Chapter Three :- Field-Effect Transistors and Applications

FETs vs. BJTs

Similarities:

- Amplifiers
- Switching devices
- Impedance matching circuits

Differences:

- FETs are voltage controlled devices. BJTs are current controlled devices.
- FETs have a higher input impedance. BJTs have higher gains.
- FETs are less sensitive to temperature variations and are more easily integrated on ICs.
- FETs are generally more static sensitive than BJTs.

FET Types

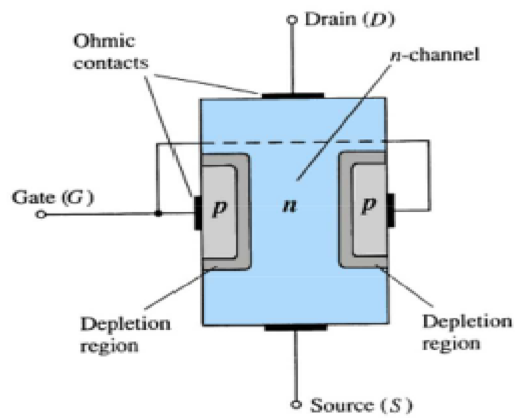
- **JFET:** Junction FET
- **MOSFET:** Metal–Oxide–Semiconductor FET
  - **D-MOSFET:** Depletion MOSFET
  - **E-MOSFET:** Enhancement MOSFET

JFET Construction

There are two types of JFETs

- ***n*-channel**
- ***p*-channel**

The n-channel is more widely used.



There are three terminals:

- **Drain (D) and Source (S) are connected to the *n*-channel**
- **Gate (G) is connected to the *p*-type material**

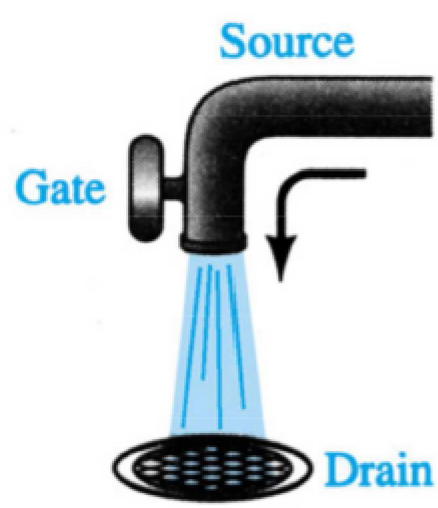
JFET Operation: The Basic Idea

JFET operation can be compared to a water spigot.

The source of water pressure is the accumulation of electrons at the negative pole of the drain-source voltage.

The drain of water is the electron deficiency (or holes) at the positive pole of the applied voltage.

The control of flow of water is the gate voltage that controls the width of the n-channel and, therefore, the flow of charges from source to drain.



JFET Operating Characteristics

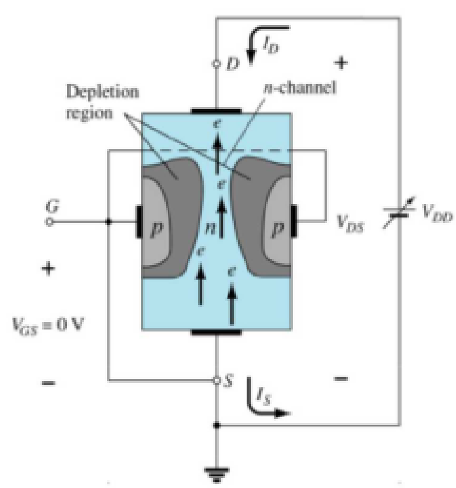
There are three basic operating conditions for a JFET:

- $V_{GS} = 0$, V_{DS} increasing to some positive value
- $V_{GS} < 0$, V_{DS} at some positive value
- Voltage-controlled resistor

JFET Operating Characteristics: $V_{GS} = 0$ V

Three things happen when $V_{GS} = 0$ and V_{DS} is increased from 0 to a more positive voltage

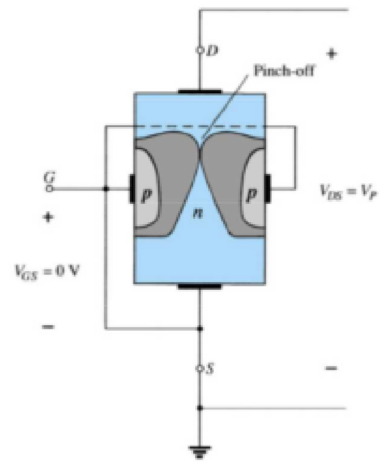
- The depletion region between p-gate and n-channel increases as electrons from n-channel combine with holes from p-gate.
- Increasing the depletion region, decreases the size of the n-channel which increases the resistance of the n-channel.
- Even though the n-channel resistance is increasing, the current (I_D) from source to drain through the n-channel is increasing. This is because V_{DS} is increasing.



JFET Operating Characteristics: Pinch Off

If $V_{GS} = 0$ and V_{DS} is further increased to a more positive voltage, then the depletion zone gets so large that it pinches off the n-channel.

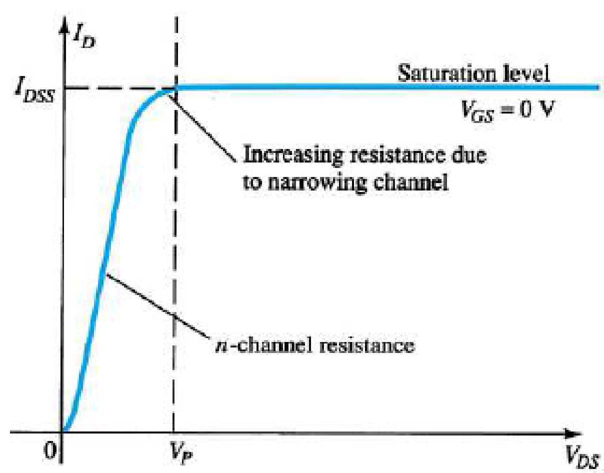
This suggests that the current in the n-channel (I_D) would drop to 0A, but it does just the opposite–as V_{DS} increases, so does I_D.



JFET Operating Characteristics: Saturation

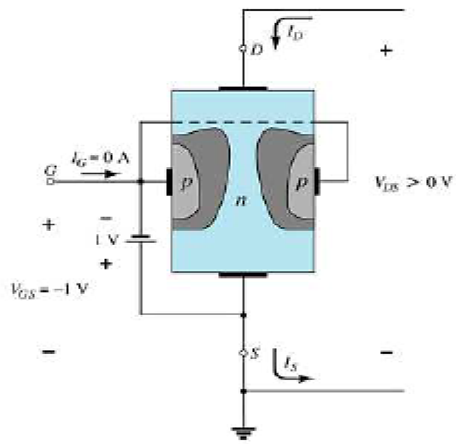
At the pinch-off point:

- Any further increase in V_{GS} does not produce any increase in I_D. V_{GS} at pinch-off is denoted as V_p.
- I_D is at saturation or maximum. It is referred to as I_{DSS}.
- The ohmic value of the channel is maximum.



JFET Operating Characteristics

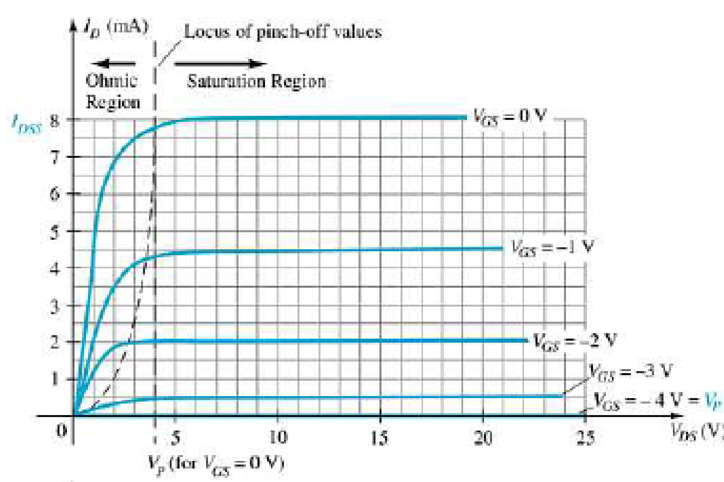
As V_{GS} becomes more negative, the depletion region increases.



JFET Operating Characteristics

As V_{GS} becomes more negative:

- The JFET experiences pinch-off at a lower voltage (V_P).
- I_D decreases ($I_D < I_{DSS}$) even though V_{DS} is increased.
- Eventually I_D reaches 0 A. V_{GS} at this point is called V_p or $V_{GS(off)}$..



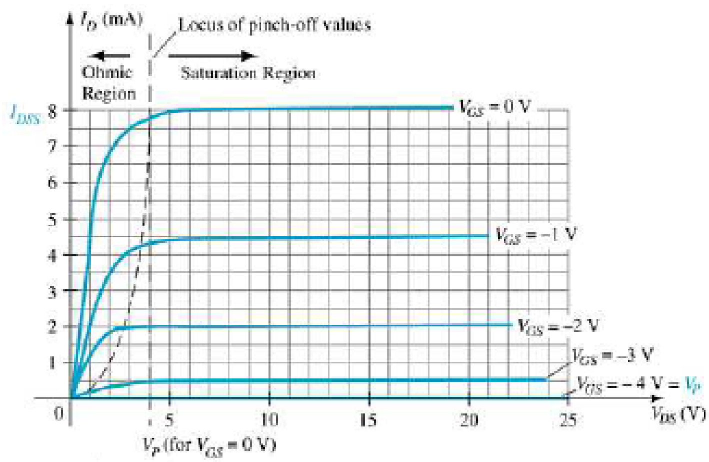
Also note that at high levels of V_{DS} the JFET reaches a breakdown situation. I_D increases uncontrollably if $V_{DS} > V_{DSmax}$.

JFET Operating Characteristics: Voltage-Controlled Resistor

The region to the left of the pinch-off point is called the **ohmic region.**

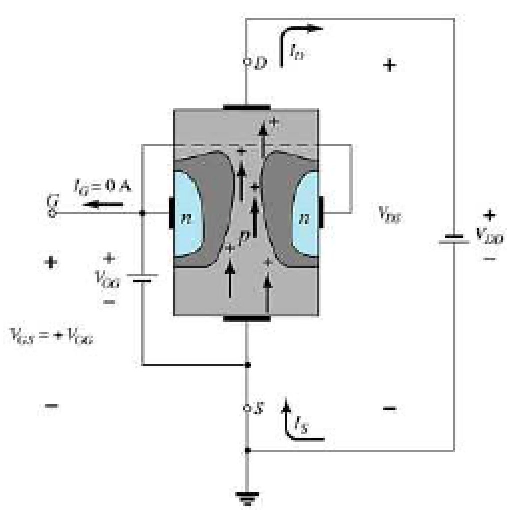
The JFET can be used as a variable resistor, where V_{GS} controls the drain-source resistance (r_d). As V_{GS} becomes more negative, the resistance (r_d) increases.

$$r_d = \frac{r_o}{\left(1 - \frac{V_{GS}}{V_P}\right)^2}$$



p-Channel JFETS

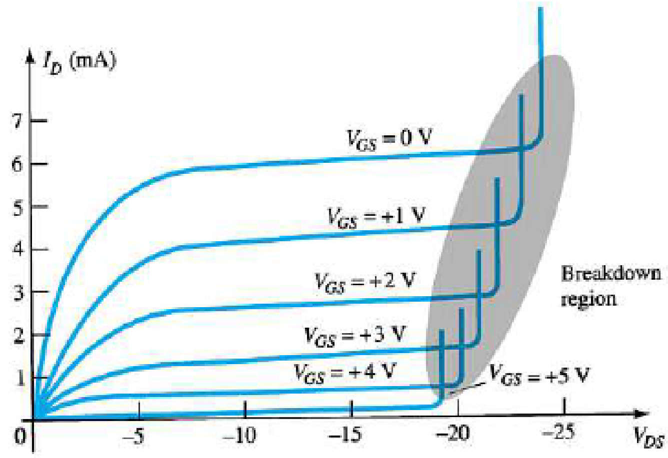
The *p*-channel JFET behaves the same as the *n*-channel JFET, except the voltage polarities and current directions are reversed.



p-Channel JFET Characteristics

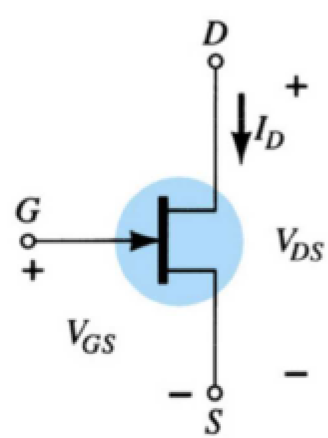
As V_{GS} increases more positively

- The depletion zone increases
- I_D decreases ($I_D < I_{DSS}$)
- Eventually $I_D = 0$ A



Also note that at high levels of V_{DS} the JFET reaches a breakdown situation: I_D increases uncontrollably if $V_{DS} > V_{DSmax}$.

N-Channel JFET Symbol



JFET Transfer Characteristics

The transfer characteristic of input-to-output is not as straightforward in a JFET as it is in a BJT.

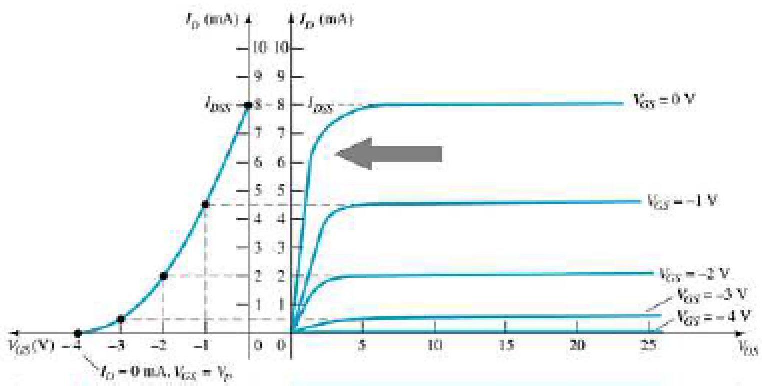
In a BJT, β indicates the relationship between I_B (input) and I_C (output).

In a JFET, the relationship of V_{GS} (input) and I_D (output) is a little more complicated:

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

JFET Transfer Curve

This graph shows the value of I_D for a given value of V_{GS}.



Plotting the JFET Transfer Curve

Using I_{DSS} and Vp ($V_{GS(off)}$) values found in a specification sheet, the transfer curve can be plotted according to these three steps:

Step 1

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

Solving for $V_{GS} = 0V$ $\quad I_D = I_{DSS}$

Step 2

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

Solving for $V_{GS} = V_p\ (V_{GS(off)})$ $\quad I_D = 0A$

Step 3

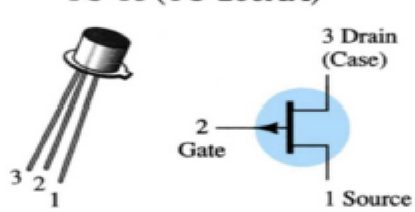
Solving for $V_{GS} = 0V$ to V_p $\quad I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$

Case and Terminal Identification

2N2844

CASE 22-03, STYLE 12

TO-18 (TO-206AA)



JFETs

GENERAL PURPOSE

P-CHANNEL

MOSFETs

MOSFETs have characteristics similar to JFETs and additional characteristics that make then very useful.

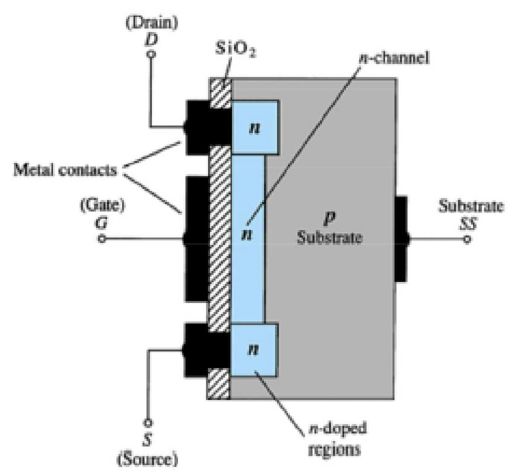
There are two types of MOSFETs:

- **Depletion-Type**
- **Enhancement-Type**

Depletion-Type MOSFET Construction

The Drain (D) and Source (S) connect to the to *n*-doped regions. These *n*-doped regions are connected via an *n*-channel. This *n*-channel is connected to the Gate (G) via a thin insulating layer of SiO_2.

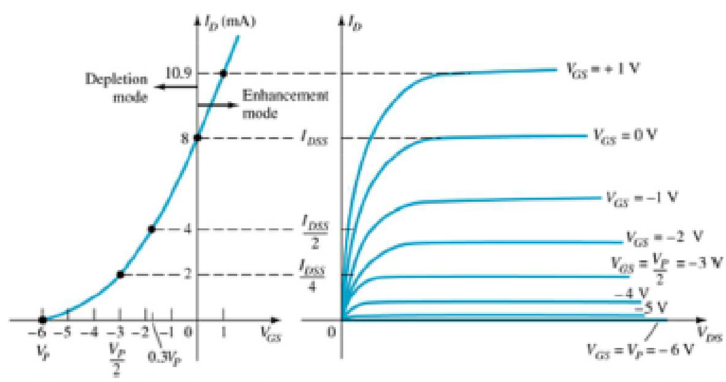
The *n*-doped material lies on a *p*-doped substrate that may have an additional terminal connection called Substrate (SS).



Basic MOSFET Operation

A depletion-type MOSFET can operate in two modes:

- **Depletion mode**
- **Enhancement mode**



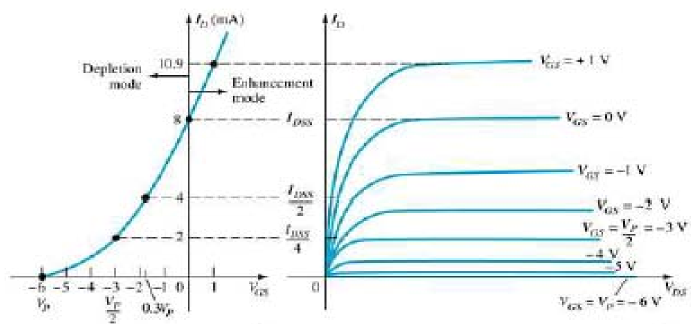
D-Type MOSFET in Depletion Mode

Depletion Mode

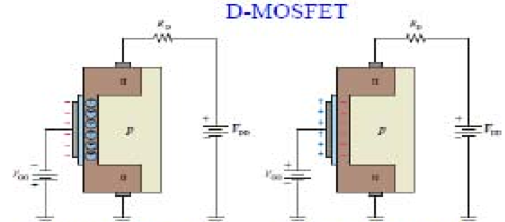
The characteristics are similar to a JFET.

- **When $V_{GS} = 0$ V, $I_D = I_{DSS}$**
- **When $V_{GS} < 0$ V, $I_D < I_{DSS}$**
- **The formula used to plot the transfer curve still applies:**

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$



D-MOSFET



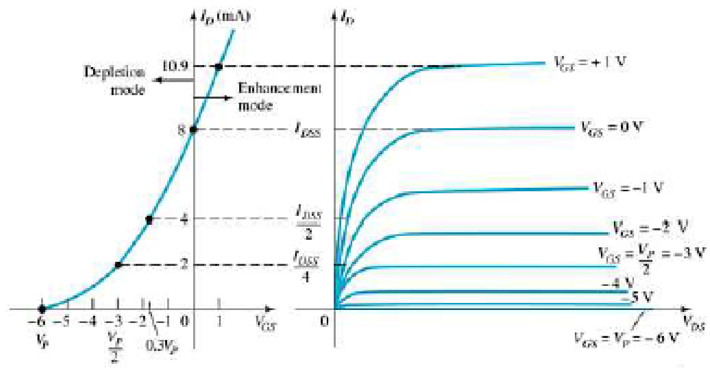
operating in D-mode operating in E-mode

D-Type MOSFET in Enhancement Mode

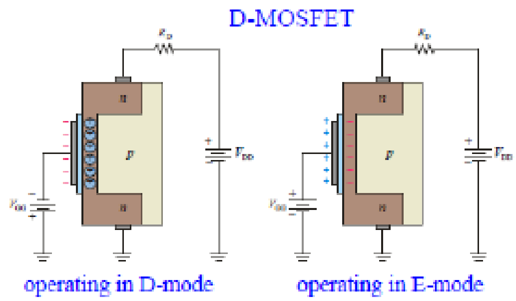
Enhancement Mode

- $V_{GS} > 0$ V
- I_D increases above I_{DSS}
- The formula used to plot the transfer curve still applies:

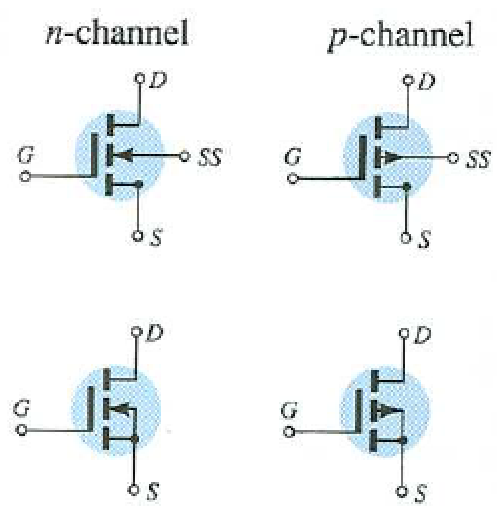
$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$



Note that V_{GS} is now a positive polarity



MOSFET Symbols



Common FET Biasing Circuits

JFET Biasing Circuits

- Fixed – Bias
- Self-Bias
- Voltage-Divider Bias

D-Type MOSFET Biasing Circuits

- Self-Bias
- Voltage-Divider Bias

E-Type MOSFET Biasing Circuits

- Feedback Configuration
- Voltage-Divider Bias

Basic Current Relationships

For all FETs:

$$I_G \cong 0A$$
$$I_D = I_S$$

For JFETS and D-Type MOSFETs:

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

For E-Type MOSFETs:

$$I_D = k(V_{GS} - V_T)^2$$

Applications

Voltage-controlled resistor
JFET voltmeter
Timer network
Fiber optic circuitry
MOSFET relay driver

Chapter Three (part two) : BJT small signal ac analysis

COMMON-EMITTER FIXED-BIAS CONFIGURATION

- the common-emitter *fixed-bias* network is shown in Fig. 8.1.

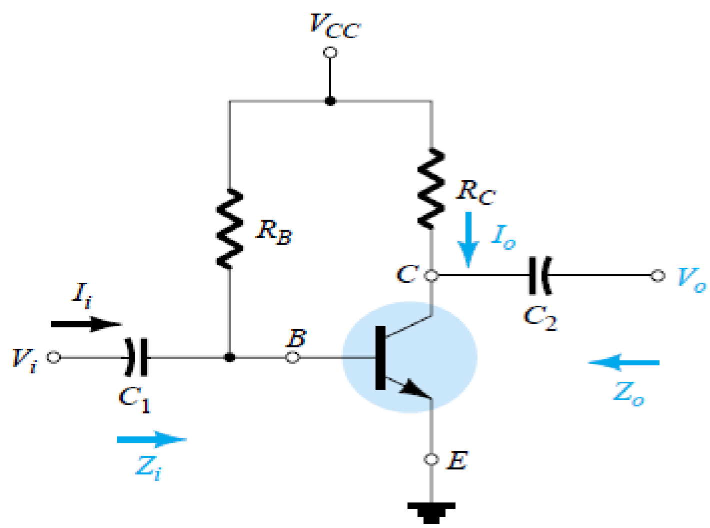


Figure 8.1 Common-emitter fixed-bias configuration.

- The small-signal ac analysis begins by removing the dc effects of *VCC* and replacing the dc blocking capacitors *C*1 and *C*2 by short-circuit equivalents, resulting in the network of Fig. 8.2.

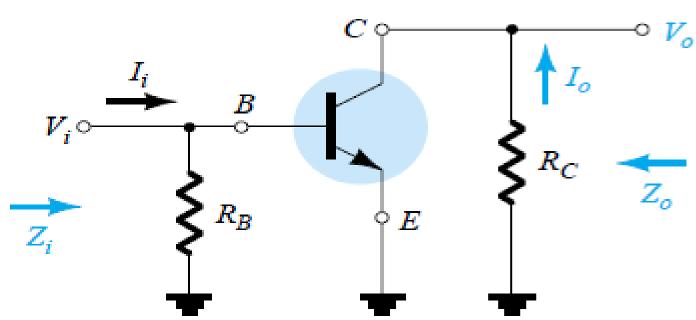


Figure 8.2 Network of Figure 8.1 following the removal of the effects of V_{CC}, C_1, and C_2.

- Substituting the *re* model for the common-emitter configuration of Fig. 8.2 will result in the network of Fig. 8.3.

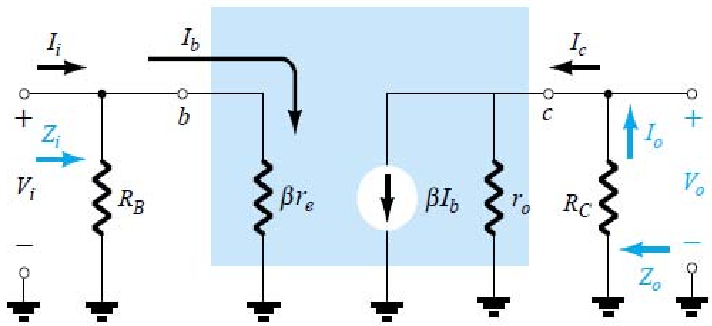


Figure 8.3 Substituting the r_e model into the network of Fig. 8.2.

- The next step is to determine β, *re*, and *ro* .Figure 8.3 clearly reveals that

$$\boxed{Z_i = R_B \| \beta r_e} \quad \text{ohms} \qquad (8.1)$$

- For the majority of situations *RB* is greater than β*re* by more than a factor of 10,

$$\boxed{Z_i \cong \beta r_e}_{R_B \geq 10\beta r_e} \quad \text{ohms} \qquad (8.2)$$

- For Fig. 8.3, when *Vi* = 0, *Ii* =0, *Ib* =0, resulting in an open-circuit equivalence for the current source. The result is the configuration of Fig. 8.4.

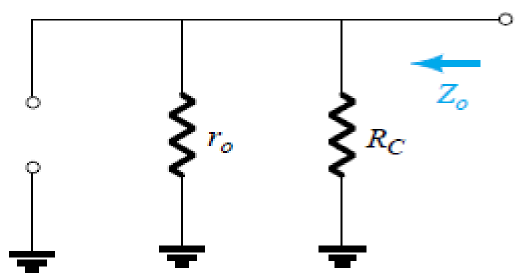


Figure 8.4 Determining Z_o for the network of Fig. 8.3.

$$\boxed{Z_o = R_C \| r_o} \quad \text{ohms} \qquad (8.3)$$

If $r_o \geq 10\ R_D$, the approximation $R_C \| r_o \cong R_C$ is frequently applied and

$$\boxed{Z_o \cong R_C}_{r_o \geq 10 R_C} \qquad (8.4)$$

- ***Av*:** The resistors *ro* and *RC* are in parallel

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

but $$I_b = \frac{V_i}{\beta r_c}$$

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

and $$\boxed{A_v = \frac{V_o}{V_i} = -\frac{(R_C \| r_o)}{r_e}}$$

If $r_o \geq 10R_C$,

$$\boxed{A_v = -\frac{R_C}{r_e}}_{r_o \geq 10R_C}$$

- ***Ai*:** The current gain is determined in the following manner: Applying the current- divider rule to the input and output circuits,

$$I_o = \frac{(r_o)(\beta I_b)}{r_o + R_C} \quad \text{and} \quad \frac{I_o}{I_b} = \frac{r_o \beta}{r_o + R_C}$$

with $$I_b = \frac{(R_B)(I_i)}{R_B + \beta r_e} \quad \text{or} \quad \frac{I_b}{I_i} = \frac{R_B}{R_B + \beta r_e}$$

The result is

$$A_i = \frac{I_o}{I_i} = \left(\frac{I_o}{I_b}\right)\left(\frac{I_b}{I_i}\right) = \left(\frac{r_o \beta}{r_o + R_C}\right)\left(\frac{R_B}{R_B + \beta r_e}\right)$$

and $$\boxed{A_i = \frac{I_o}{I_i} = \frac{\beta R_B r_o}{(r_o + R_C)(R_B + \beta r_e)}} \qquad (8.7)$$

which is certainly an unwieldy, complex expression.

However, if $r_o \geq 10R_C$ and $R_B \geq 10\beta r_e$, which is often the case,

$$A_i = \frac{I_o}{I_i} \cong \frac{\beta R_B r_o}{(r_o)(R_B)}$$

and $$\boxed{A_i \cong \beta}_{r_o \geq 10R_C,\ R_B \geq 10\beta r_e} \qquad (8.8)$$

$$\boxed{A_i = -A_v \frac{Z_i}{R_C}} \qquad (8.9)$$

- **Phase Relationship:** The negative sign in the resulting equation for *Av* reveals that a 180° phase shift occurs between the input and output signals, as shown in Fig.8.5.

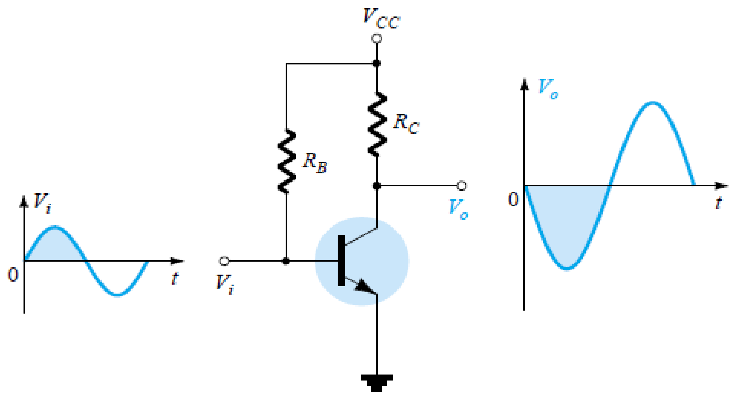


Figure 8.5 Demonstrating the 180° phase shift between input and output waveforms.

EXAMPLE:

For the network of Fig. 8.6:

(a) Determine r_e.

(b) Find Z_i (with $r_o = \infty\ \Omega$).

(c) Calculate Z_o (with $r_o = \infty\ \Omega$).

(d) Determine A_v (with $r_o = \infty\ \Omega$).

(e) Find A_i (with $r_o = \infty\ \Omega$).

(f) Repeat parts (c) through (e) including $r_o = 50\ \text{k}\Omega$ in all calculations and compare results.

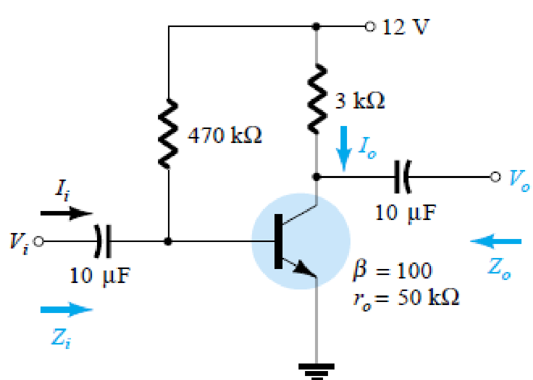


Figure 8.6 Example

Solution

(a) DC analysis:

$$I_B = \frac{V_{CC} - V_{BE}}{R_B} = \frac{12\text{ V} - 0.7\text{ V}}{470\text{ k}\Omega} = 24.0$$

$$I_E = (\beta + 1)I_B = (101)(24.04\ \mu\text{A}) = 2.42$$

$$r_e = \frac{26\text{ mV}}{I_E} = \frac{26\text{ mV}}{2.428\text{ mA}} = \mathbf{10.71\ \Omega}$$

(b) $\beta r_e = (100)(10.71\ \Omega) = 1.071\text{ k}\Omega$

$$Z_i = R_B \| \beta r_e = 470\text{ k}\Omega \| 1.071\text{ k}\Omega = \mathbf{1.069\text{ k}\Omega}$$

(c) $Z_o = R_C = \mathbf{3\text{ k}\Omega}$

(d) $A_v = -\dfrac{R_C}{r_e} = -\dfrac{3\text{ k}\Omega}{10.71\ \Omega} = \mathbf{-280.11}$

(e) Since $R_B \geq 10\beta r_e (470\text{ k}\Omega > 10.71\text{ k}\Omega)$

$$A_i \cong \beta = \mathbf{100}$$

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

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

$$A_i = \frac{\beta R_B r_o}{(r_o + R_C)(R_B + \beta r_e)} = \frac{(100)(470\text{ k}\Omega)(50\text{ k}\Omega)}{(50\text{ k}\Omega + 3\text{ k}\Omega)(470\text{ k}\Omega + 1.071\text{ k}\Omega)}$$

$$= \mathbf{94.13}\text{ vs. } 100$$

As a check:

$$A_i = -A_v \frac{Z_i}{R_C} = \frac{-(-264.24)(1.069\text{ k}\Omega)}{3\text{ k}\Omega} = \mathbf{94.16}$$

EMITTER-FOLLOWER CONFIGURATION

When the output is taken from the emitter terminal of the transistor as shown in Fig. 8.17, the network is referred to as an *emitter-follower*.

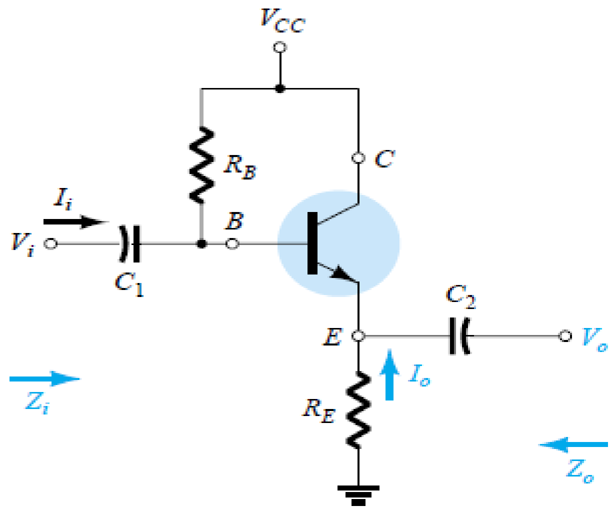


Figure 8.17 configuration.

- Substituting the *re* equivalent circuit into the network of Fig. 8.17 will result in the network of Fig. 8.18.

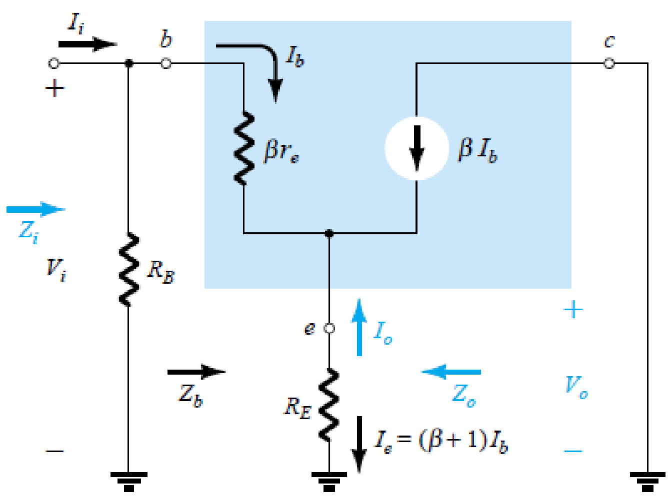


Figure 8.18 Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 8.17.

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

$$\boxed{Z_i = R_B \| Z_b} \tag{8.37}$$

$$\boxed{Z_b = \beta r_e + (\beta + 1)R_E} \tag{8.38}$$

$$\boxed{Z_b \cong \beta(r_e + R_E)} \tag{8.39}$$

$$\boxed{Z_b \cong \beta R_E} \tag{8.40}$$

- ***Zo***: The output impedance is best described by first writing the equation for the current *Ib*:

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

and then multiplying by $(\beta + 1)$ to establish I_e. That is,

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

Substituting for Z_b gives

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

or

$$I_e = \frac{V_i}{[\beta r_e/(\beta + 1)] + R_E}$$

but

$$(\beta + 1) \cong \beta$$

and

$$\frac{\beta r_e}{\beta + 1} \cong \frac{\beta r_e}{\beta} = r_e$$

so that

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

- If we now construct the network defined by Eq. (8.41), the configuration of Fig. 8.19 will result.

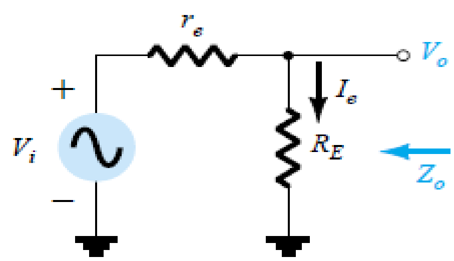


Figure 8.19 Defining the output impedance for the emitter-follower configuration.

To determine Z_o, V_i is set to zero and

$$\boxed{Z_o = R_E \| r_e}$$

Since R_E is typically much greater than r_e, the following appro plied:

$$\boxed{Z_o \cong r_e}$$

***A_v*:** Figure 8.19 can be utilized to determine the voltage g cation of the voltage-divider rule:

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

and

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

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

$$\boxed{A_v = \frac{V_o}{V_i} \cong 1}$$

- ***Ai*:** From Fig. 8.18,

$$I_b = \frac{R_B I_i}{R_B + Z_b}$$

or

$$\frac{I_b}{I_i} = \frac{R_B}{R_B + Z_b}$$

and

$$I_o = -I_e = -(\beta + 1)I_b$$

or

$$\frac{I_o}{I_b} = -(\beta + 1)$$

so that

$$A_i = \frac{I_o}{I_i} = \frac{I_o}{I_b}\frac{I_b}{I_i}$$

$$= -(\beta + 1)\frac{R_B}{R_B + Z_b}$$

and since

$$(\beta + 1) \cong \beta,$$

$$\boxed{A_i \cong -\frac{\beta R_B}{R_B + Z_b}} \tag{8.46}$$

or

$$\boxed{A_i = -A_v \frac{Z_i}{R_E}} \tag{8.47}$$

- **Phase relationship:** As revealed by Eq. (8.44) and earlier discussions of this section, *Vo* and *Vi* are in phase for the emitter-follower configuration.
- **Effect of *ro*:**

***Zi*:**

$$\boxed{Z_b = \beta r_e + \frac{(\beta + 1)R_E}{1 + \dfrac{R_E}{r_o}}} \tag{8.48}$$

If the condition $r_o \geq 10R_E$ is satisfied,

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

which matches earlier conclusions with

$$\boxed{Z_b \cong \beta(r_e + R_E)}_{r_o \geq 10R_E} \tag{8.49}$$

Z_o:

$$\boxed{Z_o = r_o \| R_E \| \frac{\beta r_e}{(\beta + 1)}} \tag{8.50}$$

Using $\beta + 1 \cong \beta$,

$$Z_o = r_o \| R_E \| r_e$$

and since $r_o \gg r_e$,

$$\boxed{Z_o \cong R_E \| r_e}_{\text{Any } r_o} \tag{8.51}$$

A_v:

$$\boxed{A_v = \frac{(\beta + 1)R_E/Z_b}{1 + \dfrac{R_E}{r_o}}} \tag{8.52}$$

If the condition $r_o \geq 10R_E$ is satisfied and we use the approximation $\beta + 1 \cong \beta$,

$$A_v \cong \frac{\beta R_E}{Z_b}$$

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

so that $$A_v \cong \frac{\beta R_E}{\beta(r_e + R_E)}$$

and $$\boxed{A_v \cong \frac{R_E}{r_e + R_E}}_{r_o \geq 10R_E} \tag{8.53}$$

EXAMPLE:-

For the emitter-follower network of Fig. 8.20, determine:

(a) r_e.

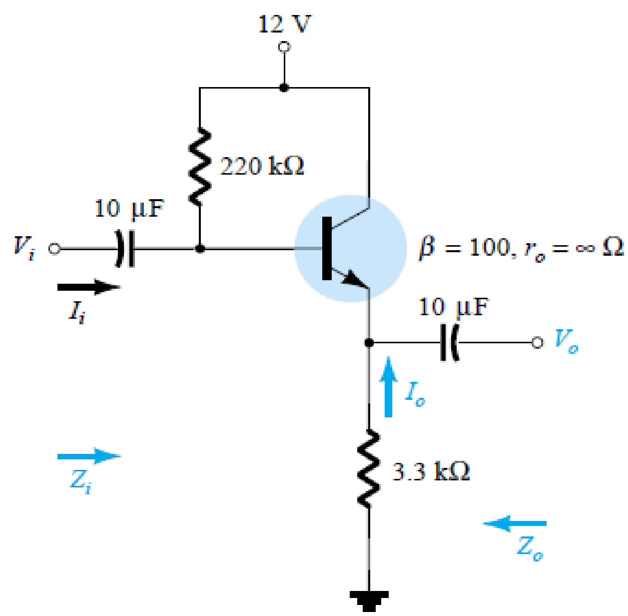
(b) Z_i.

(c) Z_o.

(d) A_v.

(e) A_i.

(f) Repeat parts (b) through (e) with $r_o = 25\ \text{k}\Omega$ and co



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

$$= (100)(12.61\ \Omega) + (101)(3.3\ \text{k}\Omega)$$
$$= 1.261\ \text{k}\Omega + 333.3\ \text{k}\Omega$$
$$= 334.56\ \text{k}\Omega \cong \beta R_E$$

$$Z_i = R_B \| Z_b = 220\ \text{k}\Omega \| 334.56\ \text{k}\Omega$$
$$= \mathbf{132.72\ k\Omega}$$

(c) $Z_o = R_E \| r_e = 3.3\ \text{k}\Omega \| 12.61\ \Omega$

$$= \mathbf{12.56\ \Omega} \cong r_e$$

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

(e) $$A_i \cong -\frac{\beta R_B}{R_B + Z_b} = -\frac{(100)(220\ \text{k}\Omega)}{220\ \text{k}\Omega + 334.56\ \text{k}\Omega} = \mathbf{-39.67}$$

versus

$$A_i = -A_v \frac{Z_i}{R_E} = -(0.996)\left(\frac{132.72\ \text{k}\Omega}{3.3\ \text{k}\Omega}\right) = \mathbf{-40.06}$$

(f) Checking the condition $r_o \geq 10R_E$, we have

$$25\ \text{k}\Omega \geq 10(3.3\ \text{k}\Omega) = 33\ \text{k}\Omega$$

which is *not* satisfied. Therefore,

$$Z_b = \beta r_e + \frac{(\beta + 1)R_E}{1 + \dfrac{R_E}{r_o}} = (100)(12.61\ \Omega) + \frac{(100 + 1)3.3\ \text{k}\Omega}{1 + \dfrac{3.3\ \text{k}\Omega}{25\ \text{k}\Omega}}$$
$$= 1.261\ \text{k}\Omega + 294.43\ \text{k}\Omega$$
$$= 295.7\ \text{k}\Omega$$

with $Z_i = R_B \| Z_b = 220\ \text{k}\Omega \| 295.7\ \text{k}\Omega$

$$= \mathbf{126.15\ k\Omega}\ \text{vs. } 132.72\ \text{k}\Omega \quad \text{obtained earlier}$$
$$Z_o = R_E \| r_e = \mathbf{12.56\ \Omega} \quad \text{as obtained earlier}$$

$$A_v = \frac{(\beta + 1)R_E/Z_b}{\left[1 + \dfrac{R_E}{r_o}\right]} = \frac{(100 + 1)(3.3\ \text{k}\Omega)/295.7\ \text{k}\Omega}{\left[1 + \dfrac{3.3\ \text{k}\Omega}{25\ \text{k}\Omega}\right]}$$
$$= \mathbf{0.996} \cong 1$$

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. 8.23, with the commonbase *re* equivalent model substituted in Fig. 8.24.

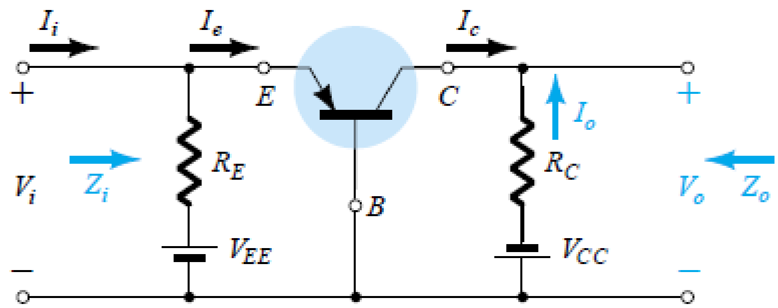


Figure 8.23 Common-base configuration.

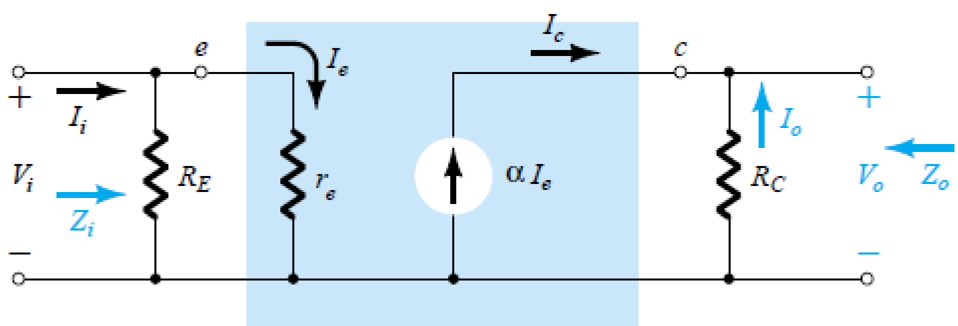


Figure 8.24 Substituting the r_e equivalent circuit into the ac equivalent network of Fig. 8.23.

***Z_i*:**

$$\boxed{Z_i = R_E \| r_e} \tag{8.54}$$

***Z_o*:**

$$\boxed{Z_o = R_C} \tag{8.55}$$

***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 $$\boxed{A_v = \frac{V_o}{V_i} = \frac{\alpha R_C}{r_e} \cong \frac{R_C}{r_e}}$$

A_i: Assuming that $R_E \gg r_e$ yields

$$I_e = I_i$$

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

with $$\boxed{A_i = \frac{I_o}{I_i} = -\alpha \cong -1}$$

- **Phase relationship:** The fact that *Av* is a positive number reveals that *Vo* and *Vi* are in phase for the common-base configuration.
- **Effect of *ro*:** For the common-base configuration, *ro* = 1/*hob* is typically in the megohm range and sufficiently larger than the parallel resistance *RC* to permit the approximation *ro*‖*RC* ≈ *RC*.

EXAMPLE:-

For the network of Fig. 8.25, determine:

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

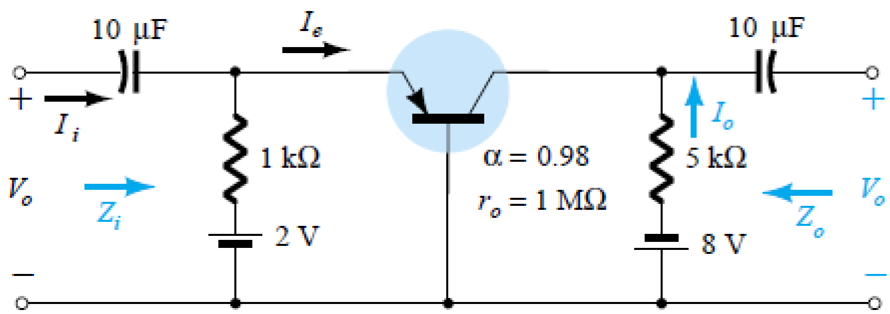


Figure 8.25 Example 8.8.

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) $A_i = \mathbf{-0.98} \cong -1$