to follow in Figs. 139-11 to 139-24 are taken with the internal sweep of the oscilloscope adjusted for 7,875 cycles per second or to the frequency which produces two horizontal line periods. All these traces show waveforms associated with horizontal sync pulses and horizontal deflection voltages.

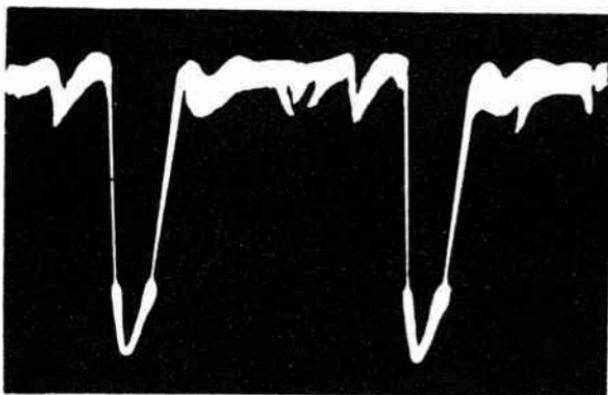


Fig. 139-11.

*Fig. 139-11:* From the top of the video detector load resistor. Here are two horizontal sync pulses of negative polar-

ity, also the horizontal blanking intervals, and the picture variations occurring between the blanking intervals.

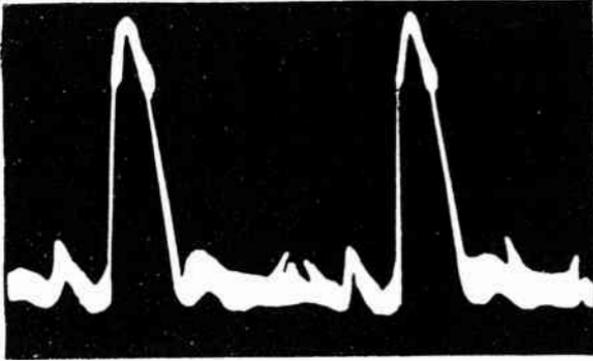


Fig. 139-12.

*Fig. 139-12:* From the plate of the video amplifier tube. Except for inversion of polarity this trace is similar to the one taken from the grid of this tube. Peak-to-peak voltage has been increased about nine times.

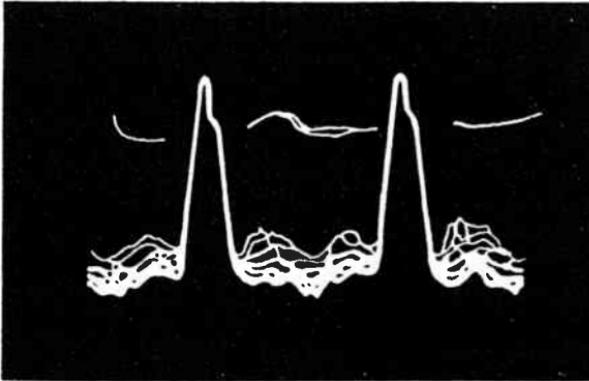


Fig. 139-13.

*Fig. 139-13:* From the grid of the sync amplifier, the first tube in the sync section. This signal comes from the output of the video amplifier, and is of the same polarity as in Fig. 139-12.

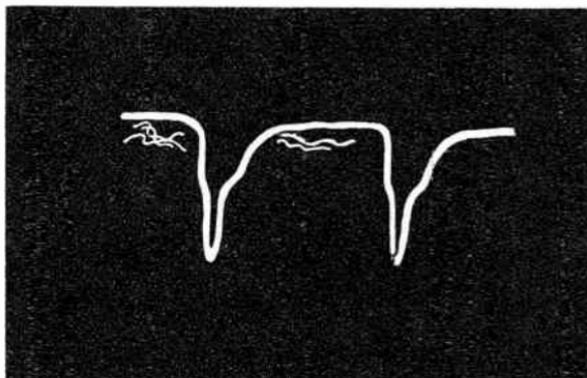


Fig. 139-14.

*Fig. 139-14:* From the plate of the sync amplifier. Polarity has been inverted. Voltages for picture variations have very nearly disappeared, while horizontal sync pulses have become distinct.

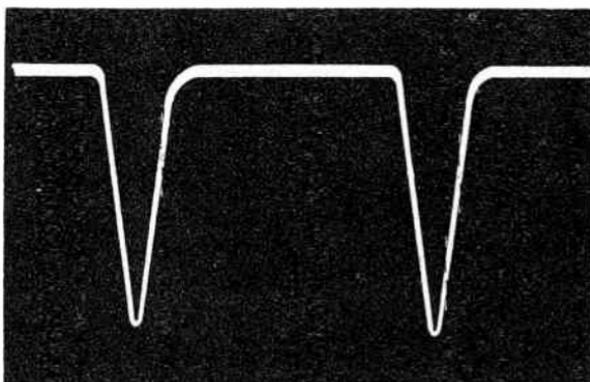


Fig. 139-15.

*Fig. 139-15:* From the plate of the sync separator tube. Only the horizontal sync pulses now remain. There has been no inversion of polarity, due to use of cathode input to this tube.

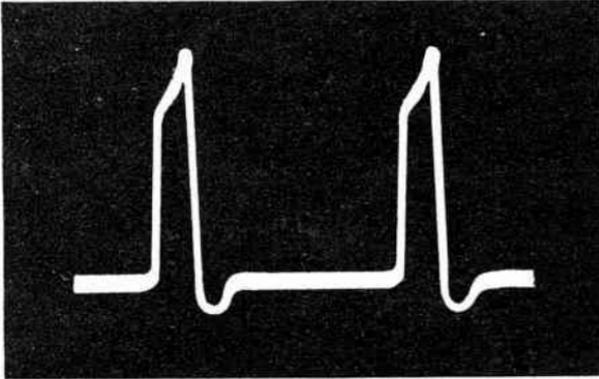


Fig. 139-16.

*Fig. 139-16:* From the plate of the sync clipper tube. This waveform is the input to the differentiating filter located between the clipper and the horizontal oscillator control tube of the horizontal afc system.

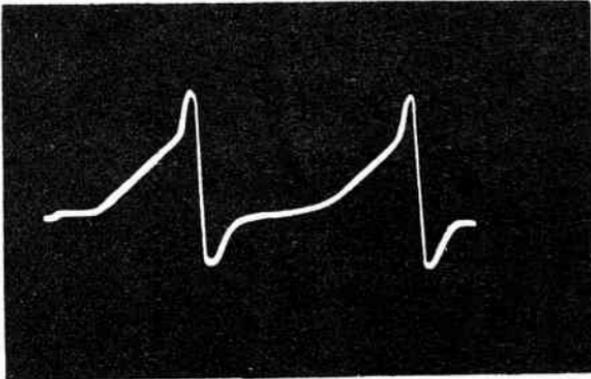


Fig. 139-17.

*Fig. 139-17:* From the top (ungrounded side) of the lock-in control capacitor in the grid circuit of the control tube of the horizontal afc system. This voltage results from combination of the output from the differentiating filter and a feedback voltage from the horizontal sweep output circuit, as required for this method of oscillator control. The waveform shown here is taken while a transmitted television signal is being received. The sharp or narrow positive peaks

represent synchronizing voltages which result from horizontal sync pulses in the signal.

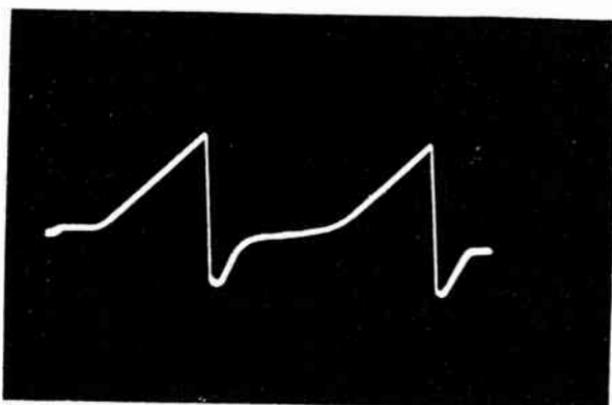


Fig. 139-18.

Fig. 139-18: This is the same as the previous trace, except that it is taken while no transmitted signal is being received. Note the absence of positive synchronizing peaks at the tops of the sawtooth portions of the wave.

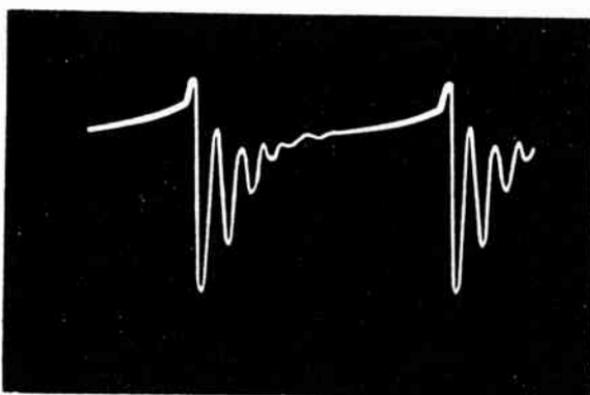


Fig. 139-19.

Fig. 139-19: From the grid of the horizontal sweep oscillator. Average grid voltage is controlled by biasing effect developed in the grid-cathode circuit of the control tube as explained in the article *Oscillator, Television Sweep, Control of* under the sub-heading *Triode Afc Control Tube*. The trace is taken while no signal is being received.

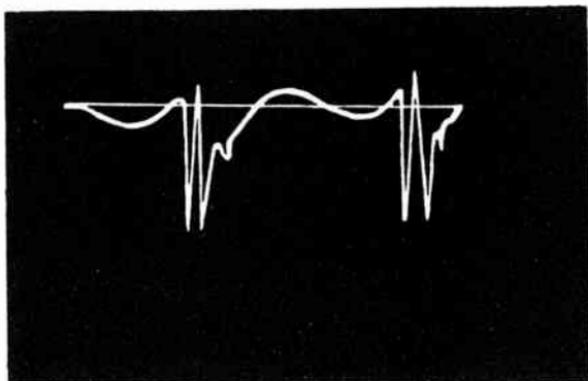


Fig. 139-20.

*Fig. 139-20:* From the plate of the horizontal sweep oscillator. This is the input for the transformer which is between the oscillator and the horizontal sweep amplifier tube. This trace is taken with no received signal.

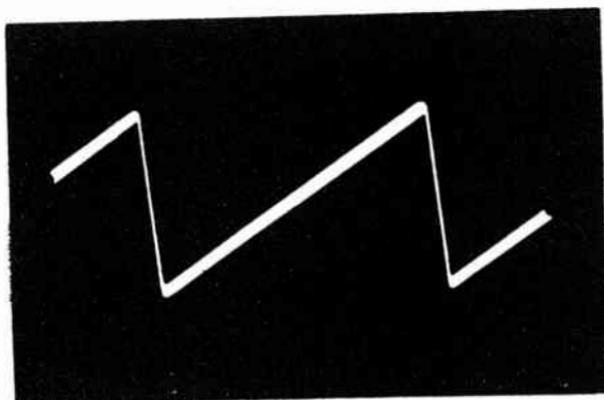


Fig. 139-21.

*Fig. 139-21:* This sawtooth voltage appears at the output of the transformer between oscillator and sweep amplifier. It is fed to the grid-cathode circuit of the sweep amplifier. The trace is taken with no received signal.

## WAVEFORMS

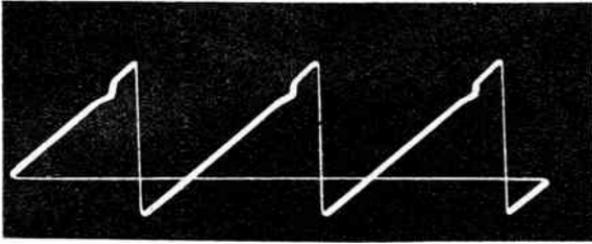


Fig. 139-22.

*Fig. 139-22:* From the grid of the horizontal sweep amplifier or output amplifier. No trace is taken from the plate of this tube because voltage there is too great for application to an oscilloscope without some suitable reducing means such as a high-voltage capacitive voltage divider.

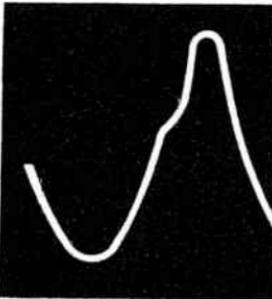


Fig. 139-23.

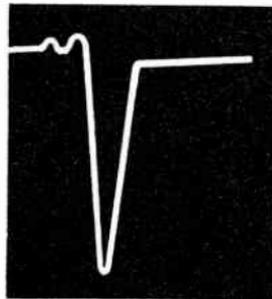


Fig. 139-24.

*Fig. 139-23:* From the cathode of the damper tube. This is the voltage applied to the input of the linearity control, which in this receiver is a variable inductor between the damper cathode and the primary winding of the sweep output transformer. No transmitted signal is being received.

*Fig. 139-24:* This is the voltage pulse appearing in the circuit containing the horizontal deflection coils of the yoke and the secondary winding of the horizontal output transformer. It is taken from the low side of the secondary winding. There is no received signal. A voltage of this form accompanies the sawtooth current in the deflection coils.

**WIRING, PRINTED.** — Printed wiring as used in television receivers consists of copper strips no more than a few thousandths of an inch thick which are permanently attached to a sheet of solid insulating material. Fig. 1 shows both sides of an antenna coupler in which inductors and circuit connections are formed by printed wiring. In other designs small

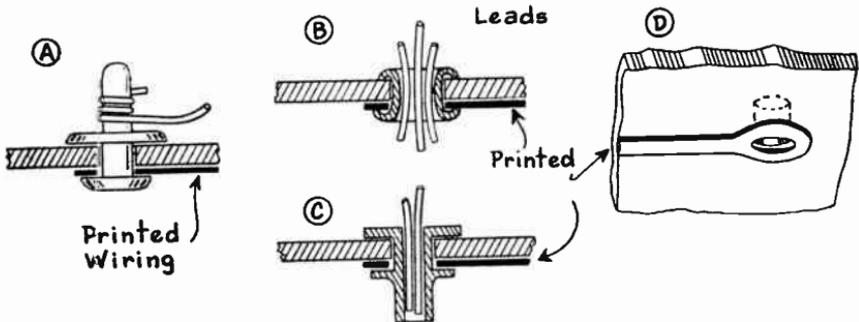


Fig. 2.—Methods of making terminal connections to printed wiring.

capacitors and resistors of the common pigtail types are mounted on the insulating base which carries printed conductors for circuit connections. Printed wiring is used for many small components operating at high frequencies, such as inter-stage coupling transformers, also for entire receiver sections such as i-f amplifiers.

One process for forming printed wiring is carried out as follows: To one entire side of a sheet of suitable plastic insulation is securely bonded a thin layer of copper. Circuit wiring or connections have previously been etched by photo-engraving on what amounts to a printing plate, and on this plate have the form of raised lines. The etched plate is used to print onto the copper a film of protective cement where the circuit lines are raised. Portions of the copper thus protected are a reproduction of the desired circuit connections.

The copper is dissolved or eaten away from all areas not protected by the cement. Then the cement is eaten away in another solution which does not attack copper. The thin copper remaining on the insulation is in the form of desired circuit connections or elements.

It is common practice to mount all resistors, capacitors, tube sockets, and other components on one side of the insulating sheet, with all printed wiring conductors on the other side. To allow for mounting and terminal connections, holes are punched through the insulation either before or after the printing process. Electrical connections may be made through the holes with metal pins as at *A* of Fig. 2, with hollow metal eyelets as at *B* or *C*, by means of rings formed at the terminal ends of printed conductors, as at *D*, or by some combination of these methods. In any case the joints are made electrically complete with solder.

Pigtails of small resistors, capacitors, and inductors may be wound around pin connectors as at *A* of Fig. 2, or passed through eyelets or rings as at *B*, *C*, or *D*. When passed through openings the pigtailed or other leads may be bent over to make the joint mechanically secure before soldering.

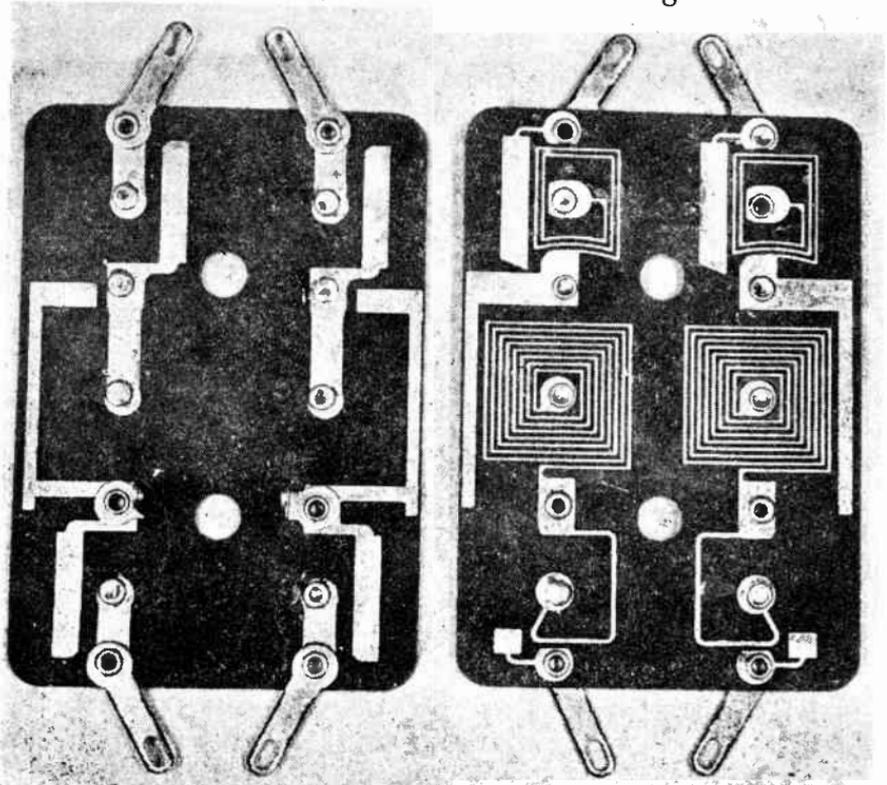


Fig. 1

In some instances, depending on mechanical and electrical requirements, solid rivets are used for connections through the insulation. Such a case is illustrated by Fig. 3, where inductors and connections formed by printed wiring are on one side of channel strips for a turret tuner, while the heads

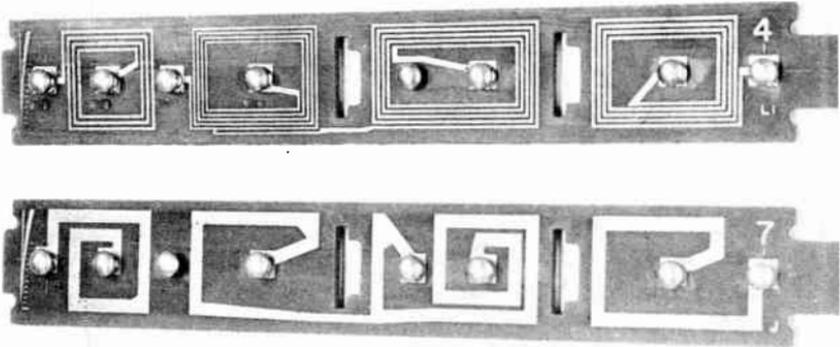


Fig. 3

of rivets on the opposite side engage contact springs mounted on the tuner frame.

After all terminals connections are in place the joints are soldered. During original manufacture, soldering is by the dip process which completes this work on an entire assembly at one operation. The side of the assembly on which the ends of leads are exposed is dipped first into liquid flux. Then this side is dipped into a bath of molten solder. The solder flows into and fills openings through which terminal leads extend, or flows around pins on which leads are wound. After the soldered assembly is cleaned it may be coated with moisture-proof varnish or lacquer, or sometimes with wax.

When a printed wiring assembly includes an entire receiver section; for example, a video amplifier or sound amplifier with demodulator, there may be a ground strip of copper all around the edges of the insulation with extensions inward where needed. Where there should be no coupling between lines for plates, grids, screens, or other circuits a live conductor strip may run alongside a grounded strip or between two grounded strips. Large areas of grounded copper conductor may act as partial electrostatic shielding.

Removal and replacement of defective parts, repair of loose solder joints, reconnection of broken conductor strips, and other service operations must be performed in such manner as to avoid overheating or blistering of the insulating base or separation of printed conductors from this base. The soldering iron or gun should have a small pencil-type tip with no sharp edges or points. Ordinary small-diameter rosin core solder may be used, although a variety having relatively low melting point is preferable.

Joints to be worked on should be freed of lacquer or other coating by using a solvent or by careful scraping. Original solder should be softened by applying the hot iron to pigtailed or other solid wire leads rather than to an eyelet, ring, or pin held in the insulating base. If a lead is bent over on the printed wiring side it is advisable to clip off the bent part before attempting to pull the lead out of the joint.

Printed wiring conductors which are open circuited anywhere between soldered joints at the ends may be repaired by soldering a length of bare or insulated hookup wire to the two joints on opposite sides of the break. If wiring conductors are fairly thick it may be possible to sweat solder a small strip of thin copper over a break, but there is danger of overheating the insulating base and causing additional damage.

Severe overheating of very thin printed wiring conductors, as may result from short circuits or other heavy overloads, may carbonize the base insulation or a moisture-proof coating and allow high-resistance leakage. It may be possible to carefully scrape away the carbonized area. Otherwise this area may be drilled out to leave air space between conductors. If scraping or drilling is difficult because of nearby conductors and circuit elements, one of the conductors from which there is leakage may be opened near its soldered joints and replaced with a jumper of insulated hookup wire.

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