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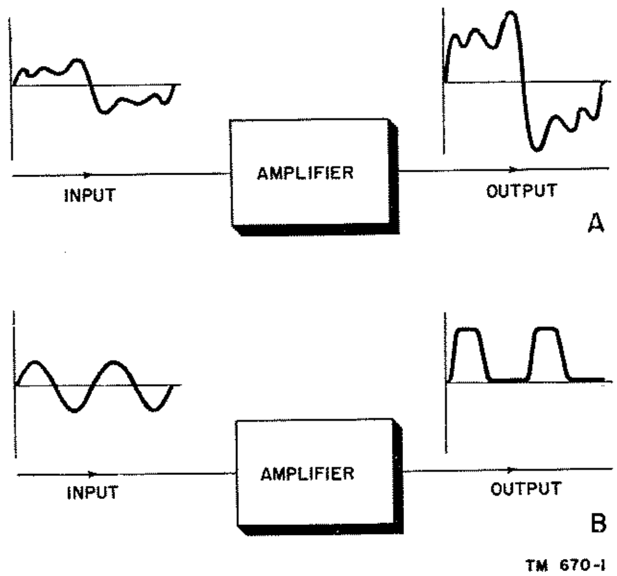

Figure 1. Amplification of signal.

CHAPTER 1

AMPLIFIER FUNDAMENTALS

1. Definition of Amplifier

A *vacuum-tube amplifier* is a device used to increase to usable strength the power or voltage of the signals developed in electronic circuits. For example: The output of a microphone must be amplified (fig. 1) before it can produce an audible sound from a loudspeaker; a radar echo signal must be amplified to make it strong enough to operate an indicator. The output signal, as in A, may have the same waveform as the input, or it may have a different waveform as in B; but in either case the amplitude of the output is greater than that of the input signal.

2. Principles of Operation of Amplifier

An amplifier consists of one or more vacuum tubes together with their associated circuits (fig. 2). It works in accordance with the fundamental principles of vacuum-tube operation explained in TM 11–662.

a. Basic Amplifier Circuit.

(1) The heart of the amplifier stage is the *tube*. The tube shown in figure 2 is a triode. Power for the output circuit is supplied by battery E_{bb}, connected in series with load resistor R_L between cathode K and plate P of the tube. The battery places the plate at a positive voltage in respect to the cathode, and causes a direct current, i_b, through the tube and R_L. The arrow indicates the direction of electron flow.

(2) The bias battery, E_{cc}, connected as shown, places grid G at a negative voltage in respect to the cathode. Usually, no current exists in the grid circuit. The alternating signal voltage, e_g, is applied to the grid across grid resistor R_g. The grid signal causes variations in the plate current, which consists of a d-c (direct-current) component caused by E_{bb}, and an a-c (alternating-current) component caused by e_g. The a-c component of the plate current develops an alternating voltage drop across R_L (in addition to the d-c voltage drop). This alternating voltage drop is of greater amplitude than e_g, and is the output voltage. If load resistor R_L were omitted from the circuit, no useful output voltage could be obtained. The grid signal would cause the plate *current* to vary, but the plate *voltage* would

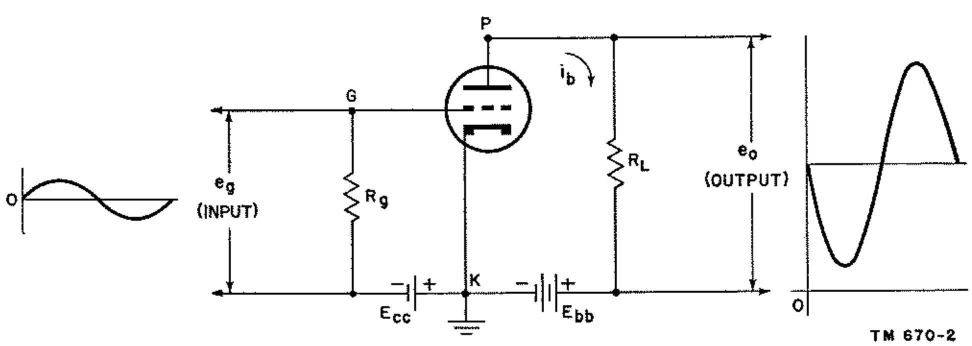

Figure 2. Simple amplifier stage.

remain constant, since at every instant it would be identical with E_{bb}.

b. Additional Amplifier Stages. Often a single stage will not provide sufficient amplification of a very weak signal. It then becomes necessary to connect two or more stages in *cascade,* as shown in the block diagram of figure 3. The output voltage of one stage is used as the input signal for the following stage, thus providing a greater over-all amplification than either stage could provide by itself. Several methods of *coupling* between stages are available (par. 4*d*). The polarity of the signal is reversed by each stage, as explained in paragraph 6*b*.

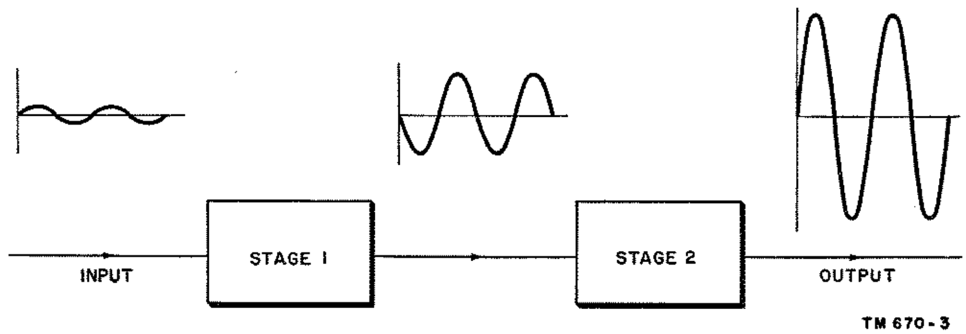

Figure 3. Two stages in cascade.

3. Amplifier Considerations

a. Gain. Amplification is expressed by comparing the amplitudes of the output and input voltages. The ratio of these two quantities is called the *voltage gain,* or simply the *gain,* of the amplifier. The output voltage may, or may not, deliver more power to the load than the stage receives at its input.

b. Frequency Response. Most amplifiers provide substantially the same gain for signals of slightly different frequencies. Over a band of frequencies for which the gain is constant, the *response* of an amplifier is said to be *flat.* For signals of frequencies above and below this band, the gain usually falls off more or less sharply. It is of interest in amplifier application to know over what range the response is flat, and how it may be expected to fall off beyond this range. This characteristic is called the *frequency response* of the amplifier.

c. Distortion. It often is desired that the amplifier reproduce in its output the exact waveform of the input signal (A of fig. 1). The extent to which this is accomplished depends partly on the frequency response of the amplifier. An amplifier that does not accomplish this result is said to introduce *distortion.* For many applications, the amplifier is designed purposely to introduce a large degree of distortion, as in B.

d. Efficiency. Any additional a-c power in the output not supplied by the input signal must come from the d-c supply. However, a part of the d-c power is wasted, in that it does not appear as a useful output. The extent to which this occurs is expressed as the *efficiency* of the amplifier. In large radio transmitters, efficiency may be an important consideration. In devices employing only small amounts of power, however, gain, frequency response, and distortion are of more importance.

4. Classification of Amplifiers

Amplifiers may be classified according to whether they are intended to provide increased voltage or increased power, whether plate current flows during the entire signal cycle or only a part of it as determined by the d-c bias, according to the frequency range over which flat response may be expected, or according to the coupling circuits used between stages or for coupling to the load. The first three of these bases of classification are discussed in this paragraph; the fourth is covered in later paragraphs.

a. Type of Service.

(1) *Voltage amplifier.* Voltage amplifiers are used to supply amplified voltages to the high-impedance grid circuits of power amplifiers, to cathode-ray tubes,

and to high-impedance vacuum-tube voltmeters. A voltage amplifier is concerned primarily with delivering large, varying output voltages to its load circuit. Therefore, the load impedance for a voltage amplifier is usually large, to develop a large voltage across its terminals. The ratio of output voltage to input voltage is called the *voltage gain* of the amplifier.

(2) *Power amplifier.* A power amplifier is used to deliver power to its load circuit. *Power amplification* is the ratio of output power to driving power—that is, to the a-c power consumed in the grid circuit. *Power sensitivity* describes the power amplification when no grid power is consumed in the grid circuits. Power sensitivity is the ratio of power output to the square of the rms (root mean square) grid-signal voltage. The load impedance for power amplification is chosen to supply maximum power output at specified distortion and plate efficiency. *Plate efficiency* is the ratio of a-c power output to d-c plate power input. In amplifiers designed for low distortion, the plate efficiency is generally low, but high efficiency is possible where distortion is permissible.

(3) *Application of amplifier.* The sound system of figure 4 contains typical applications of both types of amplifier. The weak voltage developed in the microphone is amplified by the voltage amplifier. Since very little signal current flows in the output circuit of a voltage amplifier, the output of the voltage amplifier has insufficient power to cause an audible sound from the loudspeaker. Therefore, the voltage-amplifier output is applied to a power amplifier. The output voltage of the power amplifier may even be less than its input voltage, but it delivers a much larger current. Thus, since $P = I^2R$, sufficient power is obtained to produce sound of the required loudness.

b. D-C Bias. The grid bias determines operating conditions of an amplifier by controlling the portion of the cycle during which plate current flows. In figure 5, an a-c signal voltage is projected on the plate-current grid-voltage characteristic curve, to show the waveform of plate current resulting when different bias voltages and different amplitudes of grid signal are used.

(1) In *Class A operation,* grid bias and a-c grid voltages are adjusted so that plate current in a given tube flows at

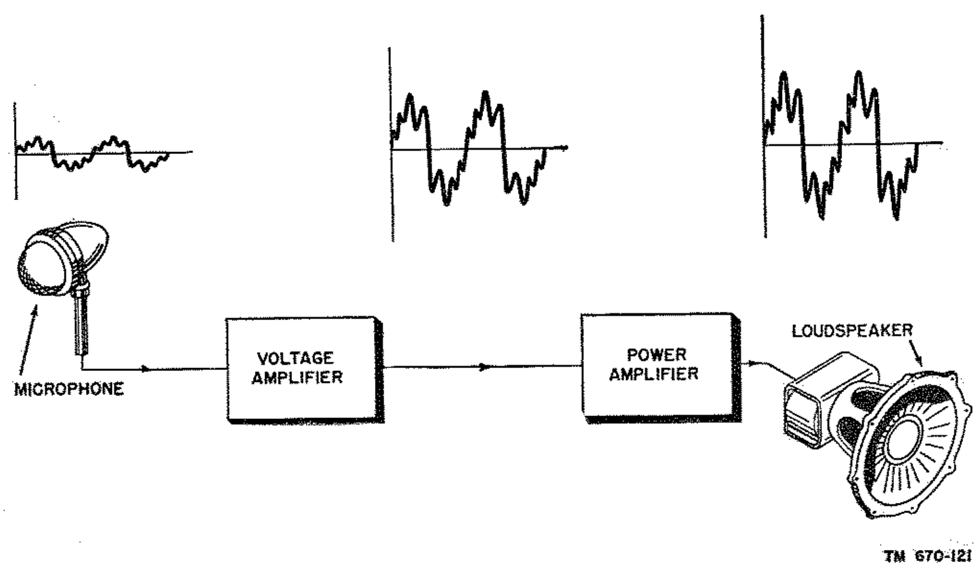

Figure 4. Sound-amplifier system.

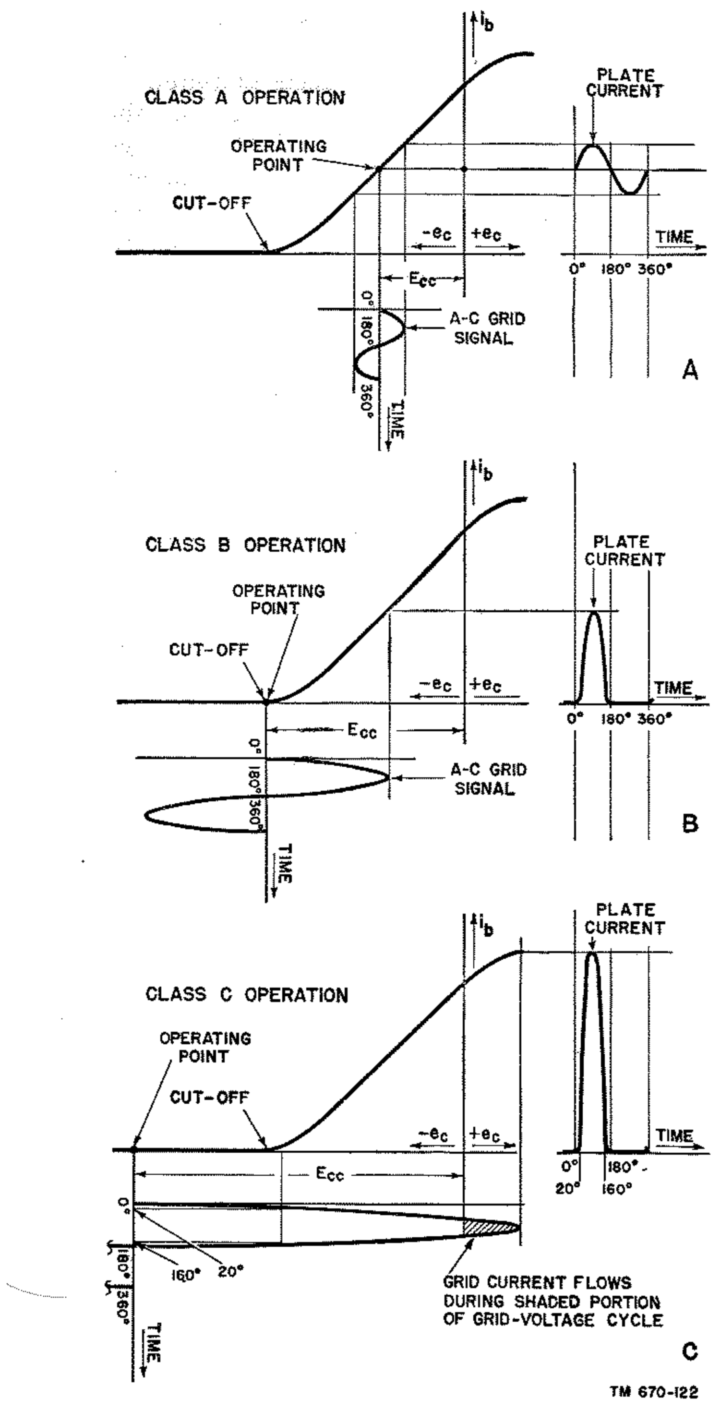

Figure 5. Classes of amplifier operation.

all times, as in A. To keep the distortion low, the grid signal swing is kept small, confining operation to the linear part of the characteristic curve. This also requires that d-c bias E_{cc} be such as to place the *operating point* near the middle of the straight part of the curve, as shown. Voltage amplifiers usually are operated under Class A conditions to obtain low distortion. Power output is small, and efficiency relatively low (10 to 20 percent).

(2) In *Class B operation,* the grid bias is made approximately equal to the cut-off value, which places the *operating point* at the cut-off point in B. Thus the plate current is close to 0 when no exciting grid voltage is applied. The plate current flows for approximately half of each cycle when an a-c grid voltage is applied. Class B power amplifiers give larger power output than Class A, better plate efficiency (50 to 60 percent), and a moderate ratio of power amplification. Class B operation introduces high distortion which is indicated by the difference between the output and input waveforms. Distortion may be minimized by operating two tubes in push-pull (paragraph 100). The grid circuit of a Class B power amplifier usually consumes power supplied by a *driver power amplifier.*

(3) In *Class AB operation,* the grid bias is less than cut-off, but greater than in Class A; the a-c grid voltage is such that plate current flows for appreciably more than half but less than the entire electrical cycle. For low signal levels these amplifiers have characteristics similiar to those of Class A, but produce somewhat more distortion; at high signal levels they operate like Class B power amplifiers with medium plate efficiency.

(4) In *Class C operation,* the grid bias is appreciably greater than the cut-off value, so that plate current flows for less than half of each cycle. In C, plate current flows from 20° to 160°. Class C amplifiers are characterized by high power output, high plate efficiency (60 to 70 percent), and very high distortion. Distortion usually is reduced to tolerable limits by using a parallel tuned circuit as a load instead of a

resistor. Class C amplifiers require moderate grid driving power. They are used to obtain large power output in the output stages of radio transmitters.

(5) Whenever the grid signal exceeds the bias, the grid is driven positive near the positive peaks of the signal, as shown by the shaded portion of the grid cycle in C. Grid current flows during this time, requiring additional driving power. Such operation is indicated by adding the number 2 to the class letter designation; thus, class C2 of class AB2 indicates that grid current flows during part of the cycle. Class AB1 or B1 indicates that no grid current flows.

c. Frequency Response. Amplifiers are classified also according to the frequency range over which they operate. These ranges are *a-f* (*audio-frequency*), *r-f* (*radio-frequency*), and *video-frequency.* The *response* of an amplifier is its gain at a particular frequency, or the manner in which the gain varies over a range of frequencies.

(1) A-f amplifiers are used to operate loudspeakers or motors. R-f amplifiers are used in radio transmitters to raise the power supplied to the antenna, and in radio receivers to increase the strength received by the antenna. Both often employ tuned circuits as loads, in order to suppress the output over all except the fairly narrow band of frequencies it is desired to amplify. Video-frequency amplifiers are designed with gain characteristics which are flat over a very wide frequency range. Video amplifiers are also called *broad-band* and *wide-band* amplifiers, and are used when the entire frequency range up to several megacycles is desired in the output. Because a signal consisting of repeated square pulses contains very high harmonics, video amplifiers are needed to give accurate reproduction of the input signal. For this reason they are known also as *pulse amplifiers.*

(2) Figure 6 compares the gain of a narrow-band amplifier with that of a broad-band amplifier. Note that the broad-band amplifier is *flat* (a term indicating constant gain) from 1 cycle to 1,000,000 cycles, whereas the narrow-band amplifier is flat only from 200 cycles to 5,000 cycles. The maximum gain of the narrow-band amplifier is 400, however, and the maximum gain of the wide-band amplifier is only 2. The considerable reduction in the gain of the wide-band amplifier is the penalty that is paid for the greater bandwidth. This is discussed later, in sections dealing with broad-band amplifiers.

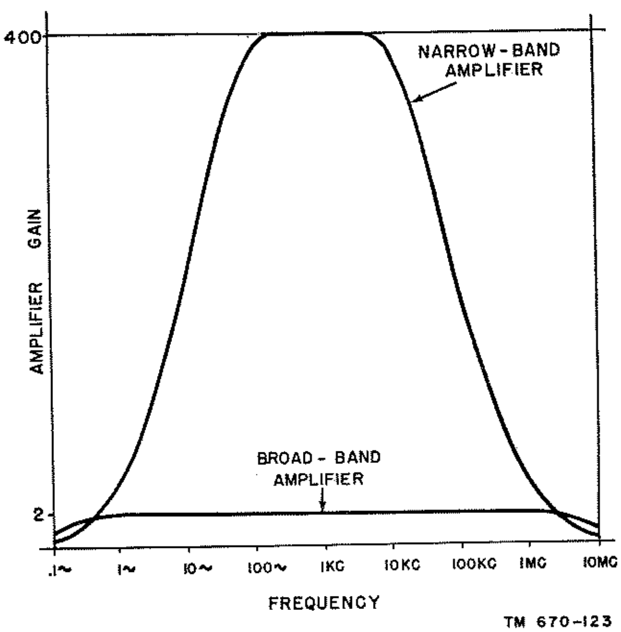

Figure 6. Gain of typical narrow-band and broad-band amplifier.

d. Coupling Methods. Four methods of coupling an amplifier to a load, or to a following stage, are available. They are (1) resistance coupling, (2) transformer coupling, (3) impedance coupling, and (4) direct coupling. The advantages of each method will be discussed later.

5. Distortion in Amplifiers

a. Distortion and Harmonic Frequencies. As already pointed out, whenever the output waveform of an amplifier differs from the input waveform, the amplifier is said to have *distorted* the signal. The waveform of any signal depends on its *harmonic content*—the relative amplitude and phase of the harmonic-frequency sine-wave components of the signal. Any device that changes the relative amplitude and phase of these harmonic components distorts the signal. Conversely, any device that adds new frequencies not present in the input signal changes the harmonic content and distorts the waveform. Amplifiers distort waveforms in both ways.

b. Types of Distortion. Four types of distortion may occur—(1) frequency distortion, (2) phase distortion, (3) amplitude distortion, and (4) intermodulation distortion. The first two types occur because the gain of an amplifier is not the same for all frequencies. This variation in gain is caused largely by the coupling circuits. Frequency and phase distortion cannot occur when the signal is a pure sine wave, because only one frequency is present. In amplitude distortion, the amplifier tube itself introduces new frequencies, so that even a sine wave input is distorted. Intermodulation distortion introduces new frequencies, but occurs only when more than a single frequency is present in the input.

(1) *Frequency distortion.* This type of distortion occurs when some components of a complex signal are amplified more than others. For example, figure 7 illustrates how frequency distortion may alter a signal consisting of a fundamental and its third harmonic. If the fundamental of the signal falls within the flat range of the amplifier response curve, and the third harmonic falls far down the slope at the high-frequency end of the curve, the fundamental is amplified, but the amplitude of the third harmonic may be negligible in the output. The result is an entirely different waveform in the output. Frequency distortion usually occurs where the amplifier gain falls off—at both low frequencies and very high frequencies. Within the flat range of an amplifier, however, there is no frequency distortion. Note that no *new* frequencies are produced by frequency distortion.

(2) *Phase distortion.* When signals pass through an amplifier they encounter a delay, known as *delay distortion* or *phase distortion,* which varies with frequency. It is caused chiefly by the reactive coupling circuits between amplifier stages. When a single pure sine wave is amplified, the delay, or phase shift, does not affect the amplified waveshape, and consequently there is no distortion. Similarly, when a complex wave is amplified, the output wave has the same shape as the input wave if the phase of each is shifted an amount proportional to each frequency on its own scale. In other words, the relative phase angles of the harmonics are not shifted in respect to the fundamental where distortion does not occur. However, when complex waveshapes are amplified, each component frequency of the waveshape may be shifted by an amount not proportional to the frequency, so that the output waveshape is not a faithful

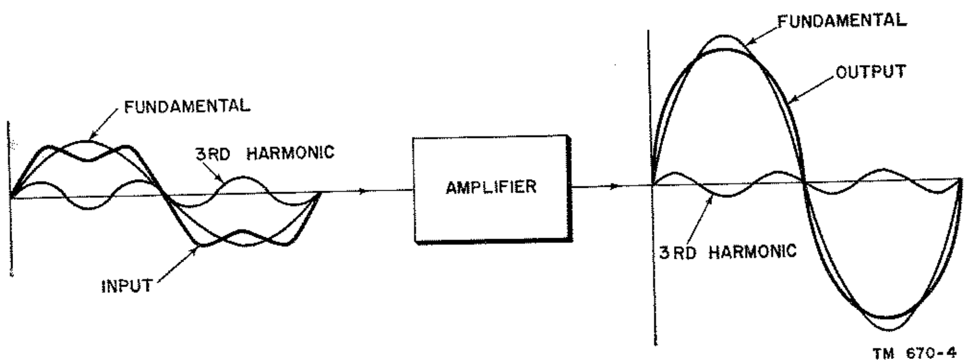

Figure 7. Frequency distortion.

representation of the input. In figure 8, a signal consisting of fundamental and third harmonic is passed through an amplifier producing phase distortion. Since both frequency components are amplified by identical ratios, their relative amplitudes are unchanged; the phase of the third harmonic, however, has been shifted by 90° in respect to the fundamental. So shown, the output waveform has been noticeably changed. In practice, frequency distortion and phase distortion almost invariably occur together. They have been separated in figures 7 and 8 to clarify the principles involved. Only one simple case is illustrated, out of the infinite variety of ways in which phase distortion can occur. In video amplifiers, special coupling circuits are used to minimize phase distortion.

even greater than is shown in B, figure 5. The output waveform resulting from a sine-wave grid signal is shown in A, figure 9. Below cut-off, the current through the tube remains zero, no matter how far the grid goes negative; above saturation, the plate current cannot increase, no matter how far positive the grid is driven. Such amplitude distortion often is required for producing special waveshapes used in radar and other applications. The new frequencies introduced by distortion are represented in B. Amplitude distortion can be produced in lesser degree, even though the tube is not driven beyond cut-off, if its characteristic is curved. Even in such cases, new frequencies are introduced, especially the second harmonic of each input frequency. Amplitude distortion

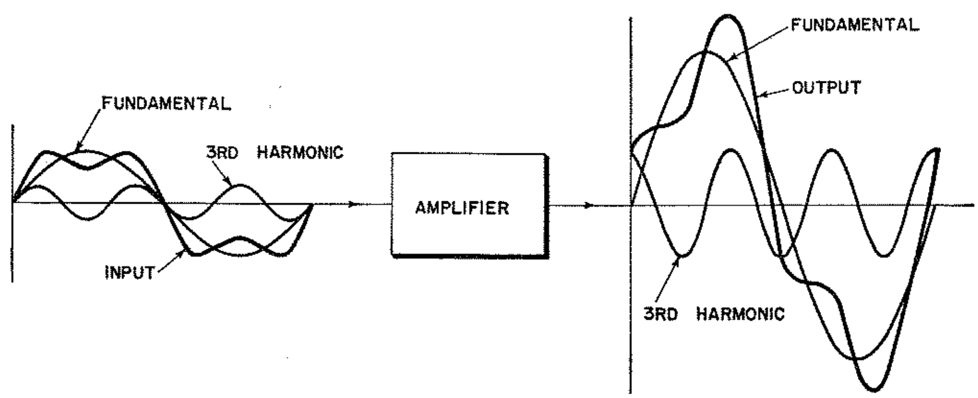

Figure 8. Phase distortion.

(3) *Amplitude distortion.* If a vacuum-tube amplifier is operated on any nonlinear part of its characteristic, a change in instantaneous grid voltage results in a change in instantaneous plate current which is not directly in proportion. The resulting distortion is *amplitude distortion,* or *nonlinear distortion.* Harmonic components are generated by the amplifier, and appear in the output in addition to those present in the input. As an extreme example, consider a Class B2 amplifier, whose grid-signal amplitude is can be reduced greatly by operating amplifier tubes within the linear region of their characteristics.

(4) *Intermodulation distortion.* A complex signal contains at least two frequency components. If such a signal is applied to an amplifier operating on any nonlinear part of its characteristic, *intermodulation distortion* results. Amplitude distortion causes harmonic components in the output waveshapes; intermodulation distortion causes, in addition, sum and difference frequencies of every pair of components of the

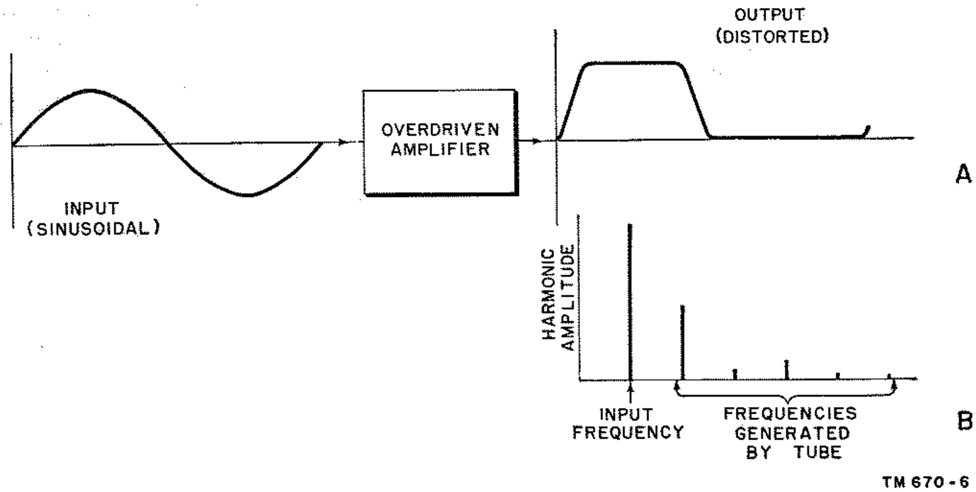

Figure 9. Amplitude distortion.

input waveshape. As an example, figure 10 shows the two new frequencies which appear when 60 cycles and 1,000 cycles are applied simultaneously to an amplifier which produces intermodulation distortion. Note that the two new frequencies have smaller amplitudes than either original signal, and that they appear at 60 cycles above and at 60 cycles below the 1,000-cycle signal. The two new frequencies, 940 cycles and 1,060 cycles, are not harmonics of either 60 cycles or 1,000 cycles. Harmonics of the two input frequencies would also be present as a result of amplitude distortion. In sound-system amplifiers, intermodulation distortion produces disagreeable, harsh sounds in the loudspeaker. Overloaded tubes and iron-core transformers are nonlinear and cause intermodulation distortion. Intermodulation distortion is minimized by operating apparatus on the most nearly linear portion of its characteristics.

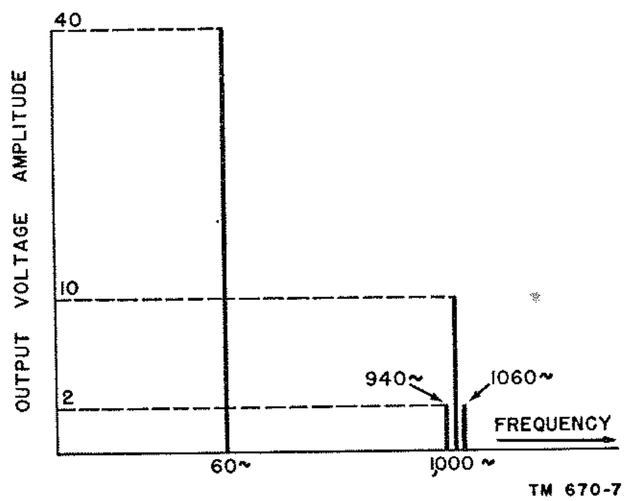

Figure 10. Intermodulation distortion.

6. Single-Stage Operation

a. Dynamic, Quiescent, and Static Conditions.

(1) In practical use, a single amplifier stage is operated in a manner similar to that illustrated in figure 2. A d-c voltage is applied to the plate through a load resistance or other impedance, and the varying signal to be amplified is applied to the grid. Grid bias is provided either by a battery, as shown, or by a method of self-bias, explained in TM 11–662. When so connected, the tube is said to be operating under *dynamic* conditions, because current and all voltages are changing from one instant to another: plate current, grid voltage, voltage from plate to cathode, and voltage drop across the load. Each of the quantities that changes continuously in dynamic operation does so by varying above and below some middle value. Thus (fig. 11), the in-

stantaneous voltage from cathode to grid e_c varies above and below the d-c grid-bias voltage, E_{cc}. This causes a corresponding variation in plate current i_b, as shown.

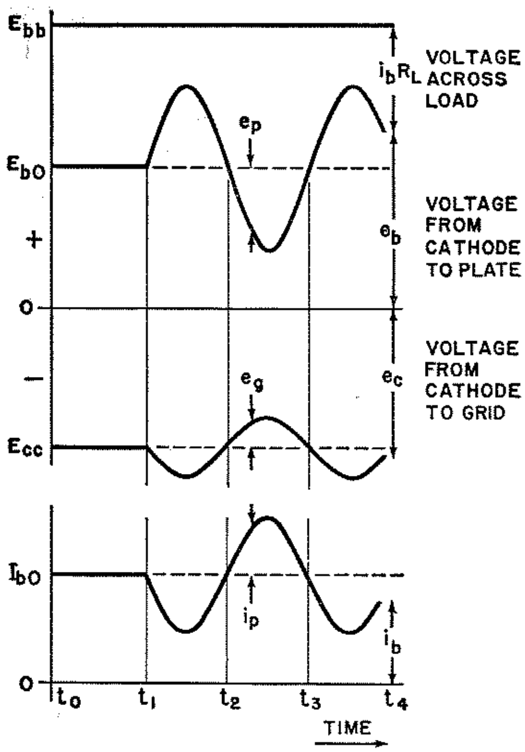

Figure 11. Voltage and current variations in Class A amplifier.

(2) In figure 2, the plate current flowing in load resistor R_L causes a varying voltage drop, i_bR_L. The instantaneous voltage across the tube, e_b, is equal to the difference between the plate-supply voltage, E_{bb} and i_bR_L. Since the latter varies, so does e_b (fig. 11).

(3) Consider what would happen if the signal voltage in figure 2 were short-circuited. The grid-bias voltage, E_{cc}, would appear as a steady voltage from cathode to grid, as is shown in figure 11 between the instants t_0 and t_1. The plate current would not be cut off. A direct plate current, designated I_{bo}, would flow. It would produce a direct voltage drop, $I_{bo}R_L$, across the load, and place a direct voltage, E_{bo}, across the tube. These values—E_{cc}, I_{bo}, and E_{bo}—are known as the *quiescent values* of the grid voltage, plate current, and plate voltage, respectively. When the tube is operated with zero signal, but otherwise is the same as in figure 2, it is operating under *quiescent conditions*. The quiescent values are the values about which e_c, i_b, and e_b vary when an alternating voltage, e_g, is applied to the grid in series with E_{cc} (fig. 11). Each horizontal solid line represents a quiescent value. Each instantaneous value is seen to vary above and below the broken-line extension of the solid line. The excursions of the instantaneous plate current and plate voltage about I_{bo} and E_{bo} are the a-c components of these quantities. They are designated i_p and e_p, as shown. It is to be noted that the a-c component of the load voltage is identical with e_p, the a-c component of the plate voltage.

(4) Quiescent values and quiescent conditions should not be confused with *static conditions*. Static conditons prevail when the plate-supply voltage is connected directly to the plate, without any intervening load impedance. Under such conditions, varying the grid voltage would vary the plate current, but not the plate voltage. This is not a practical method of operating a tube, as no output voltage can be obtained. It is used by engineers to measure the effect of changing the grid and plate voltages separately, in steps. Such measurements may be used to determine the tube coefficients, μ (amplification factor), r_p (a-c plate resistance), and g_m (transconductance), and to plot *static plate-characteristic curves*, such as the curves ordinarily shown in tube manuals and explained in TM 11–662. These curves are useful in designing amplifiers, but are not needed for the present discussion.

b. Phase Relations. Operating under dynamic conditions, the output voltage of an R-C amplifier stage is 180° out of phase with its input (of opposite polarity), as in figure 11. When the grid voltage decreases (becomes more negative), the plate current also decreases. The drop across the load resistor, therefore, decreases. But, with this decrease, more supply voltage E_{bb} appears across the tube. Thus, a *decrease* in e_c causes an *increase* in e_b. Therefore, as e_c continuously decreases and increases with the a-c signal, the a-c variations of e_b are 180° out of phase with e_g (of opposite polarity). Note that the a-c component of the load voltage, measured as a voltage rise from cathode to plate, is identical with the a-c component of e_b. This alternating component of the voltage across R_L is the desired output voltage. Thus the output voltage is also 180° out of phase with the signal voltage (of opposite polarity).

c. Amplification.

(1) Consider a tube operated under static conditions. A small increase in plate voltage causes a corresponding increase in plate current. If the grid voltage is decreased (made more negative) by a small amount, the plate current is restored to its original value. The change in grid voltage required to accomplish this is, in general, much smaller than the change in plate voltage causing the increase. The ratio of these two changes—the plate-voltage change and the grid-voltage change—is defined as the *amplification factor,* or μ, of the tube. It expresses the relative effect of changes in plate and grid voltage to produce corresponding changes in plate current.

(2) Since an a-c voltage is a continuously varying voltage, the amplification factor also expresses how much greater an a-c plate voltage would have to be, as compared with the a-c grid signal, to produce the same a-c component of plate current. Such an a-c plate voltage is not actually applied, but the concept is important in understanding the following paragraph.

7. Voltage Equivalent Circuit

The gain of an amplifier is frequently a question of primary interest. It can be calculated more or less exactly, for any class of operation, from the static characteristic curves. The calculation of gain is greatly facilitated by the use of an *equivalent circuit,* however. Two types of equivalent circuit may be used. The purpose of this paragraph is to develop the *constant-voltage* equivalent circuit, and to explain the assumptions on which it is based. The constant-current equivalent circuit is discussed in paragraph 12.

a. It has been shown (fig. 11) that the plate current of an amplifier tube is composed of a d-c component, I_{bo}, and an a-c component, i_p. The plate circuit behaves as though it were energized by a battery, which causes I_{bo}, and an a-c generator, which causes i_p. Any circuit thus energized by two or more voltage sources can be analyzed by considering the effects of the two sources separately, then adding up the results. Figure 12 shows how the plate circuit of a simple amplifier stage may be represented by two equivalent circuits.

b. In the d-c circuit of figure 12, resistor R_b represents the d-c resistance of the tube through which the quiescent plate current, I_{bo}, flows because of battery E_{bb}. In the a-c circuit, r_p is the a-c plate resistance of the tube. The

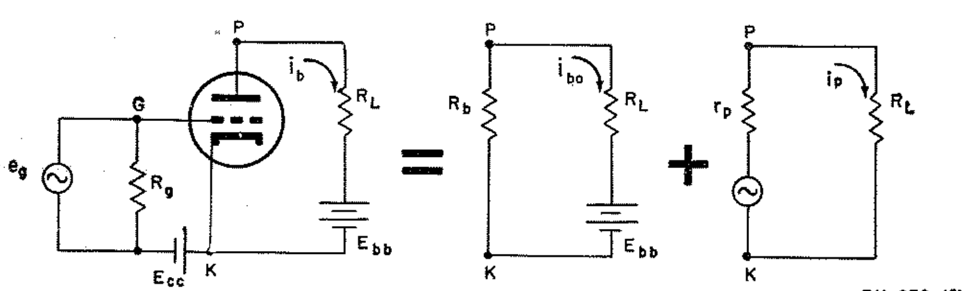

Figure 12. D-c and a-c equivalent circuits of simple amplifier.

a-c component of plate current i_p flows in both the tube and R_L. The generator that causes i_p is such that the sum of I_{bo} and i_p in the equivalent circuits will be the same as i_b, the varying plate current that flows in the actual tube circuit. (The d-c circuit is now of no further interest. It was introduced only to show that the a-c components of current and voltage in the plate circuit may be considered separately, for purposes of calculation.)

c. The generator of the alternating current in figure 12 has yet to be identified. If it were actually an alternating voltage applied to the plate of the tube, as represented in the equivalent circuit, it would cause the a-c plate current component, i_p. It follows from paragraph 6*c* (2) that this assumed generator voltage in the plate circuit must be μ times the grid signal, e_g, which actually produces i_p. The a-c equivalent circuit may, therefore, be drawn as in figure 13. It should be understood that this is *not* an actual circuit. It is a fictitious circuit having the same *alternating* current as the actual circuit. It is a useful tool for computing amplifier gain, as will be shown.

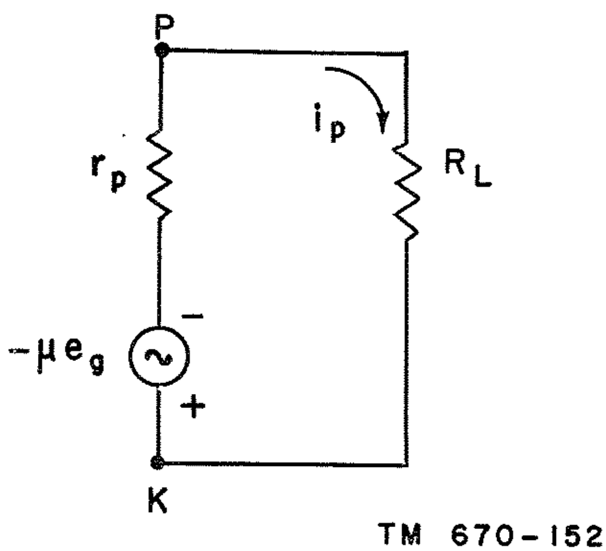

Figure 13. A-c equivalent circuit.

d. The equivalent plate generator has an instantaneous voltage of $-\mu e_g$ (fig. 13). The minus sign indicates the polarity reversal that takes place between the plate and grid voltages of the actual circuit. This is represented by assigning polarities to the generator terminals as shown, with the plate terminal negative in respect to the cathode terminal—just opposite from their actual polarities. This indicates that when the plate current is in its positive half-cycle, electrons are moving toward the positive generator terminal. This occurs when the grid voltage is in its positive half-cycle.

e. From the equivalent circuit of figure 13, the a-c plate current and the output voltage may be calculated by the methods applicable to simple series circuits:

$$i_p = \frac{-\mu e_g}{r_p + R_L}$$

$$\text{output voltage} = i_p R_L = -\mu e_g \left(\frac{R_L}{r_p + R_L}\right)$$

This expression shows that the output voltage is not simply μ times the applied signal, as in static operation, but less than this value.

f. The equivalent circuit gives the exact performance of the vacuum-tube amplifier only if certain assumptions hold true. These are (1) that the plate resistance and amplification factor of the tube (r_p and μ) remain constant under dynamic operating conditions, and (2) that distortion generated by the tube is negligible. In practice, both μ and r_p do vary somewhat. The variation usually is so small that published tube constants may be used for designing an amplifier, but performance should then be checked in the laboratory. Distortion is negligible only in a Class A amplifier, as its operation usually is confined to a fairly linear part of the characteristic.

8. Summary

a. Voltage amplifiers are used to increase the voltage of feeble signals. Power amplifiers are used to supply power to a load such as a loudspeaker or an antenna.

b. In Class A operation, plate current flows at all times; in Class B operation, plate current flows for approximately half of each cycle; in Class C operation, plate current flows for less than half of each cycle.

c. Audio-frequency amplifiers operate from about 50 cycles to 20,000 cycles; radio-frequency amplifiers operate from about 20,000 cycles up to many millions of cycles; and video-frequency amplifiers operate from about 20 cycles to 5,000,000 cycles.

d. The four types of distortion which occur in amplifiers are frequency, phase, amplitude, and intermodulation distortion. Frequency distortion occurs when the gain of an amplifier is different for signals of different frequency. Phase distortion occurs when the reactive coupling networks shift the phase of the signal components in respect to each other. Amplitude distortion results from operating on any nonlinear part of the tube characteristic, thereby adding new frequency components to the signal. Intermodulation distortion causes sum and difference frequency components to appear in the output of an amplifier when it is amplifying two or more frequencies simultaneously.

e. A tube operates under static conditions when its load impedance is zero. An amplifier operates under quiescent conditions when no external signal is impressed on it. An amplifier operates under dynamic conditions when it has an external signal impressed on its grid, and its plate voltage changes because of the changing drop across the load resistor.

f. Amplifier operation is analyzed by means of the constant-voltage equivalent circuit or the constant-current equivalent circuit, which shows the a-c component of current and voltages for a single-stage amplifier. The constant-voltage equivalent circuit shows that the output voltage of an amplifier is—

$$e_o = -\mu e_g \left(\frac{Z_L}{r_p + Z_L}\right)$$

g. The networks with which amplifier stages are coupled are known as resistance-capacitance, impedance, transformer, and direct-coupling circuits.

9. Review Questions

a. What is the purpose of an amplifier?

b. What is the difference between a voltage amplifier and a power amplifier?

c. What is the meaning of power sensitivity? Plate efficiency?

d. How are amplifiers generally classified?

e. Name four types of distortion and describe each type.

f. Compare static, quiescent, and dynamic operating conditions.

g. What is the constant-voltage equivalent circuit? Why is it used?

h. What is a single-stage amplifier? A cascade amplifier?

i. What is the theoretical maximum voltage gain obtainable from a tube?

j. What is the phase relationship between applied grid-signal voltage and plate-load voltage in an amplifier having a pure resistance load?

k. Name three basic coupling methods.

CHAPTER 2

NARROW-BAND AMPLIFIERS

Section I. R-C COUPLED AMPLIFIERS

10. Resistance-Capacitance Coupling

As pointed out earlier, four general methods are available for coupling the output of an amplifier stage to a load or to a following stage. They are (1) resistance coupling, or resistance-capacitance coupling; (2) impedance coupling; (3) transformer coupling; and (4) direct coupling. The operating characteristics of each method will be analyzed and discussed in this chapter.

a. Advantages. Amplifier stages which are coupled by combinations of resistances and capacitors are known as *resistance-capacitance coupled,* or *R-C coupled, amplifiers.* Their outstanding characteristics are high fidelity over wide frequency ranges, relatively high gain, low hum pickup from nearby a-c fields, small space requirements, and low cost. A minor disadvantage is the higher B-supply voltage which is needed to compensate for the voltage drop across the coupling resistor (load resistor) in the plate circuit of the tube.

b. Coupling Circuit.

(1) Typical circuits of triode and pentode amplifiers are shown in figure 14, together with the names of the various circuit elements. It will be noted, in A, that the elements of the triode circuit, within the broken line, comprise a single stage practically identical with that in figure 2. An exception is the substitution of self-bias, provided by $R2$ and $C2$ in figure 14, in place of bias-battery E_{cc} in figure 2. The corresponding parts of the pentode circuit are similarly inclosed in a broken line in B.

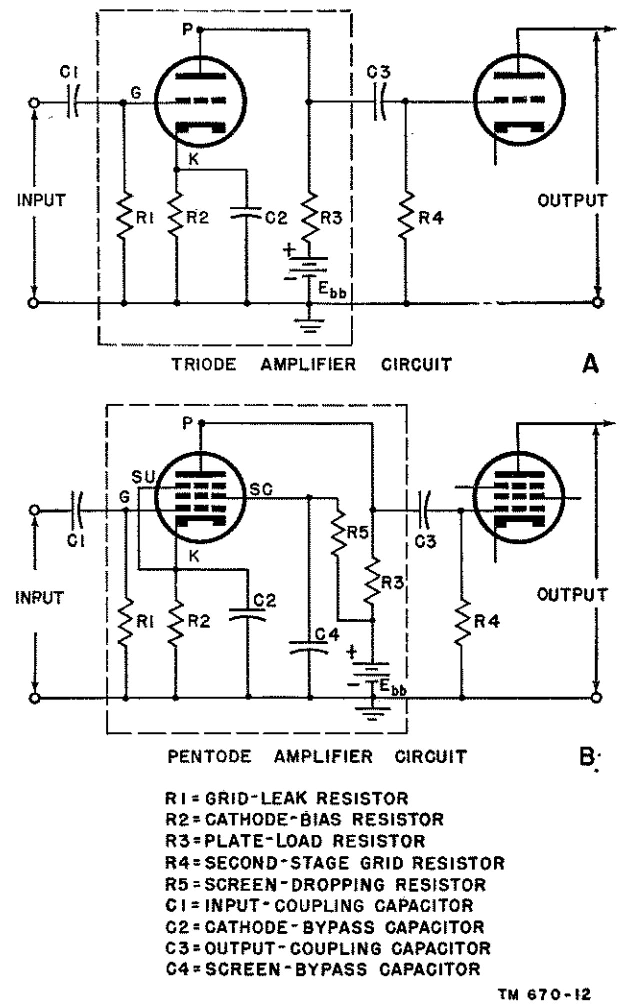

Figure 14. Typical R-C coupled amplifier.

(2) The elements outside the broken-line boxes are part of the coupling networks. *C*1 is the input coupling capacitor. It is inserted to prevent any d-c component of the input signal from appearing across the grid circuit. *C*3

serves a similar purpose for the second stage, blocking the d-c plate-supply voltage of the first stage from the grid of the second. Grid-leak resistors $R1$ and $R4$ serve two purposes. They provide from grid to ground a d-c path through which electrons striking a grid may leak off to ground. Even though no grid current flows, the grid-leak resistors are needed to place the grid at the same potential as the lower end of the cathode-bias resistor, thus producing the required bias voltage.

c. Frequency Characteristics.

(1) The frequency characteristics of any amplifier largely depend on the output impedance across which the output voltage is developed. In figure 12, this output impedance was shown as consisting only of the load resistor, R_L. Actually, it is a complex network (fig. 15). In this figure, R_L and R_g correspond to $R3$ and $R4$ of figure 14, and C_c corresponds to $C3$. C_d represents the distributed capacity of the coupling network; C_o and C_i are the interelectrode capacities of the tubes.

(2) In order to determine the frequency characteristics of an amplifier, it is customary to divide the frequency spectrum into three distinct regions known as low-, middle-, and high-frequency regions. The simplified equivalent circuits for these ranges appear in figure 16. Each circuit is based on one or more of the following assumptions: In the low-frequency range the reactances of the small shunt capacitance, C_o, C_d, C_i, are so high that they are comparable to an open circuit and are, therefore, negligible. The series reactance of the large coupling capacitor C_c, however, is appreciable at low frequencies. C_c, therefore, is shown, in A, in the low-frequency equivalent circuit. At high frequencies, the reactances of the shunt capacities are comparable to the shunt resistors. The shunt capacities, C_o, C_d, and C_i, therefore, are shown in the equivalent circuit, and their combined effect as C_s, in C. Since the reactance of coupling capacitor C_c is equivalent to a short circuit at high frequencies, its effect is negligible. For this reason, C_c does not appear in the high-frequency equivalent circuit. Between the low- and the high-frequency range lies the middle-frequency range at which both the series and shunt reactances have negligible effects upon the frequency response. Therefore, in B, no capacitors appear on the equivalent circuit for the middle-frequency range.

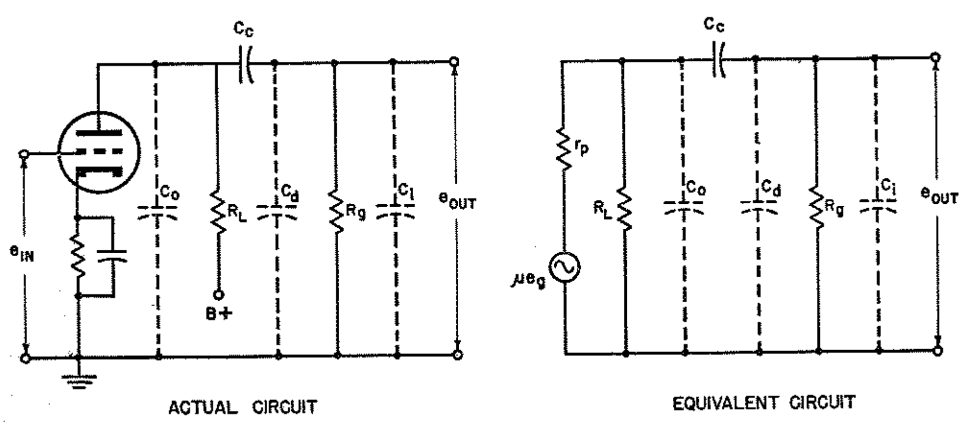

Figure 15. Output impedance of R-C coupled amplifier.

(3) At low frequencies, C_c and R_g act as a-c voltage dividers, as in A, so that only part of the voltage across R_L ap-

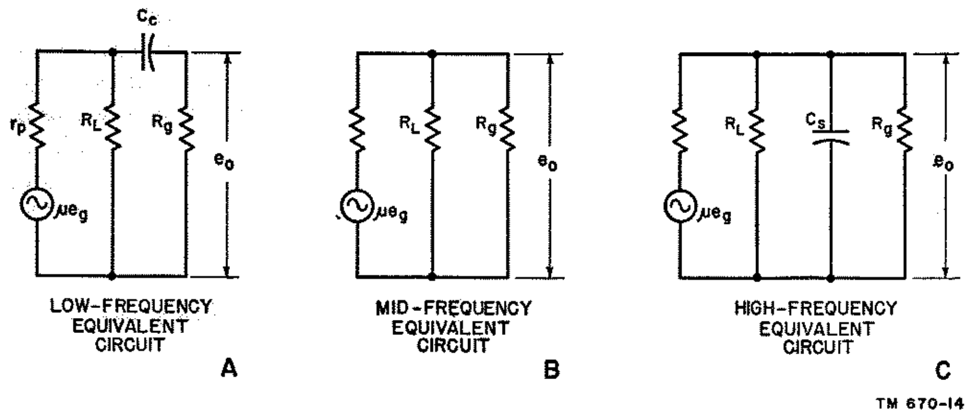

Figure 16. Low-, middle-, and high-frequency equivalent circuits of R-C coupled amplifier.

pears across the output. At high frequencies, the output terminals are shunted by C_s, as in C. Therefore, the output impedance has its highest value, and the gain is greatest, in the midfrequency range. The response at low and high frequencies is expressed in terms of the *midfrequency gain*, A_M.

11. R-C Amplifier Voltage Gain

The *voltage gain* of an amplifier is the ratio of the output voltage to the input voltage. At middle frequencies, voltage gain A_M of the R-C coupled amplifier is derived from B of figure 16 as

$$A_M = \frac{e_o}{e_g} = -\mu \left(\frac{Z_L}{r_p + Z_L}\right)$$

The term Z_L refers to the load impedance, which consists of R_L in parallel with R_g. It is helpful to rearrange the gain equation into a form which includes the g_m of the tube. This is done by multiplying numerator and demoninator by the same quantity, r_p. The value of the gain equation remains unchanged by this operation. Thus the voltage gain of the amplifier at midfrequency is

$$A_M = \frac{-\mu}{r_p} \quad \frac{r_p Z_L}{r_p + Z_L}$$

The quantity μ/r_p represents g_m. The fraction $r_pZ_L/(r_p + Z_L)$ is the equivalent resistance, R_{eq}, formed by the parallel combination, R_{eq}, of the dynamic plate resistance, r_p, load resistor R_L, and grid resistor R_g. Therefore, the voltage gain, A_M, is given by

$$A_M = -g_m R_{eq}$$

12. Constant-Current Equivalent Circuit

a. The gain equation, $A_M = -g_mR_{eq}$, leads to a very useful revision of the constant-voltage equivalent circuit. The output voltage, e_o, at midfrequencies is A_M times the grid-signal voltage, e_g, or $e_o = -g_me_gR_{eq}$, or $-(\mu\ e_g/r_p)/R_{eq}$. It is evident that quantity $\mu\ e_g/r_p$ represents a current since it consists of a voltage, $\mu\ e_g$, divided by a resistance, r_p. The amplifier output voltage, e_o, thus is produced by a current, $\mu\ e_g/r_p$, or g_me_g flowing through R_{eq}, which consists of three parallel impedances—r_p, R_L, and R_g.

b. Figure 17 shows the equivalent circuit which represents these circuit conditions for the R-C coupled amplifier in the low-, middle-, and high-frequency regions. This equivalent circuit frequently is known as *constant-current equivalent circuit*, because the generated current, $\mu\ e_g/r_p$, or g_me_g, consists of the three factors, μ, e_g, and r_p, which are assumed constant when an amplifier circuit is analyzed.

c. Assume, as a numerical illustration of the merits of the constant-current equivalent circuit (fig. 17), that a certain pentode tube has a μ of 500, an r_p of 100,000 ohms, and therefore, a $g_m = \mu/r_p = 5{,}000$ umhos. This tube is to be used as an amplifier with a load resistor of 4,000 ohms. It is required to find the midfrequency voltage gain of this amplifier using both the constant-voltage equivalent circuit and the constant-current equivalent circuit. Using the constant-voltage equation, $A_M = -\mu\ Z_L/(r_p + Z_L)$, the gain is $-(500$ times $4{,}000)/\ 104{,}000 = -19.2$. Using the constant-current equation,

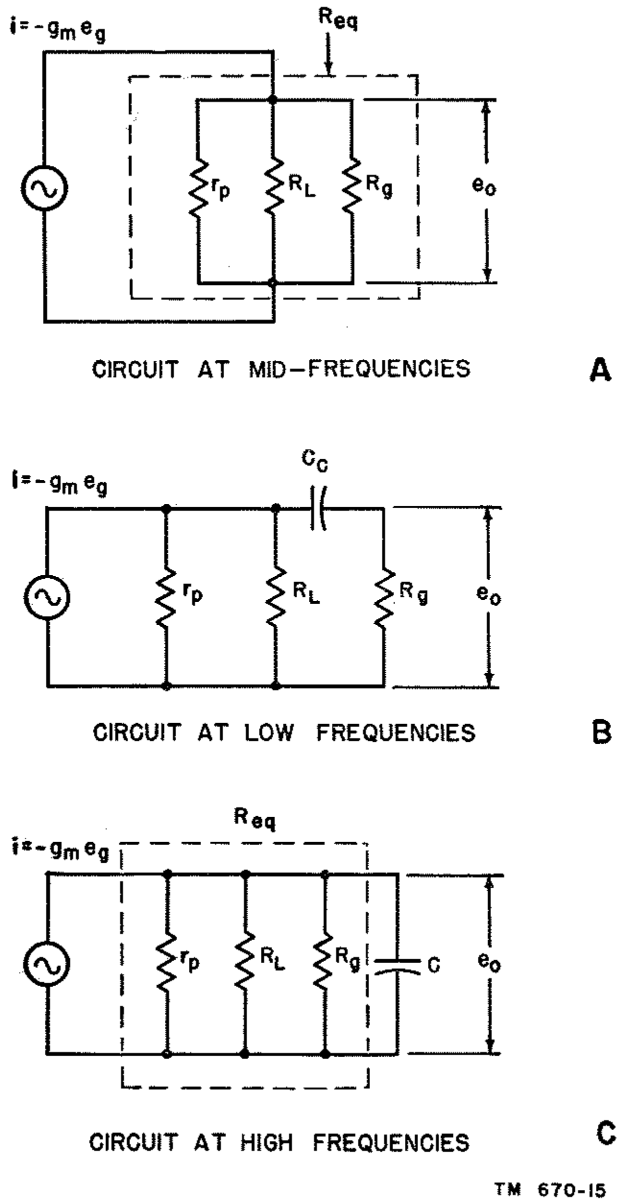

Figure 17. Constant-current equivalent circuits of R-C coupled amplifier.

$A_M = -g_m R_{eq}$, it is necessary first to find R_{eq}, which in this case consists of r_p in parallel with R_L, or (100,000 times 4,000) divided by 104,000, or 3,840 ohms. The gain $A_M = -(5{,}000 \text{ times } 10^{-6})\ (3{,}840)$, or —19.2, as before.

d. It usually is not necessary to calculate R_{eq} with great accuracy when the load resistor is much smaller than the plate resistance of the tube (fig. 17). In the illustration just given, the R_{eq} of 3,840 ohms is practically equal to the load resistor of 4,000 ohms. Hence, the mid-frequency gain can be written as approximately $-g_m R_L$. Using this approximation, the gain comes out to $-(5{,}000 \text{ times } 10^{-6})\ (4{,}000) = -20$, which is in error by a negligible 4 percent. This example shows the convenience of the constant-current equivalent diagram, especially when the plate resistance is considerably larger than the load resistance.

13. Elements Influencing Gain

a. Elements Influencing Gain at Midfrequencies. The gain expression, $A_M = -g_m R_{eq}$, shows that a high-gain R-C amplifier uses a high $-g_m$ tube having a large dynamic plate resistance, as in pentodes, and large values of plate load and grid resistors. The d-c operating conditions of the tube control g_m and r_p directly (fig. 17). The dynamic plate resistance, r_p, is controlled also by the size of load resistor R_L, since R_L determines the d-c potential applied to the plate. R_L usually is fixed for a given g_m. The grid resistor, R_g, is made as large as possible to obtain high gain at middle frequencies. Grid resistors are usually limited to 1 megohm for many tube types, in order to prevent excessive biasing as a result of rectified grid current.

b. Elements Influencing Gain at Low Frequencies. Coupling capacitor C_c and grid resistor R_L form a voltage divider which reduces voltage e_o appearing across grid resistor R_g (fig. 17). The voltage gain of the amplifier at *low frequencies* is, therefore, lower than in the middle-frequency range. The reactance of the capacitor increases as frequency goes down, since $X_C = \frac{1}{2}\pi fC$. It is essential to use a large value of C_c which offers negligible reactance at the lowest frequency that must be amplified. Good low-frequency gain is obtained also by using a grid resistor, R_g, which is much larger than the reactance of C_c at the lowest required frequency. Often it is useful to consider the product of $R_g C_c$, known as the time-constant product. Good low-frequency response is obtained by using a *large* time-constant product since R_g and C_c each must be large.

c. Elements Influencing Gain at High Frequencies.

(1) The load impedance of the tube at *high frequencies* includes the parallel combination of three capacitors: plate-to-

cathode output capacitance C_o of the tube, the distributed, or stray, capacitance, C_d, and input capacitance C_i of either a following stage, or of a load. The small series reactance of the coupling capacitor, C_c, is negligible at high frequencies, as it is equivalent to a short circuit (fig. 17). The voltage gain of the amplifier at high frequencies is reduced because the reactance of C_s reduces the total load impedance of the tube as the frequency increases. It is essential to keep the total shunting capacitance, C_s, at a value which offers negligible shunting reactance to the load at the highest frequency which must be amplified. The tube interelectrode capacitances, C_o and C_i (included in C_s), are inherent in a given tube type and cannot be altered. The distributed capacitance, C_d, is held to a minimum by using short leads, by using high-quality low-capacitance tube sockets, and by mounting all coupling elements above the chassis and away from it.

(2) Another method of reducing the shunting effect on the load impedance of the reactance of C_s is to reduce the value of shunting resistor R_L. This extends the flat range at its end, but greatly reduces the gain at all frequencies, since R_{eq} is lower in the equation $A_M = -g_m R_{eq}$.

(3) It often is useful to consider the product of $R_{eq}C_s$ the time-constant product for high frequencies. Good high-frequency response is obtained by using a small high-frequency time-constant product, since R_{eq} and C_s each must be small.

14. Universal Response of R-C Coupled Amplifier

It is possible to draw a single-frequency response curve that applies to all R-C coupled amplifiers (fig. 18). This curve helps to clarify R-C coupled amplifier operation and is known as the *universal response curve.* It is based on the assumptions made in developing the equivalent circuits of figures 13, 16, and 17.

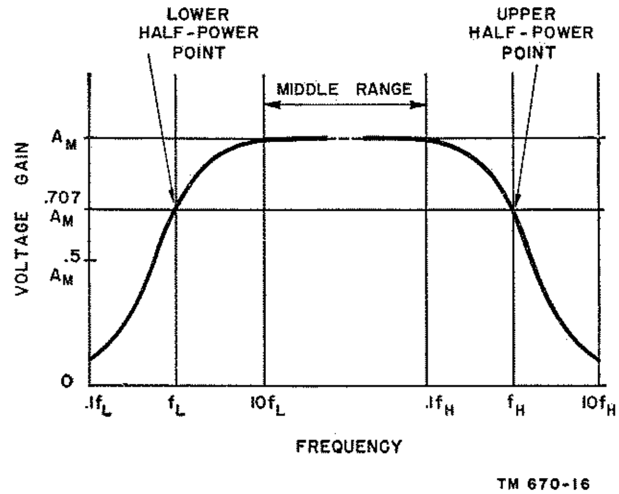

Figure 18. Universal response curve of R-C coupled amplifier.

a. The high-frequency portion of the universal response curve (fig. 18) is developed as follows. The voltage gain of the R-C amplifier drops to 70.7 percent of the midfrequency gain at the high-frequency point, f_H, where the reactance of shunt capacitor C_s has the same value as the resistance of R_{eq}. At this frequency, the resultant load impedance of the tube is 70.7 percent of R_{eq}. The voltage appearing across grid resistor R_g at this frequency is .707, or $1/\sqrt{2}$, times the voltage appearing across R_g at midfrequency. The a-c power developed in R_g is proportional to the square of the voltage impressed across R_g. Hence, the power developed in the grid resistor at frequency f_H is one-half of the power developed in R_g at middle frequencies. For this reason frequency f_H is known as the *high-frequency half-power point,* or the *upper half-power point,* or the *upper half-power frequency.*

b. The frequency response of an R-C amplifier is considered flat so long as it remains within ½ of 1 percent of the maximum- or middle-frequency response. The boundaries of flat response are located, therefore, at the frequencies where the response is 99.5 percent of middle-frequency response. The high-frequency boundary of flat response in the R-C amplifier lies at one-tenth of f_H (fig. 18), because at this frequency the total load impedance is 99.5 percent of its maximum, or middle-frequency, value.

c. The low-frequency portion of the universal response curve is developed in the same manner as the high-frequency portion. At the lower half-power frequency, f_L (fig. 18), the reactance of coupling capacitor C_c has the magnitude of R_g plus the parallel combination of r_p and R_L. The parallel combination of r_p and R_L frequently is neglected because it usually is much smaller in magnitude than R_g. At the half-power frequency, f_L, the voltage across grid resistor R_g is .707, or $1/\sqrt{2}$, times the middle-frequency value. The power developed in R_g at frequency f_L is, therefore, one-half of the power developed in R_G at middle frequencies. Hence f_L is called the *low-frequency half-power point*, or the *lower-half-power point*, or the *lower half-power frequency*. The low-frequency boundary of flat response lies at $10f_L$, because at this frequency the voltage across R_G is 99.5 percent of its maximum, or midfrequency, value.

d. Figure 18 shows that the middle-frequency portion of the universal response curve extends between $10f_L$ and $.1f_H$. It is drawn horizontal or flat, because in this range the response varies by less than ½ of 1 percent. The exact midfrequency is the geometric mean of the half-power point frequencies $f_M = \sqrt{f_L f_H}$.

e. A numerical example will make these distinctions clear. A resistance-coupled amplifier is flat between 100 and 10,000 cycles. Therefore, since 100 cycles corresponds to $10f_L$, 10 cycles is the lower half-power frequency at which the voltage gain is .707 times the midfrequency gain. Similarly, since 10,000 cycles corresponds to $.1f_H$, 100,000 cycles is the upper half-power frequency at which the voltage gain is .707 times the midfrequency gain. The exact midfrequency is $\sqrt{(10)\ (100{,}000)}$, or 1,000 cycles.

f. Logarithmic scales often are used to plot frequency and a-c voltage, because the response of human senses is very nearly logarithmic. A logarithmic plot of response shows a more realistic picture of amplifier characteristics under operating conditions. The *exact midfrequency* is the average of the half-power points taken logarithmically. The output voltage of an R-C coupled amplifier is expressed frequently in *db* (*decibels*) relative to the midfrequency output voltage. At the half-power points the response is said to be *down to 3 db* from midfrequency response because the power delivered to R_g at the half-power points is one-half of the power delivered at midfrequency. The db response is plotted on linear ordinates, since *db* is 10 times the logarithm of the power ratio. So plotted, the universal response curve of figure 18 appears as in figure 19.

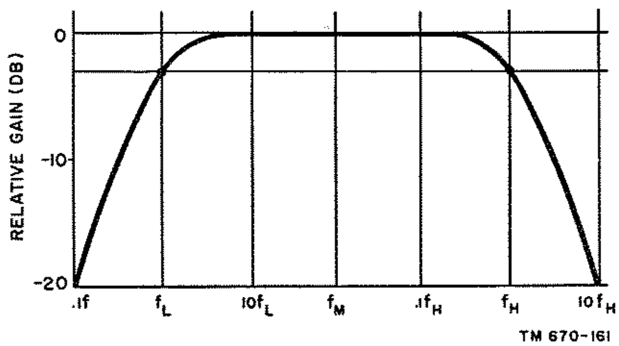

Figure 19. Decibel response curve.

15. Bandwidth

a. The bandwidth of an amplifier is defined as the difference between its two half-power frequencies. It thus is a measure of the frequency range of the amplifier.

b. The flat range of the low-frequency portion of the response curve is extended downward by increasing the low-frequency time-constant product, R_gC_c (fig. 19). The coupling capacitor is limited customarily to $.1\mu f$ (microfarad), however, to minimize d-c leakage to the grid circuit of the following stage, and to avoid excessive shunting capacitance to ground. Grid resistor, R_g, usually is made as large as tube ratings permit. Thus, the low-frequency response of an R-C coupled amplifier cannot be improved beyond a certain point by using this method of coupling.

c. The flat range of the high-frequency portion of the response curve (fig. 19) is extended upward by decreasing the time-constant product, $R_{eq}C_s$. Resistor R_L is the only element which is reduced in order to extend the high-frequency response. The shunting capacitance, C_s, cannot be reduced below a fixed minimum value, after the tube is chosen. R_g determines the extent of the low-frequency response, and it also cannot be altered. The voltage gain of the R-C coupled amplifier thus is lowered as the high-frequency range is extended. Typical single-stage R-C coupled amplifiers using pentode tubes have voltage gains up to 370 and a response range which is flat between approximately 20 cycles and 5,000 cycles. However, the voltage gain drops to unity as the high-frequency response is extended into the megacycle range by reducing the value of R_L. Other methods of extending the bandwidth are discussed in the sections on wide-band (video) amplifiers.

16. Effect of Bypass Capacitors on Low-Frequency Response

In analyzing amplifier circuits it usually is assumed that the power-supply feed lines, including bypass capacitors as well as the power supply itself, present no series impedance to the signal variations. A cathode bypass capacitor is used, in the case of cathode bias, to reduce to a small value the a-c impedance between cathode and ground. The cathode-bias circuit impedance is, therefore, usually not shown in a-c equivalent circuits. The plate- and screen-voltage supply circuits are bypassed in a similar manner, and their impedances usually are negligible. At very low frequencies, however, the reactance of a bypass capacitor becomes appreciable, and the very-low-frequency response of an amplifier falls off more rapidly than when the power-supply impedance is actually close to zero. A method for minimizing the effects of insufficient bypassing is discussed in chapter 3.

17. Response of Multistage Amplifier

The voltage gain of a multistage amplifier is the product of the individual stage gains. The individual stages of a cascade amplifier must be flat over a much wider range than the entire amplifier chain. Assume, for example, that a cascade amplifier consists of four identical R-C coupled stages. The midfrequency gain of each stage is 10 at 5,000 cycles, and the half-power points of each stage are at 50 cycles and at 500,000 cycles. The over-all gain of the amplifier at midfrequency is the product of the individual gains, or 10,000. However, the over-all gain drops to 2,500, or one-quarter of midfrequency gain, at 50 cycles and 500,000 cycles, the half-power point frequencies of the individual stages.

18. Pentodes Versus Triodes

Pentodes generally are preferred over triodes as vacuum tubes in R-C coupled amplifier stages. The plate resistance of a pentode permits greater midfrequency gain. This is seen from the gain expression $A_M = -g_m R_{eq}$, in which a high value of r_p contributes to high values of both g_m and R_{eq}. The total shunting capacitance, C_s, is lower for a pentode because the dynamic input capacitance of a pentode is considerably lower than the dynamic input capacitance of a triode. The high-frequency boundary of the flat frequency-response region is, therefore, higher in an R-C amplifier using a pentode than in one using a triode. A pentode amplifier requires a screen-voltage supply and an adequate bypass capacitor. This is a minor complication, considering the benefits of a greater gain and greater bandwidth obtained with the pentode.

Section II. IMPEDANCE-COUPLED AMPLIFIERS

19. Impedance Coupling

a. Advantages. One disadvantage of the R-C coupled amplifier is that the midfrequency gain is limited by the size of the plate load resistor, R_L. A large plate load resistor produces a large d-c voltage drop between the plate-power supply and the plate of the amplifier tube, thus reducing the plate current which in turn lowers the g_m of the tube. A reduction of g_m lowers the midfrequency gain, A_M, as is seen from the equation $A_M = -g_m R_{eq}$. It is possible to minimize this large voltage drop produced by the load resistor by replacing R_L with an iron-core inductor (fig. 20) having an extremely large value of inductance and a negligibly small value of resistance. Amplifier circuits using this method of coupling are known as *impedance-coupled amplifiers.*

b. Analysis. The operation of impedance-coupled amplifiers is analyzed in the same manner as the operation of R-C coupled amplifiers, since the circuits are quite similar. Thus, the

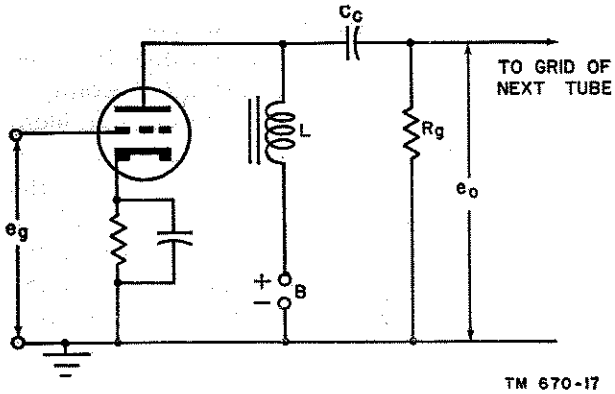

Figure 20. Impedance-coupled amplifier.

frequency spectrum over which impedance-coupled amplifiers operate is again divided into three convenient ranges—low, middle, and high frequency. Impedance-coupled circuits are used chiefly for narrow-band amplifiers because their response drops rapidly at both low and high frequencies.

c. Midfrequency Response.

(1) The gain of an impedance-coupled amplifier in the midfrequency range is $A_m = -g_m Z_{eq}$, where Z_{eq} consists of the parallel combination of plate resistance r_p, load impedance Z_L, and grid resistor R_g. The reactance of a large-value inductor, L, is generally very high compared with either R_g or r_p. Therefore, the shunting effect of L is considered negligible at middle frequencies. This assumption simplifies the gain equation at midfrequencies to $A_M = -g_m R_{eq}$, where R_{eq} consists of the parallel combination of plate resistance r_p and grid resistor R_g.

(2) A numerical example will clarify these statements. Assume that it is desired to find the voltage gain at 1,000 cycles of an impedance-coupled amplifier using a triode with a μ of 20, a dynamic plate resistance of 10,000 ohms, a 500-henry load inductor, and a 1-megohm grid leak. It simplifies matters to determine first whether the reactance of the inductor may be neglected compared with the parallel combination of r_p and R_g. At 1,000 cycles, $X_L = 2\pi f_L = 3.14$ megohms; however, r_p in parallel with R_g is approximately 10,000 ohms; consequently, the shunting influence of the inductor is negligible, since its reactance is more than 100 times as large as r_p. The gain in the middle region of frequencies is, therefore, $A_M = -g_m r_p$. The quantity $g_m r_p$ is equal to μ. Thus, the gain of the impedance-coupled amplifier at midfrequencies is the same as the μ of the tube.

(3) Typical inductance values of inductor L lie between 50 henrys and 1,000 henrys. The d-c resistance of the inductor windings is usually less than 1,000 ohms.

d. Low-Frequency Gain.

(1) The gain of the impedance-coupled amplifier drops sharply at low frequencies. This is because the reactance of the inductor, L, becomes smaller as the frequency is reduced, thus increasing its shunting effect. This effect is in addition to that of the coupling capacitor, C_c.

(2) In order to maintain a fairly uniform response down to relatively low frequencies, it is important to use a large value of inductor and a tube having a small dynamic plate resistance, such as a triode. This type of impedance-coupled amplifier is uneconomical for low-frequency applications because of the high cost of large-value iron-core inductors having small distributed capacitances and low losses.

e. High-Frequency Gain. The gain of the impedance-coupled amplifier drops at high frequencies because the relatively large inductor needed for good low-frequency response generally has an unavoidable large distributed capacitance which shunts the load impedance at high frequencies. This distributed capacitance is so much greater than the interelectrode capacitance of the following tube that the latter is negligible by comparison. The shunting effect of the inductance itself is entirely negligible above the middle range of frequencies. The response curves of impedance-coupled and of R-C coupled amplifiers are very similar in the high-frequency range.

20. Double Impedance Coupling

Inductors sometimes are used to replace both load resistor R_L and grid resistor R_g in order to accentuate the low-frequency response of the amplifier. Amplifiers which use both a plate inductor and a grid inductor are known as *double impedance-coupled amplifiers.* A resonant peak occurs in the low-frequency range at the series-resonant frequency of the grid inductor and the coupling capacitor. This method of peaking the low-frequency response is uneconomical because it requires a second inductor, and it is undesirable from an operating viewpoint because of the wide phase-angle fluctuations which take place in the vicinity of series resonance.

21. Tuned-Impedance Coupling

A very high voltage gain is obtained at one frequency by using a shunting capacitor, C, to tune the load inductance to antiresonance (fig. 21). This method of tuned-impedance coupling is employed at low frequencies for amplifying only one frequency, as required for sending telegraph code signals or control signals for servo systems. In high-frequency applications, tuned-impedance coupling is used in radio-frequency amplifiers. At antiresonance, the impedance of the tuned-load circuit is very high compared with r_p and R_g. Consequently, the gain of the tuned-impedance amplifier at antiresonance is very nearly equal to $-g_m R_{eq}$, where R_{eq} is the parallel combination of r_p and R_g. Thus, pentodes are preferred in tuned-impedance-coupled amplifiers because of their inherent high plate resistance which makes R_{eq}, of the gain equation, large. Tuned-impedance-coupled amplifiers are discussed in connection with radio-frequency amplifier circuits.

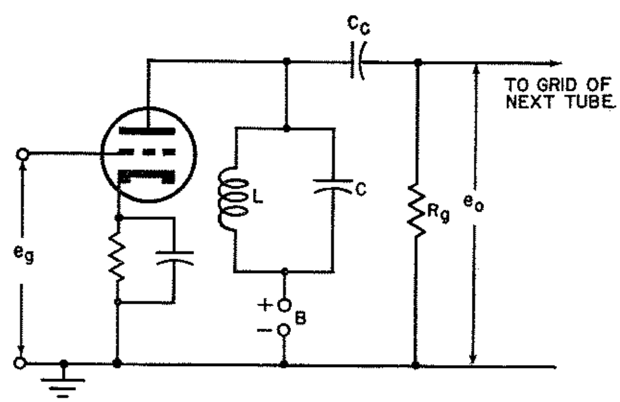

Figure 21. Tuned-impedance-coupled amplifier.

Section III. TRANSFORMER-COUPLED AMPLIFIERS

22. Transformer Coupling

a. Amplifier circuits using transformers as coupling elements are called transformer-coupled amplifiers. Figure 22 shows the circuit of a typical amplifier using both transformer coupling elements and R-C coupling elements. It is seen that transformers function either as *voltage transformers* or as *power transformers.*

b. As shown in figure 22, the first stage of the amplifier uses a voltage transformer known as an *input transformer* in order to step up the feeble signal voltage of a microphone or a telephone transmission line. Input transformers usually have a large *step-up turns ratio.*

c. The output of the second voltage amplifier stage is coupled to the grid of the power amplifier stage by a voltage transformer identified as an *interstage transformer.* Typical interstage transformers usually have a step-up turns ratio of one to three.

d. The output of the power amplifier stage is coupled with an output transformer to a load resistor, which is used (fig. 22) to represent, for example, a loudspeaker. Output transformers are power transformers since they deliver the relatively high power drawn by the load. Output transformers generally have step-down ratios, because in reducing the output voltages developed by the power-amplifier stage they increase the current to the larger values required by the load.

e. The analysis of any of the types of coupling transformer mentioned above is facilitated by an equivalent circuit that makes use of a theoretical transformer. The properties of this *ideal transformer* will be considered first, and the ideal transformer concept then will be

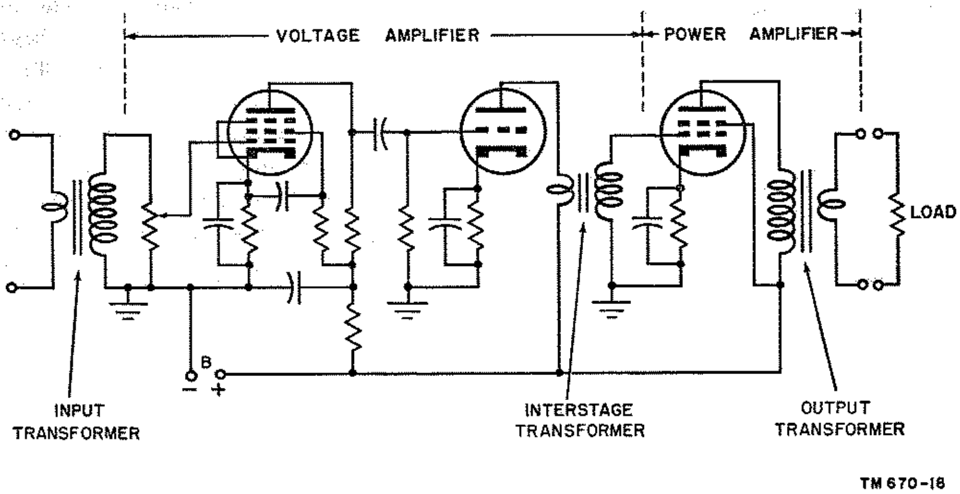

Figure 22. Typical multistage audio amplifier.

applied to explain the performance of actual transformers.

23. Ideal Transformer

a. Definition and Purpose. An *ideal transformer* is a theoretical transformer which has no losses whatever—that is, it consumes no power. Of course, such a transformer does not exist, but its ideal characteristics are set up for the purpose of simplifying the analysis of practical transformers. (The ideal transformer is very nearly attainable in practice.) The use of an ideal transformer in an equivalent circuit permits the representation of the appreciable losses of an actual transformer as though they took place in a simple circuit which is entirely separate from the transformer itself.

b. Characteristics. It follows from the definition of the ideal transformer that the load connected to its secondary terminals (fig. 23) receives all the power delivered to the primary terminals by the generator. The input power, P_i, is equal to E_1I_1, and the output power, P_o, is equal to E_2I_2. Since the input and output powers are equal, it follows that E_1I_1 is equal to E_2I_2.

c. Turns Ratio. The ratio of the primary to the secondary voltage of an ideal transformer is the same as the ratio of the primary to the secondary turns; that is—

$$\frac{E_1}{E_2} = \frac{N_1}{N_2}$$

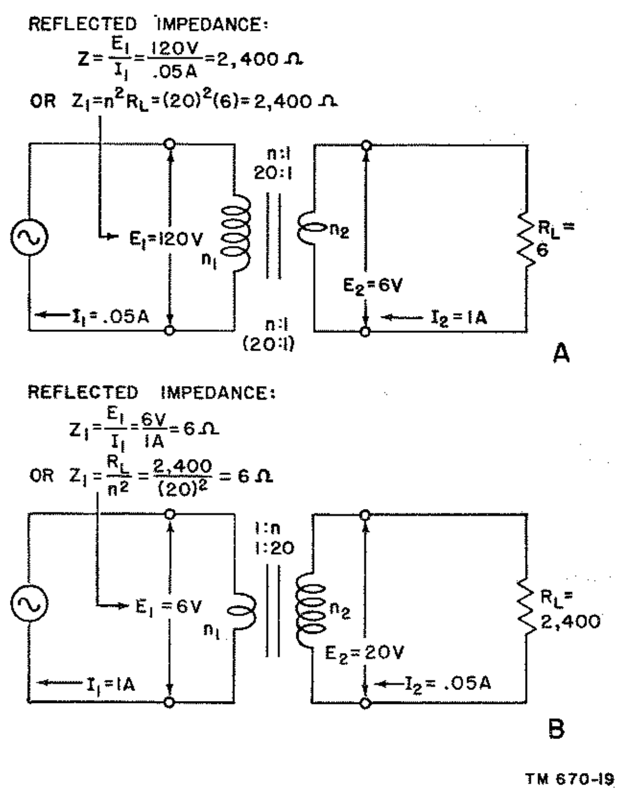

Figure 23. Ideal-transformer impedances.

If an ideal transformer in an equivalent circuit is to serve only to isolate the secondary from the primary circuit, it has a turns ratio of one to one (also written 1:1). If, however, the secondary voltage is lower than the voltage applied to the primary, the transformer has a step-down turns ratio; and if the secondary voltage is higher than the primary voltage, the transformer has a step-up turns ratio.

d. Current Ratio. Since the product of E times I is constant on both sides of the ideal transformer—E_1I_1 equaling E_2I_2, as shown above—if the voltage is stepped up, the current is stepped down, and vice versa. Thus, the current ratio is the inverse of the voltage ratio; that is,

$$\frac{I_2}{I_1} = \frac{E_1}{E_2}$$

e. Reflected Impedance. The ratio of the primary voltage and current, E_1/I_1, is known as the *reflected impedance* of the secondary as seen in the primary. It is *not* equal to the actual secondary impedance. A numerical example will make this relation clear. Assume that in the transformer of A, figure 23, the 6-ohm load resistor, R_L, has 6 volts impressed across it (E_2) and, therefore, 1 ampere flowing through it (I_2). Let the step-down turns ratio of the transformer, N_1/N_2, be represented by n. It has a value of 20:1. The generator, therefore, is supplying the primary terminals of this ideal transformer with 20 times 6 volts, or 120 volts at 1/20 ampere, or 50 milliamperes. The generator thus *operates into,* or *sees,* an impedance which is calculated by Ohm's law as 120 volts divided by 50 milliamperes, or 2,400 ohms. Therefore, the impedance seen by the generator is 20^2 times, or 400 times, as great as the 6-ohm load impedance. The 2,400-ohm impedance seen by the generator is known as the impedance reflected by the load into the primary, or briefly, the *reflected impedance.* Stated generally, the reflected impedance seen by the generator is equal to the square of the turns ratio, multiplied by the actual secondary impedance:

$$Z_1 = \frac{(N_1)^2 Z_2}{N_2} = n^2 Z_2$$

The example shows that the reflected impedance is larger than the load impedance because it is reflected across the larger number of turns of the transformer. In the step-up transformer shown in B, figure 23, the reflected impedance seen by the generator is smaller than the load impedance because it is reflected across the smaller winding of the transformer.

24. Practical Output Transformer

a. An *output transformer* is used to couple the last stage of an amplifier to a load such as a loudspeaker. Practical output transformers differ from ideal transformers in that the practical units necessarily have a finite winding resistance resulting from the wire used in the primary and secondary windings. In practical transformers, the losses caused by eddy currents and hysteresis are generally negligible, because they are very small compared with the winding resistance losses.

b. The fact that typical output transformer efficiencies range between 80 percent and 95 percent, instead of being the theoretical 100 percent of the ideal transformer, introduces only a small error when computing reflected impedances. The efficiency, therefore, is assumed as 100 percent when calculating reflected impedances, and the output transformer is treated as if it were an ideal transformer.

c. The terminals of actual transformers usually are stamped with numbers that refer to the impedances between which the transformer is designed to operate, not to the transformer-turns ratio. An output transformer intended to have characteristics approaching those of the ideal transformer represented in figure 23 would be marked as shown in A, figure 24. Transformers sometimes are made with tapped windings, so that several different impedance ratios are possible. Thus, the transformer in B could be used as an output transformer supplying either a 6-ohm load or a 15-ohm load. The load would be connected to the correspondingly numbered terminals; in either case, the impedance reflected into the primary would be 2,400 ohms.

d. Output transformers often are incorrectly referred to as impedance-matching transformers. An output transformer does not necessarily match impedances. It serves primarily to present to the power-amplifier tube that value of load impedance which is required for obtaining the maximum undistorted power output. To meet this requirement, the reflected impedance which appears across the power-amplifier tube usually is not equal to the internal dynamic plate resistance of the tube; hence, a mismatch

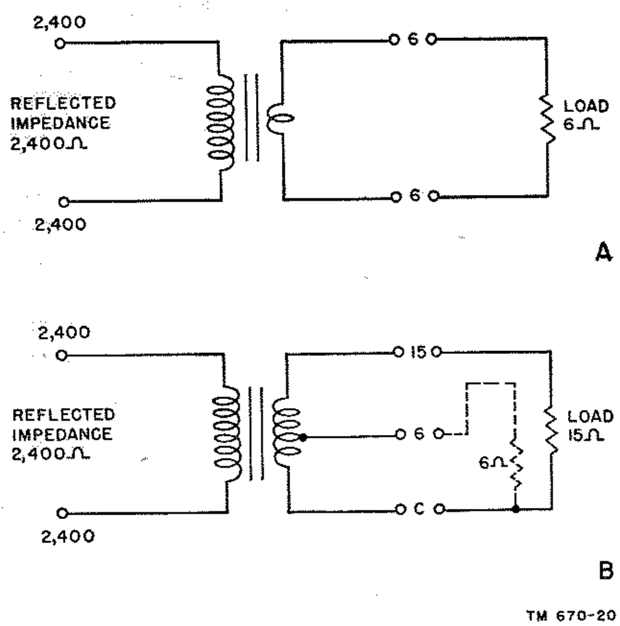

Figure 24. Output-transformer terminal markings.

exists between the internal-plate resistance of the tube and the reflected-load impedance.

25. Interstage Transformer

a. Advantages. Transformers used to couple successive stages of an amplifier generally are known as *interstage transformers.* Interstage-transformer coupling is superior to other coupling means in some respects. The step-up ratio of the transformer permits the amplifier voltage gain to exceed the tube μ; a lower plate-supply voltage is required, since the d-c drop in the transformer primary is so small that almost the full power-supply voltage is applied to the plate of the tube; the circuit is adapted readily to push-pull operation which results in a substantial reduction in even-harmonic distortion.

b. Disadvantages. Disadvantages of interstage-transformer coupling are as follows: The first cost of a transformer is considerably higher than the cost of R-C coupling elements; the transformer frequency-response characteristic extends over a relatively narrow band and is less uniform than with other coupling means; stray a-c fields induce undesirable stray voltages in the transformer; and, finally, interstage-transformer coupling requires an amplifier tube having a low dynamic plate resistance, such as a triode, in order to maintain at least a reasonably uniform response.

c. Frequency Characteristics. The response curve of figure 25 shows the distinguishing characteristics of a typical transformer-coupled amplifier response. These are a relatively uniform midfrequency response, a gradual roll-off at low frequencies, and at higher frequencies a sharper decay in response which is sometimes preceded by a resonant peak.

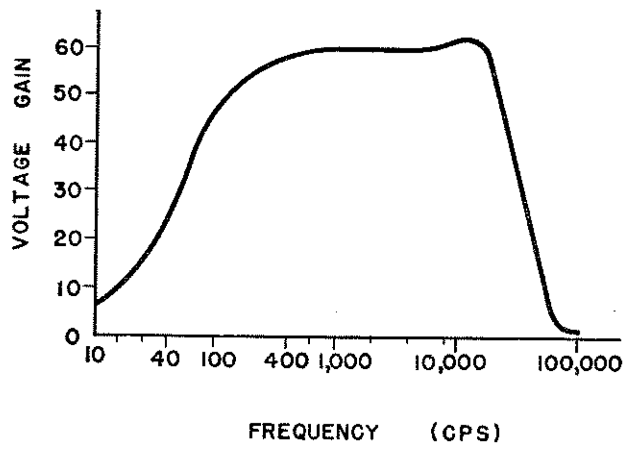

Figure 25. Response curve of typical transformer-coupled stage.

26. Equivalent Circuit of Interstage Transformer

In analyzing an interstage transformer-coupled amplifier, it is desirable to have the largest practical step-up ratio, and at the same time to maintain low losses at all frequencies.

a. In the equivalent circuit of the transformer, shown in B, figure 26, the losses and the step-up are represented by separate circuit sections: the equivalent T section and the ideal transformer. In the equivalent T are shown the components which cause all of the losses in the actual transformer; the ideal transformer is used to show only the step-up ratio. Considered together, the two units comprise an equivalent circuit which represents quite accurately the electrical circuit of the interstage transformer at all frequencies. The proof of its accuracy is lengthy and complicated and is, therefore, omitted from this manual.

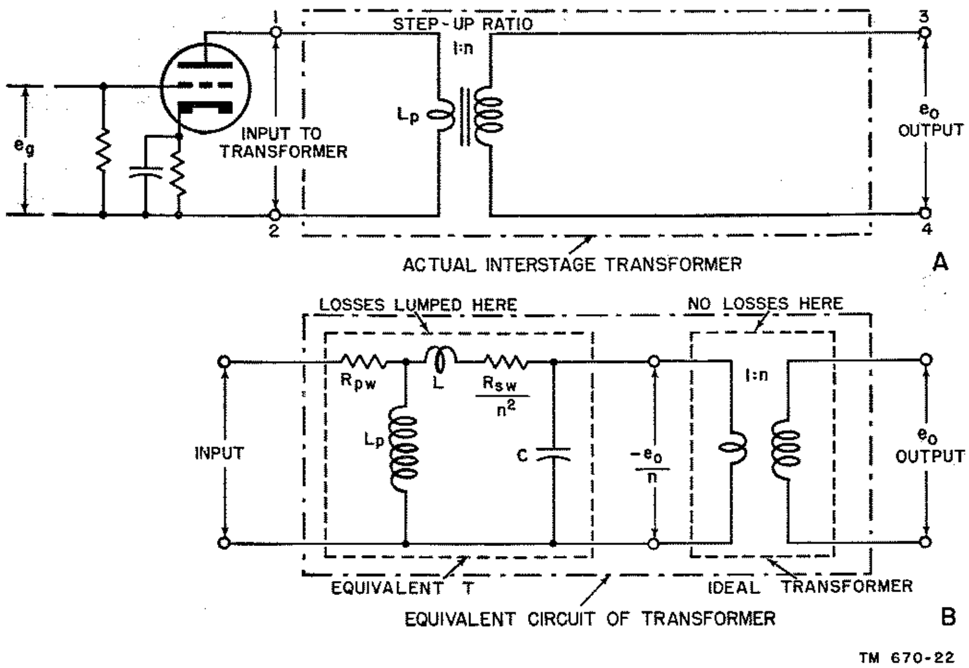

Figure 26. Simplified equivalent circuit of interstage transformer.

b. The shunt arm of the T network contains an inductor, L_p, whose magnitude is almost exactly equal to the primary inductance of the transformer. The left-hand series arm of the T contains a resistor, R_{pw}, which represents the d-c winding resistance of the transformer primary. The right-hand series arm of the T contains two elements. One is $R_{sw}/n2$, the secondary-winding d-c resistance as it appears reflected through a transformer turns ratio; the other is the reflected secondary leakage inductor, L_L. Leakage inductance represents that portion of the total magnetic flux of a transformer which does not link both the primary and the secondary windings. In the interstage transformer the largest leakage flux occurs on the side of the largest number of turns—namely, on the secondary side of the transformer. This leakage inductance is unavoidable, because it is physically difficult to couple perfectly a relatively large winding to a small core. The primary leakage flux is so small in the interstage transformer that it is generally neglected, and therefore, it is not shown in the T circuit. Capacitor C represents, in one lump, all the capacitances inherent in a transformer, including the capacitance between primary and secondary and the capacitances between turns of the windings. The output terminals of the equivalent T feed the input of the ideal interstage transformer.

c. Figure 26 shows that the output voltage of the T is $-e_o/n$, since it appears on the primary of the ideal step-up transformer. The minus sign signifies a phase shift of 180° (polarity reversal) in the ideal transformer. The performance of the interstage transformer is analyzed in a manner similar to that of the R-C coupled amplifier, by investigating the response characteristics at low, middle, and high frequencies.

27. Response of Transformer-Coupled Amplifier

a. Middle-Frequency Gain.

(1) The effects of all elements of the equivalent T are negligible in the middle-frequency range of the interstage transformer. A numerical example will clarify this statement. The equivalent T of B, figure 27, shows numerical values for the reactances at 1,000 cycles of the typical interstage transformer-coupled amplifier stage shown

in A. It is seen that the impedances of the series arms are negligibly small compared with the shunt reactances of L and C. The effects of the shunt reactances of L_p and C, in turn, are also negligible at 1,000 cycles because they are much greater in magnitude than the 10,000-ohm internal impedance, r_p, of the generator. Thus, the equivalent circuit of the transformer-coupled amplifier is reduced to a simple circuit (A, fig. 28), in which all the losses of the T are neglected, and the equivalent generator voltage appears as $-\mu e_g$ across the primary terminals of the ideal transformer, making $-\mu e_g = -e_o/n$.

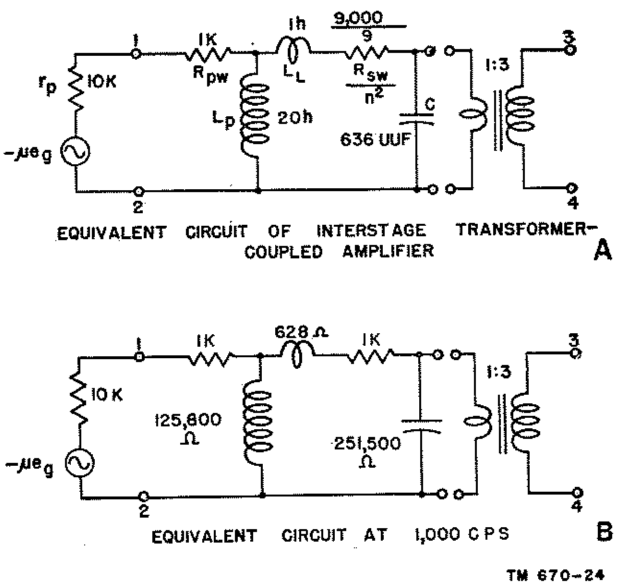

Figure 27. Transformer reactance values at 1,000 cps.

(2) The midfrequency gain, A_M, is found by multiplying both sides of this equation by n, giving $\mu n e_g = e_o$, and then dividing both sides by e_g. The midfrequency gain therefore is—

$$A_M = \frac{e_o}{e_g} = \mu n$$

This shows that the midfrequency voltage gain of the interstage transformer-coupled amplifier is n times as great as the μ of the tube, where n is the transformer step-up ratio. For example, assume that a 6C5 tube is used in a one-stage amplifier with a one-to-three step-up transformer. The μ of the 6C5 tube is 20. The midfrequency gain of the amplifier is, therefore, μn or (20) (3) = 60. Therefore, if a 1-volt signal is applied to the grid of the tube, a 60-volt output appears at the secondary terminals of the transformer.

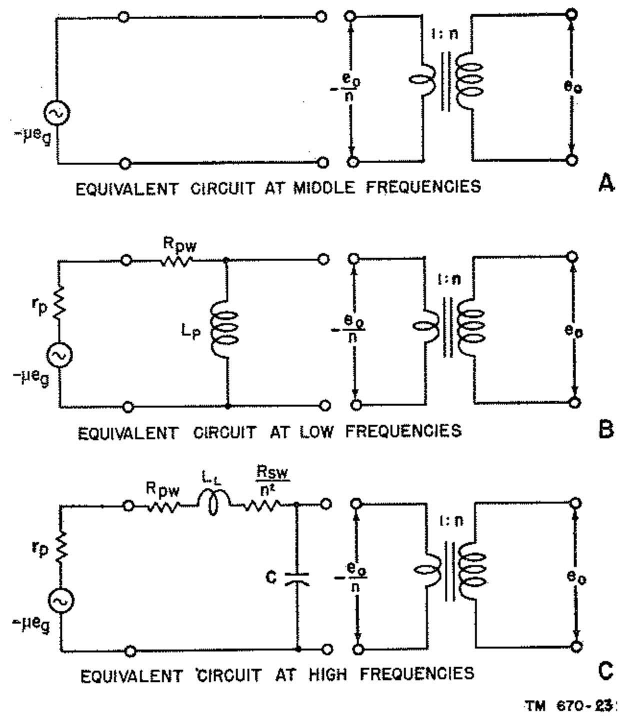

Figure 28. Equivalent circuits of amplifier using interstage transformer.

(3) The polarity of the output voltage of the transformer-coupled amplifier is the same as the polarity of the input voltage applied to the grid of the tube, because the tube and the transformer each contributes a polarity reversal, making the net change zero.

b. Low-Frequency Gain.

(1) The shunt inductive reactance of the primary, L_p (A, fig. 27), is not negligible at low frequencies, because it drops rapidly as the frequency of the applied signal is reduced. L_p is, therefore, shown on the low-frequency equivalent circuit of B, figure 28. The shunt capacitive reactance of C (A,

fig. 27) is negligible, however, because it is even larger at low frequencies than at midfrequencies. Thus, since the current through C is negligibly small, the output of the equivalent T is practically an open circuit at low frequencies. For this reason, it is permissible, at low frequencies, to replace the second series arm of the T by a short circuit (B, fig. 28), since the negligible capacitor current produces a negligible drop in the series elements, L_L and R_{sw}/n^2. The combination of R_p, R_{pw}, and L_p acts as a voltage divider connector across the equivalent generator, $-\mu e_g$.

(2) A good low-frequency response is obtained by minimizing the voltage drops in r_p and R_{pw}. These requirements compel the use of a triode having a low plate resistance and a transformer having negligible d-c losses.

(3) The permissible range of values of inductance and resistance of the primary winding of a transformer are limited by the gain requirements at high and low frequencies. A small primary inductance reduces the low-frequency response, and a very large primary inductance reduces the high-frequency response.

c. High-Frequency Gain.

(1) The shunt reactance of the primary inductance L_p (A, fig. 27) rises with frequency ($X_L = 2\pi fL$) and, therefore, at high frequencies has negligible effect as compared with the plate resistance, r_p. For example, the reactance of primary L_p is 125,600 ohms at 1,000 cps, as shown in the illustration (B, fig. 27). As the frequency is increased to 10 kc, the reactance of primary L_p rises to 1,256,000 ohms. However, the reactance of C, which decreases with frequency ($X_C = 1/2\pi fC$), is only 25,150 ohms at 10 kc (kilocycles). This is no longer negligible, compared with r_p. Thus, the high-frequency equivalent circuit reduces to a series circuit consisting of r_p, R_{pw}, R_{sw}/n^2, inductor L_L, and capacitor C (C, fig. 28).

(2) At the high-frequency end of the response curve, the voltage across C rises as L_L and C become series resonant. This rise depends on the Q of the circuit. For example, in the circuit shown in B, figure 27, the resonant frequency of L_L and C is 20 kilocycles and the Q at resonance is, therefore, X_L ($2\pi fL$ = 6.28 times 20,000 times .1) divided by the total series resistance, or 12,560 divided by 12,000 (the sum of 10,000 plus 1,000 plus 1,000), which equals 1.046. The voltage across C at resonance thus rises to 1.045 times the midfrequency value, and slightly higher just below resonance. Note the rise in the curve of figure 25.

(3) Above resonance, the response drops rapidly as the reactance of X_{LL} rises and the reactance of C drops. The use of a pentode with its inherent large dynamic plate resistance lowers the Q of the circuit at the series-resonant frequency; hence, the high-frequency response drops rapidly and there is no peak in the response curve of the pentode transformer-coupled amplifier.

(4) This discussion demonstrates that the high-frequency response of a transformer-coupled amplifier using an interstage transformer drops off as the value of C is increased, since the resulting lower resonant frequency lowers the effective Q and causes the response curve to decay at a lower frequency. A small value of capacitance C is required to obtain a good high-frequency response. Since C is controlled by the size of the windings, a small primary winding is desirable for good high-frequency response. This requirement conflicts with the requirements of a large value of primary inductance for good low-frequency response. A one-to-three step-up ratio transformer usually is chosen as a compromise.

28. Response of Output Transformer

a. The equivalent circuit of the output transformer shown in figure 29 differs from the equivalent circuit of the interstage transformer primarily because the output transformer is usually a step-down transformer. It differs also in that the leakage inductance appears chiefly on the primary side because of the larger primary-winding inductance. The capacitances are neglected and not shown because their reactances appear as small values of C divided by n^2, shunting the output of the T. For example, the turns ratio of a typical output transformer is 30 to 1, and a capacitor, C, appears in the equivalent T as divided by 30^2 or 900, resulting in a negligibly small reflected capacitance.

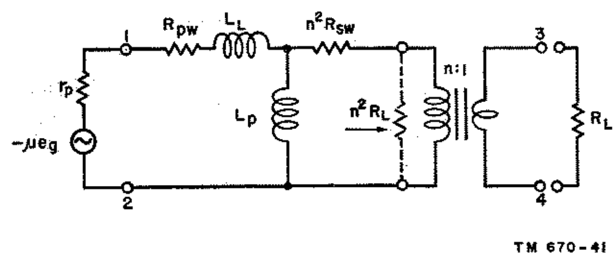

Figure 29. Equivalent circuit of amplifier using output transformer.

b. The resonant peak obtained with the interstage transformer is, therefore, absent in the output transformer, and the response curve is symmetrical on both sides of the midfrequency. Its response curve is identical with that of the R-C coupled amplifier shown in figure 18.

c. A good low-frequency response is obtained by using a large value of primary inductance, L_p, whereas a good high-frequency response is obtained by using a small leakage inductance, L_L. These two requirements conflict, since a large primary inductance produces a large leakage inductance which in turn reduces the high-frequency response. Compromises are made in the manufacture of output transformers in order to provide good response characteristics at both low and high frequencies.

Section IV. DIRECT-COUPLED AMPLIFIERS

29. Direct-Coupled Amplifier

a. Basic Circuit. Vacuum-tube circuits whose coupling networks consist of resistive elements and direct connections are known as *direct-coupled amplifiers* and *d-c amplifiers.* Figure 30 shows the basic circuit of a d-c amplifier consisting of two stages connected in cascade. The chief difference between this method of coupling and R-C coupling is the direct connection between the plate of $V1$ and the grid of $V2$, which eliminates the coupling capacitor used in R-C coupling.

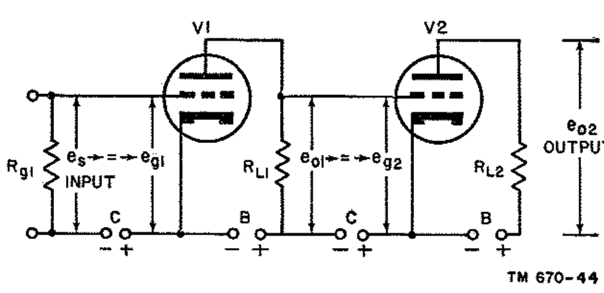

Figure 30. Basic d-c amplifier circuit.

b. Response. Since the reactance of the coupling capacitor is primarily responsible for poor low-frequency response in R-C amplifiers, its omission in direct-coupled amplifiers results in a perfect low-frequency response. The direct-coupled circuit of figure 30 amplifies any voltage applied to its input terminals, since a change in grid voltage e_g results in an amplified change in output voltages e_{o1} and e_{o2}. Consequently, the applied signal voltage may be a direct voltage or it may be a varying voltage. The response of the d-c amplifier is, therefore, the same for slow variations, or signals of very low frequencies, as it is for nonvarying signals, or signals whose frequency is 0 cps (cycle per second). Thus, a direct-coupled amplifier is suitable for measuring and amplifying both d-c and a-c signals. In the high-frequency range, however, the response of d-c amplifiers drops off for the same reasons as in R-C coupled amplifiers; this is because of the appreciable shunt reactances which are formed by the interelectrode capacitances of the vacuum tubes, and the stray and distributed capacitances of the resistive coupling elements.

c. Applications. D-c amplifier circuits are used in vacuum-tube voltmeters, oscilloscope deflection amplifiers, and servomechanism amplifiers. The amplifier load may be some sort of mechanical device such as a relay, a counter, or a meter.

30. Operation of D-C Amplifier

a. Power-Supply Requirements.

(1) In the coupling circuits previously considered the coupling device isolates the d-c voltage in the plate circuit from the d-c voltage in the grid circuits, allowing only the a-c components of the amplifier-stage output to pass through the coupling device. Thus, in the resistance-capacitance and impedance-coupled amplifiers, the coupling capacitor prevents the B-supply voltage from reaching the grid of a succeeding stage. In the transformer-coupled amplifier, there is no direct connection between the primary and secondary windings, again isolating the B-supply voltage. In both cases, only the variations in voltage or current which constitutes the signal are passed on to the succeeding stage.

(2) In the direct-coupled amplifier, however, the plate of one tube is connected directly to the grid of the next tube without any intervening capacitor, transformer, or other coupling device. The plate of a tube requires a positive potential in respect to its cathode; however, the grid of the following tube requires a negative potential in respect to its cathode. This conflicting requirement may be met by connecting a bucking battery in series with the grid-cathode circuit of the second tube (A, fig. 31). This provides the grid with the correct operating potential in respect to its cathode.

(3) A more practical system for supplying the proper voltages to a d-c amplifier is discussed in paragraph 31. The bucking batteries shown in figure 31 are introduced merely to emphasize the power-supply problem to be overcome. They are not employed in practice.

b. Quiescent Operation (fig. 31).

(1) In A, the plate-supply voltage of $V1$ is 300 volts. The quiescent plate current is such that the drop across load resistor R_{L1} is 100 volts. This places the $V1$ plate at 200 volts above ground potential. However, the $V2$ cathode is at ground potential, and the $V2$ grid must be 8 volts below this. The differ-

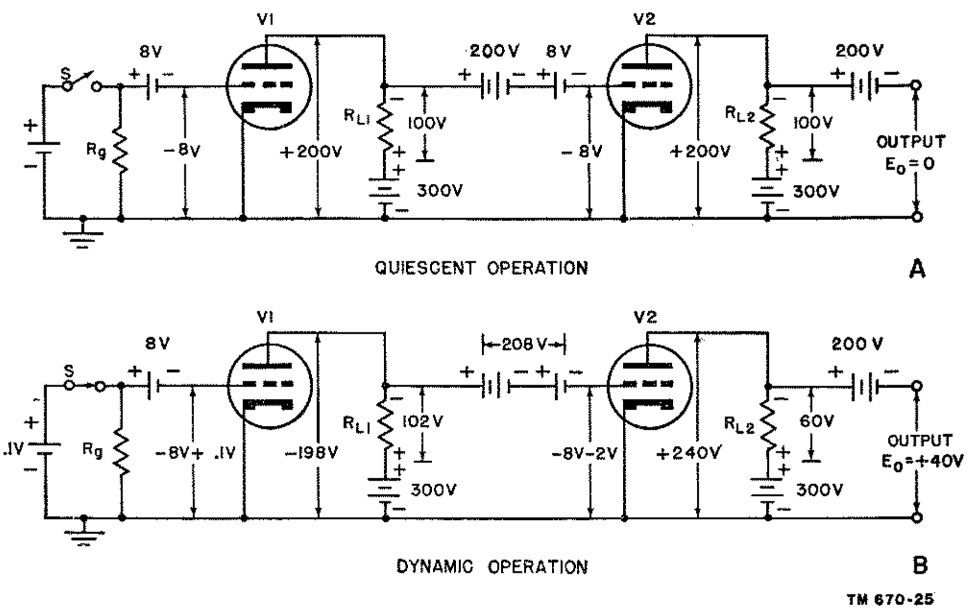

Figure 31. D-c amplifier.

ence of 208 volts between the $V1$ plate and the $V2$ grid is provided by the two bucking batteries shown—a 200-volt battery to overcome the $V1$ plate voltage, and an 8-volt battery to provide the required bias on $V2$.

(2) The output voltage of $V2$ is also 200 volts above ground, but another 200-volt bucking battery in series with the output lead reduces this potential to zero.

c. Dynamic Operation.

(1) Assume, now, that .1-volt positive d-c signal is applied to the grid of $V1$ by closing switch S, in B. This increases the plate current of $V1$ and thus reduces the plate potential from its quiescent value of 200 volts to 198 volts, a negative change of 2 volts. The bucking batteries hold the $V2$ grid at a constant 208 volts below the $V1$ plate. Hence, the $V2$ grid potential must also change by 2 volts, going from its quiescent value of —8 volts to —10 volts.

(2) If the voltage gain of the second stage is the same as the voltage gain of the first stage, namely 20, then a 2-volt *negative* change in its grid voltage produces a 40-volt *positive* change in its plate voltage. Thus, the plate voltage of $V2$ rises from its quiescent value of 200 volts to 240 volts, since the plate current drops as the grid is made more negative. This *change* in the $V2$-plate voltage is the output E_o. Its value is +40 volts. These changes in voltages are shown clearly in B.

(3) This numerical example shows that a direct-coupled amplifier amplifies direct-voltage signals and varying signals, and that the output voltage of a two-stage direct-coupled amplifier is in phase with the input voltage. However, the phase shift in a one-stage d-c amplifier is 180°, since the plate voltage of a tube drops as the grid voltage is increased. It is very important to note that the phase shift in the d-c amplifier remains at a constant value, that is, 180° and 360° (or 0°), in the frequency range where the shunt reactances of the circuit capacitances are negligible.

d. Over-All Characteristics. In the preceding example, a direct-coupled amplifier finds application for amplifying d-c signals, slowly varying signals, and a-c signals. Unfortunately, the d-c amplifier shown is quite unstable because it does not distinguish between a drift in battery potentials and a change in input signals. The .1-volt change in grid voltage can well be caused by a slight drift in bias-battery voltage or in filament temperature. Thus, a false large output voltage is obtained as the amplifier operating conditions vary slightly. This disadvantage is compensated to some extent by using considerably more complex circuits. Another great disadvantage of the circuit of figure 31 is the large number of batteries required. This problem is solved by a direct-coupled circuit called a Loftin-White amplifier.

31. Loftin-White D-C Amplifier (fig. 32)

The *Loftin-White circuit* reduces to one battery or power supply the large number of batteries required by the d-c amplifier of the preceding discussion.

a. Power-Supply Voltages. Use of a voltage divider to supply the various plate and bias voltages required is illustrated in the figure. It is assumed in this example that each tube is operating Class A with 8 volts d-c bias, a quiescent voltage of 200 volts between plate and cathode, and a 100-volt drop in each plate-load resistor, R_{L1} and R_{L2}. The voltage distribution is traced most easily from the negative end of the voltage divider, point A. The point of lowest potential is the grid of the first tube, $V1$. It is connected through grid resistor R_g to point A. The proper grid bias is obtained by connecting the cathode of tube $V1$ to point B on the voltage divider, so that the required voltage drop of 8 volts is produced between points B and A when the proper current flows through the voltage divider. The plate of tube $V1$ is connected through R_{L1} to point D, which is 308 volts above point A. The 100-volt drop across R_{L1} reduces the $V1$ plate voltage to 208 volts. This is the potential to ground for both the plate of $V1$ and the grid of $V2$. The cathode of $V2$ is connected to point C, 216 volts above point A. This places the $V2$ cathode 8 volts positive in respect to the $V2$ grid. Thus, the grid of $V2$ is 8 volts negative in respect to its cathode, as required. The plate load resistor is connected to point E, 516 volts above ground, and 300 volts above the $V2$ cathode. Thus, the quiescent plate-to-cathode potential of tube $V2$ is 200 volts, since the plate potential is 416 volts and the cathode potential is 216 volts.

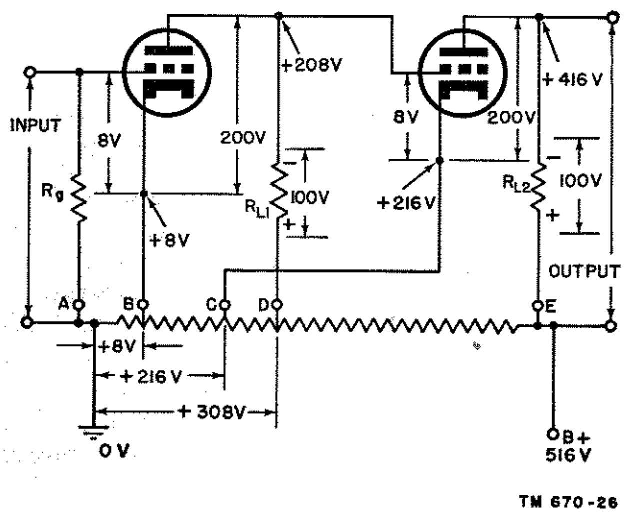

Figure 32. Practical direct-coupled amplifier.

b. Over-all Characteristics.

(1) The entire circuit of the Loftin-White d-c amplifier is a complex resistance network which requires very careful adjustment of cathode- and plate-voltage taps for proper operation.

(2) A power supply is required which is approximately twice the supply voltage for a single stage. Thus, the cathode of the second stage is electrically at a high potential of 216 volts above ground, which is a breakdown hazard in tubes having close spacing between cathode and heater elements. This type of d-c amplifier thus is limited to two stages because the cathodes of successive stages must be raised to successively higher potentials and the plate-supply voltage must be increased as stages are added. For example, the total plate supply voltage for a three-stage amplifier of this type would be 724 volts, and the cathode of the third stage would be connected to a 424-volt tap.

(3) When the tube voltages are adjusted properly for Class A operation, however, the circuit serves as a practically distortionless amplifier, the frequency response of which is uniform starting at 0 cycle up to relatively high frequencies. The circuit is effective as a low-frequency amplifier because the impedances of the coupling elements do not vary with frequency.

(4) Battery power supplies are preferred for this type of d-c amplifier. If a rectifier power supply is used, its output must be entirely free from ripple voltages. Otherwise, the ripples appear in both the grid and plate circuits and are amplified by the successive stages to a disturbing level.

32. D-C Vacuum-Tube Voltmeter

a. One important application of direct-coupled amplifiers is the direct-current or direct-voltage *vacuum-tube voltmeter.* Figure 33 shows such a circuit, which contains only resistors and direct connections. The direct voltage to be measured is applied through terminals 1 and 2 to the voltage divider made up of resistors $R1$, $R2$, and $R3$. The ratio of voltage division is varied by the voltage-range switch, so that several ranges of voltage can be measured. Resistor $R4$ serves primarily as a protective resistor to prevent damage to the meter if too high a voltage is applied to the grid.

b. A milliammeter, M, is inserted in the plate circuit as shown (fig. 33). Shunted across it is a zero-adjustment potentiometer, $R5$, and a bucking battery, E_b. Their purpose is to cause the meter to read zero when no voltage is applied to input terminals 1–2. Without $R5$ and E_b, the quiescent plate current, I_{bo}, would flow through meter M. $R5$, however, is adjusted so that E_b causes a current, I_o, equal in magnitude to I_{bo}, to flow through M in the opposite direction. The net current through M is thus zero. An alternative way of looking at this *balancing*

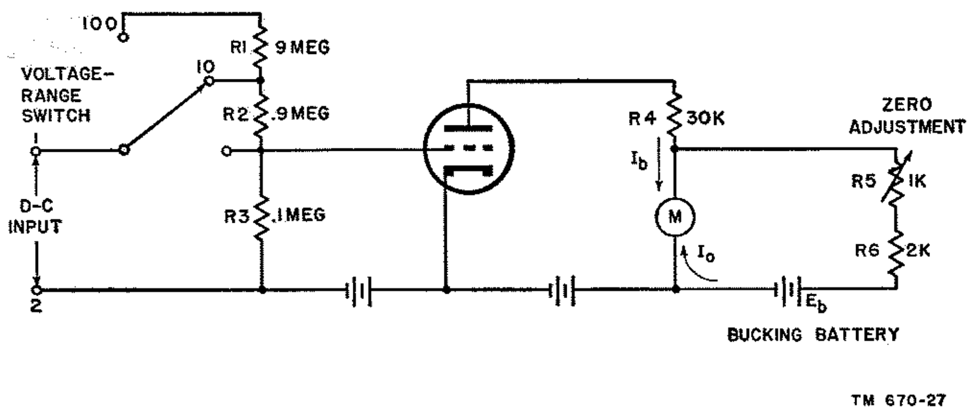

Figure 33. D-c amplifier used as vacuum-tube voltmeter.

of the quiescent plate current is as follows: The balancing elements, $R5$ and E_b, place across the meter a voltage equal and opposite to the drop across M that would be caused by the quiescent plate current flowing through it; all the quiescent current, therefore, flows through $R5$ and E_b, and none of it through the meter.

c. When a positive direct voltage is applied to the input, the meter, M, shows only the resulting change in plate current. The meter reading is proportional to the applied voltage, which is read directly on a suitably calibrated scale. As a practical example, assume that a 1-volt change in grid voltage produces a change in plate current of 1 ma (milliampere), causing meter M to deflect full scale. Thus, this d-c vacuum-tube voltmeter has a range of 1 volt, and the meter scale would be so calibrated.

d. The range of the meter is extended from 1 volt to 100 volts by a simple voltage divider, $R1$, $R2$, and $R3$, having a total resistance of 10 megohms (fig. 33). The divider has taps at 10 megohms, 1 megohm, and .1 megohm, which are selected by the voltage-range switch. With the range switch in the 100 position, only 1/100 of the applied voltage appears across $R3$ because $R3$ is .1 megohm, or 1/100 of the total divider resistance of 10 megohms. Thus, 100 volts applied to d-c input terminals 1–2 appear as 1 volt across $R3$. The actual d-c input voltage is equal to the meter reading times 100. For example, if the scale switch is on 100 and the meter reads .72 volt, the d-c input voltage is 72 volts, or 100 times .72. In a similar manner, the d-c input voltage is divided by 10 when the voltage-range switch is in the 10 position, as shown, because $R3$ is $1/10$ of 1 megohm, or $R2$ plus $R3$.

e. A disadvantage of this d-c voltmeter is its poor stability of calibration. The plate current of the tube varies in a somewhat unpredictable manner with variations in filament temperature, battery potentials, age of the tube, and with variations in resistance of the coupling element with temperature variations. These variations are especially evident when it is attempted to read small voltages accurately. Many complex circuits have been devised for reducing this instability in d-c amplifiers. One method consists of converting the direct voltage into an alternating voltage by interrupting it with a simple *chopper*. The resulting pulsations are more readily amplified, and are measured on an a-c vacuum-tube voltmeter.

33. Balanced D-C Amplifier

Another important use of a d-c amplifier is to measure the difference between two d-c voltages. This is done by means of a bridge circuit with two direct-coupled amplifier tubes serving as the two legs of the bridge (fig. 34). This circuit is known as a *balanced d-c amplifier.*

a. Meter M reads zero as long as no signal is applied between terminals 1 and 2, or between terminals 2 and 3, provided tubes $V1$ and $V2$ are matched properly so that the IR drops across $R1$ and $R2$ are identical.

b. Assume that a positive signal, applied between terminals 1 and 2 (fig. 34), increases the plate current in $V1$ so that the voltage drop

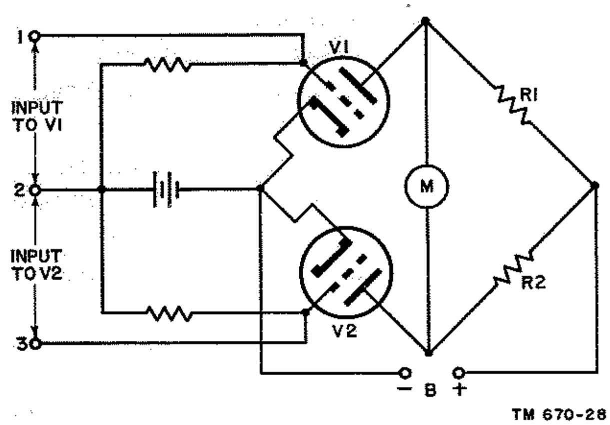

Figure 34. Balanced d-c amplifier.

across $R1$ rises from a quiescent value of 200 volts to 210 volts. Since the drop across $R2$ remains at its quiescent value of 200 volts, meter M reads 10 volts, which is the difference between the two drops across $R1$ and $R2$. Meter M may be a zero-center instrument so that the direction of signal unbalance may be read directly and without changing any connections.

c. In the same manner, it is possible to compare two direct voltages, if one is applied between points 2 and 1, and the other is applied between points 2 and 3 (fig. 34). If the two applied voltages are equal, they produce equal IR drops in resistors $R1$ and $R2$, and the bridge remains balanced. Meter M then reads zero—the difference between the two voltages. If, however, one of the applied voltages is greater than the other, the bridge is unbalanced and the meter indicates the difference between the voltage drops which appear across the plate-load resistors. The direction of meter deflection shows which voltage is greater, since the current flowing through the meter is proportional to the difference between the two applied voltages.

34. Summary

a. The constant-current equivalent circuit shows that the output voltage of an amplifier is $-g_m e_g Z_{eq}$.

b. Resistance-capacitance coupled amplifiers are characterized by a constant gain of $-g_m R_{eq}$ in the middle-frequency range and a gradual roll-off of gain at low and high frequencies. A good low-frequency response is obtained by using a large R_g and C_c. A good high-frequency response is obtained by using small values of R_L and C_s. The lower half-power point frequency, f_L, is where the reactance of coupling capacitance C_c equals R_g plus the parallel combination of r_p and R_L. The upper half-power frequency, f_H, is where the reactance of the total shunting capacitance, C_s, equals R_{eq}. The flat-or middle-frequency range of an R-C coupled amplifier stage extends from $10f_L$ to $.1f_H$. Pentodes have a high gain at middle frequencies, and their high-frequency response drops off less sharply than that of triodes.

c. In impedance-coupled amplifiers, the load resistance, R_L, is replaced by an inductor, thus producing a greater midfrequency gain for a given plate-supply voltage. However, the gain drops rapidly at low frequencies unless a large inductor is used in conjunction with a tube having a low plate resistance, such as a triode. The gain falls off at high frequencies because of the stray and distributed capacitances of the windings of the inductor. Impedance coupling generally is suited to amplification of a single frequency or of a narrow band of frequencies.

d. Transformers are used as input-, interstage-, and output-coupling elements. The input and interstage types of transformers have similar characteristics. Interstage transformers usually have a one-to-three step-up ratio, whereas input transformers have a much higher step-up ratio. The gain of a transformer-coupled amplifier at middle frequencies is μn, and falls off at low frequencies. The gain generally reaches a peak near the high frequency where the leakage inductance is in series resonance with the total capacitance. Beyond series resonance, the response drops off quickly.

e. An output transformer is used to transform a low-impedance load to the proper value of impedance required by the plate circuit of a power amplifier. The efficiency of output transformers usually exceeds 80 percent. The load impedance, seen in the primary of the output transformer, is usually larger than the load impedance and has the value n^2R_L (Where n is the turns ratio and $R1$ is the output load). The frequency-response curve of practical output transformers is similar to that of R-C coupled amplifiers.

Note. See figure 23 for sample problem.

f. Direct-coupled amplifiers, or d-c amplifiers, use coupling networks which consist of direct connections or resistors. They have a flat frequency response from 0 cycle up to relatively high frequencies. Direct-coupled amplifiers are inherently unstable. The phase shift of direct-coupled amplifiers is 180° per stage, and is independent of frequency up to relatively high frequencies. Multistage direct-coupled amplifiers require large batteries or relatively high-voltage power supplies which are free of ripple.

35. Review Questions

a. In an R-C coupled amplifier, what elements control the response at middle, low, and high frequencies?

b. What is the gain of a single-stage R-C amplifier at middle frequencies?

c. At the low-frequency half-power point, what is the magnitude of the reactance of the coupling capacitor?

d. At the high-frequency half-power point, what is the magnitude of the reactance of the total shunting capacitance?

e. Assume that a single-stage R-C amplifier has half-power points at 10 cycles and at 100,000 cycles. Between what frequencies is the amplifier response curve flat?

f. In the amplifier of Review Question 18, what is the exact midfrequency?

g. Why does the gain of an R-C amplifier decrease as the bandwidth is increased?

h. Why is the bandwidth of a cascade amplifier narrower than the bandwidth of any one stage?

i. Why are pentodes generally preferred over triodes in R-C amplifiers?

j. What is the advantage of impedance coupling as compared with resistance-capacitance coupling?

k. What elements and factors control the response of an impedance-coupled amplifier at low and at high frequencies?

l. What is the voltage gain of the impedance-coupled amplifier?

m. What are the advantages and disadvantages of double-impedance coupling? Tuned-impedance coupling?

n. What is the difference between a power transformer and a voltage transformer?

o. Name and briefly describe the use of three types of audio transformers.

p. Which of the three types of audio transformers usually are step-up transformers and which usually are step-down transformers?

q. What is an ideal transformer? Does it exist?

r. Assume that an ideal output transformer with a turns ratio of 20:1 has a 5-ohm load resistor connected across its secondary terminals. What is the impedance reflected across the primary terminals?

s. Each of the two primary terminals of a commercial output transformer is marked 5,000; each of the two secondary terminals is marked .5. What is the approximate turns ratio of the transformer?

t. What are the advantages and disadvantages of interstage transformers?

u. Why is it permissible to use a T network to represent a transformer?

v. What is the voltage gain of a transformer-coupled amplifier in the middle range of frequencies?

w. What elements control the response of the transformer-coupled amplifier at low and at high frequencies?

x. How is the voltage gain of a transformer-coupled amplifier improved at low frequencies?

y. Why are triodes preferable to pentodes in transformer coupling?

z. How does the response of a pentode transformer-coupled amplifier differ from a triode transformer-coupled amplifier at low, middle, and high frequencies?

aa. Why does the frequency-response curve of an output transformer differ from that of an input transformer?

ab. What is a d-c amplifier?

ac. What is the distinguishing characteristic of a d-c amplifier?

ad. Why is the response of a d-c amplifier flat from 0 cycle?

ae. What factors limit the high-frequency response of a d-c amplifier?

af. What are some uses of d-c amplifiers?

ag. Is a direct-coupled amplifier reliable as an amplifier of d-c? Why?

ah. What are the advantages and disadvantages of d-c amplifiers?

ai. What is the purpose of the Loftin-White d-c amplifier?

aj. Referring to figure 32, explain the operation of the Loftin-White amplifier.

ak. Why is a d-c amplifier best operated from batteries?

al. Describe two methods for measuring small values of d-c voltages.

am. Name and describe a d-c circuit used to compare two d-c voltages.

CHAPTER 3

WIDE-BAND AMPLIFIERS

36. Need for Wide-Band Pass

a. The amplifiers discussed in chapter 2 usually are known as narrow-band amplifiers because their gain is uniform for only a relatively small frequency range. For example, the range of uniform gain of some high-quality audio amplifiers extends from 100 cycles to 10,000 cycles. Of course, such amplifiers can pass signals lying in the frequency ranges below 100 cycles and considerably above 10,000 cycles; however, all signals outside the range of uniform, or flat, frequency response are subject to frequency and phase distortion caused by the reactive coupling networks and stray capacitance.

b. Uniform gain over a very wide range of frequencies is required in many electronic applications which must amplify nonsinusoidal signals. For example, the sawtooth voltages used in oscilloscope sweep circuits are considered nonsinusoidal waveforms which consist of a fundamental component and a large number of harmonics. All the components of the sawtooth signal must be amplified with uniform gain if the sawtooth signal applied to the amplifier input is to be reproduced faithfully at the output. Rectangular pulses also require broadband amplifiers for faithful amplification (fig. 35). Television and radar circuits use pulses which may last for periods between about 1 microsecond and $\frac{1}{10}$ second.

c. Experiment shows that amplifiers for such pulses must have a uniform gain from about $\frac{1}{10}$ of the lowest pulse-repetition frequency to about 10 times the highest pulse frequency if they are to be reproduced faithfully. The frequency of pulses occurring $\frac{1}{10}$ second apart is 10 cps; and that of pulses 1 microsecond apart is 1 megacycle. The flat range of the amplifier, therefore, extends from 1 cycle to 10 megacycles. An amplifier having such a wide frequency range is known as a *wide bandpass amplifier*. It might also be called a pulse amplifier.

37. Coupling Circuits for Wide-Band Amplification

Although transformer and impedance coupling may be used successfully in an audio am-

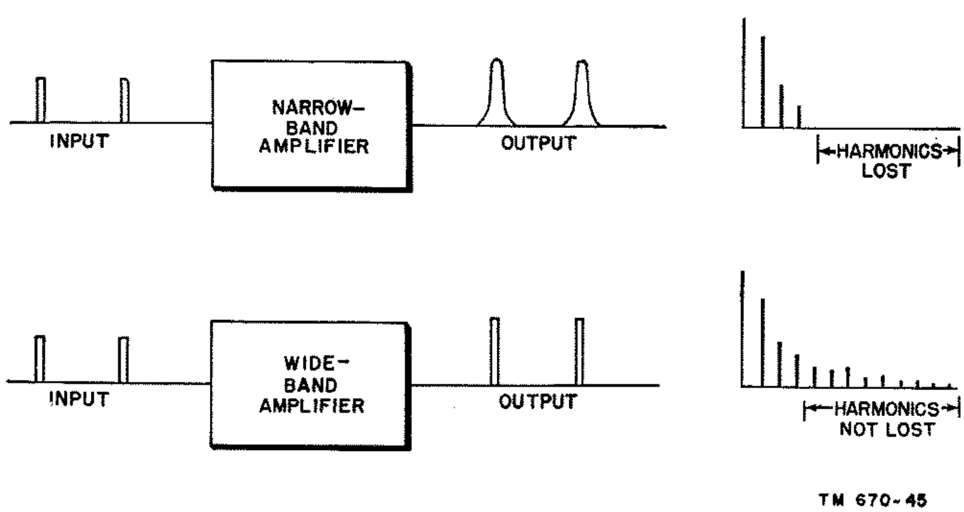

Figure 35. Waveforms in narrow- and wide-band amplifier.

plifier, a more uniform frequency response is obtained, even for the limited audio range, by the use of resistance-capacitance coupling. Wide-band amplifiers generally use resistance-capacitance coupling circuits which are modified to maintain a constant gain from much lower frequencies to considerably higher frequencies than are possible ordinarily when using conventional R-C coupling.

a. Figure 36 shows three basic types of coupling circuits used to compensate for the reduced gain at high frequencies: *shunt compensation, series compensation,* and *shunt-series* or *combination compensation.*

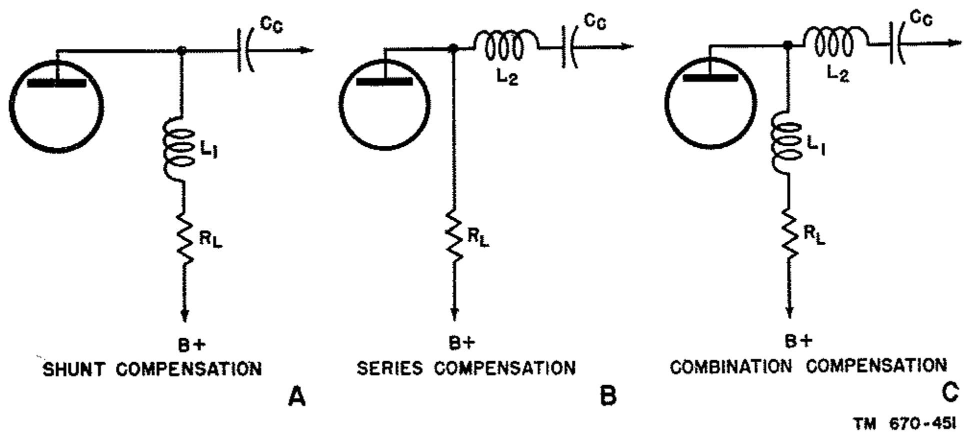

Figure 36. Coupling circuits.

b. Figure 41 illustrates a special filter-coupling circuit used to compensate for the reduced gain at low frequencies. Its operation will be described in a later paragraph on low-frequency compensation.

c. Wide-band frequency response may be obtained by using both *high-* and *low-frequency compensation* in the same circuit. These circuits are treated in detail in paragraphs 39 through 47.

38. Frequency Limitations of R-C Coupled Amplifiers

The resistance-capacitance coupled amplifier shown in figure 37 presents two limitations which must be minimized before the circuit can be employed for wide-band amplification. The high-frequency response is limited by the output capacitance, C_o, distributed capacitance, C_d, and input capacitance, C_i. These three capacitances appear in parallel and, therefore, may be represented by a single capacitance, C_S, which shunts load resistor R_L with a reactance which decreases as the frequency is increased. Thus, because of these shunt capacitances, the gain falls off as the frequency increases. The low-frequency response is limited, primarily, by the time constant, R_gC_c, which must be long compared with the period of the lowest frequency to be amplified.

39. Limit of Gain

a. Influence of High-Frequency Range.

(1) The gain of an R-C coupled amplifier at midfrequency is $A_M = -g_mR_{eq}$, as discussed in detail in chapter 2. (The minus sign in the expression for gain indicates the polarity reversal between the input and output signals.) The gain drops to 70.7 percent of its midfrequency value at the upper half-power frequency, f_H, where the reactance of C_S equals R_{eq}; that is, $X_{CS} = 1/2\pi f_H C_S = R_{eq}$. This simple equation supplies important information regarding the high-frequency limitations of an R-C coupled amplifier. The expression $1/2\pi f_H C_S$ is substituted in place of R_{eq} in the gain equation $A_M = -g_mR_{eq}$, so that it becomes $A_M = -g_m/2\pi f_H C_S$.

(2) This expression shows that for a specific amplifier the midfrequency gain, A_M, is dependent upon the variable

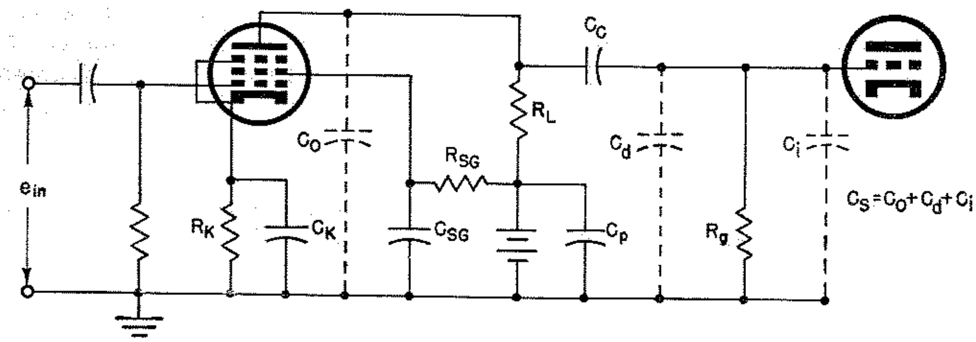

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Figure 37. R-C coupled amplifier.

term f_H, and upon the constant terms g_m, C_S, and 2π.

(3) As a numerical example of the maximum gain obtainable using a single-stage R-C coupled amplifier, assume the following constants: $g_m = 6{,}000$ umho (micromhos), $C_S = 15.9\ \mu\mu f$ (micromicrofarads). The amplifier is to have a high-frequency half-power point at 3,000,000 cycles. Substituting these values in the revised gain equation shows that maximum gain A_M is —20, since:

$$A_M = \frac{-g_m}{2\pi f_H C_S} = \frac{-(6{,}000 \times 10^{-6})}{(2\pi \times 3 \times 10^6 \times 15.9 \times 10^{-12})} = -20$$

If the half-power frequency, f_H, is to be twice as high, or 6,000,000 cycles, then the maximum possible gain of this amplifier drops from —20 to half of —20, or —10, since f_H appears in the denominator of the gain equation, $A_M = -g_m/2\pi f_H C_S$.

(4) Thus, it is seen that the required frequency range limits the maximum gain obtainable in a single stage of amplification. Increasing the high-frequency limit reduces the maximum gain obtainable per stage; on the other hand, reducing the high-frequency limit increases the maximum gain obtainable.

b. Figure of Merit. The revised gain equation, $A_M = -g_m/2\pi f_H C_S$, provides another important clue to the amplifying abilities of an R-C coupled amplifier at high frequencies. Constants g_m and C_S determine the gain of an amplifier if the half-power frequency, f_H, is fixed by the design requirements. For example, increasing g_m increases gain A_M. The importance of the constants is clarified by rewriting the gain expression in the form $(\frac{1}{2}\pi f_H)$ (g_m/C_S). This expression shows that the gain of a single-stage amplifier depends on the ratio g_m/C_S, known as the *figure of merit.* For example, if the half-power frequency, f_H, occurs at 5 mc (megacycles) in a certain amplifier, its gain is increased by using a tube with a higher figure of merit. Thus, it is important in wide-band amplifiers to use vacuum tubes which combine a high g_m with low values of interelectrode and stray capacitance, such as is found in pentodes. Triodes are not suited for use in wide-band amplifiers since they have large values of input capacitance. Therefore, the figures of merit for triodes are lower than for pentodes.

40. Low-Frequency Phase Distortion

Good low-frequency response is especially important for reproducing signals whose amplitudes remain constant for a given period of time. Examples of such signals are the flat tops of square waves used in amplifier testing, or radar reflections received from a long object such as a cloud.

a. Effects of Poor Low-Frequency Response.

(1) Even small variations of the low-frequency response in an R-C coupled amplifier produce noticeable phase distortion because the fundamental frequency and the low-frequency components of a complex wave usually have

the largest amplitudes. Phase shift often is negligible at higher frequencies because the higher-frequency components usually grow progressively smaller in amplitude as the harmonic number increases. For example, the third, fifth, and seventh harmonics of a perfect square wave are respectively only one-third, one-fifth, and one-seventh as strong as the fundamental.

(2) A variation in low-frequency gain of an R-C amplifier is much less noticeable than the corresponding variation in phase shift. For example, when the phase shift of an R-C amplifier varies by 2° from the phase at midfrequency, the low-frequency gain drops by only .6 of 1 percent to 99.4 percent of its midfrequency value. A phase shift of 2° is the maximum shift tolerated in many electronic applications.

b. Correction of Phase Distortion.

(1) One method of minimizing phase distortion is to use an amplifier which introduces no phase shift whatever at any frequency. However, an amplifier with such a characteristic is difficult to build.

(2) A more practical method of minimizing phase distortion consists in maintaining the delay time through the amplifier constant for all frequency components by permitting the phase shift to vary directly as the frequency. Figure 38 shows that this condition introduces no distortion. Two signals are applied simultaneously to an amplifier which introduces such phase shift. The amount of phase shift produced is purposely exaggerated in the illustration in order to clarify the result. Notice that a delay of 1/400 of a second corresponds to one quarter of 1 cycle of the 100-cycle signal, but it corresponds to two quarters, or ½, of 1 cycle of a 200-cycle signal. Hence, the delay time is the same for 100 cycles and for 200 cycles, because the phase shift is doubled when the frequency is doubled. Thus, the 100-cycle signal and the 200-cycle signal remain in the same relative phase relationship, and there is no phase distortion.

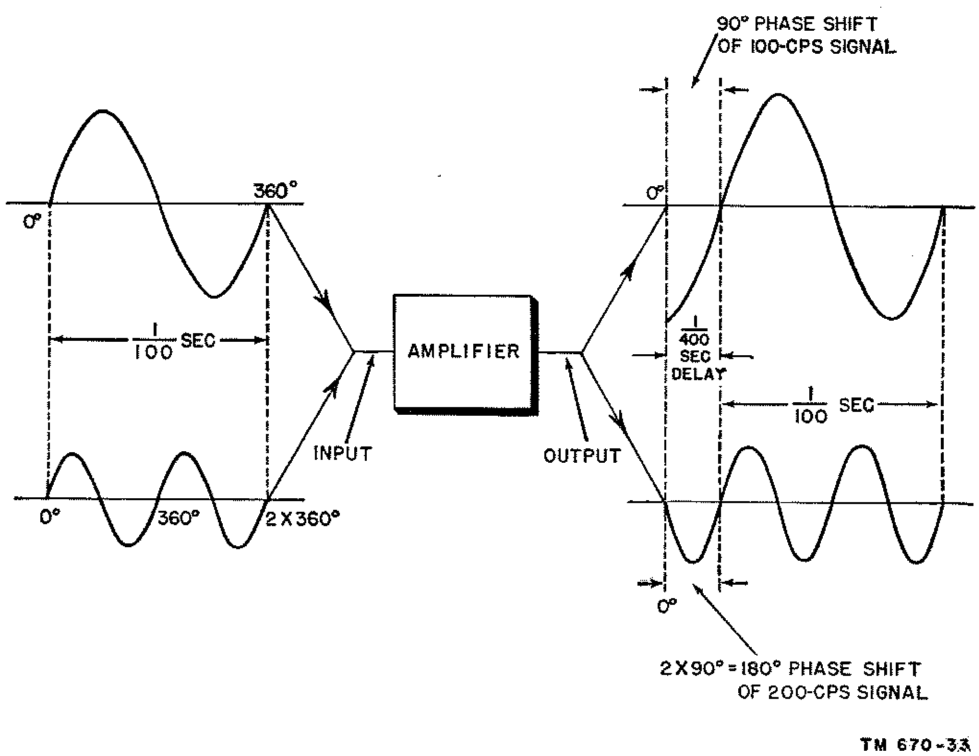

Figure 38. Constant delay time.

41. Shunt Compensation

a. Circuit Operation. The simplest coupling circuit used in compensating for the loss in gain at high frequencies is known as *shunt compensation* (A, fig. 39). In this circuit, a small inductor, L_1, is added in series with load resister R_L to compensate at high frequencies for the shunting effect of C_S, the total shunting capacitance. This compensating circuit is also known as *shunt peaking* because the shunt combination of C_S with L_1 plus R_L constitutes a parallel-resonant circuit. (Remember that at high frequencies C_c is practically a short circuit.) The resonant peak serves to maintain the gain practically uniform in the high-frequency range where the reactance of C_S tends to reduce the gain of the uncompensated amplifier. For proper operation, the Q of the circuit should equal about .5, where Q is the ratio of $2\ \pi f_H L_1$ to R_L. Using this value of Q, the impedance of the parallel-coupling circuit has a value almost equal to R_L, up to the frequency f_H, as shown in the figure. The reactance of C_S decreases as the frequency increases, but this effect is compensated almost exactly by the rise in reactance of L_1, which is connected in series with R_L. Thus, the gain of the compensated amplifier remains practically uniform up to frequency f_H, at which the gain of the uncompensated amplifier drops to 70.7 percent of midfrequency value. Response curves of compensated wide-band amplifiers are shown in figure 40. The amplifier frequency response is extended by a factor of 10, since its response without compensation is flat only to $1/10$ of f_H, or 300,000 cycles, whereas with compensation, it is flat to 3,000,000 cycles, sometimes with a slight rise near the high-frequency end.

b. Numerical Example of Shunt Peaking.

(1) As a numerical example, assume that the wide-band amplifier shown in A, figure 39, uses a tube with a g_m of 6,000 umho, that the circuit has a total shunt capacitance, C_S, 15.9 $\mu\mu$f, and that the upper half-power frequency is 3,000,000 cycles. The maximum gain therefore is —20, since $A_M = -g_m/(2\ \pi\ f_H C_S)$.

(2) The value of R_L which produces this gain is found from $A_M = -g_m R_{eq}$ by substituting —20 for A_M, and 6,000

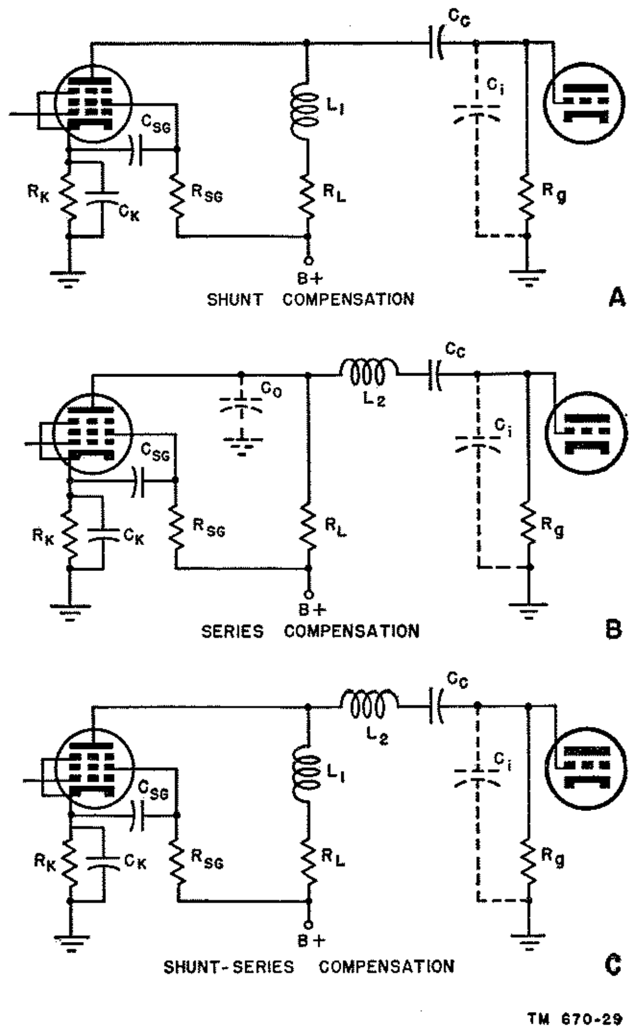

Figure 39. High-frequency compensation.

umho for g_m. Thus, R_{eq} equals 20/-(6,000 times 10^{-6}), or 3,333 ohms. As a check on the calculations, note that the reactance of C_S at the half-power frequency of 3,000,000 cycles equals R_{eq}, or 3,333 ohms. Load resistor, R_L, is usually very small compared with the r_p of a pentode and with R_g. Hence, R_L is made the same value as R_{eq}, or 3,333 ohms, since the shunting effects of r_p and R_g are negligible.

(3) The size of inductor L_1, required for flat response, now is determined easily. The reactance of L_1 is (.5) (R_L), since, for uniform response, $Q = X_L/R_L = .5$. Hence, $X_L = (Q)(R_L)$ or (.5) (3,333), or 1,667 ohms. The size of the inductor is found from the formula

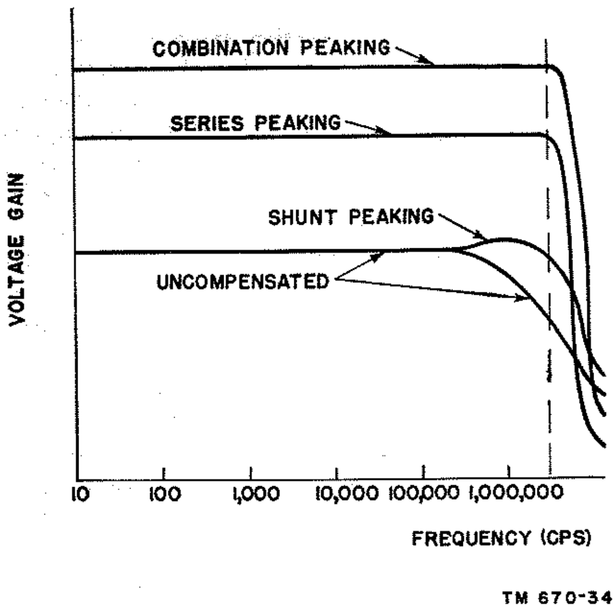

Figure 40. Response curves of wide-band amplifiers.

$X_L = 2\ \pi\ f_H L_1$, or $1667 = (2\ \pi)\ (3$ times $10^6)\ (L_1$, or $L_1 = 88.3$ microhenrys, as shown in the circuit of A.

42. Series Compensation

A wide-band amplifier using a compensating inductor, L_2, *in series* with coupling capacitor C_c is shown in B, figure 39. This method of coupling two stages of amplification is known as *series compensation* or *series peaking.* The gain of an amplifier using series peaking is about 50 percent greater than that using shunt peaking.

a. The small inductor, L_2, is chosen to resonate at a frequency *above* f_H with input capacitance C_i of the following stage, where C_i includes the stray wiring capacitance on the right of C_c. Thus, as the frequency increases and approaches resonance, the current through C_i increases, causing the voltage across C_i to increase with frequency. This resonant rise across C_i is sufficient to compensate for the decrease of voltage across R_L, caused by the shunting effect of C_o, the output capacitance of the tube, where C_o includes the stray wiring capacitance on the left of C_c. Thus, the high-frequency gain of the series-compensated amplifier remains constant up to frequency f_H, at which the gain of the uncompensated amplifier drops to 70.7 percent of its midfrequency value (fig. 40).

b. Experiment and calculation show that the circuit operates best when load resistor R_L is 1.5 times as large as the reactance of *total* shunting capacitance C_s, at f_H, and when the reactance of L_2 is about .7 R_L. The gain of the amplifier increases as the resistance of load resistor R_L increases, since $A_M = -g_m R_{eq}$. Consequently, as shown in the figure 40, the gain is 1.5 times as great as that with shunt compensation because R_L is 1.5 times larger in this case.

43. Shunt-Series Compensation

A wide-band amplifier using both series and shunt compensation is shown in C, figure 39. This coupling circuit is known as *combination compensation* or *combination peaking.* The gain of an amplifier using combination peaking is about 80 percent greater than one using shunt peaking (fig. 40). Combination peaking produces nearly uniform gain up to frequency f_H, at which the gain of the uncompensated amplifier drops to 70.7 percent of its midfreauency value. Experiment and calculation show that the circuit (C, fig. 39) operates satisfactorily when R_L is about 1.8 times as large as the reactance of the total shunting capacitance C_S at f_H, the reactance of L_1 is about .1 times R_L, and the reactance of L_2 is about .5 times R_L.

44. Comparison of Shunt, Series, and Combination Peaking

a. Shunt compensation for high-frequency peaking usually is preferred to the series and the combination methods, for the simple shunt circuit produces uniform response up to f_H and has the most nearly constant time delay (phase shift proportional to frequency). It is easy to adjust, and its operation primarily depends on the total capacitance which shunts the coupling circuit at high frequencies.

b. Series compensation produces a higher gain per stage than shunt peaking, but, for best operation, the ratio of C_i to C_o is critical and should preferably be about two to one. In B of figure 39, for example, if C_i is 10 $\mu\mu$f, C_o should be about 5 $\mu\mu$f. This requirement necessitates careful arrangement of parts so that the circuit capacitances plus the stray capacitances add up to the values required.

c. Combination peaking produces the greatest gain for the given frequency, f_H. However, the circuit is more difficult to adjust than either shunt peaking or series peaking. In particular, it is difficult to minimize the distributed capacitance of the winding of inductor L_2, and to adjust the ratio of C_o to C_i.

d. If the gain is held constant in a given stage, it follows that shunt peaking increases the bandwidth of an uncompensated R-C amplifier by a factor of about 15, and combination peaking increases the bandwidth by a factor of about 18. For example, if f_H of an uncompensated R-C amplifier is 3,000,000 cycles, then its response is flat up to 300,000 cycles. Series peaking extends the flat range response of this amplifier to 3,000,000 cycles by a factor of about 10. Shunt peaking extends the flat range of response to 4,500,000 cycles by a factor of about 15. Combination peaking extends the flat range response of this amplifier to 5,400,000 cycles by a factor of about 18.

45. Limitations at Low Frequencies

A large time constant, R_gC_c (figs. 37 and 39), usually is required in order to prevent the response at low frequencies from falling off, and to minimize phase distortion. There are, in addition, other reactances which come into play at the extremely low frequencies. For example, in figure 39, the reactance of C_K, the cathode bypass capacitor, increases as the frequency decreases. Thus, the impedance of the bias circuit, R_KC_K, which is connected in series between cathode and ground, is increased. The signal voltage applied between grid and cathode is, therefore, reduced at low frequencies because of this series impedance, and as a result the low-frequency gain is reduced. One remedy for this effect is a very large bypass capacitor, C_K. The plate-supply terminals and the screen-grid-supply resistor, similarly, must be bypassed with large capacitors to keep all series impedance as low as possible at the low frequencies to be amplified. It is physically impossible to provide capacitors which serve as bypasses for *all* frequencies, however. An additional corrective measure is thus required which can consist, for example, of the *low-frequency compensating circuit* discussed in the following paragraph.

46. Low-Frequency Compensation

a. Circuit Operation (fig. 41). The loss of gain at low frequencies is minimized by adding in series with load resistor R_L a two-element compensating filter consisting of capacitor C_F and resistor R_F. The purpose of the filter is twofold: (1) It increases the plate-load impedance of the amplifier at low frequencies, thus maintaining a more uniform gain, and (2) it compensates for the phase shift produced by R_KC_K and R_gC_c. Capacitor C_F bypasses the additional load resistor, R_F, for high-frequency signals. This makes the total load impedance, as seen by the plate circuit, equal to R_L. For very low-frequency signals, however, the capacitive reactance of C_F is practically an open circuit, thus bringing resistor R_F into play in series with R_L. Consequently, for very low-frequency signals, the load impedance in the plate circuit increases to a value of R_L and R_F. This condition maintains the gain and the phase response more uniform down to very low frequencies.

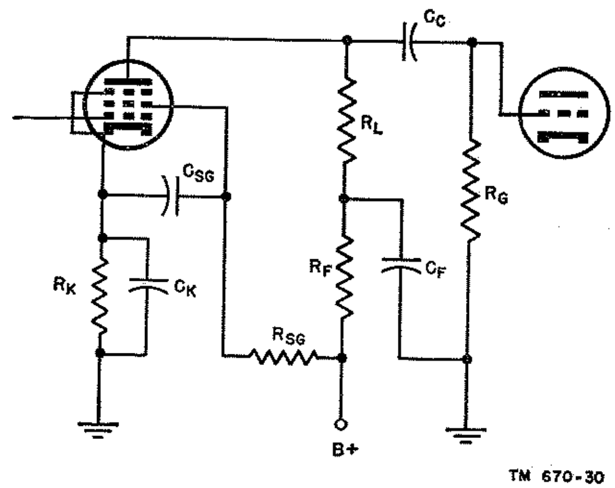

Figure 41. Low-frequency compensation.

b. Practical Values of Elements.

(1) Low-frequency compensation usually is adjusted first to eliminate the effects of the cathode-bias circuit impedance. This is done by making the time constant of the bias circuit equal to the time constant of the compensating filter, so that R_FC_F equals R_KC_K. *It is also necessary for proper compensation that R_F exceed R_K by a factor equal to the gain of the stage.* Capacitor C_K, therefore, must exceed C_F by the gain of the stage in order to main-

tain equal time constants; that is, $R_F C_F = R_K C_K$.

(2) As a numerical example, assume that the one-stage amplifier of figure 41 is to be compensated at low frequencies. Further, assume that it has a midfrequency gain of 20 and uses a cathode-bias resistor of 200 ohms. The value of R_F must, therefore, be 20 times 200 ohms, or 4,000 ohms. A practical value of bias capacitor C_K is 100 μf; hence, C_F must be 100 μf divided by 20, or 5 μf.

c. Square-Wave Testing.

(1) In a practical circuit, the effects of $R_g C_c$ are compensated by adjusting the value of resistor R_F for best over-all low-frequency response and most uniform time delay. This is done by observing the output waveshape of the amplifier on an oscilloscope while a low-frequency square wave signal is supplied to the amplifier input terminals.

(2) A perfect square wave is considered to consist of a fundamental combined with a large number of odd harmonics which are arranged in a very specific phase relationship in respect to the fundamental. The shape of the square wave becomes distorted even for small values of frequency distortion or phase distortion. Distortion of a square wave results in rounded corners, sloping tops and bottoms, and nonlinear horizontal and vertical portions (fig. 42). Thus, resistor R_F is adjusted so that the shape of the square wave in the amplifier output is a faithful reproduction of the square wave which is fed into the amplifier.

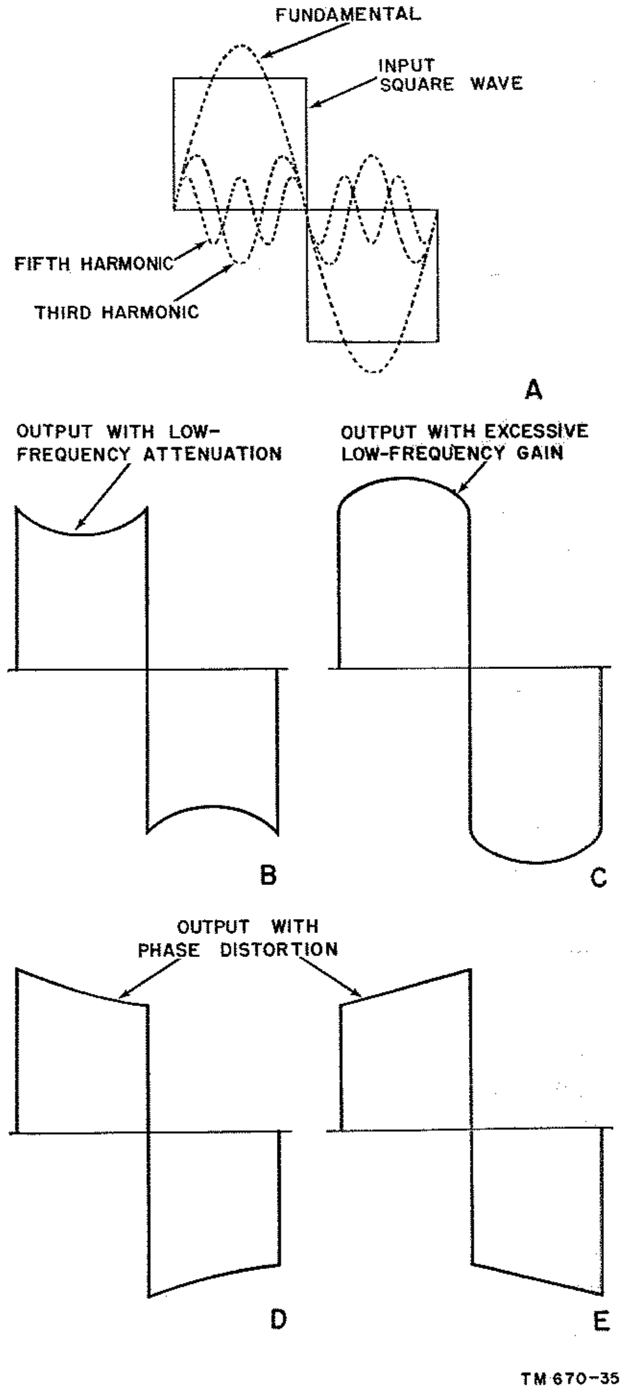

Figure 42. Distortion of square wave.

47. Simultaneous High- and Low-Frequency Compensation

Figure 43 shows a circuit for wide-band amplification which employs both high- and low-frequency compensation. The two compensating circuits operate independently and do not interfere with each other. For example, at very low frequencies, the series reactance of L_1 is negligibly small and, similarly, the shunt reactance of C_o (not shown) is so large that C_o can be assumed to be out of the circuit. Hence, the shunt-compensating circuit is ineffective in the low-frequency range. Similarly, at high frequencies, the reactances of bypass capacitors

C_K and C_F are negligibly small. Thus, the low-frequency compensation circuit is ineffective at high frequencies.

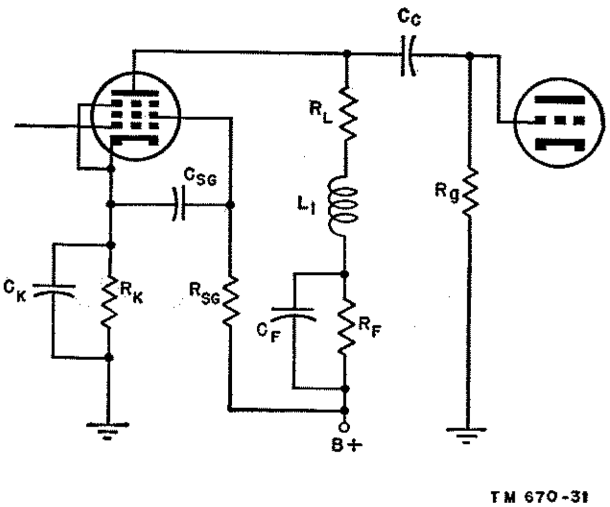

Figure 43. Wide-band amplifier compensated both at high and low frequencies.

48. Summary

a. Wide-band amplifiers are required in television, in radar, and in many other electronic applications in which it is necessary to amplify irregular waveshapes with the least possible distortion.

b. Wide-band amplifiers are R-C coupled amplifiers whose frequency range may be extended at both the low- and high-frequency ends by means of special compensating networks. The low-frequency limit is lowered by adding an R-C network in series with the load circuit. The high-frequency limit is increased by adding one or more inductors to the coupling elements.

c. Three circuits used for high-frequency compensation are shunt peaking, in which the shunting effect of C_S is compensated by connecting an inductor, L_1, in series with R_L; series peaking, in which an inductor, L_2, is connected in series between C_o and C_i, to compensate for their effects at high frequencies; and combination peaking, in which both circuits are employed in the same stage.

d. The gain is maintained nearly constant by all three high-frequency peaking circuits up to f_H. However, the voltage gain is lowest with shunt peaking. The gain with series peaking exceeds by 50 percent the gain obtained with shunt peaking. The gain obtained with combination peaking exceeds by 80 percent the gain obtained with shunt peaking.

e. An R-C filter circuit is used in series with R_L for compensating the loss in low-frequency response caused by the effects of R_gC_c and the impedances of the bypass capacitors of cathode screen and plate supply. Satisfactory compensation is obtained if $R_FC_F = R_KC_K$ when R_F exceeds R_K by a factor equal to the gain.

49. Review Questions

a. What is a wide-band amplifier?

b. Why are wide-band amplifiers needed?

c. What prevents the amplification of a wide range of frequencies in transformer coupling? In resistance coupling?

d. How does the gain of an R-C amplifier change when the high-frequency limit is increased? Why?

e. What is the figure of merit of an amplifier tube?

f. Why should the figure of merit be large if the bandwidth is to be large?

g. Are pentodes preferred over triodes in wide-band amplifiers? Why?

h. Why is a good low-frequency response essential in many circuits used in radar and in television?

i. Discuss two methods of minimizing low-frequency phase distortion.

j. With the aid of a sketch, explain why phase distortion is reduced to zero by maintaining the delay time constant.

k. Why are high-frequency compensating circuits also known as peaking circuits?

l. Explain the operation of shunt, series, and combination peaking.

m. Which type of peaking is preferable from the viewpoint of simplicity? Gain? Phase shift? Bandwidth?

n. Explain the operation of low-frequency compensation.

o. In what way, if any, does low-frequency compensation affect the gain of an amplifier which uses high-frequency compensation? Why?

p. Draw a sketch of an amplifier stage using a combination of high-frequency peaking and low-frequency compensation.

q. What is the effect of low-frequency distortion in an amplifier on the shape of a square wave which is fed to its input?

r. How is the low-frequency response of an amplifier checked and corrected?

CHAPTER 4

FEEDBACK AMPLIFIERS

50. Need for Feedback Amplifiers

a. The waveshape of the output voltage of an amplifier is usually not a perfect reproduction of the waveshape of the signal applied to the amplifier input. Consequently, the output of any amplifier invariably contains a certain amount of nonlinear distortion (amplitude distortion). As explained in paragraph 5, this distortion is caused by the vacuum tubes used in the amplifier circuits; the vacuum-tube plate current does not vary linearly with changes in grid voltage.

b. It is possible to minimize the distortion produced in a vacuum tube by applying to its grid a signal which has the same general distortion characteristics as the output, but which opposes it in polarity. The opposing signal is obtained by feeding to the amplifier input a portion of the amplifier output voltage (fig. 44). Such an amplifier, in which a portion of the output voltage is fed back to its input terminals, is known as a *feedback amplifier.*

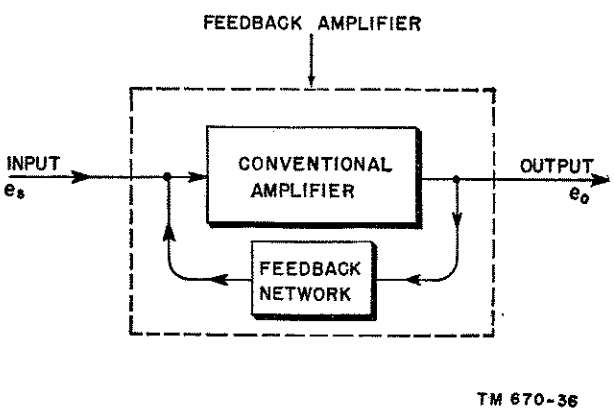

Figure 44. Feedback amplifier.

c. The drop in amplifier voltage gain occurring when the feedback voltage opposes the applied signal is compensated by other advantages of the feedback connection which are taken up later in this chapter.

d. The voltage gain of an amplifier can be increased by applying the feedback voltage in phase with the input signal voltage. This type of circuit has many useful applications, which are discussed in this chapter and in chapter 8.

51. Negative and Positive Feedback

a. Negative Feedback. When the feedback voltage decreases the gain of the amplifier, it is known as *negative feedback, inverse feedback, degenerative feedback,* or, briefly, *degeneration.* Many advantages result from its use. It reduces frequency distortion, harmonic distortion, and phase distortion, and results in an extended range of flat frequency response. The use of negative feedback provides greater operating stability by making the amplifier practically independent of variations in tube characteristics and of variations in supply voltages. Negative feedback is used also to alter the effective internal resistance r_p of a tube, increasing or decreasing r_p according to the way the feedback is applied.

b. Positive Feedback. When the feedback voltage increases the gain of the amplifier it is known as *positive feedback, regenerative feedback,* or, briefly, *regeneration.* Positive feedback is used in amplifier circuits primarily to accentuate the gain markedly for one particular frequency, or for a band of frequencies. Under certain conditions this positive feedback may be excessive and cause the amplifier to oscillate. This means that the amplifier produces an output signal at some one frequency even though no external input signal is applied. Such positive feedback applications are discussed in the chapter on oscillators.

c. Combination Feedback. Positive and negative feedback are used simultaneously in combination feedback by applying positive feedback to restore the gain lost from the application of negative feedback. Thus, the primary advantage of positive feedback is employed to overcome the chief disadvantage of negative feedback.

52. Voltage and Current Feedback

Circuits using negative or positive feedback usually are classified as using *voltage feedback* or *current feedback,* depending on how the feedback voltage is obtained.

a. The circuit in A of figure 45 consists of a conventional R-C coupled amplifier in which a portion of the output voltage, e_{fb}, is fed back in series with the input voltage. Voltage e_{fb} is known as the *feedback voltage* and the electrical path EFBD over which it is transferred to the input circuit is called the *feedback loop.* This feedback amplifier is said to employ voltage feedback because feedback voltage e_{fb} is proportional to the output voltage.

b. It is important to differentiate between voltage feedback and current feedback. The feedback used in figure 47 is known as current feedback, because feedback voltage e_{fb} is proportional to output current i_L, which flows through feedback resistor R_{fb}. Note that this circuit is simply that of an amplifier using cathode bias, but with the cathode-bypass capacitor omitted.

c. Voltage and current feedback differ also in the effects they produce and in the complexity of the circuits by which the feedback is obtained. An amplifier using negative-voltage feedback behaves like a constant-voltage generator; an amplifier using current feedback behaves like a constant-current generator. These and other effects are discussed later.

53. Voltage-Feedback Circuit

Conventional R-C amplifiers and feedback amplifiers are similar so far as their output terminals are concerned. The amplifiers differ with regard to their input circuits.

a. A conventional amplifier is readily converted to a feedback amplifier. In A, figure 45, for example, terminals *C* and *D* constitute the input terminals of a conventional R-C coupled amplifier. A voltage, e_g, is applied across terminals *C* and *D*, and is amplified by the tube. The amplified voltage, e_o, appears across resistor R_g. The conventional amplifier is converted to a feedback amplifier by applying the input signal, e_s, between grid terminals *A* and *B*, instead of connecting it in the conventional manner between grid and ground terminals *C* and *D*. To complete the circuit, the feedback voltage, e_{fb}, is applied to terminals *B* and *D*. It is thus in series with the signal voltage. The feedback voltage is a portion of output voltage, e_o, obtained from a tap on resistor R_g. Feedback is transferred to the input side of the amplifier via feedback-loop EFBD.

b. The operation of the feedback amplifier may be clarified by studying the waveshapes and the phase relationships of the input and output voltages in A, figure 45. For this purpose, it is assumed that the applied signal lies in the middle range of frequencies, at which the large capacitors, C_c and C_k, have negligibly low reactances and appear as short circuits. The small capacitor, C_s, has a relatively high reactance in this frequency range and appears as an open circuit. Thus, the a-c plate-load impedance of the tube consists of two parallel elements, resistors R_L and R_g. A voltage, e_g, applied to the tube between terminals *C* and *D*, is amplified and appears as the output voltage, e_o, across the parallel combination of R_L and R_g. The output voltage is of opposite polarity in respect to e_g for the reasons given in paragraph 6. The feedback voltage, e_{fb}, a portion of the output voltage, is fed back in series with the input voltage. The feedback voltage, therefore, is also of opposite polarity in respect to e_g. Consequently, point B becomes increasingly negative in respect to point D, or ground, as grid voltage e_g increases. This may be clarified by pointing out that, for the polarities shown, point *B* is negative in respect to ground-point *D*, and, conversely, that point *D* is positive in respect to point *B*. Similarly, point *C*, the grid, is positive in respect to both points *B* and *D*. Thus, if point *B* is taken as a reference point, the feedback voltage and grid voltage are added in phase. Their sum is identical with the external signal voltage, e_s, applied between points *A* and *B*; that is, e_s is equal to the sum of voltages, e_{fb} and e_g, since they appear across the same terminals.

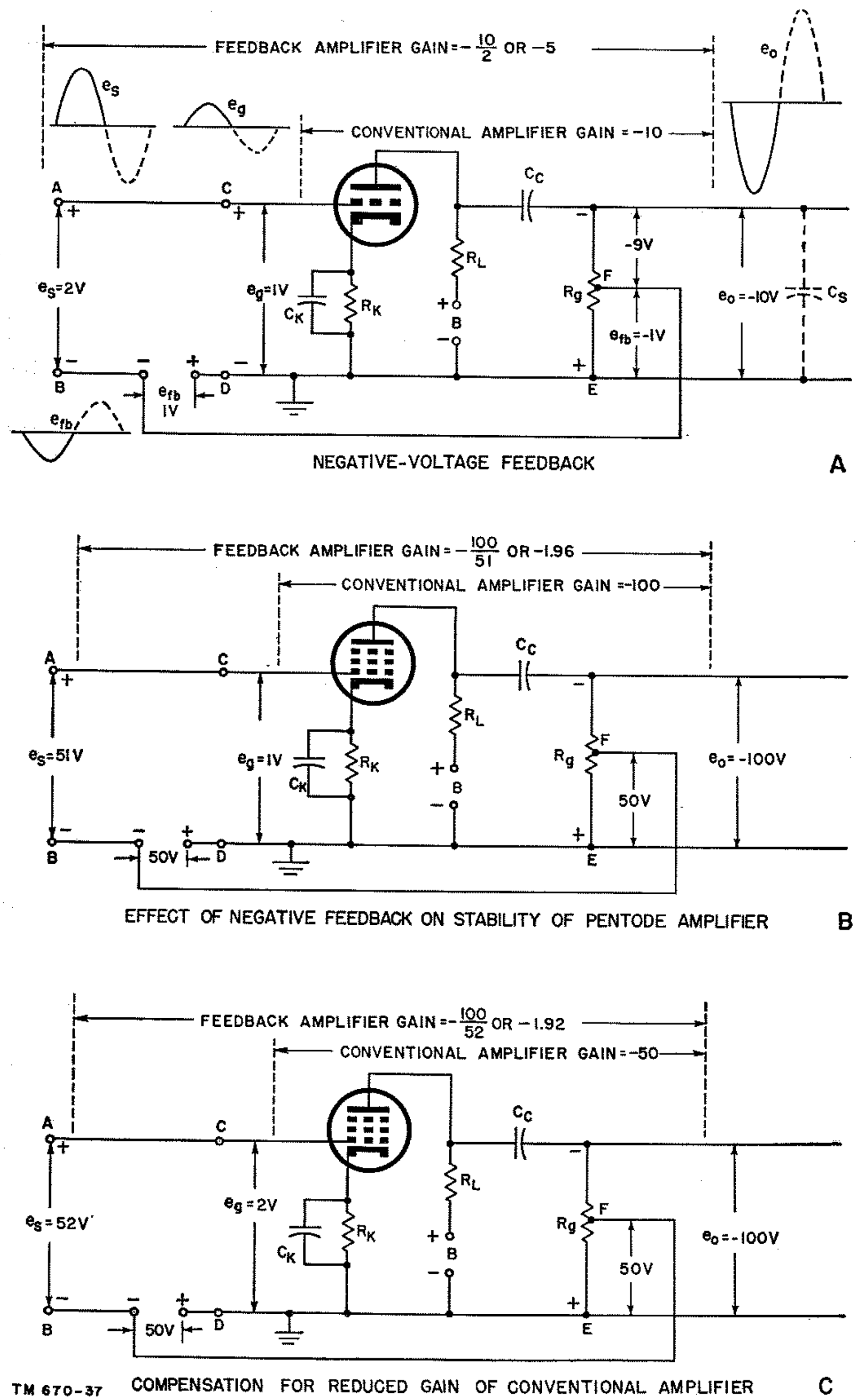

Figure 45. Negative-voltage feedback amplifier.

c. A numerical example will further clarify these relationships. Assume that the midfrequency gain of the conventional amplifier, in A, figure 45, is —10. (The minus sign before the amplified voltage signifies a polarity reversal in respect to grid voltage e_g.) A 1-volt rms signal, applied between terminals *C* and *D* of the conventional amplifier, appears as a —10-volt rms output signal e_o across R_g. Assume further that the 1-megohm resistor, R_g, is tapped at 10 percent, or 100,000 ohms, above ground. The tapped resistor acts as a voltage divider for e_o, and supplies 10 percent of e_o, or —1 volt, between the tap on R_g and ground. This voltage is e_{fb}, the feedback voltage. It has the same polarity as e_o, since both voltages are proportional to the current through resistor R_g. Like e_o, the feedback voltage, e_{fb}, is opposite in polarity to e_g. The feedback voltage of 1 volt is transferred via the feedback loop to terminals *B* and *D*. As explained in *b* above, and illustrated in A, figure 45, the signal voltage, e_s, is e_g plus e_{fb}, or 1 volt plus 1 volt, which equals 2 volts, with the indicated polarity. Since the polarity of the feedback voltage opposes the signal voltage as discussed above, the resulting grid voltage is e_s minus 1 volt, or 1 volt, as shown. The *effective* grid voltage, e_g, therefore, is less than the signal voltage, e_s, because of the opposing effect of the feedback voltage, e_{fb}.

d. In the example just given, the gain, *A*, of the conventional amplifier is —10. However, the gain *A′* (*A* prime) of the feedback amplifier is e_o divided by e_s, or —10 divided by 2, which equals —5. This is only half of the gain without feedback. It is possible to compute the gain with feedback by applying this simple formula:

$$\text{Gain with feedback } A' = A/(1 - \beta A).$$

In this formula, *A* is the voltage gain without feedback, and β represents the fraction e_{fb}/e_o. In the numerical example above, $A = -10$; and $\beta = e_{fb}/e_o = -1/-10 = .1$. These values of *A* and β are substituted in the formula for *A′*, the gain with feedback. Therefore,

$$A' = A/(1 - \beta A) = -10/(1 - (.1)\ (-10)) = -5.$$

This value of gain agrees with the values of e_o/e_s used in connection with A in figure 45. In using these formulas, it is essential to attach the minus sign to the numerical value of the gain, *A*.

54. Effects of Negative Feedback

Although the use of negative feedback results in a reduction in the voltage gain of an amplifier stage, this reduction in gain is justified by many advantages. Among the effects of negative feedback are: (1) extended frequency response, (2) reduced distortion, and (3) improved stability of operation. The last of these advantages may be interpreted as meaning that the output voltage remains nearly constant in spite of wide fluctuations in the supply voltages to the grid, plate, and filament of the tube, changes in tube characteristics caused by aging, differences in characteristics of individual tubes of the same type, and wide variations in the load current taken from the output of the stage. The manner in which these results are effected by the use of negative feedback is discussed in succeeding paragraphs.

55. Independence of Tube Characteristics

a. The gain of an amplifier using a large amount of feedback is almost completely independent of tube characteristics and of applied supply voltages. This fact is deduced from the formula for gain with feedback, $A' = A/(1 \text{ minus } \beta A)$. If the product, $-\beta A$, is much greater than 1, (1 minus βA) has practically the same value as the quantity $-\beta A$. In such a case, the formula for *A′* reduces to $A' = -1/\beta$, since $A/-\beta A = -1/\beta$. This formula shows that the amplifier gain depends only on the feedback network when $-\beta A$ is large. The feedback network in the figure is the tapped resistor, R_g; the location of the tap point determines β. For example, β is .5 when the tap point is 500,000 ohms, or halfway up on the 1-megohm resistor, R_g, for at this point, $e_{fb}/e_o = .5$. β is further increased by raising the tap to a position nearer to the top of resistor R_g; conversely, β is reduced by lowering the tap to a position on R_g closer to ground.

b. A numerical example will show that when a large amount of feedback is used, the amplifier practically depends on β alone, and is independent of the tube used. In B of figure 45, assume that β is .5 and that A is —100. Using the equation, $A' = e_o/e_s$, the gain with feedback is $A' =$ —100/51, or —1.96. In this case, it is permissible to use the approximate formula, $A' = -1/\beta$, because the quantity, $-\beta A$, or 50 (.5 times 100), is much greater than the quantity 1. Using the approximate formula to calculate the gain, $A' = -1/\beta$, or $-1/.5 \doteq -2$. (The error in gain is only 2 percent in using the approximate formula, $A' = -1/\beta$, since —2 is 2 percent larger than —1.96.) This example shows that the gain with large values of feedback depends primarily on the feedback network. The gain, therefore, is practically independent of tube characteristics and applied voltages, since the value of β, and thus of the gain, $A' = -1/\beta$, depends only on the position of a tap on a resistor.

56. Stability

a. B, figure 45, illustrates the effect of negative feedback on the stability of a pentode amplifier. Here, the gain of the conventional amplifier is 100, and 50 percent of its output voltage is fed back in series with a signal voltage. Consequently, β is .5. Thus, an input of 51 volts is required to operate the amplifier.

b. Assume now that the *B*-supply voltage suddenly drops to such a low value that the voltage gain of the *conventional amplifier,* whose input terminals are *C* and *D*, is halved, going from 100 to 50. A grid voltage of 1 volt, then, produces an output of 25 volts, or 50 percent less than before. One way of compensating for the reduced gain would be by doubling the grid voltage, e_g, increasing it from 1 volt to 2 volts, as shown in C. To accomplish this, however, the input voltage to terminals *A* and *B* of the *negative-feedback amplifier* need be raised only 1 volt, from 51 volts to 52 volts, since the signal voltage, e_s, is the sum of e_{fb} plus e_g. This would restore the output voltage and the feedback voltage to their original values.

c. The example shows that in this negative feedback amplifier a very small change in signal voltage of about 2 percent compensates for a 50-percent change in gain. If this small increase in e_s were not made, it would mean only that all other voltages would be reduced to 51/52 of the values shown in C. Thus the gain would be 51/52 times 1.96, or 1.92—a drop of only 2 percent in spite of a 50-percent drop in the gain of the conventional amplifier.

d. The penalty paid for obtaining this stability is a considerable loss in gain. In B, figure 45, the gain with feedback is only —100/51, or —1.96; that is, the gain is reduced by a factor of 51, since (1 minus βA) = (1 minus (.5) (—100)), or 51.

57. Independence of Load Variations

Another useful characteristic of the negative-voltage feedback amplifier is its relatively constant output voltage, a voltage almost independent of wide variations in load impedance.

a. Assume that the amplifier shown in A, figure 45, is operating in the high-frequency range where capacitor C_s has an appreciable shunting effect across R_g. In this high-frequency range, the output voltage of the conventional amplifier ordinarily drops; but the output voltage of the feedback amplifier remains almost unchanged. For example, grid voltage e_g rises the instant that output voltage e_o drops, because a drop in output voltage results in less feedback voltage. Thus, a drop in output voltage is counteracted instantly by a rise in grid voltage, which in turn restores the output voltage to practically its full original value.

b. The output-voltage stabilization becomes less pronounced, however, as the load impedance drops to very small values, since the compensating effect of the feedback voltage grows progressively smaller as e_{fb} becomes small in comparison to e_g. The output voltage is stabilized in a similar manner when an increase in load impedance tends to cause the output voltage to rise.

c. (1) In a broad sense, the negative-voltage feedback connection tends to maintain the output voltage constant in a manner which resembles the operation of an ideal *constant-voltage generator.* An ideal constant-voltage generator is a theoretical one which maintains its output voltage absolutely constant under all load-impedance variations. It has an internal impedance of 0 ohms in which there is no voltage drop. Of course, such an ideal generator does not exist; but its ideal characteristics are approached by a low-impedance generator. The internal-plate resistance, r_p, of an amplifier using negative-voltage feedback does not drop to 0, but it is known that r_p is reduced to a low value by the factor (1 plus $\beta\mu$).

(2) For example: A certain pentode amplifier uses a negative-voltage feedback circuit with a feedback factor, β, of .5. The pentode-tube plate resistance is 100,000 ohms and its μ is 500. What is the effective apparent plate resistance of the tube? This problem is solved by use of the factor, (1 plus $\beta\mu$). The apparent plate resistance r'_p is r_p/(1 plus $\beta\mu$), or 100,000/(1 plus (.5) (500)), which equals 398 ohms, a low value indeed. A low, effective, internal-tube impedance frequently is desired in electronic applications.

58. Frequency Response

The bandwidth of an amplifier is increased by using negative-voltage feedback, since negative-voltage feedback tends to maintain the output voltage constant.

a. In the curves of figure 46, the maximum gain of the amplifier without feedback (curve A, β equals 0) is —100; its response is flat from 1,000 cycles to 10,000 cycles; hence, the half-power points occur at 100 cycles and at 100,000 cycles. In a single-stage amplifier with feedback, however, the low-frequency limit is reduced by the factor 1/(1 minus βA), and the high-frequency limit is raised by the factor (1 minus βA). In curve B, β is .01; therefore, (1 minus βA) is (1 minus (.01)(—100)), or 2. Therefore, the lower half-power frequency, f_L, moves from 100 cycles to 50 cycles, since f_L/(1 minus βA) is 100/2, or 50 cycles. Similarly, in curve C, f_L moves from 100 cycles down to 10 cycles as a result of negative feedback and a β of .09. The upper half-power frequency, f_H, in B, is raised from 100,000 cycles to 200,000 cycles, and in the upper half-power frequency, f_H, is raised from 100,000 cycles to 1,000,000 cycles. The curves show clearly that the range of flat response is extended considerably by feedback, but at the expense of gain. However, it is usually possible to increase the input signal by the factor (1 minus βA), to make up for the loss in gain.

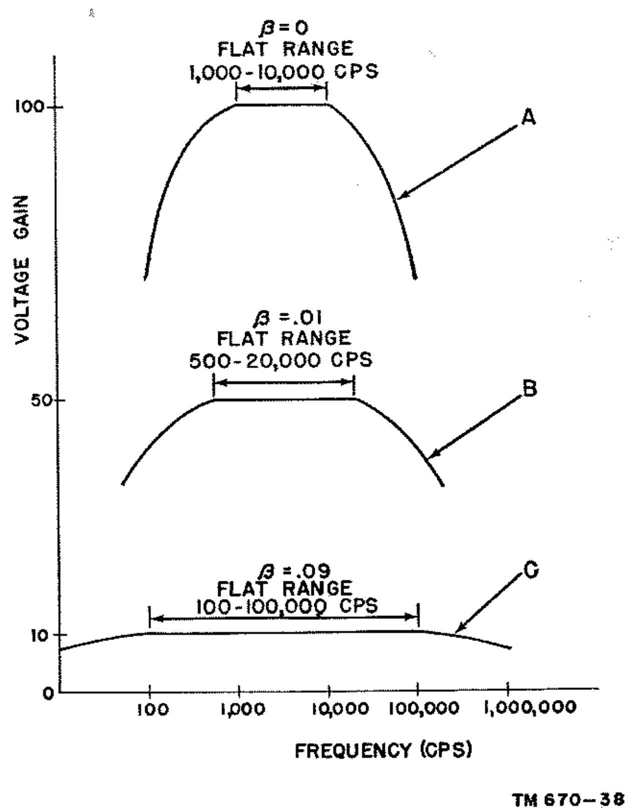

Figure 46. Response curves of R-C amplifier, with and without feedback.

b. For example, assume that the output voltage obtained with the amplifier having the response shown in B, figure 46, is to be the same as the output voltage obtained using the amplifier whose response curve is shown in A, where an input voltage of 1 volt produces 100 volts output. Using the amplifier in B, the input voltage must be raised, by a factor of two, to 2 volts, in order to obtain an output of 100 volts. The larger input signal is obtained either by adding another stage of amplification or by increasing the gain of a preceding stage.

59. Distortion and Phase Shift

Negative feedback reduces the distortion generated by a vacuum tube by the factor (1 minus

βA). Assume, for example, that a certain tube produces 5 percent harmonic distortion when operating in an amplifier without feedback. This distortion drops to one-half of 1 percent with use of the feedback amplifier ($\beta = .09$), the response curve of which is shown in C, figure 46. This is readily checked, since 5 percent divided by the feedback reduction factor 10 ((1 minus βA) = 1 minus (.09) (—100 = 10) equals ½ of 1 percent. The negative-feedback connection produces, also, a considerable reduction in phase shift, since gain and phase shift are closely related.

60. Current Feedback

Figure 47 shows two circuits which use a feedback connection known as *current feedback,* wherein the feedback voltage, e_{fb}, is proportional to the output current which flows in the load circuit. In A, the feedback voltage appears across a resistor placed in the coupling circuit to the next stage; in B, the feedback voltage is merely the a-c drop across an unbypassed cathode resistor.

a. The effects of negative-current feedback and negative-voltage feedback are the same with regard to reduced distortion, greater stability, and improved frequency and phase response. However, negative-current feedback *increases* the effective internal plate resistance, r_p, of the tube by the term ((1 plus $\beta\mu$) (R_{fb})). It is this increase of internal resistance that is responsible for the important characteristic of negative-current feedback—namely, that the

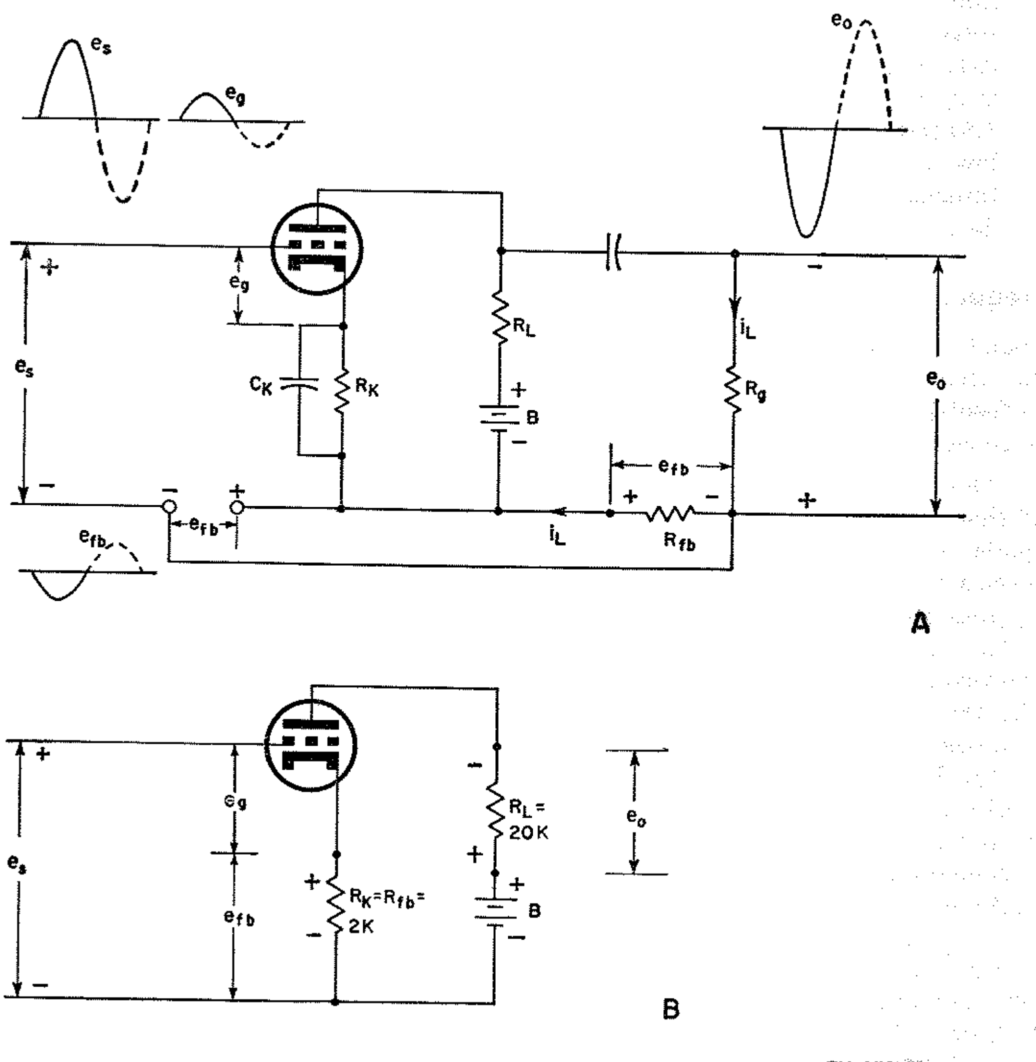

Figure 47. Negative-current feedback amplifiers.

load-circuit current remains relatively constant, practically independent of wide variations in load impedance.

b. In A, figure 47, the output current is held constant by using negative-*current* feedback, derived in a manner similar to that in B, figure 45, where negative-*voltage* feedback is used to stabilize the output voltage. In a broad sense, the negative-current feedback connection tends to maintain the output current constant in a way that resembles the operation of a *constant-current generator.* An ideal constant-current generator is a theoretical one which maintains its output current absolutely constant under all load-impedance variations. It has an internal impedance which is extremely large compared with the load impedance, so that the current in the load remains constant irrespective of the value of the load impedance.

c. In practical negative-current feedback circuits, r_p rises only moderately by the additional term ((1 plus $\beta\mu$) (R_{fb})) (par. 60*a*). For example, in B, figure 47, e_{fb} appears across the cathode resistor, R_K. Thus, r_p is increased because the cathode resistor is not bypassed. For the element values shown, β equals R_K/R_L, or .10. If the tube has a plate resistance of 10,000 ohms and a μ of 20, its effective internal-plate resistance, using negative-current feedback, rises to 16,000 ohms, since r_p plus ((1 plus $\beta\mu$) (R_K)) equals 10,000 plus (1 plus 2) (2,000), or 16,000 ohms. A high internal plate resistance may be undesirable in some applications. However, this is of negligible consequence where the many advantages of negative feedback can be realized by omitting the cathode bypass capacitor.

61. Feedback in Multistage Amplifiers

a. Cause of Oscillation. The advantages of feedback can be utilized in multistage amplifiers, but it is sometimes difficult to prevent undesired oscillation resulting from positive, or regenerative, feedback. For example, positive feedback results in the two-stage R-C amplifier of figure 48 when operating at midfrequency, if the feedback voltage is introduced in series with the input-signal voltage. This follows from the fact that the signal polarity of the amplifier is reversed twice; the feedback voltage is of the same polarity as the input-signal voltage, and the gain of the amplifier with feedback is greater than without feedback. Under these conditions, the amplifier may oscillate.

b. Prevention of Oscillation. Oscillation in a feedback amplifier is avoided by use of the circuit shown in figure 49. Here, the feedback obtained is negative, even though the polarity of the output voltage is the same as the polarity of the input voltage. Note that the feedback voltage, e_{fb}, is fed in series with the bias circuit of the first amplifier stage. An increase in amplifier output voltage, e_o, increases the voltage drop, e_{fb}, between cathode and ground, and so reduces the voltage, e_g, applied to the grid of tube V1. The feedback is negative since it reduces the gain of the amplifier. To clarify the manner in which e_{fb} is obtained, the dotted portion of grid resistor R_{g2} is shown as though it were moved to the cathode circuit of the first tube, where it appears as R_{fb}. In practice, the identical circuit is obtained by connecting the low end of R_K to a point on R_{g2} instead of to ground. Thus, the part of R_{g2} between the feedback connection and ground is R_{fb}. The feedback resistor, R_{fb}, produces a small amount of cur-

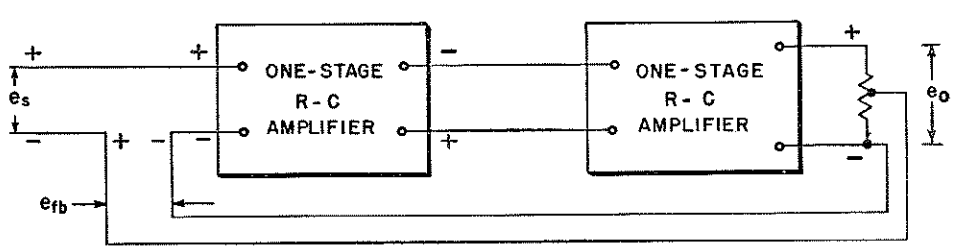

Figure 48. Feedback amplifier which may oscillate.

rent feedback also, but its effect is negligible because the resistance of R_{fb} is usually very small. It is difficult to design stable multistage amplifiers having negative-voltage feedback at all frequencies, because positive feedback may result at very low and at very high frequencies, where the phase shifts differ considerably from the phase shift in the middle range of frequencies.

feedback is opposite in respect to the polarity of the voltage, e_{fb}, caused by the negative-voltage feedback.

b. For some applications, it is desirable to accentuate the low frequencies alone. This is done by connecting a capacitor, C_1, across the feedback line. The large positive-feedback voltage which appears across C_1 at low frequencies increases the amplifier gain. At high frequencies, however, the small reactance of the capacitor serves to reduce the positive feedback to practically zero, and the high-frequency gain is affected only by the negative-voltage feedback. The high-frequency response alone may be increased in a similar manner by connecting a capacitor in series with the positive-feedback loop.

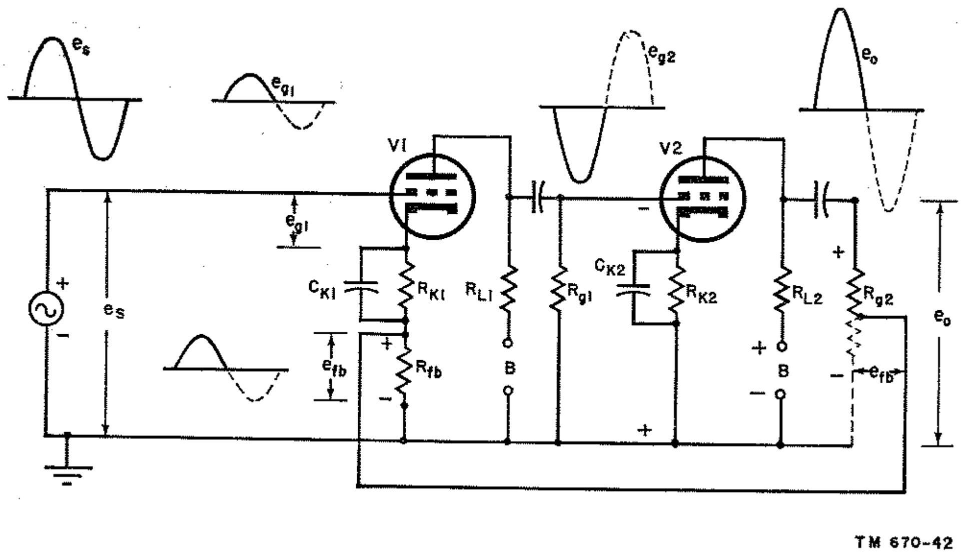

Figure 49. Two-stage feedback amplifier.

62. Combination Feedback

a. Figure 50 shows how positive feedback is used to restore the gain which is lost in introducing negative feedback in an amplifier. For the instantaneous polarities shown, the voltage feedback, obtained from the output transformer, is negative, and is fed over two stages in series with the input. The feedback obtained by connecting the two cathodes is positive because the amplifier-input grid voltage, e_g, and the over-all amplifier gain increase as the current feedback is applied. This is shown in figure 50, where for the polarities indicated the current feedback makes the cathode more negative in respect to ground. In other words, the grid becomes more positive in respect to the cathode. Thus, the gain increases as a result of the positive-current feedback. This increase in gain compensates for the gain lost because of the negative-voltage feedback. The polarities show that this is so, since the polarity of the voltage, e_{fb}, resulting from positive-current

63. Summary

a. In a feedback amplifier, a fraction of the output voltage is fed back in series with the input-signal voltage in order to improve the operating characteristics of the amplifier. The feedback voltage, e_{fb}, equals β times the output voltage, where β is a fraction less than 1.

b. Negative feedback reduces the gain of an amplifier. Positive feedback increases the gain of an amplifier. Combination feedback uses both negative and positive feedback in order to combine their advantages.

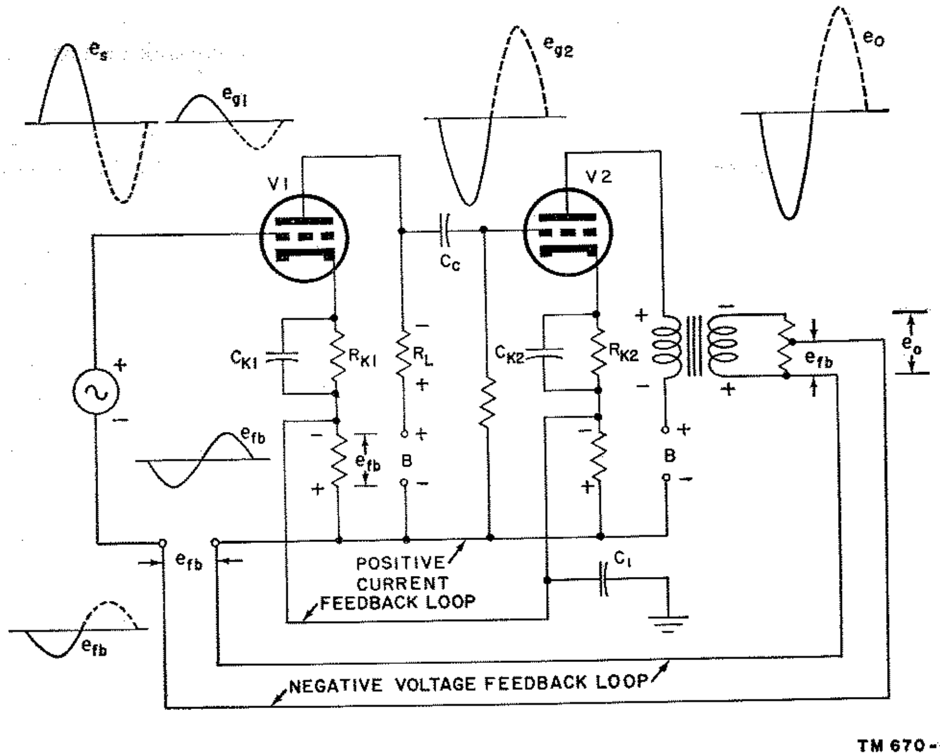

Figure 50. Combination feedback amplifier.

c. Negative feedback, applied to an amplifier, reduces harmonic and phase distortion, improves circuit stability, extends the flat range of frequency response, and alters the effective internal-plate resistance of a tube.

d. Negative feedback reduces the gain of an amplifier by the factor (1 minus βA). In voltage feedback, the feedback voltage is proportional to the amplifier-output voltage. In current feedback, the feedback voltage is proportional to amplifier-output current.

e. The internal-plate resistance, r_p, of a tube decreases by the factor (1 plus $\beta\mu$), using negative-voltage feedback; r_p increases by the term ((1 plus $\beta\mu$) (R_{fb})), using negative current feedback.

f. With negative-voltage feedback, an amplifier acquires the characteristics of a constant-voltage generator. With a negative-current feedback, an amplifier acquires the characteristics of a constant-current generator.

g. Negative feedback, applied to a single-stage amplifier, lowers the low-frequency limit by the factor 1/(1 minus βA), and raises the high-frequency limit by the factor (1 minus βA).

h. In multistage amplifiers, oscillations may result because the voltage fed back is of the same polarity as the input-signal voltage. Oscillation is avoided by introducing the fed-back energy in such a manner that it reduces the gain of an amplifier at all frequencies. Positive feedback sometimes is used in combination with negative feedback in order to restore the amplifier gain to the value obtained without feedback.

64. Review Questions

a. What is a feedback amplifier?

b. What is the difference between positive and negative feedback?

c. Compare voltage feedback with current feedback.

d. What are the advantages and disadvantages of negative feedback?

e. What advantages are gained by using positive feedback?

f. Sketch a simple one-stage R-C amplifier using negative-voltage feedback.

g. Sketch a simple one-stage transformer-coupled amplifier using negative voltage feedback.

h. Sketch a simple one-stage R-C amplifier using negative-current feedback.

i. Sketch a simple one-stage transformer-coupled amplifier using negative-current feedback.

j. Calculate β in an amplifier where the output voltage is —20 volts and the feedback voltage is —2 volts.

k. Calculate the gain of an amplifier using negative feedback if β is .1, and gain A, without feedback, is —10. Repeat for β of .01 and A of —100.

l. Using the results of the preceding review question by what factor is the gain reduced in each case? Explain.

m. A single-stage amplifier has a gain of —100 without feedback, and its response is flat from 100 cycles to 10,000 cycles. Negative feedback is applied to increase the range of flat response. Find the new limits of the flat response, if β is .01.

n. Using a simple sketch, show the effect of negative-voltage feedback on the stability of an amplifier.

o. In what manner does an amplifier using negative feedback behave like a constant-voltage generator?

p. In what manner does an amplifier using current feedback behave like a constant-current generator?

q. The output voltage of a certain amplifier has 4 percent harmonic distortion. Find the distortion if feedback is applied to this amplifier, with β of .01, and if the gain of the amplifier without feedback is —100.

r. Why does negative feedback reduce phase distortion?

s. In a voltage-feedback circuit, what happens to feedback voltage e_{fb} if the load is short-circuited?

t. In a current-feedback circuit, what happens to feedback voltage e_{fb} if the load is short-circuited?

u. Sketch a simple two-stage R-C amplifier circuit using negative-voltage feedback.

v. Sketch a simple two-stage amplifier circuit using combination negative feedback and positive feedback.

CHAPTER 5

CATHODE FOLLOWERS

65. Application of Cathode Followers

a. In radar and other electronic applications it is frequently necessary to transmit sharp pulses, or similar broad-band signals, to widely separated apparatus. For example, a pulse generator and its associated pulse amplifier may be mounted in a heavy rack which stands in a shelter on the ground, but its pulses may be required by auxiliary apparatus located a distance away, on a high tower. This creates the problem of how to transmit the sharp pulses over a cable or a transmission line without altering the pulse waveshape. The problem entails difficulties, because sharp pulses contain large numbers of high-frequency harmonics; the cable, therefore, must have a uniform response over a wide band of frequencies. It would not be practical simply to connect a transmission line directly to the pulse amplifier output. The pulse waveshape would be altered, because at high frequencies the capacitance of the line is a relatively low-impedance shunt load across the high-impedance output of the pulse amplifier. This would attenuate the higher harmonic components of the pulse, and change its shape.

b. One possible remedy would be to use a step-down transformer between the output of the pulse amplifier and the transmission line; the low-impedance line would have a negligible shunting effect on the stepped-down output impedance of the pulse amplifier. However, it is difficult, and uneconomical, to build a wide-band step-down transformer.

c. A practical solution to the problem is obtained by using a special amplifier known as a *cathode follower,* which in this application takes the place of a wide-band transformer. The cathode follower is connected between the high-impedance output of the pulse amplifier and the low-impedance line (fig. 51). Like a transformer, it has a very high input impedance and a very low output impedance. However, it is superior to a transformer for several reasons: (1) The output power of the cathode follower greatly exceeds the input power. (2) Its frequency response is flat from practically 0 cycle up to high frequencies in the megacycle range. (3) The circuit (A, fig. 55) is extremely simple and its cost is low. The operation of a cathode follower will be understood readily by anyone familiar with the operation of conventional amplifiers. These similarities are shown in the paragraphs which follow.

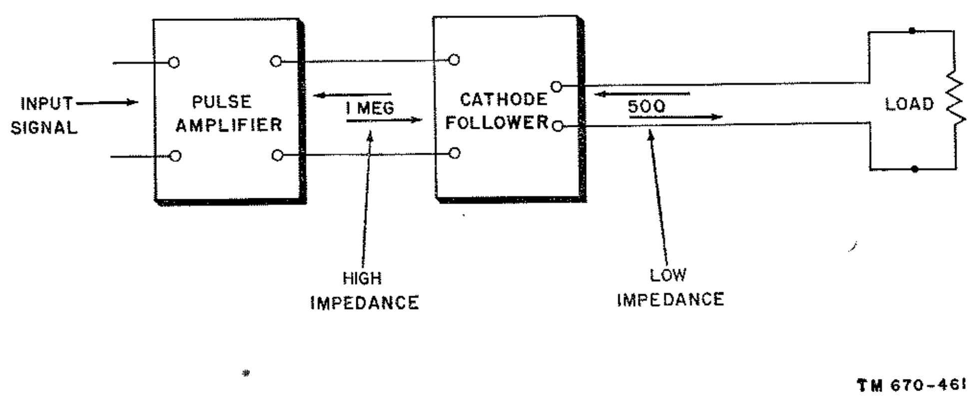

Figure 51. Cathode follower as matching device.

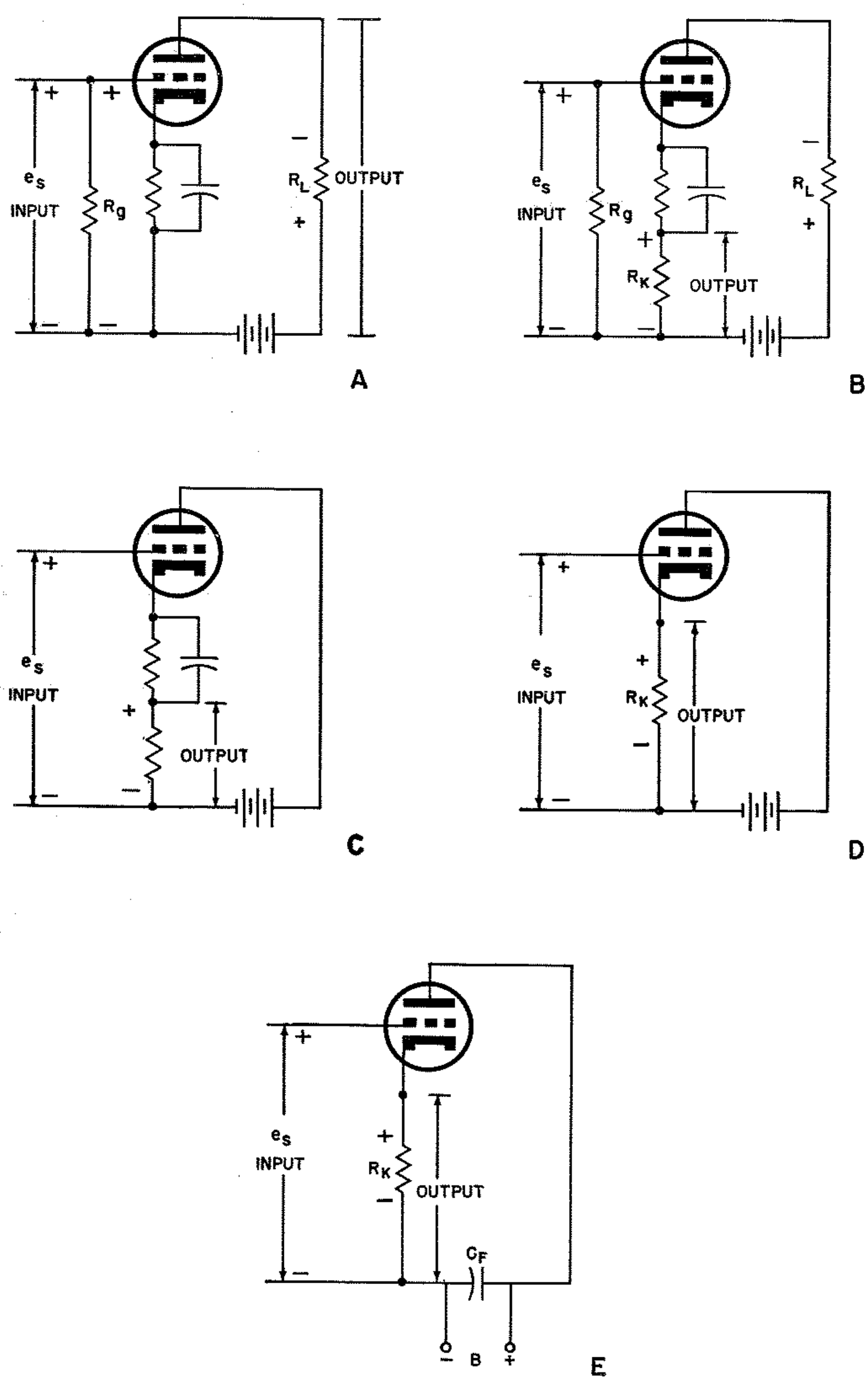

Figure 52. Evolution of cathode follower.

66. Development of Cathode-Follower Circuit

a. Figure 52 shows the close similarity between a conventional amplifier and a cathode follower.

(1) The conventional triode amplifier shown in A uses a cathode bias resistor and bypass capacitor. The d-c plate current which flows through the cathode resistor produces the required d-c grid bias. The a-c component of plate current is bypassed around the resistor by the capacitor. When an a-c signal is applied to the grid of the tube, the a-c plate-current variations flowing through the load resistor produce an a-c output voltage across it, as discussed in chapter 1.

(2) An alternative location from which an a-c output voltage could be obtained is shown in B. Here a resistor, R_K, is connected in the plate-current path in series between the cathode-bias circuit and the negative terminal of the d-c power supply. Thus, the a-c plate-current variations in the tube flow through resistors R_K and R_L and produce an a-c output voltage across each. Note, however, the opposite polarities of the instantaneous a-c voltages across the two resistors, as referred to the negative end of the plate-supply battery. The polarity of the a-c output voltage across the cathode resistor is the same as that of the grid-input voltage, whereas the a-c output voltage across the plate-load resistor is of opposite polarity to the signal voltage. This is desirable in some electronic applications such as phase splitters, where two opposite-going output voltages are required from a single input voltage.

(3) In some applications only one output voltage, having the same polarity as the input voltage, is required. In such cases the plate-load resistor can be omitted, as in C, because it has no useful function. This circuit is known as a *cathode follower,* because the a-c output voltage *follows* the polarity of the input signal voltage.

(4) The circuit is simplified further, in D, by omitting the cathode resistor and bypass capacitor, thus obtaining the required d-c grid bias from the d-c drop in the load resistor.

(5) In a further simplification, in E, the plate-supply battery is replaced by a power supply, *B*, bypassed by a large filter capacitor, C_F. Since this capacitor is practically an a-c short circuit across the power supply, substantially the entire a-c voltage developed in the plate circuit appears as a useful output across the cathode resistor.

b. The shift of the load resistor from plate circuit to cathode converts the conventional amplifier into one using negative feedback, shown in E, since the output voltage opposes the signal voltage (par. 73).

67. Characteristics of Cathode Follower

Although the cathode follower is similar to the conventional amplifier in its circuit and operation, it possesses certain distinguishing characteristics, stated briefly here and discussed in succeeding paragraphs.

a. Since the cathode follower is a negative-feedback amplifier, it has all the advantages of the negative-feedback voltage amplifier. These include an excellent frequency response, low values of harmonic (frequency) and phase distortion, high circuit stability, and low effective internal plate resistance. (Effective internal plate resistance is that value of generator resistance which is seen, or measured, when looking back from the load into the generator terminals.) In addition, the cathode follower circuit is extremely simple (fig. 53) for it consists merely of a vacuum tube, two resistors (grid resistor R_g and cathode load resistor R_K), and a bypass capacitor C_F.

b. The output voltage, e_o, of the cathode follower is always less than the input signal voltage, e_s; that is, the voltage gain of the cathode-follower circuit is always less than 1. The cathode follower has considerable power gain, however, since the power developed across the small

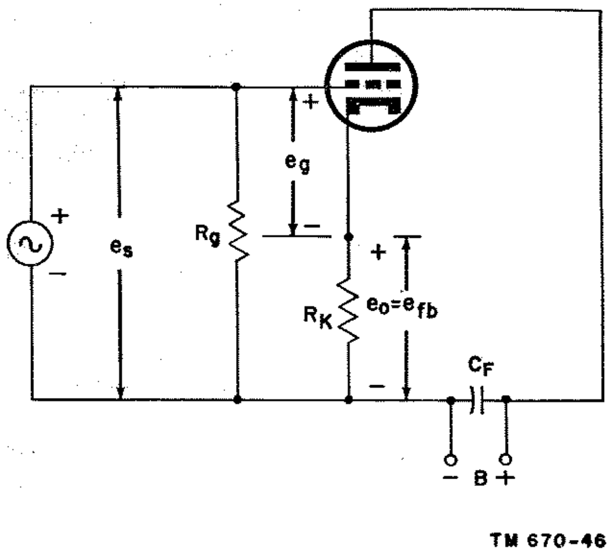

Figure 53. Cathode-follower circuit.

load resistor, R_K, is large compared with the power consumed by the high impedance of the grid circuit. Because of this, the cathode-follower circuit is often called the *cathode-follower power amplifier*. In the conventional amplifier the voltage gain is invariably greater than 1, whereas the power output may be made small or large, depending on the circuit application.

68. Operation of Cathode-Follower Circuit

a. The conventional cathode follower is readily understood if its operation first is analyzed with no signal applied. Under this condition, the d-c bias is produced across load resistor R_K by the steady value of the plate current which flows through it (ch. 1).

b. In order to obtain an output-voltage variation, e_o, it is necessary to apply a grid voltage, e_g, between grid and cathode. For the polarities shown, a positive increase in grid voltage e_g results in a positive increase in output voltage, e_o, whereas a decrease in grid voltage produces a decrease in output voltage. Thus, the output voltage, e_o, *follows* the grid voltage, e_g. The sum of grid voltage e_g plus output voltage e_o constitutes the signal voltage e_s—which is, therefore, always greater than either grid voltage e_g or output voltage e_o. The voltage gain (the output voltage divided by the input voltage) is, therefore, always less than 1, since output voltage e_o is always less than input signal voltage e_s.

c. To see that the cathode follower may be regarded as a form of negative feedback amplifier, trace the grid circuit from the cathode to the grid. The output voltage, e_o, is in series with the signal voltage, e_s, and is of opposite polarity. Thus the entire output voltage is a negative feedback voltage, e_{fb}, as designated.

69. Constant-Voltage Equivalent Circuit of Cathode Follower

a. The characteristics of the cathode follower are analyzed by the use of an equivalent circuit (C, fig. 54), similar to that of the conventional single-stage amplifier explained in chapter 1.

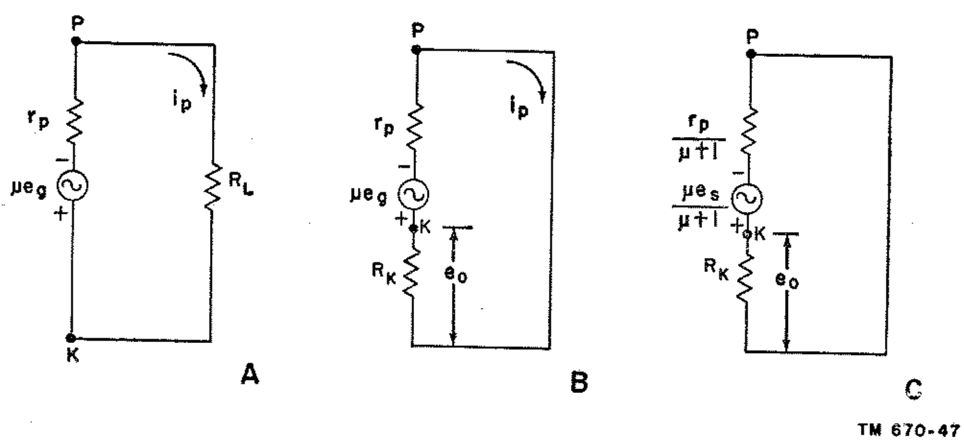

Figure 54. Equivalent circuits of conventional amplifier and cathode follower.

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b. The equivalent circuit of the conventional amplifier is shown in A of figure 54. In B, the load resistor, R_L, is replaced by a cathode resistor, R_K. This is an equivalent circuit of the cathode follower as it stands. However, figure 53 shows that in the cathode follower the grid voltage is only a part of the applied signal voltage. To facilitate calculation of the output voltage and the gain, it is more convenient to use an equivalent circuit in which the signal voltage, e_s, appears rather than the grid voltage, e_g. This transformation has been made in C of figure 54. The equivalent generator voltage, μe_s, has been divided by the quantity (μ plus 1); similarly, the actual plate resistance of the tube has been divided by (μ plus 1). A simple formula for the output voltage of the cathode follower is found more readily from the constant-current equivalent circuit which will be given in paragraph 75.

c. The circuit in C shows that the internal impedance of the cathode follower is smaller than the plate resistance of the tube by the factor $1/(\mu$ plus 1), and the circuit is useful for coupling to a low-impedance load. This will be clarified in the following paragraph.

70. Voltage Gain of Cathode Follower

a. The gain of a cathode follower is found from the equivalent circuit in C, figure 54. The equivalent generator voltage $\mu e_s/(\mu$ plus 1) always is less than the signal voltage, e_s, since the denominator of the fraction $\mu/(\mu$ plus 1) exceeds by 1 the numerator μ; consequently, the value of the equivalent generator voltage is always less than the signal voltage. For example, figure 55 shows a practical cathode follower and its equivalent circuit in which the amplification factor, μ, is 19. The equivalent generator voltage $\mu e_s/(\mu$ plus 1) equals (19 times e_s)/20, or .997 e_s, which is slightly less than the signal voltage.

b. This example shows that the output voltage is always less than the input voltage, since the equivalent generator voltage divides across the series combination of cathode resistor R_K and the effective internal plate resistance $r_p/(\mu$ plus 1). Again consider, for example, the practical cathode-follower and equivalent circuits of figure 55. The plate resistance of the tube is 10,000 ohms, the amplification factor is 19, and the cathode resistance is 500 ohms.

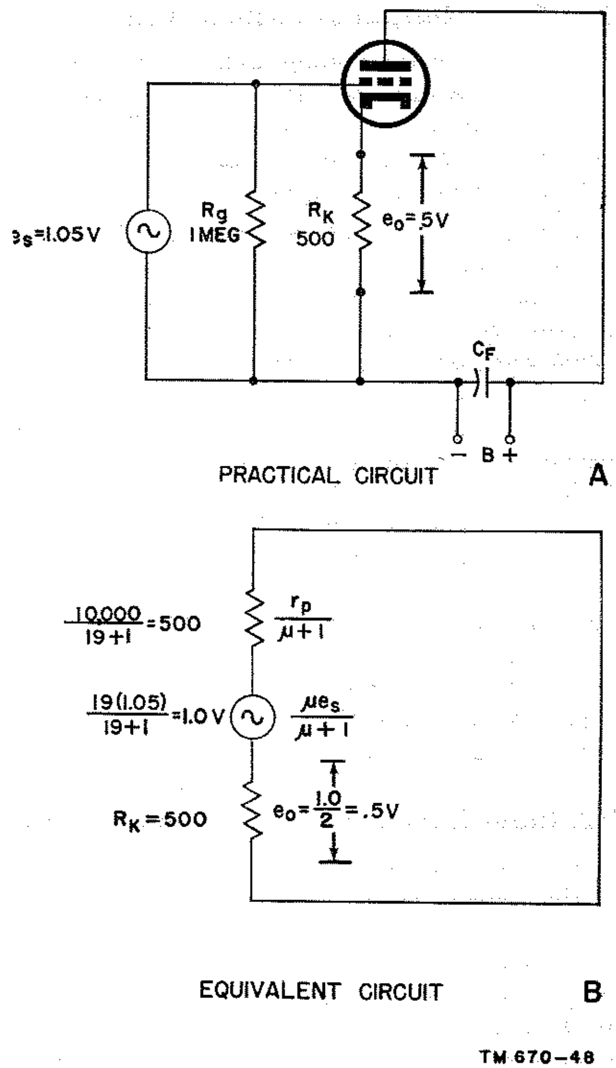


Figure 55. Practical cathode follower.

The *effective* plate resistance is $r_p/(\mu$ plus 1), or 10,000/(19 plus 1), which equals 500 ohms. When a signal of 1.05 volts is applied to the input of the cathode follower, an equivalent generator voltage of approximately 1 volt is developed, as shown in B. This voltage divides across the series combination of the cathode resistor and the effective plate resistance—that is, across two 500-ohm resistances in series. The output voltage is the voltage across the cathode resistor, R_K—in this case, one-half of the equivalent generator voltage, or .5 volt.

c. The voltage gain of any amplifier, including cathode followers, is the output voltage divided by the input voltage. In the present example, the gain is .5/1.05, or .476.

71. Power Output of Cathode Follower

a. Although the voltage gain of the cathode follower is always less than 1, its power gain may be, and usually is, very high. This is one reason why a cathode follower is superior to a transformer in matching unlike impedances.

b. For a numerical example of this, again consider the cathode follower shown in figure 55. The power delivered to the 500-ohm load resistor is e_o^2/R_K = (.5 volt)^2/500 ohms = .0005 watt, or 500 microwatts. The power consumed by the input circuit of the tube is the power consumed by grid resistor R_g alone. The vacuum-tube grid circuit consumes no power. It has an input resistance which is nearly an open circuit since the grid draws no current because it is biased with a negative voltage. Thus, the input power is only 1.1 microwatts, since E_s^2/R_g = (1.05 volts)^2/(10^6 ohms), or $1.1(10^{-6})$ watts. The power output of the cathode follower, 500 microwatts, is 455 times this input power. Thus, the power gain of the cathode follower is very high.

72. Input Impedance of Cathode Follower

The input impedance of a cathode follower is the impedance between the input terminals 1 and 2 (figs. 56 and 57). In general, it consists of the resistance of the grid resistor, R_g, in parallel with the input capacitances, as in B of figure 56. This input capacitance consists almost entirely of the interelectrode capacitances of the tube. The input impedance of the cathode followers discussed thus far cannot be greater than the resistance of the grid resistor. By a modification of the circuit, however, the input impedance of a cathode follower can be made very large—much larger than grid resistance R_g. This is useful property in electronic applications where the cathode follower serves as an impedance-matching device (fig. 51). The high input impedance produces negligible loading on the preceding stage to which it is connected; consequently, the cathode follower has no appreciable tendency to distort the waveshapes which it amplifies.

73. Input Resistance of Cathode Follower

a. At low frequencies where the reactances of the interelectrode capacitances are negligible,

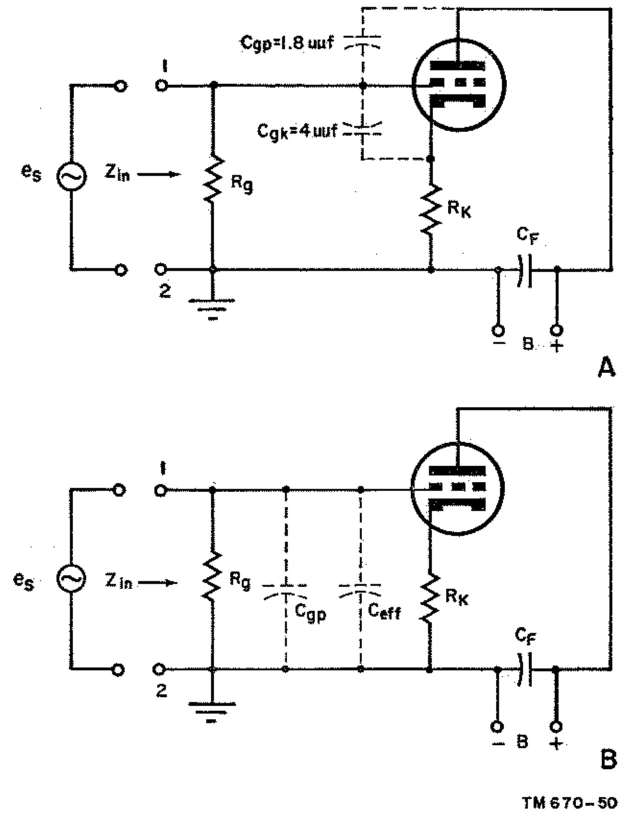


Figure 56. Input impedance of cathode follower.

the input impedance, Z_{in}, is simply the resistance of R_g (fig. 56).

b. The input resistance of a cathode follower can be increased to an effective value that is much greater than the value of the grid resistor alone, by connecting the grid resistor, R_g, to the top of the cathode load resistor, R_K (fig. 57). The desired value of operating bias for the tube is obtained from a separate bias resistor, R_B, which is bypassed by capacitor C_B. The input resistance of the cathode follower is the resistance, R_{in}, which the generator sees, or into which it operates. By Ohm's law it has the value e_x/i_{in}, where i_{in} is the current which flows through grid resistor R_g. This current is equal to grid-to-cathode voltage e_{gk} divided by grid resistance R_g.

c. (1) A numerical example will illustrate why the effective input resistance exceeds grid resistance R_g in this circuit. Assume that the grid resistance is 1 megohm, and the gain of the stage is

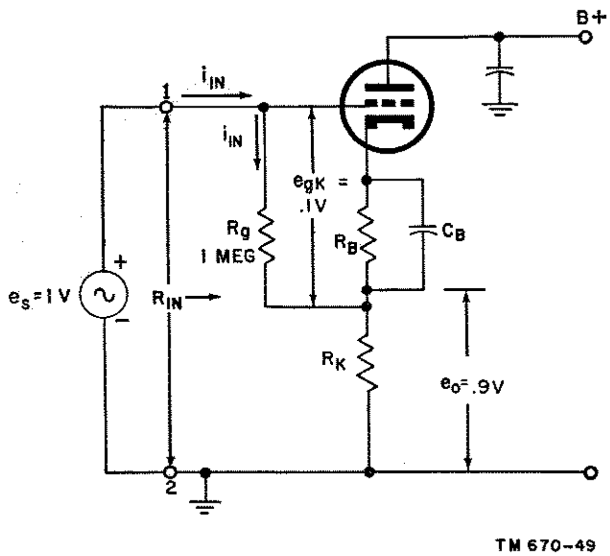


Figure 57. Cathode follower with high input resistance.

.9. For a signal voltage, e_s, of 1 volt, there appears an output voltage, e_o, of .9 volt across the cathode resistor, R_K. The grid-to-cathode voltage across the 1-megohm grid resistor, R_g, is the difference between 1 volt and .9 volt, or .1 volt. Current i_{in} is .1 μa (microampere), since e_{gK}/R_g is .1 volt/10^6 ohms, or .1μa. The input resistance seen by the generator is 10 megohms, since signal voltage e_s is 1 volt and current i_{in} is .1 μa. Thus, the input resistance is the signal voltage divided by i_{in}, or 1 volt/.1μa, or 10 megohms. This example shows that the effective input resistance is increased from a value of R_g to $10R_g$, because the input signal voltage is 10 times as large as the voltage between grid and cathode.

(2) Similarly (fig. 57), if the voltage gain of the cathode follower is .75, the effective input resistance, R_{in}, is four times as great as the actual grid resistance, or 4 megohms, since the voltage between grid and cathode is one-fourth the signal voltage; that is, the input signal voltage is four times as great as the grid-cathode voltage.

(3) A simple formula for effective input resistance is $R_{in} = R_g/(1 \text{ minus A})$, where A is the voltage gain of the cathode-follower stage. This formula and the numerical illustrations show that a large input resistance is obtained by using a large value of grid resistance, R_g, and a gain, A, as close to unity as possible.

74. Input Capacitance of Cathode Follower

a. The input capacitance of a cathode follower (fig. 56) consists of the interelectrode capacitance, C_{gp}, between grid and plate, shunted by a fraction of the interelectrode capacitance, C_{gk}, between grid and cathode. As viewed from input terminals 1 and 2 (A, fig. 56), this statement does not appear true. However, the *effective* capacitive reactance of the grid-to-cathode capacitance is the actual reactance divided by the quantity (1 minus A). This result is arrived at by exactly the same method used for determining the effective value of the input resistance (par. 73). Further analysis shows that the effective input capacitance resulting from the grid-to-cathode capacitance is the actual value of the grid to cathode capacitance times (1 minus A). Since in the cathode follower the term (1 minus A) is always less than unity, the *effective* input capacitive value caused by the grid-to-cathode capacitance is always lower than the *actual* value.

b. The effective grid capacitance is shunted by the grid-to-plate capacitance, C_{gp}. In A, this interelectrode capacitance is drawn in the conventional manner. It may, however, be drawn in shunt with the effective grid capacitance, C_{eff}, as in B, since the large bypass capacitor, C_F, grounds the plate for a-c signals. Therefore, the total capacitance seen by the generator at input terminals 1 and 2 is the sum of the effective input capacitance plus the actual grid-to-plate capacitance.

c. For example, assume the gain of the cathode follower illustrated in figure 56 to be .9. The 4-$\mu\mu$f grid-cathode capacitance appears to shunt the input terminals with a capacitance having *one-tenth* its actual value—that is, .4 $\mu\mu$f. This result is obtained as follows: Effective capacitance equals grid-to-cathode capacitance times (1 minus A), or (4) (1 minus .9), or .4 $\mu\mu$f. The total input capacitance, therefore, is the sum of the grid-to-plate capacitance plus the effective grid capacitance, or (1.8 plus .4) $\mu\mu$f, or 2.2 $\mu\mu$f. Thus, a simple formula for finding the input capacitance of a cathode follower is: Input capacitance equals grid-to-plate capacitance plus the quantity grid-to-cathode capacitance times (1 minus A). This formula and the numerical illustration demonstrate that a small input capacitance is obtained by making gain A large, and by using a tube having small interelectrode capacitances.

75. Constant-Current Equivalent Circuit of Cathode Follower

The behavior of a practical cathode follower is estimated quickly if the gain and the output impedance are known. Simple relations for the gain are found from the constant-current equivalent circuit, by a procedure similar to the one described in chapter 2, in connection with R-C amplifiers.

a. The constant-current equivalent circuit of the R-C coupled amplifier is reproduced in A of figure 58 along with the constant-current equivalent circuit of the cathode follower amplifier, shown in B. The plate resistance of the R-C amplifier is divided by (μ plus 1) in the cathode follower, and the load resistance of the R-C amplifier becomes the cathode resistor, R_K, in the cathode follower. (For simplification, the grid resistor of the following stage is omitted.)

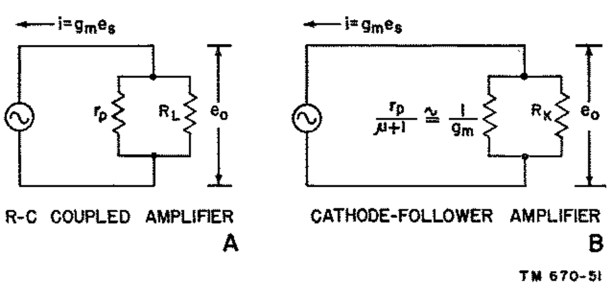


Figure 58. Constant-current equivalent circuits.

b. The equation for output voltage may be rearranged in a form which includes the transconductance, g_m, of the tube. It is

output voltage $e_o = g_m R_o e_s$

Thus, the voltage gain is e_o/e_s, which equals $g_m R_o$. Here the resistance term, R_o, is the parallel combination of the internal effective-plate resistance and the cathode-load resistor. This equation for output voltage is the basis for the constant-current equivalent circuit. The term $g_m e_s$ may be written $(\mu e_s)/r_p$, since $g_m = (\mu/r_p)$. This must represent a *current* since it consists of a voltage, μe_s, divided by a resistance, r_p. Thus, the amplifier output voltage, e_o, is produced by a current, $(\mu e_s)/r_p$, or $g_m e_s$, flowing through R_o, which consists of two parallel resistances, $r_p/(\mu$ plus 1) and R_K. In B is shown the equivalent circuit which represents the circuit conditions for the cathode-follower amplifier.

76. Output Impedance of Cathode Follower

The output impedance of a cathode follower must be known if the cathode follower is used to match an impedance such as that of the transmission line in figure 51. The output impedance of the cathode follower is the impedance seen, or measured, across the output terminals—that is, across R_K. It is found readily from the constant-current equivalent circuit (B, fig. 58).

a. It also is shown in B that the output impedance consists of two parallel impedances, the internal plate resistance $r_p/(\mu$ plus 1) and the total external load, R_K. The output impedance of a cathode follower usually is small in comparison with the actual plate resistance of the vacuum tube used. Assume, for example, that a cathode follower uses a triode having an amplification factor of 19, a plate resistance of 10,000 ohms, and a cathode load resistor of 1,000 ohms. The effective internal resistance is $r_p/(\mu$ plus 1), or 10,000/20, or 500 ohms. The output impedance, therefore, is 500 ohms in parallel with 1,000 ohms, or 333 ohms.

b. (1) Note that the output impedance is always less than the internal resistance, since load resistor R_K is in parallel with it. Thus, the largest possible value of output impedance for a given tube is the internal resistance $r_p/(\mu$ plus 1). This would occur if load R_K were exceedingly large, or an open circuit. The effective internal-generator resistance then becomes nearly equal to r_p/μ, or $1/g_m$, if μ is much greater than 1. Because of this, it is common practice to represent the output impedance as a resistor of value, $1/g_m$, connected in parallel with load resistor R_K. This representation emphasizes that a low impedance is obtained by using a tube having a large value of transconductance, g_m.

(2) For example, if a cathode follower uses a tube with a transconductance of 2,000 umho, the effective internal resistance is approximately $1/(2{,}000)(10^{-6})$, or 500 ohms. If the tube is replaced by another having a transconductance of 4,000 umho, the effective internal resistance drops to 250 ohms.

(3) The low value of effective internal resistance of the cathode follower is a desirable characteristic in many electronic applications, because a constant output voltage results from such a characteristic (par. 60).

77. Frequency Response of Cathode Follower

a. One of the outstanding characteristics of the cathode follower is its unusually wide frequency response. This makes the cathode follower useful for amplifying wide-band signals, as required in radar, television, and other electronic applications.

b. (1) The frequency response of the cathode follower is flat over an unusually wide range. It falls off at very low frequencies because of the reactance of C_F, the bypass capacitor of the *B* supply (fig. 53). This is true since a large reactance between plate and cathode reduces the varying component of the plate current which flows through load R_K, and thus reduces the output voltage. For a good low-frequency response, therefore, it is essential to use a capacitor, C_F, of large value.

(2) The response drops at very high frequencies where any capacitive reactance across the load, R_K, viewed from the output terminals, may have an appreciable shunting effect. Such a case may exist because of the plate-to-cathode capacitance, but it is usually negligible, and therefore is not shown in figure 57. For example, if the plate-to-cathode capacitance is 10 $\mu\mu$f, it has a reactance of 1,000 ohms at about 16 megacycles. Below this frequency, its shunting effect across a 1,000-ohm cathode resistor may thus be neglected. However, the high-frequency response falls off appreciably if a high capacitance cable is connected across the load resistor; therefore, in order to maintain the response uniform up to very high frequencies, it is essential to use connectors having negligibly small capacitance.

78. Distortion in Cathode Follower

a. Amplitude Distortion. Amplitude distortion in the cathode follower is equal to the distortion in a conventional amplifier divided by the factor (1 plus μ), as discussed in chapter 4 on negative-voltage feedback amplifiers. For example, assume that a single-stage R-C coupled amplifier uses a triode with an amplification factor of 19, a plate resistance of 10,000 ohms, and a load resistor of 1,000 ohms, or 1/10 of r_p. Since the load resistor is small compared with the plate resistance, the distortion is high—approximately 20 percent. In a cathode-follower circuit this distortion drops from 20 percent to 1 percent if the same tube uses a 1,000-ohm cathode resistor, since 20 percent divided by (1 plus μ) is 20 percent/20, or 1 percent. Thus, the cathode follower circuit operates into low-impedance loads with a low amount of amplitude distortion.

b. Phase Distortion. Phase distortion is extremely low in the cathode follower. It is closely related to frequency distortion. There is no appreciable phase distortion in the flat range of an amplifier. This was explained in the discussion of conventional amplifiers, and it holds equally true for cathode followers.

79. Summary

a. Cathode followers are useful for coupling a high-impedance circuit, such as the output of an amplifier, to a low-impedance load, such as a transmission line.

b. The cathode follower is characterized by simplicity of circuit, high power gain, wide range of flat-frequency response, and low distortion.

c. The voltage gain of the cathode follower is $g_m R_o$ and is always less than 1.

d. The internal resistance of the cathode follower is $r_p/(\mu$ plus $1)$, or approximately $1/g_m$. The output impedance, R_o, consists of the internal resistance, $r_p/(\mu$ plus $1)$ in parallel with R_K.

e. The input impedance of the cathode follower consists of the input resistance in parallel with the input capacitance. The formula for input resistance is $R_{in} = R_g/(1$ minus $A)$; R_{in} is infinite when there is no grid resistor. The formula for the input capacitance is $C_{in} = C_{gp}$ plus $C_{gk}(1$ minus $A)$.

80. Review Questions

a. Sketch a simple cathode-follower circuit.

b. Why are cathode followers useful?

c. What is the meaning of the term cathode follower?

d. Is the cathode follower a voltage amplifier or a power amplifier? Explain.

e. What kind of feedback is used in the cathode-follower circuit?

f. Why is the voltage gain of the cathode follower always less than 1?

g. What is the ratio of the feedback voltage to the output voltage in the cathode follower?

h. Assume that a cathode follower uses a tube having a μ of 24, an r_p of 10,000 ohms, and R_K of 400 ohms. Find (1) the output impedance; (2) the voltage gain.

i. In review question *h*, what is the largest possible value of output impedance?

j. Assume that the amplifier of figure 56 has a voltage gain of .5. Find R_{in}.

k. Assume that the amplifier of figure 57 has a gain of .5. Find C_{in}.

l. In a cathode follower, how is the input resistance made very large? How is the input capacitance made very small? How is the internal resistance made very small? What are the reasons for the wide range of flat response?

m. What is the effect on the amplifier response of connecting a long cable to the output terminals?

n. Compare the distortion produced in a cathode follower with the distortion produced in an amplifier using no feedback.

CHAPTER 6

PHASE INVERTERS

81. Fundamentals of Phase Inverters

a. Many electronic devices require an output signal voltage whose polarity is opposite to (the inverse of) the input voltage. For example, the cathode-ray tube used in certain radar and television circuits requires that a positive-going voltage be applied to one plate of a pair of deflection plates at the same time that a negative-going voltage is applied to the other plate. In many cases, only a positive-going voltage is available from the supply source. This is applied to the deflection plate requiring the positive-going voltage. The negative-going voltage for the other plate is obtained by inverting the polarity of the positive-going signal, making it a negative-going one. The device which inverts the polarity of a signal is known as a *phase inverter*. Figure 59 shows graphs of such a transposition.

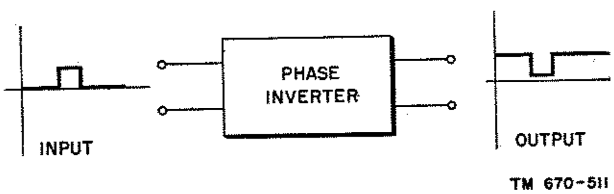


Figure 59. Phase inversion.

b. A phase inverter produces an output voltage of opposite polarity, or opposite phase, in respect to the input voltage. The meaning of *phase* in the term *phase inverter* applies to the *general* definition of phase rather than to the *special* meaning of a lag or a delay in time. Phase inversion here does *not* refer to a time delay between output and input voltages, since there is usually no appreciable time difference between these voltages in the ordinary phase inverter. *Polarity inverter* is a more descriptive name. For a sinusoidal input signal, the inverter output voltage is 180° out of phase in respect to the input voltage, as in A of figure 60. Here, as with the pulse signal of figure 59, the polarity of the output voltage is negative-going during the times when the polarity of the input voltage is positive-going, and vice versa. (Only with a sine wave could this result be obtained by a time-delay or phase shift.) In either case, the waveform is *inverted*—its polarity *reversed*—by the phase inverter, and the waveshape of the negative excursion of the output voltage is at all instants the same as the corresponding positive excursion of the input voltage. Except for the inversion, the waveshapes are unaffected by the phase inverter.

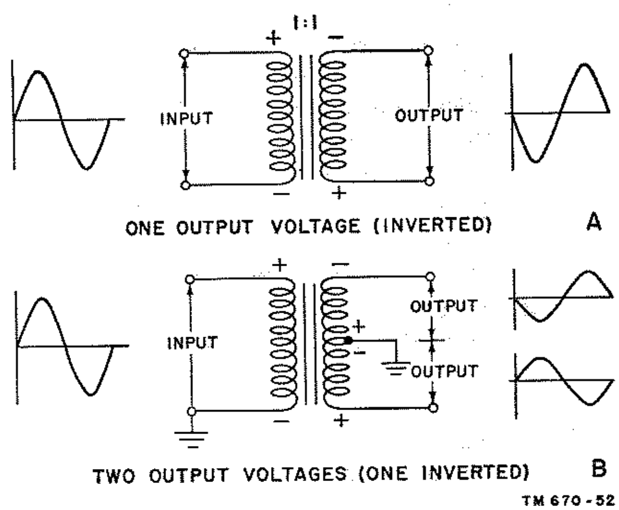


Figure 60. Transformer phase inversion.

82. Types of Phase Inverters

A phase inverter may be a transformer or a vacuum-tube circuit.

a. Transformer Phase Inverter.

(1) In figure 60, A shows a *transformer* used as a simple phase inverter for

sinusoidal signals. Note that the output voltage wave reaches its negative peak at the same instant that the input voltage wave reaches its positive peak. Thus, the output and input voltages are 180° out of phase.

(2) In B, a transformer is used to provide two voltages of opposite phase, or polarity. It splits the input voltage into two output voltages which vary in opposite directions, or are opposite in phase. Circuits which do this often are used to supply the signals of opposite polarity required by the two tubes of push-pull amplifiers.

b. Vacuum-Tube Phase Inverters.

(1) Figures 63 and 65 show *vacuum-tube phase-inverter circuits.* These utilize the fact that the polarity of the output voltage of an amplifier stage is inverted in respect to its input; that is, the instantaneous plate voltage rises when the grid voltage falls.

(2) Figure 66 shows two vacuum-tube circuits used as phase splitters. Note that *two* output voltages are developed in each case—voltage e_{o1} between plate and ground and voltage e_{o2} between cathode and ground. Voltage e_{o2} is of the same polarity as the input, e_s, and e_{o1} is of opposite polarity of e_s. Thus, e_{o1} and e_{o2} are of opposite polarity, and may be used to supply the two grids of a push-pull amplifier stage.

(3) In the phase inverters of figures 67 and 68, one tube, $V1$, acts as a conventional amplifier, and the other, $V2$, acts as phase inverter and amplifier. Tube $V1$ delivers voltage e_{o1} to its output terminals. A fraction of this voltage is amplified and inverted by the phase inverter, $V2$. The inverted output voltage, e_{o2}, therefore, is of opposite polarity to e_{o1}. Thus e_{o1} and e_{o2} may be utilized to drive a push-pull amplifier stage. Vacuum-tube circuits which split an input voltage into two output voltages that vary in opposite directions are known as *phase splitters,* or *paraphase amplifiers.* These and other circuits are discussed in detail later in this chapter.

83. Transformer Phase Inversion

A transformer is a good example of a simple phase inverter for range of the frequencies in which the secondary voltage is opposite in polarity in respect to the applied, or primary, voltage.

a. Figure 60 shows the effects of phase inversion in a transformer. A sine-wave voltage is applied to the primary terminals of the transformer. The secondary voltage of the transformer is negative during the interval when the primary voltage is positive, and, conversely, the secondary voltage is positive during the interval when the primary voltage is negative. Thus, for the sine-wave signal, the primary and secondary voltages are always opposite in polarity.

b. A transformer primarily is suited for inverting audio-frequency signals, or pure sine waves, which fall in the middle range of the frequency characteristic of the particular transformer. A transformer, therefore, is not suited for inverting sharp pulses, since these generally contain a large number of high-frequency harmonics. A transformer shifts the phase of the harmonics in respect to the fundamental frequency components, which results in phase distortion of the output pulse. The phase shift is negligible in the flat, or middle, range of the transformer response characteristic; consequently, in the middle range, the secondary voltage is a reasonably faithful inversion of the primary voltage. The phase shift becomes leading for signals whose frequency is much less than the exact midfrequency; conversely, it becomes lagging for signals whose frequency is far above the middle range. Thus, a transformer is useful as a phase inverter only in the middle range of the transformer response curve.

c. A transformer inverter is desirable as a coupling device when the following amplifier stage draws grid current. This usually occurs in amplifiers operating in Class AB2 or in Class B. Grid current, flowing through the secondary of a transformer, produces only a negligible drop in the relatively low resistance of the secondary winding. In this case, operation with a transformer inverter is superior to operation with a vacuum-tube type of phase inverter, a type to be discussed later.

d. As a rule, a step-down ratio is used between primary and secondary of transformers which couple to a power amplifier operating in Class AB2 or in Class B. This is done to insure that the power amplifier grid is driven from a low impedance source. The grid-to-cathode impedance is low during that portion of the cycle when the a-c signal voltage drives the grid of the power amplifier positive. The reflected source impedance must also be low, to minimize distortion and assure that the major portion of the a-c signal appears at the power-amplifier grid, and that only a negligible portion is lost in the generator impedance. Practical values of step-down ratios of such a coupling transformer are in the order of 1.5 to 1, or 2 to 1.

e. Triode amplifiers using transformers as phase inverters usually produce higher output voltages and powers than triode amplifiers using resistance coupling. A higher gain is obtained because nearly the entire B-supply voltage is applied to the plate of the tube. There is only a negligible drop in the winding. Tetrodes or pentrodes as a rule are not used, for considerations of frequency response.

f. The disadvantages of the transformer phase inverter are the same as those of transformer coupling, discussed in chapter 2. They include intermodulation distortion, space requirements, weight, hum pickup, and cost of high-quality transformers.

84. Transformer Phase Splitter

Transformers are excellent phase splitters. A phase splitter, it will be recalled, splits an input voltage into two output voltages that vary in opposite directions at all times. Figure 61 shows an amplifier in which the transformer, T1, acts as phase splitter. The secondary winding of the transformer is split to provide the two voltages required to drive the separate grids of tubes *V2* and *V3* of the push-pull amplifier.

a. In figure 61, the push-pull voltages are developed as follows: An a-c grid signal voltage, e_{g1}, is applied to the grid of tube *V1*. The amplified voltage, e_{p1}, appears across the primary terminals 1 and 2 of the transformer. The transformer secondary splits this voltage into two equal voltages, e_{g2} and e_{g3}, and applies them to the grids of *V2* and *V3*. These two voltages always vary in opposite directions. For example, if, at one instant, the a-c signal voltage makes terminal 3 negative in respect to the center tap, terminal 4, at the other end of the transformer secondary, is positive in respect to the center tap, and vice versa. Consequently, the two voltages, e_{g2} and e_{g3}, are exactly 180° out of phase at all times. This phase difference is independent of frequency if the tap is at the exact electrical center of the secondary winding of the transformer.

b. It is important to note that the transformer phase splitter (fig. 61) is suited primarily for splitting signals which lie in the middle range of the transformer frequency characteristic, such as audio-frequency signals

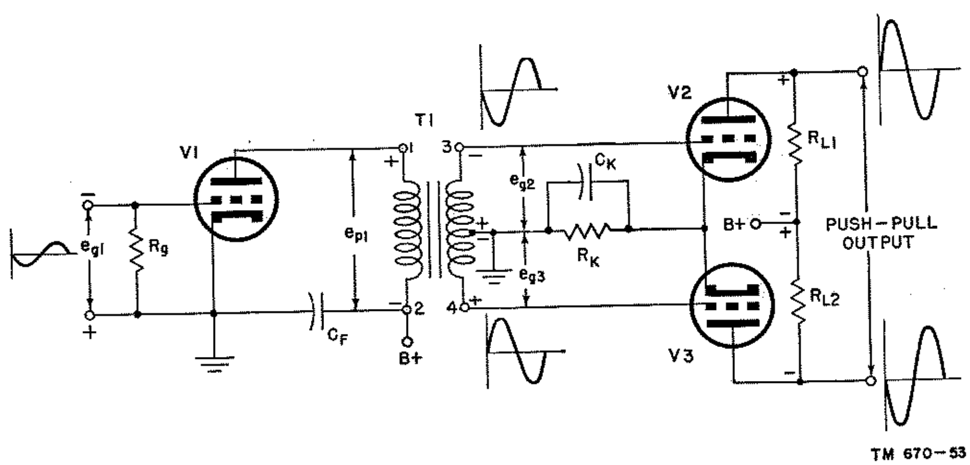


Figure 61. Transformer phase splitter.

or sine waves. Although the polarity of the two output voltages are exactly opposite to *each other* at all frequencies, the phase shift between the secondary voltage, e_{g2}, and the primary voltage, e_{p1}, is zero only at the exact midfrequency of the transformer. When the input signal contains a large number of harmonics, the higher harmonics are appreciably shifted in respect to the fundamental in the output. The waveshape of e_{g3} is undistorted only if the waveshape of e_{g2} is undistorted, since e_{g3} is the exact inverse of e_{g2}. Conversely, the waveshape of e_{g3} is distorted if the waveshape of e_{g2} is distorted, as can happen if a sharp pulse is applied to the primary of the transformer. Thus, a transformer used as a phase splitter for complex waves has the same limitations as the transformer phase inverter.

c. The transformer phase splitter has the same advantages and disadvantages as the simple transformer inverter. The advantages include moderate amplifier gain at low distortion resulting from the low d-c resistances in the windings. The disadvantages include restriction to narrow-band operation, space requirements, hum pickup, and high cost.

d. (1) Figure 62 shows a transformer phase inverter using a resistance-type voltage divider instead of a tapped transformer. This circuit suffers from several additional disadvantages and, therefore, is used only as a makeshift circuit in the event that a center-tapped transformer is needed but is not immediately available. The circuit can operate as a perfect phase inverter if the two resistors, R_{g2} and R_{g3}, are identical, and thus serve to split the secondary voltages into two equal voltages at all frequencies, like the center tap on a push-pull transformer. The windings of a center-tapped push-pull transformer usually are so arranged that there are equal, very small, distributed capacitances between the grounded transformer core and the two outside terminals; consequently, the two output voltages have equal magnitudes at all frequencies.

(2) The untapped transformer of the makeshift circuit (fig. 62), however, usually is designed for use only in single-ended amplifier circuits. One terminal normally is connected to the grid of an amplifier tube, and the other to ground. The ground connection therefore short-circuits the large distributed capacitance between ground and the low end of the transformer winding. One end of the transformer winding is usually the innermost part of the winding, close to the core, and has a large resultant capacitance to ground. The other end connects to the outermost part of the winding far from the core, and has a small resultant capacitance to ground. Resistors R_{g2} and R_{g3} are shunted by the distributed capacitances (shown by dashed lines) with reactances which

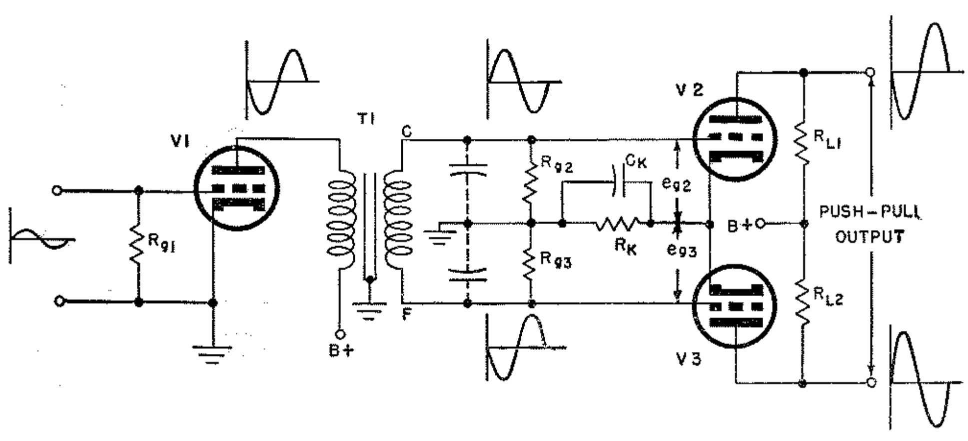


Figure 62. Transformer phase splitter using voltage divider.

are unequal at all frequencies. Thus, voltages e_{g2} and e_{g3} are unequal, even though the two resistors are identical. This inequality or *unbalance* of voltages can be minimized by shunting a capacitor across R_{g2} to compensate for the unequal capacitances, but the response of the transformer nevertheless falls off rapidly at the high-frequency end.

85. Vacuum-Tube Phase Inverter

Vacuum tubes can be used as phase inverters, because the output voltage of a single-stage amplifier is of opposite polarity in respect to the input voltage. Vacuum-tube phase inverters in general are superior to the transformer types because a vacuum-tube circuit can be designed to produce negligible values of frequency, phase, and amplitude distortion. Thus, a vacuum-tube phase inverter can have a flat frequency response and a negligible phase shift over a much wider range than a transformer.

86. Amplifier and Phase Inverter

Figure 63 shows that a simple vacuum-tube phase inverter is merely a single-stage R-C coupled amplifier. A positive-going signal, e_1, applied to the grid appears as an amplified, negative-going signal, e_2, across the output resistor R_g, and vice versa.

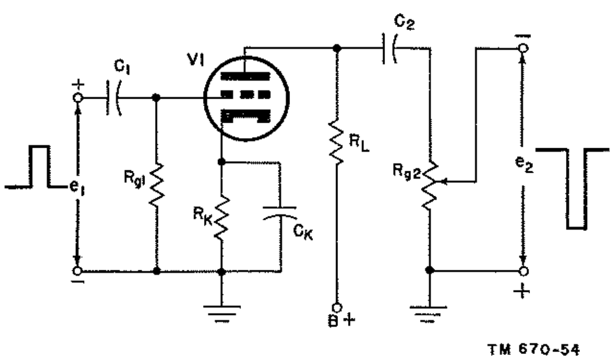


Figure 63. Phase inverter using vacuum tube.

a. (1) The circuit of figure 63 illustrates several advantages of the vacuum-tube phase inverter over the transformer inverter. One advantage is that its output voltage is considerably greater than the input voltage, because the tube amplifies the applied signal. Another is the isolation which its tube provides between output and input circuits. This is important, because variations in the output circuits of the inverter have no effect on the input circuit. Even in the extreme case, when the output of a vacuum-tube phase inverter is accidentally short-circuited, the grid circuit of the inverter is not affected. A transformer used as a phase inverter would reflect a short circuit occurring in the secondary into the circuits connected to the primary.

(2) The voltage or power output of a vacuum-tube phase inverter is variable over a wide range. The output is varied by using a variable grid resistor (fig. 63) to vary the grid voltage applied to the tube, or by altering the d-c plate and grid voltages.

b. (1) The circuit of figure 63 illustrates one disadvantage of vacuum-tube phase inverters: It could not be used to supply an amplifier stage which draws d-c grid current. Grid current flowing through resistor R_{g2} might cause considerable amplitude distortion by altering the bias on the succeeding stage. As an illustration, assume that the output of the vacuum-tube phase inverter feeds the grid circuit of a power amplifier operating in Class B. Assume also that the tube draws 10 microamperes of d-c grid current when the a-c grid signal reaches its peak amplitude. The d-c grid current, i_g, flowing through R_{g2}, produces a d-c voltage, $-i_gR_{g2}$. If R_{g2} is 1 megohm, the resulting d-c bias across R_{g2} is —10 volts. This bias voltage adds to the fixed bias of the stage. It may alter the d-c operating conditions of the power-amplifier stage in a manner which causes considerable amplitude distortion.

(2) The use of transformer T1 (fig. 61) remedies this difficulty. Consequently,

the resistance of the secondary winding of transformer T1 usually is low (a few hundred ohms). There is only a negligible d-c drop in its winding resistance, therefore, from flow of grid-current between grid and cathode. Thus, there is no significant change in d-c grid bias when an a-c signal is applied to the power amplifier.

87. Phase Inverter with Cathode Degeneration

Cathode bypass capacitor C_K is omitted from the phase inverter shown in figure 64, to provide degenerative current feedback, as discussed in chapter 4. This feedback connection is used when it is desired to obtain phase inversion with little or no amplification. The gain of the phase inverter is reduced more readily by this method than by using a voltage divider—which may introduce frequency distortion—and the frequency response of the phase inverter is improved.

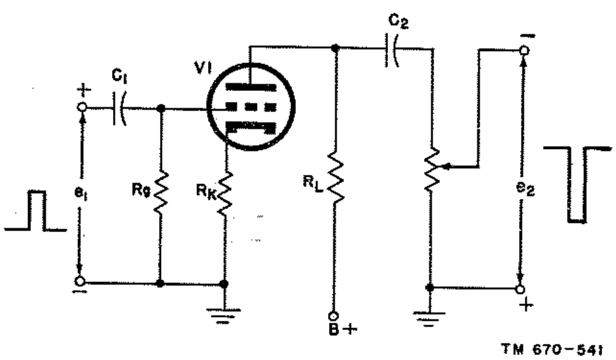


Figure 64. Phase inverter using degenerative feedback.

88. Unity Gain Phase Inverter

a. (1) Figure 65 shows a phase inverter in which the amplitude of the inverted output voltage is exactly equal to the amplitude of the input voltage. This circuit is used where it is desired to invert a signal without changing its amplitude. The voltage gain of the amplifier, e_o/e_s, is reduced to unity by using a voltage divider consisting of R_{g1} and R_{g2}, connected in series, to reduce the voltage applied to the grid of the tube. For example, if the gain, e_o/e_g, of the tube circuit is 30, the voltage divider is designed so that e_g is 1/30 of the available signal voltage, e_s. Consequently, the resistance of R_{g2} is made 1/30 of the resistance of the voltage divider.

(2) Assume that design considerations require a voltage divider of 300,000 ohms. The value of R_{g2} then is 300,000 divided by 30, or 10,000 ohms; the value of R_{g1} is the difference between 300,000 ohms and 10,000 ohms, or 290,000 ohms. If the available signal voltage is 30 volts, it is reduced by the voltage divider to 1/30 of 30 volts, or 1 volt. The tube amplifies the 1 volt back to its original value of 30 volts. Thus, the output voltage, e_o, has the same amplitude as the input signal, with its output inverted, as required.

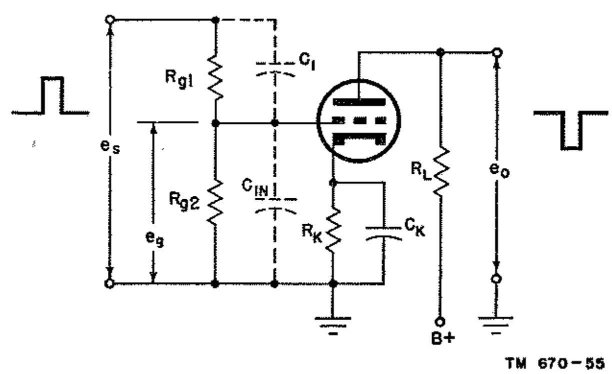


Figure 65. Unity-gain phase inverter.

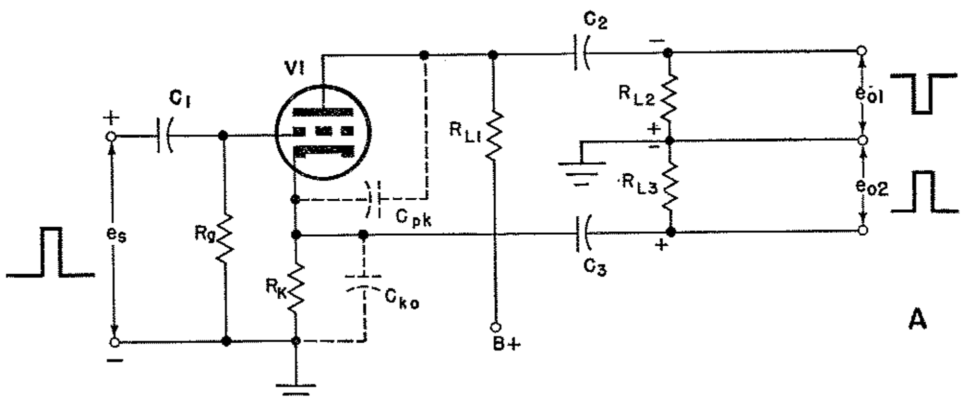
b. If the waveform to be inverted contains many harmonics, as in the case of a square wave, the voltage divider must be compensated to minimize the shunting effect of the input capacitance of the tube, C_{in}, across R_{g2}. This is done (fig. 65) by connecting across resistor R_{g1} a capacitor, C_1, the reactance of which bears the same numerical relation to the reactance of the input capacitance C_{in} as R_{g1} does to R_{g2}. Under these conditions, the signal voltage is divided in the required ratio at *all* frequencies. For example, assume that R_{g1} is 29 times as large as R_{g2}, as before, and that C_{in} is 29 $\mu\mu$f. Then, C_1 must be C_{in} divided by 29, or 1 $\mu\mu$f (since capacitive reactance increases as capacitance is decreased).

89. Single-Tube Paraphase Amplifier

Vacuum-tube phase splitters, or *paraphase amplifiers,* are superior to the transformer phase splitters discussed in paragraph 4, because they usually are cheaper than high-quality transformers, and because they can be designed to produce negligible values of frequency, phase, and amplitude distortion. However, it usually is difficult to build paraphase amplifiers which maintain perfect balance, or equality, of the two opposite-going output voltages at all frequencies.

a. In A, figure 66, a single-tube phase splitter, or paraphase amplifier, splits a signal voltage, e_s, into two equal output voltages, e_{o1} and e_{o2}. The positive-going square-wave signal, e_s, causes the plate-current flow to increase across R_K, which results in a positive-going voltage across R_K. Resistor R_K is coupled to the output load resistor, R_{L3}, by capacitor C_3, which has a negligible reactance compared with the resistance of R_{L3} at the lowest signal frequency to be amplified. Thus, the output voltage from the cathode side of the circuit is e_{o2}. In a similar manner, the rising plate current produces a negative-going voltage across R_{L1} which appears as e_{o1} across output load resistor, R_{L2}.

b. The two output voltages, e_{o1} and e_{o2}, are identical in magnitude and opposite in phase, provided resistors R_K and R_{L1} are identical and resistors R_{L2} and R_{L3} also are identical, as in A. Thus, the amplifier tube *V1* acts as a perfect phase splitter in the circuit shown. The balance of this phase splitter falls off at high frequencies. One reason for this is that resistors R_{L1} and R_K are shunted by unequal reactances because of the interelectrode capacitances of the tube, C_{pk} and C_{ko}, shown by dashed lines. R_{L1} is shunted by the reactance of the plate-to-cathode capacitance, C_{pk}; R_K is shunted by the reactance



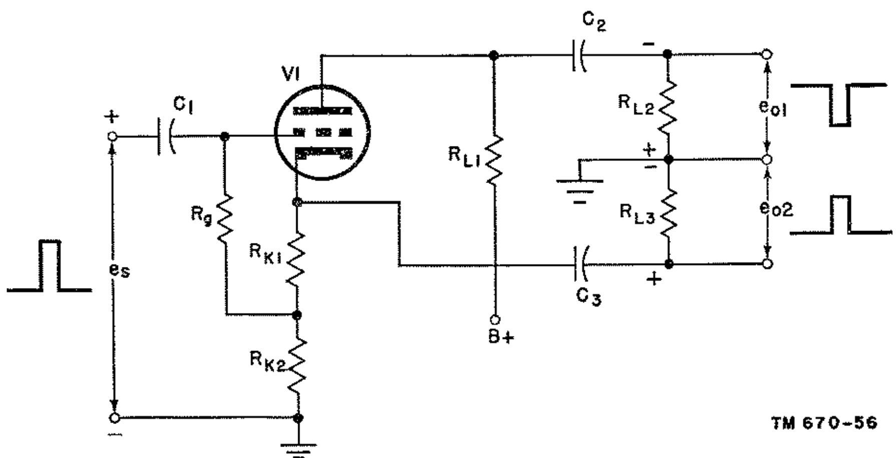


Figure 66. Single-tube paraphase amplifier.

of the cathode-to-ground capacitance, C_{ko}. These two reactances load the resistors unequally, causing unbalanced output voltages at high frequencies. This effect, alone, can be compensated; however, the reactances by themselves are not the only cause of unbalance.

c. Another cause of high-frequency unbalance is that the output voltage, e_{o1}, is derived from the plate of the tube, using a negative-*current* feedback circuit, and e_{o2} is derived from the cathode of the same tube, but using negative-*voltage* feedback. These differences in the type of feedback used are important at high frequencies, where it is necessary to consider the effects of reactances which shunt the two output circuits of the tube. A varying reactance shunting the cathode circuit has a negligible effect on the constancy of the output voltage, e_{o2}, since voltage feedback tends to maintain the output voltage constant, as shown in chapter 4. Output voltage, e_{o1}, however, varies with variations in load impedance. Its equivalent generator tends to maintain a constant load current. As a result, output voltage, e_{o1}, varies with frequency, since the constant load current flows through a parallel combination consisting of R_{L2} and a capacitance whose reactance varies with frequency. It is not practical to compensate for this unbalance in this type of paraphase amplifier.

d. The voltage gain of this paraphase amplifier (defined as e_{o1}/e_s and e_{o2}/e_s) is always less than unity, because the output voltage, e_{o2}, appearing across load resistor R_K of the cathode-follower circuit, is always less than the input voltage, e_s, as explained in chapter 5. Consequently, e_{o1} is also always less than e_s, since the two output circuits are designed to deliver approximately equal voltages.

e. The paraphase amplifier shown in B of figure 66 differs in several respects from the amplifier in A: The d-c grid bias in B is obtained across only portion R_{K1} of the cathode load resistance. The a-c output voltage, however, is taken from the series combination of the two cathode load resistors, R_{K1} and R_{K2}. This circuit is useful when the total d-c drop between cathode and ground exceeds the d-c bias required for proper operation of the tube. Assume, for example, that the d-c drop across the two resistors in B is 100 volts, that the required d-c bias is —10 volts, and that R_{K1} plus R_{K2} must have a combined resistance of 100,000 ohms. The proper value of bias is obtained by using 10,000 ohms for R_{K1} and 90,000 ohms for R_{K2}, since these values of resistance divide the d-c voltage between cathode and ground in a ratio of 1 to 10. Large values of load resistance are used in this circuit to obtain an output voltage as nearly equal to the input voltage as possible.

90. High-Gain Paraphase Amplifier Using Voltage Divider

Figure 67 shows a two-tube paraphase amplifier circuit which provides greater gain than the single-tube paraphase amplifiers.

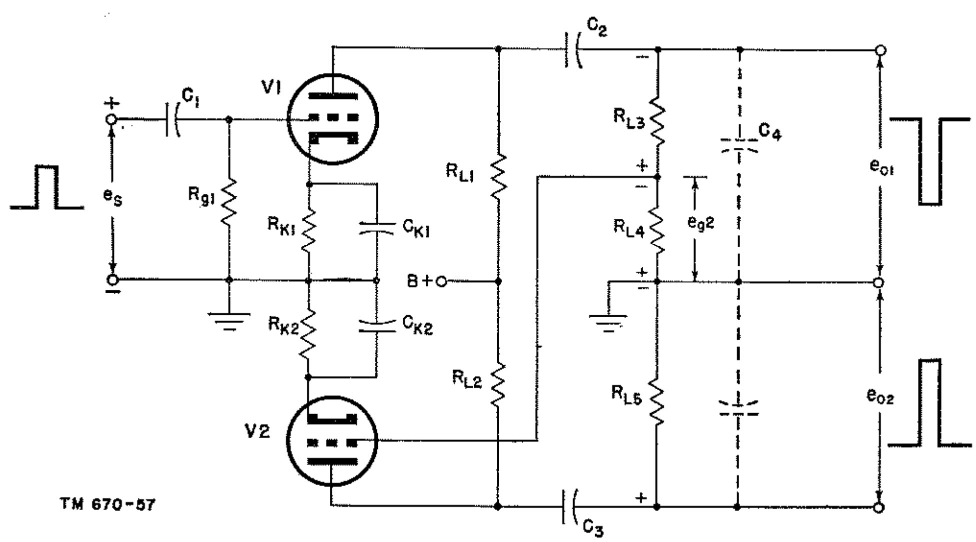


Figure 67. High-gain paraphase amplifier.

a. Tubes $V1$ and $V2$ split the signal input voltage, e_s, into two opposite-going output voltages e_{o1} and e_{o2} (fig. 67). Tube $V1$ acts as a conventional R-C amplifier. Tube $V2$ is a combined phase inverter and amplifier having the same gain as $V1$. $V1$ develops a large, inverted voltage, e_{o1}, across the output terminals when a small signal voltage, e_s, is applied to its grid. The second opposite-going output voltage, e_{o2}, is generated by supplying the grid of $V2$ with a voltage, e_{g2}, a portion of the output of $V1$, which has the same polarity as the output voltage, e_{o1}. That is, a small fraction of e_{o1} is obtained from a voltage divider consisting of R_{L3} in series with R_{L4}. The resistance of R_{L4} is such that the grid voltage, e_{g2}, applied to $V2$, has exactly the same magnitude as grid voltage e_s, applied to $V1$. Under this condition, if the two amplifier circuits are otherwise identical, $V1$ and $V2$ develop two equal output voltages, e_{o1} and e_{o2}, which vary in opposite directions, as required.

b. For a numerical example of the operation of this circuit, assume that $V1$ and $V2$ each has a gain of 30, R_{L5} has a resistance of 30,000 ohms, and the resistance of the combination of R_{L3} plus R_{L4} also is 30,000 ohms. Output voltage e_{o1} is 30 volts, if e_s is 1 volt. Therefore, grid voltage e_{g2} is also 1 volt since the two grid voltages, e_{g1} and e_{g2}, must be identical in order to produce identical output voltages. Thus, voltage e_{g2} across R_{L4} must be $\frac{1}{30}$ of e_{o1}, or 1 volt. Consequently, R_{L4} has a resistance of 1,000 ohms, or $\frac{1}{30}$ of the total load resistance of R_{L3} in series with R_{L4}. This example emphasizes why triodes usually are preferred to pentodes in this circuit. A pentode usually has a high gain, and therefore requires a small input voltage—a requirement which makes the size of R_{L4} too critical for practical applications.

c. This paraphase amplifier has several disadvantages. One disadvantage is that the circuit can be balanced perfectly over only a narrow band of frequencies. This is because phase shift is introduced at the low- and high-frequency ends of the band. For example, at high frequencies, phase shift is caused by the shunt reactances of stray capacitances across the output circuit of $V1$. At low frequencies, phase shift is caused by the series reactance of coupling capacitor C_2. These effects are multiplied in the output of $V2$, because its output circuit also is shunted by stray capacitance, which results in phase shifts at high frequencies. The series reactance of coupling capacitor C_3 results in phase shift at low frequencies. Thus, the phase shift between e_{o1} and e_{o2} is considerable at low and at high frequencies.

d. The output voltage, e_{o2}, has more amplitude distortion than e_{o1}, because tubes $V1$ and $V2$ are connected in series as far as e_{o2} is concerned, as discussed above. Thus, the distortion produced by $V1$ is fed to $V2$, which amplifies and distorts it an additional amount because of its own nonlinear operating characteristics.

91. High-Gain Paraphase Amplifier Using Differential

A second form of the two-tube paraphase amplifier appears in figure 68. Here, the input signal to the phase-inverter section is derived from the differential voltage between the outputs of the two tubes. This circuit provides high gain with reasonable balance of output voltages. Distortion is minimized by omitting the cathode bypass capacitor of $V1$ and introducing degenerative current feedback.

a. To understand the operation of this circuit assume that the two tubes and the output circuits are identical, and that R_{L5} has the same resistance as R_{L3} and R_{L4}. First consider the circuit with tube $V2$ removed, when an output voltage, e_{o1}, is developed across the series combination of resistors R_{L3} and R_{L5} when a signal e_s is applied to the grid of $V1$.

b. When $V2$ now is placed in operation in the circuit, it develops a voltage, e_{o2}, across the series combination of resistor R_{L4} and R_{L5}. The grid voltage for $V2$ is derived from the voltage drop across resistor R_{L5}. This grid voltage is half of e_{o1} when $V1$ operates alone, since R_{L3} is equal to R_{L5}; however, at the instant $V2$ is placed in operation, its output voltage opposes the voltage across R_{L5} and reduces it to a very small value.

c. This happens because the output voltage of $V1$ is of opposite polarity in respect to $V2$. Thus, tube $V1$ develops a voltage, e_{o1}, divided by 2 across R_{L3}, and $V2$ develops a voltage,

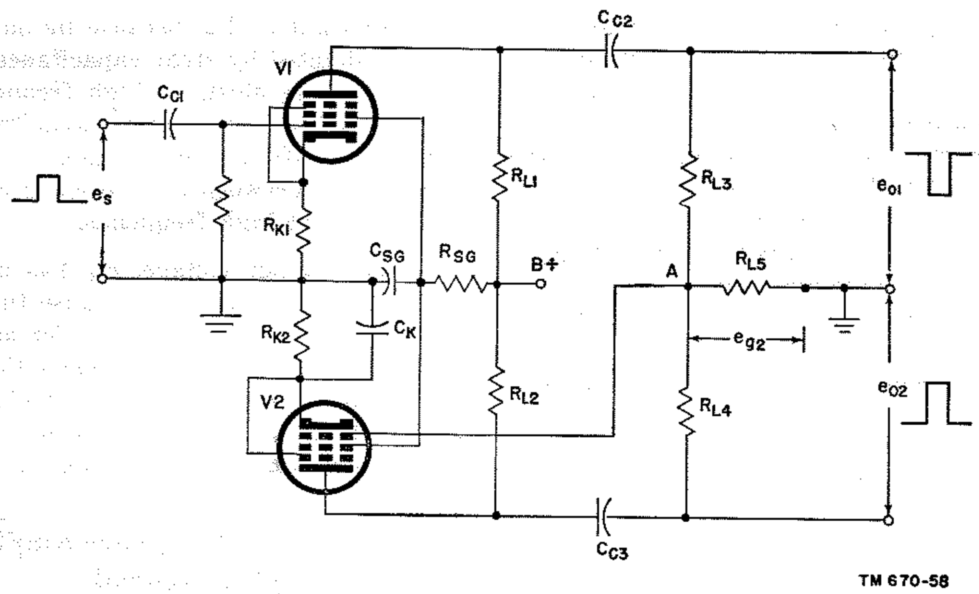


Figure 68. High-gain paraphase amplifier using differential voltage divider.

$-e_{o2}$, divided by 2. (The minus sign indicates opposite polarity between the two voltages.) The grid voltage of $V2$ is the sum of these two voltages. Hence, $e_{g2} = \frac{e_{o1}}{2}$ minus $\frac{e_{o2}}{2}$. To investigate the balance of the output voltages, it is useful to divide both sides of this equation by $e_{o2}/2$. It then reads

$$\frac{e_{o1}}{e_{o2}} - 1 = \frac{e_{g2}}{\frac{e_{o2}}{2}} = \frac{e_{o1}}{e_{o2}} = 1 + \frac{e_{g2}}{\frac{e_{o2}}{2}}$$

d. This equation shows that e_{o1}/e_{o2} is equal to 1, or e_{o1} equals e_{o2}, provided the second term, e_{g2} divided by $e_{o2}/2$, is made to disappear from the equation. Such a mathematical result is obtained if e_{g2} is made 0 volt. However, this is physically impossible, since tube $V2$ requires the grid voltage, e_{g2}, in order to develop an output voltage, e_{o2}. As a compromise, e_{g2} is made very small so that e_{o1}/e_{o2} is almost equal to 1. For this reason, it is necessary to use high-gain tubes (usually pentodes) in this amplifier, to develop from a very small voltage, e_{g2}, a large inverted output voltage, e_{o2}. The output of $V1$ usually is made slightly larger than the output of $V2$, since the effective voltage at the grid of $V2$ is the difference between the outputs of $V1$ and $V2$.

92. Cathode-Coupled Paraphase Amplifier

A third form of paraphase amplifier is shown in figure 69, together with the waveforms which result when a positive-going square wave is applied to the grid of tube $V1$. The circuit consists of two tubes. Tube $V1$ is used to increase the amplitude of the applied signal to the desired level. Tube $V2$ is used as an inverter and amplifier to produce a signal of the same amplitude as the output of $V1$, but of opposite polarity. This circuit provides moderate gain and good balance.

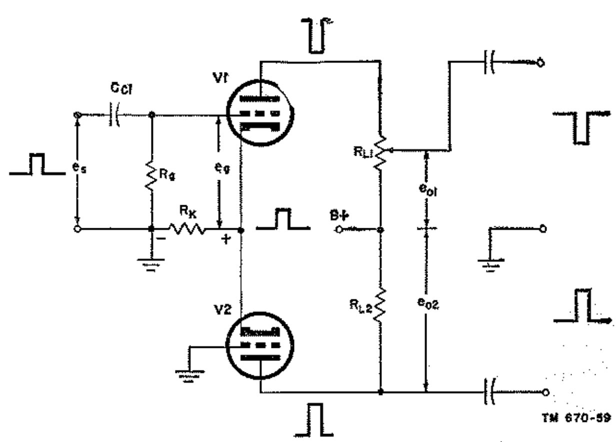


Figure 69. Cathode-coupled paraphase amplifier.

a. In figure 69, tube $V1$ is a degenerative amplifier, using an unbypassed cathode resistor, R_K, which carries the total a-c and d-c plate currents of both tubes, $V1$ and $V2$. Thus, a change in plate current caused by one tube is reflected immediately into the grid circuit of both tubes, since R_K is connected in the common cathode circuit of both tubes. The value of the cathode resistor is such that it reduces the grid-to-cathode voltage of $V1$ to a value which is approximately half the input signal voltage, or $e_s/2$. Therefore, about half of the input signal voltage, $e_s/2$, also appears across R_K. As a result, the a-c voltage developed between the cathode and a ground reference point is in phase with voltage e_g appearing between grid and cathode of $V1$, because an increase in grid voltage increases the plate current, thus making the cathode more positive in respect to ground. Conversely, the cathode voltage decreases as the grid voltage decreases.

b. The voltage developed across R_K is applied directly between the grid and cathode of tube $V2$. Consequently, $V2$ and $V1$ have voltages applied to each grid which are almost exactly equal in magnitude and opposite in polarity. Each tube amplifies these voltages. The resulting two output voltages, e_{o1} and e_{o2}, are of opposite polarity—as is required of a paraphase amplifier. Mathematical analysis and experiment show that e_{o1} is usually larger than e_{o2}, however, because the a-c plate current in $V1$ usually is larger than the a-c plate current in $V2$. Therefore, the output of $V1$ exceeds the output of $V2$. For this reason, R_{L1} is tapped or made variable, as shown, and its value is adjusted so that e_{o1} equals e_{o2}.

c. The chief advantages of this paraphase amplifier are its circuit simplicity, balance, and moderate gain. At high frequencies, however, it suffers from the unbalanced effects common to the paraphase amplifiers that have been discussed. Primarily, unbalance occurs because equal shunting reactances across the loads have unequal effects because $V1$ uses current feedback and $V2$ does not. A similar effect was discussed in paragraph 6.

93. Summary

a. A phase inverter is a device which inverts the polarity of a signal. It may be a transformer or a vacuum-tube circuit.

b. A phase splitter splits an input voltage into two output voltages which vary in opposite directions. Transformer phase splitters are used for low-frequency applications. Vacuum-tube phase splitters, also called paraphase amplifiers, are used for wide-band applications. The output voltage of single-tube phase splitters is less than the input voltage. The output voltage of two-tube paraphase amplifiers usually is much greater than the input voltage.

c. Three types of paraphase inverters are the single-tube type, the high-gain type, and the cathode-coupled type. Compared with the transformer, paraphase amplifiers are superior because they isolate input and output stages and provide reasonably well-balanced, amplified output voltages over a wide frequency range. Paraphase amplifiers are more complicated to operate than the transformer type of amplifier; however, they require a power supply, and the output voltages are unbalanced at very low and very high frequencies.

94. Review Questions

a. Briefly describe (1) a phase inverter and (2) a paraphase amplifier.

b. How is a phase inverter used?

c. How is a phase splitter used?

d. Compare transformer phase inverters with vacuum-tube inverters.

e. Sketch two simple circuits of phase inverters using (1) transformers and (2) tubes.

f. Can a transformer be used to invert a square wave? Explain.

g. For what type of operation is a transformer inverter particularly well suited?

h. Why is a step-down ratio used in a transformer-coupled stage operating in Class B?

i. Explain why triodes are preferred in transformer-coupled inverters.

j. What are the advantages and disadvantages of the transformer phase splitter?

k. Why is it necessary to use a carefully designed center-tapped transformer in a phase-splitter circuit?

l. Sketch circuits of phase inverters of simple type and unity-gain type.

m. Compare vacuum-tube phase inverters with transformer inverters.

n. What are the advantages and disadvantages of vacuum-tube phase inverters?

o. Sketch circuits of the following paraphase amplifiers: single-tube type; high-gain type using voltage divider; high-gain type using differential voltage divider; cathode-coupled type.

p. Compare the advantages and disadvantages of the phase inverters of question *o*. Consider particularly balance, gain, and circuit simplicity.

CHAPTER 7

SOUND SYSTEM AMPLIFIERS

95. Sound-System Fundamentals

a. Sound systems which amplify weak sounds so that they can be heard by large numbers of people have many applications. In the home, they amplify phonograph recordings so that they may be heard with concert-hall loudness; at meetings, they make announcements audible to all the spectators; on parade grounds and in the field they amplify instructions or commands that must be heard over a wide area.

b. (1) The level of signals derived from such sources as microphones, phonograph pickups, or radio tuners is raised by the amplifier to a level suitable for driving one or more loudspeakers. The number of stages required primarily depends on the voltage output of the sound source (microphone, phonograph pickup, or radio tuner). The output power required of a sound system depends on the type, number, size, and efficiency of the loudspeakers to be driven. This, in turn, depends on the size and dispersion of the audience and the area to be covered. Other factors which affect the power requirements are the noise level in the area, and the nature of sound—that is, whether it is speech or music.

(2) The input levels from phonograph pickups and radio tuners vary from 10 millivolts to several volts. The input levels of amplifiers vary between 30 microvolts and 10 millivolts, depending on the type of microphone for which the amplifier is designed. The output powers of sound-system amplifiers range from about 1 watt up to 100 watts, depending primarily on the size of the audience and the area to be covered. To supply a comfortable listening level in a room of 25 by 25 feet (625 square feet) with an average noise level, a 4-watt amplifier and a 10-inch wall speaker would be sufficient, but a much more powerful amplifier system would be required to supply a comfortable listening level on an outdoor athletic field or parade ground. A typical outdoor area of 50,000 square feet would require a 75-watt amplifier and a group of six loudspeakers of re-entrant type.

c. Sound-system amplifiers and their batteries or power supplies usually are mounted in convenient, portable carrying cases (figs. 70, 71, 72). The portable sound system for military applications shown in figure 70 consists of a 1.5-watt amplifier, a microphone-loudspeaker combination, and batteries. It delivers 100 percent speech intelligibility at a range of 600 feet, and is used to address small groups of personnel, usually outdoors. The public-address amplifier of figure 71 has a dual 4-stage resistance-coupled circuit which uses 14 tubes and has an output of 50 watts. The amplifier is part of a complete sound system, used to cover large areas. Associated equipment includes a microphone, a microphone stand, cables, a re-entrant horn speaker, and a transformer box assembly. Figure 72 shows a complete portable public-address system consisting of two amplifiers, six loudspeakers, two stands, a phonograph, an all-wave radio tuner, two microphones, and accessories. Each amplifier uses 11 tubes and can operate simultaneously from one or two microphones, a phonograph, and a radio tuner. The system is used for amplifying speech or music, indoors or outdoors. Three practical sound-system amplifiers will be discussed in detail in the following paragraphs. Their different designs satisfy different operating requirements.

Figure 70. Public address set.

96. Four-Watt Amplifier
(fig. 73)

a. The block and circuit diagrams of this sound-system amplifier used for home listening will deliver 4 watts of audio power to a loudspeaker. The amplifier consists of three stages, and operates from a crystal-microphone input. Output terminals are provided so that the output power may be fed to an 8-ohm speaker or to a 500-ohm transmission line. The amplifier is designed to have a flat frequency response from 20 cycles per second to 10 kilocycles per second, with a total harmonic distortion of 8 percent at 4 watts output.

b. The amplifier requires three stages of amplification to increase the feeble voltage from a crystal microphone to a power level of 4 watts. The successive a-c signal voltages at each stage are shown in the block diagram, in A. The last stage consists of a single-ended 6V6 beam pentode transformer-coupled to the load. The secondary of the output transformer delivers 4 watts to the load at either 8 ohms or 500 ohms, and the 6V6 supplies 4.5 watts to the transformer primary. The .5-watt difference is lost in the resistance of the transformer windings. The a-c signal voltage required at the amplifier input terminals is .0011 volt. It is found by dividing the output voltage of the 6V6 by the over-all voltage gain of the amplifier. The output voltage of the 6V6 tube for 4.5 watts input to the transformer primary is 151 volts. The over-all voltage gain of the amplifier is the product of the voltage gain of the individual stages, or 17.3 times 50 times 160, or 138,400. Thus, the required input voltage is 151 volts divided by 138,400, or .0011 volt. This input voltage is readily obtained from a crystal microphone.

c. The circuit diagram of this 4-watt amplifier is shown in A. The individual stages are similar in design to those discussed in the earlier chapters of this manual. The amplifier system has additional needed features, however —a tone control R1–C1, and plate-decoupling filters R2–C2 and R3–C3. The operation of these circuits is explained in the following paragraphs.

97. Tone Control

Sometimes the audio signal contains disturbing high-frequency noise components which

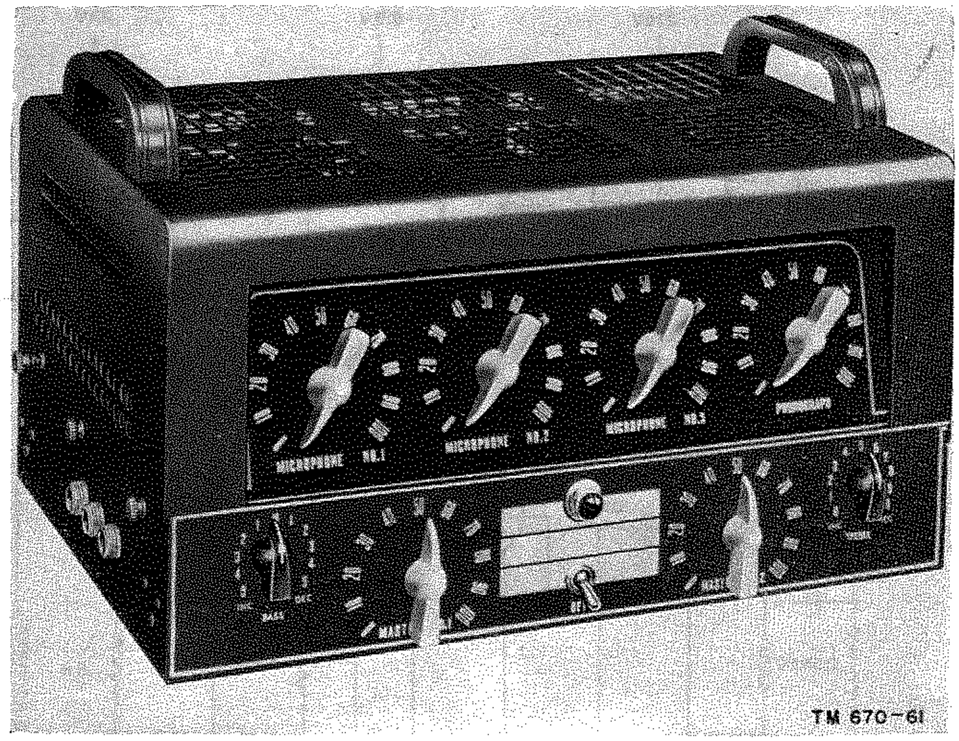


Figure 71. Typical amplifier.

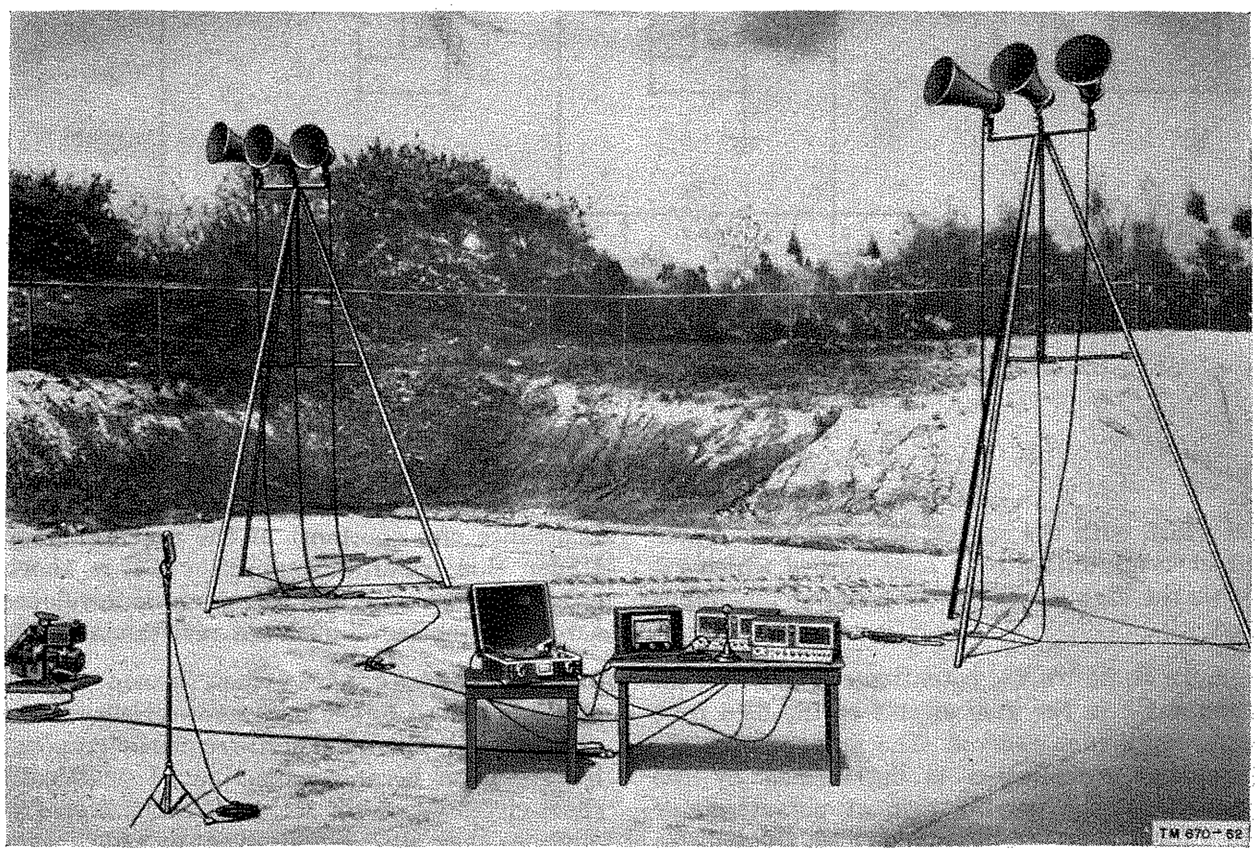


Figure 72. Typical public address set.

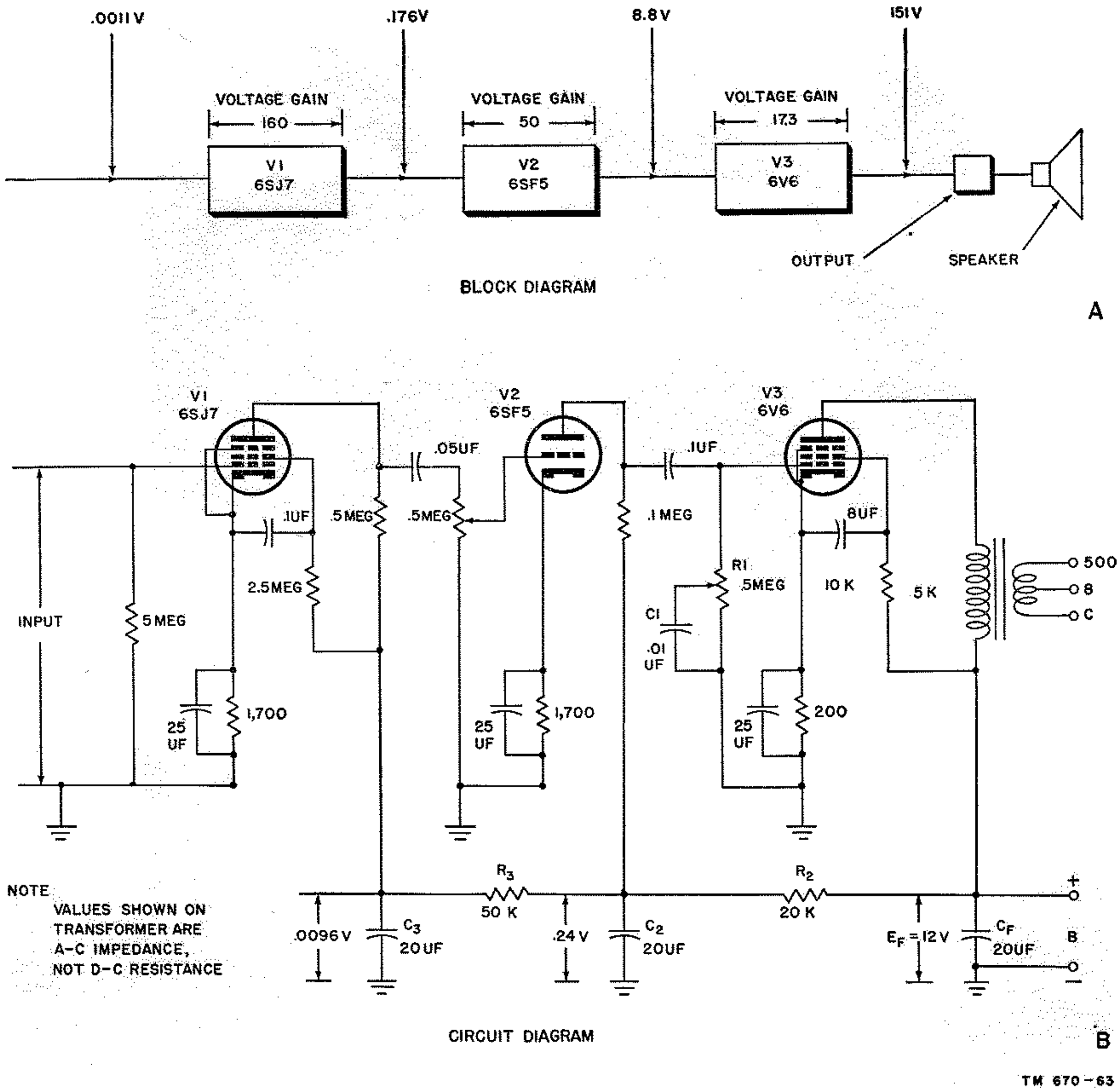


Figure 73. Single-ended power amplifier.

interfere with the speech or the music. Their effect is minimized by a *tone control* which reduces the high-frequency response of the amplifier.

a. In B, figure 73, potentiometer *R*1 and capacitor *C*1 in the grid circuit of the 6V6 tube make up a *tone control.* The amount of high-frequency attenuation is determined by the position of the moving arm of potentiometer *R*1. There is no attenuation of high frequencies when the moving arm is at the maximum-resistance end (bottom) of the potentiometer. Under this condition, capacitor *C*1 is short-circuited to ground and the signal is developed across the entire .5-megohm resistance of *R*1.

b. The opposite is true when the contact arm of *R*1 is moved to the minimum-resistance end (top) of *R*1. Then, the high-frequency response is very poor, because capacitor *C*1 practically short-circuits resistor *R*1, so that the high frequencies are attenuated over a wide range simply by changing the adjustment of *R*1. To the listener, the low frequencies are stronger because the high frequencies are attenuated. Hence, it appears as if the low frequencies are intensified, or *boosted*.

c. The tone control attenuates both the unwanted high-frequency noise components, and the higher harmonics of the speech or music itself. *Thus, it suppresses noise only at the expense of fidelity.*

98. Decoupling Filter

The purpose of a decoupling filter in a multistage amplifier is to prevent undesired feedback from later to earlier stages. Such a feedback might result in erratic performance or oscillation.

a. (1) The actual frequency response of a practical, multistage amplifier often differs considerably from its calculated response. For example, the calculated output voltage of an amplifier at 20 cycles may be 70 volts, but the output voltage of a practical amplifier built from this design may be either much higher or much lower. This is caused by undesirable feedback voltage which may be either positive or negative.

(2) Under certain circumstances (ch. 8) an amplifier may *oscillate*—that is, it may produce an output voltage at some specific frequency—even after the input voltage is removed. Such erratic frequency response is undesirable, since it is important that the amplifier gain be uniform over a wide range of frequencies.

(3) The undesired feedback voltage that causes this erratic response is coupled from the last amplifier stage into the preceding stages by an impedance common to all the stages of the system. This common impedance is the reactance of the output capacitor in the power-supply filter circuit and is appreciable at low audio frequencies. Figure 74 shows the location of this capacitance C_F in a practical circuit.

b. (1) In figure 74, A shows a simplified diagram of the last stage of the amplifier shown in B, figure 73. The plate load of the tube is the combination of the primary of the output transformer in series with the capacitor. Capacitor C_F is the output element of the power-supply filter used to smooth out the a-c ripple in the d-c power supply. An undesired feedback voltage is produced as follows: The 6V6 tube delivers an a-c output voltage, E_o, to its load (fig. 74). This voltage is divided into two unequal parts—a large voltage, E_P, across the primary of the output transformer (large reactance), and a small voltage, E_F, across capacitor C_F (small reactance). For example, assume that the a-c voltage across the primary of the output transformer is 140 volts at 20 cycles, and the voltage across capacitor C_F is 12 volts. This small voltage is unimportant for a single stage of amplification, since it is considerably less than the output voltage.

(2) The effects of this feedback voltage are serious, however, when there is more than one stage in the amplifier. It is especially important to prevent feedback from the last stage to the first, because then a voltage fed back will undergo the greatest amplification. One way of preventing this is to use a separate power supply for the first stage. A portion of a cascade amplifier in which this has been done is shown in B, figure 74. The effect of the voltage, E_F, across capacitor C_F is negligible, and there is no simple way to feed back to the first stage. Separate power supplies are expensive and heavy, however, and they duplicate the first power transformer, tubes, and filter choke.

c. In C, figure 74, the same d-c power supply is used for all the amplifier stages. The voltage across capacitor C_F now appears in series with the load resistor of the first stage. The effective output voltage of the first tube is altered by a large value, because its output voltage now consists of two unequal parts. For example, if, in B, figure 74, the a-c output voltage of the first

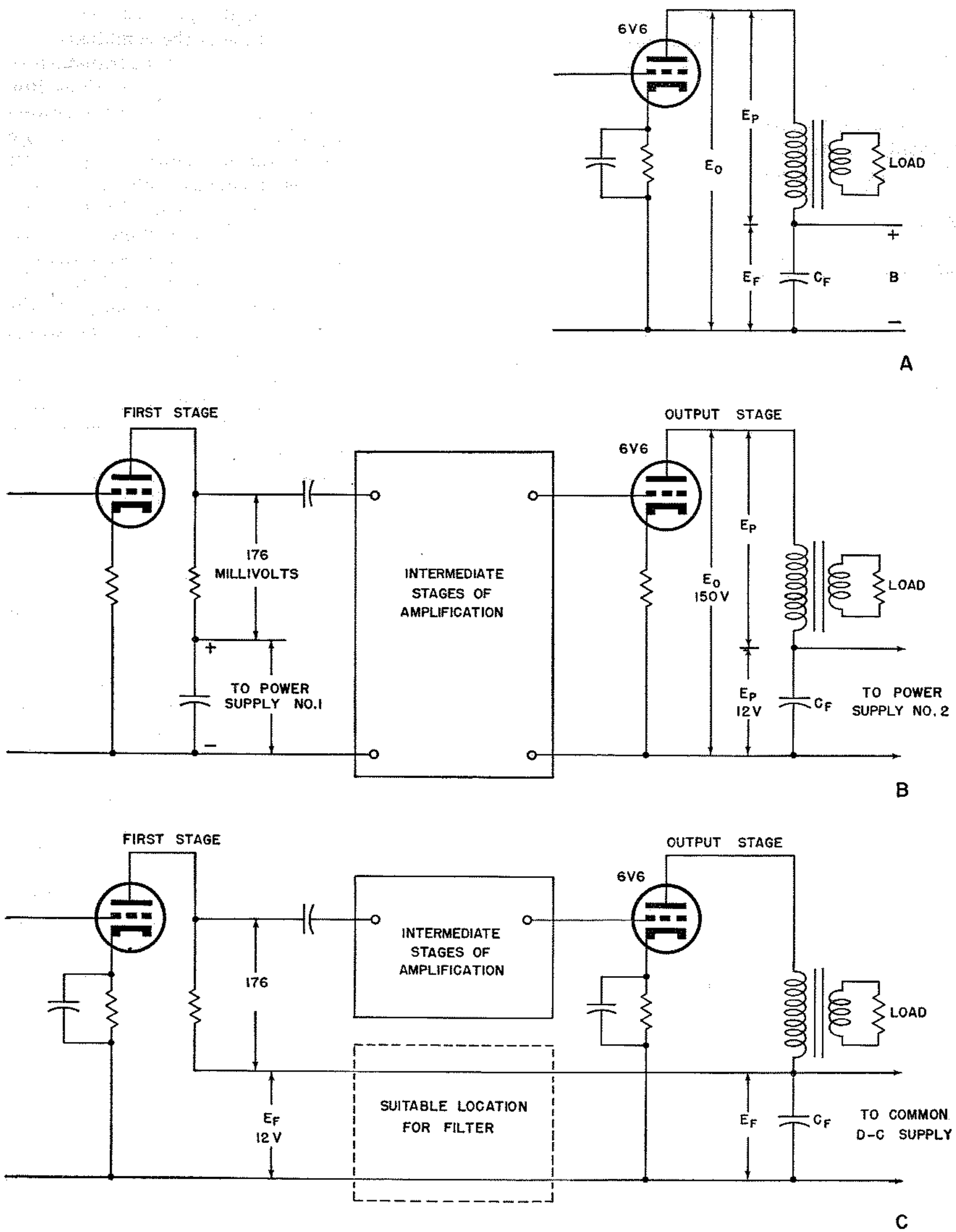


Figure 74. Common impedance feedback.

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stage is 176 millivolts, in C the output voltage of 176 millivolts is either increased or decreased by 12 volts—68 times the proper value. The mode in which the two voltages add or subtract depends on the phase shift in the intermediate stages of the amplifier. The actual response is considerably different from the calculated response at the low frequencies, since the reactance of capacitor C_F at these frequencies is large compared with the impedance of the primary of the output transformer. This type of feedback is a result of a plate impedance which several stages have in common, and is called *common impedance feedback.*

d. This feedback can be minimized by several methods. One method, in B, figure 74, employs separate power supplies and is used only in extreme cases, when the feedback cannot be eliminated by other methods. A better way to eliminate feedback effects is to insert filters between the common impedance and the earlier amplifier stages. The dashed-line rectangle in C, shows the location of such a filter for the first stage of an amplifier.

99. Filters

a. An electrical filter is a device which permits a band of frequencies, known as the *pass band,* to go through it with little or no loss, but attenuates all frequencies outside the pass band.

(1) In figure 75, A shows a simple *low-pass* filter which passes only very low frequencies. It is inclosed in a dashed-line rectangle.

(2) The filter in B, figure 75, is redrawn as a voltage divider consisting of inductor L connected in series with capacitor C. To understand its operation, assume that at 20 cps the reactance of L is 50 times as great as that of C. An a-c voltage applied to the input terminals of the filter is reduced by a factor of approximately 50, so that at 20 cps an output of .24 volt is obtained from an input of 12 volts.

(3) An advantage of this filter is the improved filtering at high frequencies. As the frequency increases, the a-c voltage drop across the filter inductor

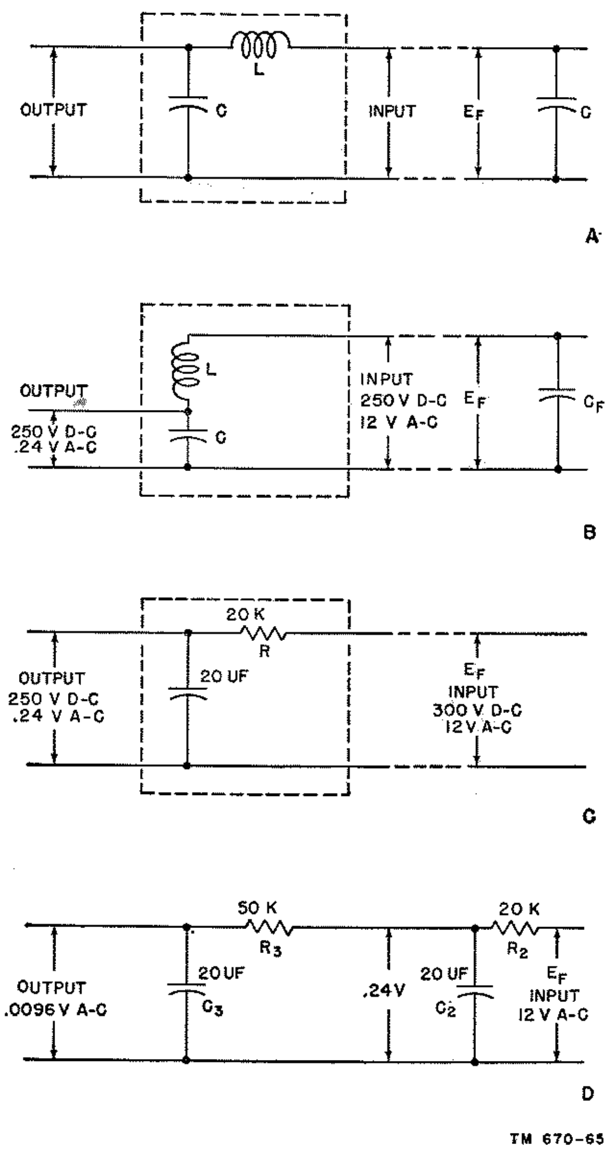


Figure 75. Low-pass filters.

increases, and that across the filter capacitor decreases. The a-c output voltage of the filter drops to a very small value at high frequencies.

(4) Another advantage is the low d-c voltage drop in inductor L. A large a-c voltage and a large d-c voltage may be applied to the input of the filter simultaneously. When this is done, the large a-c input voltage is reduced to a very small a-c output voltage; but, because of the low d-c resistance of L, the value of d-c output voltage is practically the same as that of the d-c input voltage.

The d-c voltage can serve as a supply voltage for several amplifier stages, thereby minimizing common impedance feedback. A numerical example of a-c and d-c values in such a filter is shown in B. The disadvantage of this type of L-C filter is the cost and large size of the inductor.

b. (1) The simpler filter in C, figure 75, employs a resistor in place of inductor *L*. Such a filter is known as an R-C filter. The resistance of *R* is the same as the reactance of *L* at 20 cps—in this case, 20,000 ohms. The filtering action of both filters at 20 cps is essentially the same.

(2) This R-C filter has several disadvantages. One is that the d-c voltage drop in the resistor is much larger than that in a practical inductor. If the d-c output voltage of the R-C filter is to be the same as that of the L-C filter, a larger d-c supply voltage is needed at its input terminals. For example, in *C*, the d-c output voltage required of the filter is 250 volts. Because the voltage drop in *R* is 50 volts, the d-c input voltage is 300 volts. Another disadvantage is that its filtering action over a band of frequencies is poorer than that of the L-C filter. This is so because only the capacitive reactance changes with frequency, but the series resistance remains constant at all frequencies. In the L-C filter the filtering action is more pronounced because the reactances of both elements vary in opposite directions as the frequency changes. This disadvantage is minimized by using several R-C filters in cascade, as shown in D.

c. (1) A numerical example will demonstrate the effectiveness of the two-section filter. In D, figure 75, the filter of the sound-system amplifier shown in B, figure 73, has been redrawn. The feedback voltage, E_F, is 12 volts at 20 cps and the a-c output voltage of the R2-C2 filter is .24 volt. This is negligible compared with the 8.8 volts a-c output of the second amplifier stage. The first stage of filter R2-C2 is effective in reducing the feedback. Similarly, the a-c output voltage of the R3-C3 filter section is .0096 volt, or 9.6 millivolts. This voltage is negligibly small compared with the 176 millivolts a-c output of the first amplifier stage.

(2) The two-section filter minimizes the effects of common impedance feedback at 20 cps. The feedback is completely negligible at frequencies above 20 cps, since the reactances of C2 and C3 decrease as the frequency increases.

100. Sound System, Push-Pull-Ended

Sound system amplifiers should produce large amounts of audio power and negligible amounts of distortion. These requirements often conflict, since the distortion in a power amplifier usually rises as the power output increases. This distortion can be minimized in a single-ended power amplifier (B, fig. 73) by operating it at an output level below its maximum capability. For example, distortion effects can be minimized if the 4-watt amplifier is operated at an output of about 2 watts. A more efficient method for reducing distortion is to use an amplifier with a push-pull output stage (fig. 76), which is superior to the single-ended amplifier shown in B, figure 73. The total output of the two tubes connected in push-pull is more than twice the power output of a single stage using a tube of the same type, and, in addition, the distortion is reduced. A brief review of push-pull amplification follows. For a detailed discussion of the principle and advantages see TM 11–662.

a. Advantages of Push-Pull Amplifier. The push-pull connection has many advantages. They include greater power output, lower distortion, improved low-frequency response, less hum, and simpler circuits.

(1) The power output is more than twice that obtained from a single-ended stage using a tube of the same type, because it is possible to use a higher grid drive and lower plate-load impedance. In a single-ended amplifier, most of the distortion results from second and higher even harmonics, but these are balanced out in the push-pull am-

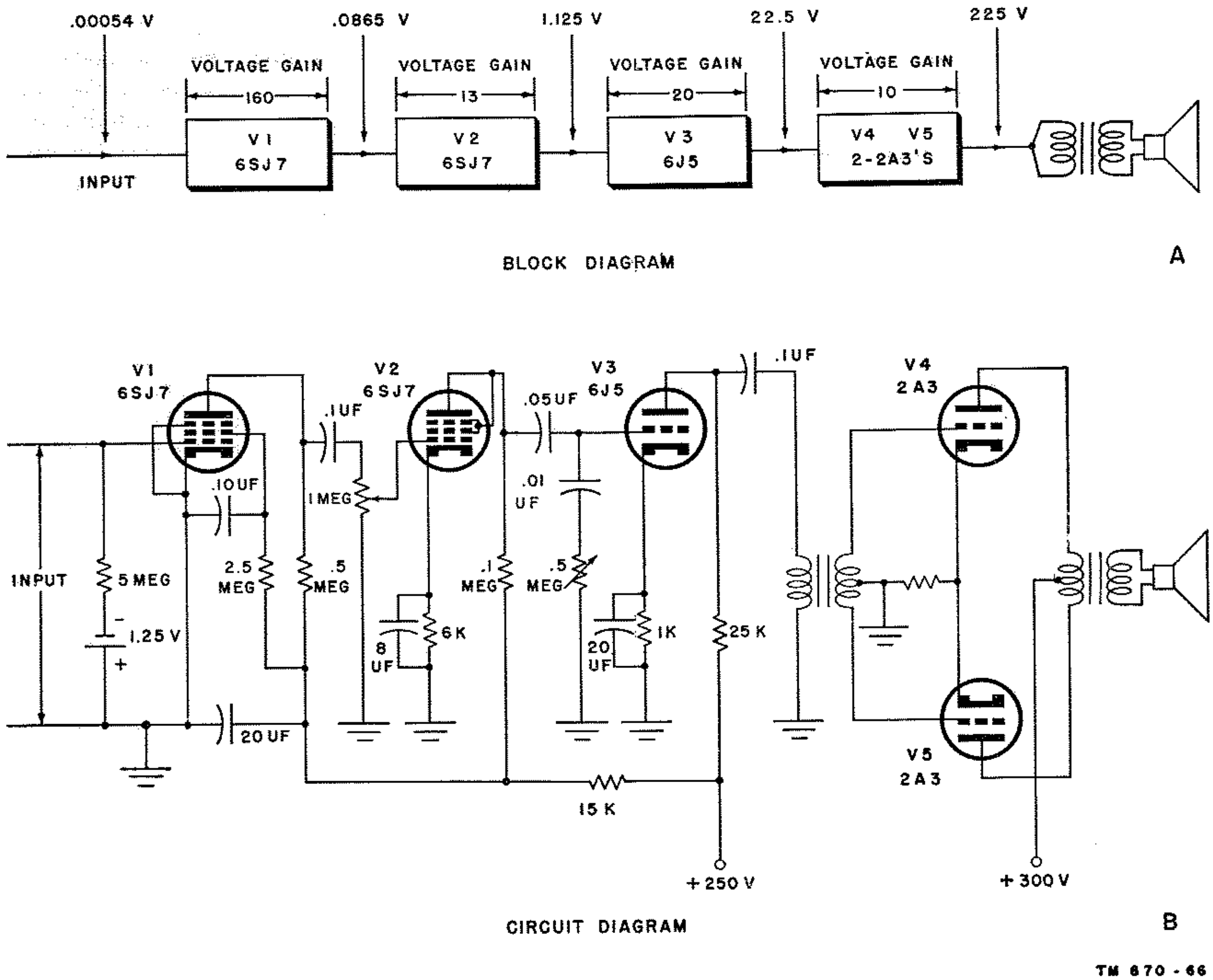


Figure 76. Push-pull-ended power amplifier.

plifier and do not appear at the loudspeaker.

(2) The low-frequency response of a push-pull stage is better than that of a single-ended stage. The d-c flows to the tubes through the primary windings in opposite directions. This produces opposing magnetic fields in the output transformer, which cancel each other. The output transformer, therefore, is not saturated by the d-c and has a high primary inductance.

(3) A-c hum may cause an objectionable interference in the signal output of an amplifier. A-c ripple in the d-c power supply causes less a-c hum in a push-pull amplifier than in a single-ended amplifier. The two push-pull tubes are supplied with the same d-c through the common center tap on the primary of the output transformer. The polarity of any a-c ripple in the d-c supply is the same on both plates, at opposite ends of the transformer, and so the a-c hum is balanced out and does not appear at the loudspeaker.

(4) The cathode bias bypass capacitor can be omitted in push-pull amplifiers, if the tubes are matched and the transformers have good electrical balance. A single cathode bias resistor supplies bias for both tubes. Only the even harmonics of plate current flow in the cathode circuit, and these have no de-

generative effect on the frequency response of the amplifier because the output contains only odd harmonics. When the circuit is not perfectly balanced, however, it is necessary to use individual cathode resistors and bypass capacitors for each tube.

b. Application of Push-Pull Amplifier.

(1) A of figure 76, is a block diagram of an amplifier with a push-pull output stage. The voltage gain of each of its four stages is shown over the block representing the stage. The amplifier is suitable for a public-address system for reproducing speech or music. When operating from a crystal microphone, the maximum output power obtainable at full volume is 10 watts at 5-percent total distortion. The input and output voltages existing when the amplifier is operating at full volume are shown stage by stage. With the tone control (not shown) at maximum for high-frequency response, the amplifier is flat from 20 to 10,000 cps.

(2) In the circuit diagram of the same amplifier, shown in B, the power stage consists of two 2A3 power triodes connected in push-pull. This stage requires 22.5 volts a-c driving voltage at the primary of its input transformer. This is three times the voltage required by the 6V6 beam pentode in the single-ended power amplifier of figure 73. The 22.5 volts is supplied by a 6J5 triode stage having a voltage gain of 20, and requiring an input voltage of 1.125 volts. Because this is more than 1,000 times the output of a crystal microphone, and one tube cannot produce a gain of 1,000, at least two stages of amplification are needed. Of these, the first is *V*1, a 6SJ7 high-gain pentode stage having a gain of 160, which amplifies the feeble output of the microphone to .0865 volt, as in A. The stage is designed to minimize hum pickup and noise. An additional gain of 13 is needed to supply the voltage required by the 6J5; therefore *V*2, a second 6SJ7, is used, connected as a triode in B. Triodes are used in the second and third stages for several reasons. For one, the low plate resistance of the triode is needed for proper coupling to the push-pull input transformer. For another, triodes give better low-frequency response, as explained in chapter 2.

c. Special Features of Push-Pull Amplifier.

(1) The sound-system amplifier circuit in B has several special features. There is a tone control in the grid circuit of *V*3, the 6J5 triode. There is also a plate-decoupling filter. The operation of these circuits is the same as that of the corresponding circuits of the single-ended amplifier, discussed in paragraphs 97, 98, and 99.

(2) Another feature is the *bias cell* used for the grid bias of the first stage. This is a specially built bias battery having an output of about 1.25 volts, provided no grid current flows through it. The cell is used to avoid the possibility of hum amplification, which is produced in low-level amplifier stages using cathode bias. The hum results from a-c ripple caused by incomplete filtering in the d-c plate supply, which produces a small amount of ripple voltage across the cathode resistor. This ripple voltage is negligible in the high-level stages, but it must be avoided in the low-level stages, where the small ripple voltage and the signal have about the same value.

(3) Another feature in this amplifier is the shunt, or parallel, feed used to couple the 6J5 triode to the push-pull input transformer. With this circuit, the primary inductance of the transformer is increased, which results in an improved low-frequency response. It is effected by inserting a .1-μf capacitor between the plate of the tube and the transformer primary. This keeps the d-c out of the transformer; the transformer is not saturated, and has a higher primary inductance. The direct current required by the tube is supplied through a 25,000-ohm resistor

connected in shunt, or parallel, with the capacitor-transformer combination. The value of the capacitor is so chosen that it resonates with the transformer primary at a low frequency, at which the effect is not objectionable.

101. Fifteen-Watt Amplifier

The circuit shown in figure 77 is considered a high-fidelity amplifier, because of the unusual fidelity with which it reproduces musical sounds. It has an excellent low-frequency response and also reproduces with high fidelity signals of a transient nature, such as staccato blasts from wind instruments and sudden bursts from drums. It is designed to amplify an audio signal from the detector of a radio receiver, or from a magnetic pickup and pre-amplifier, or a crystal pickup. The circuit has several unusual features which distinguish it from conventional sound-system amplifiers.

a. Block Diagram. The block diagram of the amplifier (A, fig. 77) shows several of its unusual features. The power stage consists of *V*3 and *V*4, two 807 tubes operating in push-pull. Instead of being driven by a single driver through an input transformer, however, each 807 tube is driven by half of *V*2, a 6SN7 double triode. The three tubes constitute two push-pull stages in cascade. *V*2 is a push-pull driver, its two halves operating as separate triodes. They receive their input from *V*1, another 6SN7 tube, the two halves of which are coupled to supply a balanced signal, as required for the push-pull input. This tube operates as a phase splitter, but in a manner somewhat different from the phase splitters described in chapter 6. Its circuit operation will be explained in *b* below. The magnitudes of the voltages at the input of each stage are given in A, along with waveforms of their phase relations. Feedback is provided from the secondary of the output transformer to the input of the first stage, as shown. With an input signal of 1.75 volts, the amplifier supplies 15 watts to a 16-ohm speaker.

b. Phase Splitter. To clarify circuit operation, figure 78 shows a schematic diagram of the phase splitter, redrawn from the complete schematic diagram of B in figure 77. The two halves of *V*1 are operated as independent triodes, directly coupled as shown, without a coupling capacitor; the usual attenuation at low frequencies caused by coupling capacitors thus is avoided. The plate of the first triode is connected directly to the grid of the second. This is feasible, without using an unusually large plate-supply voltage, by limiting, the d-c plate voltage of the first stage to 100 volts. By placing the cathode of the second triode 105 volts above ground, the required —5 volts bias on the second stage is provided. This value of bias is chosen for minimum distortion. The output is taken across the second triode, with coupling capacitors *C*2 and *C*3 inserted at both the plate and cathode. Both sides of the output are thus isolated from ground, providing the balanced push-pull output required. This signal comprises two a-c voltages of 6 volts each of opposite polarity, or 12 volts from cathode to plate of the second stage.

c. Complete Amplifier Circuit. The phase splitter of figure 78 is connected to the rest of the amplifier circuit as shown in B, figure 77. The components appearing in both diagrams are similarly identified.

(1) The complete circuit in B requires only four coupling capacitors—*C*1, *C*2, *C*3, and *C*7—which results in excellent low-frequency response. Each pair of coupling capacitors feeds the signal to a center-tapped voltage divider—*R*6 and *R*7 at the output of *V*1, and *R*9 and *R*10 at the output of *V*2. This serves to maintain the push-pull voltages equal and opposite in respect to ground.

(2) The power output stage uses two 807 tubes. These are employed normally as transmitting tubes. They are used here because of their high power output and power sensitivity, and because, unlike many transmitting tubes, they employ indirectly heated cathodes. The cathode currents of power output tubes *V*3 and *V*4 must be balanced before placing the amplifier in operation. They are measured with a milliammeter inserted first in jack *J*1 and then in jack *J*2. Grid-bias potentiometer *P*1 is adjusted so that the cathode currents are equal. The cathode cur-

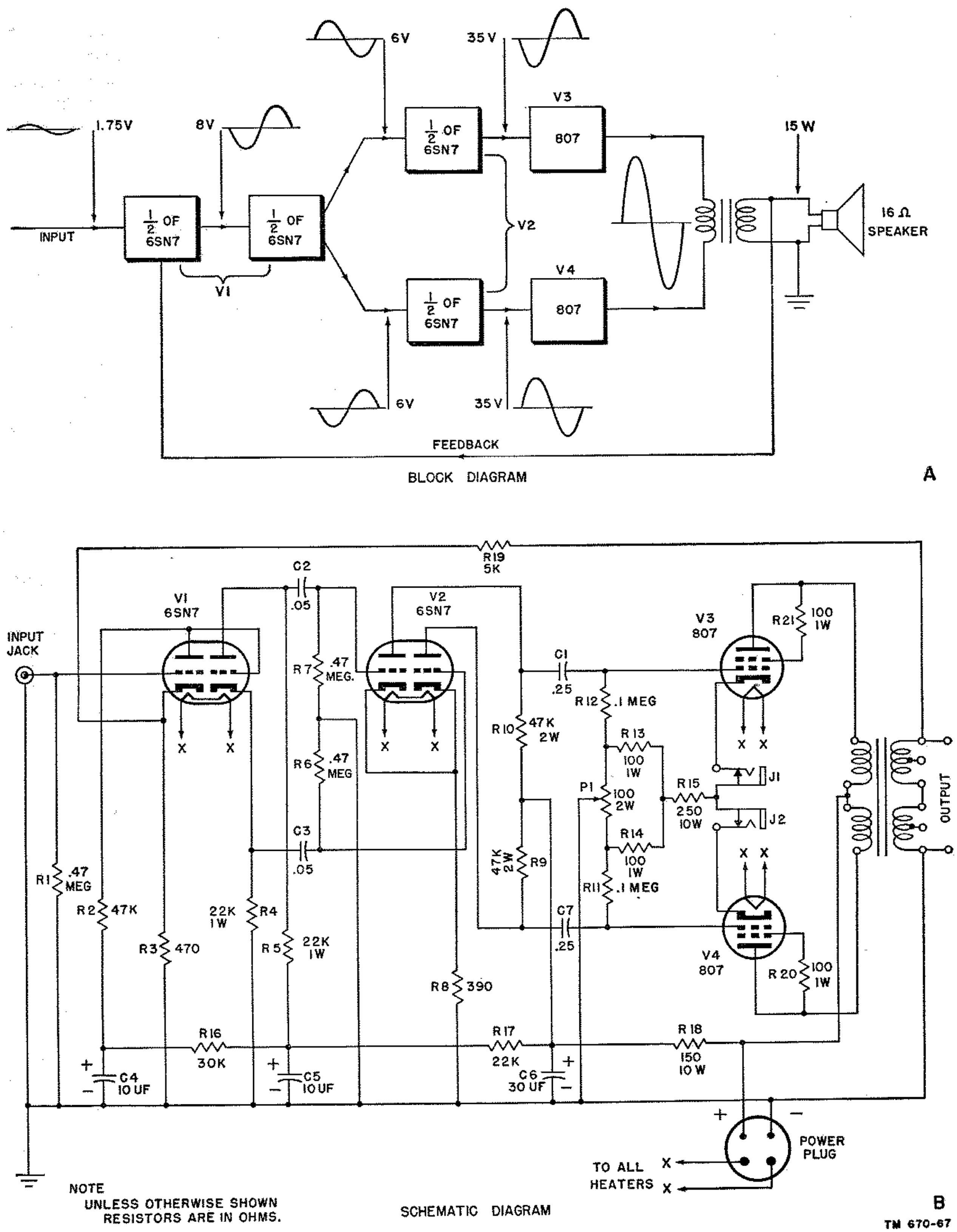


Figure 77. Fifteen-watt amplifier.

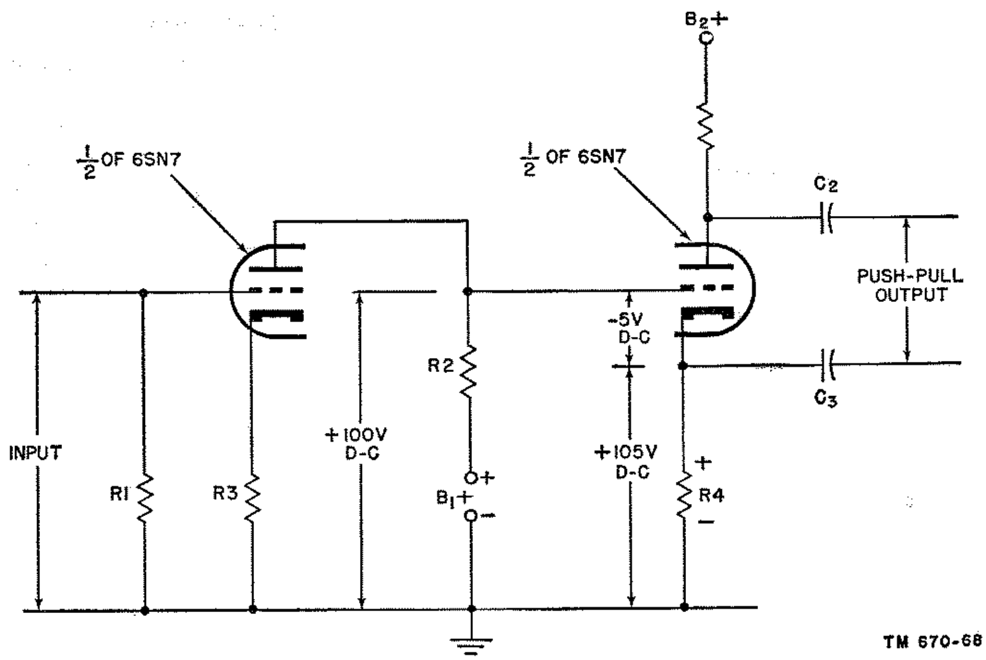


Figure 78. Phase splitter.

rent in each tube should be approximately 50 ma with no input signal. The purpose of equalizing these cathode currents is to eliminate the effect of d-c saturation of the output-transformer core.

(3) For low distortion at high power output, the power stage requires a carefully designed output transformer which presents 10,000 ohms impedance from plate to plate of the 807 tubes when the secondary is loaded with 16 ohms. A typical transformer has the following approximate specifications: primary inductance at 50 cycles, 110 henrys; total primary d-c resistance, 300 ohms; leakage inductance referred to primary, 25 millihenrys; efficiency, 95 percent. Such a transformer, operated with no d-c magnetization, as described in the preceding subparagraph, contributes to the broad frequency response and low distortion of this amplifier.

(4) The voltage feedback path is taken from the secondary of the output transformer, through *R*1, to the cathode of the first stage of *V*1, developing the feedback voltage across *R*3. The feedback circuit reduces the overall gain of the amplifier by about 20 db (decibels), but it greatly extends its flat-frequency range, and reduces both harmonic and intermodulation distortion. In addition, it lowers the internal impedance of the amplifier to only ⅔ of an ohm on the 16-ohm tap of the output transformer. This low internal impedance reduces the tendency of the loudspeaker to ring when reproducing high frequencies or transients.

(5) The high fidelity of this amplifier results from no one feature, but from a combination of several features. Its excellent transient response results largely from the phase-splitter circuit at its input, and from the reduction of the output impedance accomplished by the feedback circuit. Its low-frequency response is largely the effect of the small number of coupling capacitors. The extension of its response at the high-frequency end is caused mainly by the feedback circuit. This feature, together with the cancellation of even harmonics in the two push-pull stages,

is responsible for the practically distortion-free operation of the amplifier. The transformer is carefully chosen and operated so that it does not cancel out any of the advantages arising in the other features of the circuit.

(6) The power supply which furnishes the required d-c voltages is built on a separate chassis to reduce the inductive hum pickup. Connection of the amplifier to the power supply is made through a cable and four-prong connector that plugs into a corresponding outlet in the power-supply chassis.

102. Summary

a. Sound-system amplifiers are used to raise the levels of weak audio signals to high loudspeaker levels. The use for which an amplifier is to be put determines the power output and circuit required. Power outputs vary from a few watts to 100 watts. The last stage of practical amplifier circuits can be single-ended, or push-pull-ended.

b. Special features in the amplifier circuits include tone controls, for adjusting the over-all response of a system; decoupling filters, for reducing the effects of common impedance feedback; a bias cell in the input stage, to minimize hum; shunt feed, for improving low-frequency response; and negative feedback, for improving over-all response and maintaining a low output impedance.

103. Review Questions

a. Why are sound-system amplifiers necessary?

b. What factors determine the power-output requirements of sound-system amplifiers?

c. Compare sound-system amplifiers using output stages which are (1) single-ended, (2) push-pull-ended, (3) push-pull-ended with feedback.

d. Sketch a circuit of a tone control, and explain its operation.

e. Describe the effects of common impedance feedback on the operation of an amplifier. How may these effects be minimized?

f. Compare the operating characteristics of decoupling filters which use (1) series resistors, (2) series inductors.

g. What are the advantages of a push-pull output stage?

h. Why is a bias cell sometimes used instead of cathode bias?

i. Sketch a circuit showing the use of shunt feed. What are the advantages and disadvantages of this circuit?

j. Sketch a circuit of a direct-coupled phase splitter and explain its operation and advantages.

k. Discuss methods used for obtaining high-quality response in sound-system amplifiers.

l. Sketch a circuit of a practical sound-system amplifier.

m. Discuss the effects of negative feedback when used in the output stage of a sound-system amplifier.

CHAPTER 8

SINUSOIDAL R-C OSCILLATORS

104. Fundamentals of R-C Oscillators

a. Definition of Oscillator Circuit. An *oscillator circuit* is one which delivers an a-c output voltage, usually having a definite desired waveform and frequency, without the use of an external input signal (A, fig. 80). The operation of an oscillator circuit depends on a special application of the principles of amplifier circuits.

b. Application of Oscillator in Signal Generator.

(1) An important application of oscillators having sinusoidal output waveforms is in the *signal generator,* a test instrument used in testing amplifiers and other electronic equipment. A signal generator contains one or more oscillator circuits, together with auxiliary circuits, for varying the magnitude and frequency of the output voltage, or otherwise contributing to the convenience or precision of the instrument in practical use. The term *oscillator* sometimes is used to designate the entire signal generator.

(2) Figure 79 shows a signal generator being used to test an amplifier. The signal generator supplies the input voltage needed to enable the amplifier to function. Its output is measured with a voltmeter, an oscilloscope, or other suitable instrument. Signal generators are needed to measure the gain, frequency response, and distortion of amplifiers. Signal generators also are used to test filter performance, measure frequency, calibrate other oscillators, and make bridge measurements of unknown reactances at specified frequencies.

c. Operation of Oscillator. Basically, an oscillator is an amplifier which derives its input signal from its own outputs. This basic principle of oscillator operation is illustrated in B and C of figure 80.

(1) The amplifier in B has a gain of 10. It produces an output of 10 volts when 1 volt is applied to its input terminals from the generator through the switch in position 1. It is important to note that the input and output voltages of the amplifier have the same polarity (in phase). In C, switch contact 2 has been connected to the load resistor. Thus, when the amplifier input switch is thrown from position 1 to position 2, the external input voltage, in B, is replaced by its exact duplicate—obtained, however, from a tap on the output load resistor of the amplifier. The new input voltage is identical to the original external input signal in both amplitude and phase; therefore, the external input signal is no longer needed, and the amplifier continues to produce an output voltage as long as the feedback path from output to input is not disturbed. Under these conditions the amplifier is said to *oscillate.*

(2) In practical oscillators, the oscillations are started by variations in plate current as the tube heats up. They build up because of the feedback path and the amplifying action of the circuit. It is not necessary to use an external generator and a switch. These elements are shown merely to illustrate the principle.

(3) The preceding explanation of oscillator operation shows that an amplifier performs as an oscillator provided a portion of its amplified output voltage is fed back with a specified *amplitude* and *phase.* These are the two basic

Figure 79. Testing amplifier with oscillator.

operating principles of many types of oscillators used as signal generators.

(4) In many practical oscillators, the requirements of phase and amplitude are fulfilled by networks of resistors and capacitors which are inserted in the feedback loop of a conventional amplifier. These networks also determine the frequency of the oscillator output. Oscillators of this type are called *R-C oscillators,* and are the subject of this chapter. Since practical oscillators are used for test purposes, their performance must meet rigorous specifications as to output voltage, frequency calibration, and distortion.

d. Requirements of Oscillator Operation.

(1) Signal generators often are equipped with attenuators or other circuits by which the amplitude of the output voltage may be varied at will; but, once the control is set, the output must remain constant over a wide range of frequencies, and for long periods of time of any one setting.

(2) The oscillator frequency must be free from drift—that is, it must remain constant within narrow limits for long periods, once the control is set. Further, this frequency should correspond exactly to the calibration of the dial associated with the frequency control. This relationship should exist over the entire range of the oscillator. The dial itself should be large and easily read.

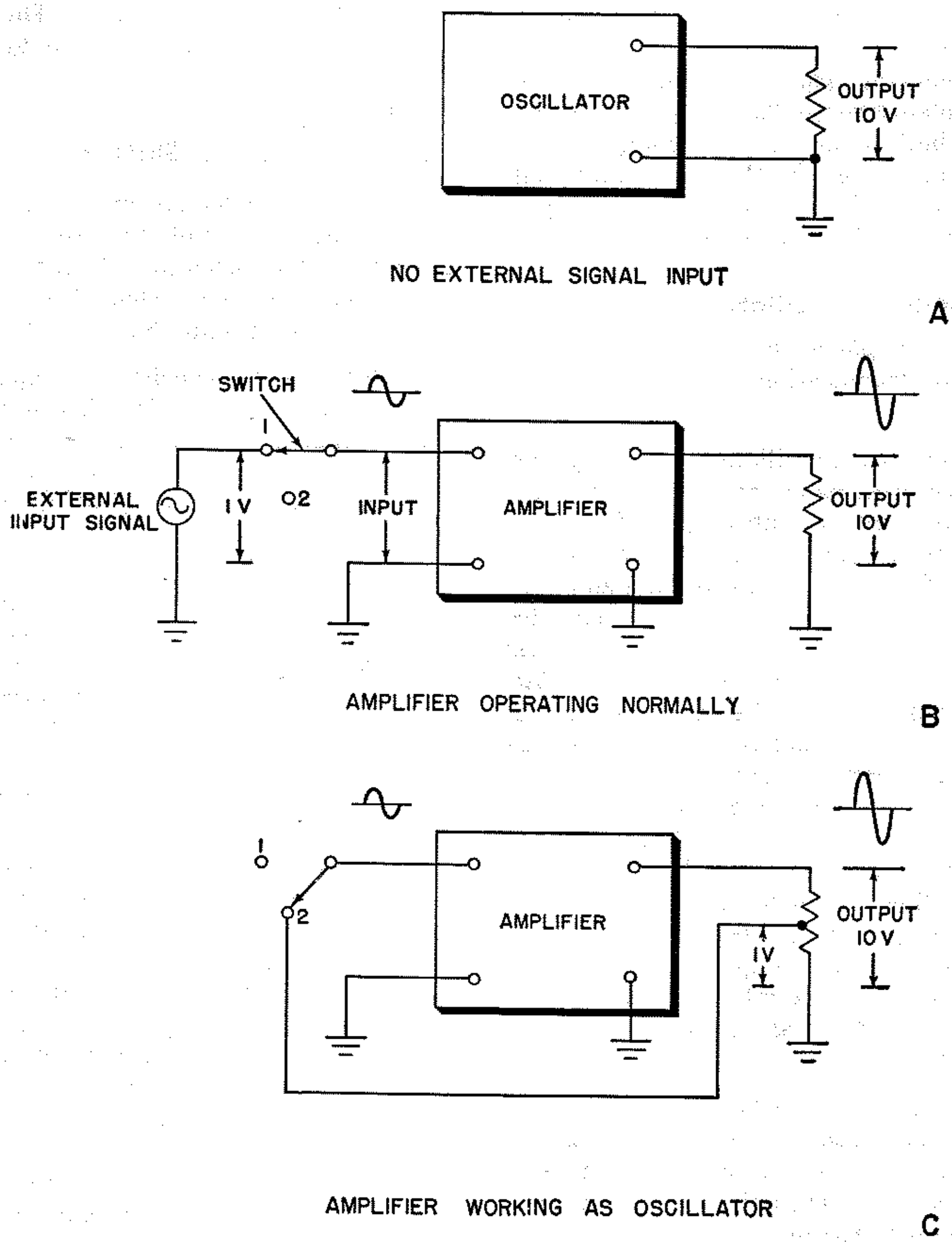


Figure 80. Operation of oscillator.

(3) The oscillator should produce an output voltage having negligible distortion, especially if it is to be used in making distortion tests on amplifiers or other equipment. Distortion products originating in the oscillator will appear in the output of the amplifier under test as well as in the circuits of the oscillator itself, and it is difficult or impossible to separate the two. If the oscillator output is appreciably distorted, even a distortion-free amplifier would appear to produce appreciable distortion. Distortion in oscillators is minimized by special circuits, and by operating all components conservatively—that is, by not driving any component with excessive current or

voltage which might produce distortion.

(4) Following paragraphs describe and explain two oscillator types which can be built to meet the foregoing requirements. They are the *phase-shift* oscillator and the *Wien-bridge* (pronounced Ween bridge) *oscillator*.

105. Phase-Shift Oscillator

The basic R-C *phase shift oscillator* may be developed by modification of a conventional *amplifier*.

a. The heart of the oscillator is the amplifier. The amplifier in A, figure 81, has a gain of 18, and, therefore, produces an output of 18 volts when 1 volt is applied to its input terminals, as shown. It differs from a conventional amplifier only in that its load resistor has a *phase shifter* connected in parallel with it. The phase shifter is an R-C network, which will be described later. This particular phase shifter reduces the 18-volt negative-going output voltage of the amplifier to a 1-volt positive-going voltage—a phase shift of 180°. It is a characteristic of the phase shifter that it produces exactly these effects at one frequency only. At any other frequency, the output will be less than 1 volt, and its phase will be shifted by more or less than 180°. At the one frequency determined by the phase shifter, however, its output is exactly in phase with the amplifier input voltage, as shown by the waveforms in A.

b. In B, figure 81, the 1-volt output from the phase shifter has been substituted in the circuit in place of the 1-volt external signal. This meets the requirements for oscillation of an amplifier explained in the preceding paragraph: A proper portion of its output voltage has been fed back to its input in proper amplitude and phase. The oscillator is redrawn in C with the phase shifter shown in conventional manner on the grid side of the amplifier.

c. It now is clear that the phase of the feedback voltage must be 0°, 360°, or a multiple of 360°, in order to satisfy the phase requirements of an oscillator. This constitutes positive voltage feedback, or regeneration, as defined in chapter 4. In the discussion, it is assumed that the one-stage amplifier introduces only 180° of the required 360° phase shift. The additional 180° of phase shift is produced by the phase shifter.

106. L-Type Phase Shifter

The phase shifter shown simply as a block in figure 81 consists of three resistance-capacitance *L-sections*, each introducing a phase shift of approximately 60°. One of these L-sections is shown in A of figure 82.

a. The L-section consists of a capacitor and a variable resistor in series. Theoretically, this combination can be made to cause a phase difference of any amount from 0° up to 90° between the input voltage, *e*, and the output voltage, e_R, simply by varying the resistor, and a phase shift of 180° should be obtainable by cascading two L-sections (since the phase shifts of cascaded networks add numerically). However, for practical reasons to be explained, it is possible to obtain with one L-section phase shifts only up to *nearly* 90°. To obtain 180°, therefore, *three* L-sections are connected in cascade, each section shifting the phase of the input voltage by 60°.

b. The operation of the L-section as a phase shifter is as follows: An alternating voltage, *e*, is applied to the input terminals. A current flows through the series R-C circuit, the magnitude of which is determined by the impedance of the capacitor and resistor. The resulting voltages are represented in the graph in B and the vector diagram in C. Since the impedance is capacitative, the current, *i*, leads the impressed voltage *e*, by an angle shown equal to 60°, in C. The resulting voltage drop, e_R, across resistor *R*, is in phase with the current, and, consequently, e_R leads the impressed voltage by the same angle.

c. The phase shift of an L-section can be varied in several ways. For example, varying resistor *R* changes the phase angle of the current flowing in the circuit. If *R* is reduced to 0, the phase angle of the current is 90°, but this action is useless, because no voltage is developed across 0 resistance. It is possible, then, to get almost, but not quite, 90° in a single L-section of this type. For this reason, two practical L-sections can give a phase shift of almost, but

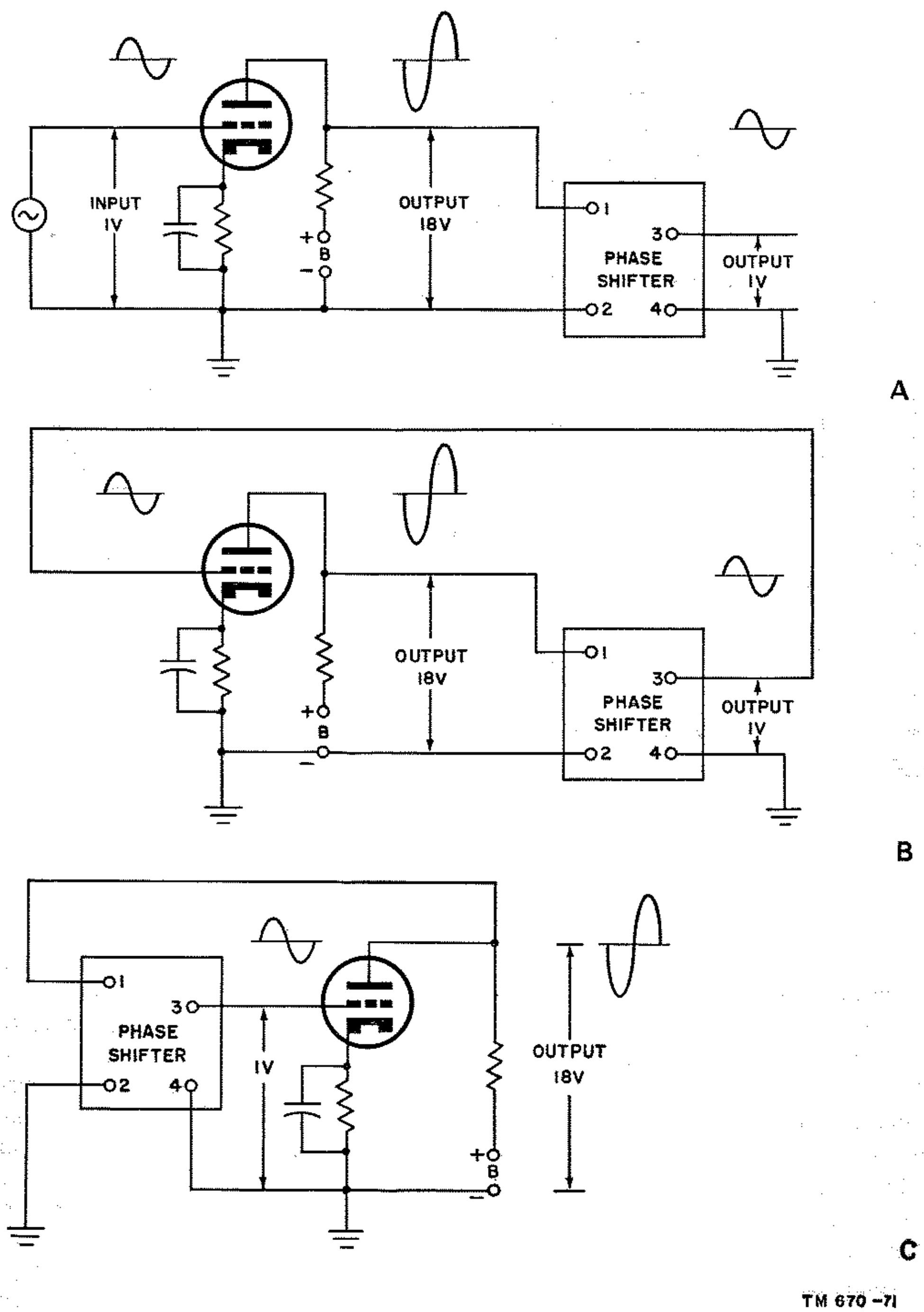


Figure 81. R-C oscillator.

not quite, 180°. For this reason, at least three L-sections are needed to obtain 180°.

d. A combination of three L-sections is shown in A of figure 83. The output of section 1 is applied to the input of section 2, and *its* output is applied to the input of section 3. Each section contributes to the total phase shift. The phase shift of each section can be adjusted so that each produces 60° of phase shift, making the total phase shift exactly 180°.

e. The phase relations of the several voltages of a three-section L-type phase shifter are

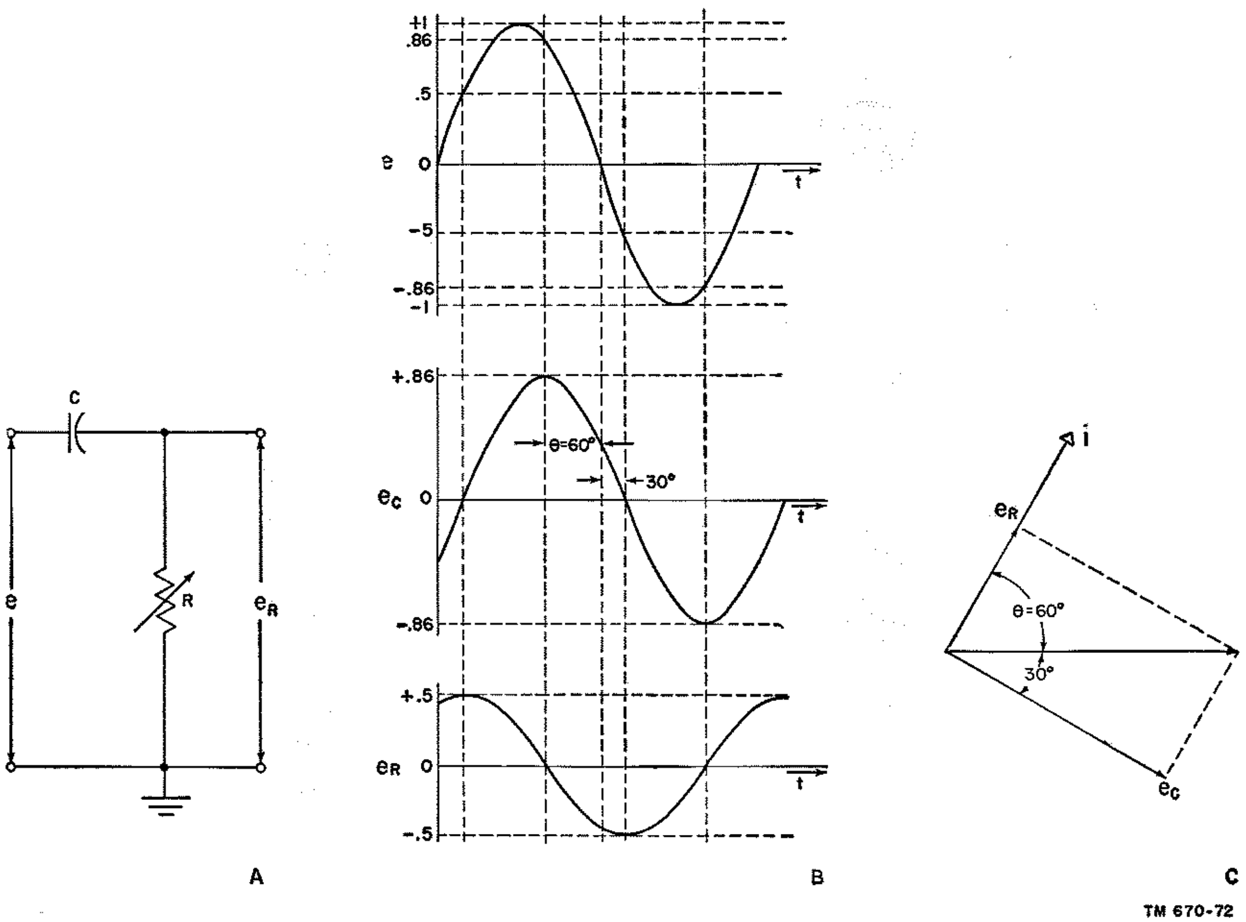


Figure 82. L-section R-C phase shifter.

shown in the graph in B. The output of section 1, e_{R1}, leads the plate voltage, e_p, by 60°. This output is impressed on section 2, output e_{R2} of which leads the plate voltage by twice 60°, or 120°. Similarly, section 3 adds another 60° phase shift, making a total of 180°. Thus, the input voltage to the grid leads the output from the plate by 180°. This phase shift is added to the 180° phase shift contributed by the tube, making the total phase shift 360°.

f. For practical reasons, to be explained, the type of phase-shift oscillator employing this three-section phase shifter usually is operated at a fixed frequency. Its frequency may be changed, however, by varying the elements of the phase shifter. For example, if the capacitance of any section is *increased,* the capacitive reactance ($X_C = \frac{1}{2} \pi fC$) is *decreased.* This causes a decrease in the angle of phase shift, provided the frequency is not changed. But at some *lower* frequency, the capacitive reactance and the phase shift will be the same as before. Oscillation can take place only at the one frequency at which the total phase shift is 180°; voltages of any other frequency are fed back out of phase, and do not result in oscillations. The frequency of oscillation depends on the ratio X_C/R of each section. This ratio equals $\frac{1}{2} \pi fCR$. If the ratio remains constant, the frequency of oscillation is constant. The *frequency* is *increased* by *decreasing* either the capacitance or the resistance, and vice versa. In practical oscillators, ganged variable capacitors usually are employed to vary frequency smoothly from the lowest to the highest frequency of one range; the resistors are varied in steps which change the frequency range.

107. Practical Phase-Shift Oscillator

a. Shortcomings of Phase-Shift Oscillator.

(1) The phase-shift oscillator shown in A,

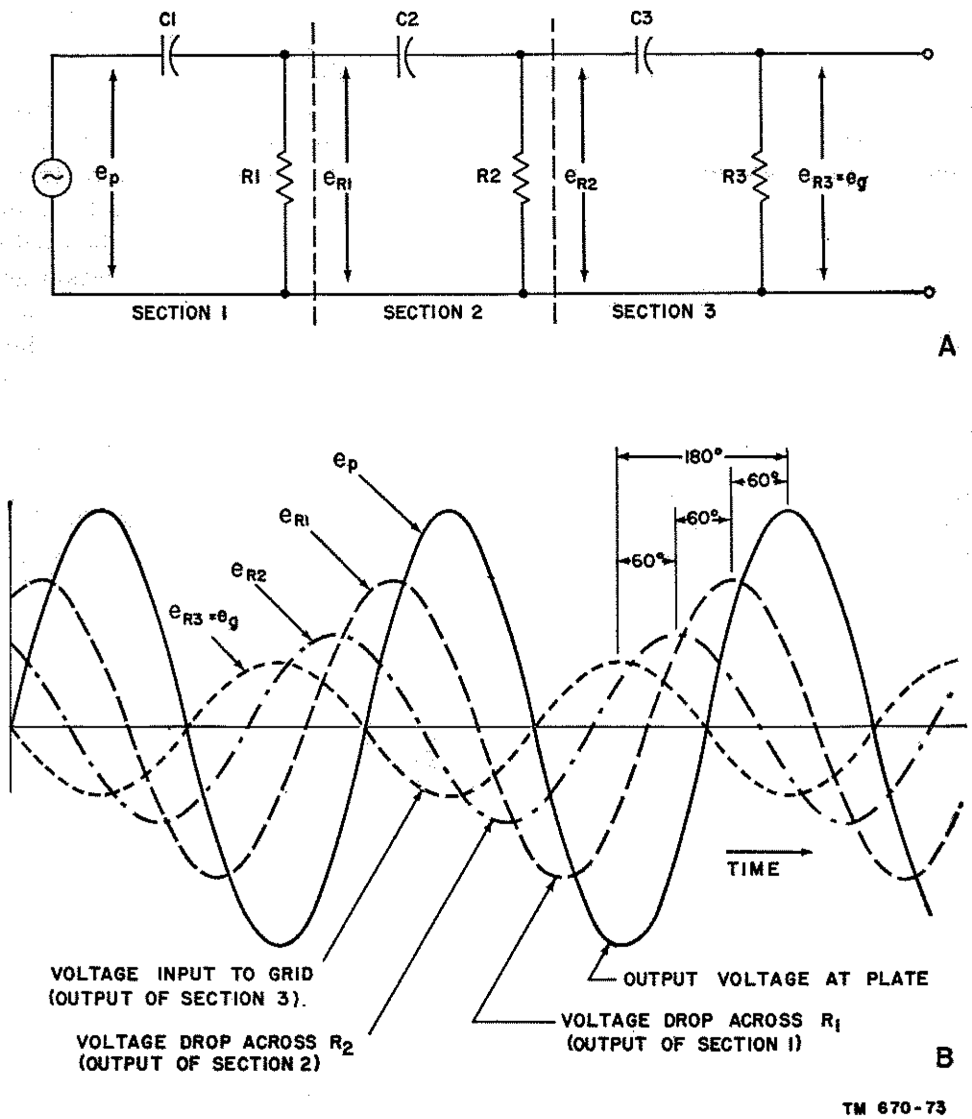


Figure 83. Phase-shifting circuit.

figure 84, uses the three-section phase shifter of figure 83. This oscillator has many advantages. It is easy to construct, requires only one tube, and operates over a wide frequency range—from below 1 cycle to about 1 megacycle. The output waveform may be almost a pure sine wave when operated at low gain to barely sustain oscillations. However, the output voltage is low in amplitude, and it varies as oscillator frequency is varied. These shortcomings are caused primarily by the type of phase-shift network used. They now will be discussed in detail in order to clarify the required circuit revisions. Distortion may be high in this oscillator because the high-frequency harmonics generated by the tube when operated on the nonlinear portion of its characteristic, are passed readily by the low-impedance series capacitors of the R-C network, applied to the grid of the tube, and reamplified. Thus, the output waveform may contain the harmonics generated by the tube plus the amplified distortion products. In order to minimize distortion, therefore, the tube is operated on the most linear portion of its characteristic, and at low output voltage.

(2) Another reason for the low output voltage of this phase-shift oscillator is the loss which the phase-shift network introduces. For example, the three-section L-type phase shifter requires an input of 29 volts in order to obtain an output of 1 volt. It will cause oscillations only if the amplifier circuit in which it is used has a gain which is the same as the loss caused by the phase-shift network, or 29. The tubes needed for this oscillator are high-mu triodes, or pentodes, but since their amplification usually is greater than 29, they are operated inefficiently at low gains. The loss in the phase-shift network can be reduced by using a larger number of sections in cascade, since the loss per section drops very rapidly as the phase shift is reduced. For example, a four-section R-C shifter requires only 18.4 volts input in order to produce 1 volt output. This phase shifter may be used with a low-mu triode for operation as an oscillator. Addition of sections cannot be continued indefinitely, however. As the number of L-sections is increased, and more C and R elements are added to the phase shifter, it becomes increasingly impractical to vary the oscillator frequency smoothly, as is required of a flexible signal generator.

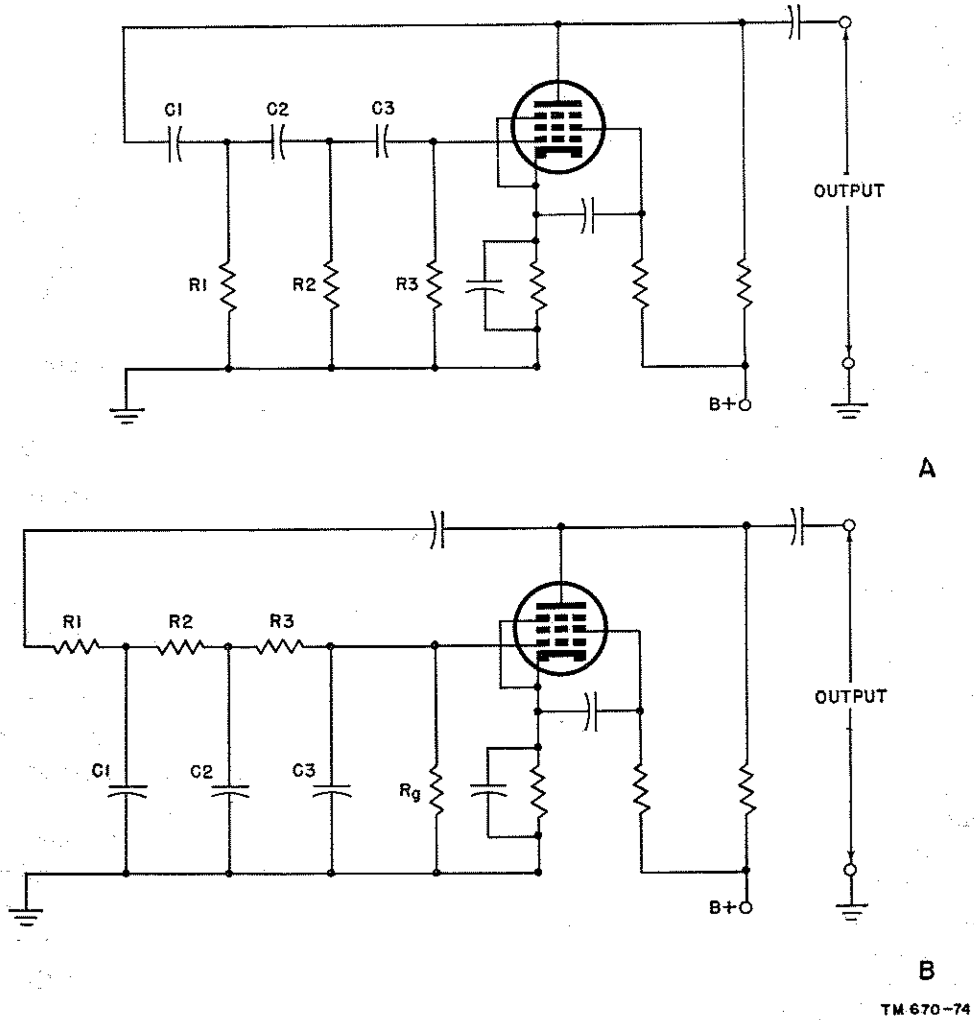


Figure 84. Practical phase-shift oscillator.

(3) The output voltage of the oscillator changes in magnitude as the frequency is varied. This effect is undesired not only because it is time-consuming to check the oscillator output voltage whenever the frequency is changed, but also because the amount of waveform distortion of an amplifier (and oscillator) usually changes with amplitude. Tests made with such an oscillator are, therefore, unreliable unless exact voltages and distortion percentages are known for every oscillator adjustment. The output voltage changes with frequency because the impedance of the phase shifter changes with frequency. For example, at high frequencies, the impedance of each L-section of the phase shifter is reduced as the reactance of the series capacitor falls off. This results in lower output voltage, because the phase shifter is part of the load in the plate circuit of the oscillator. As the load impedance decreases, the output voltage drops.

b. Modifications Which Improve Oscillator Operation. The phase-shift oscillator circuit in A, figure 84, can be modified to remedy the shortcomings discussed above. One method which minimizes waveform distortion is shown in B. Here the R and C elements of the phase shifter are interchanged. Distortion is reduced because the capacitors act practically as a-c short circuits for harmonics of the fundamental frequency. However, experiment and analysis show that the values of resistor and capacitor needed for this phase shifter are often impractical. It is preferable, therefore, to use the first phase shifter (shown in A, fig. 83) and operate it at only one frequency. Alternatively, if it is desired to vary oscillator frequency, negative feedback is applied in order to minimize amplitude and frequency distortion and to stabilize variations of output voltage. This is done more effectively by using an entirely different type of phase-shift network, having a lower loss, which is discussed in the next paragraph.

108. Wien-Bridge Oscillator

The shortcomings of the phase-shift oscillator are minimized in the *Wien-bridge oscillator* (fig. 88). This oscillator is so named because the networks of resistors and capacitors which produce the positive and the negative feedback, together make up a bridge circuit known by the name of its inventor, Wien. The Wien-bridge oscillator also operates on the basic principle discussed in paragraph 104*c*. However, the feedback voltage is obtained from a network having zero phase shift at the frequency of oscillation; and for this reason the losses of the network at the frequency of oscillation are lower than the losses of the L-type phase shifter. Since the phase shift of a single tube is only 180° (reversed polarity), it is necessary to use two tubes connected in cascade to provide the 360° of phase shift (same polarity) required for oscillation. The tubes are made part of a two-stage amplifier having two feedback loops. One of these paths provides the positive feedback which causes oscillations; the other provides the negative feedback which minimizes distortion and maintains the output voltage constant over a wide range of frequencies. Because the entire phase shift is provided by the tubes, the function of the *phase shifter* actually is to control the frequency of oscillation, rather than to shift the phase of the feedback voltage.

109. Wien Bridge

For an understanding of the operation of the Wien bridge oscillator, refer to the phase shifter illustrated in figure 85. It forms two of the four arms of a *Wien bridge*, as will be shown presently. One arm is the series combination of resistor $R1$ and capacitor $C1$; the other arm is the parallel combination of resistor $R2$ and capacitor $C2$.

a. Two voltages are of interest in the phase-shifting network illustrated. They are the input voltage, E_i, which is impressed across the entire network, and the resulting output voltage, E_o, which appears across the parallel combination of $R2$ and $C2$. The curves of figure 86 show that the output voltage is in phase with the input voltage at only one frequency, F_o. This is because the phase of the voltages across the two bridge arms varies in opposite directions as frequency changes. For example, at a frequency below F_o, the voltage across the series arm lags

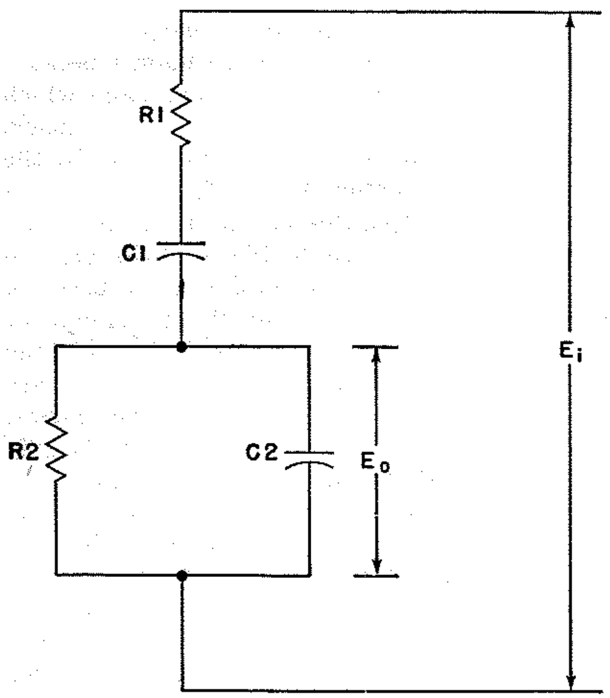


Figure 85. Wien-bridge phase shifter.

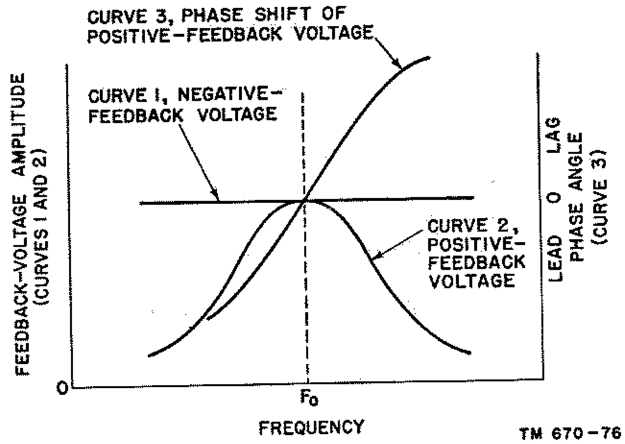


Figure 86. Frequency feedback voltages in Wien-bridge oscillator.

the input, and the voltage across the parallel arm leads the input. As the frequency increases, the phase difference between each of these voltages and the input voltage decreases, until at one critical frequency, F_o, both are in phase with the input. The voltage across the parallel arm is the output voltage. The variation of the phase of the output voltage relative to that of the input voltage is plotted as curve 3. As shown, it leads the input at frequencies below F_o, and lags at frequencies above F_o. At F_o the phase shift is zero.

b. Both experiment and analysis show that the output voltage of the phase shifter is greatest at the frequency of zero phase shift. It is customary to make the two resistors, $R1$ and $R2$, equal, and the two capacitors, $C1$ and $C2$, equal (fig. 85). Under this condition the maximum output, occurring at the frequency of zero phase shift, is exactly one-third as large as the input voltage. Consequently, only 3 volts are required at the input to the phase shifter in order to obtain 1-volt output at the oscillating frequency. (Compare this with the 18.4 volts required by the four-section L-type phase shifter.)

110. Operation of Wien-Bridge Oscillator

a. Figure 87 shows the simple steps required to develop the *Wien-bridge oscillator circuit* from a two-stage amplifier. Assume that the first stage of the amplifier, in A, has a voltage gain of 50, and the second a gain of 10. The over-all gain of the amplifier, therefore, is 500, and 1 volt applied to the input terminals of the amplifier appears as 500 volts at the output.

b. Note that the output and input voltages are in phase at the frequency for which the phase shift of the amplifier is exactly 360° (same polarity) (A, fig. 87). Because this is so, this amplifier oscillates if a portion of the output voltage is used to drive the amplifier in place of the external signal generator. One way of reducing the 500 volts of output voltage to the required 1 volt of input voltage is to use a simple resistance voltage divider. An improvement results by applying negative feedback to the amplifier, as in B. This reduces the gain to such a value that only the needed 1 volt of input is obtained. It also flattens the frequency response of the amplifier, maintaining both the phase shift and the output voltage almost perfectly constant over a wide frequency range, as explained in chapter 4. For example, the feedback network, $R3$–$R4$, reduces the over-all gain of the amplifier from 500 to about 3, if $R4$ is made twice the resistance of $R3$. The output voltage of this amplifier is now 3 volts, when 1 volt is applied to the input terminals. For reasons to be explained, a gain of 3 is just what is required. The amount of negative feedback

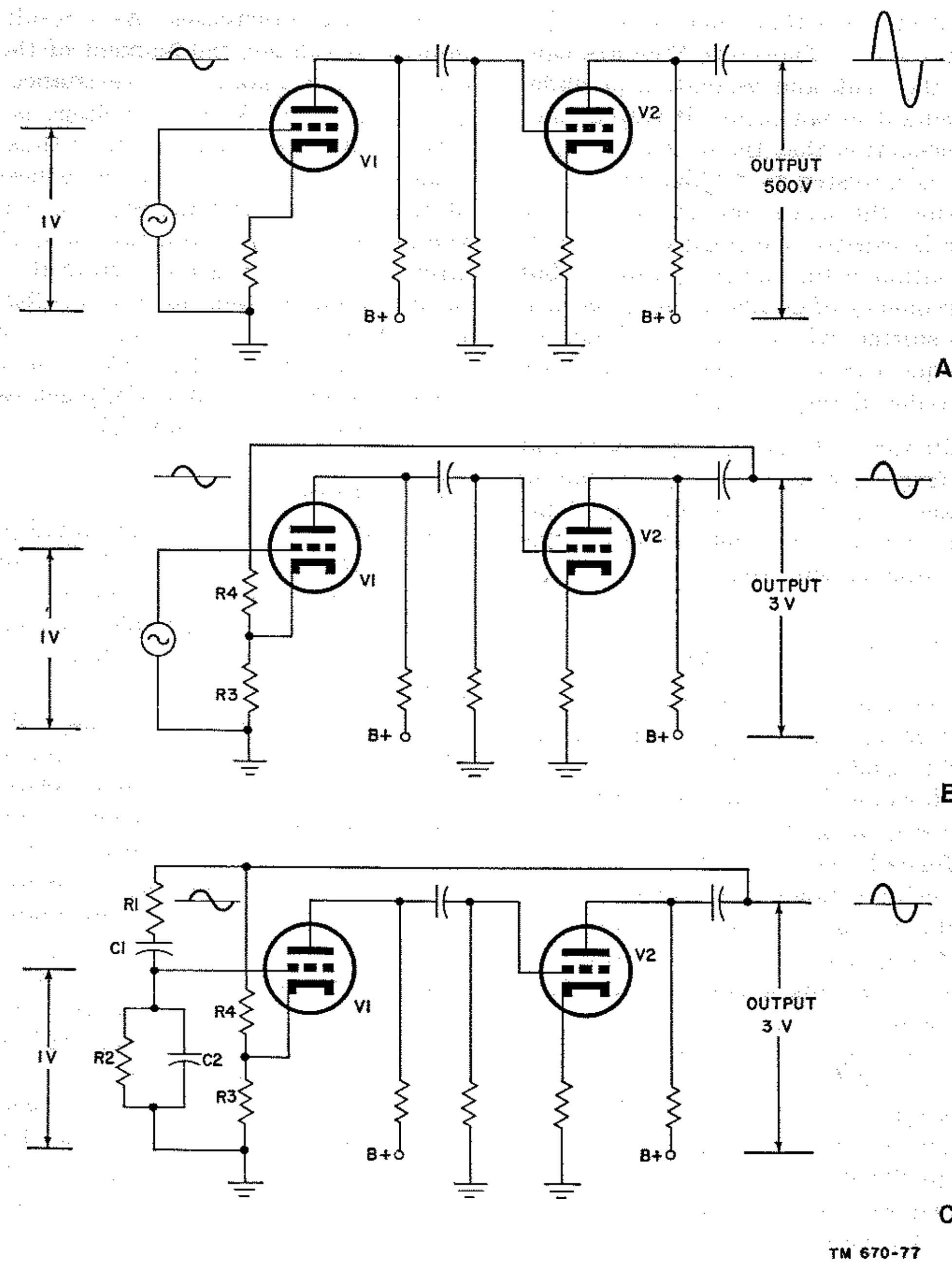


Figure 87. Development of Wien-bridge oscillator.

does not change as frequency changes, since the voltage divider is a pure resistance. Further, it is equal to, or barely less than, the maximum positive feedback (curve 1, fig. 86).

c. In order to hold the frequency of oscillation constant, the Wien-bridge phase shifter (fig. 85) is connected between the output of the amplifier and the input (fig. 87). The amplifier oscillates at one frequency because the phase shift between output and input terminals is still 360° (same polarity), since the phase shifter introduces zero phase shift at that one frequency. For example, the amplifier oscillates at 1,000 cycles if the phase shifter has zero phase shift at 1,000 cycles.

d. The amplifier amplifies voltage of other frequencies also, but these voltages are reduced by the phase shifter in the positive feedback path to amplitudes less than that of the negative-feedback voltage. Therefore they are cancelled out at the grid, and oscillation at undesired frequencies does not occur. It is to accomplish this cancellation that the over-all gain of the amplifier is adjusted to a value of 3, or a shade more, since the maximum output from the phase shifter is exactly one-third of its input. The discrimination of the circuit against all but the desired frequency of oscillation is sharpened by the phase shifter, which applies voltages of undesired frequencies to the grid out of phase as well as in reduced amplitude.

e. The oscillator, in C, is made to operate at any desired frequency by changing the values of the resistances or of the capacitances of the phase shifter. These values determine the frequency in accordance with the formula

$$\text{frequency, } F_o, = \frac{1}{2\pi R_1 C_1}$$

The formula involves only R_1 and C_1 because the resistances of R_1 and R_2 are equal, and the capacitances of C_1 and C_2 are likewise equal. Because, in the formula for F_o, $R1$ and $C1$ usually are represented as R_1 and C_1, this usage is used here. The formula shows that frequency is increased by using smaller values of resistance, or of capacitance, or of both. In a practical oscillator, the frequency is varied smoothly over a wide range by varying a gang of capacitors used for C_1 and C_2. The frequency range is selected in steps, by changing the values of resistors $R1$ and $R2$. For example, a typical oscillator has four frequency ranges: 20 to 200 cycles, 200 to 2,000 cycles, 2,000 to 20,000 cycles, and 20,000 to 200,000 cycles. Each range is selected by switching different resistors into the Wien bridge. The frequency is varied smoothly in each range by setting the variable capacitors.

f. A practical Wien-bridge oscillator usually has one other element which insures constant output voltage as frequency changes. This is a *lamp*, or thermistor. In A of figure 88, this lamp, LP, is used in place of the cathode bias resistor of $V1$ in A of figure 87. It stabilizes the amplitude of oscillations because its resistance varies with the current which flows through it. The current flowing through the lamp is determined by the feedback voltage impressed across it. This voltage increases if for some reason the output voltage increases. As a result of a high impressed voltage, the filament of the lamp becomes hotter, making its resistance larger; a greater negative-feedback voltage is developed across the increased resistance; thus, more degeneration is provided, which reduces the gain of the amplifier and thereby holds the output voltage at a nearly constant amplitude. The circuit shown in A is redrawn in B to show the bridge configuration of the feedback circuit more clearly. The two upper arms comprise the Wien phase shifter (fig. 85); the two lower arms contain the negative-feedback resistor $R3$, shown in A, and the lamp, LP.

111. Summary

a. Oscillators are used in signal generators, an important group of electronic test instruments. The requirements for a test oscillator include a smoothly variable frequency range, frequency and amplitude stability, and low distortion.

b. Phase-shift oscillators depend for their operation on the phase shift obtained in an R-C feedback network connected between output and input of an R-C amplifier. The phase-shift circuit may be either a series of L-type phase shifters, cascaded, and having a high loss, or it may be a Wien bridge having a low loss at the frequency of zero shift.

c. In the phase-shift oscillator, the phase-shift network usually produces 180° of shift, and a tube contributes another 180° to the total phase shift of 360° required for oscillation. In the Wien-bridge oscillator the phase-shift network produces zero phase shift at the frequency of oscillation. Hence, the amplifier around which the network operates contributes all of the 360° shift required for oscillation.

d. Frequency is varied in a phase-shift oscillator by varying the capacitance or resistance elements of the phase-shift network. Since this results also in a variation of the positive-feedback voltage, it is necessary to stabilize the amplitude of the output voltage by applying negative feedback.

112. Review Questions

a. Why are test oscillators necessary?

b. Under what conditions does an amplifier behave as an oscillator?

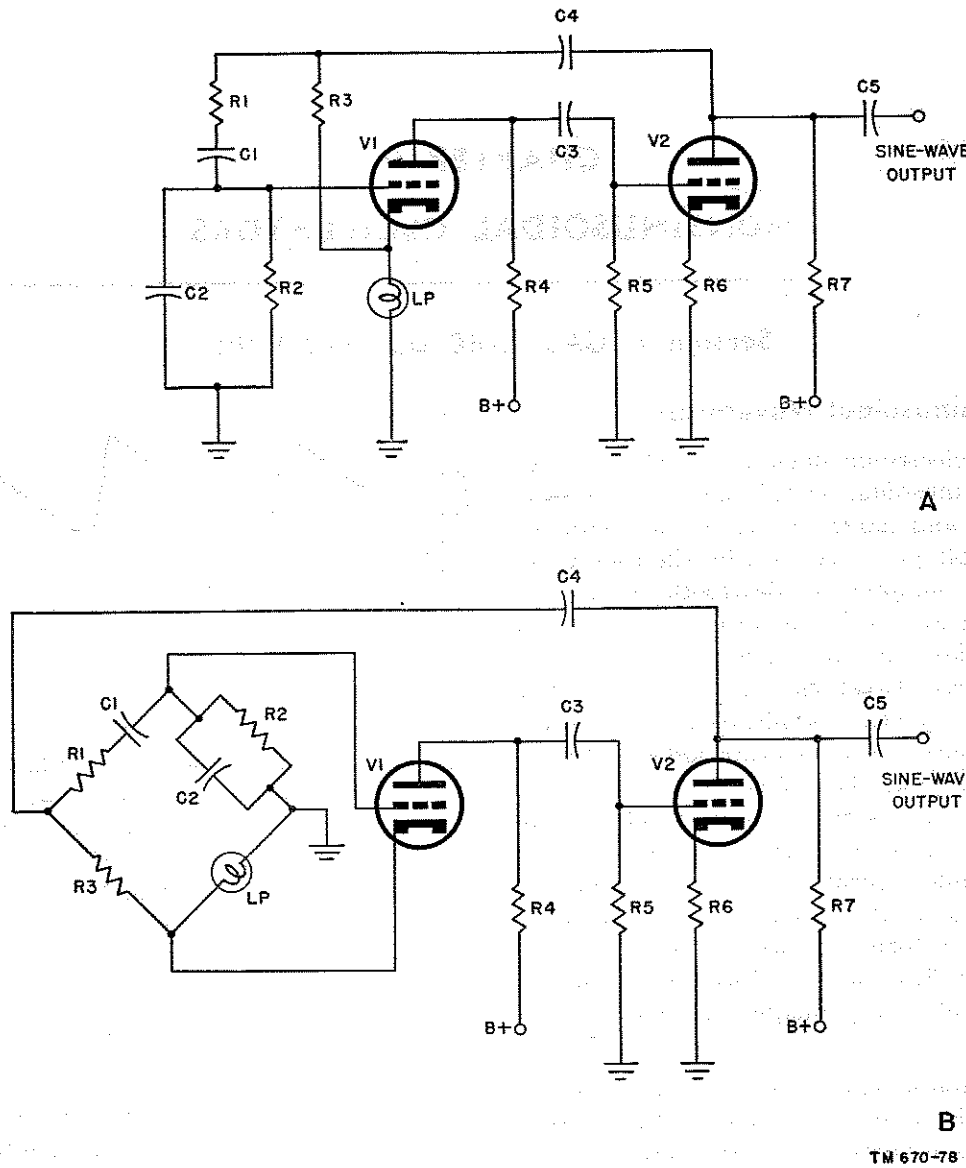


Figure 88. Wien-bridge oscillator.

c. What specifications should a test oscillator fulfill?

d. What are the operating principles of phase-shift oscillators using (1) L-type phase shifters, (2) the Wien bridge?

e. Compare the advantages and disadvantages of the L-type and Wien-bridge type of phase shifters.

f. Why is it necessary to use at least three sections of L-type phase shifters in cascade in order to obtain 180° of shift?

g. Why is distortion usually high in a phase-shift oscillator using an L-type phase-shift network? How is it minimized?

h. What is the purpose of the negative feedback in the Wien-bridge oscillator?

i. Explain how the thermistor stabilizes the output voltage of the Wien-bridge oscillator.

j. Sketch a circuit of (1) a single-tube phase-shift oscillator, (2) the Wien-bridge oscillator.

CHAPTER 9
NONSINUSOIDAL OSCILLATORS

Section I. GAS-TUBE OSCILLATORS

113. Nonsinusoidal Waveforms

a. Many electronic devices require voltages having nonsinusoidal waveforms, such as sawtooth waves and square waves, rather than the sine-wave voltages developed by the oscillators described in chapter 8. Sawtooth waves are needed in the sweep circuits of oscilloscopes and in many radar and television circuits. Square or rectangular waveforms are used in radar circuits and in other applications to develop trains of pulses, or to operate gating circuits and electronic switches. Sawtooth generators, discussed in this section, multivibrators, and blocking oscillators (sec. II), are nonsinusoidal oscillators, and operate by the charging and discharging of a capacitor through a resistance. An elementary form of sawtooth generator will be discussed first, as it affords the simplest illustration of the principle underlying each one.

b. A sawtooth waveform is used in oscilloscopes in order to *deflect,* or *sweep,* the electron beam back and forth across the screen of the cathode-ray tube at a constant rate. The resulting visible trace on the screen is known as a *sweep,* or *time base.* The beam moves from left to right across the screen at a constant and relatively slow rate, then returns much more quickly to its starting position at the left. It then repeats the same back and forth excursion, at a definite frequency. This motion of the beam is caused by a voltage which varies with time, as shown in figure 89. The instantaneous value of the voltage increases at a constant and relatively slow rate, falls much more quickly to its initial value, and then repeats this cycle indefinitely.

c. Sawtooth voltages are generated in oscillator circuits using tubes to control the charge and discharge cycle of a capacitor. The resulting voltage across the capacitor has a sawtooth waveform. The tubes may be either gas-filled or vacuum. Oscillators using vacuum tubes are more widely used, because of their greater frequency range, better linearity, and more stable operation. Gas-tube oscillators, however, are very simple.

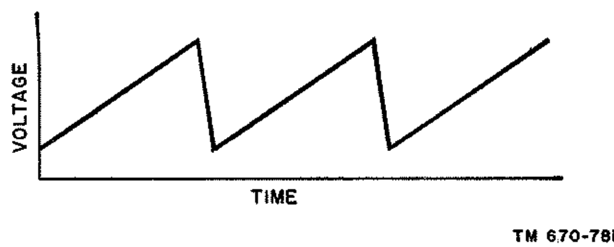


Figure 89. Sawtooth voltage waveform.

114. Gas-Diode Oscillator

The simplest sawtooth oscillator is one employing a gas diode. It is unstable in frequency, and this limits its application. An understanding of its operation, however, provides a foundation for the study of other nonsinusoidal oscillators.

a. Basic Gas-Diode Oscillator. A of figure 90 shows the three essential elements of a gas-diode oscillator circuit: a capacitor, *C,* a series resistor, *R,* and a gas-filled diode. They are connected as shown, and supplied from a d-c voltage source. The diode is a neon glow lamp, the type commonly used for indicating purposes. It consists of two electrodes mounted in a small glass envelope filled with neon gas. There is no heater or filament supply; the tube operates with a "cold" cathode. The circuit produces sawtooth oscillations when the d-c voltage, *E,* is applied. The supply voltage causes a current

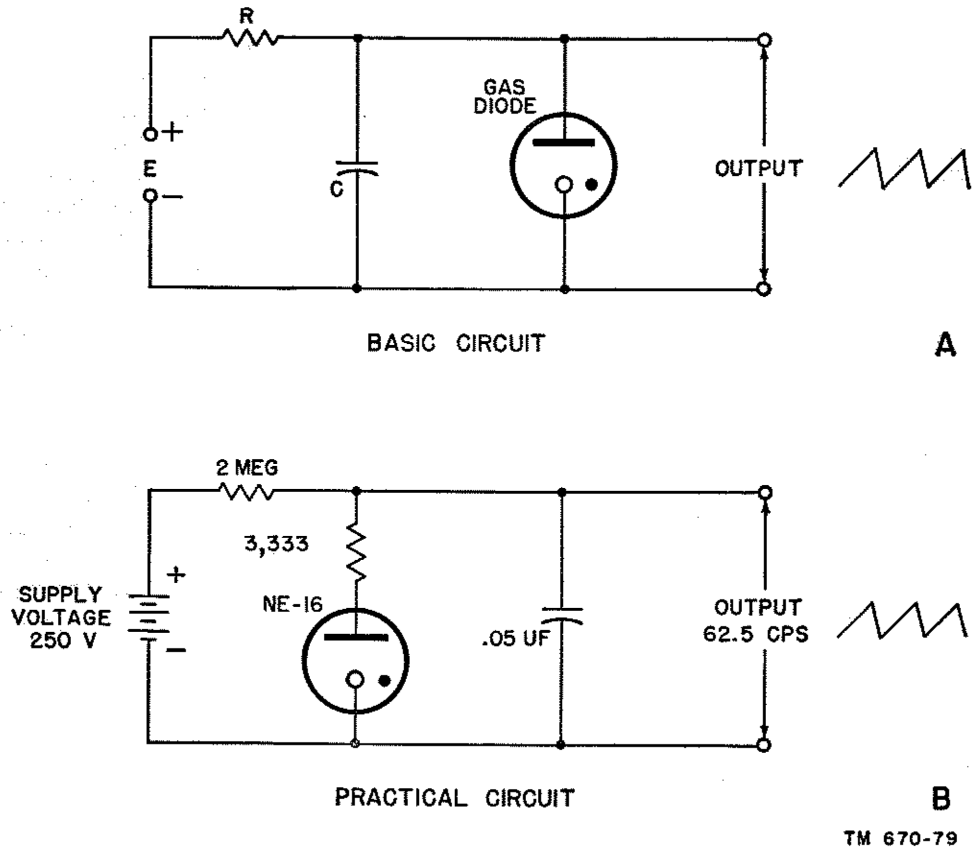


Figure 90. Gas-diode oscillators.

to flow through the series resistor, *R,* into capacitor *C.* The voltage across the capacitor rises according to the exponential charging curve shown in figure 91. As the capacitor charges, the voltage across it approaches the supply voltage, as shown by the unbroken and broken-line parts of the charge curve. The voltage across the neon tube is the same as the voltage across the capacitor, because these elements are in parallel. The neon tube acts as an open switch until the voltage across it reaches a critical value called the *firing voltage* or *firing potential.* At this point, the gas in the tube ionizes, and conducts readily, the tube acting like a very low resistance across the capacitor. The capacitor thus discharges very rapidly, as shown by the discharge curve. The discharge stops, however, when the voltage across the tube and capacitor reaches the *de-ionizing potential* of the tube. At this voltage, conduction stops, and the tube again becames an open circuit, and the capacitor begins to charge again, the voltage across it increasing as shown by the second rise in the unbroken curve. Again the voltage drops sharply when it reaches the firing potential. This process continues as long as the d-c supply is maintained, and produces the succession of sawtooth waves of voltage across the capacitor shown in the figure. Note that the capacitor voltage rises at a slightly decreasing rate. If the operation is restricted to a relatively small part of the exponential charge and discharge curve, however, the rate of voltage rise is nearly constant.

b. Practical Gas-Diode Oscillator. The simple circuit shown in A, figure 90, is impractical because, when the neon tube starts conducting, its resistance is momentarily very small, and would permit excessive current to flow, damaging the tube. This defect is eliminated in the circuit of B, where a protective resistor is inserted in series with the tube to limit the peak current to a safe value.

(1) B of figure 90 contains typical values of a practical gas-diode oscillator cir-

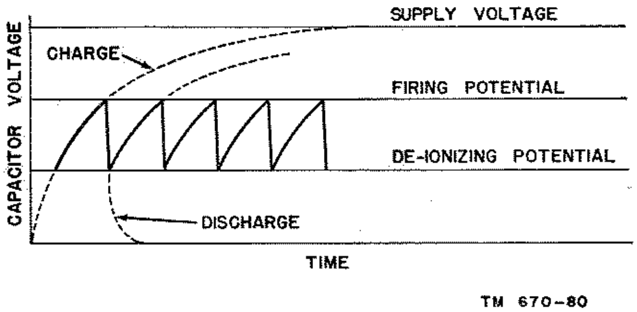


Figure 91. Variation of capacitor voltage in neon sawtooth generator.

cuit. The tube is an NE–16 neon lamp, which fires at about 77 volts and deionizes at about 48 volts. (Because these voltages may differ by several volts in different tubes, it is impossible to predict the operating characteristics of the circuit with accuracy.) The protective resistor has a resistance of 3,333 ohms, which limits the peak diode current to its proper value of 3 ma. The values of the capacitor, the larger resistor, and the supply voltage are chosen to obtain the desired frequency of oscillation.

(2) The rise in voltage across the capacitor is relatively very gradual and the decay very rapid, as is required of a sweep voltage (B, fig. 90). The capacitor charges very slowly through a 2-megohm resistor from a 250-volt source with a maximum charging current (250 volts divided by 2 megohms), of 125 microamperes. The capacitor discharges quickly with the peak discharge current of 3 ma, or 3,000 microamperes, through the conducting diode.

(3) The *frequency* of a sawtooth wave is the number of times the voltage rises and falls per second. In neon-gas diodes, the frequency range is from 1 cycle every few seconds up to about 10,000 cycles. Frequency is controlled by varying the time constant of the circuit or the magnitude of the supply voltage. The time constant depends on the values of the charging resistor and the capacitor. For example, the oscillator frequency of the circuit in B, figure 90, is 62.5 cycles, but an increase in the value of the resistor or capacitor would increase the time required to charge the capacitor from the fixed source, and result in a lower oscillator frequency (A, fig. 92). Conversely, an increased supply voltage would raise the frequency, because the capacitor would charge more quickly to the firing potential of the gas tube, as in B. The sawtooth frequency of gas-diode oscillators depends also on the values of the firing and de-ionizing voltages. These change with the age of the tube, and vary widely in different tubes of the same type number. This instability may be overcome to a large extent by the use of a gas triode instead of a diode, as will be explained in the next paragraph.

115. Thyratron Sawtooth Oscillator

a. Need for Synchronism. As a rule, whenever nonsinusoidal waves are used in electronic circuits, their frequency must be accurately controlled or *synchronized* with a signal of known frequency, if the waves are to serve a useful purpose. For example, the sawtooth waves which generate the sweep voltage of an oscilloscope must be held at a constant known frequency, in order that a steady pattern may be seen on the screen.

b. Thyratron Tube Characteristics.

(1) The frequency of a sawtooth oscillator can be controlled with great accuracy by using a special gas triode known as a thyratron. The operating characteristics of this tube differ considerably from those of vacuum triodes. Like a vacuum triode, a thyratron has a grid as its third electrode, but there are differences in both the physical structure of the two grids and in their effect on the plate current. In a vacuum triode (assuming a fixed plate voltage) the voltage applied to the grid controls not only whether plate current flows or not, but how much current flows. In a gas triode, the grid voltage can merely turn the plate

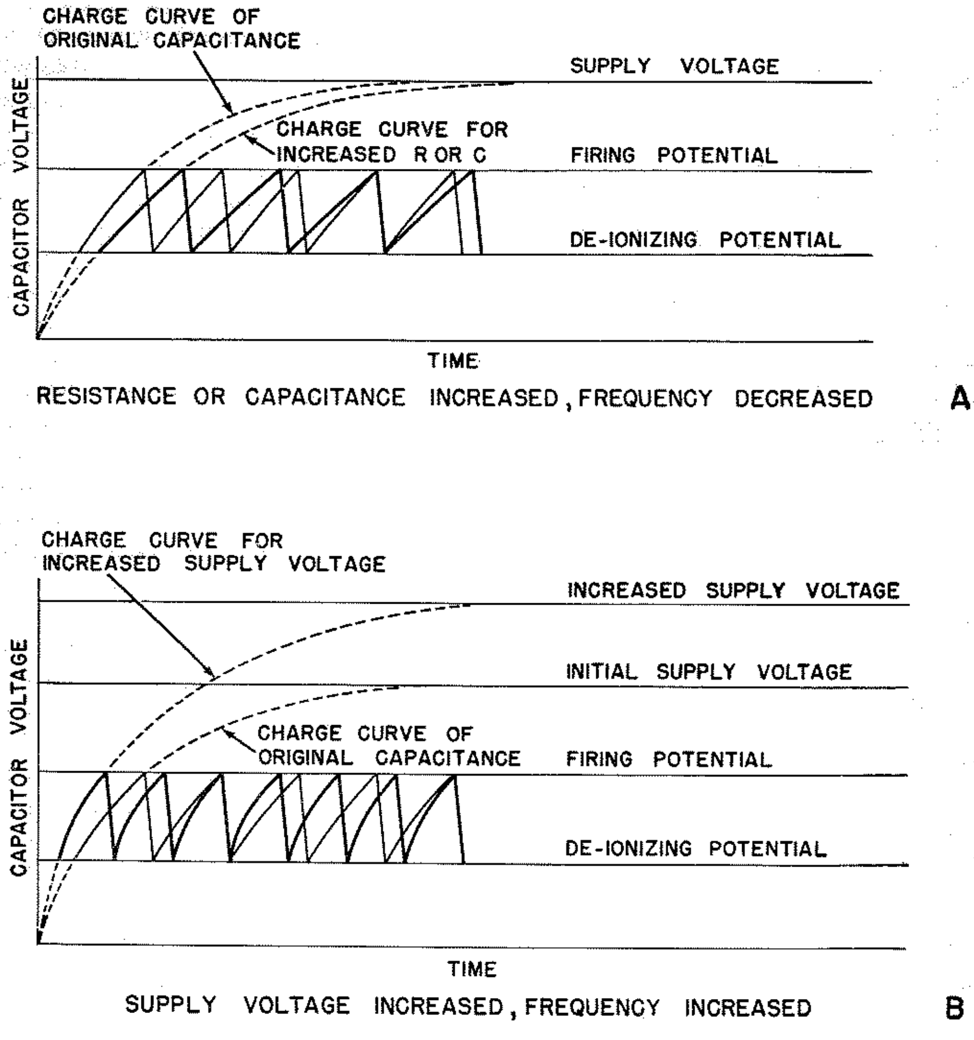


Figure 92. Frequency changes resulting from changes in circuit constants of neon sawtooth generator.

current on or off. Once plate current starts, the grid voltage exercises no control over the amount of plate current; the plate current then depends only on the plate voltage. The grid functions as a switch by altering the firing potential of the tube. The higher (more positive) the grid voltage, the lower is the plate voltage required to ionize the gas in the tube and start conduction. Thus, a thyratron that is cut off may be made to conduct by increasing sufficiently *either* the grid voltage or the plate voltage. For any given grid voltage, the thyratron functions as a gas diode. Once conduction starts, it behaves like a gas diode, regardless of grid voltage. Conduction stops only when the plate voltage falls to the de-ionization potential.

(2) The manner in which the grid and plate voltages of a typical thyratron control its functioning is shown in figure 93. The graph shows the firing plate voltage of an 885 thyratron for a range of values of negative grid voltage. For example, if the grid voltage is —20 volts, the tube will conduct or fire when the plate voltage is slightly under 160 volts. But if the grid voltage is raised to —12 volts, the tube

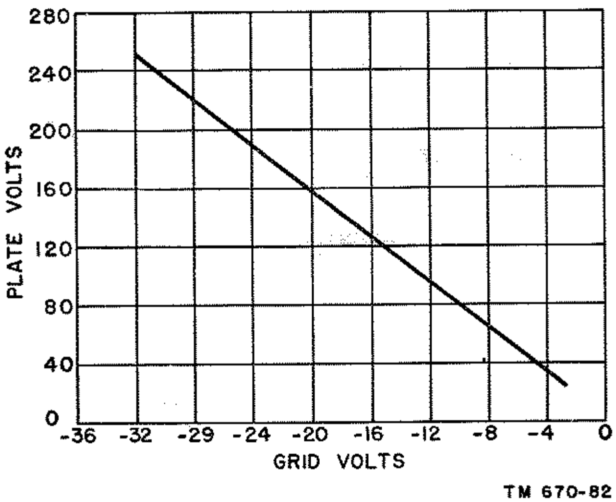


Figure 93. Plate-voltage, grid-voltage characteristic of 885 thyratron.

will fire with only 100 volts on the plate.

(3) The thyratron differs from the neon-gas tube also because the plate-current rating in thyratrons is much higher. Typical values range from 125 milliamperes in a small thyratron up to many amperes in a heavy-duty tube.

c. Practical Thyratron Oscillator.

(1) In A of figure 94, the thyratron sawtooth oscillator circuit resembles the diode-oscillator circuit of figure 90. This circuit is redrawn in B, as conventionally shown in schematic diagrams. The triode and diode circuits differ only in the values of the elements, and in the additional connection required for the bias voltage on the thyratron grid. For the circuit shown, the bias is —6 volts. The protective resistor is small, 120 ohms, because of the larger current which the thyratron can carry safely. For the operating conditions specified for this circuit, the rated peak current of the tube is 300 ma. A larger protective resistor is needed if the bias is increased, however, because this raises the firing voltage, as in figure 93. The values of the charging resistor and capacitor are chosen to fix the frequency of oscillation at 1,250 cycles.

(2) A thyratron oscillator can produce good sawtooth waves at higher frequencies than is possible with a neon-tube oscillator. The higher current rating of the thyratron enables it to discharge a capacitor many times more quickly than a neon tube discharges, and therefore, many more cycles can take place each second without violating the requirement that the discharge time be much less than the charge time. Practical thyratron oscillator frequencies range up to about 40,000 cycles. Above this frequency limit thyratrons de-ionize in erratic fashion, and other oscillator circuits are used. The frequency of this oscillator is increased if the value of either the capacitor or the resistor, or of both, is reduced. Practical minimum values of these elements for obtaining a very high frequency are .0002 μf for the capacitor and 200,000 ohms for the resistor. The frequency is lowered if the value of either the capacitor or the resistor, or of both, is increased. Practical maximum values of these elements for obtaining a very low frequency are 1 μf for the capacitor and 3 megohms for the resistor.

(3) As indicated in figure 92, a change of resistance or capacitance in the neon sawtooth oscillator has no effect on the amplitude of the sawtooth. Figure 95, however, shows that a charge of d-c grid bias of the thyratron sawtooth oscillator changes both frequency and amplitude of the sawtooth. A greater negative bias in the thyratron has these two effects: It increases the amplitude of the sawtooth voltage across the capacitor because the tube fires at a higher voltage, and it reduces the frequency because a longer time is required for the capacitor to charge to the higher firing potential. Conversely, a smaller negative bias in the thyratron reduces the output voltage and increases the frequency.

(4) Figure 96 shows the circuit of a practical thyratron oscillator, as used in

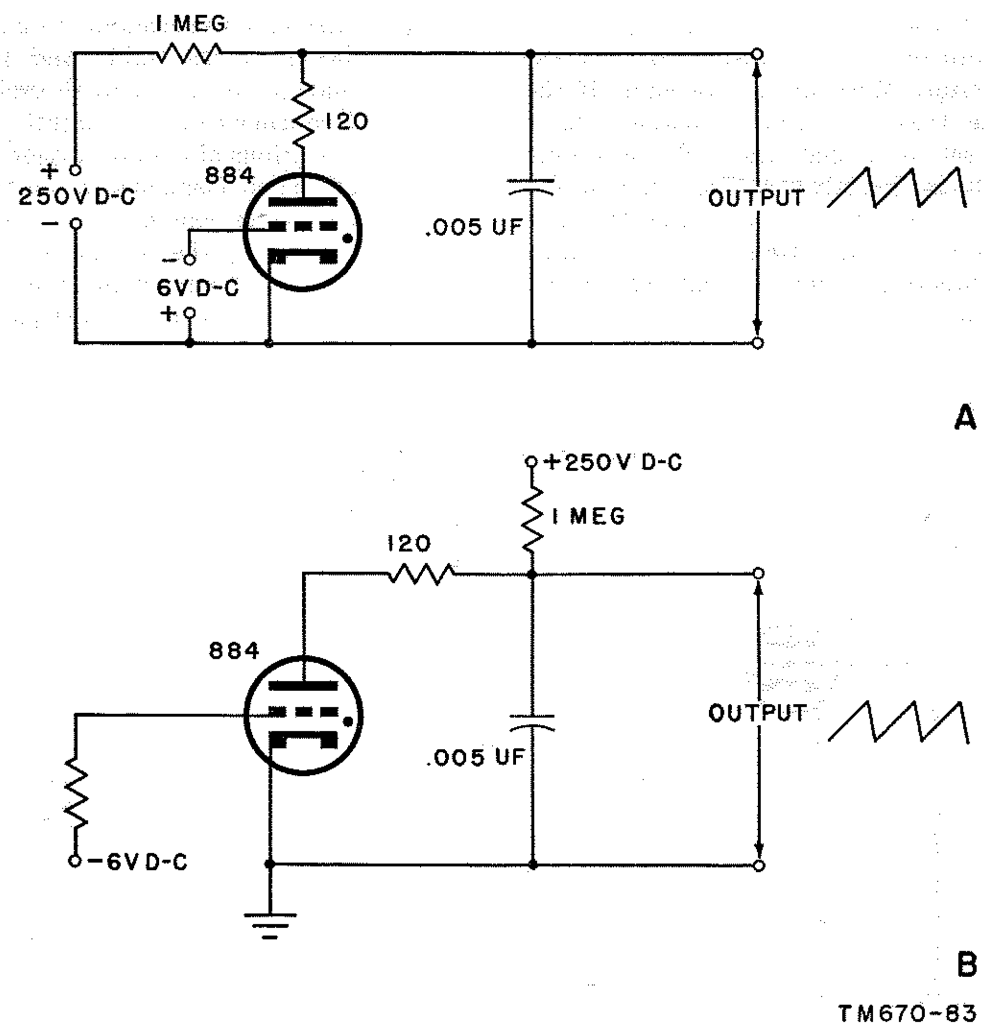


Figure 94. Thyratron sawtooth oscillator.

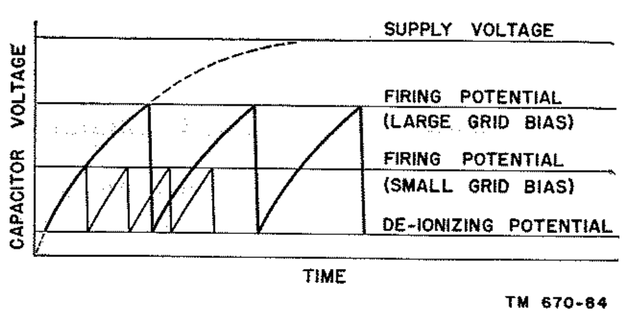


Figure 95. Change of amplitude and frequency of thyratron sawtooth generator by change of grid bias.

an oscilloscope time-base circuit. Provision is made for varying the frequency over a wide range. The differently valued frequency range is varied in steps by switching into the circuit one or another of the capacitors of the *coarse frequency selector.* Frequency is varied continuously in any one range by the potentiometer marked *fine frequency control.* The exact value of bias required depends on the output voltage and frequency range desired. Typical values range from —3 to —6 volts. Similar considerations determine the choice of the maximum resistance of the fine frequency control, which is typically between 1 and 3 megohms. An additional resistor, 200,000 to 750,000 ohms, depending on the requirements, is shown in series with the fine frequency control. Its purpose is to insure that the thyratron always goes out again once it has discharged the capacitor. Without this resistor, if the fine frequency control potentiometer is set at 0 ohm, the full d-c supply voltage, directly across the tube, would exceed the 16 volts at which the tube de-ionizes; the tube would continue conducting, and oscillations would stop.

(6) Although the circuits of figures 94 and 96 are more stable and predictable than the neon sawtooth oscillator, the frequency can vary slightly with small variations of supply voltage or heating of circuit elements. Even these small variations can be eliminated by synchronizing the oscillator with an a-c voltage of known constant frequency, as explained in the next paragraph.

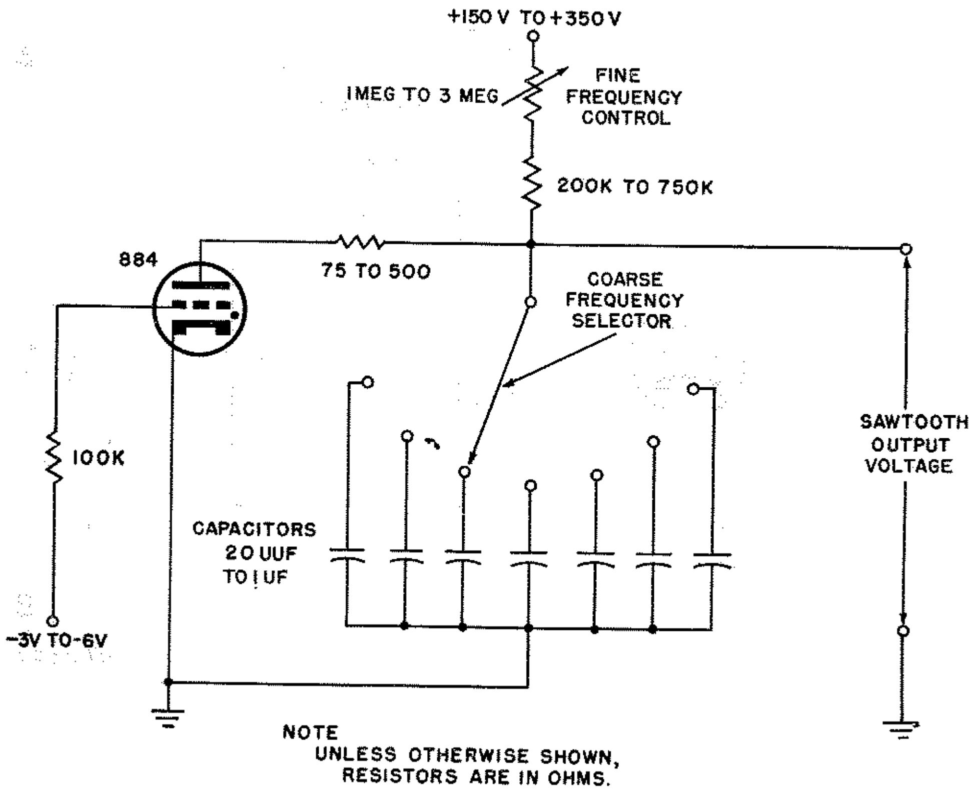


Figure 96. Time-base thyratron oscillator.

(5) Figure 95 shows that the rise in voltage of the sawtooth wave is not linear—that is, the voltage does not rise at a constant rate—but follows the exponential charging curve of the capacitor. Any small part of the lower portion of the curve is very nearly linear, however; and by using a d-c supply voltage that is much higher than the firing potential of the tube, operation is restricted to a small portion of the curve, and the output-voltage rise is sufficiently linear for most purposes.

116. Synchronized Sawtooth Oscillator

a. In time-base applications and in other electronic circuits, it is often necessary to use a sawtooth wave the frequency of which remains constant, or remains exactly in step, or in synchronism, with another frequency. The frequency at which the thyratron normally oscillates is called the *uncontrolled frequency,* or *free-running frequency.* However, if a small voltage of the desired frequency is injected at the grid of the thyratron, the circuit will oscillate at this frequency, instead of its free-running frequency. When this is done, the oscillator is said to be *synchronized* with the injected voltage.

b. The circuit of the synchronized thyratron oscillator is shown in A, figure 97. Its operation is as follows: The charging resistor, *R*1, and the capacitor, *C*2, are adjusted for a free-running frequency slightly lower than that of the *sync voltage* (synchronizing voltage). For example, if the sync-voltage frequency is 60 cycles, the free-running frequency of the thyratron is set to about 58 cycles.

c. The synchronizing action is illustrated in B. The d-c bias is set by potentiometer *R*2. Without the sync voltage, the firing potential is constant, and the tube would fire at point A, where the peak of the broken-line sawtooth intersects the constant firing potential line of the tube. When the synchronizing voltage is present on the grid, however, the firing voltage and the plate voltage vary in accordance with the grid signal. In B, it is assumed that the sync voltage is a sine wave, but this is not a necessary operating condition. At some time during the synchronizing cycle, the varying firing potential is low, so that the tube fires at a slightly earlier instant, represented by point B, where the peak of the unbroken-line sawtooth meets the varying firing-potential curve. Discharge takes place earlier, as shown by the unbroken-line sawtooth curve. On the next cycle, the capacitor voltage again reaches the firing point earlier than it would without the sync voltage, at the instant represented by point D on the unbroken-line sawtooth wave peak. The time of each oscillation thus is reduced from interval AC to interval BD and the oscillator is locked to the frequency of the injected a-c voltage. In a similar manner, the thyratron oscillator may be locked to a submultiple, or to a multiple, of the synchronizing voltage.

117. Thyratron Switch

a. General Characteristics. One of the many applications of thyratron oscillators is a circuit called a *thyratron switch*, in which two thyratrons are switched on and off alternately by an a-c input voltage. Strictly, this circuit is not an oscillator, since it delivers no output when the a-c input is removed. When the input voltage is applied, however, the thyratron switch de-

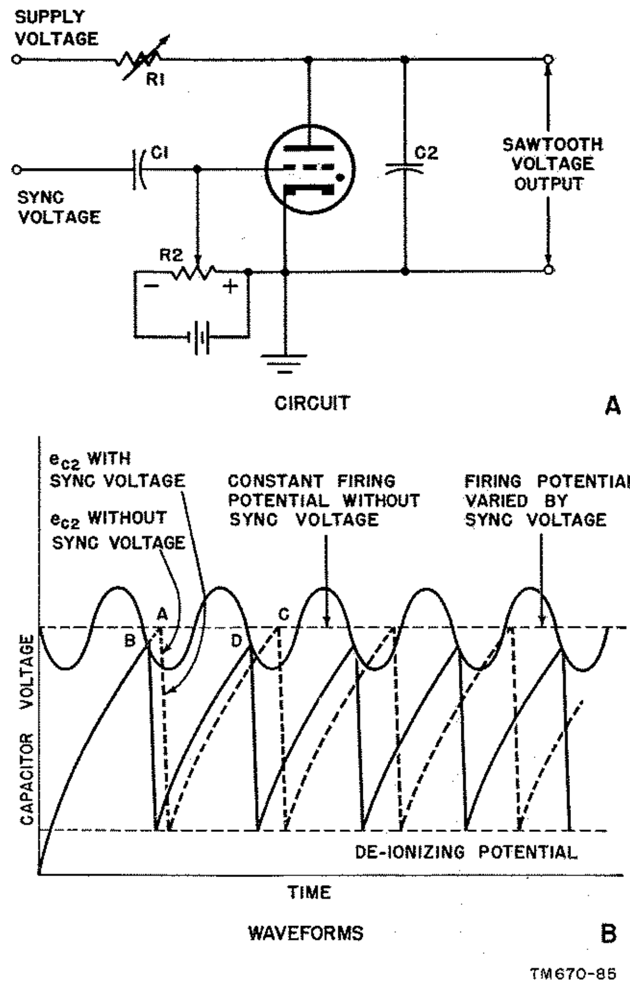


Figure 97. Synchronized thyratron saw tooth oscillator.

livers an output having a waveform substantially independent of the input waveform, much as a synchronized thyratron oscillator delivers a sawtooth output regardless of the waveform of the sync voltage. Both circuits depend on the charging and discharging of capacitors, controlled by the switching action of thyratrons. The characteristics of the two circuits differ in several respects. The frequency of the synchronized oscillator depends primarily on the circuit elements of the oscillator, and can be altered only within narrow limits by the sync voltage. The frequency of the thyratron switch is determined primarily by the frequency of the input voltage, and the output remains in synchronism with the input over wide ranges of frequency variation. The amplitude of the thyratron switch output is constant, regardless of amplitude variations at the input. Thyratron switches

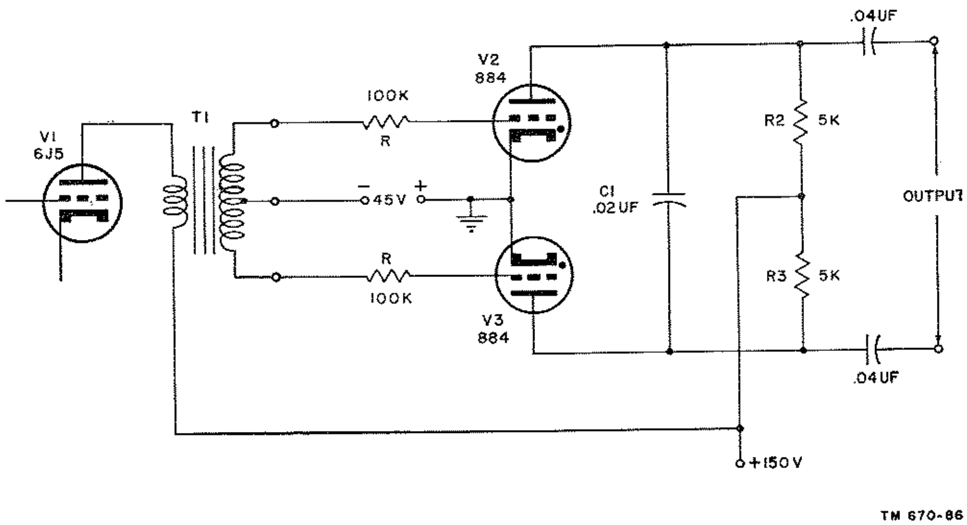


Figure 98. Circuit of thyratron switch.

are used in many military electronic equipments.

b. Circuit of Thyratron Switch. Figure 98 shows the circuit of the thyratron switch. The output of audio-amplifier tube *V*1 is coupled by transformer T1 to the grids of thyratrons *V*2 and *V*3 through protective resistors *R*. The d-c grid bias is applied between the center tap of the transformer secondary and the thyratron cathodes. Across the plates of the tubes is connected a capacitor, *C*1. The thyratron plates, *V*2 and *V*3, are each connected also to load resistors *R*2 and *R*3, respectively. To the junction of the two load resistors is applied the d-c plate potential for the two thyratrons.

c. Summary of Operation of Thyratron Switch (fig. 99). As will be shown, the output waveform depends primarily on the circuit, not on the input. Accordingly, the input voltage may be of almost any waveform. For purposes of this explanation, it is assumed that the input is a modified square wave, shown in A, derived from an audio amplifier tube *V*1 and applied across the primary of transformer T1 (fig. 98). The voltage induced in the secondary winding has approximately the same waveform and amplitude as the transformer input. The center-tapped secondary acts as a phase splitter, delivering two equal voltages, opposite-going in respect to the center tap, as in B and C. Thus during the first half-cycle of a-f voltage applied to the primary of the transformer, the grid of one thyratron tube becomes more positive and the tube fires, whereas the grid of the other tube becomes more negative. During the second half-cycle the polarities of the grid voltages reverse, and the second tube fires. The voltages at the plates of the thyratrons are substantially independent of the waveforms of the grid voltages, as in D and E. They are developed in a manner explained in detail below.

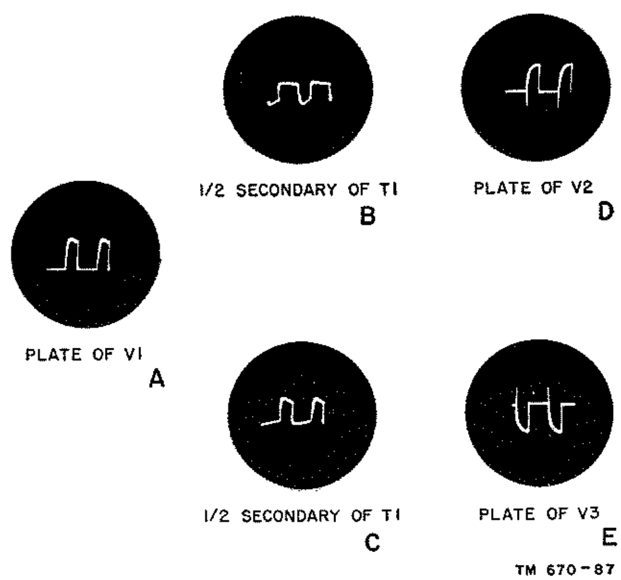


Figure 99. Waveforms in thyratron switch.

d. Step-by-Step Operation of Thyratron Switch. Both thyratrons are cut off when there is no signal input, because of the high negative d-c bias voltage applied to their grids (fig. 98). Under this condition, there is no voltage drop across resistors *R*2 and *R*3 because there is no flow of current through the resistors. As a result, both thyratron plates are at a potential of +150 volts. Also, there is no charge on capacitor *C*1 because there is no difference of potential across it. With these initial conditions, consider the sequence of events when an a-f signal voltage is applied to the transformer primary. The resulting voltages at the grids of thyratrons *V*2 and *V*3 are shown in A and B, respectively, of figure 100.

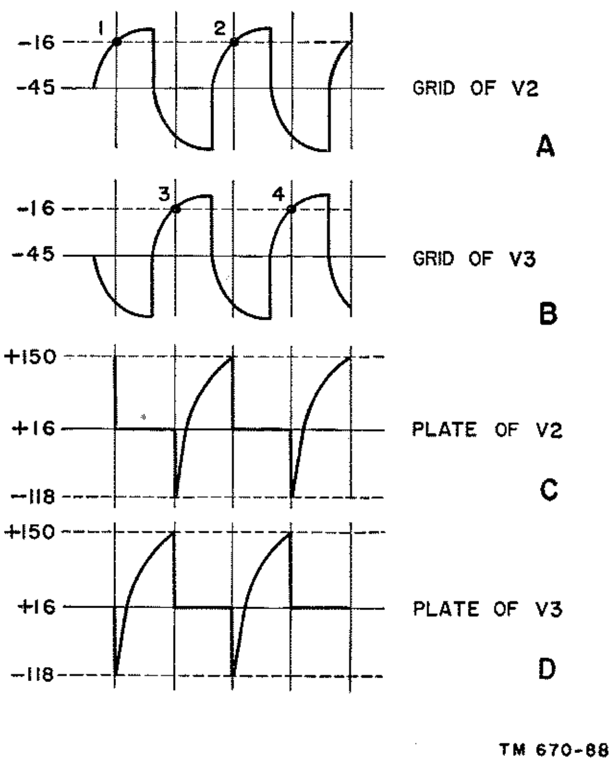


Figure 100. Theoretical waveform in thyratron switch.

(1) During the first half-cycle of a-f input voltage, the grid of tube *V*3 remains negative and this tube does not fire (fig. 98, and B, of fig. 100).

(2) During this time, the grid of tube *V*2 is driven more positive (fig. 98, and A of fig. 100). This reduces the difference of potential between the cathode and grid of the tube until this difference is approximately —16 volts. At the —16-volt level, *V*2 fires and plate current flows through resistor *R*2. The instant of firing is indicated by point 1 in A of figure 100.

(3) The waveforms of the voltages at the plates of *V*2 and *V*3 are determined by the plate-circuit elements *C*1, *R*2, and *R*3, together with the supply voltage and the thyratrons themselves (fig. 98). To clarify the discussion, the plate circuits are redrawn in figure 101, and the current paths indicated for four different instants during the cycle. In A, as tube *V*2 begins to conduct, electrons flow through the tube and resistor *R*2, along the part indicated by the unbroken line with arrows. The plate of the tube instantly drops as current starts to +16 volts which is the normal voltage drop across a type 884 tube during conduction (C, fig. 100). At this instant, the flow of current causes a voltage drop of 134 volts across *R*2.

(4) Once the flow of plate current is started, the grid voltage loses control of the plate current, and the plate current continues to flow as long as plate voltage is applied. This is shown in C by the constant level of plate voltage after the tube fires.

(5) As long as the plate voltage of *V*2 is at +16 volts, there is a voltage drop of 134 volts across *R*2 (C, fig. 100, and A, fig. 101). This voltage is applied across resistor *R*3 and capacitor *C*1 in series. The capacitor charges exponentially through *R*3. The charging path is indicated by the broken line with arrows. When the capacitor is fully charged, there is no voltage drop across *R*3, and the entire 134 volts appears across the capacitor, in the polarity shown. The capacitor plate connected to the plate of *V*2 is at +16 volts, and the other capacitor plate is at +150 volts.

(6) During the second half-cycle of a-c grid voltage, the voltage at the grid of tube *V*2 is negative, whereas the grid

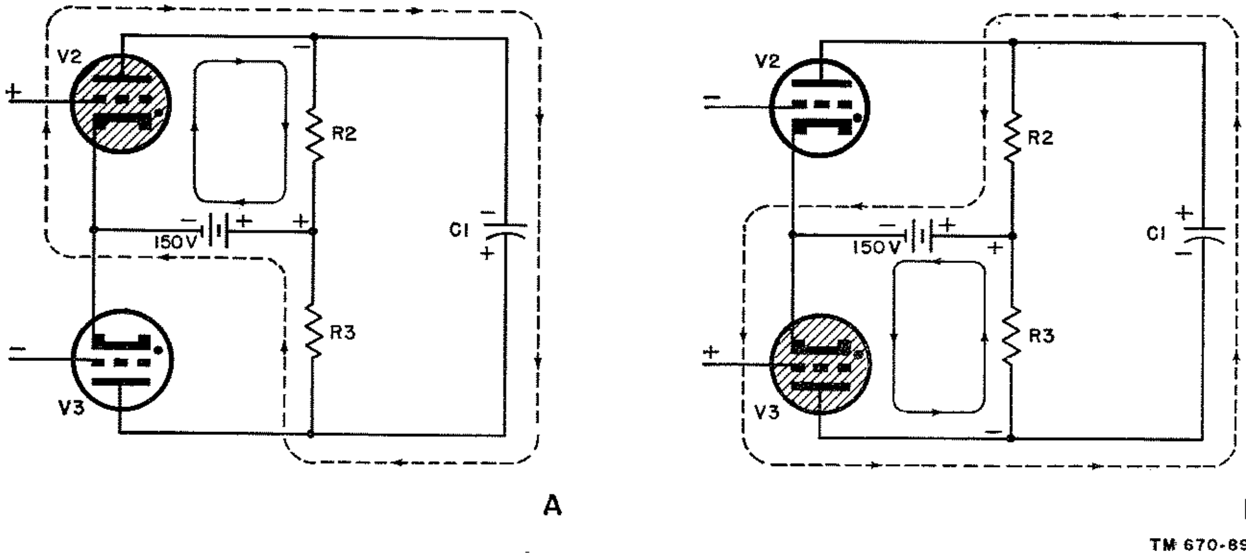


Figure 101. Functional diagrams of thyratron operation.

of tube $V3$ is driven more positive (A and B, fig. 100). When the grid voltage of tube $V3$ reaches the critical value of approximately —16 volts (point 3, in B), this tube fires and its plate voltage immediately drops to +16 volts, as in D, as current flows through resistor $R3$. The path of conduction is indicated by the unbroken line with arrows in B, figure 101.

(7) The fall of voltage at the plate of $V3$ from +150 volts to +16 volts (D, fig. 100) also places the lower plate of the capacitor at +16 volts. Since there is already a voltage of 134 volts across the capacitor, and since this voltage cannot change instantaneously, the upper plate of $C1$ falls to —118 volts. This is also the voltage at the plate of $V2$, which is immediately cut off by the negative plate voltage. The fall of the $V2$ plate to —118 volts is shown in C, figure 100.

(8) Simultaneously with these sudden voltage changes and the firing of tube $V3$, a voltage drop of 134 volts appears across $R3$ in the polarity shown in B, figure 101. This voltage is thus applied across $R3$ and $C1$ in series, in a polarity that adds to the 134 volts already existing across the capacitor. The capacitor, therefore, discharges through $R3$ and charges again in the opposite polarity, through the path indicated by the broken line. The polarity of the capacitor voltage during recharging is as shown. The plate voltage of $V2$ follows the exponential rise of the capacitor voltage from —118 volts to +150 volts (C, fig. 100). After the capacitor has attained its charge, the voltage at the plate of tube $V2$ again reaches +150 volts, but this tube does not fire until the next cycle because it is held below cut-off by the negative grid-bias voltage.

(9) As the a-c voltage across the primary of the transformer goes through its positive and negative alternations, the cycle repeats, the tube alternately firing and producing at the plate of each tube the pulsating voltage of constant amplitude and waveform shown in C and D.

Section II. VACUUM-TUBE OSCILLATORS

118. Multivibrators

a. General. Multivibrators operate on the relaxation principle which consists of building up, or storing, energy in a capacitor and then, when a certain level of voltage is reached, discharging the capacitor. Multivibrators find wide applications as pulse generators, frequency-dividing circuits, pulse counters, electronic switches, gating circuits, and time-delay circuits. A more detailed discussion of multivibrators that is given in this section is presented in TM 11–672.

b. Fundamental Characteristics Used in Multivibrators. The following fundamental characteristics of vacuum-tube circuits are repeated at this point as an aid in understanding the action of multivibrator circuits:

(1) A rise in grid voltage (in the positive direction) causes an increase in plate current through the tube; conversely, a fall in grid voltage causes a decrease in current.

(2) An increase of current through the load resistor causes the voltage on the plate of a tube to decrease. A current decrease results in a higher plate voltage.

(3) The polarity of the voltage which appears across a resistor can be determined by remembering that electron flow is from negative to positive.

(4) The voltage across a capacitor cannot change instantaneously.

(5) A capacitor requires a certain period of time to charge or discharge. This time depends on the time constant of the circuit.

119. Plate-to-Grid Free-Running Multivibrator

a. Basic Principles.

(1) The basic type of multivibrator is the plate-to-grid, or plate-coupled, free-running multivibrator shown in figure 102. This is a two-stage, resistance-capacitance coupled amplifier with the output of the second stage coupled back to the input of the first stage.

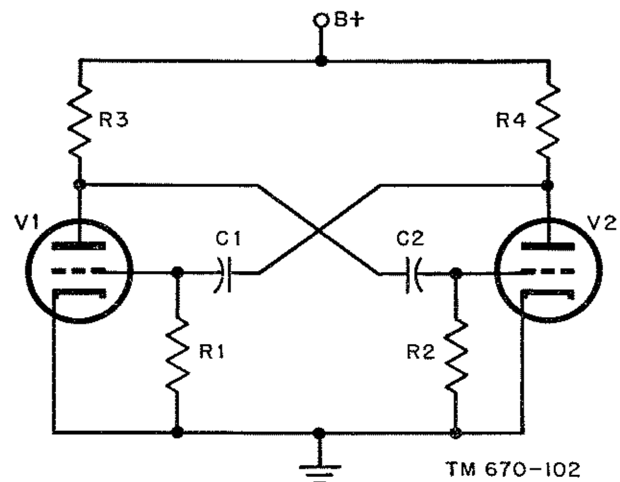


Figure 102. Plate-coupled multivibrator.

(2) Each stage of a resistance-capacitance coupled amplifier causes a polarity reversal in signal between plate and grid circuits; that is, a positive change in grid voltage causes a negative change in plate voltage, and vice versa. The phase shift caused by the R-C components usually is negligible and can be neglected. Therefore, the signal fed back from the plate of the second stage to the grid of the first stage has the same polarity and phase as the original signal. For example, a positive signal is applied to the grid of *V*1. This produces an amplified negative signal at the plate of *V*1. The negative signal is applied to the grid of *V*2. This causes a further amplification of the signal which appears as a positive voltage at the plate of *V*2. The positive voltage is fed back to the grid of *V*1. Consequently, the feedback is regenerative—that is, it reinforces the original signal, and oscillation can occur.

b. Operation of Circuit (fig. 102).

(1) When the power-supply voltage first is applied to the multivibrator, plate current begins to flow in each tube. In addition, the capacitors take on charge as the plate voltages increase. The initial currents that flow are ap-

proximately equal to each other. However, a perfect circuit balance is impossible. Any small circuit dissimilarity will cause the plate current of one tube to be slightly larger than the plate current of the other tube.

(2) It is the slightly greater plate current in one tube that starts the multivibrator action. The unbalance between plate currents of the tubes becomes greater, and the cumulative effect continues until the plate current in one tube reaches a maximum value and the plate current in the other tube is cut off.

(3) Assume that the initial plate current of tube $V1$ in figure 102 becomes slightly greater than the plate current of tube $V2$. This increase causes the voltage drop across load resistor $R3$ to increase, causing the voltage at the plate of tube $V1$ to decrease. The voltage at the plate of tube $V1$ is across the circuit composed of $C2$ in series with $R2$. Since the charge across $C2$ cannot change instantaneously, the full decrease in voltage of tube $V1$ appears across resistor $R2$ with such polarity as to drive the grid of $V2$ more negative.

(4) This decrease in grid voltage decreases the current flowing in the plate circuit of tube $V2$. The drop in voltage across $R4$, because of the decrease in plate current, increases the plate voltage of tube $V2$. The increase in plate voltage of tube $V2$ is applied to the circuit composed of $C1$ in series with $R1$. Since the charge in $C1$ cannot change instantaneously, the full increase of plate voltage appears across resistor $R1$. This resistor is connected between the grid and cathode of tube $V1$, so that its grid voltage is increased.

(5) The increase in grid voltage of tube $V1$ further increases the plate current of $V1$ and amplifies the action outlined above. Thus the unbalance is amplified, so that larger changes in voltages occur. This continues until $V2$ is completely cut off and $V1$ plate current is at maximum. Therefore, the slight initial unbalance starts a cumulative feedback action that cuts off one tube and causes maximum plate-current conduction in the other tube. Although the foregoing description might create the impression that the action takes place slowly, the truth is that the action is extremely fast—in fact, the action might occur in a fraction of a microsecond.

(6) At this point, with $V2$ cut off and tube $V1$ conducting at maximum, an action begins that sets the stage for a switching action. Coupling capacitor $C2$ must discharge now, since the plate-to-cathode voltage of tube $V1$, which is applied to the $R2$-$C2$ circuit, has been abruptly reduced. Capacitor $C2$ discharges through $R2$ and the grid voltage of $V2$ rises exponentially toward zero. When the grid voltage which is moving in a positive direction reaches the cut-off point, $V2$ fires (begins to conduct).

(7) The increase in plate current of $V2$ causes a decrease in plate-to-cathode voltage in the tube. This abruptly drives the grid of $V1$ negative, since the voltage across $C1$ cannot change instantaneously. This is a cumulative action, precisely the same as occurred during the previous half-cycle. Tube $V1$ now is cut off and tube $V2$ is conducting heavily.

(8) Since tube $V1$ is cut off, no current flows through resistor $R3$ and the plate-to-cathode voltage of tube $V1$ rises to the value of the B+ supply; this causes $C2$ to charge rapidly through the low-resistance path between the cathode and grid of $V2$. A complete cycle has now occurred, which can be briefly summarized:

  (*a*) Tube $V1$ conducts while $V2$ is cut off.

  (*b*) A switching action occurs where $V1$ is cut off and $V2$ conducts.

  (*c*) A new cycle starts with $V1$ conducting and $V2$ cut off, and so on.

(9) The output voltage taken from either plate is, therefore, a series of rectangular pulses. The pulses have a maximum amplitude when the tube is conducting heavily.

120. Waveforms and Equivalent Circuit of Plate-to-Grid Multivibrator

a. Equivalent Circuit.

(1) The analysis of multivibrator waveforms is greatly simplified through the use of equivalent circuits for the plate-to-grid resistance-capacitance coupling networks. Figure 103 shows these equivalent circuits during the charge and discharge periods of each R-C network. The equivalent circuits for the charge periods are different from those for the discharge periods. This is true because the charge path is through the low resistance cathode-to-grid circuit of the tube and the discharge path is through the high resistance of the grid resistor, the cathode-to-grid path being closed during nonconduction periods of the tube.

(2) For example, during the charging period of *C*1, the grid-to-cathode resistance, R_{g1}, of tube *V*1 is very low (A, fig. 103). Since this resistance is much lower than *R*1, the charging current flows mainly through this resistance, and *R*1 can be neglected. During this time, tube *V*2 is cut off, so that the plate resistance, R_{p2}, is infinite.

(3) During the discharge period of this capacitor, *C*1, the grid-to-cathode resistance, R_{g1}, is high compared with *R*1 and can be neglected (C, fig. 103). Tube *V*2 is conducting, and R_{p2} has a finite value. Therefore, in the equivalent circuit for the *C*1 discharging period, the plate-to-cathode resistance, R_{p2}, of tube *V*2 is included in the circuit. Therefore, as indicated in A and C, the equivalent circuits for the charge and discharge periods of these R-C circuits are different.

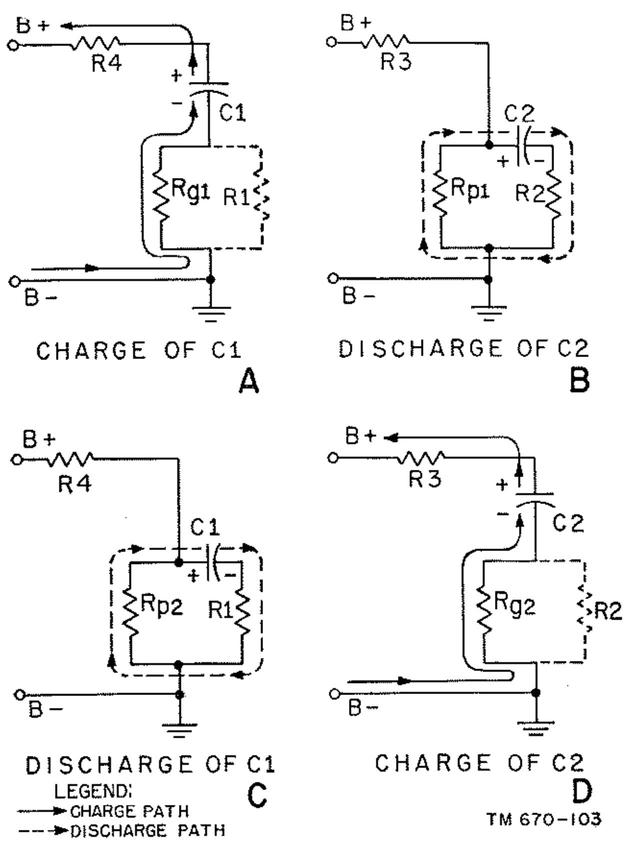


Figure 103. Multivibrator equivalent circuits.

b. Waveforms of Symmetrical Multivibrator.

(1) *General.* The waveforms in figure 104 are for a *symmetrical* or *balanced* multivibrator. A symmetrical or balanced multivibrator is one in which the circuit is completely balanced; that is, time constants *R*1 *C*1 and *R*2 *C*2, the tubes used, and the corresponding applied voltages are the same. Therefore, the conducting and nonconducting times for each tube are the same. Other types of multivibrators are used that are not symmetrical and, in these circuits, the periods of conduction of the two tubes are not the same.

(2) *Initial plate-voltage and current conditions.* The plate- and grid-voltage waveforms of the plate-coupled multivibrator begin with the assumption that the current in *V*1 is at maximum and the current in *V*2 cut off. This is indicated in A and E, figure 104, from $t = 1$ usec (microsecond) to $t = 2$ usec. During this time, the plate voltage of tube *V*1 is minimum and the

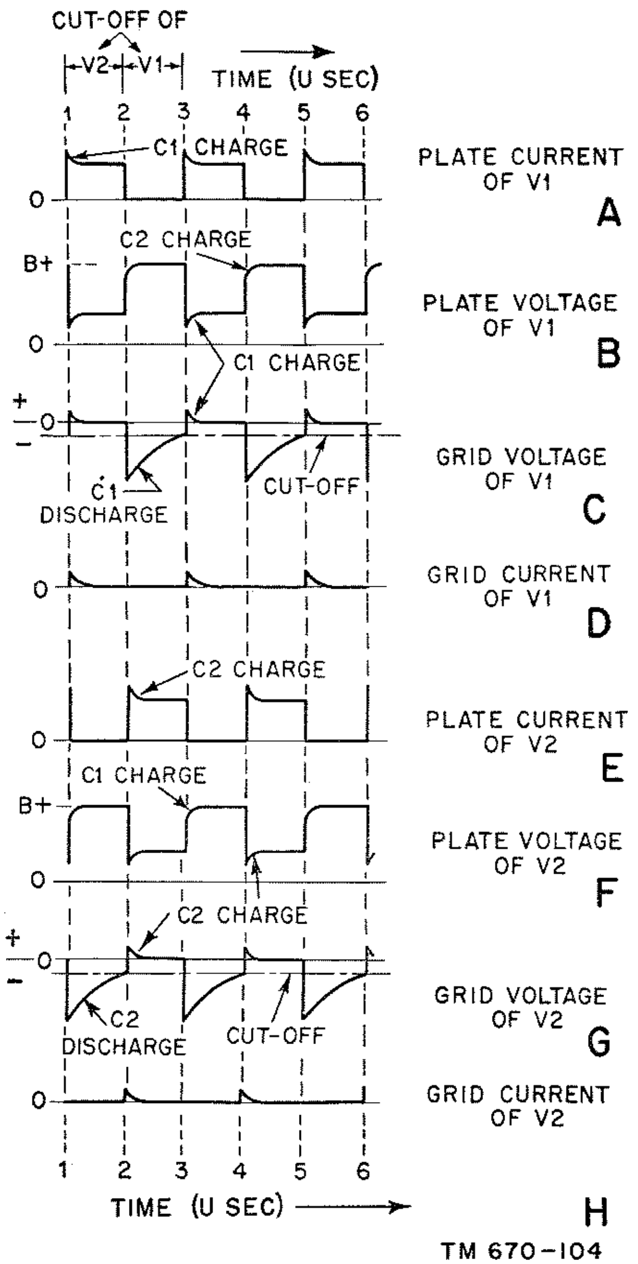


Figure 104. Waveforms in symmetrical plate-coupled multivibrator.

plate voltage of tube $V2$ is maximum, as shown in B and F, respectively.

(3) *Initial capacitor conditions.* At $t =$ 1 usec, $C1$ is being charged through the grid-to-cathode resistance, R_{g1}, to the value of the applied voltage. At this time there is no voltage drop across $R4$, since $V2$ is cut off, and the full voltage is applied. The capacitor charges very rapidly, and is fully charged within a small fraction of the total conduction period of $V1$. At this same time, $C2$ begins to discharge slowly through $R2$ and R_{p1}. The equivalent circuits in A and B, figure 103, show the current paths.

(4) *Initial grid current and voltage.* The grid voltage of $V1$ increases to a positive value momentarily, as in C. This lowers the grid-to-cathode resistance and provides a low-resistance path to charge $C1$. During this time, the grid draws current, as shown in D. As the capacitor becomes fully charged, the grid voltage and current drop rapidly to 0 and remain at 0 for the entire $V1$ conduction period. Simultaneously, the grid voltage of tube $V2$ is driven well beyond cut-off, as in G, and there is no grid-current flow at this time. The grid voltage of $V2$ starts to rise at $t = 1$ usec as $C2$ discharges through $R2$ and R_{p1}.

(5) *Waveforms from $t = 1$ usec to $t = 2$ usec.* Between $t = 1$ usec and $t = 2$ usec, tube $V2$ is cut off, and therefore, the plate current and the plate voltage of tube $V2$ remain essentially constant during this period, as in E and F. Similarly, since the grid voltage of tube $V1$ is held at 0 volt for this period, the plate voltage and the plate current of tube $V1$ remain essentially constant for the short pip during the charge of $C1$ at $t = 1$ usec, as in A and B. The only factor that is changing during this period is the grid voltage of tube $V2$. This is increasing in the positive direction in accordance with the discharge curve of capacitor $C2$, as in G. As indicated in the equivalent diagram (B, fig. 103), this capacitor discharges through $R2$ and R_{p1}. However, since $R2$ is usually much larger than R_{p1}, the time constant of the circuit is primarily a function of the value of $R2$ $C2$.

(6) *Tube V2 conditions at $t = 2$ usec.* At $t = 2$ usec the grid voltage of tube $V2$ exceeds the cut-off value and tube $V2$

begins to conduct. Within a fraction of a microsecond, the plate current of tube $V2$ reaches its maximum value; the plate voltage drops to the minimum value, and the grid voltage goes positive momentarily. Grid current flows, and then both grid voltage and current drop to 0 (E to H, fig. 104).

(7) *Tube V1 conditions at $t = 2$ usec.* When tube $V2$ starts to draw current, the grid voltage of $V1$ becomes negative, causing a reduction in plate current and an increase in plate voltage. Within a fraction of a microsecond, the grid voltage of tube $V1$ is driven beyond cut-off, the plate voltage reaches a maximum, and the plate current is cut off (A to D, fig. 104).

(8) *Capacitor conditions at $t = 2$ usec.* At $t = 2$ usec, the grid voltage of $V2$ becomes positive and capacitor $C2$ is charged through the low grid-to-cathode resistance, R_{g2}, to the full B+ voltage. At this time there is no drop across $R3$, since $V1$ is cut off so that the full B+ voltage appears at the plate. The capacitor charges very rapidly, as in G, and is fully charged within a small fraction of the total conduction period. The charging current flows through R_{g2}, which has some forward resistance. As a result, the charging current produces on the grid a small positive voltage. This positive voltage is amplified and inverted on the plate of $V2$. This explains the small peaks shown in the waveforms. $C1$ begins to discharge slowly through $R1$ and R_{p2}, as shown in the equivalent circuits C and D.

(9) *Waveforms between $t = 2$ usec and $t = 3$ usec.* Between $t = 2$ usec and $t = 3$ usec, $V1$ is cut off, and therefore its plate current and plate voltage remain essentially constant. Similarly, with the exception of small pips at $t = 2$ usec caused by the charging of $C2$, the plate voltage and plate current of tube $V2$ remain essentially constant over this same period. The only factor that is changing during this period is the grid voltage of tube $V1$. This is increasing, as in C, in accordance with the discharge curve of capacitor $C1$. Since $R1$ is usually much larger than R_{p2}, the time constant of the discharge curve is primarily a function of $R1\ C1$.

(10) *Start of new cycle at $t = 3$ usec.* At $t = 3$ usec, the grid voltage of $V1$ exceeds the cut-off value, and the entire cycle starts again with the conditions outlined at $t = 1$ usec.

(11) *Duration time.* The duration time of the symmetrical multivibrator depends on the time required for the grid voltage of the cut-off tube to reach the cut-off value. The cut-off period depends on how far beyond cut-off the grid is driven, the cut-off point of the tube, and the time constant of the capacitor discharge circuit. If the grid voltage is far beyond cut-off, a long time is required for this voltage to recover to the cut-off value. The value of grid voltage beyond cut-off depends on the i_pR_L drop of the conducting tube. The higher the drop, the greater is the decrease in plate-to-cathode voltage and the more negative is the grid voltage swing. The more rapidly the capacitor discharges, the sooner the grid reaches the cut-off value. The discharge time constant is primarily a function of $R1\ C1$ or $R2\ C2$. Therefore, the duration time of the symmetrical multivibrator is essentially directly proportional to the difference between the negative grid voltage swing and the cut-off voltage and the time constant of the discharge circuit of the grid capacitor. *A period of the multivibrator is the sum of the two duration times.*

121. Sine-Wave Synchronization of Multivibrators

Free-running multivibrators do not have good frequency stability, since a change in tube characteristics, capacitors, or resistors affects the frequency of oscillation. To obtain a highly stable frequency, it is necessary to synchronize or *lock in* the multivibrator frequency with a sine wave or a pulse series of stable frequency. The synchronizing voltage is often called the *sync* signal.

a. Basic Principles.

(1) Assume that a sine-wave synchronizing voltage is applied to the grid of tube *V*1 in a typical plate-coupled free-running multivibrator (fig. 102). A, figure 105, shows the grid-voltage waveform of this tube. At $t = 0$, the grid voltage has just been driven beyond cut-off. Without a sine-wave synchronizing voltage applied to the grid, 5 usec are required for the capacitor to discharge sufficiently to allow the grid to reach the cut-off value.

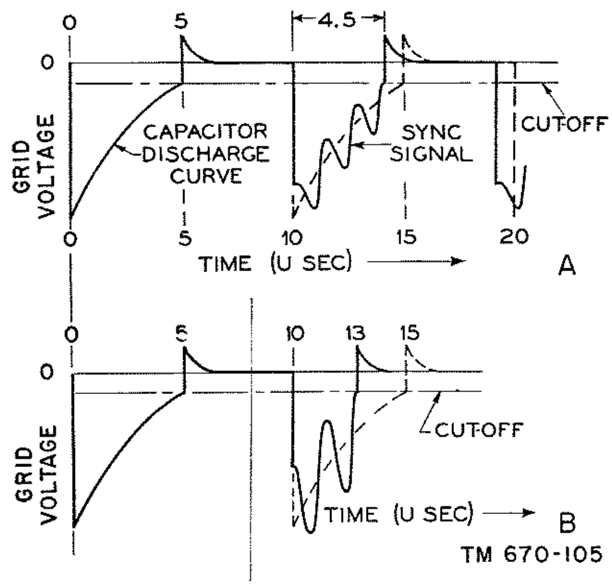


Figure 105. Sine-wave synchronization of multivibrator.

(2) However, when the sine wave is applied to the grid, the grid reaches the cut-off voltage in 4.5 usec because the positive alternation of the sine-wave signal adds to the grid voltage, as shown in B. Similarly, on the next cycle, the grid reaches the cut-off voltage 4.5 usec after the capacitor begins to discharge.

(3) Therefore, the multivibrator output voltage is synchronized with the sine-wave voltage applied to the grid of *V*1. The synchronizing signal, in A, causes the free-running multivibrator frequency to increase slightly when it is locked in with the frequency of the synchronizing signal. The frequency of the synchronizing signal is three times the frequency of the synchronized half-cycle of the multivibrator, and therefore, six times the frequency of the multivibrator. Conversely, the multivibrator frequency is one-sixth the synchronizing signal frequency. For example, assume that the free-running multivibrator frequency is normally 12,500 cps. A synchronizing sine wave whose frequency is 76,800 cps is applied to the circuit. The multivibrator frequency then is increased to 76,800/6 or 12,800 cps.

b. Effect of Increasing Sync-Signal Amplitude. The ratio of multivibrator frequency to sync signal can be changed by changing the sync-voltage amplitude. For example, if the sync-signal amplitude is increased, the cut-off point of the tube is reached at an earlier time (B, fig. 105). With an increase in sync-signal amplitude, the frequency of the multivibrator is increased. In B, the increase in sync-signal amplitude causes the cut-off value to be reached after the second cycle of the sync signal. The dashed curve in this figure is the grid-voltage waveform that exists in the absence of the sync signal. The frequency of this multivibrator is only one-fifth of the sync frequency, since three sync cycles still occur during the conduction period. If the sync frequency is 76,800 cps, the multivibrator frequency is 76,800/5, or 15,360, for the grid-voltage waveform shown in B.

c. Synchronization Applicable to all Types of Multivibrators. So far, the synchronization of a symmetrical plate-coupled multivibrator has been described. In a similar manner any type of free-running multivibrator, cathode-coupled or plate-coupled, symmetrical or assymmetrical, can be synchronized to a sine-wave input voltage. Also, the sync voltage does not necessarily have to be applied to the grid of the tube but can be applied as readily to the cathode of the tube. In this case, the grid-to-cathode voltage, and therefore, the time required to reach cut-off, again are partially governed by the sync voltage. However, the polarity of the sync signal in the cathode circuit is opposite to that used in the grid circuit.

122. Multivibrator Synchronized by Pulses

a. Basic Principles. A more common and more effective method of synchronizing multivibrators is by means of a series of pulses. This method is more effective because there is a sudden, well-defined change in voltage when a pulse is superimposed on the grid voltage (fig. 106). In sine-wave synchronization, the change in voltage is gradual; consequently, the cut-off value may be reached at slightly different times in successive cycles of operation. However, in pulse synchronization, the time at which the grid rises above cut-off is very definite, and therefore, the frequency is more stable.

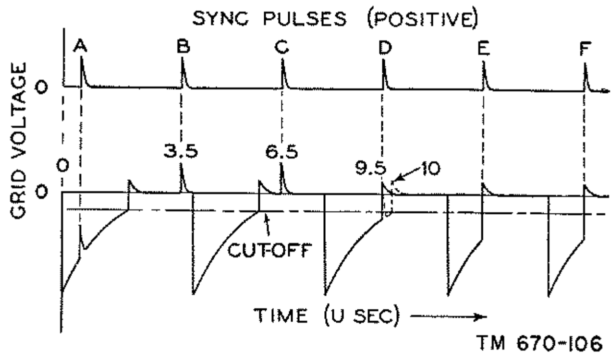


Figure 106. Waveforms at grid of multivibrator synchronized by positive pulses.

b. Positive Pulse Synchronization.

(1) Figure 106 shows the synchronization of a multivibrator output by means of positive pulses applied to the grid of tube *V*1 of a symmetrical multivibrator. Only pulses that cause the tube to go from a nonconducting to a conducting state affect the frequency and duration of the multivibrator output. For example, when pulse *A* is applied to the circuit, it does not increase the grid voltage above cut-off, so that the nonconduction period remains the same. Pulses *B* and *C* occur during the conduction periods of the tube. These pulses do not affect the frequency or duration of the multivibrator output other than to increase momentarily the grid and plate currents.

(2) Pulse *D,* however, causes the grid voltage to rise above cut-off at a time earlier than it would in the absence of the pulse. Therefore, the grid voltage rises above the cut-off value at t = 9.5 usec, instead of at t = 10 usec. Only those pulses which cause the tube to go from the cut-off to the conducting state affect the frequency of the multivibrator. For positive synchronizing pulses, synchronization can occur only during the cut-off periods of the input tube.

c. Negative Pulse Synchronization.

(1) Multivibrators also can be synchronized by means of negative sync pulses. When a negative pulse is superimposed on the grid voltage of tube *V*1, it produces synchronization when tube *V*1 is conducting. This is opposite to the action of the circuit for a positive sync pulse. One method of producing synchronization is to cause the sync pulse to have sufficient amplitude to cut off the conducting tube when it is applied. This method, however, is not generally used. A second method does not require the sync pulses to have sufficient amplitude to cut off the input tube directly.

(2) To understand the action of the circuit when negative sync pulses are used, it is necessary to consider the grid voltage waveforms of both tubes.

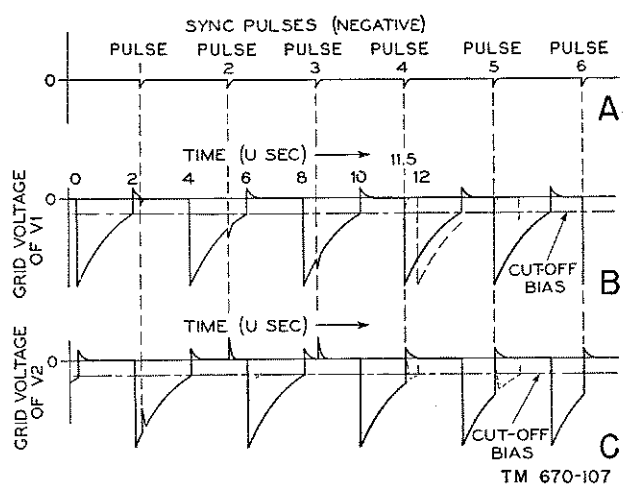


Figure 107. Waveforms at grid of multivibrator synchronized by negative pulses.

Figure 107 shows the grid-voltage waveforms of a typical multivibrator with negative sync pulse input. When a negative signal is applied to the grid of tube *V*1 while it is conducting, a positive signal appears at the grid of tube *V*2.

(3) The first negative pulse (A, fig. 107) applied to the grid of tube *V*1 reduces the grid voltage of this tube. This reduction is not sufficient to drive the tube to cut-off, as in B. The pulse is amplified and inverted by tube *V*1, whose output is applied to the grid of tube *V*2. The negative pulse therefore appears as an amplified positive pulse at the grid of tube *V*2, as shown in *C*. This pulse does not have sufficient amplitude to raise the grid voltage of *V*2 above cut-off; therefore, operation of the multivibrator is not affected.

(4) Pulses 2 and 3 are applied to the grid of tube *V*1 during the period of time when this tube is cut off. These pulses in no way affect the circuit since there is no change in the plate current of *V*1, that tube already being cut off.

(5) Negative pulse 4 reduces the grid bias of tube *V*1 but not sufficiently to cut this tube off. However, the amplified positive pulse that appears on the grid of tube *V*2 at this time causes the grid voltage to rise above cut-off, as shown in C. In the absence of this pulse, tube *V*2 remains cut off until $t = 12$ usec. However, the presence of the pulse causes this tube to conduct at $t = 11.5$ usec.

d. Multivibrator Frequency.

(1) *Sync frequency slightly higher than multivibrator frequency.* In the operation shown in figures 106 and 107, a number of cycles occurred before synchronization took place. However, after $t = 9.5$ usec (fig. 106) and $t = 11.5$ usec (fig. 107), each cycle of the multivibrator is synchronized to the frequency of the sync pulse. The multivibrator frequency equals the sync frequency. The increase in frequency is obtained by reducing the duration of one half of 1 multivibrator output cycle. The multivibrator waveforms shown are symmetrical without the synchronizing pulses and symmetrical after the synchronization occurs. The positive half-cycle of grid voltage has the same duration after synchronization, but the negative half-cycle duration is reduced.

(2) *Sync frequency a multiple of multivibrator frequency.*

(*a*) It is also possible to synchronize the multivibrator by sync pulses that are a multiple of the natural multivibrator frequency. In figure 108, the multivibrator already is synchronized by the sync pulses. Every third pulse causes the grid voltage to exceed cut-off value. Consequently, pulses 1, 4, 7, and 10 cause synchronization of the multivibrator

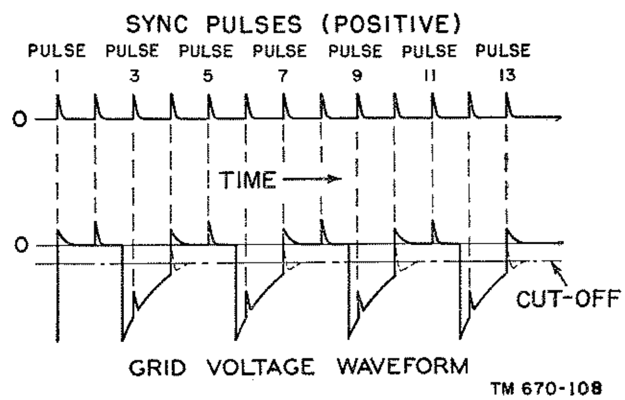


Figure 108. Synchronization of multivibrator by sync frequency three times multivibrator frequency.

(*b*) Pulses 2, 5, and 8 occur when the tube is conducting and these pulses have no effect on the frequency of the multivibrator. Pulses 3, 6, and 9 are applied during the cut-off period of the tube. However, the amplitude of the pulses applied at this time is not sufficient to raise the grid voltage above cut-off. Therefore, these pulses do not affect the multivibrator frequency.

(c) In this case, the multivibrator frequency is equal to one-third the sync frequency. For example, if the sync frequency or repetition rate is 150 kc, the multivibrator frequency is 50 kc.

(3) *Multivibrator frequency a multiple of sync frequency.* It also is possible to obtain synchronization when the multivibrator frequency is an integral or whole number multiple of the sync frequency. For example, assume that the multivibrator frequency is 150 kc and that the sync frequency is 75 kc, 50 kc, or 30 kc. When this occurs, only the particular cycle of oscillation being synchronized is controlled by the sync signal. If the multivibrator frequency is 150 kc and the sync frequency is 50 kc, every third multivibrator cycle is controlled by the sync pulse. The other cycles between the synchronized cycles tend to *fall in line.*

123. Blocking Oscillator

a. General. The blocking oscillator is another type of relaxation oscillator. Figure 109 shows a free-running type from which a continuous series of pulses is obtained at the output. With suitable modification, the blocking oscillator finds wide application in pulse generators, sweep generators, and in counter circuits. A detailed discussion is given in TM 11–672.

b. Basic Principles. Essentially, the circuit consists of a vacuum tube, usually a triode, and a transformer to provide regenerative feedback from the plate to the grid circuit. Since the

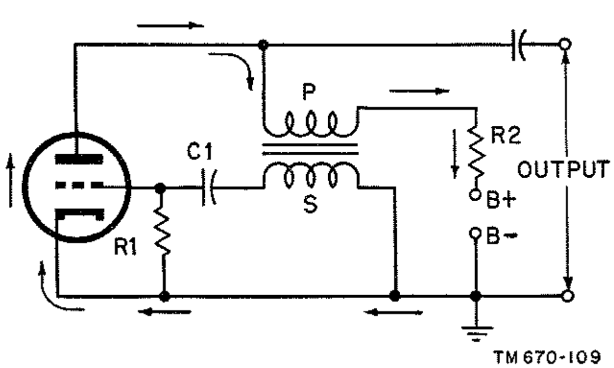
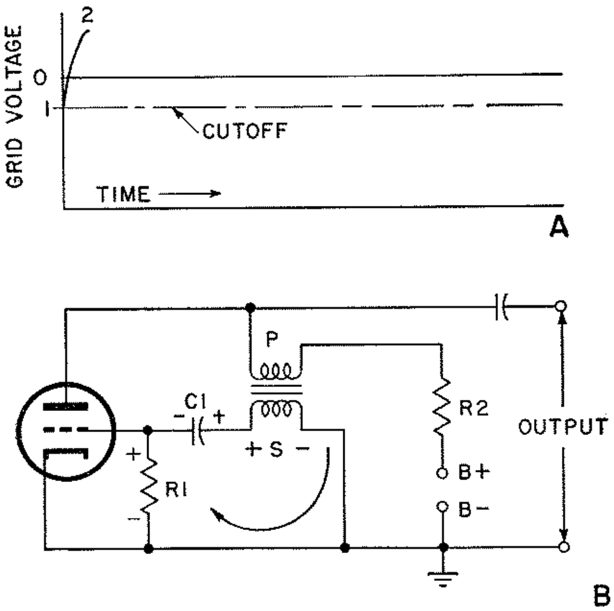


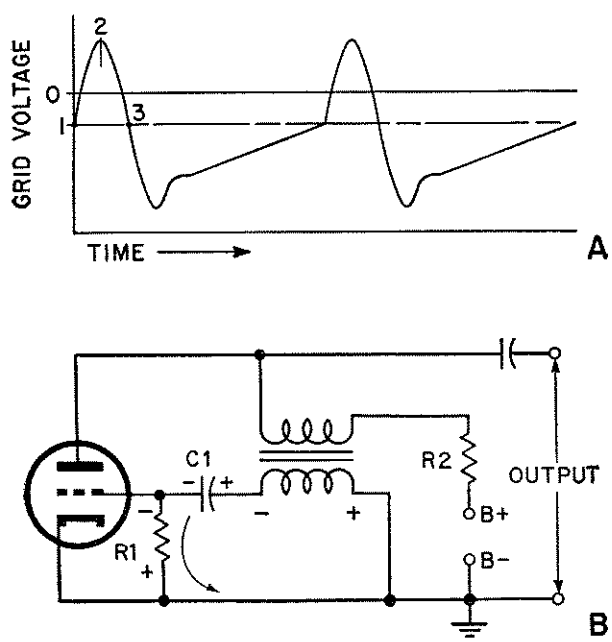
Figure 109. Free-running blocking oscillator.

grid is initially at cathode potential (with no voltage drop across $R1$), plate current flows in the direction shown in figure 109 when the plate supply voltage is applied. This current, which is increasing in value from zero to a maximum, causes an increasing voltage drop in the transformer primary, P, and hence an induced secondary voltage. The resulting secondary current begins to charge the capacitor through $R1$ (fig. 110). The voltage drop developed across $R1$ places the grid of the tube at a rising positive potential. This causes a proportional rise in plate current and also starts grid current flowing. As soon as grid current starts flowing the capacitor charge path is through the tube since the cathode-to-grid resistance is very small compared with $R1$, when the tube is conducting. Because of this low-resistance charging path, the charge time is very short. The entire action is cumulative until a further increase in grid voltage fails to cause the plate current to increase. This condition is called saturation. The momentary steady value of plate current (point 2, fig. 110), which occurs at saturation, causes the transformer field to collapse. This results in a high secondary voltage of reverse polarity (negative in respect to ground) which causes a very rapid partial discharge of capacitor $C1$ (fig. 111). The voltage developed across $R1$ drives the tube beyond cut-off and blocks the tube. The capacitor, $C1$, continues to discharge through $R1$ at a rate determined by the time constant $R1$ $C1$. The grid voltage increases (becomes less negative) until it reaches cut-off, when the entire cycle starts again. The blocking oscillator may be recognized by the reader as being similar to the ordinary regenerative, or tickler coil, oscillator, with the exception of $C1$, which blocks the tube for a number of natural cycles. The dotted line of figure 112 shows the sine wave that would result if the $R1$ $C1$ time constant were not deliberately made large. The shape of the half-cycle (output pulse) between 1 and 3, figure 112, and the dotted portion of the curve is determined by the natural resonant frequency of the transformer with its distributed capacitance and circuit capacitances. However, since the tube is blocked during most of the complete cycle, the output waveform is nonsinusoidal (B, fig. 112). The period of the blocking oscillator is the sum of the pulse duration time and the recovery time. The two prime factors involved are the $R1$ $C1$ time constant and the maximum negative grid voltage. The frequency which is the reciprocal of the period may be controlled by making $R1$ variable. An example of this control is the *hold* control on a television receiver.



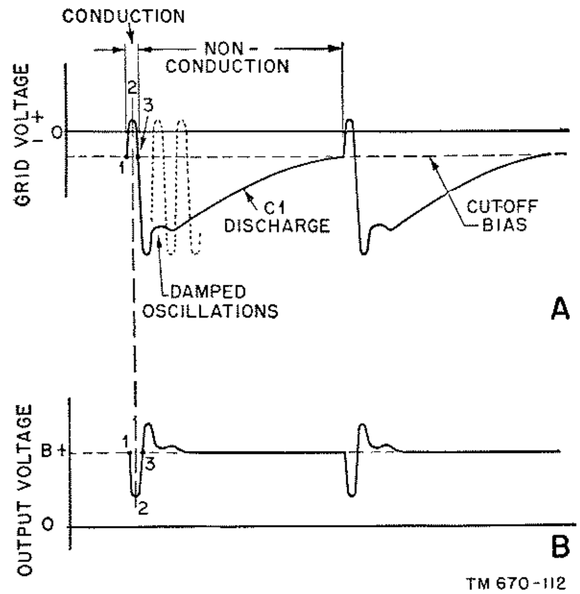
TM 670-110

Figure 110. Charge-path circuit of blocking oscillator.



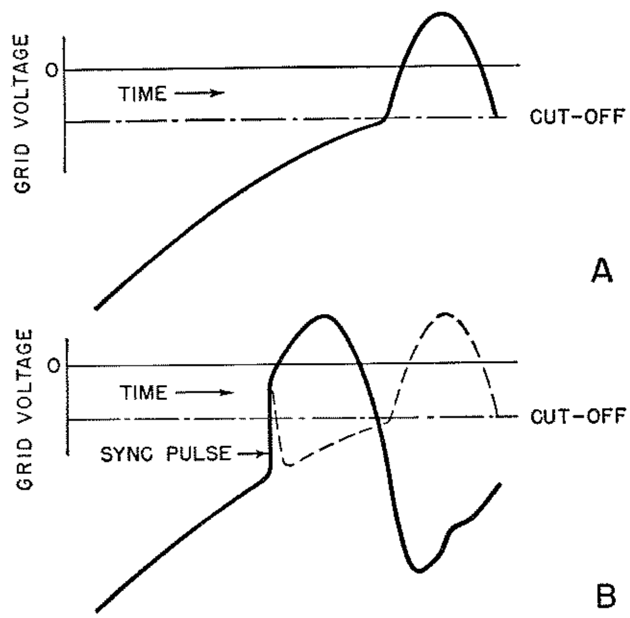
TM 670-111

Figure 111. Discharge-path circuit of blocking oscillator.



TM 670-112

Figure 112. Grid and output waveforms of blocking oscillator.



TM 670-113

Figure 113. Grid waveform of positive pulse synchronization.

c. Synchronization. The blocking oscillator may be synchronized by inserting a positive pulse at the proper time. If this pulse is applied to the grid circuit just before the grid reaches the cut-off voltage and is of sufficient magnitude to drive the grid above cut-off, the tube starts conducting at an earlier time, as in figure 113.

The output of a blocking oscillator, therefore, can be synchronized to a pulse series applied to the grid circuit. The blocking oscillator cycle is synchronized to each sync pulse that changes the tube from a cut-off to a conducting state. The sync frequency, therefore, should be slightly higher than the blocking-oscillator frequency or a multiple thereof (A and B, fig. 114).

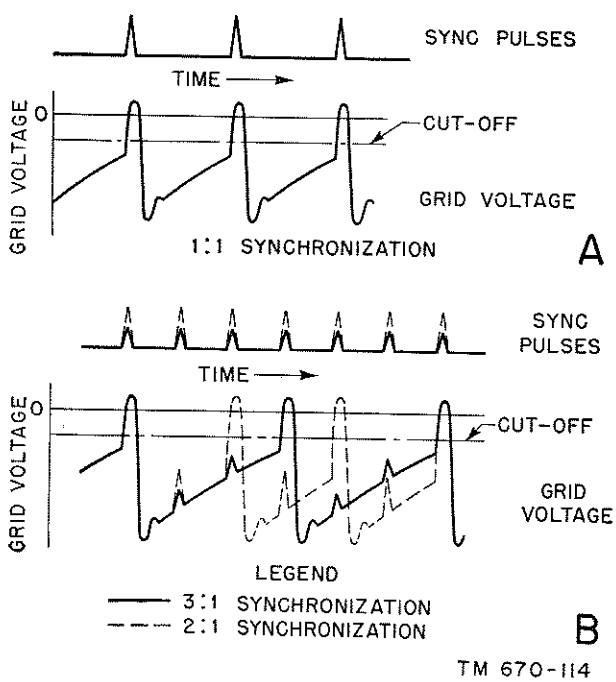


Figure 114. Multiple synchronization of blocking oscillator.

124. Summary

a. The waveforms in circuits employing gas tubes are nonsinusoidal and are characterized by sudden changes in instantaneous voltage or current values. These sudden changes result because a gas tube is either conducting or not conducting. A gas tube can be cut off only by reducing the voltage across it to a low value.

b. Sawtooth voltages may be generated by oscillators employing gas diodes or gas triodes (thyratrons). The output voltage in both circuits is taken across a capacitor which is charged through a resistance from a d-c source and discharged through the tube.

c. Important characteristics of thyratrons are their relatively high peak plate current rating and the ability of the grid to determine and control the firing voltage. The grid has no control over the magnitude of the plate current, nor can it stop conduction.

d. The characteristics of gas tubes are subject to considerable variations from tube to tube, and they vary with the age of the tube. This is especially true of glow tubes.

e. Gas-diode oscillators are less stable in frequency than thyratron oscillators. The latter may be synchronized by means of a small voltage of constant frequency, applied to the thyratron grid.

f. The thyratron switch circuit employs two thyratrons. It delivers an alternating output voltage of constant amplitude, which is locked to the input frequency over wide frequency ranges. The output waveform is substantially independent of the input waveform.

g. Multivibrators find wide application as pulse generators, frequency-dividing circuits, pulse counters, electronic switches, gating circuits, and time-delay circuits.

h. A plate-coupled, free-running multivibrator is a two-stage resistance-capacitance coupled amplifier with the output of the second stage coupled back to the input of the first stage.

i. The multivibrator is the most common type of relaxation oscillator.

j. The frequency and the duration of a free-running multivibrator are primarily a function of the R-C time constant.

k. A free-running multivibrator can be synchronized by a sine-wave sync signal.

l. Synchronization can occur at a sync frequency slightly greater than the multivibrator frequency or at a sync frequency that is a multiple of the multivibrator frequency.

m. More stable synchronization of multivibrators can be effected with pulse sync signals.

n. The blocking oscillator is a type of relaxation oscillator.

o. The transformer, used in a blocking oscillator circuit, provides regenerative feedback from the plate to the grid circuit of the tube.

p. The duration of the blocking oscillator output pulse is, primarily, a function of the resonant circuit formed by the transformer.

q. The frequency of the blocking oscillator is the reciprocal of the sum of the pulse duration time and the recovery time.

r. A blocking oscillator can be synchronized by a sync pulse that raises the grid voltage above cut-off.

s. A blocking oscillator sync frequency may be slightly higher than the blocking oscillator frequency or a multiple thereof.

125. Review Questions

a. Draw a schematic diagram of a glow-tube oscillator.

b. Why do glow tubes have few applications as control elements in electronic circuits?

c. Draw a diagram of a thyratron sawtooth oscillator.

d. What are two important characteristics of thyratrons?

e. Explain the principal application of a sawtooth oscillator.

f. To what extent does the grid exercise control over the plate current in a thyratron tube?

g. Draw a schematic diagram of a synchronized thyratron sawtooth oscillator.

h. Explain the operation of a thyratron switch.

i. Sketch the circuit of a plate-coupled free-running multivibrator.

j. Explain the operation of this multivibrator circuit.

k. Draw the equivalent circuits for the charge and discharge periods of each R-C network.

l. Why are the equivalent circuits of the resistance-capacitance coupling networks different for the charging and discharging periods?

m. On what does the duration time of a symmetrical multivibrator depend?

n. What is the period of a multivibrator?

o. Show how a free-running plate-coupled multivibrator is synchronized to a sine wave whose frequency is slightly higher than the natural multivibrator frequency.

p. How does increasing the sync amplitude affect the multivibrator frequency?

q. What is the advantage of pulse synchronization of multivibrators over sine-wave synchronization?

r. Sketch the circuit of a free-running blocking oscillator.

s. Explain the operation of a free-running blocking oscillator.

t. What is the purpose of the transformer in a blocking oscillator circuit?

u. Explain the synchronization of a blocking oscillator to a positive pulse.

APPENDIX

LETTER SYMBOLS

1. Plate Circuit

e_b instantaneous total plate voltage

e_{b1} instantaneous total plate voltage of tube $V1$ (e_{b2}—of tube $V2$ . . . etc.)

E_b total plate voltage, average

E_{bb} d-c plate-supply voltage

E_{bo} quiescent or zero signal, average value of plate voltage

e_p instantaneous value of varying component of plate voltage

e_L instantaneous total voltage across the load resistor

e_z instantaneous varying component of voltage across the load impedance

E_{Lo} quiescent or zero signal, average value of d-c voltage across the load impedance

E_{zo} varying component of average voltage across the load impedance

i_b instantaneous total plate current

I_b average total plate current

I_{bo} quiescent or zero signal, average value of plate current

i_p instantaneous value of varying component of plate current

r_p a-c plate resistance of an electron tube, defined as the ratio of a small change in plate voltage to the small change in plate current which it produces: $r_p = \frac{\triangle e_p}{\triangle i_p}$ (with e_g constant)

μ amplification factor of an electron tube, defined as the ratio of a small plate voltage change to the grid voltage change which will produce the same change in plate current: $\mu = \frac{\triangle e_p}{\triangle e_g}$ (with i_p constant)

2. Grid Circuit

$e_c (e_{c1})$ instantaneous total control-grid voltage

e_{c2} instantaneous total screen-grid voltage

g_m transconductance, also called mutual conductance, a figure of merit for an electron tube, defined as the ratio of a small change in plate current to the small change in grid voltage that produces it: $g_m = \frac{\triangle i_p}{\triangle e_g}$ (with e_p constant).

e_{c3} instantaneous total suppressor-grid voltage

E_{c1} average total control-grid voltage

E_{cc} control-grid d-c supply voltage

E_{c2} screen-grid d-c supply voltage

e_g instantaneous value of varying component of control-grid voltage

$i_c (i_{c1})$ instantaneous total control-grid current

i_{c2} instantaneous total screen-grid current

i_{c3} instantaneous total suppressor-grid current

3. Cathode Circuit

e_k instantaneous total cathode voltage

E_k average total cathode current

i_k instantaneous cathode current

I_k average d-c cathode current

4. Miscellaneous

e_{in} instantaneous input signal voltage

e_{out} instantaneous output signal voltage

E_{max} maximum value of d-c voltage

I_{max} maximum value of d-c plate current

P_1 input power to a system

P_2 output power from a system

P_o d-c output power

P_i d-c input power

P_p plate dissipation

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