Electrons, Electrodes and Electron Tubes

THE electron tube is a marvelous device. It makes possible the performing of operations, amazing in conception, with a precision and a certainty that are astounding. It is an exceedingly sensitive and accurate instrument—the product of coordinated efforts of engineers and craftsmen. Its construction requires materials from every corner of the earth. Its use is world-wide. Its future possibilities, even in the light of present-day accomplishments, are but dimly foreseen, for each development opens new fields of design and application.

The importance of the electron tube lies in its ability to control almost instantly the flight of the millions of electrons supplied by the cathode. It accomplishes this control with a minimum of energy. Because it is almost instantaneous in its action, the electron tube can operate efficiently and accurately at electrical frequencies much higher than those attainable with rotating machines.

Electrons

All matter exists in the solid, liquid, or gaseous state. These three forms consist entirely of minute divisions known as molecules, which, in turn, are composed of atoms. Atoms have a nucleus which is a positive charge of electricity, around which revolve tiny charges of negative electricity known as **electrons.** Scientists have estimated that electrons weigh only 1/30-billion, billion, billion, billionths of an ounce, and that they may travel at speeds of thousands of miles per second.

Electron movement may be accelerated by the addition of energy. Heat is one form of energy which can be conveniently used to speed up the electron. For example, if the temperature of a metal is gradually raised, the electrons in the metal gain velocity. When the metal becomes hot enough, some electrons may acquire sufficient speed to break away from the surface of the metal. This action, which is accelerated when the metal is heated in a vacuum, is utilized in most electron tubes to produce the necessary electron supply.

An electron tube consists of a cathode, which supplies electrons, and one or more additional electrodes, which control and collect these electrons, mounted in an evacuated envelope. The envelope may be made of glass, metal, ceramic, or a combination of these materials.

Cathodes

A cathode is an essential part of an electron tube because it supplies the electrons necessary for tube operation. When energy in some form is applied to the cathode, electrons are released. Heat is the form of energy generally used. The method of heating the cathode may be used to distinguish between the different forms of cathodes. For example, a directly heated cathode, or filament-cathode, is a wire heated by the passage of an electric current. An indirectly heated cathode, or heater-cathode, consists of a filament, or heater, enclosed in a metal sleeve. The sleeve carries the electron-emitting material on its outside surface and is heated by radiation and conduction from the heater.

A **filament, or directly heated cathode,** such as that shown in **Fig. 1** may be further classified by identifying the filament or electron-emitting material. The materials in regular use are tungsten, thoriated tungsten, and metals which have been coated with alkaline-earth oxides. Tungsten filaments are made from the pure metal. Because they must operate at high temperatures (a dazzling white) to emit sufficient electrons, a relatively large amount of filament power is required.

Thoriated-tungsten filaments are made from tungsten impregnated with thorium oxide. Due to the presence of thorium, these filaments liberate electrons at a more moderate temperature of about 1700°C (a bright yellow) and are, therefore, much more economical of filament power than are pure tungsten filaments.

Alkaline earths are usually applied as a coating on a nickel-alloy wire or ribbon. This coating, which is dried in a relatively thick layer on the filament, requires only a relatively low temperature of about 700-750°C (a dull red) to produce a copious supply of electrons. Coated filaments operate very efficiently and require relatively little filament power. However, each of these cathode materials has special advantages which determine the choice for a particular application.

Directly heated filament-cathodes require comparatively little heating power. They are used in tube types designed for battery operation because it is, of course, desirable to impose as small a drain as possible on the batteries. They are also used in rectifiers such as the 1G3GT/1B3GT and the 5Y3GT.

An **indirectly heated cathode,** or **heater-cathode,** consists of a thin metal sleeve coated with electron-emitting material such as alkaline-earth oxides. The emissive surface of the cathode is maintained at the required temperature (approximately 1050°K) by resistance-heating of a tungsten or tungsten-alloy wire which is placed inside the cathode sleeve and electrically insulated from it, as shown in Fig. 2. The heater is used only for the purpose of heating the cathode sleeve and sleeve coating to an electron-emitting temperature. Useful emission does not take place from the heater wire.

A new dark heater insulating coating developed by RCA has better heat transfer than earlier aluminum-oxide coatings, and makes it possible to operate heaters at lower temperatures for given power inputs. Because the tensile strength of the heater wire increases at the lower operating temperatures, tubes using **dark heaters** have increased reliability, stability, and life.

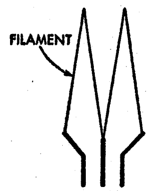

Fig. 1—Filament or directly heated cathode.

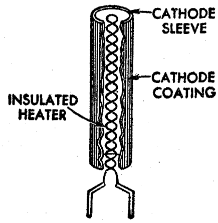

Fig. 2—Indirectly heated cathode or heater-cathode.

The heater-cathode construction is well adapted for use in electron tubes intended for operation from ac power lines and from storage batteries. The use of separate parts for emitter and heater functions, the electrical insulation of the heater from the emitter, and the shielding effect of the sleeve may all be utilized in the design of the tube to minimize the introduction of hum from the ac heater supply and to minimize electrical interference which might enter the tube circuit through the heater-supply line. From the viewpoint of circuit design, the heater-cathode construction offers advantages in connection flexibility because of the electrical separation of the heater from the cathode.

Another advantage of the heater-cathode construction is that it makes practical the design of a rectifier tube having close spacing between its cathode and plate, and of an amplifier tube having close spacing between its cathode and grid. In a close-spaced rectifier tube, the voltage drop in the tube is low, and, therefore, the regulation is improved. In an amplifier tube, the close spacing increases the gain obtainable from the tube. Because of the advantages of the heater-cathode construction, almost all present-day receiving tubes designed for ac operation have heater-cathodes.

Generic Tube Types

Electrons are of no value in an electron tube unless they can be put to work. Therefore, a tube is designed with the parts necessary to utilize electrons as well as those required to produce them. These parts consist of a cathode and one or more supplementary electrodes. The electrodes are enclosed in an evacuated envelope having the necessary connections brought out through air-tight seals. The air is removed from the envelope to allow free movement of the electrons and to prevent injury to the emitting surface of the cathode.

When the cathode is heated, electrons leave the cathode surface and form an invisible cloud in the space around it. Any positive electric potential within the evacuated envelope offers a strong attraction to the electrons (unlike electric charges attract; like charges repel). Such a positive electric potential can be supplied by an **anode** (positive electrode) located within the tube in proximity to the cathode.

Diodes

The simplest form of electron tube contains two electrodes, a cathode and an anode (plate), and is often called a diode, the family name for a two-electrode tube. In a diode, the positive potential is supplied by a suitable electrical source connected between the plate terminal and a cathode terminal, as shown in Fig. 3. Under the influence of the positive plate potential, electrons

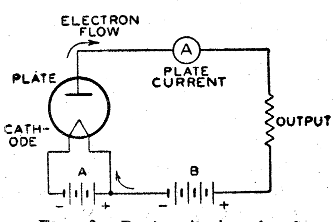

Fig. 3—Basic diode circuit.

flow from the cathode to the plate and return through the external plate-battery circuit to the cathode, thus completing the circuit. This flow of electrons is known as the **plate current.**

If a negative potential is applied to the plate, the free electrons in the space surrounding the cathode will be forced back to the cathode and no plate current will flow. If an alternating voltage is applied to the plate, the plate is alternately made positive and negative. Because plate current flows only during the time when the plate is positive, current flows through the tube in only one direction and is said to be rectified. Fig. 4 shows the rectified output current produced by an alternating input voltage.

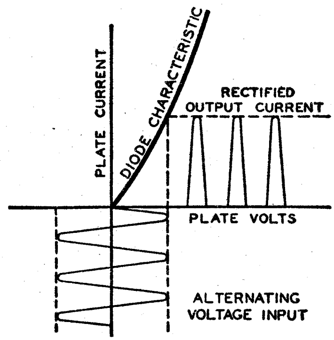

Fig. 4—Current characteristics of rectifier circuit.

Diode rectifiers are used in ac receivers to convert the ac supply voltage to dc voltage for the electrodes of the other tubes in the receiver. Rectifier tubes having only one plate and one cathode, such as the 35W4, are called **half-wave rectifiers,** because current can flow only during one-half of the alternating-current cycle. When two plates and one or more cathodes are used in the same tube, current may be obtained on both halves of the ac cycle. The 6X4, 5Y3GT, and 5U4GB are examples of this type and are called **full-wave rectifiers.**

Not all of the electrons emitted by the cathode reach the plate. Some return to the cathode, while others remain in the space between the cathode and plate for a brief period to produce an effect known as **space charge.** This charge has a repelling action on other electrons which leave the cathode surface and impedes their passage to the plate. The extent of this action and the amount of space charge depend on the cathode temperature, the distance between the cathode and the plate, and the plate potential. The higher the plate potential, the less is the tendency for electrons to remain in the space-charge region and repel other electrons. This effect may be noted by applying increasingly higher plate voltages to a tube operating at a fixed heater or filament voltage. Under these conditions, the maximum number of available electrons is fixed, but increasingly higher plate voltages will succeed in attracting a greater proportion of the free electrons.

Beyond a certain plate voltage, however, additional plate voltage has little effect in increasing the plate current because all of the electrons emitted by the cathode are already being drawn to the plate. This maximum current, illustrated in Fig. 5, is called **saturation current.** Because it is an indication of the total number of electrons emitted, it is also known as **emission current** or simply **emission.**

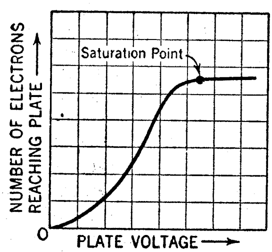

Fig. 5—Current characteristic of diode tube.

Although tubes are sometimes tested by measurement of their emission current, it is generally not advisable to measure the full value of emission because this value would be sufficiently large to cause change in the tube characteristics or even to damage the tube. Consequently, while the test value of emission current is somewhat larger than the maximum current which will be required from the cathode in the use of the tube, it is ordinarily less than the full emission current. The emission test, therefore, is used to indicate whether the cathode can supply a sufficient number of electrons for satisfactory operation of the tube.

If space charge were not present to repel electrons coming from the cathode, the same plate current could be produced at a lower plate voltage. One way to make the effect of space charge small is to make the distance between plate and cathode small. This method is used in rectifier types having heater-cathodes, such as the 5V4GA and the 6AX5GT. In these types the radial distance between cathode and plate is only about two hundredths of an inch.

Another method of reducing space-charge effect is utilized in **mercury-vapor rectifier tubes.** When such tubes are operated, a small amount of mercury contained in the tube is partially vaporized, filling the space inside the bulb with mercury atoms. These atoms are bombarded by electrons on their way to the plate. If the electrons are moving at a sufficiently high speed, the collisions tear off electrons from the mercury atoms. The mercury atom is then said to be **"ionized,"** *i.e.*, it has lost one or more electrons and, therefore, has a positive charge. Ionization is evidenced by a bluish-green glow between the cathode and plate. When ionization occurs, the space charge is neutralized by the positive mercury atoms so that increased numbers of electrons are made available. Mercury-vapor tubes are used primarily for power rectifiers.

Ionic-heated-cathode rectifiers depend on gas ionization for their operation. These tubes are of the full-wave design and contain two anodes and a coated cathode sealed in a bulb containing a reduced pressure of inert gas. The cathode becomes hot during tube operation, but the heating effect is caused by bombardment of the cathode by ions within the tube rather than by heater or filament current from an external source.

The internal structure of an ionic-heated-cathode tube is designed so that when sufficient voltage is applied to the tube, ionization of the gas occurs between the anode which is instantaneously positive and the cathode. Under normal operating voltages, ionization does not take place between the anode that is negative and the cathode, so that the requirements for rectification are satisfied. The initial small flow of current through the tube is sufficient to raise the cathode temperature quickly to incandescence, whereupon the cathode emits electrons. The voltage drop in such tubes is slightly higher than that of the usual hot-cathode gas rectifiers because energy is taken from the ionization discharge to keep the cathode at operating temperature. Proper operation of these rectifiers requires a minimum flow of load current at all times to maintain the cathode at the temperature required to supply sufficient emission.

Triodes

When a third electrode, called the **grid,** is placed between the cathode and plate, the tube is known as a triode, the family name for a three-electrode tube. The grid usually consists of relatively fine wire wound on two support rods (siderods) and extending the length of the cathode. The spacing between turns of wire is large compared with the size of the wire so that the passage of electrons from cathode to plate is practically unobstructed by the grid. In some types, a **frame grid** is used. The frame consists of two siderods supported by four metal straps. Extremely fine lateral wire (diameter of 0.5 mil or less) is wound under tension around the frame. This type of grid permits the use of closer spacings between grid wires and between tube electrodes, and thus improves tube performance.

The purpose of the grid is to control the flow of plate current. When a tube is used as an amplifier, a negative dc voltage is usually applied to the grid. Under this conditon the grid does not draw appreciable current.

The number of electrons attracted to the plate depends on the combined effect of the grid and plate polarities, as shown in Fig. 6. When the plate is positive, as is normal, and the dc grid voltage is made more and more negative, the plate is less able to attract electrons to it and plate current decreases. When the grid is made less and less negative (more and more positive), the plate more readily attracts electrons to it and plate current increases. Hence, when the voltage on the grid is varied in accordance with a signal, the plate current varies with the signal. Because a small voltage applied to the grid can control a comparatively large amount of plate current, the signal is amplified by the tube. Typical three-electrode tube types are the 6C4 and 6AF4A.

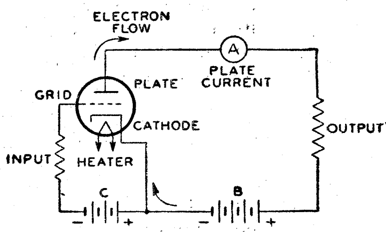

Fig. 6—Basic triode circuit.

The grid, plate, and cathode of a triode form an electrostatic system, each electrode acting as one plate of a small capacitor. The capacitances are those existing between grid and plate, plate and cathode, and grid and cathode. These capacitances are known as **interelectrode capacitances.** Generally, the capacitance between grid and plate is of the most importance. In high-gain radio-frequency amplifier circuits, this capacitance may act to produce undesired coupling between the **input circuit,** the circuit between grid and cathode, and the **output circuit,** the circuit between plate and cathode. This coupling is undesirable in an amplifier because it may cause instability and unsatisfactory performance.

Tetrodes

The capacitance between grid and plate can be made small by mounting an additional electrode, called the **screen grid** (grid No. 2), in the tube. With the addition of the grid No. 2, the tube has four electrodes and is, accordingly, called a tetrode. The screen grid or grid No. 2 is mounted between the grid No. 1 **(control grid)** and the plate, as shown in Fig. 7, and acts as an electrostatic shield between them, thus reducing the grid-to-plate capacitance. The effectiveness of this shielding action is increased by a bypass

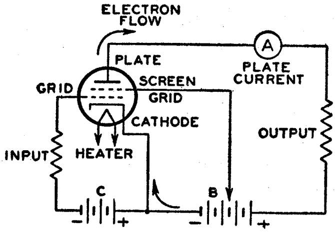

Fig. 7—Basic tetrode circuit.

capacitor connected between screen grid and cathode. By means of the screen grid and this bypass capacitor, the grid-plate capacitance of a tetrode is made very small. In practice, the grid-plate capacitance is reduced from several picofarads (pF) for a triode to 0.01 pF or less for a screen-grid tube.

The screen grid has another desirable effect in that it makes plate current practically independent of plate voltage over a certain range. The screen grid is operated at a positive voltage and, therefore, attracts electrons from the cathode. However, because of the comparatively large space between wires of the screen grid, most of the electrons drawn to the screen grid pass through it to the plate. Hence the screen grid supplies an electrostatic force pulling electrons from the cathode to the plate. At the same time the screen grid shields the electrons between cathode and screen grid from the plate so that the plate exerts very little electrostatic force on electrons near the cathode.

So long as the plate voltage is higher than the screen-grid voltage, plate current in a screen-grid tube depends to a great degree on the screen-grid voltage and very little on the plate voltage. The fact that plate current in a screen-grid tube is largely independent of plate voltage makes it possible to obtain much higher amplification with a tetrode than with a triode. The low grid-plate capacitance makes it possible to obtain this high amplification without plate-to-grid feedback and resultant instability. In receiving-tube applications, the tetrode has been replaced to a considerable degree by the pentode.

Pentodes

In all electron tubes, electrons striking the plate may, if moving at sufficient speed, dislodge other electrons. In two- and three-electrode types, these dislodged electrons usually do not cause trouble because no positive electrode other than the plate itself is present to attract them. These electrons, therefore, are drawn back to the plate. Emission caused by bombardment of an electrode by electrons from the cathode is called **secondary emission** because the effect is secondary to the original cathode emission.

In the case of screen-grid tubes, the proximity of the positive screen grid to the plate offers a strong attraction to these secondary electrons, and particularly so if the plate voltage swings lower than the screen-grid voltage. This effect reduces the plate current and limits the useful plate-voltage swing for tetrodes.

The effects of secondary emission are minimized when a fifth electrode is placed within the tube between the screen grid and plate. This fifth electrode is known as the **suppressor grid** (grid No. 3) and is usually connected to the cathode, as shown in Fig. 8. Because of its negative potential with respect to the plate, the suppressor grid retards the flight of secondary electrons and diverts them back to the plate.

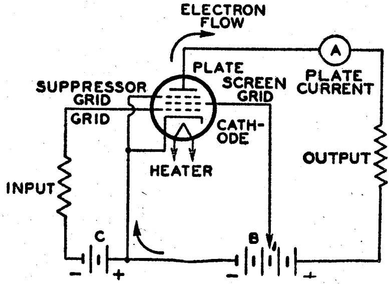

Fig. 8—Basic pentode circuit.

The family name for a five-electrode tube is "pentode." In power-output pentodes, the suppressor grid makes possible higher power output with lower grid-driving voltage; in radio-frequency amplifier pentodes, the suppressor grid makes possible high voltage amplification at moderate values of plate voltage. These desirable features result from the fact that the plate-voltage swing can be made very large. In fact, the plate voltage may be as low as, or lower than, the screen-grid voltage without serious loss in signal-gain capability. Representative pentodes used for power amplification are the 6CL6 and 6K6GT; representative pentodes used for voltage amplification are the 6AU6A, 6BA6, and 5879.

Beam Power Tubes

A beam power tube is a tetrode or pentode in which directed electron beams are used to increase substantially the power-handling capability of the tube. Such a tube contains a cathode, a control grid (grid No. 1), a screen grid (grid No. 2), a plate, and, optionally, a suppressor grid (grid No. 3). When a beam power tube is designed without an actual suppressor grid, the electrodes are so spaced that secondary emission from the plate is suppressed by space-charge effects between screen grid and plate. The space charge is produced by the slowing up of electrons traveling from a high-potential screen grid to a lower-potential plate. In this low-velocity region, the space charge produced is sufficient to repel secondary electrons emitted from the plate and to cause them to return to the plate.

Beam power tubes of this design employ beam-confining electrodes at cathode potential to assist in producing the desired beam effects and to prevent stray electrons from the plate from returning to the screen grid outside of the beam. A feature of a beam power tube is its low screen-grid current. The screen grid and the control grid are spiral wires wound so that each turn of the screen grid is shaded from the cathode by a grid turn. This alignment of the screen grid and control grid causes the electrons to travel in sheets between the turns of the screen grid so that very few of them strike the screen grid. Because of the effective suppressor action provided by space charge and because of the low current drawn by the screen grid, the beam power tube has the advantages of high power output, high power sensitivity, and high efficiency.

Fig. 9 shows the structure of a beam power tube employing space-charge suppression and illustrates how

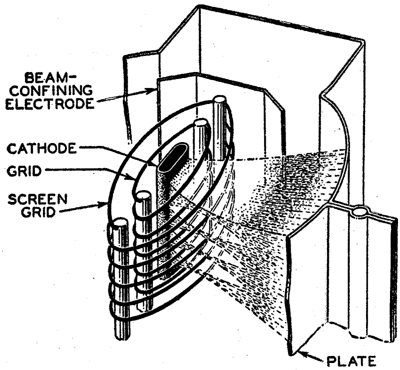

Fig. 9—Structure of beam power tube showing beam-confining action.

the electrons are confined to beams. The beam condition illustrated is that for a plate potential less than the screen-grid potential. The high-density space-charge region is indicated by the heavily dashed lines in the beam. Note that the edges of the beam-confining electrodes coincide with the dashed portion of the beam. In this way the space-charge potential region is extended beyond the beam boundaries and stray secondary electrons are prevented from returning to the screen grid outside of the beam. The space-charge effect may also be obtained by use of an actual suppressor grid. Examples of beam power tubes are 6AQ5A, 6L6GC, 6V6GTA, and 50C5.

Multi-Electrode and Multi-Unit Tubes

Early in the history of tube development and application, tubes were designed for a general service; that is, a single tube type—a triode—was used as a radio-frequency amplifier, an intermediate-frequency amplifier, an audio-frequency amplifier, an oscillator, or a detector. Obviously, with this diversity of application, one tube did not meet all requirements to the best advantage.

Later and present trends of tube design are the development of "specialty" types. These types are intended either to give optimum performance in a particular application or to combine in one bulb functions which formerly required two or more tubes. The first class of tubes includes such examples of specialty types as the 6CB6A and 6BY6. Types of this class generally require more than three electrodes to obtain the desired special characteristics and may be broadly classed as multi-electrode types. The 6BY6 is an especially interesting type in this class. This tube has an unusually large number of electrodes, namely seven, exclusive of the heater. Plate current in the tube is varied at two different frequencies at the same time. The tube is designed primarily for use as a combined sync separator and sync clipper in television receivers.

The second class includes multi-unit tubes such as the twin-diode triodes 6CN7 and 6AV6, as well as triode-pentodes such as the 6U8A and 6X8. This class also includes class A twin triodes such as the 6CG7 and 12AX7A, and types such as the 6CM7 containing dissimilar triode units used primarily as combined vertical oscillators and vertical deflection amplifiers in television receivers. Full-wave rectifiers are also multi-unit types.

A third class of tubes combines features of each of the other two classes. Typical of this third class are the pentagrid-converter types 6BE6 and 6SA7. These tubes are similar to the multi-electrode types in that they have seven electrodes, all of which affect the electron stream; and they are similar to the multi-unit tubes in that they perform simultaneously the double function of oscillator and mixer in superheterodyne receivers.

Receiving Tube Structure

Receiving tubes generally utilize a glass or metal envelope and a **base.** Originally, the base was made of metal or molded phenolic material. Types having a glass envelope and a molded phenolic base include the **"octal"** types such as the 5U4GB and the 6SN7GTB. Types having a metal envelope and molded phenolic octal base include the 6F6 and the 6L6. Many modern types utilize integral glass bases. Present-day conventional tube designs utilizing glass envelopes and integral glass bases include the seven-pin and nine-pin **miniature** types, the nine-pin **novar** and **neonoval** types, and the twelve-pin **duodecar** types. Examples of the seven-pin miniature types are the 6AU6A and 6BN6. Examples of the nine-pin miniature types are the 12AU7A and 6EA8. Examples of the novar types are the 6BH3 and 7868. The nine-pin base for the novar types has a relatively large pin-circle diameter and long pins to insure firm retention of the tube in its socket.

The **nuvistor** concept provided a new approach to electron tube design. Nuvistor tubes utilize a light-weight cantilever-supported cyclindrical electrode structure housed in a ceramic-metal envelope. These tubes combine new materials, processes, and fabrication techniques. Examples of the nuvistor are the 6CW4 and the 6DV4.

Television Picture Tubes

The picture tube, or kinescope, is a multi-electrode tube used principally in television receivers for picture display. It consists essentially of an electron gun, a glass or metal-and-glass envelope and face-plate combination, and a fluorescent screen.

The electron gun includes a cathode for the production of free electrons, one or more control electrodes for accelerating the electrons in the beam, and, optionally, a device for "trapping" unwanted ions out of the electron beam.

Focusing of the beam is accomplished either electromagnetically by means of a focusing coil placed on the neck of the tube, or electrostatically, as shown in Fig. 10, by means of a focusing electrode (grid No. 4) within the envelope of the tube. The screen is a white-fluorescing phosphor P4 of either the silicate or the sulfide type.

Deflection of the beam is accomplished either electrostatically by means of deflecting electrodes within the envelope of the tube, or electromagnetically by means of a deflecting yoke placed on the neck of the tube. Fig. 10 shows the structure of the gun section of a picture tube and illustrates how the electron beam is formed and how the beam is deflected by means of an electromagnetic deflecting yoke. In this type of tube, ions in the beam are prevented from damaging the fluorescent screen by an aluminum film on the gun side of the screen. This film not only "traps" unwanted ions, but also improves picture contrast. In many types of non-aluminized tubes, ions are separated from the electron beam by means of a tilted-gun and ion-trap-magnet arrangement.

Color television picture tubes are similar to black-and-white picture tubes, but differ in three major ways: (1) The light-emitting screen is made up of trios of phosphor dots deposited in an interlaced pattern. Each dot of a trio is capable of emitting light in one of the three primary colors (red, green, or blue). (2) A shadow mask mounted near the screen of the tube contains over 300,000 apertures, one for each of the phosphor dot trios. This mask provides color separation by shadowing two of the three phosphor dots of each trio. (3) Three closely spaced electron guns, built as a unit, provide separate beams for excitation of the three different color-phosphor-dot arrays. Thus it is possible to control the brightness of each of the three colors independently of the other two. Fig. 11 shows a cutaway view of a color television picture tube.

The three electron guns are mounted with their axes tilted toward the central axis of the envelope, and are spaced 120 degrees with respect to each other. The focusing electrodes of the three guns are interconnected internally, and their potential is adjusted to cause the separate beams to focus at the phosphor-dot screen. All three beams must be made to converge at the screen while they are simultaneously being deflected. Convergence is accomplished by the action of static and

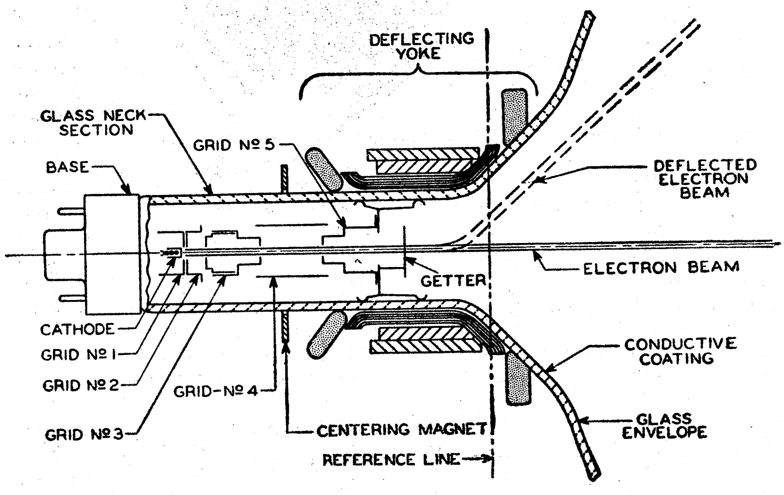

Fig. 10—Structure of television-picture-tube electron gun.

dynamic magnetic fields set up by the radial-converging magnet assembly mounted on the neck of the tube. These fields are coupled into the radial-converging pole pieces within the tube. Another pair of pole pieces in the tube is activated by the lateral-converging magnet also mounted on the neck of the tube. These pole pieces permit lateral shift in position of the blue beam in opposition to the lateral shift of the green and red beams.

A purifying magnet is used with color picture tubes to provide a magnetic field, adjustable in magnitude and direction, to effect register over the entire area of the screen. A magnetic shield is used to minimize the effects of the earth's magnetic field.

Deflection of the three beams is accomplished simultaneously by a deflecting yoke using four electromagnetic coils similar to the deflec'ng yoke used for black-and-white picture tubes.

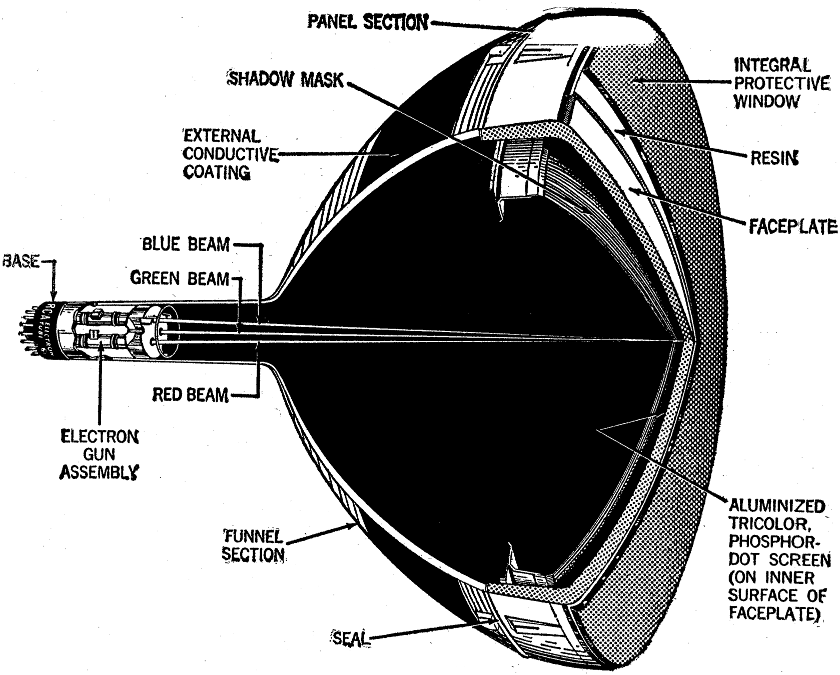

Fig. 11—Cutaway view of color television picture tube.

Electron Tube Characteristics

THE term "characteristics" is used to identify the distinguishing electrical features and values of an electron tube. These values may be shown in curve form or they may be tabulated. When the characteristics values are given in curve form, the curves may be used for the determination of tube performance and the calculation of additional tube factors.

Tube characteristics are obtained from electrical measurements of a tube in various circuits under certain definite conditions of voltages. Characteristics may be further described by denoting the conditions of measurements. For example, Static Characteristics are the values obtained with different dc potentials applied to the tube electrodes, while Dynamic Characteristics are the values obtained with an ac voltage on a control grid under various conditions of dc potentials on the electrodes. The dynamic characteristics, therefore, are indicative of the performance capabilities of a tube under actual working conditions.

Static characteristics may be shown by plate characteristics curves and transfer (mutual) characteristics curves. These curves present the same information, but in two different forms to increase its usefulness. The plate characteristic curve is obtained by varying plate voltage and measuring plate current for different grid-bias voltages, while the transfer-characteristic curve is obtained by varying grid-bias voltage and measuring plate current for different plate voltages. A plate-characteristic family of curves is shown in Fig. 12. Fig. 13 gives the transfer-characteristic family of curves for the same tube.

Dynamic characteristics include amplification factor, plate resistance, control-grid—plate transconductance, and certain detector characteristics, and may be shown in curve form for variations in tube operating conditions.

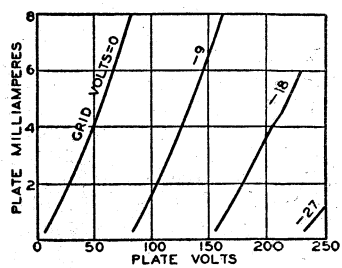

Fig. 12—Family of plate-characteristics curves.

The **amplification factor,** or μ, is the ratio of the change in plate voltage to a change in control-electrode voltage in the opposite direction, under the condition that the plate current remains

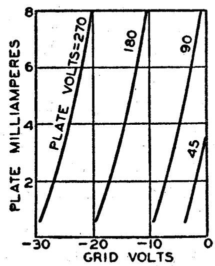

Fig. 13—Family of transfer-characteristics curves.

unchanged and that all other electrode voltages are maintained constant. For example, if, when the plate voltage is made 1 volt more positive, the control-electrode (grid-No. 1) voltage must be made 0.1 volt more negative to hold plate current unchanged, the amplification factor is 1 divided by 0.1, or 10. In other words, a small voltage variation in the grid circuit of a tube has the same effect on the plate current as a large plate-voltage change—the latter equal to the product of the grid-voltage change and amplification factor. The μ of a tube is often useful for calculating stage gain. This use is discussed in the **Electron Tube Applications** section.

Plate resistance (r_p) of an electron tube is the resistance of the path between cathode and plate to the flow of alternating current. It is the quotient of a small change in plate voltage divided by the corresponding change in plate current and is expressed in ohms, the unit of resistance. Thus, if a change of 0.1 milliampere (0.0001 ampere) is produced by a plate-voltage variation of 1 volt, the plate resistance is 1 divided by 0.0001, or 10000 ohms.

Control-grid—plate transconductance, or simply **transconductance** (g_m), is a factor which combines in one term the amplification factor and the plate resistance, and is the quotient of the first divided by the second. This term has also been known as mutual conductance. Transconductance may be more strictly defined as the quotient of a small change in plate current (amperes) divided by the small change in the control-grid voltage producing it, under the condition that all other voltages remain unchanged. Thus, if a grid-voltage change of 0.5 volt causes a plate-current change of 1 milliampere (0.001 ampere), with all other voltages constant, the transconductance is 0.001 divided by 0.5, or 0.002 mho. A "mho" is the unit of conductance and was named by spelling ohm backwards. For convenience, a millionth of a mho, or a micromho (μmho), is used to express transconductance. Thus, in the example, 0.002 mho is 2000 micromhos.

Conversion transconductance (g_c) is a characteristic associated with the mixer (first detector) function of tubes and may be defined as the quotient of the intermediate-frequency (if) current in the primary of the if transformer divided by the applied radio-frequency (rf) voltage producing it; more precisely, it is the limiting value of this quotient as the rf voltage and if current approach zero. When the performance of a frequency converter is determined, conversion transconductance is used in the same way as control-grid—plate transconductance is used in single-frequency amplifier computations.

The **plate efficiency** of a power amplifier tube is the ratio of the ac power output (P_o) to the product of the average dc plate voltage (E_b) and dc plate current (I_b) at full signal, or

$$\text{Plate efficiency } \% = \frac{P_o \text{ watts}}{E_b \text{ volts} \times I_b \text{ amperes}} \times 100$$

The **power sensitivity** of a tube is the ratio of the power output to the square of the input signal voltage (E_{in}), and is expressed in mhos as follows:

$$\text{Power sensitivity (mhos)} = \frac{P_o \text{ watts}}{(E_{in}, \text{ rms})^2}$$

Electron Tube Applications

THE diversified applications of an electron receiving tube have, within the scope of this section, been treated under seven headings. These are: Amplification, Rectification, Detection, Automatic Volume or Gain Control, Oscillation, Frequency Conversion, and Automatic Frequency Control. Although these operations may take place at either radio or audio frequencies and may involve the use of different circuits and different supplemental parts, the general considerations of each kind of operation are basic.

Amplification

The amplifying action of an electron tube was mentioned under **Triodes** in the section on **Electrons, Electrodes, and Electron Tubes.** This action can be utilized in electronic circuits in a number of ways, depending upon the results desired. Four classes of amplifier service recognized by engineers are covered by definitions standardized by the Institute of Radio Engineers (now the Institute of Electrical and Electronics Engineers). This classification depends primarily on the fraction of input cycle during which plate current is expected to flow under rated full-load conditions. The classes are class A, class AB, class B, and class C. The term "cutoff bias" used in these definitions is the value of grid bias at which plate current is very small.

Classes of Service

A **class A amplifier** is an amplifier in which the grid bias and alternating grid voltages are such that plate current in a specific tube flows at all times.

A **class AB amplifier** is an amplifier in which the grid bias and alternating grid voltages are such that plate current in a specific tube flows for appreciably more than half but less than the entire electrical cycle.

A **class B amplifier** is an amplifier in which the grid bias is approximately equal to the cutoff value, so that the plate current is approximately zero when no exciting grid voltage is applied, and so that plate current in a specific tube flows for approximately one-half of each cycle when an alternating grid voltage is applied.

A **class C amplifier** is an amplifier in which the grid bias is appreciably greater than the cutoff value, so that the plate current in each tube is zero when no alternating grid voltage is applied, and so that plate current flows in a specific tube for appreciably less than one-half of each cycle when an alternating grid voltage is applied.

The suffix 1 may be added to the letter or letters of the class identification to denote that grid current does not flow during any part of the input cycle. The suffix 2 may be used to denote that grid current flows during part of the cycle.

For radio-frequency (rf) amplifiers which operate into a selective tuned circuit, as in radio transmitter applications, or under requirements where distortion is not an important factor, any of the above classes of amplifiers may be used, either with a single tube or with a push-pull stage. For audio-frequency (af) amplifiers in which distortion is an important factor, only class A amplifiers permit single-tube operation. In this case, operating conditions are usually chosen so that distortion is kept below the conventional 5 per cent for triodes and the conventional 7 to 10 per cent for tetrodes or pentodes. Distortion can be reduced below these figures by means of special circuit arrangements such as that discussed under **inverse feedback.** With class A amplifiers, reduced distortion with improved power performance can be obtained by using a push-pull stage for audio service. With class AB and class B amplifiers, a balanced stage using two tubes is required for audio service.

Class A Voltage Amplifiers

As a class A voltage amplifier, an electron tube is used to reproduce grid-voltage variations across an impedance or a resistance in the plate circuit. These variations are essentially of the same form as the input signal voltage impressed on the grid, but their amplitude is increased. This increase is accomplished by operation of the tube at a suitable grid bias so that the applied grid input voltage produces plate-current variations proportional to the signal swings. Because the voltage variation obtained in the plate circuit is much larger than that required to swing the grid, amplification of the signal is obtained.

Fig. 14 gives a graphical illustration of this method of amplication and shows, by means of the grid-voltage vs. plate-current characteristics curve, the effect of an input signal (S) applied to the grid of a tube. The output signal (O) is the resulting amplified plate-current variation.

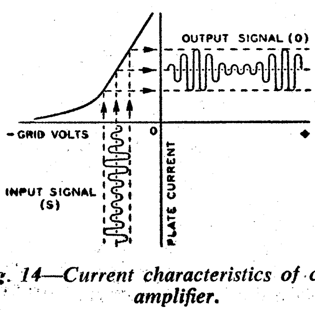

Fig. 14—Current characteristics of class A amplifier.

The plate current flowing through the load resistance (R) of Fig. 15 causes a voltage drop which varies directly with the plate current. The ratio of this voltage variation produced in the load resistance to the input signal voltage is

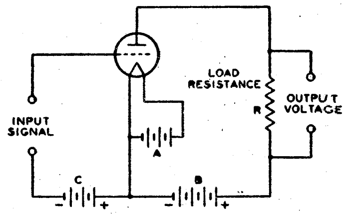

Fig. 15—Triode amplifier circuit.

the voltage amplification, or **gain,** provided by the tube. The voltage amplification due to the tube is expressed by the following convenient formulas:

$$\text{Voltage amplification} = \frac{\mu \times R_L}{R_L + r_p}$$

$$\text{or} \quad \frac{g_m \times r_p \times R_L}{1000000 \times (r_p + R_L)}$$

where μ is the amplification factor of the tube, R_L is the load resistance in ohms, r_p is the plate resistance in ohms, and g_m is the transconductance in micromhos.

From the first formula, it can be seen that the gain actually obtainable from the tube is less than the tube amplification factor, but that the gain approaches the amplification factor when the load resistance is large compared to the tube plate resistance. Fig. 16 shows graphically how the gain approaches the amplification factor of the tube as the load resistance is increased. From the curve it can be seen that a high value of load resistance should be used to obtain high gain in a voltage amplifier.

In a **resistance-coupled amplifier,** the load resistance of the tube is approximately equal to the resistance of the plate resistor in parallel with the grid resistor of the following stage. Hence, to obtain a large value of load resistance, it is necessary to use a plate resistor and a grid resistor of large

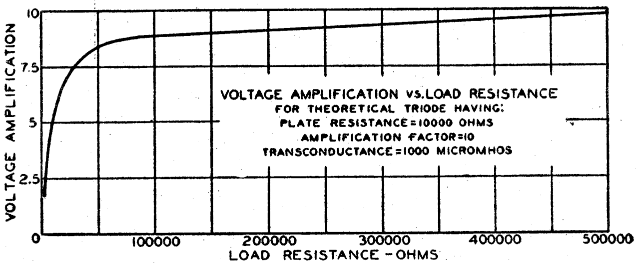

Fig. 16—Gain curve for triode amplifier circuit.

resistance. However, the plate resistor should not be too large because the flow of plate current through the plate resistor produces a voltage drop which reduces the plate voltage applied to the tube. If the plate resistor is too large, this drop will be too large, the plate voltage on the tube will be too small, and the voltage output of the tube will be too small. Also, the grid resistor of the following stage should not be too large, the actual maximum value being dependent on the particular tube type. This precaution is necessary because all tubes contain minute amounts of residual gas which cause a minute flow of current through the grid resistor. If the grid resistor is too large, the positive bias developed by the flow of this current through the resistor decreases the normal negative bias and produces an increase in the plate current. This increased current may overheat the tube and cause liberation of more gas which, in turn, will cause further decrease in bias. The action is cumulative and results in a runaway condition which can destroy the tube.

A higher value of grid resistance is permissible when cathode-resistor bias is used than when fixed bias is used. When cathode-resistor bias is used, a loss in bias due to gas or grid-emission effects is almost completely offset by an increase in bias due to the voltage drop across the cathode resistor. Typical values of plate resistor and grid resistor for tube types used in resistance-coupled circuits, and the values of gain obtainable, are shown in the **Resistance-Coupled Amplifier** section.

The **input impedance** of an electron tube (that is, the impedance between grid and cathode) consists of (1) a reactive component due to the capacitance between grid and cathode, (2) a resistive component resulting from the time of transit of electrons between cathode and grid, and (3) a resistive component developed by the part of the cathode lead inductance which is common to both the input and output circuits. Components (2) and (3) are dependent on the frequency of the incoming signal. The input impedance is very high at audio frequencies when a tube is operated with its grid biased negative. In a class A_1 or AB_1 transformer-coupled audio amplifier, therefore, the loading imposed by the grid on the input transformer is negligible. As a result, the secondary impedance of a class A_1 or class AB_1 input transformer can be made very high because the choice is not limited by the input impedance of the tube; however, transformer design considerations may limit the choice.

At the higher radio frequencies, the input impedance may become very low even when the grid is negative, due to the finite time of passage of electrons between cathode and grid and to the appreciable lead reactance. This impedance drops very rapidly as the frequency is raised, and increases input-circuit loading. In fact, the input impedance may become low enough at very high radio frequencies to affect the gain and selectivity of a preceding stage appreciably. Tubes such as the "acorn" and "pencil" types and the high-frequency miniatures have been developed to have low input capacitances, low electron-transit time, and low lead inductance so that their input impedance is high even at the ultra-high radio frequencies. **Input admittance** is the reciprocal of input impedance.

A **remote-cutoff amplifier** tube is a modified construction of a pentode or a tetrode type designed to reduce modulation-distortion and cross-modulation in radio-frequency stages. **Cross-modulation** is the effect produced in a radio or television receiver by an interfering station "riding through" on the carrier of the station to which the receiver is tuned. **Modulation-distortion** is a distortion of the modulated carrier and appears as audio-frequency distortion in the output. This effect is produced by a radio-frequency amplifier stage operating on an excessively curved characteristic when the grid bias has been increased to reduce volume. The offending stage for cross-modulation is usually the first radio-frequency amplifier, while for modulation-distortion the cause is usually the last intermediate-frequency stage. The characteristics of remote-cutoff types are such as to enable them to handle both large and small input signals with minimum distortion over a wide range of signal strength.

Fig. 17 illustrates the construction of the grid No. 1 (control grid) in a remote-cutoff tube. The remote-cutoff

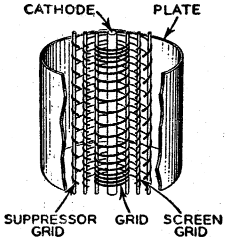

Fig. 17—Structure of remote-cutoff grid.

action is due to the structure of the grid which provides a variation in amplification factor with change in grid bias. The grid No. 1 is wound with open spacing at the middle and with close spacing at the ends. When weak signals and low grid bias are applied to the tube, the effect of the non-uniform turn spacing of the grid on cathode emission and tube characteristics is essentially the same as for uniform spacing. As the grid bias is made more negative to handle larger input signals, the electron flow from the sections of the cathode enclosed by the ends of the grid is cut off. The plate current and other tube characteristics are then dependent on the electron flow through the open section of the grid. This action changes the gain of the tube so that large signals may be handled with minimum distortion due to cross-modulation and modulation-distortion.

Fig. 18 shows a typical plate-current vs. grid-voltage curve for a remote-cutoff type compared with the curve

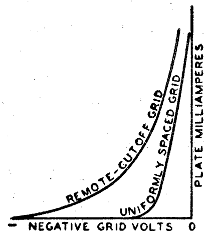

Fig. 18—Plate-current curves for triodes having remote-cutoff and uniformly spaced grids.

for a type having a uniformly spaced grid. It will be noted that while the curves are similar at small grid-bias voltages, the plate current of the remote-cutoff tube drops quite slowly with large values of bias voltage. This slow change makes it possible for the tube to handle large signals satisfactorily. Because remote-cutoff types can accommodate large and small signals, they are particularly suitable for use in sets having automatic volume control. Remote-cutoff tubes also are known as **variable-mu** types.

Class A Power Amplifiers

As a class A power amplifier, an electron tube is used in the output stage of a radio or television receiver to supply a relatively large amount of power to the loudspeaker. For this application, large power output is of more importance than high voltage amplification; therefore, gain possibilities are sacrificed in the design of power tubes to obtain power-handling capability.

Triodes, pentodes, and beam power tubes designed for power amplifier service have certain inherent features for each structure. Power tubes of the triode type for class A service are characterized by low power sensitivity, low plate-power efficiency, and low distortion. Power tubes of the pentode type are characterized by high power sensitivity, high plate-power efficiency and, usually, somewhat higher distortion than class A triodes. Beam power tubes have higher power sensitivity and efficiency than triode or conventional pentode types.

A class A power amplifier is also used as a driver to supply power to a class AB_2 or a class B stage. It is usually advisable to use a triode, rather than a pentode, in a driver stage because of the lower plate impedance of the triode.

Power tubes connected in either **parallel** or **push-pull** may be employed as class A amplifiers to obtain increased output. The parallel connection (Fig. 19 provides twice the output of a single tube with the same value of grid-signal voltage. With this connection,

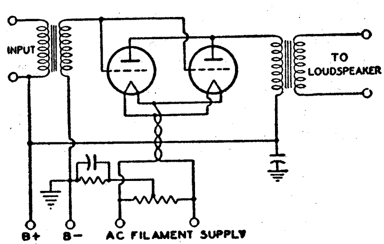

Fig. 19—Power amplifier with tubes connected in parallel.

the effective transconductance of the stage is doubled, and the effective plate resistance and the load resistance required are halved as compared with single-tube values.

The push-pull connection (Fig. 20), although it requires twice the grid-signal voltage, provides increased power and has other important advantages

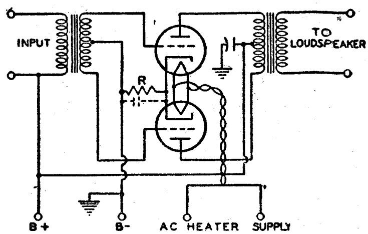

Fig. 20—Power amplifier with tubes connected in push-pull.

over single-tube operation. Distortion caused by even-order harmonics and hum caused by plate-voltage-supply fluctuations are either eliminated or decidedly reduced through cancellation. Because distortion for push-pull operation is less than for single-tube operation, appreciably more than twice single-tube output can be obtained with triodes by decreasing the load resistance for the stage to a value approaching the load resistance for a single tube.

For either parallel or push-pull class A operation of two tubes, all electrode currents are doubled while all dc electrode voltages remain the same as for single-tube operation. If a cathode resistor is used, its value should be about one-half that for a single tube. If oscillations occur with either type of connection, they can often be eliminated by the use of a non-inductive resistor of approximately 100 ohms connected in series with each grid at the socket terminal.

Operation of power tubes so that the grids run positive is inadvisable except under conditions such as those discussed in this section for class AB and class B amplifiers.

Power-Output Calculations

Calculation of the power output of a triode used as a class A amplifier with either an output transformer or a choke having low dc resistance can be made without serious error from the plate family of curves by assuming a resistance load. The proper plate current, grid bias, optimum load resistance, and per-cent second-harmonic distortion can also be determined. The calculations are made graphically and are illustrated in Fig. 21 for given conditions. The procedure is as follows:

(1) Locate the zero-signal bias point P by determining the zero-signal bias Ec_o from the formula:

$$\text{Zero-signal bias } (Ec_o) = -(0.68 \times E_b)/\mu$$

where E_b is the chosen value in volts of dc plate voltage at which the tube is to be operated, and μ is the amplification factor of the tube. This quantity is shown as negative to indicate that a negative bias is used.

(2) Locate the value of zero-signal plate current, I_o, corresponding to point P.

(3) Locate the point $2I_o$, which is twice the value of I_o and corresponds to the value of the maximum-signal plate current I_{max}.

(4) Locate the point X on the dc bias curve at zero volts, $E_c = 0$, corresponding to the value of I_{max}.

(5) Draw a straight line XY through X and P.

Line XY is known as the load resistance line. Its slope corresponds to the value of the load resistance. The load resistance in ohms is equal to $(E_{max} - E_{min})$ divided by $(I_{max} - I_{min})$, where E is in volts and I is in amperes.

It should be noted that in the case of filament types of tubes, the calculations are given on the basis of a dc-operated filament. When the filament is ac-operated, the calculated value of dc bias should be increased by approximately one-half the filament voltage rating of the tube.

The value of zero-signal plate current I_o should be used to determine the plate dissipation, an important factor influencing tube life. In a class A amplifier under zero-signal conditions, the plate dissipation is equal to the power input, *i.e.,* the product of the dc plate voltage E_o and the zero-signal dc plate current I_o. If it is found that the plate-dissipation rating of the tube is exceeded with the zero-signal bias Ec_o calculated above, it will be necessary to increase the bias by a sufficient amount so that the actual plate dissipation does not exceed the rating before proceeding further with the remaining calculations.

For power-output calculations, it is assumed that the peak alternating grid voltage is sufficient (1) to swing the grid from the zero-signal bias value Ec_o to zero bias ($E_c = 0$) on the positive swing and (2) to swing the grid to a value twice the zero-signal bias value on the negative swing. During the negative swing, the plate voltage and plate current reach values of E_{max} and I_{min}; during the positive swing, they reach values of E_{min} and I_{max}. Because power is the product of voltage and current, the power output P_o as shown by a watt-meter is given by

$$P_o = \frac{(I_{max} - I_{min}) \times (E_{max} - E_{min})}{8}$$

where E is in volts, I is in amperes, and P_o is in watts.

In the output of power-amplifier triodes, some distortion is present. This distortion is due predominantly to second harmonics in single-tube amplifiers. The percentage of second-harmonic distortion may be calculated by the following formula:

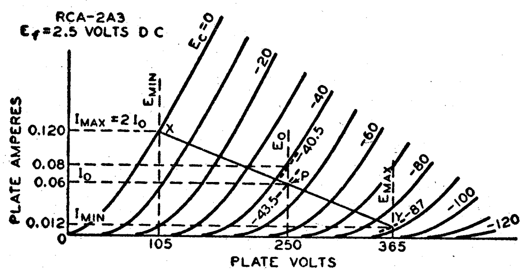

Fig. 21—Graphic calculations for class A amplifier using the 2A3 power triode.

$$\% \text{ distortion} = \frac{\frac{I_{max} + I_{min}}{2} - I_o}{I_{max} - I_{min}} \times 100$$

where I_o is the zero-signal plate current in amperes. If the distortion is excessive, the load resistance should be increased or, occasionally, decreased slightly and the calculations repeated.

Example: Determine the load resistance, power output, and distortion of a triode having an amplification factor of 4.2, a plate-dissipation rating of 15 watts, and plate-characteristics curves as shown in Fig. 21. The tube is to be operated at 250 volts on the plate.

Procedure: For a first approximation, determine the operating point P from the zero-signal bias formula, $Ec_o = -(0.68 \times 250) / 4.2 = -40.5$ volts. From the curve for this voltage, it is found that the zero-signal plate current is 0.08 ampere and, therefore, the plate-dissipation rating is exceeded ($0.08 \times 250 = 20$ watts). Consequently, it is necessary to reduce the zero-signal plate current to 0.06 ampere at 250 volts. The grid bias is then —43.5 volts. Note that the curve was taken with a dc filament supply; if the filament is to be operated on an ac supply, the bias must be increased by about one-half the filament voltage, or to —45 volts, and the circuit returns made to the mid-point of the filament circuit.

Point X can then be determined. Point X is at the intersection of the dc bias curve at zero volts with I_{max}, where $I_{max} = 2I_o = 2 \times 0.06 = 0.12$ ampere. Line XY is drawn through points P and X. E_{max}, E_{min}, and I_{min} are then found from the curves. When these values are substituted in the power-output formula, the following result is obtained:

$$P_o = \frac{(0.12 - 0.012) \times (365 - 105)}{8} = 3.52 \text{ watts}$$

The resistance represented by load line XY is

$$\frac{(365 - 105)}{(0.12 - 0.012)} = 2410 \text{ ohms}$$

When the values from the curves are substituted in the distortion formula, the following result is obtained:

$$\% \text{ distortion} = \frac{\frac{0.12 + 0.012}{2} - 0.06}{0.12 - 0.012} \times 100 = 5.5\%$$

It is customary to select the load resistance so that the distortion does not exceed five per cent. When the method shown is used to determine the slope of the load-resistance line, the second-harmonic distortion generally does not exceed five per cent. In the example, however, the distortion is excessive and it is desirable, therefore, to use a slightly higher load resistance. A load resistance of 2500 ohms will provide a distortion of about 4.9 per cent. The power output is reduced only slightly to 3.5 watts.

Operating conditions for **triodes in push-pull** depend on the type of operation desired. Under class A conditions, distortion, power output, and efficiency are all relatively low. The operating bias can be anywhere between that specified for single-tube operation and that equal to one-half the grid-bias voltage required to produce plate-current cutoff at a plate voltage of $1.4E_o$, where E_o is the operating plate voltage. Higher bias than this value requires higher grid-signal voltage and results in class AB_1 operation, which is discussed later.

The method for calculating maximum power output for **triodes in push-pull class A operation** is as follows: Erect a vertical line at 0.6 E_o (see Fig. 22, intersecting the $E_c = 0$ curve at the point I_{max}. Then, I_{max} is determined from the curve for use in the formula

$$P_o = (I_{max} \times E_o)/5$$

If I_{max} is expressed in amperes and E_o in volts, power output is in watts.

The method for determining the proper load resistance for triodes in push-pull is as follows: Draw a load line through I_{max} on the zero-bias curve and through the E_o point on the zero-current axis. Four times the resistance represented by this load line is the plate-to-plate load (R_{pp}) for two triodes in a class A push-pull amplifier. Expressed as a formula,

$$R_{pp} = 4 \times (E_o - 0.6E_o)/I_{max}$$

where E_o is expressed in volts, I_{max} in amperes, and R_{pp} in ohms.

Example: Assume that the plate voltage (E_o) is to be 300 volts, and the plate-dissipation rating of the tube is 15

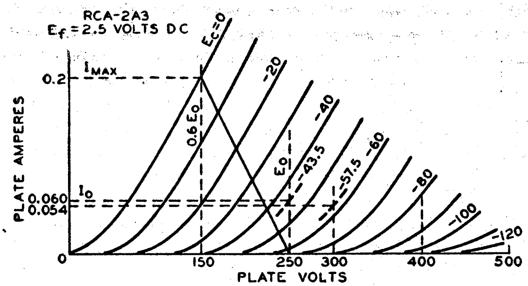

Fig. 22—Graphic calculations for push-pull class A amplifier using the 2A3 power triode.

watts. Then, for class A operation, the operating bias can be equal to, but not more than, one-half the grid bias for cutoff with a plate voltage of 1.4 × 300 = 420 volts. (Since cutoff bias is approximately —115 volts at a plate voltage of 420 volts, one-half of this value is —57.5 volts bias.) At this bias, the plate current is found from the plate family to be 0.054 ampere and, therefore, the plate dissipation is 0.054 × 300 or 16.2 watts. Since —57.5 volts is the limit of bias for class A operation of these tubes at a plate voltage of 300 volts, the dissipation cannot be reduced by increasing the bias and it becomes necessary to reduce the plate voltage.

If the plate voltage is reduced to 250 volts, the bias will be found to be —43.5 volts. For this value, the plate current is 0.06 ampere, and the plate dissipation is 15 watts. Then, following the method for calculating power output, erect a vertical line at $0.6E_o$ = 150 volts. The intersection of the line with the curve E_c = 0 is I_{max} or 0.2 ampere. When this value is substituted in the power formula, the power output is (0.2 × 250)/5 = 10 watts. The load resistance is determined from the load formula: Plate-to-plate load (R_{pp}) = 4 × (250 — 150)/0.2 = 2000 ohms.

Power output for a pentode or a beam power tube as a class A amplifier can be calculated in much the same way as for triodes. The calculations can be made graphically from a special plate family of curves, as illustrated in Fig. 23

From a point A at or just below the knee of the zero-bias curve, draw arbitrarily selected load lines to intersect the zero-plate-current axis. These lines should be on both sides of the operating point P, whose position is

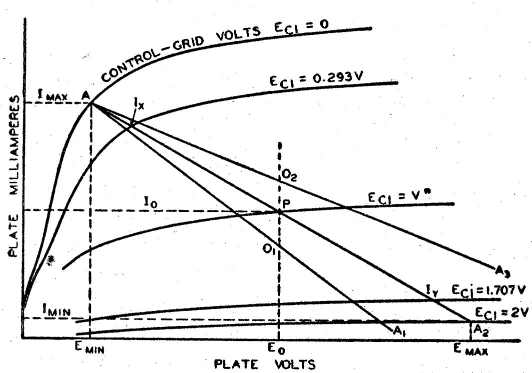

Fig. 23—Graphic calculations for class A amplifier using a pentode or beam power tube.

determined by the desired operating plate voltage, E_o, and one-half the maximum-signal plate current. Along any load line, say AA_1, measure the distance AO_1. On the same line, lay off an equal distance, O_1A_1. For optimum operation, the change in bias from A to O_1 should be nearly equal to the change in bias from O_1 to A_1. If this condition can not be met with one line, as is the case for the line first chosen, then another should be chosen. When the most satisfactory line has been selected, its resistance may be determined by the following formula:

$$\text{Load resistance } (R_L) = \frac{E_{max} - E_{min}}{I_{max} - I_{min}}$$

The value of R_L may then be substituted in the following formula for calculating power output.

$$P_o = \frac{[I_{max} - I_{min} + 1.41\,(I_x - I_y)]^2\,R_L}{32}$$

In both of these formulas, I is in amperes, E is in volts, R_L is in ohms, and P_o is in watts. I_x and I_y are the current values on the load line at bias voltages of $Ec_1 = V - 0.707V = 0.293V$ and $E_{c1} = V + 0.707V = 1.707V$, respectively.

Calculations for distortion may be made by means of the following formulas. The terms used have already been defined.

$$\%\text{ 2nd-harmonic distortion} = \frac{I_{max} + I_{min} - 2\,I_o}{I_{max} - I_{min} + 1.41\,(I_x - I_y)} \times 100$$

$$\%\text{ 3rd-harmonic distortion} = \frac{I_{max} - I_{min} - 1.41\,(I_x - I_y)}{I_{max} - I_{min} + 1.41\,(I_x - I_y)} \times 100$$

$$\%\text{ total (2nd and 3rd) harmonic distortion} = \sqrt{(\%\text{ 2nd})^2 + (\%\text{ 3rd})^2}$$

Conversion Factors

Operating conditions for voltage values other than those shown in the published data can be obtained by use of the **nomograph** shown in Fig. 24 when all electrode voltages are changed simultaneously in the same ratio. The nomograph includes conversion factors for current (F_i), power output (F_p), plate resistance or load resistance (F_r), and transconductance (F_{gm}) for voltage ratios between 0.5 and 2.0. These factors are expressed as functions of the ratio between the desired or new voltage for any electrode (E_{des}) and the published or original value of that voltage (E_{pub}). The relations shown are applicable to triodes and multigrid tubes in all classes of service.

To use the nomograph, simply place a straight-edge across the page so that it intersects the scales for E_{des} and E_{pub} at the desired values. The desired conversion factor may then be read directly or estimated at the point where the straight-edge intersects the F_i, F_p, F_r, or F_{gm} scale.

For example, suppose it is desired to operate two 6L6GC's in class A_1 push-pull, fixed bias, with a plate voltage of 200 volts. The nearest published operating conditions for this class of service are for a plate voltage of 250 volts. The operating conditions for the new plate voltage can be determined as follows:

The voltage conversion factor, F_e, is equal to 200/250 or 0.8. The dashed lines on the **nomograph** of Fig. 24 indicate that for this voltage ratio F_i is approximately 0.72, F_p is approximately 0.57, F_r is 1.12, and F_{gm} is approximately 0.892. These factors may be applied directly to operating values shown in the tube data, or to values calculated by the methods described previously.

Because this method for conversion of characteristics is necessarily an approximation, the accuracy of the nomograph decreases progressively as the ratio E_{des}/E_{pub} departs from unity. In general, results are substantially correct when the value of the ratio E_{des}/E_{pub} is between 0.7 and 1.5. Beyond these limits, the accuracy decreases rapidly, and the results obtained must be considered rough approximations.

The nomograph does not take into consideration the effects of contact potential or secondary emission in tubes. Because contact-potential effects become noticeable only at very small dc grid-No. 1 (bias) voltages, they are generally negligible in power tubes.

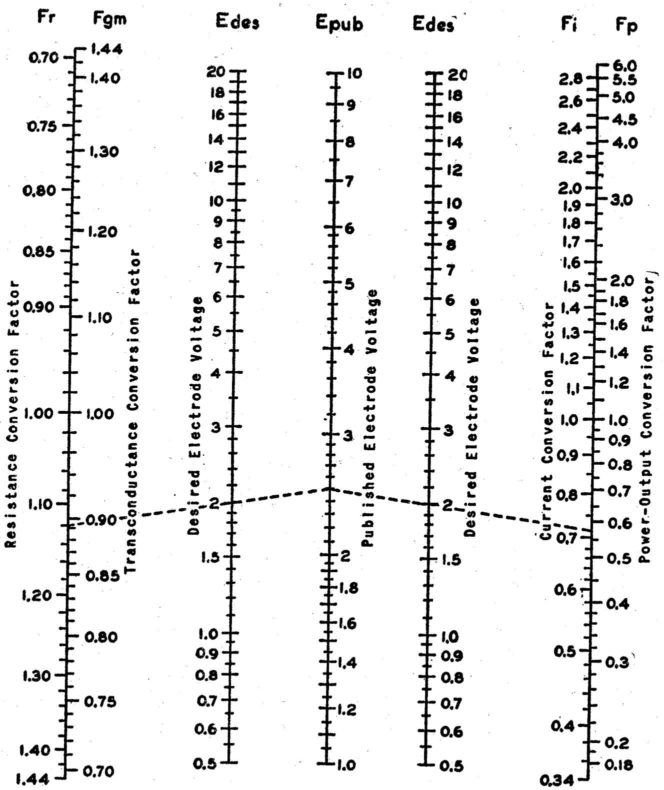

Fig. 24—Nomograph of tube conversion factors.

Secondary emission may occur in conventional tetrodes, however, if the plate voltage swings below the grid-No. 2 voltage. Consequently, the conversion factors shown in the nomograph apply to such tubes only when the plate voltage is greater than the grid-No. 2 voltage. Because secondary emission may also occur in certain beam power tubes at very low values of plate current and plate voltage, the conversion factors shown in the nomograph do not apply when these tubes are operated under such conditions.

Class AB Power Amplifiers

A class AB power amplifier employs two tubes connected in push-pull with a higher negative grid bias than is used in a class A stage. With this higher negative bias, the plate and screen-grid voltages can usually be made higher than for class A amplifiers because the increased negative bias holds plate current within the limit of the tube plate-dissipation rating. As a result of these higher voltages, more power output can be obtained from class AB operation.

Class AB amplifiers are subdivided into class AB_1 and class AB_2. In class AB_1, there is no flow of grid current. That is, the peak signal voltage applied to each grid is not greater than the negative grid-bias voltage. The grids therefore are not driven to a positive potential and do not draw current. In class AB_2, the peak signal voltage is greater than the bias so that the grids are driven positive and draw current.

Because of the flow of grid current in a class AB_2 stage, there is a loss of power in the grid circuit. The sum of this loss and the loss in the input transformer is the total driving power required by the grid circuit. The driver stage should be capable of a power output considerably larger than this required power in order that distortion introduced in the grid circuit be kept low. The input transformer used in a class AB_2 amplifier usually has a step-down turns ratio.

Because of the large fluctuations of plate current in a class AB_2 stage, it is important that the plate power supply have good regulation. Otherwise the fluctuations in plate current cause fluctuations in the voltage output of the power supply, with the result that power output is decreased and distortion is increased. To obtain satisfactory regulation, it is usually advisable to use a low-drop rectifier, such as the 5V4GA, with a choke-input filter. In all cases, the resistance of the choke and transformers should be as low as possible.

Class AB_1 Power Amplifiers

In class AB_1 push-pull amplifier service using triodes, the operating conditions may be determined graphically by means of the plate family if E_o, the desired operating plate voltage, is given. In this service, the dynamic load line does not pass through the operating point P as in the case of the single-tube amplifier, but through the point D in Fig. 25. Its position is not affected by the operating grid bias provided the plate-to-plate load resistance remains constant.

Under these conditions, grid bias has no appreciable effect on the power output. Grid bias cannot be neglected, however, since it is used to find the zero-signal plate current and, from it, the zero-signal plate dissipation. Because the grid bias is higher in class AB_1 than in class A service for the same plate voltage, a higher signal voltage may be used without grid current being drawn and, therefore, higher power output is obtained.

In general, for any load line through point D, Fig. 25, the plate-to-plate load resistance in ohms of a push-pull amplifier is $R_{pp} = 4E_o/I'$, where I' is the plate-current value in amperes at which the load line as projected intersects the plate-current axis, and E_o is in volts. This formula is another form of the one given under push-pull class A amplifiers, $R_{pp} = 4(E_o - 0.6E_o)/I_{max}$, but is more general. Power output $= (I_{max}/\sqrt{2})^2 \times R_{pp}/4$, where I_{max} is the peak plate current at zero grid volts for the load chosen. This formula simplified is $(I_{max})^2 \times R_{pp}/8$. The maximum-signal average plate current is $2I_{max}/\pi$ or $0.636\ I_{max}$; the maximum-signal average power input is $0.636\ I_{max} \times E_o$.

It is desirable to simplify these

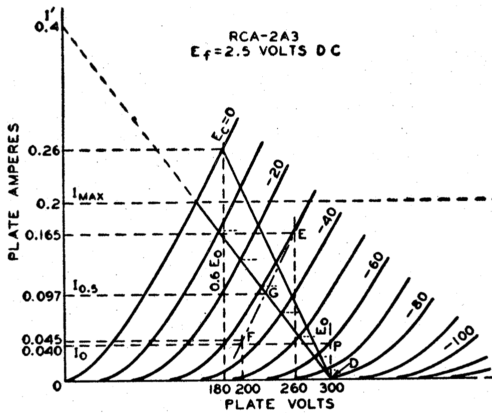

Fig. 25—Graphic calculations for class AB_1 amplifier using the 2A3 power triode.

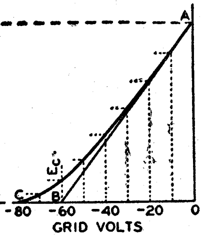

Fig. 26—Instantaneous curve for class AB_1 amplifier.

formulas for a first approximation. This simplification can be made if it is assumed that the peak plate current, I_{max}, occurs at the point of the zero-bias curve corresponding approximately to 0.6 E_o, the condition for maximum power output. The simplified formulas are:

$$P_o \text{ (for two tubes)} = (I_{max} \times E_o)/5$$
$$R_{pp} = 1.6E_o/I_{max}$$

where E_o is in volts, I_{max} is in amperes, R_{pp} is in ohms, and P_o is in watts.

It may be found during subsequent calculations that the distortion or the plate dissipation is excessive for this approximation; in that case, a different load resistance must be selected, using the first approximation as a guide, and the process repeated to obtain satisfactory operating conditions.

Example: Fig. 25 illustrates the application of this method to a pair of 2A3's operated at $E_o = 300$ volts. Each tube has a plate-dissipation rating of 15 watts. The method is to erect a vertical line at $0.6E_o$, or at 180 volts, which intersects the $E_c = 0$ curve at the point $I_{max} = 0.26$ ampere. Using the simplified formulas, the following values are obtained:

$$R_{pp} = (1.6 \times 300)/0.26 = 1845 \text{ ohms}$$
$$P_o = (0.26 \times 300)/5 = 15.6 \text{ watts}$$

At this point, it is well to determine the plate dissipation and to compare it with the maximum rated value. From the average-plate-current formula ($0.636\ I_{max}$) mentioned previously, the maximum-signal average plate current is 0.166 ampere. The product of this current and the operating plate voltage is 49.8 watts, the average input to the two tubes. From this value, subtract the power output of 15.6 watts to obtain the total dissipation for both tubes, which is 34.2 watts. Half of this value, 17 watts, is in excess of the 15-watt rating of the tube and it is necessary, therefore, to assume another and higher load resistance so that the plate-dissipation rating will not be exceeded.

It will be found that at an operating plate voltage of 300 volts the 2A3's require a plate-to-plate load resistance of 3000 ohms. From the formula for R_{pp}, the value of I' is found to be 0.4 ampere. The load line for the 3000-ohm load resistance is then represented by a straight line from the point $I' = 0.4$ ampere on the plate-current ordinate to the point $E_o = 300$ volts on the plate-voltage abscissa. At the intersection of the load line with the zero-bias curve, the peak plate current, I_{max}, can be read at 0.2 ampere. Then

$$P_o = (I_{max}/\sqrt{2})^2 \times R_{pp}/4$$
$$= (0.2/1.41)^2 \times 3000/4$$
$$= 15 \text{ watts}$$

Proceeding as in the first approximation, it is found that the maximum-signal average plate current, $0.636I_{max}$, is 0.127 ampere, and the maximum-signal average power input is 38.1 watts. This input minus the power output is $38.1 - 15 = 23.1$ watts. This value is the dissipation for two tubes; the value per tube is 11.6 watts, a value well within the rating of this tube type.

The operating bias and the zero-signal plate current may then be found by use of a curve which is derived from the plate family and the load line. Fig. 26 is a curve of instantaneous values of plate current and dc grid-bias voltages taken from Fig. 25. Values of grid bias are read from each of the grid-bias curves of Fig. 25 along the load line and are transferred to Fig. 26 to produce the curved line from A to C. A tangent to this curve, starting at A, is drawn to intersect the grid-voltage abscissa. The point of intersection, B, is the operating grid bias for fixed-bias operation. In the example, the bias is −60 volts. Refer back to the plate family at the operating conditions of plate volts = 300 and grid bias = −60 volts; the zero-signal plate current per tube is seen to be 0.04 ampere.

This procedure locates the operating point for each tube at P. The plate current must be doubled, of course, to obtain the zero-signal plate current for both tubes. Under maximum-signal conditions, the signal voltage swings from zero-signal bias voltage to zero bias for each tube on alternate half cycles. Hence, in the example, the peak of signal voltage per tube is 60 volts, or the grid-to-grid value is 120 volts.

As in the case of the push-pull class A amplifier, the second-harmonic distortion in a class AB_1 amplifier using triodes is very small and is largely canceled by virtue of the push-pull connection. Third-harmonic distortion, however, which may be larger than permissible, can be found by means of composite characteristic curves. A complete family of curves can be plotted, but for the present purpose only the one corresponding to a grid bias of one-half the peak grid-voltage swing is needed. In the example, the peak grid voltage per tube is 60 volts, and the half value is 30 volts. The composite curve, since it is nearly a straight line, can be constructed with only two points (see Fig. 25. These two points are obtained from deviations above and below the operating grid and plate voltages.

In order to find the curve for a bias of −30 volts, a deviation of 30 volts from the operating grid voltage of −60 volts is assumed. Next assume a deviation from the operating plate voltage of, say, 40 volts. Then at 300 −40 = 260 volts, erect a vertical line to intersect the (−60) − (−30) = −30-volt bias curve and read the plate current at this intersection, which is 0.167 ampere; likewise, at the intersection of a vertical line at 300 + 40 = 340 volts and the (−60) + (−30) = −90-volt bias curve, read the plate current. In this example, the plate current is estimated to be 0.002 ampere. The difference of 0.165 ampere between these two currents determines the point E on the 300 − 40 = 260-volt vertical. Similarly, another point F on the same composite curve is found by assuming the same grid-bias deviation but a larger plate-voltage deviation, say, 100 volts.

These steps provide points at 260 volts and 0.165 ampere (E), and at 200 volts and 0.045 ampere (F). A straight line through these points is the composite curve for a bias of −30 volts, shown as a long-short dash line in Fig. 25. At the intersection of the composite curve and the load line, G, the instantaneous composite plate current at the point of one-half the peak signal swing is determined. This current value, designated $I_{0.5}$ and the peak plate current, I_{max}, are used in the following formula to find the peak value of the third-harmonic component of the plate current.

$$I_{h3} = (2I_{0.5} - I_{max})/3$$

In the example, where $I_{0.5}$ is 0.097 ampere and I_{max} is 0.2 ampere, I_{h3} = (2 × 0.097 − 0.2)/3 = (0.194 − 0.2)/3 = −0.006/3 = −0.002 ampere. (The fact that I_{h3} is negative indicates that the phase relation of the fundamental (first-harmonic) and third-harmonic components of the plate current is such as to result in a slightly peaked wave form. I_{h3} is positive in some cases, indicating a flattening of the wave form.)

The peak value of the fundamental or first-harmonic component of the plate current is found by the following formula:

$$I_{h1} = 2/3 \times (I_{max} + I_{0.5})$$

In the example, I_{h1} = 2/3 × (0.2 + 0.097) = 0.198 ampere. Thus, the percentage of third-harmonic distortion is (I_{h3}/I_{h1}) × 100 = (0.002/0.198) × 100 = 1 per cent approx.

Class AB_2 Power Amplifiers

A class AB_2 amplifier employs two tubes connected in push-pull as in the case of class AB_1 amplifiers. It differs in that it is biased so that plate current flows for somewhat more than half the electrical cycle but less than the full cycle, the peak signal voltage is greater than the dc bias voltage, grid current is drawn, and, consequently, power is consumed in the grid circuit. These conditions permit high power output to be obtained without excessive plate dissipation.

The sum of the power used in the grid circuit and the losses in the input transformer is the total driving power required by the grid circuit. The driver stage should be capable of a power output considerably larger than this required power in order that distortion introduced in the grid circuit be kept low. In addition, the internal impedance of the driver stage as reflected into or as effective in the grid circuit of the power stage should always be as low as possible in order that distortion may be kept low. The input transformer used in a class AB_2 stage usually has a step-down ratio adjusted for this condition.

Load resistance, plate dissipation, power output, and distortion determinations are similar to those for class AB_1. These quantities are interdependent with peak grid-voltage swing and driving power; a satisfactory set of operating conditions involves a series of approximations. The load resistance and signal swing are limited by the permissible grid current and power and the distortion. If the load resistance is too high or the signal swing is excessive, the plate-dissipation rating will be exceeded, distortion will be high, and the driving power will be unnecessarily high.

Class B Power Amplifiers

A **class B amplifier** employs two tubes connected in push-pull, so biased that plate current is almost zero when no signal voltage is applied to the grids. Because of this low value of no-signal plate current, class B amplification has the same advantage as class AB_2, *i.e.*, large power output can be obtained without excessive plate dissipation. Class B operation differs from class AB_2 in that plate current is cut off for a larger portion of the negative grid swing, and the signal swing is usually larger than in class AB_2 operation.

Because certain triodes used as class B amplifiers are designed to operate very close to zero bias, the grid of each tube is at a positive potential during all or most of the positive half-cycle of its signal swing. In this type of triode operation, considerable grid current is drawn and there is a loss of power in the grid circuit. This condition imposes the same requirement in the driver stage as in a class AB_2 stage; *i.e.*, the driver should be capable of delivering considerably more power output than the power required for the grid circuit of the class B amplifier so that distortion will be low. Similarly, the interstage transformer between the driver and the class B stage usually has a step-down turns ratio. Because of the high dissipations involved in class B operation at zero bias, it is not feasible to use tetrodes or pentodes in this type of class B operation.

Determination of load resistance, plate dissipation, power output, and distortion is similar to that for a class AB_2 stage.

Power amplifier tubes designed for class A operation can be used in class AB_2 and class B service under suitable operating conditions. There are several tube types designed especially for class B service. The characteristic common to all of these types is a high amplification factor. With a high amplification factor, plate current is small even when the grid bias is zero. These tubes, therefore, can be operated in class B service at a bias of zero volts so that no bias supply is required. A number of class B amplifier tubes consist of two triode units mounted in one tube. The two units can be connected in push-pull so that only one tube is required for a class B stage.

High-Fidelity Amplifiers

Several high-fidelity amplifiers are shown in the **Circuits** section. The performance capabilities of such amplifiers are usually given in terms of frequency response, total harmonic distortion, maximum power output, and noise level.

To provide high-fidelity reproduction of audio program material, an amplifier should have a frequency response which does not vary more than 1 db over the entire audio spectrum. General practice is to design the amplifier so that its frequency response is flat within 1 db from a frequency below the lowest to be reproduced to one well above the upper limit of the audible region.

Harmonic distortion and intermodulation distortion produce changes in program material which may have adverse effects on the quality of the reproduced sound. **Harmonic distortion** causes a change in the character of an individual tone by the introduction of harmonics which were not originally present in the program material. For high-fidelity reproduction, total harmonic distortion (expressed as a percentage of the output power) should not be greater than about 1 per cent at the desired listening level. Types such as the 6973, 7027A and 7868 are designed to provide extremely low harmonic distortion in suitably designed push-pull amplifier circuits.

Intermodulation distortion is a change in the waveform of an individual tone as a result of interaction with another tone present at the same time in the program material. This type of distortion not only alters the character of the modulated tone, but may also result in the generation of spurious signals at frequencies equal to the sum and difference of the interacting frequencies. Intermodulation distortion should be less than 2 per cent at the desired listening level. In general, any amplifier which has low intermodulation distortion will have very low harmonic distortion.

The maximum power output which a high-fidelity amplifier should deliver depends upon a complex relation of several factors, including the size and acoustical characteristics of the listening area, the desired listening level, and the efficiency of the loudspeaker system. Practically, however, it is possible to determine amplifier requirements in terms of room size and loudspeaker efficiency.

The acoustic power required to reproduce the loudest passages of orchestral music at concert-hall level in the average-size living room is about 0.4 watt. Because high-fidelity loudspeakers of the type generally available for home use have an efficiency of only about 5 per cent, the output stage of the amplifier should therefore be able to deliver a power output of at least 8 watts. Because many wide-range loudspeaker systems, particularly those using frequency-divider networks, have efficiencies of less than 5 per cent, output tubes used with such systems must have correspondingly larger power outputs. The 6973, 7027A, 7189, and 7868 can provide ample output for most systems when used in suitable push-pull circuits.

The noise level of a high-fidelity amplifier determines the range of volume the amplifier is able to reproduce, *i.e.*, the difference (usually expressed in decibels) between the loudest and softest sounds in program material. Because the greatest volume range utilized in electrical program material at the present time is about 60 dB, the noise level of a high-fidelity amplifier should be at least 60 dB below the signal level at the desired listening level.

Cathode-Drive Circuits

The preceding text has discussed the use of tubes in the conventional **grid-drive** type of amplifier—that is, where the cathode is common to both the input and output circuits. Tubes may also be employed as amplifiers in circuit arrangements which utilize the grid or plate as the common terminal. Probably the most important of these amplifiers are the cathode-drive circuit, which is discussed below, and the cathode-follower circuit, which will be discussed later in connection with inverse feedback.

A typical **cathode-drive** circuit is shown in Fig. 27. The load is placed in the plate circuit and the output voltage is taken off between the plate and ground as in the grid-drive method of operation. The grid is grounded, and the input voltage is applied across an appropriate impedance in the cathode circuit. The cathode-drive circuit is particularly useful for vhf and uhf applications, in which it is necessary to obtain the low-noise performance usually associated with a triode, but where a conventional grid-drive circuit would be unstable because of feedback through the grid-to-plate capacitance of the tube. In the cathode-drive circuit, the grounded grid serves as a capacitive shield between plate and cathode and

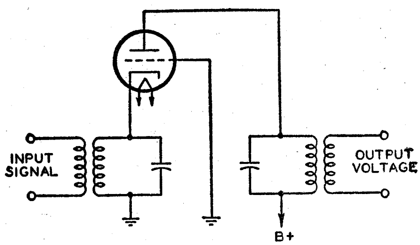

Fig. 27—Cathode-drive circuit.

permits stable operation at frequencies higher than those in which conventional circuits can be used.

The input impedance of a cathode-drive circuit is approximately equal to $1/g_m$ when the load resistance is small compared to the r_p of the tube. A certain amount of power is required, therefore, to drive such a circuit. However, in the type of service in which cathode-drive circuits are normally used, the advantages of the grounded-grid connection usually outweigh this disadvantage.

Inverse Feedback

An inverse-feedback circuit, sometimes called a **degenerative** circuit, is one in which a portion of the output voltage of a tube is applied to the input of the same or a preceding tube in opposite phase to the signal applied to the tube. Two important advantages of feedback are (1) reduced distortion from each stage included in the feedback circuit and (2) reduction in the variations in gain due to changes in line voltage, possible differences between tubes of the same type, or variations in the values of circuit constants included in the feedback circuit.

Inverse feedback is used in audio amplifiers to reduce distortion in the output stage where the load impedance on the tube is a loudspeaker. Because the impedance of a loudspeaker is not constant for all audio frequencies, the load impedance on the output tube varies with frequency. When the output tube is a pentode or beam power tube having high plate resistance, this variation in plate load impedance can, if not corrected, produce considerable frequency distortion. Such frequency distortion can be reduced by means of inverse feedback. Inverse-feedback circuits are of the **constant-voltage** type and the **constant-current** type.

The application of the **constant-voltage** type of inverse feedback to a power-output stage using a single beam power tube is illustrated in Fig. 28. In this circuit, R_1, R_2, and C are connected as a voltage divider across the output of the tube. The secondary winding of the **grid-input transformer** is returned to a point on this voltage divider. Capacitor C blocks the dc plate voltage from the grid. However, a portion of the tube af output voltage, approximately equal to the output voltage multiplied by the

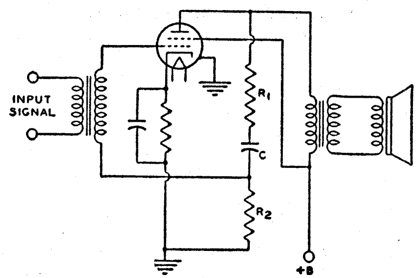

Fig. 28—Power-output stage using constant-voltage inverse feedback.

fraction $R_2/(R_1 + R_2)$, is applied to the grid. This voltage reduces the source impedance of the circuit and a decrease in distortion results which is explained in the curves of Fig. 29.

Consider first the amplifier without the use of inverse feedback. Suppose that when a signal voltage e_s is applied to the grid the af plate current i'_p has an irregularity in its positive half-cycle. This irregularity represents a departure from the waveform of the input signal and is, therefore, distortion. For this plate-current waveform, the af plate voltage has a waveform shown by e'_p. The plate-voltage waveform is inverted compared to the plate-current waveform because a plate-current increase produces an increase in the drop across the plate load. The voltage at the plate is the difference between the drop across the load and the supply voltage; thus, when plate current goes up, plate voltage goes down; when plate current goes down, plate voltage goes up.

Now suppose that inverse feedback is applied to the amplifier. The voltage fed back to the grid has the same waveform and phase as the plate voltage, but is smaller in magnitude. Hence, with a plate voltage of waveform shown by e'_p, the feedback voltage appearing on the grid is as shown by e'_{gf}. This voltage applied to the grid produces a component of plate current i'_{pf}. It is evident that the irregularity in the waveform of

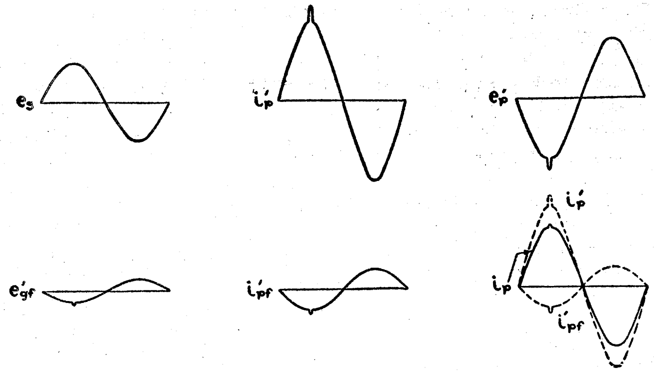

Fig. 29—Voltage and current waveforms showing effect of inverse feedback.

this component of plate current would act to cancel the original irregularity and thus reduce distortion.

After inverse feedback has been applied, the relations are as shown in the curve for i_p. The dotted curve shown by i'_{pf} is the component of plate current due to the feedback voltage on the grid. The dotted curve shown by i'_p is the component of plate current due to the signal voltage on the grid. The algebraic sum of these two components gives the resultant plate current shown by the solid curve of i_p. Since i'_p is the plate current that would flow without inverse feedback, it can be seen that the application of inverse feedback has reduced the irregularity in the output current. In this manner inverse feedback acts to correct any component of plate current that does not correspond to the input signal voltage, and thus reduces distortion.

From the curve for i_p, it can be seen that, besides reducing distortion, inverse feedback also reduces the amplitude of the output current. Consequently, when inverse feedback is applied to an amplifier there is a decrease in gain or power sensitivity as well as a decrease in distortion. Hence, the application of inverse feedback to an amplifier requires that more driving voltage be applied to obtain full power output, but this output is obtained with less distortion.

Inverse feedback may also be applied to resistance-coupled stages, as shown in Fig. 30. The circuit is conventional except that a feedback resistor, R_3, is connected between the plates of tubes T_1 and T_2. The output signal voltage of T_1 and a portion of the output signal voltage of T_2 appear across R_2. Because the distortion generated in the plate circuit of T_2 is applied to its grid out of phase with the input signal, the distortion in the output of T_2 is comparatively low. With sufficient inverse feedback of the constant-voltage type

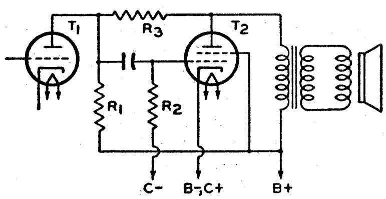

Fig. 30—Resistance-coupled stages using feedback resistor.

in a power-output stage, it is not necessary to employ a network of resistance and capacitance in the output circuit to reduce response at high audio frequencies. Inverse-feedback circuits can also be applied to push-pull class A and class AB_1 amplifiers.

Constant-current inverse feedback is usually obtained by omitting the by-pass capacitor across a cathode resistor. This method decreases the gain and the distortion but increases the source impedance of the circuit. Consequently, the output voltage rises at the resonant frequency of the loudspeaker and accentuates hangover effects.

Inverse feedback is not generally applied to a triode power amplifier, such as the 2A3, because the variation in speaker impedance with frequency does not produce much distortion in a triode stage having low plate resistance. It is sometimes applied in a pentode stage, but is not always convenient. As has been shown, when inverse feedback is used in an amplifier, the driving voltage must be increased in order to provide full power output. When inverse feedback is used with a pentode, the total driving voltage required for full power output may be inconveniently large, although still less than that required for a triode. Because a beam power tube gives full power output on a comparatively small driving voltage, inverse feedback is especially applicable to beam power tubes. By means of inverse feedback, the high efficiency and high power output of beam power tubes can be combined with freedom from the effects of varying speaker impedance.

Cathode-Follower Circuits

Another important application of inverse feedback is in the cathode-follower circuit, an example of which is shown in Fig. 31. In this application, the load has been transferred from the plate circuit to the cathode circuit of the tube.

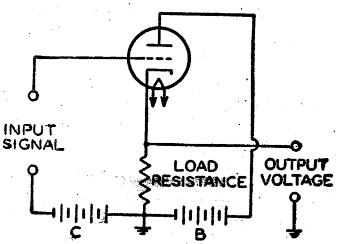

Fig. 31—Cathode-follower circuit.

The input voltage is applied between the grid and ground, and the output voltage is obtained between the cathode and ground. The voltage amplification (V.A.) of this circuit is always less than unity and may be expressed by the following convenient formulas.

For a triode:

$$V.A. = \frac{\mu \times R_L}{r_p + [R_L \times (\mu + 1)]}$$

For a pentode:

$$V.A. = \frac{g_m \times R_L}{1 + (g_m \times R_L)}$$

In these formulas, μ is the amplification factor, R_L is the load resistance in ohms, r_p is the plate resistance in ohms, and g_m is the transconductance in mhos.

The use of the cathode follower permits the design of circuits which have high input resistance and high output voltage. The output impedance is quite low and very low distortion may be obtained. Cathode-follower circuits may be used for power amplifiers or as impedance transformers designed either to match a transmission line or to produce a relatively high output voltage at a low impedance level.

In a power amplifier which is transformer coupled to the load, the same output power can be obtained from the tube as would be obtained in a conventional grid-drive type of amplifier. The output impedance is very low and provides excellent damping to the load, with the result that very low distortion can be obtained. The peak-to-peak signal voltage, however, approaches 1½ times the plate supply voltage if maximum power output is required from the tube. Some problems may be encountered, therefore, in the design of an adequate driver stage for a cathode-follower output system.

When a cathode-follower circuit is used as an impedance transformer, the load is usually a simple resistance in the cathode circuit of the tube. With relatively low values of cathode resistor, the circuit may be designed to supply significant amounts of power and to match the impedance of the device to a transmission line. With somewhat higher values of cathode resistor, the circuit may be used to decrease the output impedance sufficiently to permit the transmission of audio signals along a line in which appreciable capacitance is present.

The cathode follower may also be used as an isolation device to provide extremely high input resistance and low input capacitance as might be required in the probe of an oscilloscope or vacuum-tube voltmeter. Such circuits can be designed to provide effective impedance transformation with no significant loss of voltage.

Selection of a suitable tube and its operating conditions for use in a cathode-follower circuit having a specified output impedance (Z_o) can be made, in most practical cases, by the use of the following formula to determine the approximate value of the required tube transconductance.

$$\text{Required } g_m\ (\mu\text{mhos}) = \frac{1{,}000{,}000}{Z_o\ (\text{ohms})}$$

Once the required transconductance is obtained, a suitable tube and its operating conditions may be determined from the technical data given in the **Technical Data** section. The tube selected should have a value of transconductance slightly lower than that obtained from the above expression to allow for the shunting effect of the cathode load resistance. The conversion nomograph given in Fig. 24 may be used for calculation of operating conditions for values of transconductance not included in the tabulated data. After the operating conditions have been determined, the approximate value of the required cathode load resistance may be calculated from the following formulas. For a triode:

$$\text{Cathode } R_L = \frac{Z_o \times r_p}{r_p - [Z_o \times (1 + \mu)]}$$

For a pentode:

$$\text{Cathode } R_L = \frac{Z_o}{1 - (g_m \times Z_o)}$$

Resistance and impedance values are in ohms; transconductance values are in mhos.

If the value of the cathode load resistance calculated to provide the required output impedance does not provide the required operating bias, the basic cathode-follower circuit can be modified in a number of ways. Two of the more common modifications are shown in Figs. 32 and 33.

In Fig. 32 the bias is increased by adding a bypassed resistance between the cathode and the unbypassed load resistance and returning the grid to the low end of the load resistance. In Fig. 33 the bias is reduced by adding a bypassed resistance between the cathode and the unbypassed load resistance but, in this case, the grid is returned to the junction of the two cathode resistors so that the bias voltage is only the dc voltage drop across the added resistance. The size of the bypass capacitor should be large enough so that it has negligible reactance at the lowest frequency to be handled. In both cases the B-supply should be increased to make up for the voltage taken for biasing.

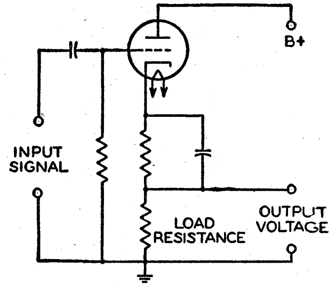

Fig. 32—Cathode-follower circuit modified for increased bias.

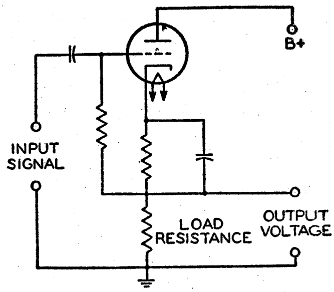

Fig. 33—Cathode-follower circuit modified for reduced bias.

Example: Select a suitable tube and determine the operating conditions and circuit components for a cathode-follower circuit having an output impedance that will match a 500-ohm transmission line.

Procedure: First, determine the approximate transconductance required.

$$\text{Required } g_m = \frac{1{,}000{,}000}{500} = 2000\ \mu\text{mhos}$$

A survey of the tubes that have a transconductance in this order of magnitude shows that type 12AX7A is among the tubes to be considered. Referring to the characteristics given in the technical data section for one triode unit of high-mu twin triode 12AX7, we find that for a plate voltage of 250 volts and a bias of −2 volts, the transconductance is 1600 micromhos, the plate resistance is 62500 ohms, the amplification factor is 100, and the plate current is 0.0012 ampere. When these values are used in the expression for determining the cathode load resistance, the following result is obtained:

$$\text{Cathode } R_L = \frac{500 \times 62500}{62500 - 500 \times (100+1)} = 2600 \text{ ohms}$$

The voltage across this resistor for a plate current of 0.0012 ampere is 2600 × 0.0012 = 3.12 volts. Because the required bias voltage is only −2 volts, the circuit arrangement given in Fig. 33 is employed. The bias is furnished by a resistance that will have a voltage drop of 2 volts when it carries a current of 0.0012 ampere. The required bias resistance, therefore, is 2/0.0012 = 1670 ohms. If 60 cycles per second is the lowest frequency to be passed, 20 microfarads is a suitable value for the bypass capacitor. The B-supply, of course, is increased by the voltage drop across the cathode resistance which, in this example, is approximately 5 volts. The B-supply, therefore, is 250 + 5 = 255 volts.

Because it is desirable to eliminate, if possible, the bias resistor and bypass capacitor, it is worthwhile to try other tubes and other operating conditions to obtain a value of cathode load resistance which will also provide the required bias. If the triode section of twin diode—high-mu triode 6AT6 is operated under the conditions given in the technical data section with a plate voltage of 100 volts and a bias of −1 volt, it will have an amplification factor of 70, a plate resistance of 54000 ohms, a transconductance of 1300 micromhos, and a plate current of 0.0008 ampere. Then,

$$\text{Cathode } R_L = \frac{500 \times 54000}{54000 - 500 \times (70 + 1)} = 1460 \text{ ohms}$$

The bias voltage obtained across this resistance is 1460 × 0.0008 = 1.17 volts. Since this value is for all practical purposes close enough to the required bias, no addition bias resistance will be required and the grid may be returned directly to ground. There is no need to adjust the B-supply voltage to make up for the drop in the cathode resistor. The voltage amplification (V.A.) for the cathode-follower circuit utilizing the triode section of type 6AT6 is

$$\text{V.A.} = \frac{70 \times 1460}{54000 + 1460 \times (70 + 1)} = 0.65$$

For applications in which the cathode follower is used to isolate two circuits—for example, when it is used between a circuit being tested and the input stage of an oscilloscope or a vacuum-tube voltmeter—voltage output and not impedance matching is the primary consideration. In such applications it is desirable to use a relatively high value of cathode load resistance, such as 50,000 ohms, in order to get the maximum voltage output. In order to obtain proper bias, a circuit such as that of Fig. 33 should be used. With a high value of cathode resistance, the voltage amplification will approximate unity.

Corrective Filters

A corrective filter can be used to improve the frequency characteristic of an output stage using a beam power tube or a pentode when inverse feedback is not applicable. The filter consists of a resistor and a capacitor connected in series across the primary of the output transformer. Connected in this way, the filter is in parallel with the plate load impedance reflected from the voice-coil by the output transformer. The magnitude of this reflected impedance increases with increasing frequency in the middle and upper audio range. The impedance of the filter, however, decreases with increasing frequency. It follows that, by use of the proper values for the resistance and the capacitance in the filter, the effective load impedance on the output tubes can be made practically constant for all frequencies in the middle and upper audio range. The result is an improvement in the frequency characteristic of the output stage.

The resistance to be used in the filter for a push-pull stage is 1.3 times the recommended plate-to-plate load resistance; or, for a single-tube stage, is 1.3 times the recommended plate load resistance. The capacitance in the filter should have a value such that the voltage gain of the output stage at a frequency of 1000 cycles or higher is equal to the voltage gain at 400 cycles.

A method of determining the proper value of capacitance for the filter is to make two measurements of the output voltage across the primary of the output transformer: first, when a 400-cycle signal is applied to the input, and second, when a 1000-cycle signal of the same voltage as the 400-cycle signal is applied to the input. The correct value of capacitance is the one which gives equal output voltages for the two signal inputs. In practice, this value is usually found to be in the order of 0.05 microfarad.

Volume Compressors and Expanders

Volume compression and expansion are used in FM transmitters and receivers and in recording devices and amplifiers to make more natural the reproduction of music which has a very large volume range. For example, in the music of a symphony orchestra the sound intensity of the soft passages is very much lower than that of the loud passages. When this low volume level is raised above the background noise for transmitting or recording, the peak level of the program material may be raised to an excessively high volume level. It is often necessary, therefore, to compress the volume range of the program content within the maximum capabilities of the FM transmitter or the recording device. Exceeding a maximum peak volume level for FM modulation corresponds to exceeding the allowed bandwidth for transmission. In some recording devices, excessive peak volume levels may cause overloading and distortion.

Volume compression may be accomplished by either manual or automatic control. The types of compression used include peak limiters, volume limiters, and volume compressors. A peak limiter limits the peak power to some predetermined level. A volume limiter provides gain reduction based on an average signal level above a predetermined level. A volume compressor provides gain reduction for only the sustained loud portions of the sound level. Only volume compressors can be correctly compensated for with volume expanders.

For faithful reproduction of the original sound, the volume expander used in the FM receiver or audio amplifier should have the reverse characteristic of the volume compressor used in the FM transmitter or recording device. In general, the basic requirements for either a volume compressor or expander are shown in the block diagram of Fig. 34. In a volume compressor, the variable-gain amplifier V_1 has greater gain for a low-amplitude signal than for a high-amplitude signal; therefore, soft passages are amplified more than loud ones. In an expander, the gain is greater for high-amplitude signals than for low-amplitude signals; therefore, loud passages are amplified more than soft ones and the original amplitude ratio is restored.

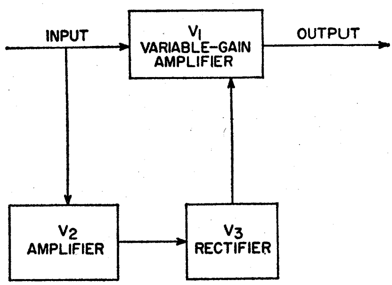

Fig. 34—Block diagram of volume compressor or expander circuit.

In the diagram shown in Fig. 34, the signal to be amplified is applied to V_1, and a portion of the signal is also applied to V_2. The amplified output from V_2 is then rectified by V_3, and applied as a negative (for compressors) or positive (for expanders) bias voltage to V_1. As this bias voltage varies with variations in signal amplitude, the gain of V_1 also varies to produce the desired compression or expansion of the signal.

Tubes having a large dynamic range provide the best results in volume compressor or expander applications. Examples of such types are the 6BJ6 and 6BE6. Push-pull operation is generally desired for the variable-gain amplifier to prevent high distortion and other undesirable effects which may occur in volume compressors and expanders.

Phase Inverters

A phase inverter is a circuit used to provide resistance coupling between the output of a signal-tube stage and the input of a push-pull stage. The necessity for a phase inverter arises because the signal-voltage inputs to the grids of a push-pull stage must be 180 degrees out of phase and approximately equal in amplitude with respect to each other. Thus, when the signal voltage input to a push-pull stage swings the grid of one tube in a positive direction, it should swing the grid of the other tube in a negative direction by a similar amount. With transformer coupling between stages, the out-of-phase input voltage to the push-pull stage is supplied by means of the center-tapped secondary. With resistance coupling, the out-of-phase input voltage is obtained by means of the inverter action of a tube.

Fig. 35 shows a push-pull power amplifier, resistance-coupled by means of a phase-inverter circuit to a single-stage triode T_1. Phase inversion in this circuit is provided by triode T_2. The output voltage of T_1 is applied to the grid of triode T_3. A portion of the output voltage of T_1 is also applied through the resistors R_3 and R_5 to the grid of T_2. The output voltage of T_2 is applied to the grid of triode T_4.

When the output voltage of T_1 swings in the positive direction, the plate current of T_2 increases. This action increases the voltage drop across the plate resistor R_2 and swings the plate of T_2 in the negative direction. Thus, when the output voltage of T_1 swings positive, the output voltage of T_2 swings negative and is, therefore, 180 degrees out of phase with the output voltage of T_1.

In order to obtain equal voltages at E_a and E_b, $(R_3 + R_5)/R_5$ should equal the voltage gain of T_2. Under the condition where a twin-type tube or two tubes having the same characteristics are used as T_1 and T_2, R_4 should be equal to the sum of R_3 and R_5. The ratio of $R_3 + R_5$ to R_5 should be the same as the voltage gain ratio of T_2 in order to apply the correct value of signal voltage to T_2. The value of R_5 is, therefore, equal to R_4 divided by the voltage gain of T_2; R_3 is equal to R_4 minus R_5. Values of R_1, R_2, R_3 plus R_5, and R_4 may be taken from the chart in the **Resistance-Coupled Amplifiers** section. In the practical application of this circuit, it is convenient to use a twin-triode tube combining T_1 and T_2.

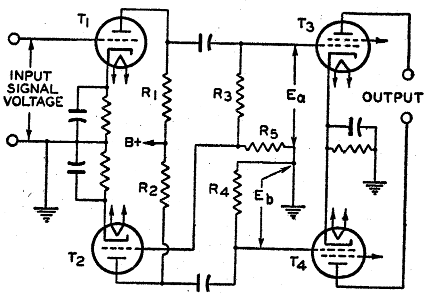

Fig. 35—Push-pull power amplifier resistance-coupled to triode by means of phase inverter.

Tone Controls

A tone control is a variable filter (or one in which at least one element is adjustable) by means of which the user may vary the frequency response of an amplifier to suit his own taste. In radio receivers and home amplifiers, the tone control usually consists of a resistance-capacitance network in which the resistance is the variable element.

The simplest form of tone control

is a fixed tone-compensating or "equalizing" network such as that shown in Fig. 36. This type of network is often used to equalize the low- and high-frequency response of a crystal phonograph pickup. At low frequencies the attenuation of this network is 20.8 dB. As the frequency is increased, the 100-picofarad capacitor serves as a bypass for the 5-megohm resistor, and the combined impedance of the resistor-capacitor network is reduced. Thus,

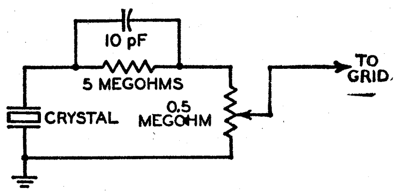

Fig. 36—Tone-control circuit for fixed tone compensation or "equalizing".

more of the crystal output appears across the 0.5-megohm resistor at high frequencies than at low frequencies, and the frequency response at the grid is reasonably flat over a wide frequency range. Fig. 37 shows a comparison between the output of the crystal (curve A) and the output of the equalizing network (curve B). The response curve can be "flattened" still more if the attenuation at low frequencies is increased by changing the 0.5-megohm resistor to 0.125 megohm.

The tone-control network shown in Fig. 38 has two stages with completely separate bass and treble controls. Fig. 39 shows simplified representations of

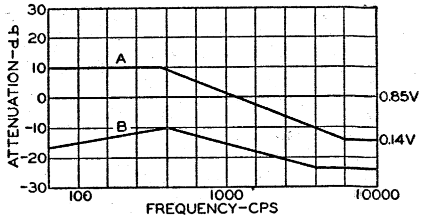

Fig. 37—Curve showing output from crystal phonograph pickup (A) and from equalizing network (B).

the bass control of this circuit when the potentiometer is turned to its extreme variations (usually labeled "Boost" and "Cut"). In this network, as in the crystal-equalizing network shown in Fig. 36, the parallel RC combination is the controlling factor. For bass "boost," the capacitor C_2 bypasses resistor R_3 so that less impedance is placed across the output to grid B at high frequencies than

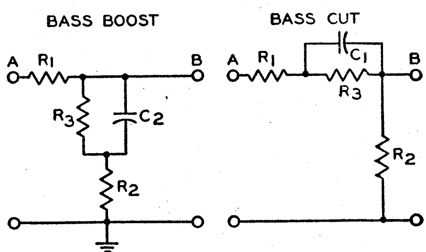

Fig. 39—Simplified representations of bass-control circuit at extreme ends of potentiometer.

at low frequencies. For bass "cut," the parallel combination is shifted so that C_1 bypasses R_3, causing more high-

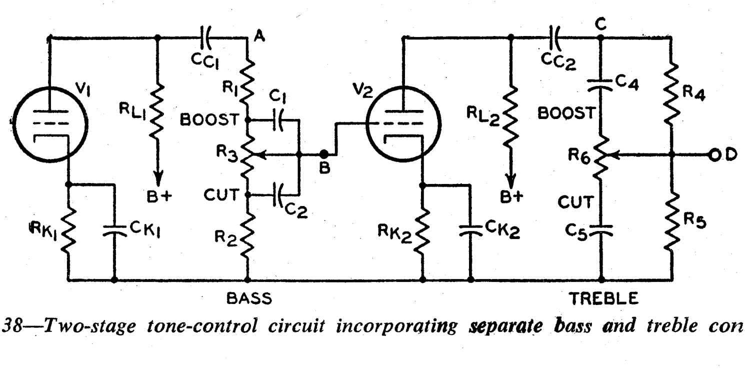

Fig. 38—Two-stage tone-control circuit incorporating separate bass and treble controls.

frequency than low-frequency output. Essentially, the network is a variable-frequency voltage divider. With proper values for the components, it may be made to respond to changes in the R_3 potentiometer setting for only low frequencies (below 1000 cycles).

Fig 40 shows extreme positions of the treble control. The attenuation of the two circuits is approximately the same at 1000 cycles. The treble "boost" circuit is similar to the crystal-equalizing network shown in Fig. 36. In the treble "cut" circuit, the parallel RC elements serve to attenuate the signal voltage further because the capacitor bypasses the resistance across the output.

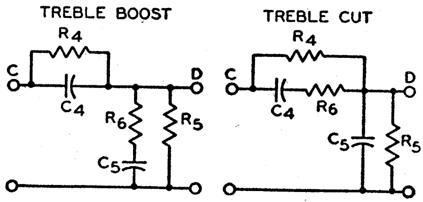

Fig. 40—Simplified representations of treble-control circuit at extreme ends of potentiometer.

The effect of the capacitor is negligible at low frequencies; beyond 1000 cycles, the signal voltage is attenuated at a maximum rate of 6 dB per octave.

The location of a tone-control network is of considerable importance. In a typical radio receiver, it may be inserted in the plate circuit of the power tube, the coupling circuit between the first af amplifier tube and the power tube, or the grid circuit of the first tube. In an amplifier using a beam power tube or pentode power amplifier without negative feedback, it is desirable to connect a resistance-capacitance filter across the primary of the output transformer. This filter may be fixed, with a supplementary tone control elsewhere, or it may form the tone control itself. If the amplifier incorporates negative feedback, the tone control may be inserted in the feedback network or else should be connected to a part of the amplifier which is external to the feedback loop. The overall gain of a well designed tone-control network should be approximately unity.

Phonograph and Tape Preamplifiers

The frequency range and dynamic range which can be recorded on a phonograph record or on magnetic tape depend on several factors, including the composition, mechanical characteristics, and speed of the record or tape, and the electrical and mechanical characteristics of the recording equipment. To achieve wide frequency and dynamic ranges, manufacturers of commercial recordings use equipment which introduces a nonuniform relationship between amplitude and frequency. This relationship is known as a "recording characteristic." To assure proper reproduction of a high-fidelity recording, therefore, some part of the reproducing system must have a frequency-response characteristic which is the inverse of the recording characteristic. Most manufacturers of high-fidelity recordings use the RCA "New Orthophonic" (RIAA) characteristic for discs and the NARTB characteristic for magnetic tape.

Some typical preamplifier stages are shown in the **Circuits** section. The location of the frequency-compensating network or "equalizer" in the reproducing system will depend on the types of recordings which are to be reproduced and on the pickup devices used.

A ceramic high-fidelity phonograph pickup is usually designed to provide proper compensation for the RIAA recording characteristic when the pickup is operated into the load resistance specified by its manufacturer. Because this type of pickup also has relatively high output (0.5 to 1.5 volts), it does not require the use of either an equalizer network or a preamplifier, and can be connected directly to the input of a tone-control amplifier and/or power amplifier.

A magnetic high-fidelity phonograph pickup, on the other hand, usually has an essentially flat frequency-response characteristic and very low output (1 to 10 millivolts). Because a pickup of this type merely reproduces the recording characteristic, it must be followed by an equalizer network, as well as by a preamplifier having sufficient voltage gain to provide the input voltage required by the tone-control amplifier and/or power amplifier. Many designs include both the equalizing and amplifying circuits in a single unit.

A high-fidelity magnetic-tape pickup head, like a magnetic phonograph pickup, reproduces the recording characteristic and has an output of only a few millivolts. This type of pickup device, therefore, must also be followed by an equalizing network and preamplifier, or by a preamplifier which provides "built-in" equalization for the NARTB characteristic.

Limiters

An amplifier may also be used as a limiter. One use of a limiter is in receivers designed for the reception of frequency-modulated signals. The limiter in FM receivers has the function of eliminating amplitude variations from the input to the detector. Because in an FM system amplitude variations are primarily the result of noise disturbances, the use of a limiter prevents such disturbances from being reproduced in the audio output. The limiter usually follows the last if stage so that it can minimize the effects of disturbances coming in on the rf carrier and those produced locally.

The limiter is essentially an if voltage amplifier designed for saturated operation. Saturated operation means that an increase in signal voltage above a certain value produces very little increase in plate current. A signal voltage which is never less than sufficient to cause saturation of the limiter, even on weak signals, is supplied to the limiter input by the preceding stages. Any change in amplitude, therefore, such as might be produced by noise voltage fluctuation, is not reproduced in the limiter output. The limiting action, of course, does not interfere with the reproduction of frequency variations.

Plate-current saturation of the limiter may be obtained by the use of grid-No. 1-resistor-and-capicitor bias with plate and grid-No. 2 voltages which are low compared with customary if-amplifier operating conditions.

As a result of these design features, the limiter is able to maintain its output voltage at a constant amplitude over a wide range of input-signal voltage variations. The output of the limiter is frequency-modulated if voltage, the mean frequency of which is that of the if amplifier. This voltage is impressed on the input of the detector.

The reception of FM signals without serious distortion requires that the response of the receiver be such that satisfactory amplification of the signal is provided over the entire range of frequency deviation from the mean frequency. Since the frequency at any instant depends on the modulation at that instant, it follows that excessive attenuation toward the edges of the band, in the rf or if stages, will cause distortion. In a high-fidelity receiver, therefore, the amplifiers must be capable of amplifying, for the maximum permissible frequency deviation of 75 kilocycles, a band 150 kilocycles wide. Suitable tubes for this purpose are the 6BA6 and 6BJ6.

Television Tuners

The vhf tuner of a television receiver selects the desired frequency channel in the range from 55 to 216 megacycles per second, amplifies it, and converts it to a lower intermediate frequency. These functions are accomplished in rf-amplifier, mixer, and local-oscillator stages employing tube types that are designed specifically for these applications. The rf-amplifier stage uses a high-transconductance tube that has small dimensions to maintain low interelectrode capacitances, particularly between grid and plate. The mixer and oscillator stages usually employ a dual-unit triode-pentode tube that has a high-transconductance pentode unit and a medium-mu triode unit.

Fig. 41 shows a simplified schematic diagram of a typical vhf television tuner. The balun converts the 300-ohm balanced antenna impedance to an unbalanced impedance of 75 ohms. The high-pass filter eliminates lower-frequency interference signals. The tuner is set to the desired frequency by simultaneous adjustment of the inductances

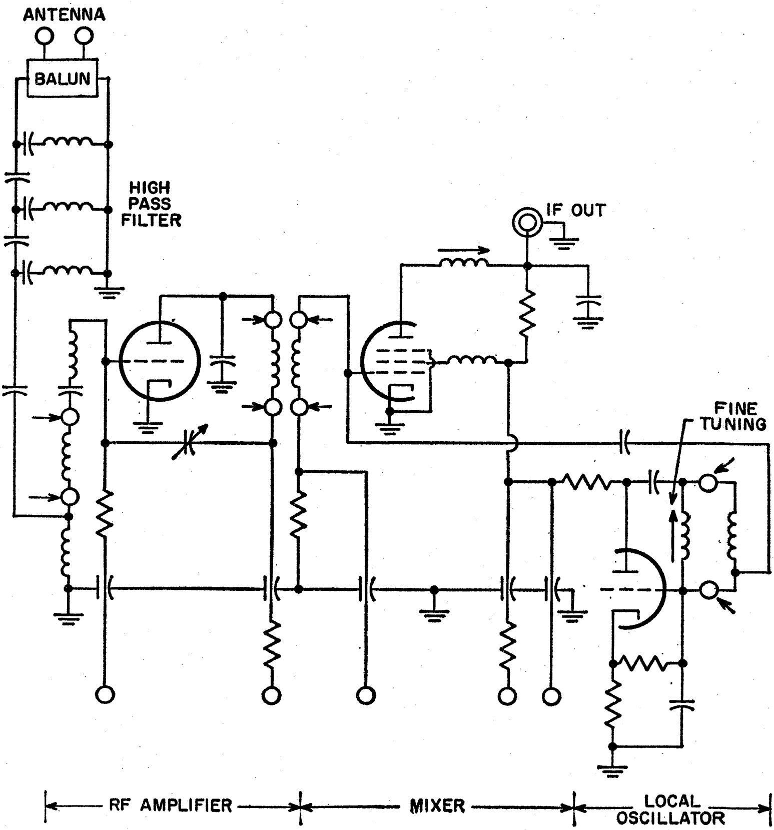

Fig. 41—Simplified schematic of typical vhf television tuner.

indicated by the several sets of arrows in Fig. 41. The inductances are either replaced completely or incremental amounts of inductance are added as the tuner is switched from high frequencies to lower frequencies. Some tuners use a combination of the two methods.

Because **noise** generated in the first amplifier stage is often the controlling factor in determining the over-all sensitivity of a radio or television receiver, the "front end" is designed with special attention to both gain and noise characteristics. The input circuit of an amplifier inherently contains some thermal noise contributed by the resistive elements in the input device. When an input signal is amplified, therefore, the thermal noise generated in the input circuit is also amplified. If the ratio of signal power to noise power **(signal-to-noise ratio, S/N)** is the same in the output circuit as in the input circuit, the amplifier is considered to be "noiseless," and is said to have a noise figure of unity, or zero dB.

In practical circuits, however, all amplifier stages generate a certain amount of noise as a result of thermal agitation of electrons in resistors and other components, minute variations in the cathode emission of tubes (shot effect), and minute grid currents in the amplifier tubes. As a result, the ratio of signal power to noise power is inevitably impaired during amplification. A measure of the degree of impairment is called the **noise figure** (NF) of the amplifier, and is expressed as the ratio of signal power to noise power at the input (S_i/N_i) divided by the ratio of signal power to noise power at the output (S_o/N_o), as follows:

$$NF = \frac{(S_i/N_i)}{(S_o/N_o)}$$

The noise figure in decibels (dB) is equal to ten times the logarithm of this power ratio. For example, a one-dB noise figure in an amplifier decreases the signal-to-noise ratio by a factor of 1.26, a 3-dB noise figure by a factor of 2, a 10-dB noise figure by a factor of 10, and a 20-dB noise figure by a factor of 100.

The over-all noise figure of a receiver is affected by the total number of stages, as shown by the following relationship:

$$NF_{receiver} = NF_1 + \frac{(NF_2 + 1)}{G_1} + \frac{(NF_3 + 1)}{G_1 G_2} \ldots$$

where G represents power gain and the subscripts indicate the number of each stage. This relationship indicates that the contribution of the second-stage noise factor to that of the over-all receiver is reduced by the gain of the first stage. Therefore, it is important that the rf amplifier have enough gain to make the effect of the second stage negligible. The third stage will then have even less effect. The maximum available power gain G of an rf stage is given by

$$G = \frac{g_m^2 R_{in} R_{out}}{4}$$

For maximum gain, therefore, the rf-amplifier tube should have high transconductance and high input and output impedances. At frequencies in the vhf television band, the input resistance is small enough to affect the gain. As mentioned previously, the rf tube is designed to have low interelectrode capacitances, small interelectrode spacings, and low lead inductances (particularly the cathode lead).

The gain of the rf stage must be reduced as the incoming-signal amplitude changes to prevent overload distortion in the following stages. As the signal amplitude increases, an automatic-gain-control (agc) circuit biases the rf tube to decrease its gain. The rf tube usually employs a semiremote-cutoff grid to reduce cross-modulation distortion.

Either a triode or a pentode can be used in the **rf-amplifier** stage of tuner input circuits of vhf television receivers. Such stages are required to amplify signals ranging from 55 to 216 megacycles per second and having a bandwidth of 4.5 megacycles per second (the tuner is usually aligned for a bandwidth of 6 megacycles per second to assure complete coverage of the band). In early rf tuners, pentodes rather than triodes were used because the grid-plate capacitance of triodes created stability problems. However, the use of twin triodes in direct-coupled cathode-drive circuits makes it possible to obtain stable operation along with the low-noise characteristics of triodes.

Pentodes or tetrodes do not provide the useful sensitivity of triodes because of the "partition noise" introduced by the screen grid. The direct-coupled cathode-drive circuit provides both the gain and the stability capabilities of the pentode, as well as the advantages of a low-noise triode input stage. Because the cathode-drive stage provides a low-impedance load to the grounded-cathode stage, the gain of the latter stage is very low and there is no necessity for neutralizing the grid-plate capacitance. An interstage impedance, usually an inductance in series with the plate of the first stage and the cathode of the second stage, is often used at higher frequencies to provide a degree of impedance matching between the units. The cathode-drive portion of the circuit is matched to the input network and provides most of the stage gain. Because the feedback path of the cathode-drive circuit is the plate-cathode capacitance, which in most cases is very small, excellent isolation is provided between the antenna and the local oscillator.

Development of single triodes having low grid-plate capacitance, such as the 6BN4, has made possible the design of neutralized triode rf circuits. Tubes such as the 6GK5 and 6CW4 are specially designed to minimize grid-plate capacitance to permit easier neutralization of a grounded-cathode circuit over the wide frequency band. Bridge-neutralized rf-amplifier stages are widely used in television tuners; in this arrangement, a portion of the output signal is returned to the grid out of phase with the feedback signal from the grid-plate capacitance. This circuit provides excellent gain and noise performance with stable operation across the band.

The **mixer** stage of a vhf tuner usually employs a pentode tube, or the pentode unit of a triode-pentode tube. Although triodes such as the 6J6 were used as mixers in early receivers, they have been replaced by pentodes because the higher output impedance of a pentode provides a higher mixer gain than can be obtained with a triode.

The amplified signal from the rf stage in Fig. 41 is applied to the mixer grid along with a local-oscillator signal of much larger amplitude. The local-oscillator signal varies the mixer grid voltage from cutoff into the grid-current region. This signal develops a grid-resistor bias, called the **injection voltage,** which is a measure of the local-oscillator voltage. Because the transfer curve of the mixer tube is nonlinear, mixing action between the rf signal and the local-oscillator signal produces sum and difference frequencies. The output circuit of the mixer is tuned to the difference frequency (about 44 megacycles per second) and rejects all other frequencies. This signal is then fed to the intermediate-frequency amplifier.

The mixer gain is a function of the amplitude of the local-oscillator signal. The gain has a broad maximum over a range of injection voltages from —2.5 to —5.0 volts for conventional-grid mixers and slightly lower for frame-grid mixers. Good impedance matching between the rf-amplifier plate and the mixer grid, consistent with bandpass requirements, is important to achieve maximum signal power transfer. A slight amount of regeneration is provided by a small screen-grid inductance. This regeneration effectively increases the mixer-grid input impedance and thus improves power gain.

The **local-oscillator** stage shown in Fig. 41 is a Colpitts type in which the tuned circuit is located between the grid and plate and the feedback path is through the tube interelectrode capacitances. A large signal is developed in the local oscillator and coupled loosely to the mixer grid to minimize the effects of changes in the mixer input on the frequency of oscillation. The circuit is designed to keep frequency shift within a very narrow range with supply-voltage and temperature changes. Fine tuning is provided by a variable inductance or capacitance across the tuned circuit. Tubes commonly used in local-oscillator and mixer circuits are the 6EA8, 6KZ8, and 6KE8.

Television IF Amplifiers

The intermediate-frequency (if) amplifier stages in a television receiver provide the additional gain required to bring the signal level to an amplitude suitable for final detection. A constant peak signal of about three to five volts is required at the input to the detector. The mixer output signal is passed through two or three stages of amplification to attain this level. High-transconductance pentodes having low grid-No.1-to-plate capacitances are normally used in if amplifiers. The coupling circuits are usually tuned transformers which may be single- or double-tuned. The transformers are either synchronously (same frequency) tuned or stagger-tuned, depending on circuit requirements. The over-all bandwidth varies from a maximum of 3.58 megacycles per second at the 6-dB points for color receivers to values in the order of 2.0 to 2.5 megacycles per second for the most inexpensive receivers. An expression for the figure of merit for a single-tuned if-amplifier tube is the gain-bandwith product G $\times$ B, which is given by

$$G \times B = \frac{g_m}{2\pi C}$$

where C is the total tuning capacitance. This relationship again demonstrates the need for high transconductance and low interelectrode capacitance.

The first stage (or first two stages in the case of a three-stage if) is gain-controlled like the rf amplifier. However, the bias applied to the if-amplifier tube varies the input resistance and capacitance of the tube and thus detunes the circuit. It is important for proper reception to maintain the frequency response of the if stages constant, particularly in the case of the color receiver. Therefore, a small unbypassed cathode resistor is used which provides degenerative feedback to minimize the effect of bias changes. In addition, the effects on input impedance caused by the grid-plate capacitance are reduced by use of a partial bypass capacitor at the screen grid to provide neutralization of the grid-to-plate capacitance.

Tubes used in the gain-controlled stages of the if amplifier have remote- or semiremote-cutoff characteristics to reduce cross-modulation or intermodulation interference. Tube types commonly used in this application include the 6BZ6, 6GM6, 6JH6, and 6JD6.

The last if-amplifier stage is a relatively-large-signal amplifier. For this reason, the tube must be biased so that it will operate over a region of linear operation for large voltage excursions. Because such a quiescent operating point provides a transconductance somewhat below the maximum value for the tube, the selection of the operating point involves a compromise between signal-handling capacity and gain. For purposes of linearity, the final if-amplifier stage is not gain-controlled, and operates with the cathode bypassed to ground. Because fixed bias is used, a sharp-cutoff tube is used to provide higher transconductance than could be obtained with an equivalent remote- or semiremote-cutoff tube. Examples of types used in this stage are the 6EW6 and 6JC6.

Video Amplifiers

The video amplifier stage in a television receiver usually employs a pentode-type tube specially designed to amplify the wide band of frequencies contained in the video signal and, at the same time, to provide high gain per stage. Pentodes are more useful than triodes in such stages because they have high transconductance (to provide high gain) together with low input and output interelectrode capacitances (to permit the broadband requirements to be satisfied). An approximate "figure of merit" for a particular tube for this application can be determined from the ratio of its transconductance, g_m, to the sum of its input and output capacitances, C_{in} and C_{out}, as follows:

$$\text{Figure of Merit} = \frac{g_m}{C_{in} + C_{out}}$$

Typical values for this figure are in the order of 500 x 10^6 or greater.

A typical video amplifier stage, such as that shown in Fig. 42, is connected between the second detector of the television receiver and the picture tube. The contrast control, R_1, in this circuit controls the gain of the video amplifier tube. The inductance, L_2, in series with the load resistor, R_L, maintains the plate load impedance at a relatively constant value with increasing

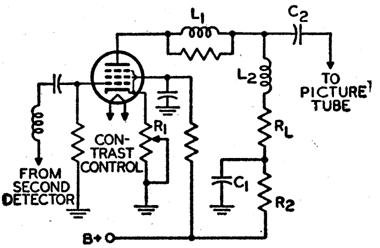

Fig. 42—Typical video amplifier stage.

frequency. The inductance L_1 isolates the output capacitance of the tube so that only stray capacitance is placed across the load. As a result, a higher-value load resistor is used to provide higher gain without affecting frequency response or phase relations. The decoupling circuit, C_1R_2, is used to improve the low-frequency response. Tubes used as video amplifiers include types 6CL6 and 12BY7A, or the pentode sections of types 6AW8A and 6AN8A.

The **luminance** amplifier in a color-television receiver is a conventional video amplifier having a bandwidth of approximately 3.5 megacycles per second. In a color receiver, the portion of the output of the second detector which lies within the frequency band from approximately 2.4 to 4.5 megacycles per second is fed to a bandpass amplifier, as shown in the block diagram in Fig. 43. The color synchronizing signal, or "burst," contained in this signal may then be fed to a "burst-keyer" tube. At the same time, a delayed horizontal pulse may be applied to the keyer tube. The output of the keyer tube is applied to the burst amplifier tube and the signal is then fed to the 3.58-megacycle oscillator and to the "color-killer" stage.

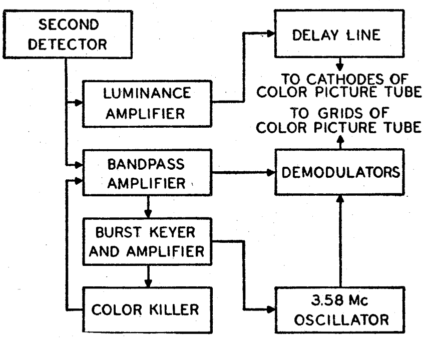

Fig. 43—Block diagram of video-amplifier section of color television receiver.

The color killer applies a bias voltage to the bandpass amplifier in the absence of burst so that the color section, or **chrominance** channel, of the receiver remains inoperative during black-and-white broadcasts. A threshold control varies the bias and controls the burst level at which the killer stage operates.

The output of the 3.58-megacycle oscillator and the output of the bandpass amplifier are fed into phase and amplitude demodulator circuits. The output of each demodulator circuit is an electrical representation of a color-difference signal, *i.e.*, an actual color signal minus the black-and-white, or luminance, signal. The two color-difference signals are combined to produce the third color-difference signal; each of the three signals then represents one of the primary colors.

The three color-difference signals are usually applied to the grids of the three electron guns of the color picture tube, in which case the black-and-white signal from the luminance amplifier may be applied simultaneously to the cathodes. The chrominance and luminance signals then combine to produce the color picture. In the absence of transmitted color information, the chrominance channel is cut off by the color killer, as described above, and only the luminance signal is applied to the picture tube, producing a black-and-white picture.

Television Sync Circuits

In addition to picture information, the composite video signal supplied to a television receiver contains information to assure that the picture produced on the receiver is synchronized with the picture being viewed by the camera or pickup tube. The "sync" pulses, which have a greater amplitude than the video signal, trigger the scanning generators of the receiver when the electron beam of the pickup tube ends each trace.

The sync pulses in the composite video signal may be separated from the video information in the output of the second or video detector by means of the triode circuit shown in Fig. 44. In this circuit, the time constant of the network R_1C_1 is long with respect to the interval between pulses. During each pulse, the grid is driven positive and draws current, thereby charging capacitor C_1. Consequently, the grid develops a bias which is slightly greater

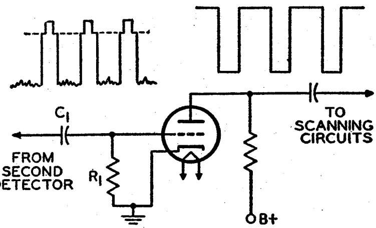

Fig. 44—Sync-separator circuit.

than the cutoff voltage of the tube. Because plate current flows only during the sync-pulse period, only the amplified pulse appears in the output. This **sync-separator** stage discriminates against the video information. Because the bias developed on the grid is proportional to the strength of the incoming signal, the circuit also has the advantage of being relatively independent of signal fluctuations.

Because the electron beam scans the face of the picture tube at different rates in the vertical and horizontal directions, the receiver incorporates two different scanning generators. The repetition rate of the vertical generator is 60 cycles per second, and the rate of the horizontal generator is approximately 15,750 cycles per second. The composite video signal includes information which enables each generator to derive its correct triggering. One horizontal sync pulse is supplied at the end of each horizontal line scan. At the end of each frame, several pulses of longer duration than the horizontal sync pulses are supplied to actuate the vertical generator. The vertical information is separated from the horizontal information by differentiating and integrating circuits.

In fringe areas, two conditions complicate the process of sync separation. First, the incoming signal available at the antenna is weak and susceptible to fading and other variations; second, the receiver is operating at or near maximum gain, which makes it extremely susceptible to interference from pulse-type noise generated by certain types of electrical equipment, ignition systems, switches, or the like. Some type of **noise-immunity** provision is almost essential for acceptable performance. Noise may be reduced or eliminated from the sync and agc circuits by gating or by a combination of gating, inversion, and cancellation. An example, of the latter method is shown in Fig. 45. In this circuit the 6GY6, which has two independent control grids, serves the dual function of agc amplifier and noise inverter. Because the sync tips of the video signal at grid No. 1 of the 6GY6 drive the tube near its cutoff region, any noise signal

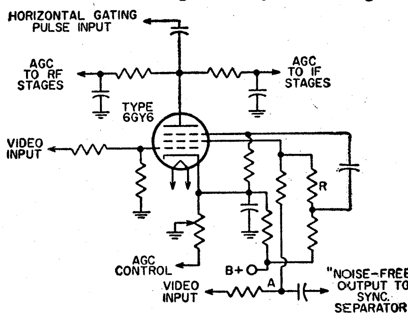

Fig. 45—Typical noise-cancellation circuit.

extending above the tip level will appear inverted across the grid-No.2 load resistor R. This inverted noise signal is re-combined with the video signal and fed to the sync separator at point "A" in Fig. 45, where noise cancellation takes place. This process leaves the sync pulses relatively free of disturbing noise and results in a stable picture. To prevent reduction of receiver gain due to the effect of noise on the agc amplifier, a portion of the inverted noise signal is fed to the second control grid, grid No.3, of the 6GY6 to cut off or gate the agc amplifier when a noise pulse occurs.

Rectification

The rectifying action of a diode finds important applications in supplying a receiver with dc power from an ac line and in supplying high dc voltage from a high-voltage pulse. A typical arrangement for converting ac to dc includes a rectifier tube, a filter, and a voltage divider. The rectifying action of the tube is explained briefly under **Diodes,** in the **Electrons, Electrodes,** and **Electron Tubes** section. High-voltage pulse rectification is described later under **Horizontal Output Circuits.**

The function of a filter is to smooth out the ripple of the tube output, as indicated in Fig. 46, and to increase rectifier efficiency. The action

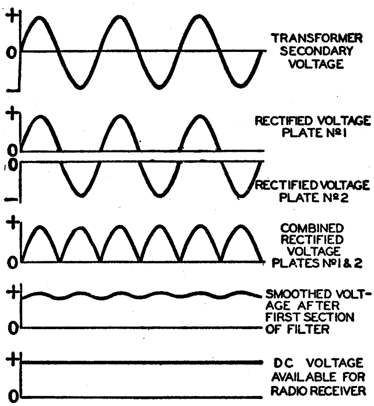

Fig. 46—Voltage waveforms of full-wave rectifier circuit.

of the filter is explained in the **Electron Tube Installation** section under **Filters.** The voltage divider is used to cut down the output voltage to the values required by the plates and the other electrodes of the tubes in the receiver.

A **half-wave rectifier** and a **full-wave rectifier** circuit are shown in Fig. 47. In the half-wave circuit, current flows through the rectifier tube to the filter on every other half-cycle of the ac input voltage when the plate is positive with respect to the cathode. In the full-wave circuit, current flows to the filter on every half-cycle, through plate No. 1 on one half-cycle when plate No. 1 is positive with respect to the cathode, and through plate No. 2 on the next half-cycle when plate No. 2 is positive with respect to the cathode.

Because the current flow to the filter is more uniform in the full-wave circuit than in the half-wave circuit, the output of the full-wave circuit requires less filtering. Rectifier operating information and circuits are given under each rectifier tube type and in the **Circuits** section, respectively.

Parallel operation of rectifier tubes furnishes an output current greater than that obtainable with the use of one tube. For example, when two full-wave rectifier tubes are connected in parallel, the plates of each tube are connected together and each tube acts as a half-wave rectifier. The permissible voltage and load conditions per tube are the same as for full-wave service but the total load-handling capability of the complete rectifier is approximately doubled.

When mercury-vapor rectifier tubes are connected in parallel, a stabilizing resistor of 50 to 100 ohms should be connected in series with each plate lead in order that each tube will carry an equal share of the load. The value of the resistor to be used will depend on the amount of plate current that passes through the rectifier. Low plate current requires a high value; high plate current, a low value. When the plates of mercury-vapor rectifier tubes are connected in parallel, the corresponding filament leads should be similarly connected. Otherwise, the tube drops will be considerably unbalanced and larger stabilizing resistors will be required.

Two or more vacuum rectifier tubes can also be connected in parallel

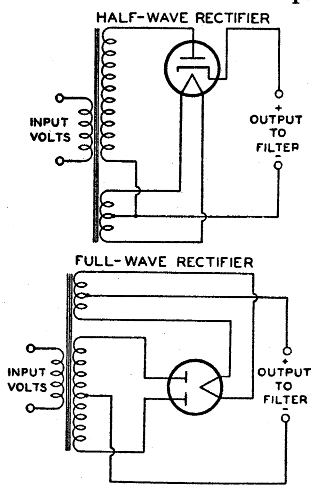

Fig. 47—Half-wave and full-wave rectifier circuits.

to give correspondingly higher output current and, as a result of paralleling their internal resistances, give somewhat increased voltage output. With vacuum types, stabilizing resistors may or may not be necessary depending on the tube type and the circuit.

A **voltage-doubler** circuit of simple form is shown in Fig. 48. The circuit derives its name from the fact that its

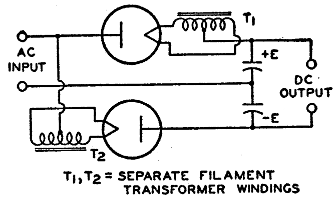

Fig. 48—Full-wave voltage-doubler circuit.

dc voltage output can be as high as twice the peak value of ac input. Basically, a voltage doubler is a rectifier circuit arranged so that the output voltages of two half-wave rectifiers are in series.

The action of a voltage doubler can be described briefly as follows. On the positive half-cycle of the ac input, that is, when the upper side of the ac input line is positive with respect to the lower side, the upper diode passes current and feeds a positive charge into the upper capacitor. As positive charge accumulates on the upper plate of the capacitor, a positive voltage builds up across the capacitor. On the next half-cycle of the ac input, when the upper side of the line is negative with respect to the lower side, the lower diode passes current so that a negative voltage builds up across the lower capacitor.

So long as no current is drawn at the output terminals from the capacitor, each capacitor can charge up to a voltage of magnitude E, the peak value of the ac input. It can be seen from the diagram that with a voltage of +E on one capacitor and —E on the other, the total voltage across the capacitors is 2E. Thus the voltage doubler supplies a no-load dc output voltage twice as large as the peak ac input voltage. When current is drawn at the output terminals by the load, the output voltage drops below 2E by an amount that depends on the magnitude of the load current and the capacitance of the capacitors. The arrangement shown in Fig. 48 is called a full-wave voltage doubler because each rectifier passes current to the load on each half of the ac input cycle.

Two rectifier types especially designed for use as voltage doublers are the 25Z6GT and 117Z6GT. These tubes combine two separate diodes in one tube. As voltage doublers, the tubes are used in "transformerless" receivers. In these receivers, the heaters of all tubes in the set are connected in series with a voltage-dropping resistor across the line. The connections for the heater supply and the voltage-doubling circuit are shown in Fig. 49.

With the full-wave voltage-doubler circuit in Fig. 49, it will be noted that the dc load circuit can not be connected to ground or to one side of the ac supply line. This circuit presents certain disadvantages when the heaters of all the tubes in the set are connected in series

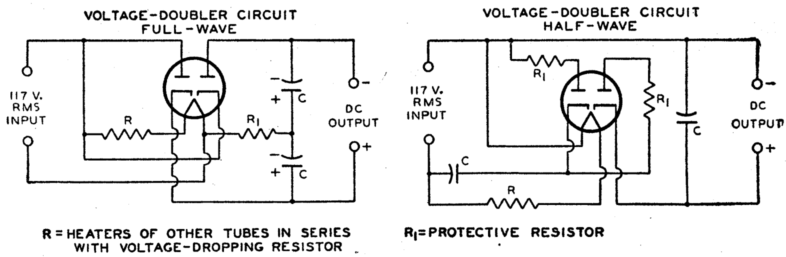

Fig. 49—Full-wave and half-wave voltage-doubler circuits showing heater-supply connections.

with a resistance across the ac line. Such a circuit arrangement may cause hum because of the high ac potential between the heaters and cathodes of the tubes.

The half-wave voltage-doubler circuit in Fig. 49 overcomes this difficulty by making one side of the ac line common with the negative side of the dc load circuit. In this circuit, one half of the tube is used to charge a capacitor which, on the following half cycle, discharges in series with the line voltage through the other half of the tube. This circuit is called a half-wave voltage doubler because rectified current flows to the load only on alternate halves of the ac input cycle. The voltage regulation of this arrangement is somewhat poorer than that of the full-wave voltage doubler.

Detection

When speech, music, or video information is transmitted from a radio or television station, the station radiates a radio-frequency (rf) wave which is of either of two general types. In one type, the wave is said to be amplitude modulated when its frequency remains constant and the amplitude is varied. In the other type, the wave is said to be frequency modulated when its amplitude remains essentially constant but its frequency is varied.

The function of the receiver is to reproduce the original modulating wave from the modulated rf wave. The receiver stage in which this function is performed is called the **demodulator** or detector stage.

AM Detection

The effect of **amplitude modulation** on the waveform of the rf wave is shown in Fig 50. There are three different basic circuits used for the detection of amplitude-modulated waves: the diode detector, the grid-bias detector, and the grid-resistor detector. These circuits are alike in that they eliminate, either partially or completely, alternate half-cycles of the rf wave. With alternate half-cycles removed, the audio variations of the other half-cycles can be amplified to drive headphones or a loudspeaker.

A **diode-detector** circuit is shown in Fig. 51. The action of this circuit when a modulated rf wave is applied is

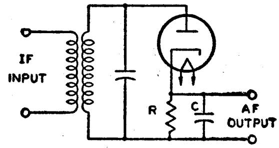

Fig. 51—Basic diode-detector circuit.

illustrated by Fig. 52. The rf voltage applied to the circuit is shown in light line; the output voltage across capacitor C is shown in heavy line.

Between points (a) and (b) on the first positive half-cycle of the applied rf voltage, capacitor C charges up to the peak value of the rf voltage. Then as the applied rf voltage falls away from its peak value, the capacitor holds the cathode at a potential more positive than the voltage applied to the anode. The capacitor thus temporarily cuts off current through the diode. While the diode current is cut off, the capacitor discharges from (b) to (c) through the diode load resistor R.

When the rf voltage on the anode rises high enough to exceed the potential at which the capacitor holds the cathode, current flows again and the capacitor charges up to the peak value of the second positive half-cycle at (d). In this way, the voltage across the capacitor follows the peak value of the applied rf

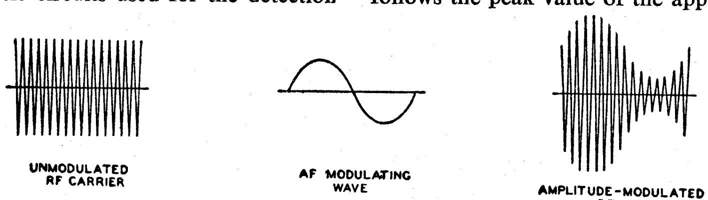

Fig. 50—Waveforms showing effect of amplitude modulation on an rf wave.

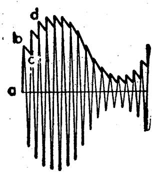

Fig. 52—Waveforms showing modulated rf input (light line) and output voltage (heavy line) of diode-detector circuit.

voltage and reproduces the af modulation.

The curve for voltage across the capacitor, as shown in Fig. 52, is somewhat jagged. However, this jaggedness, which represents an rf component in the voltage across the capacitor, is exaggerated in the drawing. In an actual circuit the rf component of the voltage across the capacitor is negligible. Hence, when the voltage across the capacitor is amplified, the output of the amplifier reproduces the speech or music originating at the transmitting station.

Another way to describe the action of a diode detector is to consider the circuit as a half-wave rectifier. When the rf signal on the plate swings positive, the tube conducts and the rectified current flows through the load resistance R. Because the dc output voltage of a rectifier depends on the voltage of the ac input, the dc voltage across C varies in accordance with the amplitude of the rf carrier and thus reproduces the af signal. Capacitor C should be large enough to smooth out rf or if variations, but should not be so large as to affect the audio variations. Two diodes can be connected in a circuit similar to a full-wave rectifier to provide full-wave detection. However, in practice, the advantages of this connection generally do not justify the extra circuit complication.

The diode method of detection produces less distortion than other methods because the dynamic characteristics of a diode can be made more linear than those of other detectors. The disadvantages of a diode are that it does not amplify the signal, and that it draws current from the input circuit and therefore reduces the selectivity of the input circuit. However, because the diode method of detection produces less distortion and because it permits the use of simple avc circuits without the necessity for an additional voltage supply, the diode method of detection is most widely used in broadcast receivers.

A typical diode-detector circuit using a twin-diode—triode tube is shown in Fig. 53. Both diodes are connected

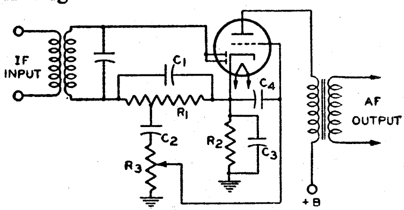

Fig. 53—Typical diode-detector circuit using a twin diode—triode tube.

together. R_1 is the diode load resistor. A portion of the af voltage developed across this resistor is applied to the triode grid through the volume control R_3. In a typical circuit, resistor R_1 may be tapped so that five-sixths of the total af voltage across R_1 is applied to the volume control. This tapped connection reduces the af voltage output of the detector circuit slightly, but it reduces audio distortion and improves the rf filtering.

DC bias for the triode section is provided by the cathode-bias resistor R_2 and the audio bypass capacitor C_3. The function of capacitor C_2 is to block the dc bias of the cathode from the grid. The function of capacitor C_4 is to bypass any rf voltage on the grid to cathode. A twin-diode—pentode may also be used in this circuit. With a pentode, the af output should be resistance-coupled rather than transformer-coupled.

Another diode-detector circuit, called a diode-biased circuit, is shown in Fig. 54. In this circuit, the triode grid

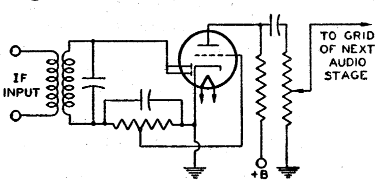

Fig. 54—Diode-biased detector circuit.

is connected directly to a tap on the diode load resistor. When an rf signal voltage is applied to the diode, the dc voltage at the tap supplies bias to the triode grid. When the rf signal is modulated, the af voltage at the tap is applied to the grid and is amplified by the triode.

The advantage of the circuit shown in Fig. 54 over the self-biased arrangement shown in Fig. 53 is that the diode-biased circuit does not employ a capacitor between the grid and the diode load resistor, and consequently does not produce as much distortion of a signal having a high percentage of modulation.

However, there are restrictions on the use of the diode-biased circuit. Because the bias voltage on the triode depends on the average amplitude of the rf voltage applied to the diode, the average amplitude of the voltage applied to the diode should be constant for all values of signal strength at the antenna. Otherwise there will be different values of bias on the triode grid for different signal strengths and the triode will produce distortion. Because there is no bias applied to the diode-biased triode when no rf voltage is applied to the diode, sufficient resistance should be included in the plate circuit of the triode to limit its zero-bias plate current to a safe value.

These restrictions mean, in practice, that the receiver should have a separate-channel automatic-volume-control (avc) system. With such an avc system, the average amplitude of the signal voltage applied to the diode can be held within very close limits for all values of signal strength at the antenna.

The tube used in a diode-biased circuit should be one which operates at a fairly large value of bias voltage. The variations in bias voltage are then a small percentage of the total bias and hence produce small distortion. Tubes taking a fairly large bias voltage are types such as the 6BF6 or 6SR7 having a medium-mu triode. Tube types having a high-mu triode or a pentode should not be used in a diode-biased circuit.

A **grid-bias detector** circuit is shown in Fig. 55. In this circuit, the grid is biased almost to cutoff, *i.e.*, operated so that the plate current with zero signal is practically zero. The bias voltage can be obtained from a cathode-bias resistor, a C-battery, or a bleeder

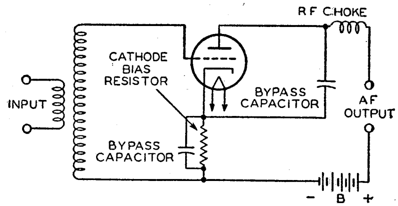

Fig. 55—Grid-bias detector circuit.

tap. Because of the high negative bias, only the positive half-cycles of the rf signal are amplified by the tube. The signal is, therefore, detected in the plate circuit. The advantages of this method of detection are that it amplifies the signal, besides detecting it, and that it does not draw current from the input circuit and therefore does not reduce the selectivity of the input circuit.

The **grid-resistor-and-capacitor method,** illustrated in Fig. 56, is somewhat more sensitive than the grid-bias method and gives its best results on weak signals. In this circuit, there is no negative dc bias voltage applied to the grid. Hence, on the positive half-cycles of the rf signal, current flows from grid to cathode. The grid and cathode thus

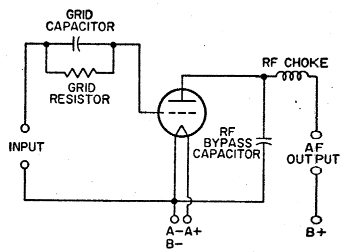

Fig. 56—Detector circuit using grid-resistor-and-capacitor bias.

act as a diode detector, with the grid resistor as the diode load resistor and the grid capacitor as the rf bypass capacitor. The voltage across the capacitor then reproduces the af modulation in the same manner as has been explained for the diode detector. This voltage appears between the grid and cathode and is therefore amplified in the plate circuit. The output voltage thus reproduces the original af signal.

In this detector circuit, the use of a high-resistance grid resistor increases selectivity and sensitivity. However, improved af response and stability are obtained with lower values of grid-circuit resistance. This detector circuit amplifies the signal, but draws current from the input circuit and therefore reduces the selectivity of the input circuit.

FM Detection

The effect of **frequency modulation** on the waveform of the rf wave is shown in Fig. 57. In this type of transmission, the frequency of the rf wave deviates from a mean value, at an rf rate depending on the modulation, by an amount that is determined in the transmitter and is proportional to the amplitude of the af modulation signal.

For this type of modulation, a detector is required to discriminate between deviations above and below the mean frequency and to translate those

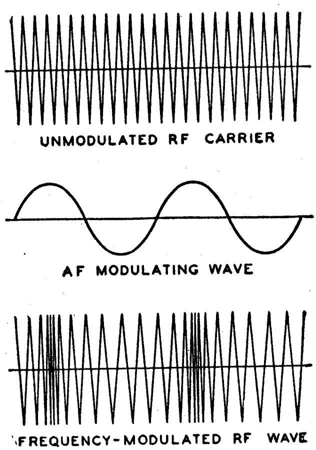

Fig. 57—Waveforms showing effect of frequency modulation on an rf wave.

deviations into a voltage whose amplitude varies at audio frequencies. Since the deviations occur at an audio frequency, the process is one of demodulation, and the degree of frequency deviation determines the amplitude of the demodulated (af) voltage.

A simple circuit for converting frequency variations to amplitude variations is a circuit which is tuned so that the mean radio frequency is on one slope of its resonance characteristic, as at A of Fig. 58. With modulation, the frequency swings between B and C, and the voltage developed across the circuit varies at the modulating rate. In order that no distortion will be introduced in this circuit, the frequency swing must be restricted to the portion of the slope which is effectively straight. Since this portion is very short, the voltage developed is low. Because of these limitations, this circuit is not commonly used but it serves to illustrate the principle.

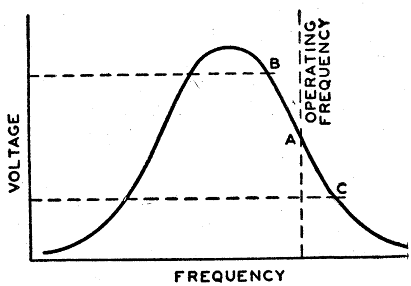

Fig. 58—Resonance curve showing desired operating range for frequency-modulation converter.

The faults of the simple circuit are overcome in a push-pull arrangement, sometimes called a **discriminator circuit,** such as that shown in Fig. 59. Because of the phase relationships between

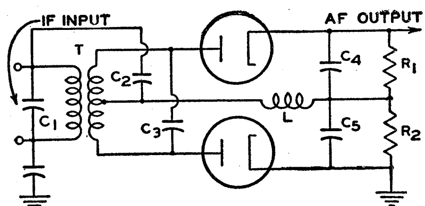

Fig. 59—Basic discriminator circuit.

the primary and each half of the secondary of the input transformer (each half of the secondary is connected in series with the primary through capacitor C_2), the rf voltages applied to the diodes become unequal as the rf signal swings from the resonant frequency in each direction.

Because the swing occurs at audio frequencies (determined by the af modulation), the voltage developed across the diode load resistors, R_1 and R_2 connected in series, varies at audio frequencies. The output voltage depends on the difference in amplitude of the voltages developed across R_1 and R_2. These voltages are equal and of opposite sign when the rf carrier is not modulated and the output is, therefore, zero. When modulation is applied, the output voltage varies as shown in Fig. 60.

Because this type of FM detector is sensitive to amplitude variations in the rf carrier, a limiter stage is frequently used to remove most of the amplitude modulation from the carrier. (See **Limiters** under **Amplification.**)

Another form of detector for frequency-modulated waves is called a **ratio detector.** This FM detector, unlike the previous one which responds to a difference in voltage, responds only to changes in the ratio of the voltage across two diodes and is, therefore, insensitive to changes in the differences in the voltages due to amplitude modulation of the rf carrier.

The basic ratio detector is given in Fig. 61. The plate load for the final if amplifier stage is the parallel resonant circuit consisting of C_1 and the primary transformer T. The tuning and coupling of the transformer are practically the

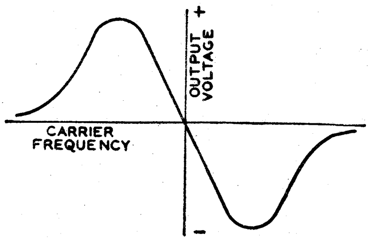

Fig. 60—Output waveform of discriminator circuit.

same as in the previous circuit and, therefore, the rf voltages applied to the diodes depend upon how much the rf signal swings from the resonant frequency in each direction. At this point the similarity ends.

Diode 1, R_2, and diode 2 complete a series circuit fed by the secondary of the transformer T. The two diodes are connected in series so that they conduct on the same rf half-cycle. The rectified current through R_2 causes a negative voltage to appear at the plate of diode 1. Because C_6 is large, this negative voltage at the plate of diode 1 remains constant even at the lowest audio frequencies to be reproduced.

The rectified voltage across C_3 is proportional to the voltage across diode 1, and the rectified voltage across C_4 is proportional to the voltage across diode 2. Because the voltages across the two diodes differ according to the instantaneous frequency of the carrier, the voltages across C_3 and C_4 differ proportionately, the voltage across C_3 being the larger of the two voltages at carrier frequencies below the intermediate frequency and the smaller at frequencies above the intermediate frequency.

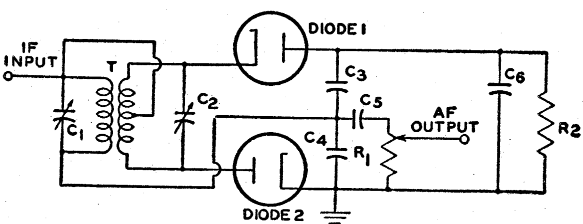

Fig. 61—Basic ratio-detector circuit.

These voltages across C_3 and C_4 are additive and their sum is fixed by the constant voltage across C_6. Therefore, while the ratio of these voltages varies at an audio rate, their sum is always constant. The voltage across C_4 varies at an audio rate when a frequency-modulated rf carrier is applied to the ratio detector; this audio voltage is extracted and fed to the audio amplifier. For a complete circuit utilizing this type of detector, refer to the **Circuits** section.

Color Demodulation

In the transmission of picture signals for color-television receivers, all the color information is contained in three signals, a luminance (black-and-white) or monochrome signal and two chrominance signals. The luminance signal, which is called the Y signal, contains brightness information only. The voltage response of the Y signal is made similar to the brightness response of the human eye by use of a composite signal that contains definite proportions of the red, green, and blue signals from the color-television camera (30 per cent red, 59 per cent green, and 11 per cent blue). This Y signal, which includes sync and blanking pulses, provides a correct monochrome picture in a conventional black-and-white television receiver.

For the generation of color-television signals, the Y signal is subtracted from the red, green, and blue signals to provide a new set of color-difference signals, which are designated as R-Y, B-Y, and G-Y. All of the original picture information is contained in the Y signal, the R-Y signal, and the B-Y signal. Therefore, the G-Y signal is not contained in the transmitted signal, but is synthesized in the receiver by proper combination of the R-Y and B-Y signals.

(Color signals transmitted under present color-television standards are not R-Y and B-Y, but a similar pair of signals designated as I and Q. In the color-television receiver, R-Y and B-Y signals are demodulated directly from the I and Q signals with negligible loss of color quality. For purposes of simplicity, only R-Y and B-Y signals are considered in this explanation. In addition, a 90-degree phase-shift network is shown; the phase-shift angle could be, and often is, some other value.)

Because the luminance signal and the two color-difference signals must be transmitted with a standard 6-megacycle channel, the two color signals are combined into one signal at the transmitter and are independently recovered at the receiver by proper detection techniques. A color subcarrier of approximately 3.58 megacycles per second is used for transmitting the color information within the 6-megacycle spectrum of the television station. As shown in Fig. 62, the 3.58-megacycle subcarrier and one of the color-difference signals are applied directly to a balanced AM modulator. The other color-difference signal is applied directly to a second balanced AM modulator, and the 3.58-megacycle subcarrier is applied to this second modulator through a 90-degree phase-shifting network. The balanced modulators effectively cancel both the individual color-

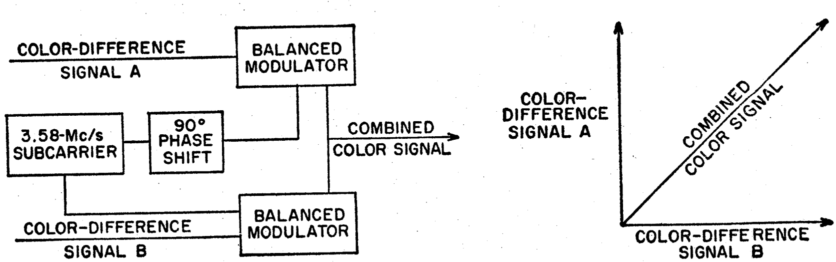

Fig. 62—Formation of combined color signal for transmission.

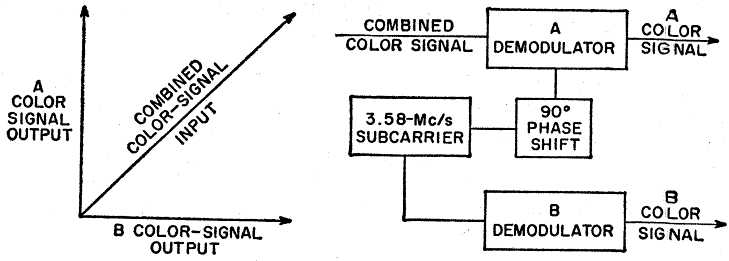

Fig. 63—Separation of combined color signal into two signals at the receiver.

difference signals and the subcarrier signal, and the output contains only the side-bands of the combined chrominance signal.

Recovery of the color information at the receiver involves a process called **synchronous detection.** In this process, two separate detectors are used to recover the separate color information, just as two separate modulators were used to combine the information at the transmitter. The 3.58-megacycle subcarrier, which was suppressed during transmission, must be reinserted at the receiver for recovery of the color information. The basis of synchronous detection is the phase relationship of this reinserted 3.58-megacycle subcarrier.

For example, the original color information is represented in Fig. 62 by the color-difference signals A and B. At the receiver, the combined color signal is fed to two demodulators A and B, as shown in Fig. 63. At the same time, a 3.58-megacycle subcarrier is also fed to the two demodulators, with the same phase relationship that was used in the modulators at the transmitter. This locally generated subcarrier essentially duplicates or replaces the original subcarrier, which was removed at the transmitter.

The local 3.58-megacycle oscillator in the color-television receiver is made to function at the proper frequency and phase by means of a synchronizing signal sent out by the transmitter. This synchronizing signal consists of a short **burst** of 3.58-megacycle signals transmitted during the horizontal blanking interval, immediately after the horizontal sync pulse, as shown in Fig. 64.

Fig. 65 shows a simplified diagram of a low-level color demodulator frequently used in color-television receivers. The locally generated 3.58-megacycle signal is applied to the grid No. 3

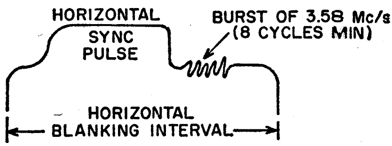

Fig. 64—Waveform for synchronizing signal.

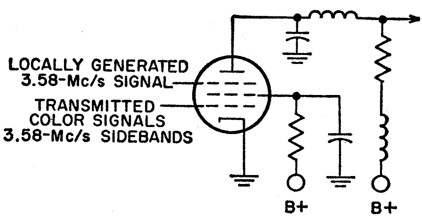

Fig. 65—Low-level color demodulator.

of the pentode. The transmitted color signal containing the 3.58-megacycle sidebands is applied to grid No. 1. The phase of the 3.58-megacycle color signal constantly changes in accordance with its color content. For example, the following table shows six variations in color (hue) as a function of subcarrier phase:

Subcarrier Phase-degrees (with respect to 3.58-megacycle local signal in phase with burst)	Hue
13	Yellow
77	Red
119	Magenta
193	Blue
257	Cyan
299	Green

The basic operating principle of the color demodulator shown in Fig. 65 is that plate current from the pentode is zero (or quite low) unless both grid No. 1 and grid No. 3 are simultaneously positive. For example, when the signals applied to the two grids are in phase, plate current can be expected to flow for 180 degrees of each ac cycle. Conversely, when the signals are 180 degrees out of phase, plate current is cut off. The output signal from the detector, therefore, is a function of the phase relationship between the transmitted color signal and the locally generated subcarrier.

In a typical color-television receiver, two color demodulators of the type shown in Fig. 65 are required. In one demodulator, the 3.58-megacycle subcarrier signal is applied directly to the pentode grid No. 3 from the local "burst" oscillator. In the other demodulator, the 3.58-megacycle signal from the burst oscillator is shifted 90 degrees in phase before it is applied to the pentode grid No. 3. As shown previously in Fig. 63, the demodulator B produces R-Y signals. These B-Y and R-Y signals are then combined (matrixed) to produce the G-Y signal, as discussed earlier. The complete luminance signal is then amplified to the required level in a conventional video-amplifier circuit.

In some color-television receivers, the demodulators are designed so that the color output signals can be applied directly to the color picture tube. In the diagram shown in Fig. 66, for example, the 6JH8 sheet-beam demodulators produce both positive and negative B-Y and R-Y signals. The positive signals are applied directly to the control grids (grid No. 1) of the blue and red guns of the color picture tube. At the same time, the negative color-difference signals are added (matrixed) in the correct proportions to produce the G-Y signal, which is applied to grid No. 1 of the green gun.

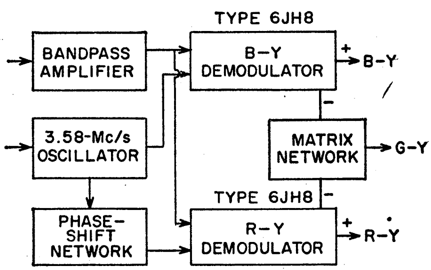

Fig. 66—Block diagram of demodulator circuit used to apply signals directly to color picture tube.

Automatic Volume or Gain Control

The chief purpose of automatic volume control **(avc)** or automatic gain control **(agc)** in a radio or television receiver is to prevent fluctuations in loudspeaker volume or picture brightness when the audio or video signal at the antenna is fading in and out.

An **automatic volume control** circuit regulates the receiver rf and if gain so that this gain is less for a strong signal than for a weak signal. In this way, when the signal strength at the antenna changes, the avc circuit reduces the resultant change in the voltage output of the last if stage and consequently reduces the change in the speaker output volume.

The avc circuit reduces the rf and if gain for a strong signal usually by increasing the negative bias of the rf, if, and frequency-mixer stage when the signal increases. A simple avc circuit is shown in Fig. 67. On each positive half-cycle of the signal voltage, when the diode plate is positive with respect to the cathode, the diode passes current.

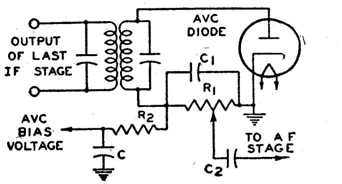

Fig. 67—Automatic-volume-control (avc) circuit.

Because of the flow of diode current through R_1, there is a voltage drop across R_1 which makes the left end of R_1 negative with respect to ground. This voltage drop across R_1 is applied, through the filter R_2 and C, as negative bias on the grids of the preceding stages. When the signal strength at the antenna increases, therefore, the signal applied to the avc diode increases, the voltage drop across R_1 increases, the negative bias voltage applied to the rf and if stages increases, and the gain of the rf and if stages is decreased. Thus the increase in signal strength at the antenna does not produce as much increase in the output of the last if stage as it would produce without avc.

When the signal strength at the antenna decreases from a previous steady value, the avc circuit acts, of course, in the reverse direction, applying less negative bias, permitting the rf and if gain to increase, and thus reducing the decrease in the signal output of the last if stage. In this way, when the signal strength at the antenna changes, the avc circuit acts to reduce change in the output of the last if stage, and thus acts to reduce change in loudspeaker volume.

The filter, C and R_2, prevents the avc voltage from varying at audio frequency. The filter is necessary because the voltage drop across R_1 varies with the modulation of the carrier being received. If avc voltage were taken directly from R_1 without filtering, the audio variations in avc voltage would vary the receiver gain so as to smooth out the modulation of the carrier. To avoid this effect, the avc voltage is taken from the capacitor C. Because of the resistance R_2 in series with C, the capacitor C can charge and discharge at only a comparatively slow rate. The avc voltage therefore cannot vary at frequencies as high as the audio range but can vary at frequencies high enough to compensate for most fading. Thus the filter permits the avc circuit to smooth out variations in signal due to fading, but prevents the circuit from smoothing out audio modulation.

It will be seen that an avc circuit and a diode-detector circuit are much alike. It is therefore convenient in a receiver to combine the detector and the avc diode in a single stage. Examples of how these functions are combined in receivers are shown in **Circuits** section.

In the circuit shown in Fig. 67, a certain amount of avc negative bias is applied to the preceding stages on a weak signal. Because it may be desirable to maintain the receiver rf and if gain at the maximum possible value for a weak signal, avc circuits are designed in some cases to apply no avc bias until the signal strength exceeds a certain value. These avc circuits are known as **delayed avc or davc** circuits.

A davc circuit is shown in Fig. 68. In this circuit, the diode section D_1 of the 6H6 acts as detector and avc diode.

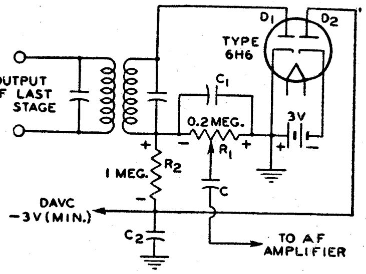

Fig. 68—Delayed avc (davc) circuit.

R_1 is the diode load resistor and R_2 and C_2 are the avc filter. Because the cathode of diode D_2 is returned through a fixed supply of —3 volts to the cathode of D_1, a dc current flows through R_1 and R_2 in series with D_2. The voltage drop caused by this current places the avc lead at approximately —3 volts (less the negligible drop through D_2). When the average amplitude of the rectified signal developed across R_1 does not exceed 3 volts, the avc lead remains at —3 volts. Hence, for signals not strongh enough to develop 3 volts across R_1, the bias applied to the controlled tubes stays constant at a value giving high sensitivity.

However, when the average amplitude of rectified signal voltage across R_1 exceeds 3 volts, the plate of diode D_2 becomes more negative than the cathode of D_2 and current flow in diode D_2 ceases. The potential of the avc lead is then controlled by the voltage developed across R_1. Therefore, with further increase in signal strength, the avc circuit applies an increasing avc bias voltage to the controlled stages. In this way, the circuit regulates the receiver gain for strong signals, but permits the gain to stay constant at a maximum value for weak signals.

It can be seen in Fig. 68 that a portion of the −3 volts delay voltage is applied to the plate of the detector diode D_1, this portion being approximately equal to $R_1/(R_1 + R_2)$ times −3 volts. Hence, with the circuit constants as shown, the detector plate is made negative with respect to its cathode by approximately one-half volt. However, this voltage does not interfere with detection because it is not large enough to prevent current flow in the tube.

Automatic gain control (agc) compensates for fluctuations in rf picture carrier amplitude. The peak carrier level rather than the average carrier level is controlled by the agc voltage because the peaks of the sync pulses are fixed when inserted on a fixed carrier level. The peak carrier level may be determined by measurement of the peaks of the sync pulses at the output of the video detector.

A conventional agc circuit, such as that shown in Fig. 69, consists of a diode

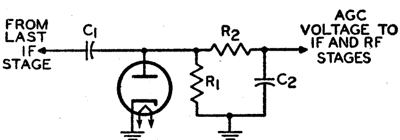

Fig. 69—Automatc-gain control (agc) circuit.

detector circuit and an RC filter. The time constant of the detector circuit is made large enough to prevent the picture content from influencing the magnitude of the agc voltage. The output voltage (agc voltage) is equal to the peak value of the incoming signal.

The diode detector receives the incoming signal from the last if stage of the television receiver through the capacitor C_1. The resistor R_1 provides the load for the diode. The diode conducts only when its plate is driven positive with respect to its cathode. Electrons then flow from the cathode to the plate and thence into capacitor C_1, where the negative charge is stored. Because of the low impedance offered by the diode during conduction, C_1 charges up to the value of the peak applied voltage.

During the negative excursion of the signal, the diode does not conduct, and C_1 discharges through resistor R_1. Because of the large time constant of R_1C_1, however, only a small percentage of the voltage across C_1 is lost during the interval between horizontal sync pulses. During succeeding positive cycles, the incoming signal must overcome the negative charge stored in C_1 before the diode conducts, and plate current flows only at the peak of each positive cycle. The voltage across C_1, therefore, is determined by the level of the peaks of the positive cycles, or the sync pulses.

The negative voltage developed across resistor R_1 by the sync pulses is filtered by resistor R_2 and capacitor C_2 to remove the 15,750-cycle ripple of the horizontal sync pulse. The dc output is then fed to the if and rf amplifiers as an agc voltage.

This agc system may be expanded to include amplification of the agc signal before detection of the peak level, or amplification of the dc output, or both. A direct-coupled amplifier must be used for amplification of the dc signal. The addition of amplification makes the system more sensitive to changes in carrier level.

A **"keyed" agc system** such as that shown in Fig. 70 is used to eliminate flutter and to improve noise immunity in weak signal areas. This system provides more rapid action than the conventional agc circuits because the filter circuit can employ lower capacitance and resistance values.

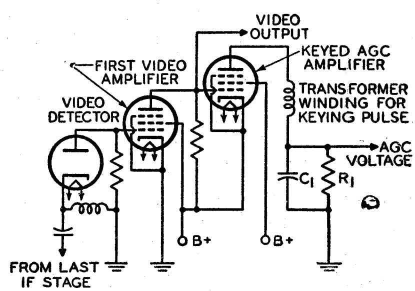

Fig. 70—"Keyed" agc circuit.

In the keyed agc system, the negative output of the video detector is fed directly to the grid No. 1 of the first video amplifier. The positive output of the video amplifier is, in turn, fed directly to the grid No. 1 of the keyed agc amplifier. The video stage increases the gain of the agc system and, in addition, provides noise clipping. The plate voltage for the agc amplifier is a positive pulse obtained from a small winding on the horizontal output transformer which is in phase with the horizontal sync pulse obtained from the video amplifier. The polarity of this pulse is such that the plate of the agc amplifier tube is positive during the retrace time. The tube is biased so that current flows only when the grid No. 1 and the plate are driven positive simultaneously. The amount of current flow depends on the grid-No. 1 potential during the pulse. These pulses are smoothed out in the RC network in the plate circuit (R_1C_1). Because the dc voltage developed across R_1 is negative, it is suitable for application to the grids of the rf and if tubes as an agc voltage.

Tuning Indication With Electron-Ray Tubes

Electron-ray tubes are designed to indicate visually by means of a fluorescent target the effects of a change in controlling voltage. One application of them is as tuning indicators in radio receivers. Types such as the 6U5, 6E5, and the 6AB5/6N5 contain two main parts: (1) a triode which operates as a dc amplifier and (2) an electron-ray indicator which is located in the bulb as shown in Fig. 71. The target is operated at a positive voltage and, therefore, attracts electrons from the cathode. When the electrons strike the target they produce a glow on the fluorescent coating

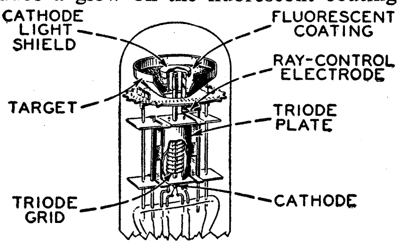

Fig. 71—Structure of electron-ray tube.

of the target. Under these conditions, the target appears as a ring of light.

A ray-control electrode is mounted between the cathode and target. When the potential of this electrode is less positive than the target, electrons flowing to the target are repelled by the electrostatic field of the electrode, and do not reach that portion of the target behind the electrode. Because the target does not glow where it is shielded from electrons, the control electrode casts a shadow on the glowing target. The extent of this shadow varies from approximately 100 degrees of the target when the control electrode is much more negative than the target to 0 degrees when the control electrode is at approximately the same potential as the target.

In the application of the electron-ray tube, the potential of the control electrode is determined by the voltage on the grid of the triode section, as can be seen in Fig. 72. The flow of the triode plate current through resistor R produces a voltage drop which determines

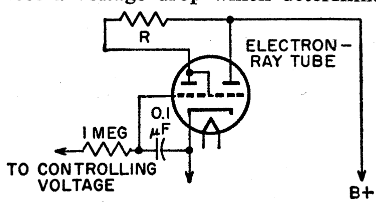

Fig. 72—Indicating circuit using an electron-ray tube.

the potential of the control electrode. When the voltage of the triode grid changes in the positive direction, plate current increases, the potential of the control electrode goes down because of the increased drop across R, and the shadow angle widens. When the potential of the triode grid changes in the negative direction, the shadow angle narrows.

Another type of indicator tube is the 6AF6G. This tube contains only an indicator unit but employs two ray-control electrodes mounted on opposite sides of the cathode and connected to individual base pins. It employs an external dc amplifier. (See Fig. 73.) Thus, two symmetrically opposite shadow angles may be obtained by connecting the two ray-control electrodes together;

or, two unlike patterns may be obtained by individual connection of each ray-control electrode to its respective amplifier.

In radio receivers, avc voltage is applied to the grid of the dc amplifier. Because avc voltage is at maximum when the set is tuned to give maximum response to a station, the shadow angle is at minimum when the receiver is tuned to resonance with the desired station.

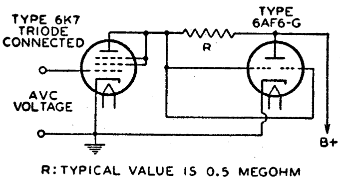

Fig. 73—Indicating circuit using 6AF6G electron-ray tube and external dc amplifier.

The choice between electron-ray tubes depends on the avc characteristic of the receiver. The 6E5 contains a sharp-cutoff triode which closes the shadow angle on a comparatively low value of avc voltage. The 6AB5/6N5 and 6U5 each have a remote-cutoff triode which closes the shadow on a larger value of avc voltage than the 6E5. The 6AF6G may be used in conjunction with dc amplifier tubes having either remote- or sharp-cutoff characteristics.

Oscillation

As an oscillator, an electron tube can be employed to generate a continuously alternating voltage. In present-day radio broadcast receivers, this application is limited practically to superheterodyne receivers for supplying the heterodyning frequency. Several circuits (represented in Figs. 74 and 75) may be utilized, but they all depend on feeding more energy from the plate circuit to the grid circuit than is required to equal the power loss in the grid circuit. Feedback may be produced by electrostatic or electromagnetic coupling between the grid and plate circuits. When sufficient energy is fed back to more than compensate for the loss in the grid circuit, the tube will oscillate. The action consists of regular surges of power between the plate and the grid circuit at a frequency dependent on the circuit constants of inductance and capacitance. By proper choice of these values, the frequency may be adjusted over a very wide range.

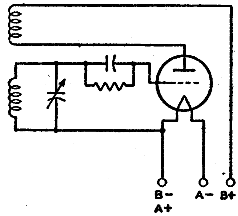

Fig. 74—Tuned-grid triode oscillator circuit using filament-type tube.

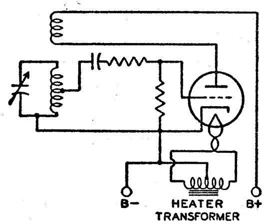

Fig. 75—Tuned-grid triode oscillator circuit using heater-cathode-type tube.

Multivibrators

Relaxation oscillators, which are widely used in present-day electronic equipment, are used to produce non-sinusoidal waveshapes such as rectangular and sawtooth pulses. Probably the most common relaxation oscillator is the multivibrator, which may be considered as a two-stage resistance-coupled amplifier in which the output of each tube is coupled into the input of the other tube.

Fig. 76 is a basic multivibrator circuit of the free-running type. In this circuit, oscillations are maintained by the alternate shifting of conduction from one tube to the other. The cycle usually starts with one tube, V_1, at zero bias, and the other, V_2, at cutoff or beyond. At this point, the capacitor C_1 is charged sufficiently to cut off V_2. C_1 then begins to discharge through the resistor R_4, and the voltage on the grid of V_2 rises until V_2 begins to conduct. The voltage on the plate of V_2 then decreases, causing V_1 to conduct less and less. At the same time, the plate voltage of V_1 begins to rise, causing V_2 to conduct still more heavily. Because of the amplification, this cumulative effect builds up extremely fast, and conduction switches from V_1 to V_2 within a few microseconds, depending on the circuit components.

In this circuit, therefore, conduction switches from V_1 to V_2 over the interval during which C_1 discharges from the voltage across R_4 to the cutoff voltage for V_2. The actual transfer of conduction does not occur until cutoff is reached. Conduction switches back to V_1 through a similar process to complete the cycle. The plate waveform is essentially rectangular in shape, and may be adjusted as to symmetry, frequency, and amplitude by proper choice of circuit constants, tubes, and voltages.

Although this type of multivibrator is free-running, it may be triggered by

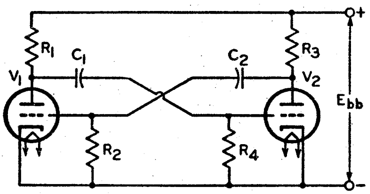

Fig. 76—Basic multivibrator circuit of the free-running type.

pulses of a given amplitude and frequency to provide a frequency-stabilized output. Multivibrator circuits may also be designed so that they are not free-running, but must be triggered externally to shift conduction from one tube to the other. Depending on the type of circuit, conduction may shift back to the first tube after a given time interval, or the second tube may continue conducting until another trigger signal is applied.

Synchroguide Circuits

The "synchroguide" is a controlled type of oscillator used in television receivers to generate and control the synchronized sawtooth voltage necessary for adequate line- or horizontal-frequency scanning. A simplified synchroguide circuit is shown in Fig. 77. This circuit provides stable, noise-free control of a blocking oscillator which generates a horizontal-frequency signal. It permits comparison of the received sync pulses and the generated sawtooth voltages so

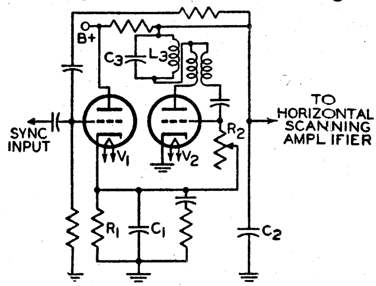

Fig. 77—Simplified synchroguide circuit.

that properly locked-in horizontal scanning results.

The triode V_2 in Fig. 77 is a conventional blocking oscillator which enables a sawtooth voltage to be developed across the capacitor C_2. A portion of this sawtooth is fed back to the grid of the control tube, V_1. The positive sync pulses are also applied to the grid of V_1. The waveforms shown in Fig. 78 illustrate the sawtooth and sync pulses (A and B) and their proper "in-sync" combination (C). The sync pulse occurs partly during the portion of the sawtooth voltage in which the triode V_1 draws current. Any shift in sync pulse as it is superimposed on the sawtooth, therefore, will affect the amount of conduction of the control tube. A change in control-tube conduction ultimately affects the bias on the oscillator-tube grid by changing the voltage to which the capacitor C_1 in the cathode circuit may charge. An increase in the positive bias increases the frequency of oscillation.

For example, waveform D in Fig. 78 illustrates a condition in which the sawtooth voltage is advanced in phase

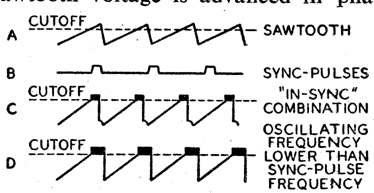

Fig. 78—Sawtooth and sync pulses in synchroguide circuit.

with respect to the sync pulses. The widening of the pulse which occurs at the corner of the sawtooth waveform allows the control tube to conduct more current and, consequently, allows the capacitor C_1 to charge to a higher voltage. This increased reference voltage also appears in the grid circuit of V_2 and makes the grid more positive. The increased grid voltage then speeds up the frequency of oscillations until proper synchronization results.

The blocking oscillator can be made more immune to changes in frequency and noise if V_2 is brought out of cutoff very sharply. This effect is obtained by sine-wave stabilization. The tuned circuit L_3C_3 in the plate circuit of Fig. 77 superimposes a shock-excited sine wave on the plate and grid waveforms, as shown in Fig. 79.

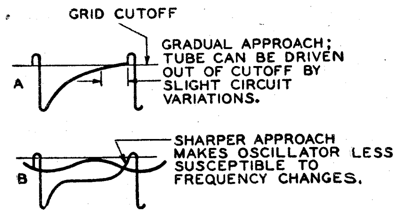

Fig. 79—Waveforms showing effect of tuned circuit L_3C_3 in Fig. 77.

Deflection Circuits

Vertical Output Circuits

A modified multivibrator in which the vertical output tube is part of the oscillator circuit is used in the vertical deflection stage of many television receivers. This stage supplies the deflection energy required for vertical deflection of the picture-tube beam. A simplified combined vertical-oscillator-output stage is shown in Fig. 80. Waveshapes at critical points of the circuit are included to illustrate the development of the desired current through the vertical output transformer and deflecting yoke.

The current waveform through the deflecting yoke and output transformer should be a sawtooth to provide the desired deflection. The grid and plate voltage waveforms of the output tube could also be sawtooth except for the effect of the inductive components in the yoke and transformer. The effect of these inductive components must be taken into consideration, however, particularly during retrace. The fast rate of current change during retrace time (which is approximately 1/15 as long as trace time) causes a high-voltage pulse at the plate which could give a trapezoidal waveshape to the plate voltage and cause increased plate current, excess damping, and lengthened retrace time. However, the grid voltage is made sufficiently negative during retrace to keep the tube close to cutoff, as described below.

The frequency, and the relative deviation of the positive and negative portions of each cycle, are dependent on the values of resistors R_1 and R_3 and the RC combination R_3C_2, as explained previously in the section on multivibrators. The desired trapezoidal waveshape at the grid of V_2 is created by capacitor C_1 and resistor R_2. If R_2 were equal to

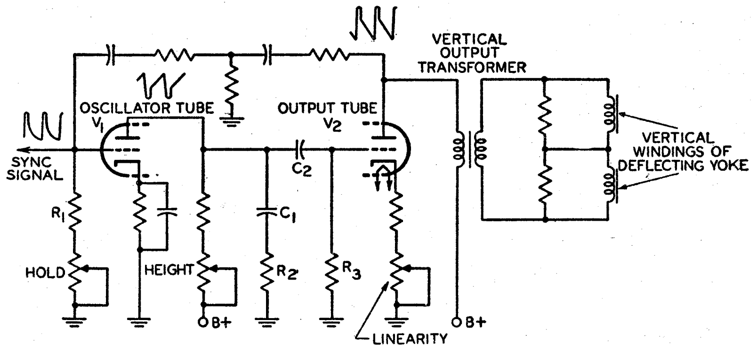

Fig. 80—Simplified combined vertical-oscillator-and-output stage.

zero, C_1 would cause the grid-voltage waveshape to take the form shown in Fig. 81(a). When R_2 is sufficiently large, C_1 does not discharge completely when V_1 conducts. When V_1 is cut off, therefore, the voltage on the grid of V_2 immediately rises to the voltage across C_1. The resulting waveshape is shown in Fig. 81(b). The negative-going pulse of the grid-voltage waveshape prevents the high plate pulse from causing excess conductance, and thereby prevents overdamping.

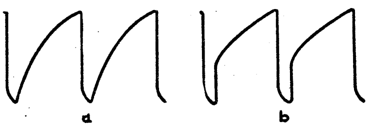

Fig. 81—Waveforms showing effect of R_2 in Fig. 80.

This vertical deflection stage utilizes twin-triode tubes such as the 6DR7 and 6EM7. The 6EM7 is particularly suitable for this application because it incorporates dissimilar units to provide for the different operating requirements of the oscillator and output sections.

Horizontal Output Circuits

Fig. 82 shows a typical horizontal-output-and-deflection circuit used in television receivers. In addition to supplying the deflection energy required for horizontal deflection of the picture-tube beam, this circuit provides the high dc voltage required for hte ultor (anode) of the picture tube and the "boosted" B voltage for other portions of the receiver. The horizontal-output tube is usually a beam power tube such as the 6DQ63, 6CD6-GA, or 6GW6.

In this circuit, a sawtooth voltage from the horizontal-oscillator tube is applied to the grid No. 1 of the horizontal-output tube. When this voltage rises above the cutoff point of the output tube, the tube conducts a sawtooth of plate current which is fed through the auto-transformer to the horizontal-deflecting yoke. At the end of the horizontal-scanning cycle, which lasts for 63.4 microseconds, the sawtooth voltage on the grid suddenly cuts off the output tube. This sudden change sets up an oscillation of about 50 to 70 kilocycles in the output circuit, which may be considered as inductor shunted by the stray capacitance of the circuit. During the first half of this oscillation, a positive voltage appears across the transformer. In the second half of the cycle, the voltage swings below the plate supply voltage, and the damper diode conducts, damping out the oscillation. At the same time, the current through the deflecting yoke reverses and reaches its negative peak. As the damper-diode current decays exponentially to zero, the output tube begins to conduct again. The yoke current, therefore, is composed of current resulting from damper-diode conduction followed by output-tube conduction.

When the output tube is suddenly cut off, the high-voltage pulse produced by shock excitation of the load circuit is increased by means of an extra winding on the transformer. This high-voltage pulse charges a high-voltage capacitor through the high-voltage rectifier. The output of this circuit is the dc high-voltage supply for the picture tube. The high-voltage rectifier also obtains its filament power through a separate winding on the horizontal-output transformer.

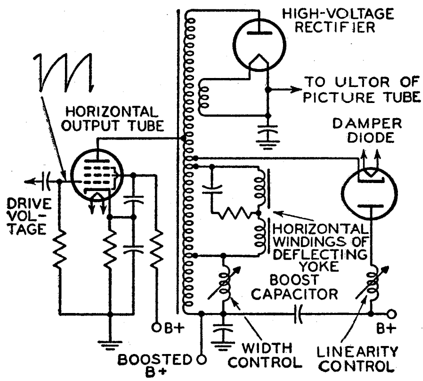

Fig. 82—Typical horizontal-deflection and high-voltage circuit.

Current flowing through the damper diode charges the "boost" capacitor through the damper portion of the transformer winding. The polarity of the charge on the capacitor is such that the voltage at the low end of the winding is increased above the plate supply voltage, or B+. This higher voltage or "boost" is used for the output-tube plate supply, and may also supply the deflection oscillators and the vertical-output circuit provided the current drain is not excessive.

High-Voltage Regulator Circuit

In color-television receivers, it is very important to regulate the high-voltage supply to the picture tube. A suitable circuit using the 6BK4 for regulation of the output of a high-voltage, high-impedance supply is shown in Fig. 83. In this circuit, the cathode is held at a fixed positive potential with respect to ground. Because the grid potential is kept slightly less positive by the voltage drop across resistor R_2, the tube operates in the negative grid region and no grid current is drawn.

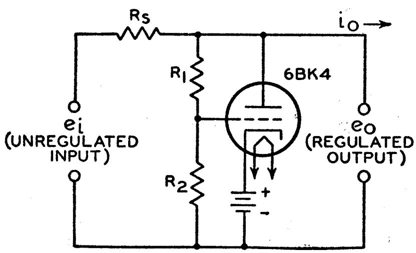

Fig. 83—High-voltage regulator circuit for color television.

When the output voltage, e_o, rises as a result of a decrease in load current, a small fraction of the additional voltage is applied to the grid of the tube by the voltage-divider circuit consisting of R_1 and R_2. This increased grid voltage causes the tube to draw an increased current from the unregulated supply. The increased current, in turn, causes a voltage drop across the high internal impedance of the unregulated supply, R_s, which tends to counteract the original rise of the voltage. If desired, the grid may be connected to a variable point on the voltage divider to allow some adjustment of the output-voltage level.

The grid voltage for the 6BK4 can also be obtained from a tap on the B-boost voltage supply. The use of this lower voltage (about 375 volts) eliminates the need for costly and troublesome high-voltage resistors. In this arrangement, variations in high voltage also vary the tapped-down B-boost voltage at the regulator grid, and the resulting variations in conduction of the regulator increase or decrease the loading of the high-voltage supply so that the total load remains nearly constant.

Frequency Conversion

Frequency conversion is used in superheterodyne receivers to change the frequency of the rf signal to an intermediate frequency. To perform this change in frequency, a frequency-converting device consisting of an oscillator and a frequency mixer is employed. In such a device, shown diagrammatically in Fig. 84, two voltages of different frequency, the rf signal voltage and the voltage generated by the oscillator, are applied to the input of the frequency mixer. These voltages beat, or heterodyne, within the mixer tube to produce a plate current having, in addition to the frequencies of the input voltages, numerous sum and difference frequencies.

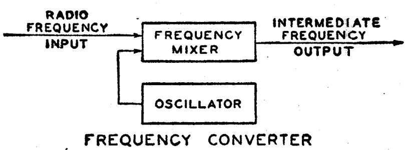

Fig. 84—Block diagram of simple frequency-converter circuit.

The output circuit of the mixer stage is provided with a tuned circuit which is adjusted to select only one beat frequency, *i.e.*, the frequency equal to the difference between the signal frequency and the oscillator frequency. The selected output frequency is known as the intermediate frequency, or if. The output frequency of the mixer tube is kept constant for all values of signal frequency by tuning the oscillator to the proper frequency.

Important advantages gained in a receiver by the conversion of signal frequency to a fixed intermediate frequency are high selectivity with few tuning stages and a high, as well as stable, overall gain for the receiver.

Several methods of frequency conversion for superheterodyne receivers are of interest. These methods are alike in that they employ a frequency-mixer tube in which plate current is varied at a combination frequency of the signal frequency and the oscillator frequency. These variations in plate current produce across the tuned plate load a voltage of the desired intermediate frequency. The methods differ in the types of tubes employed and in the means of supply input voltages to the mixer tube.

A method widely used before the availability of tubes especially designed for frequency-conversion service, and currently used in many FM, television, and standard broadcast receivers, employs as mixer tube either a triode, a tetrode, or a pentode, in which oscillator voltage and signal voltage are applied to the same grid. In this method, coupling between the oscillator and mixer circuits is obtained by means of inductance or capacitance.

A second method employs a tube having an oscillator and frequency mixer combined in the same envelope. In one form of such a tube, coupling between the two units is obtained by means of the electron stream within the tube. Because five grids are used, the tube is called a pentagrid converter.

Grids No. 1 and No. 2 and the cathode are connected to an external circuit to act as a triode oscillator. Grid No. 1 is the grid of the oscillator and Grid No. 2 is the anode. These and the cathode can be considered as a composite cathode which supplies to the rest of the tube an electron stream that varies at the oscillator frequency.

This varying electron stream is further controlled by the rf signal voltage on grid No. 4. Thus, the variations in plate current are due to the combination of the oscillator and the signal frequencies. The purpose of grids No. 3 and No. 5, which are connected together within the tube, is to accelerate the electron stream and to shield grid No. 4 electrostatically from the other electrodes.

Pentagrid-converter tubes of this design are good frequency-converting devices at medium frequencies. However, their performance is better at the lower frequencies because the output of the oscillator drops off as the frequency is raised and because certain undesirable effects produced by interaction between oscillator and signal sections of the tube increase with frequency.

To minimize these effects, several of the pentagrid-converter tubes are designed so that no electrode functions alone as the oscillator anode. In these tubes, grid No. 1 functions as the oscillator grid, and grid No. 2 is connected within the tube to the screen grid (grid No. 4). The combined two grids, Nos. 2 and 4, shield the signal grid (grid No. 3) and act as the composite anode of the oscillator triode. Grid No. 5 acts as the suppressor grid.

Converter tubes of this type are designed so that the space charge around the cathode is unaffected by electrons from the signal grid. Furthermore, the electrostatic field of the signal grid also has little effect on the space charge. The result is that rf voltage on the signal grid produces little effect on the cathode current. There is, therefore, little detuning of the oscillator by avc bias because changes in avc bias produce little change in oscillator transconductance or in the input capacitance of grid No. 1.

Examples of the pentagrid converters discussed in the preceding paragraph are the single-ended types 1R5 and 6BE6. A schematic diagram illustrating the use of the 6BE6 with self-excitation is given in Fig. 85 the 6BE6 may also be used with separate excitation. A complete circuit is shown in the **Circuits** section.

Another method of frequency conversion utilizes a separate oscillator having its grid connected to the No. 1 grid of a mixer hexode. The cathode, triode grid, and triode plate form the oscillator unit of the tube. The cathode, hexode mixer grid (grid No. 1), hexode screen

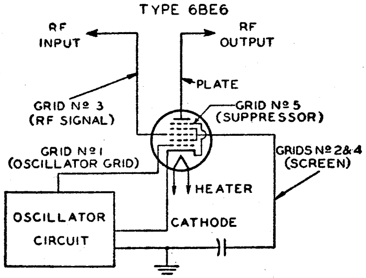

Fig. 85—Frequency-converter circuit using the 6BE6 pentagrid converter with self-excitation.

grids (grids Nos. 2 and 4), hexode signal grid (grid No. 3), and hexode plate constitute the mixer unit. The internal shields are connected to the shell of the tube and act as a suppressor grid for the hexode unit.

The action of this tube in converting a radio-frequency signal to an intermediate frequency depends on (1) the generation of a local frequency by the triode unit, (2) the transferring of this frequency to the hexode grid No. 1, and (3) the mixing in the hexode unit of this frequency with that of the rf signal applied to the hexode grid No. 3. The tube is not critical to changes in oscillator-plate voltage or signal-grid bias and, therefore, finds important use in all-wave receivers to minimize frequency-shift effects at the higher frequencies.

A further method of frequency conversion employs a tube called a pentagrid mixer. This type has two independent control grids and is used with a separate oscillator tube. RF signal voltage is applied to one of the control grids and oscillator voltage is applied to the other. It follows, therefore, that the variations in plate current are due to the combination of the oscillator and signal frequencies.

The tube contains a heater-cathode, five grids, and a plate. Grids Nos. 1 and 3 are control grids. The rf signal voltage is applied to grid No. 1. This grid has a remote-cutoff characteristic and is suited for control by avc bias voltage. The oscillator voltage is applied to grid No. 3. This grid has a sharp-cutoff characteristic and produces a comparatively large effect on plate current for a small amount of oscillator voltage. Grids Nos. 2 and 4 are connected together within the tube. They accelerate the electron stream and shield grid No. 3 electrostatically from the other electrodes. Grid No. 5, connected within the tube to the cathode, functions similarly to the suppressor grid in a pentode.

In the converter or mixer stage of a television receiver, stable oscillator operation is most readily obtained when separate tubes or tube sections are used for the oscillator and mixer functions. A typical television mixer-oscillator circuit is shown in Fig. 86. In such circuits, the oscillator voltage is applied to the mixer grid by inductive coupling, capacitive coupling, or a combination of the two. Tubes containing electrically

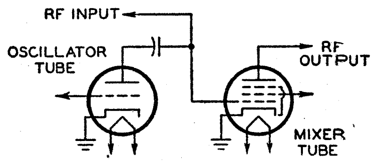

Fig. 86—Typical television mixer-oscillator circuit.

independent oscillator and mixer units in the same envelope, such as the 6U8A and 6X8, are designed especially for this application.

Automatic Frequency Control

An automatic frequency control (afc) circuit provides a means of correcting automatically the intermediate frequency of a superheterodyne receiver when, for any reason, it drifts from the frequency to which the if stages are tuned. This correction is made by adjusting the frequency of the oscillator. Such a circuit will automatically compensate for slight changes in rf carrier or oscillator frequency as well as for inaccurate manual or push-button tuning.

An afc system requires two sections: a frequency detector and a variable reactance. The detector section may be essentially the same as the FM detector illustrated in Fig. 59 and discussed under **Detection**. In the afc system, however, the output is a dc control voltage, the magnitude of which is proportional to the amount of frequency shift. This dc control voltage is used to control the grid bias of an electron tube which comprises the variable reactance section (Fig. 87).

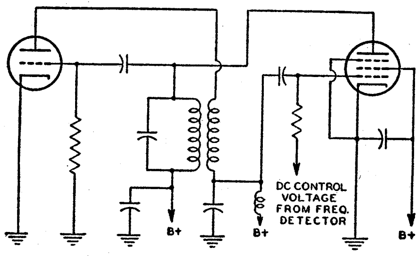

Fig. 87—Automatic-frequency-control (afc) circuit.

The plate current of the reactance tube is shunted across the oscillator tank circuit. Because the plate current and plate voltage of the reactance tube are almost 90 degrees out of phase, the control tube affects the tank circuit in the same manner as a reactance. The grid bias of the tube determines the magnitude of the effiective reactance and, consequently, a control of this grid bias can be used to control the oscillator frequency.

Automatic frequency control is also used in television receivers to keep the horizontal oscillator in step with the horizontal-scanning frequency (15,750 cps) at the transmitter. A widely used horizontal afc circuit is shown in Fig. 88. This circuit, which is often referred to as a **balanced-phase-detector** or **phase-discriminator** circuit, is usually employed to control the frequency of a multivibrator-type horizontal-oscillator circuit. The 6AL5 detector supplies a dc control voltage to the grid of the horizontal-oscillator tube which counteracts changes in its operating frequency. The magnitude and polarity of the control voltages are determined by phase relationships in the afc circuit at a given moment.

The horizontal sync pulses obtained from the sync-separator circuit are fed through a single-triode phase-inverter or phase-splitter circuit to the two diode units of the 6AL5. Because of the action of the phase-inverter circuit, the signals applied to the two diode units are equal in amplitude but 180 degrees out of

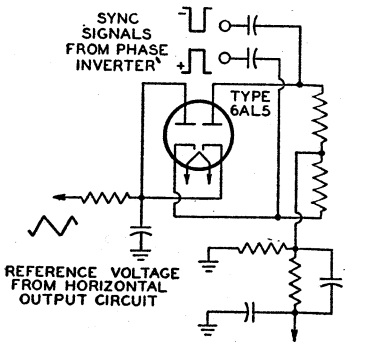

Fig. 88—Balanced phase-detector or phase-discriminator circuit for horizontal afc.

phase. A reference sawtooth voltage obtained from the horizontal output circuit is also applied simultaneously to both units. Any change in the oscillator frequency alters the phase relationship between the reference sawtooth and the incoming horizontal sync pulses, causing one diode unit of the 6AL5 to conduct more heavily than the other, and thus producing a correction signal. The system remains balanced at all times, therefore, because momentary changes in oscillator frequency are instantaneously corrected by the action of the control voltage.

The diode units of the 6AL5 are biased so that conduction takes place only during the tips of the sync pulses. The relative position of the sync pulses on the retrace portion of the sawtooth waveform at any given instant determines which diode unit conducts more heavily, and thereby establishes the magnitude and polarity of the control voltage. The network between the diode units and the grid of the horizontal-oscillator tube is essentially a low-pass filter which prevents the horizontal sync pulses from affecting the horizontal-oscillator performance.