# 7
# AMPLIFIERS

***The principal applications of transistors and vacuum tubes are based on their ability to amplify electric signals. Some circuits amplify minute voltage signals by factors of many million, while others increase the electric power of a signal in order to operate a mechanical device such as an electric motor. Still other circuits amplify currents. In each of these applications the frequency range of the input signal is important. Different circuits have been developed for dc amplification and for use at high radio frequencies.***

***Most often, the signal level is increased in several successive amplifier stages to attain the desired output signal magnitude. In this case, the interaction of amplifier stages must be considered and fairly complicated networks are involved. Fortunately, the techniques of circuit analysis developed in previous chapters, in particular the ac equivalent circuits for tubes and transistors, are sufficient for a satisfactory understanding of complete amplifier circuits.***

## VOLTAGE AMPLIFIERS

### 7-1 Cascading

The transistor and vacuum-tube circuits discussed in previous chapters are ideally suited to amplify voltage signals with minimum waveform distortion. Gain factors greater than those possible with a single-stage amplifier are obtained by *cascading* several amplifier stages. The output of one amplifier stage is amplified by another stage or stages until the desired signal voltage level is achieved.

Consider, for example, the two-stage cascaded triode amplifier, Fig. 7-1. Two individual circuits similar to those discussed in

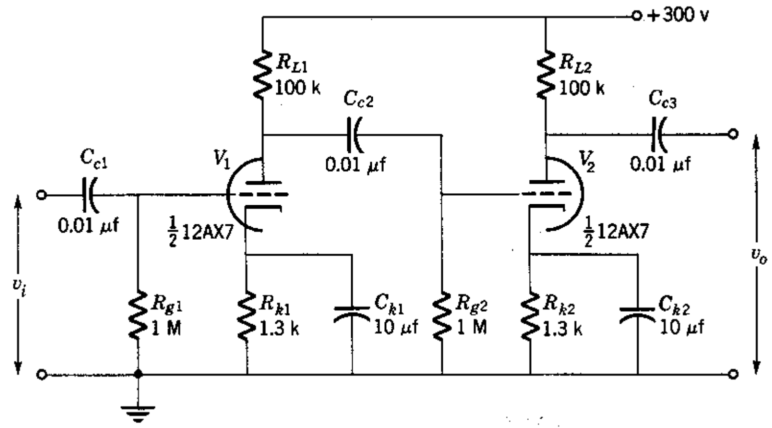



***FIGURE 7-1*** *Two-stage amplifier using triodes connected in cascade.*

Chap. 5 are connected with the coupling capacitor $C_{c2}$. This capacitor passes the amplified ac signal from $V_1$ to the grid of $V_2$. At the same time it blocks the positive plate voltage of $V_1$ from the grid of the second triode. Similarly, capacitors $C_{c1}$ and $C_{c3}$ isolate the input and output circuits insofar as dc potentials are concerned.

The entire ac equivalent circuit of this amplifier may be drawn using the principles discussed in Chap. 5. The performance of the system is determined from a complete ac-circuit analysis. Actually, this procedure is unwieldy because of the number of loops in the network and is rarely attempted. Rather, the circuit is analyzed in several separate steps, each of which has a minimum of mathematical complexity. This has the further advantage that the important effects can be isolated and more clearly examined.

For example, the reactances of both cathode bypass capacitors are assumed small enough to be negligible. Accordingly, these components are absent in the ac equivalent circuit of the amplifier, Fig. 7-2. The reactances of the coupling capacitors are also ignored, even though they are included in the equivalent circuit for clarity. With these simplifications, the output voltage may be immediately written as

$$v_o = v_{g2}\,\frac{-\mu_2}{1 + r_{p2}/R_{L2}} = v_{g1}\,\frac{-\mu_1}{1 + r_{p1}/R'_{L1}} \times \frac{-\mu_2}{1 + r_{p2}/R_{L2}} \tag{7-1}$$

where the load resistance of the first stage $R'_{L1}$ is the parallel combination of the plate resistor $R_{L1}$ and the second-stage grid resistor $R_{g2}$.

According to (7-1), the overall gain of this two-stage amplifier is simply

$$a = a_1 a_2 \tag{7-2}$$

where $a_1$ and $a_2$ are the gains of each stage. The expression for the gain of the first stage $a_1$ includes the input impedance of $V_2$ as

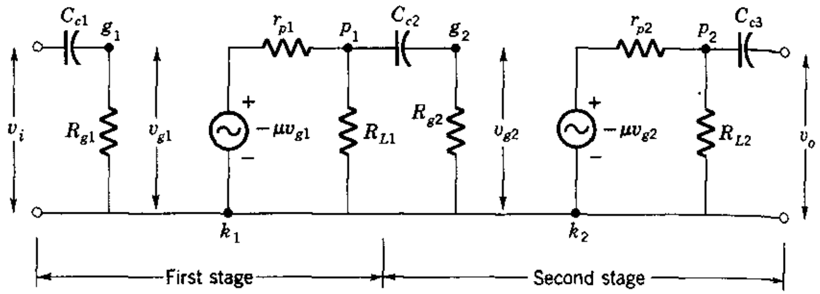



***FIGURE 7-2*** *Equivalent circuit of the two-stage cascaded amplifier of Fig. 7-1.*

part of the load resistance. Usually, $R_{g2} \gg R_{L1}$, however, so $a_1$ is essentially the gain of the isolated $V_1$ stage. Equations (7-1) and (7-2) give the *midband gain* of the amplifier since the reactances of the coupling capacitors and the cathode bypass capacitors are assumed negligible. This approximation applies to signal frequencies which are neither so low that the reactances cannot be ignored nor so high that other effects reduce the gain. These other frequency regions are discussed in the following section.

Cascaded transistor voltage amplifiers most often employ the grounded-emitter configuration because of the combined voltage and current gain of this circuit. Neither the common-base nor the emitter-follower configuration achieves as great overall voltage amplification when cascaded. This is a result of the great impedance mismatch between the output impedance of one stage and the input impedance of the succeeding stage. A typical transistor voltage amplifier, Fig. 7-3, uses interstage coupling capacitors as in the vacuum-tube case to isolate the dc bias voltages of the two stages. Note that the bias resistors of the second stage are different from those of the first stage, even though both transistors are identical. The operating points are set at different places in order to obtain most favorable values of the $h$ parameters in each stage.

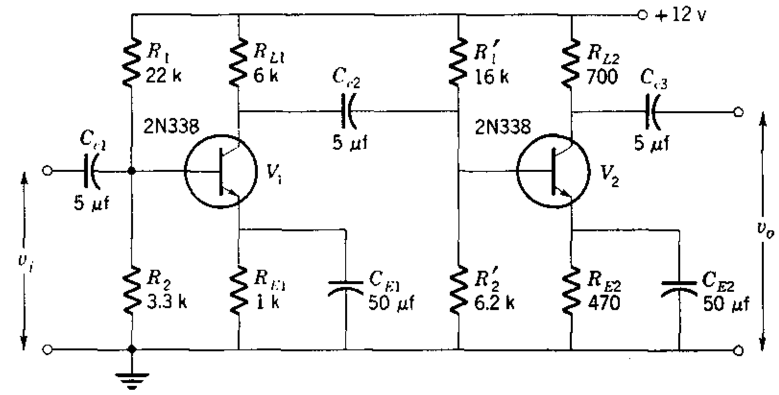



***FIGURE 7-3*** *Two-stage cascaded transistor amplifier.*

In the appropriate equivalent circuit of the amplifier, Fig. 7-4, the base bias resistors are replaced by their parallel combination, as explained in the previous chapter. Here again, the overall gain

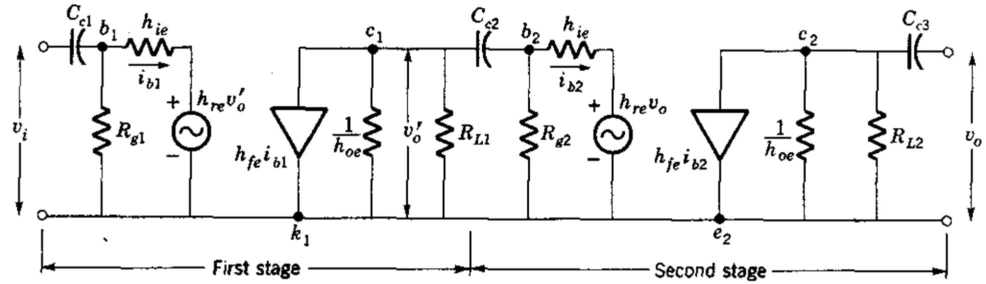



***FIGURE 7-4*** *Equivalent circuit of the two-stage transistor amplifier of Fig. 7-3.*

is the product of the individual gain of each stage, Eq. (7-2). The loading effect of the second stage upon the output of the first stage cannot be ignored in calculating the gain because of the inherently low input impedance of transistors. In fact, it is necessary to work backward through the circuit starting at the output terminals because the input impedance of a transistor amplifier depends upon the output load impedance. With the output load specified, the gain and input impedance of the second stage are calculated using the results developed in the previous chapter. This input impedance is part of the load for the preceding stage and its gain and input impedance are calculated accordingly. Thus, detailed analysis of transistor circuits is somewhat more complicated than is the case for vacuum-tube amplifiers, basically because of the input-output coupling in a transistor. Nevertheless, transistor circuits are treated quite satisfactorily by straightforward ac-circuit analysis of the equivalent circuit.

## 7-2 Low-frequency gain

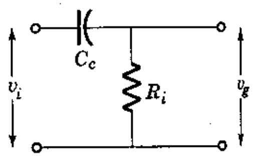
At sufficiently low frequencies the capacitive reactances may no longer be neglected. The effect of the coupling capacitors is usually of greater significance than that of the cathode bypass capacitors, although both tend to reduce the gain at low frequencies. It is not practical to make the coupling capacitors large. Large values of capacitance imply increased leakage current, which upsets grid bias of vacuum tubes. This situation is aggravated by the fact that the coupling capacitor is connected between the large positive plate potential and the low grid voltage and by the large value of grid resistance. Consequently, practical coupling capacitors are limited to values below about 0.5 $\mu$f.

No such restrictions are placed on cathode bypass capacitors since they are connected in low-impedance, low-voltage circuits where leakage currents are insignificant. Electrolytic capacitors are common in this position and values ranging up to 100 $\mu$f are used. Special low-voltage electrolytics are used in transistor amplifiers as both coupling capacitors and bypass capacitors since the impedance levels are low in both places. Nevertheless, leakage currents must be minimized into the base terminal so a limit to the capacitance exists in this case as well. The low-frequency gain of cascaded transistor amplifiers is also determined primarily by the reactance of the interstage coupling capacitors.

Because the overall gain of cascaded stages is the product of individual stage gains, it is only necessary to examine the effect of the coupling capacitor reactance for an isolated amplifier stage. According to the ac equivalent circuits of both vacuum-tube and transistor amplifiers, Figs. 7-2 and 7-4, this effect can be treated by considering the simple *RC* circuit comprising the coupling capacitor and the input impedance of the amplifier.

This part of both equivalent circuits is shown separately in Fig. 7-5 for clarity. The input impedance $R_i$ is simply the grid resistor

*FIGURE 7-5*




in the case of the tube amplifier, but in the transistor amplifier it includes the input impedance of the transistor itself. In either case, the output voltage of the stage is

$$v_o = av_g = a\frac{v_i}{R_i + 1/j\omega C_c}R_i$$

$$= \frac{av_i}{1 - j/\omega R_i C_c} \tag{7-3}$$

As discussed in Chap. 2, it is appropriate to define the characteristic frequency

$$2\pi f_0 = \omega_0 = \frac{1}{R_i C_c} \tag{7-4}$$

Substituting Eq. (7-4) into Eq. (7-3), the gain $v_o/v_i$ is

$$a(f) = \frac{a}{1 - jf_0/f} \tag{7-5}$$

where $a$ is the midband gain. Note that the gain is reduced when the signal frequency is smaller than the characteristic frequency. At the same time a phase shift is introduced between the input and the output signals. Both effects are important in determining the waveform distortion of the amplifier. Recall that in the Fourier analysis of a complex signal waveform, both the amplitudes and relative phases of all frequency components must be preserved if the output wave is to be an amplified replica of the input signal.

It is convenient to rationalize Eq. (7-5),

$$a(f) = \frac{a}{\sqrt{1 + (f_0/f)^2}} \tag{7-6}$$

so that the gain at any frequency can be immediately calculated. Note that Eq. (7-6) shows that the gain is $a/\sqrt{2}$, or about 70 percent of the midband gain, when $f = f_0$. It is important to recognize that Eq. (7-6) applies to an individual stage. The low-frequency response of the entire amplifier is always poorer than that of any individual stage because the gain of cascaded stages is the product of individual stage gains.

## 7-3 High-frequency gain

The high-frequency gain of any amplifier is reduced by stray capacitive effects that are not purposely made part of the circuit. Referring to a simple triode amplifier, Fig. 7-6, these are the grid-cathode capacitance $C_1$, the grid-plate capacitance $C_2$, and the plate-cathode capacitance $C_3$ of the tube itself. Also included in $C_1$ and $C_3$ are stray capacitances between the wires and components attached to the grid and plate terminals. All three of these capacitors shunt the signal to ground at frequencies high enough that the capacitive reactances are significant.

The effect of the grid-plate capacitance is particularly important. Consider the pertinent equivalent circuit, Fig. 7-7, in which $C_2$ is connected between the grid and plate terminals. For the moment the effect of the other capacitors is ignored. The input impedance

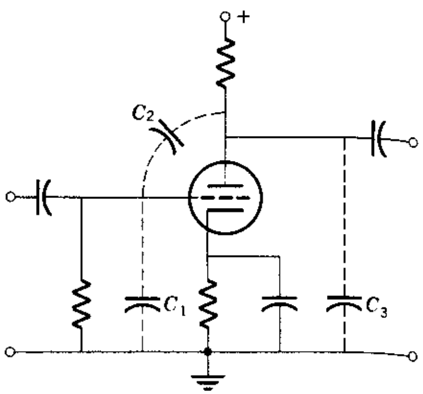


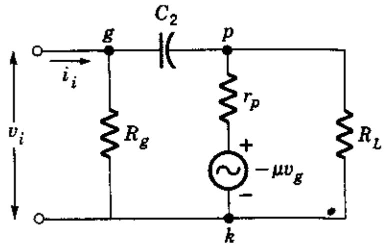

*FIGURE 7-6* *Stray capacitances in a triode amplifier.*

of the amplifier is calculated by assuming that the reactance of $C_2$ is the controlling factor,

$$Z_i = \frac{v_i}{i_i} \cong \frac{v_i}{(v_i + \mu v_i)/(1/j\omega C_2)} \qquad (7\text{-}7)$$

$$Z_i = \frac{1}{j\omega(1+\mu)C_2} \qquad (7\text{-}8)$$

*FIGURE 7-7*




This result indicates that the input impedance may be considered to be a capacitor $(1 + \mu)C_2$ connected from grid to ground. The increase in effective shunt capacitance caused by the amplification factor of the tube is called the *Miller effect* and is the dominating effect in determining the high-frequency response. The appro-

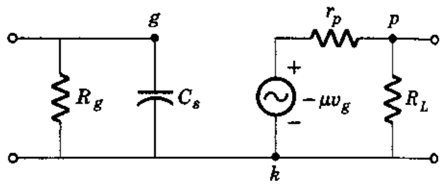



*FIGURE 7-8* *High-frequency equivalent circuit of triode amplifier includes shunt capacitance $C_s$.*

priate high-frequency equivalent circuit for the triode amplifier of Fig. 7-6 includes a shunt capacitance, as shown in Fig. 7-8. The magnitude of this capacitance,

$$C_s = C_1 + C_3 + (1 + \mu)C_2 \qquad (7\text{-}9)$$

includes the plate-cathode capacitance of the previous stage $C_3$ in the total shunt capacitance, as indicated by Eq. (7-9).

Shunt capacitance is less important in a transistor amplifier, Fig. 7-9, because of the small input impedance of the transistor com-

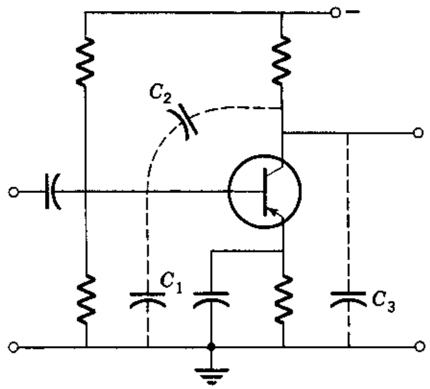



*FIGURE 7-9* *Stray capacitances in transistor amplifier.*

pared with the vacuum tube. Nevertheless, the collector-junction capacitance $C_2$ and the emitter-junction capacitance $C_1$ must be accounted for in assessing the high-frequency response. As in the case of the triode amplifier, capacitances $C_1$ and $C_3$ also include the effect of stray wiring capacitances.

The effect of the collector-junction capacitance is enhanced by the Miller effect of the transistor. The magnitude is found by a procedure identical to that used for the triode amplifier and results in a total shunt capacitance given by

$$C_s = C_1 + C_3 + (1 - h_{fe})C_2 \qquad (7\text{-}10)$$

Because $h_{fe}$ is inherently negative, the third term in Eq. (7-10) is most important. The corresponding high-frequency equivalent circuit of the transistor amplifier is illustrated in Fig. 7-10.

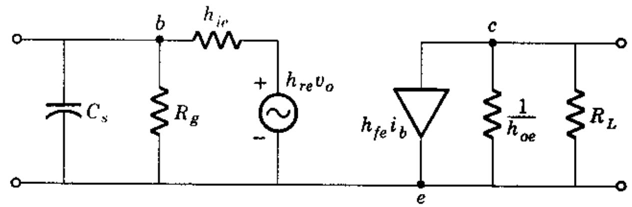



*FIGURE 7-10* *High-frequency equivalent circuit of a transistor amplifier includes shunt capacitance $C_s$.*

According to Figs. 7-8 and 7-10, the high-frequency gain of both triode and transistor amplifiers is accounted for by the input shunt capacitor $C_s$. This is evaluated using the simple circuit, Fig. 7-11, which also includes the effect of the source resistance $R_s$. The input resistance $R_i$ is essentially the grid resistance $R_g$ in the case of the vacuum tube but includes the total input impedance in the case of the transistor amplifier. The

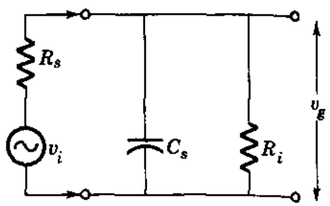



FIGURE 7-11

output voltage of either amplifier is found by analyzing the circuit of Fig. 7-11,

$$v_o = a'v_g = a'\frac{v_i}{R_s + Z}Z$$

where

$$\frac{1}{Z} = \frac{1}{R_i} + j\omega C_s$$

and $a'$ is the midband gain of the stage with no load. Substituting for $Z$ and simplifying,

$$v_o = \frac{a'v_i}{(1 + R_s/R_i) + j\omega R_s C_s} \tag{7-11}$$

The characteristic frequency of this circuit is defined as

$$2\pi f_0 = \omega_0 = \frac{1}{C_s}\left(\frac{1}{R_s} + \frac{1}{R_i}\right) \tag{7-12}$$

Introducing Eq. (7-12) into Eq. (7-11) and solving for the gain $v_o/v_i$ gives

$$a(f) = \frac{a'}{1 + R_s/R_i}\,\frac{1}{1 + jf/f_0} \tag{7-13}$$

The denominator of the first term in Eq. (7-13) accounts for the loading of the amplifier input upon the previous stage. As previously discussed, this effect is usually included in the determination of the true midband gain of the entire amplifier. Accordingly, the variation of the gain at high frequencies is conveniently written, after rationalization, as

$$a(f) = \frac{a}{\sqrt{1 + (f/f_0)^2}} \tag{7-14}$$

where $a$ is the true midband gain.

This result shows that the gain is reduced at high frequencies. Note also, Eq. (7-13), that phase shift is introduced between input and output signals, and this is equally significant in preserving the signal waveform. In the case of the vacuum-tube amplifier, the input resistance is essentially equal to the grid resistor $R_g$. Since $R_g \gg R_s$, the high-frequency performance is controlled by the output impedance of the preceding stage, according to Eq. (7-12). Conversely, $R_i < R_s$ in the case of the transistor amplifier, so the transistor input impedance is the dominating factor. As in the low-frequency case, the overall high-frequency response of the complete amplifier is poorer than that of any individual stage.

Actually, the high-frequency amplification of many transistors is limited by the transit time of carriers diffusing across the base region. This effect results in a high-frequency gain given by an expression identical to Eq. (7-14) except that $f_0$ is determined by physical constants of the transistor, such as the width of the base. This characteristic frequency is usually specified by the transistor manufacturer. In specially designed high-frequency transistors the *alpha falloff frequency* is high enough that the gain is limited by circuit parameters, as discussed above.

Using Eqs. (7-6) and (7-14), the *frequency response* of any voltage amplifier is similar to that illustrated in Fig. 7-12*a*. The *low-frequency cutoff* for each stage is determined from Eq. (7-4) while the *high-frequency cutoff* for each stage is found using Eq. (7-12). The *bandpass* of the complete amplifier is the frequency interval be-

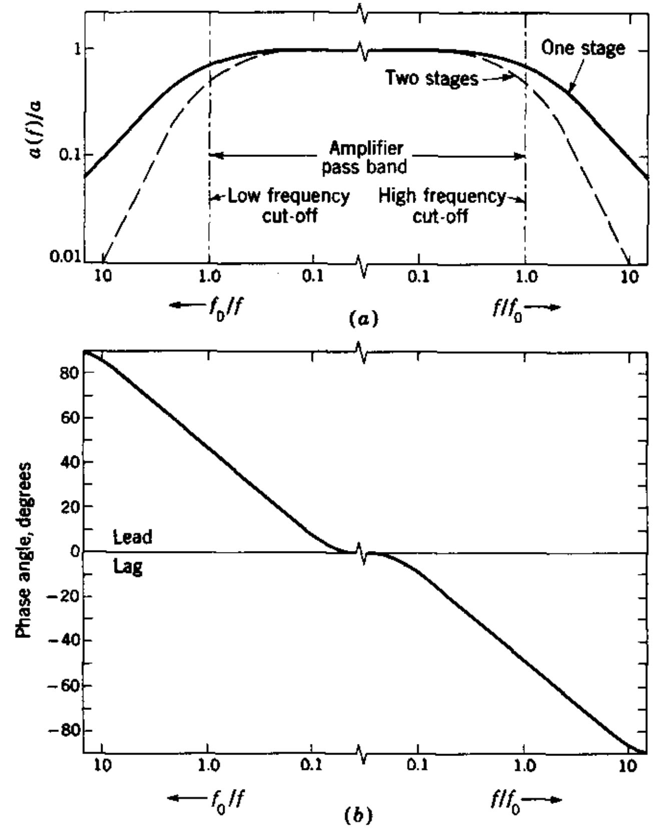



**FIGURE 7-12** *(a) Frequency-response characteristic of single amplifier stage. Characteristic of two-stage amplifier is shown dashed. (b) Phase-shift characteristic of single amplifier stage.*

tween the high- and low-frequency points where the gain falls to $1/\sqrt{2}$ of the midband gain. Since the power output is reduced to ½ of the midband value at these frequencies, they are generally referred to as the half-power points (see Chap. 2). It is conventional to employ logarithmic scales on both axes of bandpass characteristics such as Fig. 7-12*a* because the range of gains and frequencies is so great. The vertical scale is often put in terms of a unit called the *decibel,* abbreviated *db,* defined as

$$db = 20 \log \frac{a(f)}{a} \tag{7-15}$$

Correspondingly, the midband gain is often quoted in terms of decibels using the definition

$$db = 20 \log \frac{v_o}{v_i} \tag{7-16}$$

The advantage of this unit is that the total gain in db of several amplifier stages is simply the sum of the individual gains in terms of decibels. Note that, according to Eq. (7-15), the amplifier gain is down 3 db at the upper and lower half-power points.

The phase-shift characteristics of a single-stage amplifier are illustrated in Fig. 7-12*b*. The output signal leads the input at frequencies below the low-frequency cutoff and lags at frequencies above the high-frequency cutoff. The phase-shift characteristics of an entire amplifier are determined by adding the contributions from each stage.

Often an amplifier must have a wide bandpass in order to minimize waveform distortion. A number of minor circuit alterations have been developed to accomplish this end. For example, the cathode or emitter bypass capacitors may purposely be made small so that the capacitive reactance is appreciable except at frequencies near the high-frequency cutoff. This reduces the midband gain, according to Eq. (5-26), but increases the gain at high frequencies where the capacitive reactance becomes small. The net result is an extended high-frequency response, although at the expense of smaller overall amplification (see Exercise 7-3). If necessary, the loss in gain can be made up by adding another stage.

A second useful way of extending the high-frequency response is to include a small inductance as part of the plate load, Fig. 7-13. The load impedance increases at high frequencies and the gain of the amplifier is larger (see Exercise 7-4). This technique is called *peaking* since the resulting frequency-response characteristic tends to be peaked at the high-frequency end. The high-frequency response of this amplifier is further improved by the use of a pentode because of the much smaller grid-plate capacitance.

Improved low-frequency response can be achieved by adding a series resistor-capacitor combination $R_3$ and $C_3$ to the output

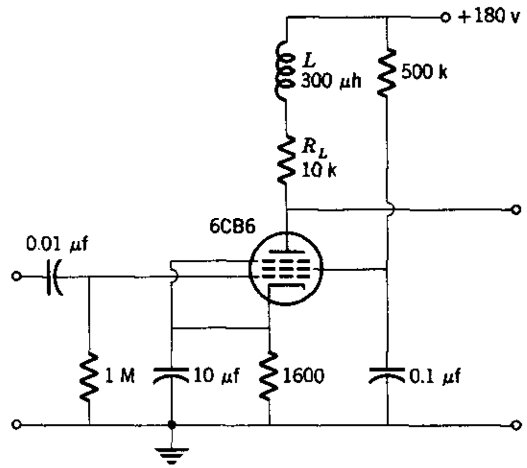



***FIGURE 7-13*** *Small inductance in load impedance improves high-frequency response because gain is increased by higher load impedance.*

circuit, Fig. 7-14. The gain at frequencies where the reactance of $C_3$ is small includes the effect of $R_3$ in the ac load resistance.

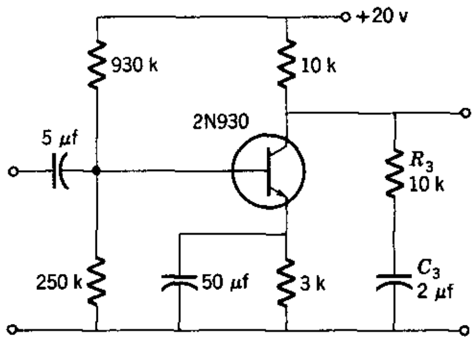



***FIGURE 7-14*** *Combination $R_3C_3$ improves low-frequency response of amplifier because reactance of $C_3$ becomes large.*

At low frequencies the gain increases as the reactance of the capacitor increases and removes $R_3$ as part of the output load. The result is an extended low-frequency response at the expense of midband gain (see Exercise 7-5).

## 7-4 Decoupling

When three or more stages of amplification are cascaded it is usually necessary to *decouple* the power supply of the input stage from the remainder of the amplifier. The reason for this is that the supply voltage changes with current because of the effective internal impedance of the power supply. Any small change in the power-supply voltage alters the bias on the first stage, and this change is amplified in the same fashion as an input signal. If the change in bias increases the current in the first stage, current in the second stage is reduced because of the 180° phase shift in the input amplifier. The current in the third stage is increased, however, because of the second 180° phase shift in the second stage. The change in the third stage is much larger than the original disturbance because of the gain of the amplifier. The additional load causes a decrease in the power-supply voltage. This, in turn, further alters the bias on the first stage and the process is cumulative. The changes continue until one tube or transistor is driven to cutoff or into saturation, which reduces the overall gain to zero. The power-supply voltage then returns to normal and the process repeats itself. The result of this *feedback* from output to input is that the amplifier rapidly oscillates from cutoff to saturation at a rate which is a function of the circuit components.

A low-pass *RC* filter inserted in the power-supply lead to the first stage, Fig. 7-15, circumvents this difficulty. The time con-

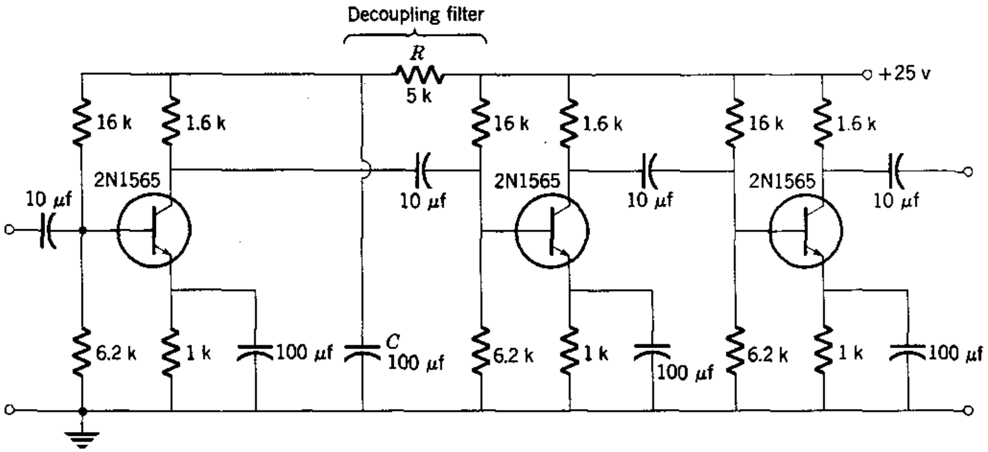



***FIGURE 7-15*** *Simple RC decoupling filter eliminates instability in multistage amplifier by reducing feedback effects resulting from common power supply.*

stant of this decoupling filter is selected so that power-supply variations are sufficiently attenuated and feedback is eliminated. Actually, the characteristic filter frequency is put below the low-frequency cutoff of the amplifier where the gain is insufficient to support feedback oscillations.

## 7-5 Transformer coupling

When transistor or vacuum-tube amplifiers deliver appreciable amounts of power, it is no longer feasible to use resistors in the collector or plate circuit. The $I^2R$ losses become significant at the high currents associated with large powers. Instead, a transformer couples the circuit to the load, Fig. 7-16. The dc collector current

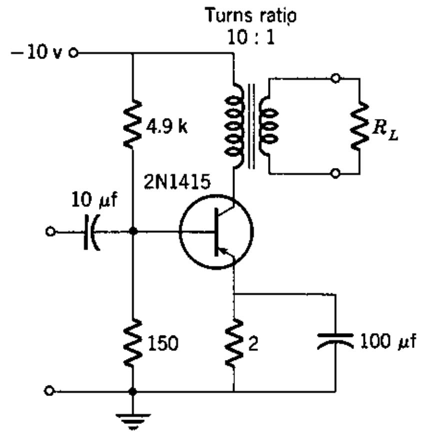



***FIGURE 7-16*** *Power amplifiers use transformer to couple transistor to load to reduce dc power lost in load resistance.*

in the winding resistance introduces only a small power loss, yet the reflected resistance of $R_L$ into the primary circuit provides the proper ac load impedance for the amplifier. Furthermore, the output impedance of the amplifier is matched to the load by the transformer and the actual load resistance can be any convenient value.

The dc load line is essentially vertical on the collector characteristic curves, Fig. 7-17, because of the small winding resistance of the transformer primary. The quiescent operating point is determined exactly as outlined in Chap. 6. The ac load line corresponding to the reflected load resistance $R_L'$ as seen from the primary side of the output transformer passes through the operating point. The slope of the ac load line is $-1/R_L'$, as shown in Fig. 7-17.

The operating point of a power amplifier is chosen to maximize the efficiency of the amplifier and to minimize the possibility of thermal runaway. Power dissipation in the transistor due to collector current is limited by the allowable temperature rise of the collector junction. If the maximum permissible temperature of the collector junction is $T_M$, the power dissipation must not exceed

$$P_M = KT_M \tag{7-17}$$

where $K$ is a constant involving the thermal conductance and other

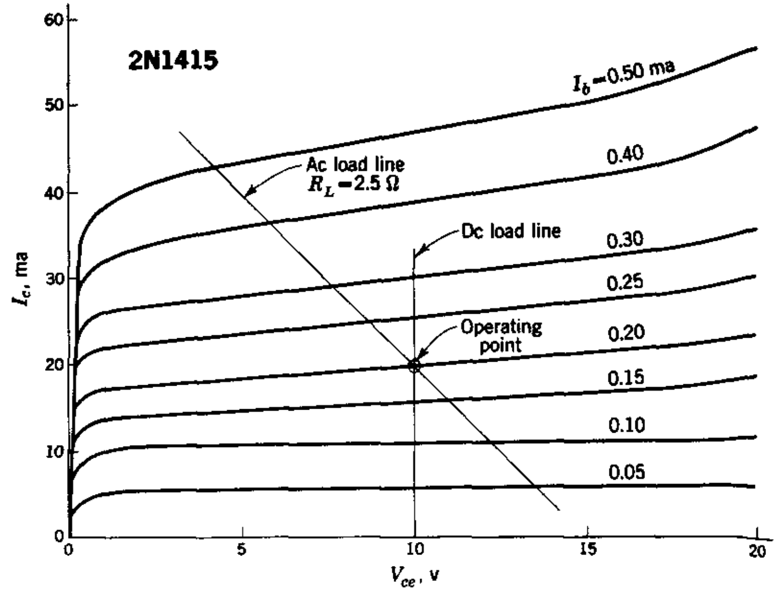



*FIGURE 7-17 Location of operating point and dc and ac load lines for power amplifier in Fig. 7-16.*

geometrical factors. Writing the power dissipation in the transistor as $I_cV_c$, the product must never exceed $KT_M$. The relation

$$I_cV_c = P_M \tag{7-18}$$

is a hyperbola on the collector characteristics, as indicated by the dashed line in Fig. 7-18. The permissible operating range of collector current and voltage is to the left of this *maximum-power hyperbola.*

Power transistors firmly mounted on a good heat conductor make $K$ in Eq. (7-17) larger. This moves the maximum-power hyperbola farther away from the origin and extends the permissible operating current and voltage range. In addition, cooling fins are often provided to maximize heat conduction away from the transistor.

The operating point is located so that the largest possible ac signals can be developed in order to maximize the power output without distortion. The maximum instantaneous collector potential is limited by reverse breakdown at the collector junction. Similarly, the maximum instantaneous transistor current corresponds to collector saturation, where the collector current no longer increases with emitter-junction current. The output waveform is badly distorted if either of these limits is exceeded because the peaks of the signal wave are clipped. Therefore, the optimum position for the operating point is in the center of the rectangle bounded by collector breakdown, collector saturation, zero collector current, and zero collector voltage, Fig. 7-18. Here, the collector current and voltage excursions on either side of the operating point are maximized without distortion.

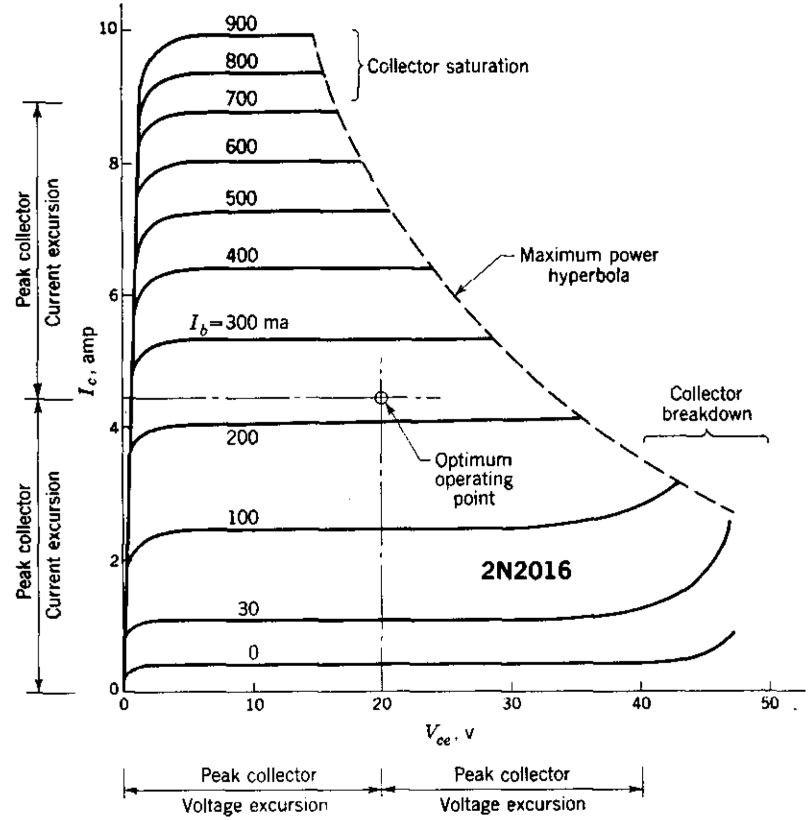



*FIGURE 7-18 Optimum location of operating point for class-A power amplifier is determined by collector saturation and collector breakdown.*

The efficiency of a power amplifier is equal to the ratio of the ac signal power to the dc or average power consumed by the amplifier. The average power is simply the product of the quiescent current times the quiescent collector voltage $I_cV_c$. If the operating point is located at the optimum position, the peak output signal current is equal to $I_c$ and the peak output signal voltage is equal to $V_c$. Consequently, the efficiency is

$$\eta = \frac{(I_c/\sqrt{2})(V_c/\sqrt{2})}{I_cV_c} = \frac{1}{2} \tag{7-19}$$

Thus, the maximum efficiency of a class-A power amplifier is 50 percent. Practical transistor amplifiers approach this ideal quite closely, even though for minimum distortion the signal excursions must be somewhat smaller than the ideal case considered above. Efficiencies of the order of 48 percent are achieved in practice.

The plate characteristics of vacuum tubes are not nearly so ideal as are transistor collector characteristics. Curvature in the characteristics is considerably greater. Consequently, peak signal voltages and currents are smaller and the efficiency is correspondingly less. Pentodes are much more satisfactory than triodes in this respect, but the efficiency of practical circuits rarely exceeds 30 percent.

An equivalent-circuit representation of power amplifiers is not feasible because of the large signal voltage excursions. Consequently, all analyses are carried out graphically. It is most useful to determine the dynamic transfer characteristic of the amplifier, which is a plot of the collector output current as a function of base input current, Fig. 7-19. The transfer characteristic is determined

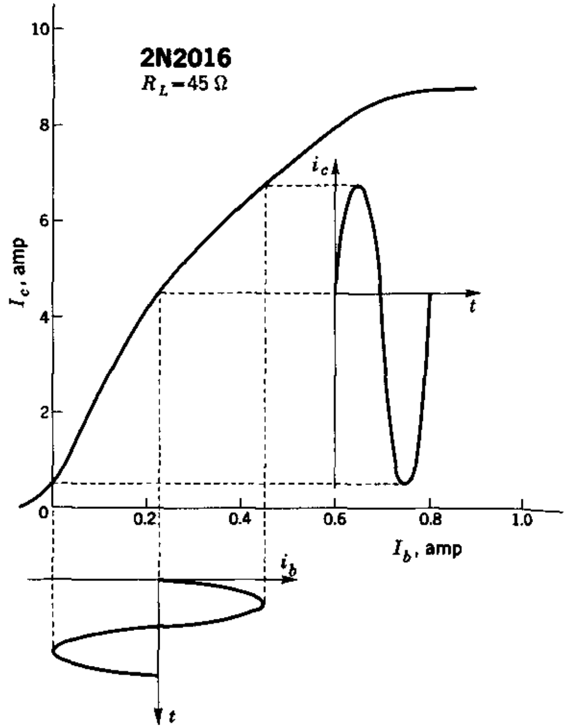


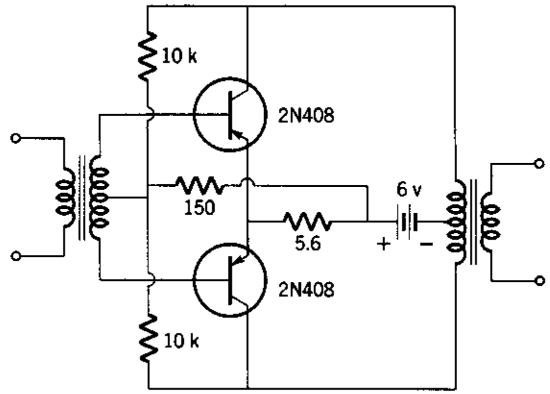

***FIGURE 7-19*** *Dynamic transfer characteristic is used to determine amplified waveform of 2N2016 power amplifier. Note distortion in output current caused by nonlinear transfer characteristic.*

from intersections of the collector characteristic curves, Fig. 7-18, with the dynamic load line. For minimum distortion the transfer characteristic should be a straight line since any curvature introduces irregularities into the output waveform.

## 7-6 Push-pull amplifier

The two-transistor *push-pull* amplifier, Fig. 7-20, has increased power output, efficiency, and less distortion than a single-transistor circuit. The center-tapped input transformer drives each of the transistors with signals 180° out of phase, which accounts for the name of the circuit. The amplified collector currents combine in the center-tapped output transformer to produce a load current

***FIGURE 7-20*** *Push-pull power amplifier.*




waveform that is a replica of the input signal. The input transformer also matches the driver stage to the input impedance of the amplifier.

Increased efficiency results when the push-pull amplifier is biased for class-B operation. Because each transistor is biased near cutoff, the quiescent current is very small and the signal voltage and current excursions can be equal to the maximum permissible collector voltage and current, Fig. 7-21. Each tube delivers one-half of a sine-wave signal to the output transformer and the output waveform is preserved even though signal currents in each transistor represent only one-half of the input signal. This action has already been noted in the complementary-symmetry amplifier discussed in Chap. 6. In push-pull operation the peak output voltage can equal the maximum collector potential, which is the same as the dc collector supply voltage (see Fig. 7-21). Correspondingly, the peak signal current is equal to the maximum collector current. The average power of the stage is equal to the power of a half-sine wave, since only one tube conducts at a time. Therefore, the efficiency of a class-B push-pull amplifier is

$$\eta = \frac{P_o}{P_{dc}} = \frac{(V_c/\sqrt{2})\,(I_c\sqrt{2})}{(2/\pi)V_cI_c} = \frac{\pi}{4} \tag{7-20}$$

According to Eq. (7-20), the maximum efficiency is 70 percent, a considerable improvement over the single-transistor class-A amplifier.

The required power output $P_o$ of any push-pull amplifier is specified by the particular application. The peak output-signal voltage is limited by the collector reverse breakdown potential $V_c$, however, which means that the collector-to-collector load resistance must be

$$R_L = \frac{V_c^2}{2P_o} \tag{7-21}$$

It usually turns out that the value of $R_L$ determined by Eq. (7-21) is smaller than the output impedance of the transistors, and maxi-

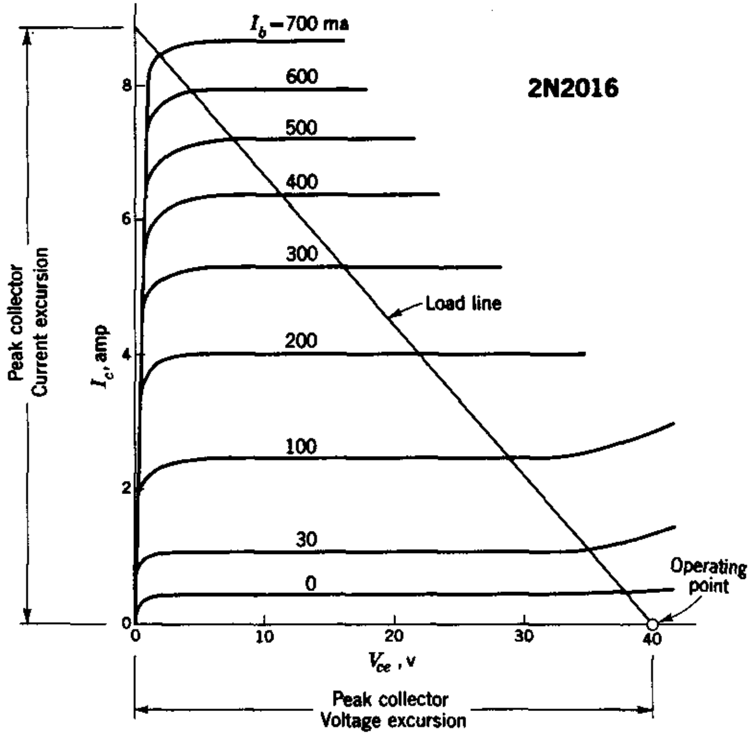



*FIGURE 7-21* *Operating point for class-B push-pull operation permits voltage to swing over maximum range of collector characteristics.*

mum power transfer conditions are not possible. Nevertheless, the turns ratio of the output transformer is selected to reflect the proper value of $R_L$ corresponding to the actual load resistance.

A small quiescent base bias current minimizes *crossover* distortion resulting from nonlinearity in the transfer characteristic of each transistor at small currents. This is illustrated in Fig. 7-22, where the transfer characteristics of the two transistors are plotted in opposite quadrants corresponding to their reversed signal polarities. The composite transfer characteristic of the entire amplifier is the average of the individual curves and is much more linear than either one. In particular, the nonlinearities cancel each other near the origin where both transistors are active. The cancellation effect means that the push-pull circuit has much less distortion than a single-ended stage. For this reason class-A push-pull amplifiers are often used, even though the power efficiency is no greater than that of the single-tube circuit.

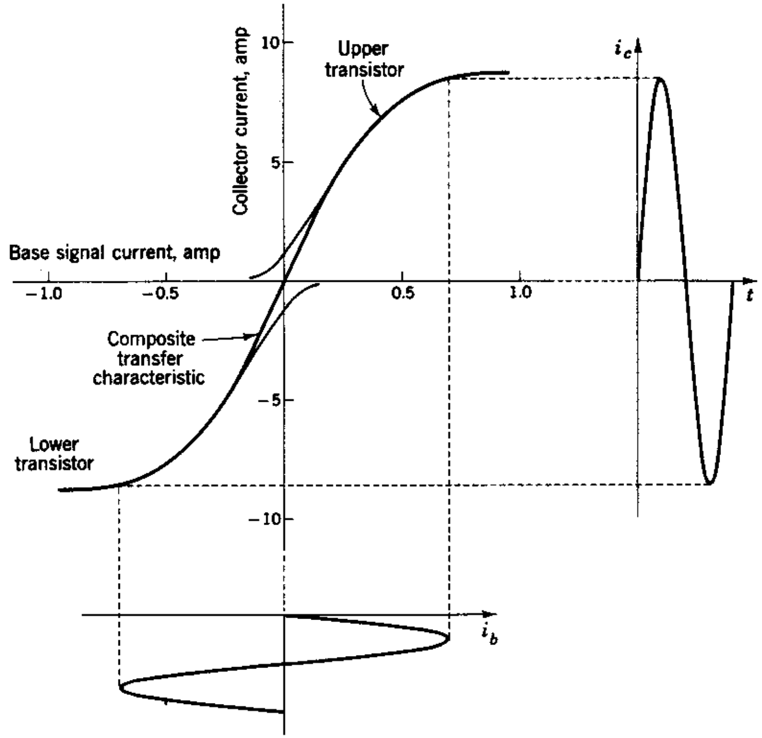



*FIGURE 7-22* *Composite transfer characteristic of push-pull power amplifier is more linear than that associated with each transistor. Compare output-signal current amplitude and waveform with single transistor case, Fig. 7-19.*

Transistors are so inherently temperature-sensitive that it is necessary to compensate for temperature changes which tend to alter the bias current. This situation is compounded in class-B power amplifiers where a rather critical value of bias is necessary to maintain class-B conditions. According to Eq. (6-25), the emitter resistance of a transistor decreases significantly with increasing temperature, which means that for a constant base bias the transistor may be near class-C operation at low temperatures and class-A operation at high temperatures. A temperature-sensitive resistor can be included in the bias circuit to counteract this undesirable variation. In the practical circuit of Fig. 7-23, a silicon diode $D_1$ performs this function. The changes in $D_1$ with temperature correspond exactly to those of the emitter junction of the transistor. The value of resistor $R_1$ in the bias network is chosen to keep $D_1$ biased in the forward direction under all conditions.

Alternatively, temperature compensation can be achieved by making resistor $R_1$ temperature-sensitive. In this case $R_1$ has a positive temperature coefficient and base bias is reduced at elevated temperatures.

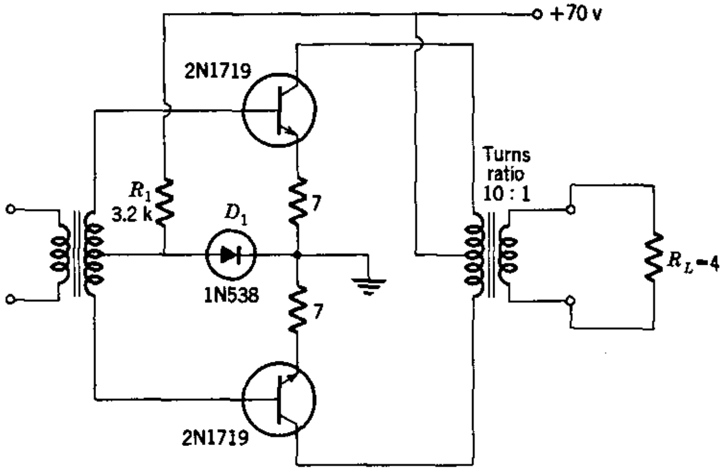



*FIGURE 7-23 A 5-watt power amplifier using diode in bias circuit to compensate for temperature changes.*

Note that the emitter resistors are not bypassed in the circuit of Fig. 7-23. The reason for this is the clamping action at the rectifying emitter junction, which is inherent in class-B operation. The clamp would charge a bypass capacitor to the peak value of the input signal and thereby maintain the transistor at cutoff at all times. The loss in gain resulting from the unbypassed emitter resistors is usually not serious, in view of the considerable improvement in bias stability achieved. Emitter bypass capacitors are permissible in class-A operation since both transistors conduct continuously.

Vacuum-tube circuits equivalent to the transistor versions are equally useful. In particular, consider the push-pull pentode amplifier illustrated in Fig. 7-24. Here the out-of-phase grid signals are provided by a single-tube *phase inverter*, which performs the same function as the difference amplifier discussed in Chap. 5. It is usually not necessary to use an input transformer in vacuum-tube circuits because of the high input impedance. Although pentodes are commonly used as power amplifiers, triodes are also satisfactory because distortions tend to cancel in the push-pull circuit. Nevertheless, the power efficiency of vacuum-tube power amplifiers is not as high as is the case for transistors.

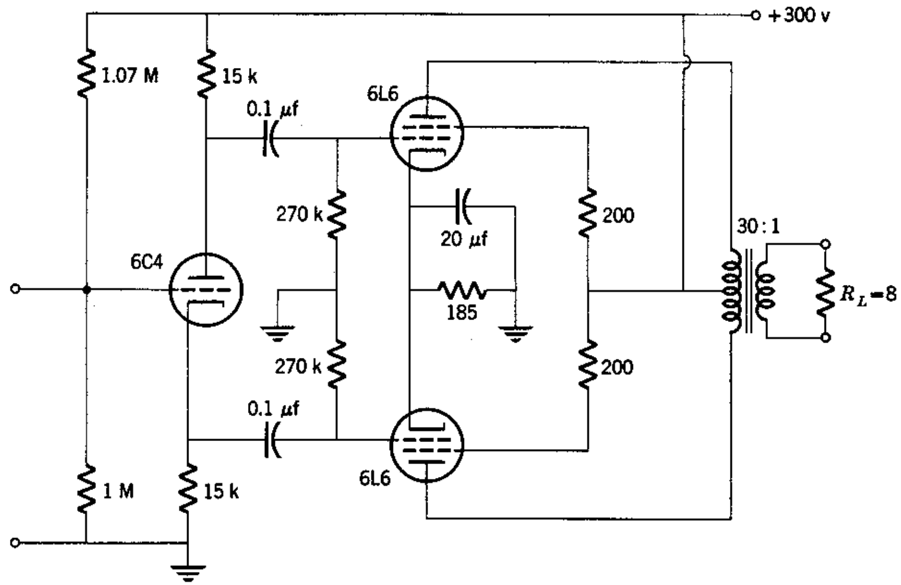



*FIGURE 7-24 Push-pull pentode power amplifier using phase-inverter driving stage.*

## TUNED AMPLIFIERS

### 7-7 Tuned coupling

Resonant circuits couple the output of one stage to the input of the next when it is only necessary to amplify signals of a single frequency or of a narrow band of frequencies. The impedance of parallel resonant circuits is very great at resonance, as discussed in Chap. 3. Therefore, appreciable gain is achieved at the resonant frequency when a *tuned circuit* is the load impedance of a vacuum-tube or transistor amplifier. A tuned amplifier also rejects signals far from the resonant frequency, which is often a considerable advantage. In addition, stray circuit capacitances are incorporated into the resonant circuit and do not shunt the signal at high frequencies.

The elementary two-stage transistor tuned amplifier, Fig. 7-25, uses parallel resonant circuits for the input circuit and output load of each transistor. The coupling capacitor $C_c$ carries the signal from one stage to the next. The operating point for each transistor is determined in the standard fashion. It is common practice to make the tuning capacitors $C_1$, $C_2$, $C_3$, and $C_4$ adjustable so that each circuit can be brought to the same resonant frequency including the effect of all stray capacitances in each stage. Tuned circuits are resistive at resonance, which means that the amplifier can be analyzed by the methods previously developed. Stray capacities can be neglected, however, and values of the $h$ parameters appropriate at the frequency of interest must be used.

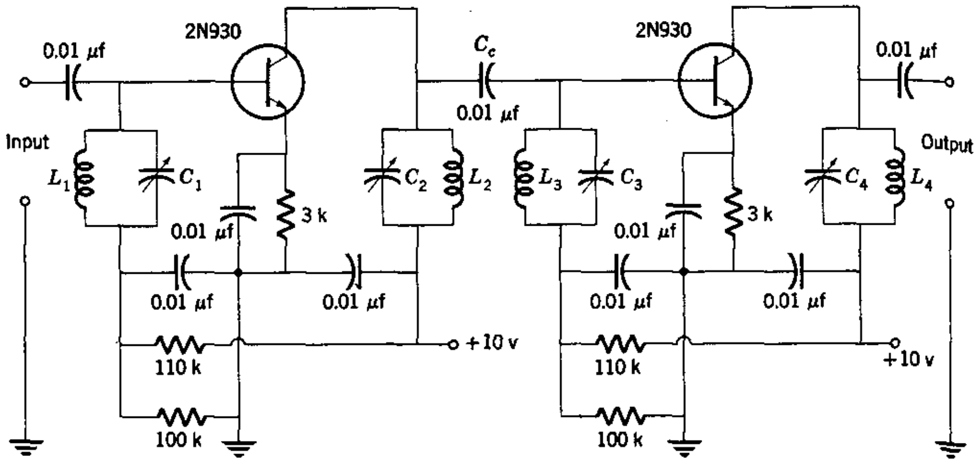
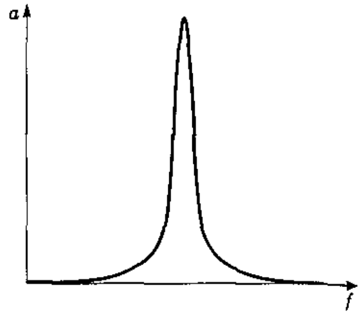



***FIGURE 7-25*** *Two-stage tuned amplifier.*

Circuit analysis at other than the resonant frequency is rather complicated because of reactance effects, but can be treated straightforwardly using the ac equivalent circuit.

If all four resonant circuits are tuned to the same frequency, the response characteristic is sharply peaked at the resonant frequency, Fig. 7-26 (also see Exercise 7-13). Such a characteristic is useful

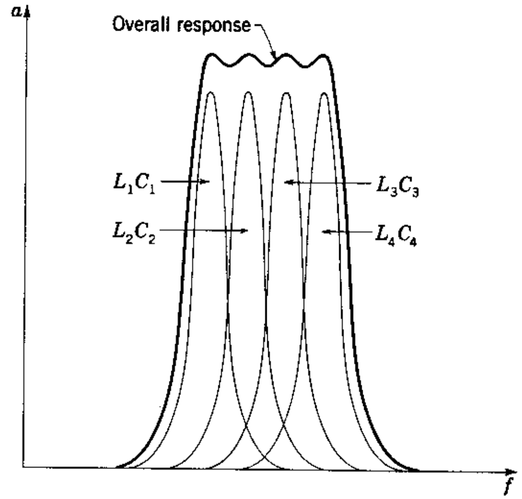
***FIGURE 7-26*** *Response characteristic of sharply tuned amplifier.*




when signals having one specific frequency are amplified. Alternatively, each circuit can be tuned to a slightly different frequency, Fig. 7-27, in which case the response characteristic becomes flat-topped. This permits amplification over a band of frequencies such as for the modulated sine wave discussed in Chap. 4. The midband gain of such a *stagger-tuned* amplifier is less than that of the single-frequency circuit since the maximum amplification of each stage occurs at different frequencies.

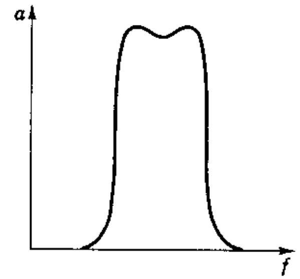
If $L_2$ and $L_3$ are wound on the same core the response characteristic may be double-peaked, Fig. 7-28, even though both primary and secondary windings are tuned to the center frequency. This is caused by mutual inductance between the windings, which are said to be *overcoupled.* The mutual inductance is altered by chang-

***FIGURE 7-27*** *Stagger-tuned amplifier has response characteristic with relatively flat top.*




ing the distance between the two coils. In this way the frequency-response curve is either sharply peaked, as in Fig. 7-26, or relatively flat-topped, as in Fig. 7-28. The overcoupled case is

***FIGURE 7-28*** *Response characteristic of overcoupled amplifier.*




particularly useful because the midband gain is greater than in the stagger-tuned amplifier and the sides of the response curve rise much more steeply. The circuit thus rejects signals at frequencies immediately outside of the passband. Note that the coupling capacitor is no longer necessary since the signal is coupled from one stage to the next by the mutual inductance. In fact, the combination $L_2C_2$ and $L_3C_3$ is looked upon as a tuned transformer.

## 7-8 Neutralization

The collector-junction capacitance in many transistors is large enough to cause an undesirable feedback effect between the collector and base. At high frequencies the capacitive reactance becomes small and the amplifier oscillates because the amplified collector signal is returned to the input where it is reamplified, etc. The circuit is useless as an amplifier when this occurs. The effect of the collector capacitance can be *neutralized* by feeding the base a signal of the same amplitude as that produced by the feedback capacitance but 180° out of phase so that the two feedback signals cancel each other. One technique for accomplishing this is illustrated in Fig. 7-29. Here, a portion of the output signal is fed back

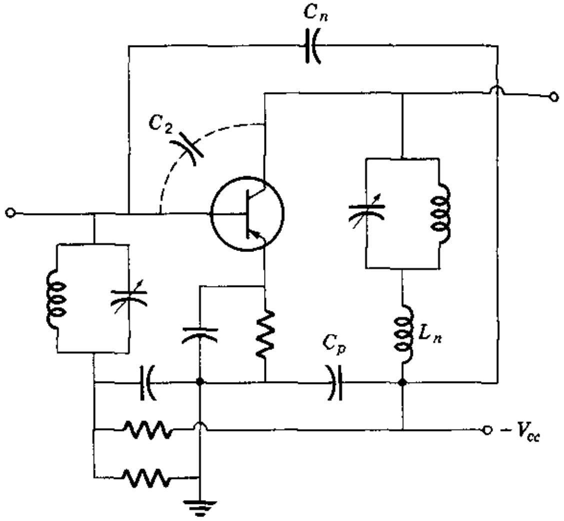



***FIGURE 7-29*** *Effect of collector-base capacitance is neutralized by signal returned from output circuit through $C_n$.*

to the input by the neutralizing capacitor $C_n$. The magnitude of the return signal is determined by the relative values of the small series inductance $L_n$ and the capacitor $C_p$, as well as the value of $C_n$. The return signal is 180° out of phase with that due to the collector capacitance because of the inductive phase shift introduced by the series inductance. Other circuit configurations are also used. In transistor design, every effort is made to minimize collector capacitance so that neutralization is unnecessary.

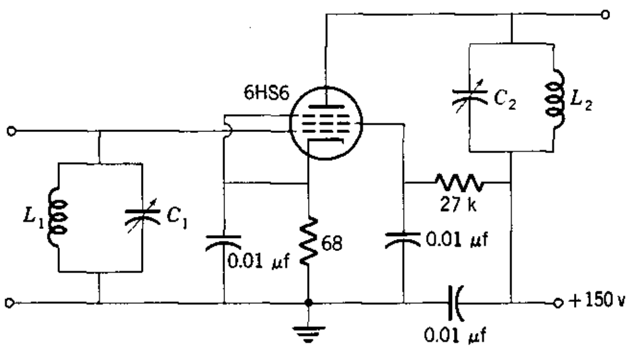



***FIGURE 7-30*** *Pentode tuned amplifier.*

A similar effect occurs in a triode vacuum tube because of the grid-plate capacitance. For this reason pentodes are almost universally used in preference to triodes as high-frequency amplifiers except for certain very-high-power class-C stages. A typical tuned amplifier stage employing a pentode is illustrated in Fig. 7-30. Analysis of the circuit using the equivalent circuit is quite direct (see Exercise 7-14).

The feedback effect in transistor amplifiers can be circumvented by employing the grounded-base configuration, Fig. 7-31. In this

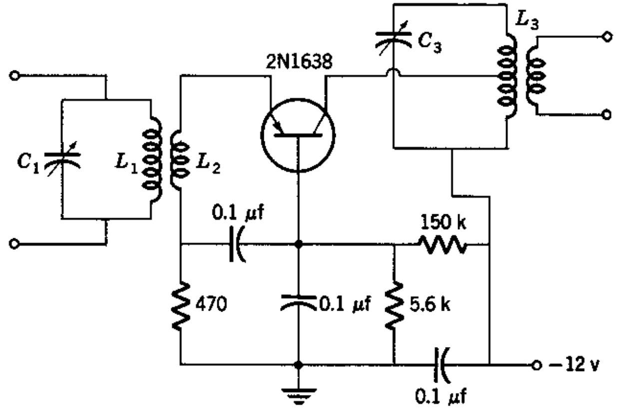



***FIGURE 7-31*** *Grounded-base tuned amplifier.*

circuit the collector-emitter capacitance is so small that neutralization is unnecessary. Furthermore, the grounded-base $h$ parameters are relatively independent of frequency so that the transistor is a useful amplifier at frequencies very close to the $\alpha$-cutoff frequency. The low input impedance of the grounded-base stage can be matched using a tuned transformer. The winding $L_2$ has only a few turns, and it is usually not advantageous to resonate the secondary of the transformer.

In the circuit of Fig. 7-31, the collector is tapped down on the inductance of the resonant circuit $L_3C_3$. This reduces the loading of the transistor output impedance upon the resonant circuit, thereby increasing the $Q$ and making the resonance curve sharper. Furthermore, the collector capacitance is much less significant in determining the resonant frequency. This is important because of the change in collector capacitance with temperature. If the collector is connected across the entire resonant circuit, the collector capacitance is effectively in parallel with the tuning capacitor $C_3$ and the resonance frequency varies with temperature.

Any spurious currents or voltages extraneous to the signal of interest are termed *noise* since they interfere with the signal. Noise voltages arise in the basic operation of electronic devices or are the result of improper circuit design and use. It is important to minimize noise effects in order to characterize the signals with the greatest possible precision and to permit the weakest signals to be amplified. A convenient measure of the influence of noise on any signal is the *signal-to-noise* ratio, the ratio of the signal power to the noise power at any point in a circuit.

## 7-9 Nyquist noise

When a number of amplifier stages are cascaded a random noise voltage appears at the output terminals, even in the absence of an input signal. This output voltage is caused by a random voltage generated in the input resistor. The noise voltages that appear across the terminals of any resistor are attributed to the random motion of the free electrons in the material of the resistance. Electrons in a conductor are free to roam about by virtue of their thermal energy and at any given instant more electrons may be directed toward one terminal of the resistor than toward the other. The result is a small potential difference between the terminals. The magnitude of the potential fluctuates rapidly as the number of electrons moving in a given direction changes from instant to instant.

Since the noise voltage across a resistor fluctuates randomly, it has Fourier components covering a wide range of frequencies. It is convenient, therefore, to specify the noise voltage in terms of the mean square noise voltage per unit cycle of bandwidth. For a resistor $R$ this quantity is

$$\langle \Delta v^2 \rangle = 4kTR \tag{7-22}$$

where $k$ is Boltzmann's constant, and $T$ is the absolute temperature. The noise voltage given by Eq. (7-22) is variously called *Nyquist noise*, after the physicist who derived this equation, or *thermal noise*, since its origin is a result of the thermal agitation of free electrons.

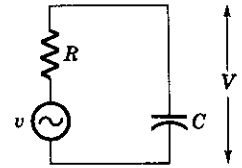
The meaning of Eq. (7-22) is as follows. A noise voltage appears between the terminals of any resistance. The magnitude of the noise voltage actually measured with any instrument depends upon the frequency response of the instrument. For example, the rms noise voltage of a 1000-Ω resistor at room temperature as measured by a voltmeter with a bandwidth of 10,000 cps is, using Eq. (7-22),

$$v = (1.65 \times 10^{-20} \times 10^3 \times 10^4)^{1/2} = 4.1 \times 10^{-7} = 0.41\ \mu\text{v} \tag{7-23}$$

This rather small voltage is not inconsequential. For example, the output voltage of an amplifier with a 10-kc bandpass and a gain of $10^6$ is nearly ½ volt if the input resistor is 1000 Ω. This output voltage is present even when no input signal is applied.

The Nyquist expression for the noise voltage of resistances, Eq. (7-22), may be understood in the following way. Replace the actual resistor by an equivalent circuit, Fig. 7-32, containing a noise voltage generator in series with a noiseless resistor and in parallel with a capacitor representing the inherent stray capacitance of the actual resistor. The square of the voltage across the capacitor is simply

*FIGURE 7-32*




$$V^2 = \frac{v^2}{1 + (\omega RC)^2} \tag{7-24}$$

The condition of the circuit is completely determined if the voltage across the capacitor is known; in thermodynamic terms the circuit is a system with 1 degree of freedom. According to the equipartition theorem in thermodynamics, the total energy of the capacitor, therefore, must equal ½$kT$. Thus, using Eq. (7-24),

$$\tfrac{1}{2}kT = \tfrac{1}{2}CV^2 = \tfrac{1}{2}C \int_0^\infty \frac{v^2 df}{1 + (\omega RC)^2} \tag{7-25}$$

where the result of Exercise 2-16 for the energy of a charged capacitor has been employed. The integration extends over all frequencies because of the random nature of the noise voltage.

Equation (7-25) determines the magnitude of the noise voltage $v^2$. We proceed by assuming that the noise voltage is independent of frequency, so $v^2$ may be brought out from under the integral sign. Therefore, Eq. (7-25) becomes

$$kT = Cv^2 \int_0^\infty \frac{df}{1 + (\omega RC)^2} = \frac{v^2}{4R} \tag{7-26}$$

Solving Eq. (7-26) for $v^2$ yields the Nyquist expression.

This development indicates that the noise voltage of resistances is independent of frequency. Accordingly, Nyquist noise is called "white" noise by analogy with the uniform spectral distribution of white light energy. As indicated by Fig. 7-32, the presence of Nyquist noise in any circuit is accounted for by including a noise generator given by Eq. (7-22) in series with a noiseless resistor. In practice it is usually necessary to consider the Nyquist noise of only those resistors in the input circuit of an amplifier. The gain of the first stage makes the amplified noise of the input resistor larger than the noise of resistors in succeeding stages.

Nyquist noise is a fundamental and unavoidable property of any resistance. An amplifier should have a bandwidth only as wide as is necessary to adequately amplify all signal components in order to minimize the ever-present Nyquist noise voltages. If it is desired to amplify a single-frequency signal, for example, the frequency response of the amplifier should be sharply peaked at that frequency, as in Fig. 7-26. The total noise voltage at the output is therefore reduced since only noise components having frequencies in the amplifier passband are amplified. The signal-to-noise ratio is enhanced and weak signals can be amplified usefully.

## 7-10 $1/f$ noise

Noise voltages in excess of Nyquist noise are observed experimentally in certain resistances when a direct current is present. Although the physical origins of this additional noise are not clear, many experiments have shown that the noise is largest at low frequencies and that it increases with the square of the current. An empirical expression for this effect is

$$\langle \Delta v^2 \rangle = K \frac{I^2}{f} \tag{7-27}$$

where $K$ is an empirical constant involving the geometry of the resistor, the type of resistance material, and other factors; $I$ is the dc current; and $f$ is the frequency. According to Eq. (7-27) the mean square noise voltage per unit bandwidth decreases inversely with frequency and the phenomenon is therefore called *1/f noise*. Since the noise also depends on $I$, it is sometimes referred to as *current noise*.

The magnitude of $1/f$ noise varies markedly with the material of the conductor and its physical form. It is absent entirely in metals, so that only Nyquist noise is observed in wire-wound resistors. Composition resistors, on the other hand, generate a large $1/f$ noise level, Fig. 7-33, which is associated with the intergranular contacts in such resistors. Although such contacts are known to be important, $1/f$ noise is also observed in single-crystal semiconductors where contact effects are negligible.

To minimize the low-frequency noise level, resistor types in which the $1/f$ noise is small, such as wire-wound units, are selected. Fortunately, at high frequencies where wire-wound resistors are unsuitable because of their inductance, the $1/f$ noise of composition resistors is usually negligible compared with Nyquist noise. In other situations the direct current in noisy components is minimized to reduce the current noise generated. The total current noise voltage in any given circuit is found by integrating Eq. (7-27) over the frequency response characteristic of the amplifier.

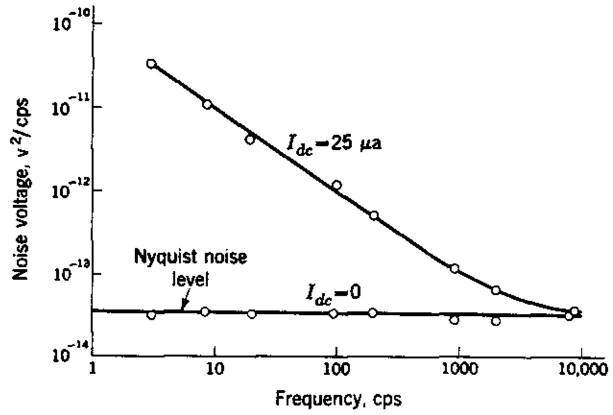



*FIGURE 7-33 Experimental noise voltage of 2.2-MΩ composition resistor. Note spectrum is 1/f noise when dc current is present and white Nyquist noise in absence of current.*

Low-frequency $1/f$ noise is also present in vacuum tubes and transistors. In the former it is often known as *flicker noise* and originates in the semiconducting cathode material, particularly at the emitting surface. The noise in transistors results from semiconductor properties, in which surface conditions are very important. In general, $1/f$ noise is more prevalent in germanium devices than in silicon units.

## 7-11 Noise in tubes and transistors

Other noise effects are also present in tubes and transistors. *Shot noise* in vacuum tubes is a result of the random emission of electrons from the cathode. Since each electron represents an increment of current, the plate current fluctuates slightly about the dc value. This effect is analogous to the noise of raindrops on a tin roof. That is, the basic reason for shot noise is that the electron is a discrete unit of electrical charge.

An expression for the magnitude of shot noise can be developed as follows. Suppose that $n$ is the average number of electrons emitted from the cathode in a time interval $t$. The direct current is then, from Eq. (1-10),

$$I = \frac{en}{t} \tag{7-28}$$

According to a general principle of statistical phenomena the variance in $n$ is equal to its average value, so that

$$\langle \Delta n^2 \rangle = n = \frac{It}{e} \tag{7-29}$$

Therefore, the current fluctuations in the time interval $t$ are

$$\langle \Delta I^2 \rangle = \left(\frac{e}{t}\right)^2 \langle \Delta n^2 \rangle = \frac{e}{t} I \tag{7-30}$$

Finally, it can be shown that the relation between the total fluctuations in a given time interval and the mean square fluctuation per unit bandwidth is given by

$$\langle \Delta i^2 \rangle = 2t \langle \Delta I^2 \rangle \tag{7-31}$$

Introducing Eq. (7-30), the current fluctuations are

$$\langle \Delta i^2 \rangle = 2eI \tag{7-32}$$

The quantity $\langle \Delta i^2 \rangle$ is analogous to $\langle \Delta v^2 \rangle$ in the Nyquist expression, Eq. (7-22), except that the noise is expressed here in terms of current fluctuations. The mean square noise voltage output of a tube is simply Eq. (7-32) multiplied by the square of the load resistance. Note that shot noise is a white noise since the right side of Eq. (7-32) is independent of frequency.

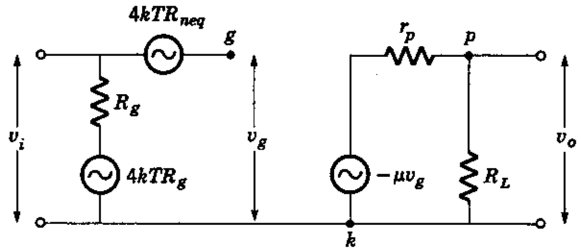
The basic expression for shot noise, Eq. (7-32), applies to the situation in which each electron emitted from the cathode proceeds to the anode independently of all other electrons. As discussed in Chap. 4 in connection with Child's law, this is not the case in practical vacuum tubes. Each electron is influenced by the presence of all the others. The result of this interaction is to make the electron current more uniform and thereby reduce the magnitude of shot noise. It turns out that the actual value is a function of the operating point. A full analysis of this effect is complicated because it depends upon subtle details of the emission current and electric field.

It is convenient to express the effective shot noise in terms of the Nyquist noise of an equivalent resistor in the grid circuit. The magnitude of the equivalent noise resistor in the case of a triode is given by

$$R_{neq} = \frac{2.5}{g_m} \tag{7-33}$$

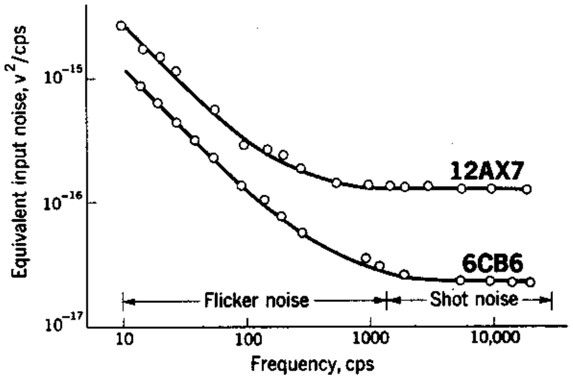
The meaning of $R_{neq}$ is simply that the effective shot noise of the triode can be represented by a noise voltage generator in the grid circuit given by the Nyquist expression, Eq. (7-22), using the resistance value determined from Eq. (7-33). Therefore, the appropriate equivalent circuit, Fig. 7-34, includes two noise generators, one associated with Nyquist noise of the grid resistor and the other associated with the effective shot noise of the tube. According to Eq. (7-33), the effective shot noise is smallest for tubes having a large mutual transconductance and is of the order of 250 Ω for the lowest-noise triodes listed in Table 5-1. The internal noise

*FIGURE 7-34 Equivalent circuit of triode including noise generators.*




as a function of frequency for typical triodes is illustrated in Fig. 7-35. In these curves the output noise level is referred to a noise voltage generator in the grid circuit. Any contribution due to the

*FIGURE 7-35 Experimental internal noise levels of 12AX7 triode and triode-connected (screen tied to plate) 6CB6. Note flicker noise at low frequencies and white shot noise at high frequencies.*




grid resistor has been eliminated simply by shorting the grid to ground. The flicker-noise and shot-noise regions can be easily discerned.

Noise voltages are treated by including noise voltage generators in the equivalent circuits. The effect of the amplifier passband, which may be determined by stray shunt capacitance or by tuned circuits, is included in assessing the magnitude of the noise voltages. The noise resulting from each generator may be treated separately since random noise voltages are independent. The total output noise voltage is therefore simply the sum of all noise effects. As mentioned previously, it is not necessary to consider the noise of the resistors $r_p$ and $R_L$ in Fig. 7-34 because the amplified noise of the input circuit is much larger than the Nyquist noise of these resistors.

An expression analogous to Eq. (7-32) applies to each electrode in pentode and other multigrid tubes where the appropriate value of current to each electrode is used. The total noise is the sum of the individual electrode noises, suitably modified by the internal conditions. The sum is known as *partition noise* since the total tube current is divided among several electrodes. Pentodes are therefore noisier than triodes and the latter are universally used when it is necessary to obtain the largest possible signal-to-noise ratio. Note that this applies principally to the input stage of an amplifier. The gain of the first stage increases the signal level sufficiently so that noise effects in succeeding stages are negligible. The signal-to-noise ratio of any circuit is determined primarily by conditions in the first amplifier stage.

The input signal to any circuit has associated with it a given signal-to-noise ratio since Nyquist noise corresponding to the source resistance is present, at least. An ideal amplifier amplifies the incoming signal and the incoming noise equally and introduces no additional noise. Therefore, the original signal-to-noise ratio is preserved at the output. Practical amplifiers are not ideal because of Nyquist noise of the input circuit and shot noise of the first stage. A useful figure of merit for any circuit is the *noise figure, NF,* which is defined as the input signal-to-noise ratio divided by the output signal-to-noise ratio. An ideal amplifier has a noise figure of unity and many practical circuits approach this value fairly closely.

The noise of transistors is a result of Nyquist noise of the semiconductor resistance, $1/f$ noise caused by current in the semiconductor crystal, and shot noise of carriers crossing the junctions. In addition, still another noise phenomenon has been observed in semiconductors. Electrons are promoted randomly from the valence band to the conduction band and also return randomly to the valence band, keeping the proper average number of carriers in each band. Random generation and recombination of carriers is caused by thermal energies and produces a conductivity fluctuation of the semiconductor. This generates a noise voltage when a direct current is present; this second type of current noise in semiconductors is termed *generation-recombination,* or *g-r noise.*

Analysis of noise phenomena in transistors is complicated by these many factors and by the inherent input-output coupling. It turns out that at intermediate frequencies the noise figure of a transistor can be expressed as

$$\mathrm{NF} = 1 + \frac{r_b}{R_s} + \frac{r_e}{2R_s} + \frac{(R_s + r_b + r_e)^2}{2h_{fe}r_eR_s} \tag{7-34}$$

where $R_s$ is the source resistance and the other symbols have their usual meaning. According to Eq. (7-34) the noise figure approaches unity if the emitter and base resistances are small with respect to the source resistance and if the forward current gain $h_{fe}$ is large. Note that the internal noise of a transistor depends upon the operating point, much as is the case for a vacuum tube.

Equation (7-34) ignores $1/f$ noise, which increases the noise level at low frequencies, Fig. 7-36. Additionally, an increase in noise at frequencies of the order of the $\alpha$-cutoff frequency is observed. The latter effect, also apparent in Fig. 7-36, can be quite well explained on the basis of the influence of current gain and the $\alpha$-cutoff frequency upon *g-r* noise.

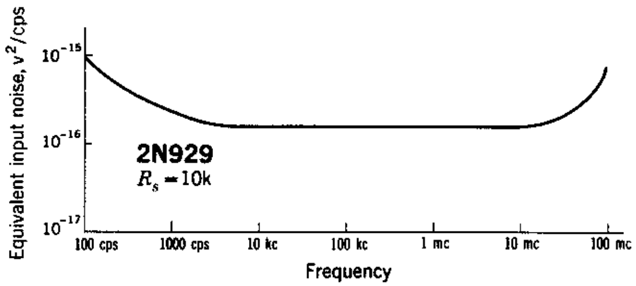



*FIGURE 7-36 Experimental internal noise level of a type 2N929 transistor with a source resistance of 10,000 Ω.*

## 7-12 Stray pickup

Noise is often introduced in practical circuits by extraneous voltage signals coupled into the circuit from the surroundings. The most common source of this noise comes from the 60-cps electric and magnetic fields produced by power mains. The 60-cps signal induced by these fields is called *hum* because it is audible as a low-frequency tone in amplifiers connected to a loudspeaker. Other *stray pickup* may result from electric fields generated by nearby electronic equipment, electric motors, lightning discharges, etc.

It is useful to *shield* those portions of a circuit where the signal level is small and, consequently, where noise voltages are most troublesome. Electric fields induce noise voltages capacitively, so it is only necessary to surround the circuit with a grounded conducting shield in order to reduce stray pickup. This is illustrated schematically in Fig. 7-37, where the capacitive coupling to external sources in Fig. 7-37*a* is interrupted by interposing a grounded conductor, Fig. 7-37*b*. Such shielding is also effective in reducing so-called *crosstalk* between different stages of the same circuit as, for example, between the input stage and the power output stage of a complete amplifier.

Additionally, it is useful to shield a circuit to minimize induced currents resulting from stray magnetic fields. This is accomplished with high-permeability ferromagnetic enclosures which reduce the intensity of the magnetic field inside. Such shielding is never complete because of the properties of ferromagnetic materials, and it is always advantageous to minimize the area of the circuit by using the shortest possible signal leads. According to

Eq. (2-66) the induced voltage in any circuit resulting from changing magnetic fields decreases if the enclosed area of the circuit is reduced. Transformers are particularly troublesome with respect

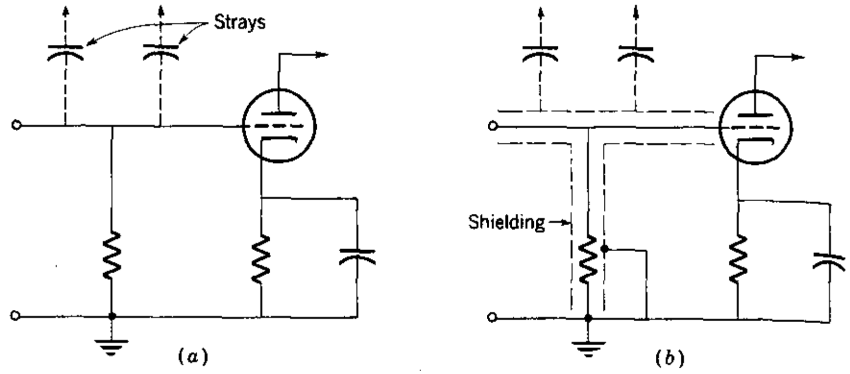



*FIGURE 7-37* *(a) Stray capacitive coupling introduces noise pickup signals into sensitive circuits. (b) Grounded conductor shields the circuit from surroundings.*

to inductive pickup because of their many turns of wire. They are kept well removed from all power transformers because of the strong magnetic fields generated by such units.

In general it is good practice to keep all circuits physically small in order to minimize stray pickup, crosstalk, and stray capacitance. All grounded components, such as bypass capacitors, pertaining to a given stage are returned to a single point. This reduces so-called *ground loops,* which are current paths through the metal chassis on which electronic circuits are often mounted. If all components of one stage are not grounded at the same point, the currents may cause undesirable signal coupling between stages.

Vacuum tubes are also *microphonic* in that noise voltages are generated by movement of the grid wires caused by mechanical vibration or shock. Transistors are much superior in this respect because of their simpler mechanical construction. On the other hand, the lower impedance level in transistor circuits makes them more susceptible to induced magnetic pickup because the induced currents are larger. Hum may also result from an inadequate power-supply filter or the ac heater-current wires. It is common practice to twist heater-current wires tightly together to reduce the net magnetic field from the current. In very sensitive circuits hum pickup from this source is reduced by heating the cathodes of the input stages with direct current.

When a number of individual electronic units are interconnected, shielded cable is used for all signal leads between units. The shield is used as the ground lead, as illustrated in Fig. 7-38. It is important that the entire system be grounded at only one point, usually the input terminal. If each unit is grounded separately, as shown by dashed lines in Fig. 7-38, large 60-cps currents can be induced in the circuit because of the large area encompassed. The current induced in loop $A$ by stray 60-cps magnetic fields introduces a large stray pickup signal into the amplifier.

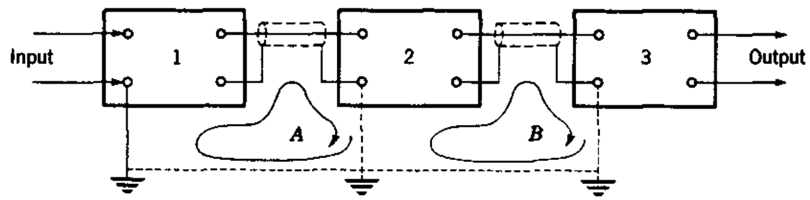



*FIGURE 7-38* *When several electronic devices are connected, system must be grounded only at one point, preferably at input. Multiple grounds can lead to large ground-loop currents.*

Note that the capacitance between the central wire and its shield tends to shunt the signal at high frequencies. For this reason such cables are kept as short as possible. The circuit output impedance is also made small since this reduces the effect of shunt capacitance. Therefore, the output stage of many electronic circuits is a cathode or emitter follower.

## SPECIAL CIRCUITS

### 7-13 Cascode amplifier

A unique circuit employing two identical triodes, shown in Fig. 7-39, combines the low-noise features of a triode with the superior high-frequency performance of a pentode. This *cascode amplifier*

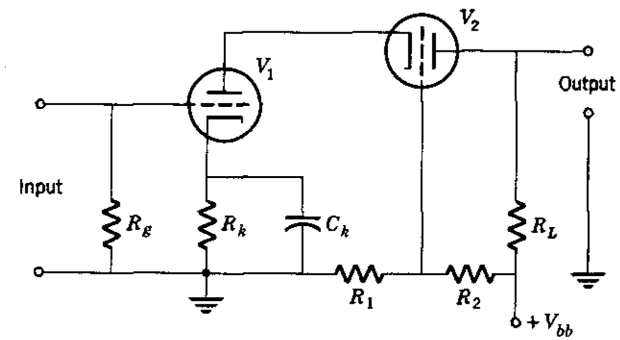



*FIGURE 7-39* *Simple cascode amplifier.*

may be thought of as a grounded-cathode stage followed by a grounded-grid amplifier. The low input impedance of the second stage effectively eliminates feedback caused by the grid-plate capacitance of the first tube while retaining the low-noise feature of a triode. The grid-plate capacitance of $V_2$ is unimportant because the grid is effectively grounded for ac signals. The voltage divider $R_1R_2$ sets the grid voltage of $V_2$ while cathode bias is employed for $V_1$. The operating point is most easily found by a series of successive approximations starting with an assumed plate current and calculating the resulting $V_1$ grid potential. If this value does not correspond to the assumed current as given by the plate characteristics, a new current value is selected and the process repeated until a satisfactory match is obtained.

The pentode-like action of the cascode amplifier is illustrated by analyzing the ac equivalent circuit (Exercise 7-19). The gain of the circuit is given by

$$a = -\frac{\mu(\mu + 1)R_L}{R_L + (\mu + 2)r_p} \tag{7-35}$$

If $(\mu + 2)r_p \gg R_L$ and $\mu \gg 1$, Eq. (7-35) reduces to

$$a = -g_m R_L \tag{7-36}$$

which is identical to the gain of a pentode amplifier, Eq. (5-21).

The cascode amplifier is most often used at high frequencies where triode grid-plate capacitance is troublesome. Also, $R_g$ and $R_L$ can be replaced with parallel resonant circuits. Improved performance is obtained by connecting a variable inductance from grid to plate of $V_1$ and adjusting it for resonance with the grid-plate capacitance at the signal frequency. The high impedance of the resonant circuit further reduces the effect of the feedback capacitance.

## 7-14 Dc amplifiers

All the amplifier circuits discussed to this point have zero gain for dc signals because of the infinite reactance of the interstage coupling circuits at zero frequency. Amplification of dc or very slowly varying signals is achieved by eliminating the interstage coupling networks entirely. In addition to the dc response of such a *direct-coupled* amplifier, the high-frequency performance is also enhanced. Stray capacitances are reduced since fewer components are associated with the signal leads.

One of the problems of a direct-coupled amplifier is bringing the signal at the output of the first stage to the proper dc level for the next amplifier stage. This can be done, for example, with a Zener diode, as in the elementary dc amplifier of Fig. 7-40. The breakdown potential of the diode is selected so that the grid potential of the second stage is at the proper value. Since the voltage drop across the Zener diode is constant, variations in the plate voltage of $V_1$ resulting from a signal applied to the grid are transmitted to the grid of $V_2$. It is also possible to replace the Zener diode with a battery, but this introduces undesirable bulk.

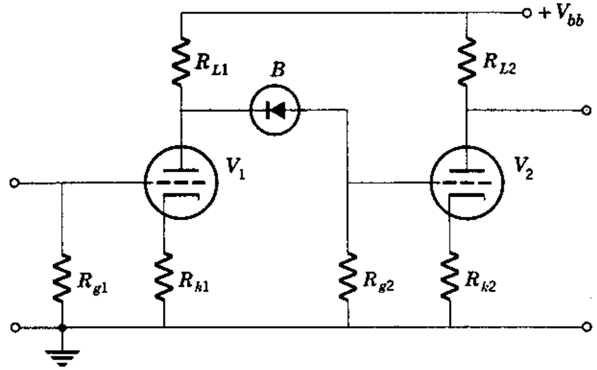



*FIGURE 7-40 Direct-coupled amplifier using Zener diode to reduce dc voltage between stages.*

This simple dc amplifier has a number of shortcomings. First, the output signal contains a quiescent dc voltage corresponding to the dc plate voltage of $V_2$. Second, any small change in the characteristics of $V_1$ is amplified by the second stage and is indistinguishable from a signal. The plate-supply voltage in all dc amplifiers must be stabilized to reduce *drift* in the circuit caused by slow changes of the electrode voltages. Drifts caused by variations in tube characteristics and supply voltages are minimized by using a balanced differential amplifier, Fig. 7-41. Changes in one side of

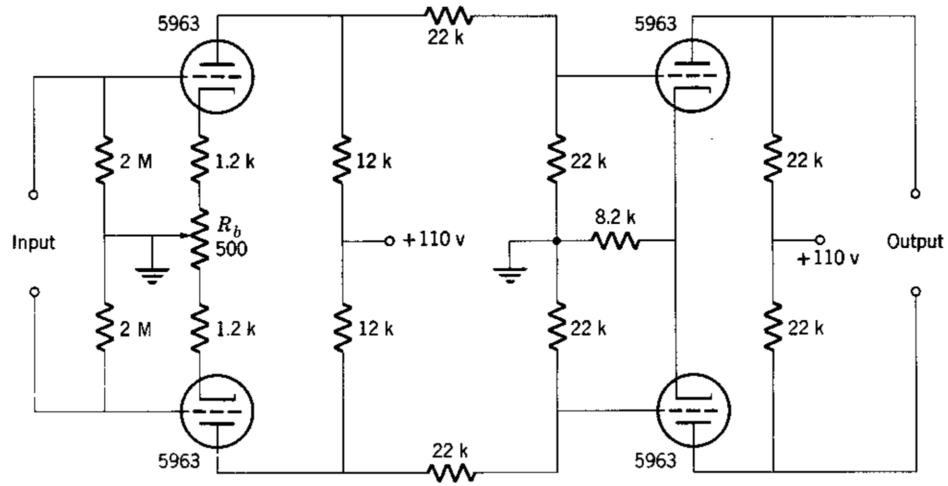



*FIGURE 7-41 Balanced dc amplifier.*

the circuit tend to be compensated by similar changes on the other side. Furthermore, the output terminals are at the same potential when the input signal is zero. The balance control $R_b$ is included to adjust for any asymmetry in the circuit.

Interstage coupling in this amplifier uses a voltage divider to reduce the plate potential to a suitable value. This results in a signal loss, however, since the voltage divider acts on the signal voltage as well. Such amplifier circuits are commonly used in oscilloscopes where their dc response and good high-frequency performance are advantageous. It proves quite complicated to determine the quiescent conditions of such circuits because of the large number of interrelated voltages and currents involved. The method of successive approximations is most effective. The ac performance of the amplifier is analyzed by the equivalent-circuit technique in the usual fashion.

Direct-coupled transistor amplifiers are very susceptible to drift because of the temperature sensitivity of transistors. Differential circuits are most often used and careful temperature compensation is employed. Such circuits become fairly elaborate, but, nevertheless, quite successful designs are possible. One compensating feature of transistors is their ability to operate satisfactorily at collector potentials from a few tenths to many tens of volts; this permits a certain amount of flexibility in circuit design.

An elementary direct-coupled transistor amplifier employing *npn* and *pnp* transistors is shown in Fig. 7-42. The reversed-bias

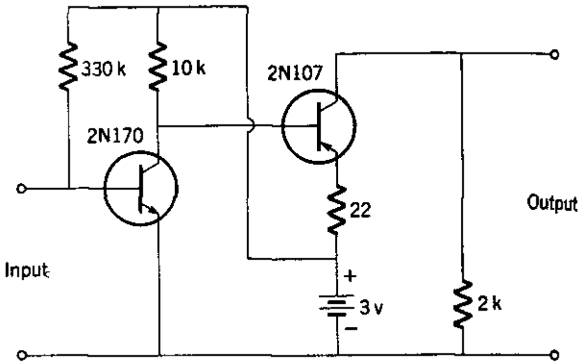



***FIGURE 7-42*** *Simple direct-coupled transistor amplifier.*

potentials inherent in the two transistor types is an example of the flexibility inherent in transistor circuits. This simple amplifier is useful to illustrate the principles of direct-coupled circuits but is not satisfactory for critical applications since no provision for compensation of drift caused by temperature changes, etc., is included. A more elaborate, practical version is considered in the next chapter.

## 7-15 Chopper amplifiers

To circumvent the drift and instabilities inherent in direct-coupled amplifiers, it is useful to convert the dc input signal to an ac voltage which can be amplified by a standard ac-coupled circuit. Subsequently, the amplifier output is rectified to recover the amplically used. Nonmechanical choppers, such as the balanced modulator, are useful because the chopping frequency can be much higher.

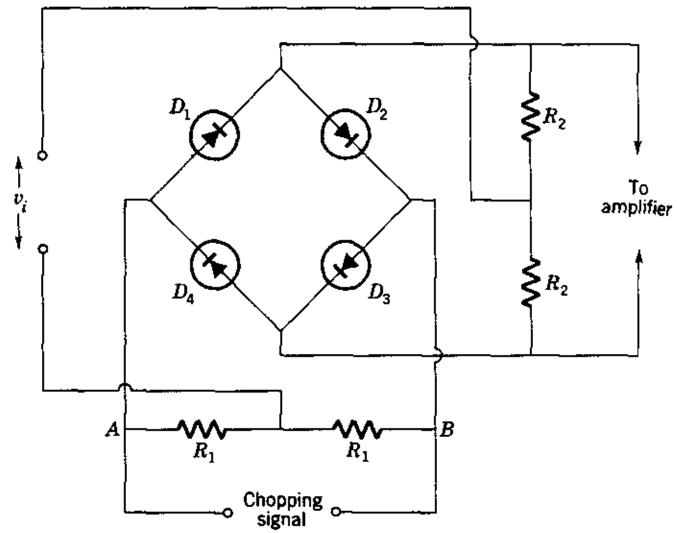



***FIGURE 7-44*** *Balanced modulator electronic chopper for use at high chopping frequencies.*

Mechanical choppers are most satisfactory for very-high-gain dc amplification because they introduce minimum noise into the circuit. Suppose, for example, that one of the four diodes of the balanced modulator is slightly different from the other three. The bridge is therefore slightly unbalanced and a portion of the chopping signal appears at the amplifier terminals even when the dc input signal is zero. This signal is amplified by the circuit and appears as a noise voltage at the output. Similarly, other semiconductor and tube devices introduce switching noises to a greater or lesser extent. The mechanical chopper has very small resistance when the contacts are closed and very high resistance when the contacts are open. It is nearly ideal in this respect. Nevertheless, chopping noise limits the amplifier sensitivity at the very highest gain applications. One source of noise is stray coupling between the driving solenoid and the signal circuits.

A major advantage of chopper amplifiers is their very low effective internal random-noise level. The reason for this is the rectifier-filter combination in the output circuit. The output voltage signal may be made as noise-free as desired by increasing the time constant of the filter. In effect, the overall bandwidth of the system is equal to the frequency interval from dc (zero cps) to $\pi\tau_0/2$, where $\tau_0$ is the filter time constant. If, for example, the time constant is 10 sec, the effective amplifier bandwidth is 0.016 cps, a very small value indeed. Since the total noise voltage increases with bandwidth [compare Eqs. (7-22) and (7-27)], the total noise is very low. Of course, when $\tau_0$ is 10 sec, a time interval of approximately 30 sec is required for the output voltage to reach its final value. Thus the amplifier responds very slowly to changes in the amplitude of the input signal. This reciprocity between bandwidth and response time is a general property of all electronic systems.

A major difficulty with the simple chopper amplifier is that the output voltage is independent of the polarity of the input signal. That is, the ac-coupled amplifier yields an ac output signal for either polarity of input voltage. This situation is corrected in chopper amplifiers which employ a second set of contacts to rectify the amplifier output, Fig. 7-45. The contacts are arranged to close

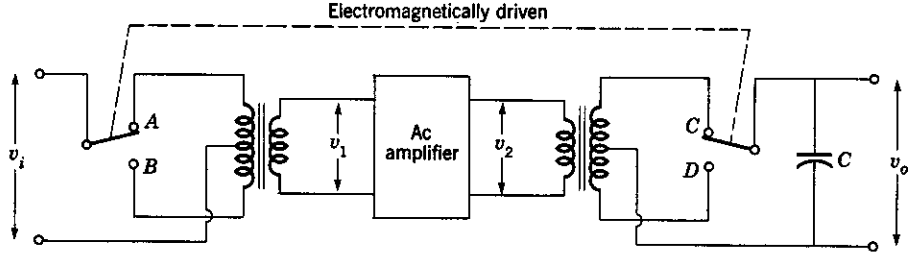



*FIGURE 7-45 Synchronous chopper amplifier preserves polarity of input voltage.*

synchronously so that as the input chopper converts the dc signal to a double-ended square wave, Fig. 7-46*a* and *b*, the output

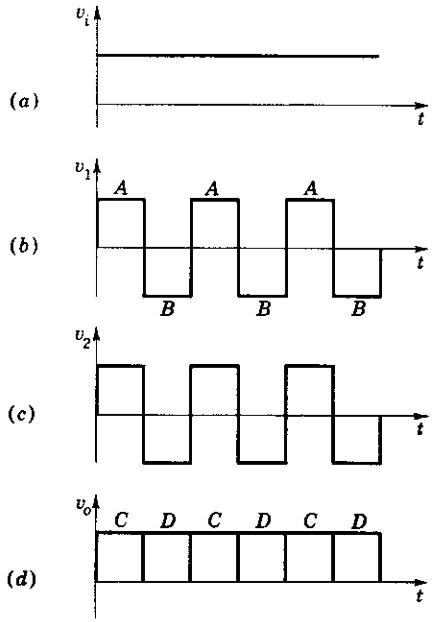



*FIGURE 7-46 Waveforms in synchronous chopper amplifier: (a) dc input signal, (b) square-wave input to amplifier, (c) amplified square-wave output, and (d) rectified square wave produced by second chopper.*

chopper reconverts the amplified square wave back to a dc signal, Fig. 7-46*c* and *d*. Comparing the waveforms in Fig. 7-46 shows that if $v_i$ is positive, $v_o$ is also positive. Similarly, if the input signal is negative, the output signal is also negative. Both sets of contacts are put on the same vibrating arm in practical synchronous choppers to assure that they open and close simultaneously. An electronic version of this *synchronous rectifier* is analyzed in the next section.

## 7-16 Lock-in amplifier

The principles of the synchronous chopper amplifier are employed in an electronic circuit which has found wide use in instrumentation systems. The version illustrated in Fig. 7-47 uses a

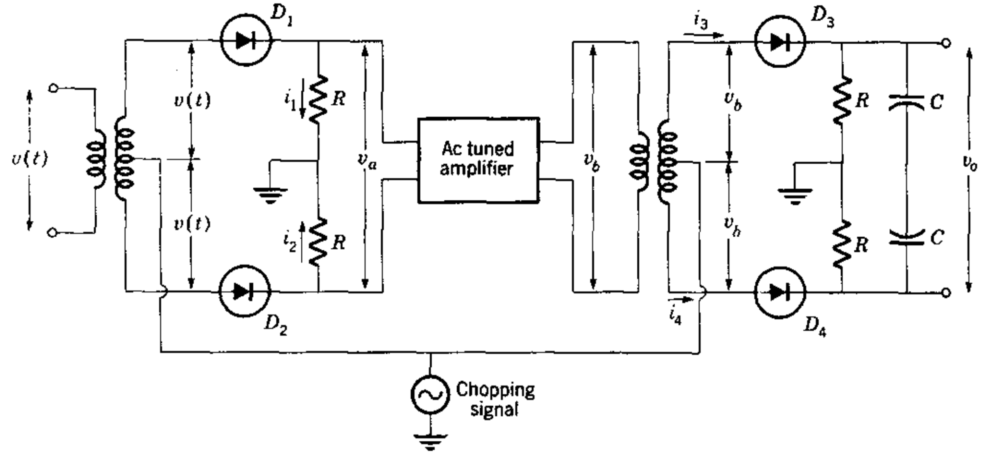



*FIGURE 7-47 Electronic synchronous chopper amplifier.*

diode modulator to convert the input signal to a form suitable for the amplifier and a diode demodulator to recover the amplified version of the input signal. For generality, we consider the input signal to be a slowly varying signal; if dc signals are important the input transformer can be replaced by a center-tapped resistor.

The input circuit is analyzed using techniques discussed in Chap. 4. It is assumed that the diode characteristic can be represented by a quadratic expression. Then the current in the diode $D_1$ is, from Eq. (4-37),

$$i_1 = a_1 v(t) + a_1 V_2 \sin \omega_2 t + a_2 v^2(t) + a_2 V_2^2 \sin^2 \omega_2 t + 2a_2 v(t) V_2 \sin \omega_2 t \qquad (7\text{-}37)$$

where $a_1$ and $a_2$ are constants related to the diode characteristic, $v_2 = V_2 \sin \omega_2 t$ corresponds to the chopping signal discussed earlier, and $v(t)$ is the input signal. An identical expression applies to the current in $D_2$, except that the polarity of $v(t)$ is reversed with respect to $v_2$. The voltage signal applied to the amplifier is

$$v_a = R(i_1 - i_2) \tag{7-38}$$

Substituting for $i_1$ and $i_2$, many terms cancel because of the reversed polarity of $v(t)$ in $D_2$. The result is

$$v_a = 2Ra_1v(t) + 4Ra_2v(t)V_2 \sin \omega_2 t \tag{7-39}$$

The first term in Eq. (7-39) is not transmitted by the amplifier since it is assumed that $\omega_2$ is much larger than any of the frequency components associated with $v(t)$. Note that the second term is simply a modulated sine wave of frequency $\omega_2$. The amplitude variations correspond to the input signal.

If the gain of the amplifier is $k$ the signal applied to the demodulator circuit is

$$v_b = 4kRa_2v(t)V_2 \sin \omega_2 t \tag{7-40}$$

Ignoring the capacitors $C$ for a moment, the current in $D_3$ is, again using Eq. (4-37),

$$i_3 = a_1v_b + a_1V_2 \sin \omega_2 t + a_2v_b^2 + a_2V_2^2 \sin^2 \omega_2 t + 2a_2v_bV_2 \sin \omega_2 t \tag{7-41}$$

The current in $D_4$ is similar except that the polarity of $v_b$ with respect to $v_2$ is reversed. Consequently, the output voltage

$$v_o = R(i_3 - i_4) \tag{7-42}$$

reduces to

$$\begin{aligned} v_o &= 2a_1Rv_b + 4a_2Rv_bV_2 \sin \omega_2 t \\ &= 2a_1Rv_b + 16a_2^2kR^2v(t)V_2^2 \sin^2 \omega_2 t \end{aligned} \tag{7-43}$$

Inserting the standard trigonometric identity $2 \sin^2 \omega t = 1 - \cos 2\omega t$,

$$v_o = 2a_1Rv_b + 8a_2^2kR^2V_2^2v(t)(1 - \cos 2\omega_2 t) \tag{7-44}$$

The filter capacitors eliminate the high-frequency terms in $\omega_2$ and $2\omega_2$. Therefore, the filtered output signal is simply

$$v_o = (8a_2^2kR^2V_2^2)v(t) \tag{7-45}$$

According to Eq. (7-45) the output voltage is an amplified replica of the input signal.

The amplifier passband is conveniently peaked at $\omega_2$ to minimize noise effects. Actually, however, the output filter circuit determines the effective bandwidth in the same way as for the chopper amplifier, and very-low-noise performance is possible. The minimum usable bandwidth depends upon the frequency components present in the signal $v(t)$.

In instrumentation applications it is often possible to produce the initial modulation in some way associated with the physical quantity being measured. For example, the infrared beam of an infrared spectrometer is chopped by means of a rotating shutter before it strikes the infrared detector. The amplified signal from the detector is demodulated by a circuit similar to Fig. 7-47 using a voltage $v_2$ derived from the shaft of the rotating shutter. The low-noise performance of the circuit permits extremely weak infrared signals to be detected. In this form the circuit is usually called a *lock-in amplifier*, since the detector is locked in step with the input signal. The system is also called a *phase-sensitive detector* because the circuit can recognize the phase of the input signal.