Tikrit university

Collage of Engineering Shirqat

Department of Electrical Engineering

Second Class

Electronic II

Chapter 5

Lec2

BJT AC Analysis

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7. Ce Emitter-Bias Configuration

The networks examined in this section include an emitter resistor that may or may not be bypassed in the ac domain. We first consider the unbypassed situation and then modify the resulting equations for the bypassed configuration.

7.1 Unbypassed

The most fundamental of unbypassed configurations appears in Fig. 29. The *re* equivalent model is substituted in Fig. 30,

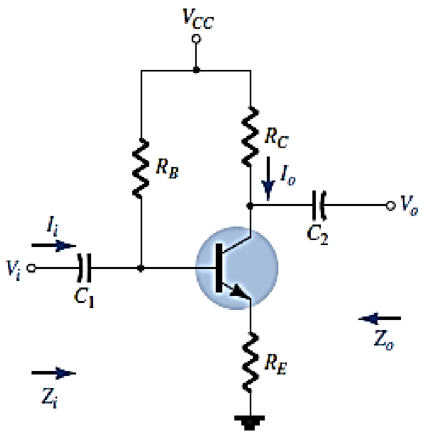

FIG. 29
CE emitter-bias configuration.

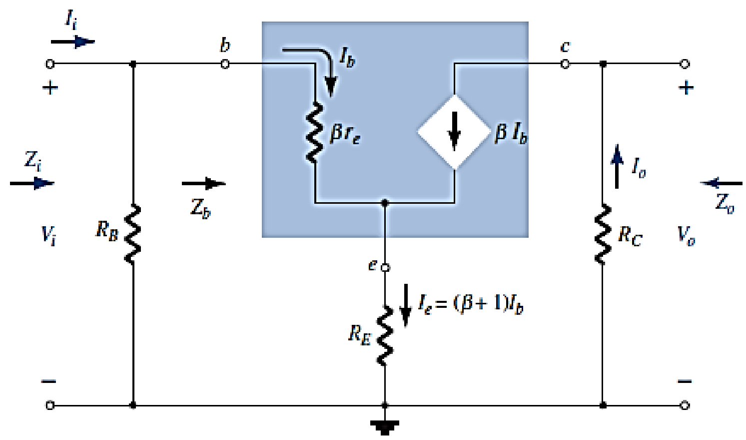

FIG. 30
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 29.

Applying Kirchhoff's voltage law to the input side of Fig. 30 results in

$$V_i = I_b\beta r_e + I_e R_E$$

or

$$V_i = I_b\beta r_e + (\beta + I)I_b R_E$$

and the input impedance looking into the network to the right of R_B is

$$Z_b = \frac{V_i}{I_b} = \beta r_e + (\beta + 1)R_E$$

The result as displayed in Fig. 31 reveals that the input impedance of a transistor with an unbypassed resistor R_E is determined by

$$\boxed{Z_b = \beta r_e + (\beta + 1)R_E} \qquad (17)$$

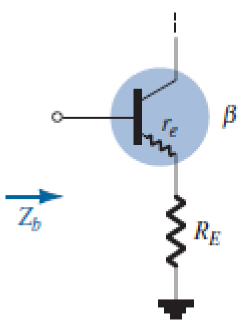

FIG. 31
Defining the input impedance of a transistor with an unbypassed emitter resistor.

Because β is normally much greater than 1, the approximate equation is

$$Z_b \cong \beta r_e + \beta R_E$$

and

$$\boxed{Z_b \cong \beta(r_e + R_E)} \tag{18}$$

Because R_E is usually greater than r_e, Eq. (18) can be further reduced to

$$\boxed{Z_b \cong \beta R_E} \tag{19}$$

Z_i Returning to Fig. 30, we have

$$\boxed{Z_i = R_B \| Z_b} \tag{20}$$

Z_o With V_i set to zero, $I_b = 0$, and βI_b can be replaced by an open-circuit equivalent. The result is

$$\boxed{Z_o = R_C} \tag{21}$$

A_v

$$I_b = \frac{V_i}{Z_b}$$

and

$$V_o = -I_o R_C = -\beta I_b R_C$$
$$= -\beta\left(\frac{V_i}{Z_b}\right)R_C$$

with

$$\boxed{A_v = \frac{V_o}{V_i} = -\frac{\beta R_C}{Z_b}} \tag{22}$$

Substituting $Z_b \cong \beta(r_e + R_E)$ gives

$$\boxed{A_v = \frac{V_o}{V_i} \cong -\frac{R_C}{r_e + R_E}} \tag{23}$$

and for the approximation $Z_b \cong \beta R_E$,

$$\boxed{A_v = \frac{V_o}{V_i} \cong -\frac{R_C}{R_E}} \tag{24}$$

Note the absence of β from the equation for A_v demonstrating an independence in variation of β.

Phase Relationship The negative sign in Eq. (22) again reveals a 180° phase shift between V_o and V_i.

$$\boxed{Z_b \cong \beta(r_e + R_E)} \quad r_o \geq 10(R_C + R_E) \tag{26}$$

$$\boxed{Z_o \cong R_C} \quad \text{Any level of } r_o \tag{28}$$

$$\boxed{A_v = \frac{V_o}{V_i} \cong -\frac{\beta R_C}{Z_b}} \quad r_o \geq 10R_C \tag{30}$$

Bypassed

If *RE* of Fig. 29 is bypassed by an emitter capacitor *CE*, the complete *re* equivalent model can be substituted, resulting in the same equivalent network as Fig. 22. Equations (5) to (10) are therefore applicable

EXAMPLE 3 For the network of Fig. 32, without C_E (unbypassed), determine:

a. r_e.
b. Z_i.
c. Z_o.
d. A_v.

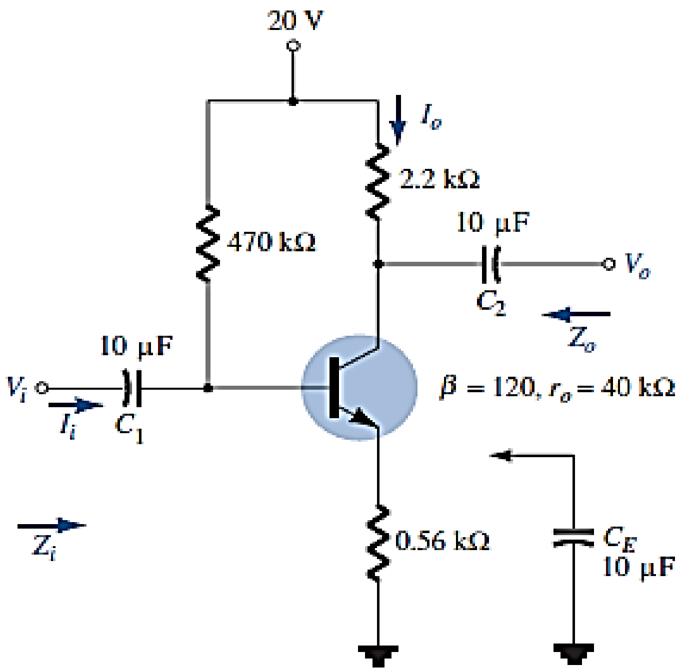

FIG. 32
Example 3.

Solution:

a. DC:

$$I_B = \frac{V_{CC} - V_{BE}}{R_B + (\beta + 1)R_E} = \frac{20\text{ V} - 0.7\text{ V}}{470\text{ k}\Omega + (121)0.56\text{ k}\Omega} = 35.89\ \mu\text{A}$$

$$I_E = (\beta + 1)I_B = (121)(35.89\ \mu\text{A}) = 4.34\text{ mA}$$

and $$r_e = \frac{26\text{ mV}}{I_E} = \frac{26\text{ mV}}{4.34\text{ mA}} = 5.99\ \Omega$$

b. Testing the condition $r_o \geq 10(R_C + R_E)$, we obtain

$$40\ \text{k}\Omega \geq 10(2.2\ \text{k}\Omega + 0.56\ \text{k}\Omega)$$
$$40\ \text{k}\Omega \geq 10(2.76\ \text{k}\Omega) = 27.6\ \text{k}\Omega\ (\textit{satisfied})$$

Therefore,

$$Z_b \cong \beta(r_e + R_E) = 120(5.99\ \Omega + 560\ \Omega) = 67.92\ \text{k}\Omega$$

and

$$Z_i = R_B \| Z_b = 470\ \text{k}\Omega \| 67.92\ \text{k}\Omega = 59.34\ \text{k}\Omega$$

c. $Z_o = R_C = 2.2\ \text{k}\Omega$

d. $r_o \geq 10R_C$ is satisfied. Therefore,

$$A_v = \frac{V_o}{V_i} \cong -\frac{\beta R_C}{Z_b} = -\frac{(120)(2.2\ \text{k}\Omega)}{67.92\ \text{k}\Omega} = -3.89$$

compared to -3.93 using Eq. (20): $A_v \cong -R_C/R_E$.

EXAMPLE 4 Repeat the analysis of Example 3 with C_E in place.

Solution:

a. The dc analysis is the same, and $r_e = 5.99\ \Omega$.

b. R_E is "shorted out" by C_E for the ac analysis. Therefore,

$$Z_i = R_B \| Z_b = R_B \| \beta r_e = 470\ \text{k}\Omega \| (120)(5.99\ \Omega) = 470\ \text{k}\Omega \| 718.8\ \Omega \cong 717.70\ \Omega$$

c. $Z_o = R_C = 2.2\ \text{k}\Omega$

d. $A_v = -\dfrac{R_C}{r_e}$

$$= -\frac{2.2\ \text{k}\Omega}{5.99\ \Omega} = -367.28\ \text{(a significant increase)}$$

8. Emitter-Follower Configuration

When the output is taken from the emitter terminal of the transistor as shown in Fig. 36, the network is referred to as an *emitter-follower*. The output voltage is always slightly less than the input signal due to the drop from base to emitter, but the approximation $AV \approx 1$ is usually a good one. Unlike the collector voltage, the emitter voltage is in phase with the signal Vi. That is, both Vo and Vi attain their positive and negative peak values at the same time. The fact that *Vo* “follows” the magnitude of *Vi* with an in-phase relationship accounts for the terminology emitter-follower.

The most common emitter-follower configuration appears in Fig. 36. In fact, because the collector is grounded for ac analysis, it is actually a common-collector configuration. Other variations of Fig. 36 that draw the output off the emitter with Vo ≈ Vi will appear later in this section. The emitter-follower configuration is frequently used for impedance-matching purposes. It presents a high impedance at the input and a low impedance at the output, which

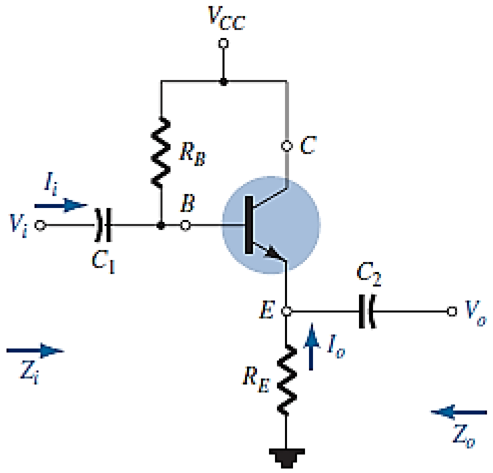

FIG. 36
Emitter-follower configuration.

is the direct opposite of the standard fixed-bias configuration. The resulting effect is much the same as that obtained with a transformer, where a load is matched to the source impedance for maximum power transfer through the system.

Substituting the *re* equivalent circuit into the network of Fig. 36 results in the network of Fig. 37.

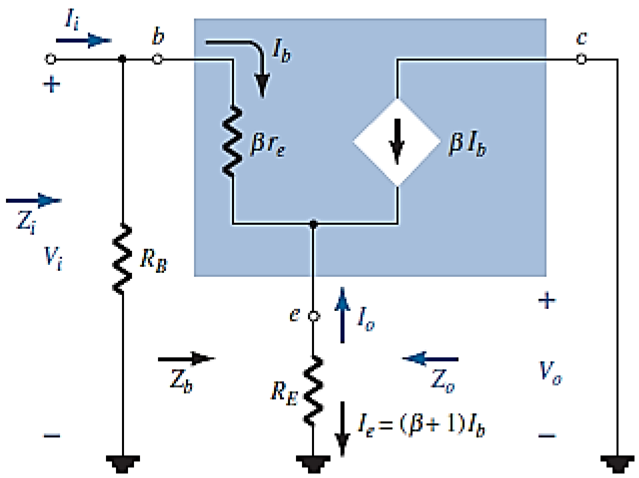

FIG. 37
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 36.

Zi The input impedance is determined in the same manner as described in the preceding section:

$$\boxed{Z_i = R_B \| Z_b} \tag{31}$$

$$\boxed{Z_b \cong \beta(r_e + R_E)} \tag{33}$$

Zo The output impedance is best described by first writing the equation for the current *Ib*, $I_b = \dfrac{V_i}{Z_b}$ and then multiplying by (β + 1) to establish Ie. That is,

$$\boxed{I_e \cong \frac{V_i}{r_e + R_E}} \tag{35}$$

$$\boxed{Z_o = R_E \| r_e} \quad (36)$$

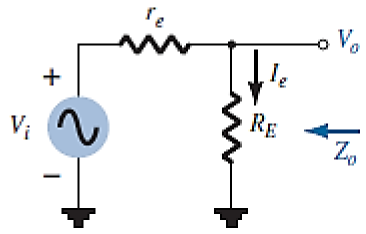

FIG. 38
Defining the output impedance for the emitter-follower configuration.

Av Figure 38 can be used to determine the voltage gain

$$V_o = \frac{R_E V_i}{R_E + r_e}$$

ication of the voltage-divider rule:

and

$$\boxed{A_v = \frac{V_o}{V_i} = \frac{R_E}{R_E + r_e}} \quad (38)$$

Because R_E is usually much greater than r_e, $R_E + r_e \cong R_E$ and

$$\boxed{A_v = \frac{V_o}{V_i} \cong 1} \quad (39)$$

phase relationship As revealed by Eq. (38) and earlier discussions of this section, *Vo* and *Vi* are in phase for the emitter-follower configuration.

EXAMPLE 7 For the emitter-follower network of Fig. 39, determine:

a. r_e.
b. Z_i.
c. Z_o.
d. A_v.
e. Repeat parts (b) through (d) with $r_o = 25\ \text{k}\Omega$ and compare results.

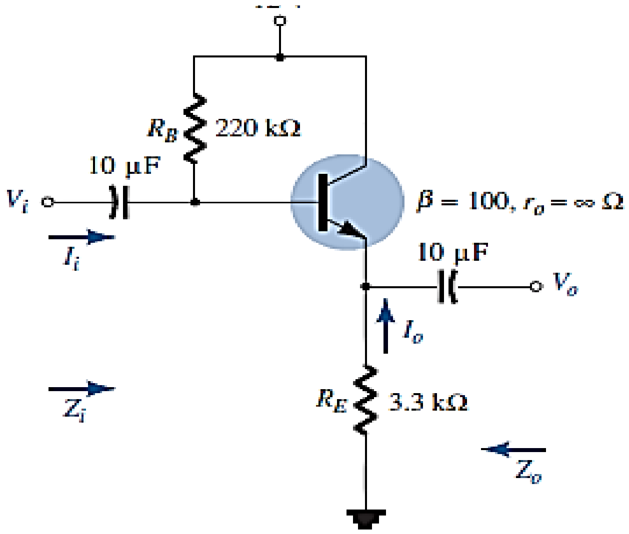

FIG. 39
Example 7.

Solution:

a. $$I_B = \frac{V_{CC} - V_{BE}}{R_B + (\beta + 1)R_E}$$
$$= \frac{12\text{ V} - 0.7\text{ V}}{220\text{ k}\Omega + (101)3.3\text{ k}\Omega} = 20.42\ \mu\text{A}$$
$$I_E = (\beta + 1)I_B$$
$$= (101)(20.42\ \mu\text{A}) = 2.062\text{ mA}$$
$$r_e = \frac{26\text{ mV}}{I_E} = \frac{26\text{ mV}}{2.062\text{ mA}} = \mathbf{12.61\ \Omega}$$

b. $$Z_b = \beta r_e + (\beta + 1)R_E$$
$$= (100)(12.61\ \Omega) + (101)(3.3\text{ k}\Omega)$$
$$= 1.261\text{ k}\Omega + 333.3\text{ k}\Omega$$
$$= 334.56\text{ k}\Omega \cong \beta R_E$$
$$Z_i = R_B \| Z_b = 220\text{ k}\Omega \| 334.56\text{ k}\Omega$$
$$= \mathbf{132.72\text{ k}\Omega}$$

c. $$Z_o = R_E \| r_e = 3.3\text{ k}\Omega \| 12.61\ \Omega$$
$$= \mathbf{12.56\ \Omega} \cong r_e$$

d. $$A_v = \frac{V_o}{V_i} = \frac{R_E}{R_E + r_e} = \frac{3.3\text{ k}\Omega}{3.3\text{ k}\Omega + 12.61\ \Omega}$$
$$= \mathbf{0.996 \cong 1}$$

9 Common-Base Configuration

The common-base configuration is characterized as having a relatively low input and a high output impedance and a current gain less than 1. The voltage gain, however, can be quite large. The standard configuration appears in Fig. 42, with the common-base re equivalent model substituted in Fig. 43.

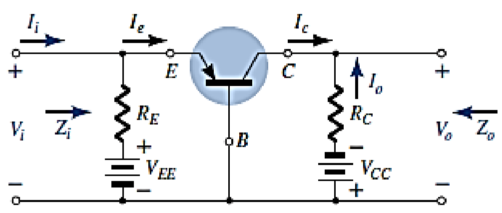

FIG. 42
Common-base configuration.

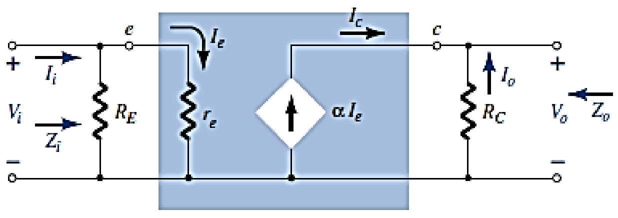

FIG. 43
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 44.

$$\boxed{Z_i = R_E \| r_e} \qquad (46)$$

$$\boxed{Z_o = R_C} \qquad (47)$$

A_v

$$V_o = -I_o R_C = -(-I_c)R_C = \alpha I_e R_C$$

with
$$I_e = \frac{V_i}{r_e}$$

so that
$$V_o = \alpha\left(\frac{V_i}{r_e}\right)R_C$$

and
$$\boxed{A_v = \frac{V_o}{V_i} = \frac{\alpha R_C}{r_e} \cong \frac{R_C}{r_e}} \qquad (48)$$

A_i Assuming that $R_E \gg r_e$ yields

$$I_e = I_i$$

and
$$I_o = -\alpha I_e = -\alpha I_i$$

with
$$\boxed{A_i = \frac{I_o}{I_i} = -\alpha \cong -1} \qquad (49)$$

phase relationship The fact that *AV* is a positive number shows that *Vo* and *Vi* are in phase for the common-base configuration.

EXAMPLE 8 For the network of Fig. 44, determine:

a. r_e.
b. Z_i.
c. Z_o.
d. A_v.
e. A_i.

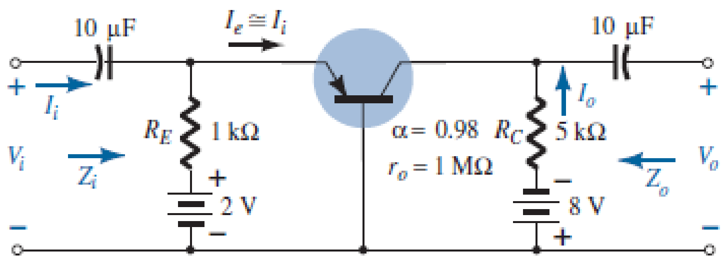

FIG. 44
Example 8.

Solution:

a. $I_E = \dfrac{V_{EE} - V_{BE}}{R_E} = \dfrac{2\text{ V} - 0.7\text{ V}}{1\text{ k}\Omega} = \dfrac{1.3\text{ V}}{1\text{ k}\Omega} = 1.3\text{ mA}$

$r_e = \dfrac{26\text{ mV}}{I_E} = \dfrac{26\text{ mV}}{1.3\text{ mA}} = \mathbf{20\ \Omega}$

b. $Z_i = R_E \| r_e = 1\text{ k}\Omega \| 20\ \Omega$
$= \mathbf{19.61\ \Omega} \cong r_e$

c. $Z_o = R_C = \mathbf{5\text{ k}\Omega}$

d. $A_v \cong \dfrac{R_C}{r_e} = \dfrac{5\text{ k}\Omega}{20\ \Omega} = \mathbf{250}$

e. $A_i = \mathbf{-0.98} \cong -1$

10 Collector Feedback Configuration

The collector feedback network of Fig. 45 employs a feedback path from collector to base to increase the stability. Substituting the equivalent circuit and redrawing the network results in the configuration of Fig. 46.

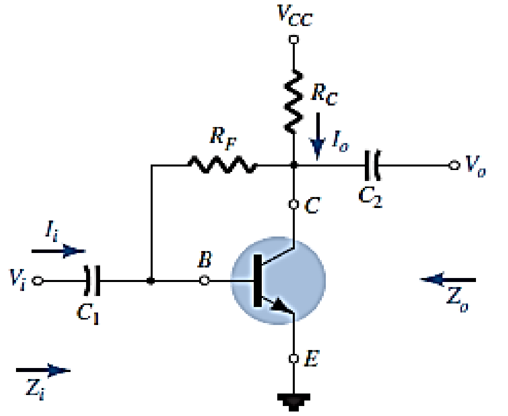

FIG. 45
Collector feedback configuration.

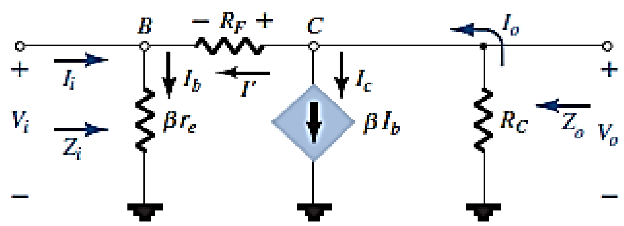

FIG. 46
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 45.

Z_i

$$I_o = I' + \beta I_b$$

and $$I' = \frac{V_o - V_i}{R_F}$$

but $$V_o = -I_o R_C = -(I' + \beta I_b)R_C$$

with $$V_i = I_b \beta r_e$$

so that $$I' = -\frac{(I' + \beta I_b)R_C - I_b\beta r_e}{R_F} = -\frac{I'R_C}{R_F} - \frac{\beta I_b R_C}{R_F} - \frac{I_b \beta r_e}{R_F}$$

which when rearranged in the following:

$$I'\left(1 + \frac{R_C}{R_F}\right) = -\beta I_b \frac{(R_C + r_e)}{R_F}$$

and finally,
$$I' = -\beta I_b \frac{(R_C + r_e)}{R_C + R_F}$$

$$\boxed{Z_i = \frac{r_e}{\dfrac{1}{\beta} + \dfrac{R_C}{R_C + R_F}}} \qquad (50)$$

Zo If we set Vi to zero as required to define Zo, the network will appear as shown in Fig. 47. The effect of *βre* is removed, and RF appears in parallel with RC and.

$$\boxed{Z_o \cong R_C \| R_F} \qquad (51)$$

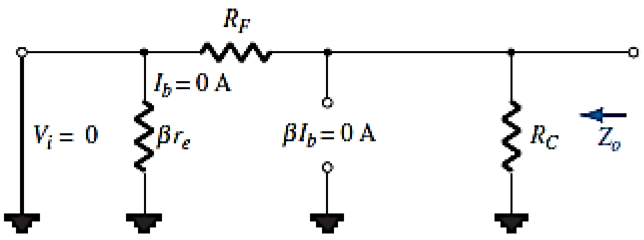

FIG. 47
Defining Z_o for the collector feedback configuration.

and
$$\boxed{A_v = -\left(\frac{R_F}{R_C + R_F}\right)\frac{R_C}{r_e}} \qquad (52)$$

For $R_F \gg R_C$
$$\boxed{A_v \cong -\frac{R_C}{r_e}} \qquad (53)$$

phase relationship The negative sign of Eq. (52) indicates a 180° phase shift between *Vo* and *Vi*.

EXAMPLE 9 For the network of Fig. 48. determine:

a. r_e.
b. Z_i.
c. Z_o.
d. A_v.

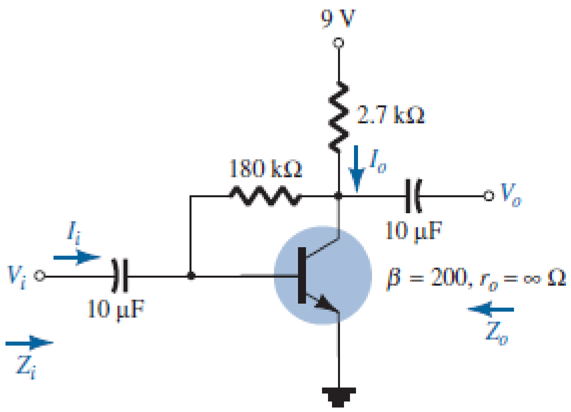

FIG. 48
Example 9.

Solution:

a. $I_B = \dfrac{V_{CC} - V_{BE}}{R_F + \beta R_C} = \dfrac{9\text{ V} - 0.7\text{ V}}{180\text{ k}\Omega + (200)2.7\text{ k}\Omega}$

$= 11.53\ \mu\text{A}$

$I_E = (\beta + 1)I_B = (201)(11.53\ \mu\text{A}) = 2.32\text{ mA}$

$r_e = \dfrac{26\text{ mV}}{I_E} = \dfrac{26\text{ mV}}{2.32\text{ mA}} = \mathbf{11.21\ \Omega}$

b. $Z_i = \dfrac{r_e}{\dfrac{1}{\beta} + \dfrac{R_C}{R_C + R_F}} = \dfrac{11.21\ \Omega}{\dfrac{1}{200} + \dfrac{2.7\text{ k}\Omega}{182.7\text{ k}\Omega}} = \dfrac{11.21\ \Omega}{0.005 + 0.0148}$

$= \dfrac{11.21\ \Omega}{0.0198} = \mathbf{566.16\ \Omega}$

c. $Z_o = R_C \| R_F = 2.7\text{ k}\Omega \| 180\text{ k}\Omega = \mathbf{2.66\text{ k}\Omega}$

d. $A_v = -\dfrac{R_C}{r_e} = -\dfrac{2.7\text{ k}\Omega}{11.21\ \Omega} = \mathbf{-240.86}$

For the configuration of Fig. 49, Eqs. (61) through (63) determine the variables of interest.

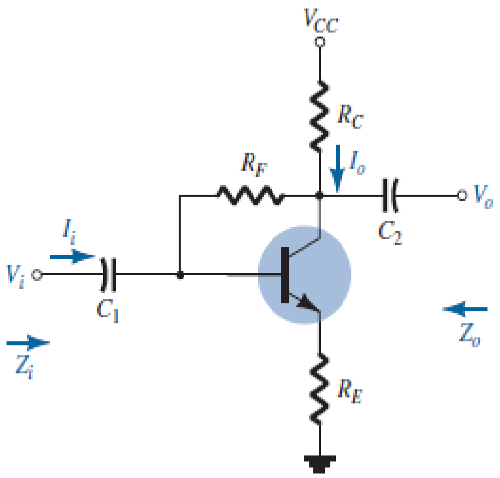

FIG. 49
Collector feedback configuration with an emitter resistor R_E.

$$\boxed{Z_i \cong \frac{R_E}{\left[\dfrac{1}{\beta} + \dfrac{(R_E + R_C)}{R_F}\right]}} \quad (62)$$

$$\boxed{Z_o = R_C \| R_F} \quad (63)$$

$$\boxed{A_v \cong -\frac{R_C}{R_E}} \quad (64)$$

12 Effect of RL And RS

All the parameters determined in the last few sections have been for an unloaded amplifier with the input voltage connected directly to a terminal of the transistor. In this section the effect of applying a load to the output terminal and the effect of using a source with an internal resistance will be investigated. The network of Fig. 54a is typical of those investigated in the previous section.

Because a resistive load was not attached to the output terminal, the gain is commonly referred to as the no-load gain and given the following notation:

$$\boxed{A_{v_{NL}} = \frac{V_o}{V_i}} \qquad (70)$$

In Fig. 54b a load has been added in the form of a resistor RL, which will change the overall gain of the system. This loaded gain is typically given the following notation:

$$\boxed{A_{v_L} = \frac{V_o}{V_i}} \text{ with } R_L \qquad (71)$$

In Fig. 54c both a load and a source resistance have been introduced, which will have an additional effect on the gain of the system. The resulting gain is typically given the following notation:

$$\boxed{A_{v_s} = \frac{V_o}{V_s}} \text{ with } R_L \text{ and } R_s \qquad (72)$$

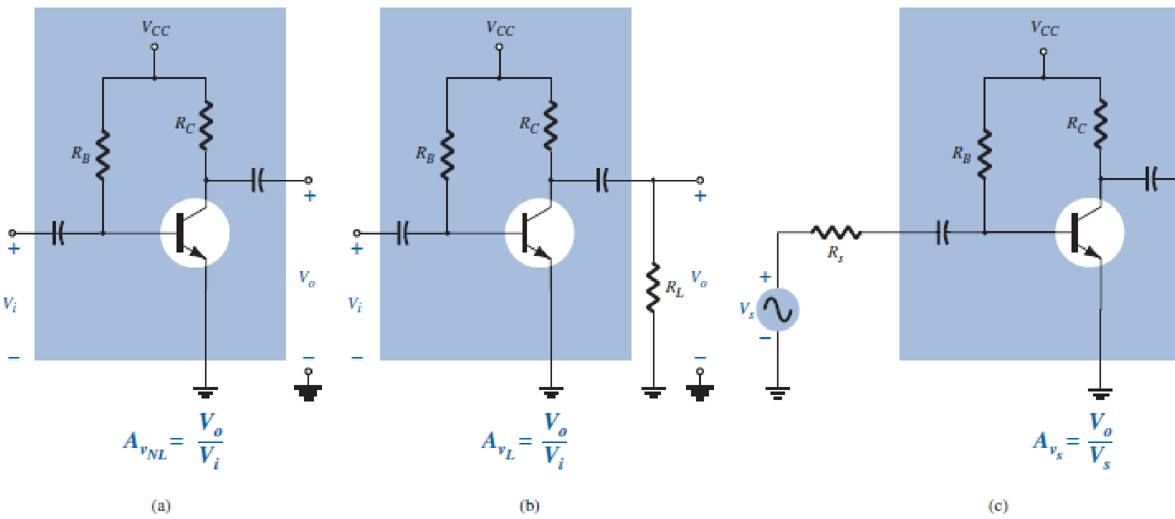

FIG. 54
Amplifier configurations: (a) unloaded; (b) loaded; (c) loaded with a source resistance.

The analysis to follow will show that:

- ***The loaded voltage gain of an amplifier is always less than the no-load gain.***
- ***The gain obtained with a source resistance in place will always be less than that obtained under loaded or unloaded conditions due to the drop in applied voltage across the source resistance.***

- ***For the same configuration $A_{V_{NL}} > A_{V_L} > A_{V_s}$***
- ***For a particular design, the larger the level of RL, the greater is the level of ac gain.***
- ***For a particular amplifier, the smaller the internal resistance of the signal source, the greater is the overall gain.***
- ***For any network, such as those shown in Fig. 54 that have coupling capacitors, the source and load resistance do not affect the dc biasing levels.***

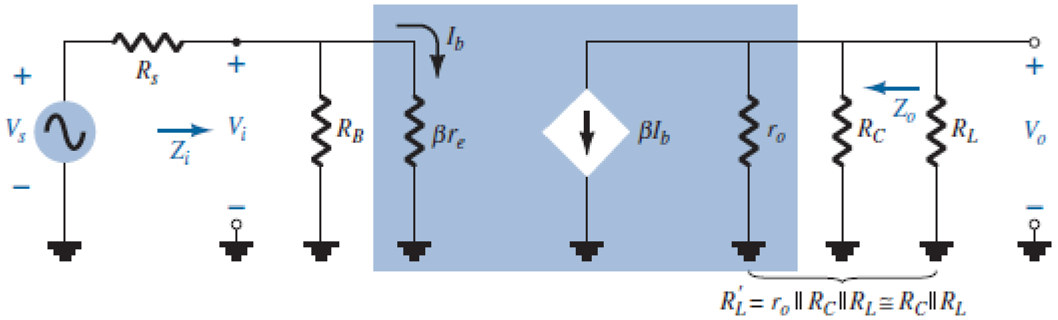

FIG. 55
The ac equivalent network for the network of Fig. 54c.

It is particularly interesting that Fig. 55 is exactly the same in appearance as Fig. 22 except that now there is a load resistance in parallel with RC and a source resistance has been introduced in series with a source Vs. The parallel combination of

$$R_L' = r_o \| R_C \| R_L \cong R_C \| R_L$$

and
$$V_o = -\beta I_b R_L' = -\beta I_b (R_C \| R_L)$$

with
$$I_b = \frac{V_i}{\beta r_e}$$

gives
$$V_o = -\beta\left(\frac{V_i}{\beta r_e}\right)(R_C \| R_L)$$

so that
$$\boxed{A_{v_L} = \frac{V_o}{V_i} = -\frac{R_C \| R_L}{r_e}} \qquad (73)$$

$$\boxed{Z_i = R_B \| \beta r_e} \qquad (74)$$

$$\boxed{Z_o = R_C \| r_o} \qquad (75)$$

as before.

If the overall gain from signal source Vs to output voltage Vo is desired, it is only necessary to apply the voltage-divider rule as follows:

$$V_i = \frac{Z_i V_s}{Z_i + R_s}$$

and $$\frac{V_i}{V_s} = \frac{Z_i}{Z_i + R_s}$$

or $$A_{v_s} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \cdot \frac{V_i}{V_s} = A_{v_L} \frac{Z_i}{Z_i + R_s}$$

so that $$\boxed{A_{v_s} = \frac{Z_i}{Z_i + R_s} A_{v_L}} \qquad (76)$$

EXAMPLE 11 Using the parameter values for the fixed-bias configuration of Example 1 with an applied load of 4.7 kΩ and a source resistance of 0.3 kΩ, determine the following and compare to the no-load values:

a. A_{v_L}.
b. A_{v_s}.
c. Z_i.
d. Z_o.

Solution:

a. Eq. (73): $A_{v_L} = -\frac{R_C \| R_L}{r_e} = -\frac{3\,\text{k}\Omega \| 4.7\,\text{k}\Omega}{10.71\,\Omega} = -\frac{1.831\,\text{k}\Omega}{10.71\,\Omega} = \mathbf{-170.98}$

which is significantly less than the no-load gain of −280.11.

b. Eq. (76): $A_{v_s} = \frac{Z_i}{Z_i + R_s} A_{v_L}$

With $Z_i = 1.07\,\text{k}\Omega$ from Example 1, we have

$$A_{v_s} = \frac{1.07\,\text{k}\Omega}{1.07\,\text{k}\Omega + 0.3\,\text{k}\Omega}(-170.98) = \mathbf{-133.54}$$

which again is significantly less than $A_{v_{NL}}$ or A_{v_L}.

c. $Z_i = \mathbf{1.07\,k\Omega}$ as obtained for the no-load situation.

d. $Z_o = R_C = \mathbf{3\,k\Omega}$ as obtained for the no-load situation.

The example clearly demonstrates that $A_{v_{NL}} > A_{v_L} > A_{v_s}$.

For the voltage-divider configuration of Fig. 56 with an applied load and series source resistor the ac equivalent network is as shown in Fig. 57.

First note the strong similarities with Fig. 55, with the only difference being the parallel connection of R1 and R2 instead of just RB. Everything else is exactly the same. The following equations result for the important parameters of the configuration:

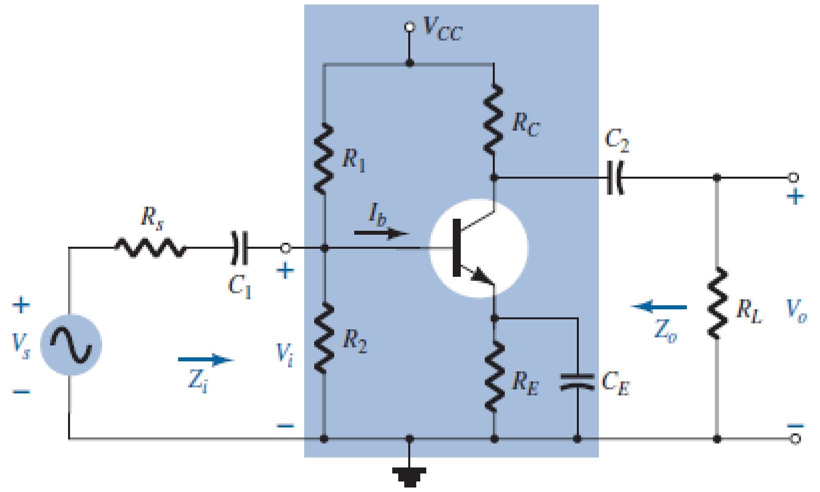

FIG. 56
Voltage-divider bias configuration with R_s and R_L.

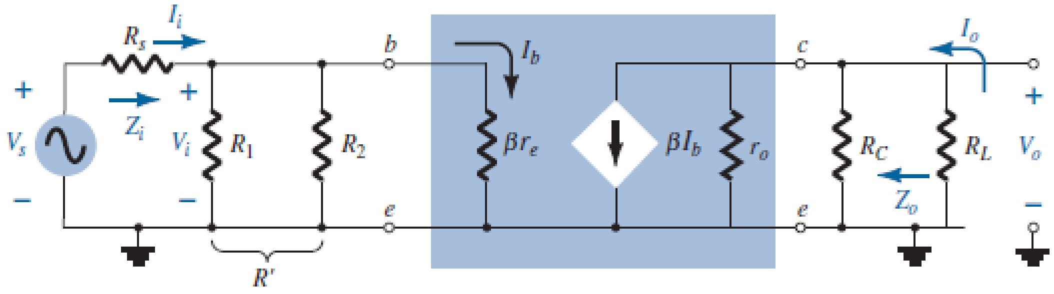

FIG. 57
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 56.

$$\boxed{A_{v_L} = \frac{V_o}{V_i} = -\frac{R_C \| R_L}{r_e}} \quad (77)$$

$$\boxed{Z_i = R_1 \| R_2 \| \beta r_e} \quad (78)$$

$$\boxed{Z_o = R_C \| r_o} \quad (79)$$

For the emitter-follower configuration of Fig. 58 the small-signal ac equivalent network is as shown in Fig. 59. The only difference between Fig. 59 and the unloaded configuration of Fig. 37 is the parallel combination of RE and RL and the addition of the source resistor Rs. The equations for the quantities of interest can therefore be determined by simply replacing RE by RE ∥ RL wherever RE appears. If RE does not appear in an equation, the load resistor RL does not affect that parameter. That is,

$$A_{v_L} = \frac{V_o}{V_i} = \frac{R_E \| R_L}{R_E \| R_L + r_e} \quad (80)$$

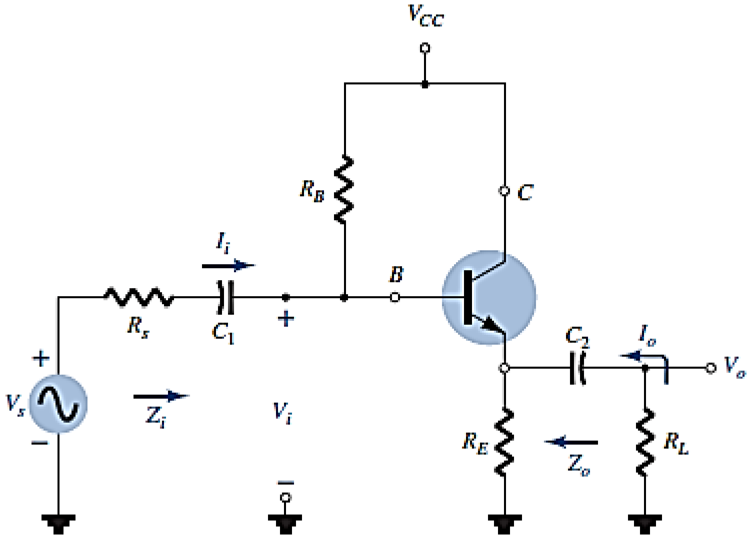

FIG. 58
Emitter-follower configuration with R_s and R_L.

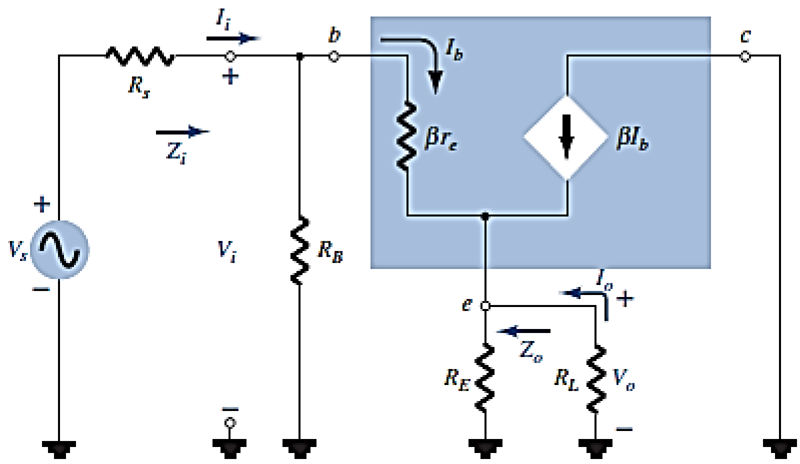

FIG. 59
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 58.

$$Z_i = R_B \| Z_b \quad (81)$$

$$Z_b \cong \beta(R_E \| R_L) \quad (82)$$

$$Z_o \cong r_e \quad (83)$$

13 Determining The Current Gain

You may have noticed in the previous sections that the current gain was not determined for each configuration. In reality the voltage gain is usually the gain of most importance. The absence of the derivations should not cause concern because:

For each transistor configuration, the current gain can be determined directly from the voltage gain, the defined load, and the input impedance.

The derivation of the equation linking the voltage and current gains can be derived using the two-port configuration of Fig. 60.

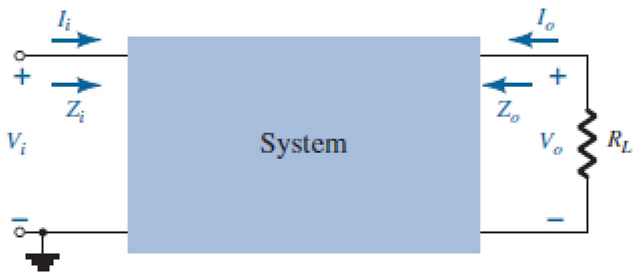

FIG. 60
Determining the current gain using the voltage gain.

The current gain is defined by

$$\boxed{A_i = \frac{I_o}{I_i}} \qquad \textbf{(84)}$$

Applying Ohm's law to the input and output circuits results in

$$I_i = \frac{V_i}{Z_i} \quad \text{and} \quad I_o = -\frac{V_o}{R_L}$$

$$A_{i_L} = \frac{I_o}{I_i} = \frac{-\dfrac{V_o}{R_L}}{\dfrac{V_i}{Z_i}} = -\frac{V_o}{V_i}\cdot\frac{Z_i}{R_L}$$

$$\boxed{A_{i_L} = -A_{v_L}\frac{Z_i}{R_L}} \qquad \textbf{(85)}$$

The value of *RL* is defined by the location of *Vo* and *Io*.

$$I_i = \frac{V_i}{Z_i} = \frac{V_i}{1.35\ \text{k}\Omega} \text{ and } I_o = -\frac{V_o}{R_L} = -\frac{V_o}{6.8\ \text{k}\Omega}$$

so that
$$A_{i_L} = \frac{I_o}{I_i} = \frac{\left(\dfrac{V_o}{6.8\ \text{k}\Omega}\right)}{\dfrac{V_i}{1.35\ \text{k}\Omega}} = -\left(\frac{V_o}{V_i}\right)\left(\frac{1.35\ \text{k}\Omega}{6.8\ \text{k}\Omega}\right)$$

$$= -(-368.76)\left(\frac{1.35\ \text{k}\Omega}{6.8\ \text{k}\Omega}\right) = \mathbf{73.2}$$

Using Eq. 82:
$$A_{i_L} = -A_{v_L}\frac{Z_i}{R_L} = -(-368.76)\left(\frac{1.35\ \text{k}\Omega}{6.8\ \text{k}\Omega}\right) = \mathbf{73.2}$$