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Department of Electronics and Communication Engineering

Course Name & Code: Analog Electronic Circuits & 17EC03

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Multistage Amplifiers

In practical applications, the output of a single state amplifier is usually insufficient, though it is a voltage or power amplifier. Hence they are replaced by **Multi-stage transistor amplifiers**.

In Multi-stage amplifiers, the output of first stage is coupled to the input of next stage using a coupling device. These coupling devices can usually be a capacitor or a transformer. This process of joining two amplifier stages using a coupling device can be called as **Cascading**.

The following figure shows a two-stage amplifier connected in cascade.

The overall gain is the product of voltage gain of individual stages.

$$A_V = A_{V1} \times A_{V2} = (V_2 / V_1) \times (V0 / V2) = V0/V1$$

Where A_V = Overall gain, A_{V1} = Voltage gain of 1st stage, and A_{V2} = Voltage gain of 2nd stage.

If there are **n** number of stages, the product of voltage gains of those **n** stages will be the overall gain of that multistage amplifier circuit.

Types of Coupling:

Resistance-Capacitance Coupling:

This is the mostly used method of coupling, formed using simple **resistor-capacitor** combination. The capacitor which allows AC and blocks DC is the main coupling element used here.

The coupling capacitor passes the AC from the output of one stage to the input of its next stage. While blocking the DC components from DC bias voltages to effect the next stage.

Impedance Coupling:

The coupling network that uses **inductance** and **capacitance** as coupling elements can be called as Impedance coupling network.

In this impedance coupling method, the impedance of coupling coil depends on its inductance and signal frequency which is **jwL**. This method is not so popular and is seldom employed.

Transformer Coupling

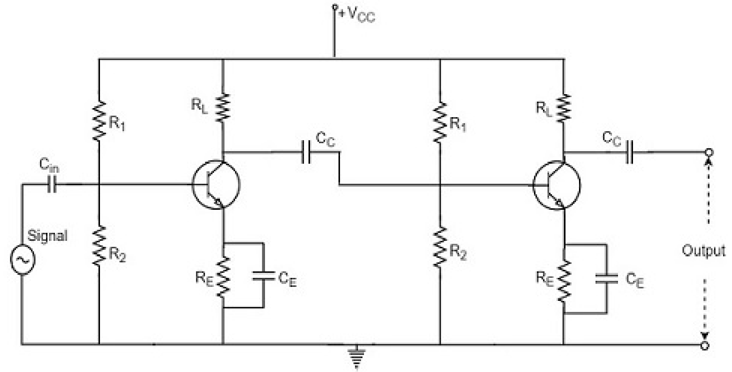
The coupling method that uses a **transformer as the coupling** device can be called as Transformer coupling. There is no capacitor used in this method of coupling because the transformer itself conveys the AC component directly to the base of second stage.

The secondary winding of the transformer provides a base return path and hence there is no need of base resistance. This coupling is popular for its efficiency and its impedance matching and hence it is mostly used.

Direct Coupling:
If the previous amplifier stage is connected to the next amplifier stage directly, it is called as **direct coupling**. The individual amplifier stage bias conditions are so designed that the stages can be directly connected without DC isolation.
The direct coupling method is mostly used when the load is connected in series, with the output terminal of the active circuit element. For example, head-phones, loud speakers etc.

Amplifier Consideration
For an amplifier circuit, the overall gain of the amplifier is an important consideration. To achieve maximum voltage gain, let us find the most suitable transistor configuration for cascading.
CC Amplifier
Its voltage gain is less than unity.
It is not suitable for intermediate stages.
CB Amplifier
Its voltage gain is less than unity.
Hence not suitable for cascading.
CE Amplifier
Its voltage gain is greater than unity.
Voltage gain is further increased by cascading.
The characteristics of CE amplifier are such that, this configuration is very suitable for cascading in amplifier circuits. Hence most of the amplifier circuits use CE configuration.

Two-stage RC Coupled Amplifier



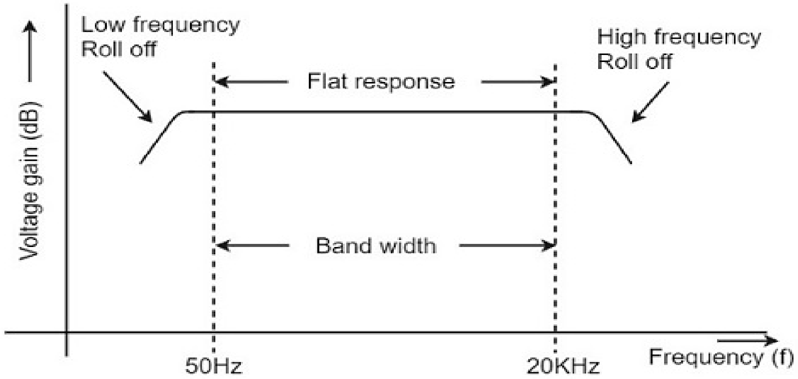
Two CE amplifiers are connected using R & C elements as shown in fig. This type of coupling is known as RC coupled amplifier.

When an AC input signal is applied to the base of first transistor, it gets amplified and appears at the collector load R_L which is then passed through the coupling capacitor C_C to the next stage. This becomes the input of the next stage, whose amplified output again appears across its collector load. Thus the signal is amplified in stage by stage action.

The important point that has to be noted here is that the total gain is less than the product of the gains of individual stages. This is because when a second stage is made to follow the first stage, the **effective load resistance** of the first stage is reduced due to the shunting effect of the input resistance of the second stage. Hence, in a multistage amplifier, only the gain of the last stage remains unchanged.

As we consider a two stage amplifier here, the output phase is same as input. Because the phase reversal is done two times by the two stage CE configured amplifier circuit.

Frequency Response of RC Coupled Amplifier



From the above graph, it is understood that the frequency rolls off or decreases for the frequencies below 50Hz and for the frequencies above 20 KHz. whereas the voltage gain for the range of frequencies between 50Hz and 20 KHz is constant.

We know that, $X_C = 1/2\pi fc$

At low freq. gain reduces because of coupling capacitor present in ckt. At high gain reduces because of junction capacitances.

Advantages of RC Coupled Amplifier:

The frequency response of RC amplifier provides constant gain over a wide frequency range, hence most suitable for audio applications.

The circuit is simple and has lower cost because it employs resistors and capacitors which are cheap.

It becomes more compact with the upgrading technology.

Disadvantages of RC Coupled Amplifier:

The voltage and power gain are low because of the effective load resistance.

They become noisy with age.

Due to poor impedance matching, power transfer will be low.

Applications of RC Coupled Amplifier:

They have excellent audio fidelity over a wide range of frequency.

Widely used as Voltage amplifiers

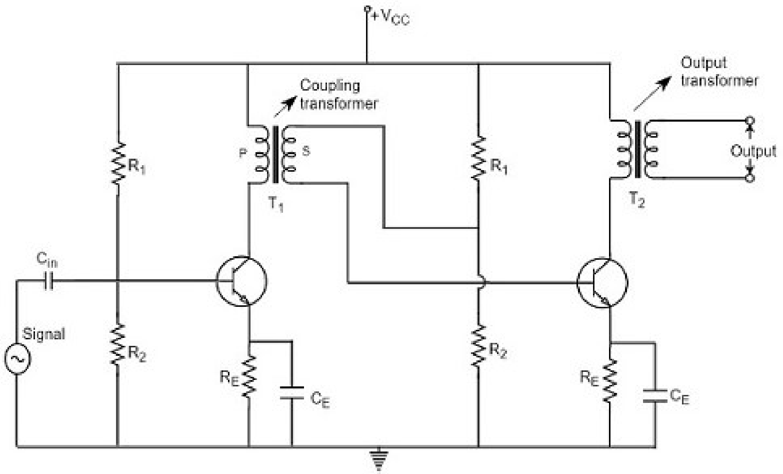
Due to poor impedance matching, RC coupling is rarely used in the final stages.

Transformer Coupled Amplifier

We have observed that the main drawback of RC coupled amplifier is that the effective load resistance gets reduced. This is because, the input impedance of an amplifier is low, while its output impedance is high.

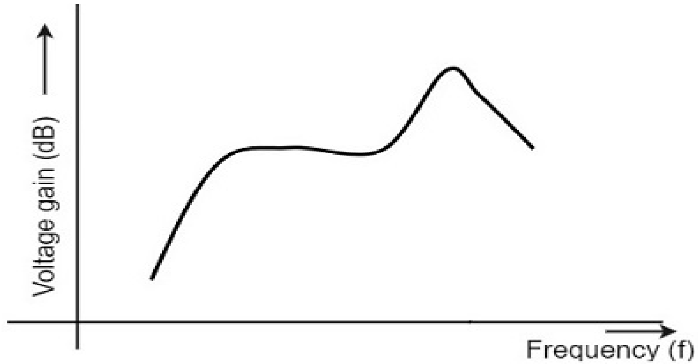
When they are coupled to make a multistage amplifier, the high output impedance of one stage comes in parallel with the low input impedance of next stage. Hence, effective load resistance is decreased. This problem can be overcome by a **transformer coupled amplifier**. In a transformer-coupled amplifier, the stages of amplifier are coupled using a transformer.

This transformer coupling provides good impedance matching between the stages of amplifier. The transformer coupled amplifier is generally used for power amplification.



The transformer which is used as a coupling device in this circuit has the property of impedance changing, which means the low resistance of a stage (or load) can be reflected as a high load resistance to the previous stage. Hence the voltage at the primary is transferred according to the turns ratio of the secondary winding of the transformer.

Frequency Response of Transformer Coupled Amplifier



The gain of the amplifier is constant only for a small range of frequencies. The output voltage is equal to the collector current multiplied by the reactance of primary.

At low frequencies, the reactance of primary begins to fall, resulting in decreased gain. At high frequencies, the capacitance between turns of windings acts as a bypass condenser to reduce the output voltage and hence gain.
So, the amplification of audio signals will not be proportionate and some distortion will also get introduced, which is called as **Frequency distortion**.

Applications
The following are the applications of a transformer coupled amplifier −
Mostly used for impedance matching purposes.
Used for Power amplification.
Used in applications where maximum power transfer is needed.

Advantages of Transformer Coupled Amplifier

The following are the advantages of a transformer coupled amplifier −

An excellent impedance matching is provided.
Gain achieved is higher.
There will be no power loss in collector and base resistors.
Efficient in operation.

Disadvantages of Transformer Coupled Amplifier

The following are the disadvantages of a transformer coupled amplifier −

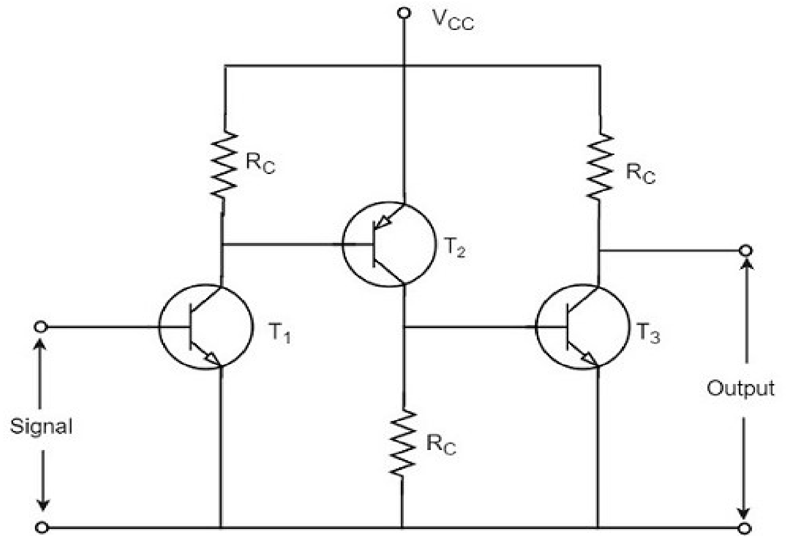
Though the gain is high, it varies considerably with frequency. Hence a poor frequency response.
Frequency distortion is higher.
Transformers tend to produce hum noise.
Transformers are bulky and costly.

Direct coupled amplifier:

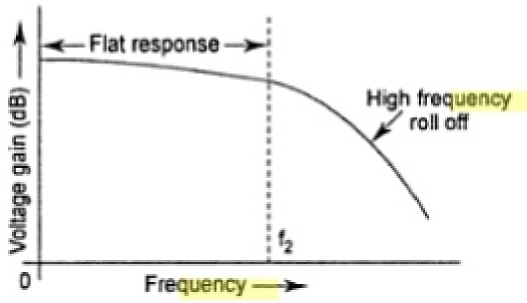
which is especially used to amplify lower frequencies, such as amplifying photo-electric current or thermo-couple current or so.

As no coupling devices are used, the coupling of the amplifier stages is done directly and hence called as **Direct coupled amplifier**.

The transistor in the second stage will be an NPN transistor, while the transistor in the first stage will be a PNP transistor and so on. This is because, the variations in one transistor tend to cancel the variations in the other. The rise in the collector current and the variation in β of one transistor gets cancelled by the decrease in the other.



The input signal when applied at the base of transistor T_1, it gets amplified due to the transistor action and the amplified output appears at the collector resistor R_c of transistor T_1. This output is applied to the base of transistor T_2 which further amplifies the signal. In this way, a signal is amplified in a direct coupled amplifier circuit.



Advantages

The advantages of direct coupled amplifier are as follows.

The circuit arrangement is simple because of minimum use of resistors.

The circuit is of low cost because of the absence of expensive coupling devices.

Disadvantages

The disadvantages of direct coupled amplifier are as follows.

It cannot be used for amplifying high frequencies.

The operating point is shifted due to temperature variations.

Applications

The applications of direct coupled amplifier are as follows.

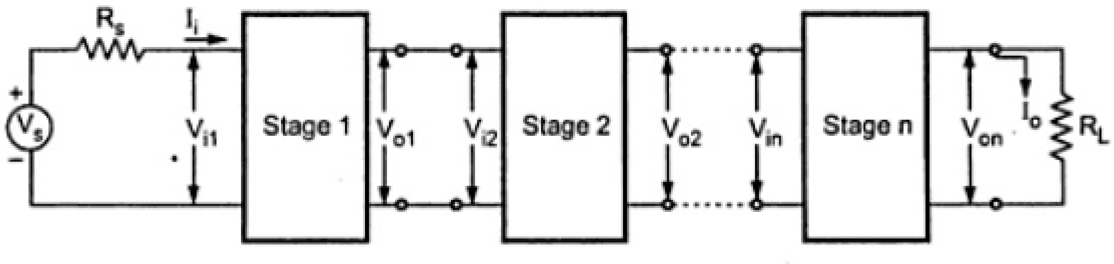
Low frequency amplifications.

Low current amplifications.

Comparisions : compare the characteristics of different types of coupling methods discussed till now.

S.No	Particular	RC Coupling	Transformer Coupling	Direct Coupling
1	Frequency response	Excellent in audio frequency range	Poor	Best
2	Cost	Less	More	Least
3	Space and Weight	Less	More	Least
4	Impedance matching	Not good	Excellent	Good
5	Use	For voltage amplification	For Power amplification	For amplifying extremely low frequencies

n-Stage Cascaded Amplifier



Voltage gain :

The resultant voltage gain of the multistage amplifier is the product of voltage gains of the various stages.

$$A_v = A_{v1} A_{v2} \dots A_{vn}$$

Gain in Decibels:

In many situations it is found very convenient to compare two powers on logarithmic scale rather than on a linear scale. The unit of this logarithmic scale is called decibel (abbreviated dB). The number N decibels by which a power P2 exceeds the power P1 is defined by

$$N = 10 \log \frac{P_2}{P_1}$$

$$P_1 = \frac{V_i^2}{R_i} \text{ and } P_2 = \frac{V_o^2}{R_o}$$

$$N = 10 \log_{10} \frac{V_o^2}{V_i^2} = 10 \log_{10} \left(\frac{V_o^2}{V_i^2} \right) = 10 \times 2 \log_{10} \frac{V_o}{V_i} = 20 \log_{10} \frac{V_o}{V_i}$$

Ex. Let us consider two stage amplifier circuit shown in fig. The first stage in the circuit is a CE amplifier and Second stage is a common collector amplifier. Calculate input and output Impedances , voltage gain and current gain with the following h-parameter

h_{ie}=2k, h_{fe}=50 , h_{re}=0, h_{oe}=0

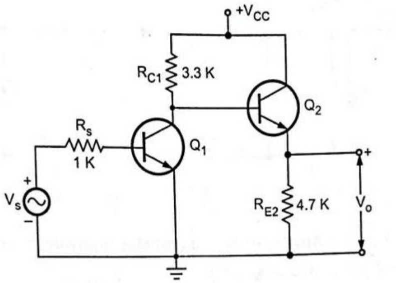


Fig. 2.2 Common emitter-common collector amplifier

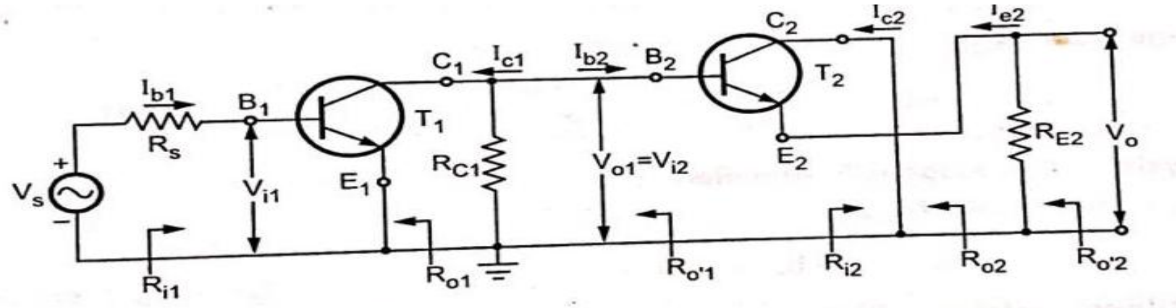


Fig. 2.3 ac equivalent circuit

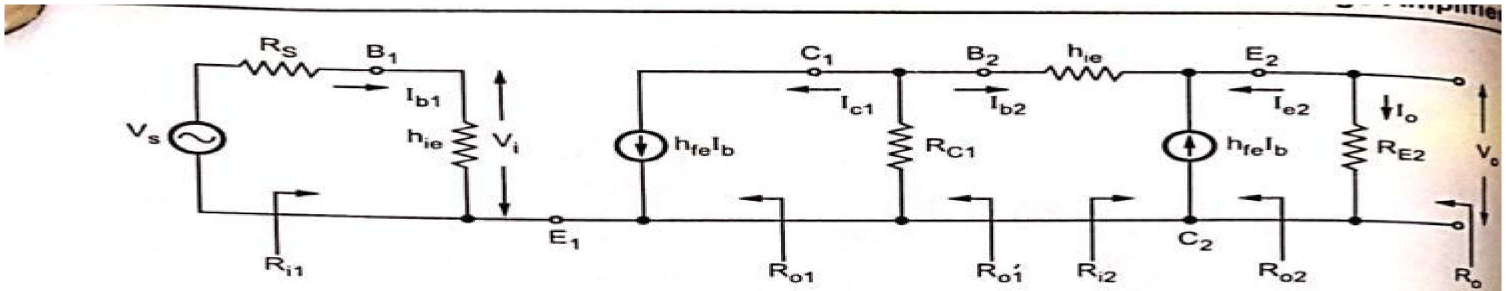


Fig. 2.4 Approximate h-parameter equivalent circuit

Analysis of second stage (CC amplifier) :

a) Current gain (A_{i2}) :

$$A_{i2} = 1 + h_{fe} = 1 + 50 = 51$$

b) Input resistance (R_{i2}) :

$$R_{i2} = h_{ie} + (1 + h_{fe}) R_{E2} = 2\text{ K} + (1 + 50) \times 4.7 \times 10^3 = 241.7\text{ K}\Omega$$

Voltage Gain (A_{v2}) :

$$A_{v2} = \frac{A_{i2} \times R_{L2}}{R_{i2}} = \frac{A_{i2} \times R_{E2}}{R_{i2}} = \frac{51 \times 4.7 \times 10^3}{241.7 \times 10^3} = 0.991$$

Analysis of first stage (CE amplifier) :

a) Current Gain (A_{i1}) :

$$A_{i1} = -h_{fe} = -50$$

b) Input resistance (R_{i1})

$$R_{i1} = h_{ie} = 2\ K$$

c) Voltage gain (A_{v1}) :

$$A_{v1} = \frac{A_{i1} \times R_{L1}}{R_{i1}}$$

where $R_{L1} = R_{C1} \parallel R_{i2}$

$= 3.3\ K \parallel 241.7\ K = 3.25\ K$

$$\therefore A_{v1} = \frac{-50 \times 3.25\ K}{2\ K} = -81.25$$

Overall voltage gain (A_v) :

$$A_v = A_{v1} \times A_{v2} = -81.25 \times 0.991 = -80.51$$

Overall voltage gain (A_{VS}) :

$$A_{vs} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s}$$

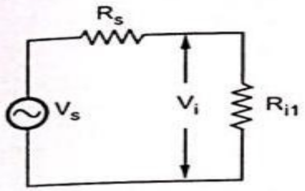


Fig. 2.5

Looking at Fig. 2.5 we can write

$$V_i = V_s \times \frac{R_{i1}}{R_s + R_{i1}}$$

$$\therefore A_{vs} = \frac{V_o}{V_i} \times \frac{R_{i1}}{R_s + R_{i1}} = A_v \times \frac{R_{i1}}{R_s + R_{i1}}$$

$$= -80.51 \times \frac{2\ K}{1\ K + 2\ K} = -53.67$$

Output Resistance (R_o) :

$$R_{o1} = \infty$$

$$R'_{o1} = R_{o1} \parallel R_{C1} = \infty \parallel 3.3K = 3.3\ K$$

$$R_{o2} = \frac{R_s + h_{ie}}{1 + h_{fe}}$$

$$= \frac{R'_{o1} + h_{ie}}{1 + h_{fe}} = \frac{3.3K + 2K}{1 + 50} \qquad \because R_s = R'_{o1}$$

$$= 103.9\ \Omega$$

$$R_o = R_{o2} \parallel R_{E2} = 103.9 \parallel 4.7\ K = 101.65\ \Omega$$

Overall Current gain (A_i) :

$$A_i = \frac{I_o}{I_{b1}} = \frac{I_o}{I_{e2}} \times \frac{I_{e2}}{I_{b2}} \times \frac{I_{b2}}{I_{c1}} \times \frac{I_{c1}}{I_{b1}}$$

where $$\frac{I_o}{I_{e2}} = -1, \frac{I_{e2}}{I_{b2}} = -A_{i2}$$

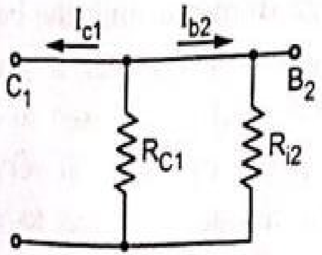


Fig. 2.6

$$\frac{I_{b2}}{I_{c1}} = -\frac{R_{c1}}{R_{i2} + R_{c1}} = \frac{3.3\,K}{241.7\,K + 3.3\,K}$$

$$= -0.01346$$

$$\frac{I_{c1}}{I_{b1}} = A_{i1}$$

$$\therefore A_i = -1 \times -A_{i2} \times -0.01346 \times -A_{i1}$$

$$= -1 \times -51 \times -0.01346 \times -(-50) = -34.323$$

Two-stage RC Coupled Amplifier analysis(**Two Stage CE-CE Cascade Amplifier)**

2.6 Two Stage RC Coupled CE-CE Cascade Amplifier

Fig. 2.14 shows that CE-CE cascade amplifier, with their biasing arrangements.

Assuming all capacitors arbitrarily large and act as a short circuit for ac signal we can draw h-parameter equivalent circuit for CE-CE cascade amplifier, as shown in Fig. 2.14.

Let us calculate R_i, A_i, A_v, R'_i, A_{vs} and A_{is} if circuit parameters are : $R_s = 1$ K, $R_{C1} = 15$ K, $R_{E1} = 100\ \Omega$, $R_{C2} = 4$ K, $R_{E2} = 330\ \Omega$ with $R_1 = 200$ K and $R_2 = 20$ K for first stage and $R_1 = 47$ K and $R_2 = 4.7$ K for second stage. Assume that $h_{ie} = 1.2\ K\Omega$, $h_{fe} = 50$, $h_{re} = 2.5 \times 10^{-4}$ and $h_{oe} = 25 \times 10^{-6}$ A/V.

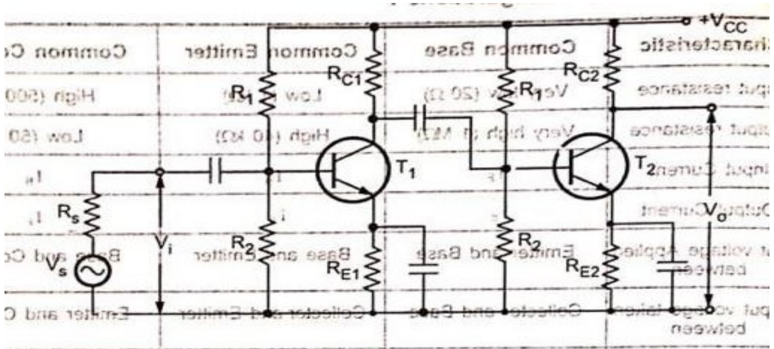


Fig. 2.14 CE-CE Cascade amplifier

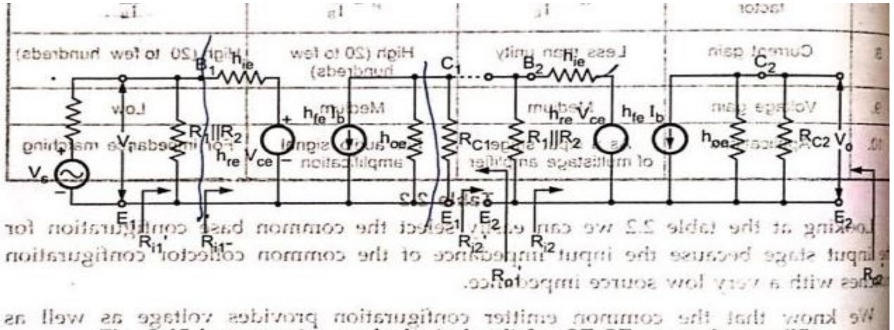


Fig. 2.15 h-parameter equivalent circuit for CE-CE cascade amplifier

Analysis of second stage

$$h_{oe}R_L = 25x10^{-6}\ X\ 4\ X10^3 = 0.1\ we\ can\ use\ apprximate\ analysis$$

Current gain (A_{I2}) = - h_{fe} = -50

Input Impedance (R_{i2}) = h_{ie} = 1.2KΩ

$$Voltage\ gain A_{v2} = \frac{A_{i2} X R_L}{R_{i2}} = \frac{-50X4X10^3}{1.2X10^3} = -166.67$$

Analysis of first stage:

$R_L' = R_{c1} // R_1 // R_2 // R_{i2} = 15k // 47k // 4.7k // 1.2k = 881.8\Omega$

$$h_{oe}R_L' = 25x10^{-6}\ X\ 881.8 = 0.022$$

$$h_{oe}R_L' < 0.1, so\ we\ can\ use\ approximate\ analysis\ for\ first\ stage$$

Current gain (A_{I1}) = - h_{fe} = -50

Input Impedance (R_{i1}) = h_{ie} = 1.2KΩ

$$Voltage\ gain A_{v1} = \frac{A_{i2} X R_L'}{R_{i2}} = \frac{-50 X 881.8}{1.2 X 10^3} = -36.74$$

Over all Voltage gain (A_v) = A_{v1} XA_{v2} = (-166.67) X(-36.74) =6123.45

Over all Voltage gain include source (A_{vs}) = A_v X [(R_{i1}) / (R_{i1}+R_s)] = 3248.6

R_{i1}'= R1 // R2 // Ri1 =220k //20k// 1.2k =1.13 k

Output Resistance (R_0) :

$R_{01}' = R_{01} // R_{c1}$ = infinity // 15k=15k

$R_{02}' = R_{02} // R_{c2}$ = infinity // 4k=4k

CE-CB cascode amplifiers: In cascode two amplifier circuits are vertically connected. Which provides high input impedance as well as good voltage gain and good frequency response

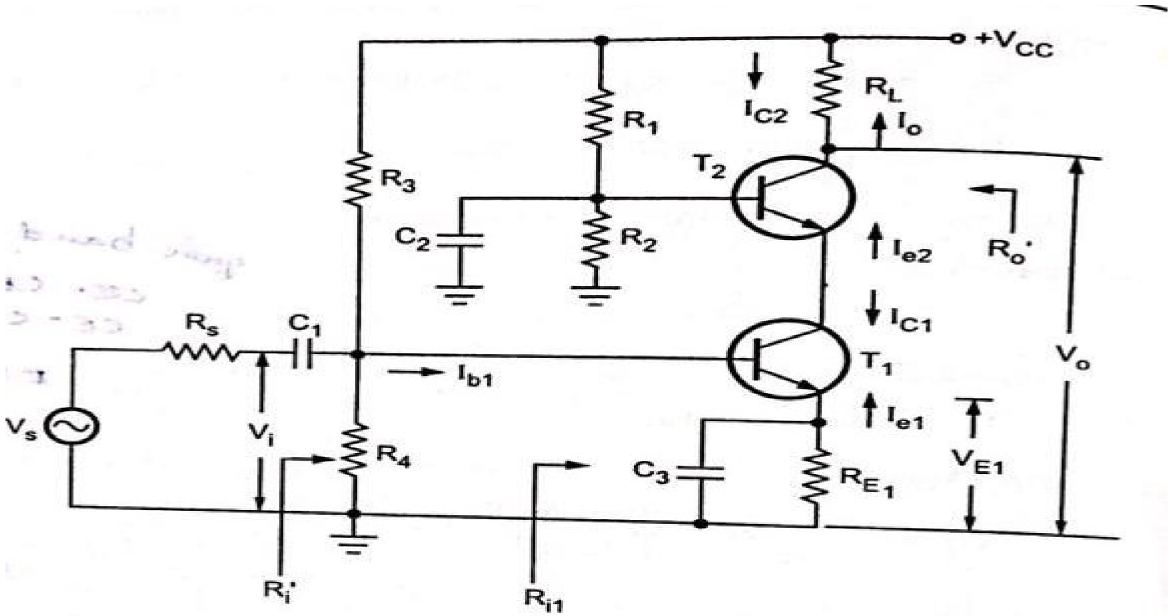


Fig. 2.16 Cascode amplifier

The simplified h-parameter equivalent circuits for cascode amplifier is drawn by replacing transistors with their simplified equivalent circuits, as shown in the Fig. 2.18.

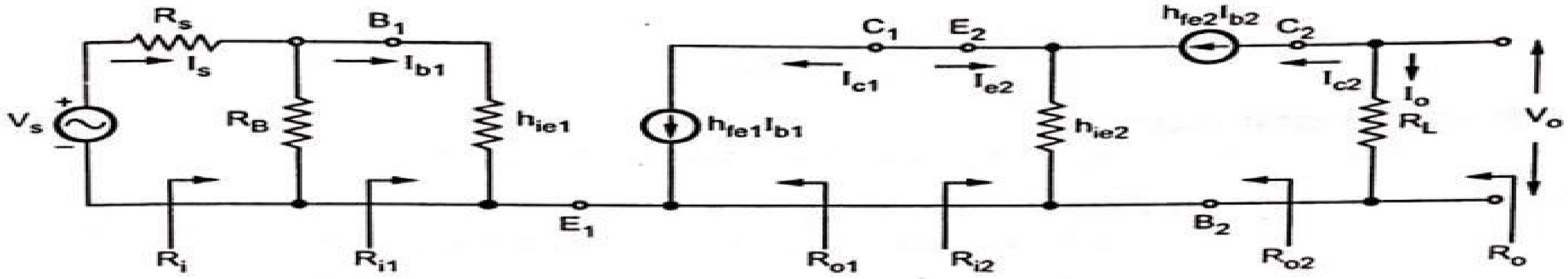


Fig. 2.18 Simplified h-parameter equivalent circuit for CE-CB amplifier

Let us consider the circuit parameters are : R_S = 1K, R_3 = 200 K, R_4 = 10 K and R_L = 3 K, and transistor parameters for both transistors are : h_{ie} = 1.1 K and h_{fe} = 50.

Analysis of second stage (CB amplifier) :

a) Current gain (A_{i2}) :

$$A_{i2} = \frac{h_{fe}}{1+h_{fe}} = \frac{50}{1+50} = 0.98$$

b) Input resistance (R_{i2}) :

$$R_{i2} = \frac{h_{ie}}{1+h_{fe}} = \frac{1.1\,K}{1+50} = 21.56\ \Omega$$

c) Voltage gain (A_{v2}) :

$$A_{v2} = \frac{A_{i2}\,R_{L2}}{R_{i2}} = \frac{0.98\times 3\,K}{21.56} = 136.36$$

Analysis of first stage (CE amplifier) :

a) Current gain (A_{i1}) :

$$A_{i1} = -h_{fe} = -50$$

b) Input resistance (R_{i1}) :

$$R_{i1} = h_{ie} = 1.1\ K$$

$$Voltage\ gain(A_{v1}) = \frac{A_{i1} X R_{L1}}{R_{i1}} = \frac{-50X21.56}{1.1k} = 0.98$$

$$Overall\ Voltage\ gain(A_v) = A_{v1} X A_{v2} = -0.98X136.36 = -133.63$$

Overall Input Impedance (R_i) = R_{i1} // R_B = R_{i1} // R_3 // R_4 =1.1k // 200k // 4.7k // 10k=986.1Ω

Output Impedance (R_0) = R_{02} // R_L =∞ // 3k=3k

$R_{01} = \infty$ $R_{02} = \infty$

Darlington pair (or) CC-CC amplifier: The cascade connection of two emitter followers(CC) is called Darlington Connection.

Figure shows direct connection of two stages of emitter follower amplifier circuit, its gives large current gain and increases the input impedances

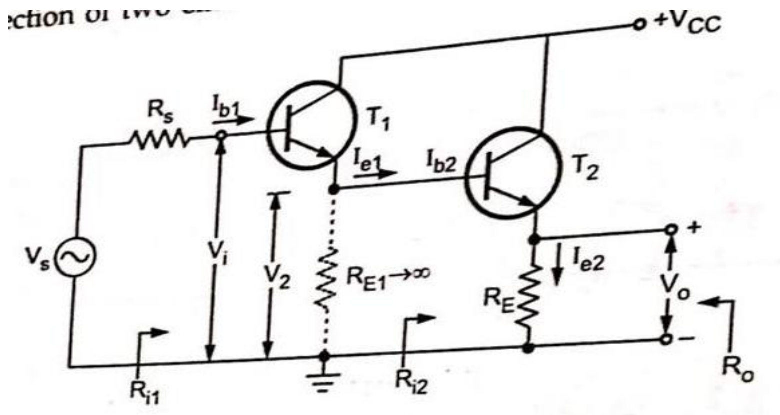
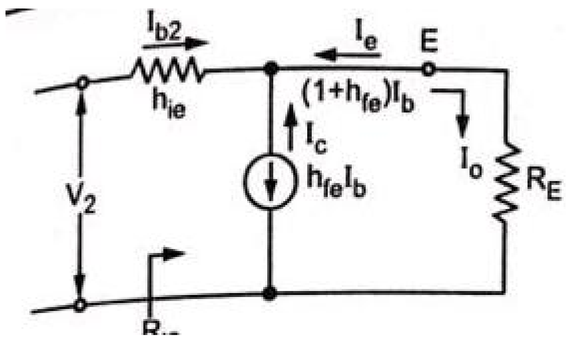


Fig. 2.74 Darlington emitter follower circuit

Assume load Resistance R_L, $h_{oe}\ R_L<0.1$, Therefore we can use approximate analysis for analysing second stage. Below fig. Show h-parameter model for second stage(CC stage).



Analysis of second stage :

a) Current Gain (A_{i2}) : $A_{i2} = \frac{I_o}{I_b} = -\frac{I_e}{I_b} = \frac{I_b + h_{fe} I_b}{I_b} = \frac{I_b(1+h_{fe})}{I_b}$

$$\therefore \boxed{A_{i2} = 1 + h_{fe}} \quad \dots (1)$$

b) Input Resistance (R_{i2}) :

$$R_{i2} = \frac{V_2}{I_{b2}}$$

Applying KVL to outer loop we get,

$$V_2 - I_{b2} h_{ie} - I_o R_E = 0$$

$$\therefore V_2 = I_{b2} h_{ie} + I_o R_E$$

$$\therefore \frac{V_2}{I_{b2}} = h_{ie} + \frac{I_o}{I_{b2}} R_E$$

$$\therefore R_{i2} = h_{ie} + A_{i2} R_E \quad \text{since, } \frac{I_o}{I_{b2}} = A_{i2}$$

$$\therefore \boxed{R_{i2} = h_{ie} + (1 + h_{fe}) R_E} \quad \dots (2)$$

$$R_{i2} = (1 + h_{fe}) R_E \quad \because h_{ie} << (1 + h_{fe}) R_E \quad \dots (3)$$

Analysis of first stage : We can see load resistance of first stage is the input resistance of the second stage i.e $R_{i2}=R_{L1}$. R_{i2} is high, usually it does not meet the requirement $h_{oe}R_L<0.1$.So we use exact analysis method for analysis of the first stage.

Exact h-parameter equivalent circuit for CC amplifier

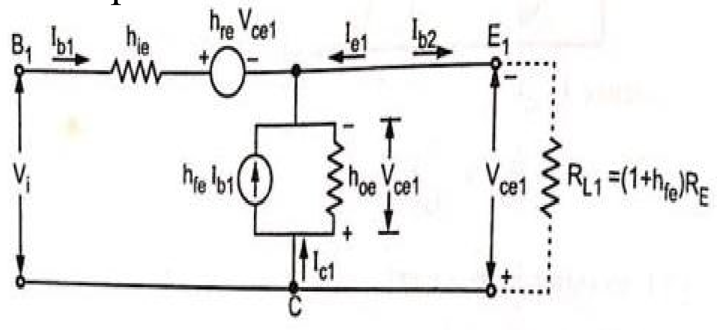


Fig 2.78

a) Current Gain (A_{i1}) :

$$A_{i1} = \frac{I_{b2}}{I_{b1}}$$

$$A_{i1} = \frac{I_{e1}}{I_{b1}}$$

$$I_{e1} = -(I_{b1} + I_{c1})$$

and

$$I_{c1} = h_{fe}I_{b1} + h_{oe}V_{ce1} = h_{fe}I_{b1} + h_{oe}(-I_{b2}R_{L1})$$

$$= h_{fe}I_{b1} + h_{oe}I_{e1}R_{L1}$$

Substituting value of I_{c1} in equation 4 we get,

$$\therefore I_{e1} = -(I_{b1} + h_{fe}I_{b1} + h_{oe}I_{e1}R_{L1}) = -I_{b1} - h_{fe}I_{b1} - h_{oe}I_{e1}R_{L1}$$

$$\therefore I_{e1} + h_{oe}R_{L1}I_{e1} = -I_{b1}(1 + h_{fe})$$

$$-\frac{I_{e1}}{I_{b1}} = \frac{1+h_{fe}}{1+h_{oe}R_{L1}}$$

We know that, $R_{L1} = (1 + h_{fe})R_E$

$$\therefore \quad A_{i1} = -\frac{I_{e1}}{I_{b1}} = \frac{1+h_{fe}}{1+h_{oe}(1+h_{fe})R_E} \quad \ldots (5)$$

$$= \frac{1+h_{fe}}{1+h_{oe}h_{fe}R_E} \quad \because h_{fe} >> 1$$

b) Input resistance (R_i) : $R_{i1} = \frac{V_i}{I_{b1}}$

Applying KVL to output loop we get

$$V_i - I_{b1} h_{ie} - h_{re} V_{ce1} + V_{ce1} = 0$$

$$\therefore V_i = I_{b1} h_{ie} + h_{re} V_{ce1} - V_{ce1}$$

The terms $h_{re} V_{ce1}$ is negligible since h_{re} is in the order of 2.5×10^{-4}

$$= I_{b1} h_{ie} - (-I_{b2} R_{L1}) = I_{b1} h_{ie} + I_{b2} R_{L1}$$

$$\therefore R_{i1} = \frac{V_i}{I_{b1}} = h_{ie} + \frac{I_{b2}}{I_{b1}} R_{L1} = h_{ie} + A_{i1} R_{L1}$$

$$\therefore \quad \boxed{R_{i1} = h_{ie} + A_{i1} (1 + h_{fe}) R_E} \quad \ldots (6)$$

Substituting value of A_{i1} we get,

$$R_{i1} = \frac{V_i}{I_{b1}} = h_{ie} + \frac{(1 + h_{fe})(1 + h_{fe}) R_E}{1 + h_{oe} h_{fe} R_E}$$

$$\therefore \quad \boxed{R_{i1} = h_{ie} + \frac{(1 + h_{fe})^2 R_E}{1 + h_{oe} h_{fe} R_E}} \quad \ldots (7)$$

$$\therefore \quad \boxed{R_{i1} \approx \frac{(1 + h_{fe})^2 R_E}{1 + h_{oe} h_{fe} R_E}} \quad \because h_{ie} << \frac{(1 + h_{fe})^2 R_E}{1 + h_{oe} h_{fe} R_E} \quad \ldots (8)$$

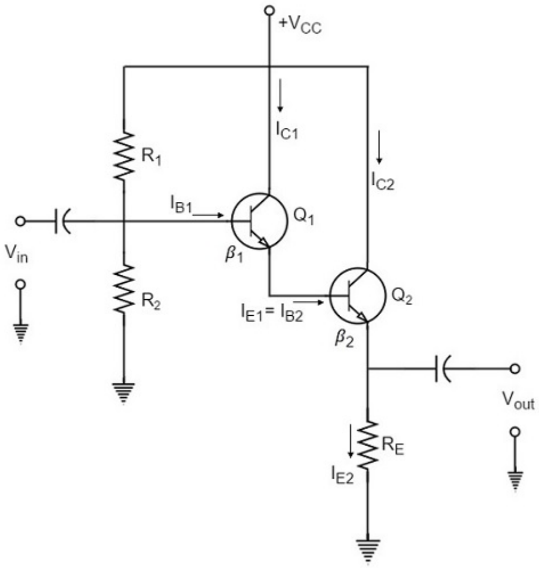
Overall Current Gain (A_i) :

$$A_i = A_{i1} \times A_{i2}$$

$$= \frac{1 + h_{fe}}{1 + h_{oe} (1 + h_{fe}) R_E} \times (1 + h_{fe})$$

$$\therefore \quad \boxed{A_i = \frac{(1 + h_{fe})^2}{1 + h_{oe} (1 + h_{fe}) R_E}}$$

Darlington pair with biasing Resistors:



The effective input impedance $R_i' = R_i // R_B = R_B$

The intension behind the designing of Darlington pair is to get the higher input impedance. Due to biasing resistors the input resistance is going to be decreases.

The solution of this problem is to use a Bootstrap connection at input side.

Darlington pair with Bootstrap connection: It is used to improve the input Impedance.

C is high value such that at all operating frequencies it is short circuited

R3 is high value

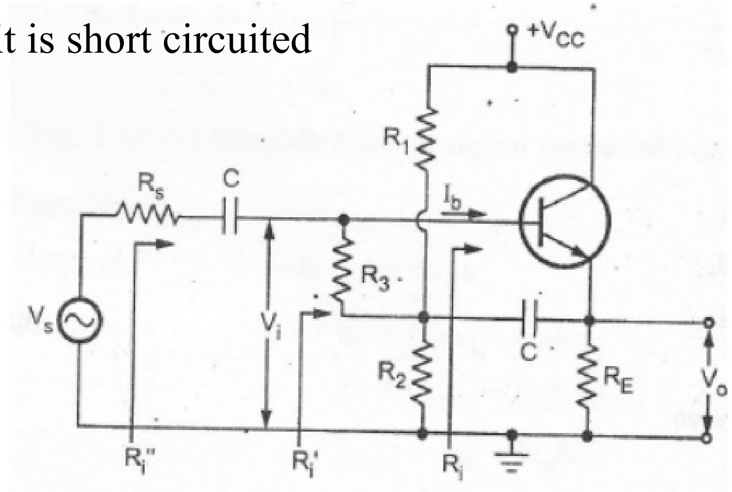


Fig. Bootstrap emitter follower

Analysis of amplifiers at high frequencies:

Transistors at high frequencies: At low frequencies, we have assumed that the response of the transistors to change of input voltage or current is instantaneous and hence we neglect the effect of shunt capacitance in the transistor.

But at high frequencies, we consider the effect of shunt capacitance. The parasitic or stray or junction capacitance is invisible but exist between parts of components or junctions.

So at high frequencies, the transistors are analysed by considering junction capacitance.

At low frequencies transistors are analysed by using h-parameter model, but it is not suitable at high frequencies because of the following reasons.

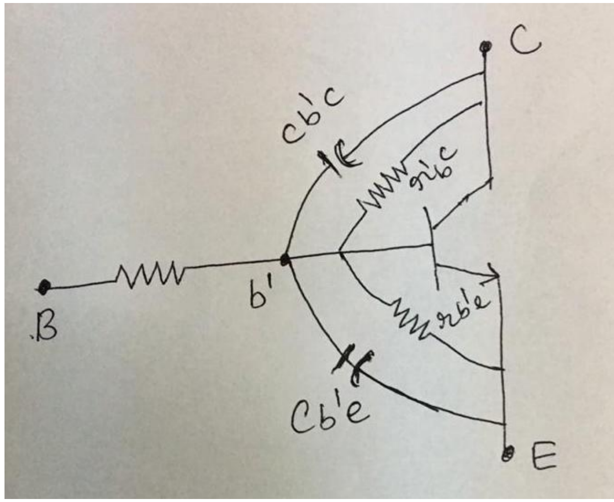
1. The values of h-parameters are not constant at high frequencies.
2. At high frequencies h-parameters are having complex nature.

CE- transistor analysis at high frequencies : At high frequencies transistor are analysed by using **hybrid - π model.**

At high frequencies , the base is divided into actual base B and virtual base B^I , in between B and B^I , resistance ($r_{bb}{}^I$) exist it is known as base spreading resistance.

We consider the effect of capacitances $C_{b'e}$, $C_{b'c}$.

We know, forward bias P-N junction exhibits a capacitance effect called diffusion capacitance i.e. $C_{b'e}$ or C_e and reverse bias P-N junction exhibits transition capacitance denoted as $C_{b'c}$ or C_c.



To analyse the transistor, consider the base spreading resistance rb'b, resistance and capacitance between base and collector i.e. $r_{b'c}$ and C_c and resistance and capacitance between base and emitter i.e. $r_{b'e}$ and C_e .

The Hybrid-π or Giacoletto Model for the Common Emitter amplifier circuit

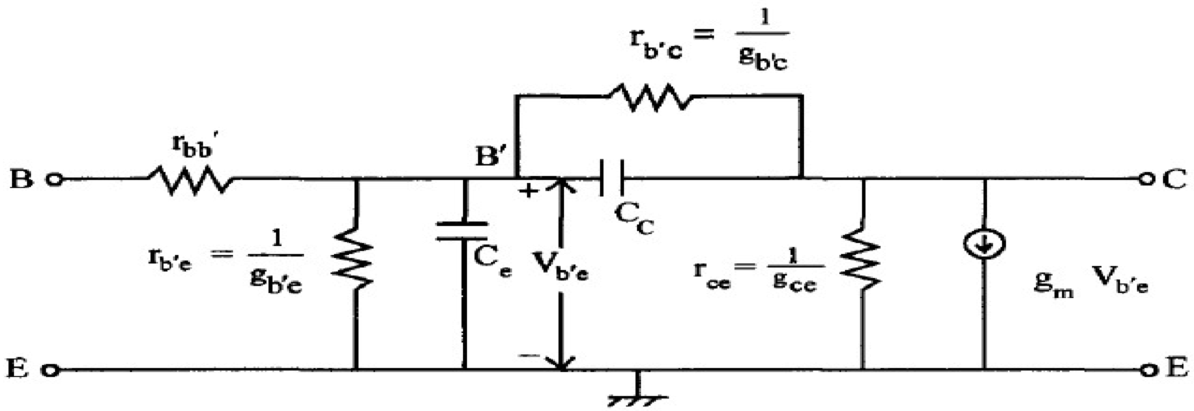


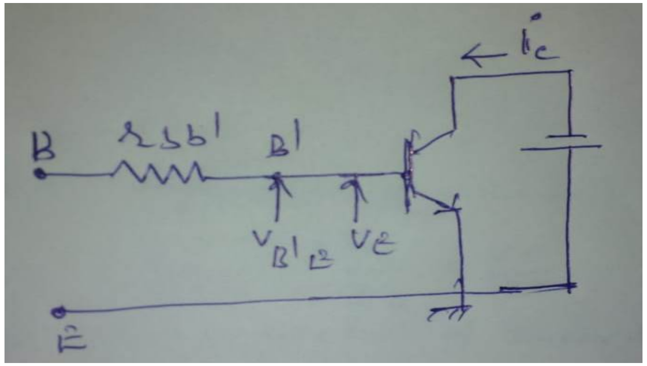
Fig. 3.1 Hybrid - π C.E BJT Model

Typical values of hybrid-π parameters at I_c=1.3 mA

Parameter	Meaning	Value
g_m	Mutual conductance of transistor	50 mA/V
$r_{bb'}$	Base spreading resistance	100 Ω
$r_{b'e}$ or	Resistance between B' and E	1 kΩ
$g_{b'e}$	Conductance between B' and E	1 m mho
$r_{b'c}$ or	Resistance of reverse biased P-N junction between base and collector	4 MΩ
$g_{b'c}$	Conductance of reverse biased P-N junction between base and collector	0.25×10^{-6} mho
r_{ce} or	Output resistance between C and E	80 kΩ
g_{ce}	Conductance between C and E	12.5×10^{-6} mho
C_e	Junction capacitance between B and E	100 pF
C_c	Junction capacitance between base and collector	3 pF

Typical values depends on **temperature and I_c**.

Determination of Hybrid-π Transconductance (g_m).:



Transconductance (g_m) is the ratio of change in the collector current due to change in voltage across base-emitter junction.

$$g_m = \frac{\partial I_c}{\partial V_{B'E}} \quad /V_{CE} - constant$$

Collector current in active region $I_C = \alpha I_E + I_{C0}$

Differentiating above equation $\partial I_C = \alpha \partial I_E$

Substitute ∂I_C in above equation

$$g_m = \frac{\alpha \partial I_E}{\partial V_{B'E}} = \alpha \frac{\partial I_E}{\partial V_E}$$

$V_{B'E} = V_E$

The emitter diode resistance r_e is given as

$$r_e = \frac{\partial V_E}{\partial I_E}$$

$$g_m = \frac{\alpha}{r_e}$$

The emitter diode is a forward biased diode and its dynamic resistance is given as

$$r_e = \frac{V_T}{I_E} \quad \ldots($$

where V_T is the "volt equivalent of temperature", defined by

$$V_T = \frac{kT}{q}$$

where k is the Boltzmann constant in joules per degree kelvin (1.38×10^{-23} J/°K) the electronic charge (1.6×10^{-19} C).

$$g_m = \frac{\alpha I_E}{V_T} \qquad g_m = \frac{I_c - I_{c0}}{V_T} \qquad I_C = \alpha I_E + I_{C0}$$

For npn or pnp, g_m should be positive. $I_c >> I_{CO}$

$$\alpha I_E = I_C - I_{C0}$$

$$g_m = \frac{I_c}{V_T}$$

$$g_m = \frac{I_C}{\frac{KT}{q}} = \frac{qI_C}{KT} = \frac{1.6\,X10^{-19}}{1.38\,X\,10^{-23}}\,X\,\frac{I_C}{T} = 11600\,\frac{I_C}{T}$$

At room temperature , 300^0K

$$g_m = 11{,}600\,\frac{I_C}{T} = \frac{11600}{300}XI_C = \frac{I_C}{26X10^{-3}} = \frac{I_C\,(mA)}{26}$$

I_c = 1.3 mA, g_m= 0.05 *A / V*

I_c = 10 mA, gm = 400 *mA /V*

3.3.2 The Input Conductance $g_{b'e}$

Fig. 3.3 (a) and (b) shows the hybrid-π model and the h-parameter model for CE configuration at low frequency, respectively. At low frequency, all capacitors are negligible and hence not drawn in Fig. 3.3 (a)

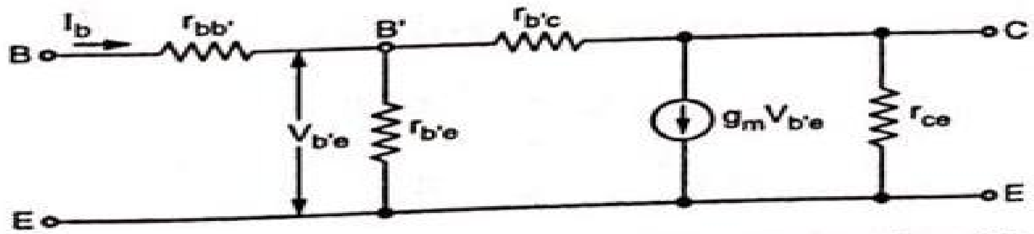
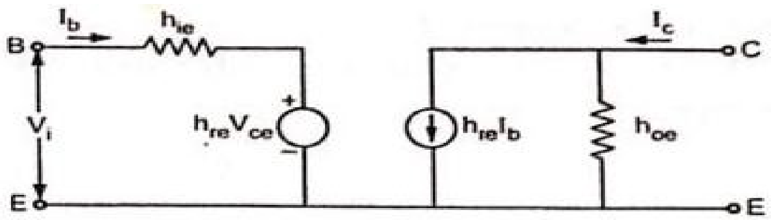


Fig. 3.3 (a) Hybrid-π model for CE configuration at low frequency



3 (b) h-parameter model for CE configuration at low frequency

From the circuit

$$I_C = h_{fe} I_b + h_{oe} V_{ce}$$

Making $V_{ce} = 0$, ie. output short circuit, the short circuit current gain h_{fe} is define as

$$h_{fe} = \frac{I_C}{I_b} \quad \text{... (}$$

Now, consider the hybrid-π model for CE configuration. Looking at Table 3.1 w have $r_{b'c} = 4\ M\Omega >> r_{b'e}$. Hence I_b flows into $r_{b'e}$ and $V_{b'e} \approx I_b\, r_{b'e}$. Similarly, as $r_{b'c}$ i very large $I_C = g_m V_{b'e}$.

$$\therefore \quad I_C = g_m V_{b'e}$$

$$= g_m I_b r_{b'e} \qquad \because V_{b'e} = I_b r_{b'e}.$$

$$\therefore \quad \frac{I_C}{I_b} = g_m \ r_{b'e} \quad \text{... (}$$

Substituting value of I_C/I_b in equation (8) we get

$$h_{fe} = g_m \, r_{b'e}$$

or

$$r_{b'e} = \frac{h_{fe}}{g_m} \quad \text{or} \quad g_{b'e} = \frac{g_m}{h_{fe}} \qquad \ldots(1$$

From equation (5) we know that $g_m = I_C / V_T$

$\therefore$

$$r_{b'e} = \frac{h_{fe} V_T}{|I_C|}$$

or

$$g_{b'e} = \frac{|I_C|}{V_T h_{fe}} \qquad \ldots($$

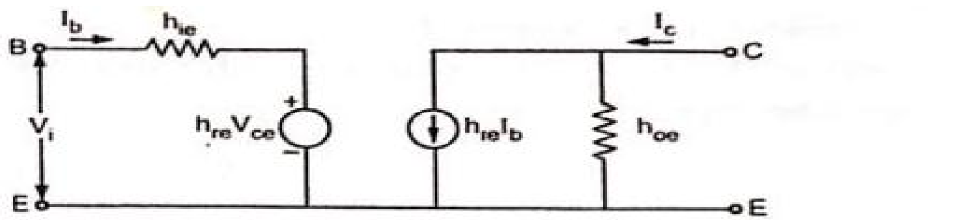
Looking at equation (11), we can say that, over the range of currents for which remains fairly constant, $r_{b'e}$ is directly proportional to temperature and inversely proportional to collector current.

The feedback conductance($g_{b'c}$):

Let us consider h-parameter model for CE-configuration with input is open circuit i.e. I_b=o

If I_b=o

$V_i = V_{BE} = h_{re} V_{CE}$



3 (b) h-parameter model for CE configuration at low frequency

Now consider the hybrid-π model for CE configuration as shown in Fig. 3.4.

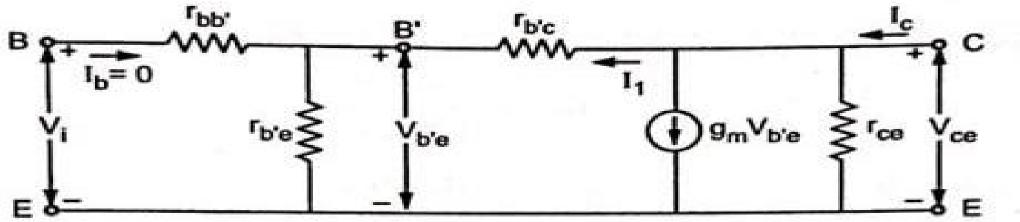


Fig. 3.4

With $I_b = 0$, V_{ce} can be given as

$$V_{ce} = I_1 (r_{b'c} + r_{b'e})$$

$$\therefore \quad I_1 = \frac{V_{ce}}{r_{b'c} + r_{b'e}} \qquad \ldots (13)$$

The voltage between B' and E, $V_{b'e}$ can be given as

$$V_{b'e} = I_1 \, r_{b'e}$$

$$\therefore \quad V_{b'e} = r_{b'e} \frac{V_{ce}}{r_{b'c} + r_{b'e}} \qquad \ldots (14)$$

With $I_b = 0$

$$V_i = V_{b'e}$$

$$= \frac{r_{b'e} V_{ce}}{r_{b'c} + r_{b'e}}$$

Substituting value of V_i in equation (12) we get,

$$h_{re} V_{ce} = \frac{r_{b'e} V_{ce}}{r_{b'c} + r_{b'e}}$$

$$\therefore \quad h_{re} = \frac{r_{b'e}}{r_{b'c} + r_{b'e}}$$

$$\therefore \quad r_{b'e} = h_{re} r_{b'c} + h_{re} r_{b'e}$$

$$\therefore \quad (1 - h_{re}) r_{b'e} = h_{re} r_{b'c}$$

$$r_{b'c} = \left(\frac{1-h_{re}}{h_{re}}\right) r_{b'e} = \frac{r_{b'e}}{h_{re}}$$

$$g_{b'c} = \frac{h_{re}}{r_{b'e}} = h_{re} g_{b'e}$$

Substitute $r_{b'e}$ in above equation

$$r_{b'c} = \frac{h_{fe} V_T}{|I_c| h_{re}}$$

$$g_{b'c} = \frac{|I_c| h_{re}}{h_{fe} V_T}$$

3.3.4 The Base Spreading Resistance $r_{bb'}$

Let us consider h-parameter model for CE configuration. The input resistance wi output shorted (V_{ce} = 0) is h_{ie}. With hybrid-π model input resistance with outp shorted is $r_{bb'} + r_{b'e}$.

$$\therefore \quad h_{ie} = r_{bb'} + r_{b'e}$$

$$\therefore \quad r_{bb'} = h_{ie} - r_{b'e} \quad \text{... (1}$$

Substituting value of $r_{b'e}$ from equation (11) we get,

$$r_{bb'} = h_{ie} - \frac{h_{fe} V_T}{I_C} \quad \text{... (1}$$

The Output conductance g_{ce}:

Using h-parameters the output conductance is given as

$$h_{oe} = \frac{I_C}{V_{ce}} \quad \ldots$$

Now, consider hybrid-π model for CE configuration, shown in Fig. 3.4. App KCL to the output circuit we get

$$I_C = \frac{V_{ce}}{r_{ce}} + g_m V_{b'e} + I_1$$

Substituting value of I_1 from equation (13) we get,

$$I_C = \frac{V_{ce}}{r_{ce}} + g_m V_{b'e} + \frac{V_{ce}}{r_{b'c} + r_{b'e}}$$

Substituting value of $V_{b'e}$ from equation (14) we get,

$$I_C = \frac{V_{ce}}{r_{ce}} + g_m \left(\frac{r_{b'e} V_{ce}}{r_{b'c} + r_{b'e}} \right) + \frac{V_{ce}}{r_{b'c} + r_{b'e}}$$

Dividing both sides by V_{ce} we get

$$\frac{I_C}{V_{ce}} = \frac{1}{r_{ce}} + \frac{g_m r_{b'e}}{r_{b'c} + r_{b'e}} + \frac{1}{r_{b'c} + r_{b'e}}$$

$$= \frac{1}{r_{ce}} + \frac{h_{fe}}{r_{b'c} + r_{b'e}} + \frac{1}{r_{b'c} + r_{b'e}} \qquad \because h_{fe} = g_m r_{b'e} \text{ (eq. 10)}$$

$$= \frac{1}{r_{ce}} + \frac{(h_{fe} + 1)}{r_{b'c} + r_{b'e}} \qquad \ldots (20)$$

$$= \frac{1}{r_{ce}} + \frac{h_{fe}}{r_{b'c} + r_{b'e}} \qquad \because h_{fe} >> 1 \qquad \ldots (21)$$

Substituting value of $\frac{I_C}{V_{ce}}$ in equation (19) we get

$$h_{oe} = \frac{1}{r_{ce}} + \frac{h_{fe}}{r_{b'c} + r_{b'e}} \quad \ldots (22)$$

$$= \frac{1}{r_{ce}} + \frac{h_{fe}}{r_{b'c}} \qquad \because r_{b'c} >> r_{b'e} \quad \ldots (23)$$

$$\therefore \quad h_{oe} = g_{ce} + g_{b'c} h_{fe}$$

$$\therefore \quad \frac{1}{r_{ce}} = g_{ce} = h_{oe} - g_{b'c} h_{fe} \quad \ldots (24)$$

Hybrid – π capacitances: This model have two capacitances those are diffusion capacitance and transition capacitance.

3.6 CE Short-Circuit Current Gain

Consider a single stage CE transistor amplifier with load resistor R_L, as shown in the Fig. 3.6.

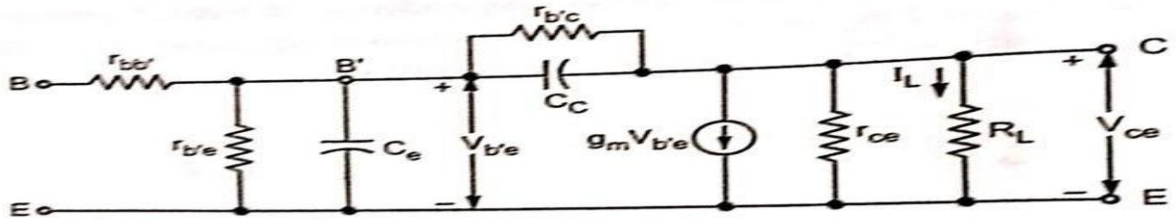
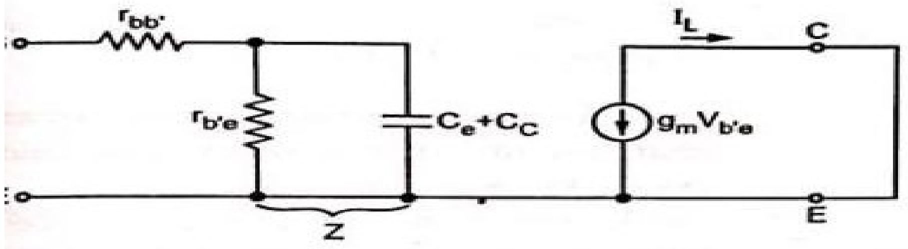


Fig. 3.6 The hybrid-π circuit for a single transistor with a resistive load R_L

For the analysis of short circuit current gain we have to assume $R_L = 0$. With $= 0$, i.e. output short circuited r_{ce} becomes zero, $r_{b'e}$, C_e and $C_{b'C}$ appear in parallel. ...en $C_C(C_{b'c})$ appears between base and emitter, it is known as **Miller capacitance** (C_M). Its admittance is given as $j\omega C_M = \frac{i_{Cb'C}}{V_{b'e}} j\omega C_{b'c}(1 + g_m R_L)$

Hence, the Miller capacitance is $C_M = C_{b'c}(1 + g_m R_L)$



g. 3.7 Simplified hybrid-π model for short circuit CE transistor

Here, $R_L = 0$

$\therefore \quad C_M = C_{b'c}\ (C_C)$

As $r_{b'c} >> r_{b'e}$, $r_{b'c}$ is neglected. With these approximation we get simplified hybrid-π model for short circuit CE transistor, as shown in the Fig. 3.7.

Parallel combination of $r_{b'e}$, and $(C_e + C_C)$ is given as

$$Z = \frac{r_{b'e} \times \frac{1}{j\omega(C_e + C_C)}}{r_{b'e} + \frac{1}{j\omega(C_e + C_C)}}$$

$$= \frac{r_{b'e}}{1 + j\omega r_{b'e}(C_e + C_C)} \quad \ldots (1)$$

This simplifies hybrid-π model as shown in the Fig. 3.8.

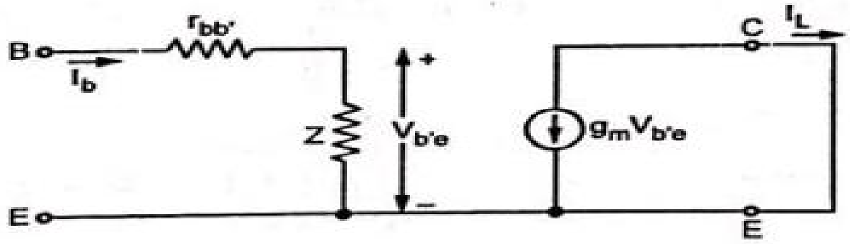


Fig. 3.8 Further simplified hybrid-π model

Look at Fig. 3.8 we can write $\quad V_{b'e} = I_b Z$

$$\therefore \quad Z = \frac{V_{b'e}}{I_b} \quad \ldots (2)$$

The current gain for the circuit shown in Fig. 3.8 can be given as

$$A_i = \frac{I_L}{I_b} = \frac{-g_m V_{b'e}}{I_b} \quad \because I_L = -g_m V_{b'e}$$

Substituting value of $V_{b'e} / I_b$ from equation 2 we get,

$$A_i = -g_m Z$$

$$= \frac{-g_m r_{b'e}}{1 + j\omega r_{b'e} (C_e + C_C)} \quad \ldots (3$$

From equation (10) of section 3.3 we know that $h_{fe} = g_m r_{b'e}$.

$$\therefore \quad A_i = \frac{-h_{fe}}{1 + j\omega r_{b'e} (C_e + C_C)} \quad \ldots (4)$$

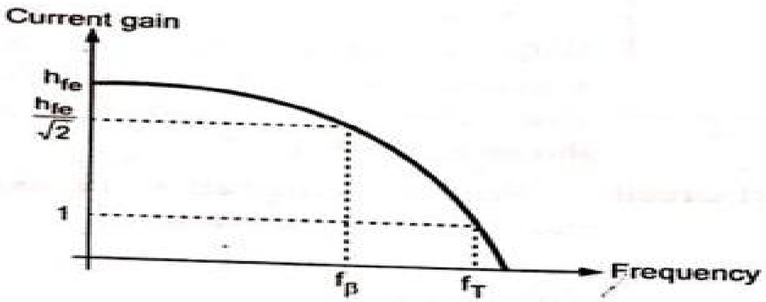


Fig. 3.9

Looking at equation (4) we can say that current is not constant. It depends on frequency. When frequency is small, the term containing f is very small compared to 1 and hence at low frequency, $A_i = -h_{fe}$. But as frequency increases A_i reduces as shown in Fig. 3.9.

Let us put

$$f_\beta = \frac{1}{2\pi r_{b'e}(C_e + C_C)}$$

Substituting value of f_β in equation (4) we get,

$$A_i = \frac{-h_{fe}}{1 + j\frac{f}{f_\beta}}$$

$$|A_i| = \frac{h_{fe}}{\sqrt{1 + \left(\frac{f}{f_\beta}\right)^2}}$$

At

$$f = f_\beta$$

$$|A_i| = \frac{h_{fe}}{\sqrt{2}}$$

3.7 Current Gain with Resistive Load

Consider a single stage CE transistor amplifier with load resistance R_L, as shown in the Fig. 3.10.

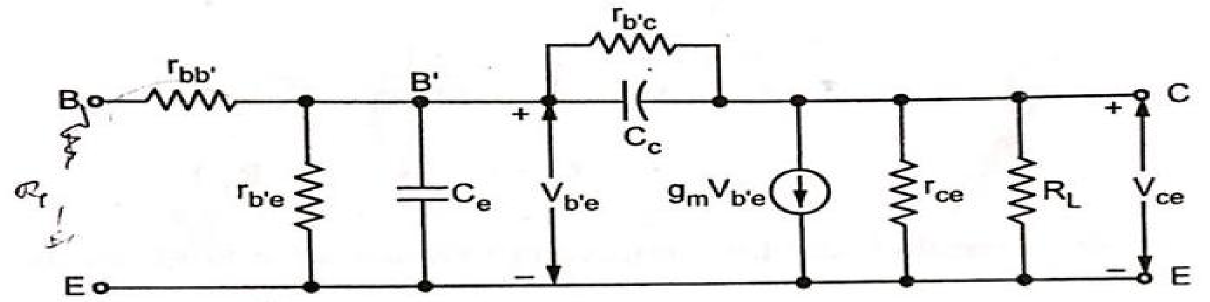


Fig. 3.10 The hybrid-π model for a single transistor with resistive load

The output circuit r_{ce} is in parallel with R_L. For high frequency amplifiers R_L ... as compared to r_{ce} and hence we can neglect r_{ce}. Using Miller's theorem, we split $r_{b'c}$ and C_C to simplify the analysis.

Fig. 3.11 shows the simplified hybrid-π model for a single transistor with resistive load.

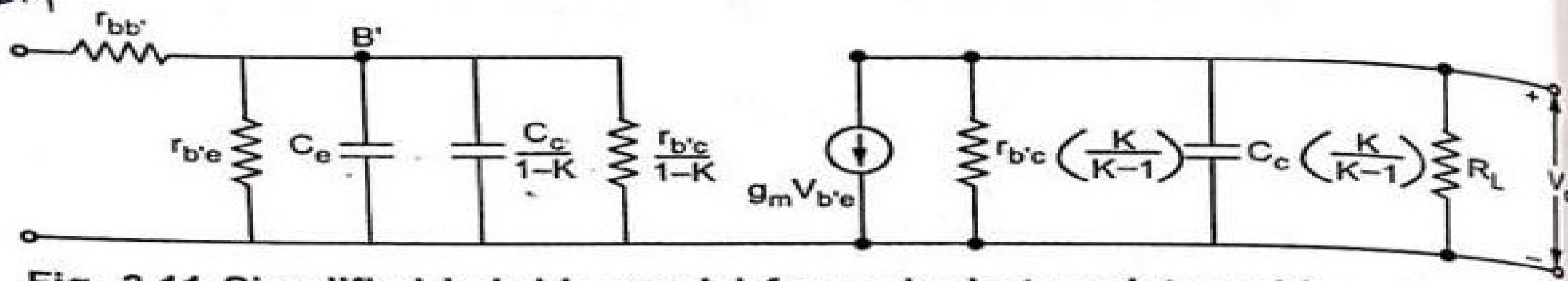


Fig. 3.11 Simplified hybrid-π model for a single transistor with resistive load

Further simplification of input circuit

The value $r_{b'c}/(1 - K) >> r_{b'e}$ (1 K) and hence $r_{b'c}/(1 - K)$ which is in pa
with $r_{b'e}$ can be neglected.

C_C also resolved by Miller's theorem.

$$\therefore \qquad \frac{\frac{1}{j\omega C_C}}{1-K} = \frac{1}{j\omega C_C\,(1+g_m R_L)}$$

$$\therefore \qquad \frac{C_C}{1-K} = C = C_C\,(1 + g_m R_L)$$

As C_e and C are in parallel, the total equivalent capacitance is given as

$$C_{eq} = C_e + C_C\,(1 + g_m R_L)$$

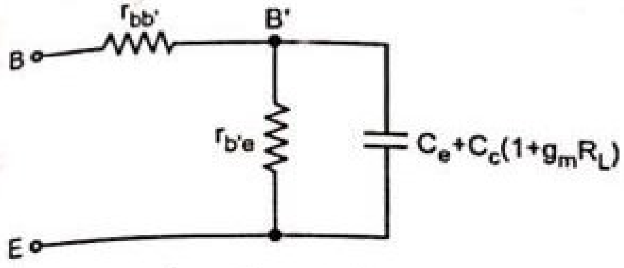


Fig. 3.12

From equation (2) we can say that input capacitance is increased. $C_C\,(1 + g_m R_L)$ is called **Miller capacitance** (C_M). With these approximations input circuit becomes, as shown in Fig. 3.12.

Further simplification for output circuit

At output circuit value of C_C can be calculated as

$$\frac{\frac{1}{j\omega C_C}}{\frac{K-1}{K}} \approx \frac{1}{j\omega C_C} \qquad \because K = -100$$

$$\therefore \qquad C_C\left(\frac{K}{K-1}\right) \approx C_C$$

This value of $r_{b'c}$ is very high in comparison with load resistance R_L which is parallel with $r_{b'c}$. Hence $r_{b'c}$ can be ignored.

Fig. 3.13 shows the further simplified hybrid-π model of single transistor in CE configuration with load resistance.

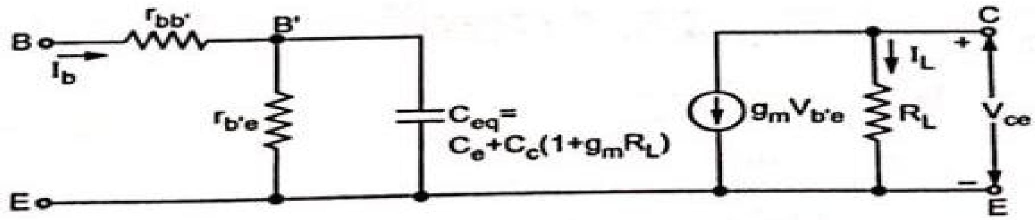


Fig. 3.13 Further simplified hybrid-π model for CE with R_L

Parallel combination of $r_{b'e}$ and C_{eq} is given as

$$Z = \frac{r_{b'e} \times \dfrac{1}{j\omega C_{eq}}}{r_{b'e} + \dfrac{1}{j\omega C_{eq}}}$$

$$= \frac{r_{b'e}}{1 + j\omega\, r_{b'e}\, C_{eq}}$$

This gives equivalent circuit as shown in Fig. 3.14.

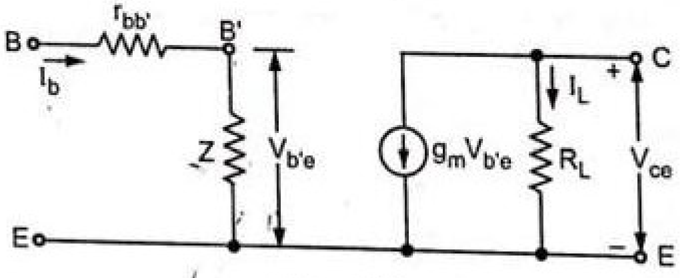


Fig. 3.14

Looking at Fig. 3.14 we can write

$$V_{b'e} = I_b Z$$

$$\therefore \quad Z = \frac{V_{b'e}}{I_b}$$

The current gain for the circuit shown in Fig. 3.14 can be given as

$$A_i = \frac{I_L}{I_b} = \frac{-g_m V_{b'e}}{I_b}$$

Substituting value of $V_{b'e}/I_b$ from equation (4) we get,

$$A_i = -g_m Z$$

Substituting value of Z from equation (3) we get,

$$A_i = \frac{-g_m r_{b'e}}{1 + j\omega r_{b'e} C_{eq}} = \frac{-h_{fe}}{1 + j 2\pi f r_{b'e} C_{eq}} \quad \because h_{fe} = g_m r_{b'e}.$$

Let

$$f_H = \frac{1}{2\pi r_{b'e} C_{eq}}$$

$$A_i = \frac{-h_{fe}}{1 + j\left(\frac{f}{f_H}\right)}$$

$$|A_i| = \frac{h_{fe}}{\sqrt{1 + \left(\frac{f}{f_H}\right)^2}}$$

At $f = f_H$

$$A_i = \frac{h_{fe}}{\sqrt{2}}$$

The f_H is the frequency at which the transistor's gain drops by 3 dB or $1/\sqrt{2}$ ıes from its value at low frequency. It is given as

$$f_H = \frac{1}{2\pi r_{b'e} C_{eq}}$$

$$= \frac{1}{2\pi r_{b'e} [C_e + C_C (1 + g_m R_L)]} \quad \dots (7)$$

At $R_L = 0$

$$f_H = \frac{1}{2\pi r_{b'e} [C_e + C_C]} = f_\beta \quad \dots (8)$$

From equation we can say that maximum possible value for f_H is f_β. As R_L creases C_{eq} increases decreasing f_H, as shown in Fig. 3.15.

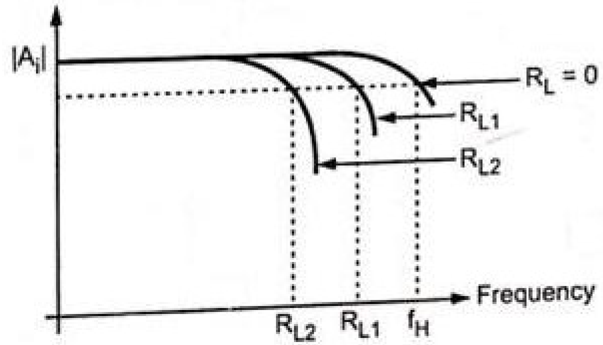


Fig. 3.15 Variation f_H with R_L

Parameter f_β: The frequency at which the transistor short circuit CE current gain drops by 3db or $1/\sqrt{2}$ times from its value at low frequency. It is given by

... given as

$$f_\beta = \frac{1}{2\pi r_{b'e}(C_e + C_C)}$$

or

$$= \frac{g_{b'e}}{2\pi(C_e + C_C)} \qquad \text{... (7)}$$

or

$$= \frac{1}{h_{fe}} \frac{g_m}{2\pi(C_e + C_C)} \qquad \because g_{b'e} = \frac{1}{r_{b'e}} = \frac{g_m}{h_{fe}}$$

5.3 Parameter f_T

It is the frequency at which short circuit CE current gain becomes unity.

At $f = f_T$, equation (6) becomes

$$1 = \frac{h_{fe}}{\sqrt{1+\left(\frac{f_T}{f_\beta}\right)^2}} \qquad \text{... (10)}$$

6.2 Parameter f_α

It is the frequency at which the transistor's short circuit CB current gain drops by dB or $1/\sqrt{2}$ times from its value at low frequency. The expression for f_α can be erived is the similar manner as for f_β.

The current gain for CB configuration is given as

$$A_i = \frac{-h_{fb}}{1 + j\frac{f}{f_\alpha}}$$

where

$$f_\alpha = \frac{1}{2\pi r_{b'e}(1 + h_{fb})C_e}$$

$$= \frac{1 + h_{fe}}{2\pi r_{b'e} C_e} \approx \frac{h_{fe}}{2\pi r_{b'e} C_e} \quad \dots (8)$$

$$|A_i| = \frac{h_{fb}}{\sqrt{1 + \left(\frac{f}{f_\alpha}\right)^2}} \quad \dots (9)$$

At

$$f = f_\alpha$$

$$|A_i| = \frac{h_{fb}}{\sqrt{2}}$$

9 Gain Bandwidth Product

9.1 Gain Bandwidth Product for Voltage

The gain bandwidth product for voltage gain is given as

$$|A_{vs\,low}\, f_H| = |A_{vso}\, f_H| = \frac{-h_{fe} R_L}{R_s + h_{ie}} \times \frac{1}{2\pi R_{eq} C_{eq}}$$

$$= \frac{-h_{fe} R_L}{R_s + h_{ie}} \times \frac{1}{2\pi C_{eq}\left[\dfrac{r_{b'e}(r_{bb'} + R_s)}{r_{b'e} + r_{bb'} + R_s}\right]}$$

$$= \frac{-h_{fe} R_L}{R_s + h_{ie}} \times \frac{1}{2\pi C_{eq}\left[\dfrac{r_{b'e}(r_{bb'} + R_s)}{(R_s + h_{ie})}\right]}$$

$\because \quad h_{ie} = r_{b'e} + r_{bb'}$

$$= \frac{-h_{fe} R_L}{2\pi C_{eq}\, r_{b'e}\,(r_{bb'} + R_s)}$$

$$= \frac{-g_m\, r_{b'e}\, R_L}{2\pi C_{eq}\, r_{b'e}\,(r_{bb'} + R_s)} \qquad \because h_{fe} = g_m\, r_{b'e}$$

$$= \frac{-g_m R_L}{2\pi C_{eq}\,(r_{bb'} + R_s)} \qquad \ldots (1)$$

This equation can be further simplified as follows

$$|A_{vso} \times f_H| = \frac{g_m}{2\pi[C_e + C_C[1 + g_m R_L]]} \times \frac{R_L}{R_s + r_{bb'}}$$

$$\because \quad C_{eq} = C_e + C_C(1 + g_m R_L)$$

$$= \frac{g_m}{2\pi[C_e + C_C g_m R_L]} \times \frac{R_L}{R_s + r_{bb'}} \quad \because g_m R_L >> 1$$

$$= \frac{R_L}{R_s + r_{bb'}} \times \frac{2\pi f_T C_e}{2\pi[C_e + C_C(2\pi f_T C_e)R_L]}$$

$$\because \quad g_m = 2\pi f_T C_e$$

$$= \frac{R_L}{R_s + r_{bb'}} \times \frac{2\pi C_e f_T}{2\pi C_e[1 + 2\pi f_T C_C R_L]}$$

$$= \frac{R_L}{R_s + r_{bb'}} \times \frac{f_T}{1 + 2\pi f_T C_C R_L}$$

$$\therefore \quad |A_{vso} \times f_H| = \frac{R_L}{R_s + r_{bb'}} \times \frac{f_T}{1 + 2\pi f_T C_C R_L}$$

3.9.2 Gain Bandwidth Product for Current

The gain bandwidth product for current gain is given as

$$|A_{is\,low} \times f_H| = |A_{iso} \times f_H| = \frac{-h_{fe} R_s}{R_s + h_{ie}} \times \frac{1}{2\pi R_{eq} C_{eq}}$$

By similar analysis with replacement of $-h_{fe} R_s$ instead of $-h_{fe} R_L$ we ge

$$|A_{iso} \times f_H| = \frac{g_m R_s}{2\pi C (R_s + r_{bb'})} = \frac{f_T}{1 + 2\pi f_T C_C R_L} \cdot \frac{R_s}{R_s + r_{bb'}}$$

The quantities f_H, A_{iso} and A_{vso} which characterize the transistor stage, on both R_L and R_s. The dependence between R_s and R_L with quantities f_H, A A_{vso} is shown in Fig. 3.18.

Individual lower cut-off frequencies produced by an external capacitors for BJT

We know that $$f = \frac{1}{2\pi RC}$$

$$f_c = \frac{1}{2\pi RC}$$

lower cut-off frequencies produced by an external capacitors C_B, C_C, C_E

Overall cut-off frequency of an amplifier $$f_L = 1.1\sqrt{f_{LB}^2 + f_{LC}^2 + f_{LE}^2}$$

Individual cut-off frequencies produced by an external capacitors C_B, C_C, C_E is f_{LB}, f_{LC}, f_{LE}

$$f_{LB} = \frac{1}{2\pi(R_S + R_i')C_B}$$

$$f_{Lc} = \frac{1}{2\pi(R_0 + R_L\ )C_C}$$

$$f_{LE} = \frac{1}{2\pi\left[\left(\frac{h_{ie} + R_{th}}{\beta}\right) // R_E\right] C_E}$$

$R_{th} = R1//R2//Rs$

For FET

$$f_{LG} = \frac{1}{2\pi(R_S + R_G)C_G}$$

$$f_{LD} = \frac{1}{2\pi(R_L + R_S')C_B}$$

$$f_{LS} = \frac{1}{2\pi C_S}$$

Overall lower cutoff frequency FET

$$f_L = 1.1\sqrt{f_{LG}^2 + f_{LD}^2 + f_{LS}^2}$$

Overall higher cutoff frequency:

$$f_H = \frac{1.1}{\frac{1}{f_{H1}^2} + \frac{1}{f_{H2}^2} + \frac{1}{f_{H3}^2} + \cdots\ldots}$$

The gain of multistage amplifiers at any frequency below lower cut-off frequency of Individual stage can be given as

$$A_V = \frac{A_{mid}^n}{\left[1 + \left(\frac{f_L}{f}\right)^2\right]^{n/2}}$$

The gain of multistage amplifiers at any frequency above higher cut-off frequency of Individual stage can be given as

$$A_V = \frac{A_{mid}^n}{\left[1 + \left(\frac{f}{f_H}\right)^2\right]^{n/2}}$$

Lower 3dB frequency:

Lower 3db frequency of individual stages

$$f_L(n) = \frac{f_{L(individual)}}{\sqrt{2^{1/n} - 1}}$$

f_L _is lower 3dB of single stage

Upper 3db frequency:

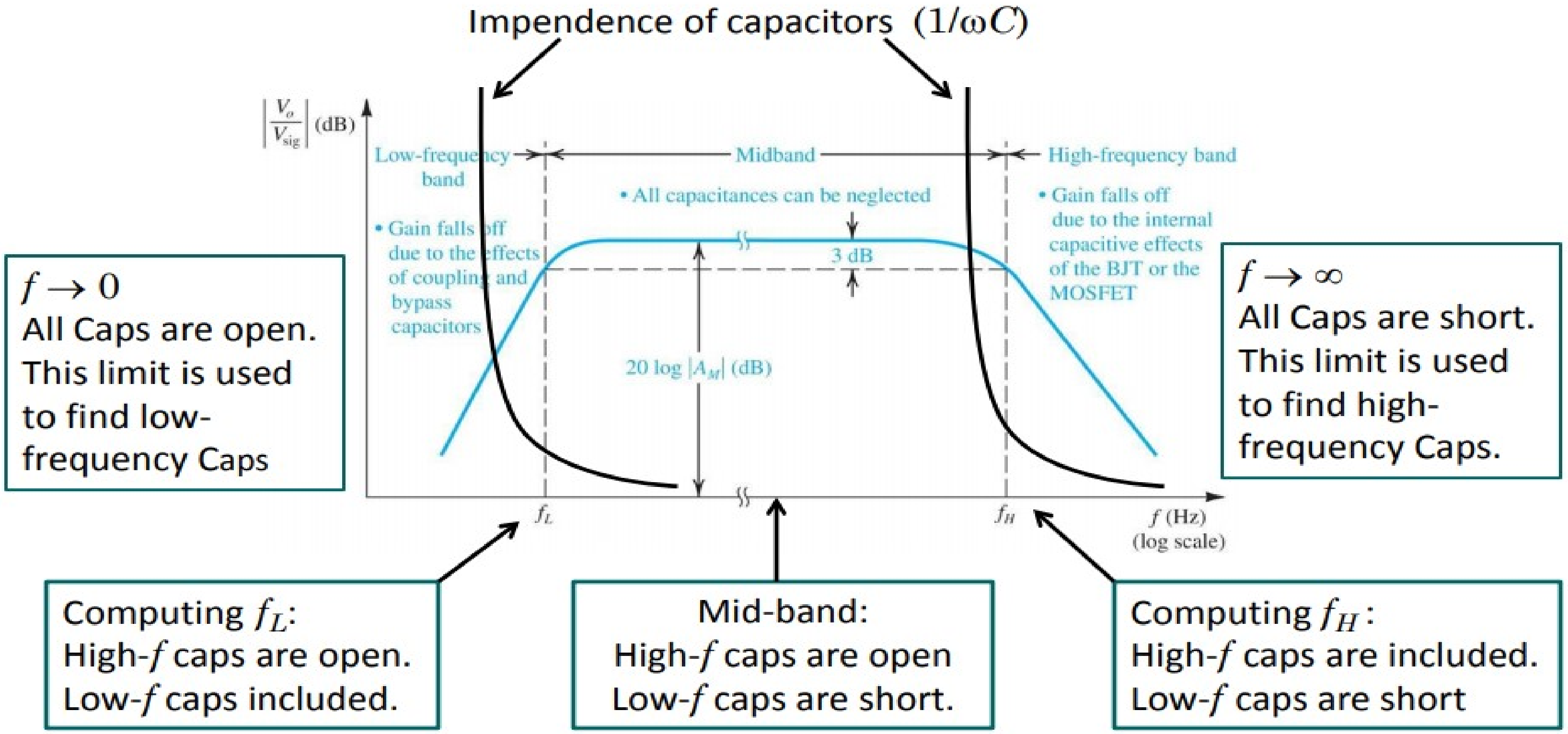
$$f_H(n) = f_{H(ind)}\sqrt{2^{1/n} - 1}$$

Overall Bandwidth:

$$B.W(n) = B.W_{(ind)}\sqrt{2^{1/n} - 1}$$

Frequency response of amplifier:

Impact of various capacitors depend on the frequency of interest



Frequency Range and Capacitance

$X_C = 1/(\omega C)$

	coupling & bypass (high)	parasitic cap (low) pF
low frequency (LF)	consider	open
mid frequency (MF)	short	open
high frequency (HF)	short	consider