CHAPTER 6

Small-Signal Midfrequency BJT Amplifiers

6.1. INTRODUCTION

For sufficiently small emitter-collector voltage and current excursions about the quiescent point (*small signals*), the BJT is considered linear; it may then be replaced with any of several two-port networks of impedances and controlled sources (called *small-signal equivalent-circuit models*), to which standard network analysis methods are applicable. Moreover, there is a range of signal frequencies which are large enough so that coupling or bypass capacitors (see Section 3.7) can be considered short circuits, yet low enough so that inherent capacitive reactances associated with BJTs can be considered open circuits. In this chapter, all BJT voltage and current signals are assumed to be in this *midfrequency range*.

In practice, the design of small-signal amplifiers is divided into two parts: (1) setting the dc bias or Q point (Chapters 3 and 5), and (2) determining voltage- or current-gain ratios and impedance values at signal frequencies.

6.2. HYBRID-PARAMETER MODELS

General hybrid-parameter analysis of two-port networks was introduced in Section 1.7. Actually, different sets of h parameters are defined, depending on which element of the transistor (E, B, or C) shares a common point with the amplifier input and output terminals.

Common-Emitter Transistor Connection

From Fig. 3-3(b) and (c), we see that if i_C and v_{BE} are taken as dependent variables in the CE transistor configuration, then

$$v_{BE} = f_1(i_B, v_{CE}) \tag{6.1}$$

$$i_C = f_2(i_B, v_{CE}) \tag{6.2}$$

If the total emitter-to-base voltage v_{BE} goes through only *small* excursions (ac signals) about the Q point, then $\Delta v_{BE} = v_{be}$, $\Delta i_C = i_c$, and so on. Therefore, after applying the chain rule to (*6.1*) and (*6.2*), we have, respectively,

$$v_{be} = \Delta v_{BE} \approx dv_{BE} = \left.\frac{\partial v_{BE}}{\partial i_B}\right|_Q i_b + \left.\frac{\partial v_{BE}}{\partial v_{CE}}\right|_Q v_{ce} \tag{6.3}$$

$$i_c = \Delta i_C \approx di_C = \left.\frac{\partial i_C}{\partial i_B}\right|_Q i_b + \left.\frac{\partial i_C}{\partial v_{CE}}\right|_Q v_{ce} \tag{6.4}$$

The four partial derivatives, evaluated at the Q point, that occur in (*6.3*) and (*6.4*) are called *CE hybrid parameters* and are denoted as follows:

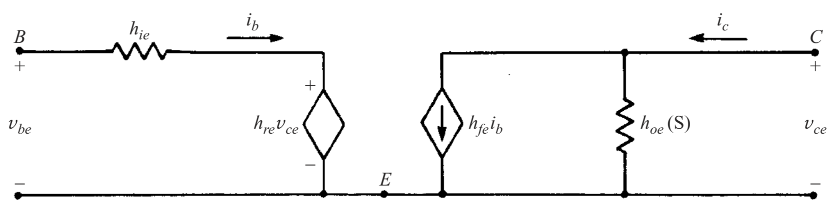
$$\textit{Input resistance} \qquad h_{ie} \equiv \left.\frac{\partial v_{BE}}{\partial i_B}\right|_Q \approx \left.\frac{\Delta v_{BE}}{\Delta i_B}\right|_Q \tag{6.5}$$

$$\textit{Reverse voltage ratio} \qquad h_{re} \equiv \left.\frac{\partial v_{BE}}{\partial v_{CE}}\right|_Q \approx \left.\frac{\Delta v_{BE}}{\Delta v_{CE}}\right|_Q \tag{6.6}$$

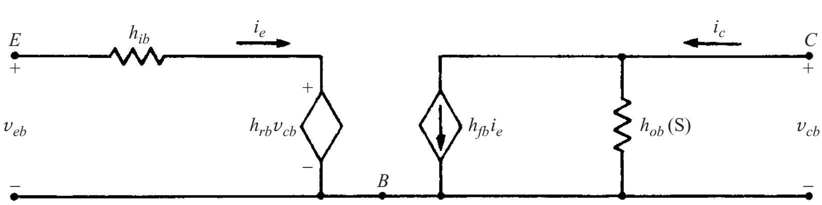
$$\textit{Forward current gain} \qquad h_{fe} \equiv \left.\frac{\partial i_C}{\partial i_B}\right|_Q \approx \left.\frac{\Delta i_C}{\Delta i_B}\right|_Q \tag{6.7}$$

$$\textit{Output admittance} \qquad h_{oe} \equiv \left.\frac{\partial i_C}{\partial v_{CE}}\right|_Q \approx \left.\frac{\partial \Delta i_C}{\Delta v_{CE}}\right|_Q \tag{6.8}$$

The equivalent circuit for (*6.3*) and (*6.4*) is shown in Fig. 6-1(*a*). The circuit is valid for use with signals whose excursion about the Q point is sufficiently small so that the h parameters may be treated as constants.



(*a*) CE small-signal equivalent circuit



(*b*) CB small-signal equivalent circuit

Fig. 6-1

Common-Base Transistor Connection

If v_{EB} and i_C are taken as the dependent variables for the CB transistor characteristics of Fig. 3-2(*b*) and (*c*), then, as in the CE case, equations can be found specifically for small excursions about the Q point. The results are

$$v_{eb} = h_{ib}i_e + h_{rb}v_{cb} \tag{6.9}$$

$$i_c = h_{fb}i_e + h_{ob}v_{cb} \tag{6.10}$$

The partial-derivative definitions of the CB h-parameters are:

$$\textit{Input resistance} \qquad h_{ib} \equiv \left.\frac{\partial v_{EB}}{\partial i_E}\right|_Q \approx \left.\frac{\Delta v_{EB}}{\Delta i_E}\right|_Q \tag{6.11}$$

$$\textit{Reverse voltage ratio} \qquad h_{rb} \equiv \left.\frac{\partial v_{EB}}{\partial v_{CB}}\right|_Q \approx \left.\frac{\Delta v_{EB}}{\Delta v_{CB}}\right|_Q \tag{6.12}$$

$$\textit{Forward current gain} \qquad h_{fb} \equiv \left.\frac{\partial i_C}{\partial i_E}\right|_Q \approx \left.\frac{\Delta i_C}{\Delta i_E}\right|_Q \tag{6.13}$$

$$\textit{Output admittance} \qquad h_{ob} \equiv \left.\frac{\partial i_C}{\partial v_{CB}}\right|_Q \approx \left.\frac{\Delta i_C}{\Delta v_{CB}}\right|_Q \tag{6.14}$$

A small-signal, h-parameter equivalent circuit satisfying (*6.9*) and (*6.10*) is shown in Fig. 6-1(*b*)

Common-Collector Amplifier

The *common-collector* (CC) or *emitter-follower* (EF) amplifier, with the universal bias circuitry of Fig. 6-2(*a*), can be modeled for small-signal ac analysis by replacing the CE-connected transistor with its h-parameter model, Fig. 6-1(*a*). Assuming, for simplicity, that $h_{re} = h_{oe} = 0$, we obtain the equivalent circuit of Fig. 6-2(*b*).

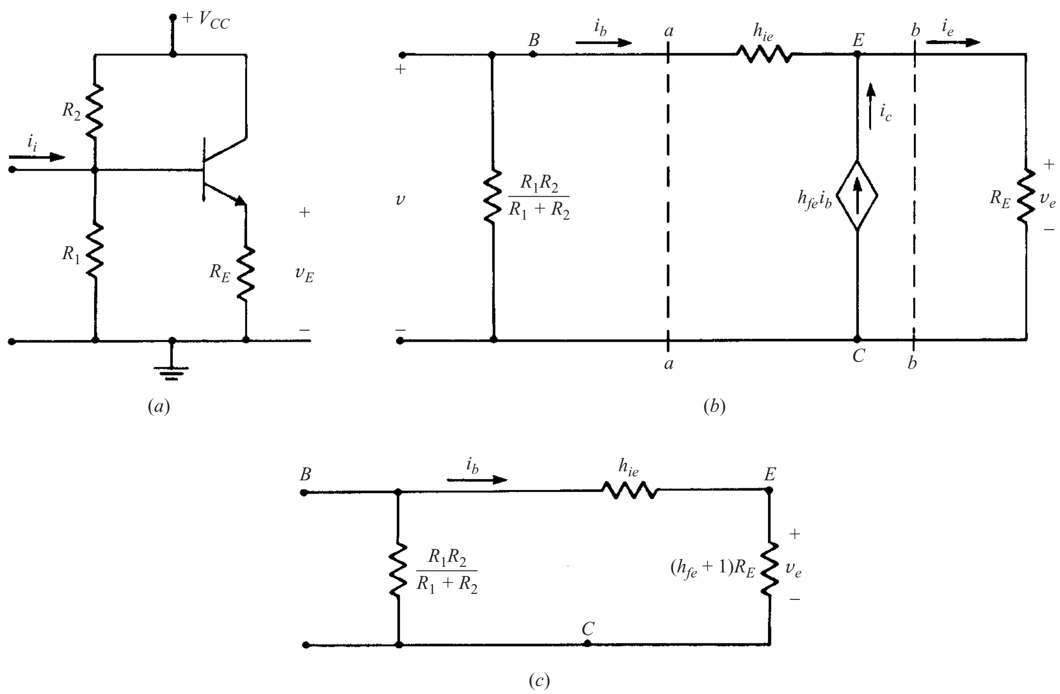


Fig. 6-2 CC amplifier

An even simpler model can be obtained by finding a Thévenin equivalent for the circuit to the right of a, a in Fig. 6-2(*b*). Application of KVL around the outer loop gives

$$v = i_b h_{ie} + i_e R_E + i_b h_{ie} + (h_{fe} + 1) i_b R_E \tag{6.15}$$

The Thévenin impedance is the driving-point impedance:

$$R_{Th} = \frac{v}{i_b} = h_{ie} + (h_{fe} + 1) R_E \tag{6.16}$$

The Thévenin voltage is zero (computed with terminals *a*, *a* open); thus, the equivalent circuit consists only of R_{Th}. This is shown, in a base-current frame of reference, in Fig. 6-2(*c*). (See Problem 6.13 for a development of the CC *h*-parameter model.)

6.3. TEE-EQUIVALENT CIRCUIT

The *tee-equivalent circuit* or *r-parameter model* is a circuit realization based on the *z* parameters of Chapter 1. Applying the *z*-parameter definitions of (*1.10*) to (*1.13*) to the CB small-signal equivalent circuit of Fig. 6-1(*b*) leads to

$$z_{11} = h_{ib} - \frac{h_{rb}h_{fb}}{h_{ob}} \tag{6.17}$$

$$z_{12} = \frac{h_{rb}}{h_{ob}} \tag{6.18}$$

$$z_{21} = -\frac{h_{fb}}{h_{ob}} \tag{6.19}$$

$$z_{22} = \frac{1}{h_{ob}} \tag{6.20}$$

(See Problem 6.17.) Substitution of these *z* parameters into (*1.8*) and (*1.9*) yields

$$v_{eb} = \left(h_{ib} - \frac{h_{rb}h_{fb}}{h_{ob}}\right)i_e + \frac{h_{rb}}{h_{ob}}(-i_c) \tag{6.21}$$

$$v_{cb} = -\frac{h_{fb}}{h_{ob}}\, i_e + \frac{1}{h_{ob}}(-i_c) \tag{6.22}$$

If we now define

$$r_b = \frac{h_{rb}}{h_{ob}} \tag{6.23}$$

$$r_e = h_{ib} - \frac{h_{rb}}{h_{ob}}(1 + h_{fb}) \tag{6.24}$$

$$r_c = \frac{1 - h_{rb}}{h_{ob}} \tag{6.25}$$

$$\alpha' = -\frac{h_{fb} + h_{rb}}{1 - h_{rb}} \tag{6.26}$$

then (*6.21*) and (*6.22*) can be written

$$v_{eb} = (r_e + r_b)i_e - r_b i_c \tag{6.27}$$

and

$$v_{cb} = (\alpha' r_c + r_b)i_e - (r_b + r_c)i_c \tag{6.28}$$

Typically, $-0.9 > h_{fb} > -1$ and $0 \le h_{rb} \ll 1$. Letting $h_{rb} \approx 0$ in (*6.26*), comparing (*6.13*) with (*3.1*) while neglecting thermally generated leakage currents, and assuming that $h_{FB} = h_{fb}$ (which is a valid assumption *except* near the boundary of active-region operation) result in

$$\alpha' \approx -h_{fb} = \alpha \tag{6.29}$$

Then the tee-equivalent circuit or *r*-parameter model for CB operation is that shown in Fig. 6-3. (See Problems 6.3 and 6.5 for *r*-parameter models for the CE and CC configurations, respectively.)

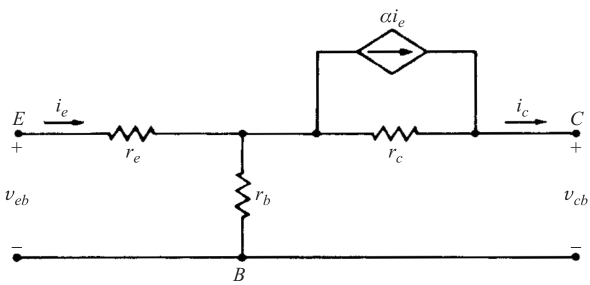


Fig. 6-3

6.4. CONVERSION OF PARAMETERS

Transistor manufacturers typically specify h_{FE} ($\approx h_{fe}$) and a set of input characteristics and collector characteristics for either CE or CB connection. Thus the necessity arises for conversion of h parameters among the CE, CB, and CC configurations or for calculation of r parameters from h parameters. Formulas can be developed to allow ready conversion from a known parameter set to a desired parameter set.

Example 6.1. Apply KVL and KCL to Fig. 6-1(*a*) to obtain $v_{eb} = g_1(i_e, v_{cb})$ and $i_c = g_2(i_e, v_{eb})$. Compare these equations with (*6.9*) and (*6.10*) to find the CB h parameters in terms of the CE h parameters. Use the typically reasonable approximations $h_{re} \ll 1$ and $h_{fe} + 1 \gg h_{ie}h_{oe}$ to simplify the computations and results.

KVL around the E, B loop of Fig. 6-1(*a*) (with assumed current directions reversed) yields

$$v_{eb} = -h_{ie}i_b - h_{re}v_{ce} \tag{6.30}$$

But KCL at node E requires that

$$i_b = -i_e - i_c = -i_e - h_{fe}i_b - h_{oe}v_{ce}$$

or

$$-i_b = \frac{1}{h_{fe}+1}\, i_e + \frac{h_{oe}}{h_{fe}+1}\, v_{ce} \tag{6.31}$$

In addition, KVL requires that

$$v_{ce} = v_{cb} - v_{eb} \tag{6.32}$$

Substituting (*6.31*) and (*6.32*) into (*6.30*) and rearranging give

$$\frac{(1-h_{re})(h_{fe}+1) + h_{ie}h_{oe}}{h_{fe}+1}\, v_{eb} = \frac{h_{ie}}{h_{fe}+1}\, i_e + \left(\frac{h_{ie}h_{oe}}{h_{fe}+1} - h_{re}\right) v_{cb} \tag{6.33}$$

Use of the given approximations reduces the coefficient of v_{eb} in (*6.33*) to unity, so that

$$v_{eb} \approx \frac{h_{ie}}{h_{fe}+1}\, i_e + \left(\frac{h_{ie}h_{oe}}{h_{fe}+1} - h_{re}\right) v_{cb} \tag{6.34}$$

Now KCL at node C of Fig. 6-1(*a*) (again with assumed current directions reversed) yields

$$i_c = h_{fe}i_b + h_{oe}v_{ce} \tag{6.35}$$

Substituting (*6.31*), (*6.32*), and (*6.34*) into (*6.35*) and solving for i_c give

$$i_c = -\left[\frac{h_{fe}}{h_{fe}+1} + \frac{h_{oe}h_{ie}}{(h_{fe}+1)^2}\right] i_e - h_{oe}\left[\frac{h_{ie}h_{oe}}{(h_{fe}+1)^2} - \frac{h_{re}+1}{h_{fe}+1}\right] v_{cb} \tag{6.36}$$

Use of the given approximations then leads to

$$i_c \approx -\frac{h_{fe}}{h_{fe}+1}\, i_e + \frac{h_{oe}}{h_{fe}+1}\, v_{cb} \tag{6.37}$$

Comparing (*6.34*) with (*6.9*) and (*6.37*) with (*6.10*), we see that

$$h_{ib} = \frac{h_{ie}}{h_{fe}+1} \tag{6.38}$$

$$h_{rb} = \frac{h_{ie}h_{oe}}{h_{fe}+1} - h_{re} \tag{6.39}$$

$$h_{fb} = -\frac{h_{fe}}{h_{fe}+1} \tag{6.40}$$

$$h_{ob} = \frac{h_{oe}}{h_{fe}+1} \tag{6.41}$$

6.5. MEASURES OF AMPLIFIER GOODNESS

Amplifiers are usually designed to emphasize one or more of the following interrelated performance characteristics, whose quantitative measures of goodness are defined in terms of the quantities of Fig. 6-4:

1. *Current amplification*, measured by the current-gain ratio $A_i = i_o/i_{\text{in}}$.
2. *Voltage amplification*, measured by the voltage-gain ratio $A_v = v_o/v_{\text{in}}$.
3. *Power amplification*, measured by the ratio $A_p = A_v A_i = v_o i_o / i_o i_{\text{in}}$.
4. *Phase shift of signals*, measured by the phase angle of the frequency-domain ratio $A_v(j\omega)$ or $A_i(j\omega)$.
5. *Impedance match or change*, measured by the input impedance Z_{in} (the driving-point impedance looking into the input port).
6. *Power transfer ability*, measured by the output impedance Z_o (the driving-point impedance looking into the output port with the load removed). If $Z_o = Z_L$, the maximum power transfer occurs.

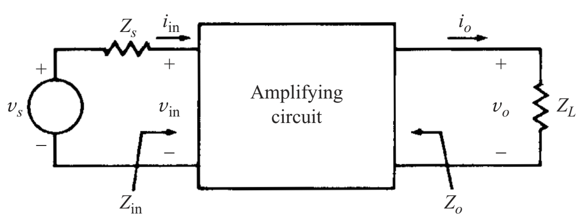


Fig. 6-4

6.6. CE AMPLIFIER ANALYSIS

A simplified (bias network omitted) CE amplifier is shown in Fig. 6-5(*a*), and the associated small-signal equivalent circuit in Fig. 6-5(*b*).

Example 6.2. In the CE amplifier of Fig. 6-5(*b*), let $h_{ie} = 1\,\text{k}\Omega$, $h_{re} = 10^{-4}$, $h_{fe} = 100$, $h_{oe} = 12\,\mu\text{S}$, and $R_L = 2\,\text{k}\Omega$. (These are typical CE amplifier values.) Find expressions for the (*a*) current-gain ratio A_i, (*b*) voltage-gain ratio A_v, (*c*) input impedance Z_{in}, and (*d*) output impedance Z_o. (*e*) Evaluate this typical CE amplifier.

(*a*) By current division at node *C*,

$$i_L = \frac{1/h_{oe}}{1/h_{oe} + R_L}(-h_{fe} i_b) \tag{6.42}$$

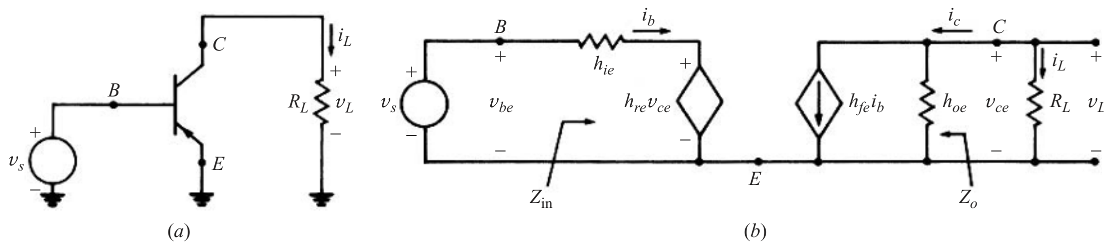


Fig. 6-5

and $$A_i = \frac{i_L}{i_b} = -\frac{h_{fe}}{1 + h_{oe}R_L} = -\frac{100}{1 + (12 \times 10^{-6})(2 \times 10^3)} = -97.7 \qquad (6.43)$$

Note that $A_i \approx -h_{fe}$, where the minus sign indicates a 180° phase shift between input and output currents.

(*b*) By KVL around B, E mesh,

$$v_s = v_{be} = h_{ie}i_b + h_{re}v_{ce} \qquad (6.44)$$

Ohm's law applied to the output network requires that

$$v_{ce} = -h_{fe}i_b\left(\frac{1}{h_{oe}} \| R_L\right) = \frac{-h_{fe}R_L i_b}{1 + h_{oe}R_L} \qquad (6.45)$$

Solving (*6.45*) for i_b, substituting the result into (*6.44*), and rearranging yield

$$A_v = \frac{v_s}{v_{ce}} = -\frac{h_{fe}R_L}{h_{ie} + R_L(h_{ie}h_{oe} - h_{fe}h_{re})}$$
$$= -\frac{(100)(2 \times 10^3)}{1 \times 10^3 + (2 \times 10^3)[(1 \times 10^3)(12 \times 10^{-6}) - (100)(1 \times 10^{-4})]} = -199.2 \qquad (6.46)$$

Observe that $A_v \approx -h_{fe}R_L/h_{ie}$, where the minus sign indicates a 180° phase shift between input and output voltages.

(*c*) Substituting (*6.45*) into (*6.44*) and rearranging yield

$$Z_{in} = \frac{v_s}{i_b} = h_{ie} - \frac{h_{re}h_{fe}R_L}{1 + h_{oe}R_L} = 1 \times 10^3 - \frac{(1 \times 10^{-4})(100)(2 \times 10^3)}{1 + (12 \times 10^{-6})(2 \times 10^3)} = 980.5\,\Omega \qquad (6.47)$$

Note that for typical CE amplifier values, $Z_{in} \approx h_{ie}$.

(*d*) We deactivate (short) v_s and replace R_L with a driving-point source so that $v_{dp} = v_{ce}$. Then, for the input mesh, Ohm's law requires that

$$i_b = -\frac{h_{re}}{h_{ie}}\, v_{dp} \qquad (6.48)$$

However, at node C (with, now, $i_c = i_{dp}$), KCL yields

$$i_c = i_{dp} = h_{fe}i_b + h_{oe}v_{dp} \qquad (6.49)$$

Using (*6.48*) in (*6.49*) and rearranging then yield

$$Z_o = \frac{v_{dp}}{i_{dp}} = \frac{1}{h_{oe} - h_{fe}h_{re}/h_{ie}} = \frac{1}{12 \times 10^{-6} - (100)(1 \times 10^{-4})/(1 \times 10^3)} = 500\,\text{k}\Omega \qquad (6.50)$$

The output impedance is increased by feedback due to the presence of the controlled source $h_{re}v_{ce}$.

(*e*) Based on the typical values of this example, the characteristics of the CE amplifier can be summarized as follows:

1. Large current gain
2. Large voltage gain
3. Large power gain (A_iA_v)
4. Current and voltage phase shifts of 180°
5. Moderate input impedance
6. Moderate output impedance

6.7. CB AMPLIFIER ANALYSIS

A simplified (bias network omitted) CB amplifier is shown in Fig. 6-6(*a*), and the associated small-signal equivalent circuit in Fig. 6-6(*b*).

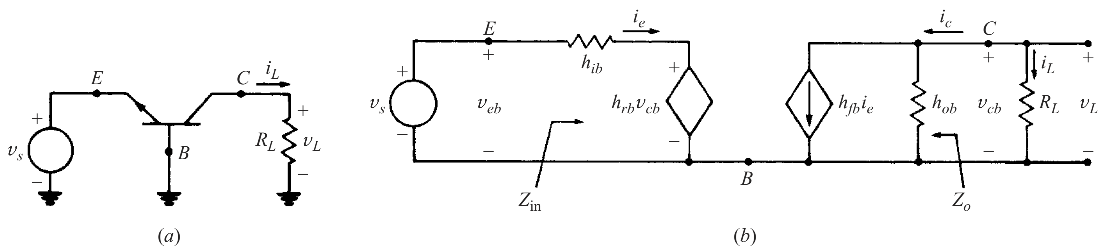


Fig. 6-6 CB amplifier

Example 6.3. In the CB amplifier of Fig. 6-6(*b*), let $h_{ib} = 30\,\Omega$, $h_{rb} = 4 \times 10^{-6}$, $h_{fb} = -0.99$, $h_{ob} = 8 \times 10^{-7}$ S, and $R_L = 20\,\text{k}\Omega$. (These are typical CB amplifier values.) Find expressions for the (*a*) current-gain ratio A_i, (*b*) voltage-gain ratio A_v, (*c*) input impedance Z_{in}, and (*d*) output impedance Z_o. (*e*) Evaluate this typical CE amplifier.

(*a*) By direct analogy with Fig. 6-5(*b*) and (*6.43*)

$$A_i = -\frac{h_{fb}}{1 + h_{ob}R_L} = -\frac{-0.99}{1 + (8 \times 10^{-7})(20 \times 10^3)} = 0.974 \tag{6.51}$$

Note that $A_i \approx -h_{fb} < 1$, and that the input and output currents are in phase because $h_{fb} < 0$.

(*b*) By direct analogy with Fig. 6-5(*b*) and (*6.46*),

$$A_v = -\frac{h_{fb}R_L}{h_{ib} + R_L(h_{ib}h_{oc} - h_{fb}h_{rb})} = -\frac{(-0.99)(20 \times 10^3)}{30 + (20 \times 10^3)[(30)(8 \times 10^{-7}) - (-0.99)(4 \times 10^{-6})]} = 647.9 \tag{6.52}$$

Observe that $A_v \approx -h_{fb}R_L/h_{ib}$, and the output and input voltages are in phase because $h_{fb} < 0$.

(*c*) By direct analogy with Fig. 6-5(*b*) and (*6.47*)

$$Z_{in} = h_{ib} - \frac{h_{rb}h_{fb}R_L}{1 + h_{ob}R_L} = 30 - \frac{(4 \times 10^{-6})(-0.99)(20 \times 10^3)}{1 + (8 \times 10^{-7})(20 \times 10^3)} = 30.08\,\Omega \tag{6.53}$$

It is apparent that $Z_{in} \approx h_{ib}$.

(*d*) By analogy with Fig. 6-5(*b*) and (*6.50*),

$$Z_o = \frac{1}{h_{ob} - h_{fb}h_{rb}/h_{ib}} = \frac{1}{8 \times 10^{-7} - (-0.99)(4 \times 10^{-6})/30} = 1.07\,\text{M}\Omega \tag{6.54}$$

Note that Z_o is decreased because of the feedback from the output mesh to the input mesh through $h_{rb}v_{cb}$.

(*e*) Based on the typical values of this example, the characteristics of the CB amplifier can be summarized as follows:

1. Current gain of less than 1
2. High voltage gain
3. Power gain approximately equal to voltage gain
4. No phase shift for current or voltage
5. Small input impedance
6. Large output impedance

6.8. CC AMPLIFIER ANALYSIS

Figure 6-7(*a*) shows a CC amplifier with the bias network omitted. The small-signal equivalent circuit is drawn in Fig. 6-7(*b*).

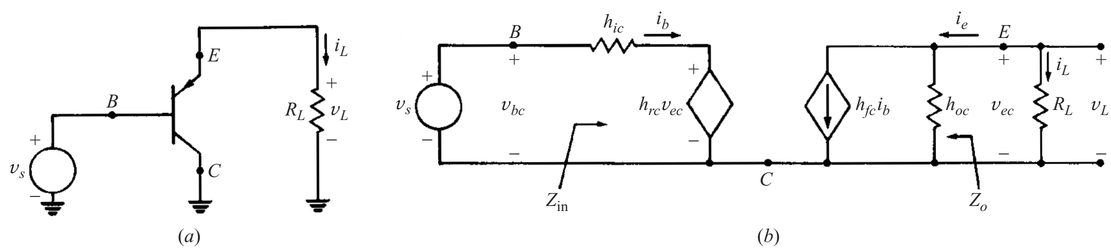


Fig. 6-7 CC amplifier

Example 6.4. In the CC amplifier of Fig. 6-7(*b*), let $h_{ic} = 1\,\text{k}\Omega$, $h_{rc} = 1$, $h_{fc} = -101$, $h_{oc} = 12\,\mu\text{S}$, and $R_L = 2\,\text{k}\Omega$. Drawing direct analogies with the CE amplifier of Example 6.2, find expressions for the (*a*) current-gain ratio A_i, (*b*) voltage-gain ratio A_v, (*c*) input impedance Z_{in}, and (*d*) output impedance Z_o. (*e*) Evaluate this typical CC amplifier.

(*a*) In parallel with (*6.43*),

$$A_i = \frac{h_{fc}}{1 + h_{oc}R_L} = -\frac{-101}{1 + (12 \times 10^{-6})(2 \times 10^3)} = 98.6 \qquad (6.55)$$

Note that $A_i \approx -h_{fc}$, and that the input and output currents are in phase because $h_{fc} < 0$.

(*b*) In parallel with (*6.46*),

$$A_v = -\frac{h_{fc}R_L}{h_{ic} + R_L(h_{ic}h_{oc} - h_{fc}h_{rc})} = -\frac{(-101)(2 \times 10^3)}{1 \times 10^3 + (2 \times 10^3)[(1 \times 10^3)(12 \times 10^{-6}) - (-101)(1)]} = 0.995 \qquad (6.56)$$

Observe that $A_v \approx 1/(1 - h_{ic}h_{oc}/h_{fc}) \approx 1$. Since the gain is approximately 1 and the output voltage is in phase with the input voltage, this amplifier is commonly called a *unity follower*.

(*c*) In parallel with (*6.47*),

$$Z_{\text{in}} = h_{ic} - \frac{h_{rc}h_{fc}R_L}{1 + h_{oc}R_L} = 1 \times 10^3 - \frac{(1)(-101)(2 \times 10^3)}{1 + (12 \times 10^{-6})(2 \times 10^3)} = 8.41\,\text{M}\Omega \qquad (6.57)$$

Note that $Z_{\text{in}} \approx -h_{fc}/h_{oc}$.

(*d*) In parallel with (*6.50*),

$$Z_o = \frac{1}{h_{oc} - h_{fc}h_{rc}/h_{ic}} = \frac{1}{12 \times 10^{-6} - (-101)(1)/(1 \times 10^3)} = 9.9\,\Omega$$

Note that $Z_o \approx -h_{ic}/h_{fc}$.

(*e*) Based on the typical values of this example, the characteristics of the CB amplifier can be summarized as follows:

1. High current gain
2. Voltage gain of approximately unity
3. Power gain approximately equal to current gain
4. No current or voltage phase shift
5. Large input impedance
6. Small output impedance

6.9. BJT AMPLIFIER ANALYSIS WITH SPICE

Since SPICE models of the BJT (see Chapter 3) provide the device terminal characteristics, a transistor amplifier can be properly biased and a time-varying input signal can be directly applied to the completely modeled amplifier circuit. Any desired signal that results can be measured directly in the time domain to form signal ratios that yield current and voltage gains. With such modeling, any signal distortion that results from nonlinear operation of the BJT is readily apparent from inspection of signal-time plots. Such an analysis approach is the analytical equivalent of laboratory operation of the amplifier where the time plot of signals is analogous to oscilloscope observation of the amplifier circuit signals.

SPICE capabilities also lend themselves to BJT amplifier analysis using the small-signal equivalent circuits. In such case, the voltage-controlled voltage source (VCVS) and the current-controlled current source (CCCS) introduced in Section 1.3 find obvious application in the small-signal equivalent circuits of the type shown in Fig. 6-1. Either time-varying analysis (.TRAN command statement) or sinusoidal steady-state analysis (.AC command statement) can be performed on the small-signal equivalent circuit.

Example 6.5. For the amplifier of Fig. 3-10(*a*), let $v_i = 0.25\sin(2000\pi t)\,\text{V}$, $V_{CC} = 15\,\text{V}$, $CC_1 = CC_2 = CC = 100\,\mu\text{F}$, $R_1 = 6\,\text{k}\Omega$, $R_2 = 50\,\text{k}\Omega$, $R_C = R_L = 1\,\text{k}\Omega$, and $R_i = R_E = 100\,\Omega$. The transistor is characterized by the model of Problem 5.4. Use SPICE methods to determine the CE hybrid parameters of (6.5) through (6.8) for this transistor at the point of operation.

The netlist code below describes the circuit.

```
EX6_5.CIR
vi  1 0 SIN(0V 250mV 10kHz)
Ri  1 2 100ohm
CC1 2 3 1000uF
CC2 4 7 1000uF
R1  3 0 6kohm
R2  3 6 50kohm
RC  6 4 1kohm
RE  5 0 100ohm
RL  7 0 1kohm
VCC 6 0 15V
Q 4  3 5 QNPNG
.MODEL QNPNG NPN(Is=10fA Ikf=150mA Isc=10fA Bf=150
+ Br=3 Rb=1ohm Rc=1ohm Va=75V Cjc=10pF Cje=15pF)
.TRAN 1us 0.1ms
.PROBE
.END
```

After executing ⟨Ex6_5.CIR⟩, the plots of Fig. 6-8 can be generated by use of the Probe feature of PSpice. The resulting *h*-parameter value is indicated on each of the four plots of Fig. 6-8.