TUNED POWER AMPLIFIERS

12-1. Introduction. In common with the operation of the classes of tube circuits being studied, it is the function of an r-f power amplifier to convert d-c power from the power supply into r-f power. Owing to the amounts of power that may be involved, it is essential that this conversion be effected at the highest possible efficiency. Essentially, therefore, the power amplifier may be regarded as a power converter, as contrasted with the r-f and i-f potential amplifiers that are used to raise a potential level. The settings of the r-f power amplifier are chosen to ensure a high conversion efficiency.

The basic circuit of a tuned power amplifier is substantially that of the single-tuned direct-coupled type discussed in Sec. 11-1. The essential differences are in the magnitude of the grid-bias supply potential E_{cc}, the corresponding value of the grid input signal e_g, and the amount of power involved. A schematic diagram of a tuned power amplifier is given in Fig. 12-1.

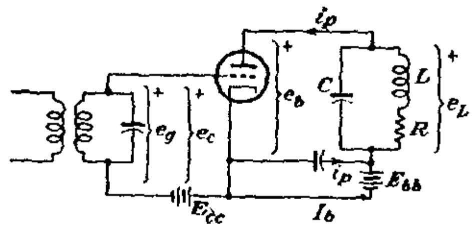

FIG. 12-1. Schematic diagram of a tuned power amplifier.

Owing to the negative bias on the tube, which is adjusted approximately to plate-current cutoff in the class B amplifier and which is adjusted beyond plate-current cutoff in the class C amplifier, harmonic currents are generated in the plate which are comparable in amplitude with the fundamental component. However, if the Q of the tuned plate circuit has a value of 10 or more, the impedance of the tank circuit to the second or higher harmonics will be very low. As a result, the higher-harmonic potentials across the tank will be very small compared with the fundamental potential. That is, the effect of the harmonic generation in the tube plate current is largely suppressed by the tuned plate load.

But the requirement that the Q of the tank circuit must be high in order to suppress harmonics in the output imposes a limitation on the frequency-response characteristics of the amplifier, since then the gain is constant only over a very narrow band of frequencies. Consequently such amplifiers are confined in their operation to narrow frequency bands.

In fact, as will be discussed in some detail, the class B amplifier may be used to amplify a narrow band of frequencies of differing amplitudes, whereas the class C amplifier is confined to a narrow band of frequencies of constant amplitudes. Despite these severe restrictions, both classes of amplifier are extensively used in restricted applications, the class B amplifier to amplify an a-m r-f carrier wave, the class C amplifier as a frequency multiplier or as a source for the production of an a-m carrier wave.

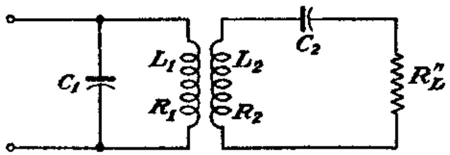

FIG. 12-2. A typical tuned-amplifier tank circuit.

12-2. Properties of the Tank Circuit. The tuned plate load in the diagram of Fig. 12-1 is drawn as a simple parallel resonant circuit. Ordinarily the load is coupled inductively to the plate tank, and a more typical coupling network is that shown in Fig. 12-2. The capacitor C_2 is assumed to be so adjusted that $1/2\pi\sqrt{L_2C_2}$, the resonant frequency of the secondary circuit, is equal to the operating frequency of the amplifier. Because of the resonance in the secondary circuit, only a resistive component $R_L' = (\omega M)^2/(R_L'' + R_2)$ is reflected into the primary of the tuned circuit. The equivalent circuit then becomes that shown in Fig. 12-3.

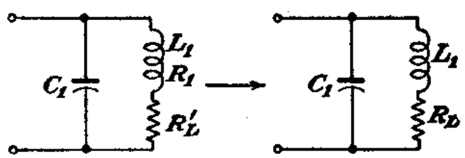

FIG. 12-3. The equivalent circuits of Fig. 12-2.

If the characteristics of the tank circuit were ideal, the impedance at resonance would be resistive and equal to the shunt resistance R_0 of the resulting network. The impedance would be zero at any of the harmonic frequencies. That is, the impedance would be

$$\mathbf{Z}(\omega_0) = R_0 \qquad \mathbf{Z}(n\omega_0) = 0 \qquad n = 2, 3, 4, \ldots \tag{12-1}$$

These ideal conditions do not prevail in practice, although it is possible to achieve relatively low impedance for $Z(n\omega_0)$. To examine this, refer to Eq. (11-6) for the impedance function of the simple tuned circuit,

$$\mathbf{Z} = R_LQ^2 \frac{1 + \delta - j(1/Q)}{1 + \delta + jQ\delta(2 + \delta)} \tag{12-2}$$

At resonance $\omega = \omega_0$, and $\delta = 0$. Equation (12-2) reduces to

$$\mathbf{Z}(\omega_0) = R_LQ^2\left(1 - j\frac{1}{Q}\right) = R_LQ^2\sqrt{1 + \frac{1}{Q^2}}\,\underline{/\tan^{-1}\frac{1}{Q}}$$

Note, however, that, if $Q = 10$, then

$$\mathbf{Z}(\omega_0) = R_LQ^2 \times 1.005\underline{/-5.7^\circ}$$

which shows that the impedance of the tank circuit is essentially resistive and is given by

$$\mathbf{Z}(\omega_0) = R_0 \doteq R_LQ^2 \tag{12-3}$$

Under these conditions it follows that

$$R_0 = R_LQ^2 = \omega_0L_1Q = \frac{L_1}{R_LC_1} = Q\sqrt{\frac{L_1}{C_1}} \tag{12-4}$$

Now consider the situation at the second-harmonic frequency. When $\omega = 2\omega_0$, $\delta = 1$ and Eq. (12-2) reduces to

$$\mathbf{Z}(2\omega_0) = R_LQ^2 \frac{1 - j\dfrac{1}{2Q}}{1 + j1.5Q} = R_LQ^2 \frac{0.25 - j\left(\dfrac{1}{2Q} + 1.5Q\right)}{1 + 2.25Q^2} \tag{12-5}$$

For $Q = 10$ this reduces to

$$\mathbf{Z}(2\omega_0) \doteq R_LQ^2\frac{1}{j1.5Q} = -j\frac{1}{1.5}\sqrt{\frac{L_1}{C_1}} \tag{12-6}$$

The ratio of the second harmonic to the fundamental-frequency impedance is then

$$\frac{Z(2\omega_0)}{Z(\omega_0)} = \frac{R_LQ^2(1/1.5Q)}{R_LQ^2} = \frac{1}{1.5Q}$$

In fact, under the extreme conditions when $I_{p2} = I_{p1}$, the relative power ratio is

$$\frac{P_{L1}}{P_{L2}} = \frac{I_{p1}^2 \operatorname{Re} Z(\omega_0)}{I_{p2}^2 \operatorname{Re} Z(2\omega_0)} = \frac{R_LQ^2I_{p1}^2}{R_LQ^2I_{p2}^2/4(1 + 2.25Q^2)} = 4(1 + 2.25Q^2)$$

where Re denotes "the real part of." With $Q = 10$, this reduces to

$$\frac{P_{L1}}{P_{L2}} = 900$$

Clearly, therefore, the second-harmonic power is negligible under these conditions.

Obviously, there will be losses in the tank circuit owing to the resistive component of the coils, and perhaps the capacitor. The power delivered to the load is

$$P_L'' = (QI_{p1})^2\frac{\omega^2M^2}{R_L'' + R_2}\frac{R_L''}{R_L'' + R_2} \tag{12-7}$$

and the power lost in the tank circuit is

$$P_T = (QI_{p1})^2\left(R_1 + \frac{\omega^2M^2}{R_L'' + R_2}\frac{R_2}{R_L'' + R_2}\right) \tag{12-8}$$

The circuit transfer efficiency, which is defined as the ratio of the power delivered to the load to that supplied to the tank circuit, is given by

$$\eta = \frac{P''_L}{P''_L + P_T} \times 100\% = \frac{P'_L - P_T}{P'_L} \times 100\% \qquad (12\text{-}9)$$

An interesting and informative form for the circuit transfer efficiency is possible by writing it as follows:

$$\eta = \eta_1\eta_2 = \frac{\text{power delivered to secondary}}{\text{power delivered to primary}} \times \frac{\text{power delivered to load}}{\text{power delivered to secondary}}$$

where η_1 is associated with the first ratio and η_2 is associated with the second ratio. These may be written as

$$\eta_1 = \frac{I_1^2 R'_L}{I_1^2(R_1 + R'_L)} = \frac{R'_L}{R_1 + R'_L}$$

Similarly

$$\eta_2 = \frac{I_2^2 R''_L}{I_2^2(R_2 + R''_L)} = \frac{R''_L}{R_2 + R''_L}$$

The expression for η_1 may be written in the following forms:

$$\eta_1 = \frac{\omega_0 L_1/R_1 - \omega_0 L_1/(R'_L + R_1)}{\omega_0 L_1/R_1} = \frac{Q_{01} - Q_{01L}}{Q_{01}} = 1 - \frac{Q_{01L}}{Q_{01}} \qquad (12\text{-}10a)$$

where $Q_{01} = \omega_0 L_1/R_1$ is the unloaded Q of the primary coil at resonance
$Q_{01L} = \omega_0 L_1/(R'_L + R_1)$ is the loaded Q of the primary circuit at resonance, including the reflected resistance of the secondary in the primary circuit

In an entirely similar way, the expression for η_2 may be written in the form

$$\eta_2 = 1 - \frac{Q_{02L}}{Q_{02}} \qquad (12\text{-}10b)$$

where $Q_{02} = \omega_0 L_2/R_2$ is the unloaded Q of the secondary coil at resonance
$Q_{02L} = \omega_0 L_2/(R_2 + R''_L)$ is the loaded Q of the secondary coil at resonance but without any effect of the primary circuit on the secondary

The complete expression for the circuit transfer efficiency becomes

$$\eta = \left(1 - \frac{Q_{01L}}{Q_{01}}\right)\left(1 - \frac{Q_{02L}}{Q_{02}}\right) \qquad (12\text{-}11)$$

For high circuit transfer efficiency, the loaded values Q_{01L} and Q_{02L} must be low, and the unloaded values Q_{01} and Q_{02} should be high. Ordinarily the loaded Q's must be 10 or greater in order to provide for a low harmonic content in the output. The unloaded Q's are subject to purely practical limitations; the possible values depend on the power output, the character of construction of the coil, and the frequency of operation. Typical values for coils of conventional design vary somewhat as follows for frequencies in the range from 500 to 1,500 kc:

Unloaded $Q \sim$ 100–200 for low-power coils
$\sim$ 500–800 for high-power coils

12-3. Choice of Q_L. It is of some interest to examine the factors which influence the choice of Q_L. Several of the factors have already been considered, but for completeness these will also be included in the tabulation below. The following conditions prevail for *low* Q_L:

1. High circuit transfer efficiency η.
2. Broader bandwidth.
3. Higher harmonic components.
4. Greater L/C ratio.

Factor 1 has been considered in considerable detail in Sec. 12-2. Factor 2 relates to the width of the pass band. This must be adequate to pass the desired frequency band but must attenuate the frequencies outside the specified band. A measure of the response is obtained from Eq. (12-2), which becomes, for frequencies near resonance

$$\frac{Z(\omega)}{Z(\omega_0)} = \frac{1}{(\omega Q/\omega_0)[1 - (\omega_0/\omega)^2]} \qquad (12\text{-}12)$$

Factor 3 was discussed in some detail in Sec. 12-2, where it was shown that the harmonic output is small if Q_L is fairly high. When Q_L is low, the harmonic output is not negligible and might result in troublesome harmonic potentials in the circuit.

Factor 4 is examined through Eq. (12-4) for the lowest Q_L for a specified R_0; this demands that the L/C ratio must be high. The highest L/C ratio exists when C is a minimum, which, in the extreme, is the tube plus stray wiring capacitances. If a capacitor is used, it should be relatively small, in parallel with a large inductor. In any design considerations Q_L is established by the allowable harmonic content and by power considerations. Normally, as already discussed, Q_L will range from 10 to 20. The unloaded Q_u is determined by requiring that the circuit transfer efficiency should be high, perhaps 90 per cent, at the lower powers and should be higher for high powers. With Q_L and Q_u known, the circuit constants can be determined.

Example. Evaluate the approximate circuit constants of a tank circuit which is to deliver 500 watts to a 72-ohm load at 2 Mc from an a-c supply of 2,000 volts.

Solution. Choose $Q_L = 12$; $\eta = 90$ per cent. Also given, $R''_L = 72$ ohms, $P''_L = 500$ watts.

a. Power input

$$P_1 = \frac{500}{0.9} = 556 \text{ watts}$$

b. From expression (12-11)

$$Q_u = \frac{Q_L}{0.1} = 120$$

c. Since

$$Q_L = \frac{\omega_0 L_1}{R_L} = \frac{(\omega_0 L_1 I_1) I_1}{I_1^2 R_L} = \frac{E I_1}{I_1^2 R_L}$$

then

$$I_1 = \frac{12 \times 556}{2{,}000} = 3.33 \text{ amp}$$

Also

$$L \doteq \frac{E}{\omega I_1} = \frac{2{,}000}{2 \times 2 \times 10^6 \times 3.33} = 47.8 \times 10^{-6} \text{ henry}$$

$$C = \frac{I_1}{\omega E} = \frac{3.33}{2 \times 2 \times 10^6 \times 2{,}000} = 132.4 \times 10^{-12} \text{ farad}$$

d. To find M, note that

$$\omega M I_1 \doteq I_2 R_L'' \qquad P = I_2^2 R_L''$$

Hence

$$M \doteq \frac{I_2 R_L''}{I_1 \omega} = \frac{\sqrt{P R_L''}}{I_1 \omega}$$

$$= \frac{\sqrt{500 \times 72}}{3.33 \times 2 \times \pi \times 10^6} = 4.53 \times 10^{-6} \text{ henry}$$

e. Current I_{p1}

$$I_{p1} = \frac{556}{2{,}000} = 0.277 \text{ amp}$$

f. Loaded R_0

$$\text{Loaded } R_0 = \frac{2{,}000}{0.277} = 7{,}220 \text{ ohms}$$

g. Unloaded R_0

$$\text{Unloaded } R_0 = {}^{120}\!/_{12} \times 7{,}220 = 72{,}200 \text{ ohms}$$

12-4. Class B Tuned Amplifiers. Considerations regarding the actual choice of tube will be given in Sec. 12-16. Transmitters may employ high- or low-impedance triodes, tetrodes, or pentodes. It will be found that the plate-circuit efficiency, i.e., the ability of the tube to convert d-c power from the supply into a-c power, is not particularly dependent on the type of tube that is used. This fact will become clearer in the light of subsequent discussions.

Under class B operation, the grid-bias supply potential E_{cc} in Fig. 12-1 is made negative by an amount sufficient to reduce the plate current to zero for zero signal potential e_g. If the dynamic characteristic of the amplifier is linear over the range of operation, then for sinusoidal input signal potential the current will consist of half-wave rectified pulses

The construction for deducing the output waveshape is sketched in Fig. 12-4.

It is important that it be recognized that Fig. 12-4 represents an idealized picture which depends upon a linear dynamic curve. This is not completely true, although, in the analysis to follow, it will be assumed that the linear relation does apply. If the dynamic curve is not linear, then a graphical solution must be used in order to determine the shape of the plate-current curve and the linear class B analysis is not valid.

To find the operating path of an amplifier with a tuned load, a special construction is required, since the conditions are different from those of an amplifier with a pure resistance load. This is so because of the interrelation of a number of factors and the different manner of operation of the circuit. Among the important factors that must be considered are

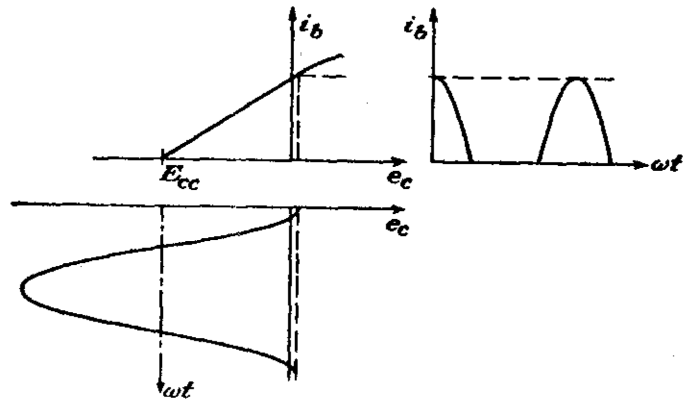

FIG. 12-4. The output waveshape from a class B stage, with a linear dynamic curve.

the allowable plate dissipation of the tube, the Q of the circuit, the effective shunt resistance of the tank circuit, the grid driving potential, the shape of the plate-current wave, and the corresponding harmonic components in the plate current. Ordinarily a method of successive approximations is necessary in which a given set of conditions is assumed and a calculation is made. If a consistent solution is not found, a second trial must be made. This procedure must be continued until a consistent solution is found.

Although the determination of the operating path is not essential for the linear analytical solution to follow, the method will be discussed here, since it will permit a check on the validity of the linear assumptions. Moreover, it is a general method and will also be used later in the discussion of the tuned class C amplifier. The details of the construction are illustrated in Fig. 12-5.

To find the operating path, it is assumed that the plate-potential swing sinusoidal when the grid input signal is sinusoidal. Also, as a starting point, it is assumed that $e_{b,\,min}$ is approximately 10 per cent of E_{bb}. The value of $e_{c,\,max}$ must not be allowed to reach an instantaneous positive potential that is higher than the plate potential $e_{b,\,min}$; otherwise the current to the grid will increase very rapidly. This may cause serious damage to the tube. Even if no damage results, the increasing grid current is accompanied by a decreasing plate current, and in consequence the analysis will no longer be valid owing to the resulting nonlinearity of the dynamic curve. With the indicated choice of conditions, the analysis can be completed, and a calculation can be made of the following: the d-c power from the plate-supply source, the a-c power output to the load,

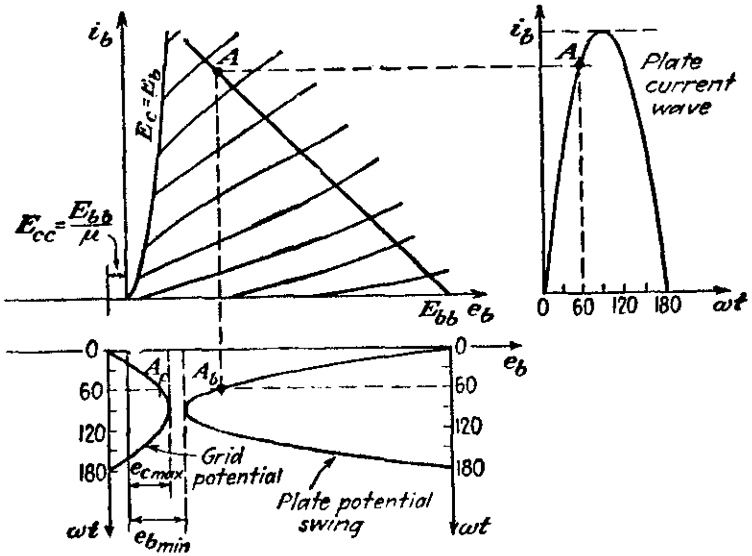

FIG. 12-5. The construction for determining the plate-current waveshape graphically from the plate characteristics.

and the plate dissipation. If the plate dissipation is within the rating of the tube, then the resulting calculations will indicate the adjustments of the circuit parameters that are necessary to achieve the indicated results.

The specific procedure is the following (refer to Fig. 12-5): Select any particular instantaneous grid potential e_c, such as that corresponding to the point A_c. Determine the corresponding instantaneous plate potential e_b by locating the point A_b at the same phase angle in the operating cycle. By projecting A_b up to its intersection with the curve for the selected grid potential, the point A on the operating path will be located. Other points are determined in a similar manner. For class B operation, the operating path should be approximately linear and should intersect the plate-potential axis at E_{bb}, approximately.

To determine the shape of the plate-current pulse as a function of the phase angle, the current corresponding to each point A on the operating path is plotted as a function of the appropriate phase angle. The corresponding plate-current pulse is plotted in Fig. 12-5 as $(i_b,\omega t)$. The curves of Fig. 12-6 illustrate the important waveshapes of the amplifier.

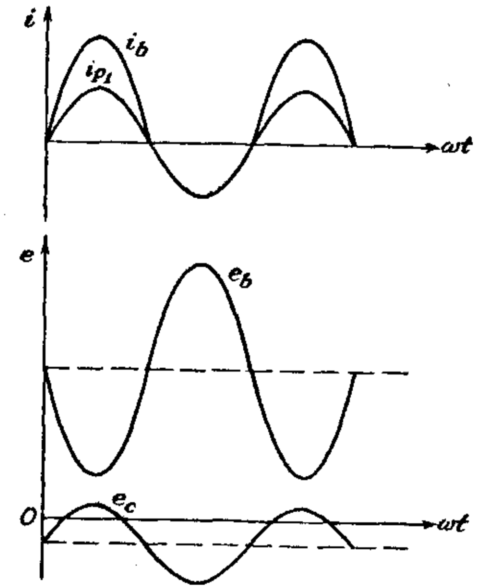

FIG. 12-6. The important waveshapes in a class B tuned amplifier.

12-5. Analytic Solution of Tuned Class B Amplifier.[1] An analytic solution of the tuned class B amplifier is based on finding an analytic form for the tube characteristics. From Eq. (1-14), the general relationship between the plate current and the plate and grid potentials is of the form

$$i_b = k\left(e_c + \frac{e_b}{\mu}\right)^{\alpha} \qquad e_c + \frac{e_b}{\mu} > 0$$

Actually, it is found that for power triodes over a wide range of parameters the plate current is of the form

$$i_b = k\left(e_c + \frac{e_b}{\mu}\right)$$

which may be written in the more complete form

$$i_b = g_m\left(e_c + \frac{e_b}{\mu}\right) \tag{12-13}$$

This is, of course, simply the first term in the Taylor expansion for the current.

The instantaneous potentials are of the form

$$\begin{aligned} e_c &= E_{cc} + E_{gm} \cos \omega t \\ e_b &= E_{bb} - E_{p1m} \cos \omega t \end{aligned} \tag{12-14}$$

But since the current is zero when the grid signal is zero, then, for $i_b = 0$,

$$e_c + \frac{e_b}{\mu} = 0$$

which requires that

$$E_{cc} + \frac{E_{bb}}{\mu} = 0$$

or, for cutoff,

$$E_{cc} = -\frac{E_{bb}}{\mu} \tag{12-15}$$

By combining Eqs. (12-13) and (12-14), the expression for plate current becomes

$$i_b = g_m\left(E_{cc} + E_{gm}\cos\omega t + \frac{E_{bb}}{\mu} - \frac{E_{p1m}}{\mu}\cos\omega t\right)$$
$$= g_m\left(E_{gm}\cos\omega t - \frac{E_{p1m}}{\mu}\cos\omega t\right)$$
$$= g_m\left(E_{gm} - \frac{E_{p1m}}{\mu}\right)\cos\omega t \qquad (12\text{-}16)$$

which is written in the form

$$i_b = I_{bm}\cos\omega t \qquad -\frac{\pi}{2} < \omega t < \frac{\pi}{2}$$
$$i_b = 0 \qquad \frac{\pi}{2} < \omega t < \frac{3\pi}{2} \qquad (12\text{-}17)$$

where $\qquad I_{bm} = g_m\left(E_{gm} - \frac{E_{p1m}}{\mu}\right)$

The average value of the plate current is

$$I_b = \frac{1}{2\pi}\int_0^{2\pi} i_b\, d(\omega t) \qquad (12\text{-}18)$$

or

$$I_b = \frac{2}{2\pi}\int_0^{\pi/2} I_{bm}\cos\omega t\, d(\omega t) = \frac{I_{bm}}{\pi} \qquad (12\text{-}19)$$

Also, by Fourier analysis, the amplitude of the fundamental component of the plate current is

$$I_{p1m} = \frac{1}{\pi}\int_0^{2\pi} i_b\cos\omega t\, d(\omega t) \qquad (12\text{-}20)$$

or

$$I_{p1m} = \frac{2}{\pi}\int_0^{\pi/2} I_{bm}\cos^2\omega t\, d(\omega t) = \frac{I_{bm}}{2} \qquad (12\text{-}21)$$

But at resonance $\mathbf{Z}(\omega_0) = R_0$ is resistive, and the fundamental-frequency potential difference across the load is

$$E_{p1m} = I_{p1m}R_0 \qquad (12\text{-}22)$$

Combining Eq. (12-22) with (12-21) and (12-17),

$$E_{p1m} = \frac{R_0 I_{bm}}{2} = \frac{R_0}{2}g_m\left(E_{gm} - \frac{E_{p1m}}{\mu}\right)$$

It follows from this that

$$E_{p1m} + \frac{R_0}{2}g_m\frac{E_{p1m}}{\mu} = \frac{R_0}{2}g_m E_{gm}$$

or

$$E_{p1m} = R_0\frac{\mu E_{gm}}{R_0 + 2r_p} \qquad (12\text{-}23)$$

which yields, for the rms value of the fundamental-frequency component of current, the expression

$$\mathbf{I}_{p1} = \frac{\mu\mathbf{E}_g}{2r_p + R_0} \qquad (12\text{-}24)$$

Also, from Eqs. (12-21) and (12-24),

$$I_b = \frac{I_{bm}}{\pi} = \frac{2\sqrt{2}\,I_{p1}}{\pi} = \frac{2\sqrt{2}}{\pi}\frac{\mu E_g}{2r_p + R_0} \qquad (12\text{-}25)$$

The gain of the amplifier is given in Eq. (12-23) and is

$$\mathbf{K} = -\frac{\mu R_0}{2r_p + R_0} \qquad (12\text{-}26)$$

The d-c power input to the plate circuit, which is equal to the average power furnished by the plate supply when the d-c power dissipated in the plate load resistance is negligible, is given by

$$P_{bb} = \frac{1}{2\pi}\int_0^{2\pi} E_{bb}i_b\, d(\omega t)$$

This becomes

$$P_{bb} = E_{bb}\frac{1}{2\pi}\int_0^{2\pi} i_b\, d(\omega t) = E_{bb}I_b \qquad (12\text{-}27)$$

The a-c power output of importance is that at the fundamental frequency and is given by

$$P_L = \frac{1}{2\pi}\int_0^{2\pi} e_L i_p\, d(\omega t)$$

which becomes

$$P_L = \frac{1}{2\pi}\int_0^{2\pi} E_{p1m}\cos\omega t\, I_{p1m}\cos\omega t\, d(\omega t)$$
$$P_L = E_{p1}I_{p1} = I_{p1}^2 R_0 \qquad (12\text{-}28)$$

The plate-circuit efficiency, which is the ratio of P_{ac} to P_{bb}, is

$$\eta_p = \frac{P_L}{P_{bb}} \times 100\% = \frac{E_{p1}I_{p1}}{E_{bb}I_b} \times 100\%$$
$$\eta_p = \frac{E_{p1}I_{p1}}{E_{bb}(2\sqrt{2}/\pi)I_{p1}} = \frac{\pi}{2\sqrt{2}}\frac{E_{p1}}{E_{bb}} = \frac{\pi}{4}\frac{E_{p1m}}{E_{bb}}$$
$$\eta_p = 78.5 \times \frac{E_{p1m}}{E_{bb}} \quad \% \qquad (12\text{-}29)$$

The plate dissipation is given by

$$P_p = \frac{1}{2\pi}\int_0^{2\pi} e_b i_b \, d(\omega t)$$

or

$$P_p = \frac{1}{2\pi}\int_0^{2\pi} (E_{bb} - e_L) i_b \, d(\omega t) = E_{bb} I_b - P_L \tag{12-30}$$

which becomes, by virtue of Eqs. (12-27) to (12-29),

$$P_p = (1 - \eta_p) P_{bb} \tag{12-31}$$

It is of some interest to calculate the results corresponding to the optimum conditions $e_{c,\max} = e_{b,\min}$. For this condition

$$\begin{aligned} e_{c,\max} &= E_{cc} + E_{gm} \\ e_{b,\min} &= E_{bb} - E_{p1m} \end{aligned} \tag{12-32}$$

from which

$$E_{gm} + E_{p1m} = E_{bb} - E_{cc}$$

By Eqs. (12-14) and (12-23), this yields

$$E_{gm} + E_{gm}\frac{\mu R_0}{2r_p + R_0} = E_{bb} + \frac{E_{bb}}{\mu}$$

or

$$E_{gm} = E_{bb}\frac{\mu + 1}{\mu}\,\frac{2r_p + R_0}{2r_p + (\mu + 1)R_0} \tag{12-33}$$

The corresponding expressions for the fundamental-frequency component and the d-c components of current are, respectively,

$$I_{p1} = \frac{E_{bb}(\mu + 1)}{\sqrt{2}}\,\frac{1}{2r_p + (\mu + 1)R_0} \tag{12-34}$$

and

$$I_b = \frac{2}{\pi} E_{bb}(\mu + 1)\,\frac{1}{2r_p + (\mu + 1)R_0} \tag{12-35}$$

The corresponding values of the optimum P_{bb}, P_{ac}, and η_p are readily calculated from these expressions for I_{p1} and I_b. The expression for the plate-circuit efficiency is found to be

$$\eta_p = \frac{I_{p1}^2 R_0}{E_{bb} I_b} = \frac{\left[\dfrac{E_{bb}(\mu + 1)}{\sqrt{2}}\,\dfrac{1}{2r_p + (\mu + 1)R_0}\right]^2 R_0}{\dfrac{2}{\pi} E_{bb}^2(\mu + 1)\,\dfrac{1}{2r_p + (\mu + 1)R_0}}$$

which reduces to

$$\eta_p = 78.5 \times \frac{R_0(\mu + 1)}{2r_p + R_0(\mu + 1)} \quad \% \tag{12-36}$$

Ordinarily the plate dissipation will be a fixed rating of the amplifier and is the limiting factor on the output power. The appropriate value of R_0 is then specified, since all aspects of the circuit may be expressed in terms of it. To examine this, note that

$$P_p = E_{bb} I_b - I_{p1}^2 R_0$$

which may be written as

$$P_p = E_{bb}^2 \frac{2}{\pi}(\mu + 1)\frac{1}{2r_p + (\mu + 1)R_0} - R_0\left[\frac{E_{bb}}{\sqrt{2}}(\mu + 1)\frac{1}{2r_p + (\mu + 1)R_0}\right]^2$$

This expression may be rearranged and yields the following quadratic expression for R_0, from which R_0 may be evaluated:

$$R_0^2 + \left[\frac{4r_p}{\mu + 1} - \frac{E_{bb}^2}{P_p}\left(\frac{2}{\pi} - \frac{1}{2}\right)\right] R_0 + \left[\frac{4r_p^2}{(\mu + 1)^2} - \frac{E_{bb}^2}{P_p}\,\frac{4r_p}{(\mu + 1)\pi}\right] = 0 \tag{12-37}$$

12-6. Analysis of Class C Amplifiers. An analysis of the operation of the tuned class C amplifier can be made on the basis of the assumption of a linear tube characteristic, essentially as an extension of the method of Sec. 12-4.[2] This analysis is considerably complicated by the fact that E_{cc} is no longer the single value chosen to yield a zero current for zero excitation but is now a parameter. Moreover, it is no longer valid to assume that the operating characteristic is linear. Hence, although such a linear-tube-characteristic analysis is possible, it is a poor approximation. It does have the advantage over other methods of giving an explicit solution for the optimum operating conditions. Owing to its approximate nature, other methods are preferred.

To see that the operating path is not linear, the construction of Fig. 12-5 is again employed. The only differences that exist arise because the grid bias E_{cc} is adjusted beyond the cutoff value. With such values of E_{cc} and with the appropriately increased value of grid driving potential, the results have the form illustrated in Fig. 12-7. The curves of Fig. 12-8 illustrate the important waveshapes in such an amplifier.

A comparison of these curves with those of Fig. 12-6 indicates that in the class C amplifier the plate current consists of pulses the duration of which is less than 180 deg of the cycle. Also, it is not possible, in general, to derive easily an analytic expression for the shape of the plate-current pulse.

Some progress can be made in finding an approximate analytical solution if the curves of Fig. 12-7 are idealized. The idealization made is in the assumption of linear curves, as illustrated in Fig. 12-9. This approximation permits the operating path to be represented by two straight-line segments. It is now possible to write an expression for the plate-current

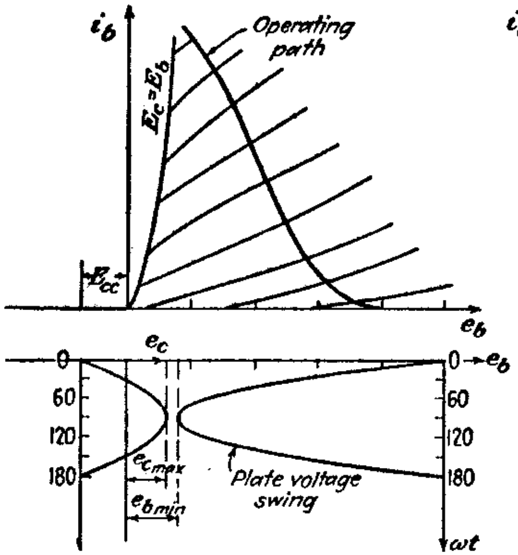

FIG. 12-7. The construction for determining the plate-current pulses in a class C amplifier.

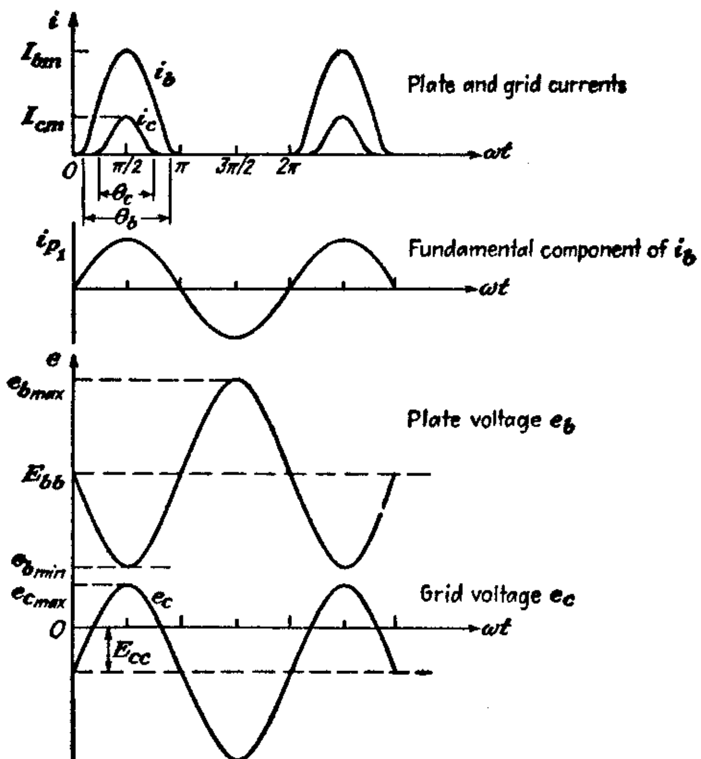

FIG. 12-8. The waveshapes at various points in the tuned amplifier.

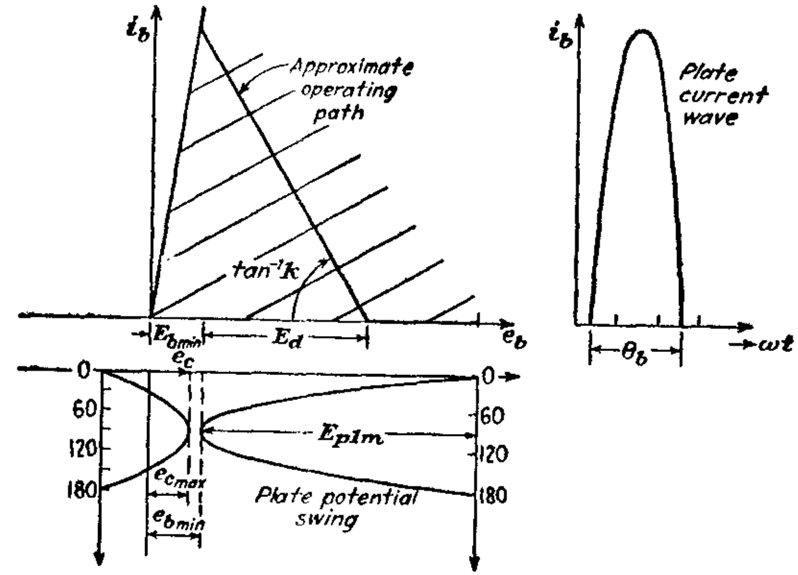

FIG. 12-9. The conditions in an idealized class C amplifier.

pulse. This is given by the relation

$$i_b = k[E_{p1m} \cos \omega t - (E_{p1m} - E_d)] \qquad \text{for } i_b > 0 \qquad (12\text{-}38)$$

where, by definition, for the condition of zero current

$$\frac{\theta_b}{2} = \omega t_b = \cos^{-1} \frac{E_{p1m} - E_d}{E_{p1m}} = \cos^{-1}\left(1 - \frac{E_d}{E_{p1m}}\right) \qquad (12\text{-}39)$$

Note that the maximum tube current is given by

$$I_{b,\max} = kE_d \qquad (12\text{-}40)$$

With the shape of the current pulse known, it is possible to compute plate-circuit information. The average value of the plate-current pulse is

$$I_b = \frac{1}{2\pi}\int_0^{2\pi} i_b \, d(\omega t)$$

which is given by the relation

$$I_b = \frac{k}{\pi}\int_0^{\theta_b/2} [E_{p1m} \cos \omega t - (E_{p1m} - E_d)]\, d(\omega t)$$

This integrates to the value

$$I_b = \frac{k}{\pi}\left[E_{p1m} \sin \frac{\theta_b}{2} - (E_{p1m} - E_d)\,\frac{\theta_b}{2}\right] \qquad (12\text{-}41)$$

Similarly, the amplitude of the fundamental component is given by the integral

$$I_{p1m} = \frac{1}{\pi} \int_0^{2\pi} i_b \cos \omega t \, d(\omega t)$$

which may be written in the form

$$I_{p1m} = \frac{2k}{\pi} \int_0^{\theta_b/2} [E_{p1m} \cos \omega t - (E_{p1m} - E_d)] \cos \omega t \, d(\omega t)$$

This integrates to the value

$$I_{p1m} = \frac{2k}{\pi} \left[\frac{E_{p1m}}{4} (\theta_b + \sin \theta_b) - (E_{p1m} - E_d) \sin \frac{\theta_b}{2} \right] \qquad (12\text{-}42)$$

It is quite possible to continue with this analysis and obtain expressions for the power transferred to the load, the plate dissipation in the tube, the power supplied by the plate power supply, and the plate-circuit efficiency, in a manner analogous to that for the class B amplifier. However, it is noted that the construction of Fig. 12-9 is necessary in order to deduce the operating path before the approximate operating path may be obtained. The results will be in error consequently, owing to the approximations. Moreover, once the construction of Fig. 12-9 is available, a semigraphical solution may be effected directly without the approximations involved in the foregoing. Because of this, the above method of analysis will not be continued, but the semigraphical method will be discussed in detail.

Attention is called to the fact that, with the class C amplifier, there will be no output for small grid signals, since the plate current is zero Consequently, the output potential is not proportional to the input potential, and these amplifiers cannot be used where such a linear relation must be maintained. They are used extensively for amplifying a signal of fixed amplitude. They are also used extensively in radio communications as either low-level or high-level modulation stages. This latter application will be examined in detail in Chap. 17. When the amplifier is biased to class B operation, a linear relation between the output and input potentials does exist and such amplifiers find extensive use in those applications requiring this characteristic. The most important application is to increase the power level of a modulated carrier wave.

12-7. Semigraphical Analysis of Class C Amplifiers. Before carrying out the details of the analysis, attention is called to a second method of obtaining the operating path of a tuned power amplifier. This makes use of the fact that the operating line appears as a straight line on the constant-current (e_b, e_c) characteristics of the tube. These constant-current tube characteristics are available for transmitting-type tubes and are provided for this particular purpose.

To verify that the dynamic characteristic is a straight line on the constant-current characteristics, use is made of Eqs. (12-13) for the grid and plate potentials, viz.,

$$\begin{aligned} e_c &= E_{cc} + E_{gm} \cos \omega t \\ e_b &= E_{bb} - E_{p1m} \cos \omega t \end{aligned} \qquad (12\text{-}43)$$

This latter expression is valid when the Q of the tank circuit is 10 or greater. Now combine these expressions by writing

$$\begin{aligned} \frac{e_c}{E_{gm}} &= \frac{E_{cc}}{E_{gm}} + \cos \omega t \\ \frac{e_b}{E_{p1m}} &= \frac{E_{bb}}{E_{p1m}} - \cos \omega t \end{aligned}$$

Adding these expressions gives

$$\frac{e_c}{E_{gm}} + \frac{e_b}{E_{p1m}} = \frac{E_{cc}}{E_{gm}} + \frac{E_{bb}}{E_{p1m}}$$

This may be written in the form

$$e_c = -\frac{E_{gm}}{E_{p1m}} e_b + E_{cc} + \frac{E_{gm}}{E_{p1m}} E_{bb} \qquad (12\text{-}44)$$

which is the slope-intercept form of the equation of a straight line. The results are illustrated in Fig. 12-10.

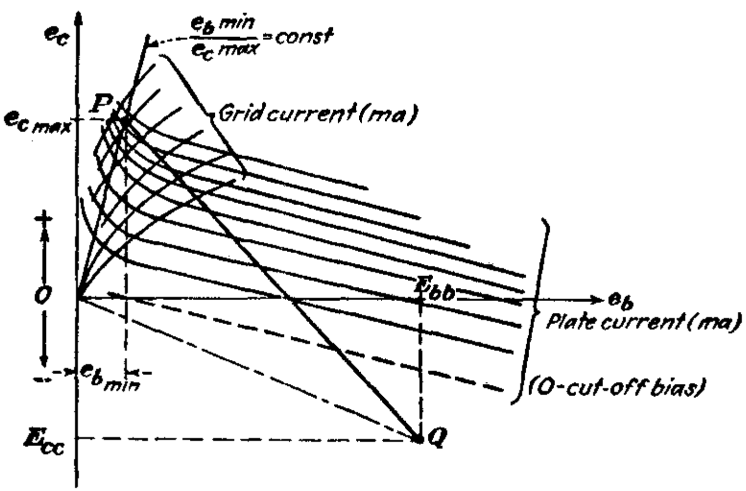

FIG. 12-10. The operating line on the constant-current curves of a power tube.

In order to establish the range of operation, it is necessary to specify the end points of the region of operation. Ordinarily this is done by specifying E_{bb}, $e_{b,\min}$, $e_{c,\max}$, quantities which are determined from considerations of economy, power output desired, efficiency, and tube ratings. The manner of this dependence will be investigated below.

With these factors specified, the operating characteristics of the amplifier are obtained from the curves in the manner illustrated in Fig. 12-11.

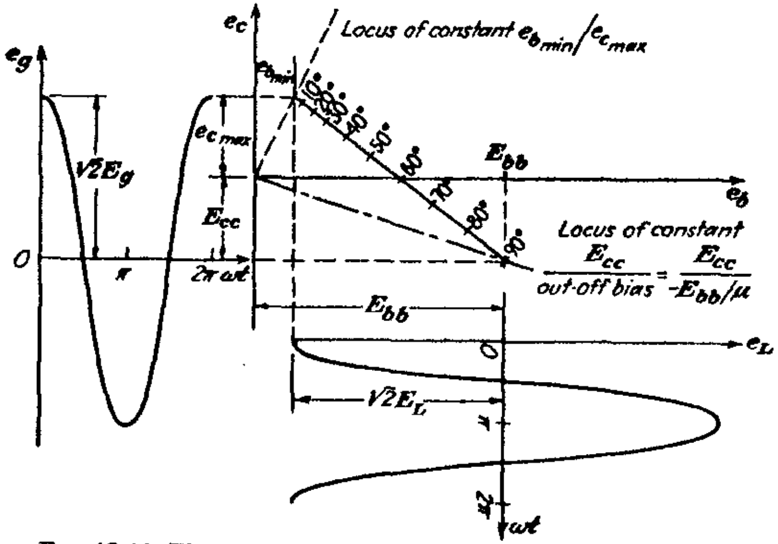

FIG. 12-11. The operating characteristics of a class C amplifier.

12-8. Grid and Plate Currents in Class C Amplifiers.[8] In order to obtain a numerical solution of the operational features of the amplifier, such as power output, efficiency, grid driving power, and plate dissipation, the average and rms values of the grid and plate currents are required. These must be deduced from the plate- and grid-current pulses as obtained from the curves, as discussed above. It is well to examine this matter before considering a detailed analysis of the amplifier operation.

An inspection of Figs. 12-7 and 12-8 shows that the plate- and grid-current pulses possess zero-axis symmetry. Consequently, these recurring waves may be represented by a Fourier series involving only cosine terms. In particular, the plate- and grid-current pulses may be represented analytically by series of the form

$$\begin{aligned} i_b &= I_b + I_{p1m} \cos \omega t + I_{p2m} \cos 2\omega t + \cdots \\ i_c &= I_c + I_{g1m} \cos \omega t + I_{g2m} \cos 2\omega t + \cdots \end{aligned} \tag{12-45}$$

The average or d-c value of the plate current is given by the integral

$$I_b = \frac{1}{2\pi} \int_0^{2\pi} i_b \, d(\omega t)$$

which becomes, by virtue of the zero-axis symmetry and the fact that conduction proceeds over the angle θ_b,

$$I_b = \frac{1}{\pi} \int_0^{\theta_b/2} i_b \, d(\omega t) \tag{12-46}$$

This integral expresses the area under the plate-current pulse. Since, however, an analytic expression for the current pulse is not available, recourse is had to any of the available methods of numerical integration, e.g., through the use of a planimeter; by dividing the base of the wave into equal parts, approximating the mean ordinates of the resulting rectangles, and then summing the areas of these rectangles; or through the use of other methods devised for numerical integration.

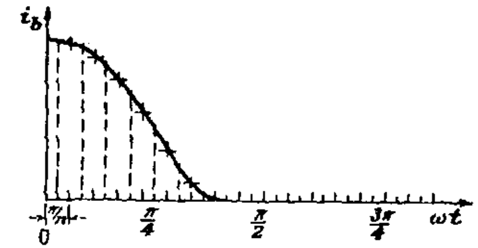

FIG. 12-12. Current waveform and its approximate representation.

The details of the second method are given. Suppose that Fig. 12-12 is the current waveform, certain features of which are to be examined. Suppose that the half recurrence period is divided into n equal parts; hence each division is $\pi/n = 180/n$ deg long. Since the current flow will proceed for less than 90 deg in each half period, and taking account of the symmetry, the integral for I_b is then given with good approximation by the expression

$$I_b \doteq \frac{1}{n} \left[\frac{i_b(0)}{2} + \sum_{k=1,2,3,\ldots} i_b \left(\frac{k\pi}{n} \right) \right] \tag{12-47}$$

where $i_b(k\pi/n)$ denotes the value of the current at the angle $k\pi/n$.

The average value of the grid current is found in a similar manner from the graph of the grid-current pulse. It is

$$I_c = \frac{1}{2\pi} \int_0^{2\pi} i_c \, d(\omega t)$$

which has the form

$$I_c = \frac{1}{\pi} \int_0^{\theta_c/2} i_c \, d(\omega t) \tag{12-48}$$

where θ_c denotes the grid-current conduction angle. In terms of the approximate calculation, this becomes

$$I_c \doteq \frac{1}{n} \left[\frac{i_c(0)}{2} + \sum_{k=1,2,3,\ldots} i_c \left(\frac{k\pi}{n} \right) \right] \tag{12-49}$$

The amplitude of the fundamental-harmonic component of the plate current is obtained from considerations of the general Fourier series representation of the current. This leads to the form

$$I_{p1m} = \frac{1}{\pi} \int_0^{2\pi} i_b \cos \omega t \, d(\omega t)$$

which may be written, in view of the existing symmetry, in the form

$$I_{p1m} = \frac{2}{\pi} \int_0^{\theta_b/2} i_b \cos \omega t \, d(\omega t) \tag{12-50}$$

This integral may be expressed as a summation by the approximate methods that have been employed above. This becomes

$$I_{p1m} \doteq \frac{2}{n} \left[\frac{i_b(0) \cos 0}{2} + \sum_k i_b\left(\frac{k\pi}{n}\right) \cos \frac{k\pi}{n} \right] \tag{12-51}$$

The amplitude of the fundamental-harmonic component of the grid current is obtained in the same way as the corresponding component of plate current. It is given by

$$I_{g1m} = \frac{1}{\pi} \int_0^{2\pi} i_c \cos \omega t \, d(\omega t)$$

which reduces to the form

$$I_{g1m} = \frac{2}{\pi} \int_0^{\theta_c/2} i_c \cos \omega t \, d(\omega t) \tag{12-52}$$

In general, the grid current flows for a relatively small portion of the cycle in the neighborhood of $\theta_c = 0$. But the value of $\cos \omega t$ does not appreciably differ from unity during this interval. Then approximately

$$I_{g1m} \doteq \frac{2}{\pi} \int_0^{\theta_c/2} i_c \, d(\omega t)$$

from which it follows that

$$I_{g1m} \doteq 2I_c \tag{12-53}$$

In general, it is not necessary to plot the grid- and plate-current waveforms, since the information may be taken directly from the curve of Fig. 12-11 and combined in a table like Table 12-1 to yield the desired results.

12-9. Power Considerations in Class C Amplifiers. A number of the results are the same as those considered in Sec. 12-3 for the class B amplifier. Here too the d-c power input to the plate circuit, which is equal to the average power furnished by the plate supply when the d-c power dissipated in the plate load resistance is negligible, is given by

$$P_{bb} = \frac{1}{2\pi} \int_0^{2\pi} E_{bb} i_b \, d(\omega t) = E_{bb} I_b \tag{12-54}$$

The a-c power output of importance is that at the fundamental frequency and is given by

$$P_L = \frac{1}{2\pi} \int_0^{2\pi} e_L i_p \, d(\omega t) = \frac{1}{2\pi} \int_0^{2\pi} E_{p1m} \cos \omega t I_{p1m} \cos \omega t \, d(\omega t)$$

which is

$$P_L = \frac{E_{p1m} I_{p1m}}{2} = E_{p1} I_{p1} \tag{12-55}$$

The plate-circuit efficiency is

$$\eta_p = \frac{P_L}{P_{bb}} \times 100\% = \frac{E_{p1} I_{p1}}{E_{bb} I_b} \times 100\% \tag{12-56}$$

The plate-circuit efficiency depends, of course, on the value of $e_{b,\min}$, since, for a specified E_{bb}, E_{p1m} is dictated by $e_{b,\min}$. A calculation of this dependence may be accomplished, using the results of Sec. 12-6. The

TABLE 12-1

ANALYSIS OF CLASS B AND CLASS C TUNED AMPLIFIER

Tube________

E_{bb}____	E_{cc}____	E_{gm}____	$e_{c,\max}$____
$\frac{e_{b,\min}}{e_{b,\max}}$____	$e_{b,\min}$____	E_{p1m}____	$i_{b,\max}$____
$i_{b,\max}$____	n____	k____	θ_b____

$i_b = l =$ length of line PQ____

1	k	0	1	2	3	4	5	6	7	8	9
2	θ_k										
3	$\cos \theta_k$										
4	$l \cos \theta_k$										
5	$i_b(\theta_k)$										
6	$i_c(\theta_k)$										
7	$i_b(\theta_k) \cos \theta_k$										

$$I_b = \frac{1}{n} \left[\frac{i_b(0)}{2} + \sum i_b\left(\frac{k\pi}{n}\right) \right]$$

$$I_c = \frac{1}{n} \left[\frac{i_c(0)}{2} + \sum i_c\left(\frac{k\pi}{n}\right) \right]$$

$$I_{p1m} = \frac{2}{n} \left[\frac{i_b(0) \cos 0}{2} + \sum i_b\left(\frac{k\pi}{n}\right) \cos\left(\frac{k\pi}{n}\right) \right]$$

general form of the relationship is best presented graphically, as in Fig. 12-13, which shows the plate-circuit efficiency vs. the plate-current conduction angle θ_b, with E_{p1m}/E_{bb} as a parameter. It might be noted that typical values for class C operation are θ_b in the range 120 to 150 deg, with corresponding plate-circuit efficiencies approximately from $\eta_p \sim 80$ to 60 per cent.

The power dissipated in the plate of the tube is given by

$$P_p = \frac{1}{2\pi}\int_0^{2\pi} e_b i_b \, d(\omega t) = \frac{1}{2\pi}\int_0^{2\pi} (E_{bb} - e_L) i_b \, d(\omega t)$$

which reduces to

$$P_p = E_{bb} I_b - E_{p1} I_{p1} = P_{bb} - P_L \tag{12-57}$$

By combining this with Eq. (12-56), there results

$$P_p = (1 - \eta_p) P_{bb} \tag{12-58}$$

This expression shows that the plate dissipation decreases as the output power increases, for a given plate power input.

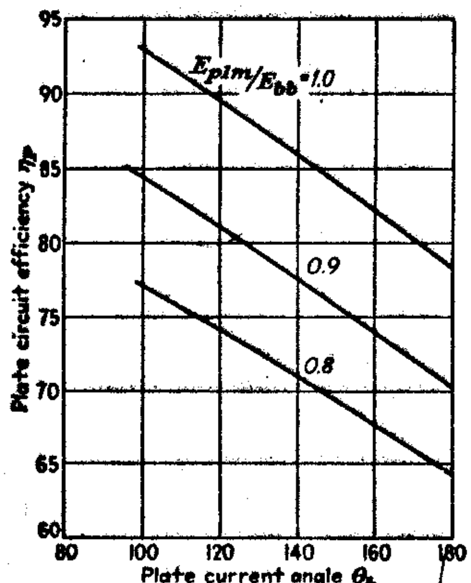

Fig. 12-13. Approximate plate-circuit efficiency for different angles of current flow. (*After A. W. Ladner and C. R. Stoner, "Short Wave Wireless Communication,"* chap. 10, *John Wiley & Sons, Inc., New York,* 1950.)

The average grid power supplied by the driving source is given by

$$P_g = \frac{1}{2\pi}\int_0^{2\pi} e_g i_c \, d(\omega t)$$

This reduces, under the assumption that the grid potential is at its maximum value when the grid current flows and does not vary appreciably during this interval, to

$$P_g \doteq E_{gm} \frac{1}{2\pi}\int_0^{2\pi} i_c \, d(\omega t)$$

which is

$$P_g \doteq E_{gm} I_c \tag{12-59}$$

The results of Thomas[4] have shown that the grid driving power is given more accurately by the expression

$$P_g = 0.9 E_{gm} I_c \tag{12-60}$$

A somewhat better approximation is given by Maling,[5]

$$P_g = E_{gm} I_c \left(0.85 + 0.16 \cos\frac{\theta_c}{2}\right) \qquad \text{for triodes}$$

$$P_g = E_{gm} I_c \left(0.81 - 0.20 \cos\frac{\theta_c}{2}\right) \qquad \text{for tetrodes and pentodes} \tag{12-61}$$

The average grid dissipation is given by the expression

$$P_c = \frac{1}{2\pi}\int_0^{2\pi} e_c i_c \, d(\omega t)$$

This may be written as

$$P_c = \frac{1}{2\pi}\int_0^{2\pi} (E_{cc} + e_g) i_c \, d(\omega t) = E_{cc} I_c + E_{gm} I_c \tag{12-62}$$

But the first term gives a measure of the amount of power that the grid battery is absorbing from the input driving source, since

$$P_{cc} = \frac{1}{2\pi}\int_0^{2\pi} E_{cc} i_c \, d(\omega t) = E_{cc} I_c \tag{12-63}$$

and E_{cc} is inherently negative. Hence the power dissipated in the grid circuit is

$$P_c = P_g - |P_{cc}| \tag{12-64}$$

Example. In order to illustrate the calculations for a typical transmitting tube, consider the following specific problem: A type 806 triode having the constant-current characteristics shown in Fig. 12-14 is used as a class C amplifier, under the following conditions:

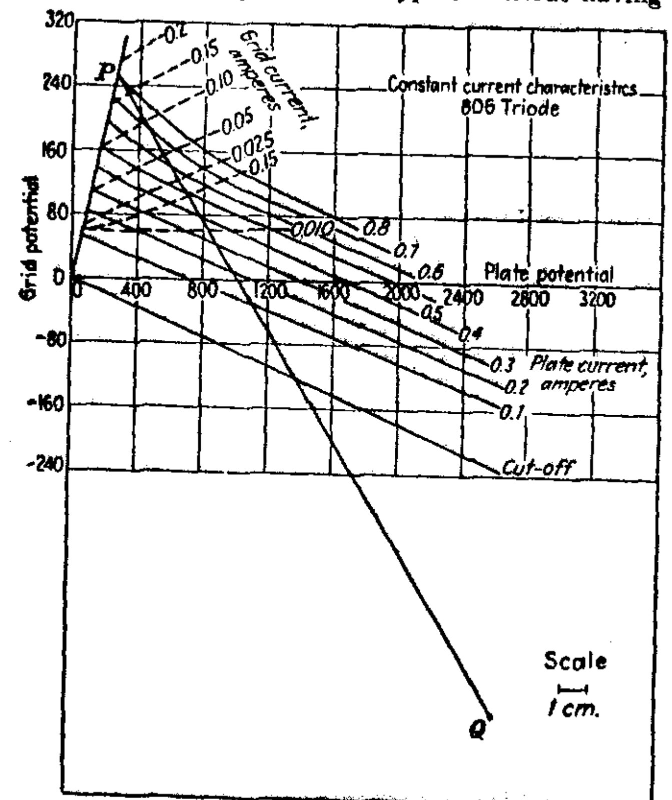

Fig. 12-14. Constant-current characteristics of an 806 triode.

$E_{bb} = 2{,}500$ volts $\qquad E_{cc} = -500$ volts

$\dfrac{e_{b,\min}}{e_{c,\max}} = 1.0 \qquad E_{gm} = 755$ volts

Determine the following:

a. Power supplied by the plate power supply.
b. A-c power output.
c. Plate-circuit efficiency.
d. Plate dissipation.
e. Grid driving power.

(Note: The details of the solution are given in Table 12-2.)

TABLE 12-2

ANALYSIS OF CLASS C AMPLIFIER

Tube—806

$E_{bb} = 2{,}500$ $E_{cc} = -500$ $E_{gm} = 755$ $e_{c,max} = 255$

$\frac{e_{b,min}}{e_{c,max}} = 1.0$ $e_{b,min} = 255$ $E_{p1m} = 2{,}245$ $I_{b,max} = 825$ ma

$I_{c,max} = 185$ ma $n = 18$ $k = 9$ $\theta_b = 120$ deg

Length of line $PQ = 27.8$ cm

1	k	0	1	2	3	4	5	6	7	8	9
2	θ_k, deg	0	10	20	30	40	50	60	70	80	90
3	$\cos\theta_k$	1.0	0.985	0.94	0.86	0.76	0.64	0.50	0.34	0.17	0.00
4	$l \cos\theta_k$	27.8	27.4	26.1	24.1	21.3	17.9	13.9	9.5	4.8	0.0
5	$i_b(\theta_k)$	825	800	750	640	410	150	0	0	0	0
6	$i_c(\theta_k)$	185	170	120	55	12	0	0	0	0	0
7	$i_b(\theta_k)\cos\theta_k$	825	788	710	555	314	96	0	0	0	0

$$I_b = \tfrac{1}{18}\left(\tfrac{825}{2} + 2{,}750\right) = 176 \text{ ma}$$
$$I_c = \tfrac{1}{18}\left(\tfrac{185}{2} + 357\right) = 25 \text{ ma}$$
$$I_{p1m} = \tfrac{1}{9}\left(\tfrac{825}{2} + 2{,}463\right) = 319 \text{ ma}$$
$$P_{bb} = 2{,}500 \times 176 = 440 \text{ watts}$$
$$P_L = \frac{2{,}245 \times 319}{2} = 357 \text{ watts}$$
$$\eta = \tfrac{357}{440} \times 100\% = 81\%$$
$$P_p = (1 - 0.81) \times 440 = 83.5 \text{ watts}$$
$$P_g = 0.9 \times 755 \times 0.025 = 17 \text{ watts}$$

12-10. Design Considerations for Class C Amplifiers. The analysis presented above is based on the assumption that the locus of the operating point of the tube characteristic is known. Frequently, however, the engineering design carries with it the requirement for the selection of the tube and the selection of the operating conditions that govern the locus to give a high plate-circuit efficiency, and other specified results. A number of factors are important in such a design, and it is desirable to examine the influence of these.

The important factors that are involved in the engineering design of a class C amplifier are the following:

1. The peak space current that should be demanded of a given tube. This is usually controlled by the values of $e_{b,min}$ and $e_{c,max}$, since the total peak-space-current demand is given by

$$I_{s,max} = I_{b,max} + I_{c,max}$$
$$= f(e_{b,min}, e_{c,max})$$

2. The minimum potential to which the plate falls, $e_{b,min}$.
3. The maximum value of the instantaneous grid potential, $e_{c,max}$.
4. The angle of plate-current flow, θ_b.
5. The angle of grid-current flow, θ_c.
6. The plate supply potential, E_{bb}.

The influence of each of these factors is considered in some detail.

Item 1. In so far as the total space current that may be safely drawn in a vacuum tube is concerned, it is limited by the allowable emission from the cathode, if saturation current may be drawn from the tube. Although it might not be too unreasonable to draw emission saturation current on the current peaks in a tube that is provided with a pure-tungsten filament, it is unwise to drive a tube with either a thoriated-tungsten or an oxide-coated cathode to such extremes. Reasonable figures for the average emitter are:

Tungsten filament—$I_{s,max}$ approximately 100 per cent of total emission current.

Thoriated-tungsten—$I_{s,max}$ from 15 to 35 per cent of the total emission current.

Oxide-coated cathode—$I_{s,max}$ from 10 to 20 per cent of the total emission current.

Items 2 *and* 3. The optimum values of $e_{b,min}$ and $e_{c,max}$ will be such that the total allowable peak space current will not be exceeded. Moreover, their relative values must be so chosen that the maximum plate current occurs at $\omega t = 0$. This requires that the tube must not be driven so hard that it operates in the region of rapidly falling plate current. Such a condition is avoided by keeping $e_{b,min} > e_{c,max}$. However, high plate-circuit efficiency results when $e_{b,min} = e_{c,max}$, although for low grid driving power it is required that $e_{b,min} > e_{c,max}$. Typical values of the ratio $e_{b,min}/e_{c,max}$ usually range from 1 to 2.

Item 4. The range over which plate conduction occurs, i.e., the conduction angle θ_b, influences both the average current I_b and the first-harmonic current amplitude I_{p1m}. For a large value of the first-harmonic current amplitude, it is desirable that θ_b be made large. However, in order to provide a high value of plate-circuit efficiency, small values of θ_b are indicated. Consequently, it is necessary to compromise between plate efficiency and power output. Typical values for class C operation, as already discussed, are θ_b in the range from 120 to 150 deg, with corresponding plate efficiencies η from about 80 to 60 per cent (see Fig. 12-13).

With the choice of I_s, $e_{b,\min}$, $e_{c,\max}$, and θ_b specified, the other operating conditions are established. It is desired, therefore, to examine the relation that expresses the grid bias, E_{cc}, and also the grid conduction angle θ_c, in terms of the fixed parameters. To find an expression for E_{cc}, it is noted that the plate current becomes zero when $\omega t = \theta_b/2$. At this point, the grid signal is given by

$$e_g = E_{gm} \cos \omega t = E_{gm} \cos \frac{\theta_b}{2} \qquad (12\text{-}65)$$

But at this point it is necessary that $e_c + e_b/\mu = 0$. This follows from the fact that the plate current may be written by an expression of the form $i_b = f(e_c + e_b/\mu)$ and, for i_b to be zero, $e_c + e_b/\mu$ must be zero. By virtue of this

$$E_{gm} \cos \frac{\theta_b}{2} + E_{cc} + \frac{1}{\mu}\left(E_{bb} - E_{p1m} \cos \frac{\theta_b}{2}\right) = 0$$

But since

$$e_{c,\max} = E_{gm} + E_{cc}$$
$$e_{b\min.} = E_{bb} - E_{p1m}$$

it follows that

$$(e_{c,\max} - E_{cc}) \cos \frac{\theta_b}{2} + E_{cc} + \frac{1}{\mu}\left[E_{bb} - (E_{bb} - e_{b,\min}) \cos \frac{\theta_b}{2}\right] = 0$$

from which

$$E_{cc} = \frac{-E_{bb}}{\mu} + \left(e_{c,\max} + \frac{e_{b,\min}}{\mu}\right) \frac{\cos (\theta_b/2)}{\cos (\theta_b/2) - 1} \qquad (12\text{-}66)$$

The angle of grid flow is readily determined, since the grid current becomes zero when $\omega t = \theta_c/2$. At this point

$$e_c = E_{gm} \cos \frac{\theta_c}{2} + E_{cc} = 0$$

from which it follows that

$$\cos \frac{\theta_c}{2} = -\frac{E_{cc}}{E_{gm}} \qquad (12\text{-}67)$$

where E_{cc} is obtained from Eq. (12-66).

12-11. Grid Bias. The foregoing mathematical discussion assumed that the grid bias potential E_{cc} was constant in magnitude. Often, however, the bias potential is obtained by means of a resistor-capacitor combination in the grid line, in the manner illustrated in Fig. 12-15. The choice of grid resistance R_g is dictated by the required bias potential and the average grid current I_c. This is frequently referred to as grid-leak bias.

It might be thought that I_c would be a definite value for a given peak driving potential E_{gm}, with the result that the grid resistance would be firmly established. As a practical fact, variations of R_g are accompanied by an almost inverse variation of I_c, with the result that, for fixed E_{gm}, the potential E_{cc} remains sensibly constant. It is desirable, therefore, that the largest R_g possible be used, with stable amplifier operation. This follows from the fact that the loss in the grid resistor is due to the heating, or $I_c^2R_g$, loss. But for a given negative bias, I_cR_g is constant, and I_c varies inversely with R_g. Consequently, by increasing R_g, I_c is reduced, and the corresponding loss is reduced.

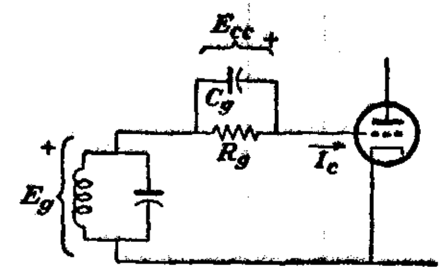

Fig. 12-15. The use of a grid resistor and grid capacitor for biasing the amplifier.

The grid driving power P_g is usually of the order of 5 to 10 per cent of the a-c power output of the amplifier P_L, when the tube is operated within its designed frequency limits. When operated above the normal frequency limits of the tube, the grid driving power increases rapidly, owing in some measure to increased dielectric losses, but principally because of transit-time loading. This latter factor is discussed at some length in Sec. 3-8. A limit is thereby set to the h-f limit of the tube.

12-12. Grid Potential and Amplifier Linearity. It has been noted on several occasions that the plate-circuit efficiency η_p depends upon the plate-current conduction angle θ_b. Moreover, the plate-current conduction angle depends upon the grid bias and the magnitude of the grid driving potential, more negative values of E_{cc} and higher E_{gm} being accompanied by smaller values of θ_b. The general character of the variation of output current and plate-circuit efficiency as a function of input grid potential is shown in Fig. 12-16.

These curves show that the a-c component of current I_{p1}, and the efficiency η_p, increase with increasing values of E_g over a wide range of E_g. A saturation value is reached beyond which there is no essential change, except that the grid current, and so the grid driving power, continue to increase. An interesting fact is that the situation remains roughly the same whether fixed bias or grid-leak bias is used. With grid-leak bias, however, the input power rises to larger values than with fixed bias. This is so because an increased E_g tends to result in a higher I_c; but this in turn causes an increase in E_{cc}. Hence, for a given output power, a larger E_{gm} is required, with a correspondingly less linear relationship between E_g and I_{p1}. Clearly, overdriving the amplifier merely results in high power dissipation in the grid circuit. Underdriving leads to a reduced amplifier output and efficiency.

The question of a linear relation between I_{p1} and E_s is of considerable importance when the tube is used for grid-circuit amplitude modulation. This matter will be discussed in Chap. 17.

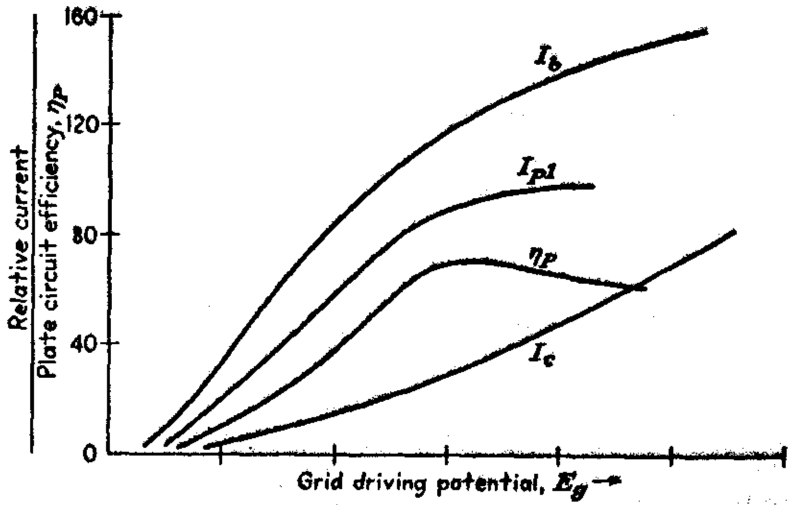

FIG. 12-16. The effect of varying the input grid potential on several of the important amplifier factors.