The AC equivalent circuit of the BJT amplifier

Is obtained according the following steps as shown in Figs. (10-5a, 10-5b, 10-5c): -

a) Setting all dc sources to zero and replacing then by short cct. equivalent.
b) Replacing all capacitors by a short cct.
c) Remove all elements by pass by the short cct. equivalents introduced by steps a and b.
d) Redrawing the circuit.

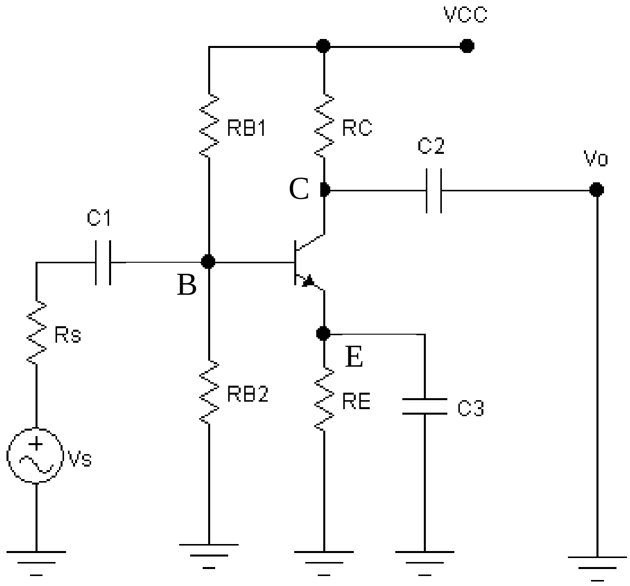


Fig.(10-5a) transistor circuit under examination.

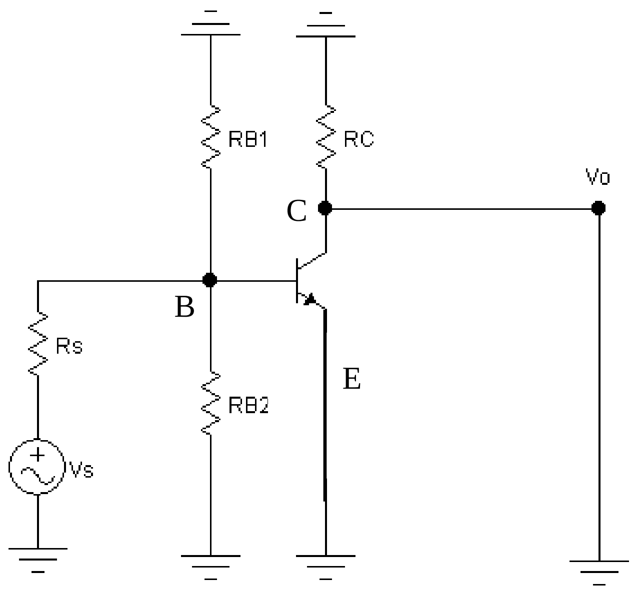


Fig.(10-5b) the network of Fig.(10-5a) following the removal of the dc

supply and inserting short cct. for capacitor.

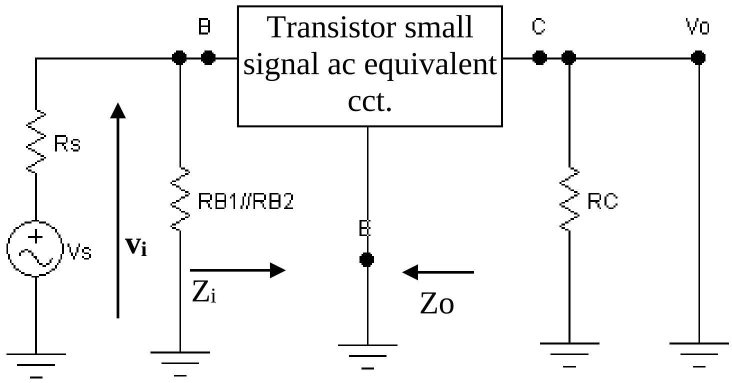


Fig.(10-5c) circuit of fig(10-1b) redrawn for small ac signal.

NOTES

1- **The hybrid parameter equivalent cct continuous to be very popular.**
2- **Manufacturers continue to specify the hybrid parameters for particular operating region on their specification sheets.**
3- **The parameter of the re-model can be derived directly from the hybrid parameters the hybrid equivalent cct suffer from being limited to a particular set of operating condition.**
4- **The hybrid equivalent cct suffer from being limited to a particular set of operating conditions.**
5- **The parameter of the other equivalent cct (re-model can be determined for any region of operation within the active region and are not limited by the single set of parameters provided by the specification sheet.**

Transistor hybrid equivalent cct (h-parameter)

- For the basic three terminal electronic device or system. It is obvious, from fig(10-6), that there are two ports(pair of terminal) interest . For our purpose the set at the left will represent the input terminal and the set at the right, the output terminal.
-For each set of terminals, there are two variables of interest.
-For the general hybrid two-port system of fig (10-6):-

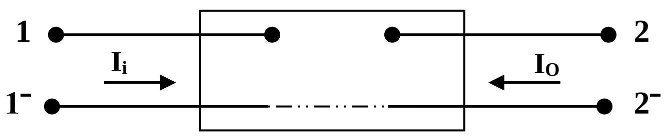


Fig. (10-6).

$V_i = h_{11}I_i + h_{12}V_o$…………..(10-12a)
$I_o = h_{21}I_i + h_{22}V_o$ …………..(10-12b)

The parameters relating the four variables are called h-parameters from the word hybrid. The term hybrid was chosen because the mixture of variables (v and i)

$h_{11} = \left.\frac{V_i}{I_i}\right|_{V_o=0}$ $=h_i$ (Ω), short cct input impedance parameter (10-13a)

$h_{12} = \left.\frac{V_i}{V_o}\right|_{I_i=0}$ = h_r (unitless), open cct reverse transfer voltage ratio parameter

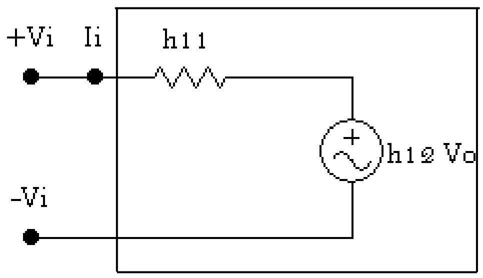
..(10-13b)

$h_{21} = \left.\frac{I_o}{I_i}\right|_{V_o=0}$ $=h_f$ (unitless), short cct. forward transfer current ratio

parameter …(10-13c)

$h_{22} = \left.\frac{I_o}{V_o}\right|_{I_i=0}$ = h_o (S), open cct. admittance parameter …(10-13d)

-Since each term of eq(10-12a) has the units of volt, let us now apply Kirchhoff's voltage law in reverse to obtain the circuit of fig (10-7).
-Since each term of eq(10-12b) has the units of volt, let us now apply Kirchhoff's voltage law in reverse to obtain the circuit of fig (10-8).



Fig(10-7) hybrid input equivalent cct

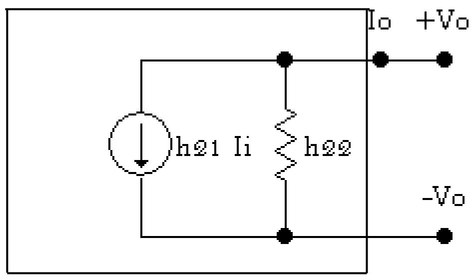
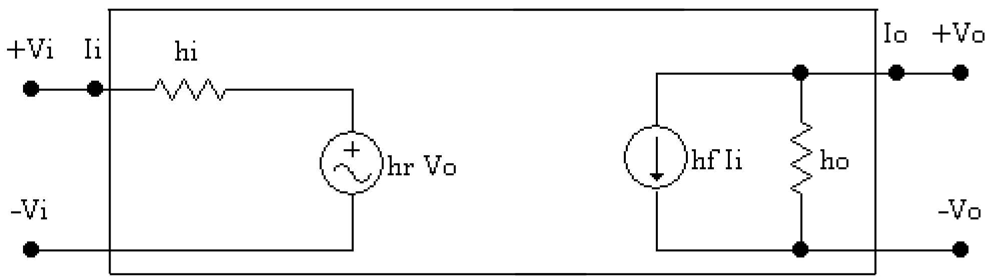


Fig (10-8) hybrid output equivalent cct

-The complete ac equivalent cct for the basic three terminal linear device is shown in fig (10-9).



Fig(10-9) complete hybrid equivalent cct.

From the BJT hybrid equivalent cct of fig (10-9) Eqs.(10-12a) and (10-12b) becomes

$V_i = h_i I_i + h_r V_o$ …….(10-14a)

$I_o = h_f I_i + h_o V_o$ ……..(10-14b)

- The circuit of fig (10-9) is applicable to any three- terminal electronic device or system with no internal independent sources.
- For the transistor, it has three basic configurations they are all three terminal configurations, so that the resulting equivalent cct will have the same format as shown in Fig. (10-9).
- In each case the bottom of the input and output section of the network of fig (10-9) can be connected as shown in Fig. (10-10), since the potential level is the same. The hybrid equivalent network for the C.E is shown in fig (10-10).Note that:

$I_i=I_b$, $I_o=I_c$, $V_i=V_{be}$, $V_o=V_{ce}$

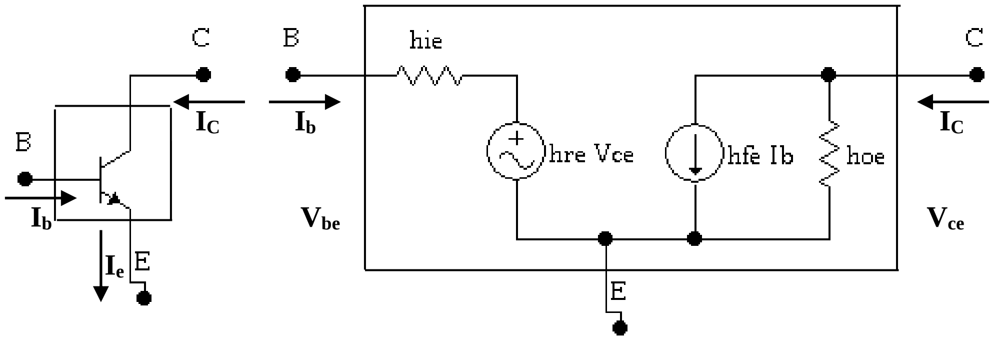


Fig. (10-10) C.E configuration.

For C.B hybrid equivalent network is shown in fig(10-11) .Note that:

$I_i=I_e$, $I_o=I_c$, $V_i=V_{eb}$, $V_o=V_{cb}$

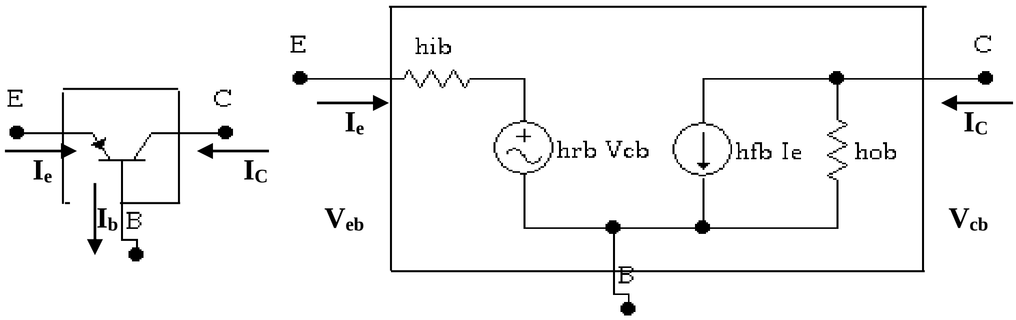


Fig. (10-11) C.B configuration.

Types of hybrid parameters

Since there are three possible configurations for BJ|T, there are three different sets of h-parameters.

A second subscript has been added to the h-parameters

h_{ie} , h_{fe} , h_{re} , h_{oe} : C.E h-parameters

h_{ic} , h_{fc} , h_{rc} , h_{oc} : C.C h-parameters

h_{ib} , h_{fb} , h_{rb} , h_{ob} : C.B h-parameters

- If all of h-parameter values in one configuration are known, then the values corresponding to any other configuration can be determined
- The C.E values are the ones most often given.

- **Table below lists typical parameter values in each of the three-transistor configuration.**

h-parameter	C.E	C.B	C.C
h_i	1400Ω	14Ω	1400Ω
h_r	$2*10^{-4}$	$4*10^{-5}$	1
h_f	100	-0.99	-101
h_o	$2*10^{-5}$S	$2*10^{-7}$S	$2*10^{-5}$S

Approximate C.E and C.B hybrid equivalent cct

Since h_{re} , h_{rb} are normally a relatively small quantity ,their removal are approximate by h_{re}=0 , h_{rb}=0 and $h_{re}V_o$=0 , $h_{rb}V_o$=0 resulting a short cct for the feedback element as shown in fig(10-12a).The resistance determined by 1/ h_{oe} and 1/ h_{ob} are large enough to be ignored ,in comparison to a parallel load ,which can be replaced by an open cct. as shown in fig(10-12b).

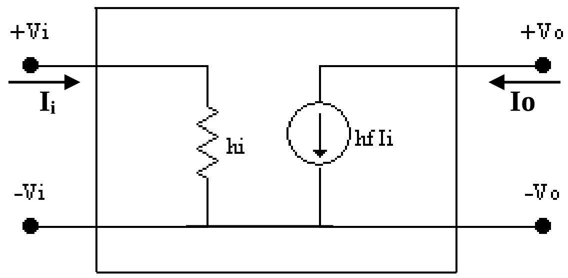


Fig.(10-12b)

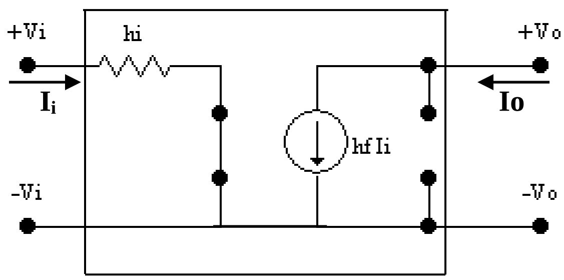


Fig.(10-12a)

For the C.E configuration the approximate equivalent model will appear as shown in fig(10-13), it is obvious that

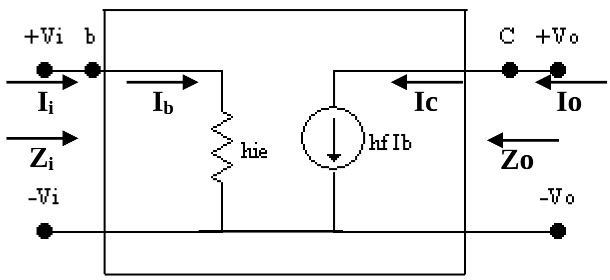


Fig. (10-13) approximate C.E hybrid equivalent model

$$I_i = I_b$$

$$I_o = I_c = I_{fe} I_b$$

$$Z_i = h_{ie}$$

$$A_i = \frac{I_c}{I_b} = h_{fe}$$

$$A_v = \frac{V_o}{V_i} = \frac{V_o}{I_b h_{ie}}$$

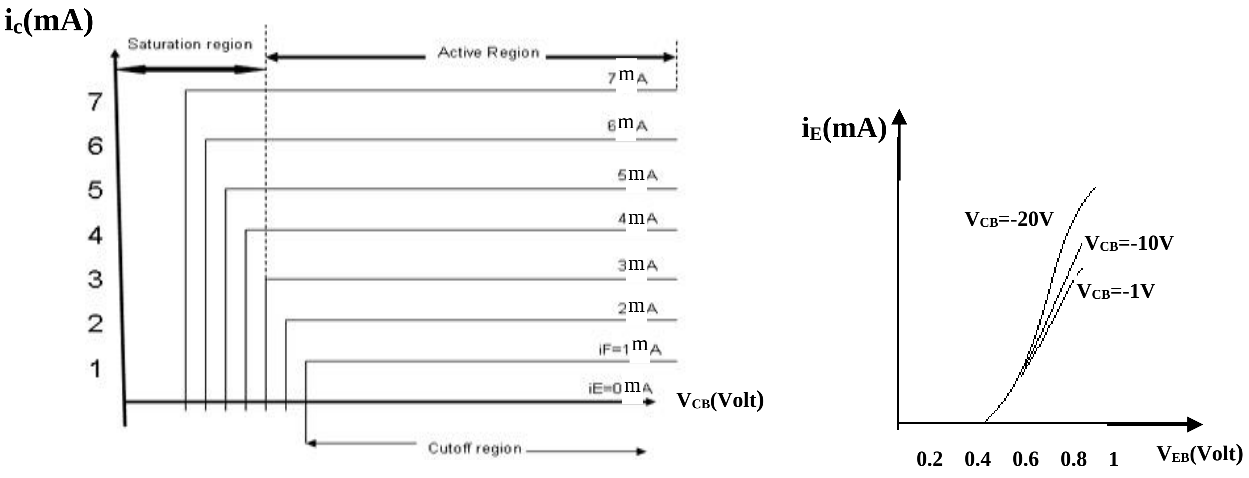
The re model

The re model derived below will permit the determination of an equivalent cct. using the dc operating point condition of the network (but the hybrid parameters specified at a particular operating point).

1- C.B transistor configuration

The derivation of the alternate equivalent cct. Begin with approximation of the input and output of the C.B transistor configuration, as redrawn in Fig. (10-14).

Note that straight line segments are used to represent the collector characteristics and a single diode characteristic for the emitter cct. resulting in equivalent cct.



O/P characteristics **I/P characteristics**

Fig. (10-14) approximate C.B characteristics

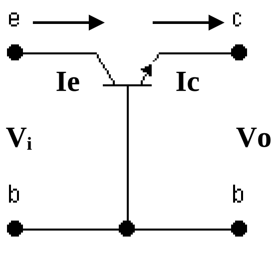
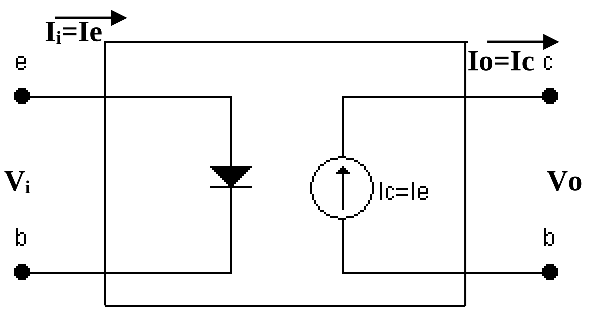


Fig. (10-15a) C.B configuration



Fig(10-15b) approximate C.B equivalent cct as defined by Fig. (10-14)

For ac condition, the input at the emitter of the C.B transistor can be determined using the dynamic resistance of the diode, which can be obtained by using the following equation

$re = \frac{26mV}{I_e}$ **ohms……………..(10-15)**

Substituting re will result in the re model of the C.B configuration as shown in Fig. (10-16).

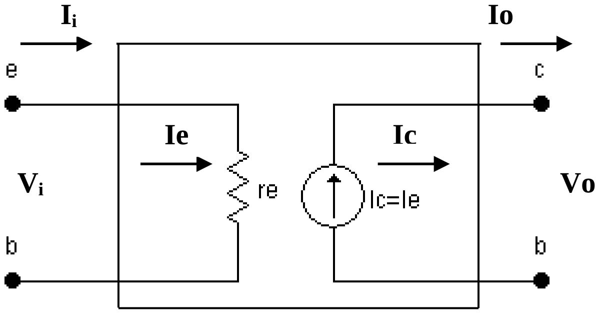


Fig. (10-16)

Note the similarities of the re model of fig (10-16) with approximate hybrid equivalent model of fig (10-12) for C.B.

A comparison of the two models (h model and re model) shows that:

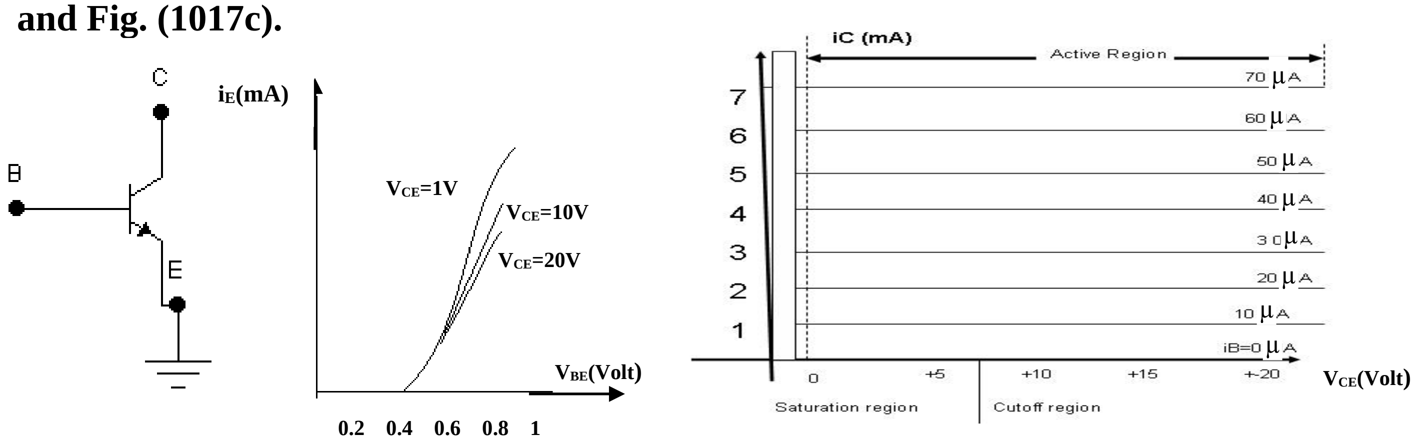
$$h_{ib} = re \quad \text{………………(10-16a)}$$
$$h_{fb} = -1 \quad \text{……………....(10-16b)}$$

For the C.B model of Fig. (10-16) the following parameters are defined:

$I_i = I_e$, $Z_i = re\,\Omega$, $Z_o = \infty\,\Omega$, $I_o = I_c = I_e$

2- C.E transistor configuraion

For C.E configuration appearing in Fig. (10-17a) the input and output characteristics have been approximated by the set appearing in Fig. (10-17b) and Fig. (1017c).



[a] [b] [c]

Fig. (10-17): [a] C.E configuration [b] I/P characteristic [c] O/P characteristic

The base characteristics are approximated to be those of a diode and

$r_{ac} = \frac{26mV}{IB}$ **……………(10-17)**

But $I_E = I_C = \beta I_B$ **and** $I_B = \frac{IE}{\beta}$

$r_{ac} = \frac{26mV}{IB} = \frac{26mV}{IE/\beta} = \beta\frac{26mV}{IE}$

$r_{ac}=\beta r_e$ ……………(10-18)

The common-emitter r_e model is shown in fig(10-18)

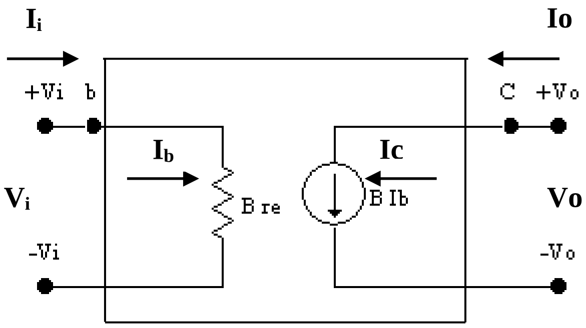


Fig. (10-18) CE r_e model

A comparison of the two models (hybrid and r_e) shows that:

$\beta=h_{fe}$ ……………..….(10-19a)

$\beta r_e=h_{ie}$. …………….(10-19b)

From fig(10-18)

$I_i=I_b$, $I_o=I_c=\beta I_b$

$Z_i=\beta r_e$, $Z_o=\infty\ \Omega$

BJT small signal analysis

1- Common-emitter fixed-bias configuration

The CE fixed bias amplifier is shown in Fig. (11-8). The small signal analysis begins by removing the dc effects of V_{cc} and replacing the dc blocking capacitors C1 and C2 by short cct. equivalent, resulting in the network of Fig. (11-9).

Note in Fig. (11-9) that the common ground of the dc supply and the emitter the relocation of R_B and R_C in parallel with the input and output section of the transistor.

Substituting the approximate hybrid small-signal equivalent cct. for the transistor of Fig. (11-9) will result in network of Fig. (11-10)

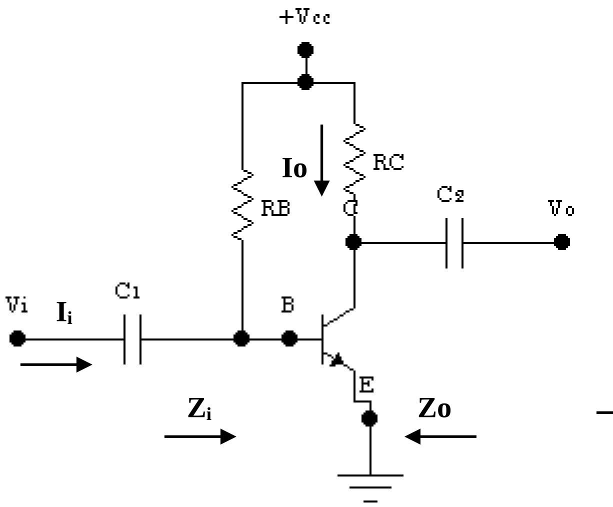


Fig. (11-8)

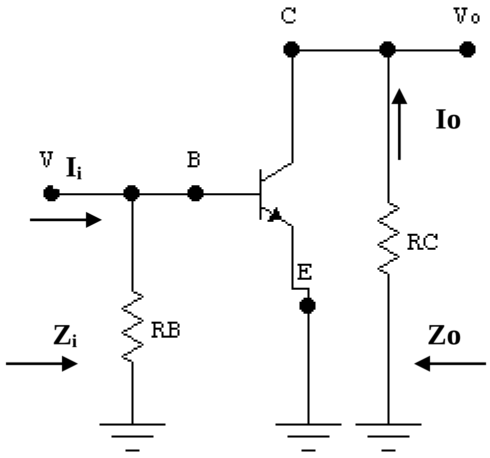


Fig. (11-9)

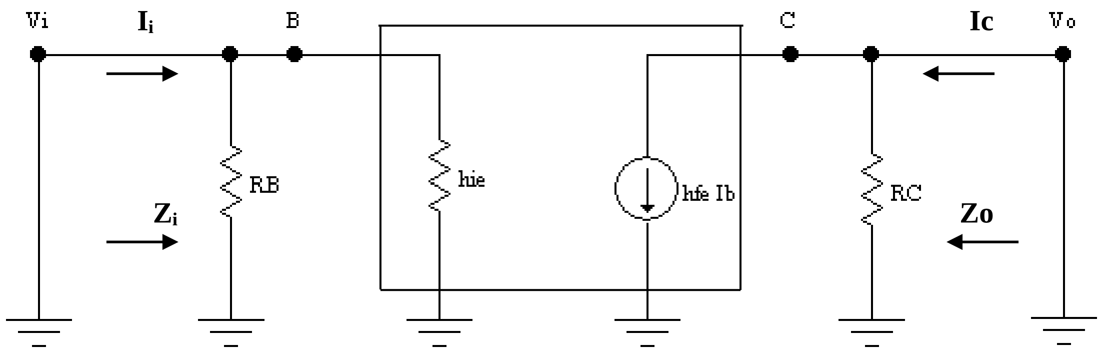


Fig. (11-10)

a- Z_i: Fig. (11-10) clearly shows that

$$Z_i = R_B // h_{ie} \quad \Omega \quad \ldots\ldots\ldots\ldots(11\text{-}3)$$

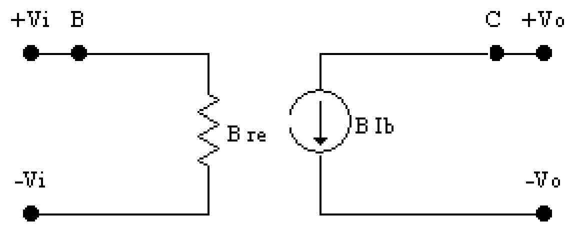
For situations where R_B is greater than hie by more than a factor 10:

$Z_i = h_{ie}$ if $R_B \geq 10\ h_{ie}$ ………..(11-4)

Using the r_e model equivalence will result in the following equation for Z_i

$Z_i = \beta r_e$ Ω ………….(11-5) Where $r_e = \frac{26mV}{I_E}$

CE r_e model



b- **Z_o: The output impedance of any network is defined when V_i=0.For Fig. (11-10) when V_i=0, I_i and therefore I_b=0 and $h_{fe}I_b$=0**

$Z_o = R_c$ Ω ….(11-6) for both the hybrid and r_e model.

c- **A_V: Voltage gain=$\frac{Vo}{Vi}$**

If $R_B \gg h_{ie}$ than $I_b \approx I_i$
$V_o = -I_oR_c = -h_{fe}I_bR_c = -h_{fe}I_iR_c$
But $I_i = \frac{Vi}{hie}$ than $V_o = -h_{fe}(\frac{Vi}{hie})R_c$
$A_V = \frac{Vo}{Vi} = -\frac{hfeRc}{hie}$

The negative sign in the resulting equation explains that an 180^0 phase shift occurs between the input and output signals. Substituting

$h_{fe} = \beta$
$h_{ie} = \beta r_e$ for r_e model
$A_V = -\frac{hfeRc}{hie} = -\frac{\beta Rc}{\beta re}$
$A_V = -\frac{Rc}{re}$ where $r_e = \frac{26mV}{I_E}$ ………..(11-8)

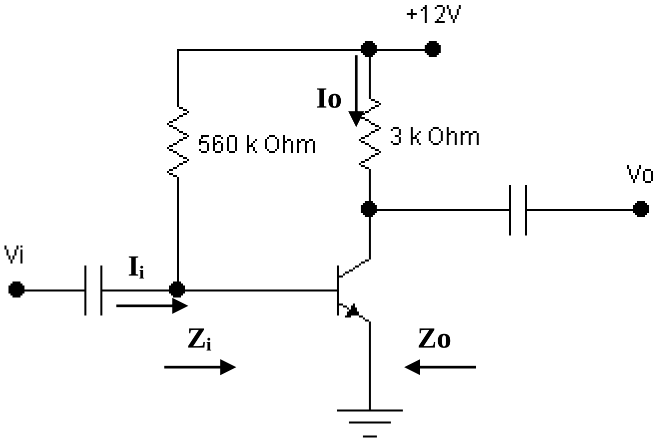
d- **A_i: Current gain=$\frac{Io}{Ii}$**

$I_o = h_{fe}I_b \approx h_{fe}I_i$

$A_i = \frac{Io}{Ii} \approx h_{fe}$
For the r_e model $A_i \approx \beta$
Note the relative simplicity of moving from one model to the other, simply recall that:

h_{fe}=β
h_{ie}=βr_e

<u>EX11-2:</u> Determine Z_i, Z_o, Av and A_i for the network of Fig. (11-1) using h and r_e models. When h_{fe}=100, h_{ie}=1300 Ω at I_C=2mA.



<u>Solution:</u>

Z_i : R_B =560kΩ >> h_{ie} =1300Ω

Z_i (E_q 11.4) Z_i =h_{ie} =1300Ω

Z_o (E 11.6) Z_o =R_c =3kΩ

A_v (E_q 11.7) A_v $= \frac{-h_{fe}R_c}{h_{ie}} = \frac{-(100)(3k\Omega)}{1.3k\Omega} = -230.77$

A_i (E_q 11.9) A_i $\approx h_{fe}$ **=100**

<u>re-model:</u>

DC analysis I_B = $\frac{V_{CC}-V_{BE}}{R_B} = \frac{12-0.7}{560k\Omega} = 20\mu A$

I_C =BI_B=(100)(20μA)=2μA=I_E

r_e $= \frac{26mv}{I_E} = \frac{26mv}{2mA} = 13\Omega$

Z_i [E_q 11.5] Z_i =Br_e =(100)(13)=1300Ω

Z_o[E_q 11.6] Z_o=R_c =3kΩ

A_v [E_q 11.8] A_v =- $\frac{R_c}{r_e} = \frac{3km}{13} = -230.77$

A_i [E_q 11.10] A_i ≈B=100

2-<u>Common-emitter voltage divider bias configuration</u>

The voltage divider bias configuration is show in Fig. (11.14). Substituting the approximate hybrid equivalent cct will result in the network of Fig. (11.15). Note the absence of R_E due to the low-impedance shorting effect of C_E. That is at the

frequency or (frequencies) of operation, the reactance of the capacitor is so small compared to R_E and the network it is treated as a short circuit across R_E.

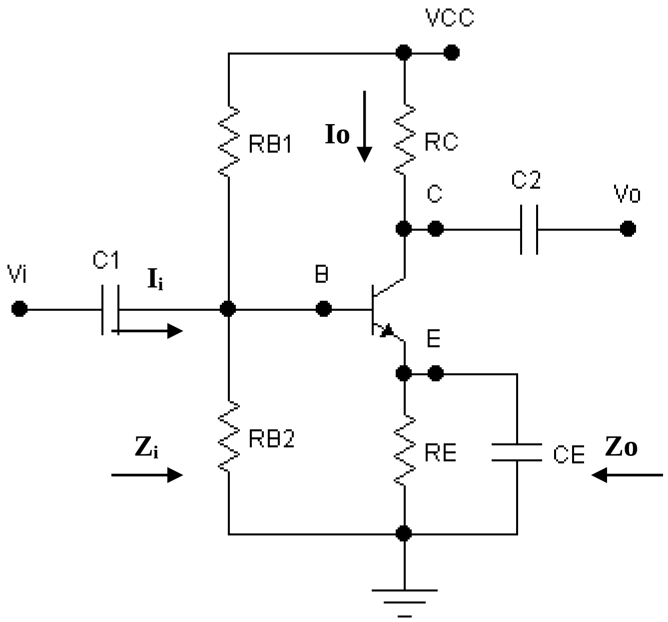


Fig. (11.14)

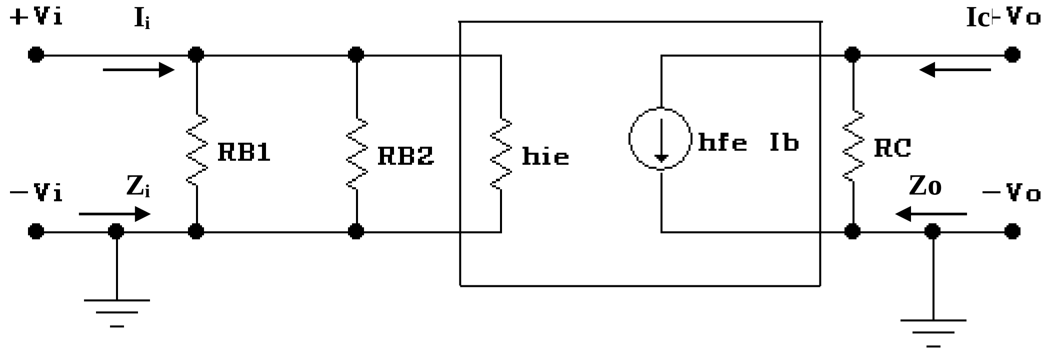


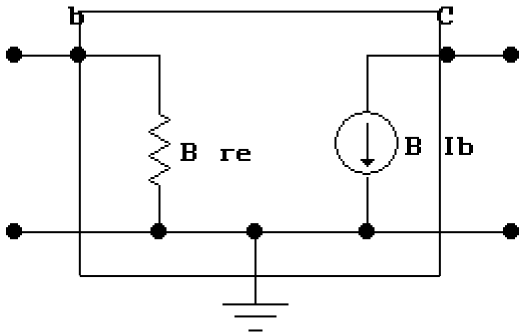
Fig. (11.15)

The parallel combination of R_{B1}, R_{B2} is defined by

$R_{BB}= R_{B1}\, R_{B2} / R_{B1}+R_{B2}$............(11.14)

a) Z_i: from Fig. (11.15)

$Z_i= R_{BB} // h_{ie}$ ....(11.15)



For re model

$Z_{i}= R_{BB}// Br_{e}$............(11.16)

b) Z_o: $Z_o= R_C$ for both model……….(11.17)

c) A_v : $V_o = -I_o R_C = - h_{fe}I_b R_C$

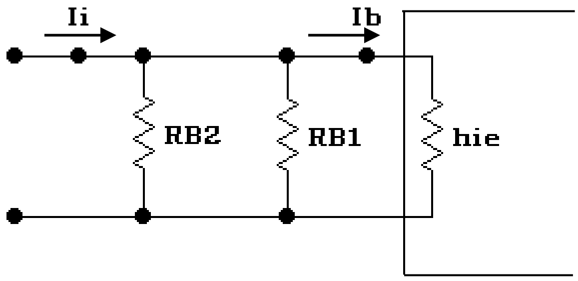
$V_o = - h_{fe} (V_i/ h_{ie}) R_C$

$A_v = V_o/ V_i = - (h_{fe}/ h_{ie}) R_C$ …………(11.18)

For re model

$V_o = -I_o R_C = - \beta I_b R_C = -(\beta V_i R_C/\beta re) = -V_i R_C /re$

$A_v = V_o/V_i = -R_C/re$ ………….(11.19)



A_i :

$$I_b = \frac{R_{BB} I_i}{R_{BB} + hie}$$

$$\therefore \frac{I_b}{I_i} = \frac{R_{BB}}{R_{BB} + hie}$$

But $I_o = hfeI_b$

$$\therefore \frac{I_o}{I_b} = hfe$$

The current gain $A_i = \frac{I_o}{I_i} = \frac{I_b}{I_i} \times \frac{I_o}{I_b}$

$$\therefore A_i = hfe\frac{R_{BB}}{R_{BB} + hie}$$

$$\therefore A_i = \frac{hfeR_{BB}}{R_{BB} + hie}$$ **………(11.20)**

If $R_{BB} >> hie$

$\therefore A_i \cong hfe$

For re-model

$$I_b = \frac{R_{BB} I_i}{R_{BB} + \beta re}$$

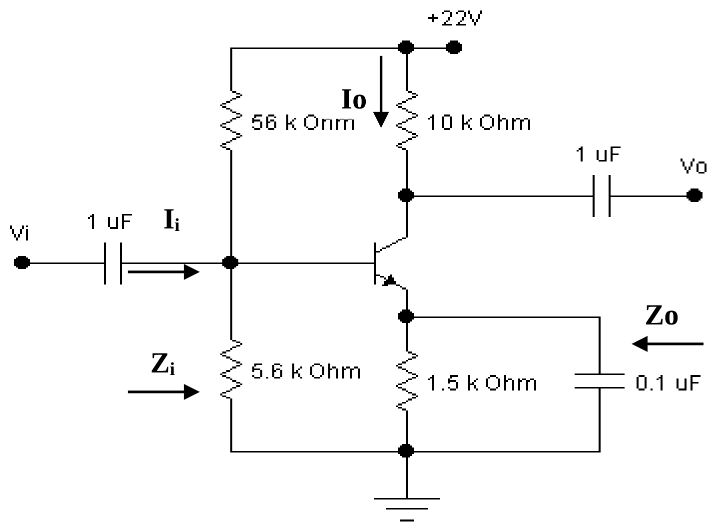
$$I_o = \beta I_b = \frac{\beta R_{BB} I_i}{R_{BB} + \beta re}$$

$hfe = \beta$

$hie = \beta re$

$$\therefore A_i = \frac{I_o}{I_i} = \frac{\beta R_{BB}}{R_{BB} + \beta re} \quad \text{............(11.21)}$$

EX 11.4: Determine Z_i, Z_o, Av and A_i for the network of Fig. (11-16) using re-model and h-model. if $\beta = hfe = 90$ **and** $hie = 2.835K\Omega$**.**



<u>Solution</u>

DC:

$$V_{BB} = \frac{R_{B2}}{R_{B1} + R_{B2}} V_{CC} = \frac{(5.6K\Omega)(22)}{56K\Omega + 5.6K\Omega} = 2V$$

$$R_{BB} = R_{B1} // R_{B2} = \frac{(56K\Omega)(5.6K\Omega)}{56K\Omega + 5.6K\Omega} = 5.09K\Omega$$

$$\boldsymbol{I_B = \frac{V_{BB} - V_{Be}}{R_{BB} + (1+\beta)R_E} = \frac{2 - 0.7}{5.09k\Omega + (1+90) \times 1.5K\Omega} = 9.18\mu A}$$

$$I_C = \beta I_B = 90 \times (9.18\mu A) = 0.836mA$$

$$re = \frac{26mV}{I_E} = 31.5\Omega$$

AC:

Z_i: from eq (11.16)

$$Z_i = R_{BB} // \beta re = \frac{5.09K\Omega(90 \times 31.5)}{5.09K\Omega + 90 \times 31.5} = 1.82K\Omega$$

Z_o: from eq (11.17)

$$Z_o = R_C = 10K\Omega$$

A_V: from eq (11.19)

$$A_V = -\frac{R_C}{re} = -\frac{10K\Omega}{31.5} = -317.5$$

A_i: from eq (11.21)

$$A_i = \frac{\beta R_{BB}}{R_{BB} + \beta re} = \frac{5.09K\Omega \times (90)}{5.09K\Omega + (90)\times(31.5)} = 57.8$$

3-CE unbypassed emitter – bias configuration

The network shown in Fig. (11.18) includes an emitter resistor unbypassed in the AC domain. substituting the approximate hybrid equivalent model will result in the configuration of Fig. (11.19).

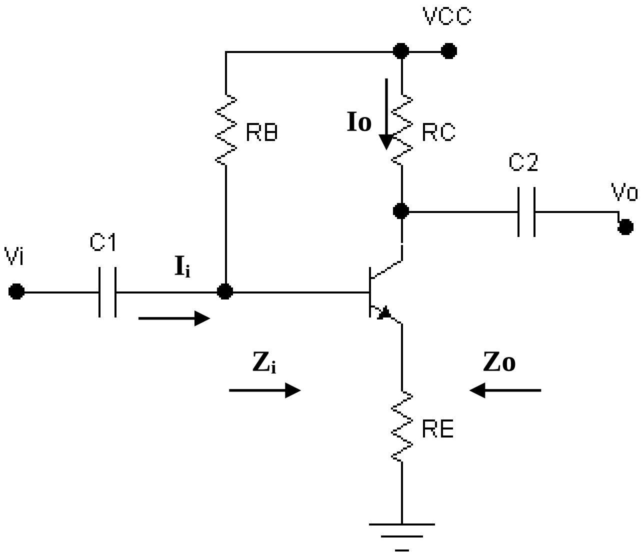


Fig. (11.18)

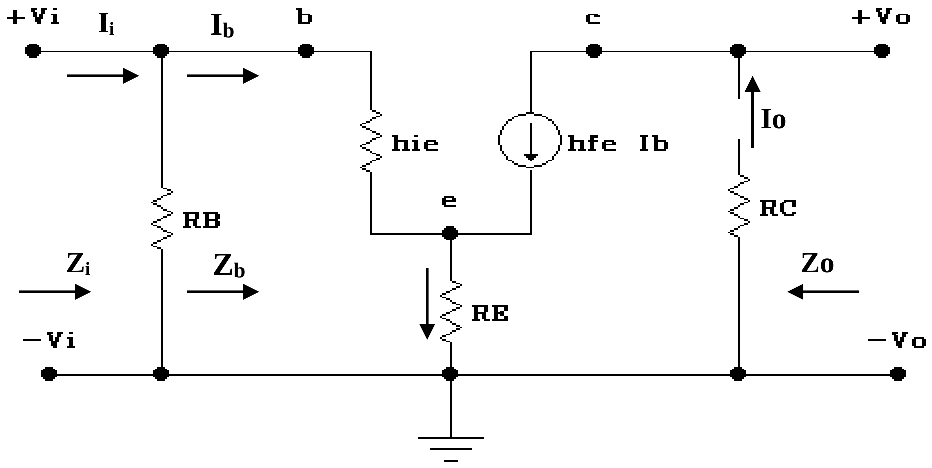


Fig. (11.19)

Applying Kirchoff's voltage law to the input side of fig (11.19) will result in

$V_i = I_b hie + R_E(1+hfe)I_b$

$Z_b = \frac{V_i}{I_b} = hie + R_E(1+hfe)$

The result as displayed in Fig. (11.19) described that the input impedance to a transistor with an unbypassed resistor R_E is determined by

$Z_b = hie + R_E(1+hfe)$ **…………(11.25)**

Since hfe is normally much greater then, the equation reduced to

$Z_b = hie + R_E hfe$**………….(11.26)**

But hfeR_E is much greater then hie

$\therefore Z_b = hfeR_E$ **………….(11.27)**

For the re-model , the following equation is normally applied

$Z_b = \beta(re + R_E)$ **…..(11.28)**

If $\beta >> 1 \therefore Z_b = \beta R_E$ **…..(11.29)**

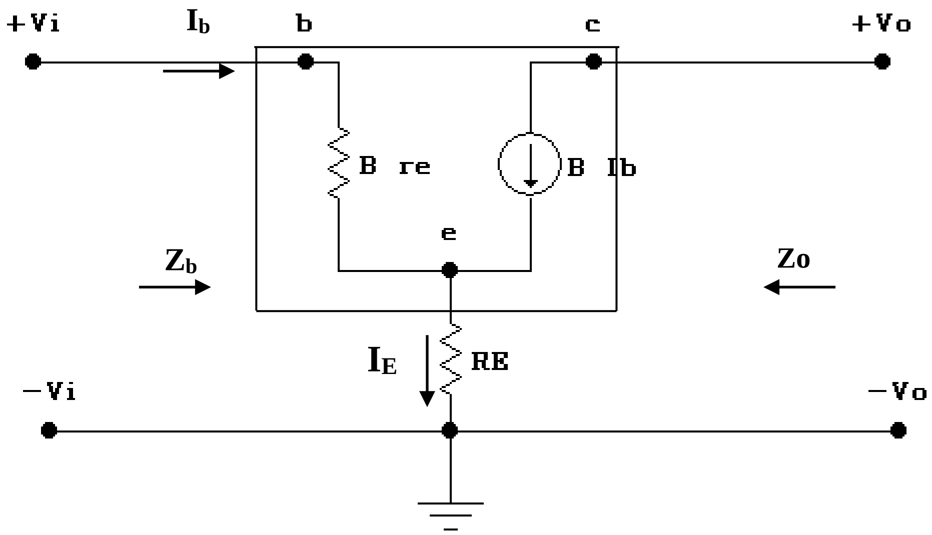


Fig. (11.20)

a) Z_i : returning to Fig. (11.19)

$Z_i = R_B // Z_b$ **…..(11.30)**

For both model

b) Z_o : with V_i set to zero ,I_b = 0 and hfeI_b can be replaced by an open circuit equivalent

$\therefore Z_o = R_C$ **…..(11.31)**

For both model

c) A_V:

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

$$V_o = -I_o R_C = -hfeI_b R_C$$

$$V_o = -hfe\frac{V_i}{Z_b}R_C$$

$\therefore A_V = \frac{V_o}{V_i} = -\frac{hfeR_C}{Z_b}$ **……(11.32)**

For the approximation $Z_b \cong hfeR_E$

$\therefore A_V = -\frac{hfeR_C}{hfeR_E} = -\frac{R_C}{R_E}$ **……(11.33)**

Which also applied to both models, due to the absence of any device parameters.

d) A_i: returning to Fig. (11.19)

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

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

Out $I_o = hfeI_b$

$$\therefore \frac{I_o}{I_b} = hfe$$

$$\therefore A_i = \frac{I_o}{I_i} = \frac{I_b}{I_i} \times \frac{I_o}{I_b} = \quad hfe\frac{R_B}{R_B + Z_b}$$

$\therefore A_i = \frac{I_o}{I_i} = \frac{hfeR_B}{R_B + Z_b}$ **…..(11.34)**

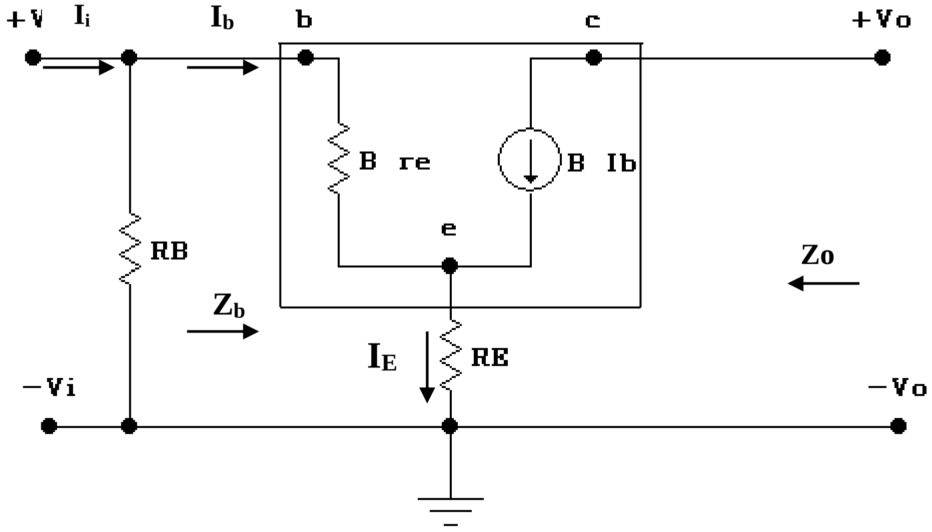
For re-model

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

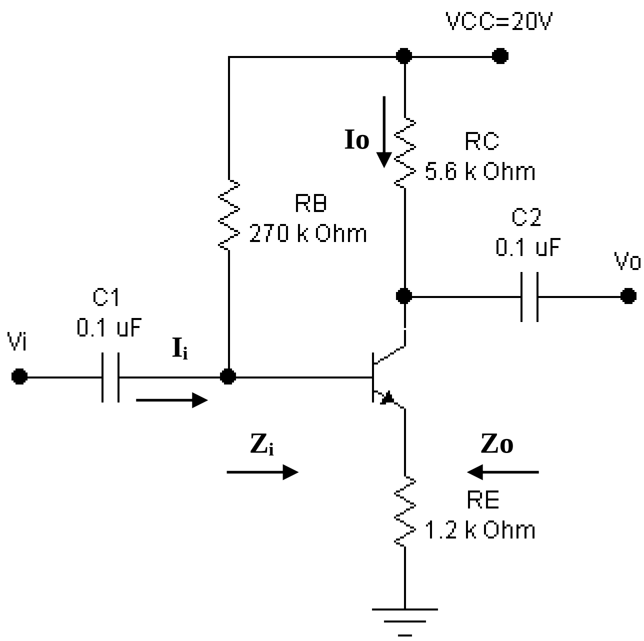
$$\therefore \frac{I_b}{I_i} = \frac{R_B}{R_B + Z_b}$$

$$I_o = \beta I_b \quad \therefore \beta = \frac{I_o}{I_b}$$

$$\therefore A_i = \frac{I_b}{I_i} \times \frac{I_o}{I_b} = \frac{\beta R_B}{R_B + Z_b}$$ **.......(11.35)**



EX 11.5 **Determine Z_i, Z_o, A_V and A_i for the network of Fig. (11.21). When hfe=120, h_{ie}=560Ω.**



Solution

Z_b from Eq. (11.25)

$Z_b = hie + R_E(1+hfe) = 0.56K\Omega + 1.2K\Omega(1+120) = 145.76K\Omega$

Z_i from Eq. (11.30)

$Z_i = Z_b // R_B = 145.76K\Omega // 270K\Omega = 94.66K\Omega$

Z_o from Eq. (11.31)

$Z_o = R_C = 5.6K\Omega$

A_V from Eq. (11.32)

$$A_V = -\frac{hfeR_C}{Z_b} = -\frac{(120)(5.6K\Omega)}{145.76K\Omega} = -4.61$$

A_i from Eq. (11.34)

$$A_i = \frac{R_B hfe}{R_B + Z_b} = \frac{(270K\Omega)(120)}{(270K\Omega)+(145.76K\Omega)} = 77.93$$

NOTE

Applied the approximate equation for Z_b, we find

$Z_b \cong hfeR_E = (120)(1.2K\Omega) = 144K\Omega$

$Z_i = R_B // Z_b = 270K\Omega // 144K\Omega = 93.91K\Omega$

$$A_V \cong -\frac{R_C}{R_E} = -\frac{5.6K\Omega}{1.2K\Omega} - 4.67$$

Which are very close to result (94.66KΩ, -4.61) obtain above

4-<u>EMITTER – FOLLOWER CONFIGURATION</u>

When the output is taken from the emitter terminal of the transistor as shown in Fig. (11-22), the network is referred to as an emitter follower. The emitter follower output voltage is in phase with input voltage. The fact that V_o follows the magnitude of V_i with an in-phase relationship. The voltage gains for emitter – follower $A_V \cong 1$ the most common emitter follower configuration as shown in Fig. (11.22), because the collector is grounded for ac analysis, it is actually a common – collector configuration. The emitter follower configuration is used for impedance at the input and a low impedance at the output, which is 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 approximate equivalent circuit to the network of Fig. (11.22) will result in the network of Fig. (11.23).

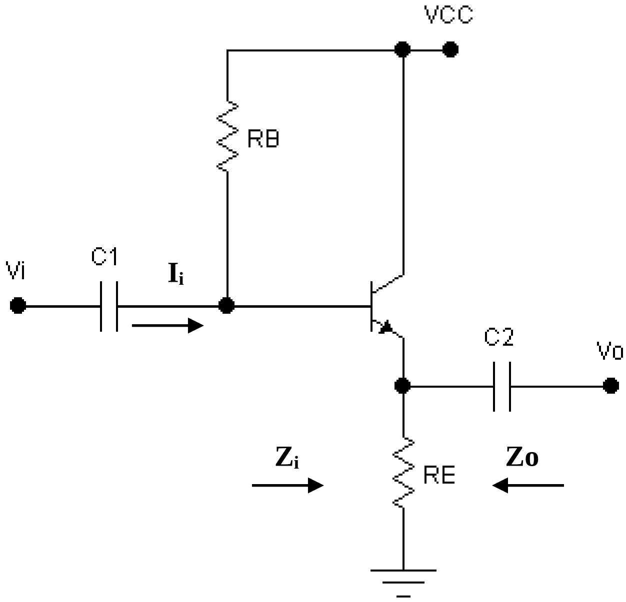


Fig. (11.22)

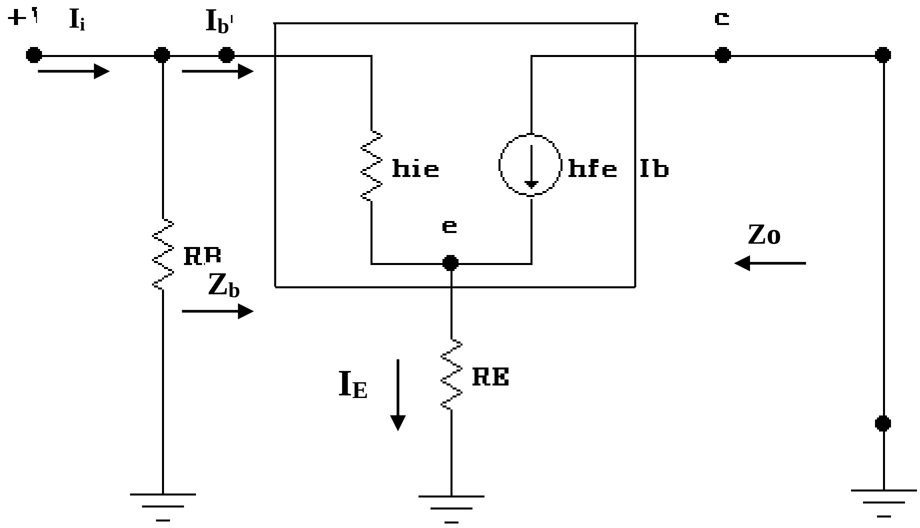


Fig. (11.23)

a) Z_i:

$Z_i = R_B // Z_b$ **…………..(11.36)**

With Z_b defined by Eqs (11.25) through (11.29)

$Z_b = hie + (1 + hfe)R_E$ **………(11.25)**

If hfe >>1

$\therefore Z_b = hie + hfeR_E$ **……….(11.26)**

If $hfeR_E$ >> hie

$\therefore Z_b = hfeR_E$ **……….(11.27)**

For the re-model

$Z_b = \beta(re + R_E)$ **………..(11.28)**

If $\beta >> 1$

$\therefore Z_b = \beta R_E$ **……….(11.29)**

b) Z_o: $I_b = \frac{V_i}{Z_b}$

$$I_e = I_b + hfeI_b = I_b(1 + hfe)$$

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

Substituting for Z_b

$$I_e = \frac{(1 + hfe)Vi}{hie + (1 + hfe)R_E}$$

$$I_e = \frac{Vi}{\frac{hie}{1 + hfe} + R_E}$$

If now construct the network defined by Eq. (11.37) the configuration of Fig. (11.24) will result: -

To determine Z_o, V_i is set to zero

$Z_o = R_E // \frac{hie}{1 + hfe}$ **………(11.38)**

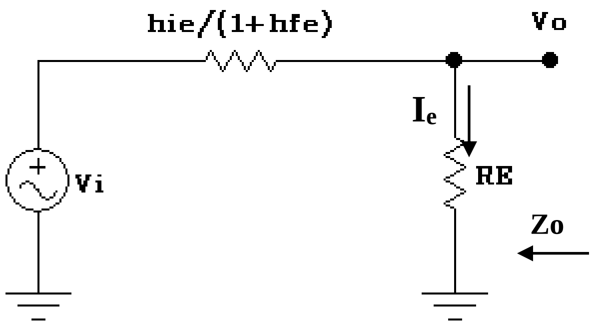


Fig. (11.24)

The network seen by the emitter branch is the input voltage V_i in series with a resistance determined by the hybrid parameter hie and hfe. The resistance ($\frac{hie}{1 + hfe}$) is usually quite small, dropping Z_o well below the R_E level.

c) A_V: Fig. (11.24) can be used to determine A_V

$$V_o = I_e R_E = \frac{R_E V_i}{R_E + (\frac{hie}{1 + hfe})}$$

$$\therefore A_V = \frac{V_o}{V_i} = \frac{R_E}{R_E + (\frac{hie}{1+hfe})}$$ **……….(11.39)**

Note the absence of a negative sign to indicate that V_o and V_i are in phase and recognize that the factor ($\frac{hie}{1+hfe}$) is the only reason V_o dose not equal V_i.

d) A_i: from Fig. (11.23)

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

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

$$I_o = I_e = (1+hfe)I_b$$

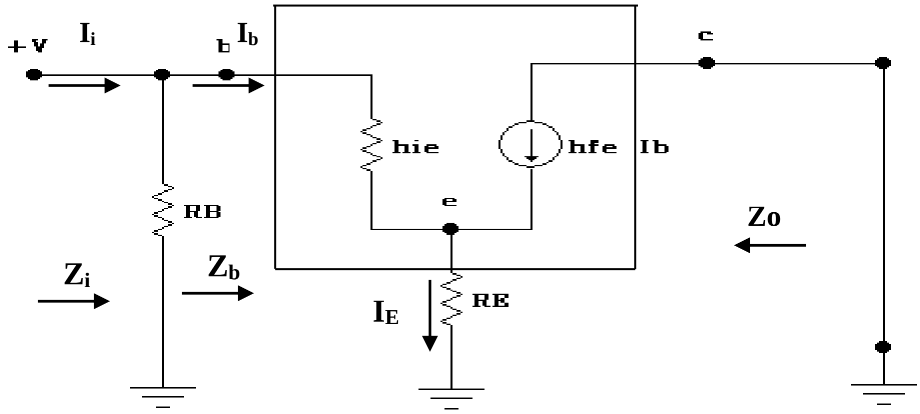
$$\therefore \frac{I_o}{I_b} = 1 + hfe$$

$$\therefore A_i = \frac{I_o}{I_i} = \frac{I_b}{I_i} \times \frac{I_o}{I_b} = (1+hfe)\frac{R_B}{R_B + Z_b}$$

$$\therefore A_i = \frac{(1+hfe)R_B}{R_B + Z_b}$$ **……...(11.40)**

re-model

The equation for the re-model can be determine directly from the above simply by substituting $hie = \beta re$ **and** $hfe = \beta$



a) Z_i

$$Z_i = R_B / Z_b$$

$$Z_b = \beta re + \beta R_E$$

$$Z_b = \beta(re + R_E)$$ **if** $\beta >> 1$ **……….(11.41)**

b) Z_o

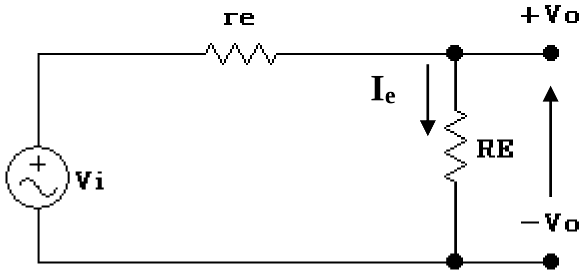
$$I_b = \frac{V_i}{Z_b}$$

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

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

$$I_e \ \frac{V_i}{\frac{\beta(re+R_E)}{1+\beta}} = \frac{V_i}{\beta(re+R_E)}$$ **……….(11.42)**

If now construct the network defined by Eq. (11.42) will result: -



To determine Z_o, V_i set to zero

$\therefore Z_o = R_E \,//\, re$ **……...(11.43)**

c) A_V:

$$V_o = I_o R_E = \frac{R_E V_i}{re+R_E}$$

$$\therefore A_V = \frac{V_o}{V_i} = \frac{R_E}{re+R_E}$$ **………(11.44)**

d) A_i:

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

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

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

$$\therefore \frac{I_o}{I_b} = 1+\beta$$

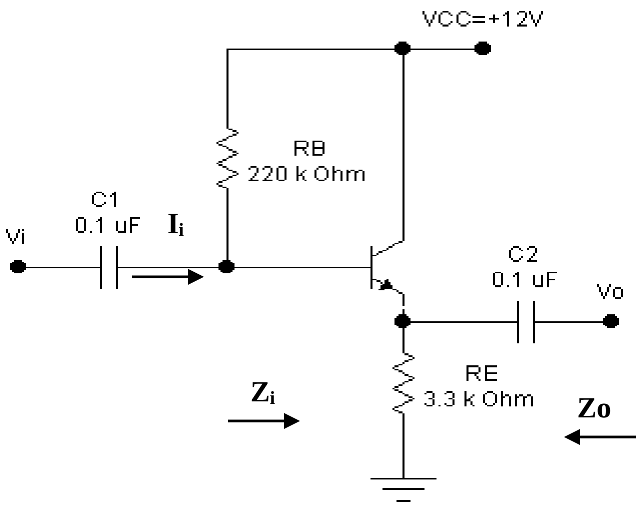
$$\therefore A_i = \frac{I_b}{I_i} \times \frac{I_o}{I_b} = \frac{R_B(1+\beta)}{R_B + Z_b}$$

$$\therefore A_i = \frac{\beta R_B}{R_B + Z_b}$$ **……...(11.45)**

EX 11.6: For the emitter follower network. When hfe=98, h_{ie}=1275 Ω.

a) Find expressions for Z_i, Z_o, A_V and A_i using h parameter.

b) Determine Z_i, Z_o, A_V and A_i.



(11.25)

Solution

$Z_b = hie + (1+hfe)R_E = 1.275K\Omega + (1+98)\times 3.3K\Omega = 327.98K\Omega$

Z_i:

$Z_i = R_B // Z_b = 220K\Omega // 327.98K\Omega = 131.68K\Omega$

Z_o:

$$Z_o = R_E // \frac{hie}{1+hfe} = 3.3K\Omega // \frac{1275\Omega}{1+98} = 12.9\Omega$$

A_V:

$$\therefore A_V = \frac{R_E}{R_E + (\frac{hie}{1+hfe})} = \frac{3300}{3300+12.9} = 0.996 \cong 1$$

A_i:

$$\therefore A_i = \frac{(1+hfe)R_B}{R_B + Z_b} = \frac{(1+98)\times 220K\Omega}{220K\Omega + 327.98K\Omega} = 39.75$$