PEARSON NEW INTERNATIONAL EDITION

Electronic Devices and Circuit Theory

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Eleventh Edition

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e. $Z_o = r_o \| R_C = 50\ \text{k}\Omega \| 3\ \text{k}\Omega = \mathbf{2.83\ k\Omega}$ vs. $3\ \text{k}\Omega$

$$A_v = -\frac{r_o \| R_C}{r_e} = \frac{2.83\ \text{k}\Omega}{10.71\ \Omega} = \mathbf{-264.24} \text{ vs. } -280.11$$

6 VOLTAGE-DIVIDER BIAS

The next configuration to be analyzed is the *voltage-divider* bias network of Fig. 26. Recall that the name of the configuration is a result of the voltage-divider bias at the input side to determine the dc level of V_B.

Substituting the r_e equivalent circuit results in the network of Fig. 27. Note the absence of R_E due to the low-impedance shorting effect of the bypass capacitor, C_E. That is, at the frequency (or frequencies) of operation, the reactance of the capacitor is so small compared to R_E that it is treated as a short circuit across R_E. When V_{CC} is set to zero, it places one end of R_1 and R_C at ground potential as shown in Fig. 27. In addition, note that R_1 and R_2 remain part of the input circuit, whereas R_C is part of the output circuit. The parallel combination of R_1 and R_2 is defined by

$$R' = R_1 \| R_2 = \frac{R_1 R_2}{R_1 + R_2} \tag{11}$$

Z_i From Fig. 27

$$Z_i = R' \| \beta r_e \tag{12}$$

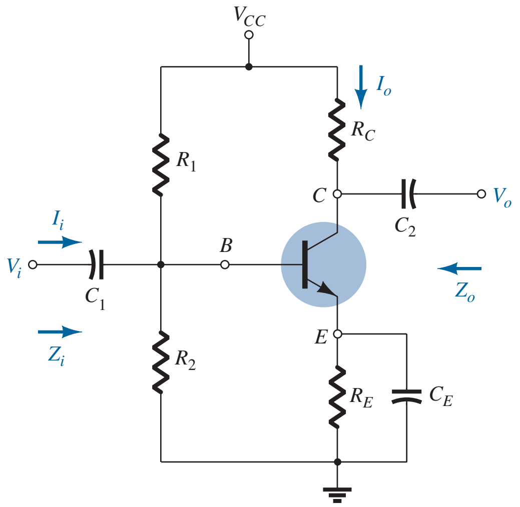

FIG. 26
Voltage-divider bias configuration.

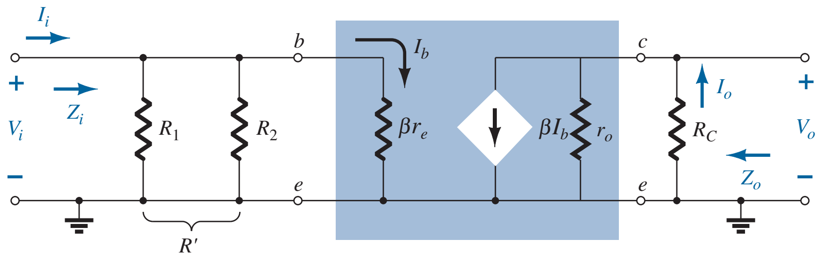

FIG. 27
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 26.

Z_o From Fig. 27 with V_i set to 0 V, resulting in $I_b = 0\ \mu A$ and $\beta I_b = 0$ mA,

$$Z_o = R_C \| r_o \tag{13}$$

If $r_o \geq 10R_C$,

$$Z_o \cong R_C \Big|_{r_o \geq 10R_C} \tag{14}$$

A_v Because R_C and r_o are in parallel,

$$V_o = -(\beta I_b)(R_C \| r_o)$$

and

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

so that

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

and

$$A_v = \frac{V_o}{V_i} = \frac{-R_C \| r_o}{r_e} \tag{15}$$

which you will note is an exact duplicate of the equation obtained for the fixed-bias configuration.

For $r_o \geq 10R_C$,

$$A_v = \frac{V_o}{V_i} \cong -\frac{R_C}{r_e} \Big|_{r_o \geq 10R_C} \tag{16}$$

Phase Relationship The negative sign of Eq. (15) reveals a 180° phase shift between V_o and V_i.

EXAMPLE 2 For the network of Fig. 28, determine:

a. r_e.
b. Z_i.
c. Z_o ($r_o = \infty\ \Omega$).
d. A_v ($r_o = \infty\ \Omega$).
e. The parameters of parts (b) through (d) if $r_o = 50$ kΩ and compare results.

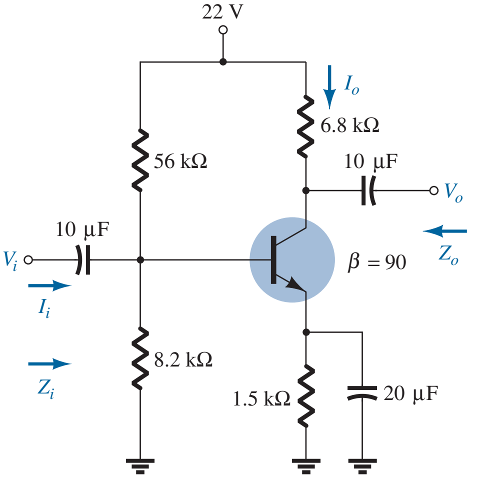

FIG. 28
Example 2.

Solution:

a. DC: Testing $\beta R_E > 10R_2$,

$$(90)(1.5\text{ k}\Omega) > 10(8.2\text{ k}\Omega)$$
$$135\text{ k}\Omega > 82\text{ k}\Omega\ (\textit{satisfied})$$

Using the approximate approach, we obtain

$$V_B = \frac{R_2}{R_1 + R_2}V_{CC} = \frac{(8.2\text{ k}\Omega)(22\text{ V})}{56\text{ k}\Omega + 8.2\text{ k}\Omega} = 2.81\text{ V}$$
$$V_E = V_B - V_{BE} = 2.81\text{ V} - 0.7\text{ V} = 2.11\text{ V}$$
$$I_E = \frac{V_E}{R_E} = \frac{2.11\text{ V}}{1.5\text{ k}\Omega} = 1.41\text{ mA}$$
$$r_e = \frac{26\text{ mV}}{I_E} = \frac{26\text{ mV}}{1.41\text{ mA}} = \mathbf{18.44\ \Omega}$$

b. $R' = R_1 \| R_2 = (56\text{ k}\Omega)\|(8.2\text{ k}\Omega) = 7.15\text{ k}\Omega$

$Z_i = R' \| \beta r_e = 7.15\text{ k}\Omega \| (90)(18.44\ \Omega) = 7.15\text{ k}\Omega \| 1.66\text{ k}\Omega$

$= \mathbf{1.35\text{ k}\Omega}$

c. $Z_o = R_C = \mathbf{6.8\text{ k}\Omega}$

d. $A_v = -\dfrac{R_C}{r_e} = -\dfrac{6.8\text{ k}\Omega}{18.44\ \Omega} = \mathbf{-368.76}$

e. $Z_i = \mathbf{1.35\text{ k}\Omega}$

$Z_o = R_C \| r_o = 6.8\text{ k}\Omega \| 50\text{ k}\Omega = \mathbf{5.98\text{ k}\Omega}$ vs. $6.8\text{ k}\Omega$

$A_v = -\dfrac{R_C \| r_o}{r_e} = -\dfrac{5.98\text{ k}\Omega}{18.44\ \Omega} = \mathbf{-324.3}$ vs. -368.76

There was a measurable difference in the results for Z_o and A_v, because the condition $r_o \geq 10R_C$ was *not* satisfied.

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.

Unbypassed

The most fundamental of unbypassed configurations appears in Fig. 29. The r_e equivalent model is substituted in Fig. 30, but note the absence of the resistance r_o. The effect of r_o is to make the analysis a great deal more complicated, and considering the fact that in

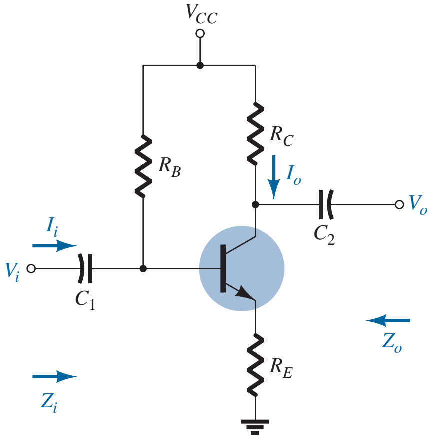

FIG. 29
CE emitter-bias configuration.

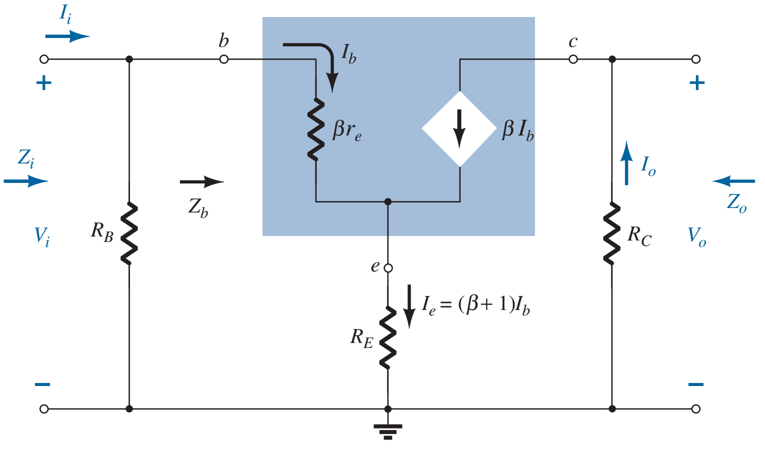

FIG. 30
Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 29.

most situations its effect can be ignored, it will not be included in the present analysis. However, the effect of r_o will be discussed later in this section.

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

$$V_i = I_b\beta r_e + (\beta + 1)I_b R_E$$

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

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

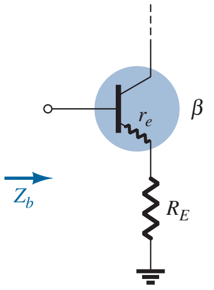

FIG. 31
Defining the input impedance of a transistor with an unbypassed emitter resistor.

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.

Effect of r_o The equations appearing below will clearly reveal the additional complexity resulting from including r_o in the analysis. Note in each case, however, that when certain conditions are met, the equations return to the form just derived. The derivation of each equation is beyond the needs of this text and is left as an exercise for the reader. Each equation can be derived through *careful* application of the basic laws of circuit analysis such as Kirchhoff's voltage and current laws, source conversions, Thévenin's theorem, and so on. The equations were included to remove the nagging question of the effect of r_o on the important parameters of a transistor configuration.

Z_i

$$Z_b = \beta r_e + \left[\frac{(\beta + 1) + R_C/r_o}{1 + (R_C + R_E)/r_o}\right] R_E \tag{25}$$

Because the ratio R_C/r_o is always much less than $(\beta + 1)$,

$$Z_b \cong \beta r_e + \frac{(\beta + 1)R_E}{1 + (R_C + R_E)/r_o}$$

For $r_o \geq 10(R_C + R_E)$,

$$Z_b \cong \beta r_e + (\beta + 1)R_E$$

which compares directly with Eq. (17).

In other words, if $r_o \geq 10(R_C + R_E)$, all the equations derived earlier result. Because $\beta + 1 \cong \beta$, the following equation is an excellent one for most applications:

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

Z_o

$$Z_o = R_C \left\| \left[r_o + \frac{\beta(r_o + r_e)}{1 + \dfrac{\beta r_e}{R_E}} \right] \right. \tag{27}$$

However, $r_o \gg r_e$, and

$$Z_o \cong R_C \left\| r_o \left[1 + \frac{\beta}{1 + \dfrac{\beta r_e}{R_E}} \right] \right.$$

which can be written as

$$Z_o \cong R_C \left\| r_o \left[1 + \frac{1}{\dfrac{1}{\beta} + \dfrac{r_e}{R_E}} \right] \right.$$

Typically $1/\beta$ and r_e/R_E are less than one with a sum usually less than one. The result is a multiplying factor for r_o greater than one. For $\beta = 100$, $r_e = 10\ \Omega$, and $R_E = 1\ \text{k}\Omega$,

$$\frac{1}{\dfrac{1}{\beta} + \dfrac{r_e}{R_E}} = \frac{1}{\dfrac{1}{100} + \dfrac{10\ \Omega}{1000\ \Omega}} = \frac{1}{0.02} = 50$$

and

$$Z_o = R_C \| 51 r_o$$

which is certainly simply R_C. Therefore,

$$Z_o \cong R_C \qquad {\scriptstyle \text{Any level of } r_o} \tag{28}$$

which was obtained earlier.

A_v

$$A_v = \frac{V_o}{V_i} = \frac{-\frac{\beta R_C}{Z_b}\left[1 + \frac{r_e}{r_o}\right] + \frac{R_C}{r_o}}{1 + \frac{R_C}{r_o}} \qquad (29)$$

The ratio $\frac{r_e}{r_o} \ll 1$, and

$$A_v = \frac{V_o}{V_i} \cong \frac{-\frac{\beta R_C}{Z_b} + \frac{R_C}{r_o}}{1 + \frac{R_C}{r_o}}$$

For $r_o \geq 10R_C$,

$$A_v = \frac{V_o}{V_i} \cong -\frac{\beta R_C}{Z_b}\Bigg|_{r_o \geq 10R_C} \qquad (30)$$

as obtained earlier.

Bypassed

If R_E of Fig. 29 is bypassed by an emitter capacitor C_E, the complete r_e 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.

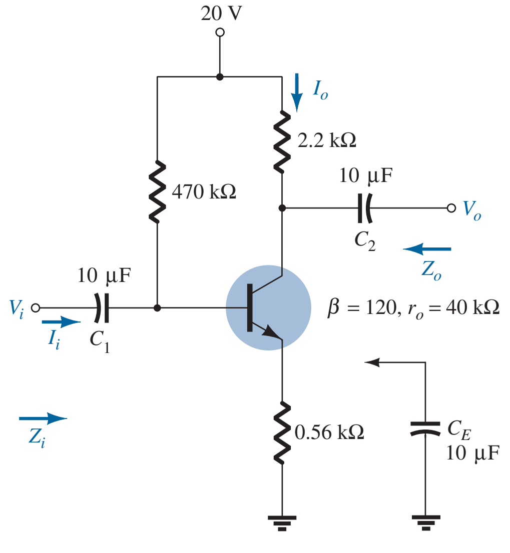

FIG. 32
Example 3.

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