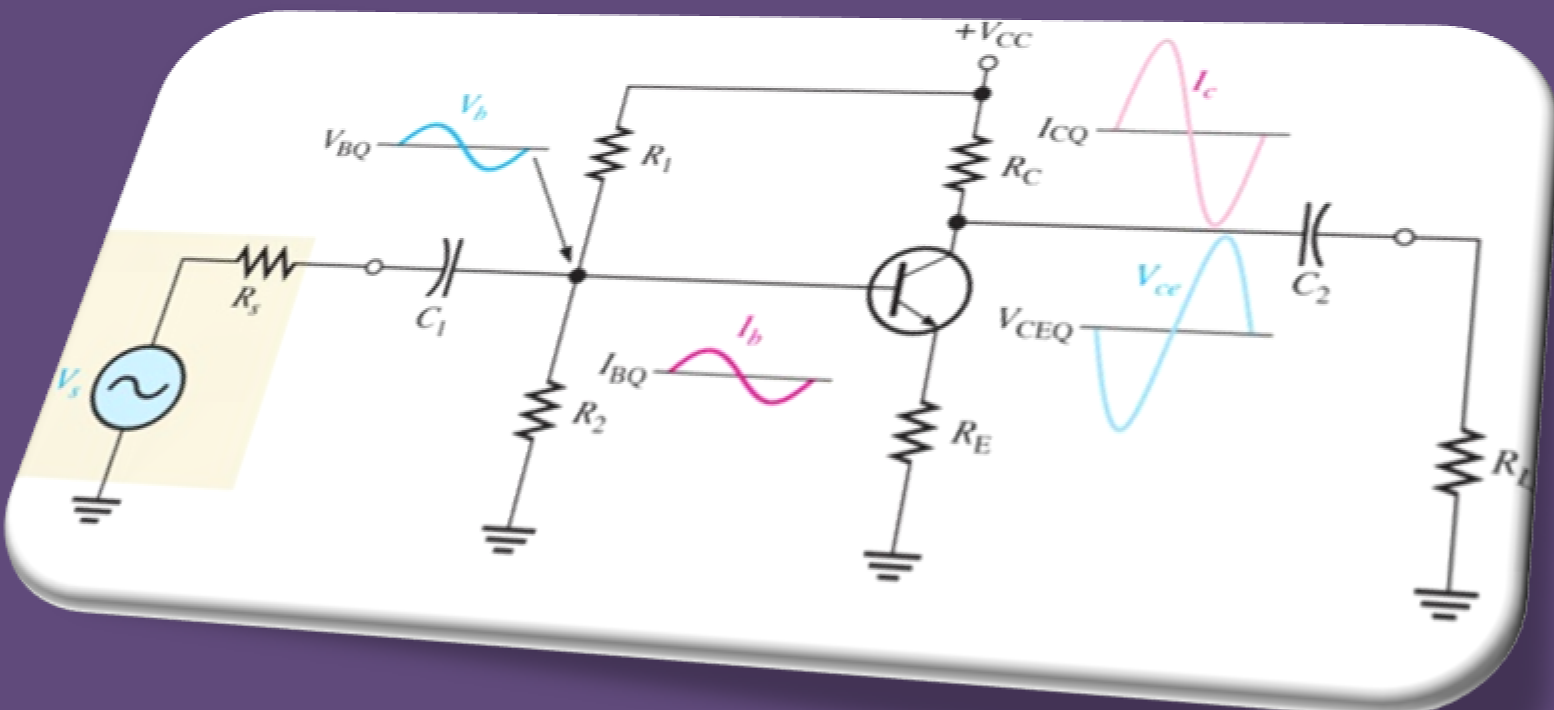
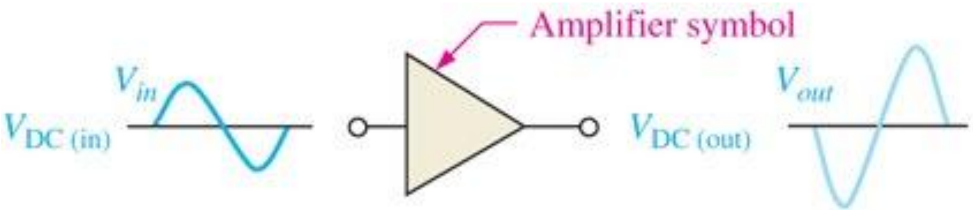
# Electronic Circuits /2/

Dr. Nidal ZAIDAN

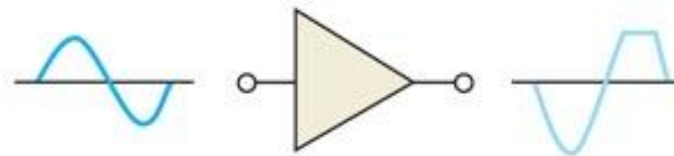
## CHAPTER /1/

## DESIGN OF BJT AMPLIFIERS

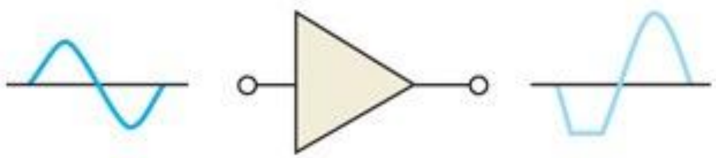





(a) Linear operation: larger output has same shape as input except that it is inverted

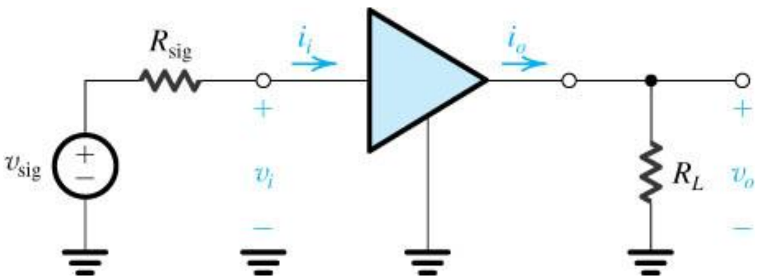


(b) Nonlinear operation: output voltage limited (clipped) by cutoff



(c) Nonlinear operation: output voltage limited (clipped) by saturation

# Characterizing BJT Amplifiers




## Definitions

- Input resistance with no load:

$$R_i \equiv \left.\frac{v_i}{i_i}\right|_{R_L=\infty}$$

- Input resistance:

$$R_{in} \equiv \frac{v_i}{i_i}$$

- Open-circuit voltage gain:

$$A_{vo} \equiv \left.\frac{v_o}{v_i}\right|_{R_L=\infty}$$

- Voltage gain:

$$A_v \equiv \frac{v_o}{v_i}$$

- Short-circuit current gain:

$$A_{is} \equiv \left.\frac{i_o}{i_i}\right|_{R_L=0}$$

- Current gain:

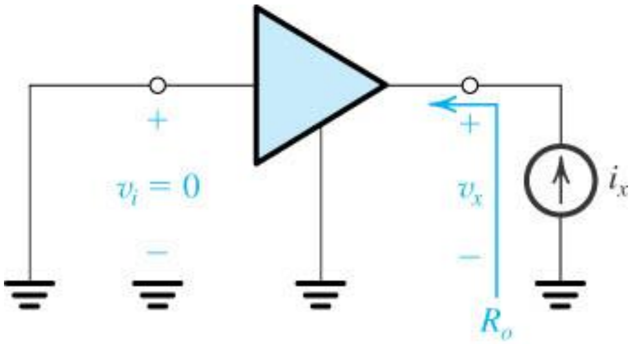
$$A_i \equiv \frac{i_o}{i_i}$$

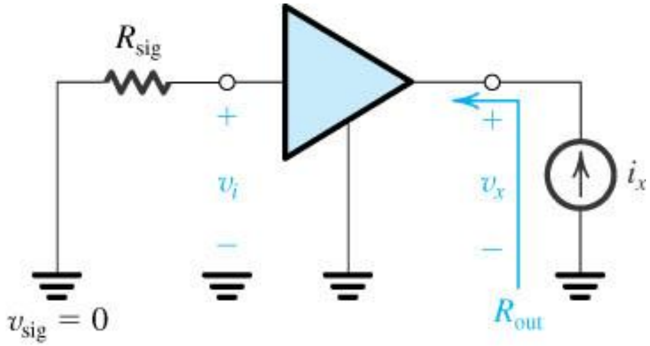
- Short-circuit transconductance:

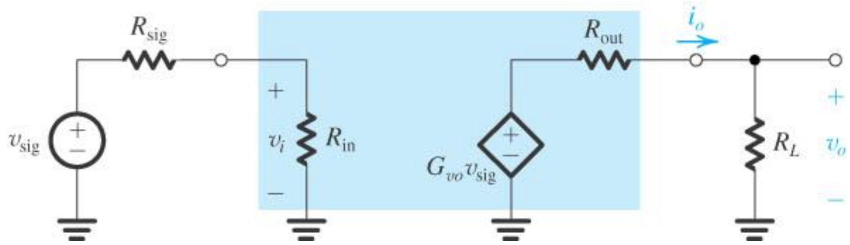
$$G_m \equiv \left.\frac{i_o}{v_i}\right|_{R_L=0}$$

- Output resistance of amplifier proper:

$$R_o \equiv \left.\frac{v_x}{i_x}\right|_{v_i=0}$$






- Output resistance:

$$R_{out} \equiv \left.\frac{v_x}{i_x}\right|_{v_{sig}=0}$$

- Open-circuit overall voltage gain:

$$G_{vo} \equiv \left.\frac{v_o}{v_{sig}}\right|_{R_L=\infty}$$

- Overall voltage gain:

$$G_v \equiv \frac{v_o}{v_{sig}}$$

- $$\frac{v_i}{v_{sig}} = \frac{R_{in}}{R_{in} + R_{sig}}$$

- $$A_v = A_{vo}\frac{R_L}{R_L + R_o}$$

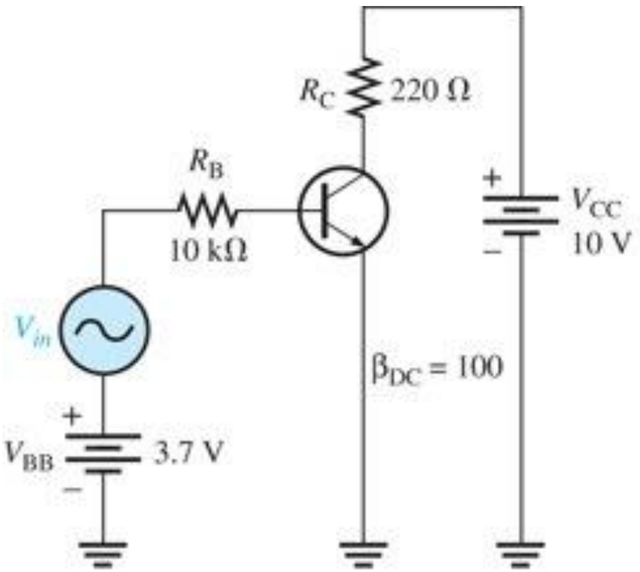
- $$A_{vo} = G_m R_o$$

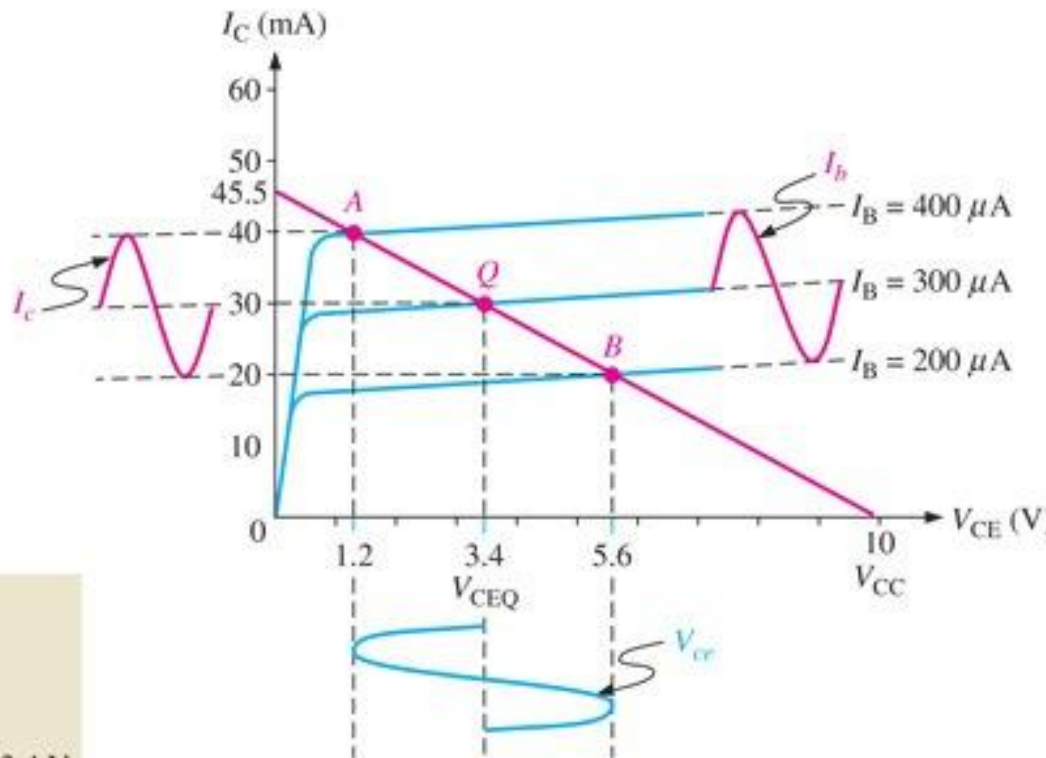
- $$G_v = \frac{R_{in}}{R_{in} + R_{sig}} A_{vo} \frac{R_L}{R_L + R_o}$$

- $$G_{vo} = \frac{R_i}{R_i + R_{sig}} A_{vo}$$

- $$G_v = G_{vo} \frac{R_L}{R_L + R_{out}}$$

## BIASING IN BJT AMPLIFIER CIRCUITS

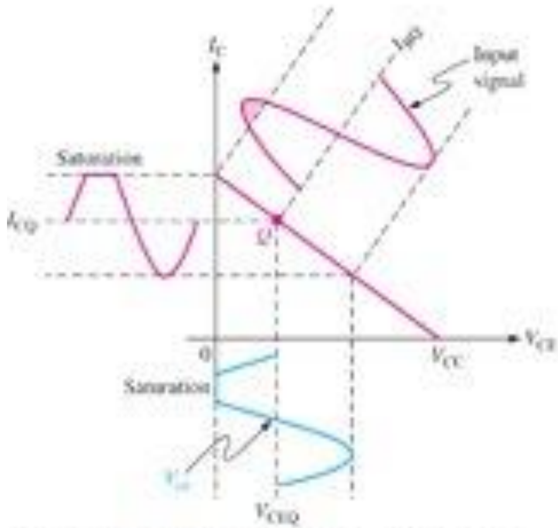


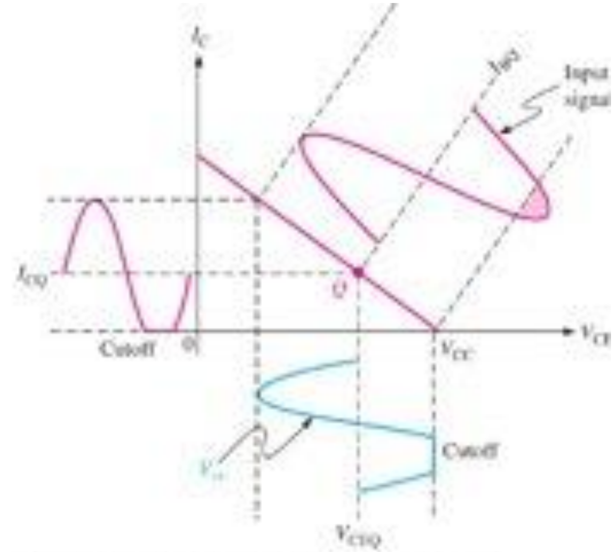

$$I_{BQ} = \frac{V_{BB} - 0.7\text{ V}}{R_B} = \frac{3.7\text{ V} - 0.7\text{ V}}{10\text{ k}\Omega} = 300\ \mu\text{A}$$

$$I_{CQ} = \beta_{DC} I_{BQ} = (100)(300\ \mu\text{A}) = 30\text{ mA}$$

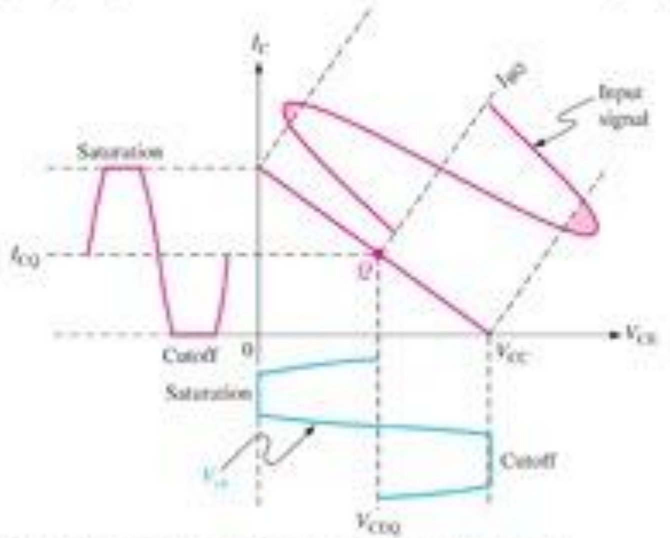
$$V_{CEQ} = V_{CC} - I_{CQ} R_C = 10\text{ V} - (30\text{ mA})(220\ \Omega) = 3.4\text{ V}$$




(a) Transistor is driven into saturation because the Q-point is too close to saturation for the given input signal.



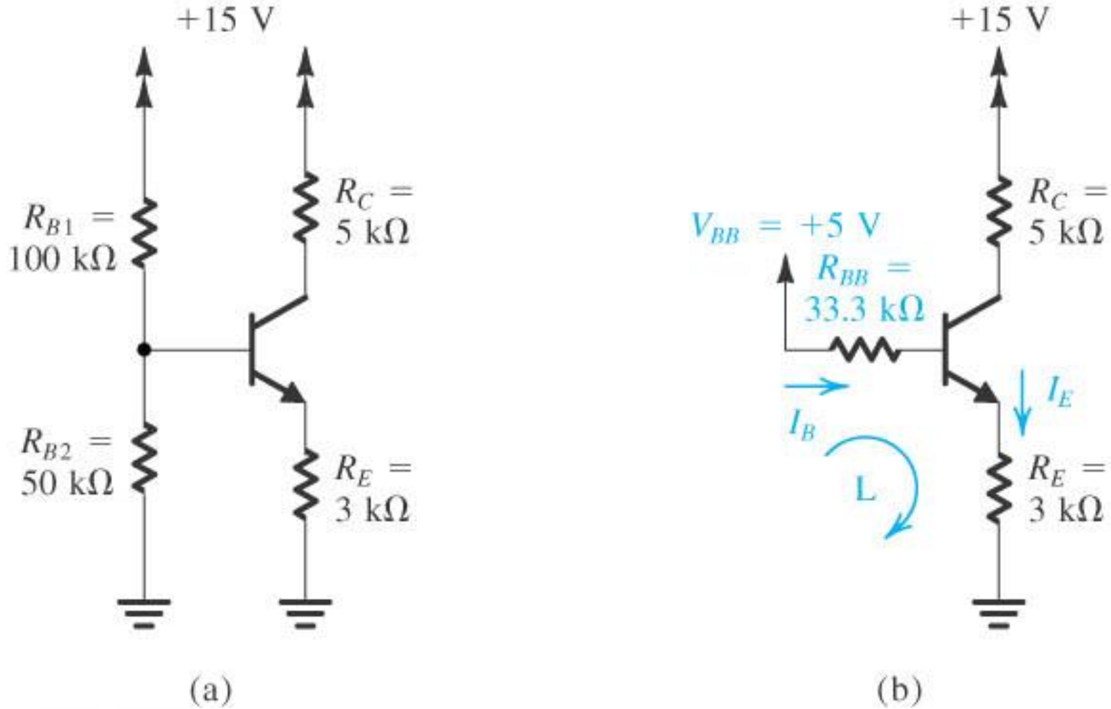

(b) Transistor is driven into cutoff because the Q-point is too close to cutoff for the given input signal.




(c) Transistor is driven into both saturation and cutoff because the input signal is too large.

## BIASING IN BJT AMPLIFIER CIRCUITS

### .1 The Classical Discrete-Circuit Bias Arrangement

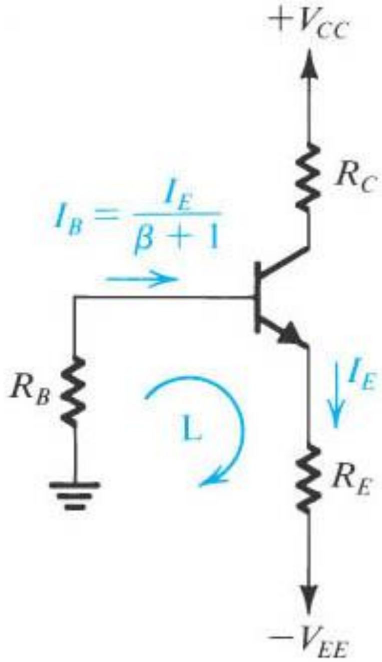



$$V_{BB} = \frac{R_2}{R_1 + R_2} V_{CC}$$

$$R_B = \frac{R_1 R_2}{R_1 + R_2}$$

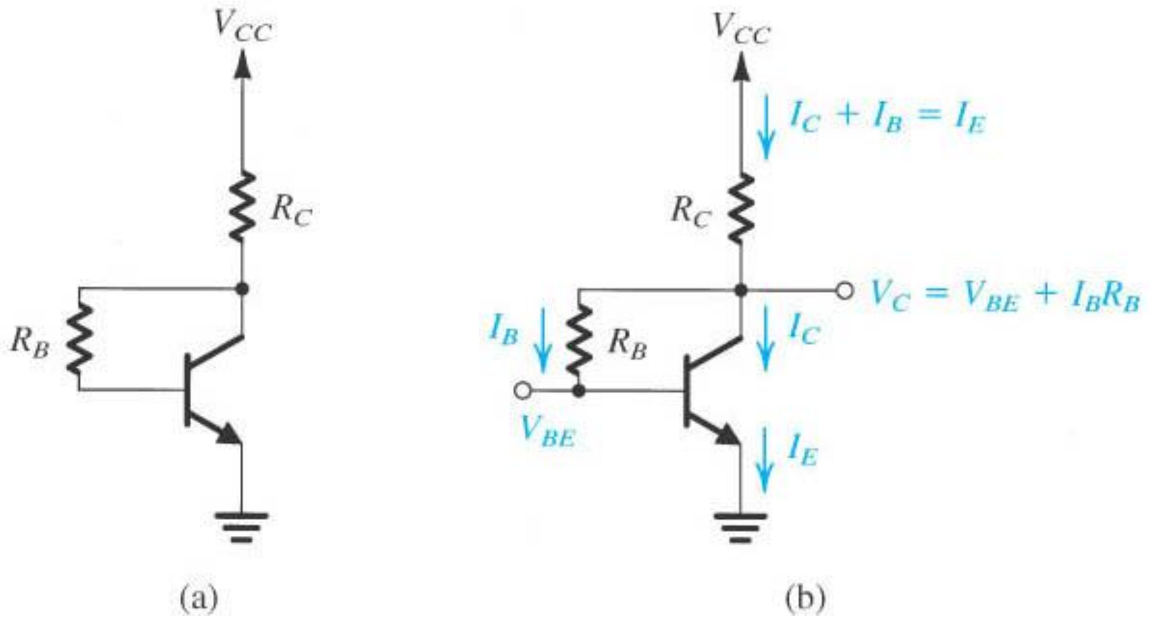
$$I_E = \frac{V_{BB} - V_{BE}}{R_E + R_B/(\beta + 1)}$$

## .2 A Two-Power-Supply Version of the Classical Bias Arrangement




$$I_E = \frac{V_{EE} - V_{BE}}{R_E + R_B/(\beta + 1)}$$

## .3 Biasing Using a Collector-to-Base Feedback Resistor



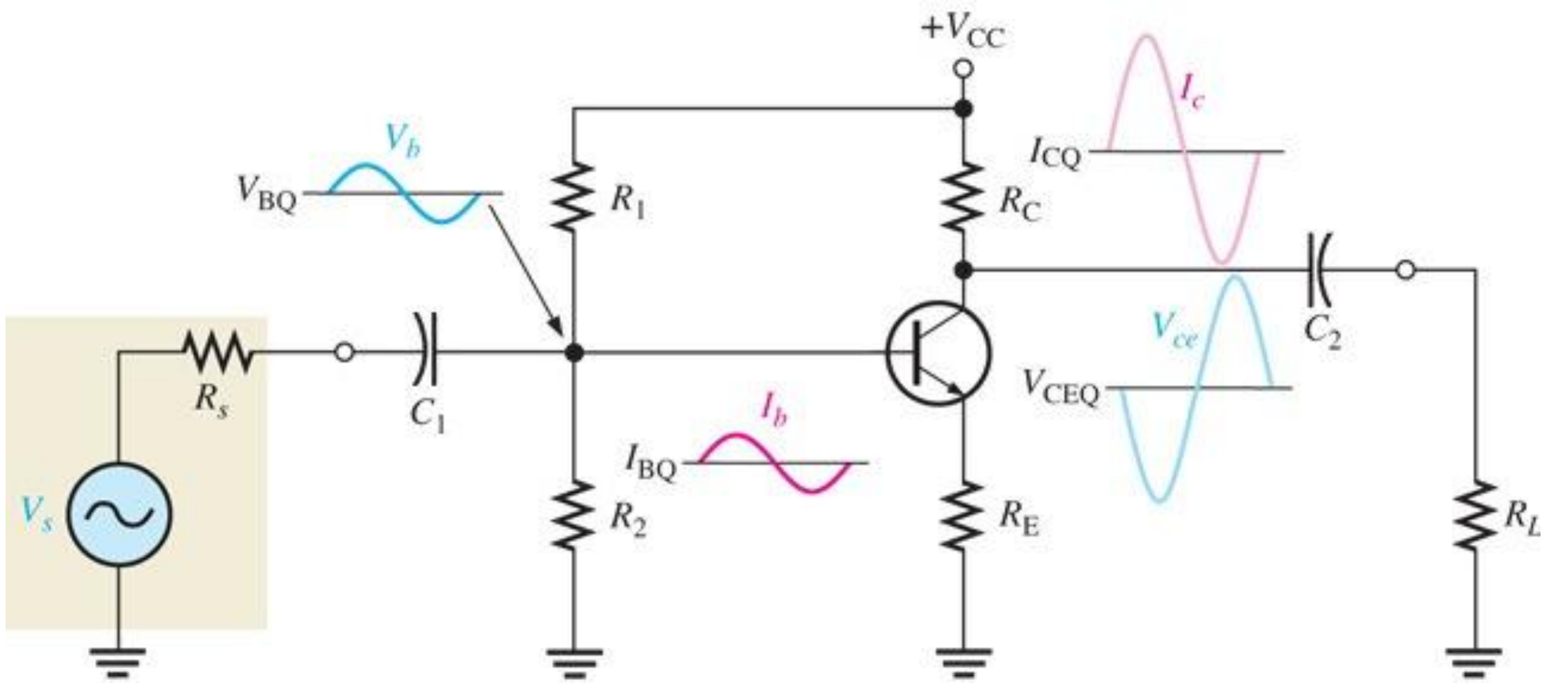

$$V_{CC} = I_E R_C + I_B R_B + V_{BE}$$

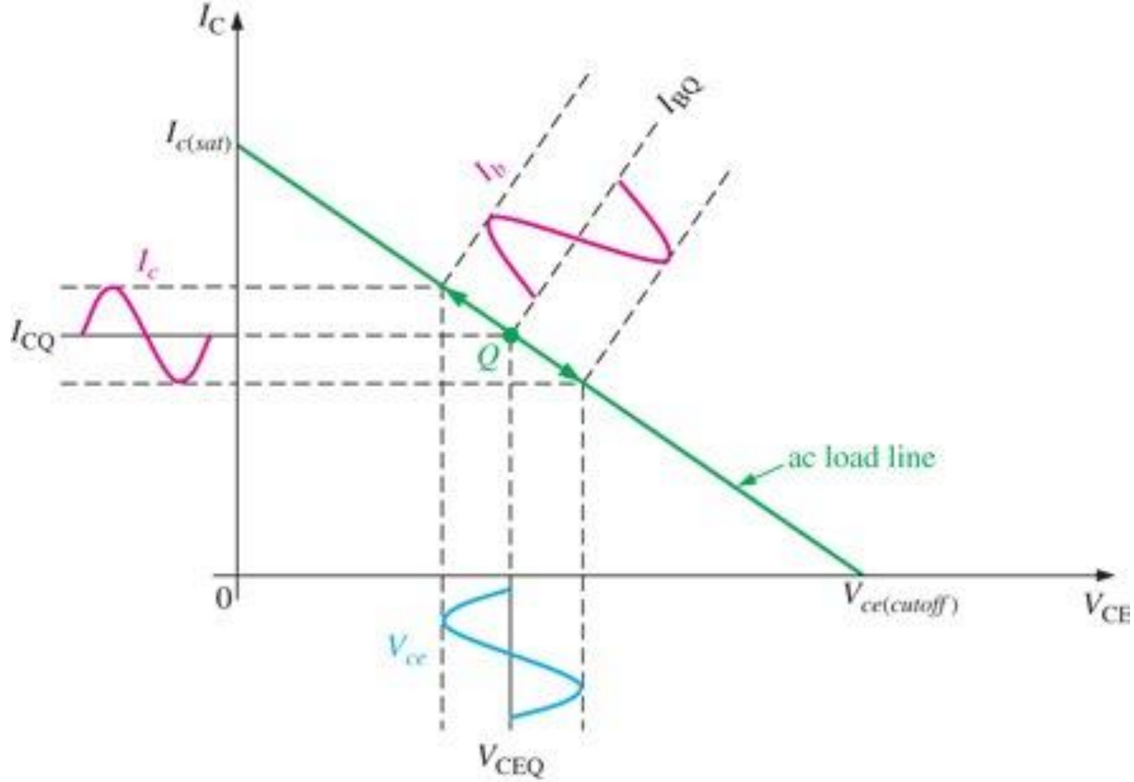
$$= I_E R_C + \frac{I_E}{\beta + 1} R_B + V_{BE}$$

Thus the emitter bias current is given by

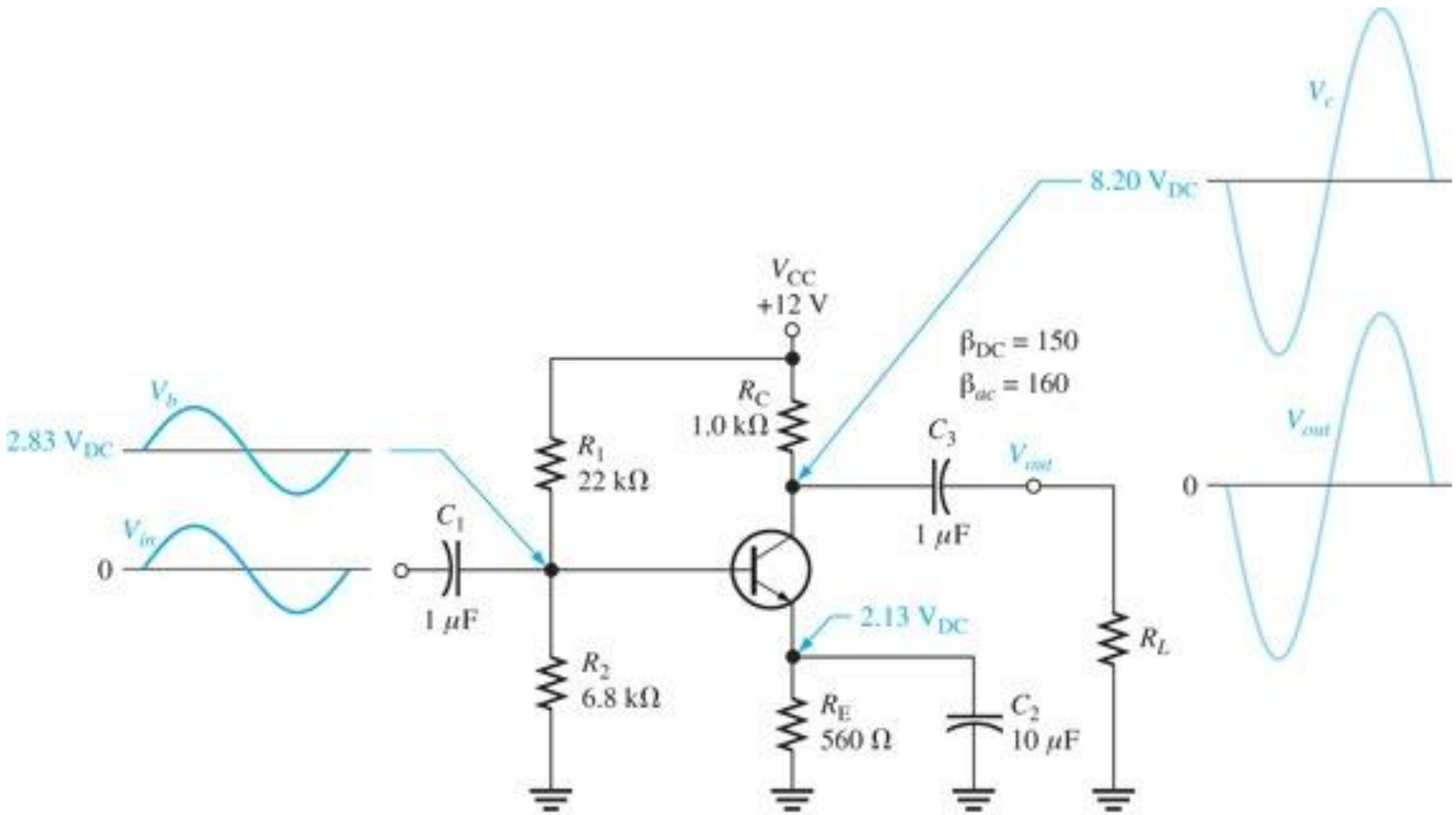
$$I_E = \frac{V_{CC} - V_{BE}}{R_C + R_B/(\beta + 1)}$$

# SMALL-SIGNAL OPERATION AND MODELS





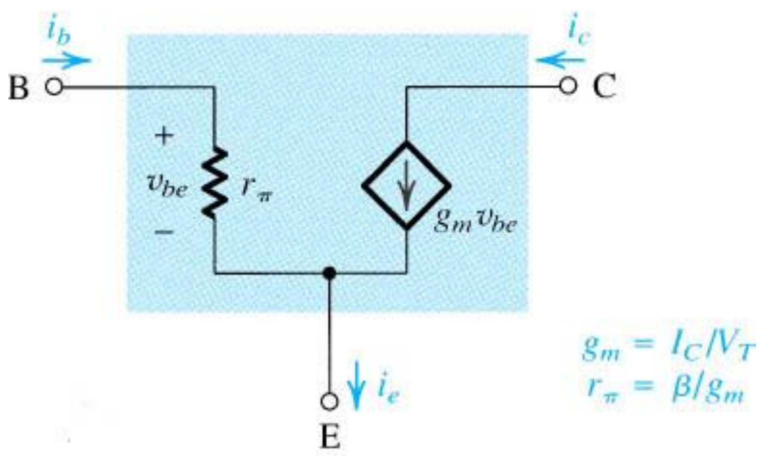
$I_C$
$I_{c(sat)}$
$I_b$
$I_{BQ}$
$I_c$
$I_{CQ}$
$Q$
ac load line
0
$V_{ce(cutoff)}$
$V_{CE}$
$V_{ce}$
$V_{CEQ}$



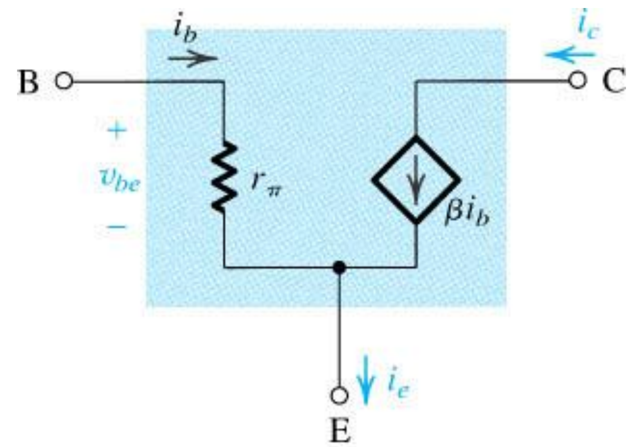
$V_c$
8.20 $V_{DC}$
$V_{CC}$
+12 V
$\beta_{DC}$ = 150
$\beta_{ac}$ = 160
$V_b$
$R_C$
1.0 kΩ
2.83 $V_{DC}$
$R_1$
22 kΩ
$C_3$
$V_{out}$
$V_{out}$
0
$V_{in}$
$C_1$
1 μF
0
1 μF
2.13 $V_{DC}$
$R_L$
$R_2$
6.8 kΩ
$R_E$
560 Ω
$C_2$
10 μF

## The Hybrid-$\pi$ Model

Two slightly different versions of the simplified *hybrid-$\pi$ model* for the small-signal operation of the BJT. The equivalent circuit in **(a)** represents the BJT as a voltage-controlled current source (a transconductance amplifier), and that in **(b)** represents the BJT as a current-controlled current source (a current amplifier).



(a)



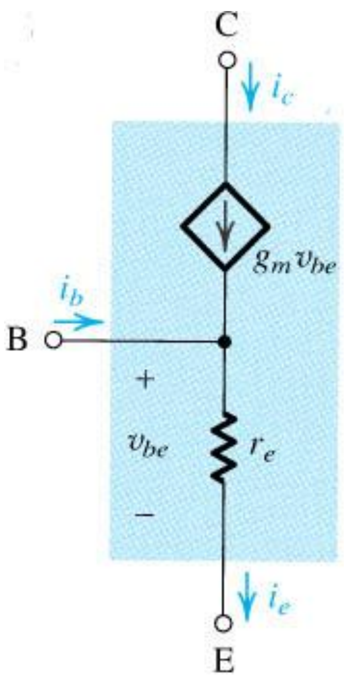
(b)

$$i_e = \frac{v_{be}}{r_\pi} + g_m v_{be} = \frac{v_{be}}{r_\pi}(1 + g_m r_\pi)$$
$$= \frac{v_{be}}{r_\pi}(1+\beta) = v_{be}\Big/\left(\frac{r_\pi}{1+\beta}\right)$$
$$= v_{be}/r_e$$

$$g_m v_{be} = g_m(i_b r_\pi)$$
$$= (g_m r_\pi) i_b = \beta i_b$$

# The T Model

Two slightly different versions of what is known as the *T model* of the BJT. The circuit in **(a)** is a voltage-controlled current source representation and that in **(b)** is a current-controlled current source representation. These models explicitly show the emitter resistance $r_e$ rather than the base resistance $r_\pi$ featured in the hybrid-π model.




(a)

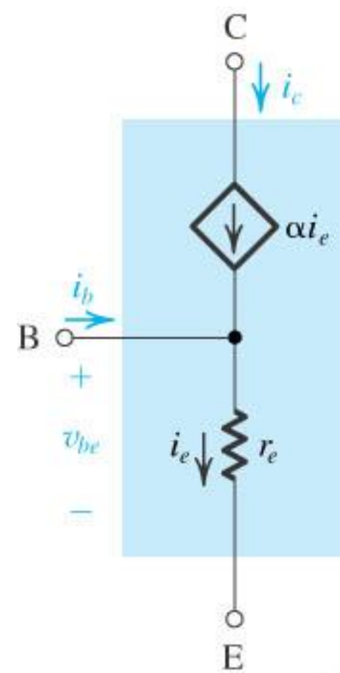
$$g_m = I_C/V_T$$

$$r_e = \frac{V_T}{I_E} = \frac{\alpha}{g_m}$$

$$i_b = \frac{v_{be}}{r_e} - g_m v_{be} = \frac{v_{be}}{r_e}(1 - g_m r_e)$$

$$= \frac{v_{be}}{r_e}(1-\alpha) = \frac{v_{be}}{r_e}\left(1 - \frac{\beta}{\beta+1}\right)$$

$$= \frac{v_{be}}{(\beta+1)r_e} = \frac{v_{be}}{r_\pi}$$




(b)

$$g_m v_{be} = g_m(i_e r_e)$$

$$= (g_m r_e) i_e = \alpha i_e$$

## Application of the Small-Signal Equivalent Circuits

The availability of the small-signal BJT circuit models makes the analysis of transistor amplifier circuits a systematic process. The process consists of the following steps:

1. Determine the dc operating point of the BJT and in particular the dc collector current $I_C$.

2. Calculate the values of the small-signal model parameters: $g_m = I_C/V_T$, $r_\pi = \beta/g_m$, and $r_e = V_T/I_E = \alpha/g_m$.

3. Eliminate the dc sources by replacing each dc voltage source with a short circuit and each dc current source with an open circuit.

4. Replace the BJT with one of its small-signal equivalent circuit models. Although any one of the models can be used, one might be more convenient than the others for the particular circuit being analyzed. This point will be made clearer later in this chapter.

5. Analyze the resulting circuit to determine the required quantities (e.g., voltage gain, input resistance). The process will be illustrated by the following examples.

# Example

We wish to analyze the transistor amplifier shown in Fig. 5.53(a) to determine its voltage gain. Assume $\beta = 100$.

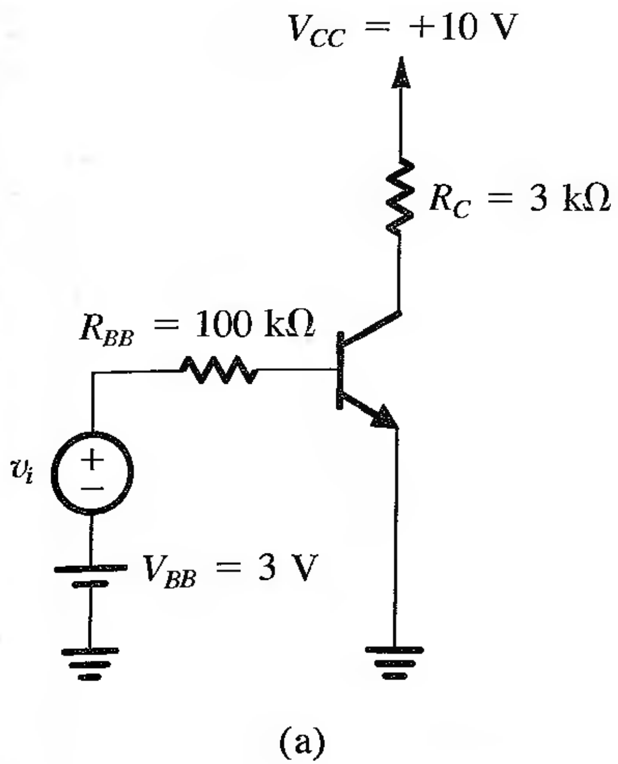


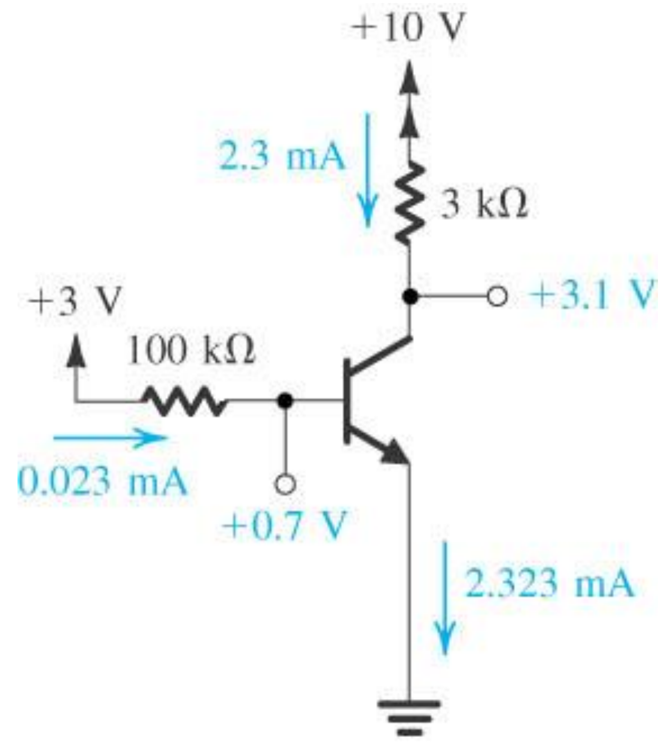

**FIGURE 5.53** Example 5.14: (a) circuit;

# Solution

The first step in the analysis consists of determining the quiescent operating point. For this purpose we assume that $v_i = 0$. The dc base current will be

$$I_B = \frac{V_{BB} - V_{BE}}{R_{BB}}$$

$$\simeq \frac{3 - 0.7}{100} = 0.023 \text{ mA}$$



(b)

The dc collector current will be

$$I_C = \beta I_B = 100 \times 0.023 = 2.3 \text{ mA}$$

The dc voltage at the collector will be

$$V_C = V_{CC} - I_C R_C$$

$$= +10 - 2.3 \times 3 = +3.1 \text{ V}$$

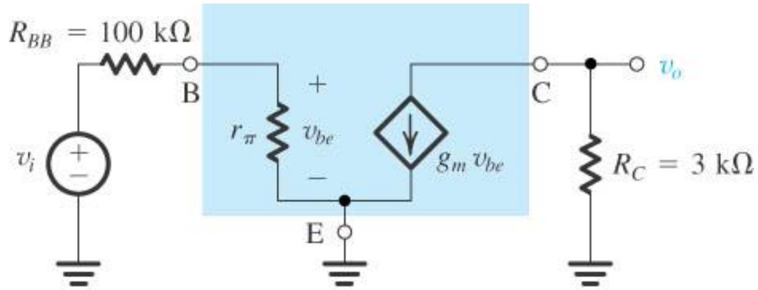
Having determined the operating point, we may now proceed to determine the small-signal model parameters:

$$r_e = \frac{V_T}{I_E} = \frac{25 \text{ mV}}{(2.3/0.99) \text{ mA}} = 10.8 \ \Omega$$

$$g_m = \frac{I_C}{V_T} = \frac{2.3 \text{ mA}}{25 \text{ mV}} = 92 \text{ mA/V}$$

$$r_\pi = \frac{\beta}{g_m} = \frac{100}{92} = 1.09 \text{ k}\Omega$$

To carry out the small-signal analysis it is equally convenient to employ either of the two hybrid-$\pi$ equivalent circuit models of Fig. 5.51. Using the first results in the amplifier equivalent circuit given in Fig. 5.53(c). Note that no dc quantities are included in this equivalent circuit. It is most important to note that the dc supply voltage $V_{CC}$ has been replaced by a *short circuit* in the signal equivalent circuit because the circuit terminal connected to $V_{CC}$ will always have a constant voltage; that is, the signal voltage at this terminal will be zero. In other words, *a circuit terminal connected to a constant dc source can always be considered as a signal ground.*




(c)

Analysis of the equivalent circuit in Fig. 5.53(c) proceeds as follows:

$$v_{be} = v_i \frac{r_\pi}{r_\pi + R_{BB}} = v_i \frac{1.09}{101.09} = 0.011 v_i \tag{5.105}$$

The output voltage $v_o$ is given by

$$v_o = -g_m v_{be} R_C$$
$$= -92 \times 0.011 v_i \times 3 = -3.04 v_i$$

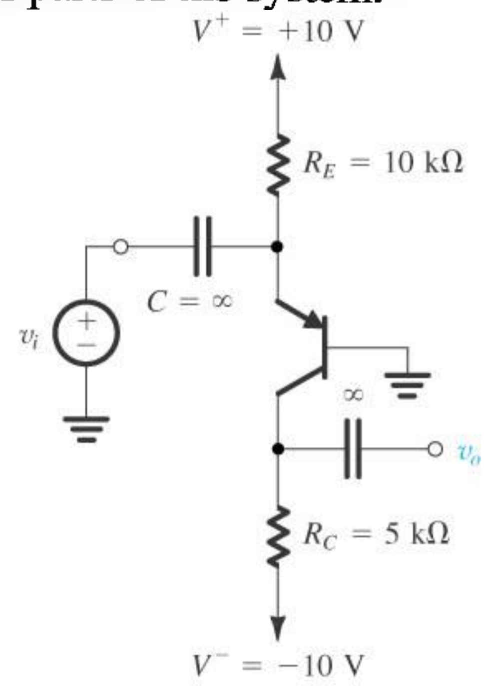
Thus the voltage gain will be

$$A_v = \frac{v_o}{v_i} = -3.04 \text{ V/V} \tag{5.106}$$

where the minus sign indicates a phase reversal.

# Example 5.16

We need to analyze the circuit of Fig. 5.55(a) to determine the voltage gain and the signal waveforms at various points. The capacitor $C$ is a coupling capacitor whose purpose is to couple the signal $v_i$ to the emitter while blocking dc. In this way the dc bias established by $V^+$ and $V^-$ together with $R_E$ and $R_C$ will not be disturbed when the signal $v_i$ is connected. For the purpose of this example, $C$ will be assumed to be very large and ideally infinite—that is, acting as a perfect short circuit at signal frequencies of interest. Similarly, another very large capacitor is used to couple the output signal $v_o$ to other parts of the system.




(a)

**FIGURE 5.55** Example 5.16: (a) circuit

# Solution

We shall start by determining the dc operating point as follows (see Fig. 5.55b):

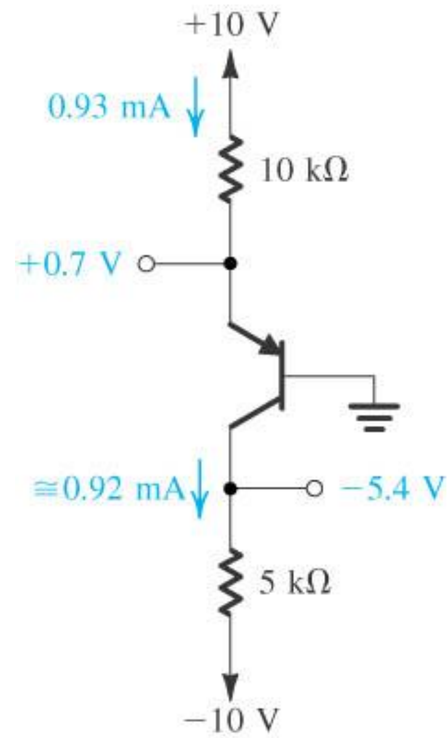
$$I_E = \frac{+10 - V_E}{R_E} \simeq \frac{+10 - 0.7}{10} = 0.93 \text{ mA}$$

Assuming $\beta = 100$, then $\alpha = 0.99$, and

$$I_C = 0.99 I_E = 0.92 \text{ mA}$$

$$V_C = -10 + I_C R_C$$

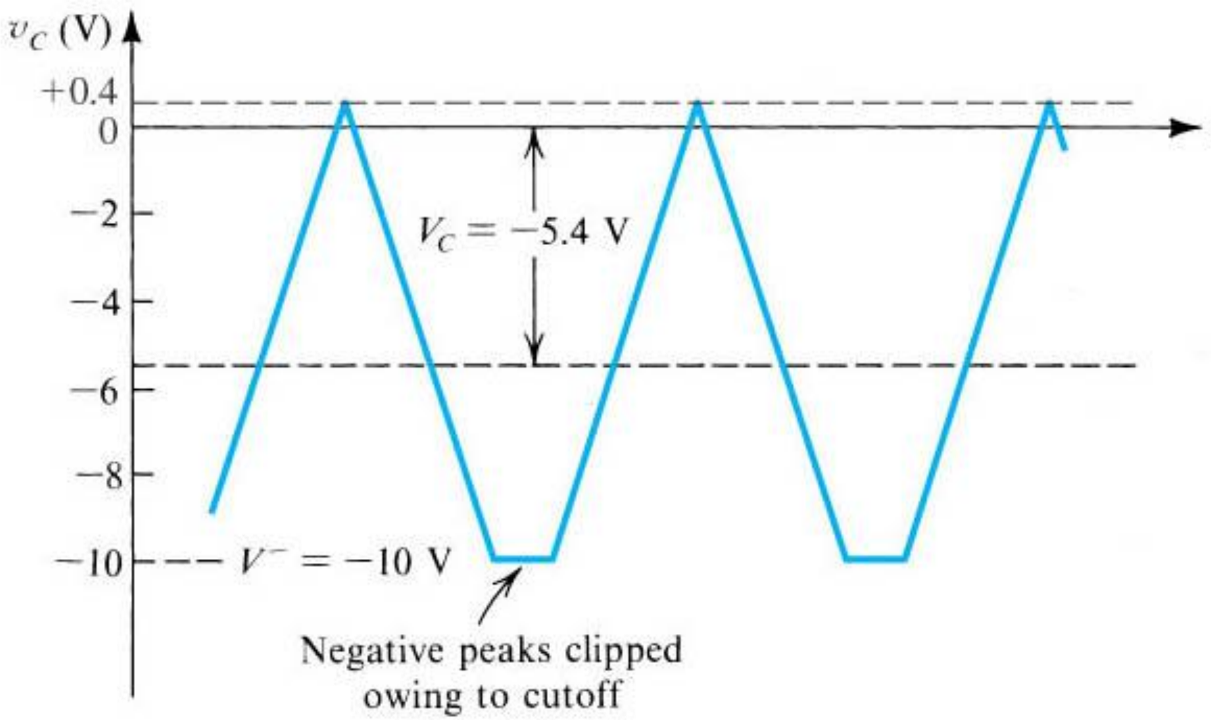
$$= -10 + 0.92 \times 5 = -5.4 \text{ V}$$




(b)

**FIGURE 5.55** Example 5.16: (a) circuit; (b) dc analysis;

Thus the transistor is in the active mode. Furthermore, the collector signal can swing from –5.4 V to +0.4 V (which is 0.4 V above the base voltage) without the transistor going into saturation. However, a negative 5.8-V swing in the collector voltage will (theoretically) cause the minimum collector voltage to be –11.2 V, which is more negative than the power-supply voltage. It follows that if we attempt to apply an input that results in such an output signal, the transistor will cut off and the negative peaks of the output signal will be clipped off, as illustrated in Fig. 5.56. The waveform in Fig. 5.56, however, is shown to be linear (except for the clipped peaks); that is, the effect of the nonlinear $i_C$–$v_{BE}$ characteristic is not taken into account. This is not correct, since if we are driving the transistor into cutoff at the negative signal peaks, then we will surely be exceeding the small-signal limit, as will be shown later.

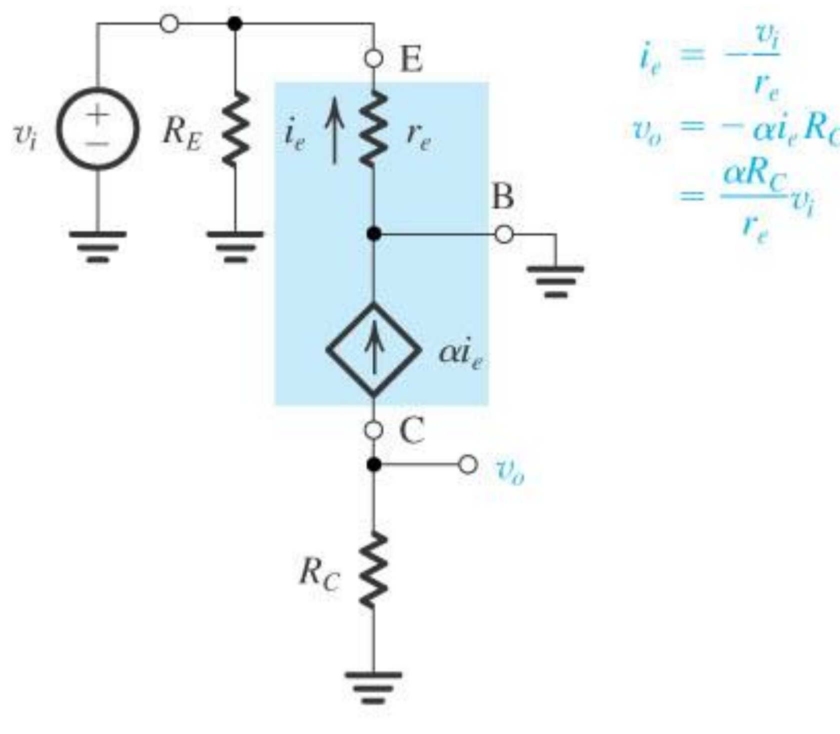


Let us now proceed to determine the small-signal voltage gain. Toward that end, we eliminate the dc sources and replace the BJT with its T equivalent circuit of Fig. 5.52(b). Note that because the base is grounded, the T model is somewhat more convenient than the hybrid-$\pi$ model. Nevertheless, identical results can be obtained using the latter.

Figure 5.55(c) shows the resulting small-signal equivalent circuit of the amplifier. The model parameters are

$$\alpha = 0.99$$

$$r_e = \frac{V_T}{I_E} = \frac{25\text{ mV}}{0.93\text{ mA}} = 27\ \Omega$$



(c)

Analysis of the circuit in Fig. 5.55(c) to determine the output voltage $v_o$ and hence the voltage gain $v_o / v_i$ is straightforward and is given in the figure. The result is

$$A_v = \frac{v_o}{v_i} = 183.3\text{ V/V}$$

Returning to the question of allowable signal magnitude, we observe from Fig. 5.55(c) that $v_{eb} = v_i$. Thus, if small-signal operation is desired (for linearity), then the peak of $v_i$ should be limited to approximately 10 mV. With $\hat{V}_i$ set to this value, as shown for a sine-wave input in Fig. 5.57, the peak amplitude at the collector, $\hat{V}_c$, will be

$$\hat{V}_c = 183.3 \times 0.01 = 1.833 \text{ V}$$

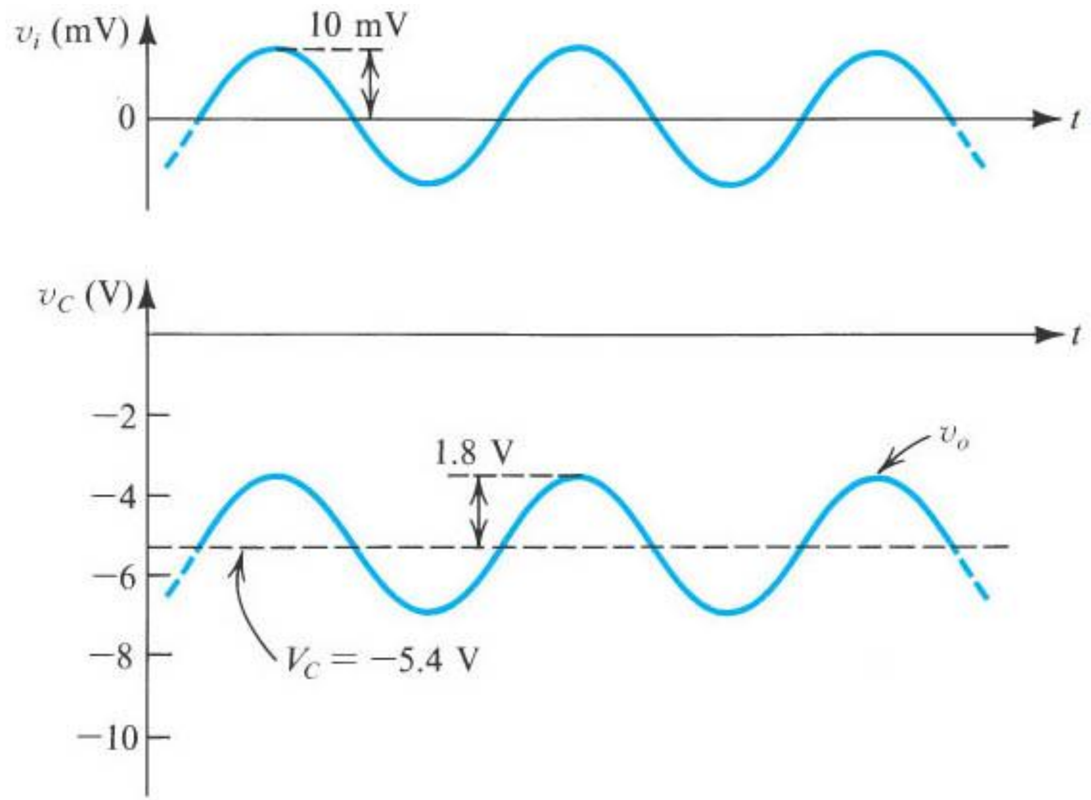



**FIGURE 5.57** Input and output waveforms for the circuit of Fig. 5.55. Observe that this amplifier is noninverting, a property of the common-base configuration.

## SINGLE-STAGE BJT AMPLIFIERS

We have studied the large-signal operation of BJT amplifiers in Section 5.3 and identified the region over which a properly biased transistor can be operated as a linear amplifier for small signals. Methods for dc biasing the BJT were studied in Section 5.5, and a detailed study of small-signal amplifier operation was presented in Section 5.6. We are now ready to consider practical transistor amplifiers, and we will do so in this section for circuits suitable for discrete-circuit fabrication. The design of integrated-circuit BJT amplifiers will be studied in Chapter 6.

There are basically three configurations for implementing single-stage BJT amplifiers: the common-emitter, the common-base, and the common-collector configurations. All three are studied below, utilizing the same basic structure with the same biasing arrangement.

### The Basic Structure

Figure 5.59 shows the basic circuit that we shall utilize to implement the various configurations of BJT amplifiers. Among the various biasing schemes possible for discrete BJT amplifiers (Section 5.5), we have selected, for simplicity and effectiveness, the one employing constant-current biasing. Figure 5.59 indicates the dc currents in all branches and the dc voltages at all nodes. We should note that one would want to select a large value for $R_B$ in order to keep the input resistance at the base large. However, we also want to limit the dc voltage drop across $R_B$ and even more importantly the variability of this dc voltage resulting from the variation in $\beta$ values among transistors of the same type. The dc voltage $V_B$ determines the allowable signal swing at the collector.

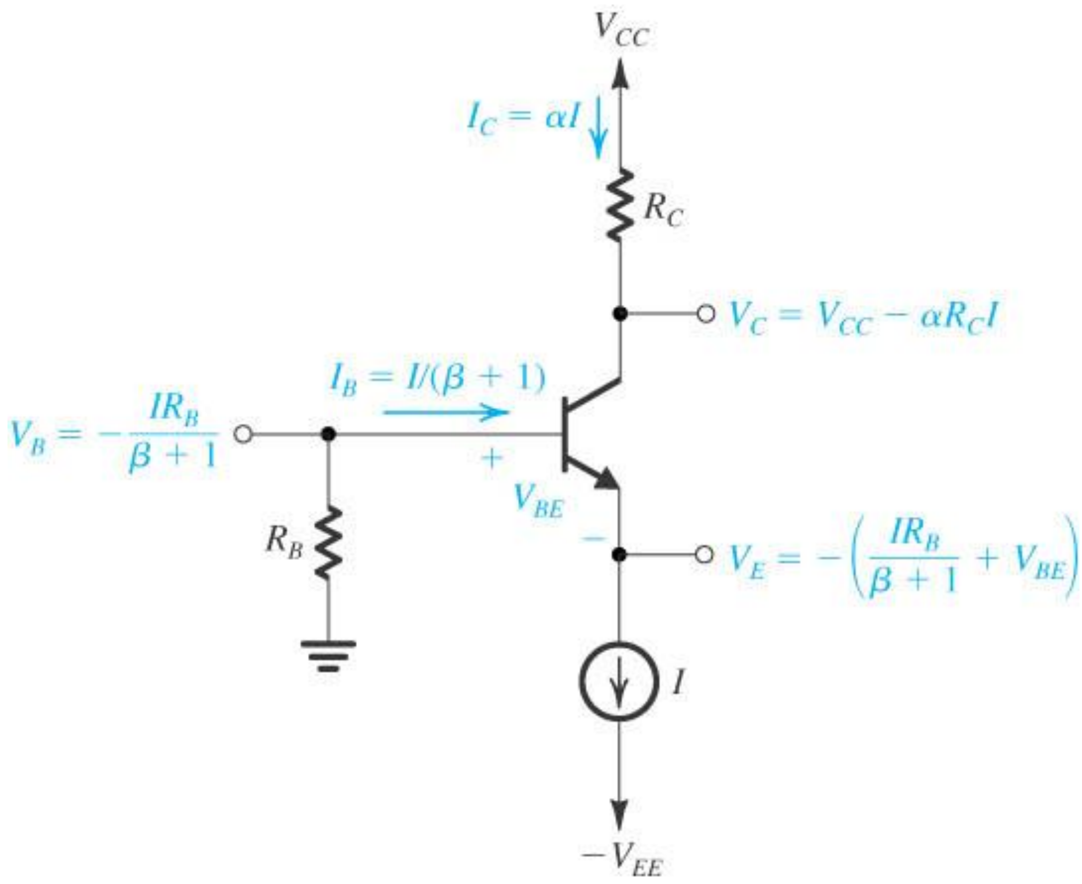



**FIGURE 5.59** Basic structure of the circuit used to realize single-stage, discrete-circuit BJT amplifier configurations.

# Exercise

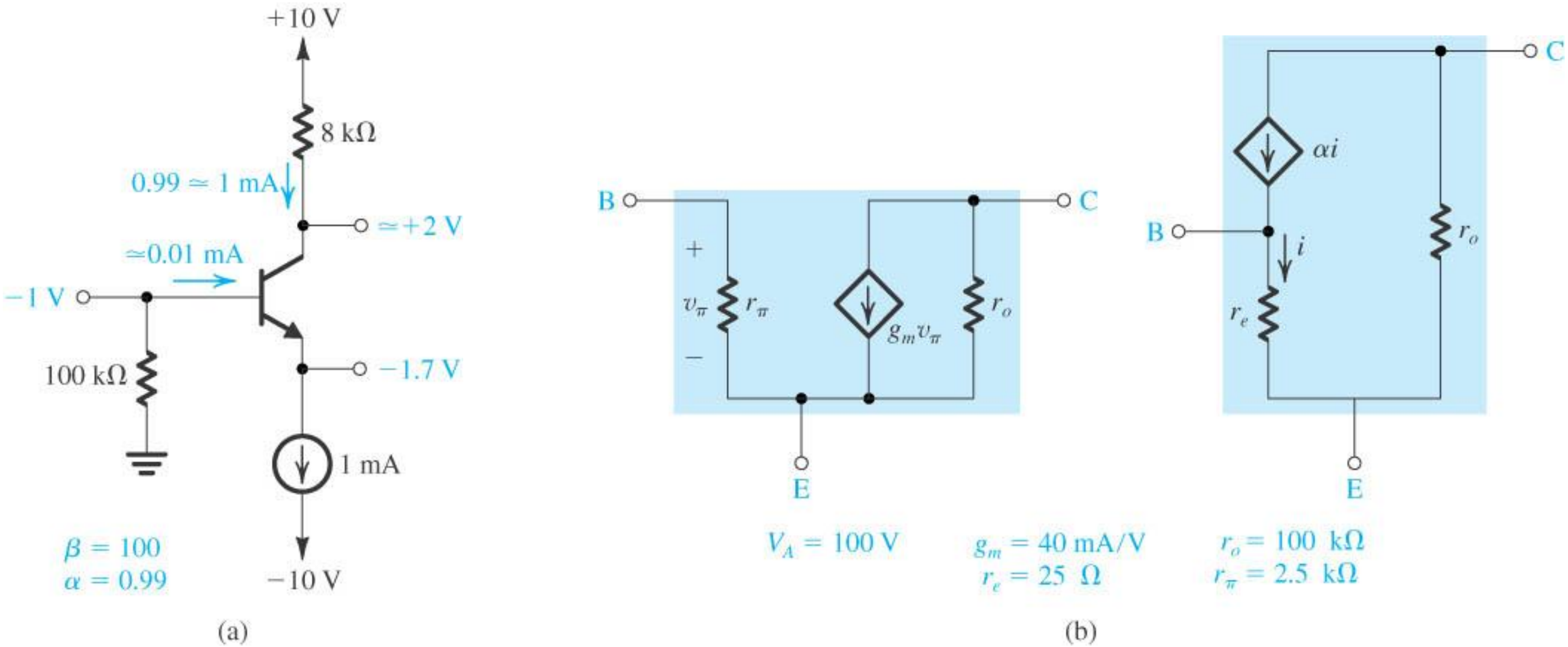
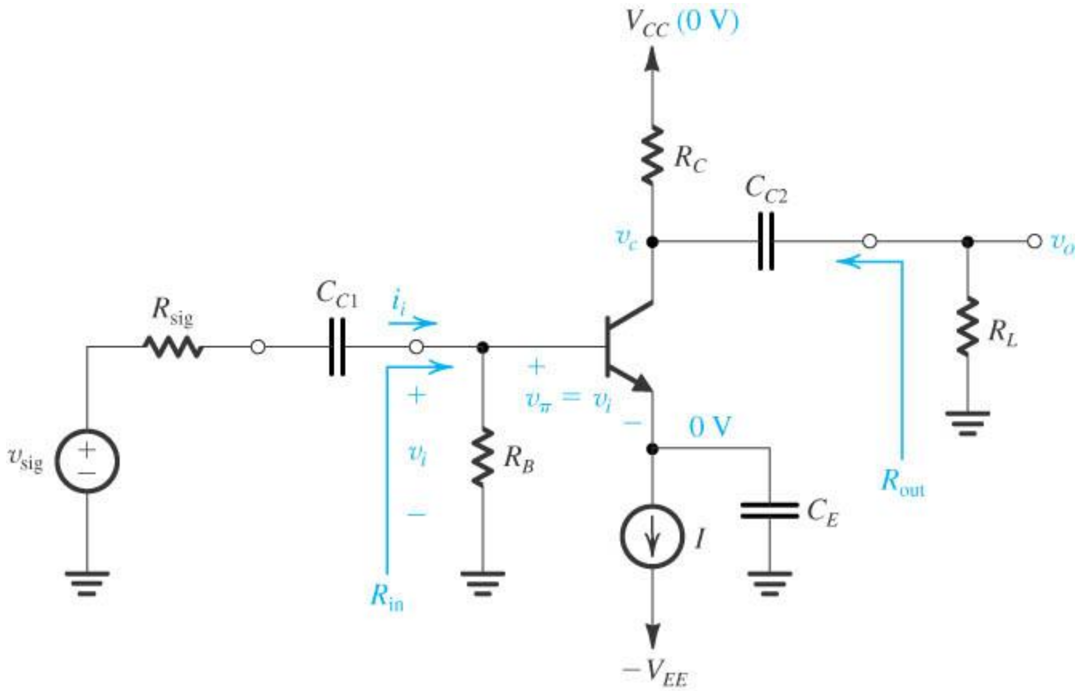
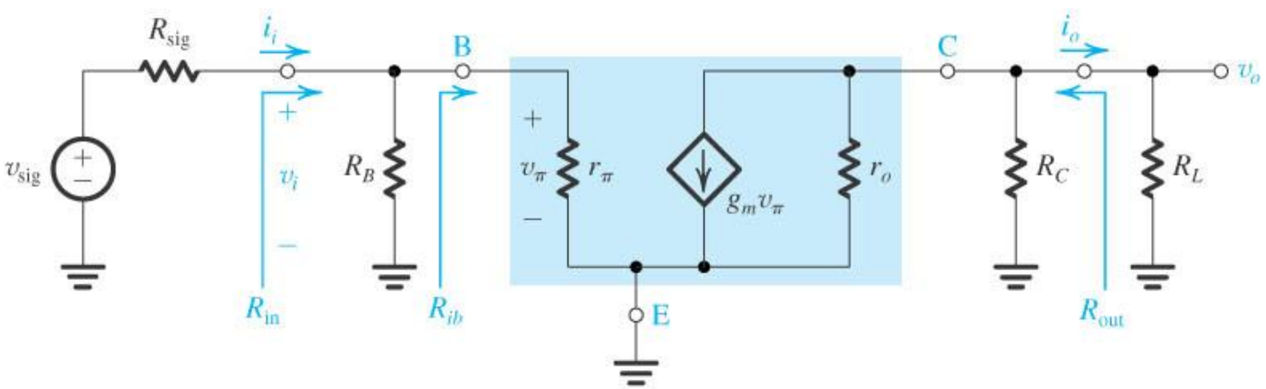



**Figure E5.41**

## The Common-Emitter (CE) Amplifier




(a)




(b)

$$R_{in} \equiv \frac{v_i}{i_i} = R_B \parallel R_{ib} \tag{5.109}$$

where $R_{ib}$ is the input resistance looking into the base. Since the emitter is grounded,

$$R_{ib} = r_\pi \tag{5.110}$$

Normally, we select $R_B \gg r_\pi$, with the result that

$$R_{in} \cong r_\pi \tag{5.111}$$

Thus, we note that the input resistance of the CE amplifier will typically be a few kilohms, which can be thought of as low to moderate. The fraction of source signal $v_{sig}$ that appears across the input terminals of the amplifier proper can be found from

$$v_i = v_{sig}\frac{R_{in}}{R_{in} + R_{sig}} \tag{5.112}$$

$$= v_{sig}\frac{(R_B \parallel r_\pi)}{(R_B + r_\pi) + R_{sig}} \tag{5.113}$$

which for $R_B \gg r_\pi$ becomes

$$v_i \cong v_{sig}\frac{r_\pi}{r_\pi + R_{sig}} \tag{5.114}$$

Next we note that

$$v_\pi = v_i \tag{5.115}$$

At the output of the amplifier we have

$$v_o = -g_m v_\pi (r_o \| R_C \| R_L)$$

Replacing $v_\pi$ by $v_i$ we can write for the voltage gain of the amplifier proper; that is, the voltage gain from base to collector,

$$A_v = -g_m (r_o \| R_C \| R_L) \tag{5.116}$$

This equation simply says that the voltage gain from base to collector is found by multiplying $g_m$ by the total resistance between collector and ground. The open-circuit voltage gain $A_{vo}$ can be obtained by setting $R_L = \infty$ in Eq. (5.116); thus,

$$A_{vo} = -g_m (r_o \| R_C) \tag{5.117}$$

from which we note that the effect of $r_o$ is simply to reduce the gain, usually only slightly since typically $r_o \gg R_C$, resulting in

$$A_{vo} \cong -g_m R_C \tag{5.118}$$

The output resistance $R_{out}$ can be found from the equivalent circuit of Fig. 5.60(b) by looking back into the output terminal while short-circuiting the source $v_{sig}$. Since this will result in $v_\pi = 0$, we see that

$$R_{out} = R_C \parallel r_o \tag{5.119}$$

Thus $r_o$ reduces the output resistance of the amplifier, again usually only slightly since typically $r_o \gg R_C$ and

$$R_{out} \cong R_C \tag{5.120}$$

Recalling that for this unilateral amplifier $R_o = R_{out}$, we can utilize $A_{vo}$ and $R_o$ to obtain the voltage gain $A_v$ corresponding to any particular $R_L$,

$$A_v = A_{vo}\frac{R_L}{R_L + R_o}$$

The reader can easily verify that this approach does in fact lead to the expression for $A_v$ in Eq. (5.116), which we have derived directly.

The overall voltage gain from source to load, $G_v$, can be obtained by multiplying $(v_i/v_{sig})$ from Eq. (5.113) by $A_v$ from Eq. (5.116),

$$G_v = -\frac{(R_B \parallel r_\pi)}{(R_B \parallel r_\pi) + R_{sig}} g_m(r_o \parallel R_C \parallel R_L) \tag{5.121}$$

For the case $R_B \gg r_\pi$, this expression simplifies to

$$G_v \cong -\frac{\beta(R_C \| R_L \| r_o)}{r_\pi + R_{sig}} \tag{5.122}$$

From this expression we note that if $R_{sig} \gg r_\pi$, the overall gain will be highly dependent on $\beta$. This is not a desirable property since $\beta$ varies considerably between units of the same transistor type. At the other extreme, if $R_{sig} \ll r_\pi$, we see that the expression for the overall voltage gain reduces to

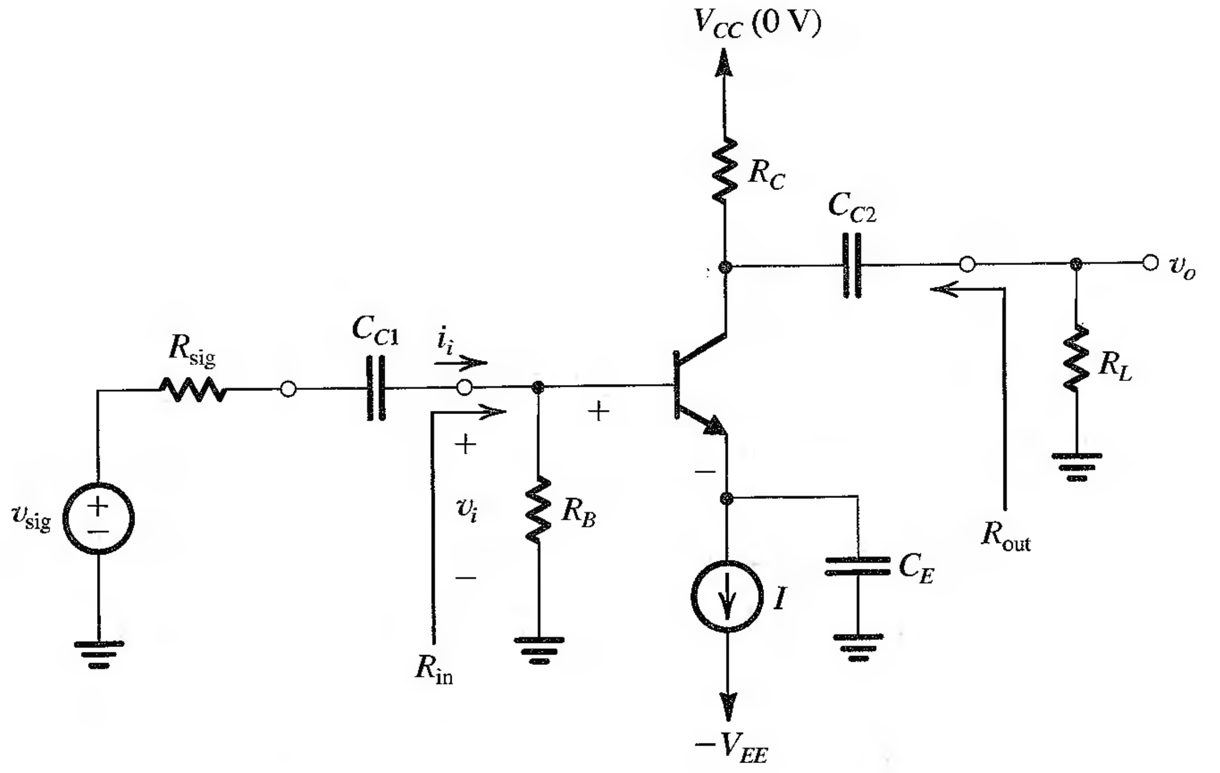
$$G_v \cong -g_m(R_C \| R_L \| r_o) \tag{5.123}$$

which is the gain $A_v$; in other words, when $R_{sig}$ is small, the overall voltage gain is almost equal to the gain of the CE circuit proper, which is independent of $\beta$. Typically a CE amplifier can realize a voltage gain on the order of a few hundred, which is very significant. It follows that the CE amplifier is used to realize the bulk of the voltage gain required in a usual amplifier design. Unfortunately, however, as we shall see in Section 5.9, the high-frequency response of the CE amplifier can be rather limited.

Before leaving the CE amplifier, we wish to evaluate its short-circuit current gain, $A_{is}$. This can be easily done by referring to the amplifier equivalent circuit in Fig. 5.60(b). When $R_L$ is short circuited, the current through it will be equal to $-g_m v_\pi$,

$$i_{os} = -g_m v_\pi$$

# Exercise




$$R_C = 8k\Omega, I = 1mA, V_{CC} = V_{EE} = 10V, R_B = 100k\Omega, \beta = 100$$

$$V_A = 100V, R_{sig} = 5k\Omega, V_T = 25mV$$

# Solution

# The Common-Emitter Amplifier with an Emitter Resistance

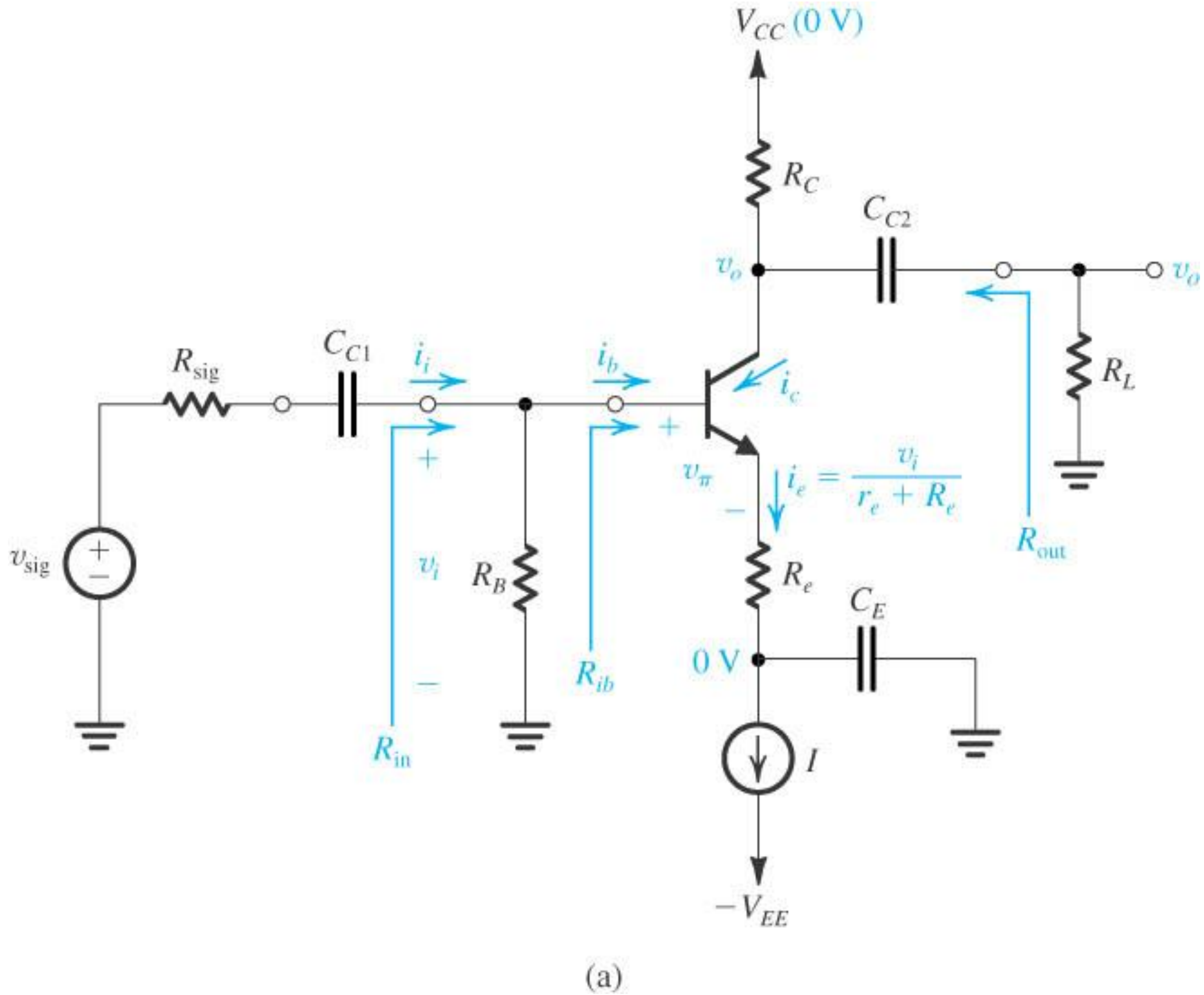



**FIGURE 5.61** **(a)** A common-emitter amplifier with an emitter resistance $R_e$.

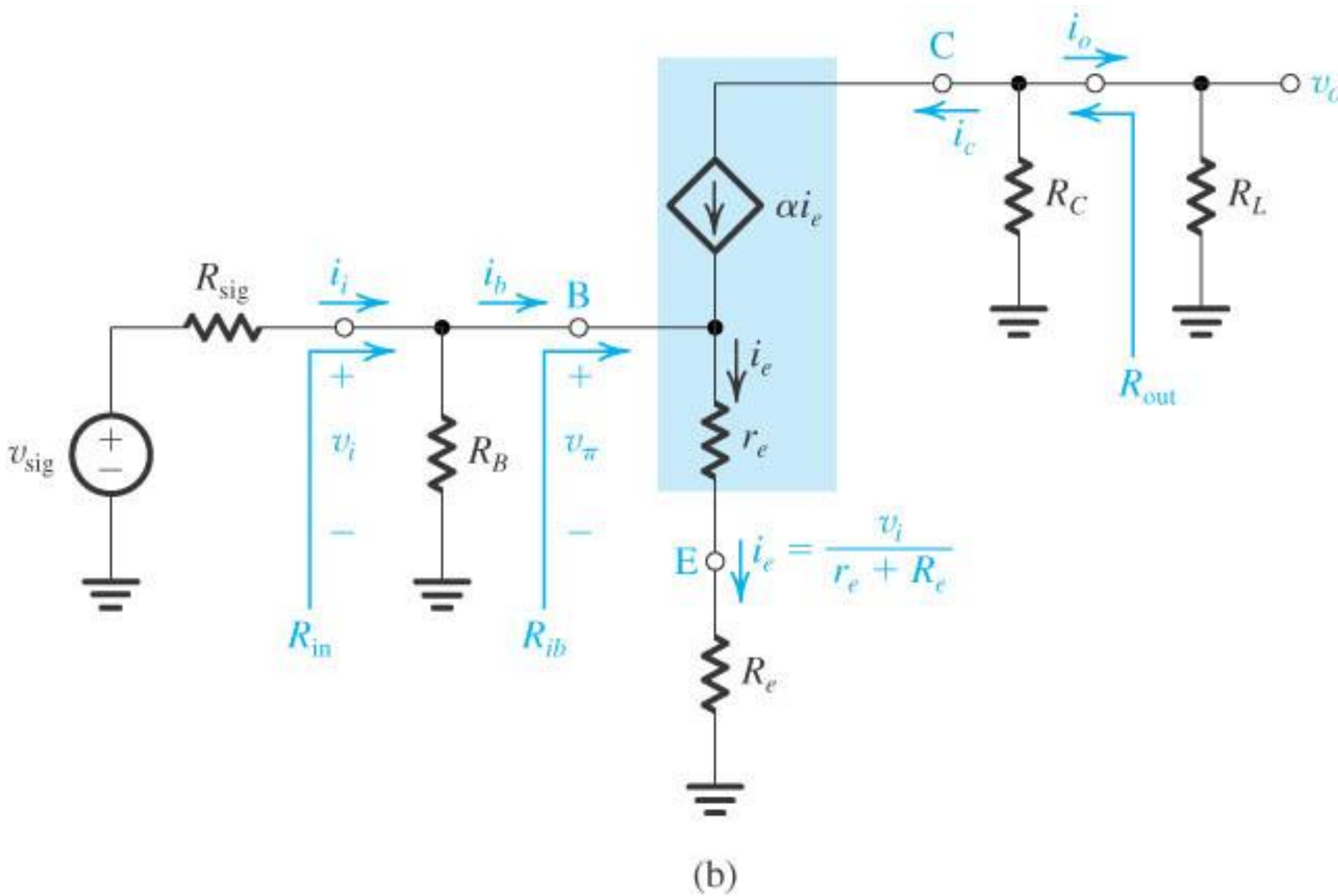



**FIGURE 5.61** **(a)** A common-emitter amplifier with an emitter resistance $R_e$. **(b)** Equivalent circuit obtained by replacing the transistor with its T model.

To determine the amplifier input resistance $R_{\text{in}}$, we note from Fig. 5.61(b) that $R_{\text{in}}$ is the parallel equivalent of $R_B$ and the input resistance at the base $R_{ib}$,

$$R_{\text{in}} = R_B \parallel R_{ib} \tag{5.125}$$

The input resistance at the base $R_{ib}$ can be found from

$$R_{ib} \equiv \frac{v_i}{i_b}$$

where

$$i_b = (1-\alpha)i_e = \frac{i_e}{\beta+1}$$

and

$$i_e = \frac{v_i}{r_e + R_e} \tag{5.126}$$

Thus,

$$R_{ib} = (\beta+1)(r_e + R_e) \tag{5.127}$$

This is a very important result. It says that *the input resistance looking into the base is $(\beta + 1)$ times the total resistance in the emitter*. Multiplication by the factor $(\beta + 1)$ is known as the **resistance-reflection rule.** The factor $(\beta + 1)$ arises because the base current is $1/(\beta + 1)$ times the emitter current. The expression for $R_{ib}$ in Eq. (5.127) shows clearly that including a resistance $R_e$ in the emitter can substantially increase $R_{ib}$. Indeed the value of $R_{ib}$ is increased by the ratio

$$\frac{R_{ib}\,(\text{with } R_e \text{ included})}{R_{ib}\,(\text{without } R_e)} = \frac{(\beta + 1)(r_e + R_e)}{(\beta + 1)r_e}$$

$$= 1 + \frac{R_e}{r_e} \cong 1 + g_m R_e \qquad (5.128)$$

Thus the circuit designer can use the value of $R_e$ to control the value of $R_{ib}$ and hence $R_{\text{in}}$. Of course, for this control to be effective, $R_B$ must be much larger than $R_{ib}$; in other words, $R_{ib}$ must dominate the input resistance.

To determine the voltage gain $A_v$, we see from Fig. 5.61(b) that

$$v_o = -i_c(R_C \parallel R_L)$$

$$= -\alpha i_e(R_C \parallel R_L)$$

Substituting for $i_e$ from Eq. (5.126) gives

$$A_v \equiv \frac{v_o}{v_i} = -\frac{\alpha(R_C \parallel R_L)}{r_e + R_e} \tag{5.129}$$

Since $\alpha \cong 1$,

$$A_v \cong -\frac{R_C \parallel R_L}{r_e + R_e} \tag{5.130}$$

This simple relationship is very useful and is definitely worth remembering: *The voltage gain from base to collector is equal to the ratio of the total resistance in the collector to the total resistance in the emitter*. This statement is a general one and applies to any amplifier circuit. The open-circuit voltage gain $A_{vo}$ can be found by setting $R_L = \infty$ in Eq. (5.129),

$$A_{vo} = -\frac{\alpha R_C}{r_e + R_e} \tag{5.131}$$

which can be expressed alternatively as

$$A_{vo} = -\frac{\alpha}{r_e}\frac{R_C}{1 + R_e/r_e}$$

$$A_{vo} = -\frac{g_m R_C}{1 + (R_e/r_e)} \cong -\frac{g_m R_C}{1 + g_m R_e} \tag{5.132}$$

The output resistance $R_{\text{out}}$ can be found from the circuit in Fig. 5.61(b) by inspection:

$$R_{\text{out}} = R_C \tag{5.133}$$

At this point we should note that for this amplifier, $R_{\text{in}} = R_i$ and $R_{\text{out}} = R_o$.

The short-circuit current gain $A_{is}$ can be found from the circuit in Fig. 5.61(b) as follows:

$$i_{os} = -\alpha i_e$$

$$i_i = v_i / R_{\text{in}}$$

Thus,

$$A_{is} = -\frac{\alpha R_{\text{in}} i_e}{v_i}$$

Substituting for $i_e$ from Eq. (5.126) and for $R_{\text{in}}$ from Eq. (5.125),

$$A_{is} = -\frac{\alpha(R_B \| R_{ib})}{r_e + R_e} \tag{5.134}$$

which for the case $R_B \gg R_{ib}$ reduces to

$$A_{is} = \frac{-\alpha(\beta+1)(r_e + R_e)}{r_e + R_e} = -\beta$$

the same value as for the CE circuit.

The overall voltage gain from source to load can be obtained by multiplying $A_v$ by $(v_i/v_{sig})$,

$$G_v = \frac{v_i}{v_{sig}} \cdot A_v = -\frac{R_{in}}{R_{sig}+R_{in}} \frac{\alpha(R_C \parallel R_L)}{r_e+R_e}$$

Substituting for $R_{in}$ by $R_B \parallel R_{ib}$, assuming that $R_B \gg R_{ib}$, and substituting for $R_{ib}$ from Eq. (5.127) results in

$$G_v \cong -\frac{\beta(R_C \parallel R_L)}{R_{sig}+(\beta+1)(r_e+R_e)} \tag{5.135}$$

We note that the gain is lower than that of the CE amplifier because of the additional term $(\beta+1)R_e$ in the denominator. The gain, however, is less sensitive to the value of $\beta$, a desirable result.

Another important consequence of including the resistance $R_e$ in the emitter is that it enables the amplifier to handle larger input signals without incurring nonlinear distortion. This is because only a fraction of the input signal at the base, $v_i$, appears between the base and the emitter. Specifically, from the circuit in Fig. 5.61(b), we see that

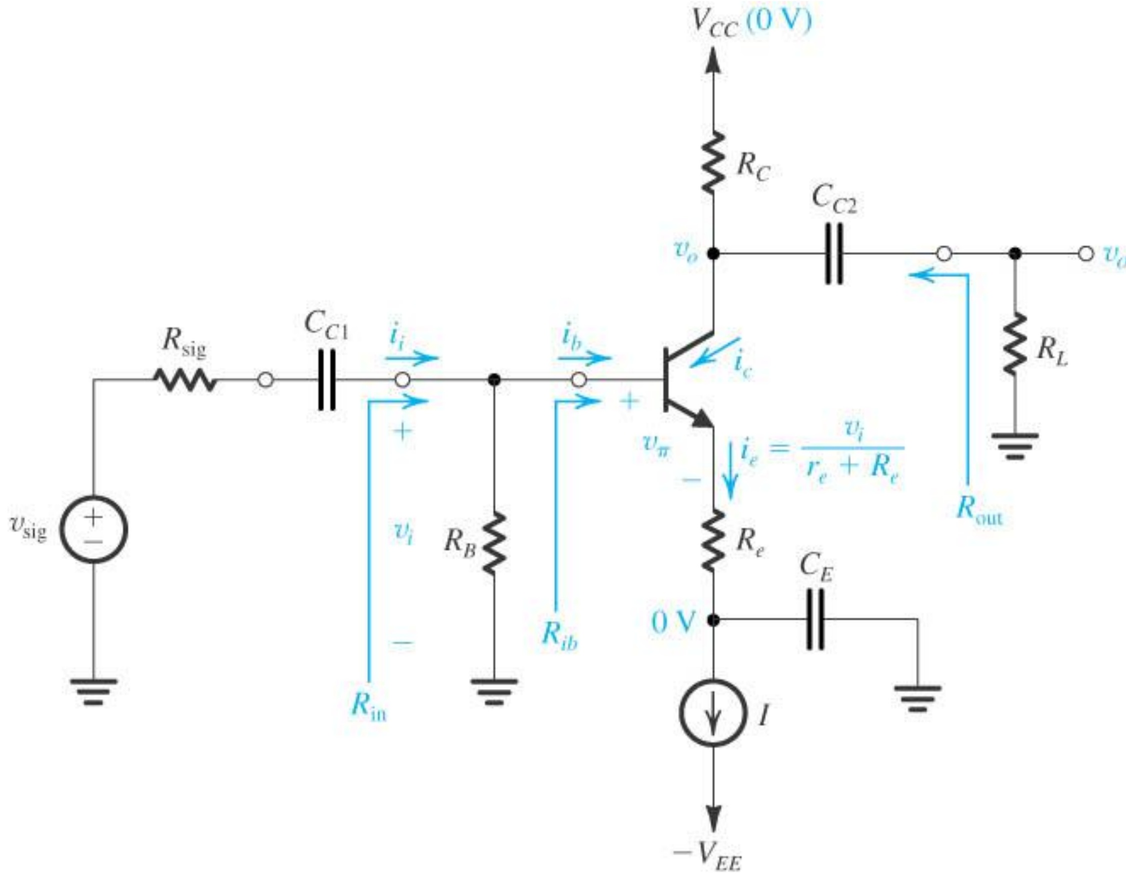
$$\frac{v_\pi}{v_i} = \frac{r_e}{r_e+R_e} \cong \frac{1}{1+g_m R_e} \tag{5.136}$$

Thus, for the same $v_\pi$, the signal at the input terminal of the amplifier, $v_i$, can be greater than for the CE amplifier by the factor $(1 + g_m R_e)$.

To summarize, including a resistance $R_e$ in the emitter of the CE amplifier results in the following characteristics:

1. The input resistance $R_{ib}$ is increased by the factor $(1 + g_m R_e)$.
2. The voltage gain from base to collector, $A_v$, is reduced by the factor $(1 + g_m R_e)$.
3. For the same nonlinear distortion, the input signal $v_i$ can be increased by the factor $(1 + g_m R_e)$.
4. The overall voltage gain is less dependant on the value of $\beta$.
5. The high-frequency response is significantly improved

# Exercise



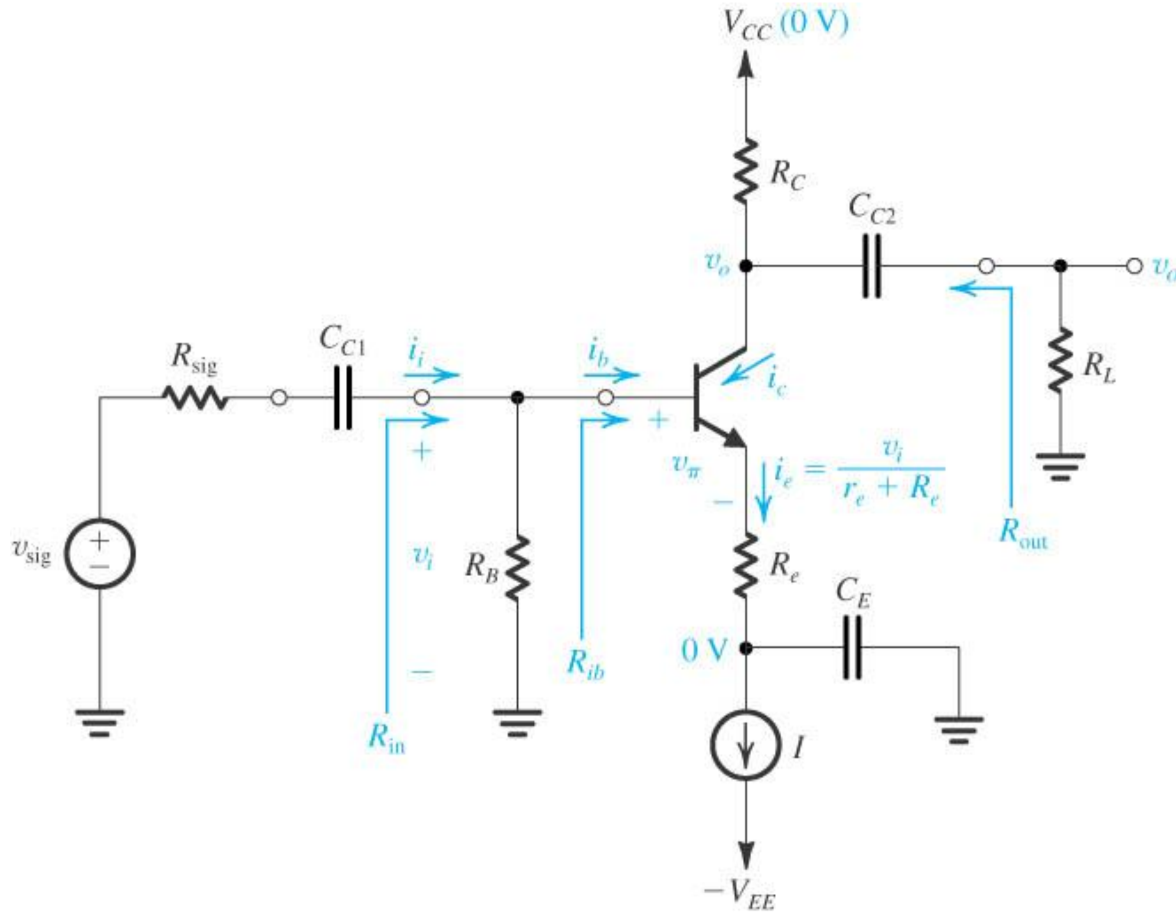

(a)

$$R_C = 8k\Omega, I = 1mA, V_{CC} = V_{EE} = 10V, R_B = 100k\Omega, \beta = 100$$

$$V_A = 100V, R_{sig} = R_L = 5k\Omega, V_T = 25mV$$

# Solution

# Exercise




$$R_C = 8k\Omega, I = 1mA, V_{CC} = V_{EE} = 10V, R_B = 100k\Omega, \beta = 100$$

$$V_A = 100V, R_{sig} = 5k\Omega, V_T = 25mV$$

**Find Re that results in *Rin* equal to four times the source resistance *Rsig*. For this value, find *Avo, Rout, Av, Gv.***

# Solution

# The Common-Base (CB) Amplifier

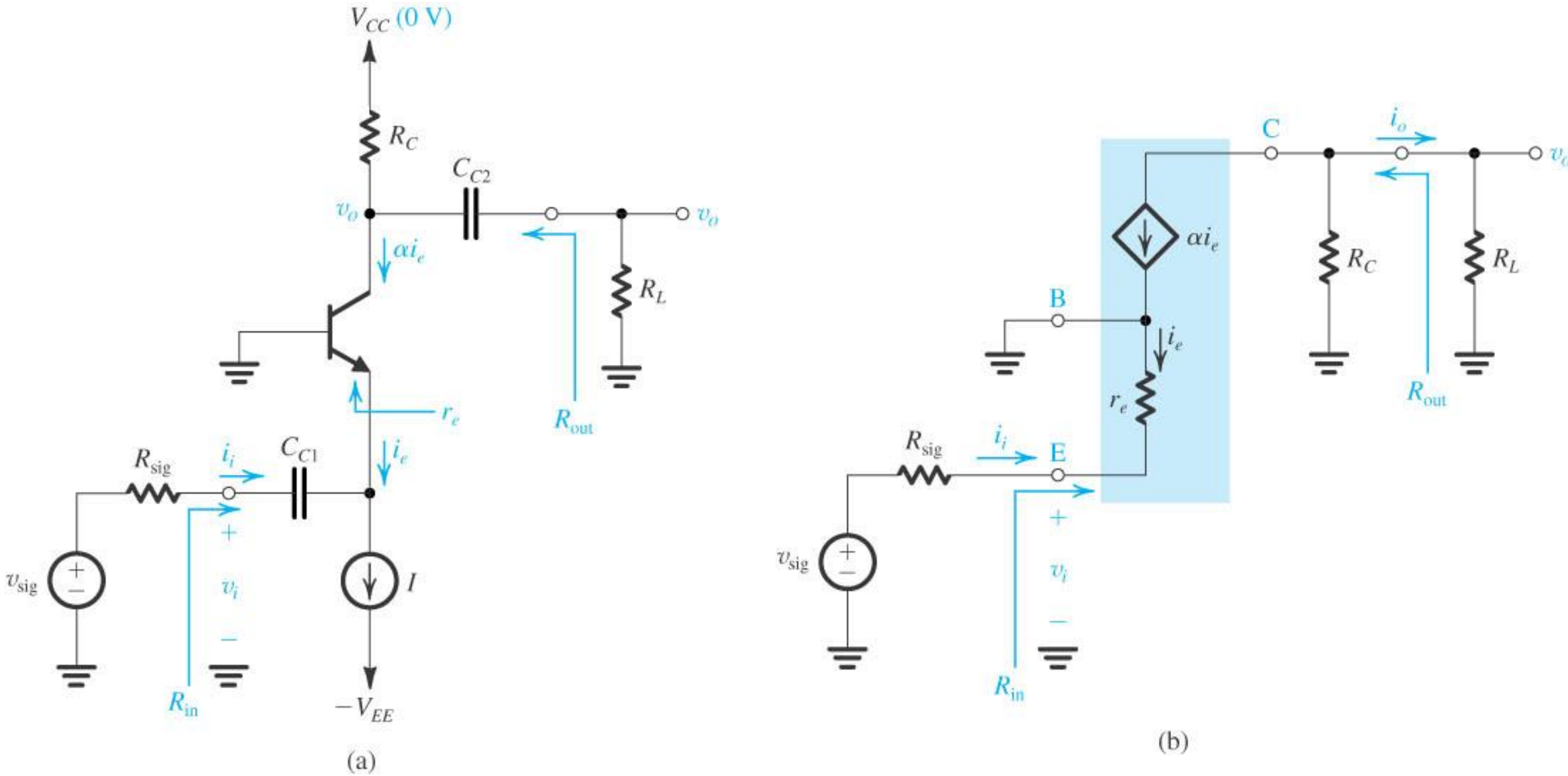



**Figure 5.62** **(a)** A common-base amplifier using the structure of Fig. 5.59. **(b)** Equivalent circuit obtained by replacing the transistor with its T model.

From inspection of the equivalent circuit model in Fig. 5.62(b), we see that the input resistance is

$$R_{\text{in}} = r_e \tag{5.137}$$

To determine the voltage gain, we write at the collector node

$$v_o = -\alpha i_e (R_C \parallel R_L)$$

and substitute for the emitter current from

$$i_e = -\frac{v_i}{r_e}$$

to obtain

$$A_v \equiv \frac{v_o}{v_i} = \frac{\alpha}{r_e}(R_C \parallel R_L) = g_m (R_C \parallel R_L) \tag{5.138}$$

The open-circuit voltage gain $A_{vo}$ can be found from Eq. (5.138) by setting $R_L = \infty$:

$$A_{vo} = g_m R_C \tag{5.139}$$

$$R_{\text{out}} = R_C$$

The short-circuit current gain $A_{is}$ is given by

$$A_{is} = \frac{-\alpha i_e}{i_i} = \frac{-\alpha i_e}{-i_e} = \alpha \tag{5.140}$$

Although the gain of the CB amplifier proper has the same magnitude as that of the CE amplifier, this is usually not the case as far as the overall voltage gain is concerned. The low input resistance of the CB amplifier can cause the input signal to be severely attenuated, specifically

$$\frac{v_i}{v_{\text{sig}}} = \frac{R_i}{R_{\text{sig}} + R_i} = \frac{r_e}{R_{\text{sig}} + r_e} \tag{5.141}$$

The overall voltage gain $G_v$ of the CB amplifier can be obtained by multiplying the ratio $v_i / v_{\text{sig}}$ of Eq. (5.141) by $A_v$ from Eq. (5.138),

$$G_v = \frac{r_e}{R_{\text{sig}} + r_e} g_m (R_C \parallel R_L)$$

$$= \frac{\alpha (R_C \parallel R_L)}{R_{\text{sig}} + r_e} \tag{5.142}$$

# The Common-Base (CB) Amplifier

In summary, the CB amplifier exhibits a very low input resistance ($r_e$), a short-circuit current gain that is nearly unity ($\alpha$), an open-circuit voltage gain that is positive and equal in magnitude to that of the CE amplifier ($g_m R_C$), and like the CE amplifier, a relatively high output resistance ($R_C$). Because of its very low input resistance, the CB circuit *alone* is not attractive as a voltage amplifier except in specialized applications, such as the cable amplifier mentioned above. The CB amplifier has excellent high-frequency performance as well, which as we shall see in Chapter 6 makes it useful together with other circuits in the implementation of high-frequency amplifiers. Finally, a very significant application of the CB circuit is as a unity-gain current amplifier or **current buffer:** It accepts an input signal current at a low input resistance and delivers a nearly equal current at very high output resistance at the collector (the output resistance excluding $R_C$ and neglecting $r_o$ is infinite).

# The Common-Collector (CC) Amplifier or Emitter Follower

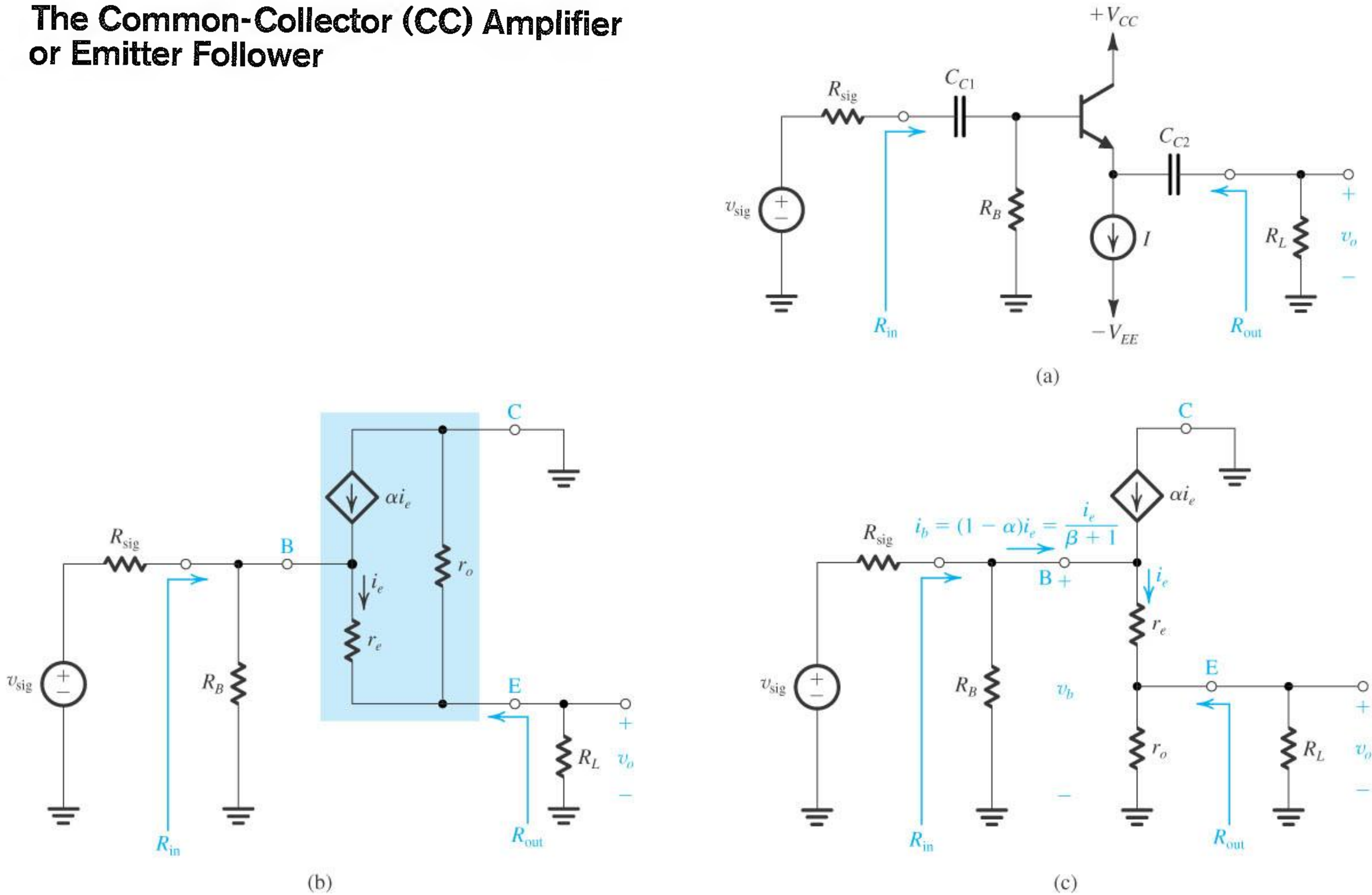


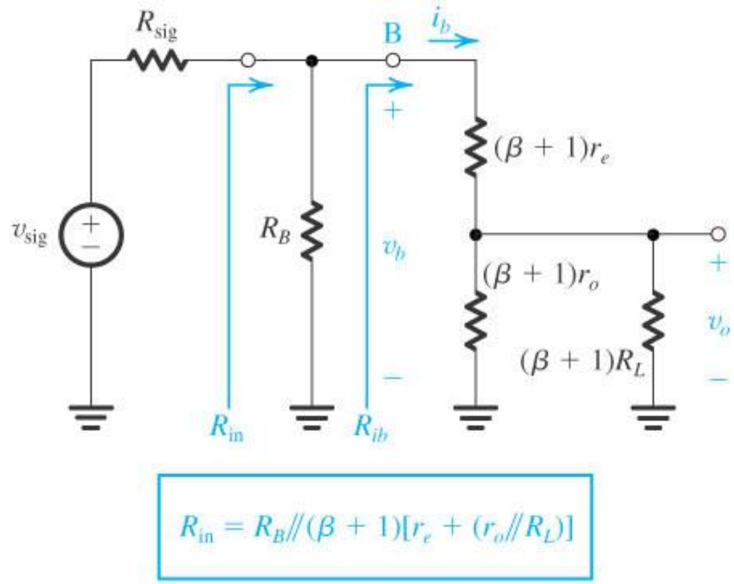

**Figure 5.63** **(a)** An emitter-follower circuit based on the structure of Fig. 5.59. **(b)** Small-signal equivalent circuit of the emitter follower with the transistor replaced by its T model augmented with $r_o$. **(c)** The circuit in (b) redrawn to emphasize that $r_o$ is in parallel with $R_L$. This simplifies the analysis considerably.

Inspection of the circuit in Fig. 5.64(a) shows that the input resistance at the base, $R_{ib}$, is

$$R_{ib} = (\beta+1)[r_e + (r_o \| R_L)] \tag{5.143}$$

from which we see that the emitter follower acts to raise the resistance level of $R_L$ (or $R_L \| r_o$ to be exact) by the factor $(\beta+1)$ and presents to the source the increased resistance. The total input resistance of the follower is

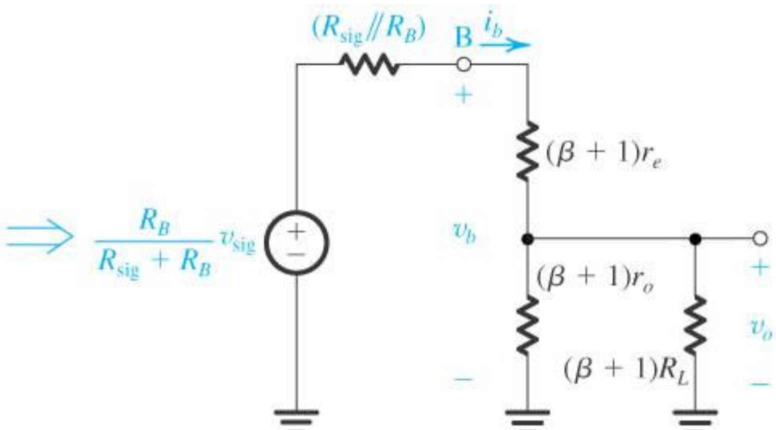
$$R_{in} = R_B \| R_{ib}$$




(a)

To find the overall voltage gain $G_v$, we first apply Thévenin theorem at the input side of the circuit in Fig. 5.64(a) to simplify it to the form shown in Fig. 5.64(b). From the latter circuit we see that $v_o$ can be found by utilizing the voltage divider rule; thus,

$$G_v = \frac{R_B}{R_{sig}+R_B}\frac{(\beta+1)(r_o \| R_L)}{(R_{sig}\| R_B)+(\beta+1)[r_e+(r_o\| R_L)]} \tag{5.144}$$




$$G_v = \frac{v_o}{v_{sig}} = \frac{R_B}{R_{sig}+R_B}\frac{(\beta+1)(r_o/\!/R_L)}{(R_{sig}/\!/R_B)+(\beta+1)[r_e+(r_o/\!/R_L)]}$$

(b)

$$R_{out} = r_o \| \left(r_e + \frac{R_{sig}\| R_B}{\beta+1}\right)$$

## The Common-Collector (CC) Amplifier or Emitter Follower

In summary, the emitter follower exhibits a high input resistance, a low output resistance, a voltage gain that is smaller than but close to unity, and a relatively large current gain. It is therefore ideally suited for applications in which a high-resistance source is to be connected to a low-resistance load—namely, as the last stage or output stage in a multistage amplifier, where its purpose would be not to supply additional voltage gain but rather to give the cascade amplifier a low output resistance. We shall study the design of amplifier output stages in Chapter 14.

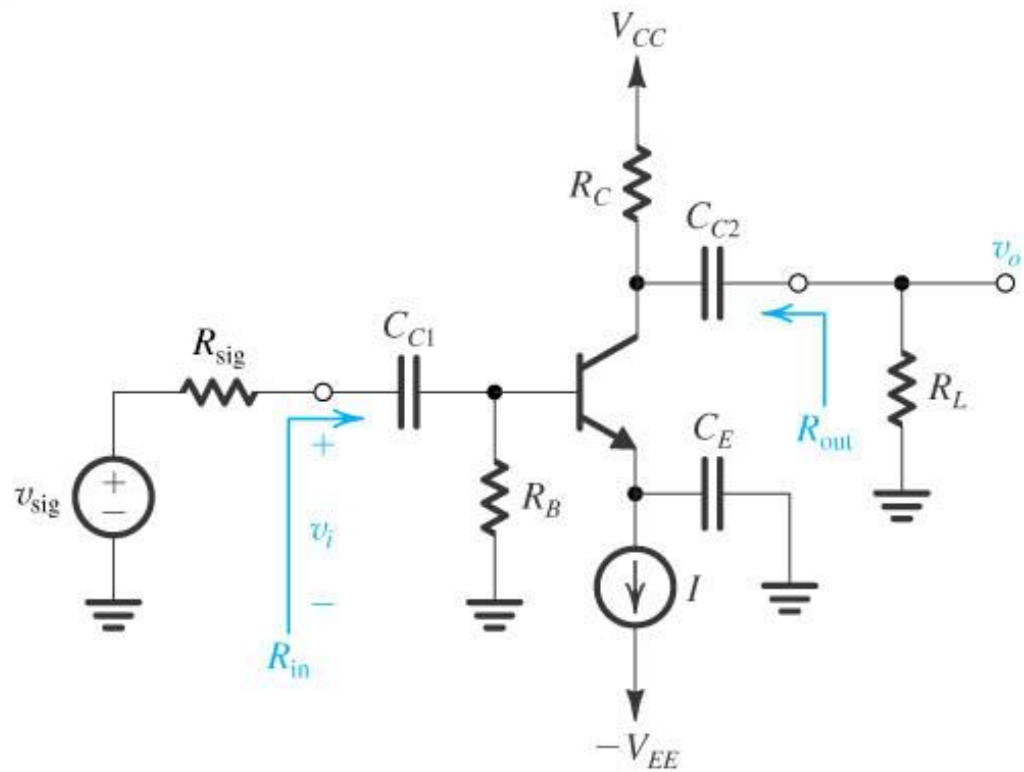
Before leaving the emitter follower, the question of the maximum allowed signal swing deserves comment. Since only a small fraction of the input signal appears between the base and the emitter, the emitter follower exhibits linear operation for a wide range of input-signal amplitude. There is, however, an absolute upper limit imposed on the value of the output-signal amplitude by transistor cutoff.

## Summary and Comparisons

For easy reference and to enable comparisons, we present in Table 5.6 the formulas for determining the characteristic parameters of discrete single-stage BJT amplifiers. In addition to the remarks already made throughout this section about the characteristics and areas of applicability of the various configurations, we make the following concluding points:

1. The CE configuration is the best suited for realizing the bulk of the gain required in an amplifier. Depending on the magnitude of the gain required, either a single stage or a cascade of two or three stages can be used.
2. Including a resistor $R_e$ in the emitter lead of the CE stage provides a number of performance improvements at the expense of gain reduction.
3. The low input resistance of the CB amplifier makes it useful only in specific applications. As we shall see in Chapter 6, it has a much better high-frequency response than the CE amplifier. This superiority will make it useful as a high-frequency amplifier, especially when combined with the CE circuit. We shall see one such combination in Chapter 6.
4. The emitter follower finds application as a voltage buffer for connecting a high-resistance source to a low-resistance load and as the output stage in a multistage amplifier.

# Common Emitter




$$R_{in} = R_B \| r_\pi = R_B \| (\beta+1)r_e$$

$$A_v = -g_m(r_o \| R_C \| R_L)$$

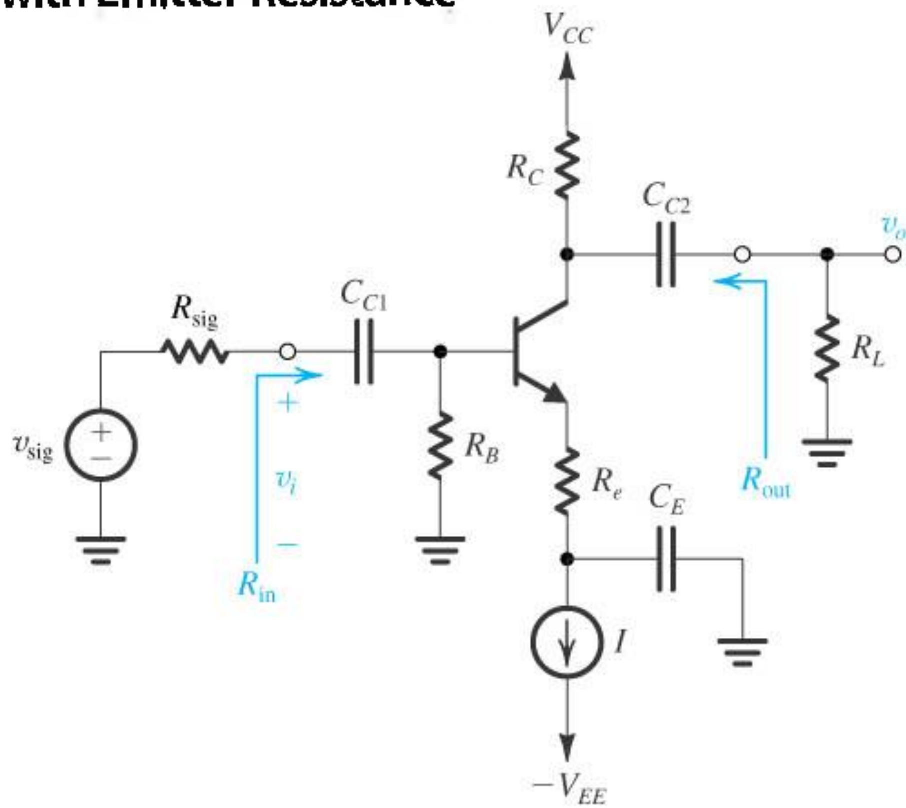
$$R_{out} = r_o \| R_C$$

$$G_v = -\frac{(R_B \| r_\pi)}{(R_B \| r_\pi) + R_{sig}} g_m(r_o \| R_C \| R_L)$$

$$\cong -\frac{\beta(r_o \| R_C \| R_L)}{r_\pi + R_{sig}}$$

$$A_{is} = -g_m R_{in} \cong -\beta$$

# Common Emitter with Emitter Resistance




Neglecting $r_o$:

$$R_{in} = R_B \parallel (\beta + 1)(r_e + R_e)$$

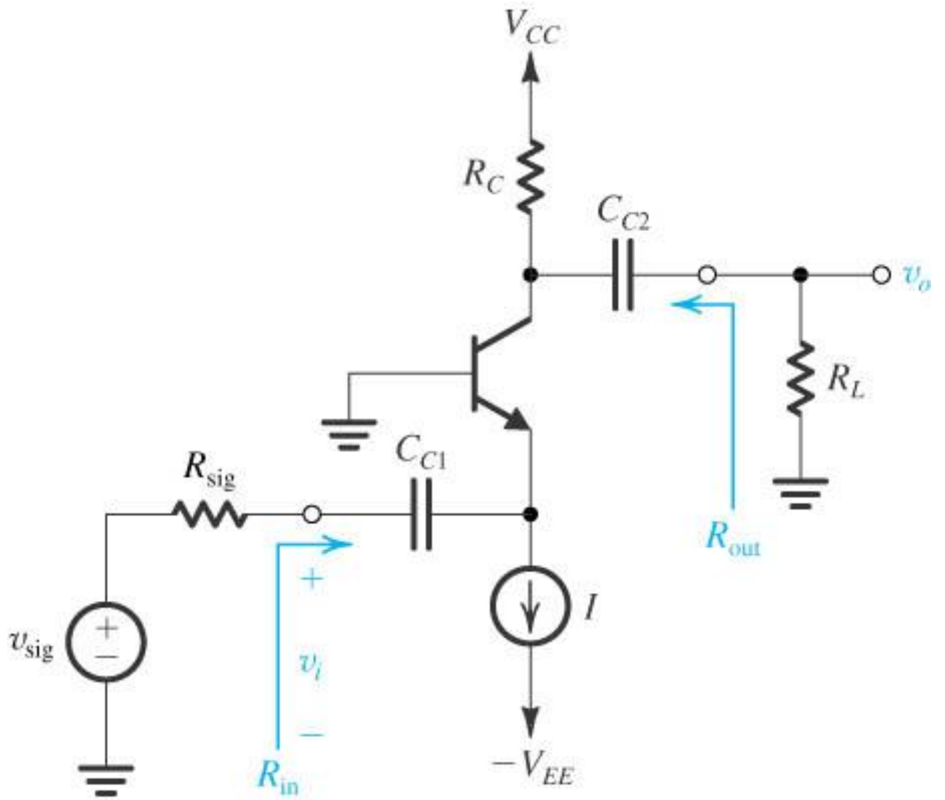
$$A_v = -\frac{\alpha(R_C \parallel R_L)}{r_e + R_e} \cong \frac{-g_m(R_C \parallel R_L)}{1 + g_m R_e}$$

$$R_{out} = R_C$$

$$G_v \cong -\frac{\beta(R_C \parallel R_L)}{R_{sig} + (\beta + 1)(r_e + R_e)}$$

$$\frac{v_\pi}{v_i} \cong \frac{1}{1 + g_m R_e}$$

# Common Base




Neglecting $r_o$:

$$R_{in} = r_e$$

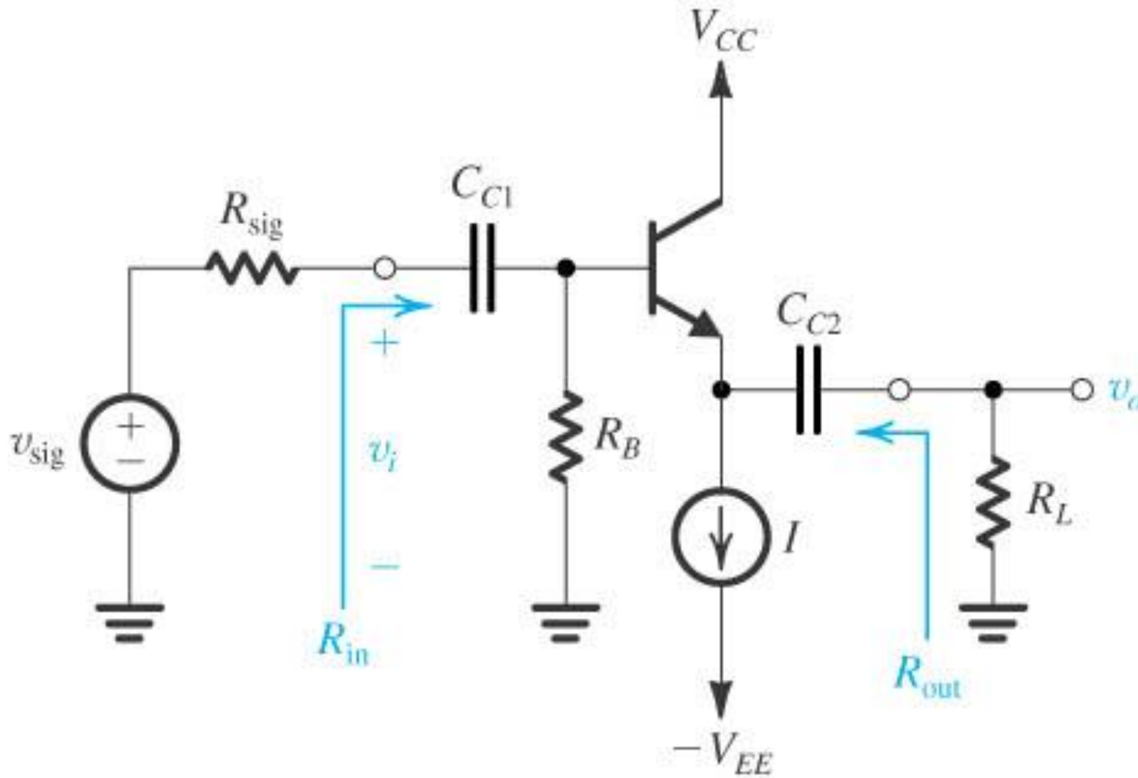
$$A_v = g_m(R_C \parallel R_L)$$

$$R_{out} = R_C$$

$$G_v = \frac{\alpha(R_C \parallel R_L)}{R_{sig} + r_e}$$

$$A_{is} \cong \alpha$$

# Common Collector or Emitter Follower




$$R_{in} = R_B \,\|\, (\beta+1)[r_e + (r_o \,\|\, R_L)]$$

$$A_v = \frac{(r_o \,\|\, R_L)}{(r_o \,\|\, R_L) + r_e}$$

$$R_{out} = r_o \,\|\, \left[ r_e + \frac{R_{sig} \,\|\, R_B}{\beta+1} \right]$$

$$G_v = \frac{R_B}{R_B + R_{sig}} \frac{(r_o \,\|\, R_L)}{\dfrac{R_{sig} \,\|\, R_B}{\beta+1} + r_e + (r_o \,\|\, R_L)}$$

$$A_{is} \cong \beta + 1$$