Chapter 1

INTRODUCTION TO THE RF POWER AMPLIFIERS DESIGN

1. BASIC TERMS AND DEFINITIONS

The general equivalent circuit of power amplifier is shown in Fig. 1-1. It consists of active device (AD), input and output networks and supply and bias circuits[1,2].

Electrical operation mode of power amplifier can be characterized by the following fundamental parameters: the first harmonic output power P_1, the dc supply power P_{DC}, the efficiency regarding the first harmonic $\eta = P_1 / P_{DC}$ (so-called electronic efficiency), the power gain K_P, the pass-band or amplitude-frequency characteristic, and the nonlinear distortions values. The power-added efficiency (PAE) is also important characteristic accounting the driving signal power.

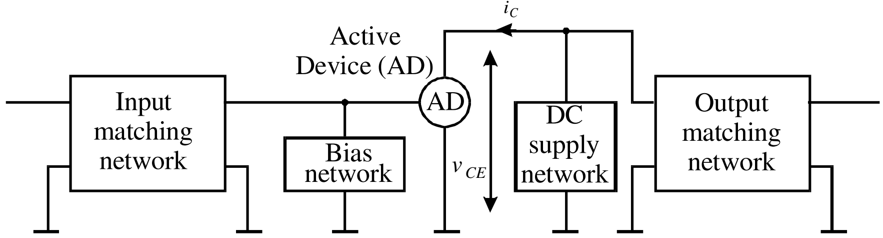


Figure 1-1. General equivalent circuit of power amplifier.

The driving signal source and supply voltage source parameters substantially influence the power amplifier operation. These external parameters are the following:

- nominal power P_{in}, frequency f and intrinsic impedance Z_i of the driving signal source;
- voltage E_C of supply voltage source, and the load impedance Z_L at the transistor output terminals.

The dependencies of the amplifier parameters on the external parameters present the sets of characteristics as follows:

- load: dependencies of P_1, P_{DC}, η, K_P on the load impedance Z_L;
- amplitude: dependencies of P_1, P_{DC}, η, K_P on the input power P_{in};
- modulation: dependencies of P_1, P_{DC}, η, K_P on the supply voltage E_C;
- frequency: dependencies of P_1, P_{DC}, η, K_P on the driving signal frequency.

The approximate views of the load, amplitude, modulation, and frequency characteristics' sets are shown in Figs. 1-2 - 1-5, respectively. The $f_1 < f_2 < f_3$ is assumed for all of the figures.

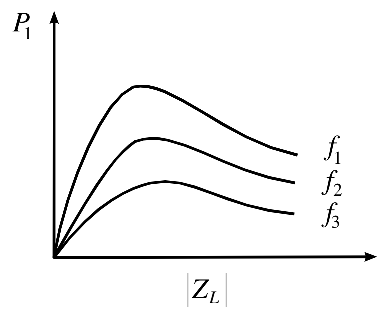


Figure 1-2. Load characteristics' set.

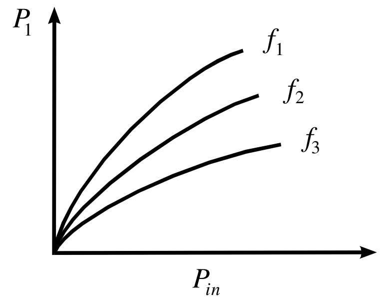


Figure 1-3. Amplitude characteristics' set.

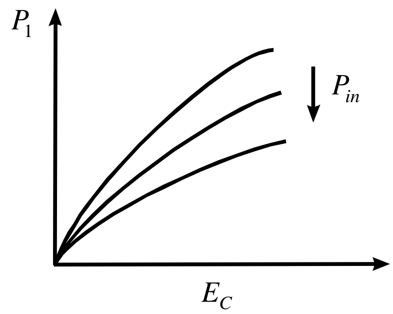


Figure 1-4. Modulation characteristics' set.

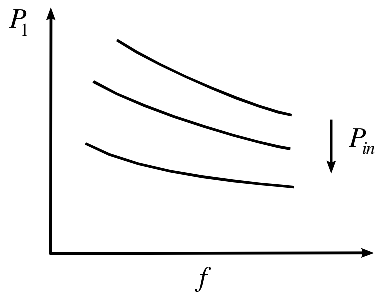


Figure 1-5. Frequency characteristics' set.

There are other characteristics' sets besides the above mentioned: for example, the bias characteristics, etc. However, the above four sets are of special importance.

The nonlinear power amplifier can operate in one of the following modes: undersaturated, critical, overloaded, switching. The latter two belong to the saturated one.

The amplifier's operating mode can be determined by the dynamic load line, which represents the operating point coordinates at the current-voltage curves' plane. The mode is undersaturated, if the dynamic load line stays within the active and cut-off regions as shown in Fig. 1-6. However, when it moves to the saturation region, the operating mode becomes overloaded. The critical mode is the boundary between the undersaturated and overloaded ones. In this case, the operating point is just touching the saturation line.

The switching mode assumes the transistor as an ideal switch that can be either in the saturation region, or in cut-off. This mode can be considered as an extreme case of overloaded one.

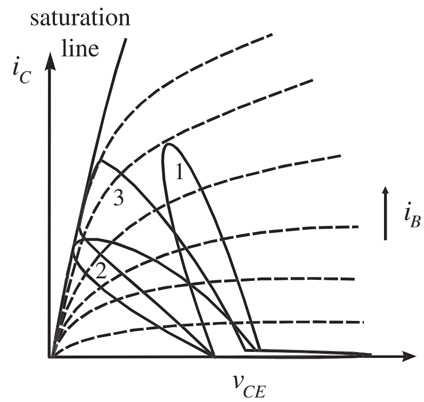


Figure 1-6. Dynamic load lines for the undersaturated (1), critical (2), and overloaded (3) operating modes.

The power amplifiers are divided into several classes within two major groups depending on the input signal amplitude, bias and supply conditions, and the properties of the input and output matching networks: 1) the sine-wave output operation, and 2) the polyharmonic operation. The detailed description of these classes is given in the further paragraphs.

2. SINE-WAVE OUTPUT OPERATION: CLASSES A, AB, B, C

There are two cases for the sine-wave output operation power amplifiers as shown in Fig. 1-7: 1) the small signal class A - linear mode, and 2) the large signal modes with cut-off.

The magnitude of RF signal is much smaller than its dc component for the linear mode (see Fig. 1-7 (a)). Typical application of such amplifiers is the input low-power stages of RF transmitters. Maximal theoretically reachable efficiency is equal 50% for this case.

The large-signal operation modes (see Fig. 1-7 (b)) give the advantage of higher efficiency. Here, the collector current is zero during the cut-off interval, so the instant parasitic power dissipation becomes zero for this region leading to the efficiency increasing.

The large signal operation modes are divided into different classes according to the conduction angle as follows: class AB with $90° < \theta < 180°$, class B with $\theta = 90°$, and class C with $\theta < 90°$.

The analytical expression for the large-signal operation collector current i_C is following:

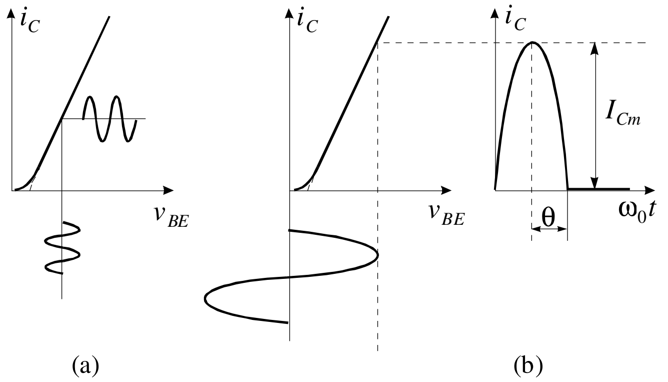


Figure 1-7. Small signal class A operation - linear mode (a), and large signal nonlinear modes with cut-off (b).

$$\left.\begin{array}{r} i_C = I_{Cm}\dfrac{\cos\omega t - \cos\theta}{1-\cos\theta}, \text{for} -\theta < \omega t < \theta \\ i_C = 0, \text{for } \theta < \omega t < 2\pi - \theta \end{array}\right\}, \tag{1-1}$$

where I_{Cm} is the maximal collector current value, θ is the conduction angle, which is equal to the half of nonzero current interval, and varies from 0° to 180°.

The dependencies of the Fourier coefficients α_0 and α_1 of the collector current dc component $I_{C0} = I_{Cm}\alpha_0$ and fundamental frequency component $I_{C1} = I_{Cm}\alpha_1$ on θ are the following:

$$\alpha_0 = \frac{\sin\theta - \theta\cos\theta}{\pi(1-\cos\theta)}, \quad \alpha_1 = \frac{\theta - \sin\theta\cos\theta}{\pi(1-\cos\theta)} \tag{1-2}$$

In case of higher harmonics, the appropriate Fourier coefficients should be determined as:

$$\alpha_n = \frac{2}{\pi}\frac{\sin n\theta\cos\theta - n\cos n\theta\sin\theta}{n(n^2-1)(1-\cos\theta)}, \quad n \geq 2$$

The α_0 - α_3 coefficients are shown in Fig. 1-8.

Assuming the ideal matching and harmonic output voltage, the collector efficiency can be written as:

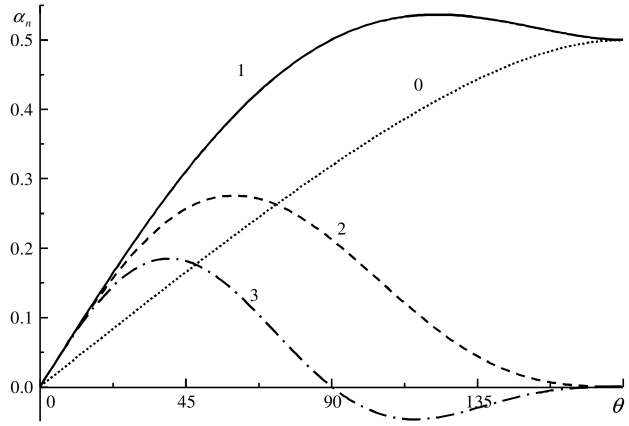


Figure 1-8. The dependencies of the Fourier coefficients α_0 - α_3 on θ.

$$\eta = \frac{P_1}{P_{DC}} = \frac{1}{2}\frac{I_{C1}V_{C1}}{I_{C0}E_C} = \frac{1}{2}\xi\frac{\alpha_1}{\alpha_0}, \qquad (1\text{-}3)$$

where the $\xi = V_{C1}/E_C$ is the transistor utilization factor, $\xi \le 1$. As it is follows from Eqs. (1-2) and (1-3), the efficiency increases when the conduction angle decreases. Furthermore, the efficiency approaches 100% for the $\theta \to 0$ and $\xi = 1$.

As the collector current becomes almost zero during the cut-off interval, the dissipated power is smaller for the decreased conduction angle. However, the fundamental component power goes down dramatically for a low θ. That is why the class B or class C with θ greater than 60° are usually used. It allows obtaining $\eta = 70..80\%$ and acceptable output power level[3-7].

3. POLYHARMONIC OPERATION: CLASS F

The polyharmonic operation is characterized by the complex output voltage waveform[8], which contains the series of harmonics. In general case, the output voltage includes infinite number of harmonics:

$$v_C = E_C - \sum_{n=1}^{\infty} V_{Cn}\cos(n\omega t + \varphi_n),$$

However, only few first harmonics are considered practically.

The simplest particular case of polyharmonic operation is so-called biharmonic[9]. Here, an output voltage contains the second or the third harmonic besides of dc and fundamental components.

The advantages of higher harmonic tuning were shown by Tyler in 1958[10] and Snider in 1967[11]. The short-circuit mode for the even harmonics and the open-circuit mode for the odd harmonics were realized at the transistor output.

For the first time, F.H. Raab[12] proposed the term "class-F" for such tuning. It is worthy to note that in this case output current contains the only even harmonics, while output voltage contains the only odd harmonics, or vice versa. In other words, either current, or voltage, but not both simultaneously, has non-zero value for a given higher harmonic, so the power dissipation is omitted.

The output network of class-F power amplifier is usually presented as the multiresonator harmonic filter. The equivalent circuit of third harmonic peaking network is shown in Fig. 1-9. This network and its varieties were considered in details by Trask[13].

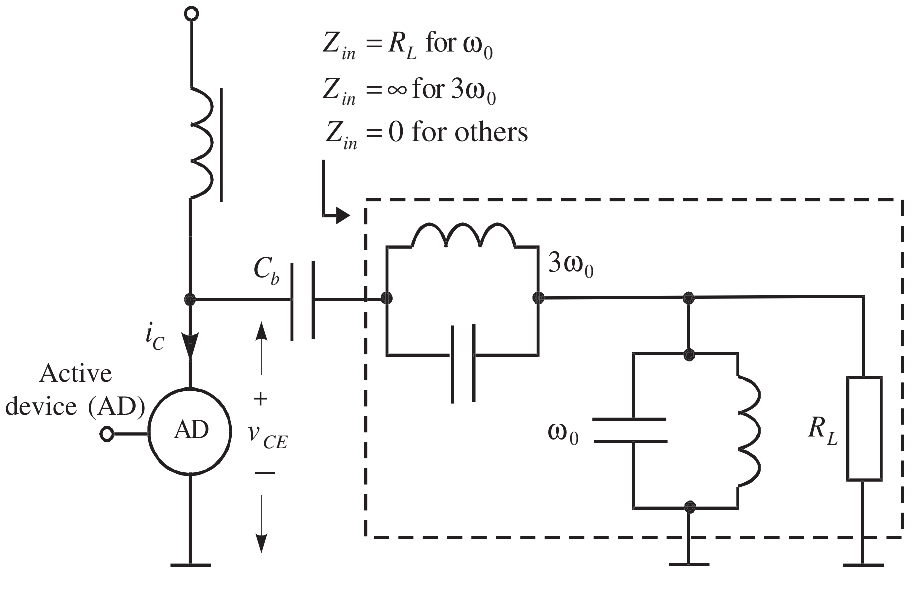


Figure 1-9. Equivalent circuit of third-harmonic peaking class F amplifier output network.

3.1 The maximally flat output current and voltage waveforms

The ideal current and voltage waveforms are shown in Fig. 1-10. They correspond to the account of infinite number of harmonics. However, tuning of output network for such case is almost impossible practically. So, usually the only few first harmonics are under consideration, as much energy efficient.

In order to choose between the circuit complexity and efficiency increasing, one have to have an idea about influence of number of harmonics taken into account on the power amplifier characteristics. Such analysis was conducted by F.H. Raab[14]. He supposed, that output voltage contains the only odd higher harmonics, while output current contains the only even higher harmonics:

$$v_C = E_C + V_{C1}\sin\omega_0 t + V_{C3}\sin 3\omega_0 t + V_{C5}\sin 5\omega_0 t + \ldots$$

$$i_C = I_{C0} - I_{C1}\sin\omega_0 t - I_{C2}\cos 2\omega_0 t - I_{C4}\cos 4\omega_0 t + \ldots,$$

The following relations were defined for voltage:

$$\gamma_V = \frac{V_{C1}}{E_C}, \quad \delta_V = \frac{V_{C\max}}{E_C}, \tag{1-4}$$

and for current

$$\gamma_I = \frac{I_{C1}}{I_{C0}}, \quad \delta_I = \frac{i_{C\max}}{I_{C0}}. \tag{1-5}$$

Here $v_{C\max}$ and $i_{C\max}$ are maximal values of voltage and current pulses, respectively. The several assumptions were made: the output network is lossless and the only fundamental frequency component can reach the load; the active device is ideal current source or ideal switch.

Raab called such the approximations of ideal current and voltage waveforms as "maximally flat". The parameters m and n define the numbers of even and odd harmonics taken into consideration, respectively.

The parameters δ_V, δ_I, γ_V, γ_I, and relations of higher harmonics to the dc component for the maximally flat waveforms are presented in Tables 1-1 and 1-2.

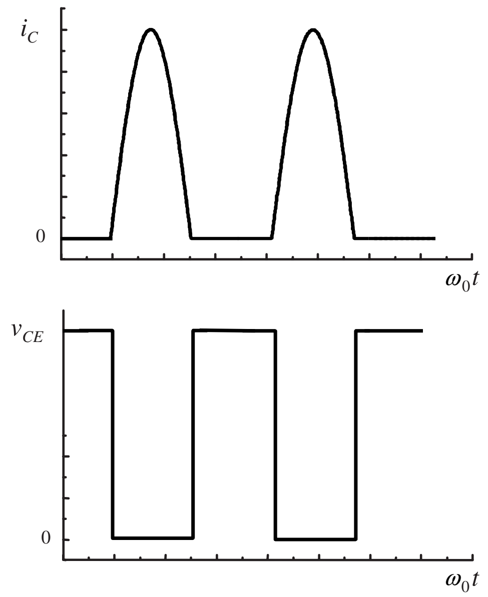


Figure 1-10. The ideal waveforms of output current and voltage for a class F power amplifier.

The collector efficiency of power amplifier with the maximally flat waveforms of current and voltage can be written using Eqs. (1-4) and (1-5) as follows:

$$\eta = \frac{\gamma_V \gamma_I}{2} \tag{1-6}$$

The efficiency values for the different combinations of m and n are summarized in Table 1-3. As one can see from the table, the increasing of number of voltage or current harmonics leads to the efficiency growing.

Table 1-1. The voltage maximally flat waveform parameters[14]

n	δ_V	$\gamma_V = V_{C1}/E_C$	V_{C3}/E_C	V_{C5}/E_C
1	2	1	–	–
3	2	9/8 = 1.125	1/8 = 0.125	–
5	2	75/64 = 1.172	25/128 = 0.195	3/128 = 0.023
∞	2	$4/\pi$ = 1.273	$4/(3\pi)$ = 0.424	$4/(5\pi)$ = 0.255

Table 1-2. The current maximally flat waveform parameters[14]

m	δ_I	$\gamma_I = I_{C1}/I_{C0}$	I_{C2}/I_{C0}	I_{C4}/I_{C0}
1	2	1	–	–
2	8/3 = 2.667	4/3 = 1.333	1/3 = 0.333	–
4	128/45 = 2.844	64/45 = 1.422	16/45 = 0.356	1/45 = 0.022
∞	π = 3.142	$\pi/2$ = 1.571	2/3 = 0.667	2/15 = 0.133

Table 1-3. Maximally flat waveform power amplifier efficiency[14]

m	η ($n = 1$)	η ($n = 3$)	η ($n = 5$)	η ($n = \infty$)
1	1/2 = 0.500, class A	9/16 = 0.563	75/128 = 0.586	$2/\pi$ = 0.637
2	2/3 = 0.667	3/4 = 0.750	25/32 = 0.781	$8/(3\pi)$ = 0.849
4	32/45 = 0.711	4/5 = 0.800	5/6 = 0.833	$128/(45\pi)$ = 0.905
∞	$\pi/4$ = 0.785, class B	$9\pi/32$ = 0.884	$75\pi/256$ = 0.920	1, class F

Herewith, the faster growing is appropriate to the simultaneous increasing of both current and voltage numbers of harmonics.

However, the above analysis did not pay any attention to the harmonic generation mechanism. In addition, it was supposed that the fundamental component and the higher harmonics are in the proper phase relation. As will be shown in the Chapters 2 and 3, this assumption can lead to the poor efficiency.

3.2 Phase relations between the first and higher harmonics

The first and the third harmonics of transistor output voltage should be out-of-phased in order to obtain its flat waveform[15,16]. In this case, the third harmonic decreases the peak value of voltage impulse and leads to its flattening while is taken in the proper magnitude relation to the fundamental component as shown in Fig. 1-11 (a). These effects allow to increase the output power capability, to decrease the power dissipation losses in the transistor, and therefore lead to the higher efficiency.

In case of incorrect phase tuning, one can obtain converse situation: the peak voltage value and the dissipation in transistor power increasing followed by efficiency decreasing (see Fig. 1-11 (b)).

The conditions of formation the out-of-phase the first and the third harmonics were considered by Colantonio et al[15,16].

Usually, the transistor of single-stage class-F power amplifier operates in the critical or slightly overloaded mode. So, the transistor output voltage is completely defined by the transistor output current and the input impedance of the output network. Therefore, for each of harmonics one can write:

$$V_{Cn} = I_{Cn} Z_{Ln} .$$

The input impedance Z_{Ln} was assumed pure resistive, and the truncated sine-wave approximation for the collector current similar to (1.1) was used under the Colantonio's et al analysis[15,16].

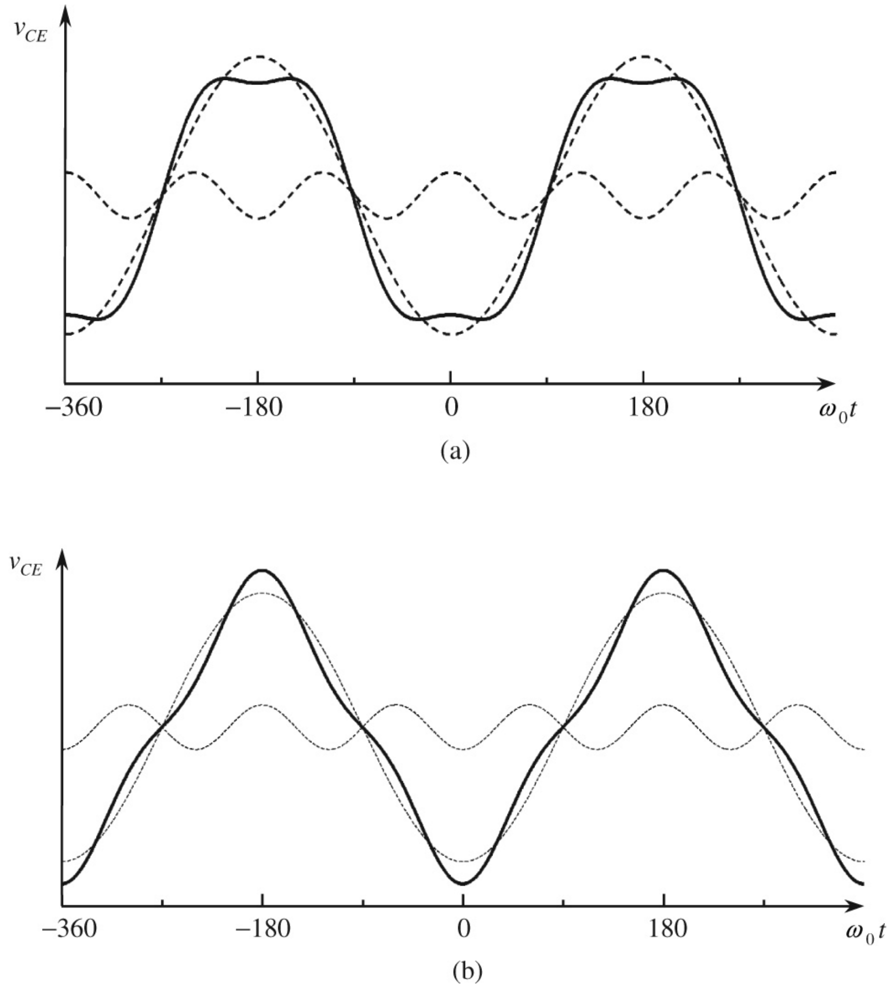


Figure 1-11. The phase relations between the first and the third harmonics: out-of-phase (a) and in-phase (b).

The first and the third harmonics are out-of-phase for above conditions if their Fourier coefficient α_n are opposite. As one can see from the Fig. 1-8, it is possible for the conduction angles $\theta > 90°$.

3.2.1 Optimum input impedance

In case of third-harmonic peaking class-F power amplifier, the transistor output voltage consists of the first, the third harmonics, and the dc-component as follows:

$$v_C = E_C - V_{C1}\cos\omega_0 t - V_{C3}\cos 3\omega_0 t$$
$$= E_C - V_{C1}\left(\cos\omega_0 t + \frac{1}{\varepsilon_3}\cos 3\omega_0 t\right) \tag{1-7}$$

where

$$\varepsilon_3 = \frac{V_{C1}}{V_{C3}} = \frac{R_{\omega 0}}{R_{3\omega 0}} \cdot \frac{I_{C1}}{I_{C3}}\,. \tag{1-8}$$

The ε_3 is defined by the relation of the first harmonic complex amplitude to the third harmonic one[15,16]. It has to be negative in order to realize voltage impulse flattening (See Fig. 1-8).

The first harmonic magnitude is greater than half of the signal swing for the signal with out-of-phase the first and the third harmonics. It allows increasing the output power without overload. The growing of the first harmonic magnitude can be expressed as:

$$V_{C1,F} = \gamma(\varepsilon_3) V_{C1,\sin}\,, \tag{1-9}$$

where the $\gamma(\varepsilon_3)$ is the class-F first harmonic growing coefficient defined by the relation of the first harmonic magnitude to the magnitude of overall voltage swing as follows:

$$\gamma(\varepsilon_3) = \frac{V_{C1}}{|v(\theta_m) - E_C|} = \frac{1}{\cos\theta_m + (1/\varepsilon_3)\cos 3\theta_m}\,.$$

Here, the θ_m is the point of minimal or maximal value of the voltage impulse v_C, which can be determined from (1-7).

In order to find θ_m, the derivative of v_C on $\omega_0 t = \theta$ should be equated to zero. Then, the obtained equation should be solved. The derivative of v_C can be written as follows:

$$v_C' = -V_{C1}(-\sin\theta - 3/\varepsilon_3 \sin 3\theta)\,. \tag{1-10}$$

The points of possible maximums of v_C within the $0 \le \theta \le \pi$ region can be found by equating of Eq. (1-10) to zero:

$$\theta_{m1} = 0\,, \tag{1-11}$$

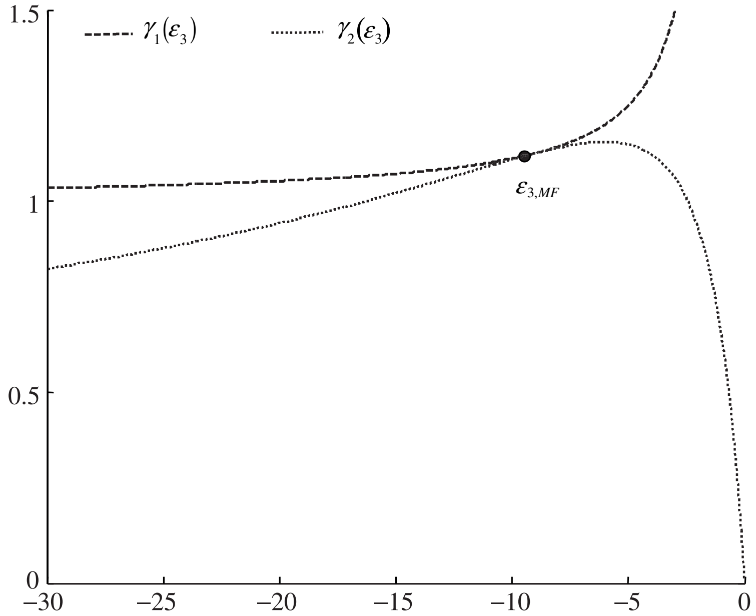


Figure 1-12. The γ_1 (ε_3) and γ_2 (ε_3) functions.

$$\theta_{m2} = \arccos\left(\frac{\sqrt{3-\varepsilon_3}}{2\sqrt{3}}\right). \tag{1-12}$$

The following expressions for the $\gamma_1(\varepsilon_3)$ and $\gamma_2(\varepsilon_3)$ functions are correspond to the θ_{m1} and θ_{m2}, respectively:

$$\gamma_1(\varepsilon_3) = \frac{1}{\cos\theta_{m1} + (1/\varepsilon_3)\cos 3\theta_{m1}} = \frac{\varepsilon_3}{1+\varepsilon_3}, \tag{1-13}$$

$$\gamma_2(\varepsilon_3) = \frac{1}{\cos\theta_{m2} + (1/\varepsilon_3)\cos 3\theta_{m2}} = -\frac{3\sqrt{3}\varepsilon_3}{(3-\varepsilon_3)^{3/2}} \tag{1-14}$$

The $\gamma_1(\varepsilon_3)$ and $\gamma_2(\varepsilon_3)$ functions are shown in Fig. 1-12.

The point of contingence of $\gamma_1(\varepsilon_3)$ and $\gamma_2(\varepsilon_3)$ functions is appropriate to the maximally flat waveform of voltage v_C. The absciss $\varepsilon_{3,MF}$ of contingence point can be determined by equating the right parts of Eqs. (1-13) and (1-14):

$$\frac{\varepsilon_{3,MF}}{1+\varepsilon_{3,MF}}=-\frac{3\sqrt{3}\varepsilon_{3,MF}}{\left(3-\varepsilon_{3,MF}\right)^{3/2}},$$

$$\varepsilon_{3,MF}=-9. \tag{1-15}$$

The waveform of voltage v_C, that is appropriate to $\varepsilon_{3,MF}$, is shown in Fig. 1-13.

The following function value $\gamma\left(\varepsilon_{3,MF}\right)$ is appropriate to the absciss $\varepsilon_{3,MF}=-9$:

$$\gamma\left(\varepsilon_{3,MF}\right)=\frac{\varepsilon_{3,MF}}{1+\varepsilon_{3,MF}}=-\frac{3\sqrt{3}\varepsilon_{3,MF}}{\left(3-\varepsilon_{3,MF}\right)^{3/2}}=\frac{9}{8}=1.125. \tag{1-16}$$

For the $-9<\varepsilon_3<0$ region, the maximal value of v_C within the $0\le\theta\le\pi$ interval can be reached for $\theta_m=\theta_{m2}$ point. Thereat, function $\gamma(\varepsilon_3)$ is the same as $\gamma_2(\varepsilon_3)$:

$$\gamma(\varepsilon_3)=\gamma_2(\varepsilon_3)=-\frac{3\sqrt{3}\varepsilon_3}{\left(3-\varepsilon_3\right)^{3/2}}, \text{ для } -9<\varepsilon_3<0. \tag{1-17}$$

The waveform of voltage v_C, that is appropriate to $\varepsilon_3=-3$, is shown in Fig. 1-14.

For the $\varepsilon_3\le-9$ region, the maximal value of v_C within the $0\le\theta\le\pi$ interval can be reached for $\theta_m=\theta_{m1}$ point. Thereat, function $\gamma(\varepsilon_3)$ is the same as $\gamma_1(\varepsilon_3)$:

$$\gamma(\varepsilon_3)=\gamma_1(\varepsilon_3)=\frac{\varepsilon_3}{1+\varepsilon_3}, \text{ для } \varepsilon_3\le-9. \tag{1-18}$$

The waveform of voltage v_C, that is appropriate to $\varepsilon_3=-15$, is shown in Fig. 1-15.

For $\varepsilon_3\to-\infty$, $\gamma(\varepsilon_3)$ is aspire to unity as follows:

$$\lim_{\varepsilon_3\to-\infty}\gamma(\varepsilon_3)=\lim_{\varepsilon_3\to-\infty}\gamma_1(\varepsilon_3)=\lim_{\varepsilon_3\to-\infty}\frac{\varepsilon_3}{1+\varepsilon_3}=1. \tag{1-19}$$

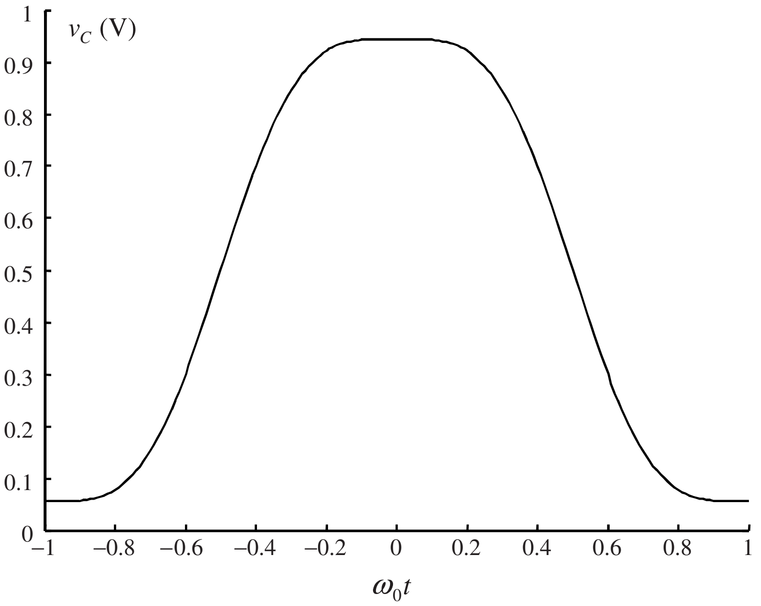


Figure 1-13. Maximally flat waveform of voltage impulse v_C, that is appropriate to $\varepsilon_{3,MF}$.

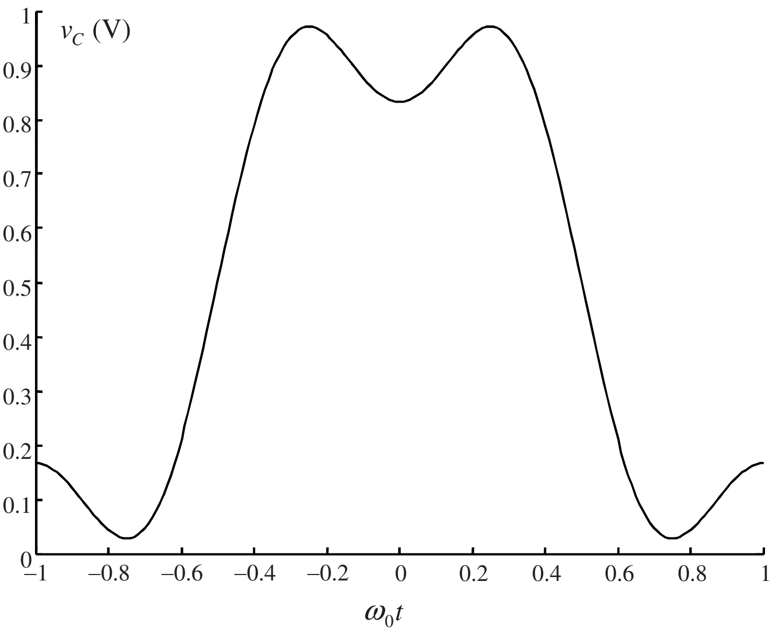


Figure 1-14. The waveform of voltage v_C, that is appropriate to $\varepsilon_3 = -3$.

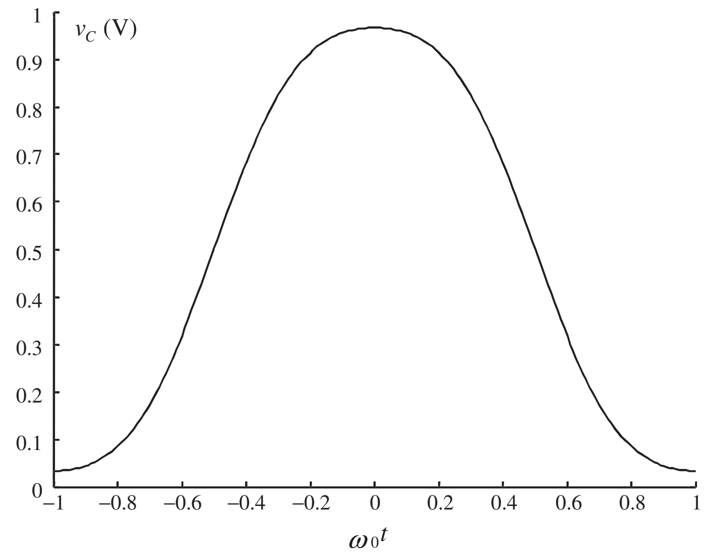


Figure 1-15. The waveform of voltage v_C, that is appropriate to $\varepsilon_3 = -15$.

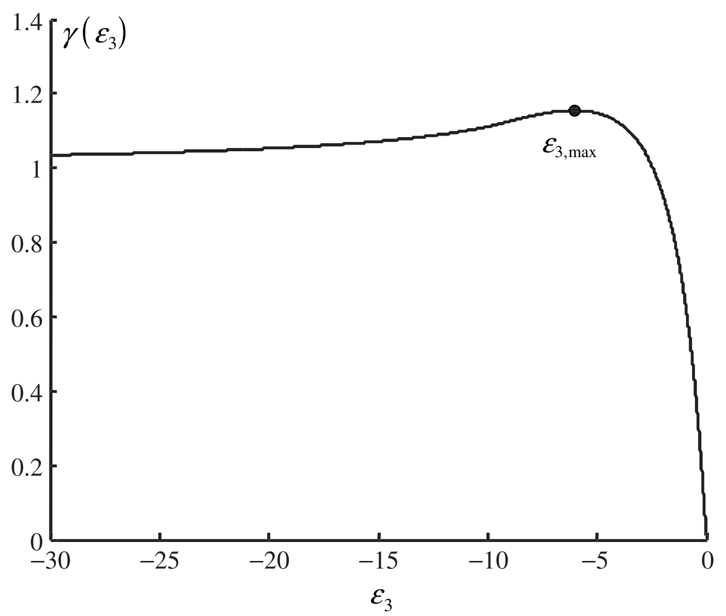


Figure 1-16. Function $\gamma(\varepsilon_3)$ for negative ε_3 values.

In this case, the third harmonic magnitude is aspire to zero, and the voltage impulse waveform is the harmonic function.

Therefore, the function $\gamma(\varepsilon_3)$ can be defined by the following set for negative ε_3 values according to the Eqs. (1-17) and (1-18):

$$\gamma(\varepsilon_3) = \begin{cases} -\dfrac{3\sqrt{3}\varepsilon_3}{(3-\varepsilon_3)^{3/2}}, \text{for} -9 < \varepsilon_3 < 0 \\ \dfrac{\varepsilon_3}{1+\varepsilon_3}, \qquad \text{for } \varepsilon_3 \le -9 \end{cases} \tag{1-20}$$

Function $\gamma(\varepsilon_3)$ for negative ε_3 values is shown in Fig. 1-16. As can be seen from Fig. 1-16, the function $\gamma(\varepsilon_3)$ reaches the maximal value within the $-9 < \varepsilon_3 < 0$ interval.

The absciss $\varepsilon_{3,\max}$ of maximum point can be found by equating the derivative of $\gamma(\varepsilon_3)$ to zero. Then, the obtained equation should be solved. The derivative of $\gamma(\varepsilon_3)$ at the $-9 < \varepsilon_3 < 0$ region can be written as follows:

$$\gamma'(\varepsilon_3) = \left(-\frac{3\sqrt{3}\varepsilon_3}{(3-\varepsilon_3)^{3/2}}\right)' = -\frac{3\sqrt{3}}{(3-\varepsilon_3)^{3/2}} - \frac{9\sqrt{3}\varepsilon_3}{2(3-\varepsilon_3)^{5/2}}. \tag{1-21}$$

The following value can be obtained by equating of Eq. (1-21) to zero:

$$-\frac{3\sqrt{3}}{(3-\varepsilon_{3,\max})^{3/2}} - \frac{9\sqrt{3}\varepsilon_{3,\max}}{2(3-\varepsilon_{3,\max})^{5/2}} = 0,$$

$$\varepsilon_{3,\max} = -6. \tag{1-22}$$

The waveform of voltage v_C, that is appropriate to $\varepsilon_{3,\max}$, is shown in Fig. 1-17.

The following function value $\gamma(\varepsilon_{3,\max})$ is appropriate to the absciss $\varepsilon_{3,\max} = -6$:

$$\gamma(\varepsilon_{3,\max}) = -\frac{3\sqrt{3}\varepsilon_{3,\max}}{(3-\varepsilon_{3,\max})^{3/2}} = \frac{2}{\sqrt{3}} \approx 1.1547. \tag{1-23}$$

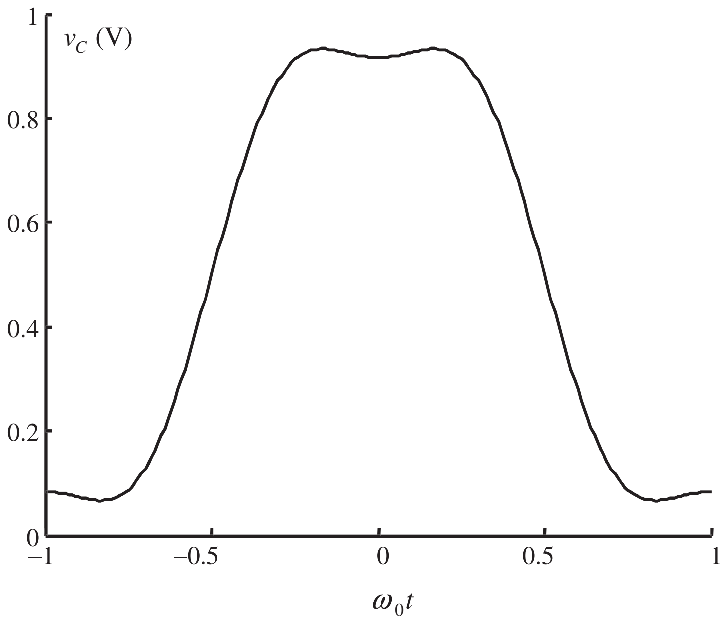


Figure 1-17. The waveform of voltage v_C, that is appropriate to $\varepsilon_{3,\,\max}$.

The first harmonic of transistor output current impulse is assumed the same for various tunings of the third harmonic in case of given input signal. Therefore, the input impedance of output network should vary in order to increase the first harmonic of transistor output voltage as follows:

$$R_{\omega 0,F} = \gamma(\varepsilon_3) R_{\omega 0,\,\sin} ,$$

where the $R_{\omega 0,\sin}$ is the input impedance of sine-wave operation power amplifier output network.

The choice of ε_3 defines the optimum load for the third harmonic:

$$R_{3\omega 0,F} = \frac{R_{\omega 0,F}}{\varepsilon_3} \frac{I_{C1}}{I_{C3}} . \tag{1-24}$$

Equation (1-24) is substantially distinct from the classical class-F condition of odd harmonics open-circuit mode. It can be explained by the fact that in practice the finite number of higher harmonics is usually taken into account.

4. POWER AMPLIFIERS' MATCHING NETWORKS

4.1 The main purposes of power amplifiers' matching networks

The electrical networks play the various roles in the transistor power amplifiers. These are: driving signal source matching with the transistor input; load matching with the transistor output; cross-product suppression in the output signal spectrum; formation of given impedance for higher harmonics at the transistor output; dividing and combining signals from the several sources.

4.1.1 Driving signal source matching with the transistor input

Suppose that the transistor input current and voltage first harmonic magnitudes are known as well as the phase shift between them. Then the equivalent impedance can be considered instead of the transistor. Its value should be selected in such a way, that the input current, voltage and their phase shift remain the same. This impedance is the input impedance of the transistor.

An input impedance of power RF transistors is a nonlinearly varying complex value, which depends on an input power level and frequency, bias conditions, parameters of the transistor itself, etc. Usually, it differs significantly from the optimum value, which allows the maximum power transfer from the driving source to the transistor. It is useful to have an amplifier input impedance equal to the standard 50Ohm, or 75Ohm. However, a typical input impedance value of a power RF transistor is just a few ohms. Therefore, the low-resistance transistor input impedance should be matched to the higher value in order to provide better power transfer.

The general look at the matching problem is the following. The two-port network is used between the driving source and the load. Here, the load Z_L means the transistor input impedance. In case of voltage driving source E_i with intrinsic impedance Z_i shown in Fig. 1-18(a) is used, the Z parameters of two-port are most convenient. The matching conditions can be written as:

$$Z_{in} = Z_i^*, \text{ and } Z_{out} = Z_L^*, \tag{1-25}$$

where asterisk means the complex conjugate; Z_{in} is an input impedance of two-port calculated at the terminals 1 and 2; Z_{out} is an output impedance of two-port calculated at the terminals 3 and 4. The Z_{in} and Z_{out} are expressed through the Z parameters as follows:

$$Z_{in} = Z_{11} - \frac{Z_{12}Z_{21}}{Z_{22} + Z_L}, \text{ and } Z_{out} = Z_{22} - \frac{Z_{12}Z_{21}}{Z_{11} + Z_i}.$$

Alternatively, in case of current driving source I_i with intrinsic admittance Y_i shown in Fig. 1-18(b) is used, the Y parameters of two-port are useful. It gives the following matching conditions:

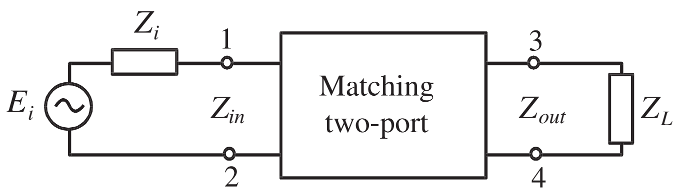
$$Y_{in} = Y_i^*, \text{ and } Y_{out} = Y_L^*, \tag{1-26}$$

where Y_L is a transistor input admittance; Y_{in} is an input admittance of two-port at the terminals 1 and 2; Y_{out} is an output admittance of two-port at the terminals 3 and 4. The Y_{in} and Y_{out} can be written as follows:

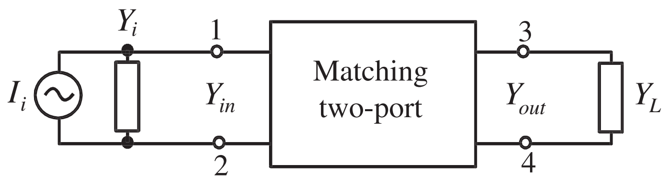
$$Y_{in} = Y_{11} - \frac{Y_{12}Y_{21}}{Y_{22} + Y_L}, \text{ and } Y_{out} = Y_{22} - \frac{Y_{12}Y_{21}}{Y_{11} + Y_i}.$$

In case of a lossless reversible reactive two-port, the Z and Y parameters can be represented as follows:

$$Z_{11} = jx_{11}; \quad Z_{12} = Z_{21} = jx_{12}; \quad Z_{22} = jx_{22};$$



(a)



(b)

Figure 1-18. General matching problem for the driving voltage (a) and current (b) sources.

$$Y_{11} = jb_{11}; \quad Y_{12} = Y_{21} = jb_{12}; \quad Y_{22} = jb_{22};$$

where the x_{11}, x_{12}, and x_{22} are the two-port reactive self-impedances; the b_{11}, b_{12}, and b_{22} are the two-port reactive self-admittances. Using these expressions, the two-port input and output impedance and admittance can be rewritten as:

$$Z_{in} = jx_{11} + \frac{x_{12}^2}{jx_{22} + Z_L}, \text{ and } Z_{out} = jx_{22} + \frac{x_{12}^2}{jx_{11} + Z_i}; \tag{1-27}$$

$$Y_{in} = jb_{11} + \frac{b_{12}^2}{jb_{22} + Y_L}, \text{ and } Y_{out} = jb_{22} + \frac{b_{12}^2}{jb_{11} + Y_i}. \tag{1-28}$$

The only one condition from Eq. (1-25) or from Eq. (1-26) is required for the matching by using the reactive reverse two-port, i.e.:

$$Z_{in} = Z_i^*, \text{ or } Y_{in} = Y_i^* \tag{1-29}$$

Matching by lossless two-port allows to provide the maximum power transfer. For this case, the power that reaches the load is the following:

$$P_{\max} = \frac{E_i^2}{8\operatorname{Re}(Z_i)}, \text{ or } P_{\max} = \frac{I_i^2}{8\operatorname{Re}(Y_i)},$$

where $\operatorname{Re}(Z_i)$ and $\operatorname{Re}(Y_i)$ are the real components of intrinsic impedance and admittance, respectively.

The ideal matching is possible at the single frequency only. The simple three-elements T -shape or Π -shape circuits can be used.

The wideband matching is a substantially difficult issue due to the theoretical limitations. The greater bandwidth gives the inferior matching. Therefore, the ladder-type filters should be used in order to increase the matching efficiency.

4.1.2 Load matching with the transistor output

The transistor output can be represented as the equivalent current source I_{C1} with known first harmonic dynamic load line. It is assumed that the current source parameters are independent of the load value.

The condition on obtaining the maximum power in the load can be found by testing the $P_1 = U_1 I_1/2$ dependence on maxima and minima. This condition is the following:

$$\frac{U_1}{I_1} = -\frac{dU_1}{dI_1}, \text{ or } R_{L1} = \left|R_{id1}\right|, \tag{1-30}$$

where $R_{id1} = dU_1/dI_1$ is intrinsic differential resistance of the current source I_{C1}.

Therefore, in order to match the transistor output and load for the given frequency and operating mode, the lossless reactive two-port can be used as described in paragraph 4.1.1. In this case, Z_i should be selected equal to the R_{L1}, while Z_L means the actual load value, that can be standard 50Ohm or 75Ohm.

4.1.3 Cross-product suppression in the output signal spectrum

The nonlinear behavior of transistor power amplifier is the cause of cross-product components in the output signal spectrum. These components are the harmonics and sub-harmonics of valid signal, and intermodulation distortions.

Matching conditions and cross-product suppression conditions can give the contradictory requirements on output electrical network. In this case, two different networks should be synthesized: one is for matching, and the other is for cross-product suppression.

4.1.4 Formation of given impedance for higher harmonics at the transistor output

The amplifier output network should provide the appropriate input impedance for the higher harmonics in order to obtain certain current and voltage waveforms. For class-F amplifier example, the output network provides short-circuit and open-circuit behaviors for even and odd harmonics, respectively.

4.1.5 Dividing and combining signals from several sources

For some applications, the generated power needs to be divided or combined. If a single amplifier cannot provide the required power level, several amplifiers can be used in the work on the single load. This can be realized with specially tuned electrical circuits.

4.2 Narrow-band power amplifiers' matching circuits

As it was mentioned above, the simple reactive three-elements Π-shape or T-shape electric circuits can be used for matching at the certain frequency.

The equations for Z and Y parameters of the Π-shape circuit shown in Fig. 1-19 are the following:

$$z_{11} = jx_{11} = \frac{j(b_2 + b_3)}{\Delta_y}; \; z_{12} = jx_{12} = \frac{jb_3}{\Delta_y};$$

$$z_{21} = jx_{21} = \frac{jb_3}{\Delta_y}; \; z_{22} = jx_{22} = \frac{j(b_1 + b_3)}{\Delta_y};$$

$$\Delta_y = y_1 y_2 + y_2 y_3 + y_1 y_3 = -(b_1 b_2 + b_2 b_3 + b_1 b_3);$$

$$y_{11} = jb_{11} = j(b_1 + b_3); \; y_{12} = jb_{12} = -jb_3;$$

$$y_{21} = jb_{21} = -jb_3; \; y_{22} = jb_{22} = j(b_2 + b_3).$$

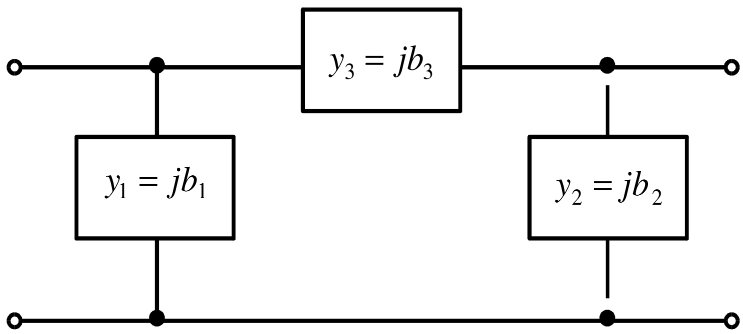


Figure 1-19. Π-shape circuit.

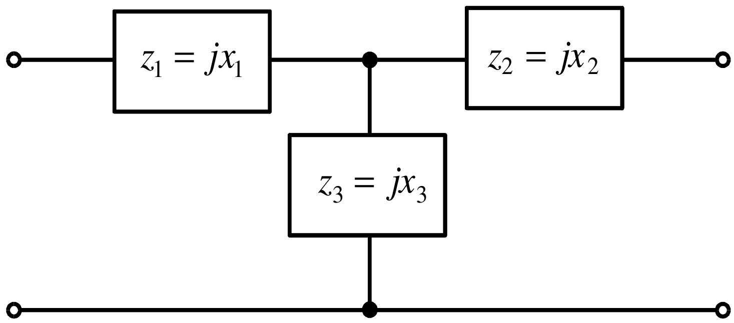


Figure 1-20. T-shape circuit.

The T-shape circuit is shown in Fig. 1-20. Its Z and Y parameters can be written as:

$$z_{11} = jx_{11} = j(x_1 + x_3)\,;\; z_{12} = jx_{12} = jx_3\,;$$

$$z_{21} = jx_{21} = jx_3\,;\; z_{22} = jx_{22} = j(x_2 + x_3)\,;$$

$$y_{11} = jb_{11} = \frac{j(x_2 + x_3)}{\Delta_z};\; y_{12} = jb_{12} = -\frac{jx_3}{\Delta_z};$$

$$y_{21} = jb_{21} = -\frac{jx_3}{\Delta_z};\; y_{22} = jb_{22} = \frac{j(x_1 + x_3)}{\Delta_z};$$

$$\Delta_z = Z_1Z_2 + Z_2Z_3 + Z_1Z_3 = -(x_1x_2 + x_2x_3 + x_1x_3)$$

The parameters of the selected electrical matching network can be found by using the known values of the load impedance $Z_L = R_L + jx_L$ or admittance $Y_L = g_L + jb_L$, and the input impedance Z_{in} or admittance Y_{in}.

Usually, the input impedance or admittance is the real number:

$$Z_{in} = R_{in}\text{, or } Y_{in} = g_{in} = 1/R_{in}\,.$$

Therefore, two pairs of equations can be obtained from the Eqs. (1-27) and (1-28) by their dividing into real and imaginary parts. The first pair is for Z parameters:

$$R_{in}R_L + x_{11}(x_{22} + x_L) - x_{12}^2 = 0\,, \tag{1-31}$$

$$R_L x_{11} - R_{in}(x_{22} + x_L) = 0\,, \tag{1-32}$$

and the other is for Y parameters:

$$g_{in}g_L + b_{11}(b_{22} + b_L) - b_{12}^2 = 0\,, \tag{1-33}$$

$$g_L b_{11} - g_{in}(b_{22} + b_н) = 0\,. \tag{1-34}$$

By substituting the x-parameters' expressions into Eqs. (1-31) and (1-32), the following equations can be obtained for the T-shape circuit:

$$R_{in}R_L + (x_1 + x_3)(x_2 + x_3 + x_L) - x_3^2 = 0\,, \tag{1-35}$$

$$R_L(x_1 + x_3) - x_{in}(x_2 + x_3 + x_L) = 0\,. \tag{1-36}$$

Here, x is positive for inductance, and is negative for capacitance.

In case of Π-shape circuit, substituting the b-parameters' expressions into Eqs. (1-33) and (1-34) gives the following:

$$g_{in}g_L + (b_1 + b_3)(b_2 + b_3 + b_L) - b_3^2 = 0\,, \tag{1-37}$$

$$g_L(b_1 + b_3) - g_{in}(b_2 + b_3 + b_L) = 0\,, \tag{1-38}$$

where b is positive for capacitance, and is negative for inductance.

There are three known and three unknown values for the each set of Eqs. (1-35), (1-36), and (1-37), (1-38). The known values are R_{in}, R_L, x_L or g_{in}, g_L, b_L, while the unknown values are x_1, x_2, x_3 or b_1, b_2, b_3.

One of x values or one of b values should be given in order to calculate the matching network parameters. The remained parameters can be found from the Eqs. (1-35), (1-36) or (1-37), (1-38).

4.2.1 Input network example

Usually, the input matching network should provide transformation of standard 50 Ohm or 75 Ohm impedance to the much lower transistor input impedance.

Example of input network is shown in Fig. 1-21. The series equivalent circuit can be used for the transistor input impedance.

The reactive part of impedance should be accounted in the X_L value. The R_1 is the required driving source impedance, R_2 is the real part of transistor input impedance. Changing the inductance L value compensates the influence of the reactive component of transistor input impedance.

The expressions for the network elements' impedances are the following:

$$Q^2 > R_1/R_2 - 1\,,\; R_1/R_2 > 1\,, \tag{1-39}$$

$$X_L = QR_2\,, \tag{1-40}$$

$$X_1 = R_1\sqrt{\frac{R_2}{R_1}\left(1+Q^2\right)-1}\,, \tag{1-41}$$

$$X_2 = \frac{R_1}{R_1/R_2 - 1}\left(Q + \sqrt{\frac{R_2}{R_1}\left(1+Q^2\right)-1}\right), \tag{1-42}$$

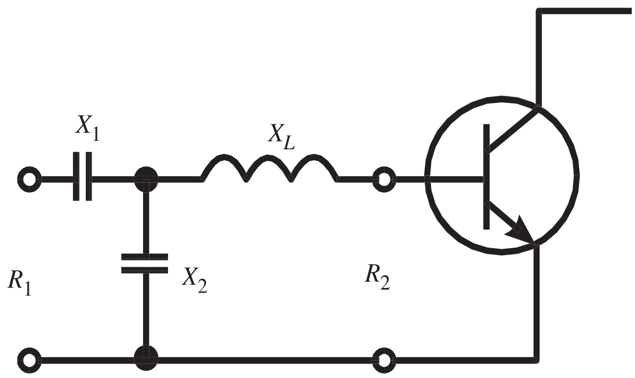


Figure 1-21. Example of input matching network.

where q is the quality factor, and should be given. The certain inductance or capacitance should be calculated for the given frequency f_0 as $L = X/(2\pi f_0)$, $C = 1/(2\pi f_0 X)$, respectively.

4.2.2 Output network example

The example of output matching network is shown in Fig. 1-22. The transistor output capacitance is accounted in the X_1 reactance. The elements' impedances can be calculated as follow:

$$Q^2 > R_1/R_2 - 1, \tag{1-43}$$

$$X_1 = R_1/Q, \tag{1-44}$$

$$X_2 = \frac{R_2}{\sqrt{\frac{R_2}{R_1}\left(1+Q^2\right)-1}}, \tag{1-45}$$

$$X_L = \frac{R_1}{1+Q^2}\left(Q + \sqrt{\frac{R_2}{R_1}\left(1+Q^2\right)-1}\right). \tag{1-46}$$

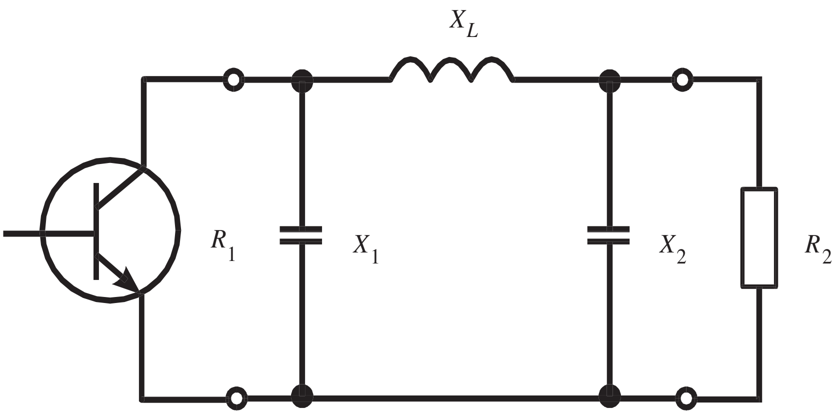


Figure 1-22. Example of output matching network.

4.2.3 Class-F amplifier output network example

The typical class-F output network provides the short-circuit behavior for the second harmonic and acts as open-circuit for the third harmonic at the transistor output. The example of such network[13] is shown in Fig. 1-23. The circuit is tuned in such a way, that the first harmonic is resonant for the L1-C1 contour; the third harmonic is resonant for the L2-C2 contour; while the second harmonic is resonant for the series contour of inductive L2-C2 and capacitive L1-C1.

The circuit elements can be calculated as follows:

$$C1 = \frac{\alpha_F}{\omega_0 \cdot RL \cdot (1-\alpha^2)}, \ \alpha_F = \frac{\omega_0 - \pi \cdot BW}{\omega_0}, \tag{1-47}$$

$$L1 = \frac{1}{\omega_0^{\ 2} \cdot C1}, \tag{1-48}$$

$$L2 = \frac{160 \cdot L1 \cdot RL^2}{81 \cdot \left[(3 \cdot RL)^2 + (2\omega_0 \cdot L1)^2\right]}, \tag{1-49}$$

$$C2 = \frac{1}{9\omega_0^{\ 2} L2}, \tag{1-50}$$

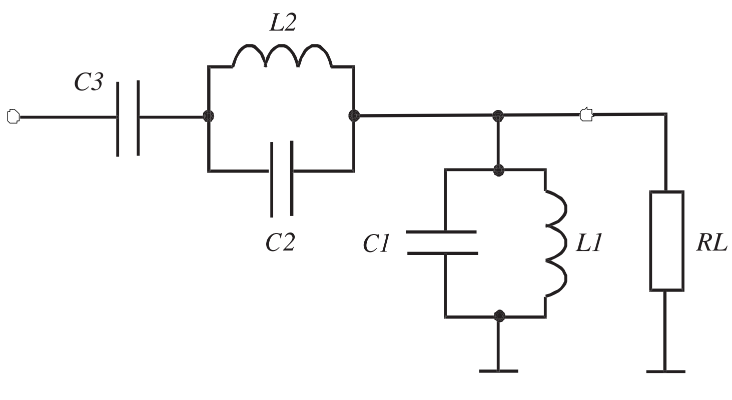


Figure 1-23. Example of class-F amplifier output network.

$$C3 = 8 \cdot C2\,, \tag{1-51}$$

where *BW* is a frequency bandwidth parameter that can be about the 40% from the operating frequency.

5. SUMMARY

In this Chapter, the classification of power amplifiers by the transistor output voltage waveform is presented. The cosinusoidal approximation of collector current is described with the Fourier coefficients' α_n dependence on the conduction angle.

The polyharmonic class-F power amplifiers are considered in details. The optimum relations between the harmonic magnitudes of transistor output current and voltage are described. The drawback of the current design method not considering the transistor lag at the relatively high frequencies is pointed out.