BASIC AMPLIFIER PRINCIPLES

3-1. Classification of Tubes and Amplifiers. The classification of an amplifier is usually somewhat involved, owing to the fact that a complete classification must include information about the tubes that are used, the conditions of the bias, the character of the circuit elements connected to the tubes, the function of the circuit, and the range of operation. Certain of these factors will be discussed here, but many will be deferred for later discussion.

Apart from the wide variety of vacuum tubes of the diode, triode, tetrode, pentode, beam, hexode, heptode, and multiunit types and the varied power capacities of each type, it is possible to classify the tubes according to their principal applications. Tubes may be classified roughly into five groups, viz., potential-amplifier tubes, power-amplifier tubes, current-amplifier tubes, general-purpose tubes, and special-purpose tubes.

1. Potential-amplifier tubes have a relatively high amplification factor and are used where the primary consideration is one of high potential gain. Such tubes usually operate into high impedance loads, either tuned or untuned.

2. Power-amplifier tubes are those which have relatively low values of amplification factor and fairly low values of plate resistance. They are capable of controlling appreciable currents at reasonably high plate potentials.

3. Current-amplifier tubes are those which are designed to give a large change of plate current for a small grid potential; i.e., they possess a high transconductance. These tubes may be required to carry fairly large plate currents. Such tubes find application as both potential and power amplifiers, depending on the tube capacity, and are used extensively in sweep generating circuits.

4. General-purpose amplifier tubes are those whose characteristics are intermediate between the potential- and the power-amplifier tubes. They must have a reasonably high amplification factor and yet must be able to supply some power.

5. Special-purpose tubes include a wide variety of types. The hexode, heptode, and multiunit tubes are of this type.

Amplifiers are classified according to their frequency range, the method of tube operation, and the method of interstage coupling. For example, they may be classed as direct-coupled amplifiers, audio-frequency (a-f) amplifiers, video amplifiers, or tuned r-f amplifiers if some indication of the frequency of operation is desired. Also, the position of the quiescent point and the extent of the tube characteristic that is being used will determine the method of tube operation. This will specify whether the

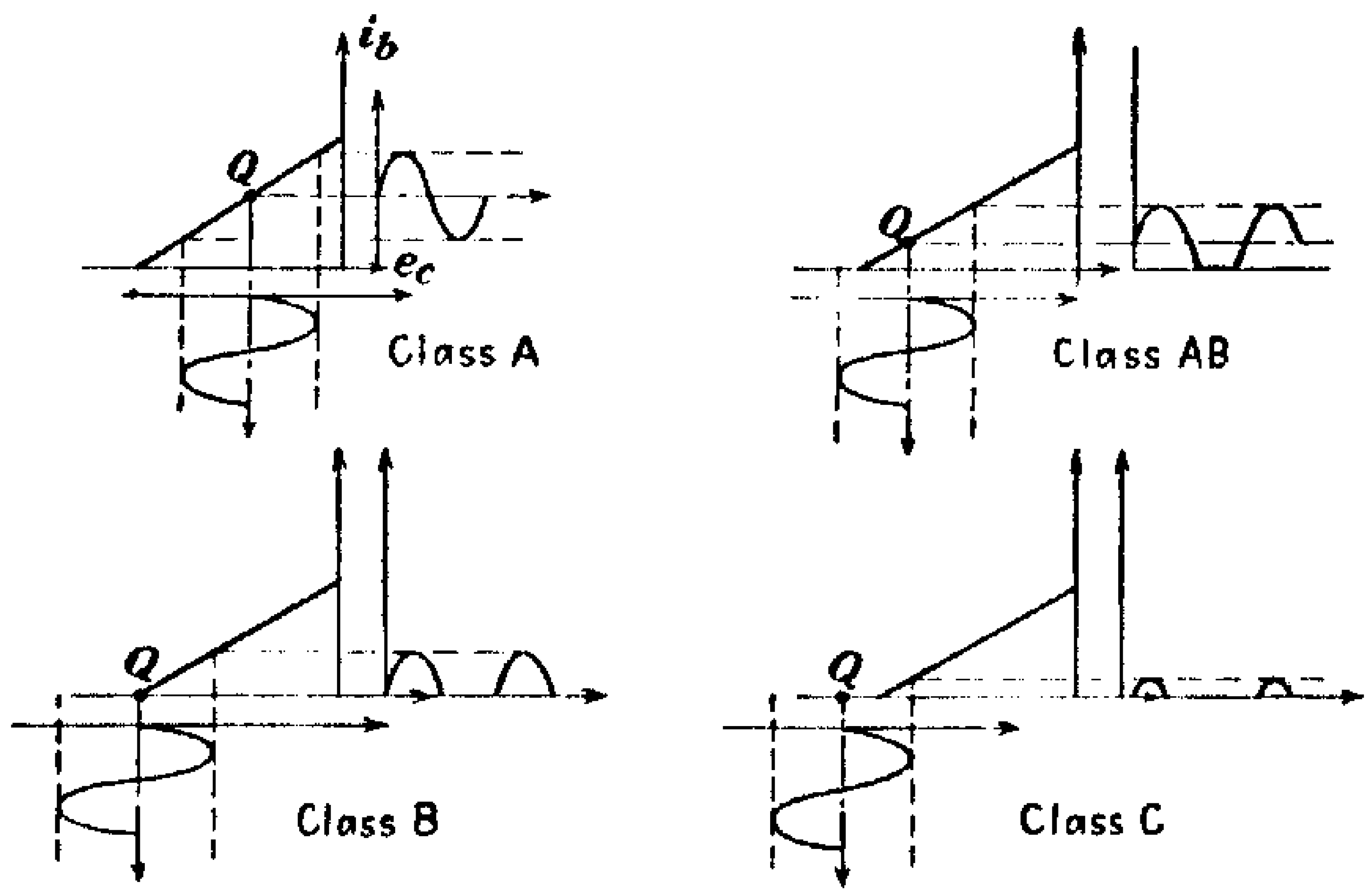


Fig. 3-1. Amplifier classification in terms of the position of the quiescent point of the tube.

tube is being operated in class A, class AB, class B, or class C. These definitions are illustrated graphically in Fig. 3-1.

1. A class A amplifier is an amplifier in which the grid bias and the a-c grid potentials are such that plate current flows in the tube at all times.
2. A class AB amplifier is one in which the grid bias and the a-c grid potentials are such that plate current flows in the tube for appreciably more than half but less than the entire electrical cycle.
3. A class B amplifier is one in which the grid bias is approximately equal to the cutoff value of the tube, so that the plate current is approximately zero when no exciting grid potential is applied, and such that plate current flows for approximately one-half of each cycle when an a-c grid potential is applied.
4. A class C amplifier is one in which the grid bias is appreciably greater than the cutoff value, so that the plate current in each tube is zero when no a-c grid potential is applied, and such that plate current flows for appreciably less than one-half of each cycle when an a-c grid potential is applied.

To indicate that grid current does not flow during any part of the input cycle, the subscript 1 is frequently added to the letter or letters of the class identification. The subscript 2 is added to denote that grid current does flow during some part of the cycle. For example, the designation class AB_1 indicates that the amplifier operates under class AB conditions and that no grid current flows during any part of the input cycle.

Potential amplifiers, whether tuned or untuned, generally operate in class A. Low-power audio amplifiers may be operated under class A and with special connections, under class AB or class B conditions. Tuned r-f power amplifiers are operated either under class B or under class C conditions. Oscillators usually operate under class C conditions. A detailed discussion is deferred until the appropriate points in the text. When a tube is used essentially as a switch, no classification is ordinarily specified.

3-2. Distortion in Amplifiers. The application of a sinusoidal signal to the grid of an ideal class A amplifier will be accompanied by a sinusoidal output wave. Frequently the output waveform is not an exact replica of the input-signal waveform because of distortion that results either within the tube or from the influence of the associated circuit. The distortions that may exist either separately or simultaneously are nonlinear distortion, frequency distortion, and delay distortion. These are defined as follows:

1. Nonlinear distortion is that form of distortion which occurs when the ratio of potential to current is a function of the magnitude of either.
2. Frequency distortion is that form of distortion in which the change is in the relative magnitudes of the different frequency components of a wave, provided that the change is not caused by nonlinear distortion.
3. Delay distortion is that form of distortion which occurs when the phase angle of the transfer impedance with respect to two chosen pairs of terminals is not linear with frequency within a desired range, the time of transmission, or delay, varying with frequency in that range.

In accordance with definition 1, nonlinear distortion results when new frequencies appear in the output which are not present in the input signal. These new frequencies arise from the existence of a nonlinear dynamic curve and were discussed in Sec. 2-8.

Frequency distortion arises when the components of different frequency are amplified by different amounts. This distortion is usually a function of the character of the circuits associated with the amplifier. If the gain vs. frequency characteristic of the amplifier is not a horizontal straight line over the range of frequencies under consideration, the circuit is said to exhibit frequency distortion over this range.

Delay distortion, also called *phase-shift distortion*, results from the fact that the phase shift of waves of different frequency in the amplifier is different. Such distortion is not of importance in amplifiers of the a-f type, since delay distortion is not perceptible to the ear. It is very objectionable in systems that depend on waveshape for their operation, as, for example, in television or facsimile systems. If the phase shift is proportional to the frequency, a time delay will occur although no distortion is introduced. To see this, suppose that the input signal to the amplifier has the form

$$e_1 = E_{m1} \sin (\omega t + \theta_1) + E_{m2} \sin (2\omega t + \theta_2) + \cdots \tag{3-1}$$

If the gain K is constant in magnitude but possesses a phase shift that is proportional to the frequency, the output will be of the form

$$e_2 = KE_{m1} \sin (\omega t + \theta_1 + \psi) + KE_{m2} \sin (2\omega t + \theta_2 + 2\psi) + \cdots$$

This output potential has the same waveshape as the input signal, but a time delay between these two waves exists. By writing

$$\omega t' = \omega t + \psi$$

then

$$e_2 = KE_{m1} \sin (\omega t' + \theta_1) + KE_{m2} \sin (2\omega t' + \theta_2) + \cdots \tag{3-2}$$

This is simply the expression given by Eq. (3-1), except that it is referred to a new time scale t'. Delay distortion, like frequency distortion, arises from the frequency characteristics of the circuit associated with the vacuum tube.

It is not possible to achieve such a linear phase characteristic with simple networks, but it may be approximated with special phase-equalizing networks.

3-3. The Decibel; Power Sensitivity. In many problems where two power levels are to be compared, it is found very convenient to compare the relative powers on a logarithmic rather than on a direct scale. The unit of this logarithmic scale is called the *bel*. A *decibel*, which is abbreviated db, is $\frac{1}{10}$ bel. By definition, two power sources are in the ratio of N bels, according to the relation

$$\text{Number of bels} = \log_{10} \frac{P_2}{P_1}$$

or

$$\text{Number of db} = 10 \log_{10} \frac{P_2}{P_1} \tag{3-3}$$

It should be emphasized that the bel or the decibel denotes a power ratio. Consequently the specification of a certain power in decibels is meaningless unless a reference level is implied or is explicitly specified. In communication applications, it is usual practice to specify 6 mw as the zero reference level. However, any power may be designated as the zero reference level in any particular problem.

Suppose that these considerations are applied to a power amplifier, with P_2 the output power and P_1 the input power. This assumes that the input circuit to the amplifier absorbs power. If the grid circuit does not absorb an appreciable power, then the term *decibel gain* of the amplifier means nothing. Under such conditions, it is customary to speak of *power sensitivity*, which is defined as the ratio of the power output to the square of the input signal potential. Thus

$$\text{Power sensitivity} \equiv \frac{P_2}{E_1^2} \quad \text{mhos} \tag{3-4}$$

where P_2 is the power output in watts and E_1 is the input signal rms volts.

If the input and output impedances are equal resistances, then $P_2 = E_2^2/R$ and $P_1 = E_1^2/R$, where E_2 and E_1 are the output and input potentials. Under this condition, Eq. (3-3) reduces to

$$\text{Number of db} = 20 \log_{10} \frac{E_2}{E_1} \tag{3-5}$$

In general, the input and output resistances are not equal. Despite this, this expression is adopted as a convenient definition of the decibel potential gain of an amplifier. It is essential, however, when the gain of an amplifier is discussed, that it be clearly stated whether one is referring to potential gain or power gain, as these two figures will be different, in general.

Many of the considerations of the foregoing sections are best illustrated by several examples.

Example 1. Calculate the gain of the grounded-grid amplifier circuit of Fig. 3-2.

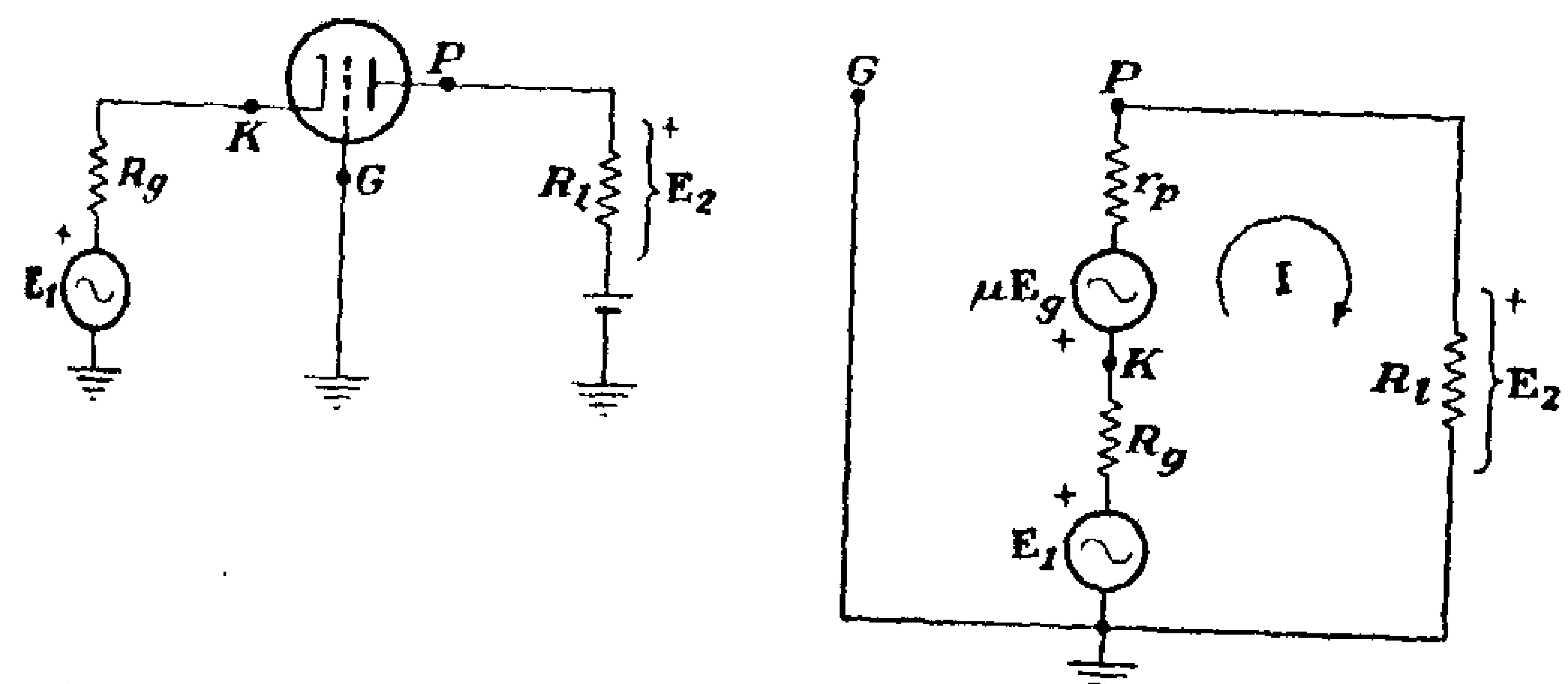


FIG. 3-2. Schematic and equivalent circuits of a grounded-grid amplifier.

Solution. The equivalent circuit of the amplifier is drawn according to the rules of Sec. 2-6 and is that shown in Fig. 3-2*b*. The application of the Kirchhoff potential law to the equivalent circuit yields

$$\mathbf{E}_1 - \mu\mathbf{E}_g - \mathbf{I}(r_p + R_g + R_l) = 0 \tag{3-6}$$

Also from the diagram

$$\mathbf{E}_g = -\mathbf{E}_1 + \mathbf{I}R_g \tag{3-7}$$

Combine the two equations to find

$$\mathbf{E}_1 - \mu(-\mathbf{E}_1 + \mathbf{I}R_g) - \mathbf{I}(r_p + R_g + R_l) = 0$$

The plate current is then given by

$$\mathbf{I} = \frac{\mathbf{E}_1(\mu + 1)}{r_p + (\mu + 1)R_g + R_l} \tag{3-8}$$

The corresponding output potential is

$$\mathbf{E}_2 = \mathbf{I}R_l = \frac{(\mu + 1)R_l\mathbf{E}_1}{(\mu + 1)R_g + r_p + R_l} \tag{3-9}$$

The gain, or potential amplification of this amplifier, which is the ratio of the output to the input potential, is

$$\mathbf{K} = \frac{\mathbf{E}_2}{\mathbf{E}_1} = \frac{R_l}{R_g + \dfrac{r_p + R_l}{\mu + 1}} \tag{3-10}$$

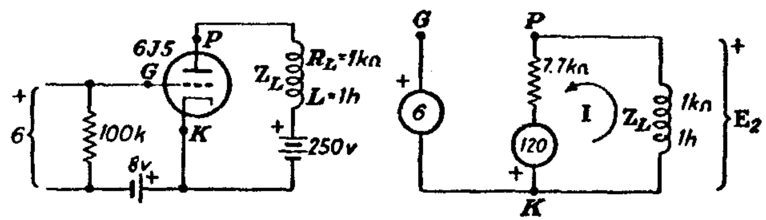
The input impedance is given as the ratio $\mathbf{E}_1/\mathbf{I}$ and is

$$\mathbf{Z}_1 = \frac{\mathbf{E}_1}{\mathbf{I}} = \frac{r_p + R_l}{\mu + 1} + R_g \tag{3-11}$$

which, for small R_g, is quite small. This means, of course, that heavy loading of the driving source may exist if R_g is small.

It is of interest to compare the results of this example with Example 1 of Sec. 2-6. Observe that it is possible to apply the signal either in the grid circuit or in the cathode circuit and still achieve operation of the tube, although the input impedance is different in the two cases.

Example 2. A type 6J5 triode for which $\mu = 20$, $r_p = 7{,}700$ ohms is employed in an amplifier, the load of which consists of an inductor for which $R_L = 1{,}000$ ohms and $L = 1$ henry. Calculate the gain and phase shift of the amplifier at $\omega = 2{,}000$ rad/sec and $\omega = 10{,}000$ rad/sec. Draw the complete sinor diagram of the system. The input signal is 6 volts rms.



Solution. The schematic and equivalent circuits are shown in the accompanying diagrams. At $\omega = 2{,}000$ rad/sec,

$$\mathbf{I} = \frac{120 + j0}{7{,}700 + (1{,}000 + j2{,}000)} = 13.1 - j3.01 \text{ ma}$$

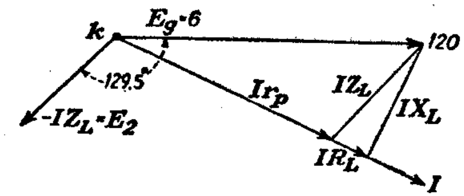
The output potential is

$$\begin{aligned}\mathbf{E}_2 &= -(1{,}000 + j2{,}000)(13.1 - j3.01) \times 10^{-3} \\ &= -(19.1 + j23.2) = 30.1\underline{/-129.5°}\end{aligned}$$

The gain is given by

$$\mathbf{K} = \frac{\mathbf{E}_2}{\mathbf{E}_1} = \frac{30.1\underline{/-129.5°}}{6\underline{/0}} = 5.01\underline{/-129.5°}$$

The potential sinor diagram has the form shown in the sketch. At $\omega = 10{,}000$ rad/sec,



$$\mathbf{I} = \frac{120 + j0}{7{,}700 + (1{,}000 + j10{,}000)} = 5.94 - j6.83 \text{ ma}$$

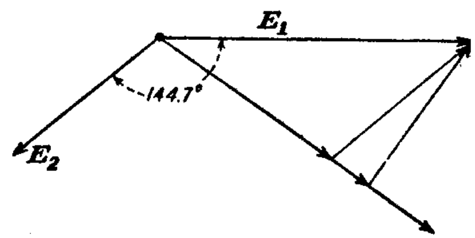
The output potential is

$$\begin{aligned}\mathbf{E}_2 &= -(1{,}000 + j10{,}000)(5.94 - j6.83) \times 10^{-3} \\ &= -(74.2 + j52.6) \\ &= 90.8\underline{/-144.7°}\end{aligned}$$

The gain is given by

$$\mathbf{K} = \frac{90.8\underline{/-144.7°}}{6\underline{/0}} = 15.1\underline{/-144.7°}$$

The potential sinor diagram has the form of the accompanying diagram.



The results are tabulated for convenience. An examination of the results indicates the presence of frequency distortion, since the gain at $\omega = 2{,}000$ rad/sec is different from that at $\omega = 10{,}000$ rad/sec. Also, phase-shift distortion exists in this amplifier.

ω	Gain and phase	Potential db gain
2,000	$5.01\underline{/-129.5°}$	14 db
10,000	$15.1\underline{/-144.7°}$	23.6 db

3-4. Interelectrode Capacitances in a Triode. It was assumed in the foregoing discussions that with a negative bias on the grid the grid driving-source current was negligible. This is generally true if one examines only the current intercepted by the grid because of its location within the region of the electron stream. Actually though, owing to the physical proximity of the elements of the tube, interelectrode capacitances between pairs of elements exist. These capacitances are important in the behavior of the circuit, as charging currents do exist.

Owing to the capacitance that exists between the plate and the grid, it is not true that the grid circuit is completely independent of the

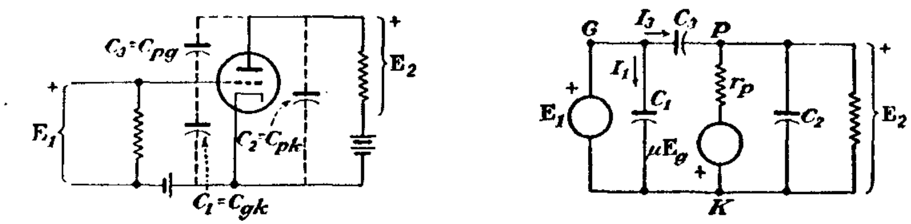


FIG. 3-3. Schematic and equivalent circuits of an amplifier, including the interelectrode capacitances.

plate circuit. Since the capacitance between plate and grid is small, the approximation that the plate circuit is independent of the grid circuit is valid at the lower frequencies. However, at the higher frequencies, interelectrode capacitances may seriously affect the operation.

A more complete schematic diagram and its equivalent circuit are given in Fig. 3-3. In this circuit, C_{pg} denotes the capacitance between the grid and the plate, C_{gk} is the grid-cathode capacitance, and C_{pk} is the capacitance between the plate and the cathode. The solution for the gain of this circuit is readily obtained with the aid of the Millman theorem (see Appendix A). The point O' corresponds to the plate terminal P, and the point O is the cathode terminal K. Four branches must be considered between these points: the load impedance with zero potential; the capacitor C_2 with zero potential; the potential rise $\mu\mathbf{E}_g$ in series with r_p; the potential $\mathbf{E}_1$ in series with C_3. The capacitor C_1 which exists across the input $\mathbf{E}_1$ does not appear in the equation. The result is

$$\mathbf{E}_2 = \frac{-\mu\mathbf{E}_g\mathbf{Y}_p + \mathbf{E}_1\mathbf{Y}_3}{\mathbf{Y}_p + \mathbf{Y}_l + \mathbf{Y}_2 + \mathbf{Y}_3} \qquad (3\text{-}12)$$

where $\mathbf{Y}_p = 1/r_p$ is admittance corresponding to r_p
$\mathbf{Y}_2 = j\omega C_2$ is admittance corresponding to C_2
$\mathbf{Y}_3 = j\omega C_3$ is admittance corresponding to C_3
$\mathbf{Y}_l = 1/\mathbf{Z}_l$ is admittance corresponsing to $\mathbf{Z}_l$
$\mathbf{E}_2$ = potential difference between P and K, or potential across load impedance

Note that $\mathbf{E}_1 = \mathbf{E}_g$. The potential gain is given by

$$\mathbf{K} = \frac{\text{output potential}}{\text{input potential}} = \frac{\mathbf{E}_2}{\mathbf{E}_1}$$

and may be written in the form

$$\mathbf{K} = \frac{\mathbf{Y}_3 - g_m}{\mathbf{Y}_p + \mathbf{Y}_l + \mathbf{Y}_2 + \mathbf{Y}_3} \qquad (3\text{-}13)$$

In this expression use has been made of the fact that $g_m = \mu/r_p$.

In this analysis a number of factors have been neglected. It has been assumed that no conduction or leakage currents exist between tube terminals. Such leakage current will depend upon many variable factors, such as the spacing between electrodes, the materials of the base, the conditions of the surface of the glass and the tube base, and perhaps the surface leakage between connecting wires. Ordinarily the error is small in neglecting the effects of this surface leakage. If this assumption is not true, the effect can be taken into account by writing for each interelectrode admittance $g_s + j\omega C_s$ instead of $j\omega C_s$, where g_s takes account of the leakage current and also dielectric losses. Interwiring and stray capacitances must be taken into account. This may be done by considering them to be in parallel with C_1, C_2, and C_3. Additional considerations are necessary at the high frequencies. These are discussed in Sec. 3-8.

The error made in the calculation of the gain by neglecting the interelectrode capacitances is very small over the a-f spectrum. These interelectrode capacitances are usually 10 $\mu\mu$f or less, which corresponds to admittances of less than 2 μmhos at 20,000 cps. This is to be compared with the mutual conductance of the tube of, say, 1,500 μmhos at the normal operating point. Likewise $\mathbf{Y}_2 + \mathbf{Y}_3$ is usually negligible compared with $\mathbf{Y}_p + \mathbf{Y}_l$. Under these conditions, the expression for the gain [Eq. (3-13)] reduces to Eq. (2-16).

3-5. Input Admittance of a Triode. Owing to the presence of the interelectrode capacitances, the grid circuit is no longer isolated from the plate circuit. In fact, with an increasing signal on the grid and with the consequent decreasing potential on the plate, an appreciable change of potential appears across the capacitance C_{pg}, with a consequent appreciable current flow. Also, the potential change across the capacitance C_{gk} is accompanied by a current flow. Clearly, therefore, the input-signal source must supply these currents. To calculate this current, it is noted from the diagram that

$$\mathbf{I}_1 = \mathbf{E}_1\mathbf{Y}_1$$

and

$$\mathbf{I}_3 = \mathbf{E}_{gp}\mathbf{Y}_3 = (\mathbf{E}_1 - \mathbf{E}_2)\mathbf{Y}_3$$

But from the fact that

$$\mathbf{E}_2 = \mathbf{K}\mathbf{E}_1$$

then the total input current is

$$\mathbf{I}_i = \mathbf{I}_1 + \mathbf{I}_3 = [\mathbf{Y}_1 + (1 - \mathbf{K})\mathbf{Y}_3]\mathbf{E}_1$$

The input admittance, given by the ratio $\mathbf{Y}_i = \mathbf{I}_i/\mathbf{E}_1$, is

$$\mathbf{Y}_i = \mathbf{Y}_1 + (1 - \mathbf{K})\mathbf{Y}_3 \tag{3-14}$$

If $\mathbf{Y}_i$ is to be zero, evidently both $\mathbf{Y}_1$ and $\mathbf{Y}_3$ must be zero, since $\mathbf{K}$ cannot, in general, be $1\underline{/0}$ deg. Thus, for the system to possess a negligible input admittance over a wide range of frequencies, the grid-cathode and the grid-plate capacitances must be negligible.

Consider a triode with a pure resistance load. At the lower frequencies, the gain is given by the simple expression [Eq. (2-16)]

$$\mathbf{K} = \frac{-\mu R_l}{R_l + r_p}$$

In this case, Eq. (3-14) becomes

$$\mathbf{Y}_i = j\omega\left[C_1 + \left(1 + \frac{\mu R_l}{R_l + r_p}\right)C_3\right] \tag{3-15}$$

Thus the input admittance is that from a capacitor between grid and cathode of magnitude

$$C_i = C_1 + \left(1 + \frac{\mu R_l}{R_l + r_p}\right)C_3 \tag{3-16}$$

Attention is called to the very large contribution to the input capacitance by the grid-plate capacitance C_3, owing to the fact that its magnitude is multiplied by the amplifier gain. As a result, the total input capacitance is very much higher than any of the interelectrode capacitances. The presence of this input capacitance will be found to affect the operation of the amplifier, and often will make operation impossible, especially at the higher frequencies. Methods of compensation have been devised to overcome this effect, and these will be examined later.

For the general case when the gain of the amplifier $\mathbf{K}$ is a complex quantity, the input admittance will consist of two terms, a resistive and a reactive term. For the case of an inductive load, the gain $\mathbf{K}$ may be written in the form (see Sec. 3-3, Example 2)

$$\mathbf{K} = -(k_1 + jk_2) \tag{3-17}$$

and Eq. (3-14) becomes

$$\mathbf{Y}_i = -\omega C_3 k_2 + j\omega[C_1 + (1 + k_1)C_3] \tag{3-18}$$

This expression indicates that the equivalent input circuit comprises a resistance (which is negative in this particular case, although it will be positive for a capacitive load) in parallel with a capacitance C_i, as shown in Fig. 3-4. The equivalent elements have the form

$$R_i = -\frac{1}{\omega C_3 k_2} \tag{3-19}$$

and the capacitor

$$C_i = C_1 + (1 + k_1)C_3 \tag{3-20}$$

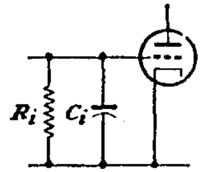


FIG. 3-4. The equivalent input circuit of a triode.

As indicated in the above development, it is possible for the term k_2 to be negative (with an inductive load). Under these circumstances the input resistance R_i will be negative. Physically, this means that power is being fed back from the output circuit into the grid circuit through the coupling provided by the grid-plate capacitance. If this feedback reaches an extreme stage, the amplifier will oscillate. These feedback effects in an amplifier will be examined in some detail in Chap. 5.

3-6. Input Admittance of a Tetrode. The equivalent circuit of the tetrode is essentially that of the triode, even though a screen grid exists in the tetrode. A schematic diagram of a simple amplifier circuit employing a tetrode is given in Fig. 3-5. In drawing the equivalent circuit, the

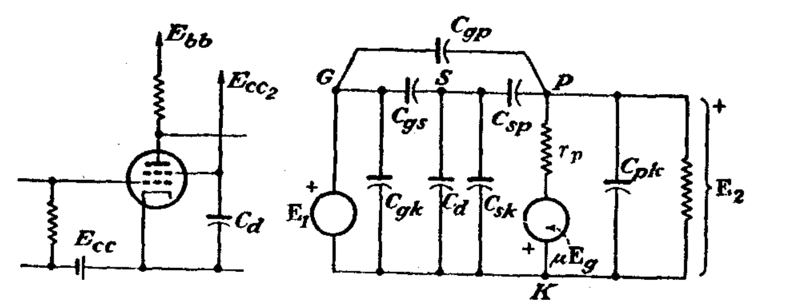


FIG. 3-5. Schematic and equivalent circuits of a tetrode in an amplifier circuit.

rules given in Sec. 2-6 have been appropriately extended and employed. This requires the introduction of a point S, the screen terminal, in addition to the points K, G, and P.

Notice that the screen potential is maintained at a fixed d-c potential with respect to cathode and is at zero potential in so far as a-c variations about the Q point are concerned. As indicated in the figure, this effectively places a short circuit across C_{ks} and puts C_{gk} and C_{gs} in parallel. This parallel combination is denoted C_1. The capacitance C_{ps} now appears from plate to cathode and is effectively in parallel with C_{pk}. This parallel combination is denoted C_2. Also, from the discussion in

Sec. 1-14, the shielding action of the screen is such that the capacitance C_{pg} between grid and plate is very small. If this capacitance is assumed to be negligible, and it is less than 0.001 $\mu\mu$f in the average potential tetrode, then Fig. 3-5 may be redrawn in the form shown in Fig. 3-6. In this figure, the capacitances have the values

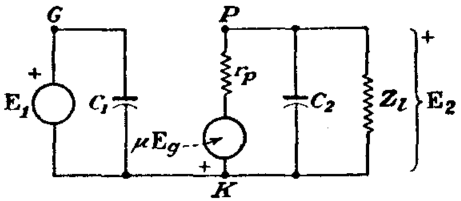


FIG. 3-6. The ideal equivalent circuit of a tetrode amplifier.

$$C_1 = C_{gk} + C_{gs} \qquad C_2 = C_{ps} + C_{pk} \tag{3-21}$$

The input admittance of the tube is then

$$\mathbf{Y}_i = j\omega C_1 \tag{3-22}$$

The mere substitution of a tetrode for a triode may not result in a very marked improvement in the amplifier response. This follows from the fact that the stray and wiring capacitances external to the tube may allow significant grid-plate coupling. It is necessary that care be exercised in order that plate and grid circuits be shielded or widely separated from each other in order to utilize the inherent possibilities of the tube.

3-7. Input Admittance of a Pentode. The discussion in Sec. 1-14 showed that, even though the tetrode had a significantly smaller grid-plate capacitance than the triode, the presence of the screen grid was accompanied by the effects of secondary emission from the plate when the instantaneous plate potential fell below the screen potential. As discussed, the effect of this is overcome by the insertion of a suppressor grid between the screen grid and the plate.

When used in a circuit as a potential amplifier, the pentode is connected in the circuit exactly like the tetrode with the addition that the suppressor grid is connected to the cathode. By drawing the complete equivalent circuit of the pentode amplifier, by appropriately extending the rules of Sec. 2-6, and by including all tube capacitances, it is easy to show that the equivalent circuit reduces to that shown in Fig. 3-7. In this diagram

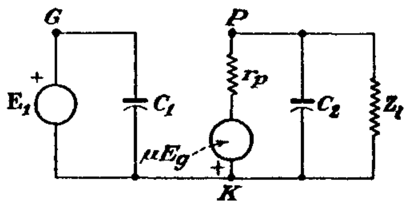


FIG. 3-7. The equivalent circuit of a pentode amplifier.

$$C_1 = C_{gk} + C_{gs} \qquad C_2 = C_{pk} + C_{ps} + C_{p3} \tag{3-23}$$

where C_{p3} is the plate-grid No. 3 capacitance

The plate load impedance $\mathbf{Z}_l$ is frequently much smaller than the plate resistance of the tube, and it is convenient to use the current-source equivalent-circuit representation of the tube, as shown in Fig. 3-8. For the range of frequencies over which the input and output capacitances C_1 and C_2 are negligible, and with $r_p \gg \mathbf{Z}_l$, the total generator current passes

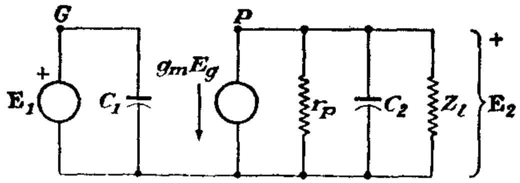


FIG. 3-8. The current-source equivalent circuit of the pentode amplifier.

through $\mathbf{Z}_l$. Under these circumstances the output potential is

$$\mathbf{E}_2 = -g_m \mathbf{E}_1 \mathbf{Z}_l$$

and the gain is given by the simple form

$$\mathbf{K} = -g_m \mathbf{Z}_l \tag{3-24}$$

If the assumed conditions are not valid, then the gain becomes

$$\mathbf{K} = -g_m \mathbf{Z} \tag{3-25}$$

where $\mathbf{Z}$ is the combined parallel impedance in the output circuit.

3-8. High-frequency (H-F) Considerations. In addition to the effects of the interelectrode capacitances in affecting the performance of an amplifier, several other factors are of importance, particularly at the higher frequencies. Some were mentioned in Sec. 3-4, plus the effects of lead inductances and also the effects due to transit time.

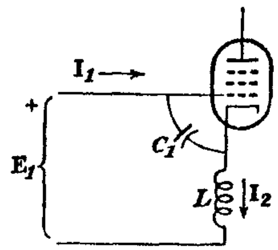


FIG. 3-9. Circuit for examining the effect of cathode lead inductance in a pentode.

To examine the effect of the cathode lead inductance, Fig. 3-9 is analyzed. For convenience, it will be assumed that the grid is negative throughout the cycle and that transit-time effects are negligible. Then for $\mathbf{I}_2 \gg \mathbf{I}_1$

$$\mathbf{I}_2 = g_m \mathbf{E}_g$$

and

$$\mathbf{E}_g = \frac{\mathbf{I}_1}{j\omega C_1}$$

Also

$$\mathbf{E}_1 = \frac{\mathbf{I}_1}{j\omega C_1} + j g_m \mathbf{E}_g \omega L$$

Combine equations to get

$$\mathbf{E}_1 = \frac{\mathbf{I}_1}{j\omega C_1} + jg_m\omega L\frac{\mathbf{I}_1}{j\omega C_1}$$
$$= \frac{\mathbf{I}_1}{j\omega C_1}(1 + jg_m\omega L)$$

The input admittance is

$$\mathbf{Y}_i = \frac{\mathbf{I}_1}{\mathbf{E}_1} = \frac{j\omega C_1}{1 + jg_m\omega L} = \frac{j\omega C_1(1 - j\omega g_m L)}{1 + \omega^2 g_m^2 L^2} \tag{3-26}$$

If $\omega^2 g_m^2 L^2 \ll 1$, then

$$\mathbf{Y}_i = \omega^2 g_m L C_1 + j\omega C_1 \tag{3-27}$$

Observe, therefore, that the cathode lead inductance introduces an input conductance of amount $\omega^2 g_m L C_1$.

A second component of input conductance arises because of the transit time of the electrons between cathode and plate. An exact calculation is difficult, but a qualitative explanation is possible which indicates the grid-loading effects involved. To understand grid loading, consider an electron that has left the cathode and is approaching the grid in its flight to the anode. Suppose that the grid potential is negative relative to the cathode so that no electrons are collected by the grid. As the electron approaches the grid, a changing image-charge density will be induced on the grid (see Sec. 1-4 for a discussion of image charges). This changing image charge represents an instantaneous grid current, the direction of flow of charge being such as to charge the bias battery. The power for this charging process is supplied by the moving electron, and as a result the electron is decelerated.

Once the electron has passed the grid, the process is reversed, and the moving electron receives energy from the grid, and it is accelerated thereby. The amount of energy lost by the electron as it approaches the grid is just equal to that which it gains as it moves away, and the net energy change is zero. As a result, the net grid loading is zero.

If the transit time of the electron in the cathode-anode space is of the order of the frequency of the applied grid potential, the grid loading becomes important, for now the electron can no longer be considered to be in a field which is constant in time. It is possible for the energy that is supplied to the grid by the moving electron to exceed the amount of energy that is returned by the grid in its interelectrode flight, with a resultant net energy loss in the grid circuit. This energy is supplied by the grid driving source, and it represents a load on this source.

From a circuits point of view, the foregoing may be described in terms of an induced current in the grid. At the lower frequencies, the induced grid current is 90 deg out of phase with the grid potential, with a consequent zero net power loss. At the higher frequencies, an inphase component exists. This inphase component reduces the input resistance, and this may produce an appreciable loading of the input circuit.

The foregoing concepts may be employed to indicate in a qualitative way the effect of the various factors on the input resistance. If T denotes the transit time, f denotes the frequency of the applied grid potential, and g_m is the mutual conductance of the tube, it is expected that the grid current I_g is proportional to T and f, since I_g is small if either of these is small. Also, I_g should be proportional to g_m, since g_m determines the a-c component of plate current for a specified E_g, and the total grid current is proportional to this a-c component of the plate current. If α denotes the transit angle, which is now less than 90 deg, then the inphase component of I_g is $I_g \cos \alpha$, which is simply $I_g\alpha$ for small deviations from 90 deg. But α is also proportional to T and f. Thus the inphase component of I_g is proportional to $g_m T^2 f^2$, or

$$g_i = kg_m f^2 T^2 \tag{3-28}$$

where k is a constant depending on the geometry of the tube and electrode potentials. This relationship agrees with the complete analyses of North, Llewellyn, and Benham.

It will be seen from Eqs. (3-27) and (3-28) that g_i and the conductance component of the cathode inductance depend on the frequency in the same way. Consequently, these components cannot be separated readily in measurement of input resistance or conductance.

Tubes for use at the high frequencies are made in a manner to reduce transit time, interelectrode capacitances, and lead inductances. This is done by means of very close electrode spacing, and generally small physical dimensions of electrodes. Among such tubes are the so-called "acorn," "doorknob," "pencil," and "disk-seal," or "lighthouse," tubes, with upper limits in frequency of approximately 2,000, 1,700, 3,000, and 3,500 Mc, respectively. These names are indicative of the external envelope shape, the first three possessing essentially cylindrical electrode structures, the last being essentially of a planar construction. The first two have the leads brought out of the envelopes at widely spaced points, in order to reduce capacitances. The latter two bring the leads out in the form of disks. At the higher frequencies these tubes are incorporated in coaxial line resonators, lead inductances being unimportant as these form part of the resonant cavities.

3-9. Potential Sources for Amplifiers. A number of different potential sources are required in an amplifier. These are the following: the filament, or A, supply; the plate, or B, supply E_{bb}; the grid-bias, or C, supply E_{cc}; the screen supply E_{cc2}. These potentials are supplied in different ways.

The Filament, or A, Supply. The most common method of heating the cathodes of indirectly heated tubes is from a low-potential winding on a transformer which operates from the a-c supply lines. Storage batteries may be used if d-c heating is necessary, but this is ordinarily not necessary except in special applications. Special low-drain tubes are available for use in portable radio sets and are fed from dry batteries.

The Plate, or B, Supply E_{bb}. Most equipments involving the use of electron tubes are operated from the a-c supply mains, and the d-c plate supply is then secured by means of a rectifier and filter unit (see Chaps. 6 and 7 for details). For applications with severe requirements on regulation or low ripple, the power supply must be electronically regulated. For low-drain requirements, dry batteries may be used.

The Grid, or C, Supply E_{cc}. The grid circuit of most amplifiers ordinarily requires very little current, and hence low-power dry batteries may be used. In most cases, however, self-bias is used (although this is restricted to class A and class AB amplifiers). Self-bias is achieved by including a resistor R_k in the cathode of the amplifier tube and shunting this resistor with a capacitor C_k, the reactance of which is small compared with R_k over the operating frequency range (see Fig. 3-10). The quiescent current I_b flows through this resistor, and the potential difference provides the grid bias. The correct self-biasing resistance $R_k = E_{cc}/I_b$.

The capacitor C_k serves to by-pass any a-c components in the plate current, so that no a-c component appears across the resistor R_k. If such an a-c component, or varying bias, does exist, then clearly there is a reaction between the plate circuit and the input circuit. Such a "feedback" effect will receive detailed consideration in Chap. 5. If this effect is to be avoided, large-capacitance capacitors may be required, particularly if the frequency is low. High-capacitance low-potential electrolytic capacitors are available for this specific service and are quite small physically.

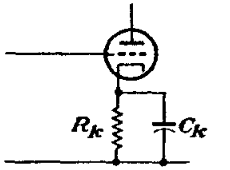


FIG. 3-10. Obtaining self-bias by means of a cathode resistor.

The Screen Supply E_{cc2}. The screen supply is ordinarily obtained from the plate-supply source. In many cases the screen potential is lower than the plate supply, and it is usual practice to connect the screen to the plate supply through a resistor. The resistor is chosen of such a size that the potential drop across it due to the screen current will set the screen at the desired potential. A capacitor is then connected from the screen to the cathode so as to maintain this potential constant and independent of B-supply variations or variations in the screen current.

It is customary to use a common B supply for all tubes of a given amplifier circuit. Because of this, the possibility for interactions among the stages through this common plate supply does exist and might be troublesome unless the effective output impedance of the power-supply unit is very small. It is necessary in some applications to include *RC* combinations known as *decoupling filters* so as to avoid this interaction.

A typical resistance-capacitance coupled-amplifier circuit which is provided with self-bias, decoupling filters, and screen dropping resistors is illustrated in Fig. 3-11.

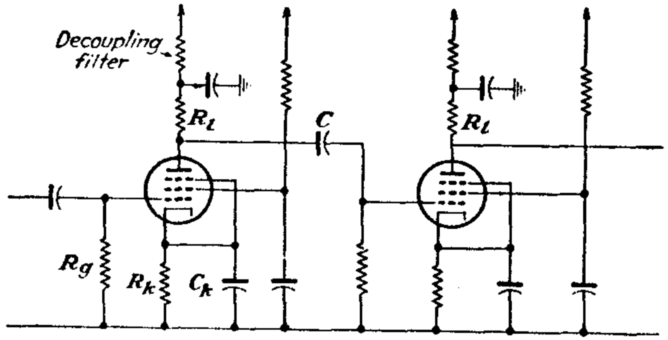


FIG. 3-11. Resistance-capacitance coupled amplifier, with self-bias, decoupling filters, and screen dropping resistors.

3-10. Power Gain of Transistors. A number of different power-gain definitions for use with transistor amplifiers are possible. They have been called (1) operating gain, (2) available gain, (3) maximum available gain, (4) insertion gain, and (5) power gain. These definitions are based on considerations of the transistor as a four-terminal network, with a potential generator $\mathbf{E}_g$ with internal impedance R_g connected to the input terminals and a resistance load R_l connected to the output terminals.

FIG. 3-12. The transistor as a four-terminal network for power-gain calculations.

Refer to Fig. 3-12, which shows the transistor as a four-terminal network, with input source and output load. R_i is the input, or driving-point, impedance with the load resistance in place. The *available generator power* is defined as the maximum power that the generator can deliver when the transistor is not in the circuit. This occurs when $R_g = R_l$. Then, according to the definitions, with the transistor connected in the circuit,

1. Source operating gain

$$= \frac{\text{actual power delivered by generator}}{\text{available generator power}}$$

2. Available gain

$$= \frac{\text{actual power delivered when } R_l \text{ is adjusted for max output}}{\text{available generator power}}$$

3. Max available gain

$$= \frac{\text{max power output (by matching } R_l \text{ and } R_g \text{ to transistor)}}{\text{available generator power}}$$

4. Insertion gain

$$= \frac{\text{power developed in } R_l \text{ when connected to transistor output}}{\text{power in } R_l \text{ when connected directly to source of power}}$$

5. Power gain

$$= \frac{\text{power developed in } R_l \text{ when connected to transistor output}}{\text{actual power delivered by generator}}$$

$$= \frac{\text{insertion gain}}{\text{source operating gain}}$$

Analytical expressions for all five gains are possible for circuits with transistors. The normal power-gain information supplied by the manufacturer is essentially that given by (5). Some details of gain calculations are given in the next section.

3-11. The Grounded-base Amplifier. The equivalent circuit which is most directly analyzed for this configuration is shown in Fig. 3-13.

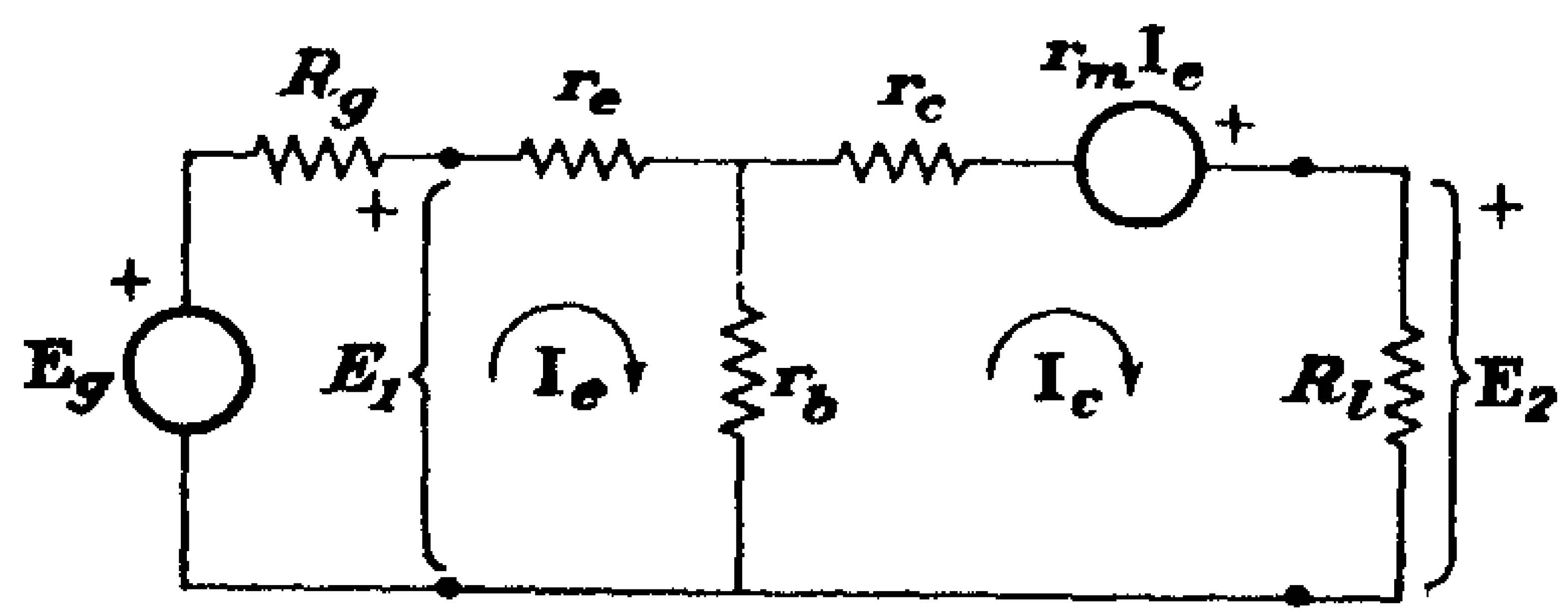


FIG. 3-13. Grounded-base amplifier, equivalent circuit.

Note that this is simply Fig. 2-20c with the driving source and the lead in the circuit. It is noted that the following analysis is valid for the junction and point-contact transistors for small-signal l-f operation.

An application of the Kirchhoff potential law yields the equations

$$\begin{aligned} \mathbf{E}_1 &= (r_e + r_b)\mathbf{I}_e - r_b\mathbf{I}_c \\ 0 &= -(r_m + r_b)\mathbf{I}_e + (r_b + r_c + R_l)\mathbf{I}_c \end{aligned} \tag{3-29}$$

The network determinant $\boldsymbol{\Delta}$ is

$$\begin{aligned} \boldsymbol{\Delta} &= \begin{vmatrix} r_e + r_b & -r_b \\ -(r_m + r_b) & r_b + r_c + R_l \end{vmatrix} \\ &= r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l) \end{aligned} \tag{3-30}$$

Expressions for the currents $\mathbf{I}_e$ and $\mathbf{I}_c$ are obtained directly from Eqs. (3-29) by an application of Cramer's rule. There results

$$\mathbf{I}_e = \frac{\begin{vmatrix} \mathbf{E}_1 & -r_b \\ 0 & r_b + r_c + R_l \end{vmatrix}}{\boldsymbol{\Delta}}$$

which is

$$\mathbf{I}_e = \mathbf{E}_1 \frac{r_b + r_c + R_l}{\boldsymbol{\Delta}} \tag{3-31}$$

Similarly, it follows that

$$\mathbf{I}_c = \frac{\begin{vmatrix} r_e + r_b & \mathbf{E}_1 \\ -(r_m + r_b) & 0 \end{vmatrix}}{\boldsymbol{\Delta}}$$

which is

$$\mathbf{I}_c = \mathbf{E}_1 \frac{r_m + r_b}{\boldsymbol{\Delta}} \tag{3-32}$$

The potential amplification $\mathbf{K}_e$ is given by

$$\mathbf{K}_e = \frac{\mathbf{I}_c R_l}{\mathbf{E}_1} = \frac{(r_m + r_b)R_l}{r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)} \tag{3-33}$$

The current amplification $\mathbf{K}_i$ is given by

$$\mathbf{K}_i = \frac{\mathbf{I}_c}{\mathbf{I}_e} = \frac{r_m + r_b}{r_b + r_c + R_l} \tag{3-34}$$

The input resistance R_i is, from Eq. (3-31),

$$R_i = \frac{\mathbf{E}_1}{\mathbf{I}_e} = \frac{r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)}{r_b + r_c + R_l}$$

which is

$$R_i = r_e + r_b \frac{r_c - r_m + R_l}{r_b + r_c + R_l} \tag{3-35}$$

The power gain is

$$K_p = \frac{I_c^2 R_l}{I_e^2 R_i} = K_i^2 \frac{R_l}{R_i}$$

By combining with Eqs. (3-34) and (3-35) there results

$$K_p = \frac{(r_m + r_b)^2 R_l}{(r_b + r_c + R_l)[r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)]} \tag{3-36}$$

The effective internal impedance is deduced in the regular way and requires an analysis of the circuit of Fig. 3-14. The result, which is left as a problem for the student (see Prob. 3-13), is found to be

$$R_o = r_c - r_b \frac{r_m - R_g - r_c}{R_g + r_e + r_b} \tag{3-37}$$

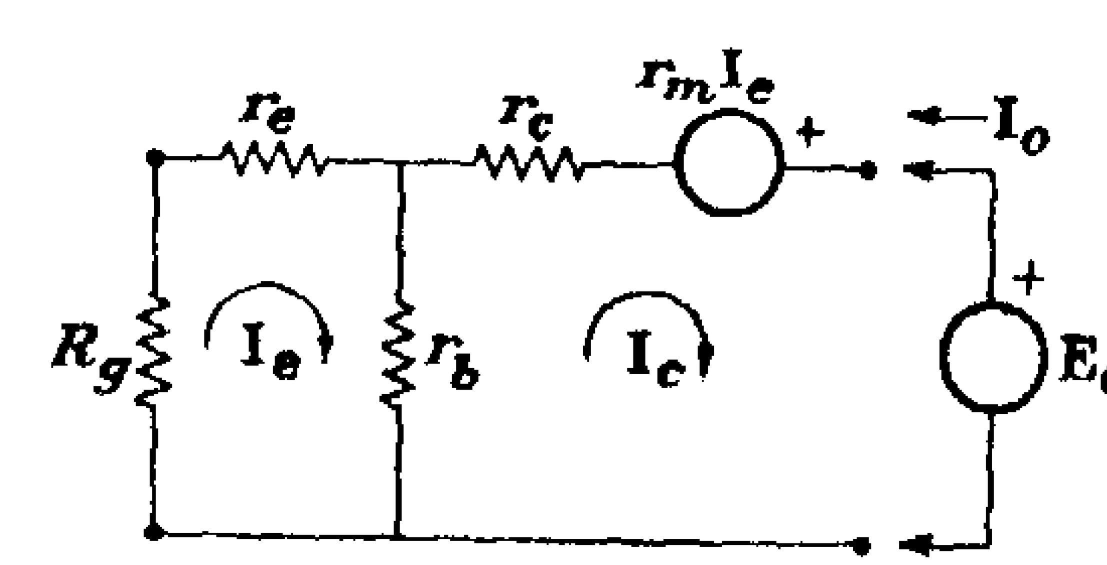


FIG. 3-14. Grounded-base amplifier arranged for determination of the effective internal impedance.

Attention is directed to Table 2-1, which gives typical circuit parameters of point-contact and junction-type transistors. Thus, although the foregoing complete expressions must be used for point-contact transistors and in some circuits using junction transistors, approximate expressions may be deduced for most

TABLE 3-1

GROUNDED-BASE FORMULAS

Factor	Exact formulas	Approximate formulas $r_c - r_m \gg r_e$; $r_c > r_b$; $r_c - r_m \gg R_l \gg r_e$
$\mathbf{K}_e$	$\frac{(r_m - r_b)R_l}{r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)}$	$\frac{aR_l}{r_e + r_b(1 - a)}$
$\mathbf{K}_i$	$\frac{r_m + r_b}{r_b + r_c + R_l}$	a
R_i	$r_e + r_b \frac{r_c - r_m + R_l}{r_b + r_c + R_l}$	$r_e + r_b(1 - a)$
R_o	$r_c - r_b \frac{r_m - R_g - r_e}{R_g + r_e + r_b}$	$r_c \frac{r_e + r_b(1 - a) + R_g}{r_e + r_b + R_g}$
K_p	$\frac{(r_m + r_b)^2 R_l}{(r_b + r_c + R_l)[r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)]}$	$\frac{a^2 R_l}{r_e + r_b(1 - a)}$

TABLE 3-2

GROUNDED-EMITTER FORMULAS

Factor	Exact formulas	Approximate formulas $r_c - r_m \gg r_e$; $r_c \gg r_b$; $r_c - r_m \gg R_l \gg r_e$
$\mathbf{K}_e$	$-\frac{(r_m - r_e)R_l}{r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)}$	$-\frac{aR_l}{r_e + r_b(1 - a)}$
$\mathbf{K}_i$	$-\frac{r_m - r_e}{r_c - r_m + R_l + r_e}$	$-\frac{a}{1 - a}$
R_i	$r_b + r_e \frac{r_c + R_l}{r_c - r_m + R_l + r_e}$	$r_b + \frac{r_e}{1 - a}$
R_o	$r_c - r_m + r_e \frac{R_g + r_b + r_m}{R_g + r_b + r_e}$	$r_c(1 - a) + r_e \frac{r_m + R_g}{R_g + r_b + r_e}$
K_p	$\frac{(r_m - r_e)^2 R_l}{(r_c - r_m + R_l + r_e)[r_b(r_c - r_m + R_l + r_e) + r_e(r_c + R_l)]}$	$\frac{a^2 R_l}{(1 - a)[r_e + r_b(1 - a)]}$

TABLE 3-3

GROUNDED-COLLECTOR FORMULAS

Factor	Exact formulas	Approximate formulas $r_c - r_m \gg r_e$; $r_c \gg r_b$; $r_c - r_m \gg R_l \gg r_e$
$\mathbf{K}_e$	$\frac{r_c R_l}{r_b(r_c - r_m + R_l + r_e) + r_c(r_e + R_l)}$	1
$\mathbf{K}_i$	$\frac{r_c}{r_c - r_m + R_l + r_e}$	$\frac{1}{1 - a}$
R_i	$r_b + r_c \frac{r_e + R_l}{r_c - r_m + R_l + r_e}$	$\frac{R_l}{1 - a}$
R_o	$r_e + (r_c - r_m) \frac{R_g + r_b}{R_g + r_b + r_c}$	$r_e + (r_b + R_g)(1 - a)$
K_p	$\frac{r_c^2 R_l}{(r_c - r_m + R_l + r_e)[r_b(r_c - r_m + R_l + r_e) + r_c(r_e + R_l)]}$	$\frac{1}{1 - a}$

junction-transistor circuits. In particular for the case where the following approximations are valid:

$$\begin{aligned} r_c - r_m &\gg r_e \\ r_c &\gg r_b \\ r_c - r_m &\gg R_l \gg r_e \end{aligned} \qquad (3\text{-}38)$$

then Eqs. (3-33) to (3-37) reduce to the following expressions:

$$\mathbf{K}_e = \frac{aR_l}{r_e + r_b(1 - a)} \qquad (3\text{-}39)$$

$$\mathbf{K}_i = a \qquad (3\text{-}40)$$

$$R_i = r_e + r_b(1 - a) \qquad (3\text{-}41)$$

$$K_p = \frac{a^2 R_l}{r_e + r_b(1 - a)} \qquad (3\text{-}42)$$

$$R_o = r_c \frac{r_e + r_b(1 - a) + R_g}{r_e + r_b + R_g} \qquad (3\text{-}43)$$

The foregoing results are tabulated for convenience in Table 3-1.

A similar series of calculations can be carried out for each of the other transistor configurations. The results of such calculations are given in Tables 3-2 and 3-3.