CHAPTER 10

# Large Signal Amplifiers

## Learning Objectives

After completing this chapter, you will learn the following:

- Large signal amplifiers and their characteristic parameters.
- Classification of large signal amplifiers: class A, class B, class AB, class C, class D and other classes.
- Principle of operation of different types of class A amplifiers including class A amplifiers with direct coupled resistive load, transformer coupled load and class A push–pull amplifiers.
- Principle of operation of different types of class B amplifiers including transformer coupled push–pull class B amplifiers, complementary-symmetry push–pull class B amplifiers and quasi-complementary push–pull class B amplifiers.
- Principle of operation of class AB amplifiers.
- Class C amplifiers.
- Class D amplifiers.
- Thermal management of power transistors.

The amplifier circuits we have studied until now are small signal amplifiers where the signal voltage and current levels are small. The focus in this chapter is on large signal or power amplifiers. Large signal or power amplifiers provide power amplification and are used in applications to provide sufficient power to the load or the power device. The output power delivered by these amplifiers is of the order of few watts to few tens of watts. They handle moderate-to-high levels of current and voltage signals as against small levels of current and voltage signals in the case of small signal amplifiers. As we have studied in earlier chapters, the main characteristic specifications in case of small signal amplifiers are amplification linearity and gain magnitude. In the case of large signal amplifiers, the main design specifications are power efficiency, maximum power-handling capability of the amplifier and distortion.

The topics covered in this chapter include classification of large signal amplifiers into different classes, namely, class A, class B, class AB, class C, class D and other classes and the main characteristic specifications of power amplifiers. The principle of operation of different types of class A amplifiers including direct coupled, transformer coupled and push–pull amplifiers is covered next. This is followed up by a discussion on transformer-coupled push–pull, complementary-symmetry push–pull and quasi-complementary push–pull class B amplifiers. Other classes of large signal amplifiers, which include class AB, class C and class D amplifiers, are covered next. The chapter ends with a discussion on thermal management of power transistors. The chapter is amply illustrated with a large number of solved examples.

# 10.1 Large Signal Amplifiers

Large signal or power amplifiers provide power amplification and are used in applications to provide sufficient power to the load or the power device. The output power delivered by these amplifiers is of the order of few Watts to few tens of Watts. In this section we will discuss different classes of power amplifiers and their characteristic specifications.

## Classification

On the basis of their circuit configurations and principle of operation, amplifiers are classified into different classes. Different classes of large signal amplifiers include class A, class B, class AB, class C, class D, class E and class F amplifiers. Each of these amplifiers offers different advantages and disadvantages as compared to each other. In this section, a brief introduction to these classes and general large signal amplifier characteristic specifications is given. Class E and F amplifiers are very rarely used. Hence, only class A, class B, class AB, class C and class D amplifiers are described in detail later in the chapter.

### Class A Amplifiers

The active device in a class A amplifier operates during the whole of the input cycle and the output signal is an amplified replica of the input signal with no clipping. In other words, in class A amplifiers, the amplifying element is so biased that it operates over the linear region of its output characteristics during the full period of the input cycle and is always conducting to some extent. Class A amplifiers offer very poor efficiency and a theoretical maximum of 50% efficiency is possible in these amplifiers. They are generally used for implementing small signal amplifiers. Figure 10.1 shows the input and output waveforms of class A amplifiers.

### Class B Amplifiers

Class B amplifiers operate only during the half of the input cycle as shown in Figure 10.2. Class B amplifiers offer much improved efficiency over class A amplifiers with a possible theoretical maximum of 78.5%.

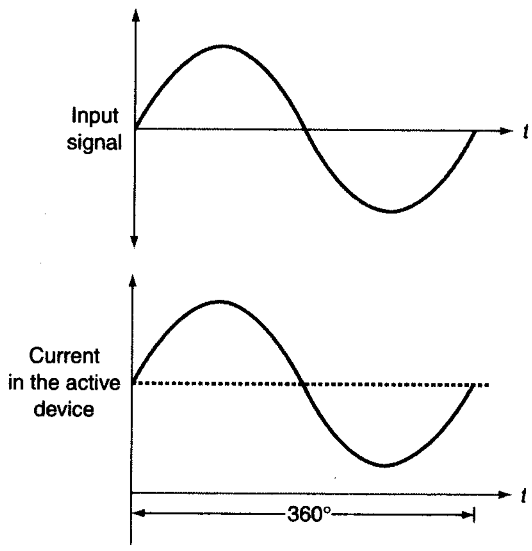



**Figure 10.1** | Waveforms of class A amplifiers.

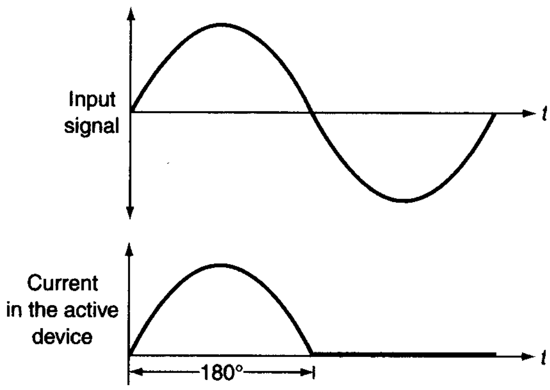



**Figure 10.2** | Waveforms of class B amplifiers.

However they also create a large amount of distortion. A single class B amplifier is rarely used in practical systems. Two class B amplifiers are used either as a complementary pair or in a push–pull arrangement. These configurations are discussed in detail in Section 10.3.

### *Class AB Amplifiers*

In a class AB amplifier, the amplifying device conducts for a little more than half of the input waveform. They sacrifice some efficiency over class B amplifiers but they offer better linearity than class B amplifiers. However, they offer much more efficiency than class A amplifiers. Figure 10.3 shows the waveforms of class AB amplifiers. Operation of class AB amplifiers is explained in Section 10.4.

### *Class C Amplifiers*

Class C amplifiers conduct for less than half cycle of the input signal (Figure 10.4) resulting in a very high efficiencies upto 90%. However, they are associated with a very high level of distortion at the output. Class C amplifiers operate in two modes, namely, the tuned mode and the untuned mode. Class C amplifiers are discussed in detail in Section 10.5.

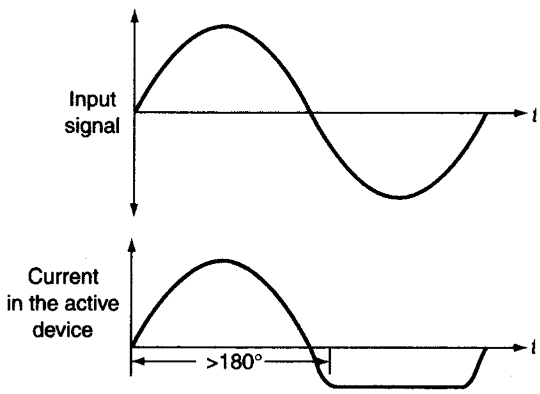



**Figure 10.3** | Input and output waveforms of class AB amplifiers.

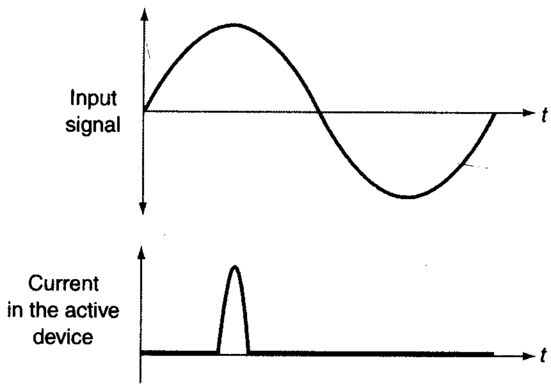



**Figure 10.4** | Input and output waveforms of class C amplifiers.

### *Class D Amplifiers*

Class D amplifiers use the active device in switching mode to regulate the output power. Hence, these amplifiers offer high efficiencies and do not require heat sinks and transformers. These amplifiers use pulse width modulation (PWM), pulse density modulation or sigma delta modulation to convert the input signal into a string of pulses. Figure 10.5 shows how a sinusoidal input waveform is converted into a string of digital pulses using PWM technique. As we can see from the figure, the pulse width of the PWM output waveform at any time instant is directly proportional to the amplitude of the input signal.

### *Other Classes*

Class E and class F amplifiers are switching power amplifiers offering very high efficiency levels. They are used at very high frequencies where the switching time is comparable to the duty time. Class G amplifiers are efficient versions of class AB amplifiers. They employ rail switching to decrease power consumption and increase efficiency. The amplifier has several power rails at different voltages and it switches between the rails as the output signal approaches each rail value. Class H amplifiers create infinite number of supply rails by modulating the supply rails. Detailed description of these classes is beyond the scope of the book.

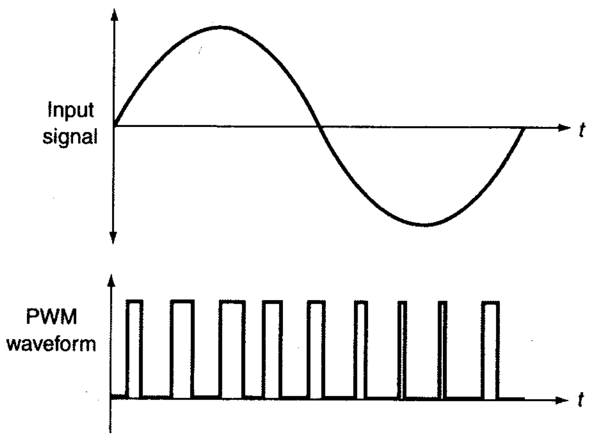



**Figure 10.5** | Class D amplifier waveforms.

## *Large Signal Amplifier Characteristics*

The main characteristics that define the performance of a power amplifier are efficiency, distortion level and output power. These characteristics are discussed in detail in this section.

### *Efficiency*

Efficiency of an amplifier is defined as the ability of the amplifier to convert the DC input power of the supply into an AC output power that can be delivered to the load. The expression for efficiency is given by

$$\eta = \frac{P_o}{P_i} \times 100\% \tag{10.1}$$

where $P_o$ is the AC output power delivered to the load and $P_i$ the DC input power.

### *Harmonic Distortion*

Distortion in large signal amplifiers is mainly caused due to harmonic distortion. Harmonic distortion refers to the distortion in the amplitude of the output signal of an amplifier caused due to the non-linearity in the characteristics of the active device used for amplification. In other words, the active device does not equally amplify all portions of the input signal over its positive and negative excursions. The distortion is more in the case of a large input signal level. Figure 10.6 shows how the non-linear dynamic transfer curve of the active device results in harmonic distortion.

The dynamic transfer curve of an active device can be generalized using the following equation

$$I_o = k_1 I_i + k_2 I_i^2 + k_3 I_i^3 + \ldots \tag{10.2}$$

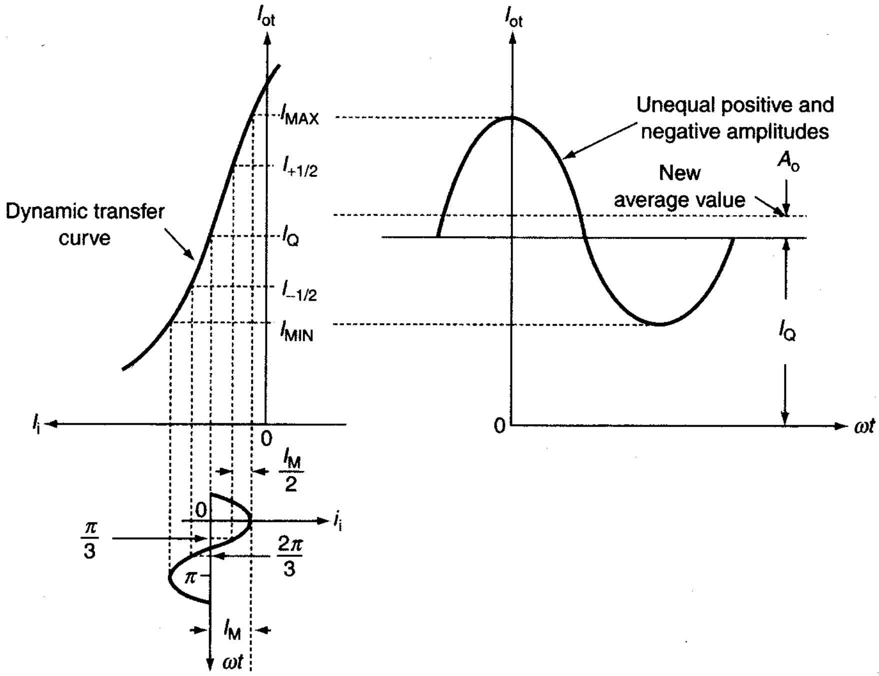



**Figure 10.6** Harmonic distortion due to the non-linearity in the transfer characteristics of the active device.

where $I_o$ is the alternating portion of the output current; $I_i$ the input current; $k_1$, $k_2$, $k_3$ are constants.

In case the input signal is not sufficiently large to drive the amplifier to the extremes of its dynamic transfer curve, then the slight curvature of the dynamic curve over the region of operation may be described by the following parabolic relation:

$$I_o = k_1 I_i + k_2 I_i^2 \tag{10.3}$$

Let us assume that the input signal is a sinusoidal signal and is expressed as

$$I_i = I_M \cos\omega t \tag{10.4}$$

Substituting Eq. (10.4) in Eq. (10.3) we get

$$I_o = k_1 I_M \cos\omega t + k_2 I_M^2 \cos^2\omega t$$

As

$$\cos^2\omega t = \frac{1}{2} + \frac{\cos 2\omega t}{2}$$

we get

$$I_o = k_1 I_M \cos\omega t + \frac{1}{2} k_2 I_M^2 + \frac{1}{2} k_2 I_M^2 \cos 2\omega t \tag{10.5}$$

Substituting $(1/2)k_2 I_M^2 = A_0$, $k_1 I_M = A_1$ and $(1/2)k_2 I_M^2 = A_2$ in Eq. (10.5) we get

$$I_o = A_0 + A_1 \cos\omega t + A_2 \cos 2\omega t$$

Total current, $I_{ot} = I_o + I_Q$, where $I_Q$ is the output DC current under quiescent conditions. Therefore,

$$I_{ot} = I_Q + A_0 + A_1 \cos\omega t + A_2 \cos 2\omega t \tag{10.6}$$

where $A_0$ is the extra DC component due to rectification of the signal; $A_1$ is the amplitude of the desired signal at the fundamental input signal frequency $\omega$; $A_2$ is the amplitude of the second harmonic frequency component at $2\omega$.

From Eq. (10.6) it is clear that there is a slight shift in the operating point when the input signal is applied. This is indicated by the extra DC component ($A_0$) which suggests that the average DC current flowing through the active device has changed. The shift in the operating point is also highlighted in Figure 10.6.

$A_2$ represents the second harmonic term which is the output component at twice the frequency of the applied input signal. Second harmonic distortion is a measure of the relative amount of second harmonic component to the fundamental frequency component and is expressed as

$$D_2 = \left|\frac{A_2}{A_1}\right| \times 100\% \tag{10.7}$$

The values of $A_0$, $A_1$ and $A_2$ can be determined in terms of the values of the output signal from the Figure 10.6.

1. When $\omega t = 0$, $I_{ot} = I_{MAX}$.
2. When $\omega t = \pi/2$, $I_{ot} = I_Q$.
3. When $\omega t = \pi$, $I_{ot} = I_{MIN}$.

Substituting these values in Eq. (10.6) and solving we get

$$A_0 = A_2 = \frac{I_{MAX} + I_{MIN} - 2I_Q}{4} = \frac{(I_{MAX} + I_{MIN})}{4} - \frac{I_Q}{2} \tag{10.8}$$

$$A_1 = \frac{I_{MAX} - I_{MIN}}{2} \tag{10.9}$$

When the magnitude of the input signal is large enough to drive the amplifier to the extremes of its dynamic transfer curve, which is mostly true for power amplifiers, the higher harmonic terms cannot be neglected. As given in Eq. (10.2), the output current $I_o$ is expressed in terms of the input signal as

$$I_o = k_1 I_i + k_2 I_i^2 + k_3 I_i^3 + \ldots$$

Let us assume that the input signal is a sinusoidal signal and is expressed as

$$I_i = I_M \cos\omega t$$

The output signal is then expressed as

$$I_o = k_1 I_M \cos\omega t + k_2 I_M^2 \cos^2\omega t + k_3 I_M^3 \cos^3\omega t + \ldots \tag{10.10}$$

Solving Eq. (10.10) in terms of multiples of the signal frequency $\omega$ and expressing the constant terms as $A_0$, $A_1$ and $A_2$ for the DC component, input signal frequency component and the second harmonic component, respectively, we get

$$I_o = A_0 + A_1\cos\omega t + A_2\cos 2\omega t + A_3\cos 3\omega t + \ldots \tag{10.11}$$

The total current $I_{ot}$ is then given by

$$I_{ot} = I_Q + A_0 + A_1\cos\omega t + A_2\cos 2\omega t + A_3\cos 3\omega t + \ldots \tag{10.12}$$

The values of the harmonic components can be determined in a manner similar to that described earlier in previous paragraphs. In order to obtain the expression for upto fourth harmonic, five values of output current are necessary to solve the equations. The output current values are chosen for maximum value of the input signal, minimum value of the input signal, zero value of the input signal, one-half the maximum positive value of input signal and one-half the maximum negative value of input signal. Let $I_{MAX}$, $I_{MIN}$, $I_Q$, $I_{+1/2}$ and $I_{-1/2}$ be the corresponding output currents for the above-mentioned values of the input signals, respectively. Solving in a manner similar as explained before, we get

$$A_0 = \frac{1}{6} \times (I_{MAX} + 2I_{+1/2} + 2I_{-1/2} + I_{MIN}) - I_Q \tag{10.13}$$

$$A_1 = \frac{1}{3} \times (I_{MAX} + I_{+1/2} - I_{-1/2} - I_{MIN}) \tag{10.14}$$

$$A_2 = \frac{1}{4} \times (I_{MAX} - 2I_Q + I_{MIN}) \tag{10.15}$$

$$A_3 = \frac{1}{6} \times (I_{MAX} - 2I_{+1/2} + 2I_{-1/2} - I_{MIN}) \tag{10.16}$$

$$A_4 = \frac{1}{12} \times (I_{MAX} - 4I_{+1/2} + 6I_Q - 4I_{-1/2} + I_{MIN}) \tag{10.17}$$

The second harmonic distortion component ($D_2$) is given by

$$D_2 = \left|\frac{A_2}{A_1}\right| \times 100\% \tag{10.18}$$

The third harmonic distortion component ($D_3$) is given by

$$D_3 = \left|\frac{A_3}{A_1}\right| \times 100\% \tag{10.19}$$

The fourth harmonic distortion component ($D_4$) is given by

$$D_4 = \left|\frac{A_4}{A_1}\right| \times 100\% \tag{10.20}$$

The total harmonic distortion ($D$) is given by the square root of the mean square values of the individual harmonic components:

$$D = \sqrt{D_2^2 + D_3^2 + D_4^2 + \ldots} \tag{10.21}$$

The power delivered at the fundamental frequency is

$$P_1 = \frac{A_1^2 R_L}{2} \tag{10.22}$$

The total power output is given by

$$P = (A_1^2 + A_2^2 + A_3^2 + \ldots) \times \frac{R_L}{2} \tag{10.23}$$

Therefore,

$$P = (1 + D_2^2 + D_3^2 + \ldots) \times P_1 = (1 + D^2) \times P_1 \tag{10.24}$$

**EXAMPLE 10.1** *Calculate the values of harmonic distortion components for an output signal having amplitude of 5 V at the fundamental frequency, second harmonic component of 0.5 V, third harmonic component of 0.2 V and fourth harmonic component of 0.05 V. Also calculate the total harmonic distortion.*

**Solution**

1. The second harmonic distortion component is given by

$$D_2 = \left|\frac{A_2}{A_1}\right| \times 100\%$$

$$D_2 = \frac{0.5\ \text{V}}{5\ \text{V}} \times 100\% = 10\%$$

2. The third harmonic distortion component is given by

$$D_3 = \left|\frac{A_3}{A_1}\right| \times 100\%$$

$$D_3 = \frac{0.2\ \text{V}}{5\ \text{V}} \times 100\% = 4\%$$

3. The fourth harmonic distortion component is given by

$$D_4 = \left|\frac{A_4}{A_1}\right| \times 100\%$$

$$D_4 = \frac{0.05\ \text{V}}{5\ \text{V}} \times 100\% = 1\%$$

4. The total harmonic distortion ($D$) is equal to

$$D = \sqrt{D_2^2 + D_3^2 + D_4^2}$$

$$D = \sqrt{(10)^2 + (4)^2 + (1)^2} = \sqrt{117} = 10.8\%$$

5. Total harmonic distortion in percentage = 10.8%.

## 10.2 Class A Amplifiers

As mentioned before, the active device in a class A amplifier operates during the whole of the input cycle. Class A amplifiers can have direct-coupled resistive load or a transformer-coupled resistive load. Class A amplifiers are also configured using push–pull arrangement. In this section, all three types of class A amplifiers are discussed in detail with emphasis on their design procedure.

### *Class A Amplifier with Direct-Coupled Resistive Load*

A simple transistor amplifier shown in Figure 10.7 is a class A amplifier with direct-coupled resistive load or a series-fed class A amplifier. The resistor $R_C$ is the load resistor. The difference between this circuit and the amplifier circuits studied earlier in Chapter 4 is that the signal levels in this case are of the order of few tens of Volts and the transistor used here is a power transistor which is capable of operating in the range of several watts to few tens of Watts.

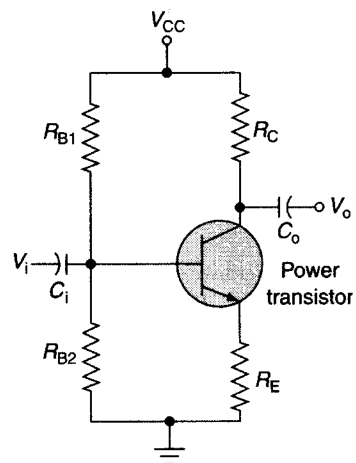



**Figure 10.7** | **Class A amplifier with direct-coupled resistive load.**

The value of the quiescent base current is given by

$$I_B = \frac{V_{TH} - V_{BE}}{R_{TH} + (\beta + 1)R_E} \tag{10.25}$$

where

$$R_{TH} = R_{B1} \| R_{B2} = \frac{R_{B1}R_{B2}}{R_{B1} + R_{B2}}$$

$$V_{TH} = \frac{R_{B2}V_{CC}}{R_{B1} + R_{B2}}$$

The value of collector current ($I_C$) is equal to $\beta I_B$ and the collector–emitter voltage ($V_{CE}$) is given by

$$V_{CE} = V_{CC} - I_C(R_C + R_E)$$

The DC bias determines the quiescent point, which in turn determines the maximum possible collector current swing and the collector–emitter voltage swing when the AC signal is applied at the input. Figure 10.8 shows the output characteristics and the current and voltage waveforms for the amplifier. Figure 10.8(a) shows the waveforms for a small input signal and Figure 10.8(b) shows the waveforms for a large input signal. The input signal applied to the base is a sinusoidal waveform and assuming the output characteristics to be equidistant, the output current and voltage waveforms are also sinusoidal. When an AC signal is applied to the base of the amplifier, the base current will vary above and below the DC bias point, which in turn will cause the collector current and the collector–emitter voltage to vary around its DC bias point. The output current and voltage varying around the bias point provide AC power to the load.

### *Output Power*

The AC output power delivered to the load (resistor $R_C$ in this case) is given by

$$P_o = V_{ce(RMS)}\, I_{c(RMS)} = I^2_{c(RMS)} R_C \tag{10.26}$$

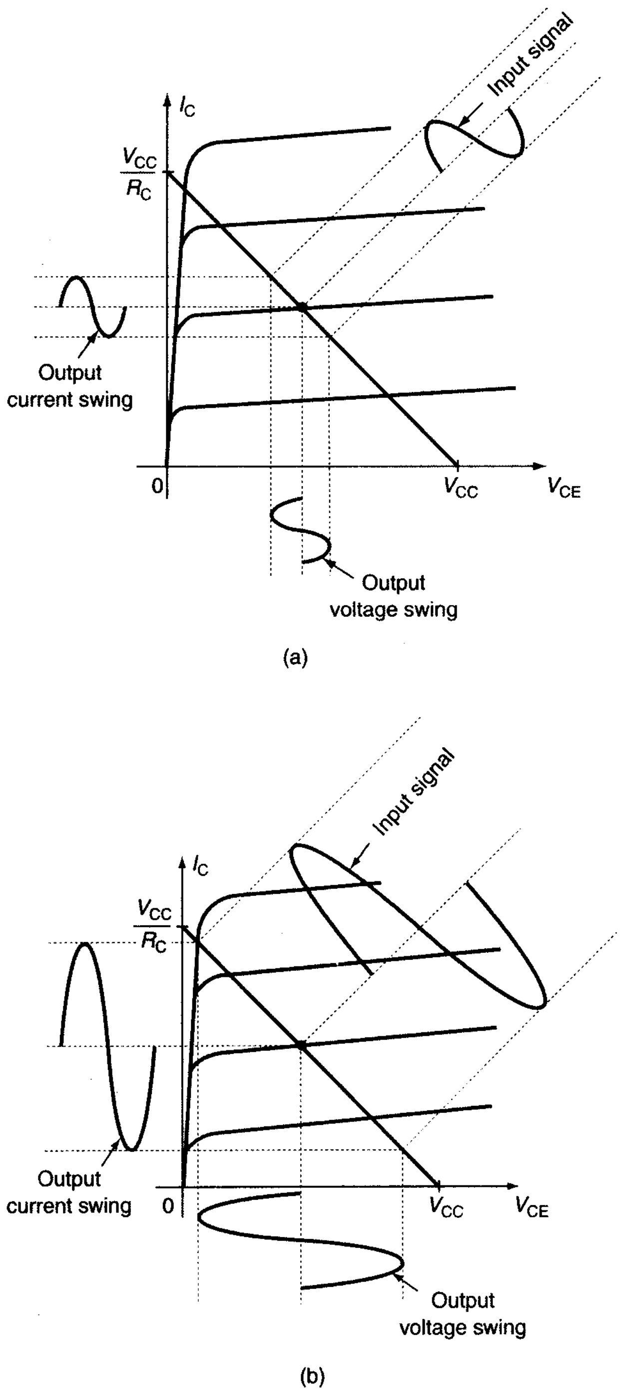



**Figure 10.8** (a) Input and output waveforms for the series-fed class A amplifier – small input signal; (b) input and output waveforms for the series-fed class A amplifier – large input signal.

where $V_{ce(RMS)}$ is the RMS value of the collector–emitter voltage; $I_{c(RMS)}$ the RMS value of the collector current; $R_C$ the load resistance.

The RMS value of a sine wave ($V_{RMS}$) can be expressed in terms of its maximum value ($V_{max}$) and minimum value ($V_{min}$) as given in the following equation:

$$V_{RMS} = \frac{V_{max} - V_{min}}{2\sqrt{2}} \tag{10.27}$$

Therefore, the RMS value of collector–emitter voltage ($V_{ce(RMS)}$) is expressed in terms of the maximum collector voltage ($V_{CE(max)}$) and minimum collector voltage ($V_{CE(min)}$) as

$$V_{ce(RMS)} = \frac{V_{CE(max)} - V_{CE(min)}}{2\sqrt{2}} \tag{10.28}$$

The RMS value of collector voltage ($I_{c(RMS)}$) is expressed in terms of the maximum collector current ($I_{C(max)}$) and minimum collector current ($I_{C(min)}$) as given by the following equation:

$$I_{c(RMS)} = \frac{I_{C(max)} - I_{C(min)}}{2\sqrt{2}} \tag{10.29}$$

Therefore, output AC power ($P_o$) can be rewritten as given by

$$P_o = \frac{(V_{CE(max)} - V_{CE(min)}) \times (I_{C(max)} - I_{C(min)})}{8} \tag{10.30}$$

### *Maximum Efficiency*

The maximum value of efficiency can be calculated by making ideal assumptions like the characteristics curves of the active device are equally spaced in the region of operation, the input signal has zero minimum value, maximum value of collector–emitter voltage ($V_{CE(max)}$) is equal to $V_{CC}$ and the maximum value of collector current ($I_{C(max)}$) is equal to $V_{CC}/R_C$.

The maximum value of power output ($P_{o(max)}$) is given by

$$P_{o(max)} = \frac{(V_{CC}) \times (V_{CC}/R_C)}{8} = \frac{V_{CC}^2}{8R_C} \tag{10.31}$$

The value of the input power ($P_i$) is equal to the product of the supply voltage ($V_{CC}$) and the quiescent value of the collector current ($I_{CQ}$). Assuming that the operating point is in the center of the output characteristics, the value of quiescent collector current ($I_{CQ}$) is equal to $V_{CC}/2R_C$. The value of input power ($P_i$) is therefore given by

$$P_i = V_{CC} \times I_{CQ} = V_{CC} \times \frac{V_{CC}}{2R_C} = \frac{V_{CC}^2}{2R_C} \tag{10.32}$$

The maximum efficiency is given by the ratio of the maximum AC output power given in Eq. (10.31) to the input power given in Eq. (10.32). The maximum value of efficiency is equal to 25%. In other words, the upper limit of the conversion efficiency in case of a series-fed class A amplifier is 25%. It may be mentioned here that in most of the practical series-fed class A amplifiers, the efficiency is much less than 25%.

**EXAMPLE 10.2**

*Calculate the efficiency of the amplifier circuit shown in Figure 10.9. The input voltage applied is such that it produces a base current of 10 mA peak. The emitter–base voltage of the transistor is equal to 0.7 V.*

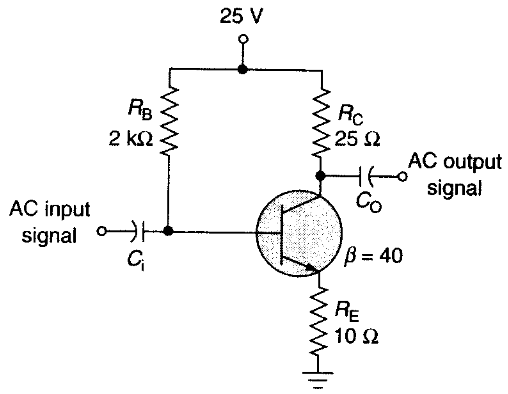



**Figure 10.9** | Example 10.2.

**Solution**

1. The value of the base current at the Q-point is given by $I_{BQ} = (V_{CC} - 0.7)/[R_B + (\beta + 1)R_E] = (25 - 0.7)/(2 \times 10^3 + 41 \times 10) = 24.3/2410 = 10.08 \times 10^{-3}$A = 10.08 mA.
2. The value of the collector current $I_{CQ}$ is given by $\beta \times I_{BQ} = 40 \times 10.08 \times 10^{-3} = 403.3 \times 10^{-3}$ A = 403.3 mA.
3. The input power $P_i = V_{CC} \times I_{CQ} = 25 \times 403.3 \times 10^{-3} = 10.08$ W.
4. The output power $P_o = [I_{C(p)}{}^2 \times R_C]/2$.
5. $I_{C(p)} = \beta \times I_{B(p)} = 40 \times 10 \times 10^{-3} = 400 \times 10^{-3} = 400$ mA.
6. $P_o = [(400 \times 10^{-3})^2 \times 25]/2 = 2$ W.
7. Efficiency $\eta = (P_o/P_i) \times 100\% = (2/10.08) \times 100\% = 19.84\%$.

## *Transformer-Coupled Class A Amplifiers*

Class A amplifier with transformer-coupled load employs a transformer-coupled output stage as shown in Figure 10.10. This configuration offers better efficiency as compared to a class A amplifier with a resistive load. This is so because in the case of direct coupling, the transistor quiescent current passes through the load resistance which results in wastage of power as it does not contribute to the AC component of the output power. In the case of a transformer-coupled load, the primary of the transformer has negligible DC resistance; therefore there is negligible power loss. In fact, the efficiency can be increased by a factor of two by using transformer coupling. However, in this case it is important to use an output-matching transformer in order to transfer a significant power to the load.

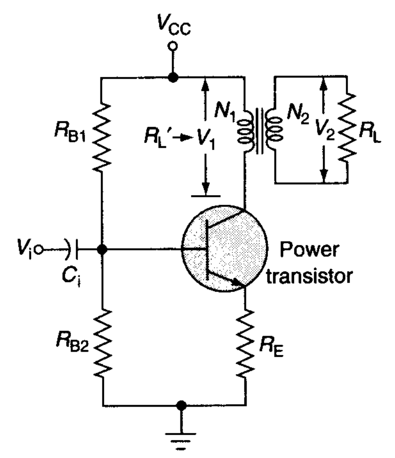



**Figure 10.10** | Transformer-coupled class A amplifier.

The relationship between the transformer's primary and secondary voltages and currents and its turn ratio are given by Eqs. (10.33) and (10.34), respectively:

$$\frac{V_1}{V_2} = \frac{N_1}{N_2} \tag{10.33}$$

where $V_1$ is the primary voltage; $V_2$ the secondary voltage; $N_1$ the number of primary turns; $N_2$ the number of secondary turns:

$$\frac{I_1}{I_2} = \frac{N_2}{N_1} \tag{10.34}$$

where $I_1$ is the primary current; $I_2$ the secondary current; $N_1$ the number of primary turns; $N_2$ the number of secondary turns.

From Eqs. (10.33) and (10.34) we get

$$\frac{V_1}{I_1} = n^2 \times \frac{V_2}{I_2} \tag{10.35}$$

where $V_1/I_1$ represents reflected load resistance $R_L'$, $V_2/I_2$ is equal to the output resistance load resistance $R_L$, $n$ is equal to the ratio of the primary turns to the secondary turns $(N_1/N_2)$.

Equation (10.35) can be rewritten as

$$R_L' = n^2 \times R_L \tag{10.36}$$

Figure 10.11 shows the transistor collector characteristics along with the DC and the AC load lines for the transformer-coupled class A amplifier. The output current and voltage waveforms are also shown.

For transformer-coupled amplifiers, the DC load line is drawn vertically from $V_{CE} = V_{CC}$. This is so as the resistance of the transformer's primary is negligible. The AC load line is drawn with a slope proportional to $-1/R_L'$.

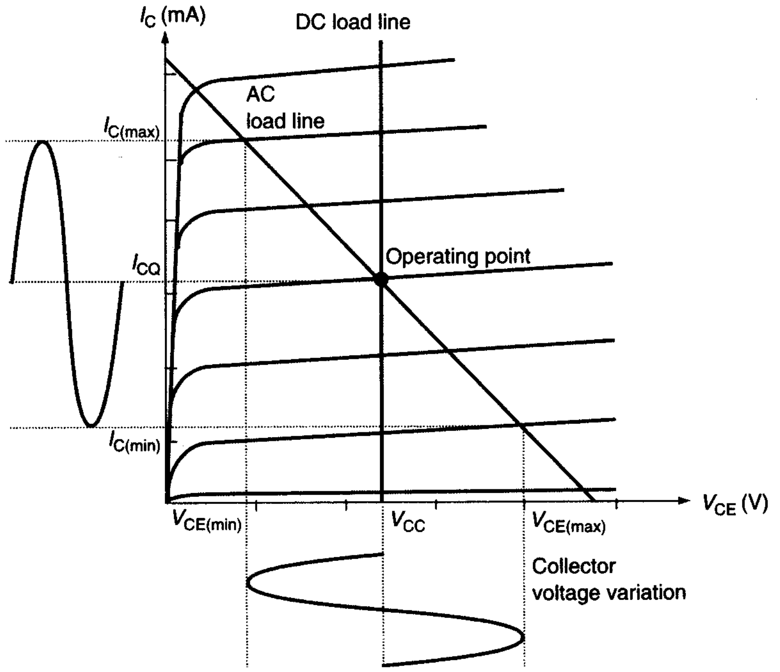



**Figure 10.11** Transistor collector characteristics along with the DC and the AC load lines for the transformer-coupled class A amplifier.

The main design objective is to design a circuit that will deliver maximum output power to the load for a small allowable distortion. To deliver maximum output power, the AC load line is drawn tangent to the hyperbola curve of maximum power dissipation. The point of tangency is where the DC load line intersects the hyperbola. As already discussed in Chapter 4, the load line must be drawn so as not to violate the maximum collector–emitter voltage rating of transistor. This is usually done by making $V_{CC} \le V_{CE(max)}/2$.

In the case of transformer-coupled class A amplifier, the maximum power dissipated ($P_{D(max)}$) is twice the AC output power ($P_o'$). Here $P_o'$ is the power developed across the transformer's primary. The Q-point is located at

$$I_{CQ} = \frac{2P_o'}{V_{CC}}, \quad V_{CEQ} = V_{CC} \tag{10.37}$$

If the Q-point is located at the mid-point of the load line then

$$R_L' = \frac{V_{CC}^2}{2P_o'} \tag{10.38}$$

$R_L'$ is the reflected load resistance seen by the transformer's primary. The transformer chosen should be such that the actual load $R_L$ on the secondary looks like $R_L'$ in the primary.

### *Output Power*

The output AC power delivered to the load can be determined using

$$P_o = \frac{V_{2(RMS)}^2}{R_L} \tag{10.39}$$

where $V_{2(RMS)}$ is the RMS value of the voltage across the transformer's secondary. If the maximum and the minimum collector–emitter voltages are $V_{CE(max)}$ and $V_{CE(min)}$, respectively, and the maximum and minimum collector current values are $I_{C(max)}$ and $I_{C(min)}$, respectively, then the AC power developed across the transformer's primary is given by

$$P_o' = \frac{(V_{CE(max)} - V_{CE(min)}) \times (I_{C(max)} - I_{C(min)})}{8} \tag{10.40}$$

The power delivered to the load ($P_o$) is then given by the product of the transformer's efficiency and the power developed across the transformer's primary given in Eq. (10.40). As the efficiency of efficient transformers is well above 90%, the power delivered to the load ($P_o$) can also be approximated by Eq. (10.40).

### *Efficiency*

The DC input power obtained from the supply is calculated by the product of the DC supply voltage ($V_{CC}$) and the quiescent collector current flowing through the circuit ($I_{CQ}$).

$$P_i = V_{CC} \times I_{CQ} \tag{10.41}$$

The efficiency of the amplifier is given by

$$\eta = \frac{P_o}{P_i} \times 100\% \tag{10.42}$$

The main source of power loss is that dissipated by the transistor in the form of heat and is approximately equal to the difference between the power that is drawn from the DC supply and the AC power delivered to the load. When the input signal is small, little AC power is delivered to the load and maximum power is dissipated by the transistor. When the input signal is large, large amount of power is delivered to the load and less power is dissipated by the transistor. In other words, in a transformer-coupled class A amplifier the power loss is minimum when the load is drawing maximum power from the amplifier and is maximum when the load is disconnected from the amplifier.

The maximum value of efficiency can be calculated by making certain assumptions like the characteristic curves of the active device are equally spaced in the region of operation, the input signal has zero minimum value and the operating point is in the center of the output characteristics.
Under the already mentioned ideal conditions,

$$I_{CQ} = \frac{I_{C(max)} - I_{C(min)}}{2} \text{ and } V_{CC} = \frac{V_{CE(max)} + V_{CE(min)}}{2}$$

Therefore,

$$P_i = \frac{(V_{CE(max)} + V_{CE(min)})}{2} \times \frac{(I_{C(max)} - I_{C(min)})}{2} \tag{10.43}$$

Substituting the values of $P_o$ and $P_i$ in Eq. (10.42) we get

$$\eta = 50\left(\frac{V_{CE(max)} - V_{CE(min)}}{V_{CE(max)} + V_{CE(min)}}\right)\% \tag{10.44}$$

Therefore, the upper limit for theoretical efficiency of a transformer-coupled class A amplifier is 50%, which is twice that of a series-fed class A amplifier.

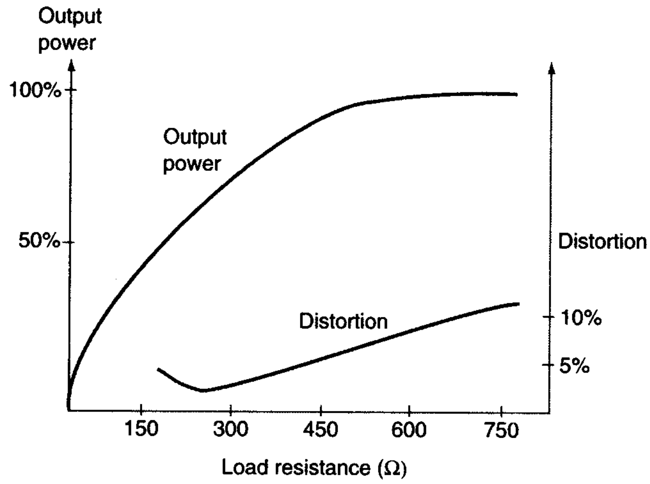



**Figure 10.12** | Variation of output power and total distortion with load resistance.

### *Variation of Output Power with Load*

Figure 10.12 shows the variation of output power and total distortion with the load resistance. As we can see from the curve, the load that provides minimum value of total distortion is different from the load that yields maximum value of output power. However, the output power curve is largely flat in the region of maximum power, so a well-designed amplifier would work properly for a load that gives slightly less than maximum output power as it will give significantly lower value of total distortion.

### *Disadvantages of Transformer-Coupled Class A Amplifier*

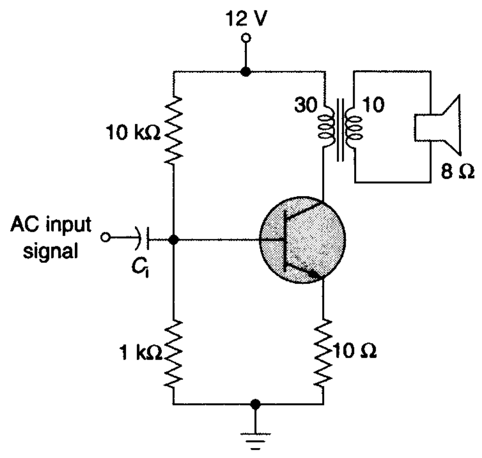
The value of total distortion is quite large in the case of these amplifiers. Moreover, the maximum output power is generated at a load resistance value that is different from the value that offers minimum harmonic distortion. Also, the output transformer is subject to saturation problems as a relatively large DC collector current flows under zero signal conditions.

**EXAMPLE 10.3**

*Determine the AC power delivered to the speaker in the circuit shown in Figure 10.13(a). The value of the quiescent base current is 8 mA and the input signal results in a peak-to-peak base current swing of 8 mA. The output characteristics of the transistor are shown in Figure 10.13(b).*

**Solution**

1. The DC load line is obtained by drawing a vertical line from the point (0, $V_{CC}$), that is, from (0, 12 V).
2. The intersection of this DC load line with the curve corresponding to quiescent base current of 8 mA gives the operating point. From Figure 10.14, we can determine that the operating point is $I_{CQ}$ = 160 mA and $V_{CEQ}$ = 12 V.
3. The effective AC resistance seen by the transformer's primary $R_L' = (N_1/N_2)^2 \times R_L = (30/10)^2 \times R_L = 9 \times 8\ \Omega = 72\ \Omega$.



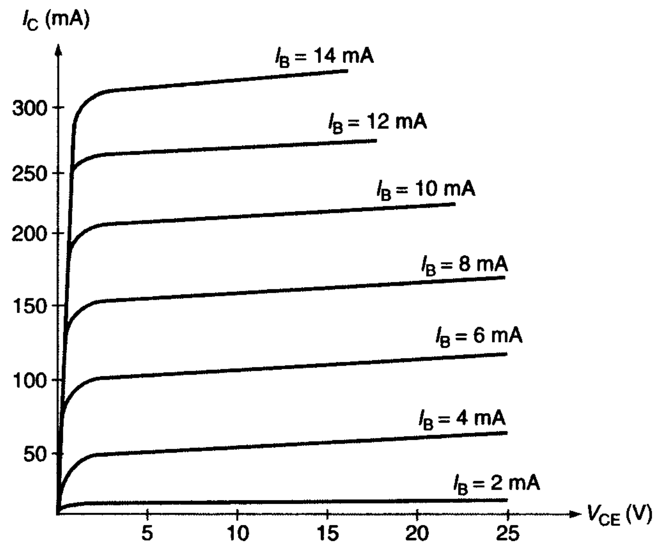




**Figure 10.13** | Example 10.3.

4. The AC load line is drawn with a slope of (−1/72) going through the operating point.

5. The intercept of the load line on the Y-axis = $I_{CQ} + V_{CC}/R_L' = [160 + (12 \times 10^3)/72]$ mA = 160 mA + 166.67 mA = 326.67 mA.

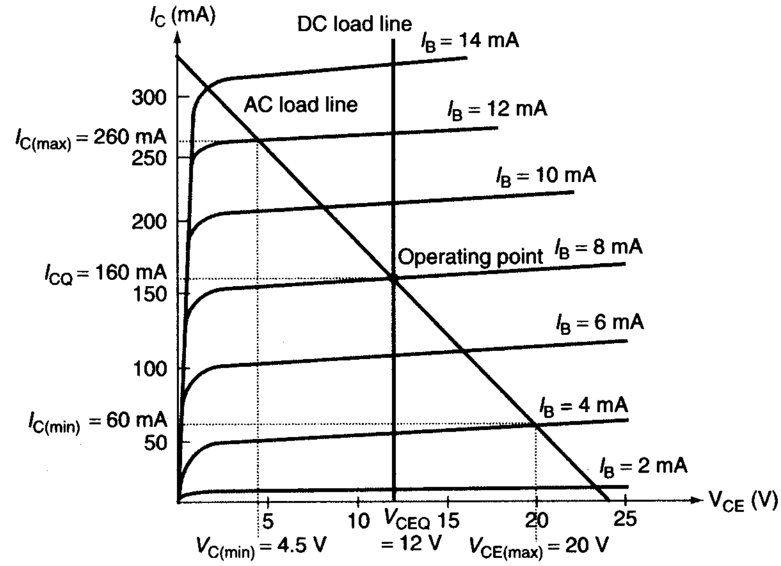



**Figure 10.14** | Solution to Example 10.3.

6. The AC load line is drawn by joining the operating point ($I_{CQ}$ = 160 mA, $V_{CEQ}$ = 12 V) and the point ($I_C$ = 326.67 mA, $V_{CE}$ = 0). The AC load line is shown in Figure 10.14.

7. The peak-to-peak base current swing is 8 mA. Therefore, the peak base current swing is 4 mA.

8. The maximum and minimum values of the collector current and collector–emitter voltage can be obtained from Figure 10.14:

$$V_{CE(min)} = 4.5 \text{ V},\ V_{CE(max)} = 20 \text{ V},\ I_{C(min)} = 60 \text{ mA and } I_{C(max)} = 260 \text{ mA}$$

9. The AC power delivered to the load is equal to

$$P_o = \frac{(V_{CE(max)} - V_{CE(min)}) \times (I_{C(max)} - I_{C(min)})}{8}$$

$$P_o = [(20 - 4.5) \times (260 - 60)]/8 \text{ mW}$$
$$= (15.5 \times 200)/8 \text{ mW} = 387.5 \text{ mW} = 0.3875 \text{ W}$$

## Class A Push–Pull Amplifiers

Push–pull amplifiers employ two active devices which are fed with input signals that are 180° out-of-phase with each other. These amplifiers usually have a center-tapped output transformer. Push–pull configuration offers advantages such as cancellation of second harmonic distortion term and reduction in the power supply ripple voltage.

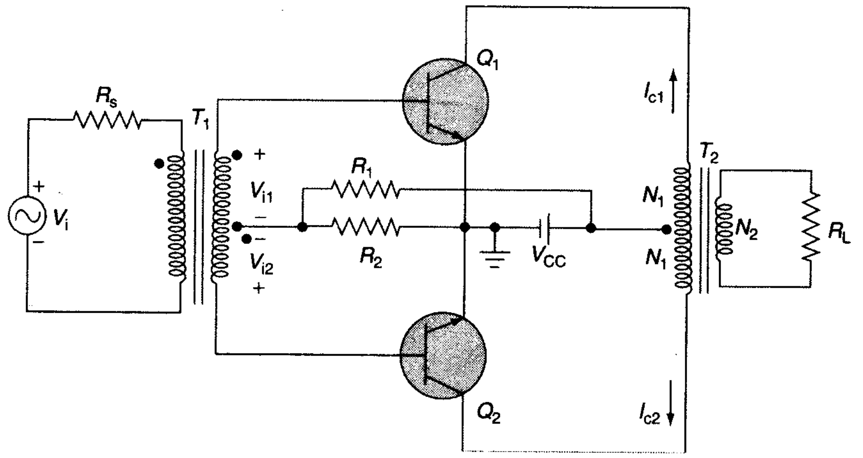



**Figure 10.15** | Class A push–pull amplifier.

Figure 10.15 shows the circuit configuration of a class A push–pull amplifier. The transformer at the input end ($T_1$) provides phase-splitting and the two voltages to the base of the transistors $Q_1$ and $Q_2$ are 180° out-of-phase w.r.t. each other. Resistors $R_1$ and $R_2$ are so chosen that both the transistors conduct for the whole input signal cycle. The collector currents of both the transistors pass through the windings of the output transformer ($T_2$) in opposite directions. Therefore, their magnetizing effects cancel with the result that there is no magnetic saturation problem in the output transformer.

When the input signal $V_i$ becomes positive, the base of transistor $Q_1$ becomes more positive and the base of the transistor $Q_2$ becomes less positive. Therefore, the collector current of transistor $Q_1$ will increase in magnitude whereas the collector current of transistor $Q_2$ will decrease in magnitude. Hence, the voltage across resistor $R_L$ which is proportional to the difference between the collector currents flowing through the two transistors will be positive.

When the input signal $V_i$ becomes negative, the reverse action takes place and the output voltage across resistor $R_L$ becomes negative. This pushing and pulling of the output currents results in decrease in the magnitude of harmonic distortion as the even harmonic terms get cancelled without increase in the values of the odd harmonic terms.

Let us assume that the input signal applied to the amplifier is a sinusoidal signal. This results in base currents that are also sinusoidal in shape. Let the base current of transistor $Q_1$ be $I_{b1} = I_B\cos\omega t$. Therefore, the base current of the transistor $Q_2$ is equal to $I_{b2} = I_B\cos(\omega t + 180) = -I_B\cos\omega t$, as the base current flowing through transistor $Q_2$ is 180° out-of-phase to the base current flowing through transistor $Q_1$.

The collector current of transistor $Q_1$ is given by

$$I_{c1} = I_{CQ} + A_0 + A_1\cos\omega t + A_2\cos 2\omega t + A_3\cos 3\omega t + A_4\cos 4\omega t + \ldots \tag{10.45}$$

The collector current through transistor $Q_2$ is given by

$$I_{c2} = I_{CQ} + A_0 + A_1\cos(\omega t + \pi) + A_2\cos 2(\omega t + \pi) + A_3\cos 3(\omega t + \pi) + A_4\cos 4(\omega t + \pi) + \ldots \tag{10.46}$$

On simplifying Eq. (10.46) we get

$$I_{c2} = I_{CQ} + A_0 - A_1\cos\omega t + A_2\cos 2\omega t - A_3\cos 3\omega t + A_4\cos 4\omega t + \ldots \tag{10.47}$$

The output voltage is directly proportional to the difference of the collector currents flowing through transistors $Q_1$ and $Q_2$. The difference between the collector currents flowing through transistors $Q_1$ and $Q_2$ is given by

$$I_{c1} - I_{c2} = 2A_1\cos\omega t + 2A_3\cos3\omega t + \ldots \quad (10.48)$$

We can see from Eq. (10.48) that the DC component and the even harmonic terms have been cancelled. The main source of distortion is the third harmonic component instead of the second harmonic component. The third harmonic distortion term ($D_3$) is given by

$$D_3 = \left|\frac{2A_3}{2A_1}\right| \times 100\% = \left|\frac{A_3}{A_1}\right| \times 100\% \quad (10.49)$$

The total harmonic distortion $D$ is given by

$$D = \sqrt{D_3^2 + D_5^2 + \ldots} \quad (10.50)$$

The discussion above has assumed that the two transistors are identically matched, which is not practically feasible. Hence, a small amount of even harmonic component will be present at the output. The push–pull configuration also results in reduction of the ripple present in the power supply voltage ($V_{CC}$).

## 10.3 Class B Amplifiers

In a class B amplifier, the active device is biased at zero DC level. Therefore, it provides an output signal varying over one-half of the input signal cycle as the active device conducts for only one-half of the input signal cycle. To obtain output for full input cycle, push–pull configuration is used, that is, two active devices are used wherein each conducts for opposite half-cycles and the combined operation provides full cycle of the output signal. Figure 10.16 shows the block schematic of a push–pull configuration. It may be mentioned here that class B amplifiers offer higher efficiency than class A amplifiers using a single active device.

A number of circuit arrangements are possible for obtaining class B operation. These include transformer-coupled push–pull configuration, complementary-symmetry push–pull configuration and quasi-complementary push–pull configuration. Each of these circuit configurations is explained in the following sections.

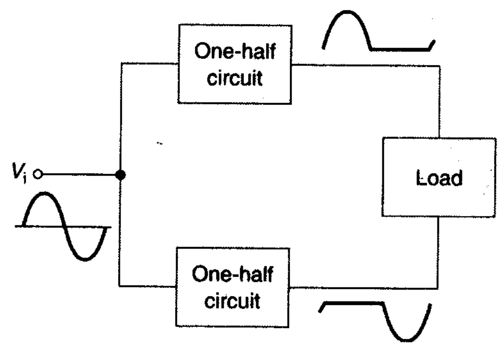



**Figure 10.16** | Block schematic of push–pull configuration.

## Transformer-Coupled Push–Pull Class B Amplifier

Figure 10.17 shows the circuit for a transformer-coupled push–pull class B amplifier. It is essentially the same as a push–pull class A amplifier except that the active devices in this case are biased in the cut-off region. If we compare the circuit of Figure 10.17 with the circuit of class A push–pull amplifier in Figure 10.15, we can see that the resistor $R_2 = 0$ in Figure 10.17, as silicon transistor, is essentially in the cut-off region if its base terminal is shorted to the emitter terminal. Class B amplifier offers higher power output, higher efficiency and there is negligible power loss under quiescent conditions, that is, when no input signal is applied. However, they have higher harmonic distortion, self-bias configuration cannot be used and power supply voltages must have good regulation.

### Conversion Efficiency

To determine the conversion efficiency, let us assume that the output characteristics of the active device are equally spaced and also that the minimum collector current is zero. Figure 10.18(a) shows the graphical schematic for determining the output waveform of a single class B transistor stage. As we can see that for a sinusoidal input excitation, the output is sinusoidal during the first half of the input cycle and is zero during the second half cycle.

The effective value of load resistance ($R_L'$) is given by

$$R_L' = \left(\frac{N_1}{N_2}\right)^2 \times R_L \tag{10.51}$$

where $N_1$ is the number of primary turns to the center tap and $N_2$ the number of secondary turns.

Figure 10.18(b) shows that the collector current waveforms of transistors $Q_1$ and $Q_2$. As is clear from the figure, the output waveform of transistor $Q_2$ is 180° out-of-phase to that of transistor $Q_1$. The load current is proportional to the difference between the collector currents flowing through the transistors $Q_1$ and $Q_2$. It is therefore a perfect sine wave for ideal conditions. The output power ($P_o$) is given by

$$P_o = \frac{I_M V_M}{2} = \frac{I_M}{2} \times (V_{CC} - V_{MIN}) \tag{10.52}$$

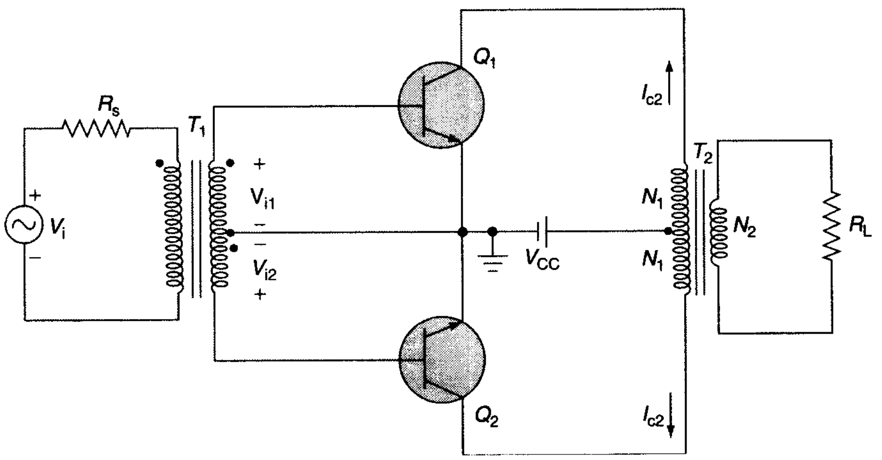



**Figure 10.17** | Transformer-coupled push–pull class B amplifier.

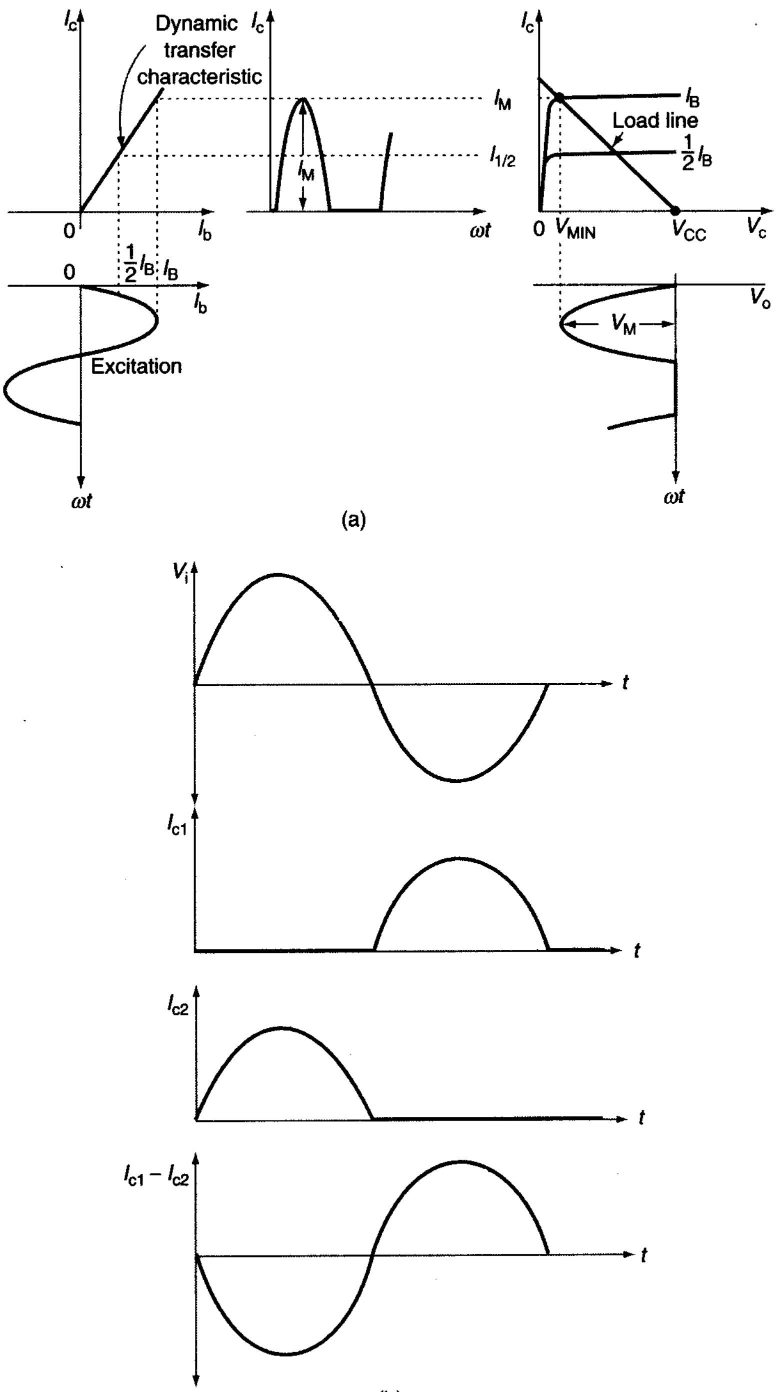



**Figure 10.18** (a) Waveform of a single class B transistor stage; (b) output waveforms for class B push–pull amplifiers.

The maximum output power ($P_{o(max)}$) that can be delivered to the load occurs for the conditions $V_M = V_{CC}$, $V_{MIN} = 0$ and the operating point is at the center of the output characteristics of the transistor. The value of $P_{o(max)}$ is given by

$$P_{o(max)} = \frac{V_{CC}^2}{2R_L'} \tag{10.53}$$

The DC input power ($P_i$) from the supply is given by

$$P_i = 2I_{DC} \times V_{CC} = \frac{2I_M \times V_{CC}}{\pi} \tag{10.54}$$

where $I_{DC}$ is the DC value of the input current and is equal to $I_M/\pi$ or $V_M/\pi R_L'$. The factor of 2 arises because the two transistors are used in push–pull configuration.

Efficiency of any amplifier is given by

$$\eta = \frac{P_o}{P_i} \times 100\% \tag{10.55}$$

Substituting the values of $P_o$ and $P_i$ given in Eqs. (10.52) and (10.54), respectively, in Eq. (10.55), we get

$$\eta = \left(\frac{\pi}{4}\right) \times \left(\frac{V_M}{V_{CC}}\right) \times 100\% = \left(\frac{\pi}{4}\right) \times \left(1 - \frac{V_{MIN}}{V_{CC}}\right) \times 100\% \tag{10.56}$$

The efficiency is maximum when $V_{MIN} = 0$. The maximum possible conversion efficiency is equal to $25\pi$ which is equal to 78.5%. Therefore, the maximum efficiency in class B amplifiers is 78.5% as compared to that of 50% in a class A amplifiers.

For practical systems, the efficiency achieved is not as high as 78.5% but the value of efficiency is higher in systems where the minimum value of the collector voltage is much less than the supply voltage (i.e., $V_{MIN} << V_{CC}$).

Class B systems offer higher efficiency as no current flows through the active device in a class B amplifier when there is no input signal excitation whereas there is a quiescent current flowing through the active device in a class A amplifier in the absence of input signal. It may be mentioned here that power dissipation in class B amplifiers is zero in the absence of input signal and it increases with excitation, whereas in the case of a class A amplifier, dissipation is maximum at zero input and decreases as the signal increases. The collector dissipation in both the transistors ($P_C$) is the difference between the input power to the collector circuit ($P_i$) and the power delivered to the load ($P_o$):

$$P_C = P_i - P_o = \frac{2V_{CC} \times V_M}{\pi R_L'} - \frac{V_M^2}{2R_L'} \tag{10.57}$$

When $V_M = 0$, $P_C = 0$. As $V_M$ increases, the value of the power dissipated in the transistors ($P_C$) increases and it is maximum for $V_M = 2V_{CC}/\pi$. The peak value of power dissipation in both the transistors ($P_{C(max)}$) is given by

$$P_{C(max)} = \frac{2V_{CC}^2}{\pi^2 R_L'} \tag{10.58}$$

Therefore,

$$P_{C(max)} = \frac{4}{\pi^2} \times P_{o(max)} \cong 0.4P_{o(max)} \tag{10.59}$$

Hence, we can obtain a power output of five times the specified power dissipation of a single transistor. As an example, to deliver 10 W from a class B push–pull amplifier, the transistors should have a collector dissipation of 2 W each.

### Harmonic Distortion

The output of a push–pull system has a mirror symmetry. Therefore, $I_Q = 0$, $I_{MAX} = -I_{MIN}$ and $I_{+1/2} = -I_{-1/2}$. The values of different harmonic distortion components are given by

$$A_0 = A_2 = A_4 = 0 \tag{10.60}$$

$$A_1 = \frac{2}{3} \times (I_{MAX} + I_{+1/2}) \tag{10.61}$$

$$A_3 = \frac{1}{3} \times (I_{MAX} - 2I_{+1/2}) \tag{10.62}$$

Therefore, there is no even-harmonic distortion. The principal contributor to the harmonic distortion is the third harmonic distortion component ($D_3$) given by

$$D_3 = \left|\frac{A_3}{A_1}\right| \times 100\% \tag{10.63}$$

The total output power ($P_o$), taking harmonic distortion into account, is given by

$$P_o = (1 + D_3^2) \times \frac{A_1^2 R_L'}{2} \tag{10.64}$$

### Crossover Distortion

Crossover distortion refers to the non-linearity in the output signal when the output signal crosses from positive to negative or from negative to positive. In other words, crossover distortion refers to the distortion resulting at low values of the input signal in the vicinity of the zero region. The output of the transistor collector current is not a perfect half-sine-wave and it results in crossover distortion. This arises as the value of the transistor's $\beta$ is quite low near the cut-off region for small values of collector current compared to the rest of the input signal. Figure 10.19 shows the output waveform with crossover distortion.

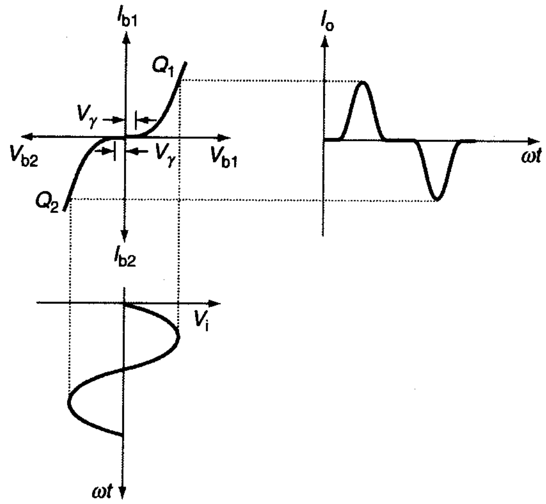



**Figure 10.19** | Crossover distortion.

**EXAMPLE 10.4** *A class B amplifier provides a 20 V peak output signal to 15 Ω load. The system operates on a power supply of 25 V. Determine the efficiency of the amplifier.*

**Solution**

1. The peak load current $I_L(p) = 20/15 = 1.33$ A.
2. The DC value of current drawn from the supply $I_{DC} = (2/\pi) \times I_L(p) = 0.847$ A.
3. Input power delivered by the supply voltage $P_i = V_{CC} \times I_{DC} = 25 \times 0.847 = 21.175$ W.
4. Output power delivered to the load $P_o = V_L(p)^2/2R_L = 20^2/(2 \times 15) = 400/30 = 13.33$ W.
5. Efficiency $\eta = (P_o/P_i) \times 100\% = (13.33/21.175) \times 100\% = 62.95\%$.

## Complementary-Symmetry Push–Pull Class B Amplifier

Complementary-symmetry push–pull class B amplifiers dispense with the input and the output transformers as shown in Figure 10.20. They make use of complementary transistors (NPN and PNP) to obtain a full-cycle output across the load with each transistor operating for half cycle. The same input is applied to the base terminals of both the transistors. The NPN transistor is so biased that it conducts during the positive half of the input signal whereas the PNP transistor conducts during the negative half of the input signal. The output appears across the load for the full cycle of the input signal.

However, there are two main drawbacks of this configuration. First, it requires two different voltage supplies. Second, the output signal suffers from crossover distortion. This is so because the transistors conduct partially near the zero-voltage condition.

A more practical version of a complementary-symmetry push–pull class B amplifier is the one employing darlington-connected transistors as shown in Figure 10.21. The circuit offers higher output current capability and lower output resistance.

## Quasi-Complementary Push–Pull Class B Amplifier

Another variation of the complementary-symmetry push–pull class B amplifier is the quasi-complementary push–pull class B amplifier shown in Figure 10.22. The push–pull operation is achieved by using a pair of

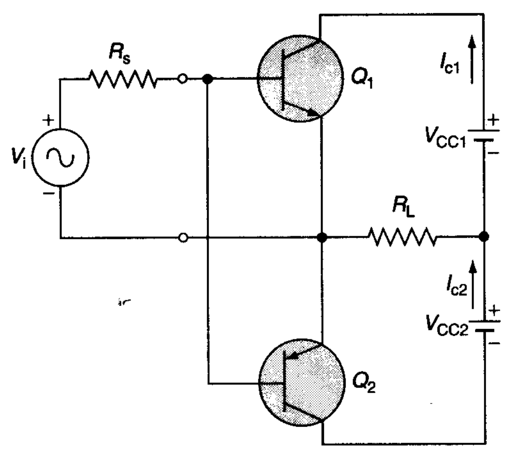



**Figure 10.20** Complementary-symmetry push–pull class B amplifier.

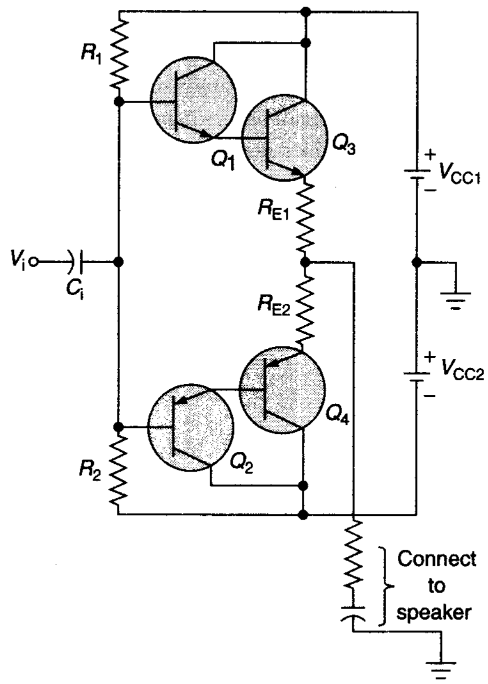



**Figure 10.21** | Complementary-symmetry push–pull class B amplifier with Darlington transistors.

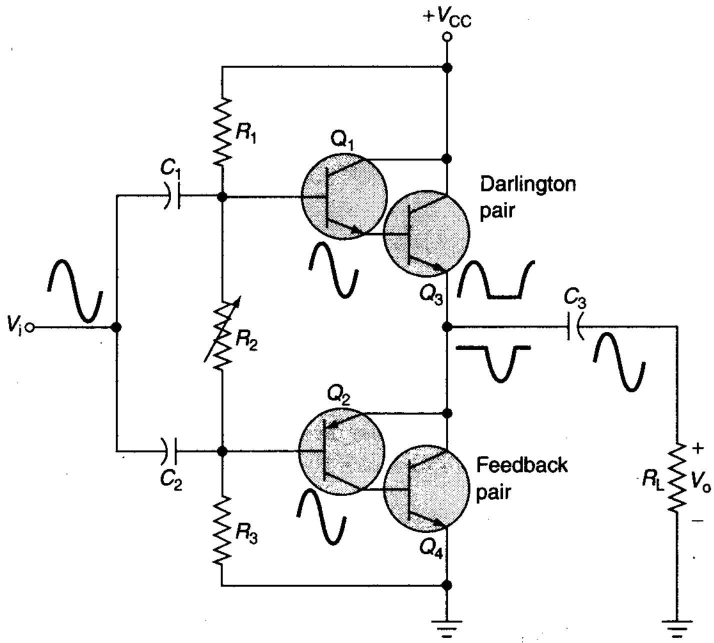



**Figure 10.22** | Quasi-complementary push–pull class B amplifier.

complementary transistors. Transistors $Q_1$ and $Q_3$ form a darlington connection that provides output from a low-impedance emitter–follower. Transistors $Q_2$ and $Q_4$ form a feedback pair and they provide a low-impedance drive to the load. The crossover distortion can be minimized by adjusting the value of the resistor $R_2$.

One of the main advantages of using the quasi-complementary configuration is that it employs matched NPN transistors as high current devices and does not require a high-power PNP transistor as required in case of complementary-symmetry amplifiers. It may be mentioned here that quasi-complementary push–pull class B amplifier is the most popular form of power amplifiers.

**EXAMPLE 10.5**

*For the circuit shown in Figure 10.23, determine the input power, output power, power handled by each transistor and the circuit efficiency for a 15 V peak input signal.*

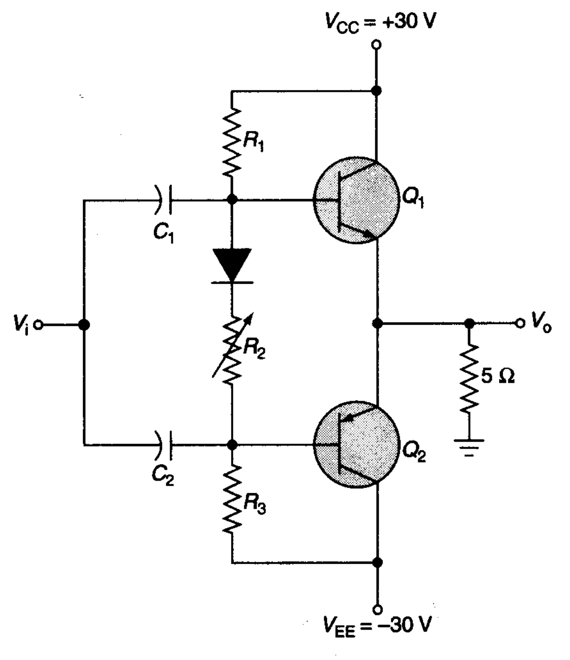



**Figure 10.23** | Example 10.5.

**Solution**

1. The ideal voltage gain of the amplifier is unity. Therefore, the resulting voltage across the load resistance is ideally the same as the input signal. That is, $V_L(p) = 15$ V.
2. The output power across the load resistor ($R_L$) is given by $P_o = [V_L(p)]^2/(2 \times R_L) = (15)^2/(2 \times 5) = 225/10 = 22.5$ W.
3. Peak load current $I_L(p) = V_L(p)/R_L = 15/5 = 3$ A.
4. The DC current drawn from the power supply $I_{DC} = (2/\pi) \times I_L(p) = (2/\pi) \times 3 = 1.91$ A.
5. Input power $P_i = V_{CC} \times I_{DC} = 30 \times 1.91 = 57.3$ W.

6. Total power dissipated $P_D = P_i - P_o = 57.3 - 22.5 = 34.8$ W.
7. Power dissipated in each transistor = $P_D/2 = 34.8/2 = 17.4$ W.
8. The efficiency at the given input signal is $\eta = (P_o/P_i) \times 100\% = (22.5/57.3) \times 100\% = 39.27\%$.

## 10.4 Class AB Amplifiers

Class AB amplifiers do not suffer from the problem of crossover distortion as in these amplifiers a small current flows even at zero input signal level. Figure 10.24 shows the configuration of a class AB push–pull amplifier. The circuit is essentially the same as that for class A amplifier. The voltage drop across resistor $R_2$ is approximately equal to the cut-in voltage of the transistor. When an AC signal is applied to the base, the collector current starts flowing immediately. This biases the transistor away from class B slightly towards class A. Resistors $R_E$ are referred to as the emitter-stabilizing resistors and reduce the amount of distortion produced. However they cause a decrease in the output power due to the negative feedback effect.

## 10.5 Class C Amplifiers

The amplifiers discussed so far, that is, class A, class B and class AB amplifiers, are linear amplifiers where the amplitude and phase of the output signal is directly related to the amplitude and phase of the input signal. In applications where linearity is not the main issue and efficiency is of primary concern, class C, class D, class E and class F amplifiers are used.

The active device in case of class C amplifiers is so biased that it conducts for less than one half of the input cycle, that is, it operates for less than 180° of the input cycle. Only a small portion of the input cycle is passed through the amplifier. This distorts the input signal and hence class C amplifiers are not used for

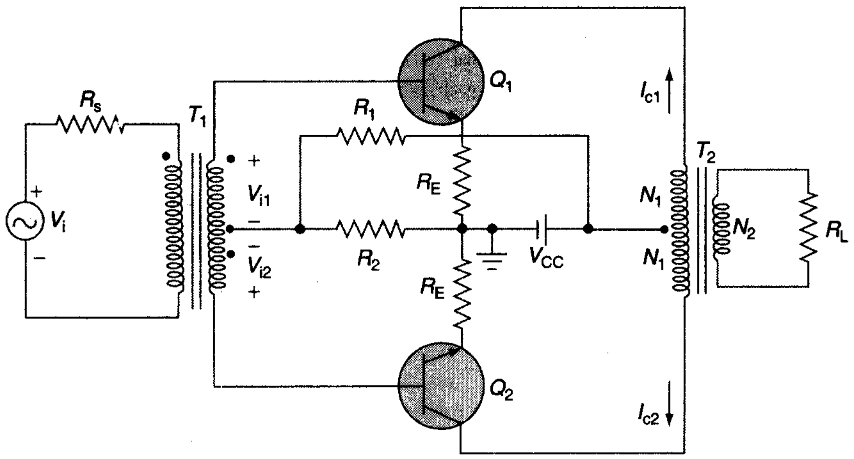



Figure 10.24 | Class AB amplifier.

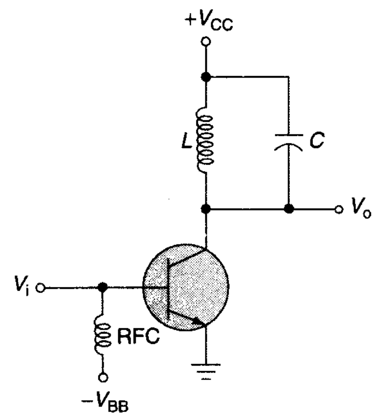



**Figure 10.25** | Tuned class C amplifier.

audio applications. They are widely used as radio frequency (RF) and intermediate frequency (IF) amplifiers. They offer very high efficiencies of the order of 90%.

Class C amplifiers operate in two modes, namely, the untuned mode and the tuned mode. However, mostly they are operated in the tuned mode. Figure 10.25 shows a tuned class C amplifier configuration. The tuned circuit provides a full cycle of the output signal at the fundamental or the resonant frequency of the tuned circuit. Hence, the unwanted frequencies are drastically suppressed. Other residual harmonics can be removed by using a filter. Class C amplifiers are generally used in RF transmitters and are primarily not intended for use as large signal or power amplifiers.

## 10.6 Class D Amplifiers

Class D amplifiers operate with digital or pulsed signals. Class D amplifiers offer very high efficiency levels of the order of 90%; hence they are quite desirable in power amplifiers and audio amplifiers. Figure 10.26 shows the typical efficiency versus output power curves for class AB and class D amplifiers. It may be mentioned here that the theoretical maximum efficiency of class D amplifiers is 100% and over 90% is achievable in practice. High efficiency results in less power consumption for a given output power and reduces the heat-sink requirements drastically. Also, the loading on the power transformer is reduced by a substantial amount, resulting in the use of smaller transformers for the same output power.

The basic concept used in class D amplifiers is to convert the input signal into a pulsed waveform. In other words, the output transistors are operated as switches, that is, they operate either in the cut-off region (open switch) or in the saturation region (closed switch). As we have studied earlier in Chapter 3 on bipolar junction transistors, when the transistor is in the cut-off region, the current through it is ideally zero and when it is in the saturation region, the voltage across it is very small, ideally zero. The power dissipation in both the cases is very low.

Class D amplifiers use PWM, pulse density modulation or sigma delta modulation to convert the input signal into a string of pulses. In the case when the PWM technique is used, the input signal modulates the duty cycle of the drive waveform applied to the input of the switching transistors. In other words, when the amplitude of the input signal increases, the time of the HIGH state increases and that of the LOW state

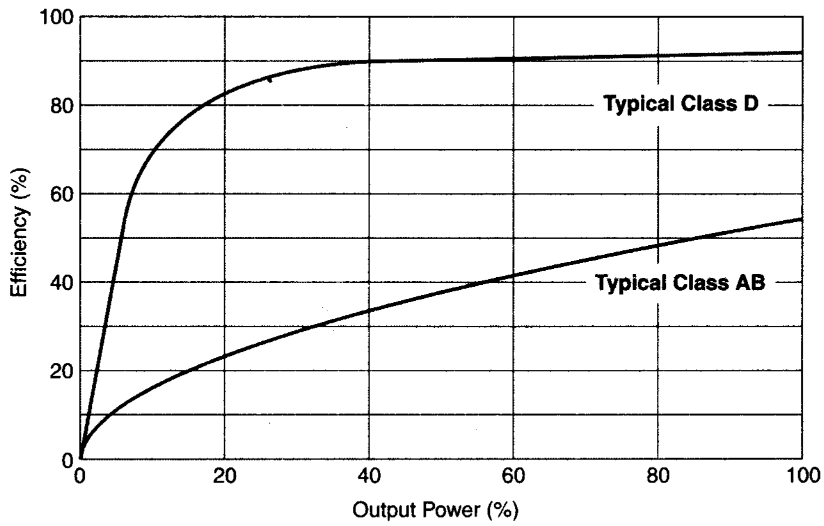



**Figure 10.26** | Efficiency versus output power curves for class AB and class D amplifiers.

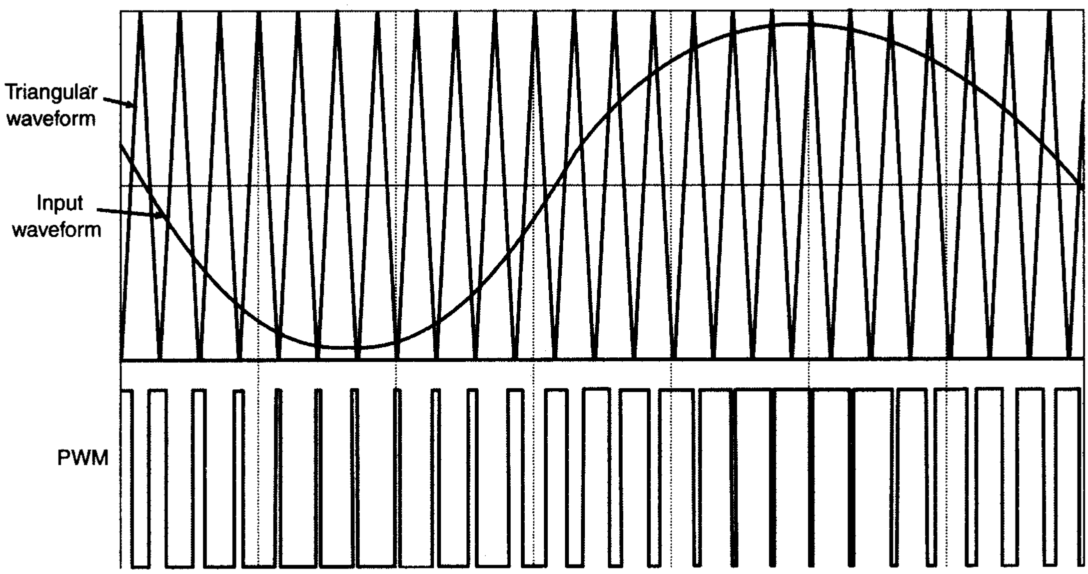



**Figure 10.27** | Generation of PWM waveform.

decreases. When the amplitude of input signal decreases, the time of HIGH state decreases and that of LOW state increases. PWM is usually generated by comparing the input signal with a triangular waveform (Figure 10.27) or a sawtooth waveform. Figure 10.28 shows the block diagram of the circuit used to generate the PWM waveform used to drive the switching transistors.

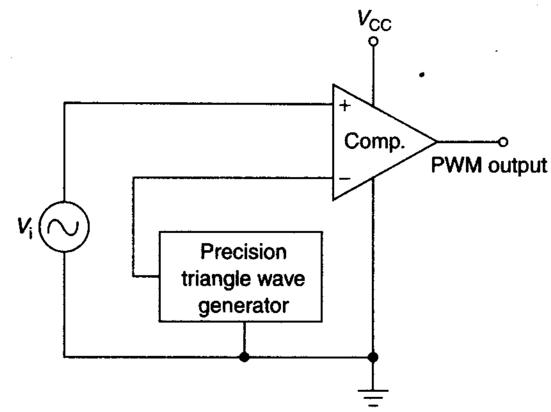
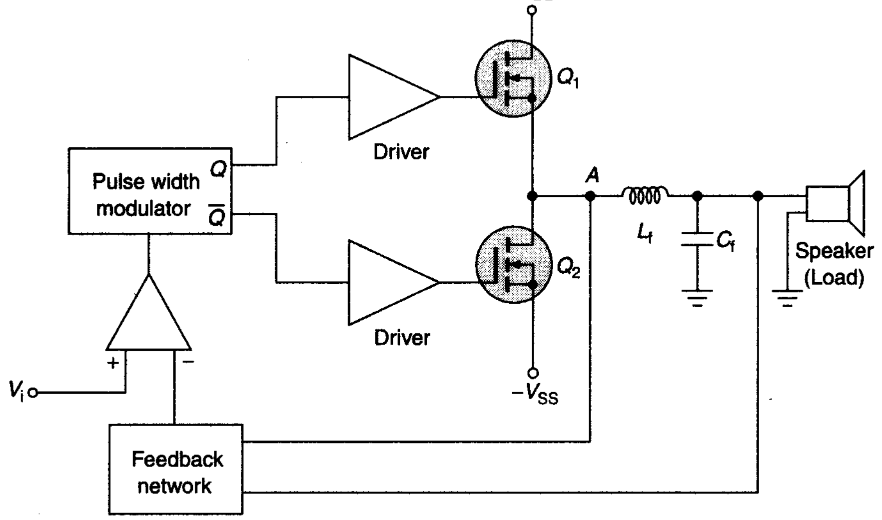



**Figure 10.28** Block diagram for PWM waveform generation.

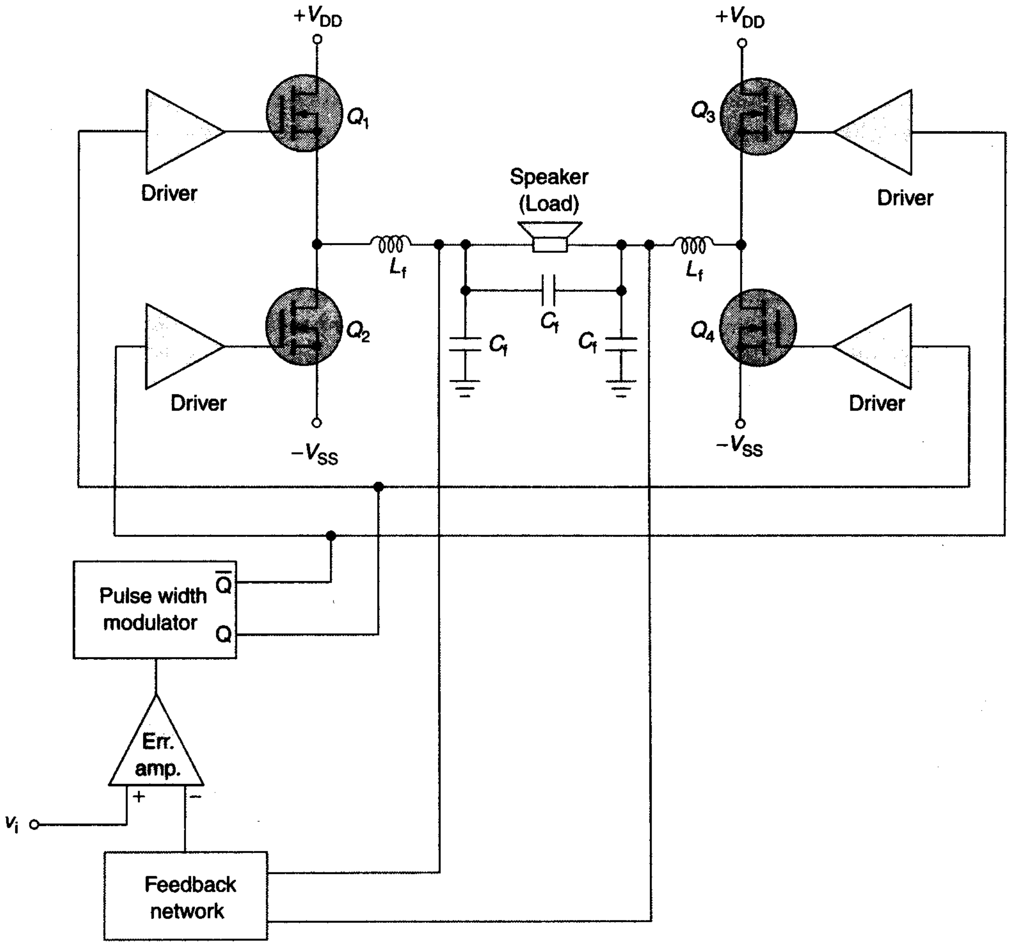
Class D amplifiers can be built using two basic topologies, namely, the half-bridge topology and the full-bridge topology. The half-bridge topology makes use of two active devices whereas the full-bridge topology makes use of four active devices. The half-bridge topology is simpler as compared to the full-bridge topology. Figure 10.29(a) shows the block diagram of a class D amplifier making use of half-bridge topology and Figure 10.29(b) shows the block diagram of a class D amplifier making use of full-bridge topology.

For the circuit in Figure 10.29(a), when the MOSFET $Q_1$ is in the ON-state and the MOSFET $Q_2$ is in the OFF-state, the inductor node (A) is connected to the supply voltage $V_{DD}$ and the current starts to increase through it. When the MOSFET $Q_2$ is in the ON-state and the MOSFET $Q_1$ is in the OFF-state,




(a)

**Figure 10.29** (a) Half-bridge topology for class D amplifiers; (b) full-bridge topology for class D amplifiers.




(b)

**Figure 10.29** *Continued.*

the current through the inductor $L_f$ starts to decrease. Therefore, the current waveform in the inductor $L_f$ is a triangular waveform. It may be mentioned here that only one of the MOSFETs is in the ON-state at any given time instant. The inductor $L_f$ in conjunction with the capacitor $C_f$ and the output load forms a low-pass filter that reconstructs the input signal by averaging the switching node voltage. Feedback is provided in order to reduce the timing errors.

## 10.7 Thermal Management of Power Transistors

Power transistors are used as active devices in power amplifiers. The maximum power that can be handled by an active device is related to its junction temperature, as the power dissipated by the device causes an increase in its junction temperature. The average power dissipated by the transistor can be approximated by the following equation:

$$P_D = V_{CE} I_C \tag{10.65}$$

This power dissipation is allowed upto a certain temperature, above which the value of power dissipation capability decreases linearly with increase in temperature such that it reduces to zero at the maximum device case temperature. Larger the power handled by the transistor, higher is the case temperature. It may be mentioned here that the silicon transistors offer larger maximum temperature ratings than germanium transistors. Figure 10.30 shows the typical maximum power derating curve for a silicon power transistor.

Therefore, the limiting factor in the power-handling capability of a transistor is its maximum permissible collector-junction temperature. Power transistors generally have large metal cases to provide a large area from which the heat generated by the device may be transferred. The power-handling capability of the device can be enhanced either by the use of heat sinks or by making use of thermoelectric coolers. With the heat sink, the transistor has a larger area to radiate heat into the air, thereby holding the case temperature to a much lower value than would be possible without the heat sink. The design of heat sink was discussed in detail in Chapter 3.

Thermoelectric coolers are solid-state heat pumps and are used in applications which require cooling below the ambient temperature, or where temperature cycling or precise temperature stabilization is required. Thermoelectric coolers are based on the Peltier effect, according to which the DC current applied across two dissimilar materials results in a temperature differential. These coolers transfer heat from one side of the active device to the other side against the temperature gradient. The cooling effect depends upon the amount of DC current and how well the heat from the hot side can be removed. Detailed description of thermoelectric cooling principle and application is beyond the scope of the present text.

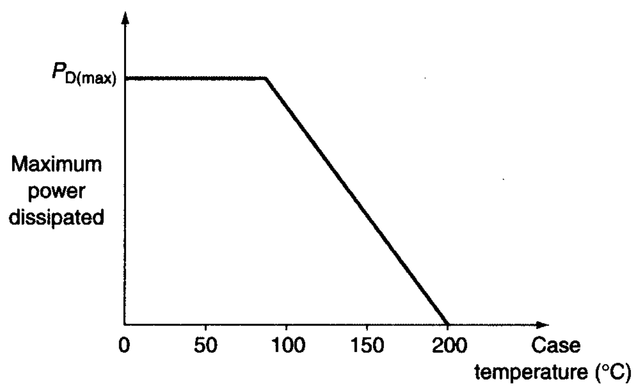



**Figure 10.30** Typical power derating curve for a silicon power transistor.

## KEY TERMS

Class A amplifier
Class AB amplifier
Class C amplifier
Class D amplifier
Crossover distortion
Efficiency
Harmonic distortion
Thermal resistance
Large signal amplifier
Power amplifier

## OBJECTIVE-TYPE EXERCISES

### Multiple-Choice Questions

1. Which of the following amplifier class suffers mainly from the problem of crossover distortion?
   a. Class A
   b. Class B
   c. Class AB
   d. Class C
2. Which of the following best describes a class A amplifier?
   a. High efficiency and high distortion
   b. Low efficiency and high distortion
   c. Low efficiency and low distortion
   d. High efficiency and low distortion
3. Heat sinks are used in power transistors to
   a. reduce the transistor power.
   b. reduce the junction temperature.
   c. increase the ambient temperature.
   d. increase the collector current.
4. In which class of amplifiers it is theoretically possible to achieve 100% efficiency?
   a. Class D
   b. Class A
   c. Class AB
   d. Class C
5. In which of the following amplifier classes, the active device operates for the whole of the input signal cycle?
   a. Class A
   b. Class AB
   c. Class B
   d. Class C
6. The maximum possible efficiency in series-fed class A and transformer coupled class A amplifiers are
   a. 50% and 25% respectively.
   b. 25% and 50% respectively.
   c. Both have 25%.
   d. Both have 50%.
7. How much percentage of maximum power can be dissipated by a transistor for an operating temperature equal to its maximum junction temperature?
   a. 100%
   b. 10%
   c. 50%
   d. 0%
8. A 10 W class B push–pull amplifier should have transistors with minimum individual collector dissipation of
   a. 2 W.
   b. 1 W.
   c. 5 W.
   d. 10 W.
9. Which of the following harmonic component has the main contribution in the total distortion in an amplifier?
   a. Third harmonic component
   b. Second harmonic component
   c. Fourth harmonic component
   d. Fifth harmonic component
10. Class C amplifiers are generally used as
    a. power amplifiers.
    b. RF amplifiers.
    c. audio amplifiers.
    d. none of these.

### Match the Following

Figure 10.31 shows six power amplifier configurations. Identify the amplifier configurations from the list given below:

a. Series-fed class A amplifier
b. Tuned class C amplifier
c. Half-bridge class D amplifier
d. Complementary-symmetry push–pull class B amplifier
e. Push–pull class A amplifier
f. Push–pull class B amplifier

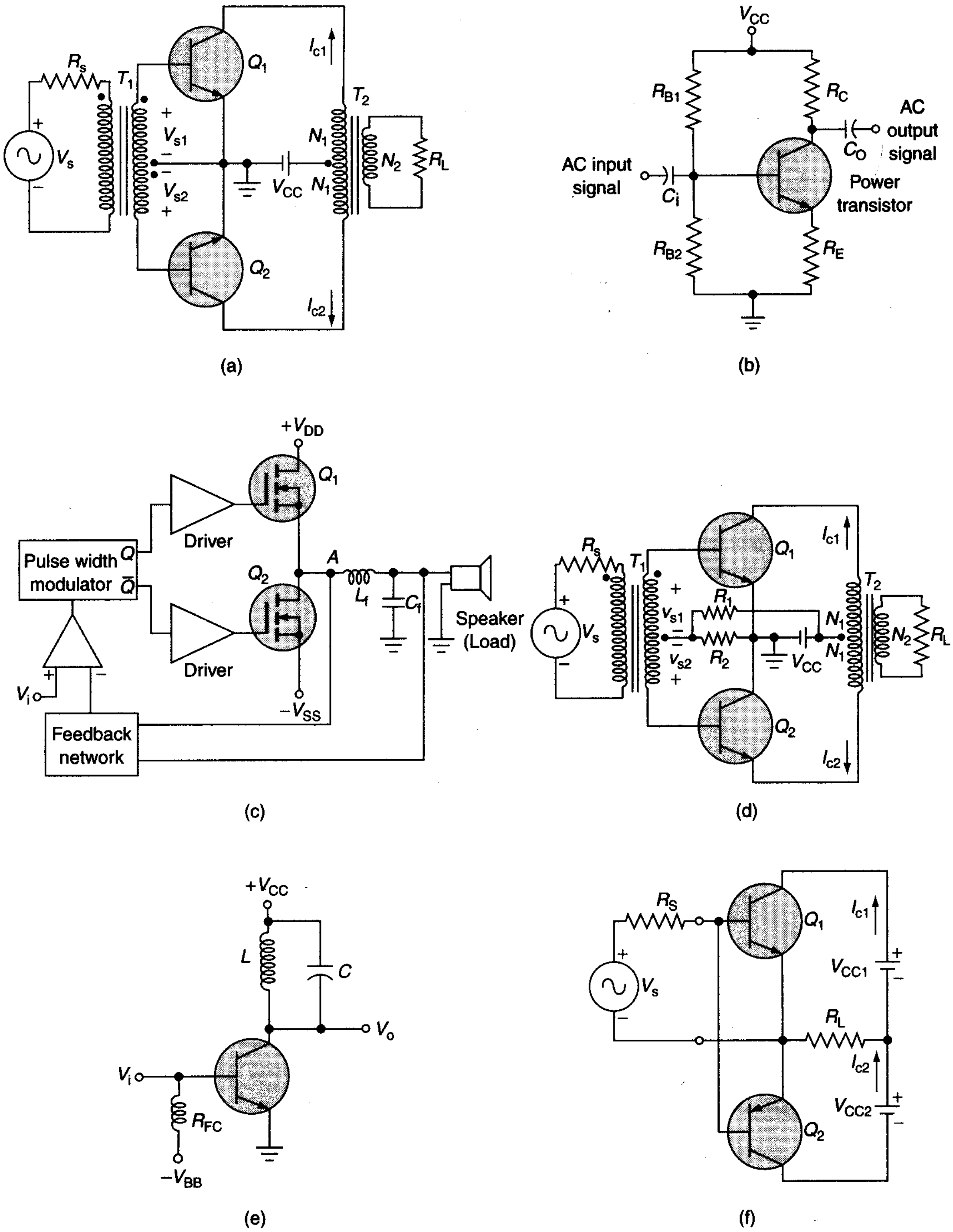



**Figure 10.31** | Match the following.

## REVIEW QUESTIONS

1. Explain how the active device in a class A amplifier dissipates less power when a signal is applied to it as compared to when there is no input signal.

2. What are the advantages offered by class A transformer-coupled amplifier over a direct-coupled class A amplifier?

3. Derive an expression to prove that the maximum efficiency in the case of a class B amplifier is 78.5%. What assumptions are made in calculating the maximum theoretical efficiency?

4. What is crossover distortion? What are the main factors that lead to crossover distortion and how can they be removed?

5. What are power amplifiers? How are they classified into different classes depending upon their mode of operation?

6. Sketch the circuit diagram of a quasi-complementary push–pull class B amplifier. Also draw the relevant current and voltage input and output waveforms.

7. Draw the circuit diagram of class A push–pull amplifier and explain its principle of operation.

8. What are class C amplifiers? How do they differ from class A and class B amplifiers?

9. Explain in detail how the use of heat sinks results in enhancement in the power dissipation capability of the transistor.

10. Write short notes on the following:
    a. Crossover distortion
    b. Harmonic distortion
    c. Complementary-symmetry push–pull class B amplifier
    d. Maximum efficiency of a class A amplifier

## PROBLEMS

1. A class B amplifier provides a 15 V peak output signal to 10 Ω load. The system operates on a power supply of 20 V. Determine the efficiency of the amplifier.

2. Calculate the harmonic distortion components for an output signal having fundamental amplitude of 6 V, second harmonic component of 0.5 V, third harmonic component of 0.1 V and fourth harmonic component of 0.04 V. Also calculate the total harmonic distortion.

3. Calculate the efficiency of the amplifier circuit in Figure 10.32. The input voltage applied is such that it produces a peak-to-peak base current of 16 mA. The emitter–base voltage of the transistor is equal to 0.7 V.

4. For the circuit of Figure 10.33, determine the input power, output power, power handled by each transistor and the circuit efficiency for an input of 15 V peak.

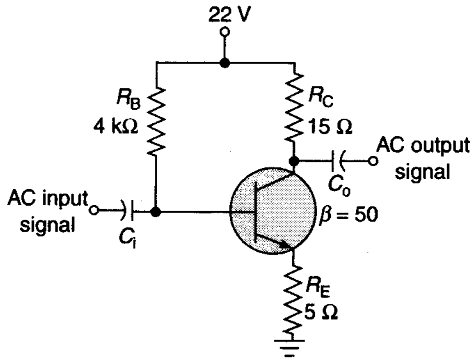



**Figure 10.32** | Problem 3.

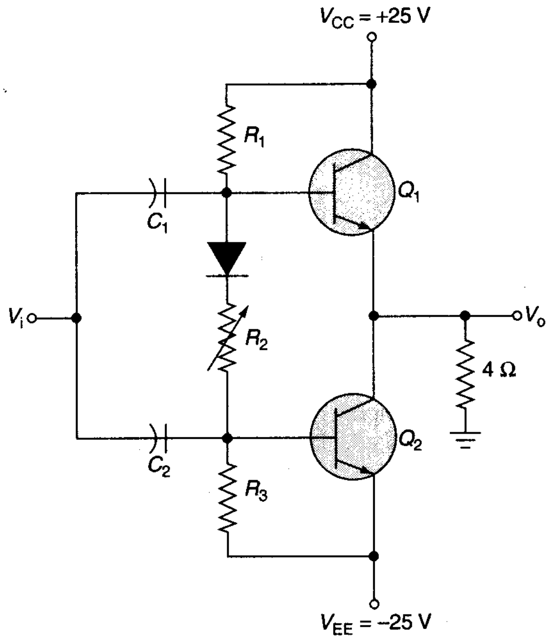



**Figure 10.33** | Problem 4.

# ANSWERS

## Multiple-Choice Questions

1. (b)
2. (c)
3. (b)
4. (a)
5. (a)
6. (b)
7. (d)
8. (a)
9. (b)
10. (b)

## Match the Following

Figure 10.31(a) – (f)
Figure 10.31(b) – (a)
Figure 10.31(c) – (c)
Figure 10.31(d) – (e)
Figure 10.31(e) – (b)
Figure 10.31(f) – (d)

## Problems

1. 58.87%
2. $D_2$=8.33%, $D_3$=1.67%, $D_4$=0.67%, $D$=8.52%
3. 21.82%
4. Input power = 59.71 W; output power = 28.125 W; power handled by each transistor = 15.794 W; efficiency = 47.1%

CHAPTER 11

# Feedback Amplifiers

## Learning Objectives

After completing this chapter, you will learn the following:

- Classification of amplifiers based on input and output parameters of interest.
- Types of negative feedback.
- Examples of voltage-series, current-series, voltage-shunt and current-shunt feedback in amplifier circuits.
- Effect of negative feedback on input and output impedances.
- Effect of negative feedback on gain (or conversion) and bandwidth parameters.
- Effect of negative feedback on noise performance.

Practical amplifier circuits always have some kind of negative feedback inherent to their operation. As we will see during the course of our discussion, presence of negative feedback in amplifiers brings along with it many advantages. In this chapter are introduced the basic concepts of negative feedback in amplifiers. The chapter begins with a brief outline on the classification of amplifiers based on input (current or voltage) and output (current or voltage) parameters of interest. This is followed by different types of feedback encountered in amplifiers. The classification of types of feedback is based on the type of output signal (current or voltage) fed back to the input and also the mode in which it is fed back (series or shunt). Also, each of the different types of feedback enhances the performance of a particular type of amplifier circuit. Effect of different types of negative feedback on the performance parameters of the amplifier is also discussed in detail. The text is adequately illustrated with practical circuits and a large number of solved examples.

## 11.1 Classification of Amplifiers

On the basis of the input and output parameters of interest, amplifiers are classified as voltage amplifiers, current amplifiers, transresistance amplifiers and transconductance amplifiers. In the case of a voltage amplifier, a small change in the input voltage produces a large change in the output voltage. Voltage gain, which is the ratio of the change in the output voltage to change in the input voltage, is the gain parameter. The input and output circuits of a voltage amplifier are represented by Thevenin's equivalent circuits as shown in Figure 11.1. In order that the circuit of Figure 11.1 acts like a true voltage amplifier, the input resistance of the amplifier represented by $R_i$ in Figure 11.1 should ideally be infinite so that the whole applied input voltage appears across the amplifier input irrespective of the value of source resistance $R_s$. In a practical voltage amplifier circuit, $R_i$ is much larger than $R_s$. Also, the output resistance $R_o$ should ideally be

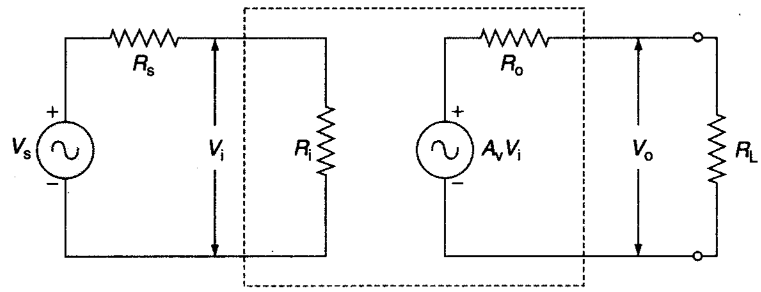



**Figure 11.1** | Equivalent circuit of voltage amplifier.

zero so that whole of amplified input voltage appears across the load resistance $R_L$ irrespective of the value of $R_L$. In a practical voltage amplifier circuit, $R_o$ is much smaller than $R_L$.

In the case of a current amplifier, a small change in the input current produces a large change in the output current. Current gain, which is the ratio of change in the output current to the change in the input current, is the gain parameter. The input and output circuits of a current amplifier are represented by Norton's equivalent circuits as shown in Figure 11.2. In order that the circuit of Figure 11.2 acts like a true current amplifier, the input resistance of the amplifier represented by $R_i$ in Figure 11.2 should ideally be zero so that the whole applied input current flows through the amplifier input irrespective of the value of source resistance $R_s$. In a practical current amplifier circuit, $R_i$ is much smaller than $R_s$. Also, the output resistance $R_o$ should ideally be infinite so that whole of amplified input current flows through the load resistance $R_L$ irrespective of the value of $R_L$. In a practical current amplifier circuit, $R_o$ is much larger than $R_L$.

In the case of a transresistance amplifier, a small change in the input current produces a large change in the output voltage. Ratio of the change in the output voltage to change in the input current is the gain parameter. The gain parameter has the units of resistance. The input and output circuits of a transresistance amplifier are, respectively, represented by Norton's and Thevenin's equivalent circuits as shown in Figure 11.3. For the circuit of Figure 11.3 to behave like a true transresistance amplifier, the input resistance of the amplifier represented by $R_i$ in Figure 11.3 should ideally be zero so that the whole of applied input current flows through the amplifier input irrespective of the value of source resistance $R_s$. In a practical transresistance amplifier circuit, $R_i$ is much smaller than $R_s$. Also, the output resistance $R_o$ should ideally be zero so that the output voltage appears across the load resistance $R_L$ irrespective of the value of $R_L$. In a practical circuit, $R_o$ is much smaller than $R_L$.

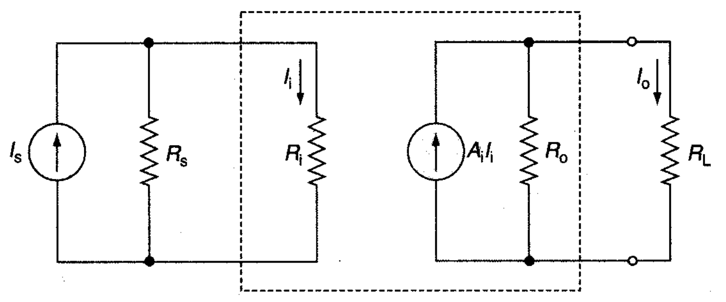



**Figure 11.2** | Equivalent circuit of current amplifier.

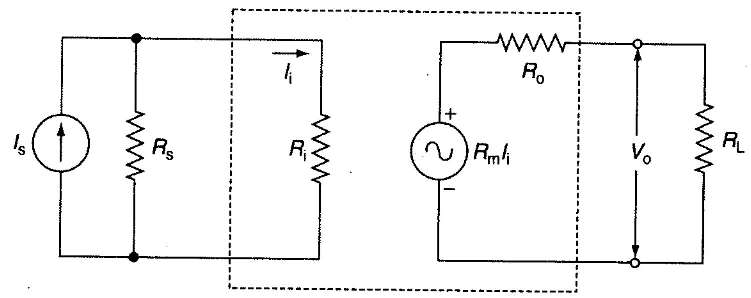



**Figure 11.3** | Equivalent circuit of transresistance amplifier.

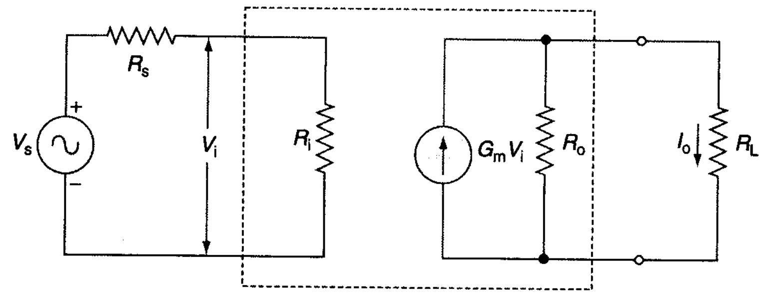



**Figure 11.4** | Equivalent circuit of transconductance amplifier.

In the case of a transconductance amplifier, a small change in the input voltage produces a large change in the output current. Ratio of the change in the output current to the change in the input voltage is the gain parameter. The gain parameter has the units of conductance. The input and output circuits of a transconductance amplifier are, respectively, represented by Thevenin's and Norton's equivalent circuits as shown in Figure 11.4. For the circuit of Figure 11.4 to behave like a true transconductance amplifier, the input resistance of the amplifier represented by $R_i$ in Figure 11.4 should ideally be infinite so that the whole of applied input voltage appears across the amplifier input irrespective of the value of source resistance $R_s$. In a practical transconductance amplifier circuit, $R_i$ is much larger than $R_s$. Also, the output resistance $R_o$ should ideally be infinite so that the output current flows through the load resistance $R_L$ irrespective of the value of $R_L$. In a practical circuit, $R_o$ is much larger than $R_L$.

## 11.2 Amplifier with Negative Feedback

In a negative-feedback amplifier, a sample of the output signal is fed back to the input and the feedback signal is combined with the externally applied input signal in a subtractor circuit as shown in Figure 11.5. As a result, the actual signal $X_i$ applied to the basic amplifier is the difference of the externally applied input signal $X_s$ and the feedback signal $X_f$. The generalized representation of input and output signals is intended to indicate that the signal could either be a current or a voltage signal.

Figure 11.6 shows a more elaborate schematic arrangement of an amplifier with negative feedback depicting different circuit blocks. The source of external input signal could either be a voltage source or a current source. The basic amplifier produces either an output current or voltage proportional to the input

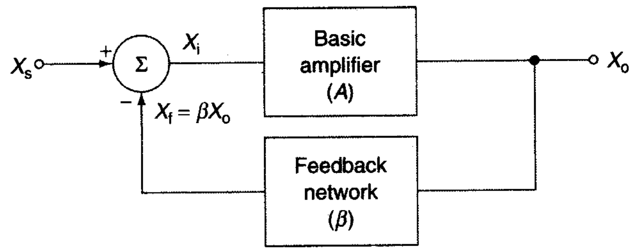



**Figure 11.5** | Block schematic of amplifier with negative feedback.

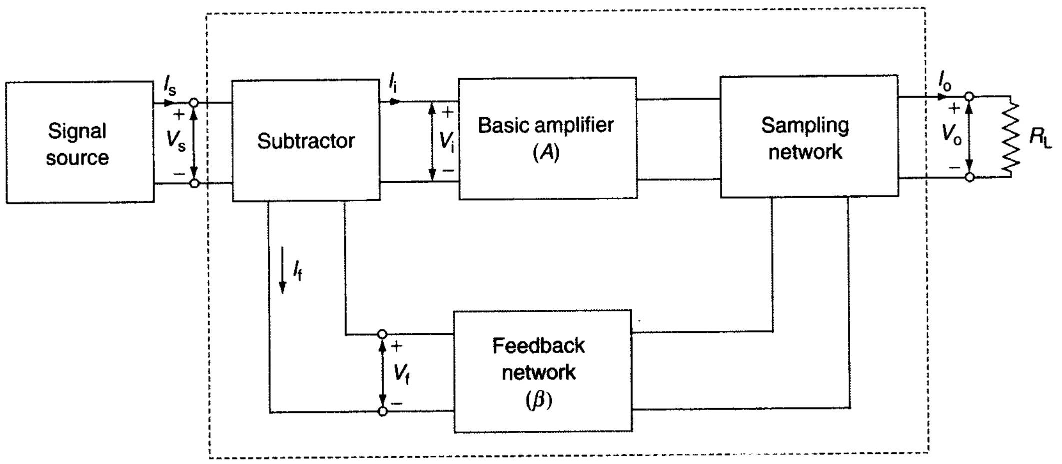



**Figure 11.6** | Schematic arrangement of various building blocks of a negative-feedback amplifier.

current or voltage. The nature of input and output parameters decides the nature of gain parameter of the amplifier as outlined earlier in Section 11.1.

The sampling network samples either the output voltage by connecting the input side of the feedback network in shunt across the output [Figure 11.7(a)] or the output current where the input side of the feedback network is connected in series with the output [Figure 11.7(b)]. The two sampling techniques are, respectively, known as voltage and current sampling.

The combining operation of externally applied input signal and the feedback signal is carried out in a differential amplifier (or subtractor) circuit. It is done in either of the two ways, namely, series connection [Figure 11.8(a)] and shunt connection [Figure 11.8(b)].

The gain parameter – which could be a voltage gain, current gain, transresistance or transconductance – has two values, namely, the gain of the basic amplifier without feedback and the gain of the amplifier with feedback. With reference to Figure 11.5, gain values without and with feedback are, respectively, given by $X_o/X_i$ and $X_o/X_s$.

## Effect of Negative Feedback on Gain

We will now derive an expression for gain with feedback in terms of gain without feedback (open-loop gain) and the feedback factor for the negative-feedback amplifier of Figure 11.5. The actual signal applied to the amplifier input $X_i$ is the difference of externally applied input signal $X_s$ and the feedback signal $X_f$. It is given by

$$X_i = X_s - X_f \tag{11.1}$$

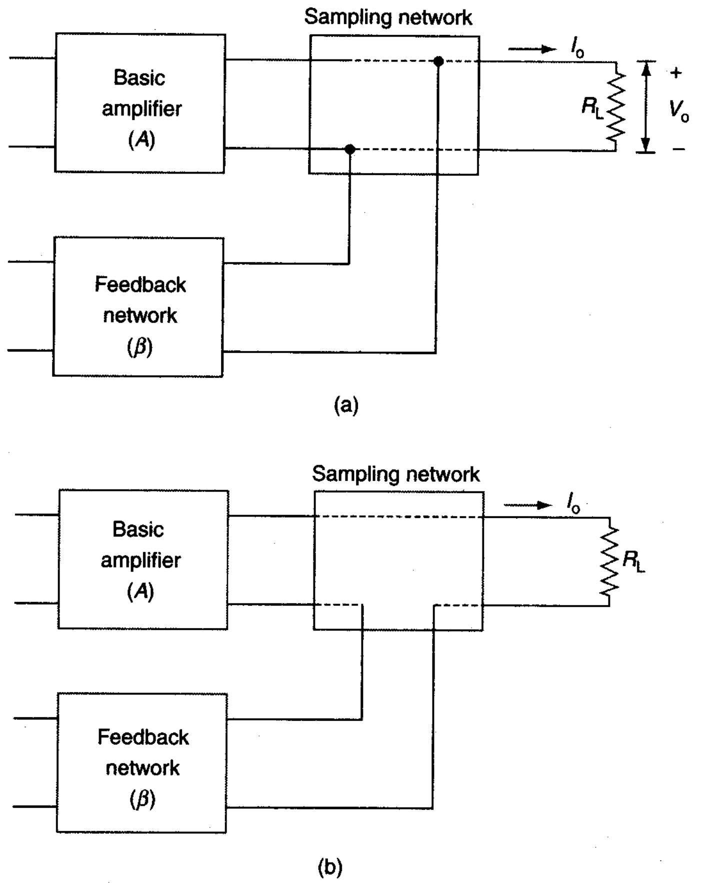



**Figure 11.7** Sampling in negative-feedback amplifiers: (a) Voltage sampling; (b) current sampling.

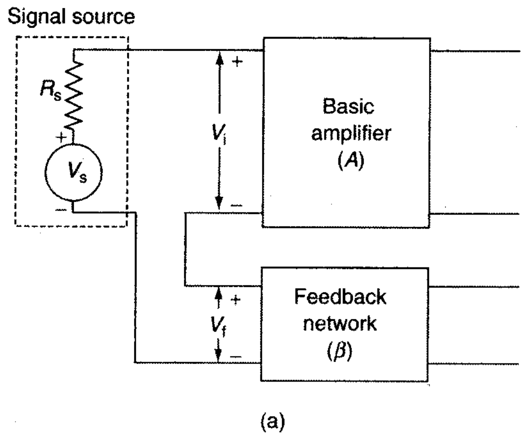



**Figure 11.8** Mixing operation in negative-feedback amplifiers: (a) Series input connection; (b) shunt input connection.

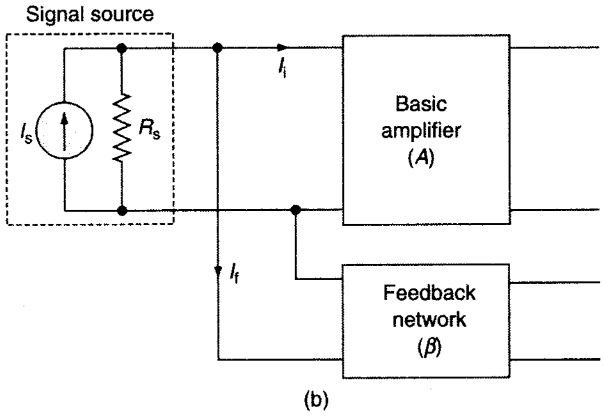



**Figure 11.8** | *Continued.*

Also,

$$X_f = \beta X_o, X_o = AX_i \text{ and } X_o = A_f X_s \quad (11.2)$$

Substituting for $X_i$ and $X_f$ in Eq. (11.1), we get

$$\frac{X_o}{A} = X_s - \beta X_o \quad (11.3)$$

Simplifying Eq. (11.3), we get the expression for $A_f$ as follows:

$$A_f = \frac{X_o}{X_s} = \frac{A}{1+\beta A} \quad (11.4)$$

An important parameter relevant to negative-feedback amplifiers is the loop gain. It is given by $(-\beta A)$ with minus sign indicating negative feedback. Another relevant term is the quantum of feedback expressed in decibels. It is given by

$$\text{dB of feedback} = 20\log\left|\frac{A_f}{A}\right| = 20\log\left|\frac{1}{1+\beta A}\right| \quad (11.5)$$

It may be mentioned here that for Eq. (11.4) and equations for input and output impedances derived in the case of amplifiers with negative feedback in the latter part of the chapter to be valid, three fundamental assumptions should be satisfied for feedback network. These include the following:

1. The input signal is transmitted to the output through the amplifier only and not through the feedback network. That is, forward transmission through feedback network is zero. This further implies that if the gain of the amplifier were reduced to zero, the output must drop to zero.
2. The feedback signal is transmitted from output to input through feedback network only. That is, reverse transmission through amplifier is zero.
3. The feedback factor is independent of load and source resistances.

**EXAMPLE 11.1** *Refer to the non-inverting amplifier circuit of Figure 11.9. The closed-loop gain of this amplifier is usually considered as being equal to $[(1 + (R_2/R_1)]$. Given that the open-loop gain of the opamp is 10,000, values of $R_1$ and $R_2$ are 10 kΩ and 100 kΩ, respectively, and the value of the applied input voltage $V_i$ is 1 V, determine the percentage error incurred in computing the output voltage by using this expression.*

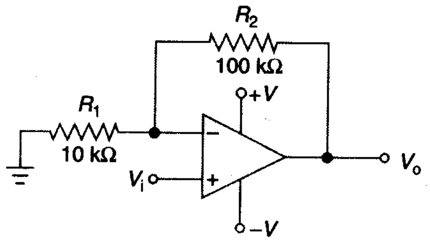



**Figure 11.9** | Example 11.1.

**Solution**

1. Closed-loop gain ignoring the effect of finite value of open loop gain = $1 + (R_2/R_1)$.
2. Substituting for $R_1$ and $R_2$, it comes out to be 11.
3. The output voltage is, therefore, 11 V.
4. The true value of closed-loop gain is given by $[A/(1 + \beta A)]$. $A$ is the open-loop gain and is equal to 10,000. $\beta$ is the feedback factor and is given by $[R_1/(R_1 + R_2)]$. This equals 1/11.
5. The true value of the closed-loop gain is, therefore, given by [10,000/{(1 + 10,000)/11}] = 10.988.
6. The true value of output voltage is, therefore, 10.988 V.
7. Percentage error = [(11 − 10.988)/10.988] × 100 = 0.11%.

## 11.3 Advantages of Negative Feedback

Introduction of negative feedback in amplifiers gives them many desirable features, which include the following:

1. It desensitizes the gain parameter to variation in values of components used in building the basic amplifier.
2. It increases the bandwidth.
3. It reduces distortion and increases linearity. It may be mentioned here that fundamentally all electronic devices (bipolar transistors, MOSFETs, etc.) that provide power gain are non-linear. Negative feedback allows the gain to be traded for higher linearity.
4. It reduces noise.
5. It can decrease or increase the input resistance depending upon the feedback topology. To be more specific, it depends upon how the feedback signal is mixed with the externally applied input signal. Series and shunt mixing, respectively, increase and decrease the input resistance.
6. It can decrease or increase the output resistance depending upon the feedback topology. More precisely, it depends upon how the output signal is sampled. Voltage sampling decreases the output resistance while current sampling increases it.

Most of the above-mentioned advantages (increased linearity, increased bandwidth, etc.) come at the cost of reduced gain. Closed-loop gain is smaller than the open-loop gain and the quantum of reduction depends upon the open-loop gain and the feedback factor. It, in fact, depends upon the product of the two called the loop gain.

Second disadvantage is the sensitivity of the input and output impedances to variation in open-loop gain. Third, negative feedback may sometimes lead to instability. But this can be taken care of in a carefully designed amplifier. Each one of these features is discussed in detail in the following sections.

## Desensitivity (or Stability) of Gain

Introduction of negative feedback makes the amplifier insensitive to variations in the values of the components and parameters of the active devices used in building the amplifier. The expression that relates percentage variation in the gain parameter of the amplifier with feedback to the percentage variation in the same without feedback can be derived as follows.

Gain with feedback $A_f$ is related to gain without feedback $A$ by Eq. (11.4) and the equation is reproduced as follows:

$$A_f = \frac{A}{1+\beta A}$$

Differentiating the above equation with respect to $A$, we get

$$\frac{dA_f}{dA} = \frac{d}{dA}\left[\frac{A}{1+\beta A}\right] = \frac{(1+\beta A)-\beta A}{(1+\beta A)^2} = \frac{1}{(1+\beta A)^2}$$

$$dA_f = \frac{dA}{(1+\beta A)^2} = \left(\frac{dA}{A}\right)\times\left[\frac{A}{(1+\beta A)^2}\right] = \left(\frac{dA}{A}\right)\times\left[\frac{A_f}{1+\beta A}\right]$$

$$\frac{dA_f}{A_f} = \left[\frac{1}{1+\beta A}\right]\times\left(\frac{dA}{A}\right)$$

$$\left|\frac{dA_f}{A_f}\right| = \left|\frac{1}{1+\beta A}\right|\times\left|\frac{dA}{A}\right| \qquad (11.6)$$

$|(1 + \beta A)|$ is called the desensitivity parameter $D$. Thus, percentage variation in gain with feedback is equal to the percentage variation in gain without feedback divided by the desensitivity parameter $D$. As a special case, when $\beta A >> 1$, $A_f$ equals $1/\beta$ and therefore becomes independent of the open-loop gain of the amplifier. In that case, the stability of the gain parameter depends only on the stability of the components used in the feedback network. By using stable passive components in the feedback network, it is indeed possible to have a stable amplifier that is immune to variations in parametric values of the amplifier. Remember that all this comes at the cost of reduced gain. This implies that the gain without feedback needs to be much larger than the desired value of the gain with feedback. If increase in instability of the amplifier on account of having a larger gain without feedback is not significantly higher than the instability of the amplifier without feedback at a lower gain value, then introduction of negative feedback certainly improves the stability by a significant margin.

## Effect on Bandwidth

Bandwidth increases with introduction of negative feedback. Increase in bandwidth results from the fact that amplifiers exhibit a constant gain-bandwidth product. Reduction in gain is, therefore, accompanied by increase in bandwidth. Bandwidth increases by the same desensitivity factor $D = 1 + \beta A$ by which the gain reduces. Bandwidth of amplifier with feedback is given by

$$(BW)_f = BW \times (1+\beta A) \qquad (11.7)$$

Figure 11.10 shows the gain-bandwidth curve of a typical operational amplifier. The gain-bandwidth product is given by the unity gain crossover frequency. We can see how bandwidth can be traded for closed-loop gain. Increase in bandwidth with introduction of negative feedback can also be explained as follows. As the gain rolls off with increase in frequency, reduced output signal amplitude means reduced feedback. Reduced negative feedback means increase in the magnitude of effective input signal which increases the output. In other words, output remains at its mid-band value up to a higher frequency.

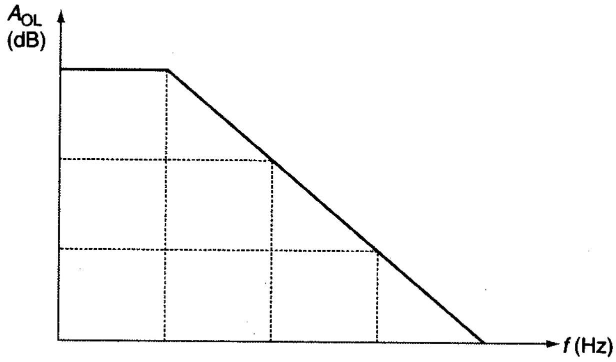



**Figure 11.10** | Effect of negative feedback on bandwidth.

## Effect on Non-Linear Distortion

As discussed in Chapter 10 on large signal amplifiers, non-linear distortion arises from the existence of non-linear transfer characteristics of the active device and large input and output signal swings that are large enough to drive the active device to operate in the non-linear region of its characteristics. Non-linear distortion can be assumed to be predominantly second harmonic distortion. Introduction of negative feedback reduces distortion provided that reduction in gain caused by negative feedback is compensated by increased gain in the preamplifier stages rather than introducing an additional gain stage. Remember that the non-linear distortion mainly occurs in the last stage of amplification and increasing the gain of previous stages does not add significantly to the overall distortion level.

It can be proved that non-linear distortion decreases by the desensitivity factor $D = 1 + \beta A$. Let us assume that the distortion levels without and with negative feedback are $D_2$ and $D_{2f}$, respectively. Suffix "2" implies second harmonic distortion. $D_2$ is the distortion contributed by the active device. In the presence of feedback, $D_{2f}$ appears as $-\beta AD_{2f}$ at the input of the amplifier. The following gives the expression for $D_{2f}$:

$$D_2 - \beta AD_{2f} = D_{2f} \tag{11.8}$$

$$D_{2f} = \frac{D_2}{1+\beta A} \tag{11.9}$$

Derivation of above expression makes use of superposition principle. Equation (11.9) is, therefore, valid only when the active device operates close to the linear region. This further means that the above expression is valid for small distortion levels. Also, small amount of additional distortion arising out of small fraction of distortion present at the output being fed back to the input is assumed to be negligible in deriving the above expression.

## Effect on Noise

Introduction of negative feedback acts on the noise generated in the amplifier in the same manner as it does on the non-linear distortion. Reduction in noise is governed by

$$N_f = \frac{N}{1+\beta A} \tag{11.10}$$

$N_f$ and $N$ are, respectively, the noise levels with and without negative feedback. However, if reduction in gain resulting from introduction of negative feedback is compensated by adding an amplifier stage, the overall system may turn out to be noisier than before. In order to benefit from the positive effect of negative feedback on noise, it is important that the reduced gain is compensated by readjustment of amplifier parameters for a higher gain without feedback so that the amplifier with feedback gives the desired gain.

## Effect on Input Resistance

Input resistance in the case of an amplifier with negative feedback is affected depending upon how the feedback signal (voltage or current) is connected to the source of external input signal. The input resistance increases if the feedback signal is connected in series with the source of input and decreases if the feedback signal is connected across it in shunt. Increase or decrease, as the case may be, is by desensitivity factor $D = 1 + \beta A$. Thus, in the case of voltage-series and current-series feedback, input resistance with feedback $R_{if}$ is given by

$$R_{if} = R_i \times (1+\beta A) \tag{11.11}$$

$R_i$ is the input resistance without feedback. Input resistance in the case of voltage-shunt and current-shunt feedback is given by

$$R_{if} = \frac{R_i}{1+\beta A} \tag{11.12}$$

Equations (11.11) and (11.12) will be derived subsequently in the relevant sections for different feedback topologies.

## Effect on Output Resistance

Negative feedback affects the output resistance of the amplifier depending upon the nature of the output signal fed back to the input. In case output voltage is sampled, output resistance decreases. In case output current or a voltage proportional to the output current is sampled, output resistance increases. Output resistance with feedback $R_{of}$ in the case of current-series and current-shunt feedback is given by Eq. (11.13). $R_o$ is the output resistance without feedback. Output resistance in the case of voltage-series and voltage-shunt feedback is given by Eq. (11.14):

$$R_{of} = R_o \times (1+\beta A) \tag{11.13}$$

$$R_{of} = \frac{R_o}{1+\beta A} \tag{11.14}$$

Equations (11.13) and (11.14) will be derived subsequently in the relevant sections for different feedback topologies.

**EXAMPLE 11.2**

*An amplifier without feedback has a voltage gain of 100. The designer decides to use 10% negative feedback to bring the non-linear distortion to an acceptable level. Determine the gain of the amplifier in the presence of feedback. If the desired value of gain with feedback is 50, what should in that case be the feedback factor?*

**Solution**

1. Gain with feedback = $A/(1 + \beta A)$, where $A$ is the gain without feedback and $\beta$ is the feedback factor.
2. Gain with feedback = $100/(1 + 0.1 \times 100) = 100/11 = 9.09$.
3. If the gain with feedback needs to be equal to 50, then $50 = 100/(1 + \beta \times 100)$.
4. Solving for $\beta$, we get $100\beta + 1 = 2$. That is, $\beta = 0.01$ or 1%.

**EXAMPLE 11.3**

*The total harmonic distortion of an amplifier reduces from 10% to 1% on introduction of 10% negative feedback. Determine the open-loop and closed-loop gain values.*

**Solution**

1. Distortion (with feedback) = Distortion (without feedback)/(1 + $\beta A$), where $A$ is the open-loop gain.
2. $0.01 = 0.1/(1 + 0.1A)$, which gives $1 + 0.1A = 10$ or $A = 90$.
3. Therefore, open-loop gain = 90.
4. Closed-loop gain = $90/(1 + 0.1 \times 90) = 9$.

**EXAMPLE 11.4** *A negative-feedback amplifier has –20 dB of feedback. Determine the loop gain.*

**Solution**

1. Negative feedback in dB = 20 log (Gain with feedback/Gain without feedback) = –20.
2. Log (Gain with feedback/Gain without feedback) = –1.
3. Gain with feedback/Gain without feedback = 0.1 = $1/(1 + \beta A)$.
4. This gives $(1 + \beta A) = 10$, or $\beta A = 9$.
5. Loop gain = $\beta A = 9$.

**EXAMPLE 11.5** *A voltage amplifier is characterized by voltage gain of 100, input resistance of 100 kΩ and output resistance of 1 kΩ in the absence of any feedback. Find the modified values of these parameters if 5% of output voltage were feedback in series and phase opposition with the input.*

**Solution**

1. Open-loop gain $A = 100$ and feedback factor $\beta = 0.05$.
2. Closed-loop gain = $100/(1 + 0.05 \times 100) = 100/6 = 16.67$.
3. Since the feedback signal is in series with input, modified input resistance is given by $100 \times 10^3 \times (1 + 0.05 \times 100) = 600$ kΩ.
4. Since it is fraction of the output voltage that is fed back to the input, modified output resistance is given by $1 \times 10^3/(1 + 0.05 \times 100) = 1000/6\ \Omega = 166.67\ \Omega$.

**EXAMPLE 11.6** *A voltage amplifier is a cascade arrangement of three identical amplifier stages each having an open loop gain of $A_1$. The output from the final stage is fed back in series and phase opposition with the input of the first stage to get an overall closed loop gain of $A_f$. Derive an expression for overall open-loop gain A in terms of open loop gain stability $|dA_1/A_1|$ of individual stages and overall closed loop gain stability $|dA_f/A_f|$.*

**Solution**

1. Open-loop gain, $A = A_1^3$. Therefore, $dA = 3A_1^2 dA_1$.
2. This gives $dA/A = 3A_1^2 \times (dA_1/A_1^3) = 3dA_1/A_1$.
3. Now, $dA_f/A_f = [1/|1+\beta A|] \times (dA/A)$.
4. $|dA_f/A_f| = [1/|1+\beta A|] \times 3 \times dA_1/A_1$.
5. Also, $1/|1+\beta A| = A_f/A$.
6. Therefore, $A = 3A_f \times [|dA_1/A| / |dA_f/A_f|]$.

## 11.4 Feedback Topologies

On the basis of the nature of sampled signal and the mode in which it is fed back to the input, there are four feedback topologies. These include the following:

1. Voltage-series feedback topology (also known as series–shunt topology).
2. Voltage-shunt feedback topology (also known as shunt–shunt topology).
3. Current-series feedback topology (also known as series–series topology).
4. Current-shunt feedback topology (also known as shunt–series topology).

Each one of these feedback topologies is described in the following sections.

## 11.5 Voltage-Series (Series–Shunt) Feedback

In the case of voltage-series (series–shunt) feedback, output voltage is sampled and mixed in series with the externally applied input signal. Figure 11.11 shows the block schematic arrangement of a generalized feedback amplifier with voltage-series feedback. Figure 11.12 shows the equivalent circuit for the schematic arrangement of Figure 11.11.

### *Gain*

The gain parameter in this case is the voltage gain. Equation (11.15) gives the expression for voltage gain with feedback in terms of gain without feedback and the feedback factor:

$$A_{Vf} = \frac{A_V}{1+\beta A_V} \tag{11.15}$$

where $A_V$ and $A_{Vf}$ are the voltage gains without and with feedback taking into account the load resistance $R_L$.

### *Input Resistance*

Expression for input resistance is derived as follows:

$$R_{if} = \frac{V_s}{I_s} = \frac{V_s}{I_i} \tag{11.16}$$

Now $I_i = V_i/R_i$ and $V_i = V_s - \beta V_o$. Substituting for $V_i$, we get

$$I_i = \frac{V_s - \beta V_o}{R_i} \tag{11.17}$$

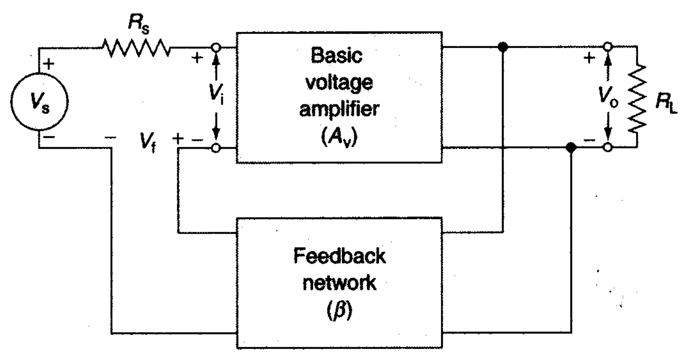



**Figure 11.11** Schematic arrangement of voltage-series (series–shunt) feedback.

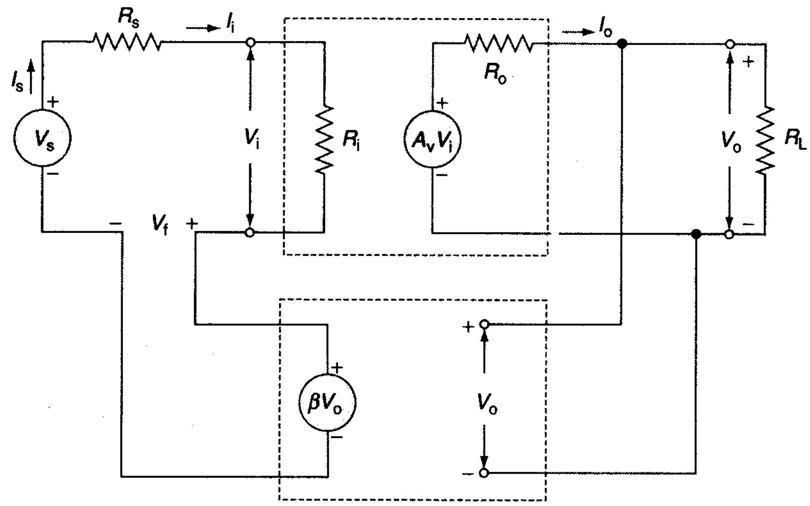



**Figure 11.12** | Equivalent circuit for schematic arrangement of Figure 11.11.

Substituting for $I_i$ from Eq. (11.17) in Eq. (11.16), we get

$$R_{if} = V_s \times \frac{R_i}{V_s - \beta V_o} = \frac{R_i}{1-\beta(V_o/V_s)}$$

Since $V_o/V_s = A_{Vf}$, where $A_{Vf}$ is the voltage gain with feedback taking load resistance ($R_L$) into account, therefore

$$R_{if} = \frac{R_i}{1-\beta A_{Vf}} \tag{11.18}$$

Now

$$A_{Vf} = \frac{A_V}{1+\beta A_V}$$

where $A_V$ is the voltage gain without feedback taking load resistance ($R_L$) into account. It is given by

$$A_V = A_v \times \left(\frac{R_L}{R_o + R_L}\right)$$

$A_v$ is the open-circuited voltage gain without feedback, that is, the voltage gain without feedback without taking load resistance ($R_L$) into account. Therefore

$$R_{if} = \frac{R_i}{1-(\beta A_V/1+\beta A_V)} = R_i \times (1+\beta A_V) \tag{11.19}$$

### *Output Resistance*

Expression for output resistance is derived by considering $V_s = 0$ or $I_s = 0$, letting load resistance $R_L = \infty$ and applying a voltage $V$ across the output. Ratio of applied voltage $V$ to the resulting current $I$ gives the output resistance.

$$I = \frac{V - A_v V_i}{R_o}$$

For $V_s = 0$, $V_i = -V_f = -\beta V$. Therefore

$$I = \frac{V + \beta A_v V}{R_o}$$

$$R_{of} = \frac{V}{I} = \frac{R_o}{1 + \beta A_v} \qquad (11.20)$$

Remember that $R_{of}$ is the output resistance with feedback with $R_L = \infty$. From Eq. (11.20) we see that it is equal to $R_o$ divided by the factor $(1 + \beta A_v)$, which contains the voltage gain $A_v$ and not $A_V$.

Considering the effect of load resistance $R_L$, the output resistance with feedback $R_{of'}$ is given by parallel combination of $R_{of}$ and $R_L$. $R_{of'}$ can also be expressed by

$$R_{of'} = \frac{R_{o'}}{1 + \beta A_V}$$

where

$$R_{o'} = \frac{R_o \times R_L}{R_o + R_L} \qquad (11.21)$$

Remember, $A_V$ is the voltage gain without feedback taking load resistance ($R_L$) into account.

### *Practical Circuits*

Common-drain amplifier (source follower) circuit as shown in Figure 11.13 is an example of voltage-series feedback as the output voltage appearing across the source resistor ($R_S$) is fed back in series with the externally applied input signal. The bipolar transistor counterpart of source follower is the emitter–follower, which is another example of voltage-series feedback. Source–follower and emitter–follower are examples of 100% feedback with the result that the closed-loop gain is approximately unity.

A non-inverting amplifier circuit configured around an opamp (Figure 11.14) is yet another example of voltage-series feedback. In this case, the feedback voltage ($V_f$) appears across $R_1$. It equals $V_o \times R_1/(R_1 + R_2)$. The feedback voltage $V_f$ is in series with externally applied input voltage $V_s$. Also, it is in phase opposition with the external input as $V_s$ and $V_f$ are, respectively, applied to non-inverting and inverting inputs of the opamp.

Another non-inverting amplifier configuration with voltage-series feedback is shown in Figure 11.15. The circuit shown is a cascade arrangement of two common-emitter amplifier stages. The emitter resistor of the first stage is split into two series-connected resistors. Relatively much smaller of the two resistance values is un-bypassed and forms a part of the potential divider arrangement with another resistor connected from the output. The feedback factor is given by $R_1/(R_1 + R_2)$.

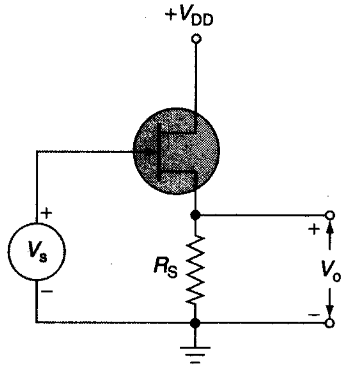



**Figure 11.13** | Source–follower amplifier circuit.

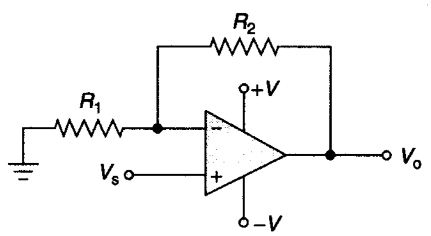



**Figure 11.14** | Voltage-series feedback in non-inverting amplifier.

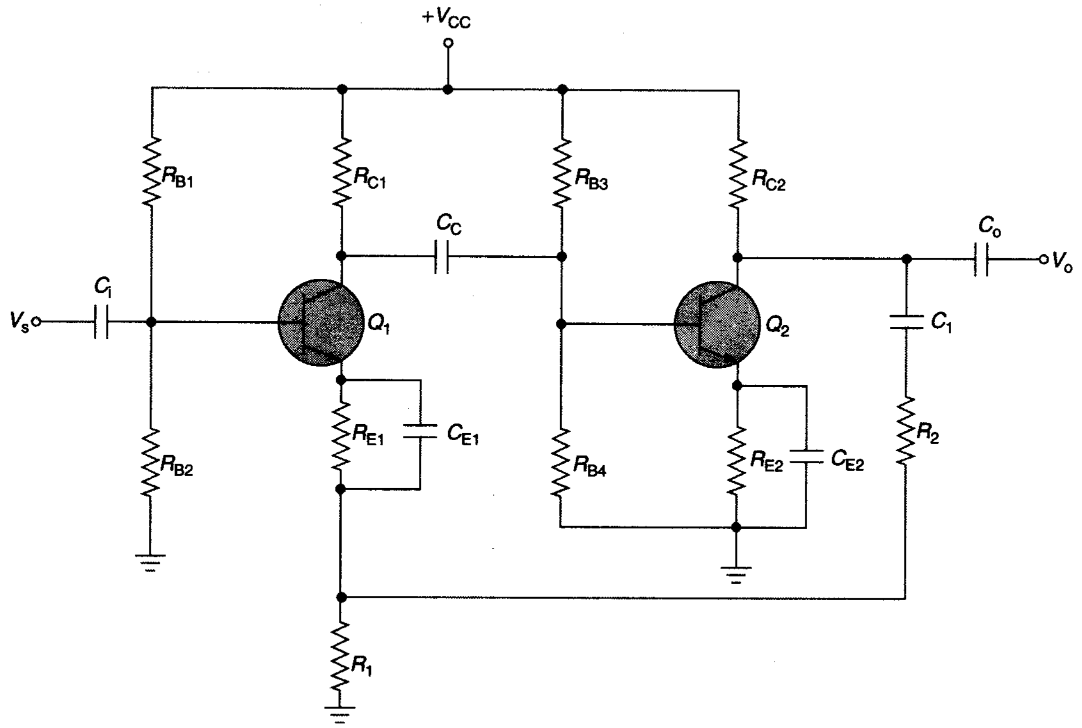



**Figure 11.15** | Two-stage common-emitter amplifier with voltage-series feedback.

**EXAMPLE 11.7** *For the common-collector circuit of Figure 11.16, determine expressions for voltage gain, input resistance and output resistance in the presence of negative feedback.*

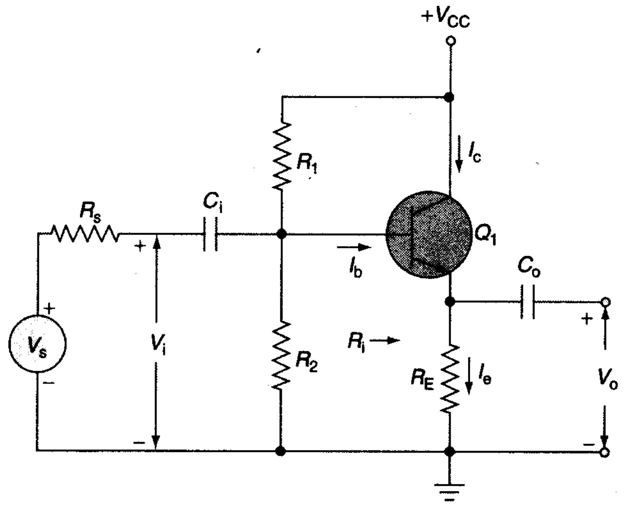



**Figure 11.16** | Example 11.7.

**Solution**

1. Voltage gain, $A_V$ = Output voltage/Input voltage.
2. Output voltage = $h_{fe} \times I_b \times R_E$ (assuming $I_c \cong I_e$) and input voltage = $V_s$.
3. Therefore, $A_V = h_{fe} \times I_b \times R_E / V_s \cong h_{fe} \times R_E/(R_s + h_{ie})$ as $V_s/I_b = R_s + h_{ie}$.
4. Desensitivity factor $D$ is given by the expression

$$D = 1 + \beta A_V = 1 + \frac{h_{fe} \times R_E}{R_s + h_{ie}}$$
$$= (R_s + h_{ie} + h_{fe}R_E) / (R_s + h_{ie})$$

as $\beta = 1$.

5. Therefore, voltage gain with feedback is given by

$$A_{Vf} = (h_{fe}R_E)/(R_s + h_{ie} + h_{fe}R_E)$$

6. Input impedance, $R_i = R_s + h_{ie}$.
7. Therefore, input resistance with feedback is given by

$$R_{if} = R_i \times D = R_s + h_{ie} + h_{fe}R_E$$

8. We are interested in the resistance looking into the emitter. Hence $R_E$ is considered as an external load. Therefore, $R_{of} = \infty$.
9. Since $R_E$ is the load resistance in the present case, therefore, $R_{o'} = R_E$.
10. This gives $R_{of'} = R_{o'}/D = R_E \times (R_s + h_{ie})/(R_s + h_{ie} + h_{fe}R_E)$.
11. $R_{of} = \lim_{R_E \to \infty} R_{of'} = (R_s + h_{ie})/h_{fe}$.

**EXAMPLE 11.8**

*For the source follower circuit of Figure 11.17, determine the voltage gain, input resistance and output resistance in the presence of voltage-series feedback.*

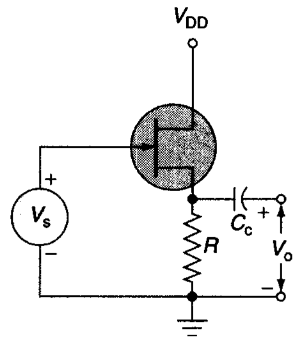



**Figure 11.17** | Example 11.8.

**Solution**

1. If there were no negative feedback, voltage gain $A_V$ would have been expressed by

$$A_V = V_o/V_s = \{g_m \times (r_d \| R) \times V_s\}/V_s$$
$$= (g_m \times r_d \times R \times V_s)/\{(r_d + R) \times V_s\}$$

Therefore, $A_V = (g_m \times r_d \times R)/(r_d + R)$.

2. Also $g_m \times r_d = \mu$. This gives $A_V = (\mu \times R)/(r_d + R)$.
3. Desensitivity factor,

$$D = 1 + \beta A_V = 1 + [(\mu \times R)/(r_d + R)]$$
$$= [\{r_d + (1+\mu) \times R\}/(r_d + R)]$$

as $\beta = 1$

4. Therefore, $A_{Vf} = A_V/D = [(\mu \times R)/\{r_d + (1+\mu)R\}]$.
5. The input impedance of an FET for all practical purposes is infinite. Therefore, input resistance with feedback $R_{if} = R_i \times D$ is also infinite.
6. Output resistance with feedback $R_{of} = R_o/(1+\beta A_v)$. Here $R_o = r_d$ considering the load resistance $R$ to be an open circuit. Also, $A_v = \mu$. Therefore, $R_{of} = r_d/(1+\mu)$.
7. Now $R_{of'} = R_{o'}/D$, $R_{o'} = (r_d \times R)/(r_d + R)$ and $D = \{r_d + (1+\mu) \times R\}/(r_d + R)$. $R_{of'}$ is the output resistance with feedback taking load resistance into account.
8. This gives

$$R_{of'} = [(r_d \times R)/\{r_d + (1+\mu) \times R\}]$$

9. The value of $R_{of}$ can also be determined by using the expression

$$R_{of} = \lim_{R\to\infty} R_{of'} = \lim_{R\to\infty}[(r_d \times R)/\{r_d + (1+\mu) \times R\}] = r_d/(1+\mu)$$

**EXAMPLE 11.9**

*For the opamp based non-inverting amplifier circuit of Figure 11.18, determine the voltage gain and the input impedance in the presence of feedback given that open-loop gain and the input impedance of the opamp are 80 dB and 1 MΩ, respectively.*

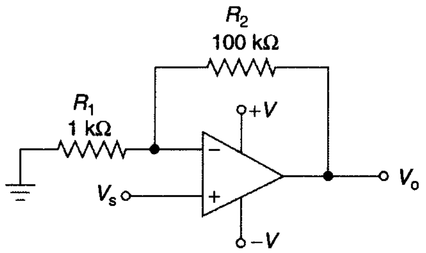



**Figure 11.18** | Example 11.8

**Solution**

1. Open-loop gain, $A$ = 80 dB = 10000.
2. Feedback factor, $\beta = R_1/(R_1 + R_2) = 1 \times 10^3/(1 \times 10^3 + 100 \times 10^3) = 1/101$.
3. Therefore desensitivity factor, $D = 1 + \beta A = 1 + 10000/101 = 100.0099$.
4. Gain with feedback = 10000/100.0099 = 99.99.
5. Input impedance of the opamp = 1 MΩ.
6. Since it is a case of series feedback, input impedance would increase by the desensitivity factor.
7. That is, input impedance with feedback = $1 \times 10^6 \times 100.0099$ = 100 MΩ.

## 11.6 Voltage-Shunt (Shunt–Shunt) Feedback

In the case of voltage-shunt (shunt–shunt) feedback, output voltage is sampled and mixed in shunt with the externally applied input signal. Figure 11.19 shows the block schematic arrangement of a generalized feedback amplifier with voltage-shunt feedback. Figure 11.20 shows the equivalent circuit for the schematic arrangement of Figure 11.19.

### *Gain*

The gain parameter in this case is the transresistance. Equation (11.22) gives the expression for transresistance with feedback $R_{Mf}$ in terms of transresistance without feedback $R_M$ and the feedback factor $\beta$.

$$R_{Mf} = \frac{R_M}{1+\beta R_M} \tag{11.22}$$

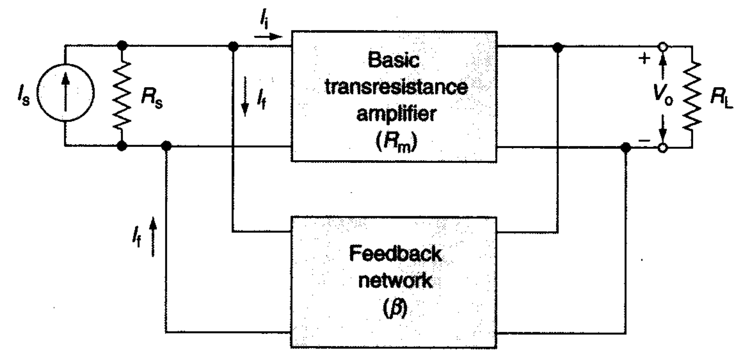



**Figure 11.19** | Schematic arrangement of voltage-shunt (shunt–shunt) feedback.

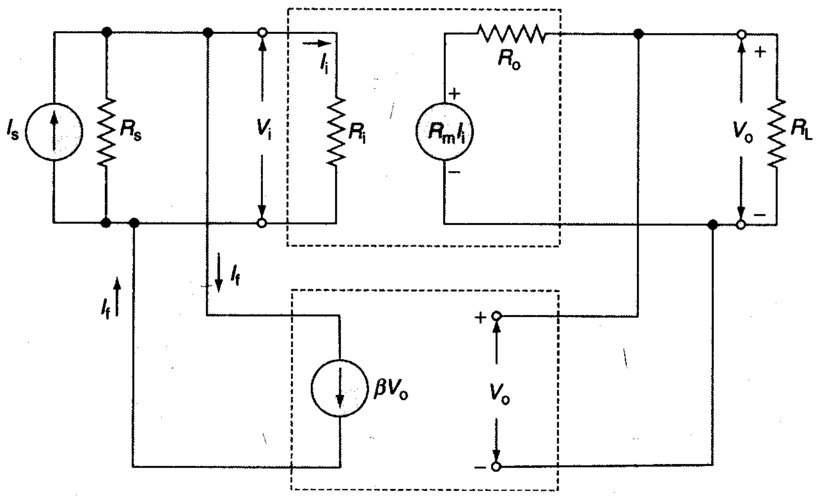



**Figure 11.20** | Equivalent circuit for schematic arrangement of figure 11.19.

### *Input Resistance*

Expression for input resistance is derived as follows.

$$R_{\text{if}} = \frac{V_{\text{i}}}{I_{\text{s}}} = \frac{R_{\text{i}} \times I_{\text{i}}}{I_{\text{s}}} \tag{11.23}$$

Substituting $I_{\text{s}} = I_{\text{i}} + I_{\text{f}}$ in Eq. (11.23), we get

$$R_{\text{if}} = \frac{R_{\text{i}} \times I_{\text{i}}}{I_{\text{i}} + I_{\text{f}}} = \frac{R_{\text{i}}}{(I_{\text{i}} + I_{\text{f}})/I_{\text{i}}} = \frac{R_{\text{i}}}{1 + (I_{\text{f}}/I_{\text{i}})}$$

Now, $I_{\text{f}} = \beta V_{\text{o}}$. Therefore, $R_{\text{if}} = R_{\text{i}} / [1 + (\beta V_{\text{o}}/I_{\text{i}})]$.

Also, $V_{\text{o}}/I_{\text{i}} = R_{\text{M}}$. This gives

$$R_{\text{if}} = R_{\text{i}}/(1 + \beta R_{\text{M}}) \tag{11.24}$$

$R_{\text{M}}$ is the transresistance taking load resistance $R_{\text{L}}$ into account and is given by

$$R_{\text{M}} = R_{\text{m}} \times \left( \frac{R_{\text{L}}}{R_{\text{o}} + R_{\text{L}}} \right)$$

where $R_{\text{m}}$ is the open-circuit transresistance, that is, without taking the load resistance $R_{\text{L}}$ into account and can be expressed as

$$R_{\text{m}} = \lim_{R_{\text{L}} \to \infty} R_{\text{M}}$$

### *Output Resistance*

Expression for output resistance is derived by considering $I_{\text{s}} = 0$, letting load resistance $R_{\text{L}} = \infty$ and applying a voltage $V$ across the output. Ratio of applied voltage $V$ to the resulting current $I$ gives the output resistance.

$$I = (V - R_{\text{m}} I_{\text{i}})/R_{\text{o}}$$

For $I_{\text{s}} = 0$, $I_{\text{i}} = -I_{\text{f}} = -\beta V_{\text{o}} = -\beta V$. Therefore,

$$I = (V + \beta R_{\text{m}} V)/R_{\text{o}} = V(1 + \beta R_{\text{m}})/R_{\text{o}}$$

$$R_{\text{of}} = V/I = R_{\text{o}}/(1 + \beta R_{\text{m}}) \tag{11.25}$$

Remember that $R_{\text{of}}$ is the output resistance with feedback with $R_{\text{L}} = \infty$. Considering the effect of load resistance $R_{\text{L}}$, the output resistance with feedback $R_{\text{of'}}$ is given by parallel combination of $R_{\text{of}}$ and $R_{\text{L}}$. $R_{\text{of'}}$ can also be expressed by

$$R_{\text{of'}} = R_{\text{o'}}/(1 + \beta R_{\text{M}}) \tag{11.26}$$

where

$$R_{\text{o'}} = (R_{\text{o}} \times R_{\text{L}})/(R_{\text{o}} + R_{\text{L}})$$

### *Practical Circuits*

Common-emitter amplifier with collector-to-base feedback as shown in Figure 11.21 is an example of voltage-shunt feedback. As is evident from the circuit diagram, a current proportional to the output voltage is fed back in the shunt with the source of input signal. Since $V_{\text{o}} >> V_{\text{i}}$, the feedback current is equal to $V_{\text{o}}/R_{\text{B}}$ and the feedback factor $\beta$ equals $1/R_{\text{B}}$. The amplifier nearly achieves the characteristics of a transresistance amplifier.

Figure 11.22 shows the opamp version of voltage-shunt feedback topology. The circuit shown is that of an inverting amplifier. The input current given by $V_{\text{s}}/R_{\text{s}}$ flows through the feedback resistance $R_{\text{f}}$ to produce an output voltage equal to $-I_{\text{s}} \times R_{\text{f}}$. The gain parameter is the transresistance $R_{\text{f}}$.

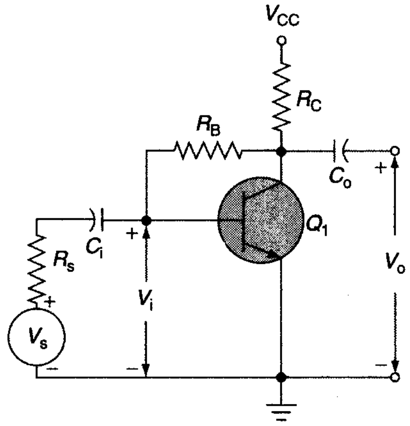



Figure 11.21 | Common-emitter amplifier circuit with voltage-shunt feedback.

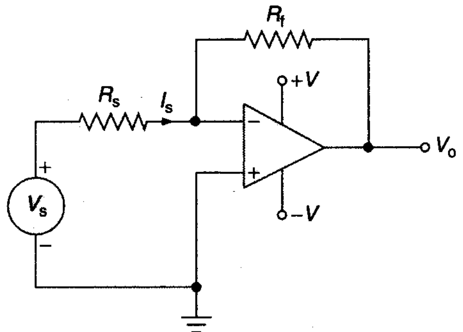



Figure 11.22 | Inverting amplifier with voltage-shunt feedback.

## EXAMPLE 11.10

*Refer to the opamp-based inverting amplifier circuit of Figure 11.23. Identify the type of negative feedback. Determine the transimpedance gain, the input impedance and output impedance of the amplifier, given that transimpedance, input impedance and output impedance parameters of the opamp are 100 MΩ, 10 MΩ and 100 Ω, respectively.*

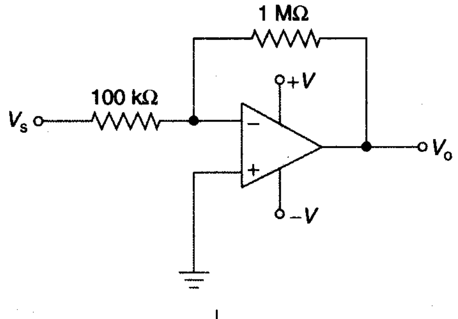



Figure 11.23 | Example 11.10.

### Solution

1. The feedback is of voltage-shunt type.
2. Transimpedance, $R_M = 100$ MΩ $= 100 \times 10^6$ Ω $= 10^8$ Ω
3. Feedback factor, $\beta = 1/10^6\ (\Omega)^{-1} = 10^{-6}\ (\Omega)^{-1}$
4. Therefore, transimpedance with feedback,

$$R_{Mf} = R_M/(1+\beta R_M)$$
$$= 10^8/(1+10^{-6}\times 10^8) = 10^8/101 = 0.99\times 10^6\ \Omega = 990\ \text{k}\Omega$$

5. Input impedance, $R_{if} = R_i/(1+\beta R_M) = 10^7/(1 + 10^{-6} \times 10^8) = 10^7/101$ $= 99$ kΩ
6. As the output is seen across an open-circuit, $R_m = R_M$
7. Output impedance, $R_{of} = R_o/(1+\beta R_m) = 100/(1 + 10^{-6} \times 10^8) = 100/101$ $= 0.99$ Ω

**EXAMPLE 11.11** *Refer to the voltage-shunt feedback circuit of Figure 11.24. Derive expressions for feedback factor, transresistance, input resistance and output resistance with feedback.*

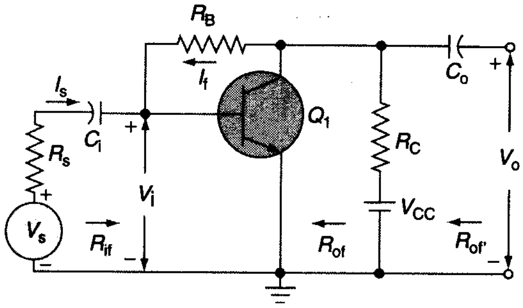



**Figure 11.24** | Example 11.11.

**Solution**

1. Feedback factor $\beta$ is given by $I_f / V_o$.
2. Now, $I_f = (V_o - V_i)/R_B$.
3. Since $V_o \gg V_i$, therefore $I_f = V_o / R_B$. That is, $\beta = I_f / V_o = 1/R_B$.
4. Transresistance with feedback, $R_{Mf} = V_o / I_s \cong 1/\beta = R_B$.
5. Input resistance with feedback is given by parallel combination of $h_{ie}$ and Miller component of $R_B$ appearing across input. That is, $R_{if} = h_{ie} \parallel R_B/(1 - A_V)$ where $A_V$ is the voltage gain.
6. $A_V = -h_{fe}R_C' / h_{ie}$, where $R_C' = R_C \parallel$ (Miller component of $R_B$ appearing across output). That is, $R_C' = R_C \parallel R_B$. Remember that the Miller component of $R_B$ appearing across output will be equal to $R_B \times [A_V/(A_V - 1)] \cong R_B$.
7. $R_{if}$ therefore can be determined by substituting the value of $A_V$ in the expression for $R_{if}$ and simplifying the resultant expression.
8. Output resistance with feedback $R_{of}$ is given by parallel combination of Miller component of $R_B$ appearing across output and $R_C$. This is approximately equal to $R_C \parallel R_B$, which is further equal to $(R_C \times R_B)/(R_C + R_B)$.

## 11.7 Current-Series (Series–Series) Feedback

In the case of current-series (series–series) feedback, output current (usually a voltage proportional to the output current) is sampled and mixed in series with the externally applied input signal. Figure 11.25 shows the block schematic arrangement of a generalized feedback amplifier with current-series feedback. Figure 11.26 shows the equivalent circuit for the schematic arrangement of Figure 11.25.

### *Gain*

The gain parameter in this case is the transconductance. Equation (11.27) gives the expression for transconductance with feedback $G_{Mf}$ in terms of transconductance without feedback $G_M$ and the feedback factor $\beta$.

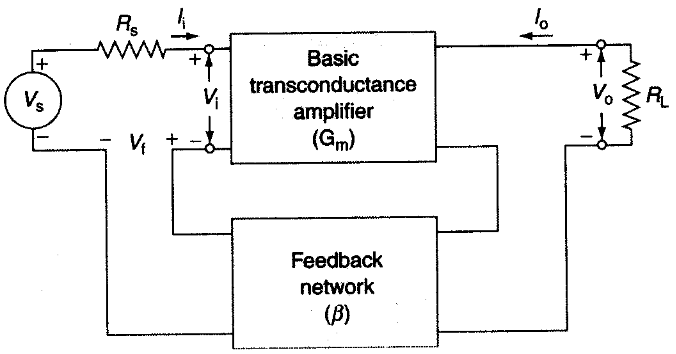



Figure 11.25 | Schematic arrangement of current-series (series–series) feedback.

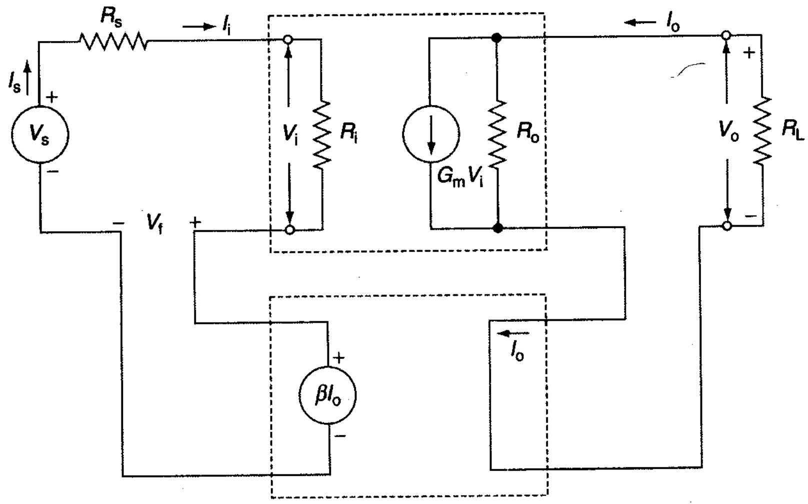


Figure 11.26 | Equivalent circuit for schematic arrangement of Figure 11.25.

$$G_{Mf} = \frac{G_M}{1+\beta G_M} \tag{11.27}$$

*Input Resistance*

Expression for input resistance is derived as follows:

$$R_{if} = V_s/I_s = V_s/I_i = (V_i + V_f)/I_i \tag{11.28}$$

Now, $I_i = V_i/R_i$. Substituting the value of $I_i$ in Eq. (11.28), we get

$$R_{if} = (V_i + V_f)\times(R_i/V_i) = R_i \times [1+(V_f/V_i)]$$

Also, $V_f = \beta I_o$. Therefore,

$$R_{if} = R_i \times [1+(\beta I_o/V_i)] = R_i \times (1+\beta G_M) \tag{11.29}$$

$G_M$ is the transconductance taking load resistance $R_L$ into account. It is expressed as

$$G_M = G_m \times \left(\frac{R_o}{R_o + R_L}\right)$$

where $G_m$ is the short-circuit transconductance, that is, without taking the load resistance $R_L$ into account. It is also expressed as

$$G_m = \lim_{R_L \to 0} G_M$$

### *Output Resistance*

Expression for output resistance is derived by considering $V_s = 0$, letting load resistance $R_L = \infty$ and applying a voltage $V$ across the output. Ratio of applied voltage $V$ to the resulting current $I$ gives the output resistance:

$$I = (V/R_o) - G_m V_i$$

For $V_s = 0$, $V_i = -V_f = -\beta I_o = +\beta I$. Therefore,

$$I = (V/R_o) - \beta I G_m$$

$$R_{of} = V/I = R_o \times (1 + \beta G_m) \qquad (11.30)$$

Remember that $R_{of}$ is the output resistance with feedback with $R_L = \infty$. Considering the effect of load resistance $R_L$, the output resistance with feedback $R_{of'}$ is given by parallel combination of $R_{of}$ and $R_L$. $R_{of'}$ can also be expressed by

$$R_{of'} = R_{o'} \times [(1 + \beta G_m)/(1 + \beta G_M)]$$

where

$$R_{o'} = R_o \| R_L \qquad (11.31)$$

### *Practical Circuits*

Common-emitter amplifier with unbypassed emitter resistor as shown in Figure 11.27 is an example of current-series feedback. As is evident from the circuit diagram, a voltage proportional to the output current is fed back in series with the source of input signal. This is true if the base current were considered as negligible. The

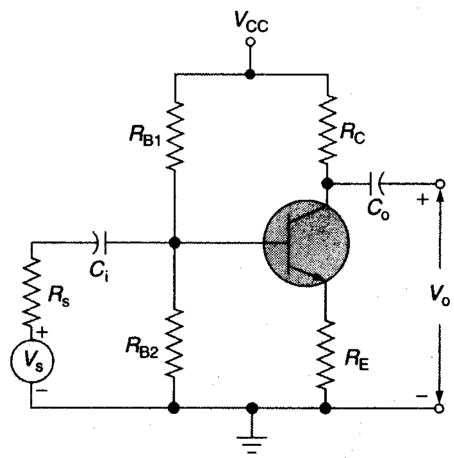



**Figure 11.27** | Common-emitter amplifier with current-series feedback.

feedback factor $\beta$ in this case equals $-R_E$. The amplifier nearly achieves the characteristics of a transconductance amplifier. Although the output current is proportional to the output voltage, this configuration should not be misunderstood as a case of voltage-series feedback. If it were considered one, then the feedback factor would equal $-R_E/R_C$. This violates one of the fundamental assumptions of negative-feedback amplifiers according to which the feedback factor should be independent of load and source resistances.

Similarly, common-source amplifier with un-bypassed source resistor is also a case of current-series feedback. Also, an opamp wired as a non-inverting amplifier where the output taken is the current $I_o$ across the feedback resistor $R_2$ is an example of current-series feedback. As shown in Figure 11.28, it is the voltage developed across $R_1$ ($= -I_o \times R_1$) that is fed back in series with the input.

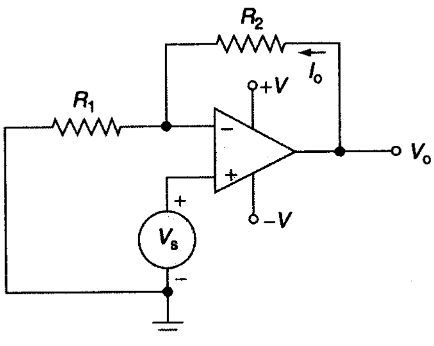



**Figure 11.28** | Opamp-based amplifier circuit with current-series feedback.

**EXAMPLE 11.12**

*Figure 11.29 shows the basic common-emitter amplifier circuit with unbypassed emitter resistor. Derive expressions for transconductance, input resistance and output resistance with and without feedback. Also prove that the transconductance parameter in this circuit is stabilized due to presence of current-series feedback.*

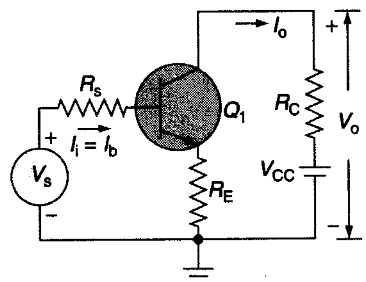



**Figure 11.29** | Example 11.12.

**Solution**

1. The feedback voltage appears across emitter resistor $R_E$. Therefore, $V_f = -I_o \times R_E$.
2. This gives feedback factor, $\beta = V_f / I_o = -R_E$.
3. In absence of any negative feedback, signal $V_i$ appearing across the input of amplifier is same as the externally applied input signal $V_s$.
4. Therefore, transconductance without feedback $G_M$ is given by

$$I_o/V_i = I_o/V_s = -(h_{fe} \times I_b)/V_s$$

5. Now, $V_s/I_b = R_s + h_{ie} + R_E$. Therefore, $G_M = -h_{fe}/(R_s + h_{ie} + R_E)$.
6. Desensitivity factor, $D = (1 + \beta G_M) = 1 + [(h_{fe} \times R_E)/(R_s + h_{ie} + R_E)]$.
   Therefore, $D = [\{R_s + h_{ie} + (1 + h_{fe}) \times R_E\}/(R_s + h_{ie} + R_E)]$.
7. This gives expression for transconductance with feedback $G_{Mf}$ as
   $$G_{Mf} = G_M/D = -h_{fe}/[R_s + h_{ie} + (1 + h_{fe}) \times R_E]$$
8. Since $(1 + h_{fe}) \times R_E >> (R_s + h_{ie})$ and also $h_{fe} >> 1$, expression for $G_{Mf}$ simplifies to $G_{Mf} \cong -1/R_E$. This proves that $G_{Mf}$ is independent of transistor parameters and is only dependent on $R_E$.
9. The input resistance in the absence of feedback is given by $R_i = R_s + h_{ie} + R_E$.
10. The input resistance with feedback is $R_{if} = R_i \times D = R_s + h_{ie} + (1 + h_{fe}) \times R_E$.
11. Output resistance in the absence of feedback ($R_o$) without considering the collector load resistance $R_C$ is infinity.
12. The output resistance with feedback $R_{of}$ without considering $R_C$ is given by $R_{of} = R_o \times (1 + \beta G_m)$. $G_m$ is short-circuited value of $G_M$. That is,
    $$G_m = \lim_{R_C \to 0} G_M$$
    As the value of $R_o$ is infinity, $R_{of}$ is also infinity.
13. Since $G_M$ is independent of $R_C$, therefore, $G_m = G_M$.
14. Output resistance with feedback $R_{of'}$ considering the effect of $R_C$ is given by the parallel combination of $R_{of}$ and $R_C$. As the value of $R_{of}$ is infinity, $R_{of'}$ equals $R_C$.
15. An alternative expression for $R_{of'}$ is given by $R_{of'} = R_{o'} \times [(1 + \beta G_m)/(1 + \beta G_M)]$.

**EXAMPLE 11.13**

*Refer to the common-source FET amplifier circuit with un-bypassed source resistor as shown in Figure 11.30. Derive expressions for transconductance, input resistance and output resistance with and without feedback*

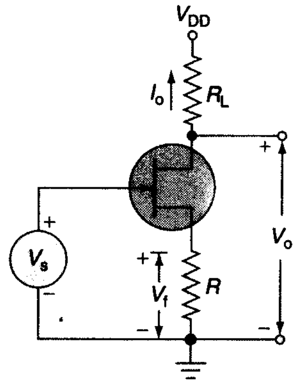



**Figure 11.30** | Example 11.13.

**Solution**

1. The feedback voltage appears across source resistor $R$. Therefore, $V_f = -I_o \times R$.
2. This gives feedback factor, $\beta = V_f / I_o = -R$.
3. In the absence of any negative feedback, the signal $V_{gs}$ appearing across the input of the amplifier is the same as the externally applied input signal $V_s$. Therefore, transconductance without feedback $G_M$ is given by $I_o / V_{gs} = I_o / V_s$.
4. $I_o$ is given by $I_o = -(g_m \times V_{gs} \times r_d)/(r_d + R_L + R)$. This gives $G_M = -(g_m \times r_d)/(r_d + R_L + R)$.
5. Now, $g_m \times r_d = \mu$. Therefore, $G_M = -\mu/(r_d + R_L + R)$.
6. Desensitivity factor,

$$D = 1 + \beta G_M = 1 + [(\mu \times R)/(r_d + R_L + R)]$$
$$= [\{r_d + R_L + (1+\mu) \times R\}/(r_d + R_L + R)]$$

7. This gives expression for transconductance with feedback $G_{Mf}$ as

$$G_{Mf} = G_M / D = -\mu / [r_d + R_L + (1+\mu) \times R]$$

8. The input resistance $R_i$ in the absence of feedback is infinity.
9. The input resistance with feedback $R_{if} = R_i \times D = \infty$.
10. If $R_L$ is considered to be an external load, the output resistance $R_o$ in the absence of feedback without considering the load resistance $R_L$ is given by $r_d + R$. Therefore, output resistance with feedback $R_{of}$ without considering $R_L$ is given by $R_{of} = R_o \times (1 + \beta G_m)$. $G_m$ is the short-circuited value of $G_M$. That is,

$$G_m = \lim_{R_L \to 0} G_M$$

11. Now, $G_m = -\mu/(r_d + R)$. This gives

$$R_{of} = R_o \times (1 + \beta G_m)$$
$$= (r_d + R) \times [\{r_d + (1+\mu) \times R\} / (r_d + R)]$$
$$= r_d + (1+\mu) \times R$$

12. Output resistance with feedback $R_{of'}$ considering the effect of $R_L$ is given by parallel combination of $R_{of}$ and $R_L$.
13. An alternative expression for $R_{of'}$ is given by

$$R_{of'} = R_{o'} \times [(1 + \beta G_m)/(1 + \beta G_M)]$$

## 11.8 Current-Shunt (Shunt–Series) Feedback

In the case of current-shunt (shunt–series) feedback, output current is sampled and mixed in shunt with the externally applied input signal. Figure 11.31 shows the block schematic arrangement of a generalized feedback amplifier with current-shunt feedback. Figure 11.32 shows the equivalent circuit for the schematic arrangement of Figure 11.31.

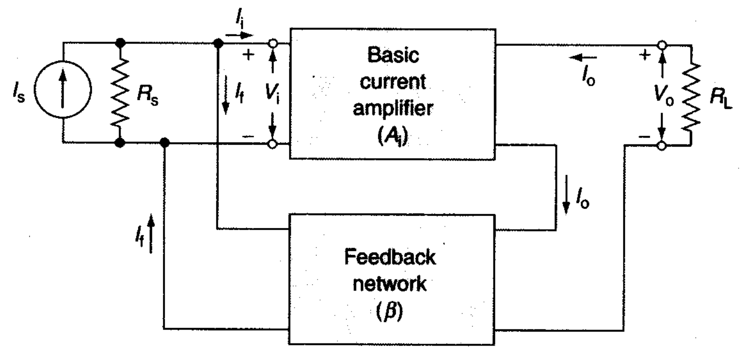



**Figure 11.31** | Schematic arrangement of current-shunt (shunt-series) feedback.

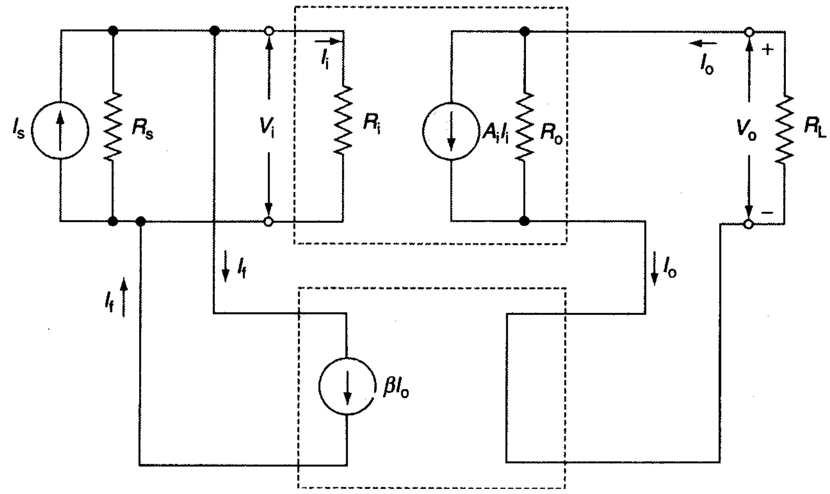



**Figure 11.32** | Equivalent circuit for schematic arrangement of Figure 11.31.

## *Gain*

The gain parameter in this case is the current gain. Equation (11.32) gives the expression for current gain with feedback $A_{If}$ in terms of current gain without feedback $A_I$ and the feedback factor $\beta$.

$$A_{If} = \frac{A_I}{1+\beta A_I} \tag{11.32}$$

## *Input Resistance*

The expression for input resistance is derived as follows:

$$R_{if} = V_i / I_s = (R_i \times I_i)/I_s \tag{11.33}$$

Substituting $I_s = I_i + I_f$ in Eq. (11.33), we get

$$R_{if} = R_i \times [I_i / (I_i + I_f)] = R_i / [(I_i + I_f)/I_i] = R_i / [1 + (I_f / I_i)]$$

Now, $I_f = \beta I_o$. Therefore,

$$R_{if} = R_i/[1+(\beta I_o/I_i)]$$

Also, $I_o/I_i = A_I$, where $A_I$ is the current gain without feedback taking load resistance into account. This gives

$$R_{if} = R_i/(1+\beta A_I) \tag{11.34}$$

$A_I$ is given by

$$A_I = A_i \times \left(\frac{R_o}{R_o + R_L}\right)$$

where $A_i$ is the current gain without taking the load resistance $R_L$ into account. In fact, it is the short-circuited current gain. It can also be expressed as

$$A_i = \lim_{R_L \to 0} A_I$$

### *Output Resistance*

Expression for output resistance is derived by considering $I_s = 0$, letting load resistance $R_L = \infty$ and applying a voltage $V$ across the output. Ratio of applied voltage $V$ to the resulting current $I$ gives the output resistance.

$$I = (V/R_o) - A_i I_i$$

For $I_s = 0$, $I_i = -I_f = -\beta I_o = +\beta I$. Therefore,

$$I = (V/R_o) - \beta I A_i$$

$$R_{of} = V/I = R_o \times (1+\beta A_i) \tag{11.35}$$

Remember that $R_{of}$ is the output resistance with feedback with $R_L = \infty$. Considering the effect of load resistance $R_L$, the output resistance with feedback $R_{of'}$ is given by parallel combination of $R_{of}$ and $R_L$. $R_{of'}$ can also be expressed by

$$R_{of'} = R_{o'} \times [(1+\beta A_i)/(1+\beta A_I)] \tag{11.36}$$

where

$$R_{o'} = R_o \| R_L$$

Also, for $R_L = \infty$, $R_{o'} = R_o$ and $A_I = 0$, Eq. (11.36) then reduces to

$$R_{of'} = R_o \times (1+\beta A_i) \tag{11.37}$$

### *Practical Circuits*

A cascade arrangement of two common-emitter amplifier stages with feedback from emitter of the second stage to the base of the first stage as shown in Figure 11.33 is an example of current-shunt feedback. The voltage across emitter resistor of the second stage is out of phase with the base voltage of the first stage, which confirms this to be a case of negative feedback. Also, it can be proved that the feedback current, that is the current flowing through resistor $R$, is proportional to the output current and the proportionality factor $\beta$ is equal to $[R_E/(R_E + R)]$. Current-shunt feedback leads to very high output impedance and very low input impedance. These are the desirable traits of a good current amplifier.

Figure 11.34 shows opamp-based inverting current amplifier. This circuit too has current-shunt feedback. One can see the similarities between this circuit and the transistorized circuit shown in Figure 11.33. The feedback factor in this case is given by $[R_1/(R_1 + R_2)]$.

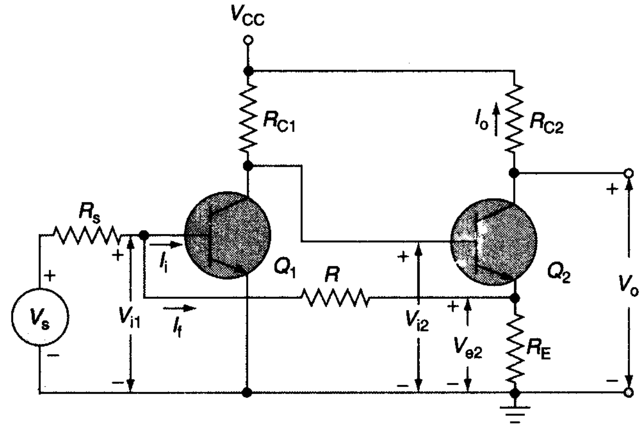



**Figure 11.33** Cascade arrangement of common-emitter amplifier stages with current-shunt feedback.

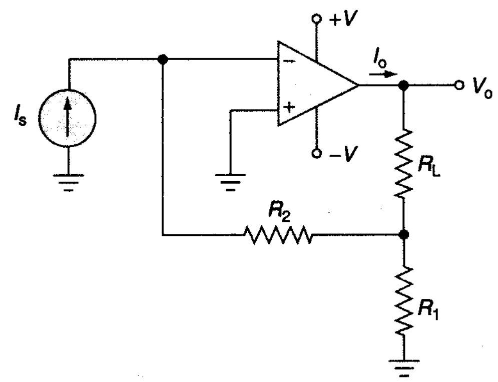



**Figure 11.34** Opamp-based current amplifier with current-shunt feedback.

## EXAMPLE 11.14

*Refer to the cascade arrangement of two CE amplifier stages as shown in Figure 11.33. Prove that the circuit has current-shunt feedback. Derive an expression for the feedback factor β. What is the current gain with feedback?*

### Solution

1. Refer to the circuit of Figure 11.33. As the output current is sampled and fed back in shunt with the input current $I_i$, it is a current-shunt feedback.
2. The feedback current $I_f$ is given by $I_f = (V_{i1} - V_{e2})/R$ .
3. Assuming $V_{e2} \gg V_{i1}$, we get $I_f = -V_{e2}/R$ .
4. $V_{e2} = (I_f - I_o) \times R_E$ considering base current of $Q_2$ to be negligible as compared to its collector current.

5. Therefore, $I_f = (I_o - I_f) \times (R_E/R)$. This gives $I_f = (R_E/R) \times [R/(R+R_E)] \times I_o$ $= [R_E/(R+R_E)] \times I_o = \beta \times I_o$, where $\beta = R_E/(R+R_E)$.
6. Current gain with feedback $A_{If}$ is approximately given by $1/\beta$. That is $A_{If} \cong (R+R_E)/R_E$.

## KEY TERMS

Current amplifier
Current-series feedback
Current-shunt feedback
Desensitivity parameter
Negative-feedback amplifier
Transconductance amplifier
Transresistance amplifier
Shunt–series feedback
Series–series feedback
Voltage amplifier
Voltage-series feedback
Voltage-shunt feedback

## OBJECTIVE-TYPE EXERCISES

### *Multiple-Choice Questions*

1. Introduction of negative feedback desensitizes or stabilizes the gain. The gain stability
   - **a.** increases with increase in open-loop gain.
   - **b.** decreases with increase in open-loop gain.
   - **c.** decreases with increase in feedback factor.
   - **d.** increases with increase in loop gain.
2. A voltage amplifier has an open-loop gain of 100. If 10% negative feedback were introduced in the amplifier, then an 11% change in open-loop gain would cause
   - **a.** 1% change in closed-loop gain.
   - **b.** 11% change in closed-loop gain.
   - **c.** 1.1% change in closed-loop gain.
   - **d.** 0.1% change in closed-loop gain.
3. In the case of a negative-feedback amplifier, which of the following is true?
   - **a.** Closed-loop gain can be less than or more than the open-loop gain depending upon the type and amount of feedback.
   - **b.** Closed-loop gain is always less than the open-loop gain.
   - **c.** Closed-loop gain is always more than the open-loop gain.
   - **d.** Closed-loop gain and bandwidth are always less than the corresponding open-loop values.
4. In which of the following feedback topologies, the input impedance increases with introduction of feedback.
   - **a.** Voltage-shunt feedback
   - **b.** Current-shunt feedback
   - **c.** Voltage-series feedback
   - **d.** None of these
5. In which of the following feedback topologies, the input impedance decreases with introduction of feedback.
   - **a.** Voltage-shunt feedback
   - **b.** Current-series feedback
   - **c.** Voltage-series feedback
   - **d.** None of these
6. In which of the following feedback topologies, the output impedance increases with introduction of feedback.
   - **a.** Voltage-shunt feedback
   - **b.** Current-shunt feedback
   - **c.** Voltage-series feedback
   - **d.** None of these

7. In which of the following feedback topologies, both input as well as output impedances decrease with introduction of feedback.
   a. Voltage-shunt feedback
   b. Current-shunt feedback
   c. Voltage-series feedback
   d. None of these

8. Which of the following amplifier configurations has an inherent current-series feedback?
   a. Emitter–follower
   b. Common-base amplifier
   c. Common-emitter amplifier by bypassed emitter resistor
   d. Common-emitter amplifier with unbypassed emitter resistor

9. It is desired to design a voltage controlled current source. What type of negative feedback should preferably be introduced to make it a stable source?
   a. Voltage-series feedback
   b. Current-series feedback
   c. Current-shunt feedback
   d. Voltage-shunt feedback

10. A voltage amplifier consists of cascade arrangement of two identical stages. Negative feedback is introduced from output of second stage to input of first stage to get a stable overall closed-loop gain of 45. In order that 10% variation in open-loop gain of each stage does not produce more than 1% variation in overall closed-loop gain, the minimum value of open-loop gain of each stage should be
    a. 30.
    b. 900.
    c. 45.
    d. none of these.

11. A voltage amplifier has 5% negative feedback. Its voltage gain is
    a. indeterminate from given data.
    b. approximately 20.
    c. approximately 5.
    d. 100.

12. For the expression of gain with feedback $A_f = A/(1 + \beta A)$ in the case of negative-feedback amplifier to be valid,
    a. feedback factor $\beta$ should be independent of load and source resistances.
    b. forward transmission through the feedback network should be zero.
    c. reverse transmission through amplifier should be zero.
    d. all the above should be true.

13. Voltage-shunt feedback stabilizes
    a. voltage gain.
    b. current gain.
    c. transresistance.
    d. transconductance.

14. An amplifier with open-loop input resistance of 100 kΩ has −20 dB of voltage-series feedback. Closed-loop input resistance would be
    a. 10 kΩ.
    b. 1000 kΩ.
    c. 2000 kΩ.
    d. 5 kΩ.

15. In a negative-feedback amplifier with a high open-loop gain, doubling the feedback factor
    a. doubles the closed-loop gain too.
    b. has no effect on closed-loop gain.
    c. reduces the closed-loop gain to one-fourth.
    d. reduces the closed-loop gain to one-half.

## Identify the Feedback Topology

Figure 11.35 shows some circuit configurations. Identify the feedback topology used in each of these circuit configurations.

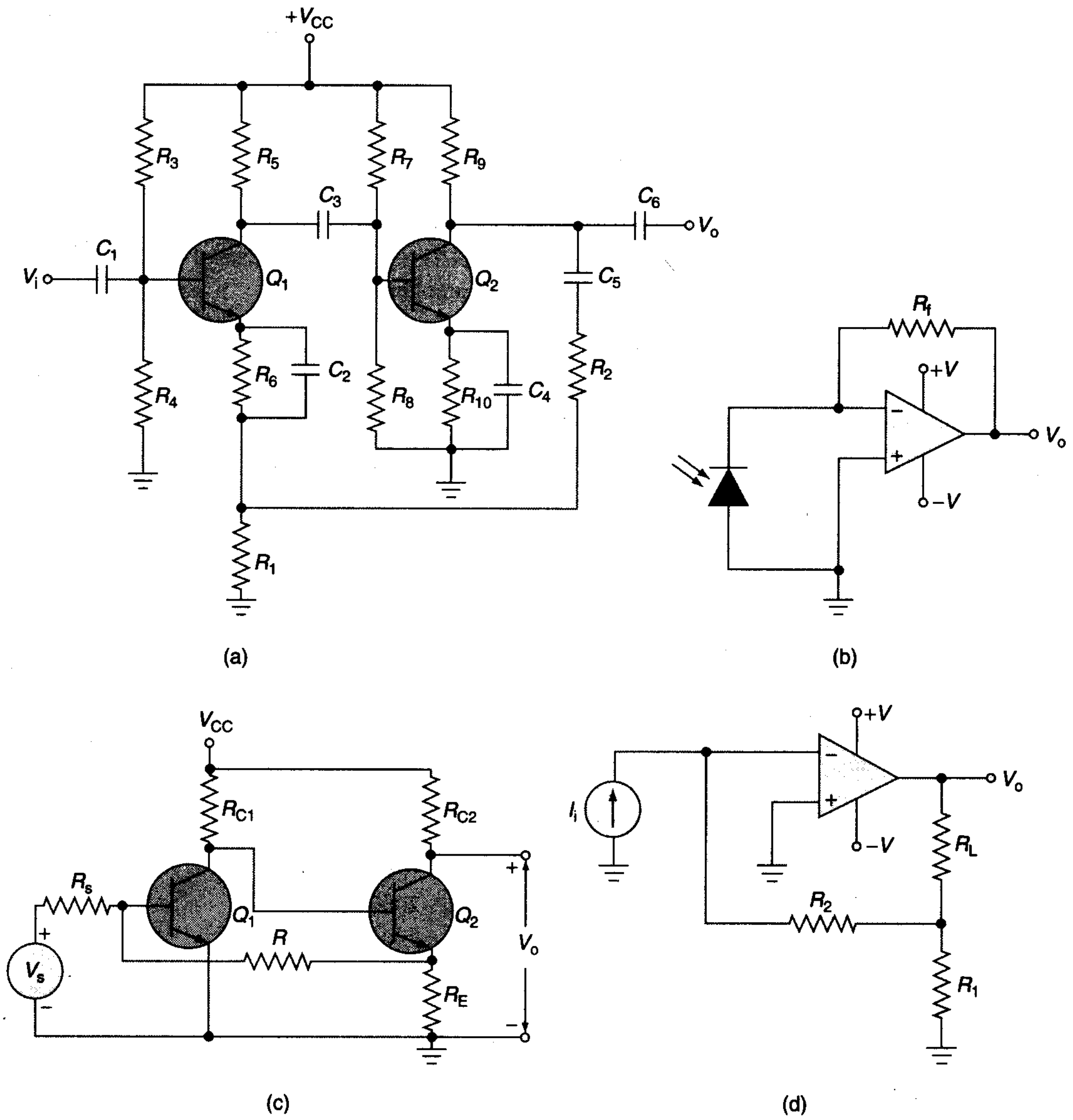



**Figure 11.35** | Identify the feedback topology.

## REVIEW QUESTIONS

1. How do you classify amplifiers based on the nature of input and output signals of interest? What is the gain parameter of interest in each of these types?

2. What are the merits/demerits of introduction of negative feedback in amplifiers?

3. What type of negative feedback should be introduced in an amplifier to make it work like

a true (a) current-to-voltage converter and (b) voltage-to-current converter? Give reasons for your answer.

4. How will the input impedance of an amplifier be affected by introduction of (a) voltage-series feedback and (b) current-shunt feedback?

5. How will the output impedance of an amplifier be affected by introduction of (a) current-series feedback and (b) voltage-shunt feedback?

6. Derive the relevant expressions to prove that output impedance reduces in the case of voltage-series and voltage-shunt feedback and increases in the case of current-series and current-shunt feedback.

7. Derive the relevant expressions to prove that input impedance reduces in the case of voltage-shunt and current-shunt feedback and increases in the case of voltage-series and current-series feedback.

8. Give two examples of practical amplifier circuits representative of the following types of negative feedback.
   a. Series–series feedback
   b. Series–shunt feedback

9. Outline the three fundamental assumptions that must be true in order that introduction of negative feedback has the desired effect on gain parameter stability, input and output impedances.

10. Justify the following.
    a. A common-emitter amplifier with unbypassed emitter resistor is representative of current-series feedback and not voltage-series feedback.
    b. The gain parameter in a negative-feedback amplifier depends exclusively on the components of feedback network when the open-loop gain becomes very large.

## PROBLEMS

1. Determine percentage reduction in gain of an amplifier due to introduction of 20 dB of negative feedback.

2. The gain of an amplifier reduces to 10 by introduction of –40 dB of feedback. Determine the open-loop gain of the amplifier.

3. A voltage amplifier is characterized by an open-loop voltage gain of 100, input resistance of 50 kΩ and output resistance of 2 kΩ. Negative feedback of 10% of output voltage is introduced in series with the input to bring the distortion below acceptable level. Find the modified values of these parameters.

4. For the opamp-based non-inverting amplifier circuit of Figure 11.36, determine the voltage gain and the input impedance in the presence of feedback. Given that open-loop gain and the input impedance of the opamp are 70 dB and 10 MΩ, respectively.

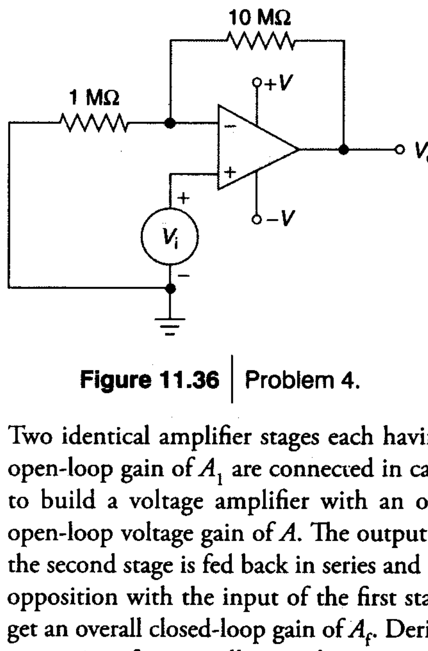



**Figure 11.36** | Problem 4.

5. Two identical amplifier stages each having an open-loop gain of $A_1$ are connected in cascade to build a voltage amplifier with an overall open-loop voltage gain of $A$. The output from the second stage is fed back in series and phase opposition with the input of the first stage to get an overall closed-loop gain of $A_f$. Derive an expression for overall open-loop gain $A$ in

terms of open-loop gain stability $|dA_1/A_1|$ of individual stages and overall closed-loop gain stability $|dA_f/A_f|$.

6. A voltage amplifier is specified to have a voltage gain of 1000 ± 50. It is desired that the voltage gain does not vary by more than ±0.1%. Find the value of feedback factor $\beta$ and the value of closed-loop gain.

7. Figure 11.37 shows an FET-based common-source amplifier circuit with voltage-series feedback provided by series combination of $R_1$ and $R_2$. FET is characterized by $g_m = 4000\ \mu S$ and $r_d = 10\ k\Omega$. Calculate the voltage gain with feedback.

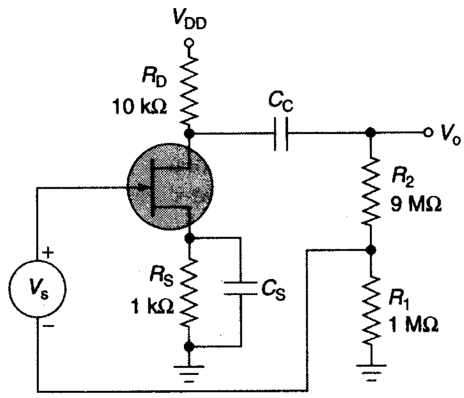



**Figure 11.37** | Problem 7.

## ANSWERS

### Multiple-Choice Questions

1. (d)
2. (a)
3. (b)
4. (c)
5. (a)
6. (b)
7. (a)
8. (d)
9. (b)
10. (a)
11. (b)
12. (d)
13. (c)
14. (b)
15. (d)

### Identify the Feedback Topology

Figure 11.35(a): Voltage-series feedback
Figure 11.35(b): Voltage-shunt feedback
Figure 11.35(c): Current-shunt feedback
Figure 11.35(d): Current-shunt feedback

### Problems

1. 90%
2. 1000
3. Gain = 9.09; input resistance = 550 kΩ; output resistance = 181.82 Ω
4. Voltage gain = 10.96; input impedance = 2884.80 MΩ
5. $A = 2A_f \times |dA_1/A_1| / |dA_f/A_f|$
6. $\beta = 0.049$; closed-loop gain = 20
7. −6.56