UNTUNED POWER AMPLIFIERS

The potential amplifiers discussed in Chap. 4 are designed to increase a potential signal from a low level to one which is adequate for operating some low-power circuit. Such amplifiers are generally operated in class A since the amplification is to be accomplished without distortion.

A power amplifier serves to supply an appreciable amount of power to some power-absorbing circuit, although in general it must be accomplished under very low grid-driving-power demands. Power amplifiers may be operated as class A, B, or C or at any point between these limits, the choice of operating conditions being determined by the ultimate purpose of the amplifier. If the amplifier is to reproduce the audio spectrum without distortion, then the amplifier must be operated in class A if a single tube is used. If two tubes are used in a push-pull circuit, then the amplifier may also be operated in class AB or class B. If the amplifier is to reproduce the input waveshape over a very narrow range of frequencies, tuned class B or tuned class C amplifiers may be used. Only a-f amplifiers will be considered in this chapter.

10-1. Class A Triode Power Amplifiers. The basic schematic diagram of a typical series-fed power amplifier and its equivalent circuit is given

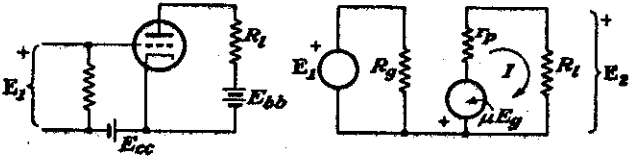


FIG. 10-1. The schematic and equivalent circuits of a simple series-fed power amplifier.

in Fig. 10-1. It is observed that this circuit is identical with that of Fig. 4-1 for the simple potential amplifier.

If it is assumed that the dynamic curve is linear over the entire range of operation, then the plate current is given by

$$I = \frac{-\mu E_1}{R_l + r_p} \tag{10-1}$$

since $\mathbf{E}_g = \mathbf{E}_1$, and the power supplied to the load is

$$P = I^2R_l = \frac{\mu^2E_1^2}{4r_p}\frac{4}{(1 + R_l/r_p)^2}\frac{R_l}{r_p} \tag{10-2}$$

A sketch showing the variation of the output power as a function of R_l/r_p is given in Fig. 10-2. It can be seen that the power curve reaches a maximum at the point at which $R_l/r_p = 1$, although this maximum is quite broad. P is a slowly varying function of R_l in the neighborhood of the maximum, and the power is at least 88 per cent of its maximum value for values of R_l/r_p ranging from 0.5 to 2.0. This condition shows that the power loss is less than 2.25 db for all values of R between $0.5r_p$ and $2r_p$.

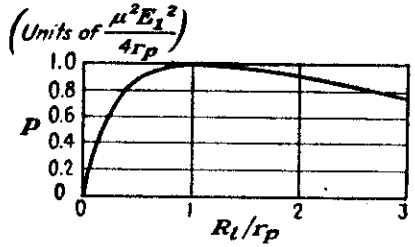


FIG. 10-2. Variation of power output as a function of resistance ratio R_l/r_p.

Since the maximum power transfer occurs when the load resistance equals the internal plate resistance of the tube, it is necessary to use tubes with low values of r_p in order to obtain reasonable amounts of power with nominal values of plate supply potential. Since the g_m of a tube cannot be designed over very wide limits, then tubes with low r_p also possess low values of μ. As a result, large grid excitation potentials are required for appreciable amounts of power output. Note from Eq. (10-2) for a given value of E_g that tubes which possess large values of

$$\frac{\mu^2}{r_p} = g_m\mu$$

possess high output-power output. In fact, the power sensitivity which was defined by Eq. (3-4) becomes, under the conditions of maximum power transfer, simply $\mu g_m/4$.

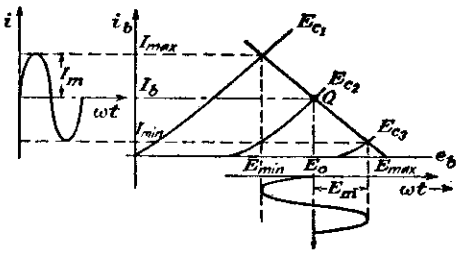


FIG. 10-3. The output current and potential waveforms in a triode power amplifier.

To determine the power output directly from the static plate characteristic of the tube, it is necessary only to draw the appropriate load line on these characteristics and read the significant information from the diagram. Thus, by referring to Fig. 10-3, it follows that

$$I = \frac{I_m}{\sqrt{2}} = \frac{I_{max} - I_{min}}{2\sqrt{2}}$$

and

$$E = \frac{E_m}{\sqrt{2}} = \frac{E_{max} - E_{min}}{2\sqrt{2}} \tag{10-3}$$

The power output is

$$P = \frac{E_mI_m}{2} = \frac{I_m^2R_l}{2}$$

which may be written in the form

$$P = \frac{(E_{max} - E_{min})(I_{max} - I_{min})}{8} \tag{10-4}$$

If distortion is not negligible, the harmonic components must be evaluated according to Eqs. (2-32). The total power output is then

$$P = (B_1^2 + B_2^2 + B_3^2 + \cdots)\frac{R_l}{2}$$

This may be written as

$$P = \left[1 + \left(\frac{B_2}{B_1}\right)^2 + \left(\frac{B_3}{B_1}\right)^2 + \cdots\right]\frac{B_1^2}{2}R_l$$

which becomes, by Eq. (2-33),

$$P = (1 + D_2^2 + D_3^2 + \cdots)P_1$$

or

$$P = (1 + D^2)P_1 \tag{10-5}$$

where D is the total distortion. Notice, however, that if the total distortion is high, say 10 per cent, then $P = 1.01P_1$. That is, a 10 per cent distortion represents a power of only 1 per cent of the fundamental. Thus, with little error, the output power is approximately that of the fundamental-frequency component only.

10-2. Output Circuits. It is not always feasible, nor it is generally desirable, to connect the load directly in the plate lead, as shown in Fig. 10-1. Among the reasons for this are: that the quiescent current through the load resistor represents a considerable waste of power, as it does not contribute to the a-c component of power; and that the quiescent current may cause a serious polarization of the output. For example, it is inadvisable to pass a large d-c current through the voice coil of a loud-speaker. For these reasons, the transformer-coupled load is used extensively, although the parallel-feed system may be used.

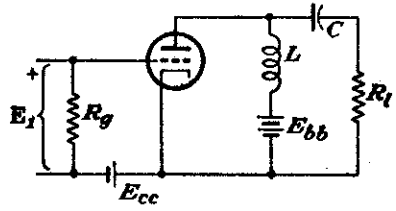


FIG. 10-4. A parallel or shunt-fed power amplifier.

The circuit of the parallel-feed system is illustrated in Fig. 10-4. It is often called the impedance-capacitance coupled system. In this system the plate supply is connected to the plate of the tube through a high

inductance L, the load resistor being connected across the output through a blocking capacitor C. The inductor L must be so chosen that $\omega L \gg R_l$, and C must be so chosen that $1/\omega C \ll R_l$ over the operating range of frequencies.

The potential and current relations for the parallel-feed system are illustrated in Fig. 10-5. Observe that the quiescent current through the tube is determined by R_L, the resistance of the inductor, although the "dynamic" resistance into which the tube is working is R_l. Since the static resistance of the choke or the transformer winding is usually small, the static load line is almost vertical.

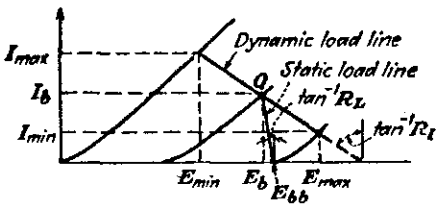


FIG. 10-5. The dynamic and static load lines of a shunt-fed or transformer-coupled amplifier.

Suppose that the load resistance into which the tube works is small; for example, the resistance of the voice coil of a dynamic loud-speaker usually ranges from about 5 to 15 ohms. If such a low resistance load were used in either the series- or the shunt-feed circuits, only a very small power output would be possible, most of the power being lost within the tube resistance. In this case, and in fact in any case in which the load resistance does not properly match the tube resistance, the use of a transformer as an impedance-matching device will permit optimum power transfer. Such a system is illustrated in Fig. 10-6.

The impedance-transforming property of an ideal transformer follows from the simple transformer relations

$$\mathbf{E}_2' = \frac{N_1}{N_2}\mathbf{E}_2 \qquad \text{and} \qquad \mathbf{I}_2' = \frac{N_2}{N_1}\mathbf{I}_2 \tag{10-6}$$

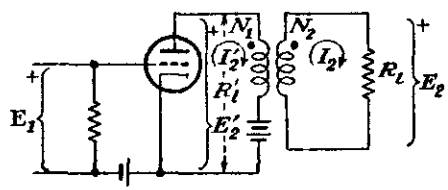


FIG. 10-6. A simple transformer-coupled load in a power amplifier.

where $\mathbf{E}_2'$ and $\mathbf{E}_2$ are the primary and secondary potentials, respectively, and $\mathbf{I}_2'$ and $\mathbf{I}_2$ are the primary and secondary currents. The ratio of the above yields

$$\frac{\mathbf{E}_2'}{\mathbf{I}_2'} = \left(\frac{N_1}{N_2}\right)^2 \frac{\mathbf{E}_2}{\mathbf{I}_2}$$

which may be written as

$$R_l' = \left(\frac{N_1}{N_2}\right)^2 R_l = \frac{1}{n^2} R_l \tag{10-7}$$

since $\mathbf{E}_2'/\mathbf{I}_2'$ and $\mathbf{E}_2/\mathbf{I}_2$ represent the effective input and output impedances, which are resistances. When the turns ratio N_1/N_2 is greater than unity, the transformer is called a *step-down* transformer; with the ratio N_1/N_2 less than unity, it is a *step-up* transformer.

Equation (10-7) is true for an ideal transformer. In general, however, the coupling is not perfect, the primary and secondary resistances are not negligible, and the core losses cannot be neglected. By taking these factors into account, the input impedance is given by

$$\mathbf{Z}_l' = R_1 + \frac{R_2}{n^2} + j\omega\left(L_1' + \frac{L_2'}{n^2}\right) + \frac{\mathbf{Z}_l}{n^2} \tag{10-8}$$

where R_1 and R_2 are the primary and secondary winding resistances, L_1' and L_2' are the primary and secondary leakage inductances, and $\mathbf{Z}_l$ is the load impedance.

The same distinction between the static and dynamic load lines must be made for transformer-coupled loads as for the shunt-feed circuit, and, as noted, Fig. 10-5 applies for both circuits. The frequency response of the transformer is not flat for all frequencies. However, the effects are less severe than for a transformer interstage coupling, since over the audio range the transformer capacitances, tube capacitances, and stray capacitances appear across a relatively low plate-resistance tube or across the low-output load resistance.

10-3. Maximum Undistorted Power. The foregoing analyses, which are based on the linear equivalent plate circuit, are not completely valid owing to the curvature of the dynamic characteristic, particularly in the region of small plate currents. In order to obtain the maximum possible power output without making the instantaneous plate current too small during the most negative part of the applied signal, and without driving the grid positive at the positive peak of the applied signal, it is necessary to maintain a careful balance among the grid bias, load impedance, plate supply potential, and plate resistance.

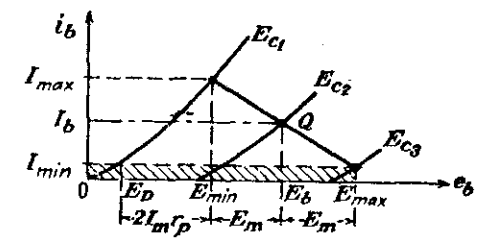


FIG. 10-7. Graphical construction for determining the operation conditions and maximum undistorted power output from an amplifier.

To find the expression for the output power under these conditions, and also to determine the appropriate conditions in order to achieve the present results, refer to the graphical construction of Fig. 10-7. Since the distortion that results at small plate currents arises from the curvature of the static characteristics, this region is eliminated by setting I_{min} at an appropriate value. This specifies the crosshatched area of the diagram. Thus the grid is allowed to swing from any point Q, corresponding to the potential E_b (which would be E_{bb} if an ideal shunt-fed or transformer-

coupled load were used) between zero grid bias and that bias which corresponds to I_{min}. It will be assumed that the distortion is negligible in this region. The characteristics are essentially linear in the operating region, with a slope corresponding to r_p.

To find the value of load resistance for which the power will be a maximum, refer to Fig. 10-7. It is noted that

$$E_b = E_D + 2I_m r_p + E_m$$

But since

$$E_m = R_l I_m$$

then it follows that

$$\begin{aligned} E_b &= E_D + 2I_m r_p + I_m R_l \\ &= E_D + I_m(R_l + 2r_p) \end{aligned}$$

Solving this expression for I_m, there results

$$I_m = \frac{E_b - E_D}{R_l + 2r_p} \tag{10-9}$$

The power to the load is then given by the expression

$$P = \frac{I_m^2 R_l}{2} = \frac{(E_b - E_D)^2}{16 r_p} \frac{4R_l/2r_p}{(1 + R_l/2r_p)^2} \tag{10-10}$$

A sketch showing the variation of the output power as a function of R_l/r_p is given in Fig. 10-8. This curve, like that of Fig. 10-2, reaches a maximum, but in this case at the point at which $R_l = 2r_p$, although the variation is not rapid in the region of the maximum. The power remains at least 88 per cent of its maximum value for load resistances R_l ranging from r_p to $4r_p$.

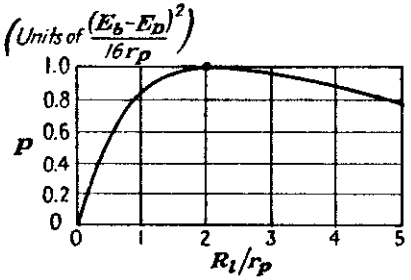


FIG. 10-8. Variation of output power as a function of load R_l.

To find the appropriate bias for these conditions, combine Eq. (10-9) with the fact that the current changes from I_b to I_m when the signal potential E_g is equal to E_c. Thus

$$I_m = \frac{\mu E_c}{R_l + r_p} \tag{10-11}$$

The result, by equating Eq. (10-9) to (10-11), yields

$$E_c = \frac{3}{4\mu}(E_b - E_D) \tag{10-12}$$

in which R_l has been set equal to $2r_p$. The value of E_D is obtained directly from the curves.

The maximum undistorted power output becomes, from Eqs. (10-10) and (10-12),

$$P_{max} = \frac{\mu^2 E_c^2}{9r_p}$$

Further, since $E_c = E_{gm}$, this becomes

$$P_{max} = \frac{\mu^2 E_{gm}^2}{9r_p} = \frac{2}{9}\mu g_m E_g^2 \tag{10-13}$$

and the power sensitivity at optimum power output is

$$\text{Power sensitivity} = \frac{P}{E_g^2} = \frac{2}{9}\mu g_m \quad \text{mhos} \tag{10-14}$$

which is slightly less than that for the conditions of Sec. 10-1.

The results showing the power output and second-harmonic distortion as a function of R_l of a type 6A3 triode are illustrated in Fig. 10-9. Optimum output is obtained at about $R_l = 2{,}500$ ohms, which is approximately three times the plate resistance r_p of the tube. Although the second-harmonic distortion is not negligible at this point, a 5 per cent distortion is usually tolerable.

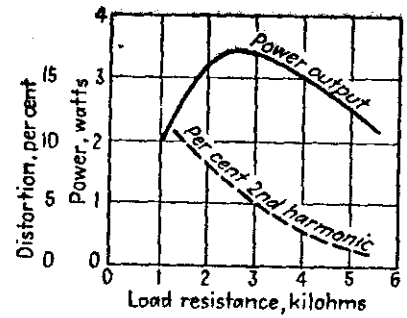


FIG. 10-9. Power output and second-harmonic distortion of a type 6A3 triode as a function of load resistance.

The above analysis is based on the use of a plate-supply source of so-called nominal value (about 300 volts). If it is assumed that a plate source of any potential is available, then with increases in the value of E_{bb} the ultimate limitation will be imposed by the allowable plate dissipation. Under these circumstances, it is found that the circuit should be operated with a value of R_l that greatly exceeds $2r_p$.[1]

10-4. Plate-circuit Efficiency. The foregoing discussion gives the methods for calculating the output power of a power amplifier. The a-c power so obtained is converted from the d-c plate supply by the vacuum tube. The ratio of these quantities is called the *plate-circuit efficiency* of the amplifier. Thus

$$\eta_p = \frac{\text{a-c power output to the load}}{\text{d-c power input to the plate circuit}} \times 100\% \tag{10-15}$$

Suppose that P_p denotes the average power dissipated by the plate. Then by the principle of the conservation of energy

$$E_{bb} I_b = I_b^2 R_1 + E_p I_p + P_p$$

where R_1 is the static load resistance. By solving for P_p, there obtains

$$P_p = E_{bb}I_b - I_b^2R_1 = E_pI_p$$

But as

$$E_{bb} = E_b + I_bR_1$$

then P_p has the form

$$P_p = E_bI_b - E_pI_p \qquad (10\text{-}16)$$

This equation expresses the amount of power that must be dissipated by the plate and represents the kinetic energy of the electrons that is converted into heat at the plate. Notice in particular that the heating of the anode is reduced by the amount of the a-c power that the tube supplies to the load. Hence, a tube is cooler when delivering power to a load

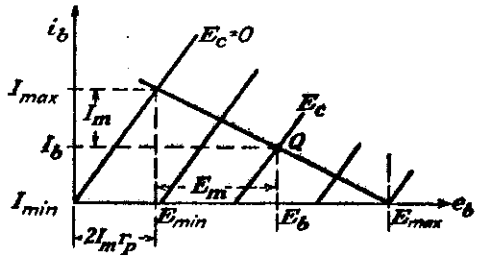


FIG. 10-10. The plate characteristics of an ideal triode.

than when there is no such a-c power transfer. This is a very important factor in the operation of high-power r-f transmitting tubes, since such tubes are ordinarily operated close to the rated allowed plate dissipation. If for any reason the output circuit becomes slightly detuned, with a consequent decrease in output power, the plate power may become dangerously high.

The plate-circuit efficiency may be written in several different forms. From Eq. (10-15),

$$\eta_p = \frac{E_pI_p}{E_{bb}I_b} \times 100\% \qquad (10\text{-}17)$$

This may also be written as

$$\eta_p = \frac{P_0}{P_0 + P_p + I_b^2R_1} \times 100\% \qquad (10\text{-}18)$$

Clearly, a large value of η_p means a small value of P_p for a given output. This means that a smaller tube with a smaller plate-supply source may be used.

It is possible to obtain an approximate expression for the theoretical value of η_p for the series-fed and the shunt-fed circuits. Consider that an ideal tube is used in an amplifier circuit. The plate characteristics of such a tube would have the form illustrated in Fig. 10-10. Suppose that the grid does not swing beyond $E_c = 0$ and may swing to give zero current. Then, by the proper choice of E_c,

$$I_b = I_m$$

and

$$\eta_p = \frac{E_pI_p}{E_{bb}I_b} = \frac{E_mI_m}{2E_{bb}I_m} = 50\frac{E_m}{E_{bb}} \quad \% \qquad (10\text{-}19)$$

For the series-fed circuit, the point marked $E_{max} = E_{bb}$, and from the diagram,

$$E_{max} = E_{bb} = 2E_m + 2I_mr_p$$

Hence Eq. (10-19) becomes

$$\eta_p = 50\frac{E_m}{2(E_m + I_mr_p)} = \frac{25}{1 + I_mr_p/E_m} \quad \%$$

from which, since

$$E_m = I_mR_l$$

then

$$\eta_p = \frac{25}{1 + r_p/R_l} \quad \% \qquad (10\text{-}20)$$

The theoretical maximum plate-circuit efficiency for the series-fed amplifier is 25 per cent. For the conditions of maximum power output, when $R_l = r_p$, $\eta_p = 12.5$ per cent. Actually, owing to the limited range of operation without distortion, η_p seldom exceeds 10 per cent in practice. Evidently, the linear vacuum-tube amplifier is an inefficient device for converting d-c into a-c power.

In the shunt-fed system a means has been devised for eliminating the d-c power loss in the load. This results in an improved plate efficiency. If the static resistance is assumed negligible, then

$$E_{bb} = E_b = E_m + 2I_mr_p$$

and

$$\eta_p = 50\frac{E_m}{E_{bb}} = 50\frac{E_m}{E_m + 2I_mr_p} \quad \%$$

This reduces to the form

$$\eta_p = \frac{50}{1 + 2r_p/R_l} \quad \% \qquad (10\text{-}21)$$

The theoretical maximum plate-circuit efficiency of the shunt-fed or transformer-fed amplifier is 50 per cent. Thus the elimination of the static power loss in the load reflects itself as an improved plate efficiency. For the conditions of maximum output power, when $R_l = 2r_p$, $\eta_p = 25$ per cent. However, since the static resistance is not negligible and since the current I_{min} cannot be taken as zero if distortion is to be avoided, the actual plate-circuit efficiency will be less than the 25 per cent figure.

10-5. Power Pentodes and Beam Power Tubes. The plate characteristics of the power pentode are markedly different from those of a triode, and the graphical analyses given above are not valid for pentodes. Since the characteristics of the beam power tubes are similar to those of the power pentode, the discussion to follow applies for both the power pentode and the beam power tube.

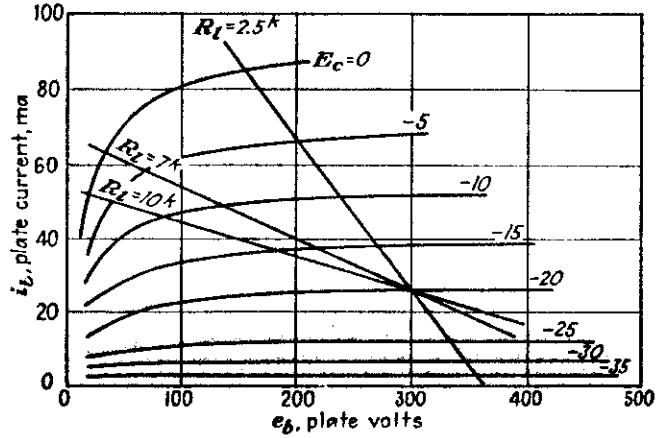


FIG. 10-11. Plate characteristics of a 6F6 pentode.

Power pentodes differ from triodes principally in the character of the dynamic curve with increasing load resistances. In the triode, the distortion decreases as the magnitude of the load resistance increases. This follows from the fact that the dynamic curve becomes increasingly linear as the load resistance becomes higher. In the power pentode, the dynamic characteristic is critically dependent on the load resistance, with excessive curvature at both the high and the low values of load resistance. Moreover, the critical load resistance to be used cannot be related analytically with the plate resistance of the tube. This resistance is always less than the plate resistance of the tube. It might appear therefore that the output-power capacity of the tube would be too low to make the tube feasible. Actually, owing to the large μg_m product, even with the appropriate R_l the power output is usually higher than with the tube connected as a triode, and this with a smaller grid driving-potential amplitude.

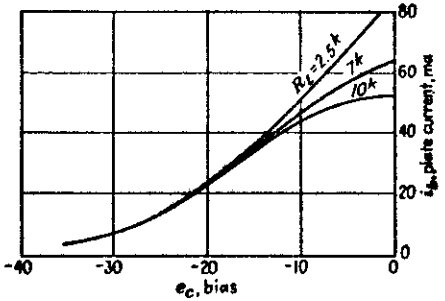


FIG. 10-12. Dynamic characteristics of a 6F6 for three values of plate load resistance.

To examine the matter in somewhat greater detail, refer to Fig. 10-11, which gives the plate characteristics of a 6F6 power pentode. It will be supposed that the load is transformer-coupled to the tube and that the plate potential is maintained at 300 volts. Three load lines are shown, 2,500, 7,000, and 10,000 ohms. The corresponding dynamic curves are given in Fig. 10-12. The following example will help clarify the situation:

Example. Calculate the output power, the plate-circuit efficiency, and the second, third, and fourth harmonics for the 6F6 that supplies power to a loudspeaker, the effective resistance of which is changed to have values of 2.5 kilohms, 7 kilohms, and 10 kilohms. A 300-volt source is available, and the tube is biased at −20 volts.

Solution. The important data from Figs. 10-11 and 10-12 are included in the tabulation. The general character of the results is given graphically in Fig. 10-13. Notice that optimum power transfer occurs somewhat above 10 kilohms, which is very small compared with the tube resistance of 78,000 ohms.

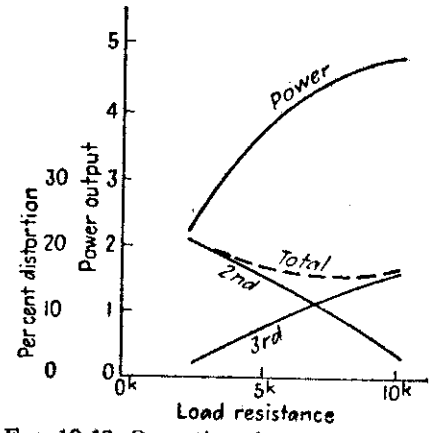


FIG. 10-13. Operating characteristics of a 6F6 pentode as a function of load resistance.

R_l / I	2.5 kilohms	7 kilohms	10 kilohms
I_{max}	86	64	53
$I_{½}$	52	49	46
I_b	25	25	25
$I_{-½}$	6	6	6
I_{min}	1	1	1

R_l / B	2.5 kilohms	7 kilohms	10 kilohms
B_0	9	4	1
B_1	44	35	31
B_2	9	4	1
B_3	−1	−4	−5
B_4	+½	−½	

R_l / D	2.5 kilohms	7 kilohms	10 kilohms
D_2	20.5	11.5	3.2
D_3	−2.3	−11.5	−16
D_4	+1	−1.4	

R_l	2.5 kilohms	7 kilohms	10 kilohms
I_{a-c}	45.0	35.5	31
P_{a-c}	2.5	4.4	4.8
I_{d-c}	31	29	26
P_{d-c}	10.2	8.7	7.8
η_p	24.5	50.6	61.6

10-6. Push-Pull Amplifiers. The use of two tubes in parallel provides the possibility of obtaining twice the output power of a single tube with the same distortion. A push-pull amplifier circuit is a much more desirable connection for two tubes. In this circuit, the two tubes are

arranged as shown in Fig. 10-14. The excitation potentials to the grids of the two tubes must be of equal magnitude, but of opposite phase. In the circuit shown, a center-tapped transformer is used to provide these two equal potentials that differ by 180 deg. A number of vacuum-tube circuits are possible for achieving these results, and several of these will be examined below.

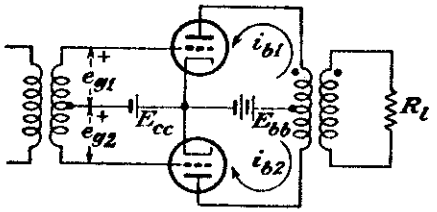


FIG. 10-14. The basic push-pull circuit.

To examine certain of the features of such an amplifier, suppose that the input grid potential to one tube is of the form

$$e_{g1} = E_{gm} \cos \omega t$$

The plate current of this tube will be represented in general by the expression (see Sec. 2-8)

$$i_{b1} = B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t + B_3 \cos 3\omega t + \cdots \quad (10\text{-}22)$$

The corresponding signal to the second tube is

$$e_{g2} = -E_{gm} \cos \omega t = E_{gm} \cos (\omega t + \pi)$$

and the output plate current is

$$i_{b2} = B_0 + B_1 \cos (\omega t + \pi) + B_2 \cos 2(\omega t + \pi) + B_3 \cos 3(\omega t + \pi) + \cdots$$

which has the form

$$i_{b2} = B_0 - B_1 \cos \omega t + B_2 \cos 2\omega t - B_3 \cos 3\omega t + \cdots \quad (10\text{-}23)$$

But, from Fig. 10-14, the currents are in opposite directions through the output-transformer windings. The total output is then proportional to the difference between the plate currents in the two tubes. This is

$$i = k(i_{b1} - i_{b2}) = 2k(B_1 \cos \omega t + B_3 \cos 3\omega t + \cdots) \quad (10\text{-}24)$$

This expression shows that the push-pull circuit balances out all even harmonics in the output and leaves the third-harmonic term as the principal source of distortion.

Another feature of importance in the push-pull system is evident from the circuit of Fig. 10-14. It is observed that the steady components of the plate currents flow in opposite directions in the windings, thus opposing each other magnetically in the transformer core. This eliminates any tendency toward core saturation and the resulting distortion that might arise from the magnetization of the transformer core.

The effects of ripple potentials that may be contained in the power supply due to inadequate filtering will be balanced out in the push-pull circuit. This is so because the currents that are produced by this ripple potential are in opposition in the output transformer and hence will not appear in the load. Of course, the effects of the ripple potential that appear on the grids of the tubes will not be balanced out and will be noticeable with the signal.

Another feature of this amplifier is that under self-biased conditions there is no need for a by-pass capacitor across the cathode resistor. This follows from the fact that the potential which appears across the self-bias resistor R_k is $(i_{b1} + i_{b2})R_k$. But this is, from Eqs. (10-22) and (10-23),

$$(i_{b1} + i_{b2})R_k = 2R_k(I_b + B_0 + B_2 \cos 2\omega t + B_4 \cos 4\omega t + \cdots)$$

But for tubes operating in class A the harmonic amplitudes are very small and are therefore not significant.

One of the particularly significant features of the push-pull amplifier is that the output power possible with the two tubes for a given total distortion is higher than twice that of the single tube. This results from the fact that with the automatic cancellation of even harmonics in the output the tubes may be driven harder until the third harmonic terms become significant.

Note also that, with the increased grid drive, the rectification component $2B_0R_k$ becomes significant and adds to the bias $2I_bR_k$, if self-bias is used. As a result, the effective bias increases, with consequent reduction of output power. This means that the output power of a push-pull amplifier under otherwise similar conditions will be higher with fixed bias than with self-bias.

10-7. Equivalent Circuit of a Class A Push-Pull Amplifier. Suppose that both tubes of the push-pull amplifier are identical and that μ and r_p are constant over the range of operation. The equivalent circuit of the system then has the form given in Fig. 10-15. Observe that the connection between the cathode terminals and the mid-point of the output transformer does not carry a fundamental-frequency component of current, owing to the cancellation that occurs. This connection may be omitted from the diagram without influencing the operation. The resulting cir-

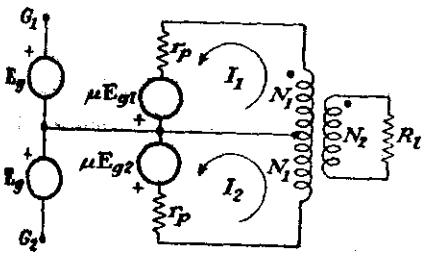


FIG. 10-15. Equivalent circuit of the class A push-pull amplifier of Fig. 10-14.

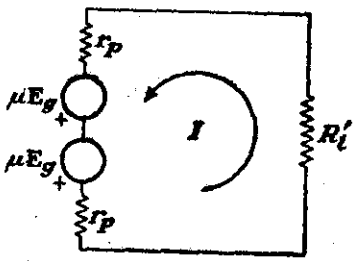


FIG. 10-16. The simplified equivalent circuit of the class A push-pull amplifier.

cuit then has the form given in Fig. 10-16. In this diagram

$$R_l' = \left(\frac{2N_1}{N_2}\right)^2 R_l \tag{10-25}$$

The resultant current is given by

$$\mathrm{I} = \frac{2\mu \mathrm{E}_g}{2r_p + R_l'} \tag{10-26}$$

which may be written in the form

$$\mathrm{I} = \frac{\mu \mathrm{E}_g}{r_p + R_l'/2} \tag{10-27}$$

The total power delivered to the load is then

$$P = I^2 R_l' = 2\left(\frac{\mu E_g}{r_p + R_l'/2}\right)^2 \frac{R_l'}{2} \tag{10-28}$$

This expression may be interpreted to mean that the total output power is twice the power of each tube considered to be working into the equivalent load resistance $R_l'/2$.

A more significant expression results by writing Eq. (10-28) in the form

$$P = \left(\frac{\mu E_g}{r_p/2 + R_l'/4}\right)^2 \frac{R_l'}{4} \tag{10-29}$$

This may be interpreted to show that the class A push-pull amplifier may be represented by a single composite generator which has emf $\mu \mathrm{E}_g$, with an internal resistance $r_p/2$, and which works into a load resistance equal to $R_l'/4$. It is possible, in fact, to derive a set of static characteristics of the composite tube from the tube plate characteristics and to obtain significant operating information from this.

10-8. Composite Static-characteristic Curves. The composite static characteristics of the push-pull amplifier may be obtained from the plate

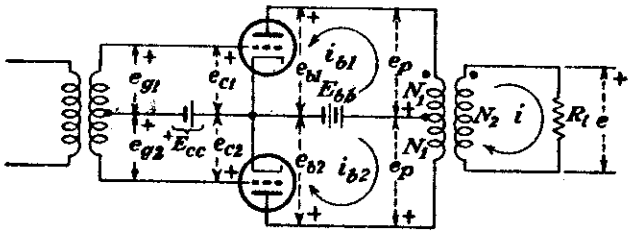


FIG. 10-17. The potentials in the push-pull amplifier.

characteristics of the individual tube by a graphical construction.[2] It is assumed that the output transformer is ideal, whence the potentials across each half of the transformer are equal. The situation is illustrated in Fig. 10-17. The load potential is given by

$$e = \frac{N_2}{N_1} e_p \tag{10-30}$$

Also, in the ideal transformer, the total primary ampere-turns are equal to the secondary ampere-turns, from which

$$N_1 i_{b1} - N_1 i_{b2} = N_2 i \tag{10-31}$$

Thus the load potential is

$$e = iR_l = (i_{b1} - i_{b2}) \frac{N_1}{N_2} R_l \tag{10-32}$$

By combining Eqs. (10-30) and (10-32) there results

$$e_p = (i_{b1} - i_{b2}) \left(\frac{N_1}{N_2}\right)^2 R_l \tag{10-33}$$

which may be written in the form

$$e_p = (i_{b1} - i_{b2}) \frac{R_l'}{4} \tag{10-34}$$

where R_l' is the plate-plate resistance.

The following relationships are evident from an inspection of the diagram of Fig. 10-17:

$$\begin{aligned} e_{b1} &= E_b - e_p \\ e_{b2} &= E_b + e_p \\ e_{c1} &= E_{cc} + e_g \\ e_{c2} &= E_{cc} - e_g \end{aligned} \tag{10-35}$$

It follows from these equations that when the plate-cathode potential e_{b1} of tube $T1$ decreases from the quiescent-point value E_b by an amount e_p, then the corresponding potential e_{b2} of tube $T2$ increases beyond E_b by a like amount. Also, when the grid-cathode potential e_{c1} increases beyond E_{cc} by the signal potential e_g, the corresponding value of e_{c2} decreases below E_{cc} by e_g. These conditions are shown in Fig. 10-18 for three different values of e_p. In these diagrams point A corresponds to conditions $e_{b1} = E_b - e_p$, $e_{c1} = E_{cc} + e_g$, the tube current being i_{b1}.

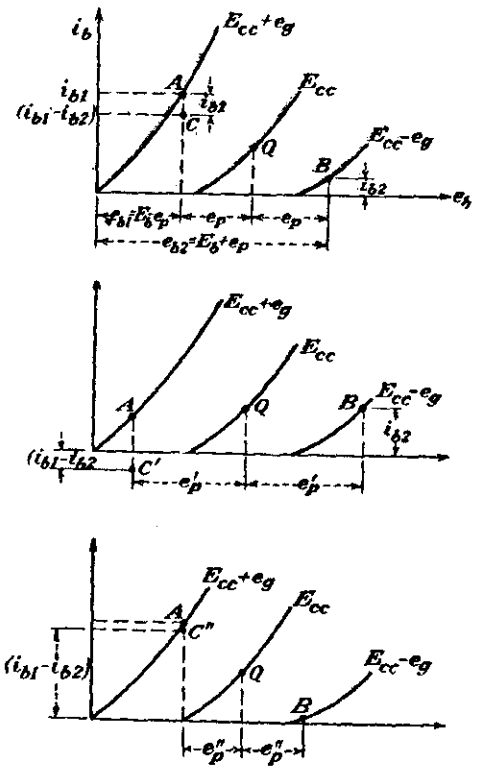


FIG. 10-18. To construct the composite static characteristics of a push-pull amplifier from the plate characteristic.

Point B is that for $e_{b2} = E_b + e_p$, $e_{c2} = E_{cc} - e_g$, and the tube current is i_{b2}. Point C has the ordinate $i_{b1} - i_{b2}$ for the chosen value of e_p.

The composite static characteristics are the family curves of i_{b1}-i_{b2} vs. e_b, with grid signal potential as a parameter. Clearly, point C is one point on the composite static curve for the signal potential e_g. Other points

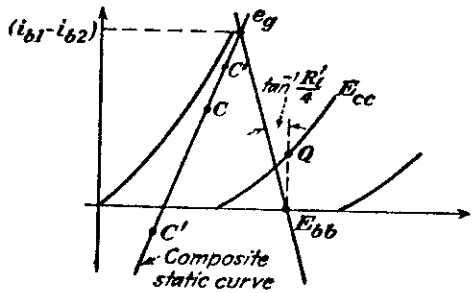


FIG. 10-19. The composite static curve derived from the constructions of Fig. 10-18.

on this characteristic are found by maintaining e_g constant and by varying e_p. The construction for two other values e_p' and e_p'' are given in Fig. 10-18*b* and 10-18*c*. These locate two other points C' and C'' on the composite static characteristic. The complete composite static is given in Fig. 10-19.

Several significant features are evident from Fig. 10-19. The composite static characteristic extends above and below the zero-current axis. Also, the composite static curve is much more linear than the plate characteristics of the individual tube.

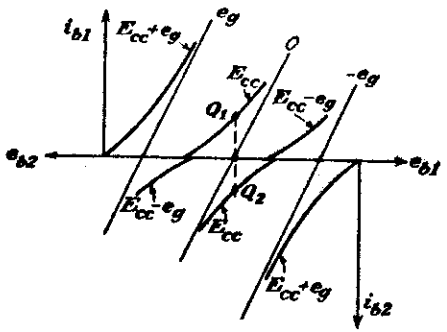


FIG. 10-20. The Thompson method of obtaining the composite static characteristics of a push-pull amplifier.

An alternative method for obtaining the composite static characteristics was described by Thompson.[3] According to this method, the plate characteristics of the tube are plotted in the usual way. The curves are also plotted in an inverted manner, with the potential scale shifted so that the potentials E_b of both sets of curves are aligned with each other. This construction is shown in Fig. 10-20. The inverted curves represent the plate family of tube $T2$. The two methods are essentially equivalent.

The foregoing discussion of the graphical construction is general and applies to any type of tube under class A, AB, or B conditions of operation. Although the illustrations are for triodes, the curves for other tube types are obtained in the same way. In fact, owing to their shape, the composite static characteristics for pentode-type tubes are more easily obtained than the triode curves. The simplification results because the current i_{b2} remains substantially constant for large variations in e_b. The composite static characteristics for the 6F6 pentode are shown in Fig. 10-21.

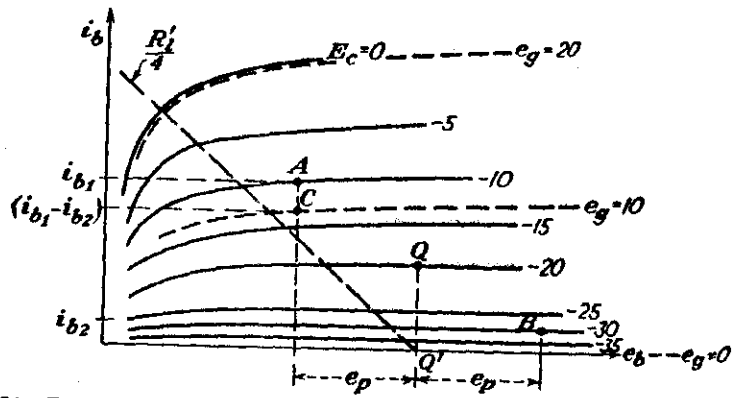


FIG. 10-21. Composite static characteristics of a 6F6 pentode push-pull amplifier.

10-9. Composite Dynamic Characteristic. The composite dynamic characteristic of a push-pull amplifier is obtained from the composite static characteristics in precisely the same way that the dynamic curve is obtained from the plate characteristics of a single-tube amplifier. This requires drawing the effective composite load line on the composite static curves and noting the points of intersection of the two, which are points on the composite dynamic curve. This construction also permits obtaining the load line into which each tube is working and hence also the dynamic characteristic of the individual tubes. The construction is shown in Fig. 10-22. Clearly, the intersection of the push-pull load line with the composite plate characteristics gives points on the composite dynamic characteristic. That is, points A, B, C, D are points on the composite dynamic curve. These points are replotted in Fig. 10-23 to give the composite dynamic curve.

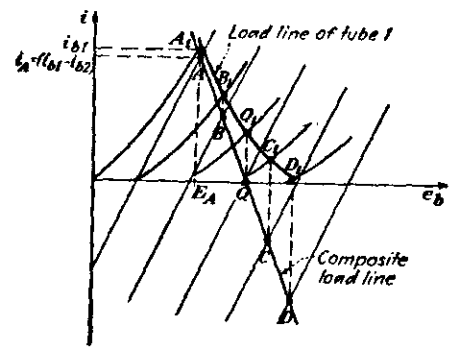


FIG. 10-22. The load line of the composite circuit, and the load line of one tube of the push-pull amplifier.

To find the dynamic characteristic of each tube, the procedure is essentially the reverse of that of Sec. 10-8. In particular, consider the point B in Fig. 10-22. This point is a representation of i_{b1}-i_{b4} corre-

sponding to the appropriately chosen value of e_p. But the point i_{b1} lies vertically above this point, by an amount i_{b2}, and must lie on the plate characteristic of the tube. This defines the point A_1. The other points are obtained in a similar way, and the results are shown in Fig. 10-22. These points are plotted also in Fig. 10-23. The points A_2, B_2, C_2, D_2 for tube $T2$ are obtained by symmetry from the corresponding points of tube $T1$.

It is interesting to note that the composite dynamic curve is practically a straight line, although the individual tube dynamic curves are markedly curved. Thus, for a sinusoidal input, the total output current is closely sinusoidal, although the current in each tube is not sinusoidal, in general.

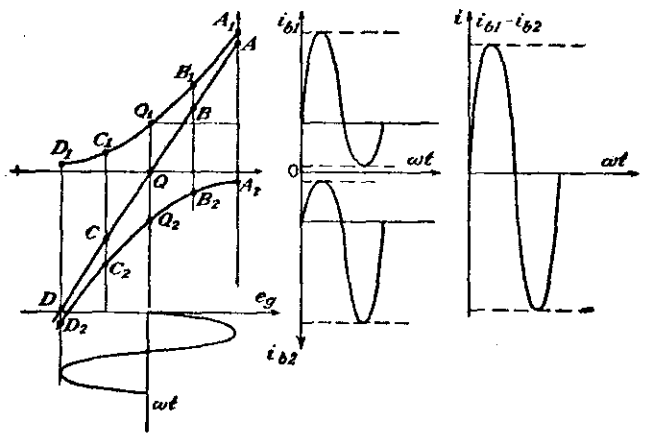


FIG. 10-23. The single-tube dynamic characteristics and the composite dynamic of the push-pull amplifier. The waveshapes in each tube and in the output are also shown.

If only the output current is desired, there is no requirement for the individual tube dynamics. If the plate-circuit efficiency is required, the individual tube dynamics are required in order to calculate the value of I_b and the d-c component B_0 that results from the partial rectification due to the curvature of the dynamic. The total d-c power input to the amplifier is $2E_{bb}(I_b + B_0)$.

10-10. Power Output and Distortion in Push-Pull Amplifiers. Owing to the fact that the dynamic curve must be an odd function, by virtue of the manner of its construction, then for the composite circuit

$$\begin{aligned} I_b &= 0 \\ I_{max} &= -I_{min} \\ I_{\frac{1}{2}} &= -I_{-\frac{1}{2}} \end{aligned} \tag{10-36}$$

Under these circumstances the five-point schedule of Eq. (2-30) reduces to

$$\begin{aligned} B_0 &= B_2 = B_4 = 0 \\ B_1 &= \tfrac{2}{3}(I_{max} + I_{\frac{1}{2}}) \\ B_3 &= \tfrac{1}{3}(I_{max} - 2I_{\frac{1}{2}}) \end{aligned} \tag{10-37}$$

The fundamental power output is given by the expression

$$P_1 = \left(\frac{B_1}{\sqrt{2}}\right)^2 \frac{R_l'}{4} = \frac{B_1^2 R_l'}{8} \tag{10-38}$$

By neglecting the harmonic components of power, the total power is given by

$$P_1 = \frac{E_m I_m}{2} \tag{10-39}$$

where E_m and I_m denote the peak values of the a-c output voltage and current, respectively. These values are obtained directly from the curves of Fig. 10-22, since $I_m = I_A$ and $E_m = E_b - E_A$, whence

$$P = \frac{(E_b - E_A)I_A}{2} \tag{10-40}$$

To find the maximum output power in the push-pull class A system utilizing triodes, use is made of the fact that the load resistance should equal the internal resistance of the equivalent or composite generator. This follows from Eq. (10-29) for the class A amplifier and requires that $R_l'/4 = r_p/2$. This requires that the slope of the effective load line must be equal to the reciprocal of the composite static characteristic, which has a value of $r_p/2$.

Suppose that the tubes are operated in push-pull class B. Now, since the tubes are biased to cutoff, then either one or the other of the two tubes will be supplying current to the circuit and each contributes power for one-half of each cycle. Consequently, the equivalent generator will be one with an internal resistance equal to r_p of the tube. The maximum power under these conditions will be obtained for $R_l'/4 = r_p$. It is reasonable to expect that the internal resistance of the equivalent source of a push-pull class AB amplifier will lie between the value for the class A circuit $r_p/2$ and that for a class B circuit r_p. In all cases, however, recourse should be had to the composite static characteristics, and then $R_l'/4$ should be set equal to the reciprocal of the slope of these lines.

The situation for pentodes is different from that discussed above for triodes and follows roughly the reasoning of Sec. 10-5. The optimum load is that which yields the maximum power with low distortion. The optimum load line is drawn through the point Q' so that it intersects the peak composite grid-voltage curve in the neighborhood of the knee of the curve. This is illustrated in Fig. 10-21 for the 6F6 tube.

10-11. Driver Stages for Push-Pull Amplifiers. The driver may be considered to comprise the circuit that supplies the two potentials of equal magnitude but in phase opposition to the grids of the push-pull power amplifier. A variety of suitable circuits exist, the most direct of which

is illustrated in Fig. 10-24. This circuit consists of a simple amplifier with a transformer in the output, the secondary of which is center-tapped. The characteristics of this stage are determined by the grid driving-power requirements of the power amplifier. If the push-pull circuit requires substantially no driving power, then there are no serious requirements imposed on the driver stage. If the driver stage is called upon to supply power, and this would seldom exceed 15 per cent of the output of the push-pull stage, although it is ordinarily less than this amount, the driver stage must have a relatively low internal resistance if no distortion is to be introduced into the grid circuit of the push-pull amplifier. It is advisable in such cases that a step-down transformer be used to couple the driver stage to the push-pull input in order to reduce the effective resistance in the grid circuit.

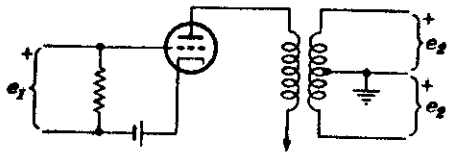


FIG. 10-24. A simple driver circuit for a push-pull amplifier.

If the power requirements are low, then any one of a wide variety of "paraphase" circuits may be used. A paraphase circuit is one which provides two equal output potentials which are 180 deg apart in phase from a single signal source.

Single-tube Paraphase Amplifier. A single-tube amplifier in which the load resistor is divided equally between the plate and cathode circuits is the simplest form of paraphase amplifier. The circuit, redrawn for convenience in Fig. 10-25, is discussed at some length in Sec. 5-4. The resistors R_l and R_k have the same value, whence the amplitude of the potential developed across each is the same, since the same current flows through each. The polarity is opposite because the cathode output is taken from the more positive end of R_k and the plate output is taken from the less positive end of R_l. The analysis of this circuit in Sec. 5-4 shows that the gain of the stage is less than unity and is given by the expression

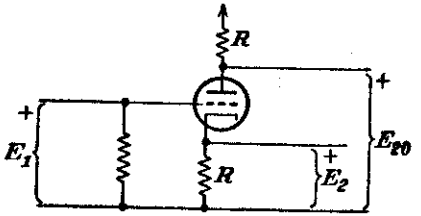


FIG. 10-25. A single-tube paraphase amplifier.

$$\mathbf{K}_f = \frac{\pm \mu R}{(\mu + 2)R + r_p} \doteq \pm \frac{\mu}{\mu + 2} \qquad (10\text{-}41)$$

Two-tube Paraphase Amplifiers. In the two-tube paraphase amplifier, one tube is used as a conventional amplifier, and a second tube is used as a phase-inverter amplifier. Figure 10-26 illustrates such a circuit. The resistors R_1 and R_2 comprise a potential divider across the output of a conventional amplifier, the ratio of the resistances being chosen so that the output of $T1$ is equal to the output of $T2$. Also, the operating conditions of the tubes are carefully chosen to allow the curvature of the characteristic of $T2$ to compensate for the curvature of $T1$. Thus the output potentials relative to ground are both slightly distorted to provide a comparatively undistorted potential difference between the output terminals. This method is difficult to apply in practice because the adjustments necessary to reduce distortion to a minimum are critical.

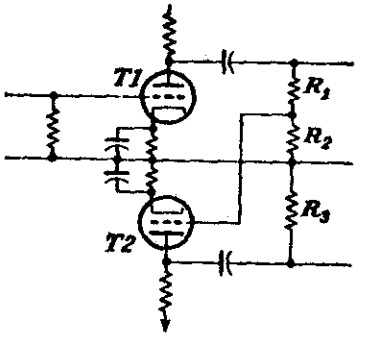


FIG. 10-26. A two-tube paraphase amplifier.

A second form of two-tube paraphase amplifier employs the differential potential between the outputs of two tubes as the input signal to the phase-inverter section. This circuit, which is also referred to as the *floating paraphase* amplifier, is illustrated in two versions in Fig. 10-27.

In the circuit of Fig. 10-27*a* tube $T1$ is an amplifier to increase the amplitude of the applied waveform. The cathode resistors R_k, if not by-passed, will provide some degeneration, which will help to reduce distortion. The output from $T1$ is coupled through C_1 to R_1 and R_3, both of which have the same value as R_2. The potential which appears across R_3 is applied to the grid of $T2$. The output of $T2$ is passed through C_2 and is applied across R_2 and R_4. Thus half the output of both $T1$ and $T2$ appears across R_3. Since these potentials are of opposite polarity, the resultant potential across R_3 is the difference between these two. The output of $T1$ is larger than the output of $T2$, and in order that this difference should be kept as small as possible, pentodes are used, so as to take advantage of their high amplification.

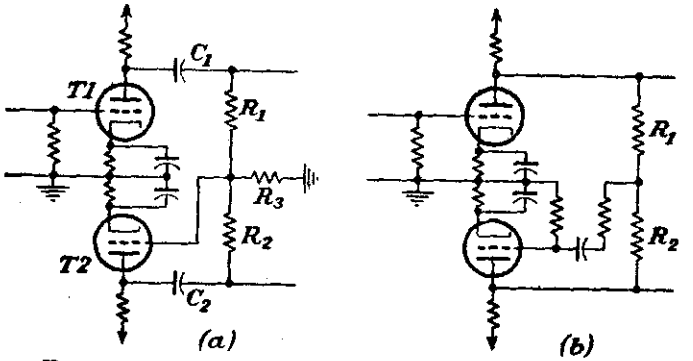


FIG. 10-27. Two forms of floating paraphase amplifiers.

The feature of the circuit of Fig. 10-27*b* is that the difference between the output potentials is taken care of in the choice of the resistors R_1 and R_2 so that the output potentials have the same amplitude. To do this requires that the following condition be satisfied:

$$\frac{K-1}{K+1} \doteq \frac{R_2}{R_1} \tag{10-42}$$

where K is the gain of the stage.

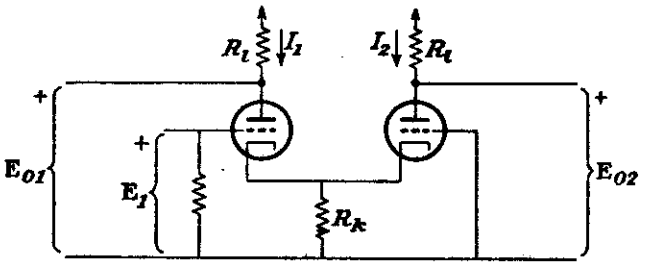


FIG. 10-28. A cathode-coupled paraphase amplifier.

The cathode-coupled paraphase amplifier, which is closely related to the cathode-coupled amplifier which was discussed in Sec. 4-10, is illustrated graphically in Fig. 10-28. This circuit is used extensively to provide push-pull deflection potentials for the plates of a cathode-ray tube, and also as the driver of a push-pull amplifier.

An analysis of this amplifier circuit is readily effected in terms of the Thévenin equivalent circuit as viewed from the cathode-ground terminals

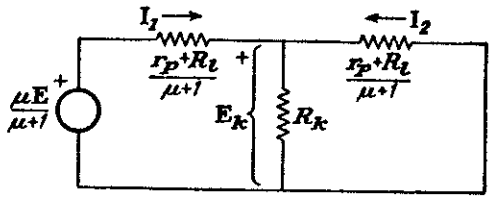


FIG. 10-29. The equivalent circuit of Fig. 10-28 as viewed from the cathode-ground terminals.

The resulting equivalent circuit is given in Fig. 10-29. It follows directly from this figure that the potential drop across R_k is given by

$$\mathbf{E}_k = -\mathbf{I}_2 \frac{r_p + R_l}{\mu + 1} = \frac{\mathbf{I}_1 R_k \dfrac{r_p + R_l}{\mu + 1}}{R_k + \dfrac{r_p + R_l}{\mu + 1}}$$

From this

$$-\frac{\mathbf{I}_1}{\mathbf{I}_2} = 1 + \frac{r_p + R_l}{R_k(\mu + 1)} \tag{10-43}$$

Notice from this expression that the current $\mathbf{I}_1$ is always greater than $\mathbf{I}_2$. For example, for a typical triode with, say, $r_p = 10{,}000$ ohms and $\mu = 19$, and with $R_l = R_k = 10{,}000$ ohms, $\mathbf{I}_1/\mathbf{I}_2 = 1.10$. In general, if the magnitude of $\mathbf{I}_1$ is not to exceed the magnitude of $\mathbf{I}_2$ by more than, say, 10 per cent (this requires that $-\mathbf{I}_1/\mathbf{I}_2 \leqq 1.1$), then

$$R_k \geqq 10 \frac{r_p + R_l}{\mu + 1}$$

It should be observed that the more nearly $\mathbf{I}_1$ and $\mathbf{I}_2$ are to be the same, the larger must R_k be. In fact, the two currents are equal only if $R_k = \infty$, an impractical condition. As a result, this circuit can never give exactly equal outputs, although by making R_k large compared with $(r_p + R_l)/(\mu + 1)$ almost equal outputs are possible.

To determine the output potentials, it is noted from the diagram that

$$\frac{\mu}{\mu + 1}\mathbf{E} + \mathbf{I}_1 \frac{r_p + R_l}{\mu + 1} - \mathbf{I}_2 \frac{r_p + R_l}{\mu + 1} = 0$$

from which

$$\mathbf{I}_1 - \mathbf{I}_2 = \frac{\mu \mathbf{E}}{r_p + R_l}$$

If it is assumed that $\mathbf{I}_1 = -\mathbf{I}_2$ approximately, as discussed above, then

$$\mathbf{I}_1 = -\mathbf{I}_2 = \frac{\mu \mathbf{E}}{2(r_p + R_l)}$$

whence

$$\mathbf{E}_{01} = -\mathbf{E}_{02} = \frac{\mu \mathbf{E}}{2(r_p + R_l)} \tag{10-44}$$