MULTISTAGE AMPLIFIERS

INTRODUCTION:

* The gain of an amplifier does depend on the parameters of the device and circuit components, there exists an upper theoretical limit for the gain obtainable from single stage.
* The Voltage amplification or power gain or frequency response obtained with a single stage of amplification is usually not sufficient to meet the needs of either a composite electronic circuit or load device.

* For Example:

A Speaker represents a heavy load in an audio amplifier system, and several amplifier stages may be required to "boost a signal" originating at a microphone or magnetic tape head to a level sufficient to provide a large amount of Power to the Speaker.

* We hear of pre-amplifiers, power amplifiers and output amplifiers, all of which constitute stages of amplification in such a system.
* It may be emphasized here that a practical amplifier is always a multistage amplifier that may provide a higher Voltage or Current gain or both.

CLASSIFICATION OF AMPLIFIERS:

* A circuit that increases the amplitude of the given i/p signal is an amplifier.

* Amplifiers can be classified as follows:

1) Based on transistor configuration:

a) Common Emitter Amplifier

b) Common collector Amplifier

c) Common Base Amplifier

2) Based on the Active device

a) BJT amplifier

b) FET amplifier

3) Based on the Q-point (operating Condition)

a) Class-A amplifier

b) Class B Amplifier

c) class AB Amplifier

d) Class C Amplifier

4) Based on the Number of Stages

a) Single stage amplifiers

b) Multi stage amplifiers

5) Based on the output

a) Voltage Amplifier

b) Power amplifier

6) Based on the Frequency Response

a) Audio frequency (AF) Amplifier

b) Intermediate frequency (IF) Amplifier

c) Radio frequency (RF) Amplifier

7) Based on the Bandwidth

a) Narrow band amplifier (normally RF amplifier)

b) Wide band amplifier (normally Video amplifier)

MULTISTAGE AMPLIFIER:

Definition: An Amplifier that produce Voltage, Current or power gain through the use of two or more stages is called multistage Amplifier.

* It is to be noted that the output of the first stage makes the input for the Second stage, the output of the second stage makes the input for third stage and so on.

* A multistage amplifier can be represented by a block diagram, as shown.

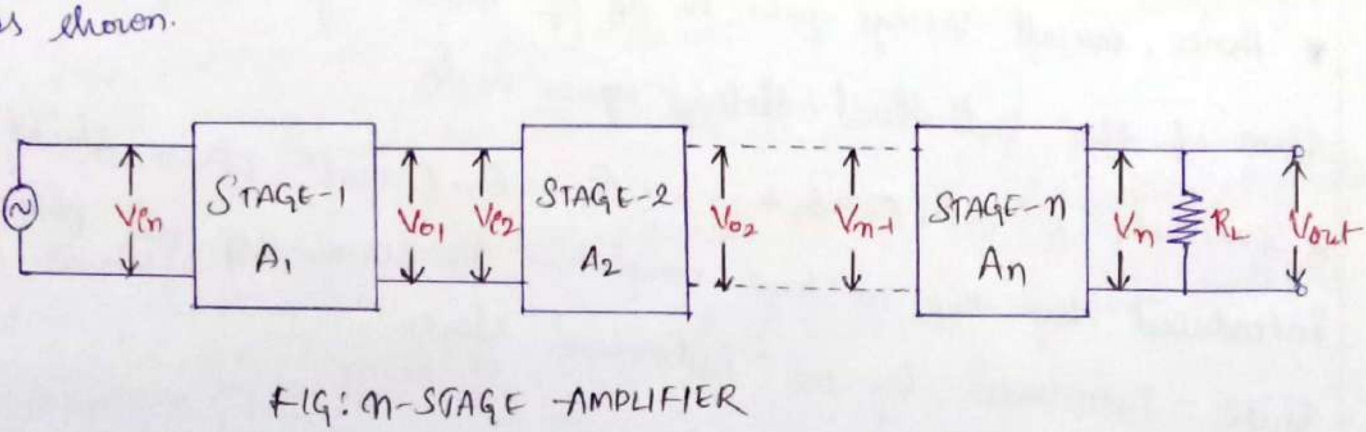


FIG: n-STAGE AMPLIFIER

* The Signal Voltage V_s is applied to the input of 1^{st} Stage and final output is available at the output terminal of last stage.

* let $A_{v1}, A_{v2}, A_{v3} \ldots A_{vn-1}, A_{vn}$ are Voltage gains of Individual stages. A_v be the overall Voltage gain.

* By definition of Voltage gain

$$A_v = \frac{V_o}{V_i}, \quad A_{v1} = \frac{V_{o1}}{V_{in1}} \quad A_{v2} = \frac{V_{o2}}{V_{i2}} \ldots \quad A_{vn-1} = \frac{V_{on-1}}{V_{in-1}} \quad A_{vn} = \frac{V_{on}}{V_{in}}$$

Here, $A_v = \frac{V_{o1}}{V_{i1}} * \frac{V_{o2}}{V_{i2}} * \ldots \ldots \frac{V_{on-1}}{V_{in-1}} * \frac{V_{on}}{V_{in}}$

$$\boxed{A_V = A_{v1} * A_{v2} \ldots \ldots A_{vn-1} * A_{vn}} \rightarrow (1)$$

* Hence, Overall Voltage gain of n-Stage amplifier is the product of Voltage gains of individual stage.

Taking logarithm on both sides for eq(1)

$$20 \log_{10} A_V = 20 \log_{10} [A_{v1} * A_{v2} \ldots \ldots A_{vn-1} * A_{vn}]$$

$$20 \log_{10} A_V = 20 \log_{10} A_{v1} + 20 \log_{10} A_{v2} \ldots \ldots 20 \log_{10} A_{vn-1} + 20 \log_{10} A_{vn}$$

$$A_V(d_B) = A_{v1}(d_B) + A_{v2}(d_B) + \ldots \ldots + A_{vn-1}(d_B) + A_{vn}(d_B)$$

* Hence, overall Voltage gain in d_B of multistage amplifier is sum of the individual Voltage gain in d_B.

* Similarly $\theta = \theta_1 + \theta_2 + \theta_3 + \ldots \theta_{n-1} + \theta_n$. Overall phase shift introduced by the n-stage amplifier is sum of the phase shifts introduced by the individual stages.

* In a multistage amplifier, the output of first stage is combined to the next stage through a coupling device. The process is known as Cascading.
* The process of joining two amplifier stages using a coupling device is called Cascading.

Need of Coupling:

→ To transfer the AC output of one stage to the i/p of next stage.

→ Block the dc to pass from one stage to the next stage i.e., to isolate dc conditions.

* For an ideal coupling network the following requirements should be fullfilled.

i. The direct currents should not pass through the coupling Network.

ii. AC signal waveform should transfer from one amplifier to next amplifier without distortion.

iii. Some voltage loss of signal cannot be avoided in coupling Network but this loss should be minimum just negligible.

iv. Coupling Network impedance should not be frequency dependent.

* Unfortunately, there is no coupling Network which fulfills all the demands.

TYPES OF COUPLING NETWORKS:

* There are three types of interstage Coupling Networks. They are,

1. R-C Coupling
2. Transformer Coupling
3. Direct Coupling.

1. R-C COUPLING:

* The Circuit diagram of RC Coupled Amplifier is as shown below.

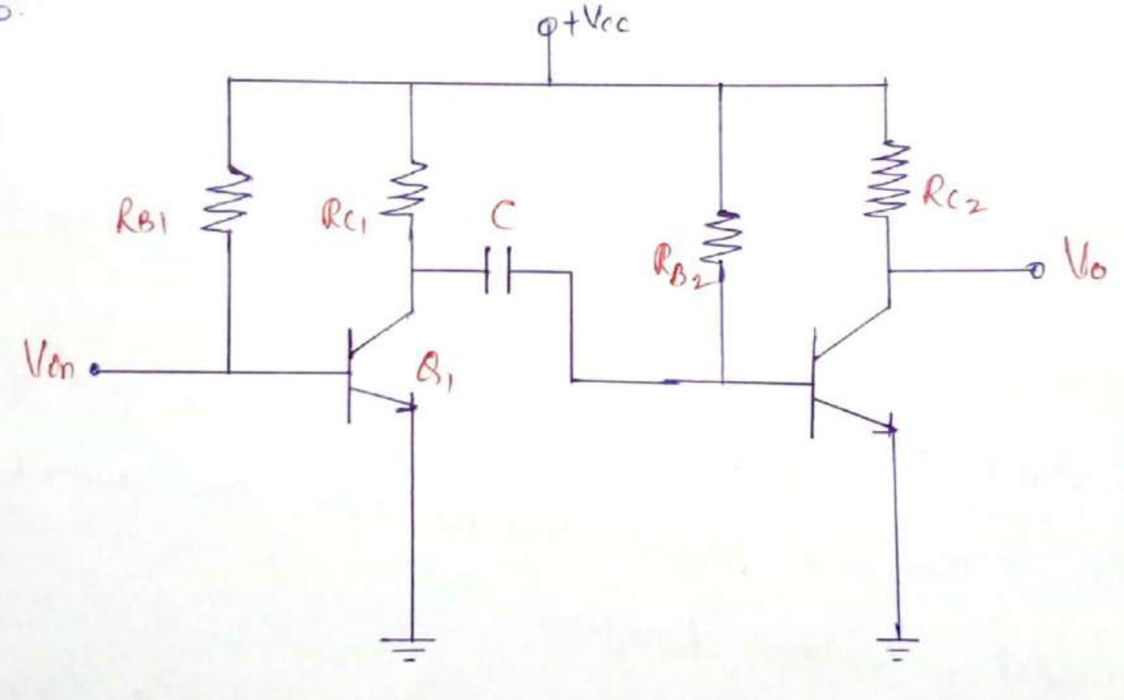


FIG: R-C COUPLED AMPLIFIER

* In RC Coupling AC output of first stage is given to the input of second stage by a collector resistor and Coupling Capacitor.
* Coupling Network consists of R and C Components.
* The Coupling Capacitor (C) isolates the DC Conditions of one stage from the next stage.

* The Coupling Capacitor (C_c) isolates the DC Conditions of one stage (7)
acts as short circuit for AC signals and open circuit for DC
Conditions.

* The amplifier using this type of RC Coupling are called
RC Coupled Amplifier.

APPLICATIONS:

1. Used in all Audio Small signal amplifiers
2. Used in tape recorders, Radio Receivers, TV receivers.

2. TRANSFORMER COUPLING:

* The Coupling device in Coupling network is a transformer.
* In this method primary winding of the transformer acts as
collector load and Secondary winding transfers ac output signal
directly to the base of next stage.
* This type of Coupling increases the overall Circuit gain and the
level of interstage impedance matching.
* This is restricted to power amplifiers where efficiency and
impedance matching are critical Requirements.

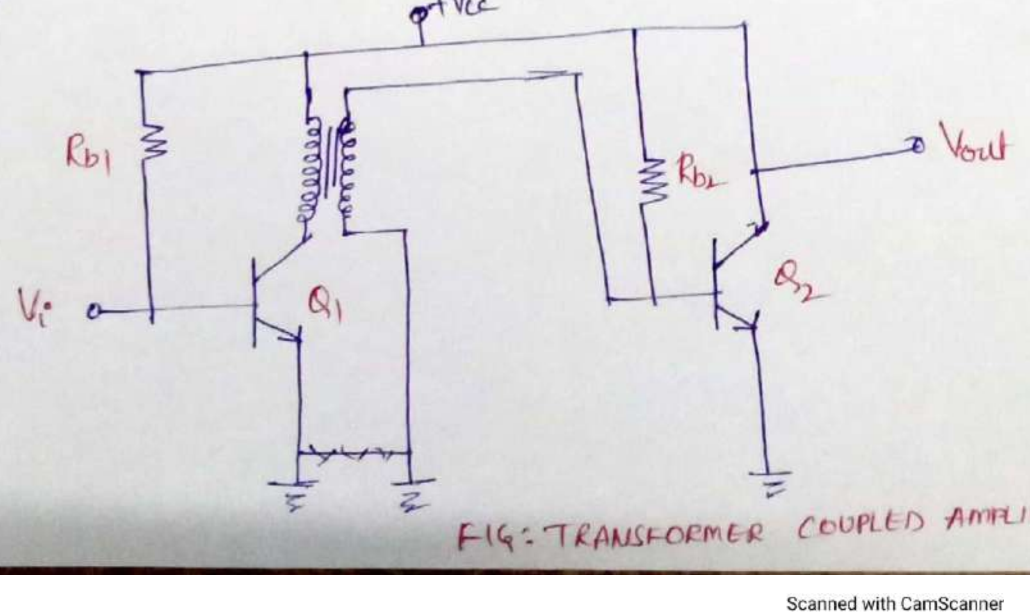


FIG: TRANSFORMER COUPLED AMPLIFIER

* The amplifiers using Transformer Coupling are called Transformer Coupled Amplifiers. (8)

3. DIRECT COUPLING:

* In this method ac o/p signal is directly fed to the input of next stage.
* Here, difficulty is included in the Coupling network, biasing conditions are disturbed to avoid this special dc Voltage load, circuits are used to match the Output dc loads.
* It is used where amplifications of low frequency is to be done.
* Coupling devices like capacitor, transformers are not used because of their size Complexity.

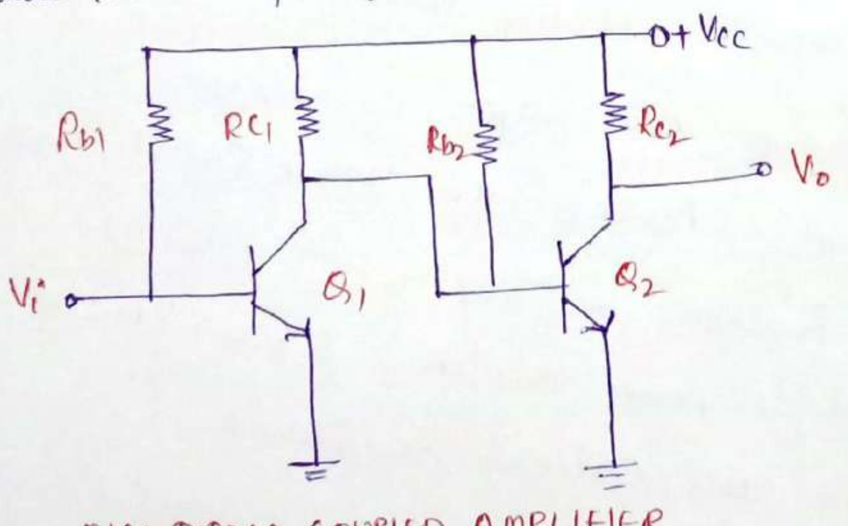
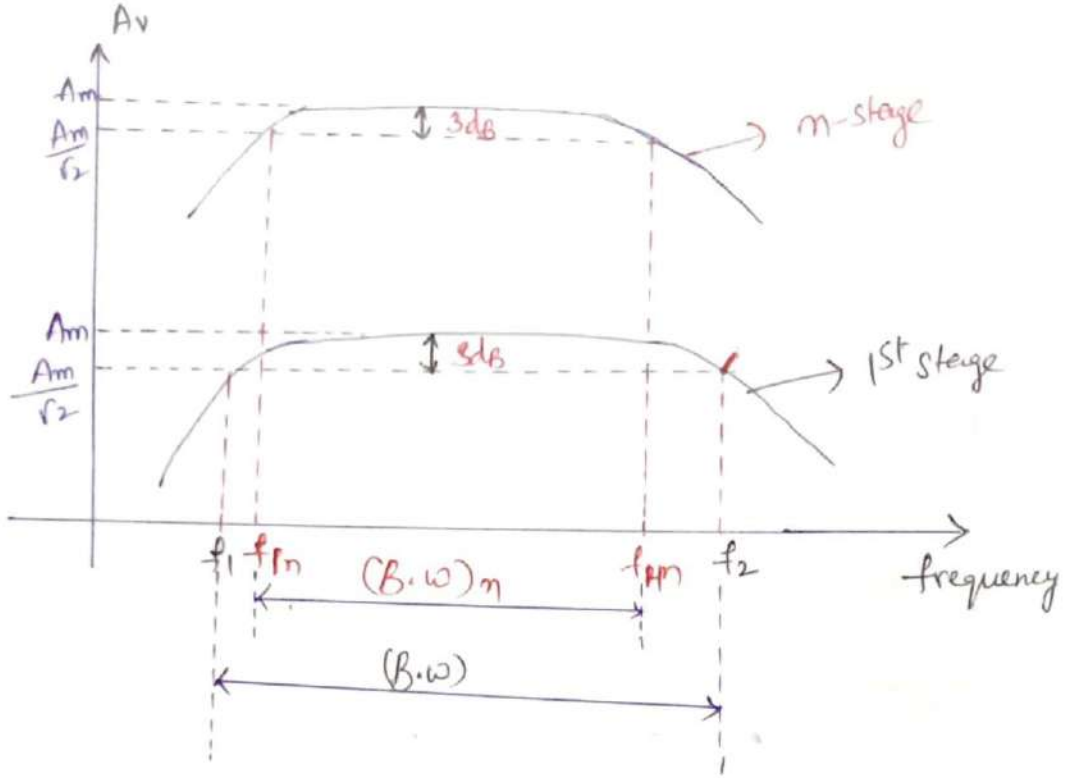


FIG: DIRECT COUPLED AMPLIFIER

FREQUENCY RESPONSE OF MULTISTAGE AMPLIFIERS:

* Frequency response of multistage amplifier is as shown below



* For a single stage CE amplifier voltage gain at low frequencies and high frequencies are given as,

$$A_L = \frac{A_m}{1 - j\left(\frac{f_1}{f}\right)} \quad , \quad A_H = \frac{A_m}{1 + j\left(\frac{f}{f_2}\right)}$$

∴ where f_1 & f_2 are lower and upper cut-off frequencies at which voltage gain falls to $1/\sqrt{2}$ of its maximum value.

$$|A_L| = \frac{A_m}{\sqrt{1 + \left[\frac{f_1}{f}\right]^2}} \quad \& \quad |A_H| = \frac{A_m}{\sqrt{1 + \left[\frac{f}{f_2}\right]^2}}$$

OVERALL LOWER CUT-OFF FREQUENCY (f_{Ln})

⇒ For n-stage Cascade amplifier

$$(A_L)^n = \left[\frac{A_m}{1 - j\left[\frac{f_1}{f}\right]}\right]^n$$

$$\left|\frac{A_L}{A_m}\right|^n = \left[\frac{1}{\sqrt{1+\left[\frac{f_1}{f}\right]^2}}\right]^n \rightarrow (1)$$

At $f = f_{Ln}$

$$\left|\frac{A_L}{A_m}\right| = \left[\frac{1}{\sqrt{2}}\right] \rightarrow (2)$$

Equating (1) & (2)

$$\left[\frac{1}{\sqrt{2}}\right] = \left[\frac{1}{\sqrt{1+\left[\frac{f_1}{f}\right]^2}}\right]^n$$

$$\frac{1}{\sqrt{1+\left[\frac{f_1}{f}\right]^2}} = \left[\frac{1}{\sqrt{2}}\right]^{1/n}$$

$$\sqrt{1+\left[\frac{f_1}{f}\right]^2} = \left[\sqrt{2}\right]^{1/n}$$

Squaring on both sides

$$1+\left(\frac{f_1}{f_{Ln}}\right)^2 = 2^{1/n}$$

$$\left(\frac{f_1}{f_{Ln}}\right)^2 = 2^{1/n}-1$$

$$\frac{f_1}{f_{Ln}} = \sqrt{2^{1/n}-1}$$

$$\boxed{f_{Ln} = \frac{f_1}{\sqrt{2^{1/n}-1}}}$$

OVERALL UPPER CUT-OFF FREQUENCY f_{Hn}:

$$\left|\frac{A_H}{A_m}\right|^n = \left[\frac{1}{\sqrt{1+\left[\frac{f}{f_2}\right]^2}}\right]^n \rightarrow ①$$

At $f_2 = f_{Hn} \Rightarrow \left|\frac{A_H}{A_m}\right|^n = 1/\sqrt{2} \rightarrow ②$

Equating ① & ②

$$\frac{1}{\sqrt{2}} = \left[\frac{1}{\sqrt{1+\left[\frac{f}{f_2}\right]^2}}\right]^n$$

$$\frac{1}{\sqrt{1+\left[\frac{f}{f_2}\right]^2}} = \left[\frac{1}{\sqrt{2}}\right]^{1/n}$$

$$\sqrt{1+\left[\frac{f}{f_2}\right]^2} = \left[\sqrt{2}\right]^{1/n}$$

Squaring on both sides

$$1+\left[\frac{f}{f_2}\right]^2 = 2^{1/n}$$

$$\left[\frac{f}{f_2}\right]^2 = 2^{1/n}-1$$

$$\left[\frac{f_{Hn}}{f_2}\right] = \sqrt{2^{1/n}-1}$$

$$\boxed{f_{Hn} = \frac{f_2}{\sqrt{2^{1/n}-1}}}$$

ANALYSIS OF CASCADED RC COUPLED AMPLIFIERS :

* The most popular cascade amplifier is formed by cascading several CE amplifier stages.
* The n-stage CE amplifier is as shown below.

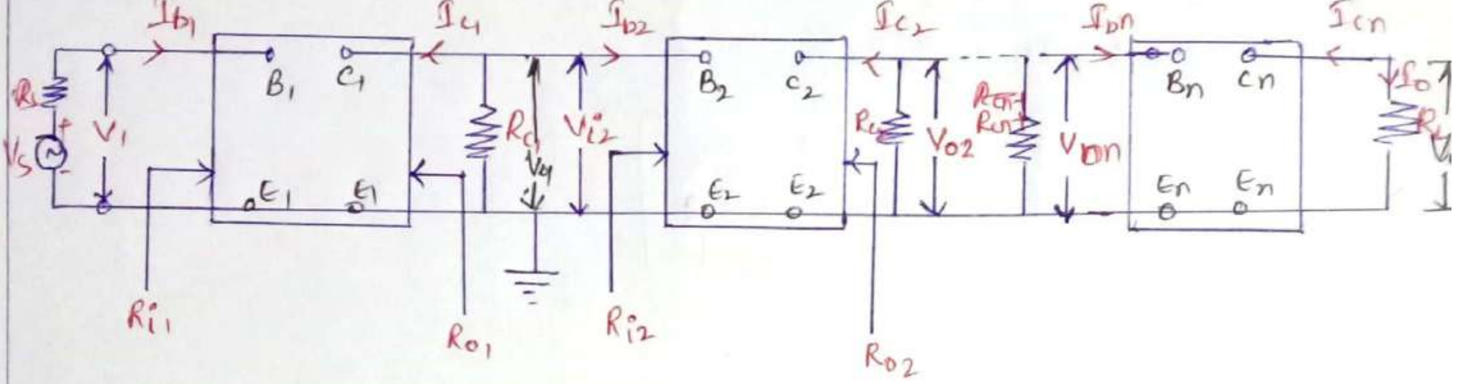


FIG: n-STAGE CE AMPLIFIER

VOLTAGE GAIN:

* In multistage amplifier the output voltage of first stage acts as the input voltage of second stage and so on. The voltage gain of the complete cascade amplifier is equal to the product of the voltage gains of the individual stages.

* The voltage gain of first stage is $A_{V1} = \frac{V_{o1}}{V_{i1}}$

* The voltage gain of second stage is $A_{V2} = \frac{V_{o2}}{V_{i2}}$

Similarly overall voltage gain $A_V = \frac{V_o}{V_i}$

$$\frac{V_o}{V_i} = \frac{V_{o1}}{V_{i1}} * \frac{V_{o2}}{V_{i2}} \;------\; \frac{V_{on-1}}{V_{in-1}} * \frac{V_{on}}{V_{in}}$$

$$A_V = A_{V1} * A_{V2} \;------\; A_{Vn-1} * A_{Vn}$$

$$\theta = \theta_1 + \theta_2 + \;----\; \theta_{n-1} + \theta_n$$

* From above expressions, we can conclude that,

i) The magnitude of the Resultant Voltage gain equals to the product of the magnitudes of the Voltage gains of the individual stages.

ii) The phase shift of the resultant voltage gain equals to the sum of the phase shifts of the individual stages.

* The following figure shows a particular stage say the k^{th} stage of the n stage Cascaded amplifier.

→ The Voltage gain of the k^{th} stage is given by,

$$A_{VK} = \frac{A_{IK} R_{LK}}{R_{IK}}$$

* Where R_{LK} is the effective load impedance at the collector of the k^{th} Stage and R_{iK} is the input impedance of the k^{th} stage.

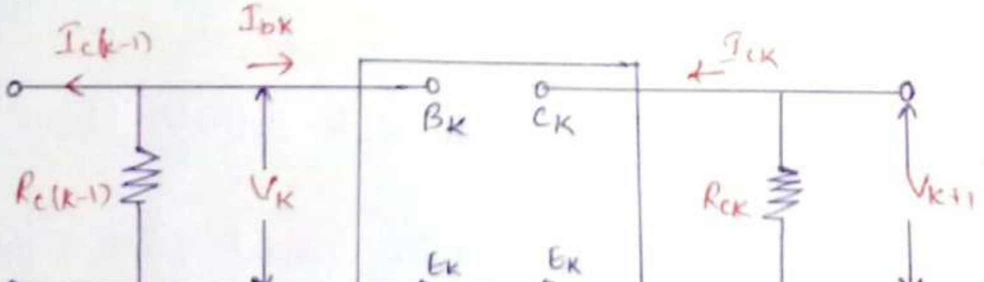


FIG: K^{TH} STAGE OF A CASCADED AMPLIFIER

→ the Current gain $A_{in} = \dfrac{-h_{fe}}{1 + h_{oe} R_{Ln}}$

$$R_{in} = h_{ie} + h_{re} A_{In} R_{Ln}$$

Where R_{Ln} is the effective load impedance for the last stage and equals R_{cn}.

where $$R_{L(n-1)} = \frac{R_{c(n-1)} * R_{in}}{R_{c(n-1)} + R_{in}}$$

$$A_{I(n-1)} = \frac{-h_{fe}}{1 + h_{oe} R_{L(n-1)}}$$

$R_{i(n-1)}$ can be found from $R_{i(n-1)} = h_{ie} + h_{re} A_{I(n-1)} R_{L(n-1)}$

CURRENT GAIN:

=) A_I is the current gain of the complete n-stage amplifier

$$A_I = \frac{I_o}{I_{b1}} = \frac{-I_{cn}}{I_{b1}}$$

$$\frac{-I_{cn}}{I_{b1}} = \frac{-I_{c1}}{I_{b1}} * \frac{I_{c2}}{I_{c1}} \cdots \frac{I_{cn}}{I_{c(n-1)}}$$

$$A_I = A_{I1} * A'_{I2} \cdots A'_{In}$$

where A_{I1} is the base to collector current gain

$A'_{I2}, A'_{I3} \ldots$ are the collector to collector current gains

* for k^{th} stage the collector to collector current gain is given by,

$$A'_{IK} = \frac{I_{cK}}{I_{c(K-1)}} \rightarrow ① \qquad A_{IK} = \frac{I_{cK}}{I_{bK}}$$

$$A'_{IK} = \frac{I_{cK}}{I_{bK}} * I_{c(K-1)}$$

multiply and divide eq ① by I_{bK}

$$A'_{IK} = \frac{I_{cK}}{I_{c(k-1)}} * \frac{I_{bK}}{I_{bK}}$$

$$= \frac{I_{cK}}{I_{bK}} * \frac{I_{bK}}{I_{c(K-1)}}$$

$$A'_{IK} = A_{IK} * \frac{I_{bK}}{I_{c(K-1)}}$$

From the figure $I_{bK} = \frac{R_{c(k-1)}}{R_{c(K-1)} + R_{iK}} * I_{c(k-1)}$

$$\frac{I_{bK}}{I_{c(K-1)}} = \frac{R_{c(K-1)}}{R_{c(K-1)} + R_{inK}}$$

$$\boxed{A'_{IK} = A_{IK} * \frac{R_{c(K-1)}}{R_{c(K-1)} + R_{inK}}}$$

Power gain : $A_P = \frac{V_0}{V_1} * \frac{I_0}{I_{b1}} = \frac{-V_0}{V_1} \frac{I_{cn}}{I_{b1}}$

$$A_P = A_V A_I$$

$$A_V = A_I \frac{R_{cn}}{R_{i1}}$$

$$\boxed{A_P = A_I^2 \frac{R_{cn}}{R_{i1}}}$$

TWO STAGE RC COUPLED AMPLIFIER:

* The two stage R-C coupled amplifier using Common Emitter Configuration is as shown.

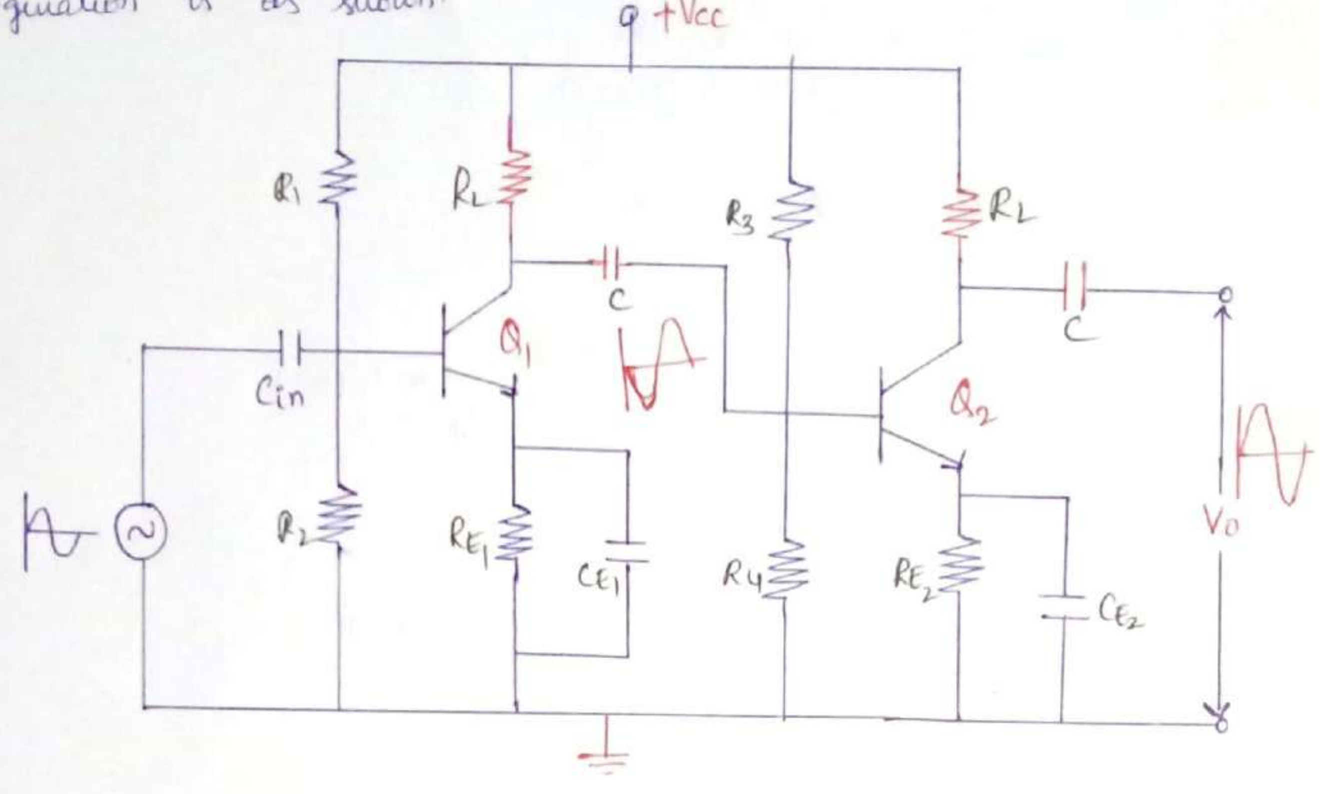


FIG: CIRCUIT DIAGRAM OF TWO STAGE RC COUPLED AMPLIFIER

* The output of the first stage is coupled to the input of second stage through coupling capacitor C followed by a shunt connection of resistor. Therefore the amplifier is known as Resistance Capacitance Coupled Amplifier.
* The resistors R_1, R_2 and R_e form the biasing and stabilization network.
* The bypass capacitor C_e prevents the loss of amplification.
* The coupling capacitors are also known as Blocking capacitors which blocks DC and allows AC Voltage.

OPERATION :

* AC Signal is applied to the base of Q_1 transistor, the amplified Signal will develop across the collector of Q_1 transistor.

* The amplified Signal is Connected to the base transistor through Coupling capacitor 'C'.

* The 2nd Stage can be used for further amplification of the input Signal so the Cascaded stages amplifies the signal and then the overall gain increases.

* The 1st stage output is out of phase with the input Signal. This out of phase signal is fed to the i/p of the next stage.

* Therefore the overall output Voltage of 2nd Stages is in phase with the i/p signal with higher amplitude

FREQUENCY RESPONSE :

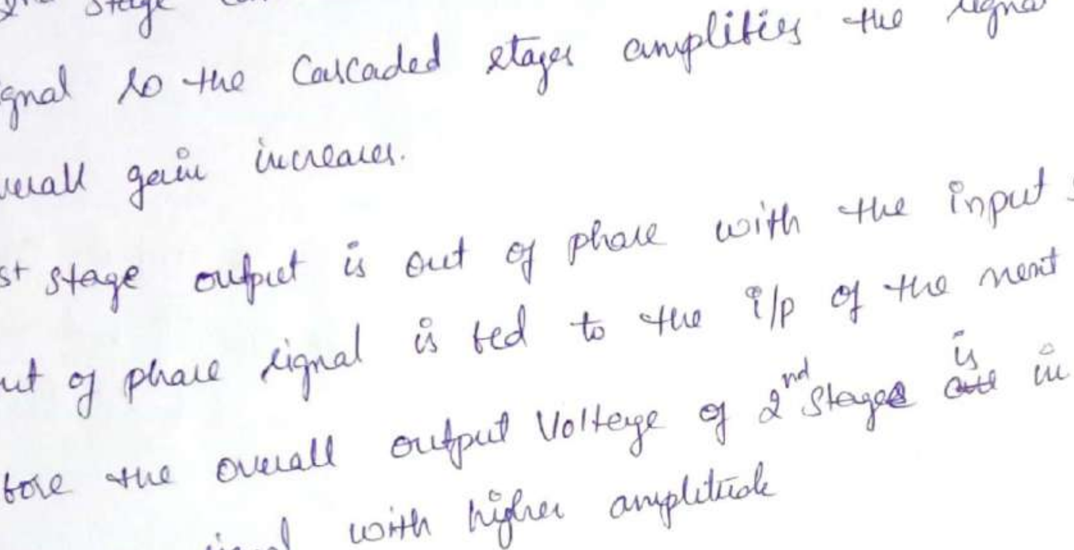


FIG: FREQUENCY RESPONSE OF TWO STAGE RC COUPLED AMPLIFIER

* It can be seen from the figure i.e., frequency response curve that the voltage gain is maximum and constant at mid frequencies.

* The Voltage gain is low at lower and higher frequencies.

AT LOW FREQUENCIES:

* At low frequencies, the reactance of the coupling capacitor will increases, so very small o/p voltage of the signal will be transfer from one stage to other stage. Due to large reactance, the capacitor C_E cannot parallel the emitter resistor R_E. Due to this 2 factors, the voltage gain will decreases at low frequencies.

AT HIGH FREQUENCIES:

* At high frequencies, the reactance of coupling capacitor will be very small and it behaves as a short circuit but due to junction capacitances the output voltage will decreases at high frequency. So the gain can be reduced at high frequencies.

AT MID FREQUENCIES:

* At mid frequencies, the voltage gain of the amplifier is constant. The effect of coupling capacitor to maintain the uniform voltage gain at mid frequencies.
* As frequency increases in this range, the reactance of the capacitor decreases which tends to increase the gain at the same time.
* The above two factors almost cancel each other resulting a uniform voltage gain at mid frequencies.

ADVANTAGES:

→ It provides a good frequency response
→ It is less expensive
→ It is small in size
→ Less complexity

DISADVANTAGES: It provides poor impedance matching.

APPLICATIONS: 1. Radio Receivers

2. Tape Recorders

3. TV Receivers

ANALYSIS:

* In the analysis of RC coupled amplifier, the following simplified assumptions are made.

1) h_{re} is so small that the voltage source $h_{re}V_o$ can be neglected.

2) $1/h_{oe}$ is so large that it can be considered as an open circuit.

3) The reactance C_e for any given input frequency is so small that the parallel combination of R_e and C_e can be effectively considered as a short circuit.

4) The bias resistors R_1 and R_2 are usually large as compared to h_{ie}.

⇒ With these assumptions, the simplified circuit is as follows,

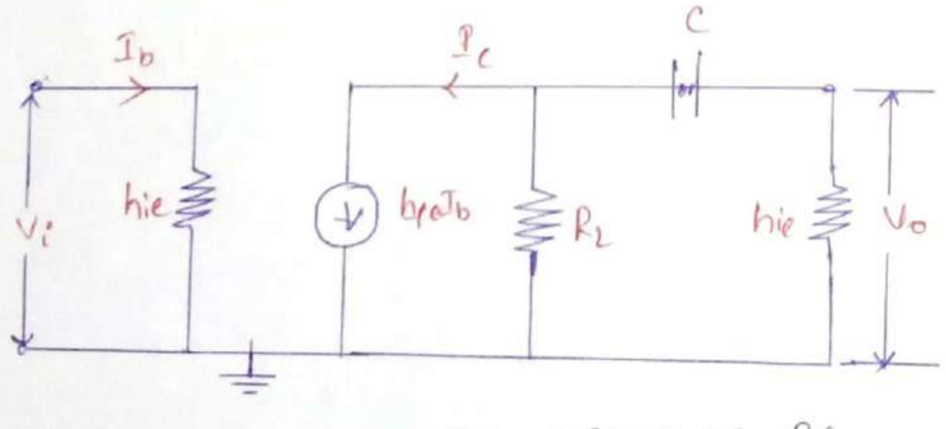


FIG: SIMPLIFIED EQUIVALENT CIRCUIT OF RC COUPLED AMPLIFIER

MIDDLE FREQUENCY RANGE:

* ~~The~~ At Mid frequencies, the impedance offered by coupling capacitor 'C' is so small, acts as an effective short circuit.

* Hence at mid frequencies, the effect of coupling capacitor 'C' can be neglected.

* The thevenin's equivalent circuit is as follows.

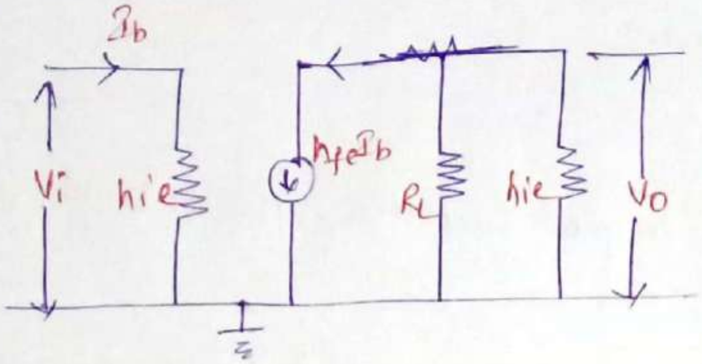


FIG: EQUIVALENT CIRCUIT AT MID FREQUENCY RANGE

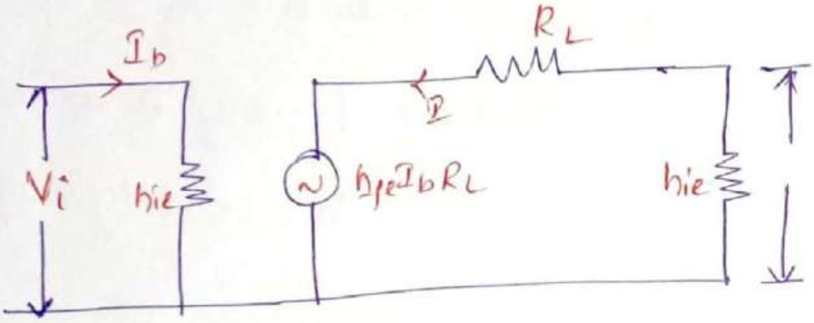


FIG: THEVENIN'S EQUIVALENT CIRCUIT

CURRENT GAIN:

By definition, Current gain $(A_I)_m = \dfrac{I}{I_b} \rightarrow (1)$

from the above figure $I = \dfrac{-h_{fe}I_bR_L}{R_L + h_{ie}} \rightarrow (2)$

Sub eq (2) in eq (1)

$$(A_I)_m = \frac{-h_{fe}\cancel{I_b}R_L}{R_L + h_{ie}} * \frac{1}{\cancel{I_b}}$$

$$A_{Im} = \frac{-h_{fe}R_L}{R_L + h_{ie}}$$

The magnitude of current gain is given by $|A_{Im}| = \dfrac{h_{fe}R_L}{R_L + h_{ie}}$

VOLTAGE GAIN:

By definition $A_{vm} = \frac{V_o}{V_i} \rightarrow (1)$

The output Voltage $V_o = h_{ie} I$

$$V_o = h_{ie} * \frac{-h_{fe} I_b R_L}{R_L + h_{ie}}$$

$$V_o = \frac{-h_{ie} h_{fe} I_b R_L}{R_L + h_{ie}} \rightarrow (2)$$

The Input Voltage $V_i = h_{ie} I_b \rightarrow (3)$

Sub eq (2) & eq (3) in eq (1)

$$A_{vm} = \frac{-\cancel{h_{ie}} h_{fe} \cancel{I_b} R_L}{R_L + h_{ie}} * \frac{1}{h_{ie} \cancel{I_b}}$$

$$A_{vm} = \frac{-h_{fe} R_L}{R_L + h_{ie}}$$

* The magnitude of voltage gain at mid-frequency is given by,

$$|A_{vm}| = \frac{h_{fe} R_L}{R_L + h_{ie}}$$

* At mid frequencies, the magnitude of current gain & voltage gain are equal i.e., $|A_{Im}| = |A_{vm}|$

Here negative sign shows the phase angle of 180°.

LOW FREQUENCY RANGE:

* In low frequency range the impedance offered by coupling capacitor is large. Hence it largely affects current amplification so it is included in the equivalent circuit

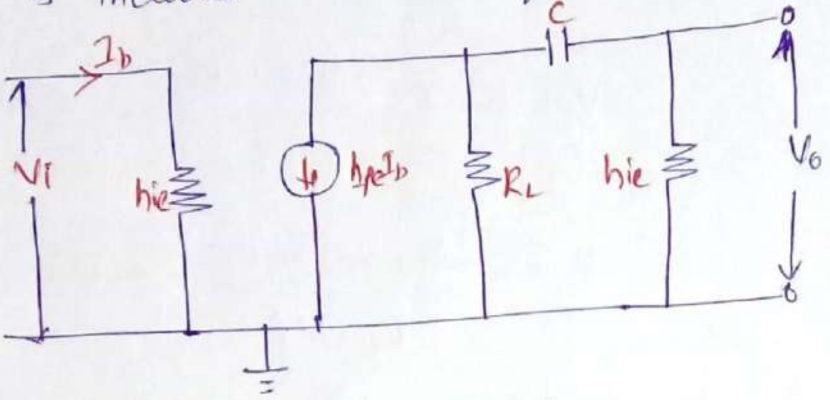


FIG: EQUIVALENT CIRCUIT

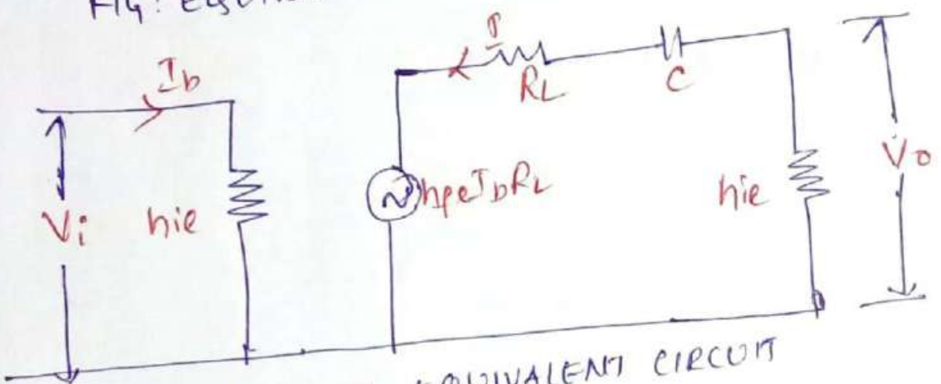


FIG: THEVENINS EQUIVALENT CIRCUIT

CURRENT GAIN:

By definition, Current gain $A_{IL} = \dfrac{I}{I_b}$

from the circuit $I = \dfrac{-h_{fe}I_bR_L}{hie+R_L+\frac{1}{j\omega C}} = \dfrac{-h_{fe}I_bR_L}{hie+R_L-j/\omega C}$

Current gain $A_{IL} = \dfrac{-h_{fe}\cancel{I_b}R_L}{hie+R_L-\frac{j}{\omega C}} * \dfrac{1}{\cancel{I_b}}$

$$A_{IL} = \frac{-h_{fe}R_L}{hie+R_L-\frac{j}{\omega C}}$$

magnitude of current gain $|A_{IL}| = \dfrac{h_{fe}R_L}{\sqrt{(R_L+hie)^2+\left(\frac{1}{2\pi fC}\right)^2}}$ $\because \omega = 2\pi f$

VOLTAGE GAIN:

By definition, $A_{VL} = \frac{V_o}{V_i}$

output Voltage $V_o = hie * I$

$$V_o = hie * \frac{-h_{fe} I_b R_L}{R_L + hie - \frac{j}{\omega C}}$$

$$V_o = \frac{-hie\, h_{fe} I_b R_L}{R_L + hie - \frac{j}{\omega C}}$$

Input Voltage $V_i = hie * I_b$

$$A_{VL} = \frac{-\cancel{hie}\, h_{fe} \cancel{I_b} R_L}{R_L + hie - \frac{j}{\omega C}} * \frac{1}{\cancel{hie * I_b}}$$

$$A_{VL} = \frac{-h_{fe} R_L}{R_L + hie - \frac{j}{\omega C}} = \frac{-h_{fe} R_L}{R_L + hie - \frac{j}{2\pi f C}} \quad (\because \omega = 2\pi f)$$

* Magnitude of Voltage gain is $|A_{VL}| = \frac{h_{fe} R_L}{\sqrt{(R_L + hie)^2 + \left(\frac{1}{2\pi f C}\right)^2}}$

* From the above expression it is obvious that as frequency increases of the i/p Voltage ~~decreases~~ increases, the magnitude of the Voltage gain decreases and vice versa.

HIGH FREQUENCY RANGE:

* In high frequency range, the reactance offered by Coupling Capacitor 'C' is very small and hence it can be considered as short circuited.
* In a bipolar transistor, there are two depletion regions across two P-N junctions. They behave like a dielectric media and hence give rise to two internal Capacitances. At high frequency range the reactance of these capacitances (C_{be} & C_{bc}) are considered.

* The C_{be} and C_{bc} may be replaced with a single capacitance C_d across the i/p resistance hie of the transistor.

* The values of shunt capacitance C_d in the input circuit of the first stage is small because it depends on the o/p impedance of the first transistor.

* But in the o/p circuit of the first stage C_d is increased by stray capacitance of the wiring. Therefore the reactance $1/\omega C_d$ will have appreciable shunting effect on R_2 & hie.

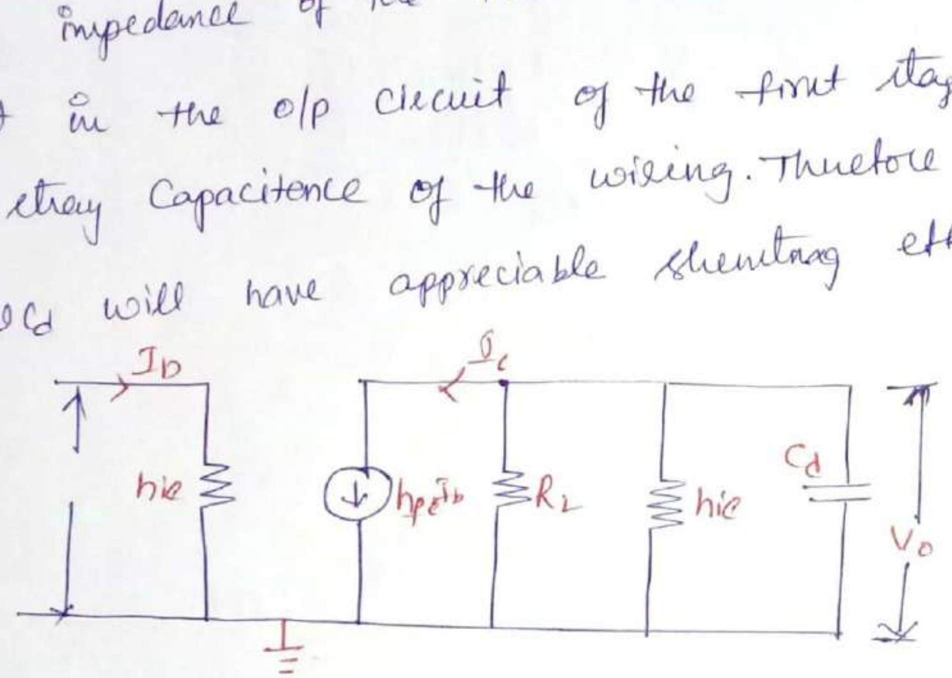


FIG: EQUIVALENT CIRCUIT AT HIGH FREQUENCIES

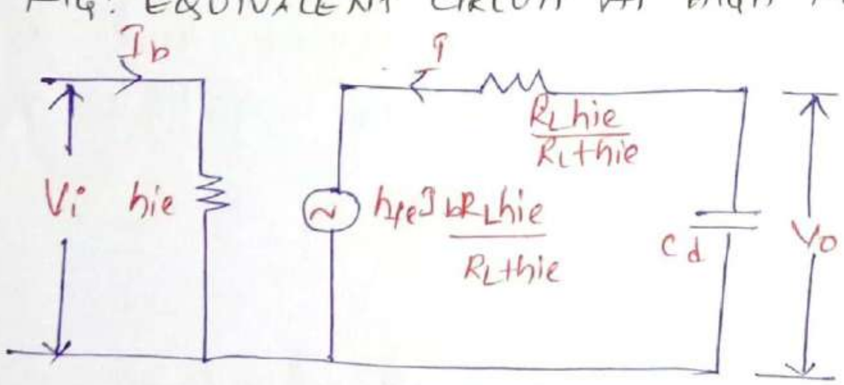


FIG: THEVENIN'S EQUIVALENT CIRCUIT

CURRENT GAIN:

* By definition $A_{IH} = \frac{I}{I_b}$

from the above figure $$I = \frac{-h_{fe} I_b \frac{R_L hie}{R_L+hie}}{\frac{R_L hie}{R_L+hie} + \frac{1}{j\omega C_d}}$$

$$I = \frac{-h_{fe} I_b R_L hie/(R_L+hie)}{(R_L+hie)\left[R_L hie + \frac{1}{j\omega C_d}(R_L+hie)\right]}$$

$$I = \frac{-h_{fe} I_b R_L hie}{R_L hie + \frac{1}{j\omega C_d}(R_L+hie)}$$

Current gain $A_{IH} = \dfrac{-h_{fe} \cancel{I_b} R_L hie}{R_L hie + \frac{1}{j\omega C_d}(R_L+hie)} * \dfrac{1}{\cancel{I_b}}$

$$A_{IH} = \frac{-h_{fe} hie R_L}{R_L hie + \frac{1}{j\omega C_d}(R_L+hie)}$$

VOLTAGE GAIN:

By definition, Voltage gain $A_{VH} = \dfrac{V_o}{V_i}$

output Voltage $V_o = \dfrac{1}{j\omega C_d} * I$

$$V_o = \frac{1}{j\omega C_d} * \frac{-h_{fe} hie R_L I_b}{R_L hie + \frac{1}{j\omega C_d}(R_L+hie)}$$

$$= \frac{1}{\cancel{j\omega C_d}} * \frac{-hie h_{fe} R_L I_b}{\frac{R_L hie j\omega C_d + (R_L+hie)}{\cancel{j\omega C_d}}}$$

$$V_o = \frac{-hie h_{fe} I_b R_L}{j\omega R_L hie C_d + (R_L+hie)}$$

Input Voltage $V_i = I_b * hie$

$$A_{VH} = \frac{-h_{fe} \cancel{hie} \cancel{I_b} R_L}{(R_L+hie) + j\omega C_d R_L hie} * \frac{1}{\cancel{I_b} * \cancel{hie}}$$

$$A_{VH} = \frac{-h_{fe} R_L}{(R_L + h_{ie}) + j\omega C_d R_L h_{ie}}$$

The magnitude of Voltage gain is given by,

$$|A_{VH}| = \frac{h_{fe} R_L}{\sqrt{(R_L + h_{ie})^2 + (\omega C_d R_L h_{ie})^2}} = \frac{h_{fe} R_L}{\sqrt{(R_L + h_{ie})^2 + (2\pi f C_d R_L h_{ie})^2}}$$

* From above expression it is obvious that as the frequency 'f' of i/p voltage increases, the magnitude of voltage gain decreases.

LOWER CUT-OFF FREQUENCY (f_1):

* The lower cut-off frequency is defined as the frequency at which the magnitude of the voltage gain in the lower frequency range falls to $1/\sqrt{2}$ or 0.707 of the magnitude of the gain in mid frequency range. Thus,

$$|A_{VL}| = \frac{|A_{Vm}|}{\sqrt{2}} \quad \text{or} \quad \frac{|A_{VL}|}{|A_{Vm}|} = \frac{1}{\sqrt{2}} \rightarrow (1)$$

In R-C Coupled amplifier

$$|A_{VL}| = \frac{h_{fe} R_L}{\sqrt{(h_{ie} + R_L)^2 + \left(\frac{1}{2\pi f C}\right)^2}} = \frac{h_{fe} R_L}{\sqrt{(h_{ie} + R_L)^2 \left[1 + \left[\frac{1}{2\pi f C (h_{ie} + R_L)}\right]^2\right]}}$$

$$= \frac{h_{fe} R_L}{(h_{ie} + R_L)\sqrt{1 + \left[\frac{1}{2\pi f C (h_{ie} + R_L)}\right]^2}}$$

$$= \frac{h_{fe} R_L}{h_{ie} + R_L} \times \frac{1}{\sqrt{1 + \left[\frac{1}{2\pi f C (h_{ie} + R_L)}\right]^2}} \quad \left(\because |A_{Vm}| = \frac{h_{fe} R_L}{h_{ie} + R_L}\right)$$

$$= |A_{Vm}| \times \frac{1}{\sqrt{1 + \left[\frac{1}{2\pi f C (h_{ie} + R_L)}\right]^2}}$$

At f_1 $\quad \frac{(A_{VL})}{(A_{Vm})} = \frac{1}{\sqrt{1+\left[\frac{1}{2\pi f C(h_{ie}+R_L)}\right]^2}} \rightarrow (2)$

Equating (1) & (2) and If f_1 be the lower cut off frequency

$$\frac{1}{\sqrt{2}} = \frac{1}{\sqrt{1+\left[\frac{1}{2\pi f_1 C(h_{ie}+R_L)}\right]^2}}$$

$$\sqrt{2} = \sqrt{1+\left[\frac{1}{2\pi f_1 C(h_{ie}+R_L)}\right]^2}$$

$$2 = 1+\left[\frac{1}{2\pi f_1 C(h_{ie}+R_L)}\right]^2$$

$$\left[\frac{1}{2\pi f_1 C(h_{ie}+R_L)}\right]^2 = 1$$

$$\frac{1}{2\pi f_1 C(h_{ie}+R_L)} = 1$$

$$2\pi f_1 C(h_{ie}+R_L) = 1$$

$$f_1 = \frac{1}{2\pi C(h_{ie}+R_L)} \rightarrow (3)$$

Sub eq (3) in eq (2), we get

$$\boxed{|A_{VL}| = \frac{|A_{Vm}|}{\sqrt{1+\left(\frac{f_1}{f}\right)^2}}}$$

UPPER CUT-OFF FREQUENCY (f_2):

* The upper cut-off frequency is defined as the frequency at which the magnitude of the voltage gain in the high frequency range falls to $1/\sqrt{2}$ & 0.707 of magnitude of the gain in the mid frequency range. Thus at $f = f_2$

$$|A_{VH}| = \frac{|A_{Vm}|}{\sqrt{2}} \Rightarrow \frac{|A_{VH}|}{|A_{Vm}|} = \frac{1}{\sqrt{2}} \rightarrow (1)$$

Here $$|A_{VH}| = \frac{h_{fe} R_L}{\sqrt{(h_{ie}+R_L)^2 + (2\pi f C_d R_L h_{ie})^2}}$$

$$= \frac{h_{fe} R_L}{\sqrt{(h_{ie}+R_L)^2\left[1+\left[\frac{1}{2\pi f C_d R_L h_{ie}(h_{ie}+R_L)}\right]^2\right]}}$$

$$= \frac{h_{fe} R_L}{(h_{ie}+R_L)\sqrt{1+\left[\frac{1}{2\pi f C_d (h_{ie}+R_L) R_L h_{ie}}\right]^2}}$$

$$= \frac{|A_{Vm}|}{\sqrt{1+\left[\frac{1}{2\pi f C_d R_L h_{ie}(h_{ie}+R_L)}\right]^2}} \quad \left(\because A_{Vm} = \frac{h_{fe} R_L}{h_{ie}+R_L}\right)$$

$$\frac{|A_{VH}|}{|A_{Vm}|} = \frac{1}{\sqrt{1+\left[\frac{1}{2\pi f C_d R_L h_{ie}(h_{ie}+R_L)}\right]^2}} \rightarrow (2)$$

Equating (1) & (2) and f_2 is the upper cut-off frequency

$$\frac{1}{\sqrt{2}} = \frac{1}{\sqrt{1+\left[\frac{1}{2\pi f_2 C_d R_L h_{ie}(R_L+h_{ie})}\right]^2}}$$

$$\sqrt{2} = \sqrt{1+\left[\frac{1}{2\pi f_2 C_d R_L h_{ie}(h_{ie}+R_L)}\right]^2}$$

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$$2 = 1+\left[\frac{1}{2\pi f_2 C_d R_L h_{ie}(h_{ie}+R_L)}\right]^2$$

$$\left[\frac{1}{2\pi f_2 C_d R_L h_{ie}(h_{ie}+R_L)}\right]^2 = 1$$

$$2\pi f_2 C_d R_L h_{ie}(h_{ie}+R_L) = 1$$

$$f_2 = \frac{1}{2\pi C_d R_L h_{ie}(h_{ie}+R_L)} \rightarrow (3)$$

Sub eq (3) in eq (2)

$$\boxed{|A_{VH}| = \frac{A_{Vm}}{\sqrt{1+\left(\frac{f}{f_2}\right)^2}}}$$

$$\theta_h = \tan^{-1}\left(\frac{f}{f_2}\right)$$

DARLINGTON PAIR AMPLIFIER:

* A very popular connection of two bipolar junction transistors for operation as one "superbeta" transistor is the Darlington Connection.
* The Composite transistor acts as a single unit with a Current gain that is the product of the Current gains of the individual transistors.

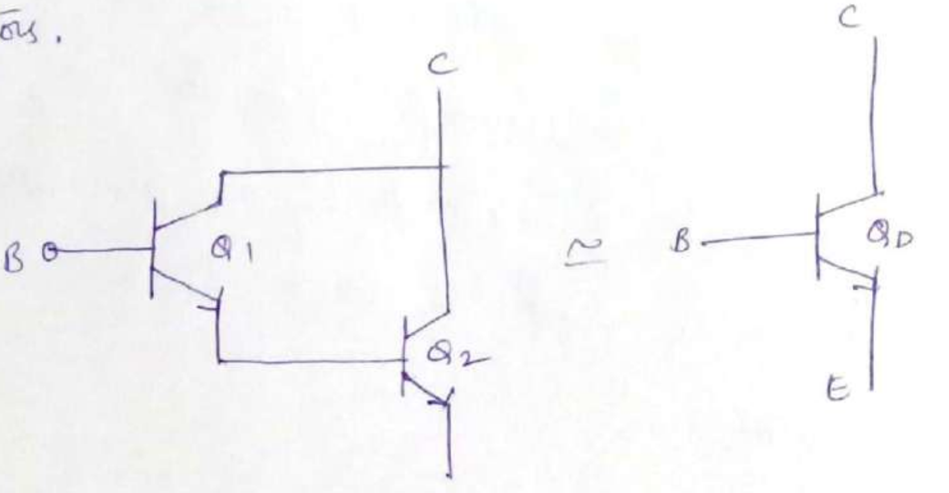


FIG: DARLINGTON TRANSISTOR CONNECTION AND SINGLE DARLINGTON TRANSISTOR

* If the two transistors are matched such that $\beta_1 = \beta_2 = \beta$, the Darlington Connection provides a current gain of

$$\beta_D = \beta^2$$

* When two transistors having high current gain are connected as a Darlington pair, the overall gain of the pair becomes very high.

DC ANALYSIS:

* A Darlington transistor offers very high current gain, β_D is used.

The base current is given by, $I_B = \dfrac{V_{CC} - V_{BE}}{R_B + \beta_D R_E}$

The Emitter current is given by, $I_E = (\beta_D + 1) I_B$

$$= \beta_D I_B$$

The dc voltages are given by, $V_E = I_E R_E$

$$V_B = V_E + V_{BE}$$

AC ANALYSIS:

* The AC input signal is applied to the base of the Darlington transistor through capacitor C_1, with the ac output V_o obtained from the emitter through capacitor C_2.
* The Darlington transistor is replaced by an ac equivalent circuit comprised of an input resistance r_i and an output current source $\beta_D I_b$.

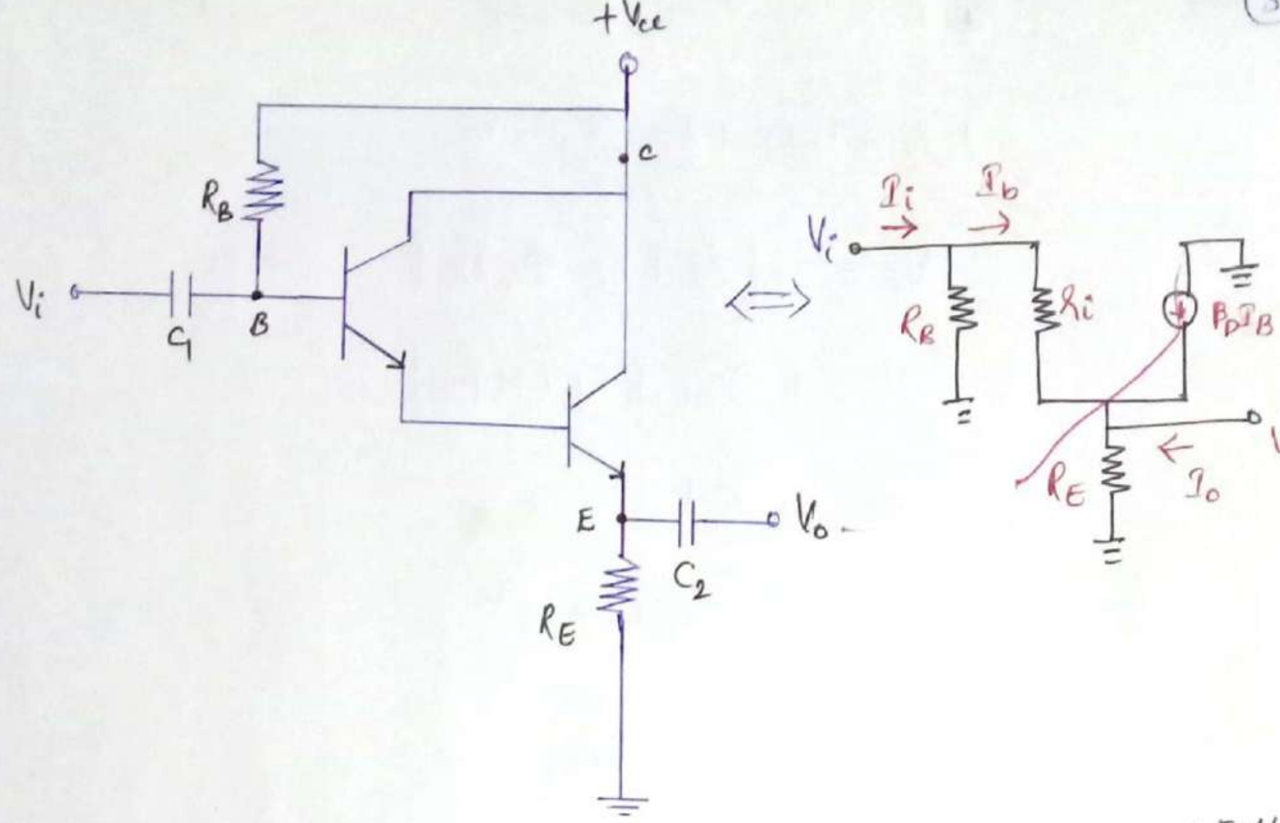
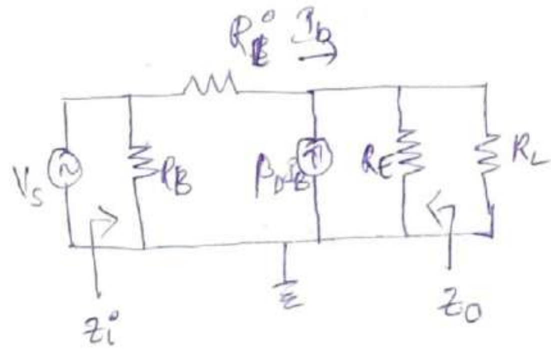


FIG: AC EQUIVALENT CIRCUIT OF DARLINGTON EMITTER FOLLOWER

AC INPUT IMPEDANCE:



* The ac base current through r_i is,

$$I_b = \frac{V_i - V_o}{R_i} \rightarrow ①$$

But output voltage is given by,

$$V_o = (I_b + \beta_D I_b) R_E \rightarrow ②$$

$\Rightarrow$ Ac Input Impedance $Z_i = \dfrac{V_i}{I_b}$

Substituting eq ② in eq ①

$$I_b = \frac{V_i - (I_b R_E + \beta_D I_b R_E)}{R_i}$$

$$I_b = \frac{V_i - I_b R_E - \beta_D I_b R_E}{R_i}$$

$$I_b r_i = V_i - I_b R_E - \beta_D I_b R_E$$

$$I_b r_i + I_b R_E + \beta_D I_b R_E = V_i$$

$$V_i = I_b [r_i + r_E + \beta_D R_E]$$

$$V_i = I_b [r_i + r_E (1+\beta_D)]$$

Since $\beta_D >> 1$

$$V_i = I_b (r_i + \beta_D R_E)$$

$$\frac{V_i}{I_b} = r_i + \beta_D R_E$$

The ac impedance, looking into the circuit is

$$\boxed{Z_i = R_B \| (r_i + \beta_D R_E)}$$

AC CURRENT GAIN:

* The ac output current through R_E is given as,

$$I_o = I_b + \beta_D I_b = (1+\beta_D) I_b$$

$$\simeq \beta_D I_b \qquad (\because \beta_D >> 1)$$

* The transistor current is then

$$\frac{I_o}{I_b} = \beta_D$$

* The ac current gain of the circuit is,

$$A_i = \frac{I_o}{I_i} \rightarrow (1)$$

Multiply numerator and denominator of eq(1) with I_b.

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$$A_i = \frac{I_o}{I_i} \times \frac{I_b}{I_b} = \frac{I_o}{I_b} * \frac{I_b}{I_i} \rightarrow (2)$$

using the current divider rule in the ac equivalent circuit

$$I_b = I_i * \frac{R_B}{(r_i + \beta_D R_E) + R_B}$$

$$\frac{I_b}{I_i} = \frac{R_B}{(r_i + \beta_D R_E) + R_B} \rightarrow (3)$$

Substituting eq (3) in eq (2)

$$A_i = \frac{I_o}{I_b} * \frac{R_B}{(r_i + \beta_D R_E) + R_B}$$

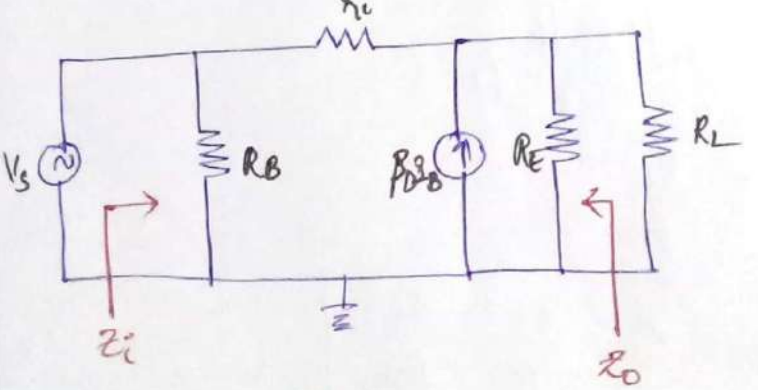
$$= \beta_D * \frac{R_B}{(r_i + \beta_D R_E) + R_B} \qquad (\because \beta_D = \frac{I_o}{I_b})$$

$$= \beta_D * \frac{R_B}{\beta_D R_E + R_B}$$

$$\boxed{\therefore \text{The current gain is given by } A_I = \beta_D \frac{R_B}{\beta_D R_E + R_B}}$$

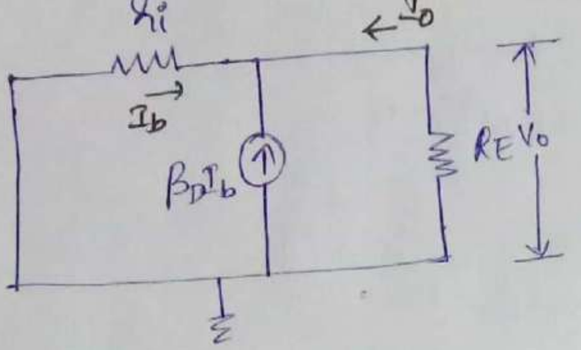
AC OUTPUT IMPEDANCE :

* The circuit for output impedance is as shown below



(a)

* The output impedance can be calculated by making $V_s = 0$ then the equivalent circuit will be



From the above circuit, $I_0 = \frac{V_0}{R_E}$ Apply KCL then,

$$I_0 = \frac{V_0}{R_E} + \frac{V_0}{r_i} - \beta_D I_b \quad \text{where } I_b = -\frac{V_0}{r_i}$$

$$= \frac{V_0}{R_E} + \frac{V_0}{r_i} - \beta_D\left(-\frac{V_0}{r_i}\right)$$

$$= \frac{V_0}{R_E} + \frac{V_0}{r_i} + \frac{V_0 \beta_D}{r_i}$$

$$= V_0\left[\frac{1}{R_E} + \frac{1}{r_i} + \frac{\beta_D}{r_i}\right]$$

$$\frac{V_0}{I_0} = \frac{1}{\frac{1}{R_E} + \frac{1}{r_i} + \frac{\beta_D}{r_i}}$$

$$= R_E \,||\, r_i \,||\, \frac{r_i}{\beta_D}$$

$$\approx \frac{r_i}{\beta_D}$$

$$\boxed{\therefore \text{Output impedance } Z_0 = \frac{V_0}{I_0} = \frac{r_i}{\beta_D}}$$

AC VOLTAGE GAIN:

* The equivalent circuit to determine Voltage gain

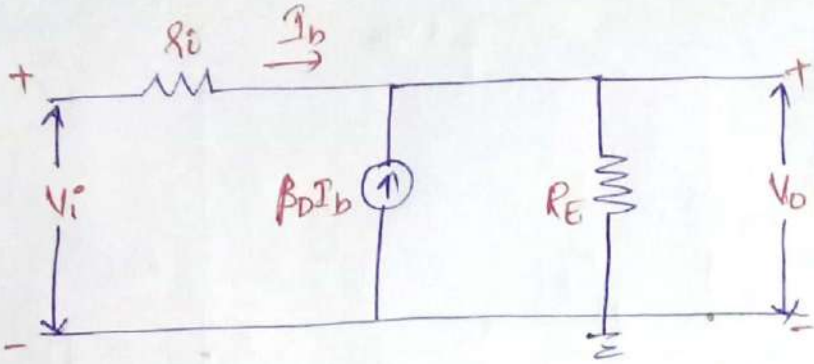


FIG: AC equivalent circuit to determine A_V

From the circuit, $V_o = (I_b + \beta_D I_b) R_E$

$$= I_b R_E + \beta_D I_b R_E$$

$$V_o = I_b (R_E + \beta_D R_E)$$

$$V_i = I_b r_i + (I_b + \beta_D I_b) R_E$$

$$V_i = I_b [r_i + R_E + \beta_D R_E]$$

$\therefore$ Voltage gain $A_V = \frac{V_o}{V_i}$

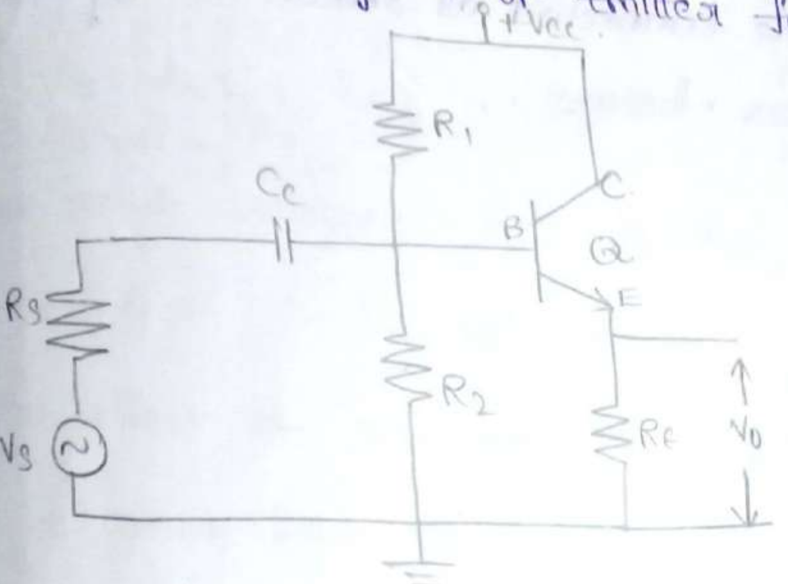
$$= \frac{\cancel{I_b}(R_E + \beta_D R_E)}{\cancel{I_b}(r_i + R_E + \beta_D R_E)}$$

$$= \frac{R_E + \beta_D R_E}{r_i + R_E + \beta_D R_E} \quad (\because r_i << R_E + \beta_D R_E)$$

$$= \frac{\cancel{R_E + \beta_D R_E}}{\cancel{R_E + \beta_D R_E}} = 1$$

$$\boxed{A_V = 1}$$

The circuit diagram of emitter follower is shown in fig(a).



fig(a): emitter follower (common collector amplifier).

Circuit description:

Vcc → Biasing voltage.

R1 & R2 → Biasing resistor.

Re → emitter resistor.

$R_S \rightarrow$ source resistor

$V_S \rightarrow$ Source voltage

$Q \rightarrow$ npn transistor

$B, E, C \rightarrow$ Base, Emitter & collector

V_i & $V_0 \rightarrow$ input & output voltage

(3) The input resistance of common collector amplifier (or) emitter follower is high generally.

i.e. $R_i = h_{ie} + h_{re} A_I R_L \ \Omega$

$\simeq h_{ie} + 1 \times A_I R_C \qquad \because h_{ie} = h_{ie}$

$R_i = h_{ie} + A_I R_C . \qquad$ & $h_{re} \simeq 1$

But current gain $(A_I) \simeq h_{fe} = 1 + h_{fe}$.

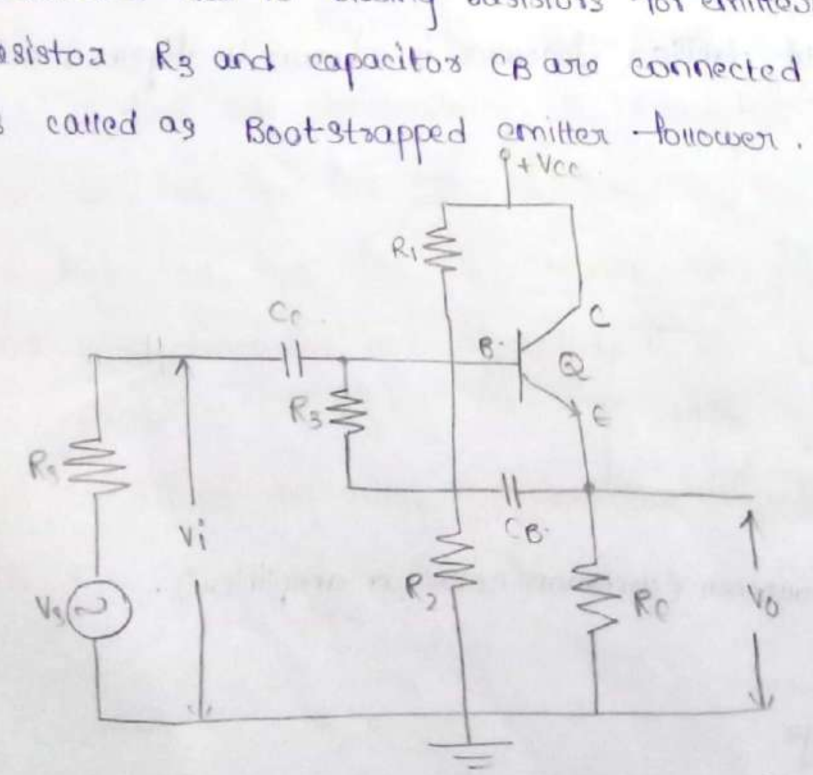
$R_i = h_{ie} + (1 + h_{fe}) R_e$

But practically Biasing resistors R_1 & R_2 reduces the input resistance as $R_i' = R_i \| R_b \ \Omega$ where.

$R_b = R_1 \| R_2 \ \Omega$.

$\simeq R_b \ \Omega$

To overcome the above disadvantage i.e. decrease of input resistance due to biasing resistors for emitter follower and additional resistor R_3 and capacitor C_B are connected as shown in fig(b) which is called as Bootstrapped emitter follower.



fig(b): Boot- ~~stapped~~ emitter follower.
Strapped.

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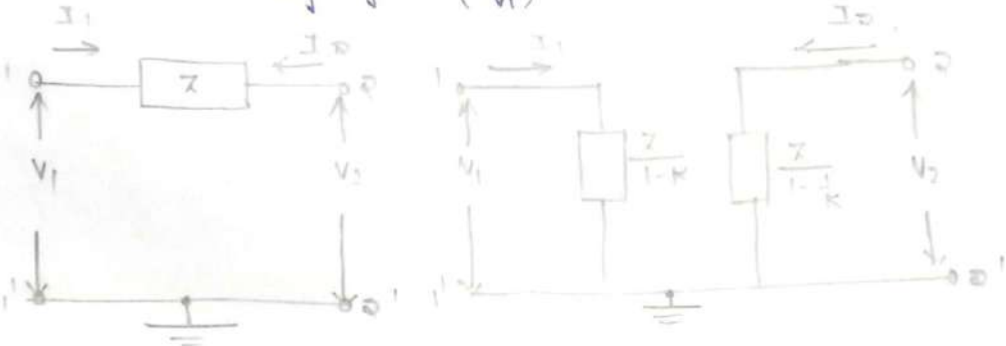
The top of R_3 is connected to Base (B) (input) and the bottom of R_3 is connected to emitter (E) (output) via capacitor C_B.

The value of capacitor C_B is selected such that it acts as short circuit for low frequency of operation.

Millers theorem states that if there is an impedence between two nodes can be replaced by two impedence $\frac{Z}{1-K}$ & $\frac{Z}{1-\frac{1}{K}}$ between input node and ground & output node and ground, where

$K =$ voltage gain $\left(\frac{V_2}{V_1}\right)$.



Here voltage gain of Cc amplifier is Av which is approximately unity i.e. $A_V \simeq 1$.

R_3 is connected b/w input & output terminals.

Now, by millers theorem,

effective input resistance $R_{eff} = \frac{R_3}{1-A_V}$

As $A_V \simeq 1$ Reff is very high $R_{effective} = R_{eff}$

for ex: $\frac{R_3}{1-0.99}$

Now input resistance $R_i' = R_i \| R_{eff} \simeq R_i$

i.e. $R_i' \simeq R_i$

≫ The effect of increasing the i/p resistance, when Av approaches unity is called Boot strapping

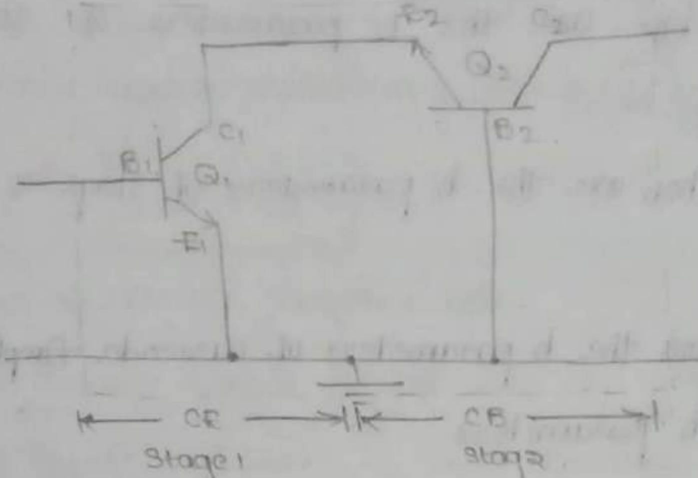
≫ The above term arises from the fact that $A_V = \frac{V_0}{V_i} \simeq 1$

$V_0 \simeq V_i$

≫ It means if one end of the resistor R_3 changes in voltage V_i then another end of R_3 moves through the same change in Voltage (V_0).

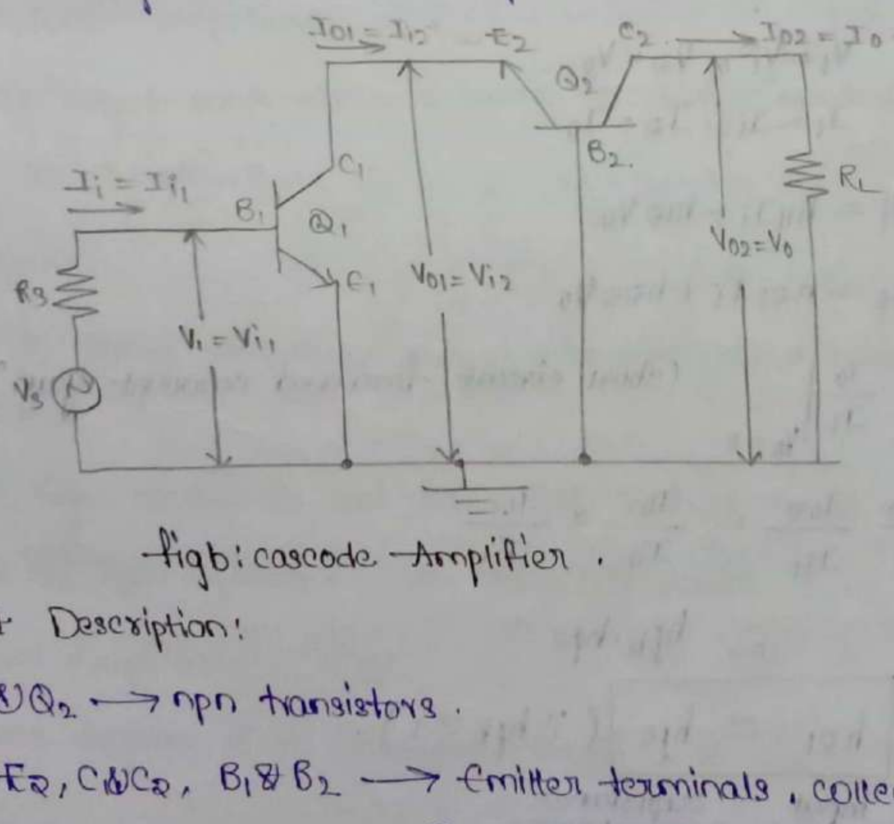
Case Code Amplifier:

The circuit diagram of CE – CB configuration (casecode config) is shown in fig(a). In which it is observed that ~~those~~ output of stage 1 (C_1) given to stage 2 ~~output~~ input (E_2)



fig(a): Casecode configuration (CE followed by CB).

Circuit diagram of cascode amplifier is shown in fig(b).



figb: cascode Amplifier.

Circuit Description:

Q_1 & Q_2 → npn transistors.

E_1 & E_2, C_1 & C_2, B_1 & B_2 → Emitter terminals, collector terminals, Base terminals of Q_1 & Q_2 respectively.

R_L, R_S, V_S → load resistance, source resistance, source voltage.

V_{01}, V_{02}, V_0 → output voltages of CE, CB & casecode.

V_{i1}, V_{i2}, V_i → input voltages of CE, CB & casecode.

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from fig $V_i = V_{i_1}$, & $I_i = I_{i_1}$.

$V_{01} = V_{i2}$, $I_{01} = I_{i2}$.

$V_{02} = V_0$, $I_{02} = I_0$

Analysis:

⇒ Let $h_{fe}, h_{oe}, h_{ie}, h_{re}$ are the h parameters of stage 1 CE Amplifier

⇒ Let $h_{fb}, h_{ob}, h_{ib}, h_{rb}$ are the h parameters of stage 2 CB Amplifier.

⇒ $h_{21}, h_{22}, h_{11}, h_{12}$ are the h parameters of casecode Amplifier.

By defination of h parameters

$$V_1 = h_{11} I_1 + h_{12} V_2.$$

$$I_2 = h_{21} I_1 + h_{22} V_2.$$

from fig: $V_1 = V_i$, $V_2 = V_0$

$I_1 = I_i$ $I_2 = I_0$

then $V_i = h_{11} I_i + h_{12} V_0$

$$I_0 = h_{21} I_i + h_{22} V_0.$$

$$h_{21} = \left.\frac{I_0}{I_i}\right|_{V_0 = 0} \quad \text{(short circuit forward current gain)}.$$

$$\simeq \frac{I_{02}}{I_{i1}} = \frac{I_{02}}{I_{i1}} \times \frac{I_{01}}{I_{i1}}$$

$$= h_{fb} \cdot h_{fe}.$$

$$\boxed{h_{21} \simeq h_{fe}} \quad (\because h_{fb} \simeq 1).$$

Short circuit ~~output~~ input ~~conductance~~ resistance.

$$\cancel{h_{22}} = h_{11} = \left.\frac{V_i}{I_i}\right|_{V_0 = 0} (\Omega). \simeq h_{ie} \text{ (stage is CE)}.$$

$$\boxed{h_{11} = h_{ie}.}$$

Open circuit reverse voltage gain $h_{12} = \left.\frac{V_i}{V_0}\right|_{I_i = 0}$

$$= \frac{V_{i1}}{V_{02}} = \frac{V_{i1}}{V_{01}} \times \frac{V_{i2}}{V_{02}}.$$

$$\boxed{h_{12} = h_{re} \cdot h_{rb}}$$

Open circuit output conductance $h_{22} = \left.\frac{I_0}{V_0}\right|_{I_i = 0}$ (v) $= h_{ob}$ (since stage 2 is CB).

$$\boxed{h_{22} = h_{ob}}$$

Advantages of cascode Amplifier are.

(1) Overall current gain is current gain of single CE Amplifier

(2) Overall input resistance is input resistance of single CE Amplifier

(3) Overall reverse voltage gain is product of reverse voltage gain of both CB & CE Amplifiers.

(4) Overall output conductance is same as output conductance of single CB Amplifier.

Applications:

(1) Used in tuned Amplifier designed because of reduction in internal feedback

(2) Has high bandwidth and less noise used as small signal amplifier

(3) Used in Rf Applications as video Amplifier.

Differential Amplifier:

Block diagram of a differential Amplifier is shown in fig (a).

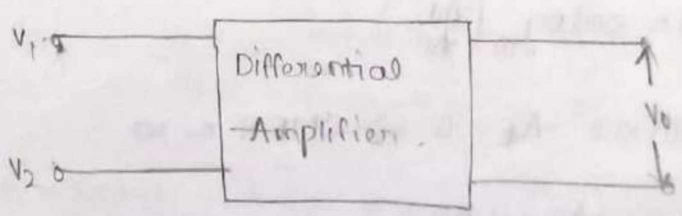


fig (a): Block diagram.

An Amplifier which Amplifies the difference between two input voltages is called as differential Amplifier.

Voltage gain of differential amplifier is

$$A_d = \frac{V_0}{V_1 - V_2} = \frac{V_0}{V_d}.$$

Where $V_1, V_2 \rightarrow$ two input voltages

$V_0 \rightarrow$ output voltage.

$V_1 - V_2 \rightarrow$ difference voltage.

$$V_0 = A_d (V_1 - V_2).$$

For an ideal differential amplifier $V_1 = V_2$.

$$V_0 = 0V.$$

But in a practical amplifier the output voltage also depends upon average signal (or) common mode signal.

i.e. $V_c = \frac{V_1 + V_2}{2}.$

Common mode voltage gain $A_c = \frac{V_0}{V_c} = \frac{V_0}{\left(\frac{V_1+V_2}{2}\right)}$

$$V_0 = A_c V_c = A_c \left(\frac{V_1 + V_2}{2}\right)$$

Total output voltage $V_0 = A_d V_d + A_c V_c$.

$$= A_d (V_1 - V_2) + A_c \left(\frac{V_1 + V_2}{2}\right)$$

CMRR (common mode rejection ratio) is defined as ratio of differential voltage gain (A_d) to the common mode voltage gain (A_c)

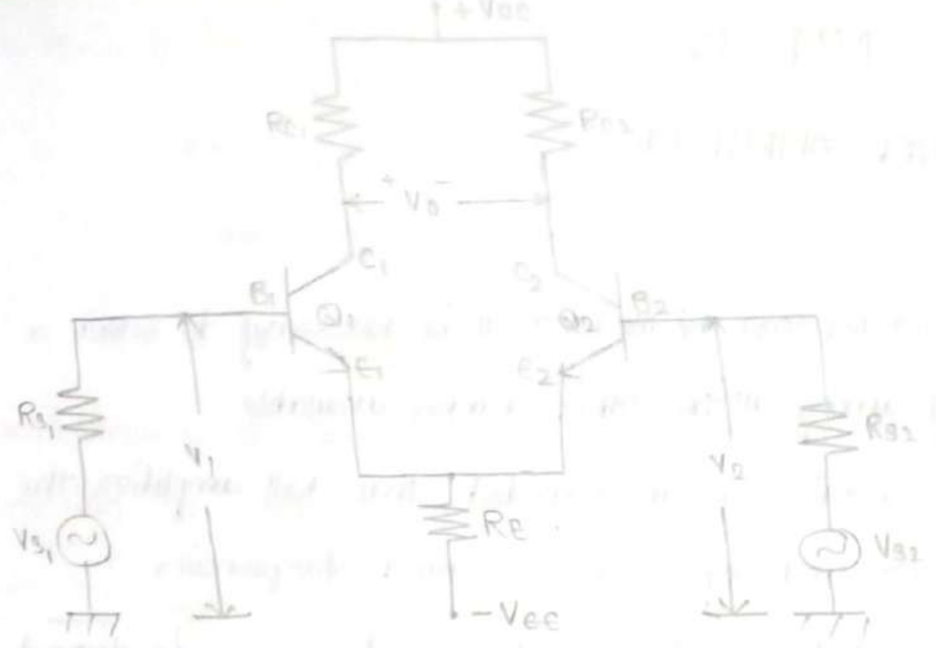
$$CMRR = \frac{A_d}{A_c}$$

$$CMRR(dB) = 20\log_{10}\left(\frac{A_d}{A_c}\right)$$

For an ideal amplifier $A_c = 0 \Rightarrow CMRR = \infty$

But in practical cases $A_c << A_d$.

The circuit diagram of differential amplifier by using BJT is shown in fig(b).



fig(b): Emitter coupled differential Amplifier.

Circuit description:

$-V_{ee}$, $+V_{cc}$ ⟶ positive biasing voltage

Q_1 & Q_2 ⟶ npn transistors.

V_1 & V_2 ⟶ input voltages.

R_E ⟶ Emitter resistance

B_1 & B_2, C_1 & C_2, E_1 & E_2 ⟶ Base terminals, collector terminals, Emitter terminals.

R_{C1} & R_{C2} ⟶ collector resistors.

R_{S1} & R_{S2} ⟶ source resistors.

V_{S1} & V_{S2} ⟶ voltage sources.

Applications:

Used as basic input stage for opamp (operation amplifier).

to provide balanced output.

$$A_d = \frac{V_0}{V_d} = \frac{-h_{fe} R_L}{R_s + h_{ie}}.$$

$$A_c = \frac{V_0}{V_c} = \frac{-h_{fe} R_L}{R_s + h_{ie} + (1+h_{fe}) 2R_e}.$$

$$CMRR = 20\log_{10}\left[\frac{A_d}{A_c}\right] = 20\log_{10}\left[\frac{R_s + h_{ie} + (1+h_{fe}) 2R_e}{R_s + h_{ie}}\right]$$

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ANALYSIS OF MULTISTAGE AMPLIFIER USING FET:

* The circuit diagram for multistage amplifier using FET.

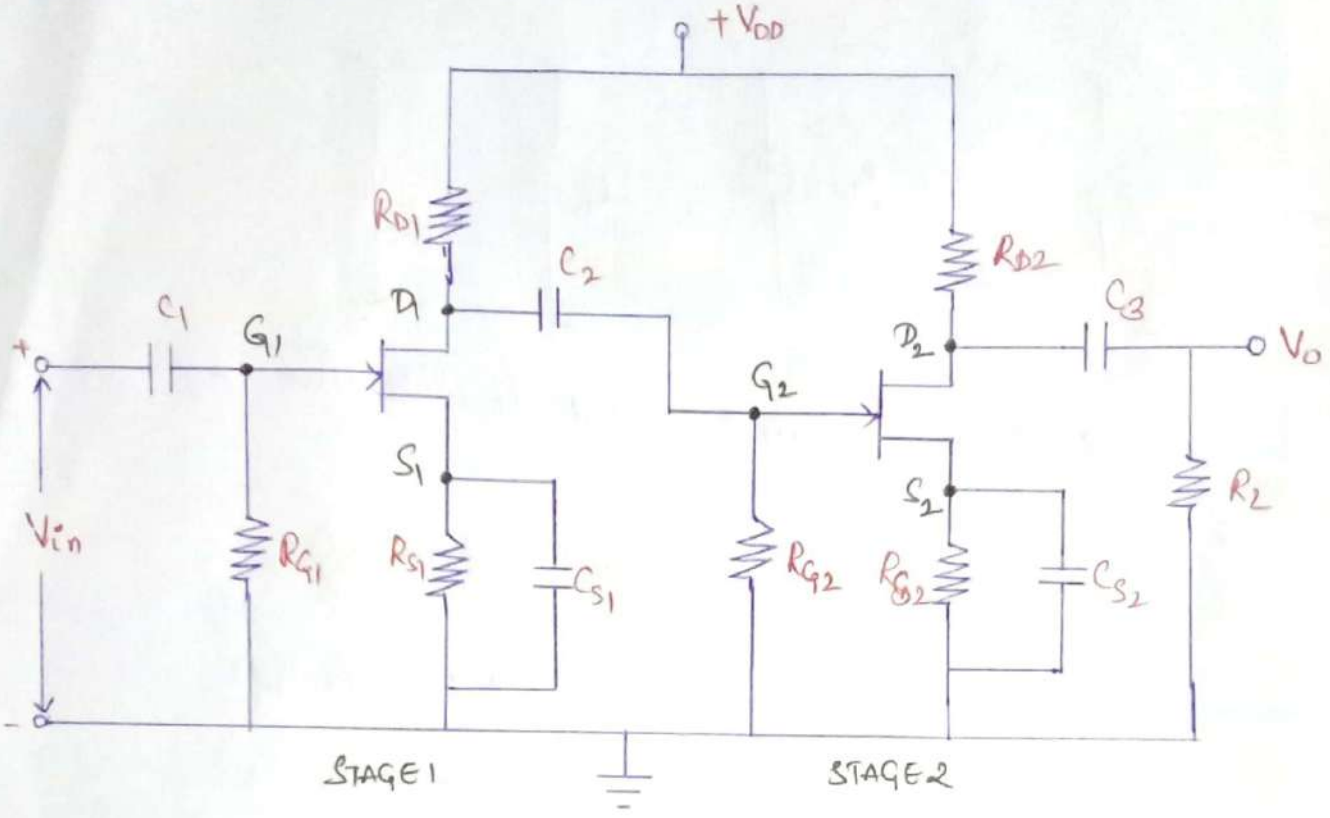


FIG: MULTISTAGE FET AMPLIFIER

* The overall gain of the multistage FET amplifier is given by the product of individual gains

i.e., $A_V = A_{V1} * A_{V2}$

$= (-g_{m1} R_{D1}) * (-g_{m2} R_{D2})$

$= g_{m1} R_{D1} g_{m2} R_{D2}$

* The Input impedance $Z_{in} = R_{G1}$

* The output impedance $Z_{out} = R_{D2}$

* The main function of cascading stages is to achieve the large overall voltage gain.

Unit - II

SMALL SIGNAL HIGH FREQUENCY TRANSISTOR AMPLIFIER MODELS

Hybrid - π model of a transistor in CE configuration:

Introduction:

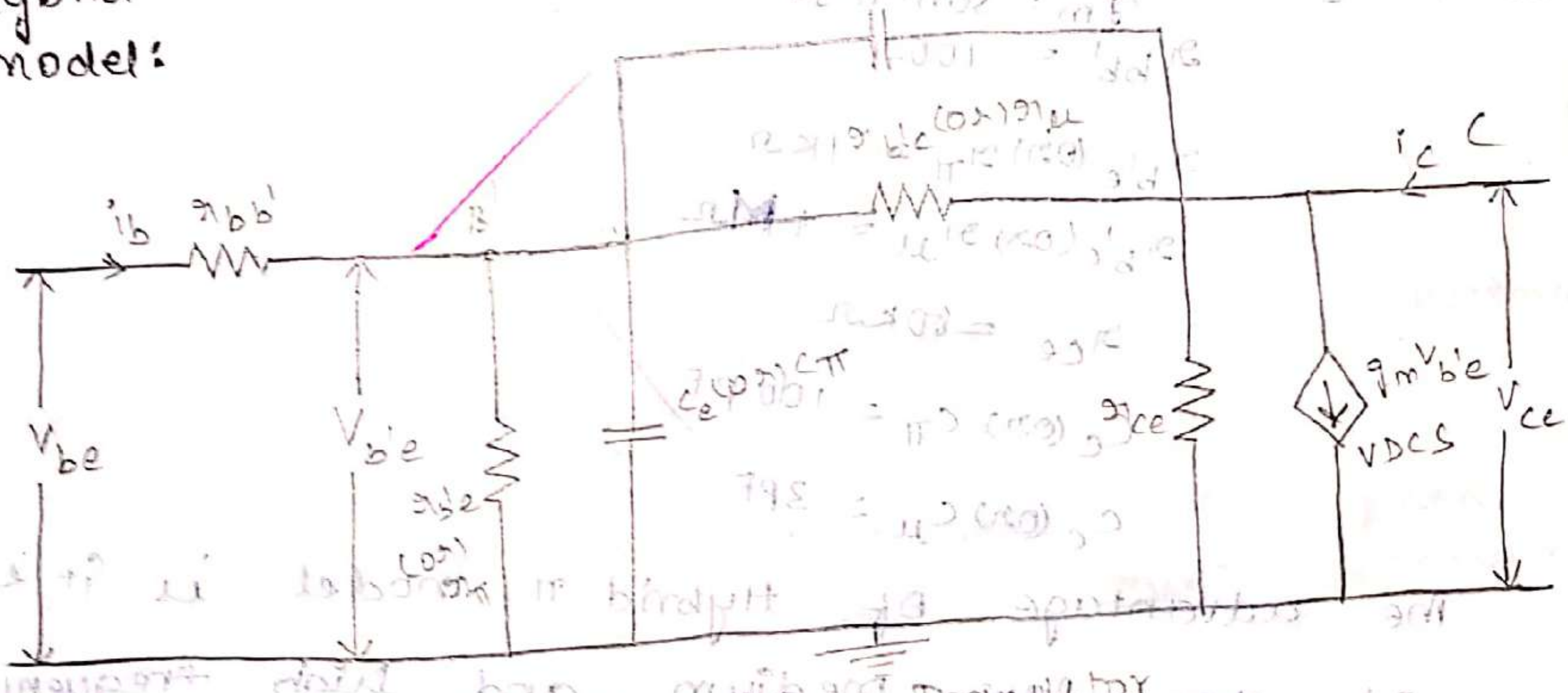
→ The low frequency small signal model (Hybrid model) of BJT works for frequencies below 1MHz

→ for frequencies greater than 1MHz, response of the transistor will be limited by internal parasitic capacitances of BJT. i.e., * the gain decreases at high frequencies

* ON and OFF switching times of BJT will be high.

→ Hence a model was introduced in 1969 by L.J. Gia colletto called as hybrid - π model (because of its shape) (or) Gia colletto model (because of the inventor)

Hybrid - π model (or) Hybrid - Pi model (or) Gia colletto model:



Hybrid π model of CE Transistor

where

B, E, C ⟶ Base, Emitter, Collector

B' ⟶ Physically inaccessable internal node i.e., virtual base

$r_{bb'}$ ⟶ Resistance b/n actual base and virtual base (or) i.e., bulk resistance of base (or) base spreading resistance

$r_{b'e}$ (or) r_π ⟶ Resistance b/n virtual base and emitter

$r_{b'c}$ (or) r_{μ} → Resistance b/n virtual base and collector

r_{ce} → Resistance b/n collector and emitter

C_e (or) C_{π} → Diffusion capacitance b/n forward biased emitter base junction

C_c (or) C_{μ} → Transition capacitance b/n reverse biased collector base junction

$g_m v_{b'e}$ → Voltage dependent current source

g_m → transconductance (or) mutual conductance

$v_{b'e}$ → Voltage b/n virtual base and emitter

v_{ce} → voltage b/n collector and emitter

i_b → base current

i_c → collector current

Assume that all parameters in this model are independent of frequency and are constant under given biased condition. Typical values of seven hybrid parameters at room temperature and $I_c = 1.3mA$ are as follows

$g_m = 50m\mho$

$r_{bb'} = 100\Omega$

$r_{b'e}$ (or) $r_{\pi} = 1K\Omega$

$r_{b'c}$ (or) $r_{\mu} = 4M\Omega$

$r_{ce} = 80K\Omega$

C_e (or) $C_{\pi} = 100\,PF$

C_c (or) $C_{\mu} = 3PF$

The advantage of Hybrid π model is it is valid for low, medium and high frequencies but it is used only for high frequencies. Because of complexity in analysis it can't be used for low and medium frequencies.

H-parameter model (Hybrid model) is used only for low and medium frequencies as it fails at high frequencies.

Derivation of Hybrid - π parameter in terms of h-parameter:

consider H-parameter model of CE transistor as shown in figure (a).

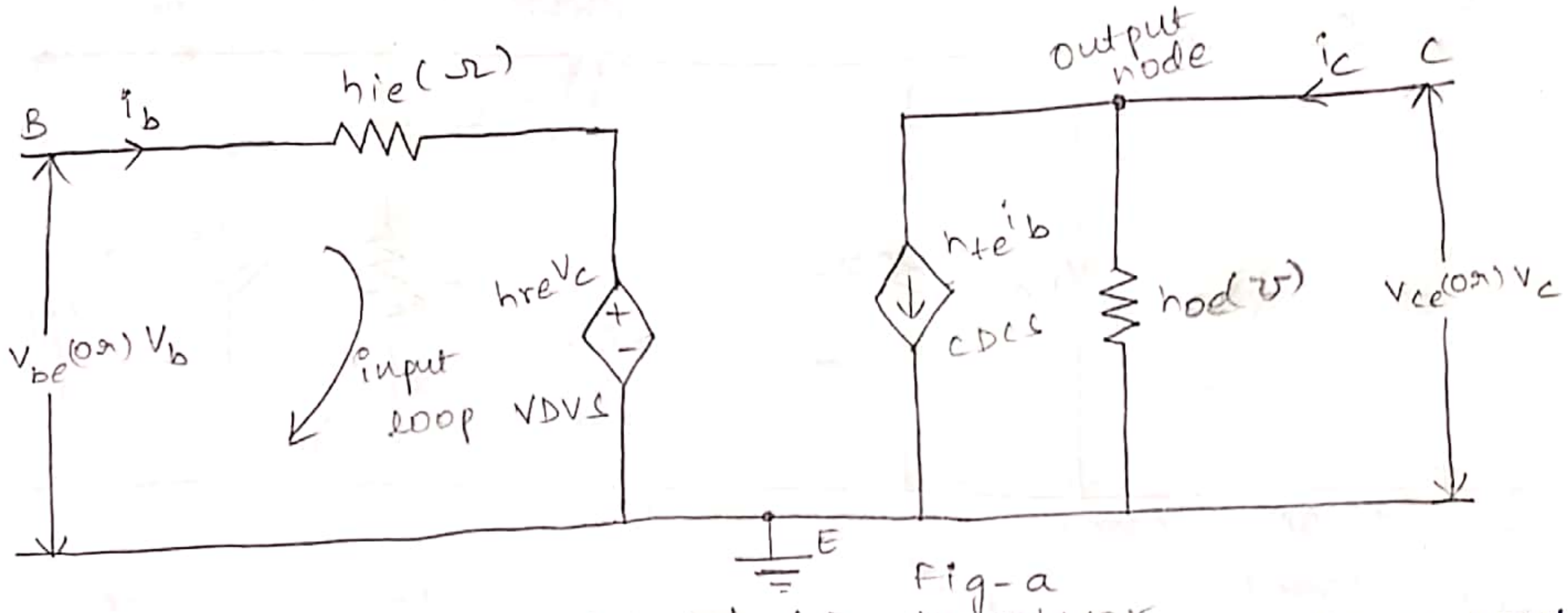


Fig-a

h-parameter model of CE transistor

By applying KVL to input loop & KCL to output node

$$V_b = h_{ie} i_b + h_{re} V_c$$

$$i_c = h_{fe} i_b + h_{oe} V_c$$

$h_{ie} = \left.\frac{V_b}{i_b}\right|_{V_c=0}$ (Ω) short circuit input impedance

$h_{fe} = \left.\frac{i_c}{i_b}\right|_{V_c=0}$ short circuit forward current gain

$h_{re} = \left.\frac{V_b}{V_c}\right|_{i_b=0}$ open circuit reverse voltage gain

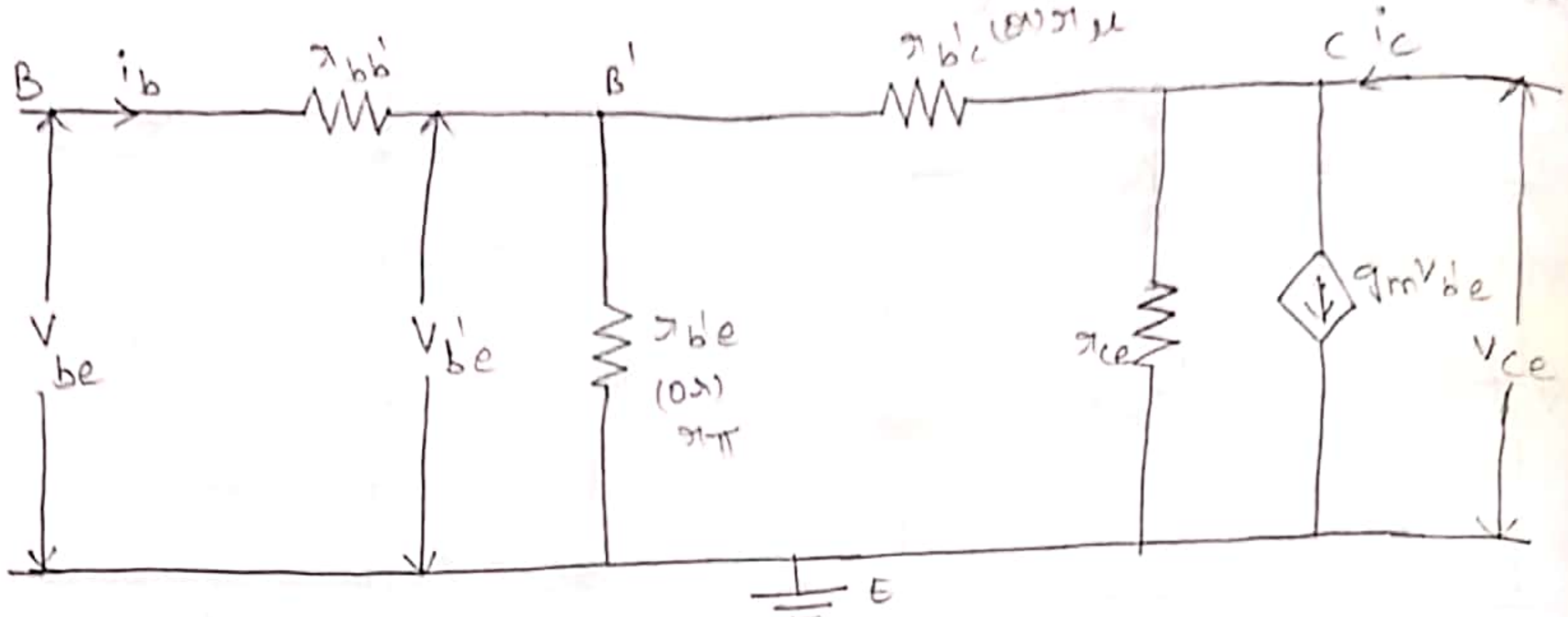
$h_{oe} = \left.\frac{i_c}{V_c}\right|_{i_b=0}$ $(\mho)$ open circuit output conductance

As above all four parameters are defined differently, they are called as Hybrid parameters (or) h-parameters.

consider hybrid - π model for CE transistor at low frequencies as shown in fig-b. It is observed that reactance of capacitor is inversely proportional to frequency

i.e., $X_C = \frac{1}{\omega C} = \frac{1}{2\pi f C}$

At low frequencies, as capacitive reactance is very high. C_e & C_c capacitors are treated as open circuit in fig-b



Hybrid-π model of CE transistor at low frequency

Where seven hybrid-π parameters are

g_m → Trans conductance

r_{ce} → Resistance between collector and emitter

$r_{b'c}$ (or) r_μ → Resistance b/n virtual base and collector

$r_{b'e}$ (or) r_π → Resistance b/n virtual base and emitter

$V_{b'e}$ → Voltage b/n virtual base and emitter

$r_{bb'}$ → Resistance b/n actual base and virtual base

V_{be} → Voltage b/n actual base and emitter.

V_{ce} → Voltage b/n collector and emitter

C_e (or) C_π → Diffusion capacitance b/n forward biased emitter base junction

C_c (or) C_μ → Transition capacitance b/n reverse biased collector base junction

Trans conductance (or) mutual conductance $g_m(\mho)$:

By definition

$$g_m = \left.\frac{\partial I_C}{\partial V_{B'E}}\right|_{V_{CE} = \text{constant}}$$

For npn transistor

$$I_C = \alpha I_E + I_{CO}$$

$$\frac{\partial I_C}{\partial V_{B'E}} = \alpha \frac{\partial I_E}{\partial V_{B'E}} + 0$$

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

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

$$= \frac{\alpha}{\left(\frac{\partial V_E}{\partial I_E}\right)} = \frac{\alpha}{r_e}$$ where r_e is dynamic resistance of forward biased emitter diode.

$$= \frac{\alpha}{\left(\frac{V_T}{I_E}\right)}$$

$$= \frac{\alpha I_E}{V_T}$$

$$\simeq \frac{|I_C|}{V_T}$$

where $V_T = \frac{KT}{q} = \frac{T}{11600} = 26mV$ (at room temperature)

$$\boxed{g_m = \frac{|I_C|}{V_T}} \quad (v)$$

It is observed at g_m directly proportional to collector current (I_C) and thermal voltage (V_T) and inversely proportional to temperature (T).

hence

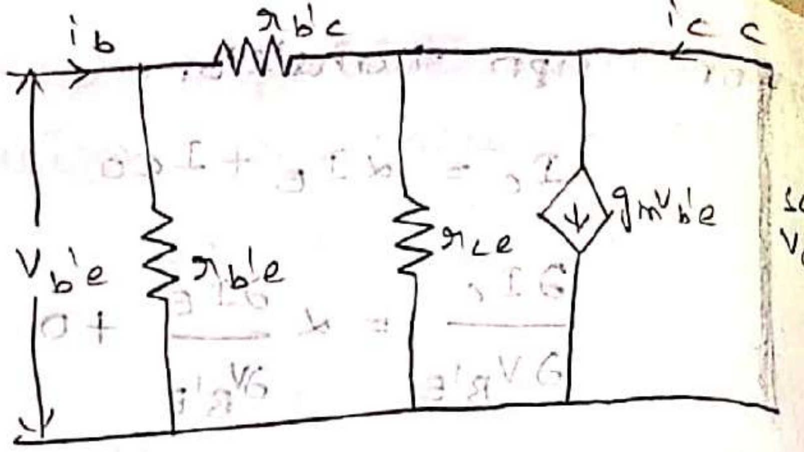
Resistance b/n virtual base (B') and emitter (E) $r_{b'e}$ (Ω):

Consider short circuit forward current gain

$$h_f = \left.\frac{i_c}{i_b}\right|_{V_c = 0}$$

From fig-(b) with $V_c = 0$ most of the current i_b flows through $r_{b'e}$ because r_{ce} is short circuited and $r_{b'e} << r_{b'c}$.

Now $V_{b'e} = i_b r_{b'e}$



from fig-(b) with short circuit $i_c = g_m V_{b'e}$

$$= g_m i_b r_{b'e}$$

$$\frac{i_c}{i_b} = g_m r_{b'e}$$

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

$$\boxed{r_{b'e} = \frac{h_{fe}}{g_m} \Omega}$$

Resistance between virtual base and collector ($r_{b'c}$)

consider open circuit reverse voltage gain

$$h_{re} = \frac{V_b}{V_c}\Big|_{i_b=0}$$

from fig(b) $i_b = 0$

$$V_{b'e} = \left[\frac{V_c}{r_{b'c} + r_{b'e}}\right] r_{b'e}$$

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

$$\frac{V_b}{V_c} = \frac{r_{b'e}}{r_{b'c}}$$

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

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

Resistance b/n actual base & virtual base :- $r_{bb'}(\Omega)$

consider short circuit input impedance

$$h_{ie} = \frac{V_b}{I_b}\bigg|_{V_c=0} \ (\Omega)$$

with $r_c = 0$, $r_{b'e}$ & $r_{b'c}$ are in parallel, since r_{ce} is short circuit.

Hence

$$r_{b'e} \parallel r_{b'c} = \frac{r_{b'e} \cdot r_{b'c}}{r_{b'e} + r_{b'c}} = r_{b'e} \quad [\because r_{b'e} << r_{b'c}]$$

Now,

$$V_b = i_b(r_{bb'} + r_{b'e})$$

$$\frac{V_b}{I_b} = r_{bb'} + r_{b'e}$$

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

$$\boxed{r_{bb'} = h_{ie} - r_{b'e}} \ \Omega$$

Relation b/n collector and emitter :- r_{ce}

consider open circuit output conductances

$$h_{oe} = \frac{I_c}{V_c}\bigg|_{i_b=0}$$

from fig(B) By KCL

$$i_c = g_m V_{b'e} + \frac{V_c}{r_{ce}} + \frac{V_c}{r_{b'e} + r_{b'c}} \quad (\because V_b = V_{b'e})$$

$$= g_m V_b + \frac{V_c}{r_{ce}} + \frac{V_c}{r_{b'e} + r_{b'c}}$$

$$i_c = g_m h_{re} V_c + \frac{V_c}{r_{ce}} + \frac{V_c}{r_{b'c}} \quad (\because h_{re} = \frac{V_b}{V_c} \ \& \ r_{b'e} << r_{b'c})$$

$$i_c = V_c\left[g_m h_{re} + \frac{1}{r_{ce}} + \frac{1}{r_{b'c}}\right]$$

$$\frac{i_c}{V_c} = g_m h_{re} + g_{ce} + g_{b'c}$$

$$h_{oe} = \frac{h_{fe}}{r_{b'e}} \times \frac{r_{b'e}}{r_{b'c}} + g_{ce} + g_{b'c}$$

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

$$h_{oe} = (1 + h_{fe})\, g_{b'c} + g_{ce}$$

$$\boxed{g_{ce} = h_{oe} - (1 + h_{fe})\, g_{b'c}} \ (\mho) \qquad r_{ce} = \frac{1}{g_{ce}} \ \Omega$$

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Transition capacitance : (C_c)

It is measured b/n reverse biased collector base junction is C_b (common base) configuration and is given by manufacturer over the data sheet as C_{ob}. Therefore $\boxed{C_c \text{ (or) } C_\mu = C_{ob}}$

Diffusion capacitance (C_e):

It is measured b/n forward biased emitter base junction in CB configuration and is given by C_e (or) C_π

$$C_e \text{ (or) } C_\pi = \frac{g_m}{2\pi f_T}$$

where f_T is the frequency at which short circuit current gain falls to unity i.e., $h_{fe}=1$

Conclusions:

1. $g_m = \frac{|I_C|}{V_T}$ (℧) where $V_T = \frac{T}{11600}$

2. $r_{b'e} = \frac{h_{fe}}{g_m}$ (Ω)

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

4. $r_{bb'} = h_{ie} - r_{b'e}$

5. $r_{ce} = \frac{1}{g_{ce}}$

6. C_c (or) $C_\mu = C_{ob}$

7. C_e (or) $C_\pi = \frac{g_m}{2\pi f_T}$

* The following low frequency parameters for given transistor $I_C = 5mA$, $V_{CE} = 10V$ at room temperature, $h_{ie} = 600\Omega$, $h_{fe} = 100$, $h_{re} = 10^{-4}$, $h_{oe} = 20\mu\Omega$. at the same operating point $f_T = 500MHz$ $C_{ob} = 3pF$ calculating values of Hybrid-π parameters.

Given data

collector current $(I_C) = 5mA$

Voltage b/n collector and emitter $(V_{CE}) = 10V$

h_{ie} = short circuit input impedance $= 600\Omega$

Short circuit forward current gain $(h_{fe}) = 100$

Open circuit reverse voltage gain $(h_{re}) = 10^{-4}$

Open circuit output conductance $(h_{oe}) = 20\mu\Omega$

Frequency at room temperature $(f_T) = 500MH$

Transition capacitance $(C_{ob}) = 3pF$

$$g_m = \frac{|I_C|}{V_T} \mho \qquad V_T = \frac{T}{11600}$$

$$= \frac{5\times10^{-3}}{26\times10^{-3}} = 0.1923\,\mho = 192\,m\mho$$

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

$$= \frac{100}{0.1923} = 520\Omega$$

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

$$= \frac{520}{10^{-4}}$$

$$= 5200000$$

$$= 5.2\,M\Omega$$

$$r_{bb'} = h_{ie} - r_{b'e}\,(\Omega)$$

$$= 600 - 520 = 80\Omega$$

$r_{ce} = \frac{1}{g_{ce}}$

$g_{ce} = h_{oe} - (1+h_{fe})\, g_{b'c}$

$g_{b'c} = \frac{1}{r_{b'c}} = \frac{1}{5.2\times10^{-6}} = 192307.6923$
$= 0.1923$

$g_{ce} = (20\times10^{-6}) - (1+100)\,192037.6$
$= -19423076.92 = -0.19418$

$r_{ce} = \frac{1}{-0.19418} = -5.1498$

Tranition capacitance C_c (or) $C_\mu = C_{ob}$

$\rightarrow C_c$ (or) $C_\mu = 3pf$

Diffusion capacitance C_e (or) $C_\pi = \frac{g_m}{2\pi f_T}$

$C_\pi = \frac{0.1923}{2\times3.14\times500\times10^{-6}}$

$C_\pi = \frac{9615}{157} = 61.24\,2F$

* The following low frequency parameters for a given $I_c = 5mA$, $V_{CE} = 8V$ at room temperature $h_{ie} = 1k\Omega$, $h_{fe} = 100$, $h_{re} = 10^{-4}$, $h_{oe} = 4\times10^{-5}\, A/V$ at the same operating point, $f_T = 10MHz$, $C_{ob} = 2pf$. Calculate the values of hybrid-π parameters (or) high frequency parameters.

Given

collector current $I_c = 5mA$

voltage b/n collector and emitter $V_{ce} = 8\Omega$

short circuit input impedance $h_{ie} = 1k\Omega$

short circuit forward current gain $h_{fe} = 100$

Open circuit reverse voltage gain $h_{re} = 10^{-4}$

Open circuit output conductance $h_{oe} = 4\times10^{-5}$

$f_T = 10MHz$

$C_{ob} = 2pf = 2\times10^{-12}F$

(i) Trans conductance $g_m = \frac{|I_c|}{V_T} = \frac{5\times10^{-3}}{26\times10^{-3}} = 0.1923\mho$

(ii) Resistance b/n virtual base and emitter

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

$$r_{b'e} = \frac{100}{0.1923} = 520.02$$

(iii) Resistance b/n virtual base & collector

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

$$r_{b'c} = \frac{100}{0.1923} \cancel{= 520.02} = \frac{520.02}{10^{-4}} = 5200200$$

(iv) Resistance b/n actual base & virtual base

$$r_{bb'} = h_{ie} - r_{b'e}$$

$$r_{bb'} = 1\times10^{3} - 520.02$$

$$= 479.98.$$

(v) Resistance b/n collector & emitter

$$r_{ce} = \frac{1}{g_{ce}}$$

$$g_{ce} = h_{oe} - (1+h_{fe})\, g_{b'c}$$

$$= h_{oe} - (1+h_{fe})\frac{1}{r_{b'c}}$$

$$g_{ce} = 4\times10^{-5} - (1+100)\frac{1}{5200200}$$

$$= 4\times10^{-5} - (101)\frac{1}{5200200}$$

$$g_{ce} = 2.0577\times10^{-5}$$

$$\Rightarrow r_{ce} = \frac{1}{2.0577\times10^{-5}} = 48597.94$$

(vi) Transmission capacitance C_c (or) $C_\mu = C_{ob} = 2pF$

(vii) Diffusion capacitance C_e (or) $C_\pi = \frac{g_m}{2\pi f_T}$

$$= \frac{0.1923}{2\times3.14\times10\times10^{-3}}$$

$$= \frac{1923}{628}$$

$$= 3.0621$$

4) Common Emitter Amplifier with short circuit load ($R_L = 0\Omega$) at high frequencies:

→ Consider a single stage CE amplifier with short circuit load (by short circuiting b/n collector and emitter to make $R_L = 0\Omega$) as shown in fig-(a).

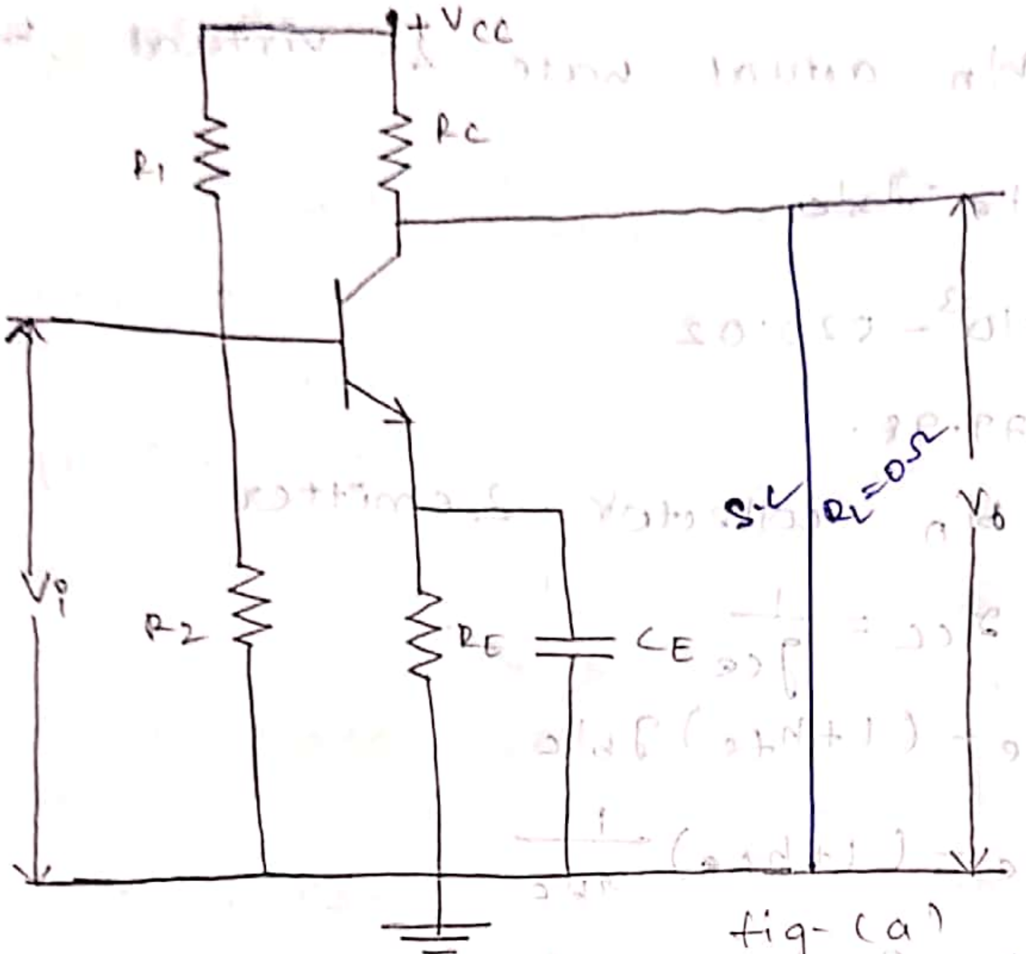


fig-(a)

CE amplifier high frequencies by using hybrid-π model at high frequencies is shown in fig.

→ Small signal equivalent circuit of CE amplifier:

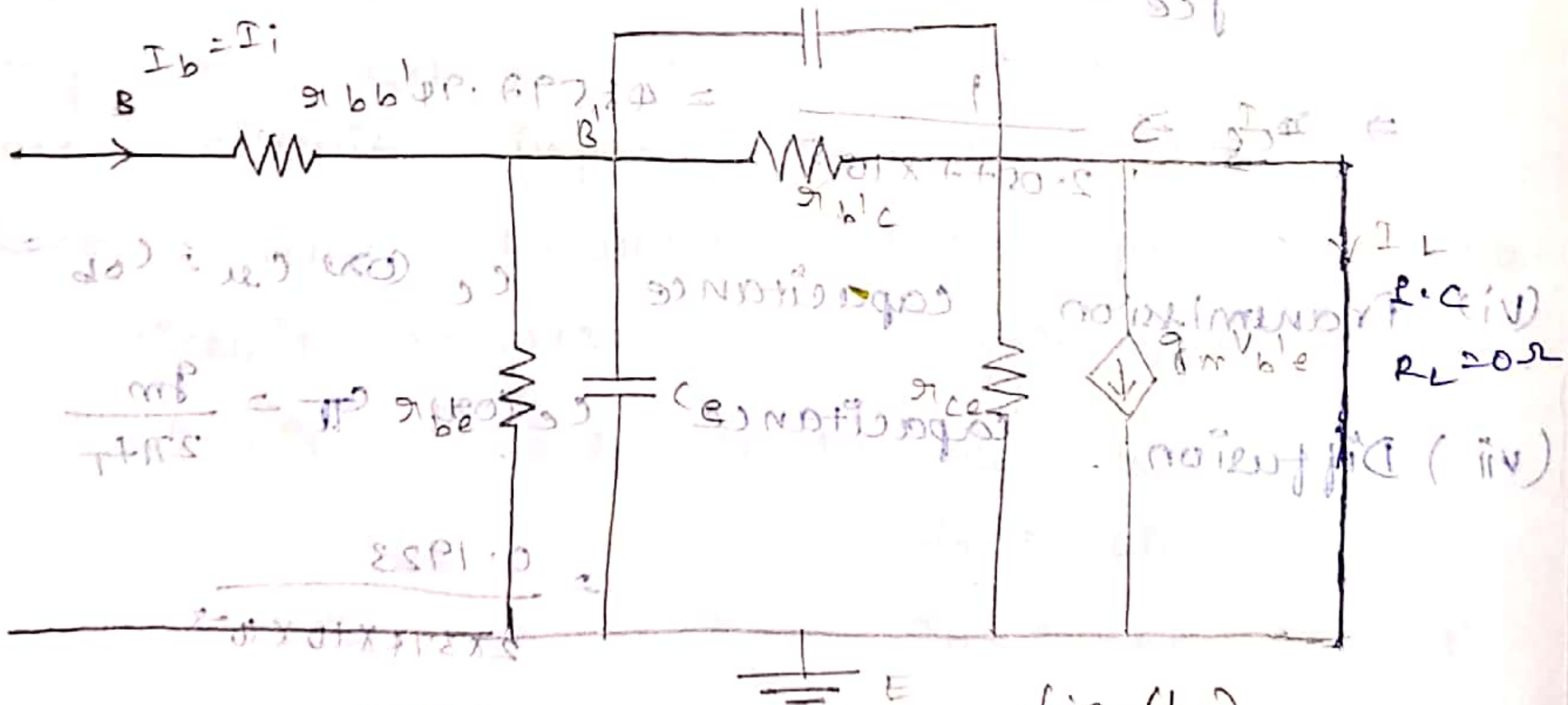


fig-(b)

Approximations:

→ R_{CE} disappear because it is short circuited

→ $r_{b'e}$ & $r_{b'c}$ are in parallel because of short circuit i.e., $r_{b'e} \| r_{b'c} = \dfrac{r_{b'e}\, r_{b'c}}{r_{b'e} + r_{b'c}} \simeq r_{b'e}$ ($\because r_{b'e} << r_{b'c}$)

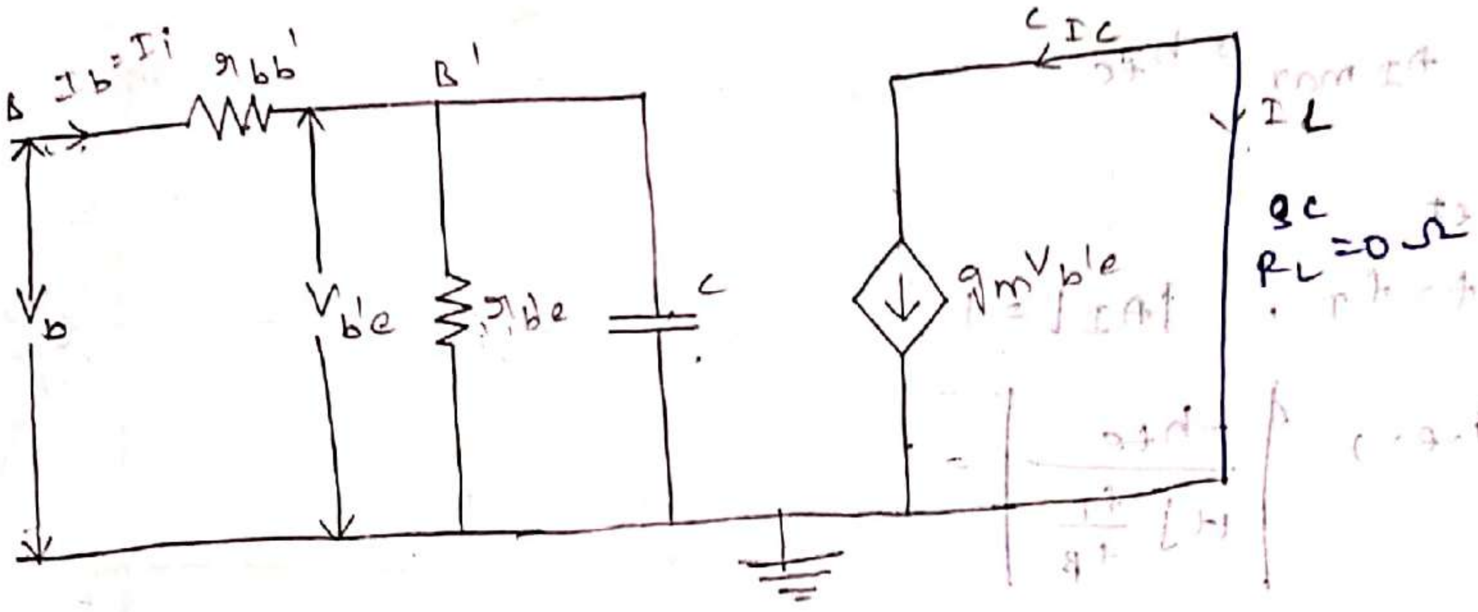
Hence $r_{b'c}$ disappear.

→ C_e & C_c are in parallel because of short circuit,

$\therefore C = C_e + C_c$

→ current to be delivered from input to output node can be neglected.

With the above four approximations, approximate equivalent circuit is shown in fig-(c)



By definition,

short circuit current gain

$$A_I = \frac{I_L}{I_i} = \frac{-I_C}{I_b} = \frac{-g_m V_{b'e}}{\left[\frac{V_{b'e}}{\left(r_{b'e} \| \frac{1}{j\omega C}\right)}\right]}$$

$$A_I = \frac{-g_m}{\left[\frac{r_{b'e} + \frac{1}{j\omega C}}{r_{b'e} \times \frac{1}{j\omega C}}\right]} = \frac{-g_m r_{b'e}}{1 + j\omega r_{b'e} C}$$

$$= \frac{-h_{fe}}{1 + j\,2\pi f\, r_{b'e} C} \qquad \left[\because r_{b'e} = \frac{h_{fe}}{g_m} \;\&\; \omega = 2\pi f\right]$$

$$\left[1 << \right] = \frac{-h_{fe}}{1 + j\left(\frac{f}{f_\beta}\right)} \qquad \text{where } f_\beta = \frac{1}{2\pi r_{b'e} C}$$

At $f = f_\beta$

$$|A_I| = \left|\frac{-h_{fe}}{1 + j\frac{f_\beta}{f_\beta}}\right| = \left|\frac{-h_{fe}}{1 + j1}\right|$$

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

Hence f_β is the frequency at which short circuit CE current gain falls to $\frac{1}{\sqrt{2}}$ of its maximum value (or) short CE circuit gain falls to below 3dB line from its maximum value in dB.

At $f =$ very small

$$|A_I| \approx \left|\frac{-h_{fe}}{1+j(0)}\right| \simeq h_{fe}$$

$$A_{I\,max} = h_{fe}$$

Let

At $f = f_T$, $|A_I| = 1$

i.e., $$\left|\frac{-h_{fe}}{1+j\frac{f_T}{f_\beta}}\right| = 1$$

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

squarring on both sides

$$\frac{h_{fe}^2}{1+\frac{f_T^2}{f_\beta^2}} = 1$$

$$h_{fe}^2 = 1 + \frac{f_T^2}{f_\beta^2}$$

$$h_{fe}^2 - 1 = \frac{f_T^2}{f_\beta^2}$$

$$h_{fe}^2 \simeq \frac{f_T^2}{f_\beta^2} \quad \left[\because h_{fe}^2 >> 1\right]$$

Square root on both sides

$$h_{fe} = \frac{f_T}{f_\beta}$$

$$\boxed{f_T = h_{fe} f_\beta}$$

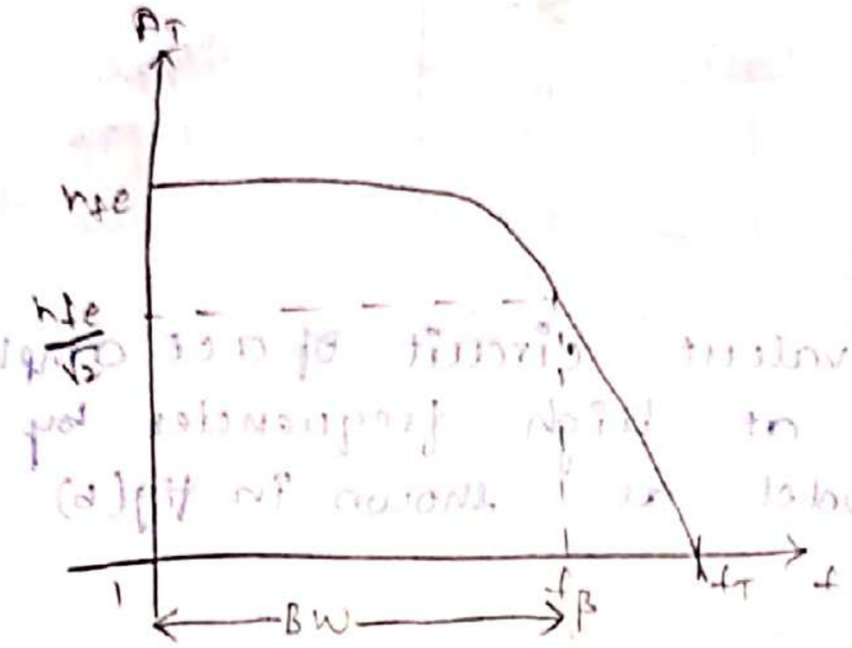
$$f_T = h_{fe}\frac{1}{2\pi r_{b'e} C}$$

$$f_T = \frac{g_m}{2\pi(C_e + C_c)} \quad ; \quad r_{b'e} = \frac{h_{fe}}{g_m}$$

$$= \frac{g_m}{2\pi C_e} \quad [\because C_e >> C_c]$$

$$\boxed{f_T = \frac{g_m}{2\pi C_e}}$$

Frequency response of CE amplifier with short circuit load is shown in figure



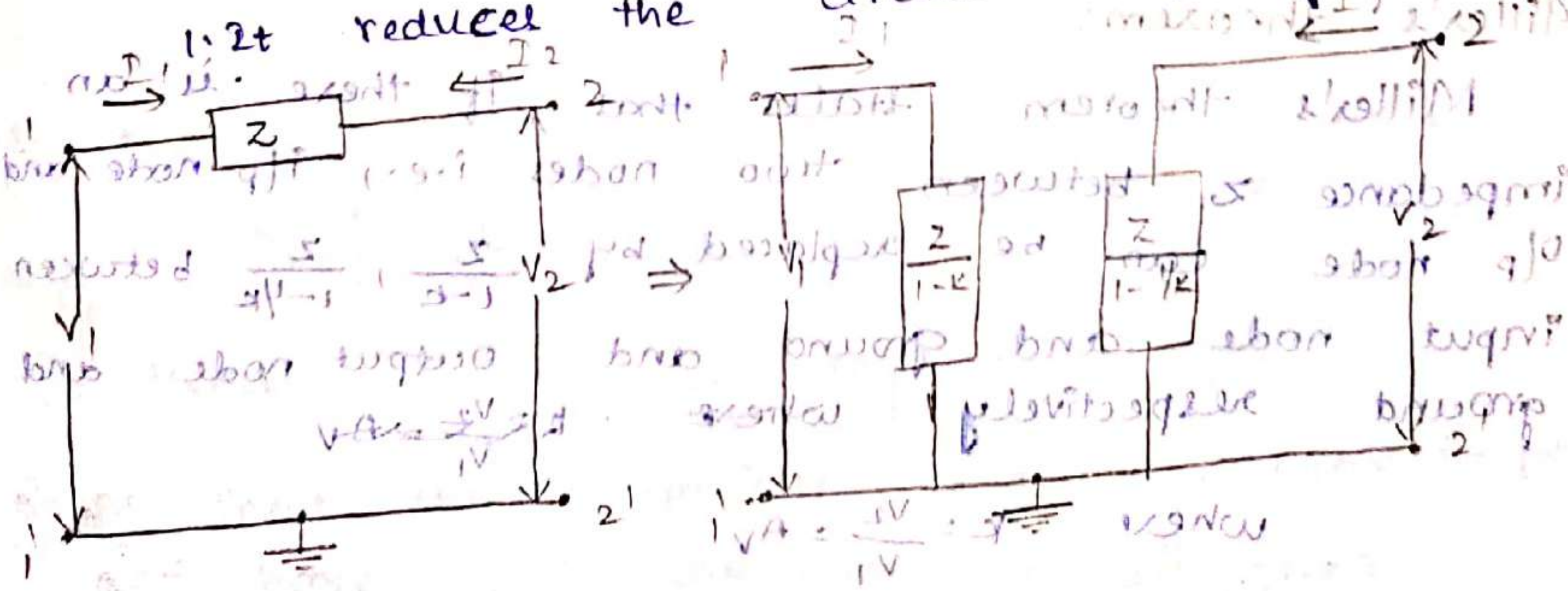
gain bandwidth product is

$$\boxed{|A_I| \times BW = h_{fe} f_\beta = f_T}$$

Miller's theorem:

Miller's theorem states that if there is an impedance z between two nodes i.e., i/p node and o/p node can be replaced by two impedances $\frac{z}{1-k}$, $\frac{z}{1-1/k}$ between input node and ground, output node and ground respectively where $k = \frac{V_2}{V_1} = A_V$

Applications of miller's theorem are
1. It reduces the circuit complexity



4) Common Emitter (CE) Amplifier with resistive load (R_L) at high frequencies :-

Consider a single stage CE amplifier with load resistor R_L as shown in figure (a)

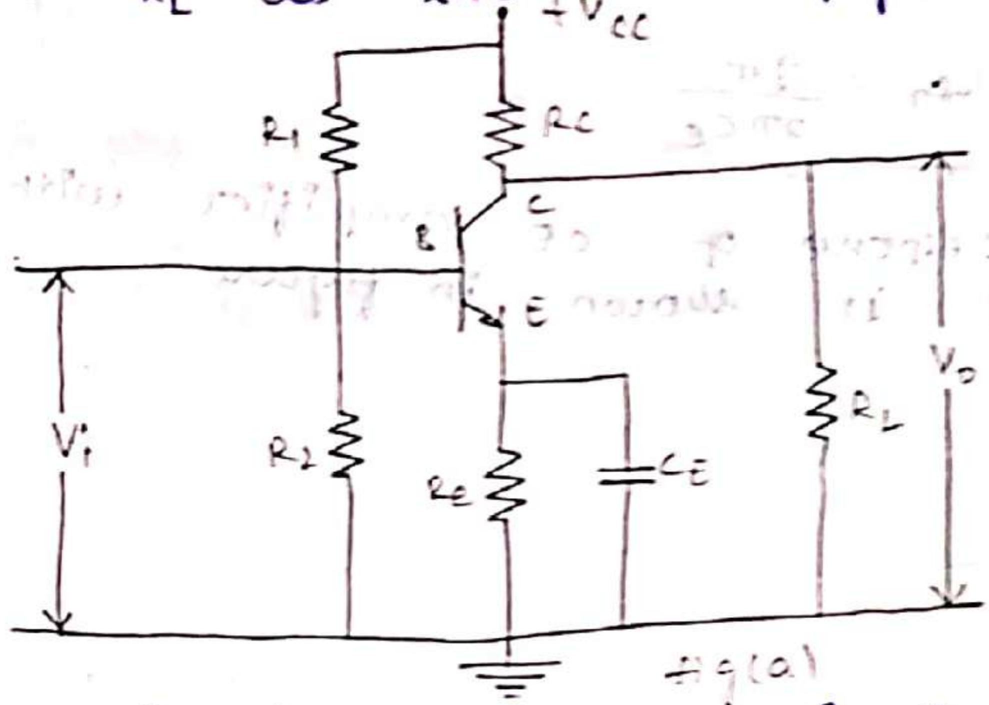


fig (a)

→ Small signal equivalent circuit of a CE amplifier with resistive load at high frequencies by using hybrid π model as shown in fig (b)

→ $I_i = I_b$

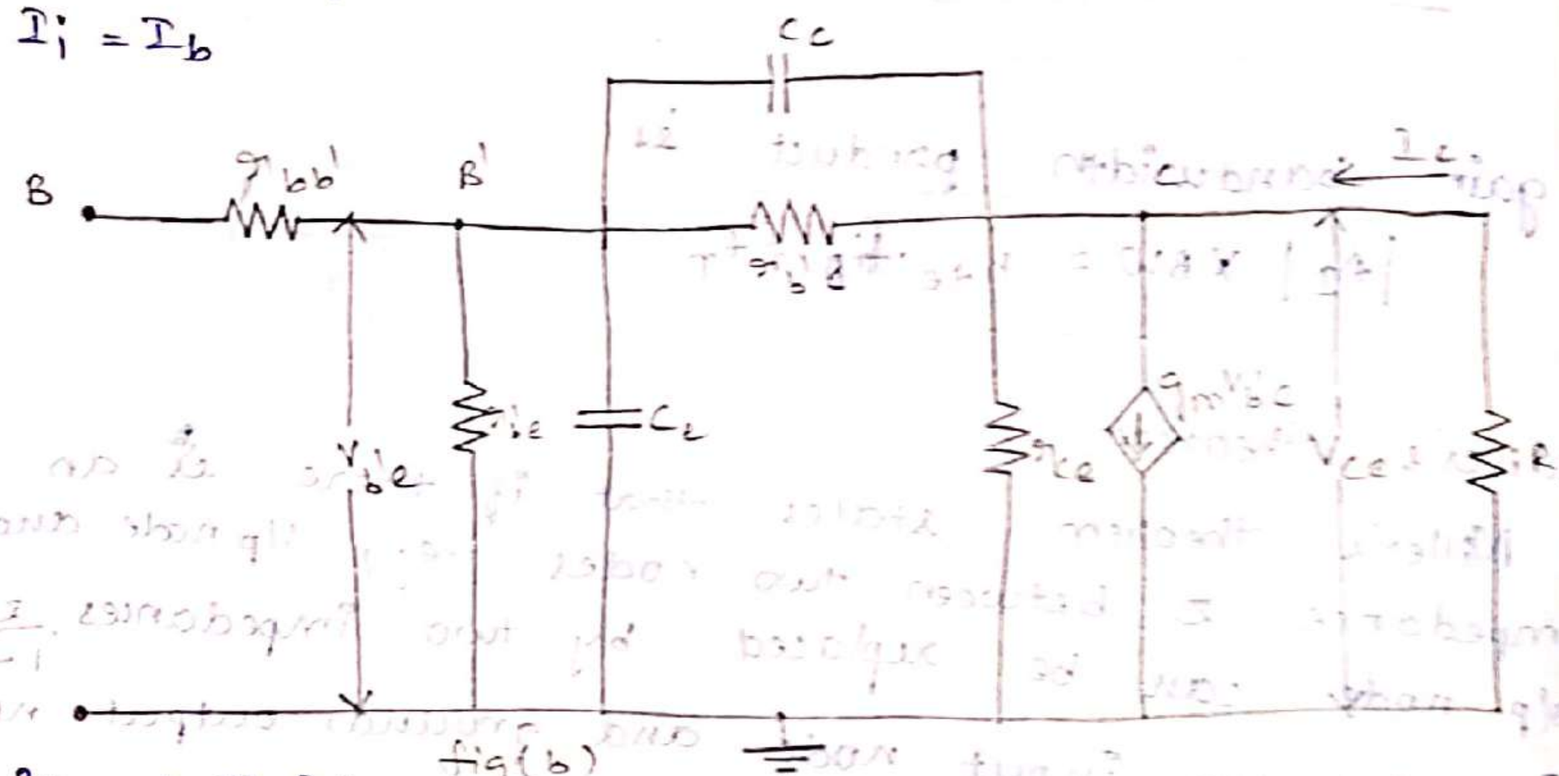


fig (b)

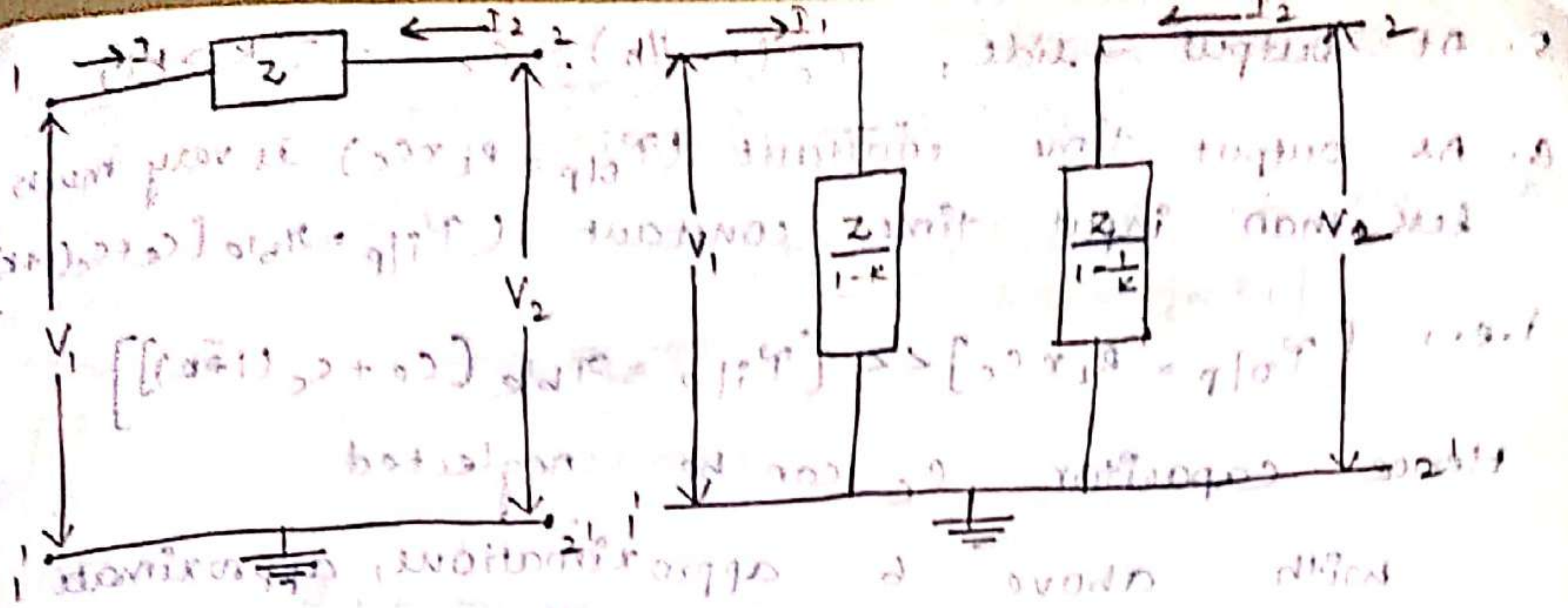
By definition

$$\text{voltage gain } (A_V) = K = \frac{V_{ce}}{V_{b'e}}$$

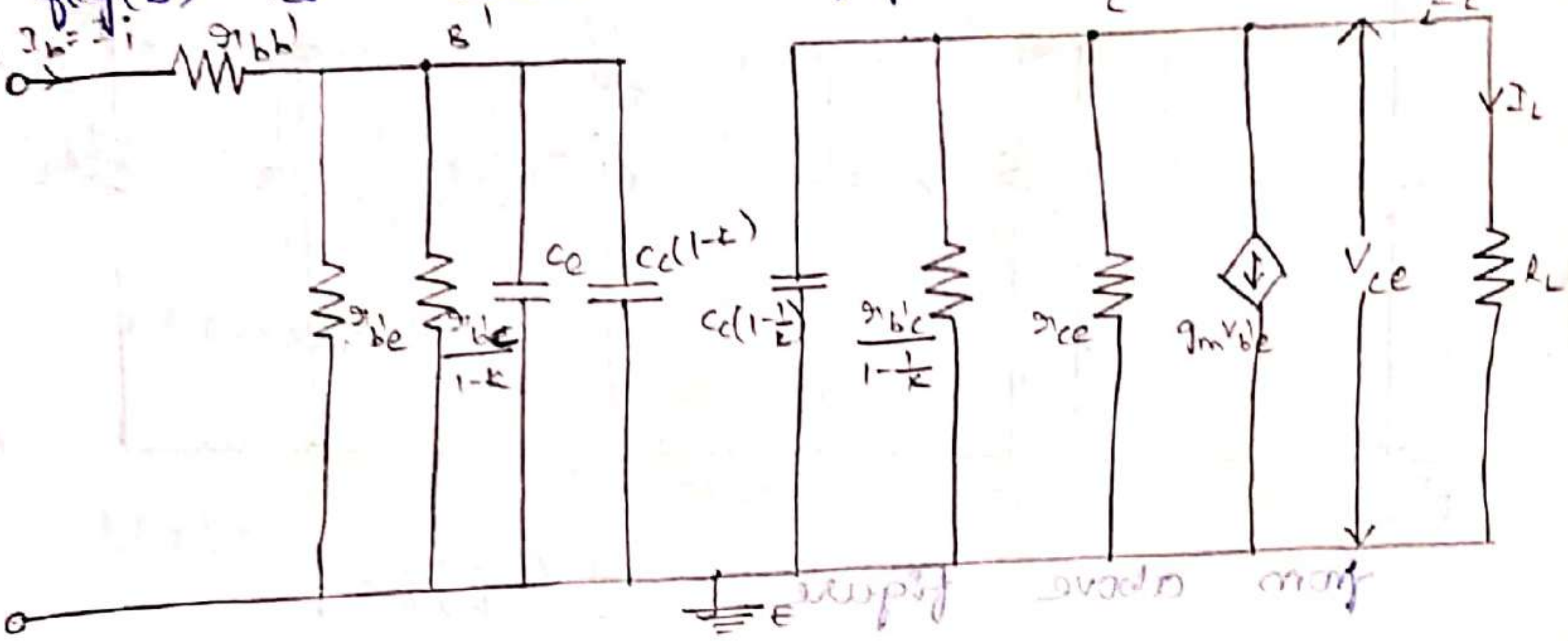
Miller's theorem:

Miller's theorem states that if there is an impedance Z between two nodes i.e., i/p node and o/p node can be replaced by $\frac{Z}{1-K}$, $\frac{Z}{1-1/K}$ between input node and ground and output node and ground respectively ~~where~~ $\cancel{K = \frac{V_2}{V_1} = A_V}$

where $K = \frac{V_2}{V_1} = A_V$



On application of Miller's theorem equivalent circuit of fig(b) is shown in fig (c)



Where $K = \frac{V_{ce}}{V_{b'e}}$

· Approximations:

1. At output side $\frac{r_{b'c}}{1-1/K}$ is approximately $r_{b'c}$ [$\because K>>1$]

$$\frac{r_{b'c}}{1-1/K} \simeq r_{b'c}$$

2. At output side three resistors are in parallel i.e.,
$r_{b'c} \| r_{ce} \| R_L \simeq R_L$ [$\because r_{b'c} >> r_{ce} >> R_L$] hence $r_{b'c}$ and r_{ce} can be neglected

3. At input side

$$r_{b'e} \| \frac{r_{b'c}}{1-K} \simeq [r_{b'e}]$$ [$\because r_{b'e} << r_{b'c}$]

Hence $\frac{r_{b'c}}{1-K}$ can be neglected

4. At input side, capacitors C_e and $C_c(1-K)$ are in parallel and hence total capacitance, $C = C_e \| C_c(1-K)$

$$C \simeq C_e + C_c(1-K)$$

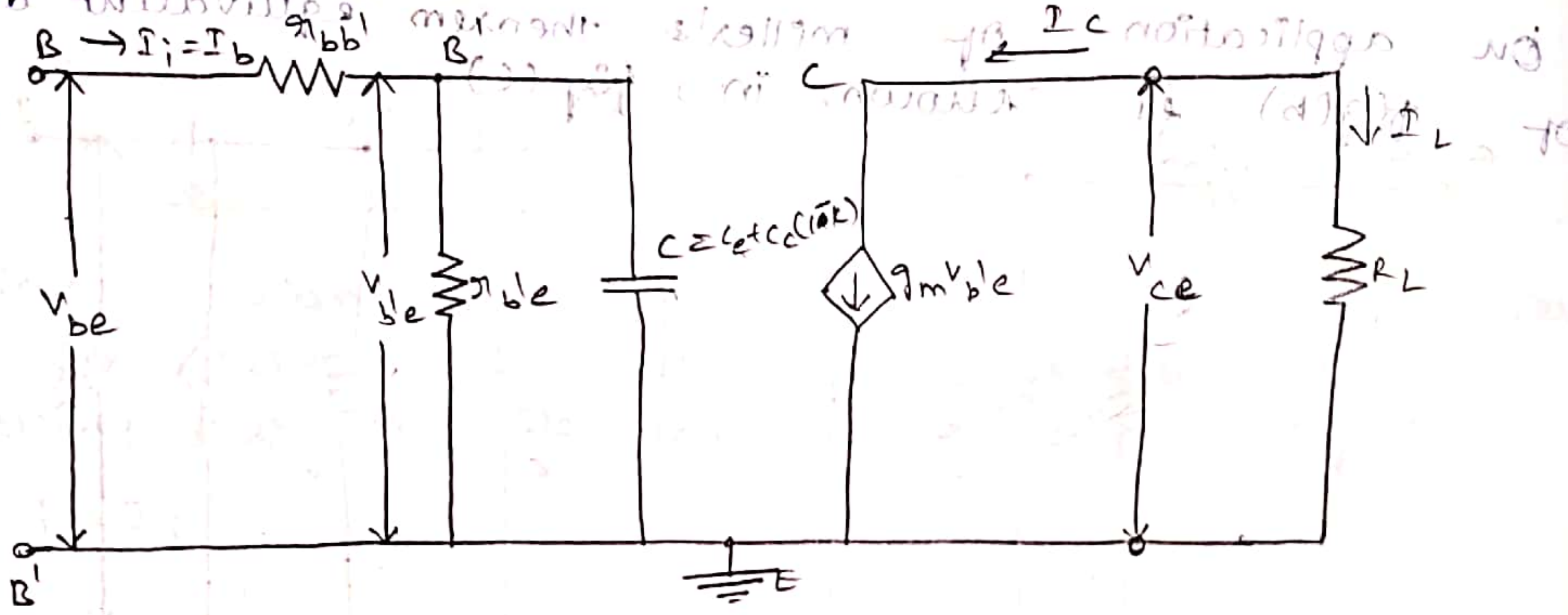
5. At output side, $C_c(1-1/K) \simeq C_c$ [$\because K >> 1$]

6. As output time constant ($\tau_{o/p} = R_L \times C_c$) is very much less than input time constant ($\tau_{i/p} = r_{b'e}[C_e + C_c(1+K)]$)

i.e., $[\tau_{o/p} = R_L \times C_c] << [\tau_{i/p} = r_{b'e}[C_e + C_c(1+K)]]$

Hence capacitor C_c can be neglected

With above 6 approximations, approximate equivalent circuit is as follows



from above figure

$$V_{ce} = (-g_m V_{b'e}) R_L$$

Voltage gain $\Rightarrow \dfrac{V_{ce}}{V_{b'e}} = -g_m R_L \Rightarrow A_v = K = -g_m R_L$

[$K < 0$ CE]

current gain with resistive load $A_I = \dfrac{I_L}{I_i} = \dfrac{I_c}{I_b}$

$$I_L = -g_m V_{b'e}$$

$$I_i = \frac{V_{b'e}}{r_{b'e} \| \frac{1}{j\omega C}}$$

$$I_i = \frac{V_{b'e}}{\left[\dfrac{r_{b'e} \times \frac{1}{j\omega C}}{r_{b'e} + \frac{1}{j\omega C}}\right]}$$

$$A_I = \frac{-g_m \cancel{V_{b'e}}}{\left[\dfrac{\cancel{V_{b'e}}}{\left[\dfrac{r_{b'e} \times 1/j\omega C}{r_{b'e} + 1/j\omega C}\right]}\right]}$$

$$A_I = \frac{-g_m r_{b'e}}{1 + j\omega r_{b'e} C}$$

$$= \frac{-h_{fe}}{1+j2\pi f r_{b'e}[C_e + C_c(1+g_m R_L)]}$$

$\therefore [h_{fe} = g_m r_{b'e}, \omega = 2\pi f, C = C_e + C_c(1-K)$ & $K = -g_m R_L]$

$$A_I = \frac{-h_{fe}}{1+j\left(\frac{f}{f_H}\right)} \quad \text{where}$$

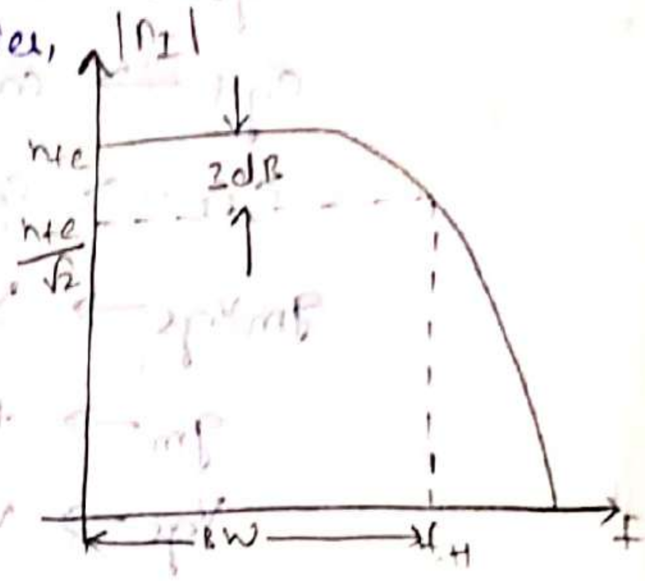
$$f_H = \frac{1}{2\pi r_{b'e}[C_e + C_c(1+g_m R_L)]}$$

from frequency response

If $f=0$ (or) very low frequencies,

$$|A_I| = \left|\frac{-h_{fe}}{1+j(0)}\right|$$

$\Rightarrow A_{I\,max} = h_{fe}$



At $f = f_H$

$$|A_I| = \left|\frac{-h_{fe}}{1+j\left(\frac{f}{f_H}\right)}\right| = \left|\frac{-h_{fe}}{1+j1}\right|$$

$$|A_I| = \left|\frac{-h_{fe}}{\sqrt{2}}\right| = \frac{h_{fe}}{\sqrt{2}}$$

from frequency response, f_H is the frequency at which CE current gain with resistive load falls to $\frac{1}{\sqrt{2}}$ of its maximum value (or) falls below 3dB of its maximum value in dB.

* current gain bandwidth product $A_I \times BW = |A_I|_{max} \times f_H$

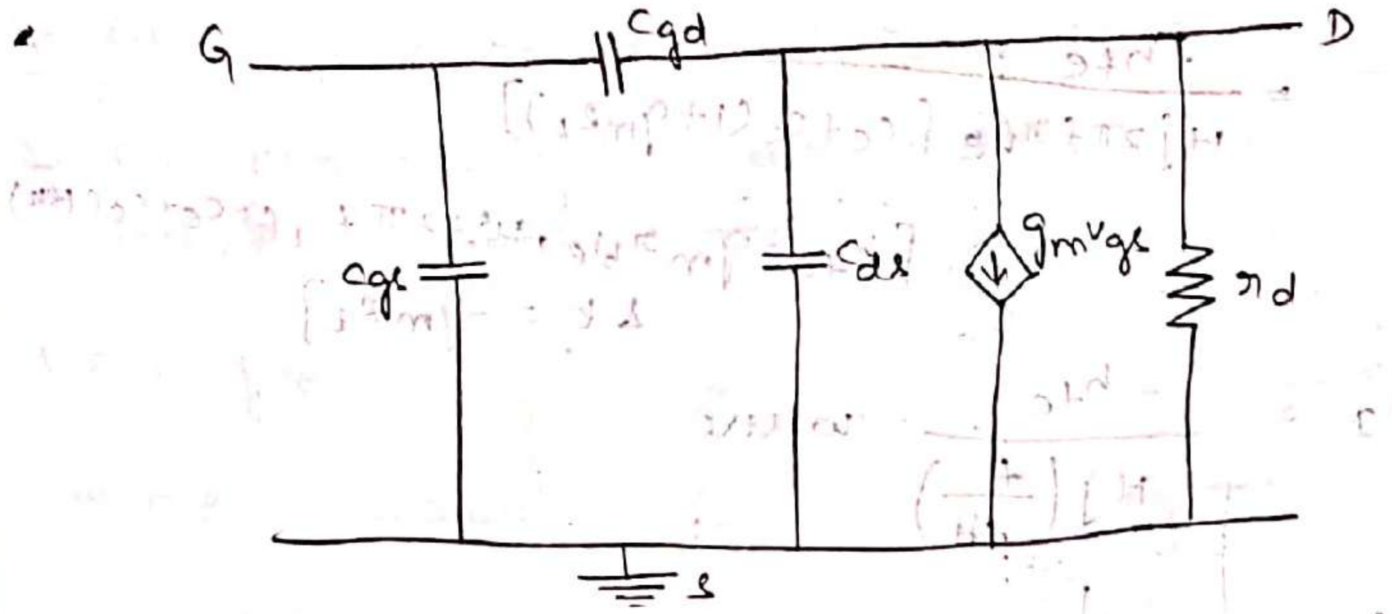
$$= h_{fe} \times \frac{1}{2\pi r_{b'e}[C_e + C_c(1+g_m R_L)]}$$

$$= \frac{g_m}{2\pi[C_e + C_c(1+g_m R_L)]} \qquad \boxed{\because r_{b'e} = \frac{h_{fe}}{g_m}}$$

Hybrid π model for CS FET:

In low frequency model of J-FET works for frequencies below 1MHz, for frequencies greater than 1MHz the response of JFET will be limited by internal paracetic capacitance of JFET.

The hybrid π-model for common source (CS) FET is shown in fig(a).



parameters of hybrid π FET model are:

where G, D, S are gate, drain, source.

$C_{gs} \rightarrow$ capacitance b/n gate and source

$C_{gd} \rightarrow$ capacitance b/n gate & drain

$C_{ds} \rightarrow$ capacitance b/n drain & source

$g_m V_{gs} \rightarrow$ voltage dependent current source

$g_m \rightarrow$ trans conductance

$V_{gs} \rightarrow$ voltage b/n gate & source

$r_d \rightarrow$ internal drain resistance

$\mu = g_m r_d \rightarrow$ Amplification factor

The three capacitors [C_{gs}, C_{gd}, C_{ds}] are indirectly given over the data sheet of JFET as

$C_{gd} = C_{iss} - C_{fs}$

$C_{gs} = C_{iss} - C_{rss}$

$C_{ds} = C_{os} - C_{rss}$

where C_{iss} = common source input capacitance

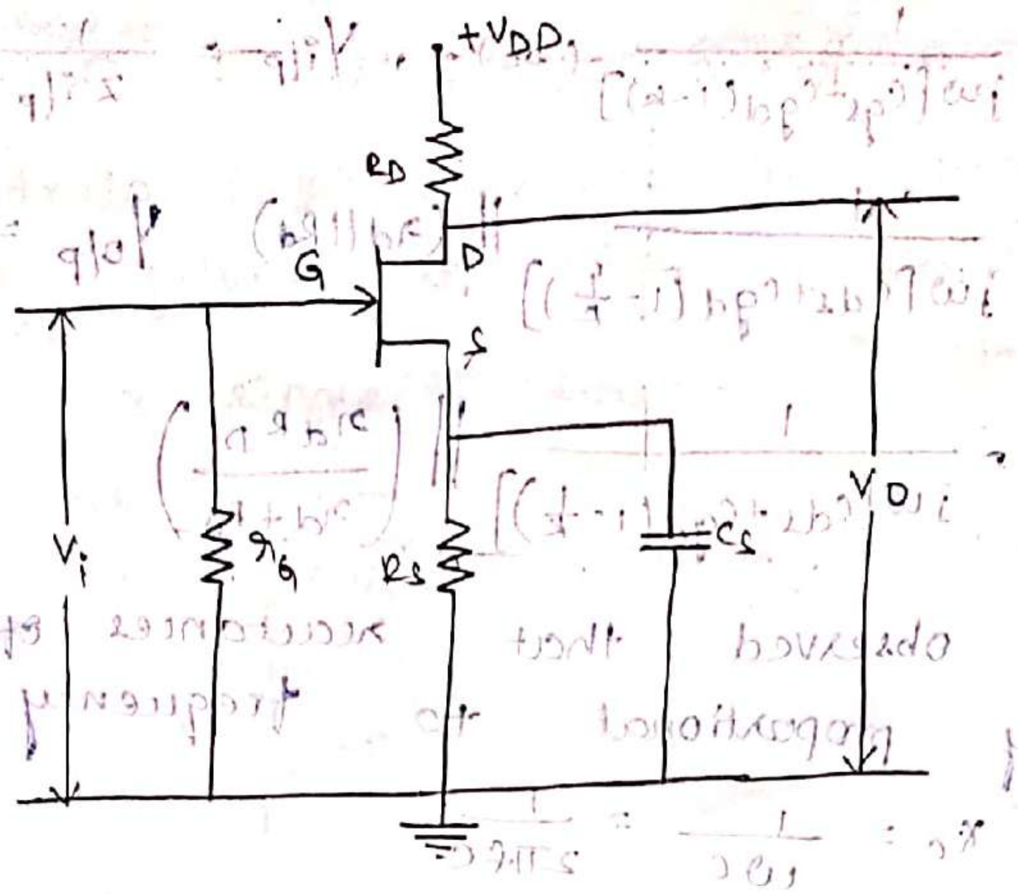
C_{fs} = forward transfer capacitance

C_{rs} = reverse transfer capacitance

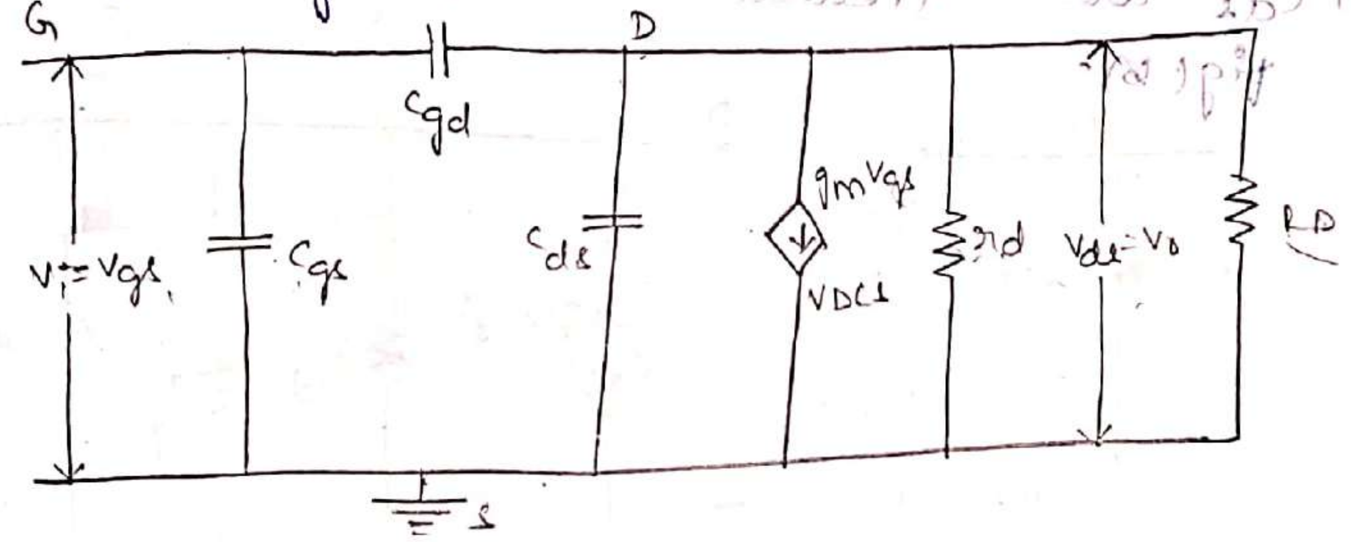
C_{os} = o/p capacitance

Common source amplifier at high frequencies:

Consider a single stage CS amplifier as shown in fig(a).



small signal equivalent circuit of CS amplifier at high frequencies by using hybrid-π mode is shown in fig(b).



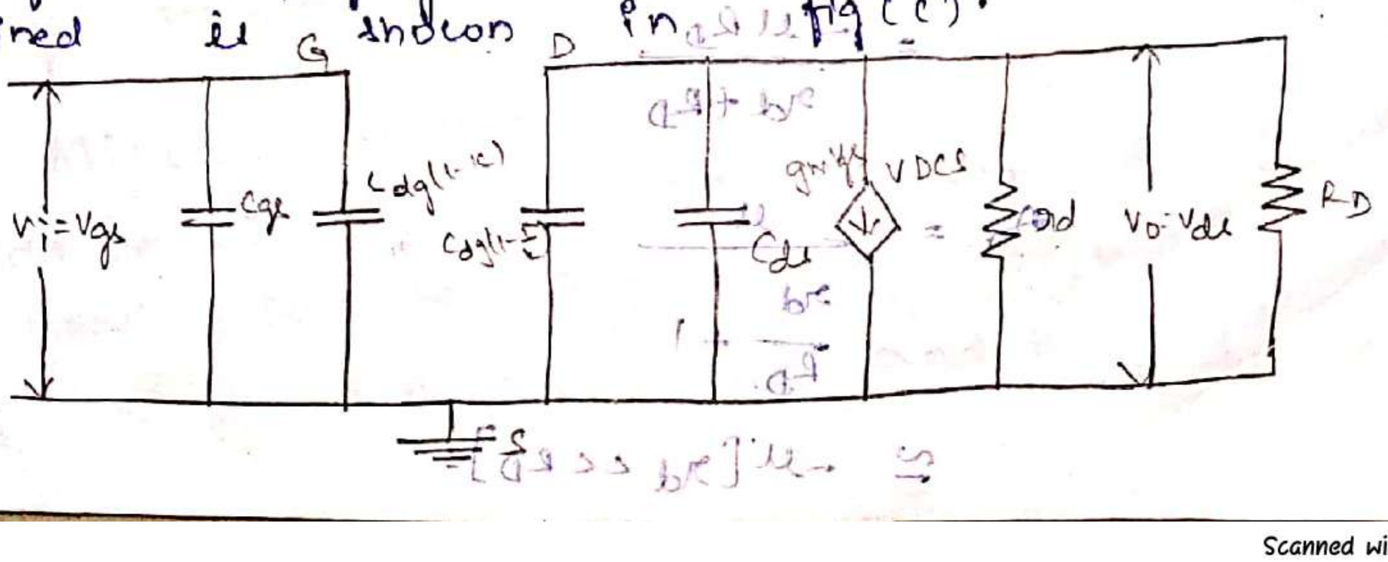
From figure $K = A_V = \frac{V_o}{V_i} = \frac{V_{ds}}{V_{gs}}$

voltage gain of CS amplifier.

Miller's theorem states that if there is impedance (Z) b/n two nodes i.e., input node & output node, can be replaced by two impedances $\frac{Z}{1-K}$, $\frac{Z}{1-1/K}$ b/n i/p node and ground, o/p node and ground respectively as shown below

where K = voltage gain.

After applying miller's theorem to fig(b). The circuit obtained is shown in fig(c).



$$Z_{i/p} = \frac{1}{j\omega[C_{gs}+C_{gd}(1-k)]} (\Omega) \quad , \quad Y_{i/p} = \frac{1}{Z_{i/p}} (\mho)$$

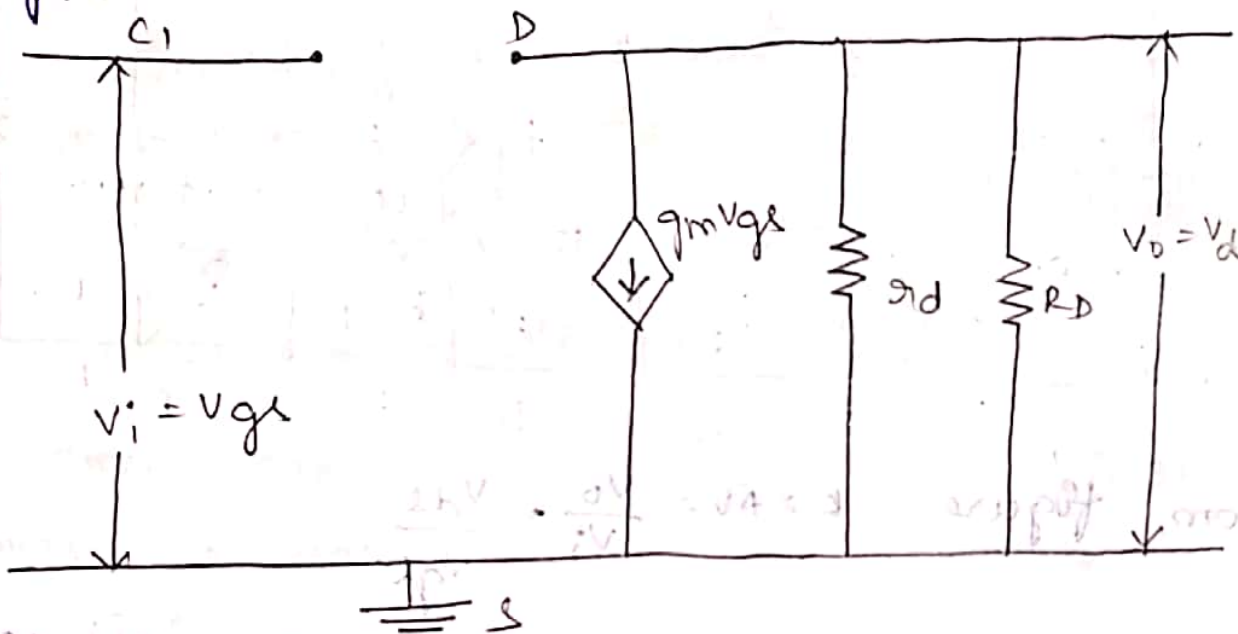
$$Z_{o/p} = \frac{1}{j\omega[C_{ds}+C_{gd}(1-\frac{1}{k})]} \parallel (r_d \parallel R_d) \quad Y_{o/p} = \frac{1}{Z_{o/p}} (\mho)$$

$$= \frac{1}{j\omega[C_{ds}+C_{gd}(1-\frac{1}{k})]} \parallel \left(\frac{r_d R_D}{r_d + R_D}\right)$$

It is observed that reactances of the capacitor is inversely proportional to frequency i.e.,

$$X_C = \frac{1}{\omega C} = \frac{1}{2\pi f C}$$

Hence at low frequencies has capacitive reactance very high the three capacitors are C_{gd}, C_{gs}, C_{ds} are treated as open circuit, as shown in fig(b).



High frequency hybrid -π model of CS FET at low frequencies

$$V_{ds} = -g_m V_{gs} (r_d \parallel R_D)$$

$$A_v = \frac{V_{ds}}{V_{gs}} = -g_m \left(\frac{r_d R_D}{r_d + R_D}\right)$$

$$= \frac{-g_m r_d R_D}{r_d + R_D}$$

$$= \frac{-\mu R_D}{r_d + R_D}$$

$$A_v = \frac{\mu}{\frac{r_d}{R_D}+1}$$

$$\simeq -\mu [r_d << R_D].$$

∴ R_D is also denoted R_L.

Common drain [CD] amplifier at high frequencies (or) source follower at high frequencies:

Consider a single stage common drain amplifier as shown in fig(a)

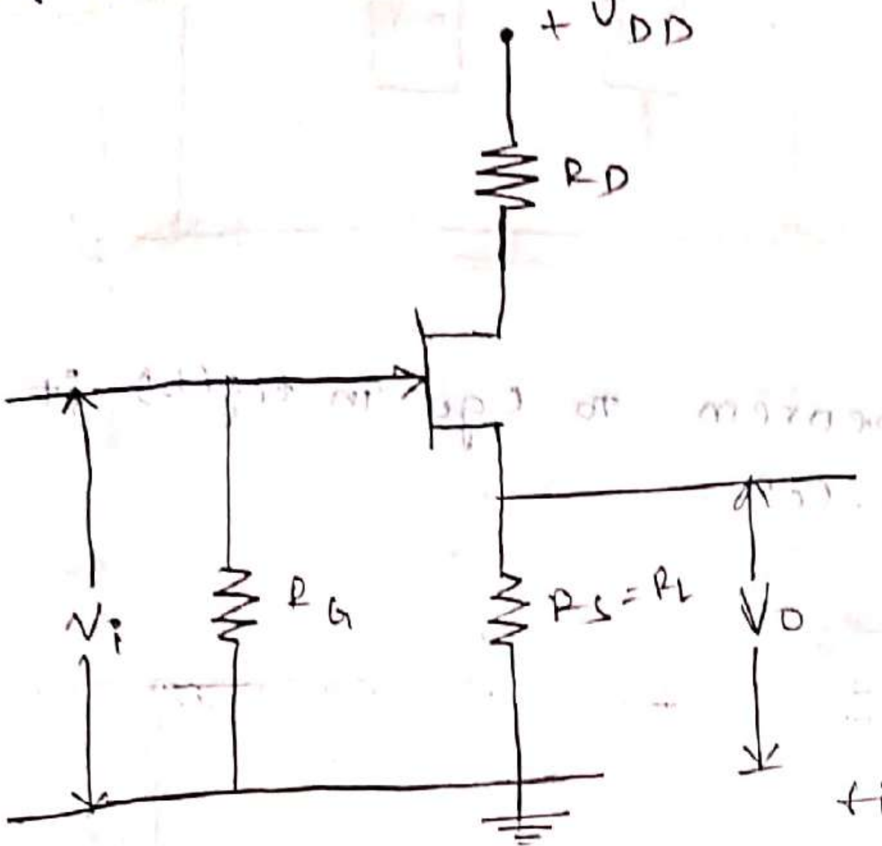


fig-(a)

Small signal equivalent circuit of CD amplifier [source follower] at high frequencies by using hybrid-π model of FET is shown in fig-(b).

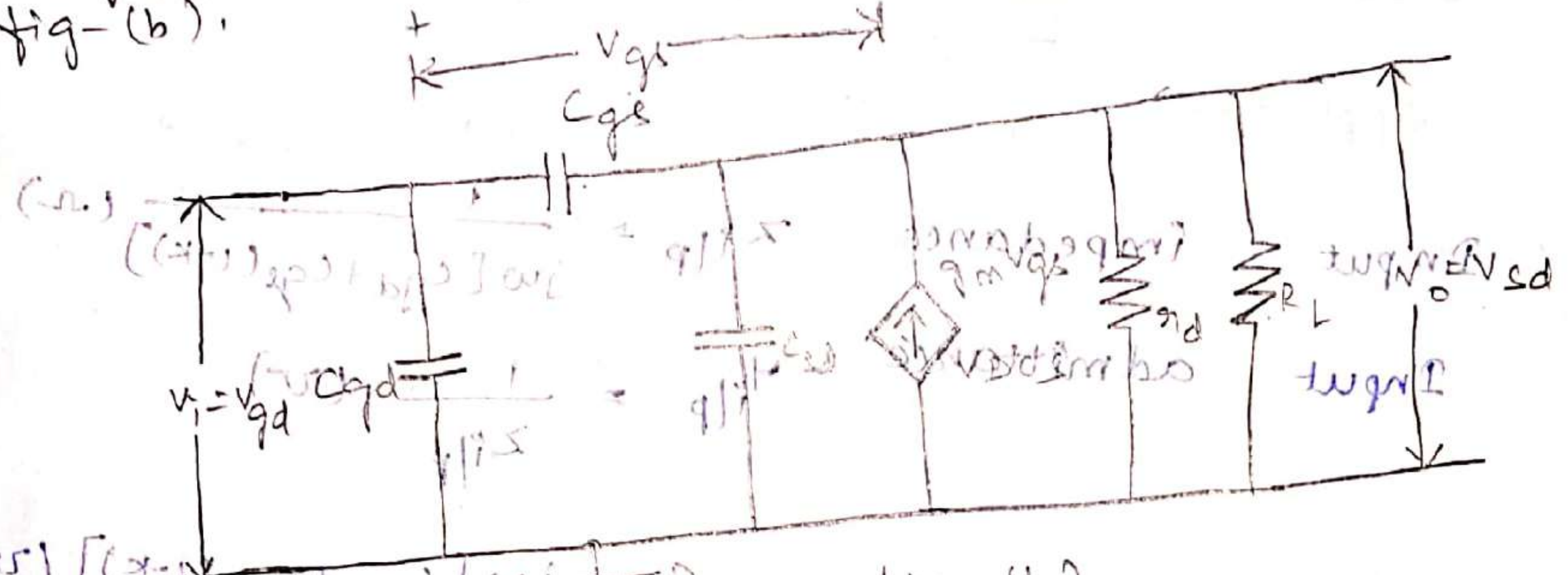


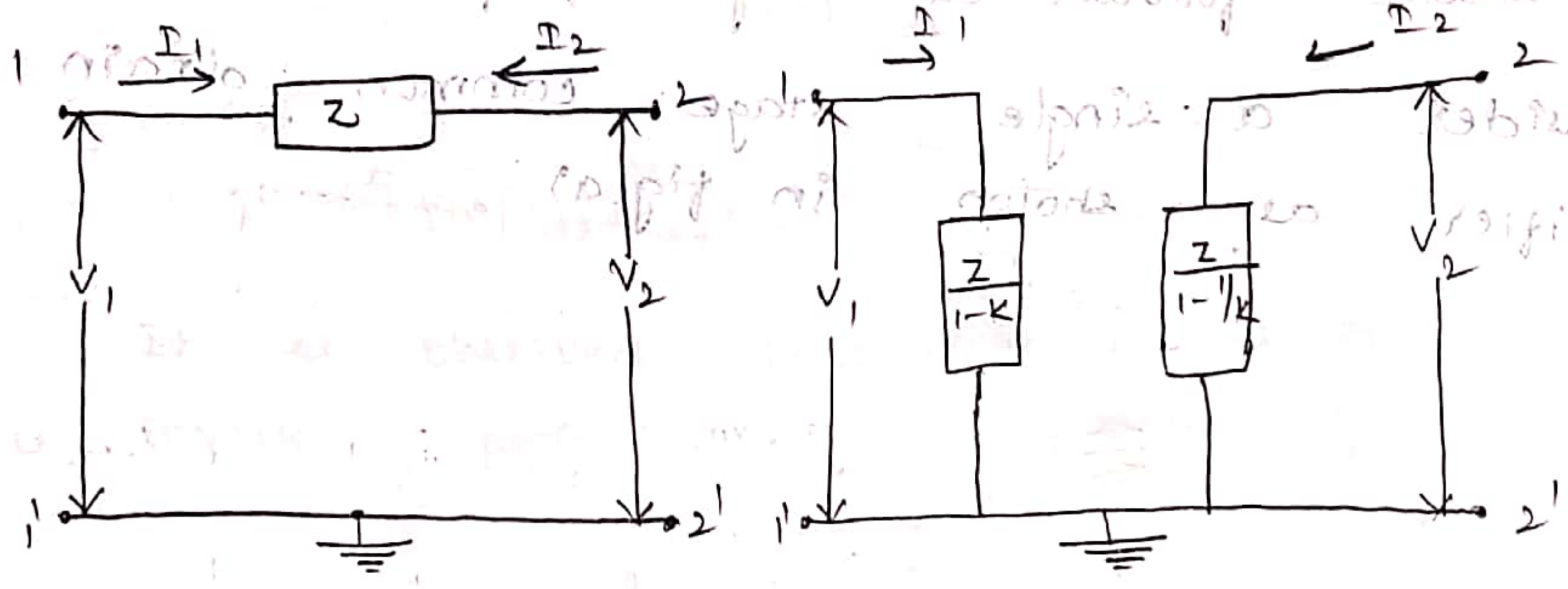
fig-(b)

~~Miller's theorem~~ voltage gain $A_V = \frac{V_o}{V_i} = \frac{V_{sd}}{V_{gd}}$

By definition,

Miller's theorem states that if there is impedance (Z) b/n two nodes i.e., i/p node and o/p node can be replaced by two impedances $\frac{Z}{1-K}$, $\frac{Z}{1-1/K}$ b/n input node and ground, output node and ground respectively

where $K = A_V = \frac{V_2}{V_1}$



Apply miller's theorem to C_{gs} in fig(b) it will gives us fig-(c)

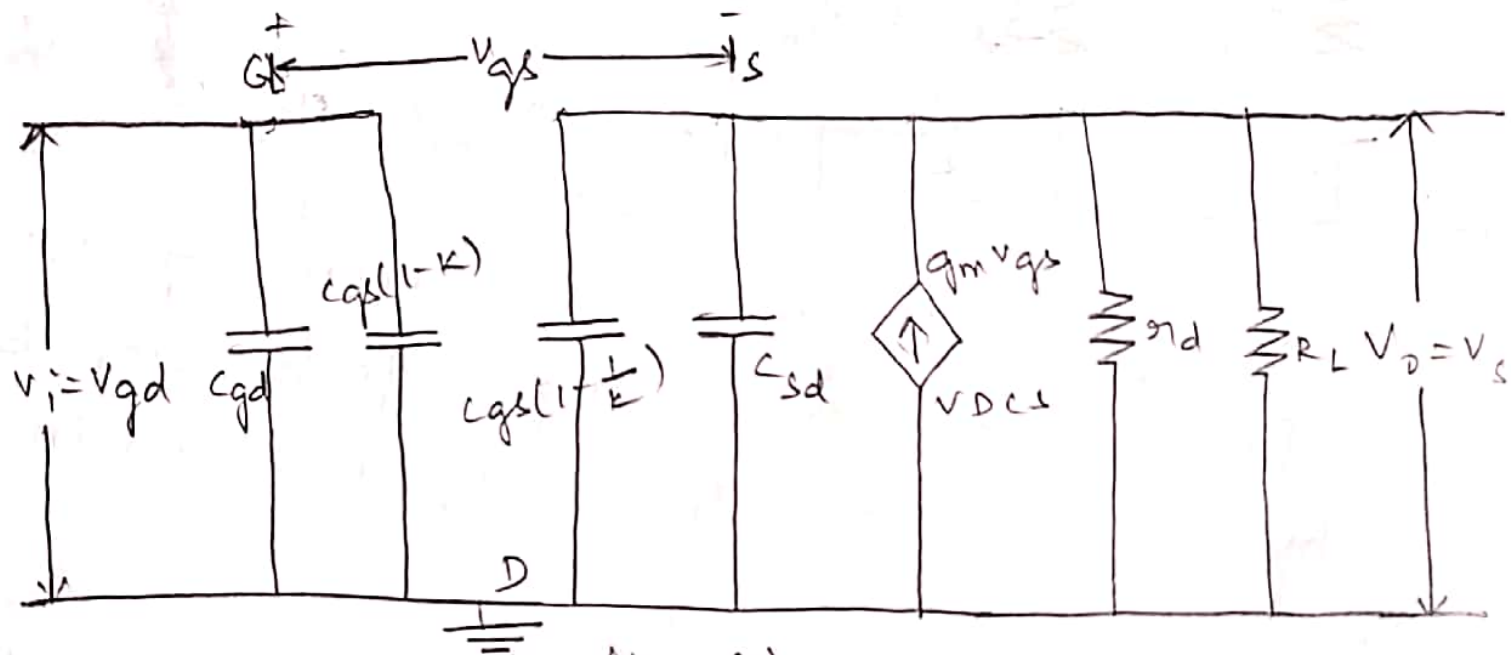


fig-(c)

Application of miller's theorem to fig-(b)

Input impedance $Z_{i/p} = \frac{1}{j\omega[C_{gd} + C_{gs}(1-K)]}$ (Ω)

Input admittance $Y_{i/p} = \frac{1}{Z_{i/p}}$ $(\mho)$

$= j\omega[C_{gd} + C_{gs}(1-K)]$ $(\mho)$

Output impedance $Z_{o/p} = \frac{1}{j\omega[C_{sd} + C_{gs}(1-\frac{1}{K})]} \| (r_d \| R_L)$

$= \frac{1}{j\omega[C_{sd} + C_{gs}(1-\frac{1}{K})]} \| \frac{r_d R_L}{r_d + R_L}$ (Ω)

Output admittance $Y_{o/p} = \frac{1}{Z_{o/p}}$ $(\mho)$

At low frequencies, all the three capacitors are having high reactances, and hence they can be disappeared in the approximate equivalent circuit [i.e., $X_C = \frac{1}{j\omega C} = \frac{1}{j2\pi f C}$ → capacitors are treated as open circuit]

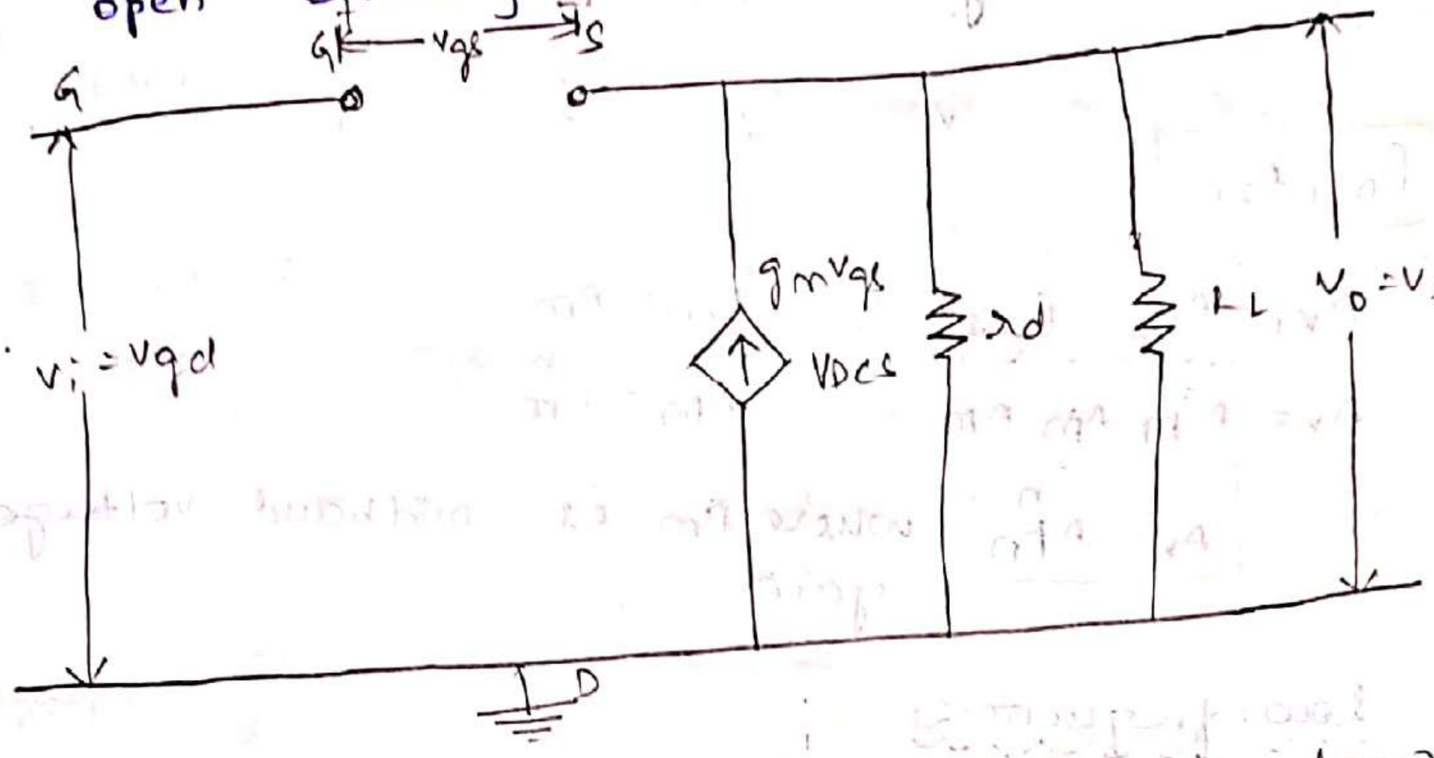


fig (d)

Approximate equivalent circuit at low frequencies

$$V_o = (g_m v_{gs})(r_d \| R_L)$$

$$= g_m (V_i - V_o)\left(\frac{r_d R_L}{r_d + R_L}\right)$$

$$= g_m (V_i - V_o) r_d \quad [\because r_d << R_L]$$

$$V_o = \mu (V_i - V_o)$$

$$V_o + \mu V_o = \mu V_i$$

$$\frac{V_o}{V_i} = \frac{\mu}{1+\mu} \simeq 1$$

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ELECTRONIC CIRCUIT ANALYSIS

UNIT [PART-A]

FEEDBACK AMPLIFIERS

1

INTRODUCTION:

* Any System whether it is electrical, mechanical, hydraulic or pneumatic may be considered to have atleast one input and one output. If the System is to perform smoothly, we must be able to measure or Control output.
* For example if the input is 10mV, gain of the amplifier is 100, output will be 1V. If the input deviates to 9mV & 11mV, output will be 0.9V or 1.1V. So there is no control over the output.
* But by introducing feedback between the output and input, there can be control over the output. If the input is increased, it can be made easy to increase by having a link between the output and input. So that Input can be made to depend on output.
* Some examples for are 1. Temperature of a Furnace
  2. Traffic light
  3. Our human eyes and mind.
* An amplifier is a device that amplifies the input signal, when we talk about ideal amplifier, there exist some parameters like Voltage gain, Input impedance, output impedance and Bandwidth

CIRCUIT DIAGRAM:

* This circuit is a Two-port Network and it represents an Amplifier.

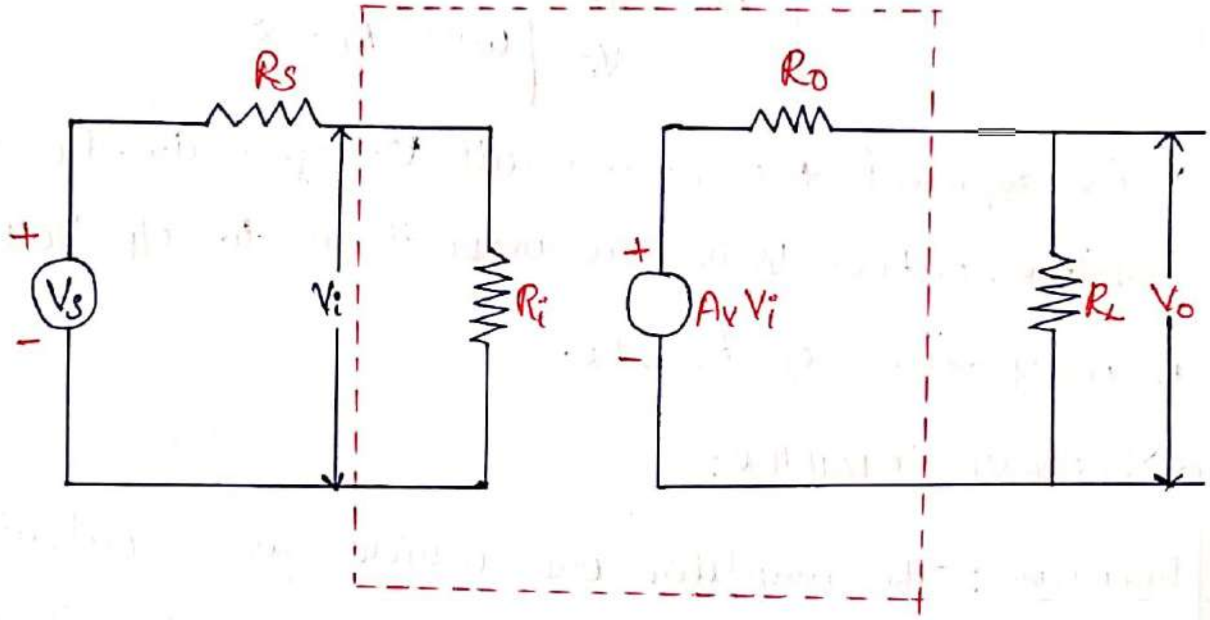


FIG: EQUIVALENT CIRCUIT OF VOLTAGE AMPLIFIERS

* The Voltage amplifier can be designed with the help of thevenins equivalent circuit on both sides.

* In the above fig, the amplifier i/p resistance R_i is large when compared to Source resistance (R_s) i.e., $R_i >> R_s$, So that drop across R_s is very small.

$$R_i >> R_s, \quad V_i \simeq V_s$$

* Similarly load resistance R_L is large compared to the o/p resistance R_o of the amplifier

If $R_L >> R_o$, $V_o = A_V V_i$

$$\boxed{V_o = A_V V_s} \quad \because V_i = V_s$$

∴ The output Voltage is proportional to the Input Voltage.

$\therefore$ For Ideal Voltage Amplifier $\boxed{R_i = \infty,\ R_O = 0}$

$$V_o = A_v V_i$$

$$\boxed{A_v = \frac{V_o}{V_i}} \text{ with } R_L = \infty$$

* A_v represent the open circuit voltage gain. For ideal voltage amplifier, output voltage is proportional to i/p voltage and is independent of R_S & R_L.

2> CURRENT AMPLIFIER:

Definition: The amplifier one which gives output current proportional to input current and the proportionality factor is independent of R_S & R_L then it is called as "Current Amplifier".

* Current amplifier can be designed with the help of Nortons equivalent circuit on both sides.

CIRCUIT DIAGRAM:

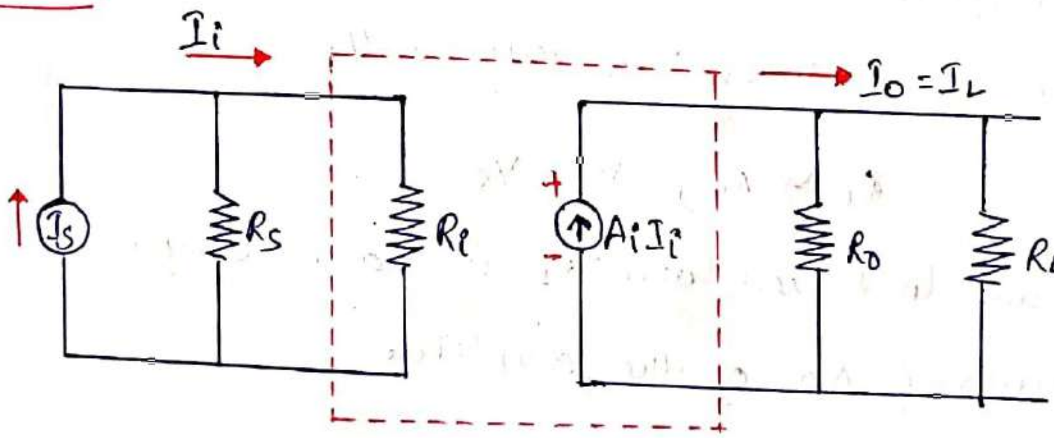


FIG: NORTON'S EQUIVALENT CIRCUIT OF CURRENT AMPLIFIER

* In the above figure, the amplifier input resistance R_i is less when compare to R_s. So we get $\boxed{I_i = I_s}$ $\because R_i << R_s$

* Current amplifier intend to amplify input current signal and provide output current signal. 5

* If $R_S >> R_i$, then $\boxed{I_i = I_S}$.

* Similarly if $R_O >> R_L$, then $I_O = A_I I_i$

$$I_O = A_I I_S \quad \because I_i = I_S$$

* For Ideal current amplifier, $R_i = 0$, $R_O = \infty$

If $R_i = 0$, $\quad I_S \simeq I_i$

$R_O = \infty \quad I_L = I_O = A_I I_i = A_I I_S$

$$\boxed{A_I = \frac{I_L}{I_i}}$$

* A_I represents the short circuit current amplification.

3) TRANSCONDUCTANCE AMPLIFIER:

DEFINITION: The Amplifier which supplies output current which is proportional to input voltage independently of the magnitude of R_S & R_L then such amplifier is called as "Transconductance Amplifier".

* In Transconductance amplifier, the input signal is a voltage signal and o/p signal is a current signal.

* Transconductance amplifier can be designed with the help of thevenin's theorem at i/p side and Nortons theorem at o/p side.

CIRCUIT DIAGRAM:

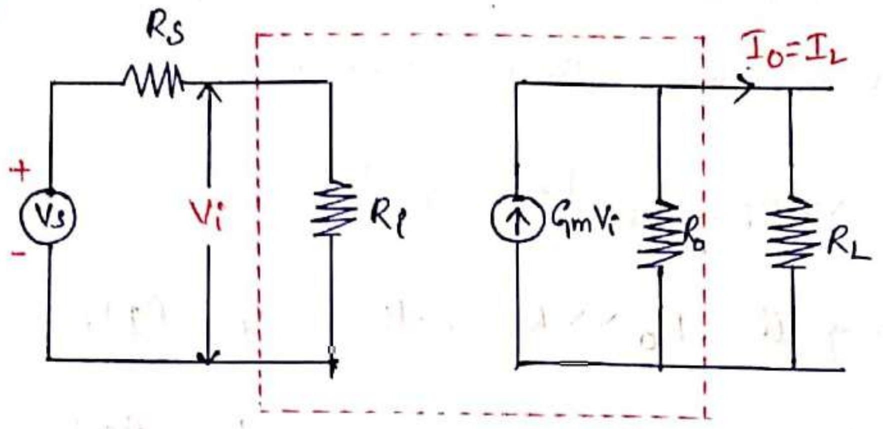


FIG: EQUIVALENT CIRCUIT FOR TRANS CONDUCTANCE AMPLIFIER.

* In the above fig, the amplifier i/p resistance R_i is high compare to R_s, So we get $\boxed{V_i = V_s}$

* Similarly R_o is high compare to R_L, So we get,

$$I_o = g_m V_i$$

$$\because V_i = V_s \quad \boxed{I_o = g_m V_s}$$

* Therefore, the o/p current is directly proportional to i/p Voltage and transconductance factor (g_m) doesnot depends on R_L & R_s.

* For ideal Trans conductance amplifier $\boxed{R_i = R_o = \infty}$

4> TRANS RESISTANCE AMPLIFIER:

Definition: The amplifier which supplies o/p Voltage, which is proportional to i/p current and independent of R_s & R_L then such amplifier is called as "Trans resistance amplifier"

* In Trans resistance amplifier the i/p signal is a current signal and o/p signal is a Voltage signal.

* Trans resistance amplifier can be designed with the help of Norton's theorem at i/p side and thevenin's theorem at o/p side.

CIRCUIT DIAGRAM:

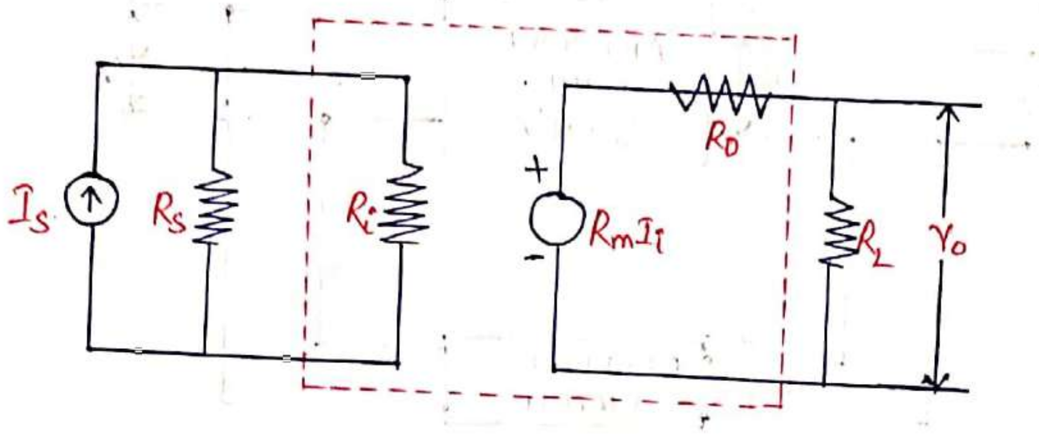


FIG: EQUIVALENT CIRCUIT FOR TRANS RESISTANCE AMPLIFIER.

* In the above figure, the amplifier i/p resistance R_i is very less ($R_i << R_s$) compare to R_s So we get $I_i = I_s$.

* Similarly 'R_o' is very less compare to 'R_L' ($R_o << R_L$) So, we get $V_o = R_m I_i$

$\because I_i = I_s$ $\boxed{V_o = R_m I_s}$

* Therefore, the o/p Voltage is directly proportional to i/p current and transresistance factor 'R_m' doesn't depend on R_L & R_s.

* For ideal Trans resistance $\boxed{R_o = R_i = 0}$

FEEDBACK PRINCIPLE:

Definition: The process of Combining a fraction of output energy (Voltage or Current) back to the input is known as "Feedback."

* The amplifier which provides feedback is called as Feedback amplifier.

* The feedback circuit amplifier has two parts
  1. Amplifier
  2. Feedback Network.

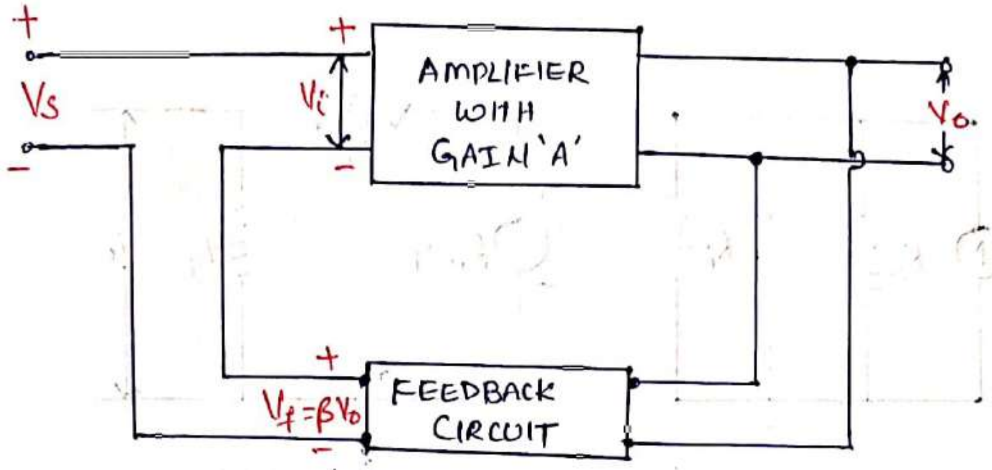


FIG: PRINCIPLE OF FEEDBACK AMPLIFIER.

* Let 'A' be the gain of the amplifier without feedback

$$\boxed{A = \frac{V_o}{V_i}}$$

* The feedback Network extracts a Voltage $V_f = \beta V_o$ from the o/p V_o of the amplifier.

$V_i = V_s + V_f = V_s + \beta V_o$ (positive feedback) ($\because V_f = \beta V_o$)

$V_i = V_s - V_f = V_s - \beta V_o$ (Negative feedback)

i.e., $V_i = V_s \pm V_f$

* the quantity $\beta = \frac{V_f}{V_o}$ is called feedback ratio & feedback fraction

$$A = \frac{V_o}{V_i}$$

$$A = \frac{V_o}{V_s \pm V_f}$$

$$V_o = A(V_s \pm V_f)$$

$$V_o = A(V_s \pm \beta V_o)$$

$$V_o = AV_s \pm \beta A V_o$$

$$V_o \mp \beta A V_o = A V_s$$

$$V_o(1 \mp \beta A) = A V_s$$

$$\frac{V_o}{V_s} = \frac{A}{1 \mp \beta A} \rightarrow ①$$

$$\boxed{A_f = \frac{A}{1 \mp \beta A}}$$

where A_f is the overall gain (gain with feedback)

A is the openloopgain (gain without feedback)

β is the feedback ratio or feedback factor.

⇒ The ratio of output voltage V_o to the applied signal voltage V_s is called overall gain i.e, gain with feedback A_f.

$$A_f = \frac{\text{Output Voltage}}{\text{Input Signal Voltage}} = \frac{V_o}{V_s} \rightarrow ②$$

from eq ① & ② $A_f = \dfrac{A}{1 \mp \beta A}$

i.e., $A_f = \dfrac{A}{1 + A\beta}$ for Negative feedback

$A_f = \dfrac{A}{1 - A\beta}$ for positive feedback.

* Feedback means the o/p signal is coupled to the Input of the same circuit. This feedback signal provides the control element of the system. (10)

Eg: 1. Temperature of device

2. Human mind & eyes.

TYPES OF FEEDBACK:

* There are two types of Feedback.

1. Positive feedback
2. Negative Feedback.

POSITIVE FEEDBACK:

* If the feedback signal V_f is inphase with the input signal V_s, then the net $V_i = V_s + V_f$. Hence, the input Voltage applied to the basic amplifier is increased, thereby increasing the V_o exponentially. This type of feedback is said to be positive & Regenerative feedback.

* Gain of the amplifier with positive feedback is,

$$\boxed{A_f = \frac{A}{1 - A\beta}}$$

* Since positive feedback causes excessive distortion and instability, it is rarely used in Amplifier circuits.

* However, because of its capability of increasing the power of increasing the original signal it is used in Oscillator circuits.

NEGATIVE FEEDBACK:

* If the feedback signal V_f is out of phase with the input signal V_s then such feedback is known as Negative feedback or Degenerative feedback.
* Then, $V_i = V_s - V_f$, so the input voltage applied to the basic amplifier is decreased and correspondingly the o/p is decreased.
* Hence the voltage gain is reduced. Gain of the amplifier with Negative feedback is,

$$\boxed{A_f = \frac{A}{1+A\beta}}$$

* The Negative feedback has various advantages like,

-> Gain Stability
-> Reduction in Non-linear distortion
-> Reduction in Noise.
-> Increase in Bandwidth or Improvement in frequency response.
-> Increase in Input Impedance.
-> Decrease in output Impedance.

* Because of its numerous advantages it is widely used in amplifier circuits.
* The drawback in Negative feedback is it reduces overall gain of the amplifier, this problem can be compensated by increasing the number of stages in amplifier circuits.

CONCEPT OF FEEDBACK:

(12)

* The below figure represents the block diagram of an amplifier with feedback.

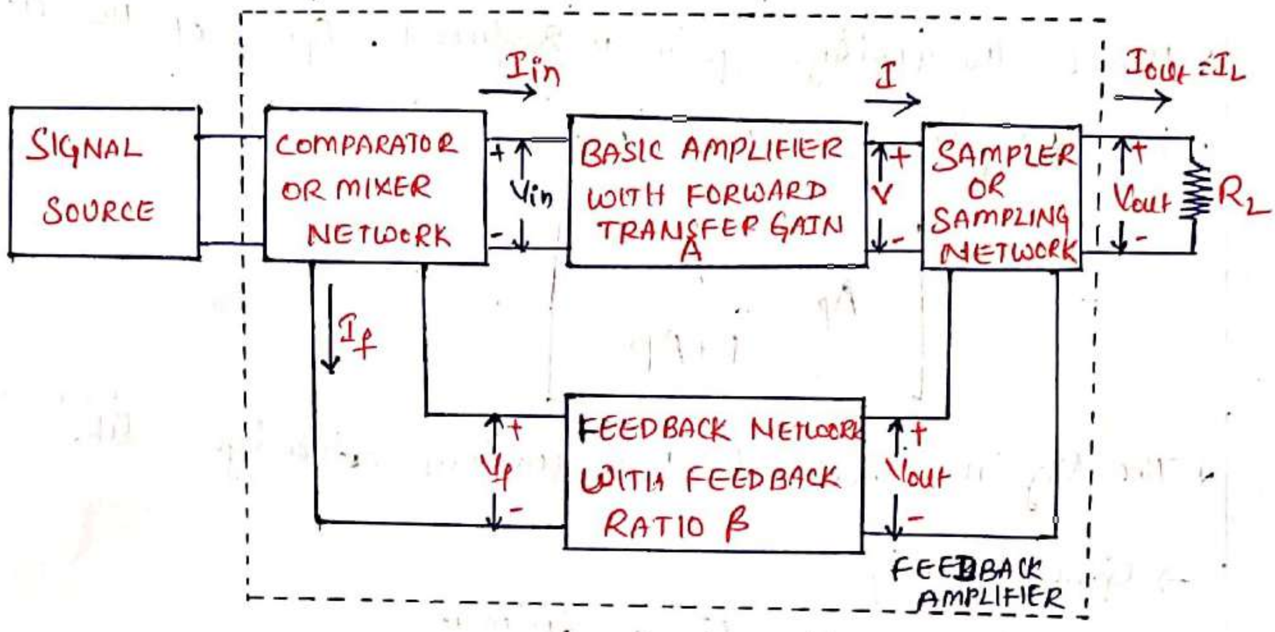


FIG: BLOCK DIAGRAM OF AN AMPLIFIER WITH FEEDBACK

* The output quantity (either Voltage or Current) is sampled by means of a suitable sampling Network (or Sampler) and is fed to the feedback Network.

* The output of feedback network, which has a fraction of o/p signal is combined with the external (source) signal through a mixer network and fed into the basic amplifier.

* The different blocks of feedback amplifier are explained below.

1. SIGNAL SOURCE: This block is either a signal Voltage V_s in series with a resistor R_s or a Signal current I_s in parallel with a Resistor R_s.

TRANSFER RATIO (OR) GAIN:

* The symbol A in the Basic amplifier represents the ratio of the output signal to the input signal.

* The transfer ratio $A_V = \frac{V_o}{V_i}$ referred to as Voltage gain

The transfer ratio $A_I = \frac{I_o}{I_i}$ referred to as current gain

The transfer ratio $G_m = \frac{I_o}{V_i}$ referred to as trans conductance.

The transfer ratio $R_m = \frac{V_o}{I_i}$ referred to as trans resistance.

* These are the transfer gains of the basic amplifier without feedback and is represented by symbol A.

* The symbol A_f is defined as the ratio of the o/p signal to the input signal of the amplifier configuration.

* A_f is the transfer gain of the amplifier with feedback.

$$A_{Vf} = \frac{V_o}{V_s},\quad A_{If} = \frac{I_o}{I_s} \quad R_{mf} = \frac{V_o}{I_s},\quad G_{mf} = \frac{I_o}{V_s}.$$

GENERAL STRUCTURE OF SINGLE LOOP FEEDBACK AMPLIFIER:

* The below figure represents the signal-flow diagram of a feedback amplifier in which quantity "X" represents either Voltage or Current signals.

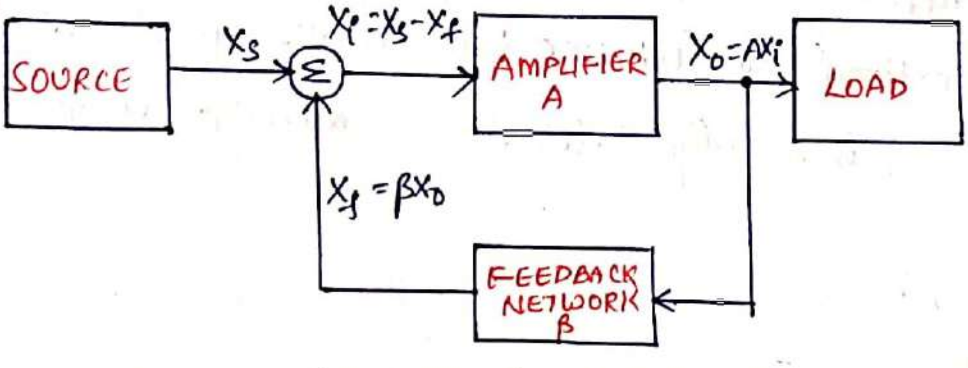


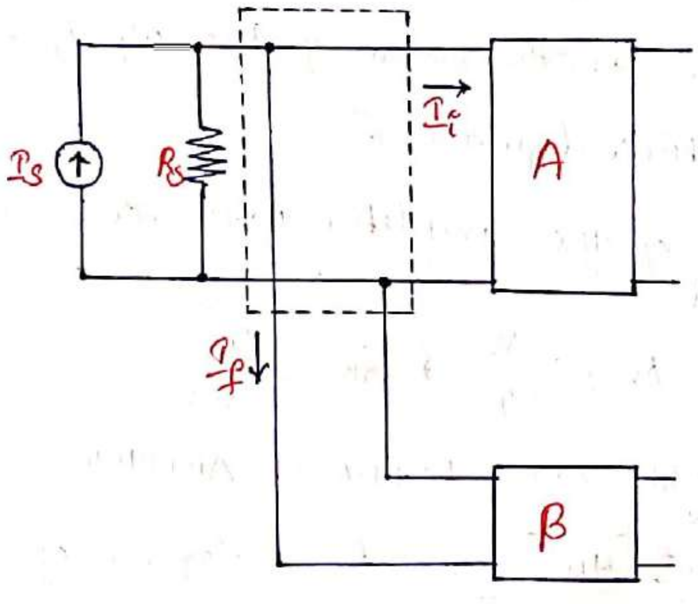
FIG: GENERAL STRUCTURE OF SINGLE LOOP FEEDBACK AMPLIFIER

* This is because in this case, feedback signal is returned to the input in series to oppose the applied voltage causing input (14) current to fall and consequently makes the input impedance to increase.

2. SHUNT MIXER:

* The connecting of feedback signal in parallel with an input current source is known as "Shunt mixer."

* In case of shunt or parallel connections, the current drawn from the signal source is increased by an amount equal to feedback current I_f and therefore, input impedance falls.



-> Voltage f/b tends to reduce the o/p impedance

-> Current f/b tends to increase the o/p impedance

FIG: SHUNT MIXER

MIXER: A differential Amplifier, which has two inputs and one output proportional to difference between the signals at the two inputs, is usually referred as a mixer or comparator.

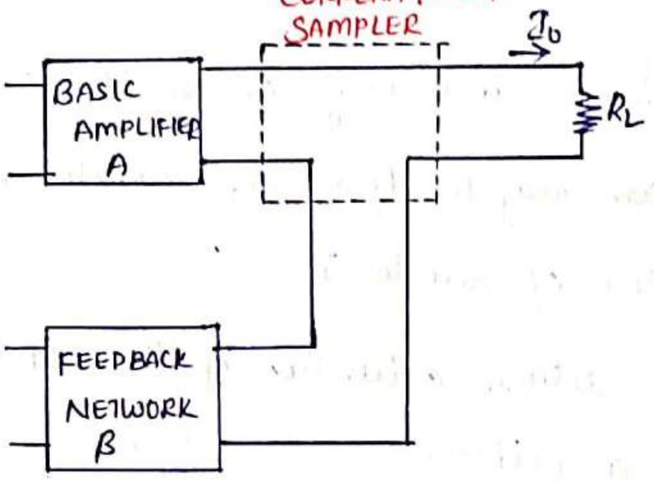


FIG: CURRENT (OR) LOOP SAMPLING

4. MIXER NETWORK:

* Like sampling, there are two ways of mixing the feedback signal. Mixer is also known as comparator
* Mixer is of two types
  1. Series mixer
  2. Shunt mixer

SERIES MIXER:

* The connection of feedback signal in series with the input signal voltage is known as "Series mixer".
* Series feedback connection tend to increase the input impedance of the amplifier

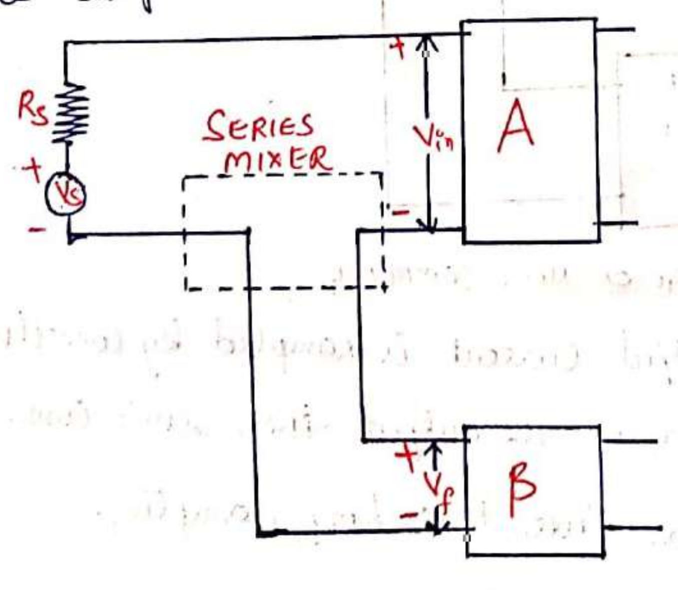


FIG: SERIES MIXING

2. FEEDBACK NETWORK:

* The feedback network is usually in the form of a passive two-port network and may be formed of : resistors, inductors and capacitors (most often of resistors).

* Its function is to return a function of the o/p energy (Voltage or current) to the input of the amplifier.

3. SAMPLING NETWORK:

* Sampling Networks are of two types
    1. Voltage Sampling
    2. Current Sampling

1. VOLTAGE SAMPLING: The output Voltage is sampled by connecting the feedback Network in shunt across the output then such connection at the output is refered to as Voltage or node sampling

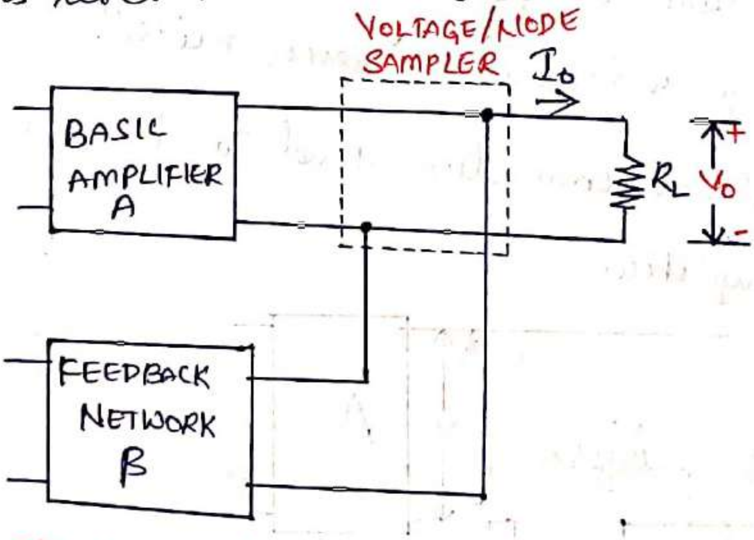


FIG: VOLTAGE (OR) NODE SAMPLING

2. CURRENT SAMPLING: The output current is sampled by connecting the feed back Network in series across the output then such connection at the output is refered to as Current or loop sampling.

* When the feedback signal X_f and the input signal X_i are out of phase, then the feedback is called negative feedback. (17)

From the above figure, we have

$$X_i = X_S - X_f = X_d$$

* Where X_d represents the difference between the applied input signal X_S and feed back signal X_f and it is called the error signal & comparison signal.

* The feedback factor $\beta = \frac{X_f}{X_o}$

where X_o is the output Voltage/current

* The transfer gain without feedback A is defined by

$$A = \frac{X_o}{X_i}$$

* The transfer gain with feedback A_f is defined by,

$$A_f = \frac{X_o}{X_S} = \frac{X_o}{X_i + X_f}$$ $(\because X_i = X_S - X_f$
$X_S = X_i + X_f)$

$$= \frac{X_o}{X_i + \beta X_o}$$ $(\because \beta = \frac{X_f}{X_o}$
$X_f = \beta X_o)$

$$= \frac{X_o}{X_i\left(1+\beta \cdot \frac{X_o}{X_i}\right)}$$

$$= \frac{X_o}{X_i} * \frac{1}{1+\beta \cdot X_o/X_i}$$

$$A_f = A \cdot \frac{1}{1+\beta A} \qquad \left(\because A = \frac{X_o}{X_i}\right)$$

$$\boxed{A_f = \frac{A}{1+\beta A}}$$

CLASSIFICATION OF FEEDBACK AMPLIFIERS :

* Based on the type of Sampling at the output Side and the type of mixing to the input side, feedback amplifiers are classified into four topologies.

1. Voltage-Series feedback ~~or series shunt feedback~~
2. Current-Series feedback ~~or series series feedback~~
3. Current-Shunt feedback ~~or shunt series feedback~~
4. Voltage-Shunt feedback

* If the feedback signal is connected in series with the i/p Signal then it is called as Series feedback amplifier.
* If the feedback signal is connected in shunt with the i/p Signal then it is called as Shunt feedback amplifier.

1. VOLTAGE SERIES FEEDBACK AMPLIFIER:

* If the ~~input~~ Feedback o/p signal is connected in series with the i/p Signal then it is called as Voltage Series feedback amplifier.

* If Voltage is sampled and mixing is in series then the type of feedback is known as Voltage Series feedback.

* The block diagram of Voltage series feedback amplifier is as shown below. (19)

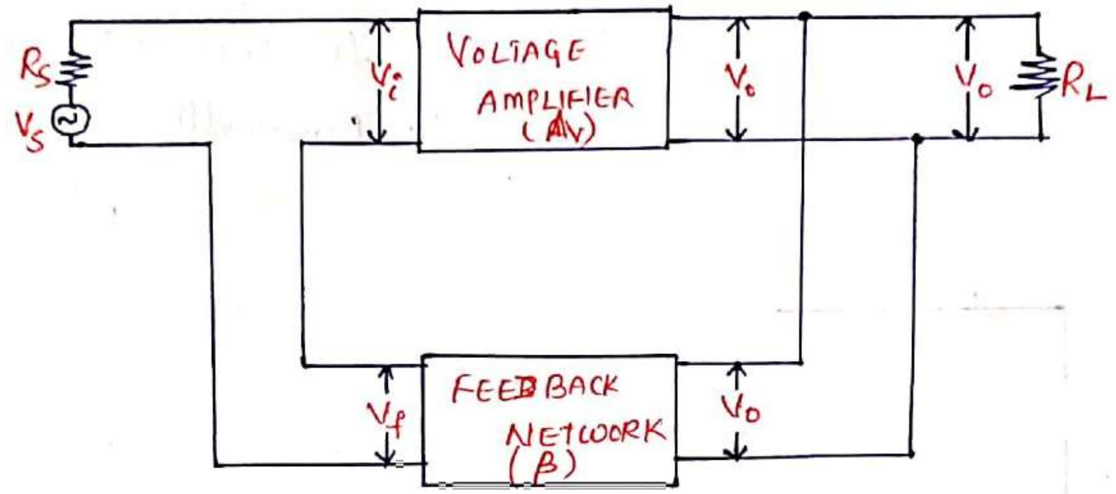


FIG: BLOCK DIAGRAM OF VOLTAGE SERIES FEEDBACK AMPLIFIER.

* Here Basic amplifier is a Voltage Amplifier

* At Input side Voltages are mixed by means of Series feedback

$V_S \rightarrow$ Source Signal

$V_i \rightarrow$ Input Signal

$V_f \rightarrow$ feedback Signal

From the figure $V_i = V_S - V_f$

$$V_S = V_i + V_f$$

* At output side Voltage is sampled by using Shunt Sampling

$V_o \rightarrow$ output Signal.

* Voltage Series feedback amplifier is also known as Shunt-Series feedback amplifier

* The circuit diagram for Voltage-Series feedback amplifier to derive amplifier parameter such as, 1. Input Impedance (R_i)

2. Output Impedance (R_o)
3. Voltage gain with feedback (A_{vf})
4. Bandwidth

*

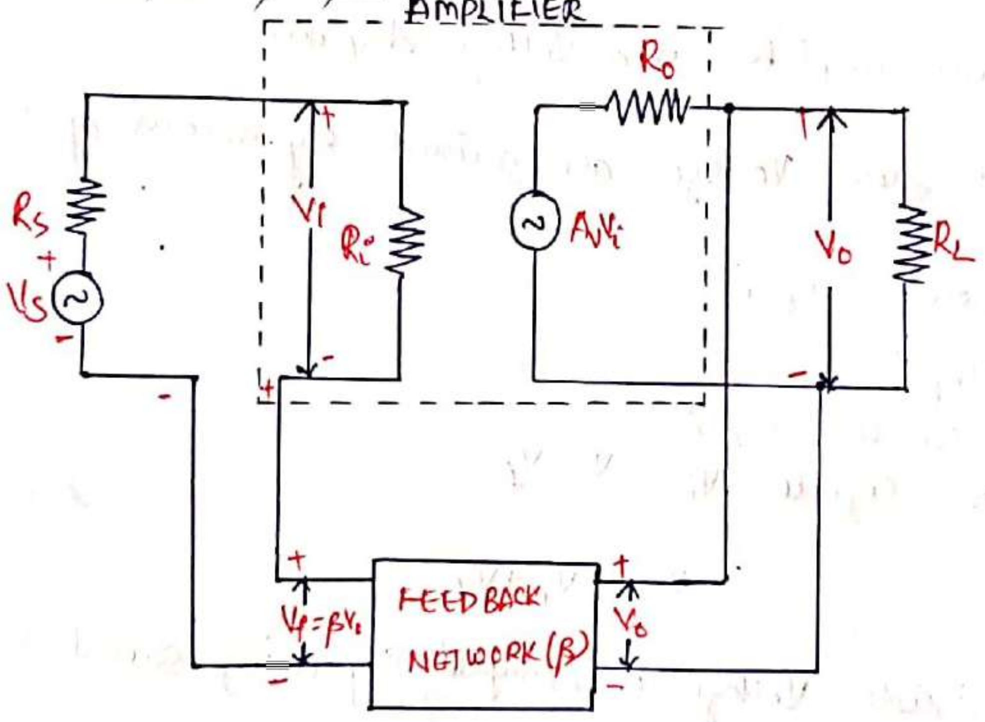


FIG: CIRCUIT DIAGRAM FOR VOLTAGE SERIES FEEDBACK AMPLIFIER.

=) By definition, open loop Voltage gain(β) gain without feedback

is $A_v = \frac{V_o}{V_i} \rightarrow (1)$

Feedback factor $\beta = \frac{V_f}{V_o} \rightarrow (2)$

* Closed loop Voltage gain (or) Voltage gain with feedback is,

(21)

$$A_{vf} = \frac{V_o}{V_s} \rightarrow (3)$$

VOLTAGE GAIN WITH FEEDBACK :

* The ratio of output voltage to source voltage is defined as voltage gain with feedback.

$$A_{vf} = \frac{V_o}{V_s} \rightarrow (1)$$

We know that $V_s = V_i + V_f \rightarrow (2)$

Substitute eq (2) in eq (1)

$$A_{vf} = \frac{V_o}{V_i + V_f}$$

$$= \frac{V_o}{V_i + \beta V_o} \quad (\because V_f = \beta V_o)$$

$$= \frac{V_o}{V_i\left[1 + \beta \cdot \frac{V_o}{V_i}\right]}$$

$$= \frac{V_o / V_i}{1 + \beta \cdot \frac{V_o}{V_i}}$$

$$= \frac{A_v}{1 + A_v \beta} \quad \left(\because A_v = \frac{V_o}{V_i}\right)$$

$$\boxed{\therefore \text{Voltage gain with feedback } A_{vf} = \frac{A_v}{1 + A_v\beta} \rightarrow A_{vf} << A_v.}$$

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BANDWIDTH WITH FEEDBACK:

* If Voltage gain with feedback decreases, it Increases the bandwidth with feedback.
* This is due to the product of gain x Bandwidth always be a constant

$$\boxed{\therefore \text{Bandwidth with feedback } (B.W)_f = BW(1+A_V\beta) \quad \Rightarrow (B.W)_f > B.W}$$

INPUT RESISTANCE WITH FEEDBACK:

Definition: The ratio of Input Voltage to Input Current is defined as Input resistance without feedback

* Input resistance without feedback $R_i = \dfrac{V_i}{I_i}$
* Input resistance with feedback is defined as the ratio of Source Voltage to the Input current.

$$R_{if} = \frac{V_s}{I_i}$$

We know that $V_s = V_i + V_f$

$$R_{if} = \frac{V_i + V_f}{I_i}$$

$$= \frac{V_i + \beta V_o}{I_i}$$

$$= \frac{V_i + \beta(A_V V_i)}{I_i} \qquad \left(\because A_V = \frac{V_o}{V_i}\right.$$

$$V_o = A_V V_i$$

$$= \frac{V_i[1+\beta A_v]}{I_i} = \frac{V_i}{I_i}[1+A_v\beta]$$

$$= R_i[1+A_v\beta) \quad (\because R_i = \frac{V_i}{I_i})$$

$\therefore$ Input Resistance with feedback $R_{if} = R_i(1+A_v\beta) \Rightarrow R_{if} > R_i$

* High Input Impedance is always desirable in an amplifier. ~~as~~ Such a desirable characteristic can be achieved with the help of Negative feedback.

OUTPUT RESISTANCE WITH FEEDBACK:

* Just as High Input Impedance is advantageous to an amplifier, is desirable. Similarly low Output Impedance is desirable.
* With lower output Impedance, the amplifier is better suited to drive a low impedance load.
* The Input terminals are short-circuited i.e., $V_s = 0$

$$\cancel{V_s} = V_i + V_f$$

$$0 = V_i + V_f$$

$$V_i = -V_f.$$

* When $V_s = 0$, $-V_f$ is the only input Voltage to the amplifier

$$R_{of} = \frac{V_o}{I_o}\Big|_{V_s=0}$$

$$V_i = V_s - V_f = 0 - \beta V_o$$

Apply KVL to the output side,

we have $V_o = I_o R_o + A_v V_i$

$$V_o = I_o R_o + A_v(-\beta V_o)$$

$$V_o + A_v \beta V_o = I_o R_o$$

$$V_o(1 + A_v\beta) = I_o R_o$$

$$\frac{V_o}{I_o} = \frac{R_o}{1 + A_v\beta}$$

$$\boxed{\therefore R_{of} = \frac{R_o}{1 + A_v\beta}} \quad (R_{of} << R_o)$$

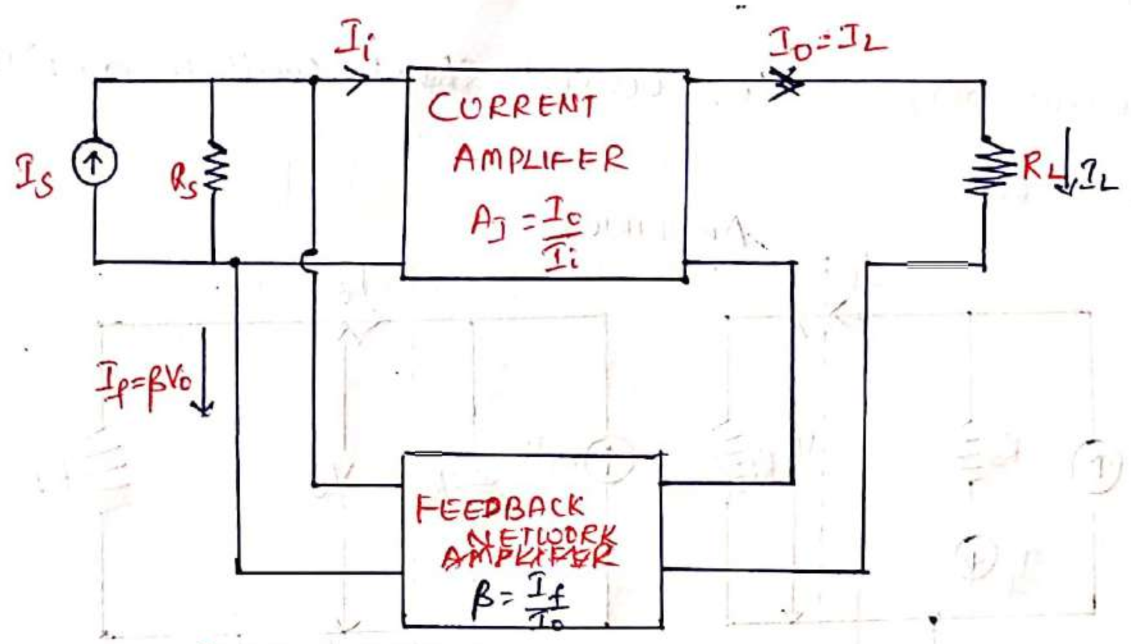
* when the A_v increases, output impedance decreases and vice versa.

For Voltage Series Feedback amplifier

1. Gain with feedback $A_{vf} = \frac{A_v}{1 + A_v\beta}$ (Decreases)
2. Bandwidth with feedback $(B.W)_f = B.W(1 + A_v\beta)$ (Increases)
3. Input resistance with feedback $R_{if} = R_i(1 + A_v\beta)$ (increases)
4. output resistance with feedback $R_{of} = \frac{R_o}{1 + A_v\beta}$ (decreases.

2. CURRENT SHUNT FEEDBACK AMPLIFIER:

* The Block diagram for Current-shunt feedback amplifier, is as shown below.



BLOCK DIAGRAM OF
FIG: CURRENT SHUNT FEEDBACK AMPLIFIER.

* In Current-shunt feedback amplifier, the feedback signal is in parallel with the Input Signal.
* This is also known as Shunt-series Feedback Amplifier
* The Current-shunt feedback amplifier works as a true current amplifier as the input signal is a current and the output signal is a current.

Basic Amplifier : Current amplifier

-> At input side currents are mixed by means of Shunt feedback.

-> At output side currents are sampled by means of ~~series~~ Sampling Current

$I_s \rightarrow$ Source Signal

$I_i \rightarrow$ Input Signal

$I_f \rightarrow$ Feedback Signal.

* The Circuit diagram for current-Shunt feedback amplifier is as shown below.

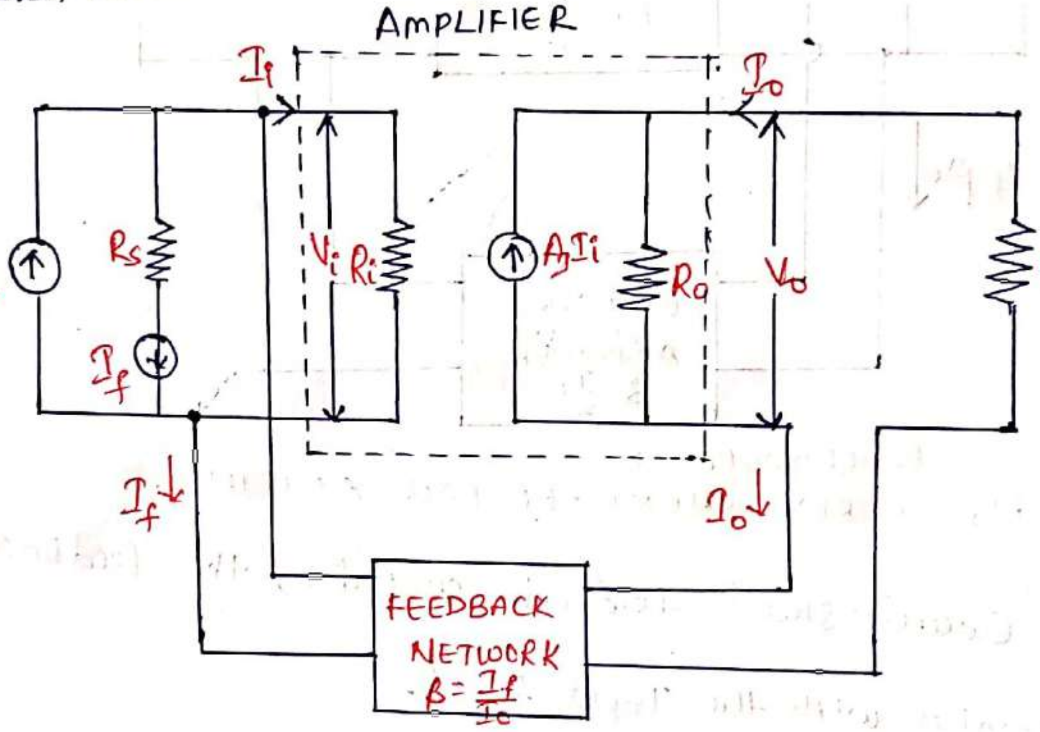


FIG: CIRCUIT DIAGRAM FOR CURRENT-SHUNT FEEDBACK AMPLIFIER.

* By definition

→ open loop current gain (or) current gain without feedback.

$$A_I = \frac{I_o}{I_i}$$

Feedback factor $\beta = \frac{I_f}{I_o}$

→ Closed loop current gain (or) current gain with feedback

$$A_{If} = \frac{I_o}{I_s}.$$

CURRENT GAIN WITH FEEDBACK: (A_{If})

Definition: The ratio of output current I_0 to the Source Current I_S is called as current gain with feedback.

$$A_{If} = \frac{I_0}{I_S}$$

We know that $I_i = I_S - I_f$

$$I_S = I_i + I_f$$

then $A_{If} = \dfrac{I_0}{I_i + I_f}$

$$= \frac{I_0}{I_i + \beta I_0} \qquad (\because I_f = \beta I_0)$$

$$A = \frac{I_0}{I_i + \beta I_0} \qquad \left(\because A_I = \frac{I_0}{I_i}\right)$$

~~$I_0 = A_I I_i$~~

$$= \frac{I_0}{I_i\left[1 + \beta \cdot \frac{I_0}{I_i}\right]}$$

$$= \frac{I_0 / I_i}{1 + \beta \frac{I_0}{I_i}} = \frac{A_I}{1 + A_I \beta}$$

$$\boxed{\therefore \text{Current gain with feedback } A_{If} = \frac{A_I}{1 + \beta A_I} \qquad A_{If} < A_I}$$

* Current gain with feedback decreases.

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BANDWIDTH WITH FEEDBACK:

* As current gain with feedback decreases, Bandwidth with feedback should increase.
* Because gain × Bandwidth product must be always constant

$$\boxed{\text{Bandwidth with feedback } (B.\omega)_f = B.\omega(1+A_I\beta) \quad B.\omega_f > B.\omega}$$

INPUT RESISTANCE WITH FEEDBACK:

* By definition Input resistance without feedback is,

$$R_i = \frac{V_i}{I_i}$$

* Input resistance with feedback is,

$$R_{if} = \frac{V_i}{I_s}$$

we know that $I_s = I_i + I_f$

then $R_{if} = \dfrac{V_i}{I_i + I_f} = \dfrac{V_i}{I_i + \beta I_o}$

$$= \frac{V_i}{I_i[1+\beta I_o]}$$

$$= \frac{V_i}{I_i[1+\beta A_I I_i} \qquad \left[\because A_I = \frac{I_o}{I_i} \quad I_o = A_I I_i\right]$$

$$= \frac{V_i}{I_i[1+\beta A_I]} = \frac{V_i/I_i}{1+\beta A_I} = \frac{R_i}{1+\beta A_I}$$

$$\boxed{\therefore \text{Input resistance with feedback } R_{if} = \frac{R_i}{1+\beta A_I} \quad R_{if} < R_i}$$

* When the negative feedback signal is feed back to the input in shunt with the applied signal, the input resistance is decreased. 29

OUTPUT RESISTANCE WITH FEEDBACK:

* In current shunt feedback amplifier, at the output side current sampling is done, it tends to increase the output resistance with feedback.

* By open circuiting the current source I_S, $-I_f$ is the only i/p to the amplifier

$$R_{of} = \left.\frac{V_o}{I_o}\right|_{\text{with } I_S = 0}$$

$$I_i = I_S - I_f = 0 - I_f = -\beta I_o$$

Apply KCL to the o/p loop.

$$I_o = \frac{V_o}{R_o} + A_I I_i$$

$$I_o = \frac{V_o}{R_o} + A_I(-\beta I_o)$$

$$I_o + A_I \beta I_o = \frac{V_o}{R_o}$$

$$I_o(1+\beta A_I) = \frac{V_o}{R_o}$$

$$R_o = \frac{V_o}{I_o(1+\beta A_I)}$$

$$\frac{V_o}{I_o} = R_o(1+\beta A_I)$$

$\therefore$ output resistance with feedback $R_{of} = R_o(1+A_I\beta) \Rightarrow R_{of} > R_o$

$\therefore$ For Current-Shunt feed back Amplifier

$\Rightarrow$ Current gain with feedback $A_{If} = \dfrac{A_I}{1+A_I\beta}$ (decreases)

$\Rightarrow$ Bandwidth with feedback $(B.\omega)_f = B.\omega(1+A_I\beta)$ [Increases]

$\Rightarrow$ Input resistance with feedback $R_{if} = \dfrac{R_i}{1+A_I\beta}$ [decreases]

$\Rightarrow$ Output resistance with feedback $R_{of} = R_o(1+A_I\beta)$ [Increases]

VOLTAGE-SHUNT FEEDBACK AMPLIFIER:

* The Block diagram for Voltage-shunt feedback amplifier is as shown below.

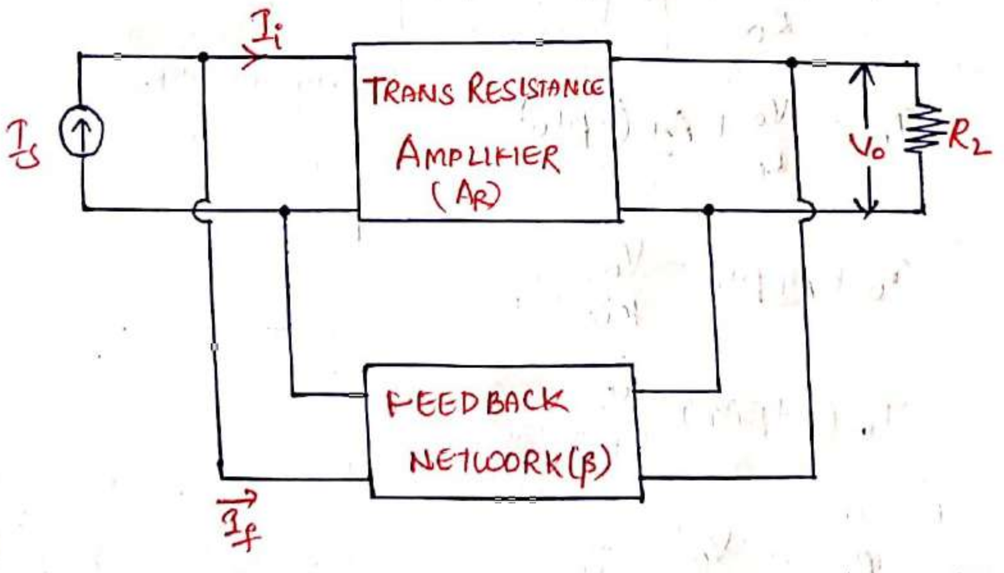

FIG: BLOCK DIAGRAM FOR VOLTAGE-SHUNT FEEDBACK

* In Voltage-shunt feedback, the feedback signal is in parallel with Input Signal. At the output side feedback signal is in parallel with output signal.

* It is also called a Shunt-derived, Shunt-fed feedback Connection.
* Here, a fraction of the output Voltage is supplied in parallel with the input voltage through the feedback network.
* The feedback signal I_f is proportional to the output Voltage V_o.

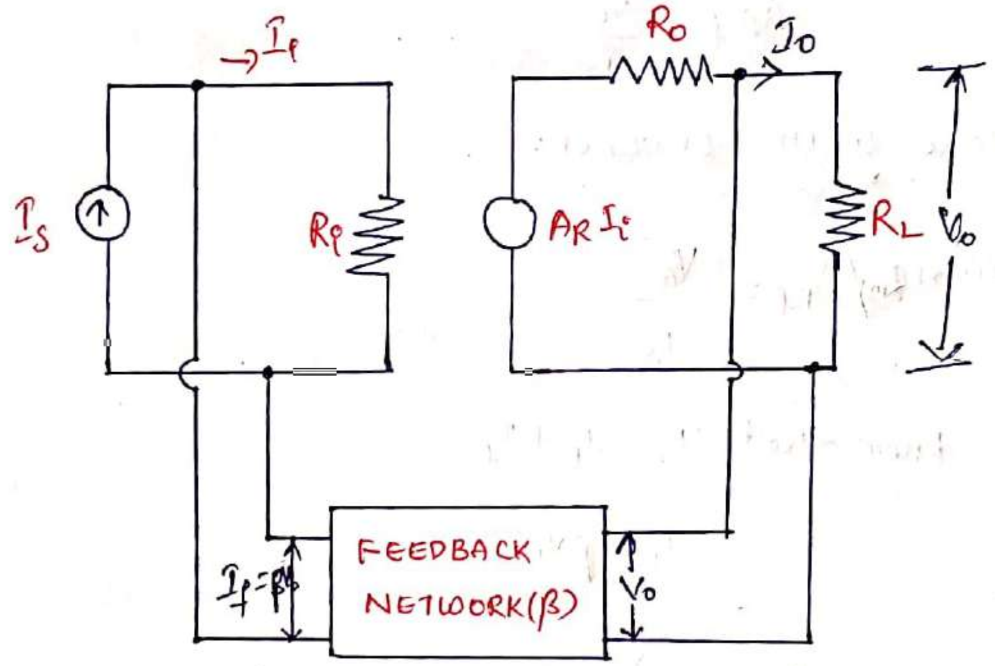

FIG: CIRCUIT DIAGRAM FOR VOLTAGE SHUNT FEEDBACK.

1. Basic Amplifier is transresistance Amplifier
* At input, Currents are mixed by means of shunt feedback.

I_s -> Source Signal.

I_i -> Input Signal

I_f -> Feedback Signal

* At output, Voltage is sampled by means of Shunt output Signal V_o.

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* open loop gain transresistance A_R or R_m without feedback (37)

$$A_R = \frac{V_o}{I_i}$$

feedback factor $\beta = \frac{I_f}{V_o}$.

* closed loop transresistance with feedback

$$A_{Rf} = \frac{V_o}{I_s}$$

TRANS RESISTANCE WITH FEEDBACK:

* By definition R_{mf}/ $A_{Rf} = \frac{V_o}{I_s}$

we know that $I_s = I_i + I_f$

$$I_f = \beta V_o$$

$$A_{Rf} = \frac{V_o}{I_i + I_f}$$

$$= \frac{V_o}{I_i + \beta V_o} \quad (\because V_o = A_R I_i)$$

$$= \frac{V_o}{I_i + \beta A_R I_i} = \frac{V_o}{I_i(1 + A_R \beta)}$$

$$\boxed{A_{Rf} \text{ or } R_{mf} = \frac{A_R}{1 + A_R \beta}} \quad A_{Rf} << R_R$$

BANDWIDTH WITH FEEDBACK:

* As transresistance with feedback decreases bandwidth with feedback should increases. The reason is gain bandwidth product should always be constant.

$$(B.W)_f = B.W(1+A_R\beta)$$

$$B.W = f_2 - f_1$$

$$f_{2f} = f_2(1+A_R\beta)$$

$$f_{1f} = \frac{f_1}{1+A_R\beta}$$

INPUT RESISTANCE WITH FEEDBACK:

* Input resistance without feedback $R_i = \frac{V_i}{I_i}$

Input resistance with feedback $R_{if} = \frac{V_i}{I_s}$

$$R_{if} = \frac{V_i}{I_s}$$

$$= \frac{V_i}{I_i + I_f} = \frac{V_i}{I_i + \beta V_o}$$

$$= \frac{V_i}{I_i + \beta A_R I_i} = \frac{V_i}{I_i(1+A_R\beta)}$$

$$\boxed{R_{if} = \frac{R_i}{1+A_R\beta}}$$

OUTPUT RESISTANCE WITH FEEDBACK (R_{of}):

* By definition, $R_o = \frac{V_o}{I_o}\Big|_{\text{with } I_s = 0}$

$$\cancel{V_i} = \cancel{V_s} - I_f \quad I_i = I_s - I_f$$

$$\cancel{V_i} = -\cancel{I_f} \quad I_i = -I_f \quad (\because I_s = 0$$

By applying KVL to the o/p ckt

$$V_o = I_o R_o + A_R I_i$$

$$V_o = I_o R_o + A_R(-I_f) \qquad (\because A_R = \frac{V_o}{I_i}$$

$$V_o = I_o R_o - A_R \beta V_o \qquad V_o = A_R I_i)$$

$$V_o + A_R \beta V_o = I_o R_o \qquad I_f = \beta V_o)$$

$$V_o[1 + A_R\beta) = I_o R_o$$

$$\frac{V_o}{I_o} = \frac{R_o}{1 + A_R\beta}$$

$$\boxed{R_{of} = \frac{R_o}{1 + A_R\beta}}$$

For Voltage-Shunt feedback

Transfer Gain, $A_{Rf} = \frac{A_R}{1 + A_R\beta}$ (decreases)

Input resistance $R_{if} = \frac{R_i}{1 + A_R\beta}$ (decreases)

Bandwidth $(B.\omega)_f = B.W(1 + A_R\beta)$ (Increases)

output resistance $R_{of} = \frac{R_o}{1 + A_R\beta}$ (decreases)

CURRENT-SERIES FEEDBACK:

* The blockdiagram for a current series feedback is as shown below.

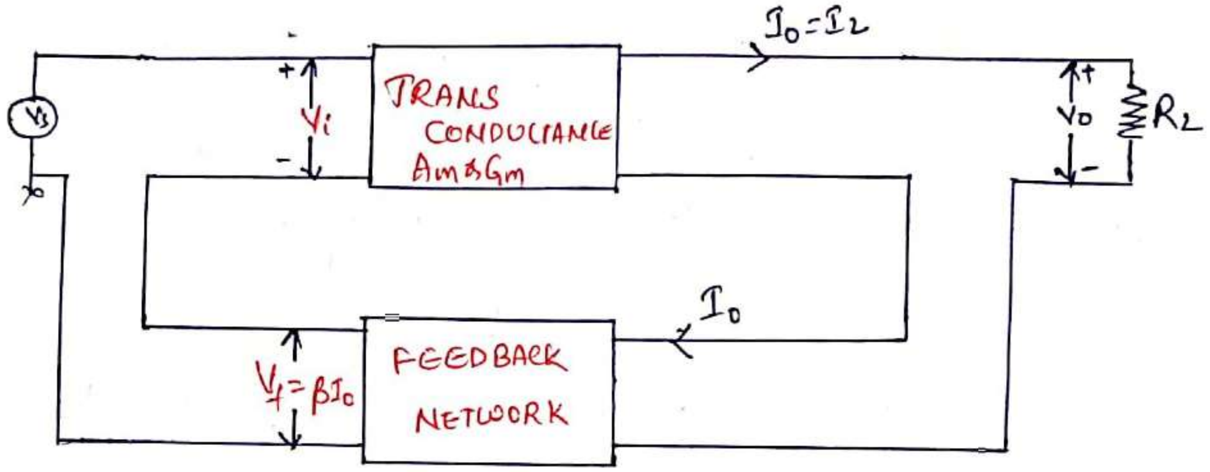

FIG: CURRENT SERIES FEEDBACK AMPLIFIER

* If current is sampled and the mixing is in series with the input then the type of feedback is known as Current series feedback
* Since the current is sampled, the output parameter monitored is current and mixing is series, the parameter affected is the Input Voltage.
* Hence the parameter that is stabilised in current series feedback is transconductance $G_m = \frac{I_o}{V_i}$.
* The feedback factor is the ratio of feedback voltage to the output current that is $\beta = \frac{V_f}{I_o}$

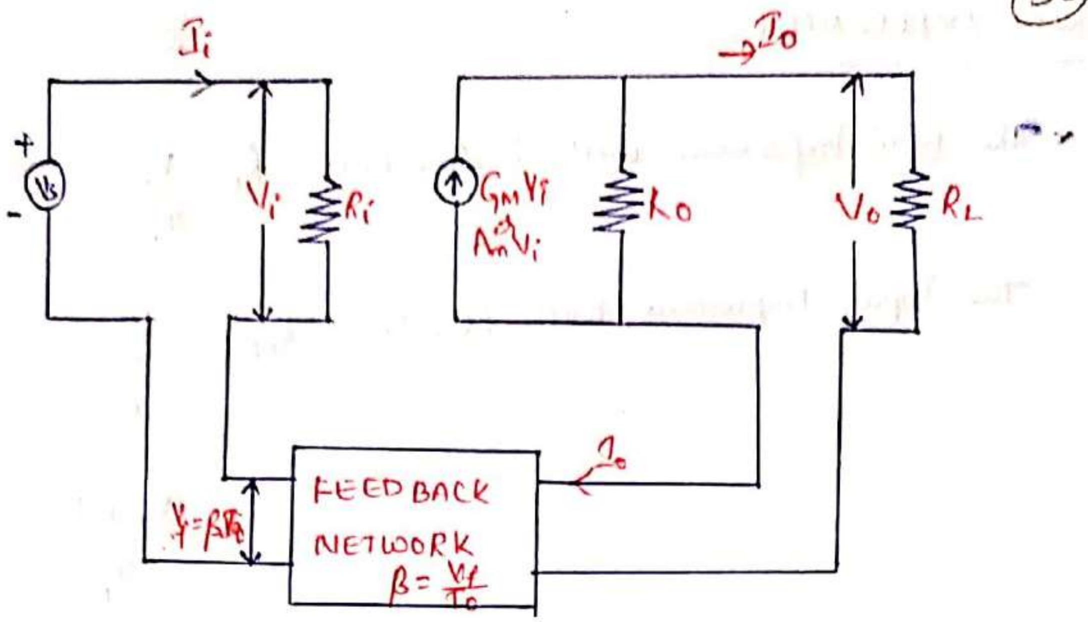

FIG: CIRCUIT DIAGRAM FOR ~~VOLTAG~~ CURRENT SERIES FEEDBACK

TRANSCONDUCTANCE WITH FEEDBACK:

* The transconductance without feedback is $G_m = \frac{I_o}{V_i}$

The transconductance with feedback is $G_{mf} = \frac{I_o}{V_s}$

We know that $V_s = V_i + V_f$

$$G_{mf} = \frac{I_o}{V_i + V_f}$$

$(\because V_f = \beta I_o)$

$$= \frac{I_o}{V_i + \beta I_o}$$

$$= \frac{I_o}{V_i\left[1 + \beta \frac{I_o}{V_i}\right]}$$

$$G_{mf} = \frac{G_m}{1 + \beta G_m}$$

$$\boxed{\therefore G_{mf} = \frac{G_m}{1 + \beta G_m}}$$

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INPUT IMPEDANCE:

* The input impedance without feedback $R_i = \frac{V_i}{I_i}$

The input impedance with feedback $R_{if} = \frac{V_s}{I_i}$

$$= \frac{V_i + V_f}{I_i}$$

$$= \frac{V_i + \beta I_o}{I_i}$$

$$R_{if} = \frac{V_i}{I_i} + \beta \cdot \frac{I_o}{I_i} \qquad \left(\because G_m = \frac{I_o}{V_i} \quad I_o = G_m V_i\right)$$

$$= R_i + \beta \cdot G_m \frac{V_i}{I_i}$$

$$= R_i + \beta G_m R_i$$

$$\boxed{R_{if} = R_i(1+\beta G_m)}$$

OUTPUT IMPEDANCE:

* The output impedance can be obtained by equating $V_s = 0$

Apply KCL to o/p loop

$$R_O = \frac{V_o}{I_O}$$

$$R_{of} = \left.\frac{V_o}{I_O}\right|_{V_s=0}$$

$$I_o + G_m V_i = \frac{V_o}{R_O}$$

$$I_O = \frac{V_o}{R_O} - G_m V_i$$

$$I_O = \frac{V_o}{R_O} - G_m(-V_f) \qquad \left(\because V_s = 0 \quad V_i = V_s - V_f \quad V_i = -V_f\right)$$

$$I_o = \frac{V_o}{I_o} + G_m V_f$$

$$I_o = R_o + G_m \beta I_o$$

* The output Impedance can be obtained by equating $V_s = 0$

$$V_s = V_i + V_f$$

$$0 = V_i + V_f \qquad V_i = -V_f$$

$$R_{of} = \left.\frac{V_o}{I_o}\right|_{V_s = 0}$$

By applying KCL

$$I_o = \frac{V_o}{R_o} + G_m V_i$$

$$I_o = \frac{V_o}{R_o} + G_m(-V_f)$$

$$I_o = \frac{V_o}{R_o} + G_m \beta I_o$$

$$I_o + G_m \beta I_o = \frac{V_o}{R_o}$$

$$I_o(1 + G_m \beta) = \frac{V_o}{R_o}$$

$$R_o(1 + G_m \beta) = \frac{V_o}{I_o}$$

$$\boxed{R_{of} = R_o(1 + G_m \beta)}$$

For Current series feedback

$$A_{Gmf} = \frac{G_m}{1+G_m\beta} \quad \text{(decreases)}$$

$$R_{if} = R_i(1+\beta G_m) \quad \text{(Increases)}$$

$$R_{of} = R_o(1+A_{Gm}\beta) \quad \text{(Increases)}$$

$$(B.\omega)_f = B.\omega(1+G_m\beta) \quad \text{(Increases)}$$

PERFORMANCE COMPARISON OF FEEDBACK AMPLIFIERS *

PARAMETER	VOLTAGE SERIES	VOLTAGE SHUNT	CURRENT SERIES	CURRENT SHUNT
1. Type of Amplifier	Voltage Amplifier	Trans-resistance Amplifier	Trans conductance Amplifier	Current Amplifier
2. Desensitivity	A_{vf}	R_{mf}	G_{mf}	A_{If}
3. Transfer gain	Decreases	Decreases	Decreases	Decreases
4. Non linear distortion	Decreases	Decreases	Decreases	Decreases
5. Noise	reduces	reduces	Reduces	Reduces
6. Bandwidth	Increases	Increases	Increases	Increases
7. Input Impedance	Increases	Decreases	Increases	Decreases
8. Output Impedance	Decreases	Decreases	Increases	Increases

General Characteristics of Negative Feedback Amplifier:

* The Negative feedback improves many desirable characteristics.

The ~~rea~~

Stabilization of Gain with Negative Feedback:

* The variations in temperature, supply voltages, ageing of components & variations in transistor parameters with replacement are some of the factors that affect the gain of an amplifier and cause it to change.

* However, the overall gain of the amplifier can be made independent of these variations if negative feedback is used, this is an ~~most~~ important advantage of negative feedback.

→ The voltage gain with negative feedback is given as,

$$A_f = \frac{A}{1+A\beta}$$

if $A\beta >> 1$ then $A_f = \frac{\cancel{A}}{\cancel{A}\beta} \approx \frac{1}{\beta}$

* Thus, gain with feedback is independent of internal gain of the amplifier and depends on the passive elements such as resistors i.e, feedback network.

* The values of resistors ~~mainly~~ remain fairly constant because they can be chosen very precisely with almost zero temperature coefficient of resistance. Thus the gain is stablized.

$$A_f = \frac{A}{1+A\beta} \rightarrow (1)$$

Differentiating eq(1) w.r.t A

$$\frac{dA_f}{dA} = \frac{(1+A\beta)(1) - \frac{d}{dA}(1+A\beta)\cdot A}{(1+A\beta)^2} \quad \left(\because \frac{d\rho}{dV} = \frac{V\cdot\frac{du}{dx} - u\cdot\frac{dv}{dx}}{V^2}\right)$$

$$= \frac{1+A\beta - (0+\beta)\cdot A}{(1+A\beta)^2} = \frac{1+\cancel{A\beta}-\cancel{A\beta}}{(1+A\beta)^2}$$

$$\frac{dA_f}{dA} = \frac{1}{(1+A\beta)^2}$$

$$dA_f = \frac{dA}{(1+A\beta)^2} \rightarrow (2)$$

dividing eq(2) by eq(1)

$$\frac{dA_f}{A_f} = \frac{dA}{(1+A\beta)^2} \Big/ \frac{A}{1+A\beta}$$

$$\frac{dA_f}{A_f} = \frac{dA}{(1+A\beta)^{\cancel{2}}} * \frac{\cancel{1+A\beta}}{A} = \frac{dA}{A} * \frac{1}{1+A\beta}$$

* The term $\frac{dA_f}{A_f}$ represents the fractional change in amplifier transfer gain with feedback and $\frac{dA}{A}$ denotes the fractional change in voltage gain without feedback.

* The term $\frac{1}{1+A\beta}$ is called sensitivity.

∴ The Sensitivity is defined as the ratio of percentage change in voltage gain with feedback to the percentage change in voltage gain without feedback.

$$\text{Sensitivity} = \frac{\left(\frac{dA_f}{A_f}\right)}{\left(\frac{dA}{A}\right)} = \frac{1}{1+A\beta}$$

∴ The Reciprocal of the term Sensitivity is called Desensitivity.

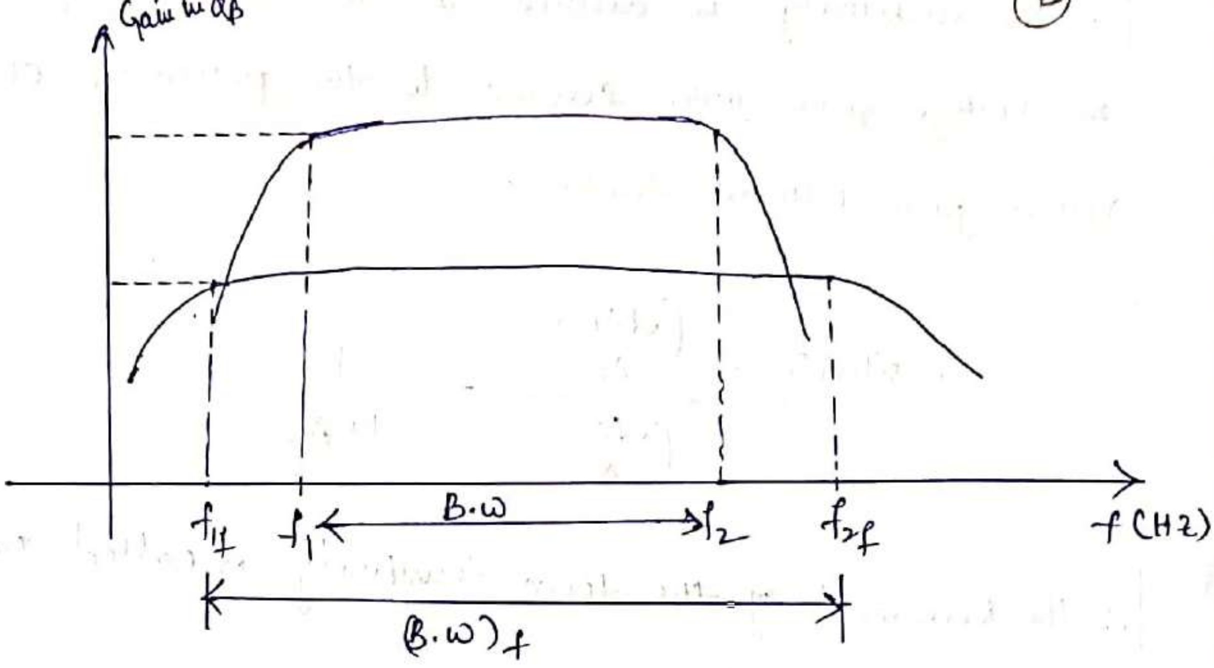
i.e., Desensitivity $D = (1+A\beta)$.

2. INCREASE OF BANDWIDTH:

* The Bandwidth of an amplifier is the difference between the upper cut-off frequency f_2 and the lower cut-off frequency f_1.
* Due to the negative feedback in the amplifier, the upper cut-off frequency f_{2f} is increased by the factor $(1+A\beta)$ and the lower cut-off frequency f_{1f} is decreased by the same factor $(1+A\beta)$.

$$f_{2f} = f_2(1+A\beta)$$

$$f_{1f} = f_1/(1+A\beta)$$



* As the voltage gain of a feedback amplifier reduces by the factor $\frac{1}{(1+A\beta)}$, its bandwidth would be increased by $(1+A\beta)$ i.e.,

$$B.W_f = BW(1+A\beta).$$

DECREASED DISTORTION:

* Consider an amplifier with an open loop voltage gain and ~~total~~ a total harmonic distortion D. Then with the introduction of negative feedback with the feedback ratio β, the distortion will reduce to

$$D_f = \frac{D}{1+A\beta}$$

DECREASED NOISE:

* There are many sources of noise in an amplifier depending upon the active device used. With using the negative feedback with the feedback ratio β, the noise N, can be reduced by a

factor of $1/(1+A\beta)$ in a similar manner to non-linear distortion (40)

Thus the noise with feedback is given by,

$$N_f = \frac{N}{1+A\beta}$$

INCREASE IN INPUT IMPEDANCE:

* An amplifier should have high input impedance so that it will not load the preceding stage or the input voltage source.

* Such a desirable characteristic can be achieved with the help of negative series voltage feedback. The input impedance with feedback is given by,

$$Z_{if} = Z_i(1+A\beta).$$

* Thus the input impedance is increased by a factor of $(1+A\beta)$.

DECREASE IN OUTPUT IMPEDANCE:

* An amplifier with low output impedance is capable of delivering power to the load without much loss. Such a desirable characteristics is achieved by employing negative series voltage feedback in an amplifier

$$Z_{of} = \frac{Z_o}{1+A\beta}$$

METHOD OF ANALYSIS OF FEEDBACK AMPLIFIERS:

* For analyzing the feedback amplifier, it is necessary to go through the following steps.

Step1: Identify Topology (type of feedback)

a) To find the type of Sampling Network.

i. By shorting the output if feedback signal becomes zero, then it is called "Voltage Sampling".

ii. By opening the output loop if feedback signal becomes zero, then it is called "Current Sampling"

b) To find the type of mixing Network.

i. If the feedback signal is subtracted from the externally applied signal as a Voltage in the Input loop, it is called "Series mixing".

ii. If the feedback signal is subtracted from the externally applied signal as a Current in the Input loop, it is called as "Shunt mixing".

* Thus, by finding the type of sampling network and mixing network, type of feedback amplifier can be identified.

Step2: To find the Input circuit.

i. For Voltage Sampling, the output Voltage is made zero by shorting the output.

ii. For Current Sampling, the output current is made zero by opening the output loop. (46)

Step-3: To find the output circuit.

i. For Series mixing, the input current is made zero by opening the input loop.

ii. For Shunt mixing, the input voltage is made zero by shorting the input loop.

Step-2 & Step-3: Ensure that the feedback is reduced to zero without altering the loading on the basic amplifier.

Step-4: Optional. Replace each active device by its h-parameter model at low frequency.

Step-5: Find the open loop gain (gain without feedback) of the amplifier

Step-6: Indicate X_f (feedback voltage or feedback current) and X_o (output voltage or o/p current) on the circuit and evaluate $\beta = X_f/X_o$.

Step-7: From A & β, find D, A_f, R_{if}, R_{of} and R'_{of}

PROBLEMS

1. The Voltage gain of an amplifier without feedback is 3,000 Calculate the Voltage gain of the amplifier if the Negative feedback is introduced in the circuit. Given that feedback fraction $\beta = 0.01$.

Sol Given data Voltage gain without feedback $A_V = 3,000$

feedback fraction $\beta = 0.01$

Find Voltage gain with -ve feedback

$A_{vf} = ?$

We know that $A_{vf} = \dfrac{A_v}{1 + A_v \beta}$

$= \dfrac{3000}{1 + 3000(0.01)}$

$= \dfrac{3000}{31}$

$= 96.774$

$\therefore$ Voltage gain with -ve feedback $A_{vf} = 96.774$

2. Calculate the gain of a -ve feedback amplifier with an internal gain $A_v = 75$ and feedback fraction $\beta = \frac{1}{15}$, what will be the gain if A_v doubles?

Sol Given data Internal gain $A_v = 75$

feedback fraction $\beta = \frac{1}{15}$

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Voltage gain with feedback $A_{vf} = \frac{A_v}{1+A_v\beta}$

$= \frac{75}{1+\cancel{75}^{5}\times\frac{1}{\cancel{15}}}$

$= \frac{75}{6}$

$$\boxed{A_{vf} = 12.5}$$

When A_v doubles i.e., $A_v = 2(75)$

$A_v = 150$

$A_{vf} = \frac{A_v}{1+A_v\beta} = \frac{150}{1+\cancel{150}^{10}\times\frac{1}{\cancel{15}}}$

$= \frac{150}{11}$

$$\boxed{A_{vf} = 13.64}$$

3. An amplifier with -ve feedback gives an o/p of 12.5V with an input of 1.5V. When feedback is removed, it requires 0.25V i/p for the same o/p. Find i) Value of Voltage gain without f/b. ii) Value of β, if the i/p & o/p are ~~in~~ out of phase and β is real.

Sol Given that $V_o = 12.5V$

$V_s = 1.5V$

$V_i = 0.25$

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i. Voltage gain without feedback $A_v = \frac{V_o}{V_i}$

$$= \frac{12.5}{0.25}$$

$$\boxed{A_v = 50}$$

Voltage gain with feedback $A_{vf} = \frac{V_o}{V_s}$

$$= \frac{12.5}{1.5}$$

$$\boxed{A_{vf} = 8.333}$$

ii.) Feedback ratio $\beta = ?$

We know that $A_{vf} = \frac{A_v}{1+\beta A_v}$

$$(1+\beta A_v) A_{vf} = A_v$$

$$1+\beta A_v = A_v / A_{vf}$$

$$\beta = \frac{A_v/A_{vf} - 1}{A_v}$$

$$= \frac{50/8.33 - 1}{50}$$

$$\boxed{\beta = 0.10}$$

OSCILLATORS

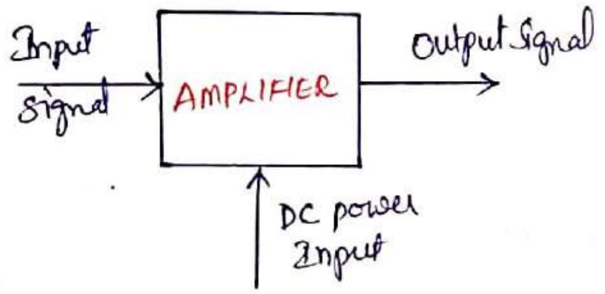
INTRODUCTION:

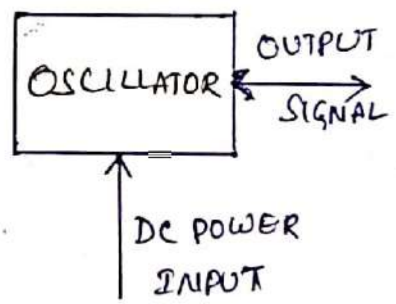
* As, we know that an amplifier strengthens the input signal without any change in its waveform and frequency. The additional power required comes from the external dc source.
* Thus an amplifier is essentially an energy converter that draws energy from a dc supply and converts it into ac energy at signal frequency, the energy conversion process being controlled by the input signal.
* On the other hand an oscillator does not require any external signal either to start or maintain the process of energy conversion and the energy conversion process is controlled by the oscillator itself.
* Oscillators finds wide applications in electronic ~~measuring~~ Communication equipment. In AM, FM super heterodyne receivers, "local oscillator" are used to assist in the reduction of the incoming Radio frequency (IF).
* other applications include their use as "clocks" in digital systems such as microcomputers, in the sweep circuits found in TV sets and oscilloscops.

OSCILLATOR:

DEFINITION: The electronic circuit which is used to generate a periodic waveform without an AC input signal is called as Oscillator.

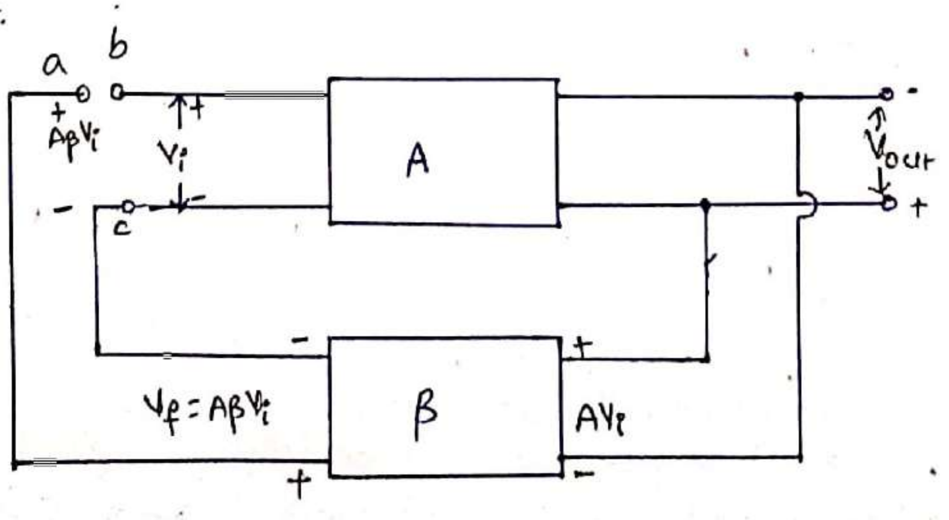
* As we know that an amplifier strengthens the input signal without any change in its waveform and frequency.
* In an oscillator, the output signal frequency depends on the passive components employed in the circuit. Oscillator may provide fixed & variable frequency.





OPERATION OF OSCILLATOR:

* To understand how an oscillator produces an output signal without an external input signal, let us consider the feedback circuit.



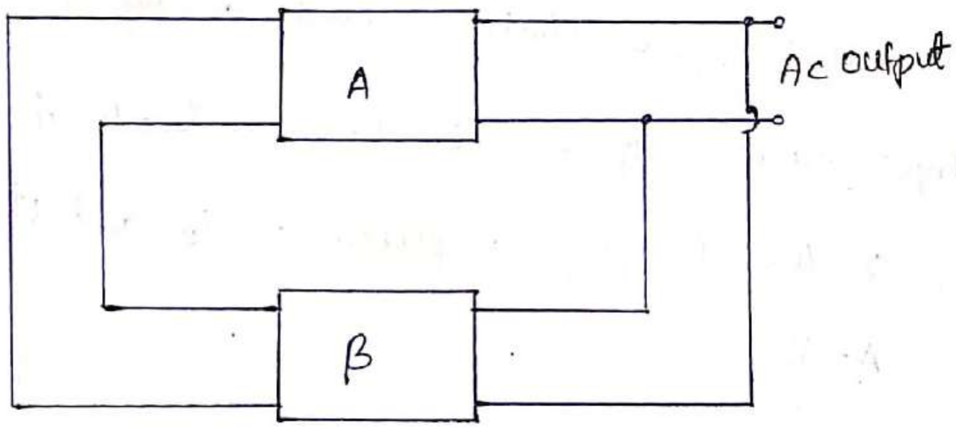
* Where V_i is the voltage of ac input driving the input terminals bc of an amplifier having voltage gain A. 3

* The amplified voltage is $V_o = AV_i$.

* This voltage drives a feedback circuit that is usually a resonant circuit, as we get maximum feedback at one frequency. The feedback voltage returning to point a is given by,

$$V_f = A\beta V_i$$

where β is the gain of feedback network.



* Here, the amplifier generates 180° phase shift.

CONDITION FOR OSCILLATIONS (or) BARKHAUSEN CRITERION:

1. The total phase shift around the loop should be zero degrees or 360°.

2. The magnitude of product of the open loop gain of the Amplifier (A) and the feedback factor is unity i.e., $|A\beta| = 1$.

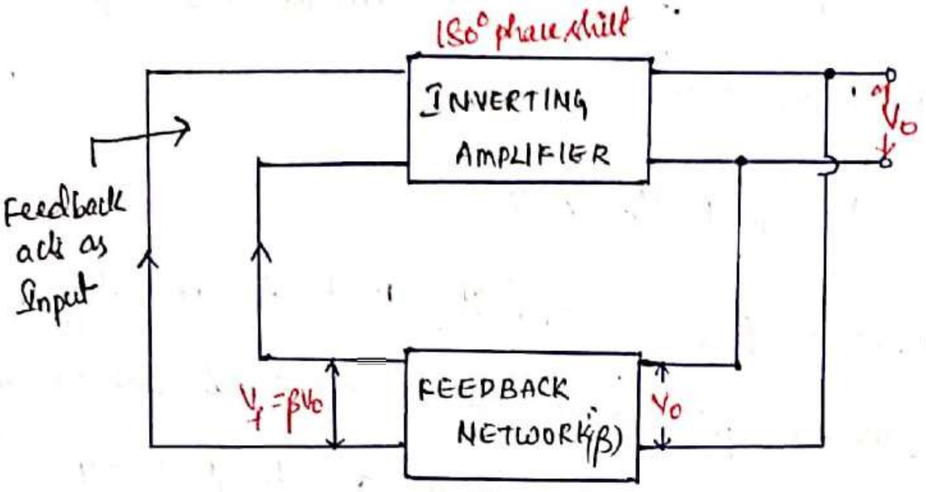


FIG: BASIC BLOCK DIAGRAM OF OSCILLATOR CIRCUIT

* For the feedback network input is 'V_o' then the feedback network produces 180° phase shift
* This feedback signal is given to the input of the inverting amplifier so that phase shift around a loop is 0° or 360°.
* Let the input voltage of the feedback network is V_o i.e. output voltage of the inverting amplifier is V_o and it is given by,

$$A = \frac{V_o}{V_i}$$

$$V_o = AV_i$$

* Feedback network provides 180° phase shift and it is given by

$$\beta = \frac{-V_f}{V_o} \qquad V_f = -\beta V_o$$

* Where -ve sign indicates the 180° phase shift provided by the feedback Network.

$$V_f = -\beta V_o \Rightarrow V_f = -\beta A V_i$$

⇒ For the oscillator V_f must acts as a input voltage of Inverting amplifier.

$$V_i = V_f$$

$$V_i = -A\beta V_i$$

$$-A\beta = 1$$

$$\boxed{|A\beta| = 1}$$

-) The above Condition is called Barkhausen criterion.

-) The inverting amplifier produces 180° phase shift and the feedback network produces 180°. So that phase shift around the loop is 360°.

* The above two Conditions are required to be satisfied by the circuit to work as an oscillator producing sustained oscillations of Constant frequency and amplitude.

* Let us see the effect of magnitude of the product of gain and feedback factor on the nature of the oscillation.

1. $|A\beta| > 1$

* The total phase shift around a loop is 0° or 360° and $|A\beta| > 1$ then the oscillations are growing type. The amplitude of oscillation goes on increasing

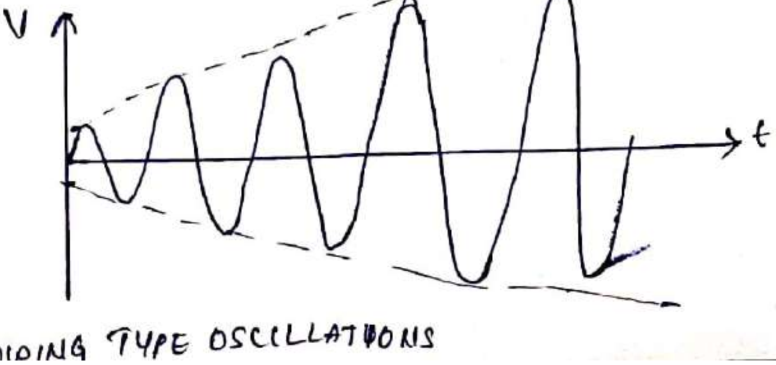


FIG: GROWING TYPE OSCILLATIONS

2. $|A\beta| = 1$

* When the total phase shift around a loop is $0°$ or $360°$ ensuring positive feedback and $|A\beta| = 1$ then the oscillations are with constant frequency and amplitude called Sustained oscillations.

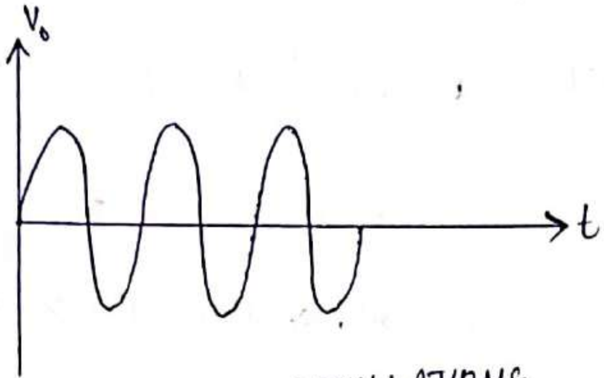


FIG: SUSTAINED OSCILLATIONS.

3. $|A\beta| < 1$

* When total phase shift around a loop is $0°$ or $360°$ but the oscillations are of decaying type i.e., the $|A\beta| < 1$ then the oscillations are of decaying type i.e., the amplitude decreases exponentially.

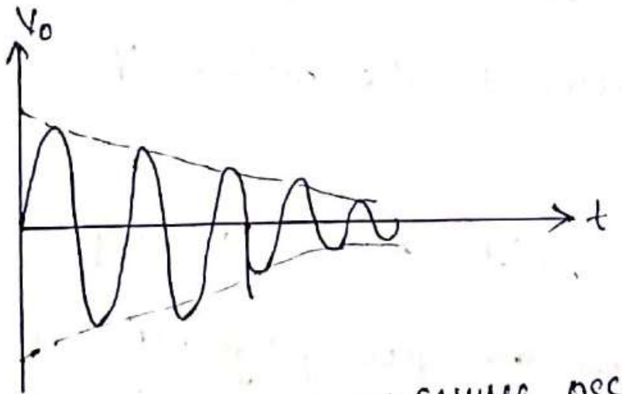


FIG: EXPONENTIALLY DECAYING OSCILLATIONS.

CLASSIFICATION OF OSCILLATORS:

* Oscillators are classified in the following different ways.

1. According to the waveforms generated
   a. Sinusoidal oscillator
   b. Relaxation oscillator

a. Sinusoidal oscillator generates sinusoidal waveforms.

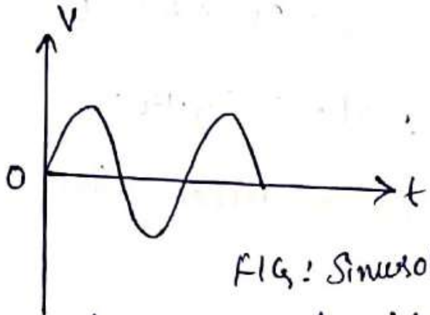


FIG: Sinusoidal wave form

b. Relaxation oscillator generates voltages or currents which vary abruptly one or more times in a cycle of oscillation.

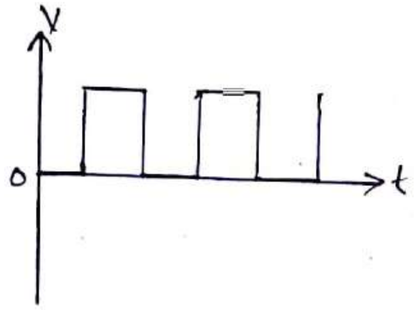


FIG: Square waveform

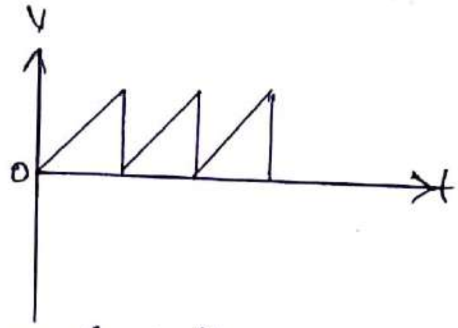


FIG: Sawtooth

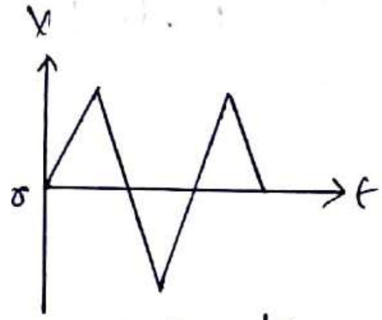


FIG: Triangular

2. According to the fundamental mechanism involved
   a. Negative Resistance oscillators
   b. Feedback oscillator

* Negative resistance oscillator uses negative resistance of the amplifying device to neutralise the positive resistance of the oscillator.

Feedback oscillator uses positive feedback in the feedback amplifier to satisfy the Barkhausen criterion.

3. According to the frequency generated:

a. Audio frequency oscillator (AFO) : up to 20KHZ.

b. Radio frequency oscillator (RFO) : 20KHZ to 30MHZ

c. Very high frequency (VHF) oscillator : 30 MHZ to 300 MHZ.

d. Ultra high frequency (UHF) oscillator : 300 MHZ to 3GHZ.

e. Microwave frequency oscillator : above 3GHZ.

4. According to the type of circuit used, Sine-wave oscillators may be classified as,

a. LC tuned oscillators

b. RC phase shift oscillator

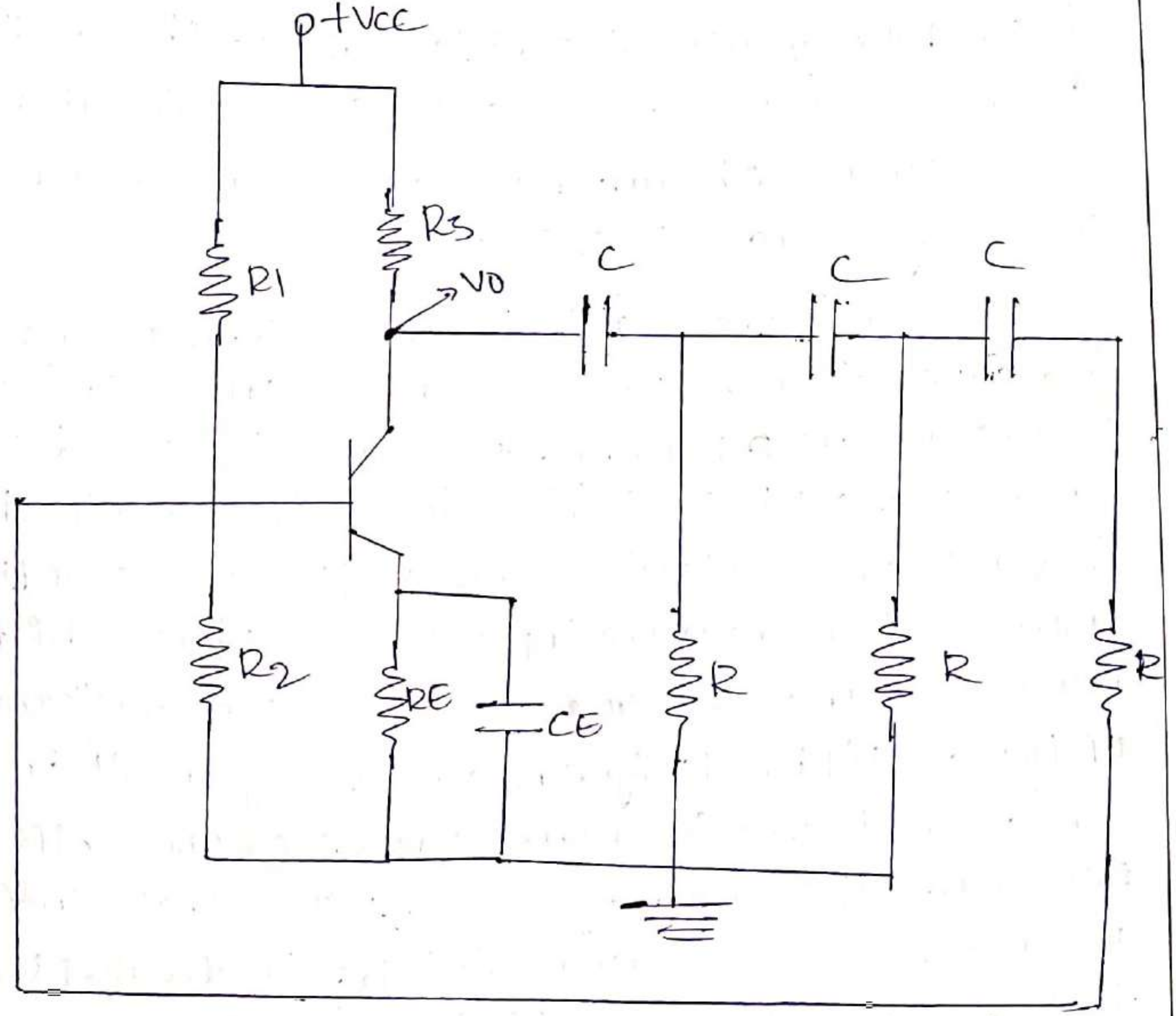
* RC phase shift oscillator *

for producing oscillations is an oscillator circuit we need positive feedback which means that the voltage signal feedback should be in phase with the input signal. for providing a position positive feedback at one particular frequency, an inverting amplifier may be used with a feedback network that causes a phase shift of 180° at the desired frequency of oscillations as shown in figure. The 180° phase shift in the feedback signal can be obtained by a suitable network consisting of three R-C sections.

When a phase shift network such as that indicated given below is used in a phase shift oscillator, the R's and C's must be selected so as to produce a phase shift of 180° at the desired frequency of oscillation. The output of the voltage amplifier is fed to the input to the phase shift network. Thus $V_1 = V_{out}$. The output resistance of the amplifier designed to be very small in comparison to the input impedance of the shift network. The output voltage of the phase shift network V_2 is fed into the input of the amplifier. i.e., $V_2 = V_{in}$. The amplifiers input impedance must be much larger than the output impedance of the phase shift network.

Alternatively, a positive feedback can be obtained by using two stages of amplifiers each giving a phase shift of 180°. A part of this output is fedback to the input through a feedback network without causing any further phase shift. wein bridge oscillator operates on this principle.

(i)



Circuit diagram.

(i) h_{re} of the transistor is usually negligibly small and therefore, h_{re} Vout is omitted from the circuit.

(ii) h_{oe} of the transistor is very small i.e. $\frac{1}{h_{oe}}$ is much larger than RC. Thus the effect of hoe can be neglected.

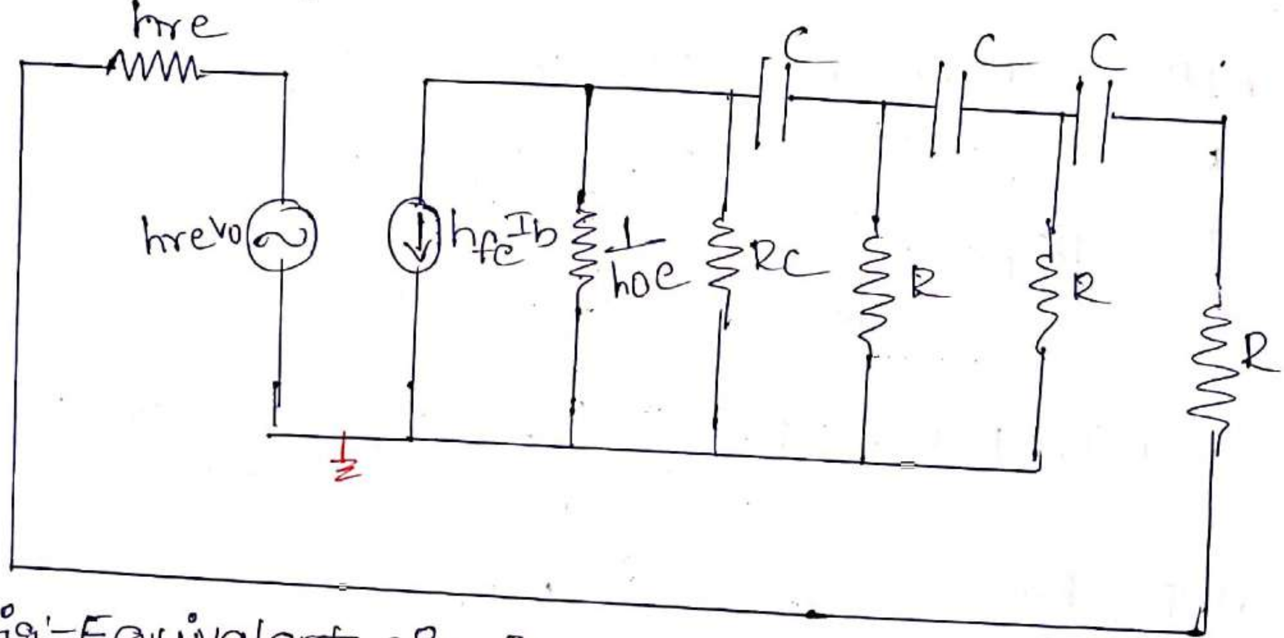
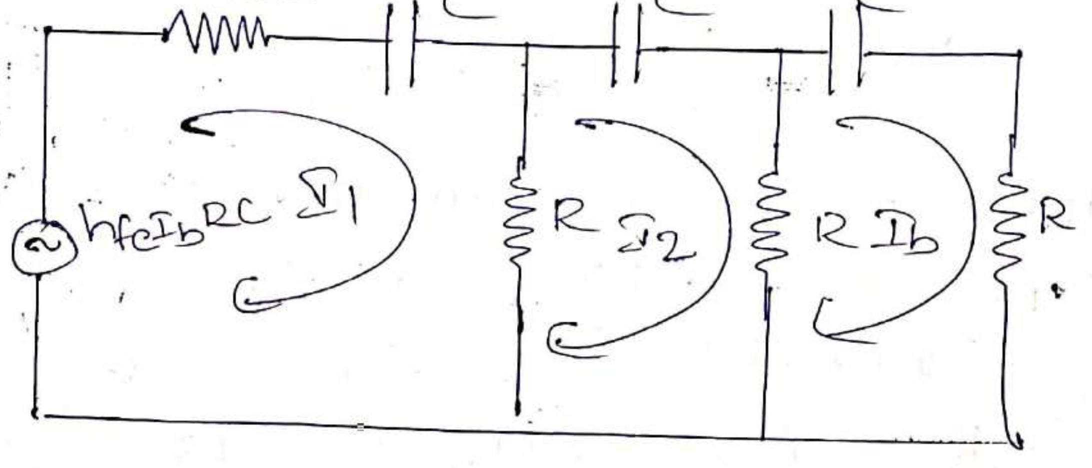


fig:- Equivalent circuit of Transistor phase-shift oscillator is shown above.

Making above asumptions and replacing current source by equivalent voltage source, the simplified equivalent circuit is shown below.



Apply KVL to the loop 1.

$$+h_{fe}I_bR_C + I_1R_C + \frac{1}{j\omega C_1}I_1 + R(I_1 - I_2) = 0$$

$$h_{fe}I_bR_C + I_1R_C + \frac{1}{j\omega C_1}I_1 + RI_1 - RI_2 = 0$$

$$h_{fe}I_bR_C + I_1\left(R_C + \frac{1}{j\omega C_1} + R\right) - RI_2 = 0 \rightarrow (1)$$

Apply KVL to the loop 2.

$$R(I_2 - I_1) + \frac{1}{j\omega C}I_2 + R(I_2 - I_b) = 0$$

$$RI_2 - RI_1 + \frac{1}{j\omega C}I_2 + RI_2 - RI_b = 0$$

$$-RI_1 + I_2\left(2R + \frac{1}{j\omega C}\right) - RI_b = 0 \rightarrow (2)$$

Apply KVL to the loop (3)

$$R(I_b - I_2) + \frac{1}{j\omega C}I_b + RI_b = 0$$

$$RI_b - RI_2 + \frac{1}{j\omega C}I_b + RI_b = 0$$

$$-RI_2 + I_b\left(R + R + \frac{1}{j\omega C}\right) = 0 \rightarrow (3)$$

where, $\frac{1}{j\omega C} = X_C$ (or) jX_C.

Apply cramer's rule from above equations.

$$I_1(R_C + R - jX_C) - RI_2 - h_{fe}R_CI_b = 0 \rightarrow (1)$$

$$-RI_1 + (2R + jX_C)I_2 - RI_b = 0 \rightarrow (2)$$

$$-RI_2 + (2R + jX_C)I_b = 0 \rightarrow (3)$$

cramer's rule,

$$\begin{vmatrix} R_C+R-jX_C & -R & h_{fe}R_C \\ -R & 2R-jX_C & -R \\ 0 & -R & 2R-jX_C \end{vmatrix} = 0$$

$$= R_C+R-jX_C\left((2R-jX_C)^2-R^2\right)-(-R)\left(-R(2R-jX_C)-0\right] + (-h_{fe}R_C)(R^2+0)=0$$

$$= R_C+R-jX_C\left[\left((2R)^2-2\cdot 2R\cdot jX_C\right)-X_C^2-R^2\right] + R(-2R^2+jRX_C)-h_{fe}R_C R^2=0$$

$$= (R_C+R-jX_C)(3R^2-j4X_CR-X_C^2)-2R^3+jR^2X_C - h_{fe}R_CR^2=0$$

$$= 3R^2R_C-j4X_CR_CR-R_CX_C^2+3R^3-j4X_CR^2-X_C^2R - j3X_CR^2-4RX_C^2+jX_C^3-2R^3+jR^2X_C-h_{fe}R_CR^2=0$$

$$= R^3+R^2R_C(3+h_{fe})-5RX_C^2-R_CX_C^2-6jR^2X_C - j4RR_CX_C+jX_C^3=0$$

equating the imaginary component of the equal to zero.

$$6R^2X_C+4RR_CX_C-X_C^3=0$$

$$X_C(6R^2+4RR_C-X_C^2)=0$$

$$6R^2+4RR_C-X_C^2=0$$

$$6R^2+4RR_C=X_C^2$$

$$X_C=\sqrt{6R^2+4RR_C}$$

$$\therefore X_C = \frac{1}{2\pi f c}$$

$$\frac{1}{2\pi f c} = \sqrt{\cancel{8}6R^2 + 4RRC}$$

$$\frac{1}{2\pi f c} = \sqrt{R^2\left(6 + \frac{4RC}{R}\right)}$$

$$\therefore RC = R$$

$$\frac{1}{2\pi f c} = \sqrt{R^2\left(6 + 4\frac{\cancel{R}}{\cancel{R}}\right)}$$

$$\frac{1}{2\pi f c} = \sqrt{R^2 (10)}$$

$$\frac{1}{2\pi f c} = R\sqrt{10}$$

$$1/f = \frac{1}{2\pi RC\sqrt{10}}$$

$$\boxed{f = \frac{1}{2\pi RC\sqrt{10}}}$$

$\therefore$ where $\frac{RC}{R} = k$.

$$\frac{1}{2\pi f c} = \sqrt{R^2 (6 + 4k)}$$

$$\frac{1}{2\pi f c} = \sqrt{R^2 (6 + 4k)}$$

$$\boxed{f = \frac{1}{2\pi RC\sqrt{6 + 4k}}}$$

Now equating real term into zero. 15

$$R^3 + (3+h_{fe}) R^2 R_C - X_C^2 R_C - 5X_C^2 R = 0$$

$$R^3 + 3R^2 R_C + h_{fe} R^2 R_C - X_C^2 R_C - 5X_C^2 R = 0$$

$$R^3 + 3R^2 R_C + h_{fe} R^2 R_C - X_C^2 (R_C + 5R) = 0$$

$\therefore X_C^2 = 6R^2 + 4RR_C$ from imaginary part.

$$R^3 + 3R^2 R_C + h_{fe} R^2 R_C - (6R^2 + 4RR_C)(R_C + 5R) = 0$$

$$R^3 + 3R^2 R_C + h_{fe} R^2 R_C - 6R^2 R_C - 30R^3 - 4RR_C^2 - 20R^2 R_C = 0$$

$$-23 R^2 R_C + h_{fe} R^2 R_C - 29 R^3 - 4RR_C^2 = 0$$

$$-29 R^3 - 23 R^2 R_C - 4RR_C^2 + h_{fe} R^2 R_C = 0$$

$$-29 R^3 - 23 R^2 R_C - 4R_C^2 R + h_{fe} R^2 R_C = 0$$

$$h_{fe} R^2 R_C = 29 R^3 + 23 R^2 R_C + 4RR_C^2$$

$$h_{fe} \cancel{R^2 R_C} = \cancel{R^2 R_C}\left[29 \cdot \frac{R^3}{R^2 R_C} + 23 \frac{R^2 R_C}{R^2 R_C} + \frac{4RR_C^2}{R^2 R_C}\right]$$

$$h_{fe} = 29 \cdot \frac{R}{R_C} + 23 + \frac{4R_C}{R}$$

$\therefore$ If $R_C = R$

$A_V = 29$

$\beta = \frac{1}{29}$

$A\beta = 29 \cdot \frac{1}{29}$

$\boxed{A\beta = 1}$

$$\therefore \frac{R_C}{R} = k \rightarrow \frac{R}{R_C} = \frac{1}{k}.$$

$$h_{fe} = \frac{29}{k} + 23 + 4k$$

$$h_{fe} = 29 + 23 + 4$$

$$\boxed{h_{fe} = 56}$$

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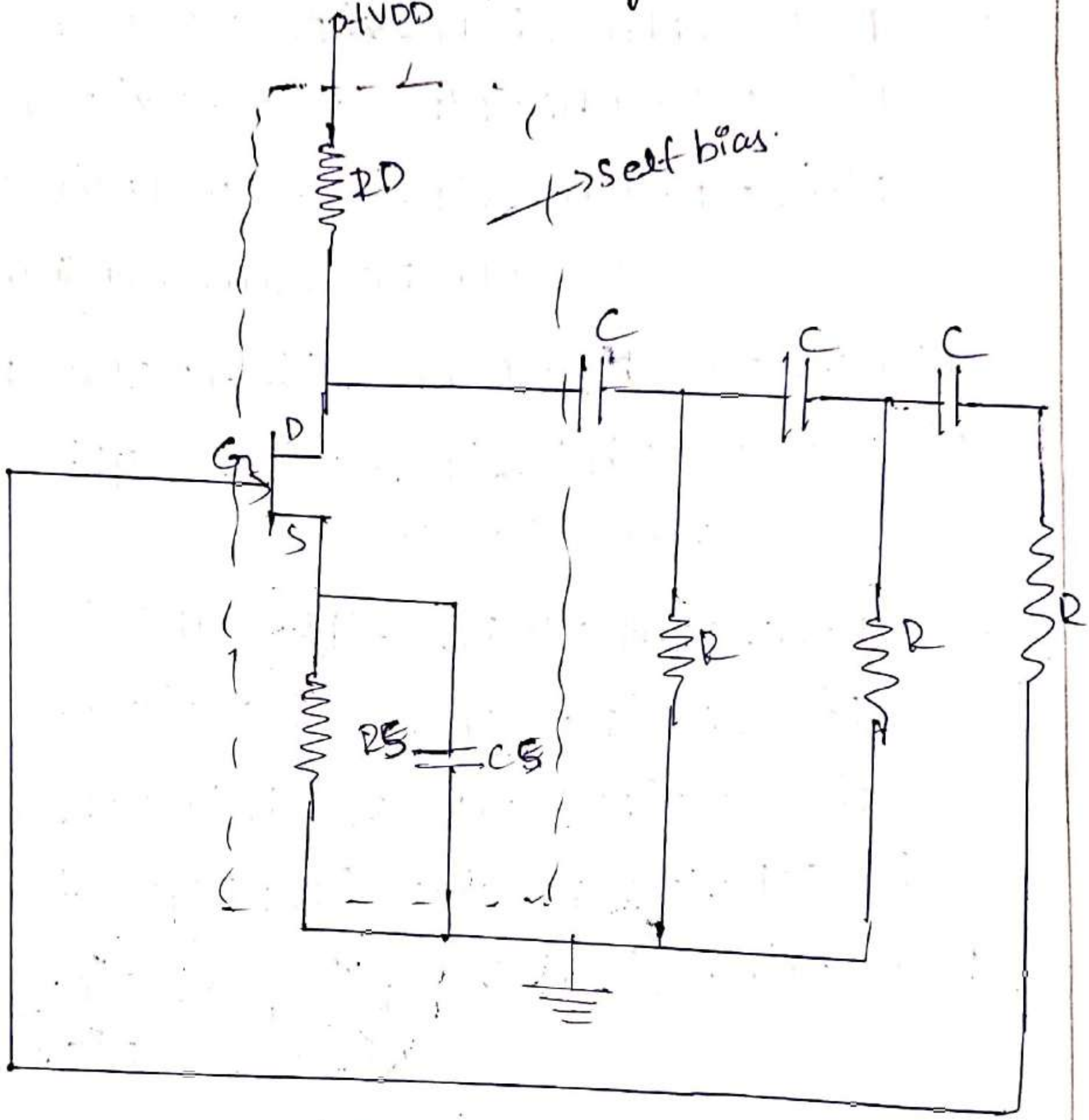
RC phase shift oscillator using FET:-



$X_C = \sqrt{6}\,R$

$\frac{1}{\omega C} = \sqrt{6}\,R$

$\frac{1}{2\pi f C} = \sqrt{6}\,R$

$2\pi f C = \frac{1}{\sqrt{6}\,R}$

$$\boxed{f_0 = \frac{1}{2\pi f C \sqrt{R}\,6}}$$

WIEN BRIDGE OSCILLATOR :

* It is one of the most popular type of oscillators used in audio and sub-audio frequency ranges (20 - 20KHZ).
* The circuit diagram of Wein bridge oscillator using BJT is shown in the below figure.

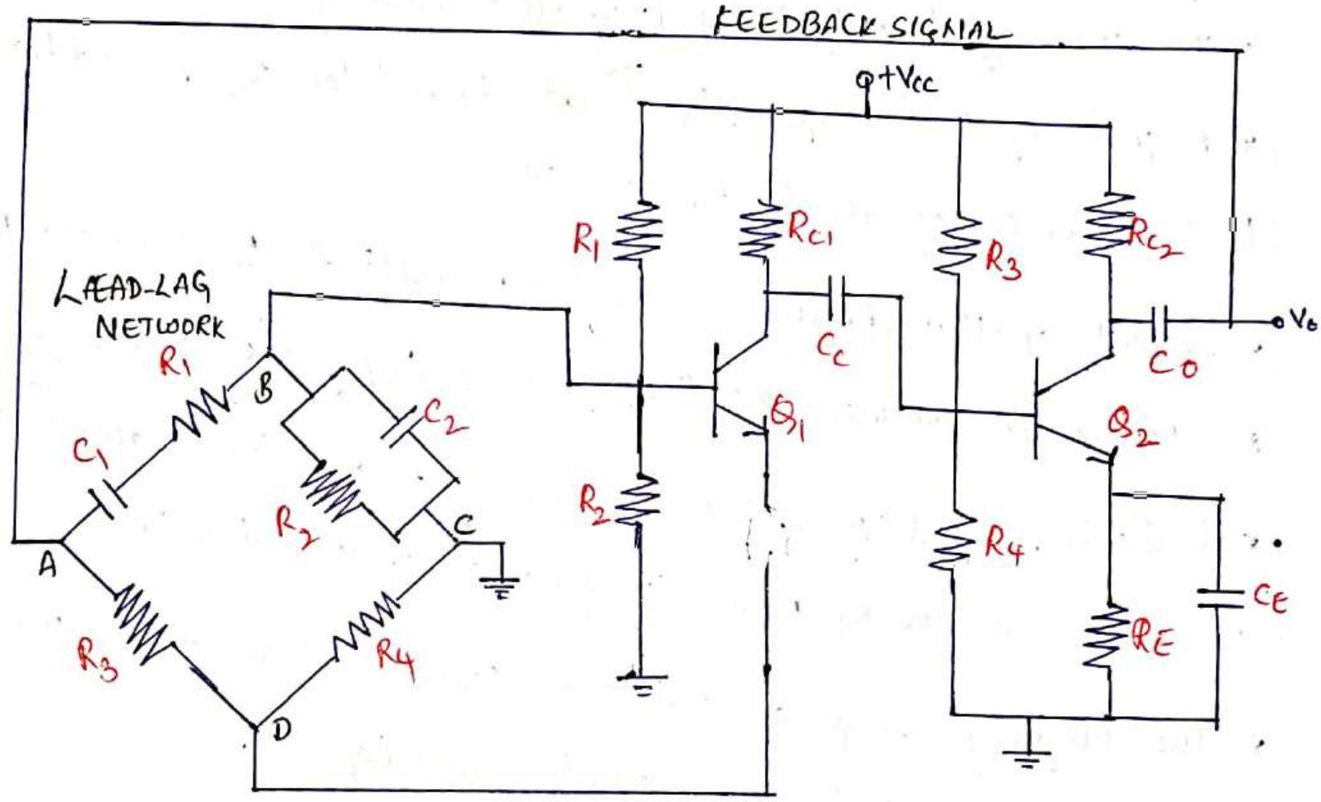


FIG: WEIN BRIDGE OSCILLATOR USING BJT

OPERATION :

* The circuit is set in oscillation by any random change in base current of transistor Q_1, that may be due to noise inherent in the transistor or variation in voltage of dc supply.
* This variation in base current is amplified in collector circuit of transistor Q_1 but with a phase shift of 180°.

* The output of transistor Q_1 is fed to the base of second transistor Q_2 through Capacitor C_4. Here Q_1 acts as both oscillator & amplifier
* Now a still further amplified and twice phase-reversed signal appears at the Collector of the transistor Q_2.
* Having been inverted twice, the output signal will be in phase with the signal ~~output~~ input to the base of transistor Q_1.
* A part of the output signal at transistor Q_2 is fed back to the input of the bridge circuit.
* A part of this feedback signal is applied to emitter Resistor R_4 where it produces negative feedback to provide Constant o/p over a Range of frequencies
* Similarly, a part of feedback signal is applied across the base bias resistor R_2 where it produces positive feedback.
* The frequency range of oscillator can be changed by varying values of resistors and capacitors R_1, R_2 and C_1, C_2

CONSTRUCTION:

* Wein bridge oscillator is essentially a two-stage amplifier with ~~an~~ R-C bridge circuit (wein bridge).

* Here, wein bridge is a lead-lag network (R_1-C_1 & R_2-C_2). The phase shift across the network lags with increasing frequency and leads with decreasing frequency.

* By adding Wein bridge feedback network, the oscillator becomes sensitive to a signal of only one particular frequency.

* This particular frequency is that at which wein bridge is balanced and for which the phase shift is 0°.

* By employing wein-bridge feedback network, frequency stability is increased.

* In the bridge circuit R_1 in series with C_1, R_3, R_4 and R_2 in parallel with C_2 form the four arms.

* From the analysis of the bridge circuit it is obvious that the bridge will be balanced only when,

$$\frac{R_3}{R_4} = \frac{R_1}{R_2}$$

$$R_3 R_2 = R_1 R_4$$

$$R_3 (R_2 \| C_2) = (R_1 \text{ Series } C_2) R_4$$

$$R_3 \left[\frac{R_2 * 1/j\omega C_2}{R_2 + \frac{1}{j\omega C_2}} \right] = \left(R_1 + \frac{1}{j\omega C_1} \right) R_4 \rightarrow ①$$

$$R_3 \left[\frac{R_2 / j\omega C_2}{j\omega C_2 R_2 + 1 / j\omega C_2} \right] = \left\{ R_1 - \frac{j}{\omega C_1} \right] R_4$$

$$R_3 \left[\frac{R_2}{1 + j\omega R_2 C_2} \right] = \left[R_1 - \frac{j}{\omega C_1} \right] R_4$$

$$R_2 R_3 = R_4 (R_1 - j/\omega C_1)(1 + j\omega R_2 C_2)$$

$$R_2 R_3 = R_4\left(R_1 - \frac{j}{\omega C_1} + j\omega C_2 R_2 R_1 - \frac{j^2 \omega C_2 R_2}{\omega C_1}\right)$$

$$R_2 R_3 = R_4\left(R_1 - \frac{j}{\omega C_1} + j\omega R_1 R_2 C_2 + \frac{C_2 R_2}{C_1}\right)$$

$$R_2 R_3 = R_1 R_4 - j\frac{R_4}{\omega C_1} + j\,\omega R_1 R_2 C_2 R_4 + R_2 R_4 \frac{C_2}{C_1}$$

$$R_2 R_3 - R_1 R_4 - \frac{C_2}{C_1} R_2 R_4 + j\left[\frac{R_4}{\omega C_1} + R_1 R_2 C_2 R_4 \omega\right] = 0$$

Separating real and imaginary terms we have

$$R_2 R_3 - R_1 R_4 - \frac{C_2}{C_1} R_2 R_4 = 0$$

$$\frac{C_2}{C_1} R_2 R_4 = R_2 R_3 - R_1 R_4$$

$$\frac{C_2}{C_1} \cancel{R_2 R_4} = \cancel{R_2 R_4}\left[\frac{R_2 R_3}{R_2 R_4} - \frac{R_1 R_4}{R_2 R_4}\right]$$

$$\frac{C_2}{C_1} = \frac{R_3}{R_4} - \frac{R_1}{R_2} \qquad \frac{C}{C} = \frac{R_3}{R_4} - \frac{R}{R}$$

Equating imaginary parts to zero $\quad 1 = \frac{R_3}{R_4} - 1$

$$\frac{R_4}{\omega C_1} + R_1 R_2 C_2 R_4 \omega = 0 \qquad \frac{R_3}{R_4} = 2$$

$$\frac{R_4 - R_1 R_2 C_2 R_4 \omega^2 C_1}{\omega C_1} = 0 \qquad \boxed{R_3 = 2R_4}$$

$$R_4 - R_1 R_2 C_1 C_2 R_4 \omega^2 = 0$$

$$R_1 R_2 C_1 C_2 \cancel{R_4} \omega^2 = \cancel{R_4}$$

$$\omega^2 = \frac{1}{R_1R_2C_1C_2}$$

$$\omega = \frac{1}{\sqrt{R_1R_2C_1C_2}}$$

$$2\pi f = \frac{1}{\sqrt{R_1R_2C_1C_2}}$$

$$\boxed{f = \frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}}$$

$C_1 = C_2 = C$, $R_1 = R_2 = R$ then eq (1) becomes

$$\frac{R_3R}{1+j\omega CR} = \frac{R_4(1+j\omega CR)}{j\omega C}$$

$$jR_3R\omega C = R_4(1+j\omega CR)^2$$

$$jR_3R\omega C = R_4(1-\omega^2R^2C^2+2j\omega RC)$$

$$jR_3R\omega C = R_4 - \omega^2R^2C^2R_4 + 2j\omega RCR_4$$

$$R_4 - \omega^2R^2C^2R_4 + j(2\omega RCR_4 - R_3R\omega C) = 0$$

equating imaginary part to zero

$$2\omega RCR_4 - R_3R\omega C = 0$$

$$2\cancel{\omega RC}R_4 = R_3\cancel{\omega CR}$$

$$\boxed{R_3 = 2R_4}$$

If $C_1 = C_2 = C$ and $R_1 = R_2 = R$, then $\boxed{f = \frac{1}{2\pi CR}}$

$$\boxed{R_3 = 2R_4}$$

* Thus, we see that in a bridge circuit the output will be in phase with the input only when the bridge is balanced i.e, at resonant frequency.

* At all other frequencies the bridge is off-balance i.e, the voltage feedback and output voltage do not have the correct phase relationship for sustained oscillations

* The amplifier voltage gain, $A = \frac{R_3+R_4}{R_4} = \frac{R_3}{R_4} + \frac{R_4}{R_4} = \frac{R_3}{R_4}+1$

$$R_3 = 2R_4 \Rightarrow A = \frac{2\cancel{R_4}}{\cancel{R_4}}+1 \quad \boxed{A=3}$$

* The above corresponds with the feedback network attenuation of 1/3. Thus in this case, voltage gain must be equal to or greater than 3 to sustain oscillations.

LC OSCILLATORS:

* The oscillators which use the elements L and C to produce the oscillations are called LC oscillators.
* The circuit using L and C is called tank circuit & oscillatory circuit or tuned circuit.
* These oscillators are used for high frequency range from 200KHz to few MHz.
* Due to high frequency range, these oscillators are used for sources of RF (Radio frequency) range.

OPERATION OF LC TANK CIRCUIT:

* The LC tank circuit consists of elements L and C connected in parallel as shown in the figure.

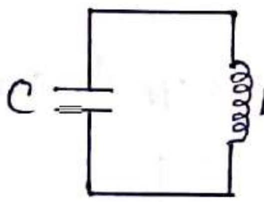


FIG: LC TANK CIRCUIT

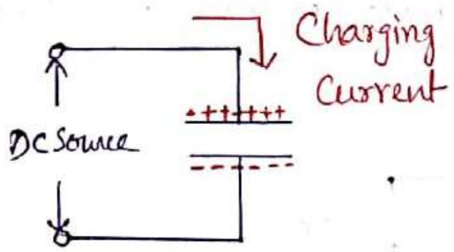


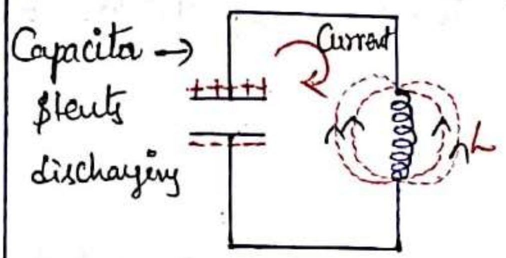
FIG: INITIAL CHARGING

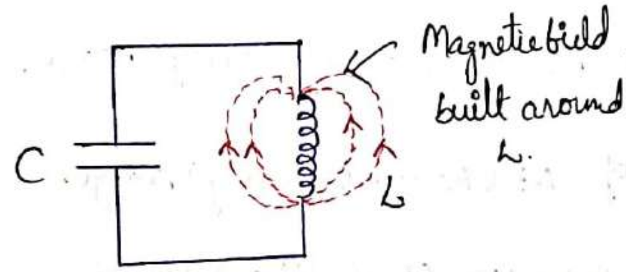
* Let capacitor is initially charged from a DC source with the polarities as shown in the figure.
* When the capacitor gets charged, the energy gets stored in a capacitor as Electrostatic Energy.
* Then such a charged capacitor is connected across Inductor in a tank circuit, then the capacitor starts discharging through Inductor.

* Then the Conventional Current flows due to this the magnetic field gets set up around the inductor L. Thus inductor stores energy in the form of Electromagnetic field. 23

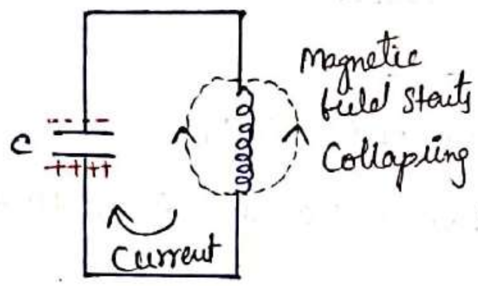
* When Capacitor is fully discharged, the maximum Current flows through the circuit.

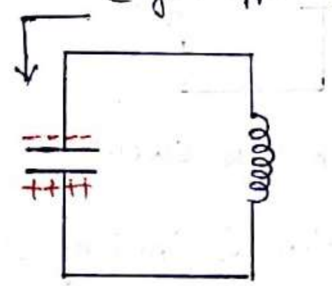
* At this instant all the electrostatic energy gets stored as a magnetic energy in the Inductor L.





* Now, the magnetic field around L starts collapsing. As per Lenz's law. This starts the charging the capacitor with opposite polarity making lower plate positive and upper plate negative as shown in figure.





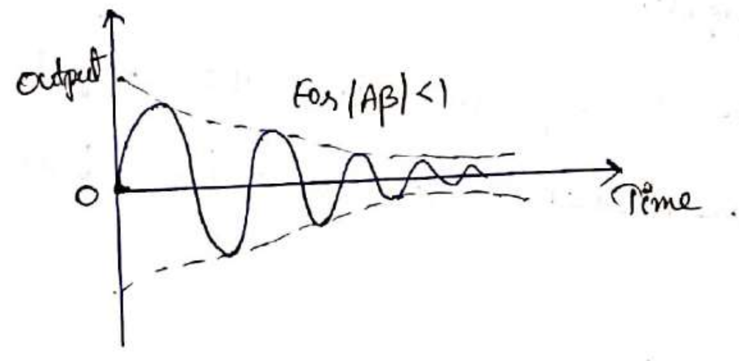
* Then Capacitor again starts discharging through inductor L, but the direction of Current through the inductor in 'L' but the direction of Current through the inductor in the 'L' but the direction as shown in figure.

* Again Electrostatic energy converted to magnetic field when the Capacitor is fully discharged and again magnetic field collapses and again the Capacitor gets charging in opposite direction 24

* Thus the Capacitor charges with alternate polarities and discharging producing alternating current in the tank circuit

* This alternating current generally Electronic oscillations.. But these Oscillations of the Capacitor are damped because every time transfer of energy from L to C and C to L dissipates energy in the form of heat in the resistance of the coil and in the connecting wires in the form of electromagnetic radiation.

* These losses decrease the amplitude of oscillating current gradually till it ceases. These are called as exponentially decaying oscillations or damped oscillations.



GENERAL FORM OF AN L-C OSCILLATOR:

* In general form of an oscillator, the amplifier section may be active devices such as vaccum tube, BJT, FET & op-amp may be used in the amplifier section.

* Z_1, Z_2 & Z_3 are the reactive elements constituting the feedback tank circuit which determines the frequency of oscillations.

* Here, Z_1 & Z_2 serve as an ac voltage divider for the output voltage and feedback signal.

* Thus, the voltage across Z_2 is the feedback signal.

* The equivalent circuit is as shown below.

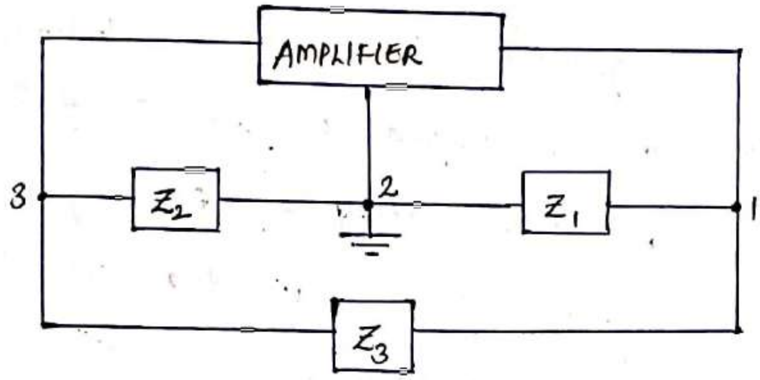


FIG: GENERAL FORM OF AN OSCILLATOR

* The equivalent circuit is drawn with the following two assumptions.

i. h_{re} of transistor is negligibly small and therefore, the feedback source $h_{re}V_{out}$ is negligible.

ii. h_{oe} of the transistor is very small i.e., the output resistance $1/h_{oe}$ is very large and, therefore $1/h_{oe}$ is omitted from the equivalent circuit.

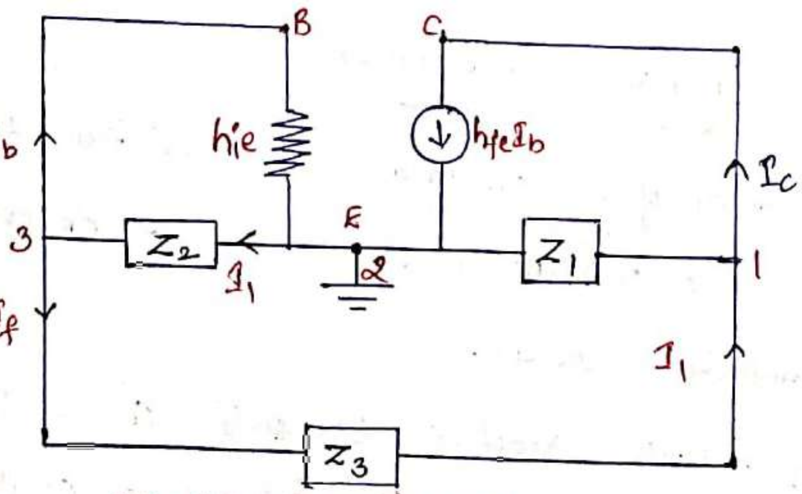


FIG: EQUIVALENT CIRCUIT

* Let us determine the load impedance between output terminals 1 and 2. Here Z_2 and h_{ie} are in parallel and their resultant impedance is in series with impedance Z_3.

26

i.e., $Z_3 + (Z_2 \| h_{ie}) \rightarrow ①$

* The equivalent impedance of eq/① is in parallel with impedance Z_1.

* Thus load impedance between output terminals is given as,

$$Z_L = Z_1 \| \left[Z_3 + (Z_2 \| h_{ie}) \right]$$

$$= Z_1 \| \left[Z_3 + \frac{Z_2 h_{ie}}{Z_2 + h_{ie}} \right]$$

$$= Z_1 \| \left[\frac{Z_3(Z_2 + h_{ie}) + Z_2 h_{ie}}{Z_2 + h_{ie}} \right]$$

$$= \frac{Z_1 * \left[\frac{Z_3(Z_2 + h_{ie}) + Z_2 h_{ie}}{Z_2 + h_{ie}} \right]}{Z_1 + \frac{Z_3(Z_2 + h_{ie}) + Z_2 h_{ie}}{Z_2 + h_{ie}}}$$

$$= \frac{Z_1 Z_3 (Z_2 + h_{ie}) + Z_1 Z_2 h_{ie} / \cancel{Z_2 + h_{ie}}}{Z_1 (Z_2 + h_{ie}) + Z_3 (Z_2 + h_{ie}) + Z_2 h_{ie} / \cancel{Z_2 + h_{ie}}}$$

$$= \frac{Z_1 Z_2 Z_3 + Z_1 Z_3 h_{ie} + Z_1 Z_2 h_{ie}}{Z_1 Z_2 + Z_1 h_{ie} + Z_2 Z_3 + Z_3 h_{ie} + Z_2 h_{ie}}$$

$$Z_L = \frac{Z_1 \left[Z_2 Z_3 + h_{ie}(Z_2 + Z_3) \right]}{h_{ie}\left[Z_1 + Z_2 + Z_3 \right] + Z_1 Z_2 + Z_2 Z_3}$$

* The Voltage gain of a CE amplifier without feedback is given as

$$A = \frac{-I_L}{I_b} \quad + \frac{V_2}{V_1} \qquad \left(\because V_2 = -I_c Z_L = -h_{fe} I_b Z_L \quad (\because I_c = h_{fe} I_b)\right)$$

$$V_1 = h_{ie} I_b$$

(2)

$$= \frac{-h_{fe} \cancel{I_b} Z_L}{h_{ie} \cancel{I_b}}$$

$$A = \frac{-h_{fe} Z_L}{h_{ie}}$$

* The output voltage between terminals 1 and 2 is given as,

$$V_{out} = [Z_3 + (Z_2 \| h_{ie})] I_1$$

$$= \left[Z_3 + \frac{Z_2 h_{ie}}{Z_2 + h_{ie}}\right] I_1$$

$$= \left[\frac{Z_3(Z_2 + h_{ie}) + Z_2 h_{ie}}{Z_2 + h_{ie}}\right] I_1$$

$$= \left[\frac{Z_3 Z_2 + Z_3 h_{ie} + Z_2 h_{ie}}{Z_2 + h_{ie}}\right] I_1$$

$$= \left[\frac{Z_2 Z_3 + h_{ie}(Z_2 + Z_3)}{Z_2 + h_{ie}}\right] I_1$$

* The Voltage feedback to the input terminals 2 and 3 is given as

$$V_f = \frac{Z_2 h_{ie}}{Z_2 + h_{ie}} * I_1$$

So, feedback fraction, $\beta = \frac{V_f}{V_{out}}$

$$\beta = \frac{\dfrac{Z_2 h_{ie}}{\cancel{Z_2 + h_{ie}}} * I_1}{\dfrac{h_{ie}(Z_2+Z_3)+Z_2 Z_3}{\cancel{Z_2 + h_{ie}}} * I_1}$$

(28)

I_B

$$\beta = \frac{Z_2 h_{ie}}{Z_2 Z_3 + h_{ie}(Z_2+Z_3)}$$

* Applying the criterion of oscillation i.e., $A\beta = 1$, we have

$$\frac{-h_{fe} * Z_L}{\cancel{h_{ie}}} * \frac{Z_2 \cancel{h_{ie}}}{Z_2 Z_3 + h_{ie}(Z_2+Z_3)} = 1$$

Substituting the value of Z_L, then

$$-h_{fe}\left[\frac{Z_1[h_{ie}(Z_2+Z_3)+Z_2 Z_3]}{h_{ie}(Z_1+Z_2+Z_3)+Z_1 Z_2+Z_2 Z_3}\right] * \frac{Z_2}{Z_2 Z_3 + h_{ie}(Z_2+Z_3)} = 1$$

$$\frac{h_{fe} Z_1 \cancel{[h_{ie}(Z_2+Z_3)+Z_2 Z_3]}}{h_{ie}(Z_1+Z_2+Z_3)+Z_1 Z_2+Z_2 Z_3} * \left[\frac{Z_2}{\cancel{Z_2 Z_3 + h_{ie}(Z_2+Z_3)}}\right] = -1$$

$$\frac{h_{fe} Z_1 Z_2}{h_{ie}(Z_1+Z_2+Z_3)+Z_1 Z_2+Z_2 Z_3} = -1$$

$$h_{fe} Z_1 Z_2 = -[h_{ie}(Z_1+Z_2+Z_3) + Z_1 Z_2 + Z_2 Z_3]$$

$$h_{fe} Z_1 Z_2 + h_{ie}(Z_1+Z_2+Z_3) + Z_1 Z_2 + Z_2 Z_3 = 0$$

$$h_{ie}(Z_1+Z_2+Z_3) + Z_1 Z_2(1+h_{fe}) + Z_2 Z_3 = 0 \longrightarrow (2)$$

* This equation 2 is the general equation for the oscillator.

HARTLEY OSCILLATOR:

* The oscillator in which Z_1 and Z_2 are inductors and Z_3 is a Capacitor then it is called Hartley oscillator.

* The following figure represents the Hartley oscillator using BJT.

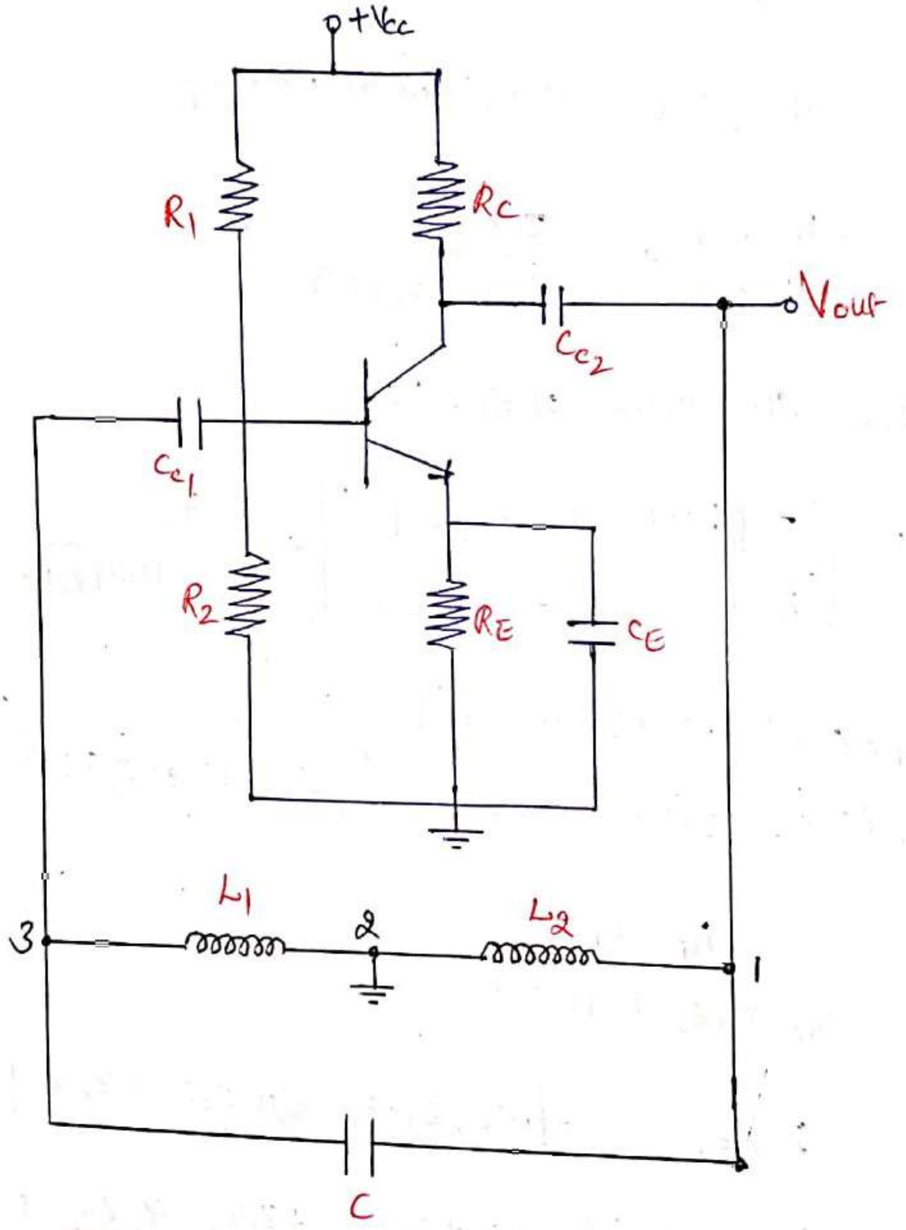


FIG: HARTLEY OSCILLATOR USING BJT

* Resistors R_1, R_2 and R_E provides the necessary dc bias to the transistor. C_E is a bypass capacitor. C_{c_1} and C_{c_2} are Coupling Capacitors.

* The feedback network consisting of inductors L_1 and L_2 and Capacitor C determines the frequency of the oscillator. 30

* When the supply voltage $+V_{cc}$ is switched ON, a transient Current is produced in the tank Circuit.

* The oscillatory current in the tank Circuit produces ac voltages across L_1 and L_2. As terminal 2 is grounded, it is at zero potential

* If terminal 3 is at positive potential with respect to 2 at any instant, terminal 1 will be at a negative potential w.r.t ~~termi~~ terminal 3 at the same instant.

* Thus the phase difference between the terminals 1 and 3 is always 180°. In the CE mode, the transistor provides the phase difference of 180° between the input and output, therefore, the total phase shift 360°.

* Thus at the frequency determined for the tank circuit, the necessary Condition for sustained oscillations is satisfied. If the feedback is adjusted so that the loop gain $A\beta = 1$, the circuit acts as an oscillator

ANALYSIS:

* In the Hartley oscillator, Z_1 and Z_2 are inductive reactances and Z_3 is the Capacitive reactance. Suppose M is the mutual inductance between the inductors, then,

$$Z_1 = j\omega L_1 + j\omega M$$

$$Z_2 = j\omega L_2 + j\omega M$$

$$Z_3 = \frac{1}{j\omega C} = \frac{-j}{\omega C}$$

The general equation for LC oscillator is,

$$h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2(1+h_{fe})+Z_2Z_3=0 \rightarrow (1).$$

Substituting Z_1, Z_2, Z_3 values in eq(1)

$$h_{ie}\left[j\omega L_1+j\omega m+j\omega L_2+j\omega m-\frac{j}{\omega C}\right]+[j\omega L_1+j\omega m][j\omega L_2+j\omega M](1+h_{fe})$$
$$+(j\omega L_2+j\omega M)\left(\cancel{j\omega L}\ \frac{-j}{\omega C}\right)=0$$

$$jh_{ie}\omega\left[L_1+M+L_2+M-\frac{1}{\omega^2 C}\right]+\left[-\omega^2L_1L_2-\omega^2L_1M-\omega^2L_2M-\omega^2M^2\right]$$
$$(1+h_{fe})+\cancel{\frac{\omega L_1}{\omega C}+\frac{\omega M}{\omega C}}=0$$
$$+\frac{L_2}{\omega^2 C}+\frac{M}{\omega^2 C}$$

$$j\omega h_{ie}\left[L_1+L_2+2M-\frac{1}{\omega^2C}\right]-\omega^2\left[L_1L_2+L_1M+L_2M+\cancel{\omega^2}M^2\right](1+h_{fe})$$
$$\frac{L_2}{C}+\frac{M}{C}=0 \quad \frac{1}{\omega^2C}[L_2+M]=0$$

Equating imaginary term to zero

$$\omega h_{ie}\left[L_1+L_2+2M-\frac{1}{\omega^2C}\right]=0$$

$$L_1+L_2+2M-\frac{1}{\omega^2C}=0$$

$$\frac{1}{\omega^2C}=L_1+L_2+2M$$

$$\omega^2C=\frac{1}{L_1+L_2+2M}. \qquad (\because L_{eq}=L_1+L_2+2M)$$

$$\omega^2C=\frac{1}{L_{eq}}$$

$$\omega^2=\frac{1}{L_{eq}C}$$

$$(2\pi f)^2=\frac{1}{L_{eq}C}$$

$$f^2 = \frac{1}{(2\pi)^2 L_{eq} C}$$

$$f = \frac{1}{\sqrt{(2\pi)^2 L_{eq} C}}$$

$$\boxed{f_0 = \frac{1}{2\pi\sqrt{L_{eq} C}}}$$ → This is the resonant frequency.

Equating real term to zero

$$-\omega^2[L_1L_2 + L_1M + L_2M + M^2]((1+h_{fe}) + \frac{1}{\omega^2 C}(L_2+M) = 0$$

$$-\omega^2(L_2+M)[\cancel{L_1} + \cancel{L_2} + \cancel{2M}\ (L_1+M)(1+h_{fe}) - \frac{1}{\omega^2 C}] = 0$$

$$(L_1+M)(1+h_{fe}) - \frac{1}{\omega^2 C} = 0$$

$$(L_1+M)(1+h_{fe}) = \frac{1}{\omega^2 C}$$

$$(1+h_{fe}) = \frac{1}{\omega^2 C(L_1+M)} \qquad \left(\because \frac{1}{\omega^2 C} = L_1 + L_2 + 2M\right)$$

$$1+h_{fe} = \frac{L_1+L_2+2M}{L_1+M} \Rightarrow 1+h_{fe} = \frac{(L_1+M)+(L_2+M)}{L_1+M}$$

$$= \frac{L_1+M}{L_1+M} + \frac{L_2+M}{L_1+M}$$

$$\cancel{1} + h_{fe} = \cancel{1} + \frac{L_2+M}{L_1+M}$$

$$\boxed{h_{fe} = \frac{L_2+M}{L_1+M}}$$

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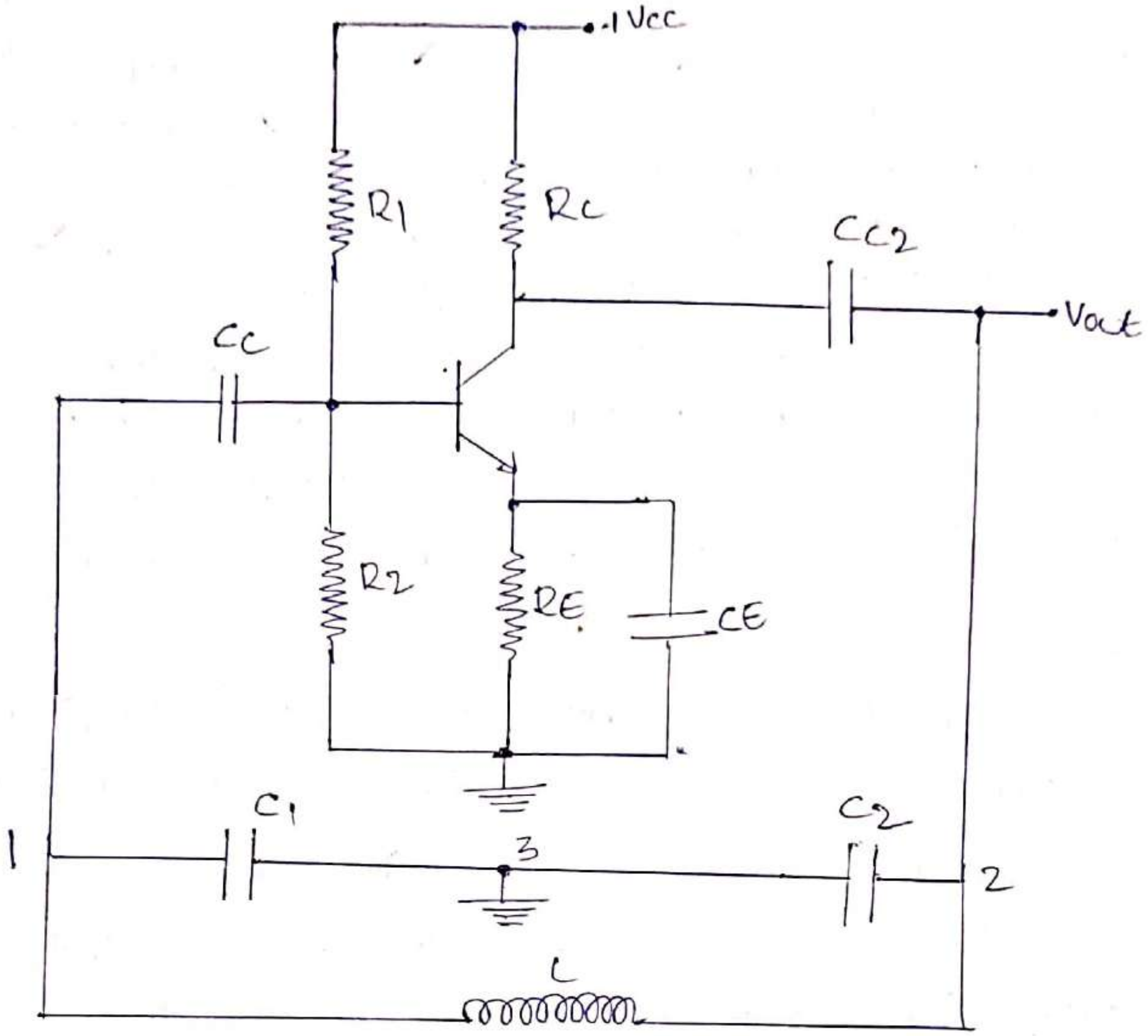
* As for other oscillator circuits, the loop gain must be greater than 1 to ensure that circuit oscillates 33

FET HARTLEY OSCILLATOR:

* Colpitts oscillator :-



Colpitts oscillator circuit diagram.

* In the Colpitt's oscillator shown in above figure, Z_1 and Z_2 are capacitors and Z_3 is an inductor.
* The Resistors R_1, R_2 and R_E provide the necessary d.c. bais to the transistor. C_E is a bypass capacitor.
* C_B and C_C are coupling capacitors. The feedback network consisting of capacitors C_1 and C_2 and an inductor L determines the frequncy of the oscillator.

When the supply voltage +Vcc is switched ON, a transient current is produced in the tank circuit and consequently, damped harmonic oscillations are set up in the circuit. The oscillator current in the tank circuit produces a.c voltages. arrows C_1 and C_2. As terminal 3 is earthed, it will be at zero potential. If terminal 1 is at a positive potential with respect to 3 at any instant, terminal 2 will be at a negative potential with respect with 3 at the same instant. Thus the phase disfference between the terminals 1 and 2 is always $180°$.

In the CE mode, the transistor provides the phase difference of $180°$ between the input and output. Therefore the total phase shift is $360°$. Thus at the frequncy determined for the tank circuit the necessary condition for sustained oscillations is satisfied. If the feedback is adjusted so that the loop gain $A\beta = 1$.

Analaysis :-

$$Z_1 = \frac{1}{j\omega C_1} = \frac{-j}{\omega C_1}$$

$$Z_2 = \frac{1}{j\omega C_2} = \frac{-j}{\omega C_2}$$

$$Z_3 = j\omega L.$$

$$h_{ie}[Z_1+Z_2+Z_3]+Z_1Z_2(1+h_{fe})+Z_2Z_3=0$$

$$=h_{ie}\left[\frac{-j}{\omega C_1}+\left(\frac{-j}{\omega C_2}\right)+j\omega L\right]+\left[\frac{-j}{\omega C_1}\cdot\frac{j}{\omega C_2}\right](1+h_{fe})$$

$$+\left(\frac{-j}{\omega C_2}\right)(j\omega L)=0$$

$$=h_{ie}\left[\frac{-j}{\omega C_1}-\frac{j}{\omega C_2}+j\omega L\right]+\left[\frac{-j}{\omega C_1}\right]\left[\frac{-j}{\omega C_2}\right](1+h_{fe})$$

$$+\left[\frac{-j}{\omega C_2}\right](j\omega L)=0$$

$$=-jh_{ie}\left[\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L\right]-\left[\frac{1}{\omega^2C_1C_2}(1+h_{fe})\right]\frac{L}{C_2}=0$$

equating imaginary term to zero.

$$-h_{fe}\left[\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L\right]=0$$

$$\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L=0$$

$$\omega L=\frac{1}{\omega C_1}+\frac{1}{\omega C_2}$$

$$\omega L=\frac{\omega C_2+\omega C_1}{\omega C_1\omega C_2}$$

$$\omega L=\frac{\omega C_2+\omega C_1}{\omega^2C_1C_2}$$

$$\omega L=\frac{\cancel{\omega}(C_1+C_2)}{\omega^{\cancel{2}}C_1C_2}$$

$$\omega L = \frac{C_2 + C_1}{\omega C_1 C_2}$$

$$\omega^2 L = \frac{C_2 + C_1}{C_1 C_2}$$

$$\omega^2 = \frac{C_1 + C_2}{(C_1 C_2) L}$$

$$\omega^2 = \frac{1}{C_{eq} L} \qquad \therefore C_{eq} = \frac{C_1 C_2}{C_1 + C_2}$$

$$(2\pi f)^2 = \frac{1}{C_{eq} L} \qquad \frac{1}{C_{eq}} = \frac{C_1 + C_2}{C_1 C_2}$$

$$\boxed{f_0 = \frac{1}{2\pi \sqrt{C_{eq} L}}}$$

$\therefore$ equating real term equal to zero.

$$-\frac{1}{\omega^2 C_1 C_2}(1 + h_{fe}) + \frac{L}{C_2} = 0$$

$$(1 + h_{fe}) \frac{1}{\omega^2 C_1 C_2} = \frac{L}{C_2}$$

$$(1 + h_{fe}) = \frac{L \omega^2 C_1 C_2}{C_2}$$

$$(1 + h_{fe}) = \left[\frac{C_1 + C_2}{\cancel{C_1} C_2}\right] L \cancel{C_1}$$

$$(1 + h_{fe}) = \frac{C_1 + C_2}{C_2}$$

$$1 + h_{fe} = \frac{C_1 + C_2}{C_2}$$

where;

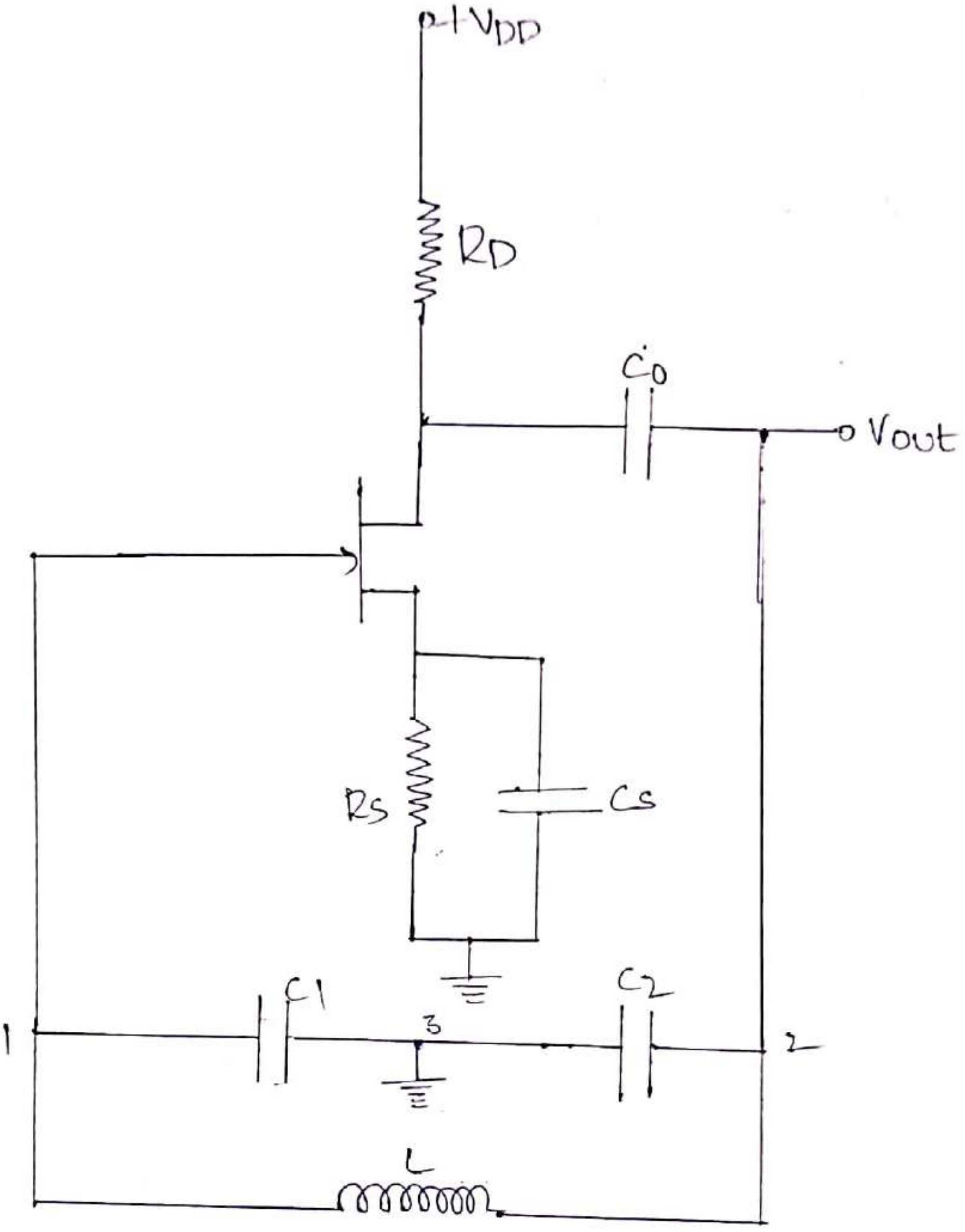
$$\omega^2 = \frac{C_1 + C_2}{C_1 C_2 \times L}$$

$$1 + h_{fe} = \frac{C_1}{C_2} + \frac{\cancel{C_2}}{\cancel{C_2}}$$

$$\cancel{1} + h_{fe} = \cancel{1} + \frac{C_1}{C_2}$$

$$\boxed{h_{fe} = \frac{C_1}{C_2}}$$

Colpitt's oscillator using FET :- (37)



Circuit diagram.

* crystal oscillators *.

A crystal oscillators, the usual electrical resonant circuit is placed by a mechanically vibrating crystal.

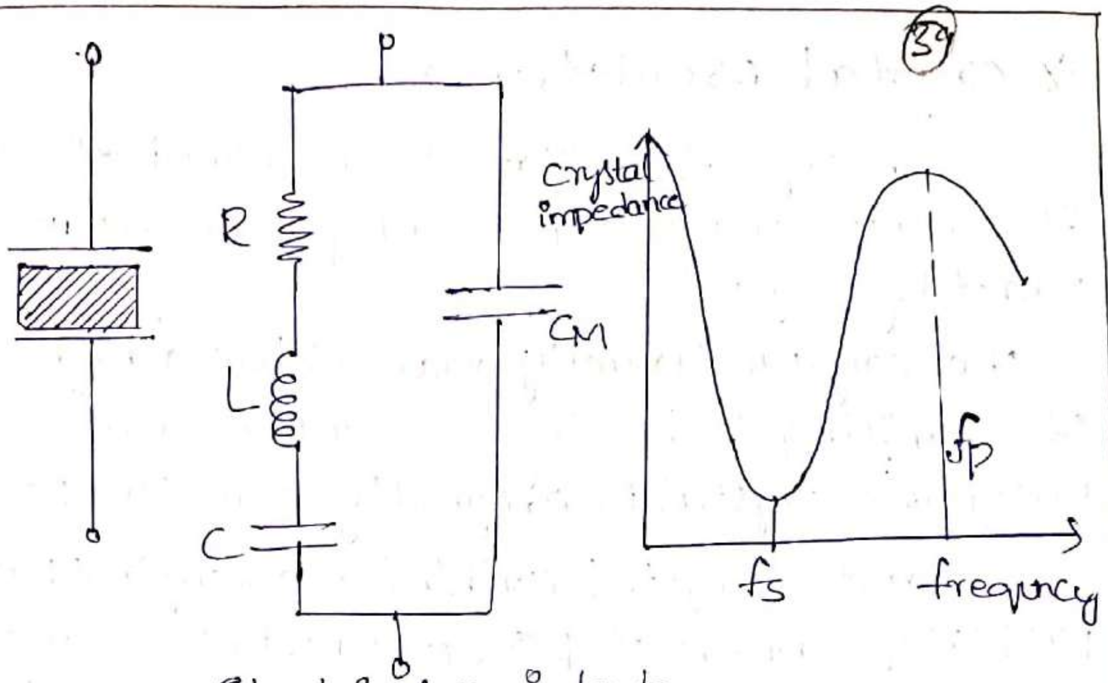
The crystal (usually quartz) has a high degree of stability in holding constant at whatever frequency the crystal is originally cut to operate.

A quartz crystal exhibits a very important property known as piezo-electric effect. when a mechanical pressure is applied across the forces of the crystal, a voltage propotional to the applied mechanical pressure appears across the crystal surfaces. the crystal is distorted by an amount propotional to the applied voltage. An alternating voltage applied to a crystal causes. it to vibrate as its natural frequncy.

Besides quartz, the other substances that exhibit the piezo-electric effect are "Rochelle salt" and "tournaline".

for use in electronic oscillators, the crystal is suitably cut and then mounted between two metal plates.

Although the crystal has electromechnical resosance but the crystal action can be represented by an electrical resonance circuit as shown in below figure.



Electrical equivalent circuit of a crystal.

The crystal actually behaves as a series R-L-C circuit in parallel, with CM where CM is the capacitive of the mounting Electrodes.

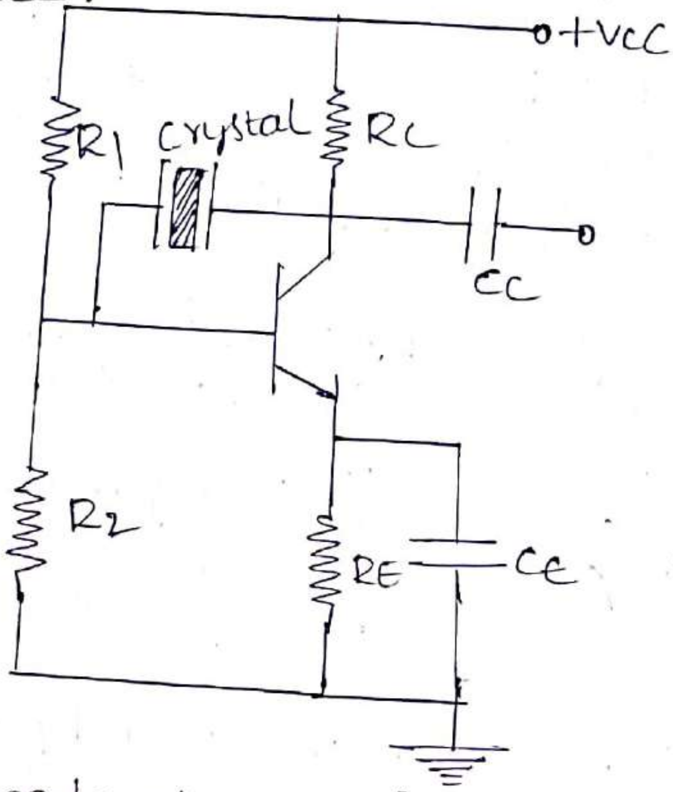
Because of CM, the crystal has two resonant frequency. One of these is the series resonant frequency fs at which $2\pi f L = \frac{1}{2\pi f C}$ and in this case crystal impedance is very low.

$$f_s = \frac{1}{2\pi\sqrt{LC}}$$

The other is parallel resonant frequency with is due to parrell resonace of capacitance of and resonance of the series circuit in. In this case crystal impedance is very high.

$$f_p = \frac{1}{2\pi}\sqrt{\frac{1+\frac{C}{C_M}}{LC}}$$

* oscillator with crystal operating in series resonance :-



In this mode of operation the crystal impedance is the smallest and the amount of positive feedback is the largest.

→ Resistors R_1 R_2 and R_E provides a voltage divider stabilized dc bias circuit.

The voltage feedback signal from the collector to the base is maximum when the crystal impedance is minimum (i.e; in series resonant mode). The coupling capacitor C_E has negligible impedance at this circuit operating frequency but blocks only dc between collector and base.

The circuit is generally called the "pierce crystal".

The resulting circuit frequency of the crystal. varations in supply voltage, transistor parameters, etc; have no Effect on the circuit operating frequency which is held stabilized by the crystal.

The frequency of vibration is inversly propotional to the thickness of the crystal.

$$f = \frac{P}{2l}\sqrt{\frac{Y}{\rho}}$$

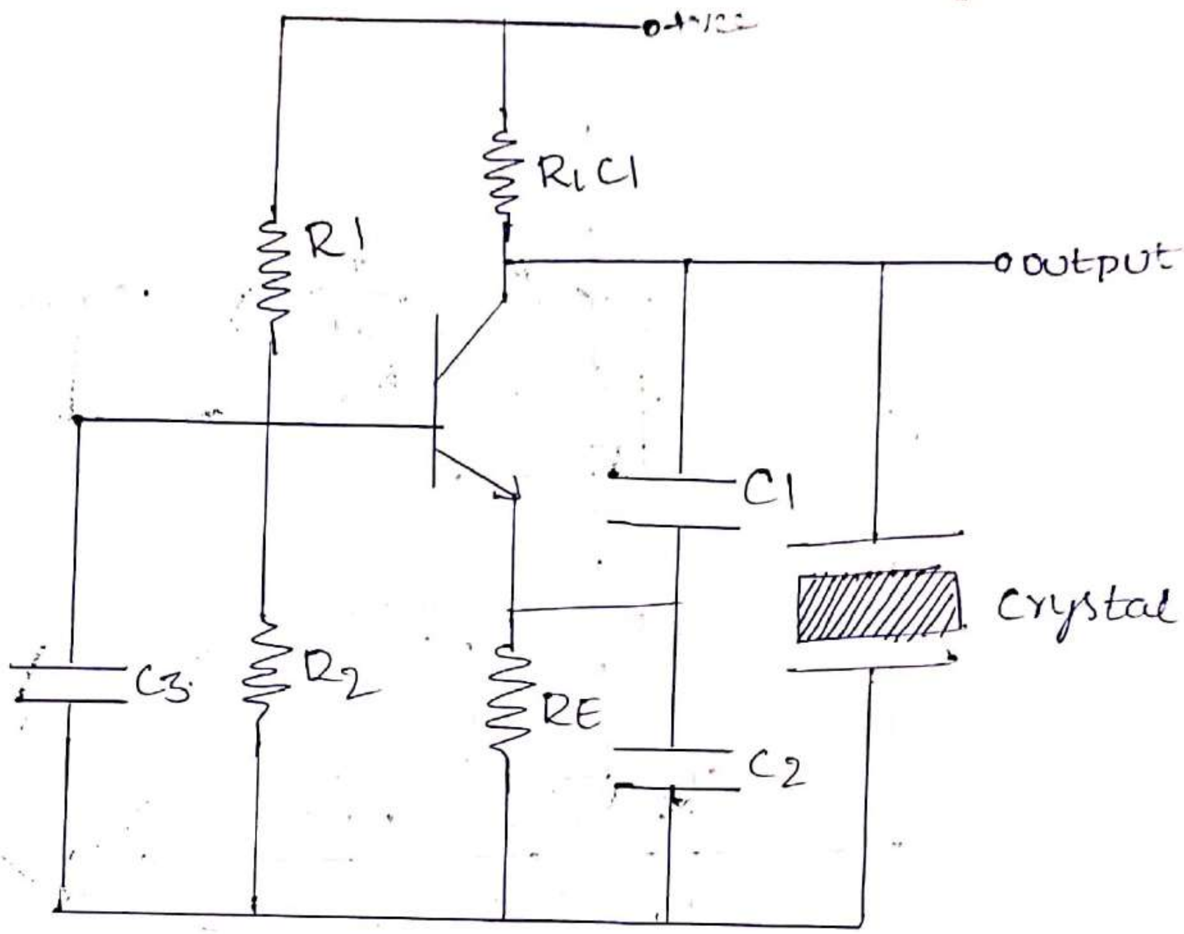
Y = young moduls.
ρ = density of material.
$P = 1, 2, 3 - - - -$

The crystal is suitably cut and polished to vibrate at a certain frequency and mounted between two metal plates.

* Oscillator with crystal operating in parallel Resonance :-

oscillator circuit with crystal operating in parallel resonance (a modified colpitt's oscillator circuit). Since the parallel-resonant impedance of a crystal is of a maximum value, C_1 and C_2 form a capacitive voltage divider which returns a portion of the output voltage divider which returns a pattern of the output voltage divider which returns a portion of the output voltage to the transistor emitter.

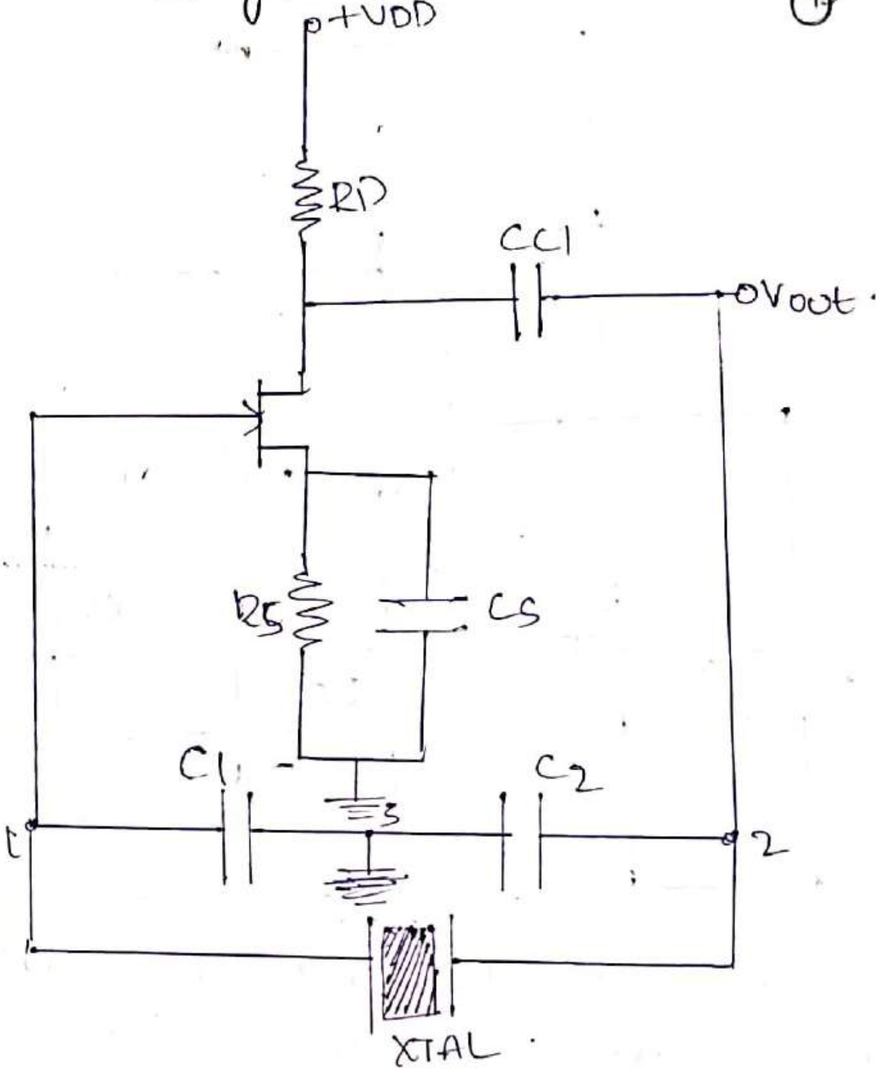
Circuit diagram:-



capacitor C_3 provides an ac short circuit across R_2 to ensure that the transistor base remains at a fixed voltage level. As the output voltage increases positives the emitter voltage also increases and since the base voltage is fixed, the base-emitter voltage is reduced.

The reduction is V_{BE} causes collector current I_C to diminishes and this is torn causes the collector voltage V_C to increases positively. Thus, the circuit is applying in our input, and a state of oscillation exists. The crystal in parallel with C_1 and C_2 permits maximum voltage feedback from the oscillator to emitter when its impedance is maximum.

Crystal oscillator using FET:-



* frequency Stability of oscillators:-

The frequency stability of an oscillator is a measure of its ability to maintain the required frequency as precisely as possible over as long a time interval as possible.

The accuracy of frequency callibration required may be any where between 10^{-2} and 10^{-10}. The main drawback in transistor oscillators is that the frequency of oscillation is not stable during a long time operation. The following are the factors which contribute to the change in frequency.

1. Due to change in temperature, the values of the frequency determining compounds, viz, resistors, inductors and capacitors charge.
2. Due to variation in the power supply, unstable transistors, parameters, change in climatic conditions and again.
3. The Effective resistance of the tank circuit is changed when the load is completed.
4. Due to variation in biasing conditions and loading condition.

The varation of frequency with temperature is given by,

$$S_{\omega,T} = \frac{\Delta\omega/\omega_0}{\Delta T/T_0}$$ ppmc (parts per million per °c).

ω_0 and T_0 are the desired frequency of oscillation and the rating temperature respectively. in the absence of automatic temperature control, the Effect of temperature on the resonant LC circuit can be reduced by selecting an inductance "L" with positive temperature coefficient and a capacitance "C" with negative temperature coefficent.

The loading Effect may be minimised if the oscillators is coupled to the load loosely or by a circuit with high input resistance and low output resistance properties. The frequency stability is defined as.

$$\boxed{S_\omega = \frac{d\theta}{d\omega}}$$

where $d\theta$ is the phase shift introduced for a small frequencies change in nominal frequency f_0, the circuit giving the largest value of $\frac{d\theta}{d\omega}$ has the more stable oscilator frequency, if the Q is infinite [an ideal inductors with zero series resistance]. this phase page changes abroptly from $-90°$ to $+90°$.

For tuned oscillators, S_ω is directly propotional to the 'Q' of a tuned circuit. A frequency stability of one part in can be achieved with 'LC' ckt for 'LC' oscillators, a tuned circuit must be lightly loaded to pressure high Q value.

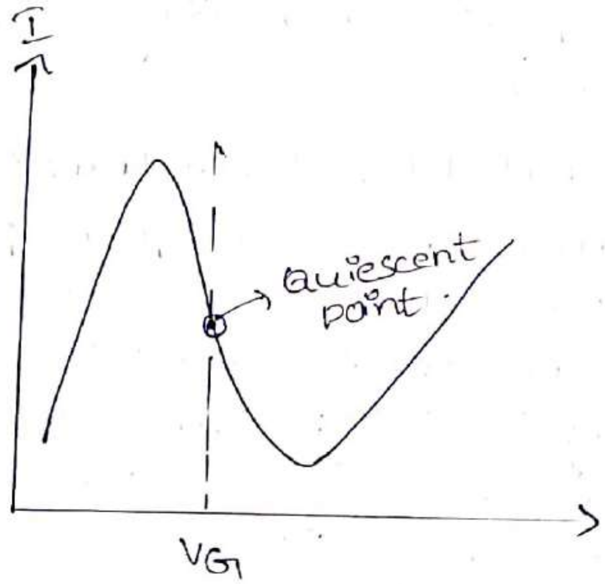
As piezo-electric crystals have high Q values of the order of 10^5, they can be used as parallel resonant circuits in oscillators to get very high frequency stability of "1 ppm" (parts per million).

* Amplitude stability of oscillators *.

An oscillators do not require positive feedback for their operation. if the positive resistance of "LC" tank circuit is cancelled by introducing the right amount of negative resistance across the tank circuit, then the steady oscillation can be maintained. The several devices such as dynatron, tranistron, thermistor UJT and tunnel diode that exhibit a region of negative resistance with in the V-I charaterstics, such devices operated in the negative resistance region are placed across a high Q for oscillation to occurs the negative resistance should be numerically less than the dynamic resistance of the tuned circuit.

In the case of "RC" circuit oscillators, the amplitude against variations due to fluctuation by aging of the transistant others components can be stabilised by replacing the resistor in bridge by senistors which are temperature dependent resistors. Thus the

Stability in amplitude of the RC oscillators can easily be maintained. (4)



V-I Charateristics of negative resistance oscillators.

Introduction :- Almost in all practical Amplifiers, Number of stage are Cascaded to Amplify the weak signals to a sufficient level to operate the output device (loud speaker.) ①

→ In such Amplifiers the function of first few stages is only to amplify the voltage but last stage is designed to provide maximum power to drive the output device (loud speaker.). This final stage is called power Amplifier stage.

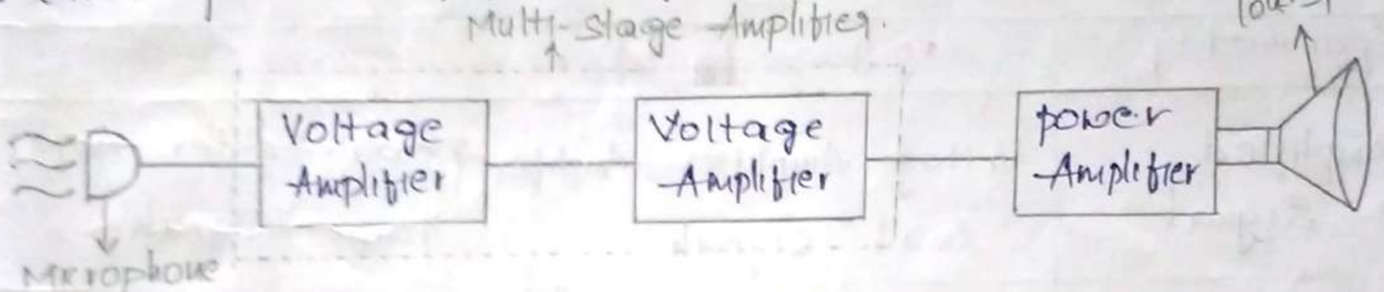


fig:- Public Address System.

→ Microphone Converts the sound to Electrical Signal and loud-speaker Converts Electrical Signal to Sound Signal again.

→ When a person speaks in microphone, it Converts Sound Signal into Electrical signal.

→ A Electrical Signal produced is of very low Voltage (a few milli volts)

→ If this Signal is fed to the speaker directly, it will not be in position to drive the Speaker.

→ Therefore voltage level of the signal is the first increased to sufficient level (a few volts) by passing it through a number of stages of voltage Amplifier.

→ This Amplified voltage signal is then fed to the final stage of multi stage Amplifier which is Capable to deliver the required power to drive the speaker.

→ The Speaker finally Converts the Electrical signal into Sound signal.

→ So, we Conclude that In final stage, we have to apply the power Amplifier to transfer maximum power (or) to deliver maximum power to the output device.

* Differences between Voltage Amplifier and power Amplifier:-

→ The main function of a voltage Amplifier is to amplifies the signals upto certain level.

→ The main function of power Amplifier is to deliver maximum power to the load (loud speaker.)

→ parameter	Voltage Amplifier	power Amplifier.
Amplified Signal	Voltage Amplifier Amplifies Small signals	Power Amplifier Amplifies Large Signals.
Output power	power output is low	Power output is high
output impedance	output impedance is low	Output impedance is Very low.
input impedance	input impedance is high	input impedance is low
Size of the Amplifier	Size of the voltage Amplifier is Small	Size of the power Amplifier is large.
Usage of Transistor	Normal transistor is used	power transistor is used.
output delivered to the load	unable to delivery maximum power to the load.	It delivers maximum power to the load.
Signals Handled	It handles Small signals	It handles large signals
Applications	used in voltage Amplifier	used in public adress System (Telephones.)

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* Classification of power Amplifier:-

We have different types of power-Amplifiers they are:

1. Class - A power Amplifier.
2. Class - B power Amplifier.
3. Class - C power Amplifier.
4. Class - AB power Amplifier.

(1.) Class- A power Amplifier:- The power Amplifier in which the operating point is adjusted as the Collector Current flows during whole Cycle of the input signal is known as class-A power Amplifier.

→ The Q-point is kept at the centre of the active region.

→ when an ac input signal is applied, the Collector Voltage and the Collector Current Varies simultaneously.

→ In class-A power Amplifier distortion is low because output signal is reproduced during full cycle of the input signal

→ Conduction Angle in class-A power Amplifier is 360°.

→ In class-A power Amplifier, efficiency is low.

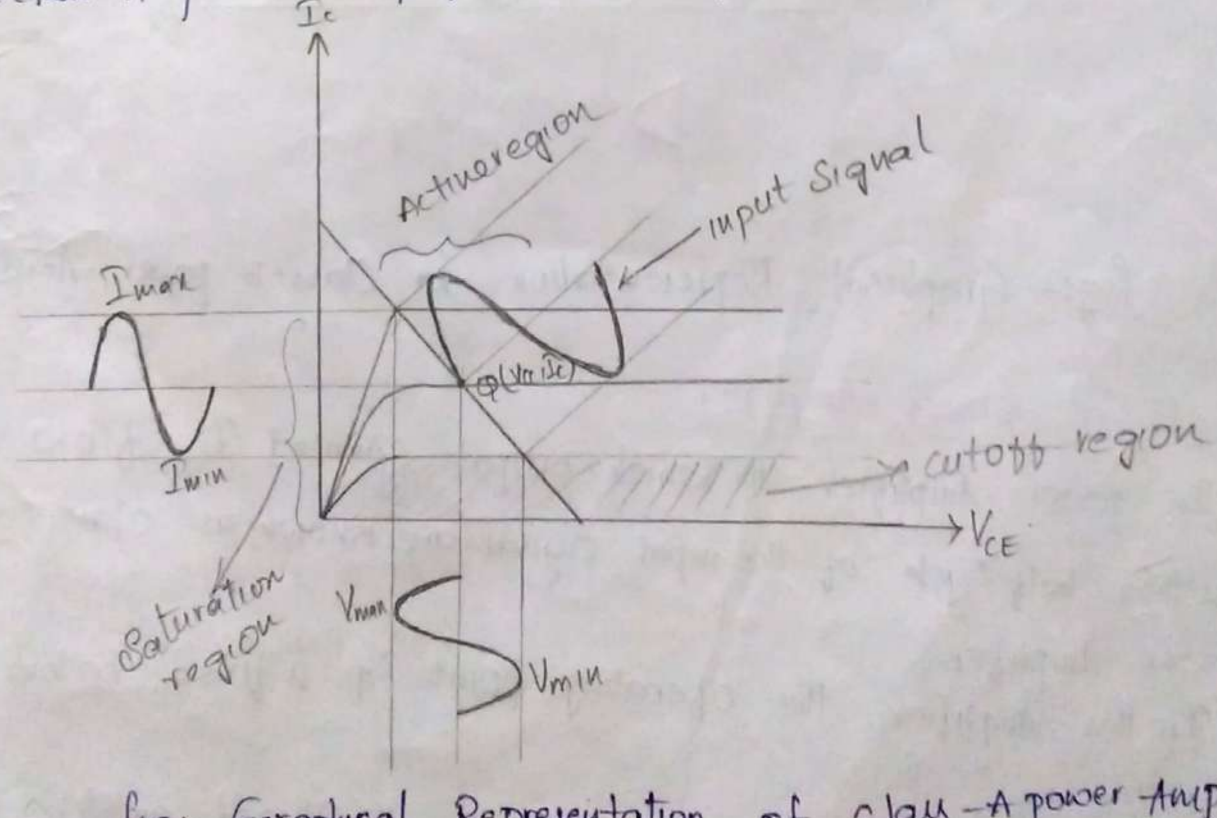

fig: Graphical Representation of class-A power Amplifier

(2.) Class-B power Amplifier:-

→ The power Amplifier in which the operating point is so adjusted that the Collector current flows during the positive half cycle of the input signal.

→ Conduction Angle in Class-B power Amplifier is 180°.

→ The operating point 'Q' is fixed near to x-axis.

i.e the cycle of the input signal is in active region and Negative cycle of the input signal is in cutoff region.

→ The Graphical representation of Class-B power Amplifier is shown in figure below.

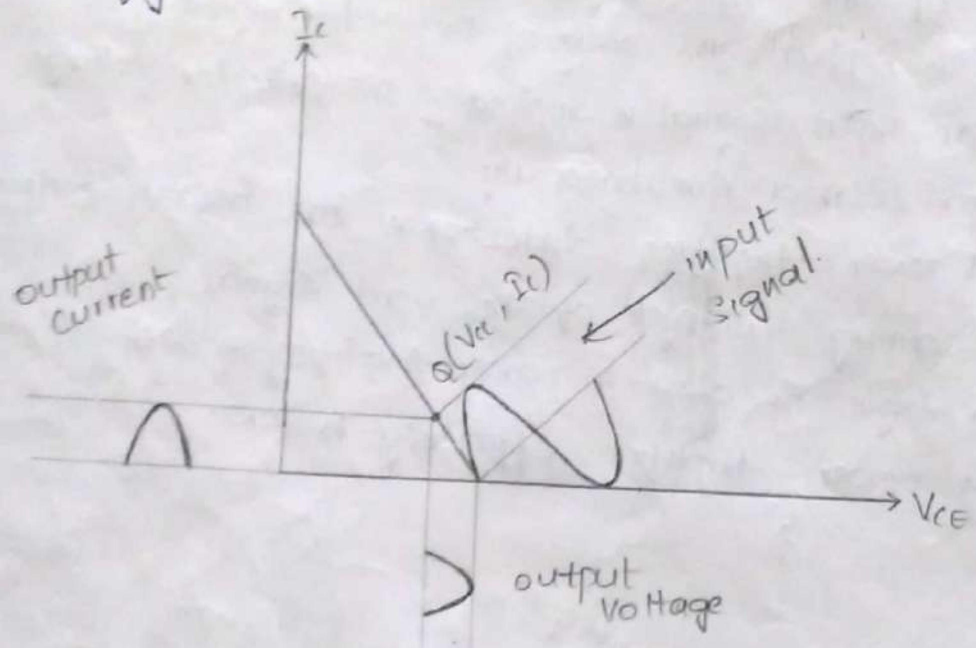

fig:- Graphical Representation for Class-B power Amplifier.

(3.) Class-C power Amplifier:-

→ In power Amplifier in which output current I_C flow for less than half cycle of the input signal are known as class-C power Amplifier.

→ In this Amplifier the operating point 'Q' is fixed below the x-axis.

→ class-C power Amplifiers are used in Tuned amplifiers to amplify the narrow band of frequencies.

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Conduction angle of Class-C power Amplifier is ~~180°~~. less than 180°.

The Graphical representation is shown below.

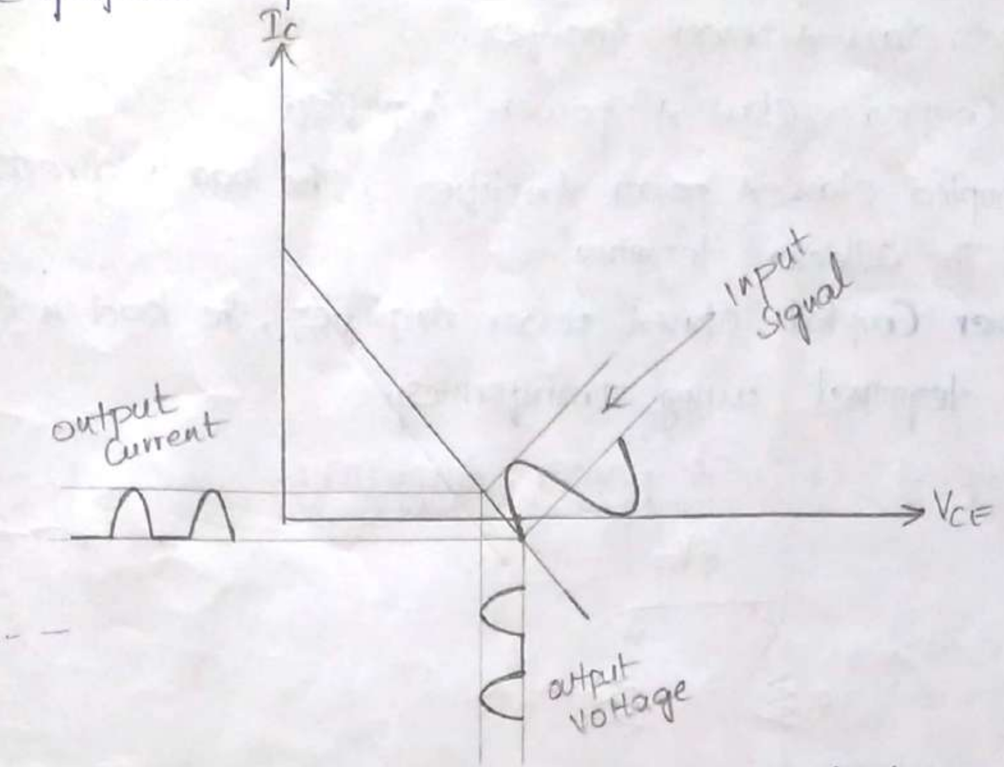

fig:- The Graphical representation of class-C power Amplifier.

4) Class-AB power Amplifier:-

→ The power Amplifier in which the output Current flows for more than half Cycle and less than full cycle of the input signal.

→ The Conduction Angle of Class AB power Amplifier is between Class-A and class-B power Amplifiers. Conduction angles (ie, $<360°$ & $>180°$)

→ The Graphical representation is shown in the graph below.

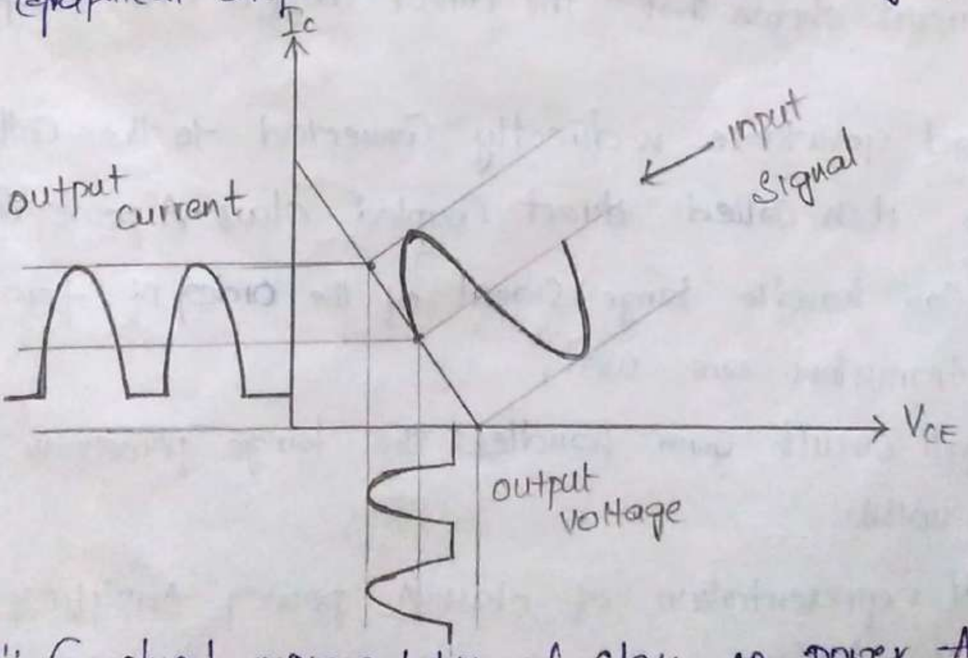

fig:- Graphical representation of Class-AB power Amplifier.

(1.) Class-A power Amplifier:-

The class-A power Amplifier is classified into two types

(a) Direct Coupled Class-A power Amplifier

(b) Transformer Coupled Class-A power Amplifier.

→ In direct Coupled Class-A power Amplifier, The load is direc[tly] Connected to the Collector terminal.

→ In Transformer Coupled Class-A power Amplifier, the load is Co[nnected] to the Collector terminal using Transformer.

(a.) Direct Coupled Class-A power Amplifier:-

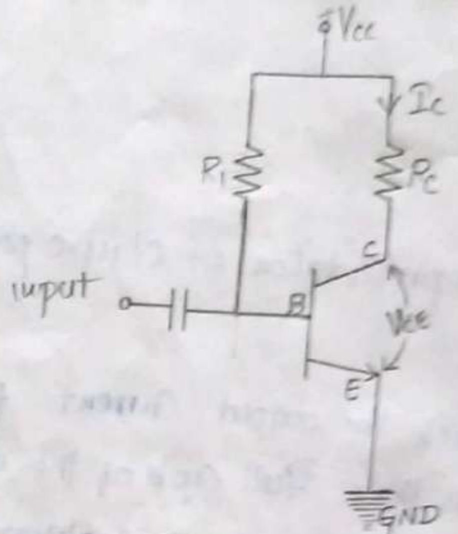

fig:- Direct Coupled Class-A power Amplifier.

DC-Operation :-

→ The above circuit shows that the direct Coupled Class-A power Amplifier.

→ Here, the load resistance is directly Connected to the Collector terminal. Hence it is called direct Coupled class-A power Amplifier

→ This circuit Can handle large Signal of the order of few volts, Hence power transistors are used,

→ So, the Overall circuit ~~gain~~ handles the large power in the range of few volts.

→ The graphical representation of class-A power Amplifier is shown in figure below.

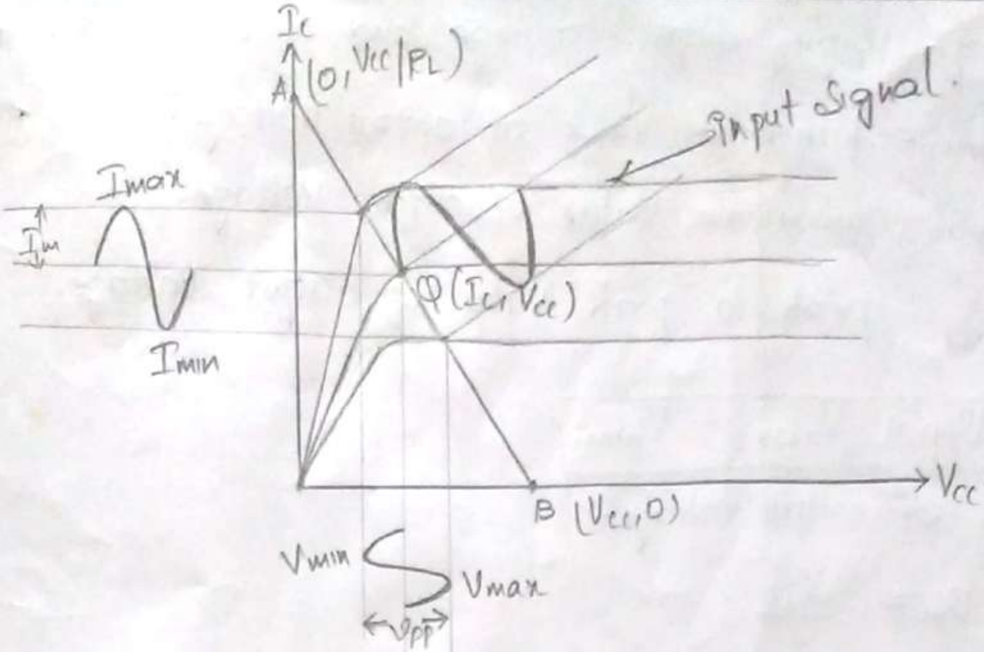

fig: Graphical representation of Direct Coupled Class - A power Amplifier.

→ The Collector Supply voltage V_{CC} under resistance R_B, decides the DC base bias Current.

→ Apply Kirchoff's voltage law to the output circuit, we get,

$$V_{CC} - I_C R_C - V_{CE} = 0$$

$$V_{CC} = I_C R_C + V_{CE}$$

$$\boxed{I_C = \frac{V_{CC} - V_{CE}}{R_C}}$$

→ The DC input power is provided by the Supply voltage.

→ The Collector Current (I_C) drawn the dc Current without AC input signal, Hence the dc input power is given by.

$$\boxed{P_{dc} = V_{CC} \times I_C}$$

AC-Operation :-

→ when an AC input signal is applied, the base Current Varies Sinusoidally, due to this Collector Current I_C and V_{CE} also Varies Sinusoidally,

→ The Varying output voltage and output Current delivers an AC power to the load.

→ From the graph Varying output voltage and output Current given by. V_{min} = minimum Value of output Voltage.

V_{max} = maximum Value of output Voltage.

V_{PP} = peak to peak Value of output Voltage.

$$\boxed{V_{PP} = V_{max} - V_{min}}$$

→ V_m = peak of output Voltage.

$$V_m = \frac{V_{PP}}{2}.$$

So, we get.

$$\boxed{V_m = \frac{V_{max} - V_{min}}{2}} \longrightarrow (1)$$

→ Similarly, the output Current Can be given by.

I_{min} = Minimum Value of output Current.

I_{max} = Maximum Value of output Current.

I_{PP} = Peak to peak Value of Output Current

$$\boxed{I_{PP} = I_{max} - I_{min}}$$

→ I_m = peak of Output Current

$$I_m = \frac{I_{PP}}{2}.$$

$$\boxed{I_M = \frac{I_{max} - I_{min}}{2}} \longrightarrow (2)$$

→ Hence, the RMS Value of Alternating output Voltage and output Current Can be given as $V_{rms} = \frac{V_m}{\sqrt{2}}$ and $I_{rms} = \frac{I_m}{\sqrt{2}}$.

∴ The RMS Value of AC power is given by.

$$P_{AC} = V_{rms} \times I_{rms}.$$

$$P_{AC} = \frac{V_m}{\sqrt{2}} \times \frac{I_m}{\sqrt{2}}$$

$$\boxed{\therefore P_{AC} = \frac{V_m I_m}{2}}$$

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from equations ① & ②

$$P_{AC} = \frac{\frac{(V_{max} - V_{min})}{2} \times \frac{(I_{max} - I_{min})}{2}}{2}$$

$$\boxed{\therefore P_{AC} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}$$

Efficiency:- Efficiency is defined as the ratio of AC output power to the DC ~~output~~ i/p power.

$$= \frac{\text{a.c power delivered to the load}}{\text{Total power delivered by d.c supply}}$$

$$\eta = \frac{P_{AC}}{P_{DC}}$$

$$\eta = \frac{\frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}{V_{cc}\, I_c}$$

$$\eta = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8\, V_{cc}\, I_c}$$

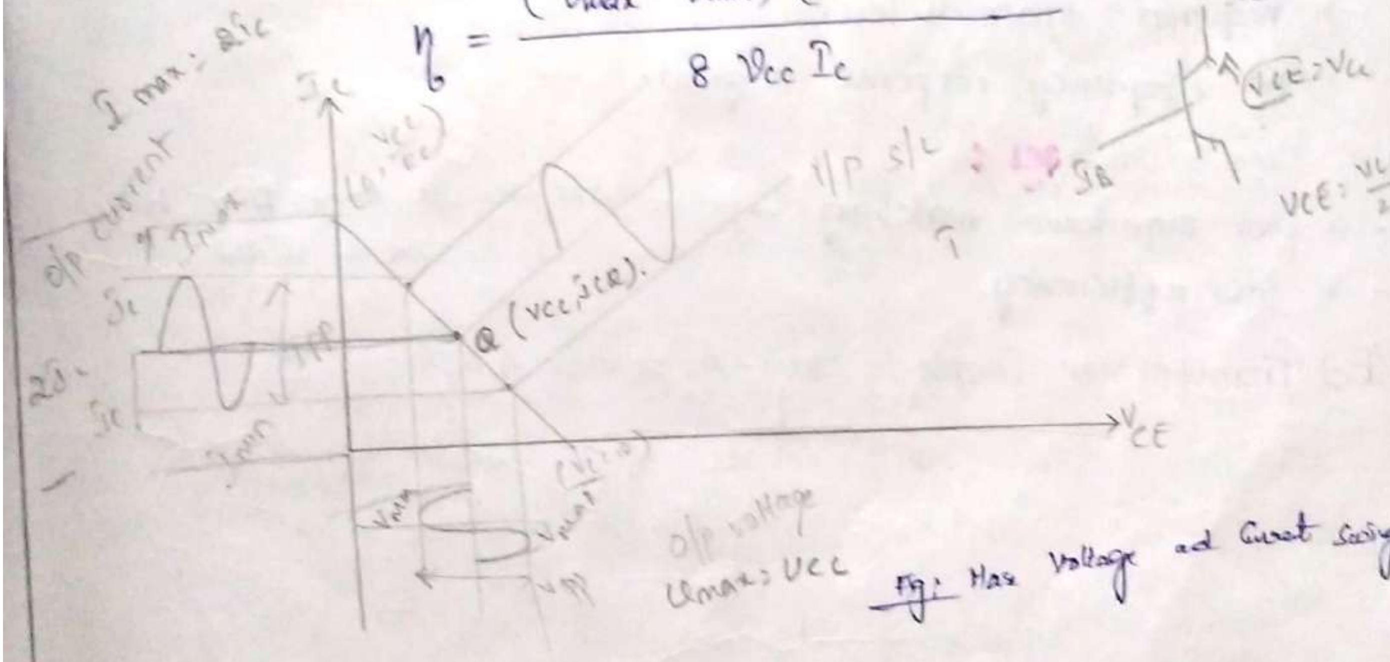


Fig: Max Voltage and Current Swings

we know that, $\eta = \frac{P_{AC}}{P_{DC}} \times 100$

from the above Graph $V_{max} = V_{cc}$ and $I_{max} = 2I_c$

Now,

$$\eta = \frac{\frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}{V_{cc} \times I_c} \times 100$$

4 ⑤

from equations ① & ②

$$P_{AC} = \frac{\frac{(V_{max} - V_{min})}{2} \times \frac{(I_{max} - I_{min})}{2}}{2}$$

$$\boxed{\therefore P_{AC} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}$$

Max. **Efficiency:-** Efficiency is defined as the ratio of AC output power to the DC ~~output~~ i/p power.

$$= \frac{\text{a.c power delivered to the load}}{\text{Total power delivered by d.c Supply}}$$

$$\eta = \frac{P_{AC}}{P_{DC}}$$

$$\eta = \frac{\frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}{V_{CC}\, I_C}$$

$$\eta = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8\, V_{CC}\, I_C}$$

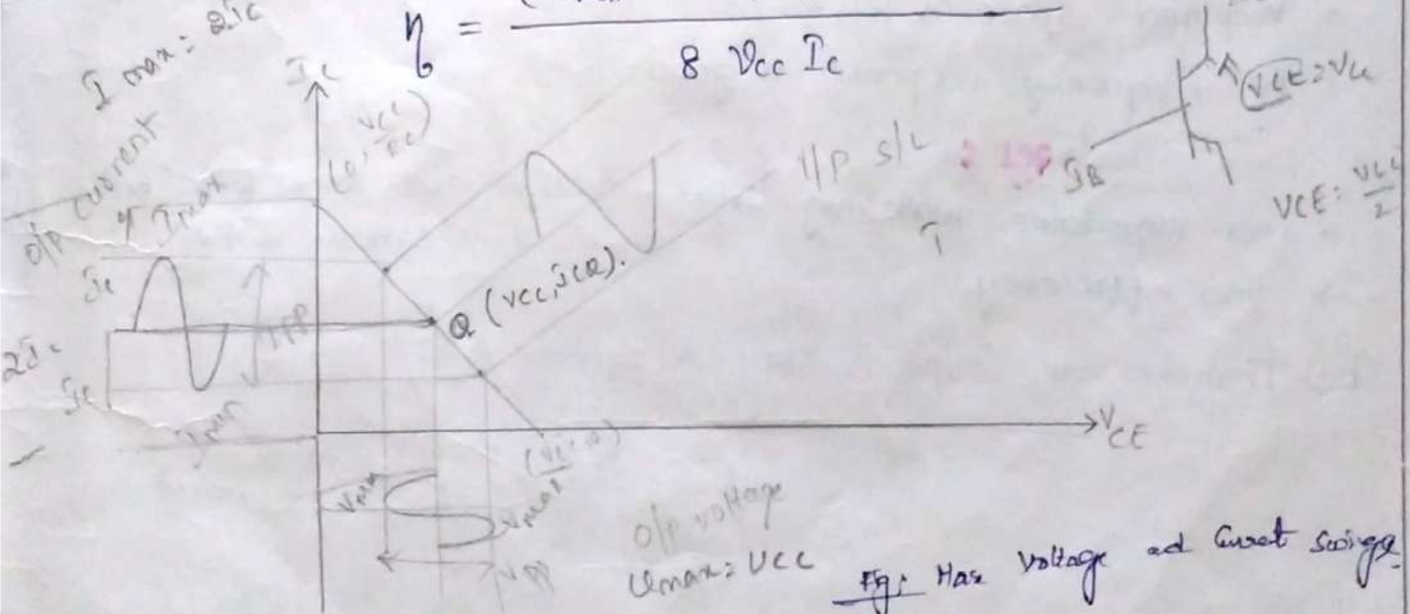


Fig: Max voltage and Current Swings.

we know that, $\eta = \frac{P_{AC}}{P_{DC}} \times 100$

from the above Graph $V_{max} = V_{CC}$ and $I_{max} = 2 \cdot I_C$

Now,

$$\eta = \frac{\frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}{V_{CC} \times I_C} \times 100$$

$$= \frac{(V_{cc}-0)(2I_c-0)}{8V_{cc} \times I_c} \times 100$$

$$= \frac{2V_{cc}I_c}{\frac{8}{4}V_{cc}I_c} \times 100$$

$$= \frac{1}{4} \times 100$$

$$= 0.25 \times 100$$

$$\boxed{\therefore \eta = 25\%}$$

* **Advantages:-**

→ Simple to design.
→ Less number of Components are required.
→ Cost is low.
→ required Space is less.
→ The frequency response is good.

* **Dis-advantages:-**

→ poor impedance matching. → power dissipation is more, large heat sinks are essential.
→ low efficiency.

(b.) Transformer Coupled class-A power Amplifier:-

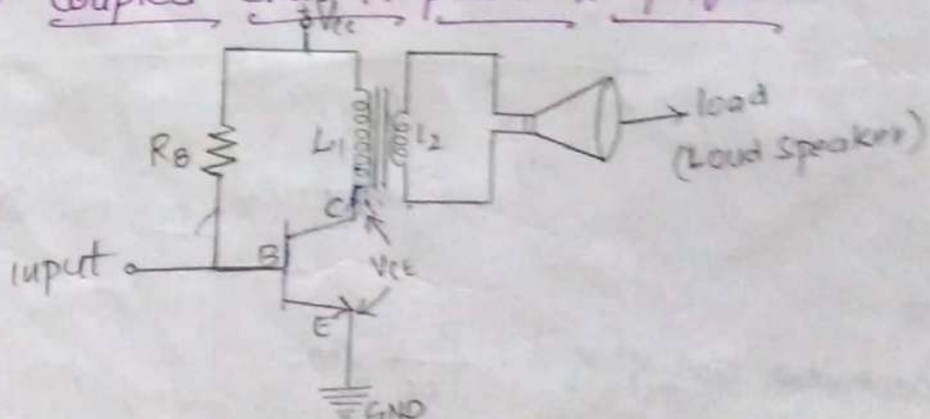
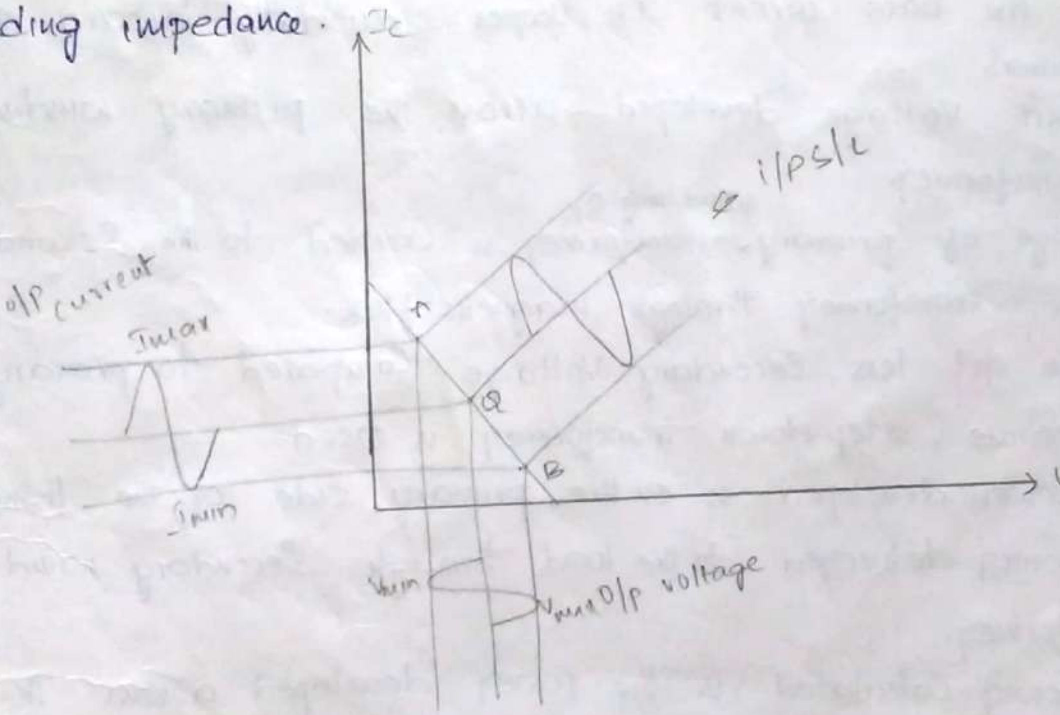


fig:- Circuit diagram for Transformer Coupled class-A power Amplifier.

→ If maximum power is transferred to the load, the impedance matching is necessary.

→ Impedance matching is poor in Case of direct Coupled class-A power Amplifier. because, the loud Speaker resistance (4-Ω to 16-Ω)

less than the output impedance of the direct Coupled class-A (2)
wer Amplifier. To Overcome this problem by using Transformer
Coupled Class-A power Amplifier.
The circuit diagram for Transformer Coupled Class-A power Amplifier
is shown in above figure.
→ Here the Transformer is directly Coupled to (or) Connected to the
Collector terminal.
→ In the above figure we use step down Transformer, in that Secondary
Voltage is less than the primary voltage. So, high voltage side has
always high impedance and the low voltage side has always low impedance
→ So, the secondary winding impedance is less than the primary
winding impedance.



* DC - Operation:-

→ it is assumed that the winding resistance is '0Ω'. because,
no resistor is connected between Supply Voltage and the Collector
terminal So, no voltage drop across the primary winding of
the Transformer.
→ The slope of the dc load line is reciprocal of the dc resistance.
→ In this circuit we have zero resistance in the winding. So, the

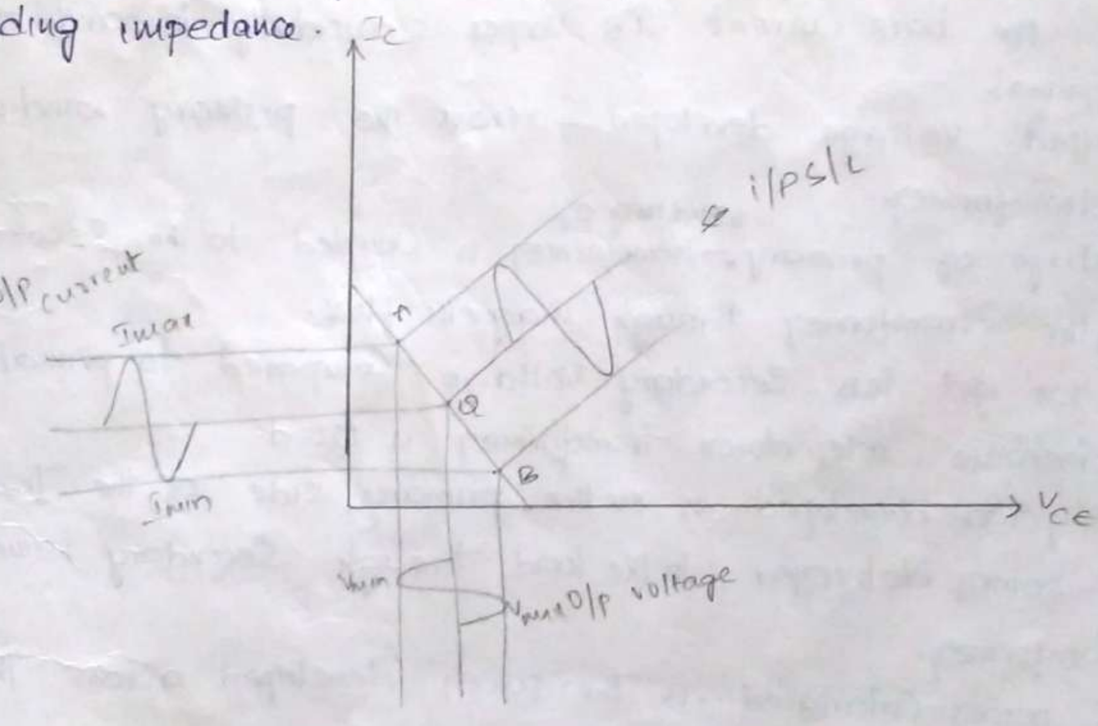
less than the output impedance of the direct Coupled class-A power Amplifier. To Overcome this problem by using Transformer Coupled Class-A power-Amplifier.

The circuit diagram for Transformer Coupled Class-A power Amplifier is shown in above figure.

→ Here the Transformer is directly Coupled to (or) Connected to the Collector terminal.

→ In the above figure we use, step down Transformer, in that Secondary Voltage is less than the primary voltage. So, high voltage side has always high impedance and the low voltage side has always low impedance

→ So, the Secondary winding impedance is less than the primary winding impedance.



* DC - Operation :-

→ it is assumed that the winding resistance is '0Ω'. because, no resistor is connected between Supply Voltage and the Collector terminal so, no voltage drop across the primary winding of the Transformer.

→ The slope of the dc load line is reciprocal of the dc resistance.

→ In this circuit we have zero resistance in the winding. So, the slope of the dc load line is ideally '∞'

→ This tells that dc load line in ideal condition is a very straight line.

→ Apply Kirchhoff's voltage law to the output circuit.

$$V_{CC} - V_{CE} = 0$$

$$\boxed{V_{CE} = V_{CC}}$$

→ The dc input power is given by.

$$\boxed{P_{dc} = V_{CC} \times I_C}$$

* AC-Operation:-

→ When an AC input signal is applied across the base of the transistor, the base current I_B varies sinusoidally towards the base terminal.

→ The output voltage developed across the primary winding of the transformer.

→ The voltage of primary winding of transformer is coupled to the secondary winding of the transformer through magnetic flux.

→ Here we get less secondary voltage compared to primary voltage because, step down transformer is used.

→ The AC power developed is on the primary side of the transformer and this power delivered to the load through secondary winding of the transformer.

→ The AC power calculated, is the power developed across the primary winding the transformer.

→ Assuming the ideal transformer, the power delivered to the load on the secondary winding is same as that of power developed across the primary.

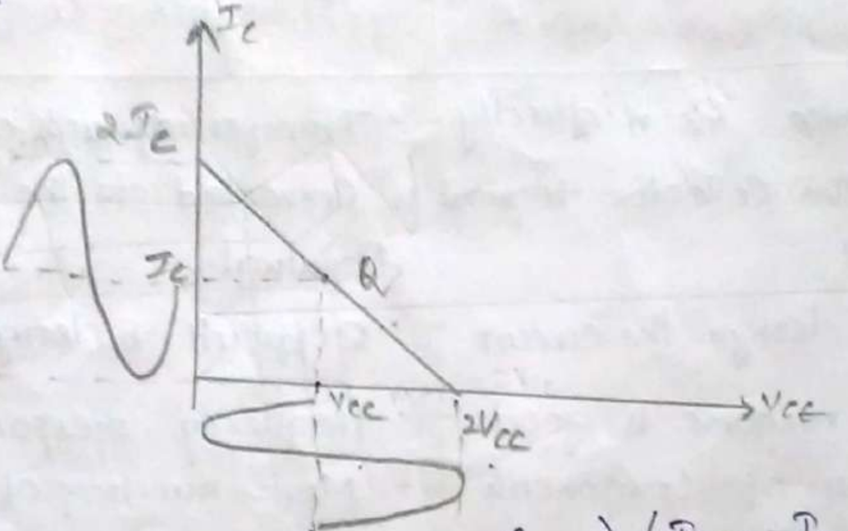
→ The AC power is given by

$$\boxed{\therefore P_{AC} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}$$

Efficiency:- The efficiency can be defined as the ratio of AC output power to the DC ~~output~~ Input power.

$$\therefore \eta = \frac{P_{AC}}{P_{DC}} \times 100$$

→ The maximum efficiency can be taken from maximum vary output voltage and output current is shown in the graph below.



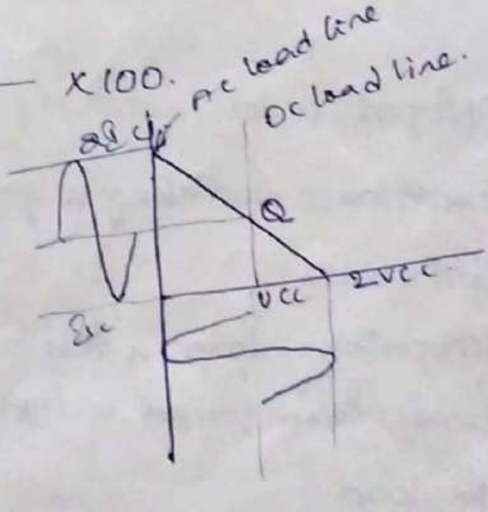
$$\therefore \eta = \frac{\dfrac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}{V_{cc} \times I_c} \times 100$$

$$\eta = \frac{(2V_{cc} - 0)(2I_c - 0)}{8 V_{cc} \times I_c} \times 100$$

$$= \frac{4 V_{cc} I_c}{8 V_{cc} I_c} \times 100$$

$$= 0.5 \times 100$$

$$\boxed{\therefore \eta = 50\%}$$



* Advantages:-

→ Efficiency is high when compare to direct coupled class-A power amplifier.

→ Impedance matching is obtained. , → power dissipation is less

* Dis-advantages:-

→ Difficult to design.

→ More number of components are required to design.

→ Cost is high
→ space required is also high.
→ The frequency response is less.

* Differences between direct coupled class-A power Amplifier and Transformer coupled class-A power Amplifier:-

Direct Couple class-A p. A	Transformer Coupled class-A p. A
→ Load Resistance R_C is directly connected to the collector terminal	→ Transformer acts as a load which is connected to the collector terminal.
→ Simple to design the circuit	→ Difficult to design the circuit
→ frequency response is good.	→ frequency response is less.
→ Less number of components are required to design the circuit.	→ More number of components are required to design the circuit.
→ Efficiency is less (25 %)	→ Efficiency is high when compared to Direct Coupled Class A power Amplifier (50 %).
→ Output is less	→ Output is high.
→ Impedance matching is poor.	→ Impedance matching is high.
→ Cost is less.	→ Cost is high.
→ Occupation space is less	→ Occupation space is high.
→ power transferred is less to the load.	→ Maximum power is transferred to the load.
→ Here, $V_{max} = V_{cc}$ (maximum Voltage is same as that of supplied voltage.)	→ Here, $V_{max} = 2V_{cc}$ (Maximum Voltage is two times of supplied voltage.)
→ Here, $I_{max} = 2I_c$ (Maximum Current is 2 times that of Collector Current.)	→ Here, $I_{max} = 2I_c$ (Maximum Current is 2 times to that of Collector Current.)

Class-B power Amplifier :-

• Class-B power Amplifier is defined as the output signal flows for the half Cycle (or) 180° of the input signal.

→ In Class-B power Amplifier, for the Negative half Cycle the transistor will be in off state. So that output Current should be zero for Negative half Cycle. So we get distortion during the Negative half Cycle.

→ To avoid this distortion we go for push-pull class-B Amplifier.

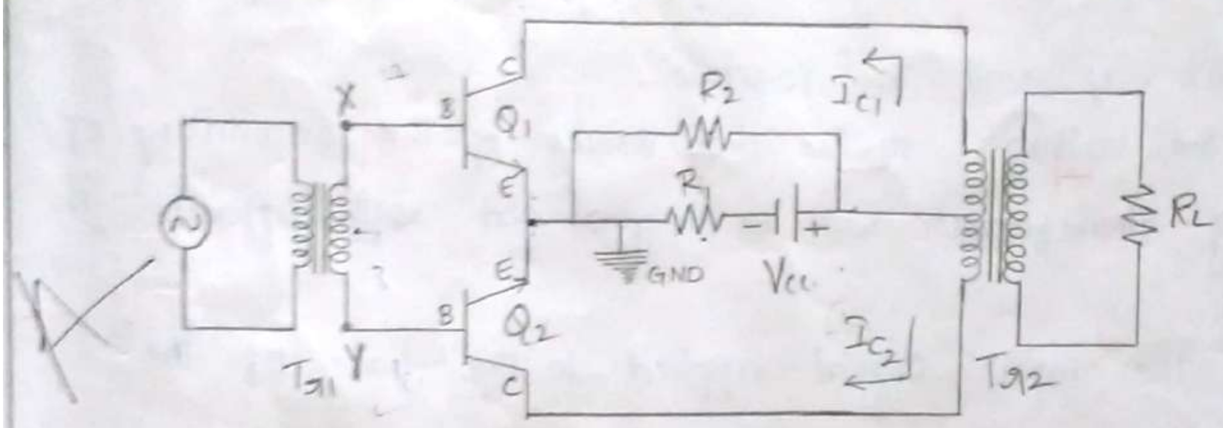


fig:- push-pull class-B power Amplifier.

→ The above figure shows the class-B push pull Amplifier.

→ In this circuit both the ~~transformers~~ transistors Q_1 & Q_2 are of N-P-N type.

→ These both transistors are in CE Configuration.

→ It Consists of two transformers T_{r1} and T_{r2} one is input transformer T_{r1} and other is output transformer T_{r2}.

→ input transformer is also Called as driving transformer which drives the circuit.

→ R_1 and R_2 resistors are the biasing resistors which provides the biasing to the transistors.

→ The input signal is applied to the primary of the driver transformer.

→ The Centre tap on the Secondary of the transformer is grounded

→ The Centre tap on the primary of the ~~d~~ output transf is Connected to the Supply voltage ($+V_{cc}$).

→ With respect to the centre tap for a positive half cy of input signal, the point 'X' shown on the secondary the driver transformer will be positive while the point 'y' will be Negative.

→ Similarly, with respect to the centre tap for a Negative half cycle of input signal, the point 'X' shown on the Secondary of the driver transformer will be Negative while the point 'y' will be positive.

→ Thus the voltages in the two halves of the secondary of the driver transformer will be equal but with opposite in polarities.

→ Hence the input signal applied to the base of the transistors Q_1 & Q_2 will be $180°$ out of phase.

→ Each transistor output is in the form of half rectified waveform

→ Hence the peak value of the output current of each transistor is 'I_m'. So, the average value (dc value) of output current of each transistor is '$\frac{I_m}{\pi}$' (due to half rectified wave form.)

→ The two currents drawn by the two transistors from the dc supply voltage. Hence, the total dc current of both transistors is given by $I_{dc} = \frac{I_m}{\pi} + \frac{I_m}{\pi}$.

$$\boxed{I_{dc} = \frac{2 I_m}{\pi}}$$

→ The dc input power is given by $P_{dc} = V_{cc} \times I_c$

where $I_c = I_{dc}$

$$\boxed{P_{dc} = V_{cc} \times \frac{2 I_m}{\pi}}$$

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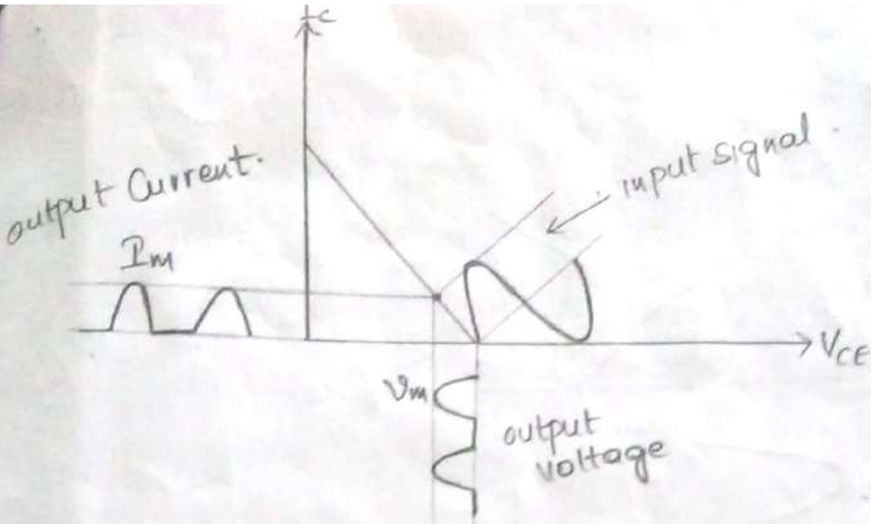


fig: Graphical Representation for push-pull Class-B power Amplifier.

→ where I_m and V_m are the peak Values of output Current and output Voltage respectively.

→ The RMS Value of Voltages and Currents are given by

$$V_{RMS} = \frac{V_m}{\sqrt{2}}. \qquad I_{RMS} = \frac{I_m}{\sqrt{2}}.$$

→ The AC power is given by.

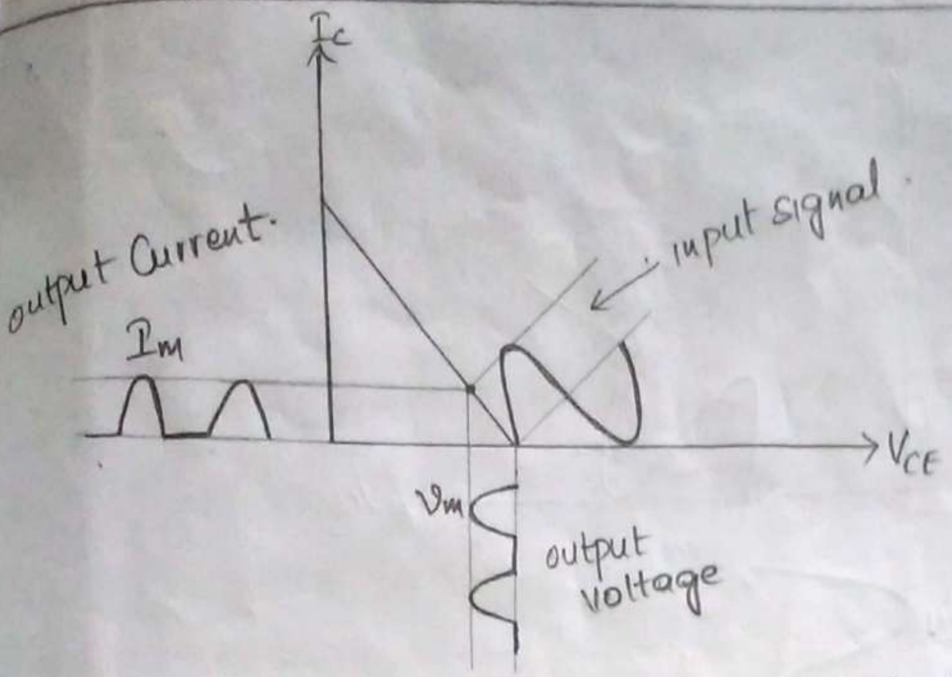
$$P_{AC} = \frac{V_m}{\sqrt{2}} \times \frac{I_m}{\sqrt{2}}$$

$$\boxed{P_{AC} = \frac{V_m I_m}{2}.}$$

Efficiency:- Efficiency is defined as the ratio of AC output power to the dc input power. & it is denoted by 'η'.

$$\eta = \frac{P_{AC}}{P_{DC}} \times 100$$

→ The maximum value of 'η' can be taken from maximum Vary of output Voltage and output Current. and is shown in the figure below.



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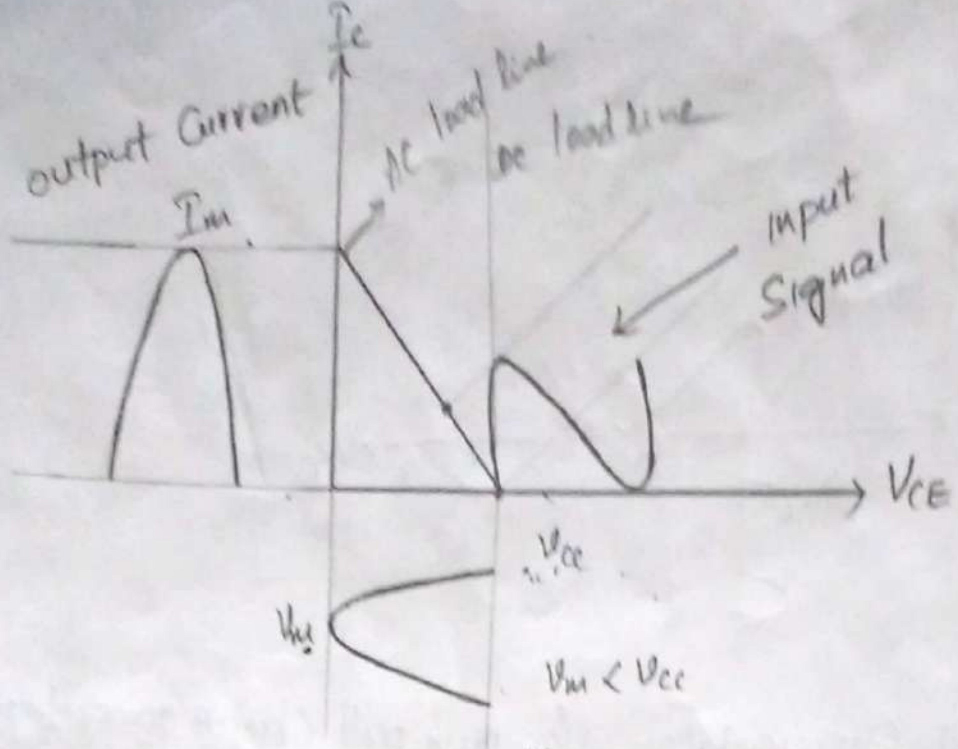
$$\eta = \frac{P_{AC}}{P_{DC}} \times 100$$

→ The maximum Value of 'η' can be taken from maximum Vary of output voltage and output Current. and is shown in the figure below.



output voltage

→ $$\eta = \frac{\frac{V_m I_m}{2}}{\frac{V_{cc} \times 2 I_m}{\pi}} \times 100$$

$$\eta = \frac{\frac{V_m}{2}\pi}{V_{cc} \times 2} \times 100$$

$$\eta = \frac{V_m \pi}{4 V_{cc}} \times 100$$

→ from graph $V_m = V_{cc}$.

Now $$\eta = \frac{\cancel{V_{cc}} \times \pi}{4 \cancel{V_{cc}}} \times 100$$

$$\eta = \pi/4 \times 100$$

$$= 0.785 \times 100$$

$$\boxed{\therefore \eta = 78.5\ \%}$$

→ thus, we Conclude that the efficiency is maximum in class-B push pull power Amplifier.

→ So, the distortion may be reduced by using the push-pull operation of transistors.

Advantages :-

- Efficiency is high
- power dissipation is low.
- Impedance matching is good.
- distortion is less.

In push-pull power Amplifier one transistor being pushed deep into conduction while the other being pulled out (Non-conducting). Hence the name push-pull Amplifier.

* Dis-advantages :-

- circuit design is Complex.
- Cost is high.
- More number of Components is required to design the circuit.
- frequency response is low.
- Occupation space is more.

* Complimentary Symmetry class-B power Amplifier :-

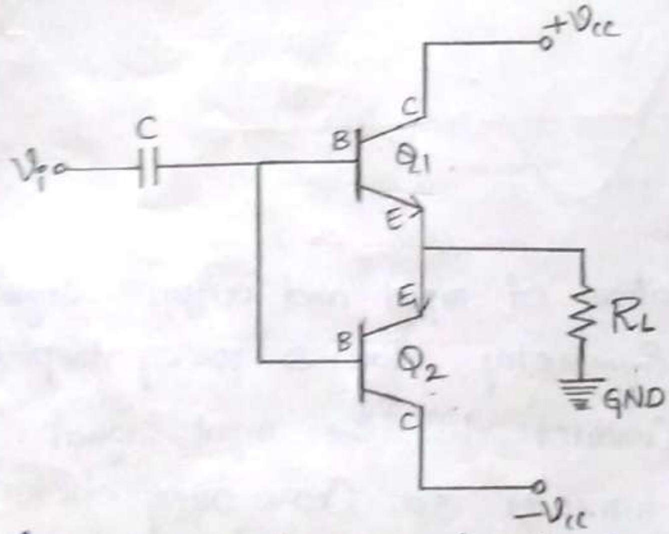


fig: Circuit diagram for Complementary Symmetry Class-B power Amplifier.

→ In class-B push-pull amplifier requires two centre trapped transformers, due to that Cost is high, Circuit design is difficult, and occupation space is more. This draw backs can be Overcome by using Complementary Symmetry class-B power Amplifier.

→ The above figure shows the Complementary Symmetry class-B power Amplifier.

→ It has one N-P-N and one P-N-P transistors Q_1 and Q_2.
→ The input voltage is applied to the base of Q_1 and Q_2 transistors.

* **Operation :-**

→ During positive half cycle of the input signal, transistor Q_1 will conduct and transistor Q_2 does not conduct. So, the positive cycle appear across the load resistance 'R_L'.
→ Similarly, during negative half cycle of the input signal, transistor Q_1 does not conduct and transistor Q_2 conducts. So, the negative half cycle appear across 'R_L'.

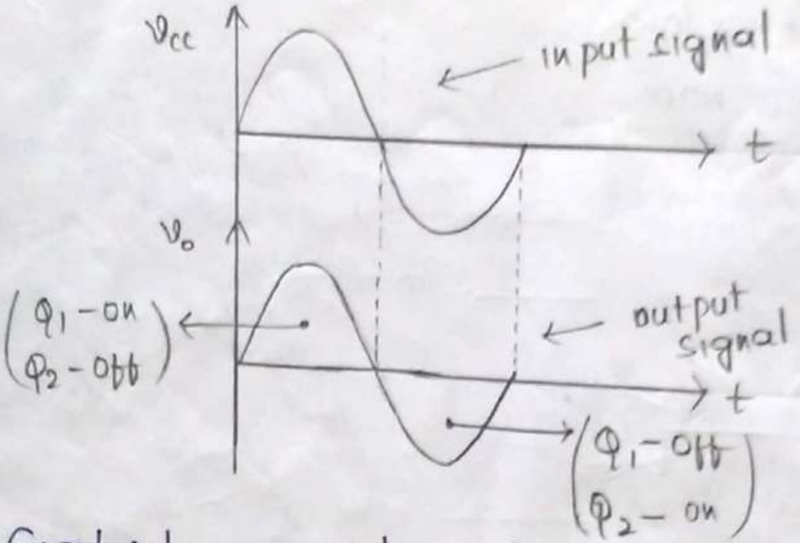


fig:- Graphical representation of input and output signals of Complementary Symmetry Class-B power Amplifier.

→ Transistor will not conduct till the input signal exceeds cut in voltage of the transistor. So, cross over distortion will be present in the output signal of complementary symmetry power Amplifier.

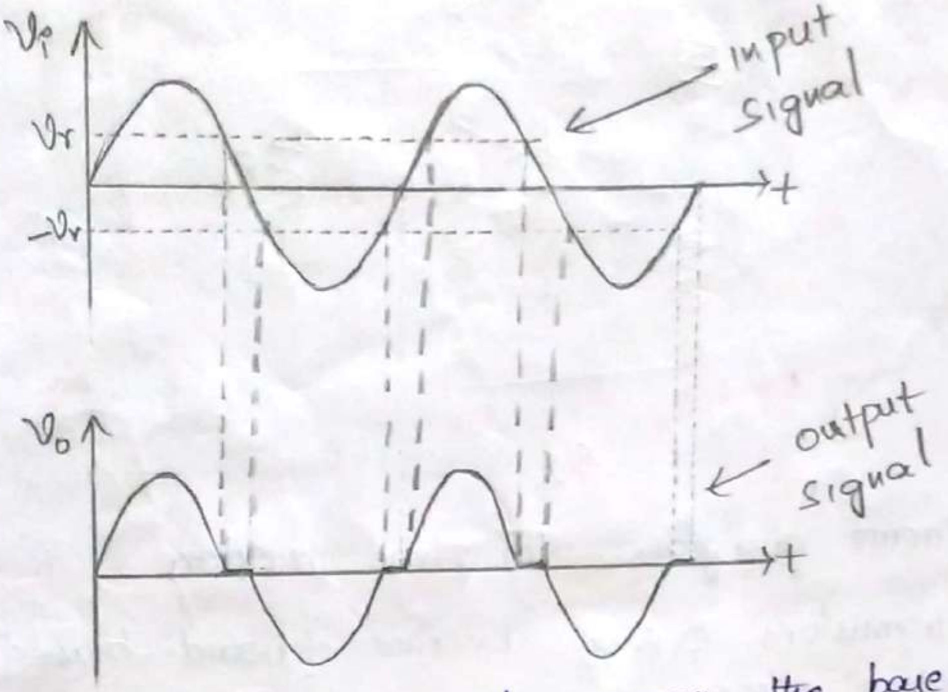
* **Advantages:-**

→ The circuit is transformer less, due to this cost is low and size is small and weight is less.
→ frequency response is improved.
→ Due to common collector configuration impedance matching is possible.

Dis-advantages:-

- Two power supplies is necessary.
- Cross over distortion will be present in the output signal.

Cross over distortion:-



→ The transistor to be in active region, the base-Emitter junction is forward biased and Collector-Base junction is reverse biased.

→ The transistor will ~~Con~~ not Conduct till the input voltage Exceeds the cut in voltage of the transistor. So, we get Cross over distortion in the output signal. It is called as Cross over distortion.

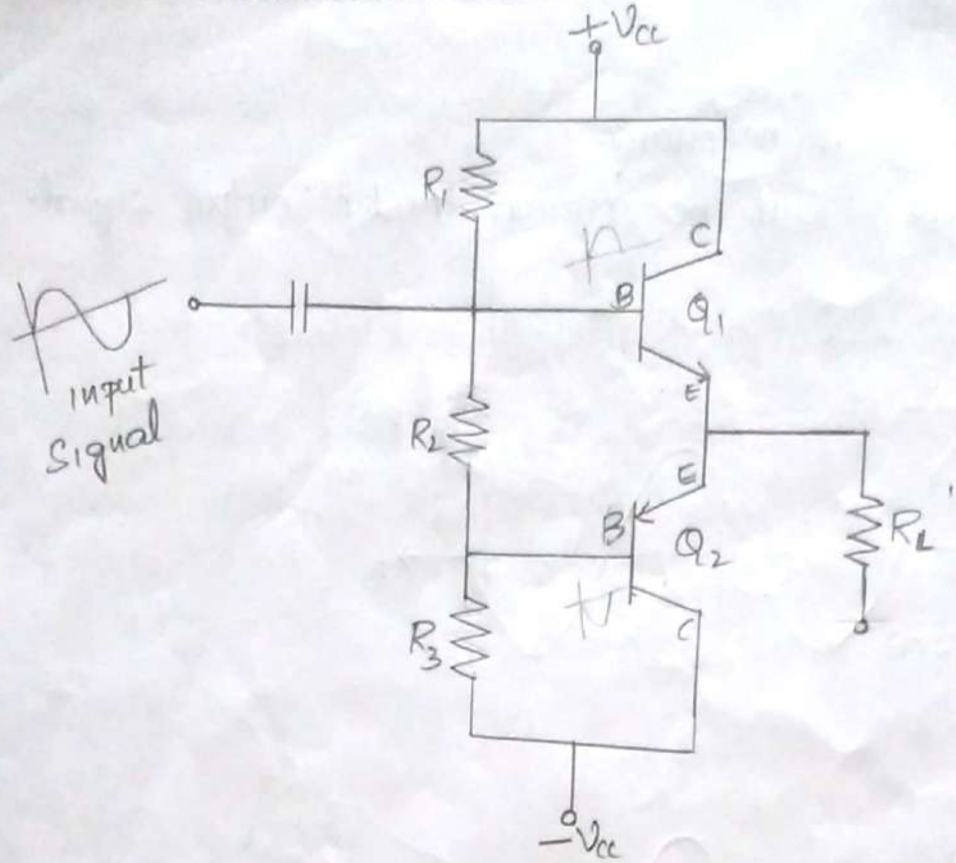
→ The Corresponding waveforms of Cross over distortion is shown in the above graph.

→ To eleminate Cross over distortion DC bias voltage should be provided for base-Emitter junction of Each transistor.

→ By using phase inverter circuits. we can overcome the Cross over distortion in the Class-B power Amplifier.

Phase Invertors:-

→ By using phase inverter circuit, we can reduce the Cross over distortion by providing dc voltage across the base of two transistors.



fig(a):- Circuit diagram for phase invertor.

→ Both the transistors Q_1 & Q_2 become forward biased due to the voltage drop across 'R_2'.

→ The phase invertor circuit is shown in the above figure (a).

→ It Consists of one N-P-N and one P-N-P transistors name Q_1 and Q_2 respectively.

→ Collector - base junction of Q_1 transistor is in reverse bias due to positive voltage is Connected by supply.

→ Collector - base junction of Q_2 transistor is in reverse bias due to Negative Voltage is Connected to the Ground.

→ Therefore, both Collector - base junctions of transistors Q_1 & Q_2 are in reverse bias.

→ During positive peak Voltage 'Q_1' transistor is turned on and 'Q_2' transistor is turned off. So, the positive Voltage appears across the load resistance 'R_L'.

→ During Negative peak voltage 'Q_1' transistor is turned off and 'Q_2' transistor is turned on. So, the Negative Voltage appears across the load resistance 'R_L'.

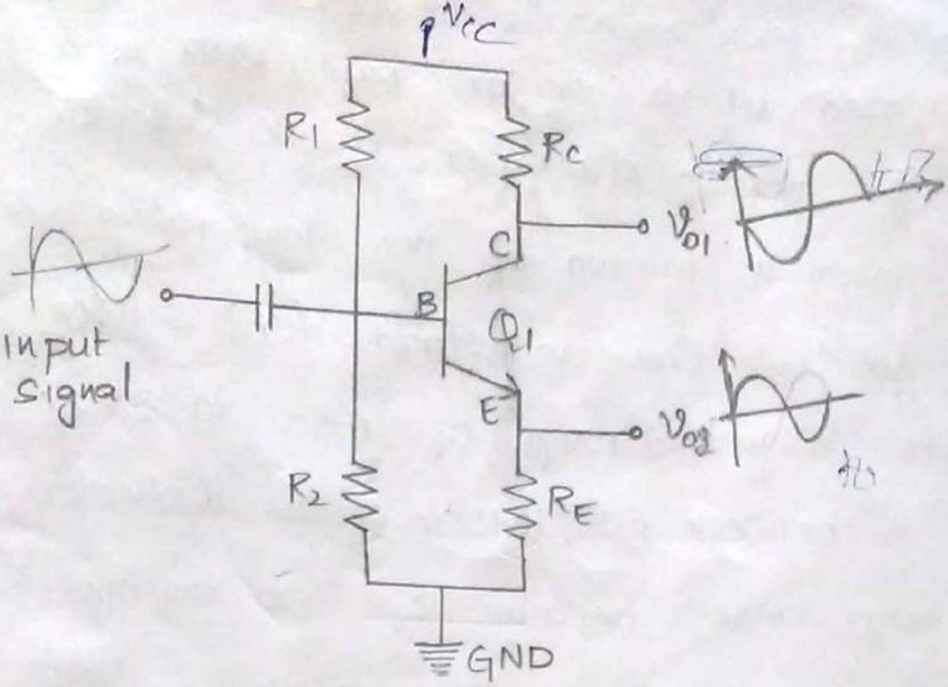


fig (b):- Circuit diagram for modified phase invertor.

→ The modified phase invertor circuit is shown in the figure (b).

→ This circuit is to obtain the two out of phase signals.

→ Here 'R1' and 'R2' are biasing resistors, 'RC' is a load resistance of Collector terminal and 'RE' is a load resistance of Emitter terminal.

→ When the input is applied to the base terminal, the output is taken at the Collector and Emitter terminals (V_{o1} & V_{o2}) will be out of phase by 180°.

→ The output impedance of the circuit from Collector terminal is that of CE Configuration.

→ The output impedance of the circuit from Emitter terminal is that of CC Configuration.

→ When the bypass Capacitor across 'RE' is not present, the Voltage V_{o2} across 'RE' is inphase to the input signal.

→ The output voltage V_{o1} at the Collector terminal will be 180° out of phase to the input signal.

→ Therefore V_{o1} and V_{o2} voltages are out of phase by 180°.

Thermal Runaway:-

→ The maximum average power $P_D(max)$, which a transistor can dissipate depends upon the transistor Construction and may lie in the range from a few milliwatts to 200W As mentioned earlier the power dissipated at its collector base with in a transistor the power dissipated at its Collector - base junction predominantly the power dissipated at its Collector - base junc.

Thus maximum power is limited by the temperature that the Collector - base junction Can withstand for silicon transistor, this temperature is in the range of 150c to 225°c and for germanium it is between 60° to 100°c. The Collector - base junction temperature may arise because of two reasons.

1. Due to rise in the ambient temperature.

2. Due to self heating.

* The self heating Can be Explained as follows:-

→ the increase in the Collector Current increases the power dissipated at the Collector junction. This in turn further increase the temperature of the junction and hence increases Collector Current. The process of the Cummulative and it is referred to as Self heating. This Excess heat produced at the Collector base junction may even burn and destroy the transistor. This Situation is Called Thermal Run away of the transistor.

* Heat Sinks:-

→ Heat sink is basically a large metallic heat Conducting device which when placed near a transistor cools it is necessary increasing it's effective surface area.

* Requirements of heat sinks:- For transistor operating a high level, the heat sinks must be designed to remove the heat by metallic Conduction (or) forced air Cooling.

→ The purpose of heat sinks is to keep the operating point temperature of the transistors prevent thermal breakdown.

...crease in temperature I_c increases & due to increase I_c,
increase which results in the increase in power dissipation and
... by temperature increases. This is a commulative process, thus
... the transistor break down occurs. In order to prevent this,
heat sinks are used which maintain low temperature and
to dissipate power.

If heat sinks are used, the heat is transferred from
due to the junction of package and from package to heat
sink and from heat sink to the ambient. Heat sink factors the
power dissipation and prevent breakdown of the device.

Types of heat sinks:—
Heat sinks are broadly classified as
* Low power transistor type * High power transistor type

* Low power transistor type:—
→ Low power transistor can be mounted directly on the metal
chassis to increase the heat dissipation capacity.
→ The casing the transistor must be insulated from metal
case to prevent short circuited.
→ Beryllium oxide, insulating washers are used for insulating
casing from the case which gives good thermal conductivity
→ Zinc oxide film, silicon compound between walls & case
improves the heat transfer from semiconductor device
to case of chassis.
→ A low power transistor heat sink is shown
in the figure below.

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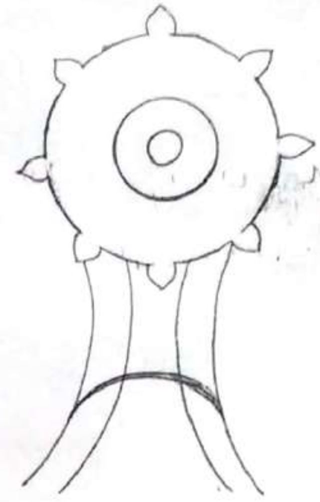


fig:- Fin-type Heat Sink.

* High power transistor type :-

→ To-3, To-66 are the two types of high power transistors these transistors are of diamond shape and dissipates power in order to ...

→ The transistor heat sinks shown in figure below perform Cooling by Conducting Convection and radiating methods : The thermal resistance of heat sinks will typically 3°C/12 .

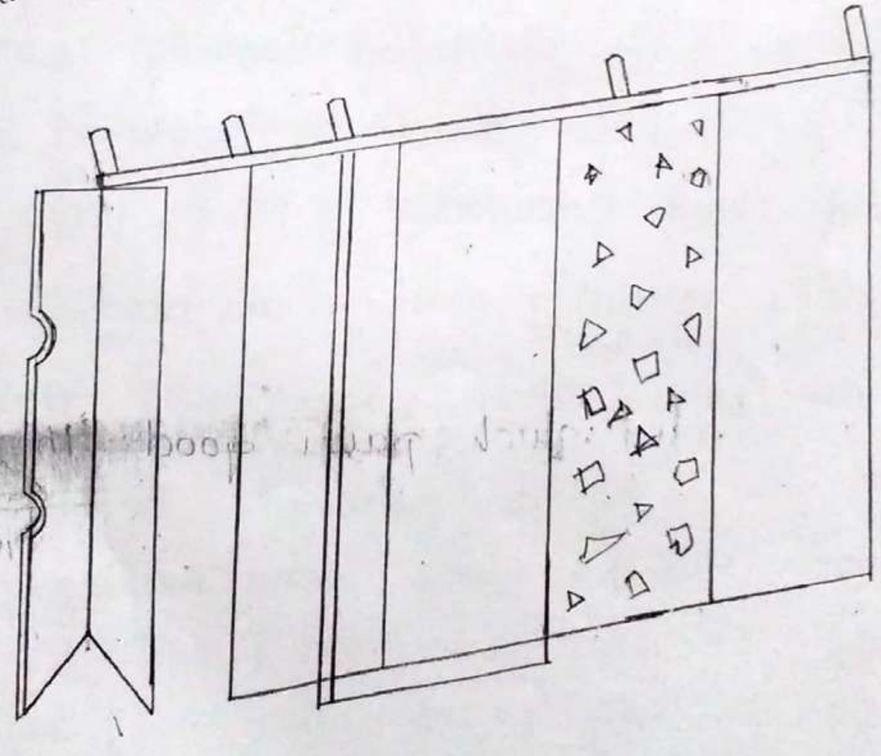


fig:- power transistor heat sink.

DISTORTION:

* In power amplifiers, the input signal applied is alternating in nature. The basic features of any alternating signals are Amplitude, Frequency and phase.

FREQUENCY DISTORTION:

* The change in gain of the amplifier with respect to the frequency is called Frequency distortion.

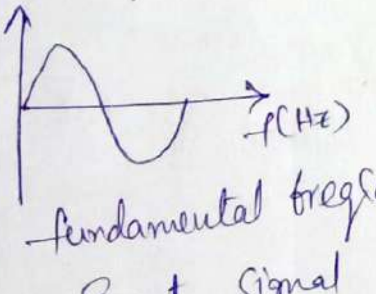
AMPLITUDE DISTORTION:

* A transistor is a perfectly linear device i.e., the dynamic characteristics of a transistor is a straight line over the operating ranges.
* But in practical cases the dynamic characteristics is not perfectly linear due to such non-linearity the wave-form of o/p voltage differs from that of the i/p signal.
* Such a distortion is called Non-linear distortion or Harmonic distortion (or) Amplitude distortion

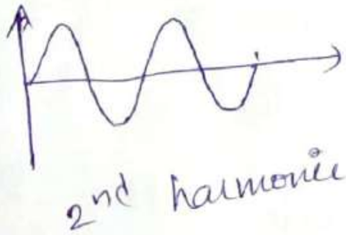
HARMONIC DISTORTION:

* The presence of the frequency components in the output waveform which are not present in the input waveform
* The frequency component in the O/P, whose frequency is same as the i/p signal frequency then it is called as Fundamental frequency.
* The additional frequency components present in the o/p signal are the integral multiples of fundamental frequency these components are called as "Harmonics".

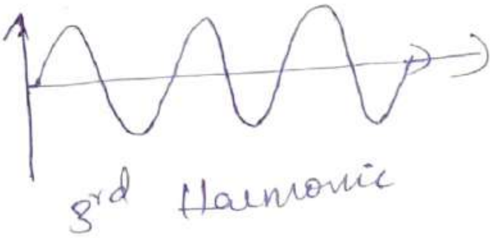
Eg: 'f' Hz is the fundamental frequency
 2f Hz is the second harmonic etc.



fundamental freq(a)
Input Signal



2nd harmonic



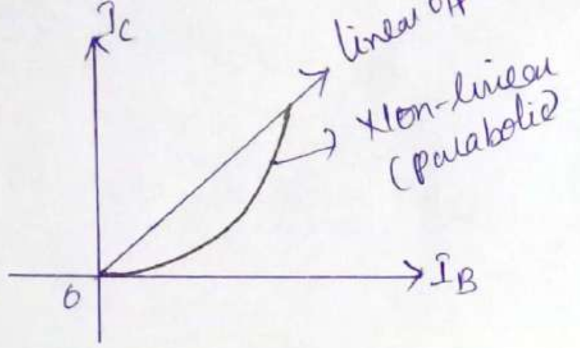
3rd Harmonic

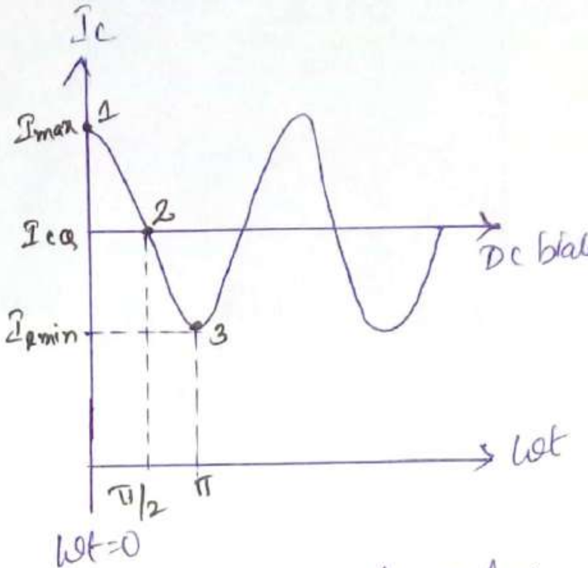
* If the number of Harmonics increases then the amplitude decreases.

SECOND-ORDER HARMONIC DISTORTION (3-POINT METHOD)

* Let us consider AC input signal causes the base current swing which is Cosine in nature.

* To analyze Second harmonic distortion, assume that the dynamic characteristics are ~~non-linear~~ non-linear in nature (parabolic).





* Due to this collector current swing, the operating point value and relation between I_c and I_B is non-linear.

$$i_b = I_{Bm} \cos \omega t$$

mathematically it can be expressed as,

$$i_c = G_1 I_b + G_2 I_b^2$$

$$i_c = G_1 (I_{Bm} \cos \omega t) + G_2 (I_{Bm} \cos \omega t)^2$$

$$= G_1 I_{Bm} \cos \omega t + G_2 I_{Bm}^2 \cos^2 \omega t \qquad \left(\because \cos^2 \omega t = \frac{1 + \cos 2\omega t}{2}\right]$$

$$= G_1 I_{Bm} \cos \omega t + G_2 I_{Bm}^2 \left[\frac{1 + \cos 2\omega t}{2}\right]$$

$$= G_1 I_{Bm} \cos \omega t + \frac{1}{2} G_2 I_B^2 m + \frac{1}{2} G_2 I_{Bm}^2 \cos 2\omega t$$

$$i_c = B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t$$

* The total current can be expressed in terms of second harmonic component, DC signal component.

i.e., $I_C = I_{CQ} + B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t \rightarrow ①$

where $(I_{CQ} + B_0)$ = DC component independent of time

B_1 = Amplitude of fundamental freq

B_2 = Amplitude of second harmonic.

* Let us find the value of total collector current at the various instances 1, 2 & 3.

At point 1, $\omega t = 0$, then eq ① becomes

$$I_C = I_{CQ} + B_0 + B_1 + B_2 \rightarrow ②$$

At point 2, $\omega t = \frac{\pi}{2}$ then eq ① becomes

$$I_C = I_{CQ} + B_0 - B_2 \rightarrow ③$$

At point 3, $\omega t = \pi$ then eq ① becomes

$$I_C = I_{CQ} + B_0 - B_1 + B_2 \rightarrow ④$$

From eq ③ If $I_C = I_{CQ}$

$$\boxed{B_0 = B_2}$$

From eq ② & ④

$$I_{max} - I_{min} = \cancel{I_{CQ}} + \cancel{B_0} + B_1 + \cancel{B_2} - \cancel{I_{CQ}} - \cancel{B_0} + B_1 - \cancel{B_2}$$

$$I_{max} - I_{min} = 2B_1$$

$$\boxed{B_1 = \frac{I_{max} - I_{min}}{2}}$$

From eq ③ & eq ④

$$I_{max} + I_{min} = I_{CQ} + B_0 + B_1 + B_2 + I_{CQ} + B_0 - B_1 + B_2$$

$$I_{max} + I_{min} = 2I_{CQ} + 2B_0 + 2B_2$$

$$I_{max} + I_{min} = 2I_{CQ} + 2(B_2) + 2B_2 \quad (\because B_0 = B_2)$$

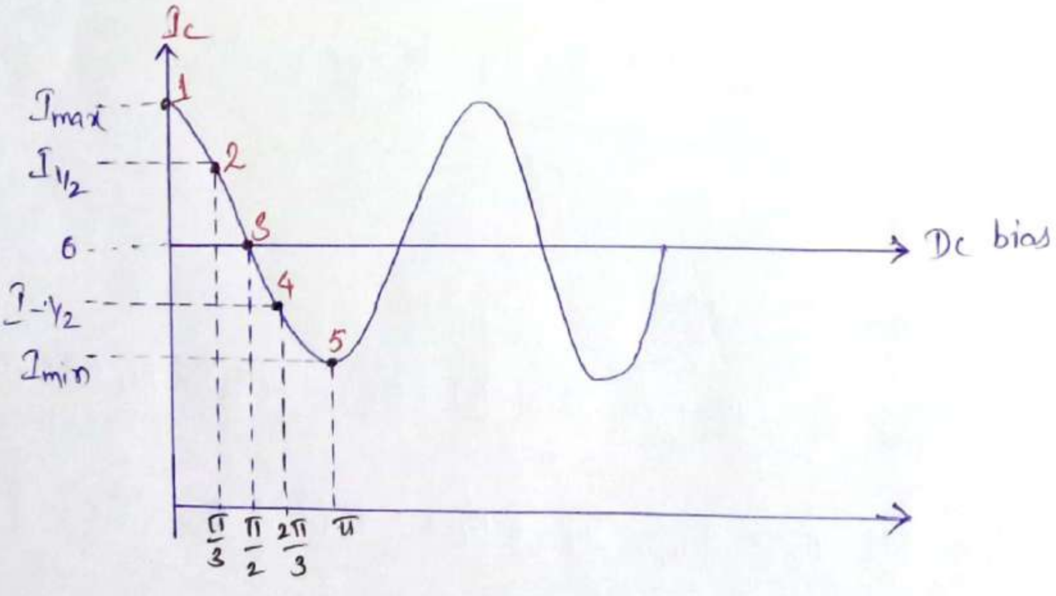
$$I_{max} + I_{min} = 2I_{CQ} + 4B_2$$

$$4B_2 = I_{max} + I_{min} - 2I_{CQ}$$

$$\boxed{B_2 = \frac{I_{max} + I_{min} - 2I_{CQ}}{4}}$$

HIGHER ORDER HARMONIC DISTORTION (5 POINT METHOD):

* As the non-linearity increased in the dynamic characteristics of a transistor then the harmonic distortion also increases.



* $I_b = I_{Bm} \cos \omega t$

* The mathematical expression for the collector current due to higher order harmonics,

$$I_C = G_1 I_{Bm}\cos\omega t + G_2 I_{Bm}^2 \cos^2\omega t + G_3 I_{Bm}^3 \cos^3\omega t + G_4 I_{Bm}^4 \cos^4\omega t \ldots$$

$$I_C = G_1 I_b + G_2 I_b^2 + G_3 I_b^3 + G_4 I_b^4 + \ldots \qquad (\because I_b = I_{Bm}\cos\omega t)$$

$$I_C = B_0 + B_1\cos\omega t + B_2\cos 2\omega t + B_3\cos 3\omega t + B_4\cos 4\omega t + \ldots$$

$$I_C = I_{CQ} + B_0 + B_1\cos\omega t + B_2\cos 2\omega t + B_3\cos 3\omega t + B_4\cos 4\omega t + \ldots \rightarrow \text{eq } 1$$

where $(I_{CQ} + B_0) \Rightarrow$ is DC component independent of time

B_1 = Amplitude of fundamental frequency

B_2 = Amplitude of second harmonic

B_3 = Amplitude of third harmonic

At point 1 $\omega t = 0$:

Then equation 1 becomes

$$I_C = I_{CQ} + B_0 + B_1\cos(0) + B_2\cos 2(0) + B_3\cos 3(0) + B_4\cos 4(0)$$

$$I_C = I_{CQ} + B_0 + B_1 + B_2 + B_3 + B_4 \rightarrow \text{eq } 2$$

At point 2 $\omega t = \pi/3$

Then equation 1 becomes

$$I_C = I_{CQ} + B_0 + B_1\cos\pi/3 + B_2\cos 2(\pi/3) + B_3\cos 3(\pi/3) + B_4\cos 4(\pi/3)$$

$$I_{1/2} = I_C = I_{CQ} + B_0 \quad I_C = I_{1/2} = I_{CQ} + B_0 + \tfrac{1}{2}B_1 - \tfrac{1}{2}B_2 - B_3 - \tfrac{1}{2}B_4 \rightarrow 3$$

At point 3 $\omega t = \pi/2$

Then equation 1 becomes

$$I_C = I_{CQ} + B_0 + B_1\cos\pi/2 + B_2\cos 2(\pi/2) + B_3\cos 3(\pi/2) + B_4\cos 4(\pi/2)$$

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$$I_c = I_{CQ} + B_0 - B_2 + B_4 \rightarrow eq\ 4$$

At point 4 $\omega t = 2\pi/3$

Then eq 1 becomes

$$I_c = I_{CQ} + B_0 + B_1 \cos 2\pi/3 + B_2 \cos 2(2\pi/3) + B_3 \cos 3(2\pi/3) + B_4 \cos 4(2\pi/3)$$

$$I_c = I_{CQ} + B_0 - \cancel{B}_{1/2} B_1 - 1/2 B_2 + B_3 - 1/2 B_4 \rightarrow 5$$

At point 5 $\omega t = \pi$

Then eq 1 becomes

$$I_c = I_{CQ} + B_0 \cancel{+} B_1 \cos\pi + B_2 \cos 2\pi + B_3 \cos 3\pi + B_4 \cos 4\pi$$

$$I_c = I_{CQ} + B_0 - B_1 + B_2 - B_3 + B_4 \rightarrow 6$$

By solving above equations

$$B_0 = \frac{1}{6}\left[I_{max} + 2I_{1/2} + 2I_{-1/2} + I_{min}\right]$$

$$B_1 = \frac{1}{3}\left[I_{max} + 2I_{1/2} - 2I_{1/2} - I_{min}\right]$$

$$B_2 = 1/4\left[I_{max} - 2I_{CQ} + I_{min}\right]$$

$$B_3 = \frac{1}{6}\left[I_{max} - 2I_{1/2} + 2I_{-1/2} - I_{min}\right]$$

$$B_4 = \frac{1}{12}\left[I_{max} - 4I_{1/2} + 6I_{CQ} \cancel{+} 4I_{-1/2} + I_{min}\right]$$

OPERATION OF CLASS AB POWER AMPLIFIER:

* The power amplifier in which the output current flows for more than half cycle and less than full cycle of i/p signal is known as Class-AB Power amplifier.
* The Conduction angle of class AB power amplifier is between class-A and class-B power amplifier i.e., (<360° & >180°).
* As, class A has the problem of low efficiency and class-B has distortion problem, this class AB is emerged to eliminate these two problems, by utilizing the advantages of both the classes.
* The cross-over distortion is the problem that occurs when both the transistors are OFF at the same instant, during the transition period.
* In order to eliminate this, the condition has to be chosen for more than one half cycle. Hence the other transistor gets into conduction, before the operating transistor switches to cut-off state.
* This is achieved by Class-AB power amplifier

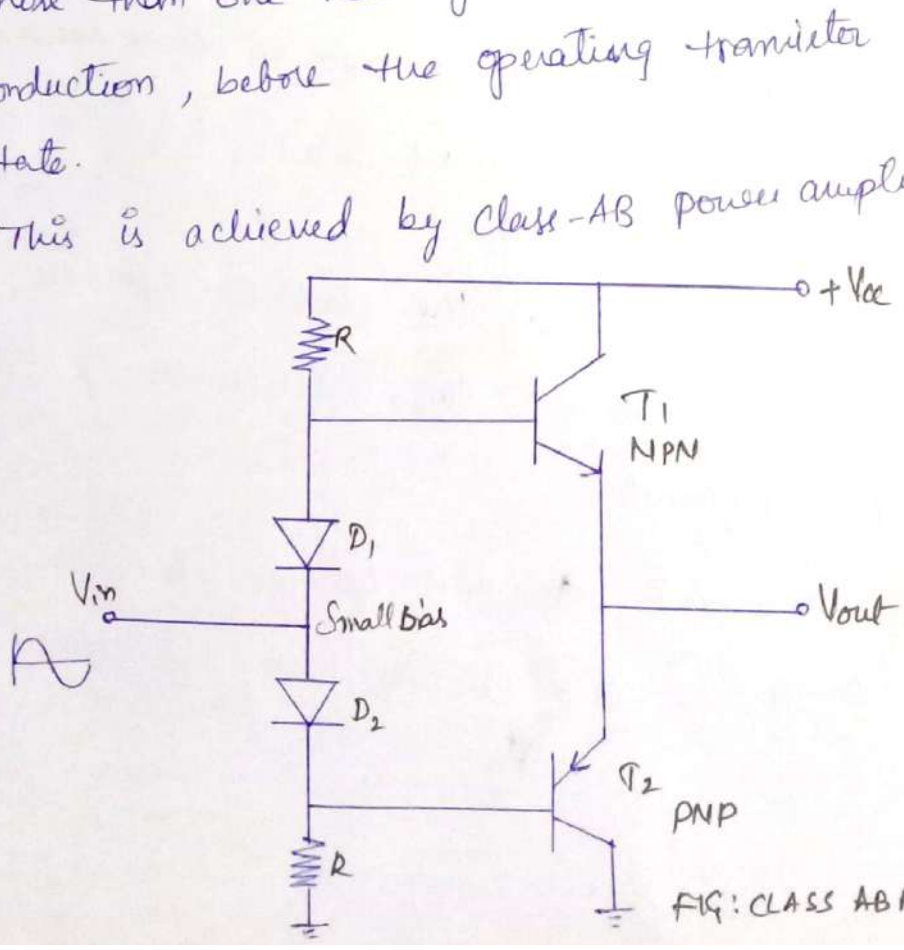

FIG: CLASS AB POWER AMPLIFIER

* In class AB amplifier design, each of the push-pull transistors is conducting for slightly more than the half cycle of conduction in class B, but much less than the full cycle of conduction of class A.

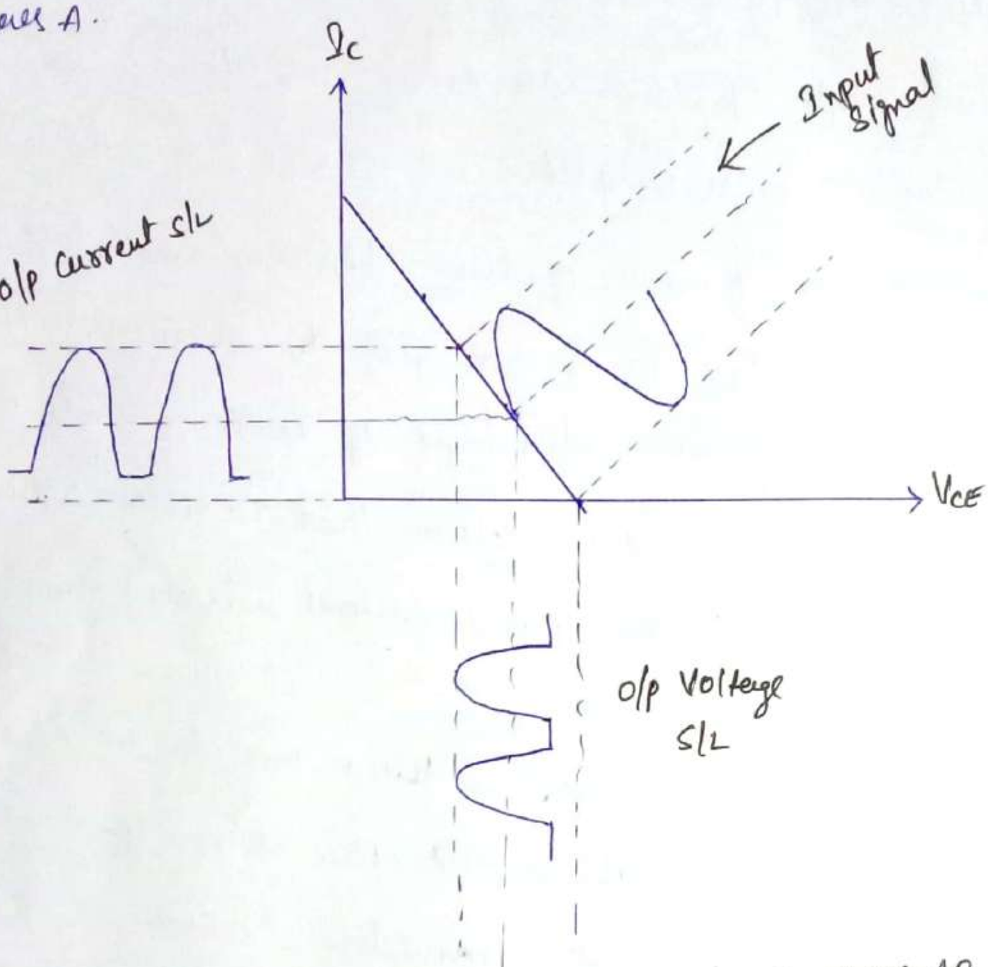

FIG: GRAPHICAL REPRESENTATION OF CLASS-AB POWER AMPLIFIER

* The small bias voltage given using diodes D_1 & D_2 helps the operating point to be above the cut-off point.
* The cross-over distortion created by the class-B amplifier is overcome by this class-AB as well the inefficiencies of class-A and B don't affect the circuit.
* So, the class AB is a good compromise between class-A and class-B in terms of efficiency and linearity having the efficiency reaching about 50% to 60%.

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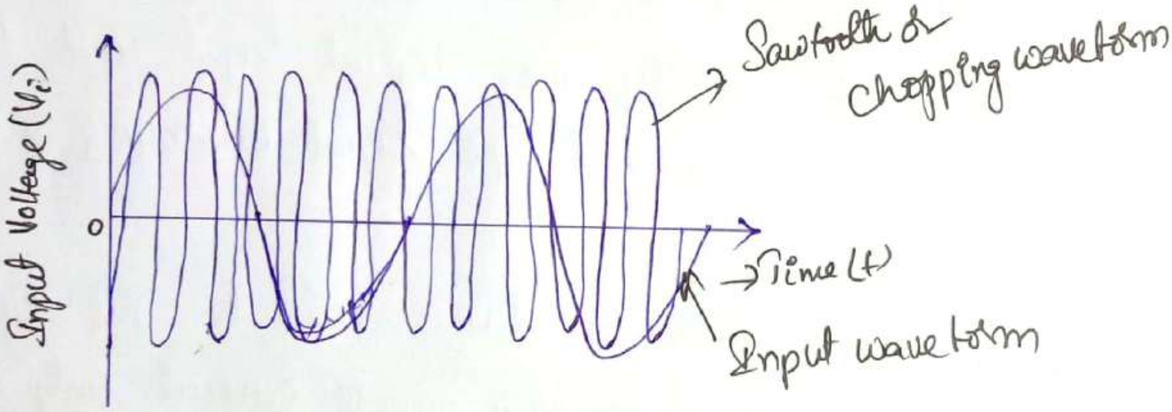
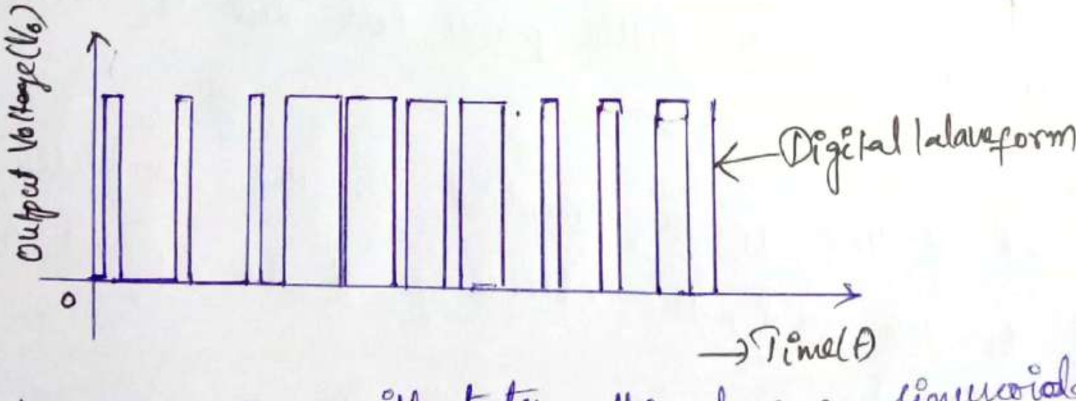
CLASS-D POWER AMPLIFIER:

* A class-D power amplifier is designed to operate with Digital or pulse type signals.

* The letter 'D' stands for "Digital" since that is the nature of signals provided to the class-D amplifier.

* It is necessary to convert any input signal into a pulse type waveform before using it to drive a large power load and to convert the signal back to the sinusoidal signal.





* The above figures illustrates the how a sinusoidal signal may be converted into a pulse type signal.

* Using some form of Sawtooth or Chopping waveform to be applied.

* The Block diagram of class-D power amplifier is as shown.

For converting Digital S/L back to sinusoidal S/L

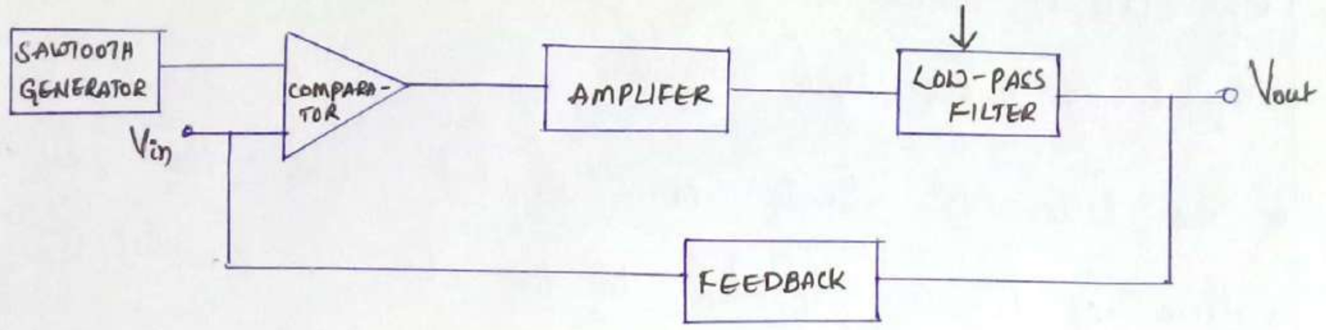


FIG: BLOCK DIAGRAM OF CLASS-D POWER AMPLIFIER

* The Sawtooth & chopping waveform to be applied with the input into a comparator type op-amp circuit so that a representative pulse-type signal is generated.
* The amplifier will amplify the digital signal and then convert back to the sinusoidal type signal employing a low-pass filter.
* Since the amplifier's transistor devices used to give the output are basically either off or ON, they provide current only when they are turned ON, with little power loss due to their low on-voltage.
* Thus most of the power supplied to the amplifier is transferred to the load, the efficiency of the circuit is typically very high.
* Class-D amplifier can attain efficiencies of 90% and with careful component choices can exceed 95% even.
* Power MOSFET's devices have been quite popular as the driver devices for the class D power amplifier.

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CLASS-S AMPLIFIER:

* The class-S amplifier has an input a pulse-width modulated (PWM) signal to turn Q_1 and Q_2 ON or OFF as switches with a switching frequency much higher than the signal frequency.

* The circuit diagram for class-S amplifier is given by,

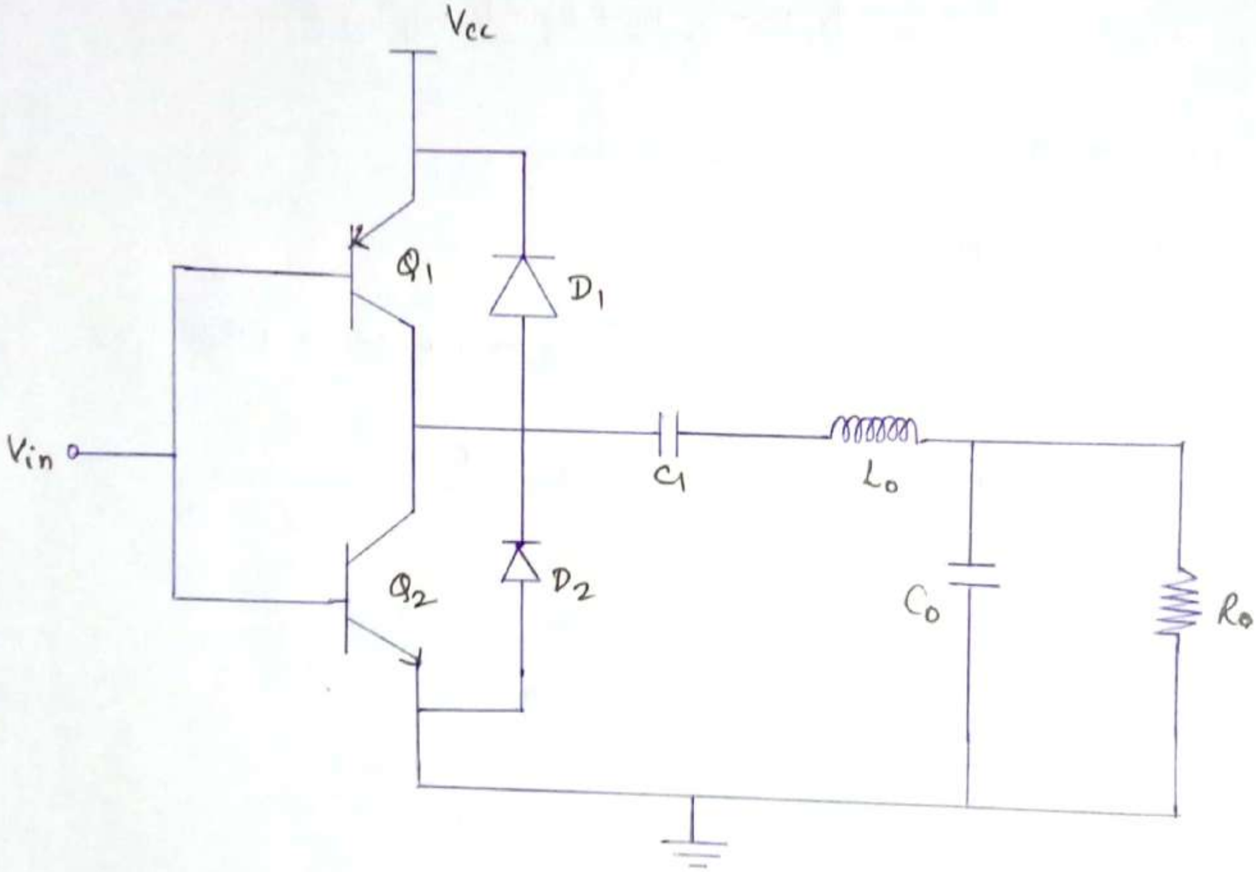
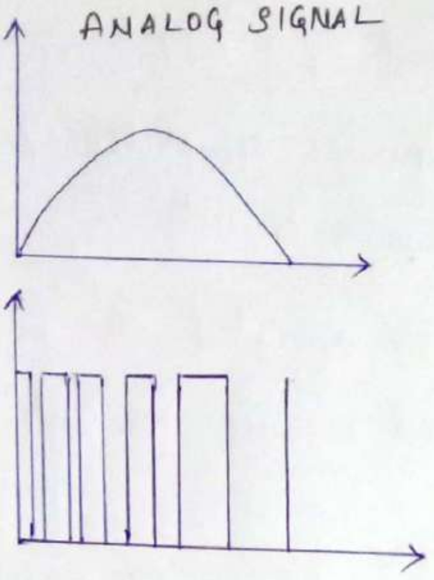


FIG: CIRCUIT DIAGRAM FOR CLASS-S AMPLIFIER

* It is used for both amplification and Amplitude modulation
* It is similar to class D except the rectangular PWM Voltage waveform is applied to a low-pass filter that allows only the slowly varying DC or average Voltage component to appear across the load.



PWM SAMPLE

* The L_0 and C_0 in the circuit forms a low pass filter that turns the PWM signal into an analog waveform.
* If only positive outputs are needed, only Q_1 and D_2 are required. For negative signals, only D_1 and Q_2 are necessary.
* The switching frequency must be significantly higher than the signal frequency, this technique is not viable for amplification in the GHz frequency range.

MOSFET POWER AMPLIFIER:

* Power amplifiers designed to switch large currents ON and OFF make use of MOSFET devices.
* MOSFET based class-D amplifiers is commonly used. Other applications include line drivers for digital switching circuits, switched mode voltage regulators.
* Power MOSFET's have several advantages over bipolar transistors for power amplifier applications

* One of their most important advantages is a transfer characteristic which is more linear than that of a BJT.
* This property makes MOSFET power amplifier to have much less output distortion than bipolar circuits.
* Power MOSFET's can readily be operated in parallel to reduce the total channel resistance and increase the output current level.
* The advantage of using MOSFET device for switching is that the "turn OFF" time is not delayed by minority - carrier storage, as it is in a bipolar transistor.
* Further current in a MOSFET is due to majority charge carriers only and they are not subjected to thermal runaway.
* In addition, very large input impedance of MOSFET, makes the design of driver circuitry less complex.
* Though MOSFET amplifiers can be used in class A mode, it is advantageous to use in class D mode.

$$= 5 + \frac{85 \times 8}{85 + 8} = 5 + 7.31 = 12.31°\text{C/W}$$

$$P_D = \frac{T_J - T_A}{Q_{J-A}} = \frac{160 - 40}{12.31} = \frac{120}{12.31} = 9.75 \text{ W}$$

Types of heat sinks

Low power transistor type The small signal low power transistors can be mounted directly on the metal chassis to increase the sufficient heat dissipation capability. Care should be taken while doing this because very often the collector of the transistor is connected to the transistor case to increase heat-dissipation capabilities. Hence, some provision for insulating the case from the chassis, which is usually at ground potential, must be provided unless a common collector is being employed.

One method of achieving this is to use a beryllium oxide insulating washer which has a good thermal conductivity, as shown in Fig. 7.14(a). By using a zinc oxide film Silicon compound between the washer and the chassis, heat transfer from the transistor case to the chassis may be improved. An insulated clamp over the top of the transistor may be used to help improve thermal dissipation and increasepressure.

When the transistor is mounted in Teflon (PTFE—Poly Tere Fluoro Ethylene) sockets, it does not provide thermal conduction from transistor case to chassis. Therefore, a press-on fin type of a black anodized heat sink may be used, as shown in Fig. 7.14(b), for mounting transistors that are encased in a metal TO-5 package.

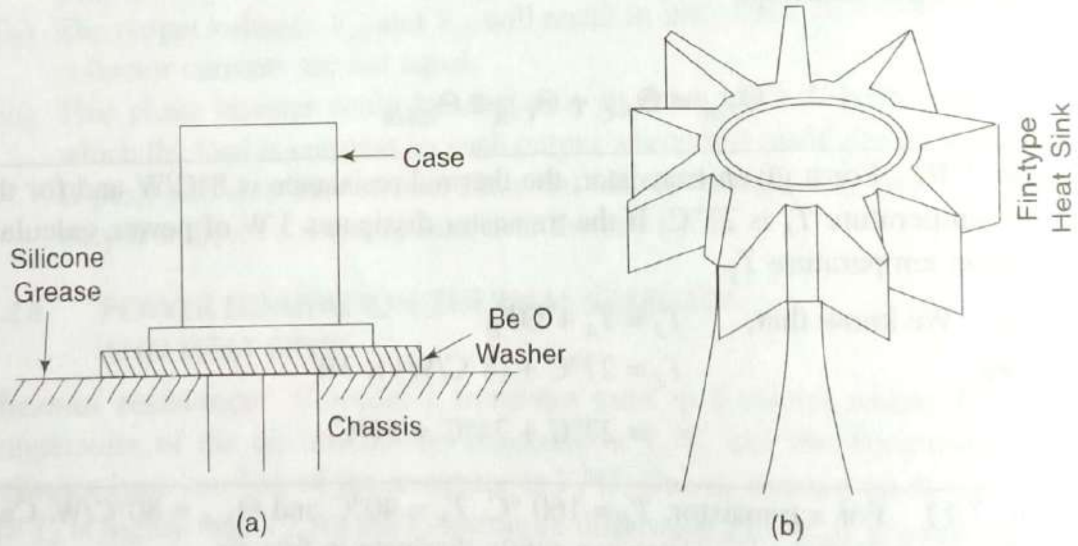


Fig. 7.14 *(a) Mounting the transistor case close to the chassis using a berylliumoxide insulating washer (b) using a separate heat sink pressed onto the transistor*

Power transistor heat sinks The diamond shaped TO-3 and TO-66 types are the popular mounting packages used for the power transistors which havedissipation in the order of 100 W. These have two leads for emitter and base but the case, or the mounting flange of the case, is the collector terminal. So, it is necessary to insulate the case from the heat sink by the use of an insulating washer. Figure 7.15 shows a typical heat sink that can accommodate a TO-3 power transistor package that provides cooling by conduction, convection and radiation. Although measuring only 11.5 cm

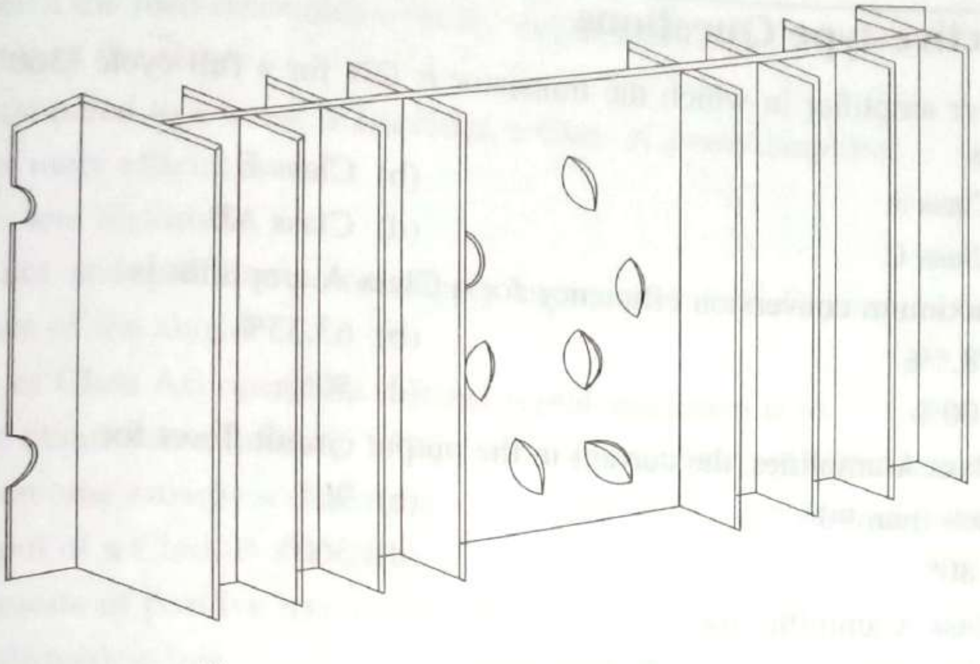


Fig. 7.15 *Power transistor heat sink*

by 7.8 cm, it has a thermal dissipation equal to that of a flataluminium sheet 25 cm $\times$ 20 cm $\times$ 0.32 cm. The thermal resistance of this heat sink is 3°C/W.

TUNED AMPLIFIERS

Introduction:-

* A audio amplifier amplifies a wide band of frequencies equally well and doesnot permit the selection of a particular desired frequency while rejecting all other frequencies.
* For instance, radio and television transmission are carried on a specific radio frequency assigned to the broadcasting station.
* The radio receiver is required to pick up and amplify the radio frequency desired while discriminating all others.
* To achieve this, the simple resistive load is replaced by a parallel tunned circuit whose impedence strongly depends upon frequency.
* Therefore, the use of tunned circuits in conjunction with a transistor makes possible the selection and efficient amplification of a particular desired radio-frequency.
* Such an amplifier is called a Tunned amplifier.

Advantages of Tuned Amplifiers:

* In high frequency applications, it is generally required to amplify a single frequency, rejecting all other frequencies present.
* For such purposes, tuned amplifiers are used. These amplifiers ~~are~~ use tuned parallel circuit as the collector load and offer the following advantages:

i) Small power loss:-

- A tunned parallel circuit employs reactive components L and C.
- Consequently, the power loss in such a circuit is quite low.
- On the other hand, if a resistive load is used in the

collector circuit, there will be considerable loss of power.

• Therefore, tuned amplifiers are highly efficient.

ii) High selectivity :-

* A tuned circuit has the Property of selectivity i.e it can select the desired frequency for amplification out of a large number of frequencies simultaneously impressed upon it.

* For instance, if a mixture of frequencies ~~simultaneously impressed upon it~~. fr is fed to the input of a tuned amplifier, then maximum amplification occurs for fr.

* For all other frequencies, the tuned circuit offers very low impedance and hence these are amplified to a little extent and may be thought as rejected by the circuit.

* On the other hand, if we use resistive load in the collector, all the frequencies will be amplified equally well i.e the circuit will not have the ability to select the desired frequency.

iii) Smaller collector supply voltage :-

* Because of little resistance in the Parallel tuned circuit. it requires small collector supply voltage VCC.

• On the other hand, if a high load resistance is used in the collector for amplifying even one frequency, it would mean large voltage drop across it due to zero signal collector current. Consequently, a higher collector supply will be needed.

why not tuned circuits for low frequency amplification:-

• The tuned amplifiers are used to select and amplify a specific high frequency or narrow band of frequencies.

• The reader may be inclined to think as to why tuned circuits are not used to amplify low frequencies. This is due to the following reasons:

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i) Low frequencies are never single:

- A tuned amplifier selects and amplifies a single frequency.
- However, the low frequencies found in practice are the audio frequencies which are a mixture of frequencies from 20 Hz to 20 kHz and are not single.
- It is desired that all these frequencies should be equally amplified for proper reproduction of the signal.
- Consequently, tuned amplifiers cannot be used for the purpose.

ii) High values of L and C:

- The resonant frequency of a parallel tuned circuit is given by;

$$f_r = 1/2\pi LC$$

- For low frequency amplification, we require large values of L and C.
- This will make the tuned circuit bulky and expensive.
- It is worthwhile to mention here that R-C and transformer coupled amplifiers, which are comparatively cheap, can be conveniently used for low frequency.

Classification:-

Tuned Amplifiers

- Small signal
  - To amplify low RF signals
  - power output is low
  - operated in class A
    - Single tuned
    - Double tuned
    - Staggered tunned
- Large signal
  - To amplify large RF signals
  - power output is more
  - operated in class B, class C or class AB modes.
  - Pushpull configuration used to further reduce harmonic distortion.

Tuned amplifiers:-

- Amplifiers which amplify a specific frequency or narrow band of frequencies are called Tuned amplifiers.
- Tuned amplifiers are mostly used for the amplification of high or radio frequencies.
- It is because radio frequencies are generally single and the tuned circuit permits their selection and efficient amplification.
- However, such amplifiers are not suitable for the amplification of audio frequencies as they are mixture of frequencies from 20 Hz to 20 kHz and not single.
- Tuned amplifiers are widely used in radio and television circuits where they are called upon to handle radio frequencies.
- The impedance of this tuned circuit strongly depends upon frequency.
- It offers a very high impedance ~~of this tuned circuit~~ strongly depends upon frequency.
- It offers a very high impedance of this at resonant frequency and very small impedance at all other frequencies.
- If the signal has the same frequency as the resonant frequency of LC circuit, large amplification will result due to high impedance of LC circuit at this frequency.
- When signals of many frequencies are present at the input of tuned amplifier, it will select and strongly amplify the signals of resonant frequency while rejecting all others.
- Therefore, such amplifiers are very useful in radio receivers to select the signal from one particular

broadcasting station when signals of many other frequencies are present at the receiving aerial.

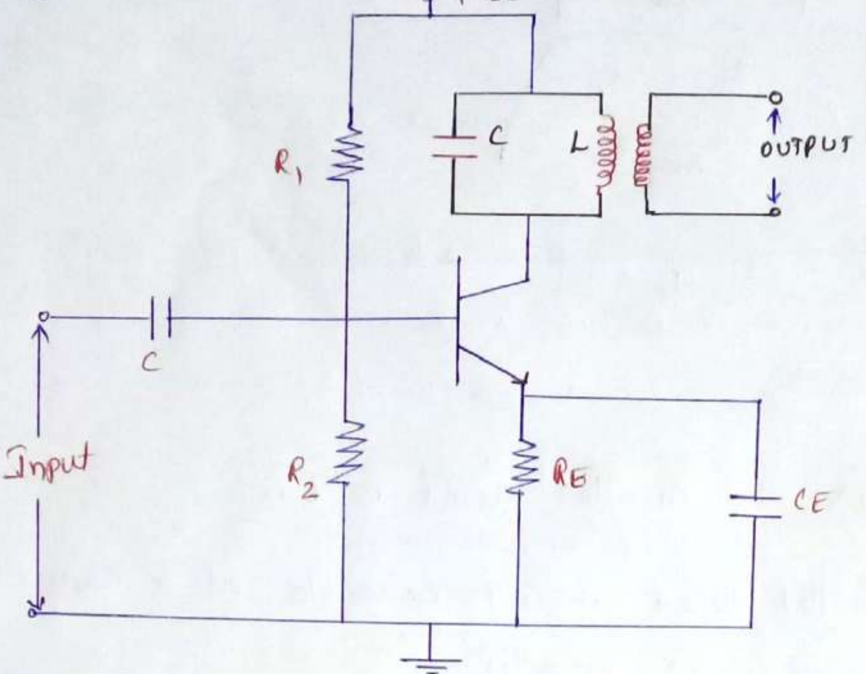
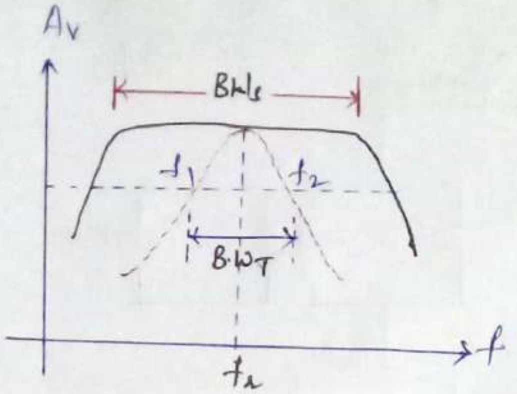


FIG: TUNED AMPLIFIER

Distinction between Tuned Amplifiers and other Amplifiers :-

- we have seen that amplifiers (e.g., voltage amplifier, power amplifier etc) provide the constant gain over a limited band of frequencies i.e., from lower cut-off frequency $f1$ to upper cut-off frequency f_2.
- Now bandwidth of the amplifier, $BW = f_2 - f_1$.
- The reader may wonder, then, what distinguishes a tuned amplifier from other amplifiers? The differences is that tuned amplifiers are designed to have specific, usually narrow brand with.
- This point is illustrated in Fig 5.2. Note that BWS is the bandwidth of standard frequency response while BWT is the bandwidth of the tuned amplifier.
- In many applications, the narrower the bandwidth of a tuned amplifier, the better it is.



Analysis of Parallel Tuned circuit :

- A parallel tuned circuit consists of a capacitor C and indicator L in parallel.

i) In Practice, some resistance R is always Present with the coil. If an alternating voltage is applied across this parallel circuit, the frequency of oscillations will be that of the ~~parall~~ applied voltage.

- However, if the frequency of applied voltage is equal to the natural or resonant frequency of LC circuit, then electrical resonance will occur.

- under such conditions, the impedance of the tuned circuit becomes maximum and the line current is minimum.

- The circuit then draws just enough energy from a.c. supply necessary to overcome the losses in the resistance R.

Parallel resonance :-

- A parallel circuit containing reactive elements (L and C) is resonant when the circuit power factor is unity i.e. applied voltage and the supply current are in phase.

- The phasor diagram of the parallel circuit.

ii) The coil current IL has two rectangular components viz active component IL cos ωL and reactive component IL sin ωL. This parallel circuit will resonate when

$$f_r = \frac{1}{2\pi L}\sqrt{\frac{L}{C} - R^2}$$

Resonant frequency $f_r = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{L^2}}$

If coil resistance R is small (as is generally the case), then, $f_r = \frac{1}{2\pi\sqrt{LC}}$

The resonant frequency will be in Hz if R, L & C are in ohms, henry and farad respectively.

Characteristics of Parallel Resonant circuit :-

It is now desirable to discuss some important characteristic of parallel resonant circuit.

1) Impedance of tuned circuit :-

- The impedance offered by the parallel LC circuit is given by the supply voltage divided by the line current i.e V/I.
- Since at resonance, line current is minimum, therefore, impedance is maximum at resonant frequency. This fact is shown by the impedance - frequency curve.
- It is clear from impedance - frequency curve that impedance rises to a steep peak at resonant frequency fr.
- However, the impedance of the circuit decreases rapidly when the frequency is changed above or below the resonant frequency.
- This characteristic of parallel tuned circuit provides it the selective properties i.e to select the resonant frequency and reject all others.

Line current, $I = I_L \cos \phi_L$

$$\frac{V}{Z_r} = \frac{V}{Z_L} \times \frac{R}{Z_L}$$

$$\frac{1}{Z_r} = \frac{R}{Z^2_L}$$

$$\frac{1}{Z_r} = \frac{R}{L/C} = \frac{CR}{L}$$

$\left[\because Z^2_L = \frac{L}{C} \text{ from eq ii}\right]$

the circuit power factor is unity. This is possible only when the net reactive component of the circuit current is zero i.e.

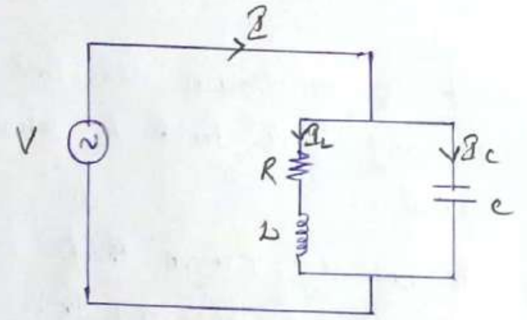
$$I_C - I_L \sin\phi_L = 0$$

$$I_C = I_L \sin\phi_L$$

- Resonance in parallel circuit can be obtained by changing the supply frequency.
- At some frequency f_r (called resonant frequency), $I_C = I_L \sin\phi_L$ and resonance occurs.

Resonant frequency:-

- The frequency at which parallel resonance occurs (i.e, reactive component of circuit current becomes zero) is called the resonant frequency f_r.



At parallel resonance, we have,

$$I_C = I_L \sin\phi_L$$

$$I_L = V/Z_L;\ \sin\phi_L = X_L/Z_L \text{ and } I_C = V/X_C$$

$$\frac{V}{X_C} = \frac{V}{Z_L} \times \frac{X_L}{Z_L}$$

$$X_L X_C = Z_L^2$$

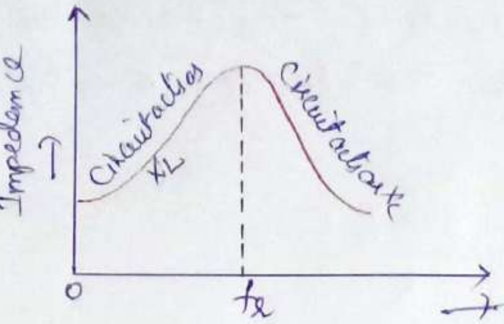
$$\frac{\omega L}{\omega C} = Z_L^2 = R^2 + X_L^2$$

$$\frac{L}{C} = R^2 + (2\pi f_r L)^2$$

$$(2\pi f_r L)^2 = \frac{L}{C} - R^2$$

$$2\pi f_r L = \sqrt{\frac{L}{C} - R^2}$$

$\therefore$ circuit impedance, $Z_r = \frac{L}{CR}$



- Thus at parallel resonance, the circuit impedance is equal to L/CR.
- It may be noted that Z_r will be in ohms if R, L & C are measured in ohms, henry and farad respectively.

ii) Circuit Current :-

At parallel resonance, the circuit or line current I is given by the applied voltage divided by the circuit impedance Z_r i.e.,

Line current, $I = \frac{V}{Z_r}$ where $Z_r = \frac{L}{CR}$

- Because Z_r is very high, the line current I will be very small.

iii) Quality factor Q :-

- It is desired that resonance curve of a parallel tuned circuit should be as sharp as possible in order to provide selectively.
- The sharp resonance curve means that impedance falls rapidly as the frequency is varied from the resonant frequency.
- The smaller the resistance of coil, the more sharp is the resonance curve. This is due to the fact that a small resistance consumes less power and draws a relatively small line current.
- The ratio of inductive reactance and resistance of the coil at resonance, therefore, becomes a measure of the quality of the tuned circuit. This is called quality factor and may be defined as

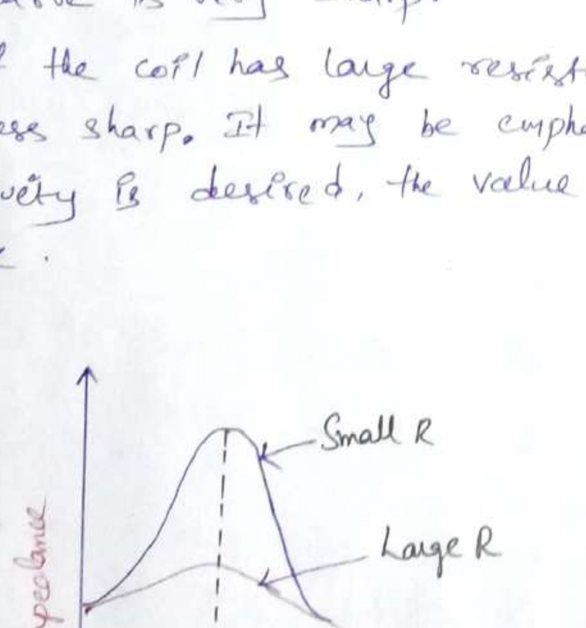
under :

- The ratio of inductive reactance of the coil at resonance to its resistance is known as quality factor Q i.e.,

$$Q = \frac{X_L}{R} = \frac{2\pi f_r L}{R}$$

- The quality factor Q of a parallel tuned circuit is very important because the sharpness of resonance curve and hence selectivity of the circuit depends upon it.
- The higher the value of Q, the more selective is the tuned circuit.
- It is clear that when the resistance is small, the resonance curve is very sharp.
- However, if the coil has large resistance, the resonance curve is less sharp. It may be emphasised that where high selectivity is desired, the value of Q should be very large.

Q



Q factor (or) Quality factor :-

It is defined as the ratio of reactance of coil to resistance of the coil.

$$Q = \frac{\omega L}{R} \text{ (or) } \frac{X_L}{R}$$

$$Q = 2\pi \times \frac{\text{maximum energy stored per cycle}}{\text{power dissipated per cycle}}$$

{Energy stored in the inductor $= \frac{1}{2} L I^2 = \frac{1}{2} L I_m^2$]

Enery stored in the capacitor $= \frac{1}{2} C V^2$

Power dissipation per cycle in inductor

$$\text{Energy} = \text{power} \times \text{time}$$

$$= \left(\frac{I_m}{\sqrt{2}}\right)^2 \times R \times T$$

$$= \frac{I_m^2}{2} \times R \times T$$

$$= \frac{I^2_m R}{2f}$$

$$Q = 2\pi \times \frac{\text{max. energy stored per cycle}}{\text{power dissipation per cycle}}$$

$$Q = 2\pi \times \frac{\frac{1}{2} L I_m^2}{\frac{I^2_m R}{2f}}$$

$$= \frac{2\pi L f}{R}$$

$$= \frac{2\pi f L}{R}$$

$$Q = \frac{\omega L}{R}$$

This is the Q-factor for inductor.

Q-factor for capacitor $= 1/\omega RC$.

Frequency Response of Tuned Amplifier.

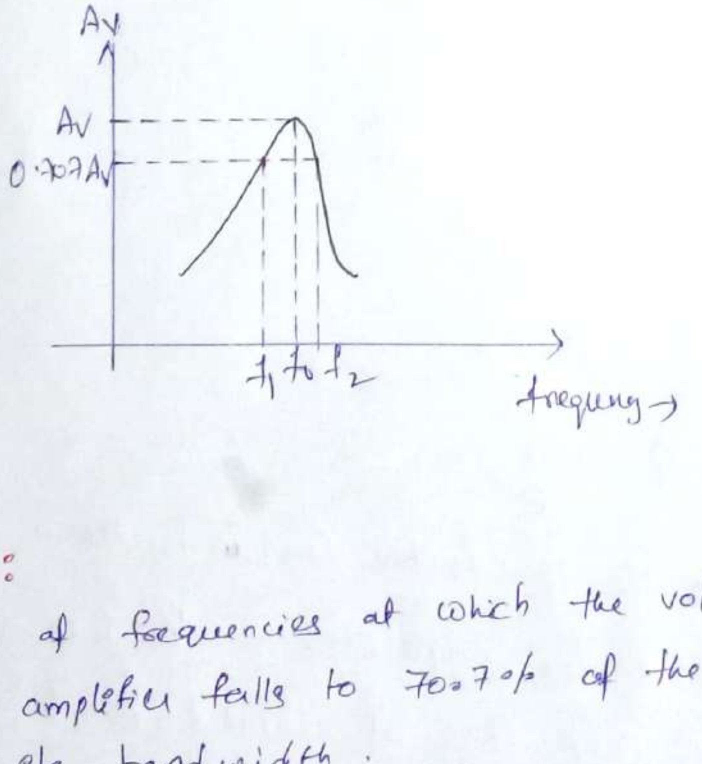
- The voltage gain of an amplifier depends upon β, input impedance and effective collector load.
- In a tuned amplifier, tuned circuit is used in the collector therefore, voltage gain of such an amplifier

is given by :

$$\text{voltage gain} = \frac{\beta Z_c}{Z_{in}}$$

where ZC = effective collector load
Z_{in} = input impedance of the amplifier.

- The value of ZC and hence gain strongly depends upon frequency in the tuned amplifier.
- As ZC is maximum at resonant frequency, therefore, voltage gain will be maximum at this frequency.
- The value of ZC and gain decrease as the frequency is varied above and below the resonant frequency.
- It is clear that voltage gain is maximum at resonant frequency and falls off as the frequency is varied in either direction from resonance.



Band width :

- The range of frequencies at which the voltage gain of the tuned amplifier falls to 70.7% of the maximum gain is called its bandwidth.
- The amplifier will amplify nearly any signal in this frequency range.
- The bandwidth of tuned amplifier depends upon the value of Q & LC circuit i.e. upon the sharpness

of the frequency response.

- The greater the value of Q of tuned circuit, the lesser is the bandwidth of the amplifier and vice-versa.
- In practice, the value of Q of LC circuit is made such so as to permit the amplification of desired narrow band of high frequencies.
- The practical importance of bandwidth of tuned amplifiers is found in communication system.
- In radio and TV transmission, a very high frequency wave, called carrier wave is used to carry the audio or picture signal.
- In radio transmission, the audio signal has a frequency range of 10 KHz.
- If the carrier wave frequency is 710 kHz, then the resultant radio wave has a frequency range between (710-5) KHz and (710+5) KHz.
- Consequently, the tuned amplifier must have a bandwidth of 705 kHz to 715 kHz (i.e 10 kHz).
- The Q of the tuned circuit should be such that bandwidth of the amplifier lies in this range.

Relation between Q & bandwidth:

- The quality factor Q of a tuned amplifier is equal to the ratio of resonant frequency (f_r) to bandwidth (BW) i.e.,

$$Q = \frac{f_r}{BW}$$

- The Q of an amplifier is determined by the circuit component values. It may be noted here that Q of a tuned amplifier is generally greater than 10.
- when this condition is met, the resonant frequency

at Parallel resonance is approximately given by:

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

SINGLE TUNED AMPLIFIERS:

* An amplifier circuit with a single tuned section being at the collector of the amplifier circuit is called as single tuned amplifier circuit.

CONSTRUCTION:

* A simple transistor amplifier circuit consisting of a parallel tuned circuit in its collector load, makes a single tuned amplifier circuit.
* The values of capacitance and inductance of the tuned circuit are selected such that its resonant frequency is equal to the frequency to be amplified
* The following circuit diagram shows a single tuned amplifier circuit.

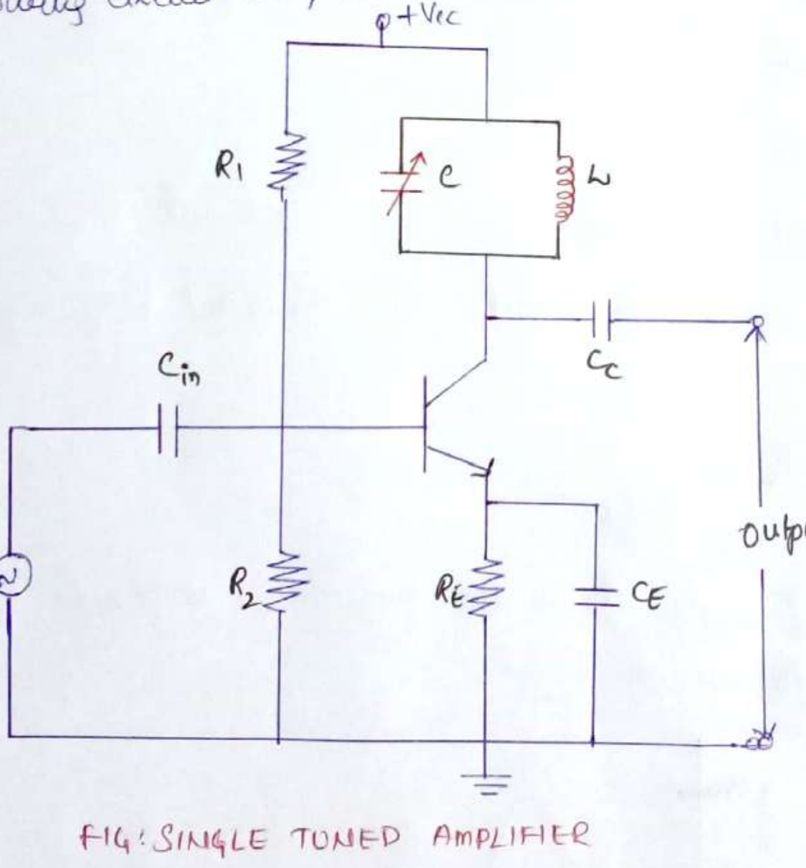


FIG: SINGLE TUNED AMPLIFIER

* The output can be obtained from the coupling capacitor C_c as shown above or from a secondary winding placed at L.

OPERATION:

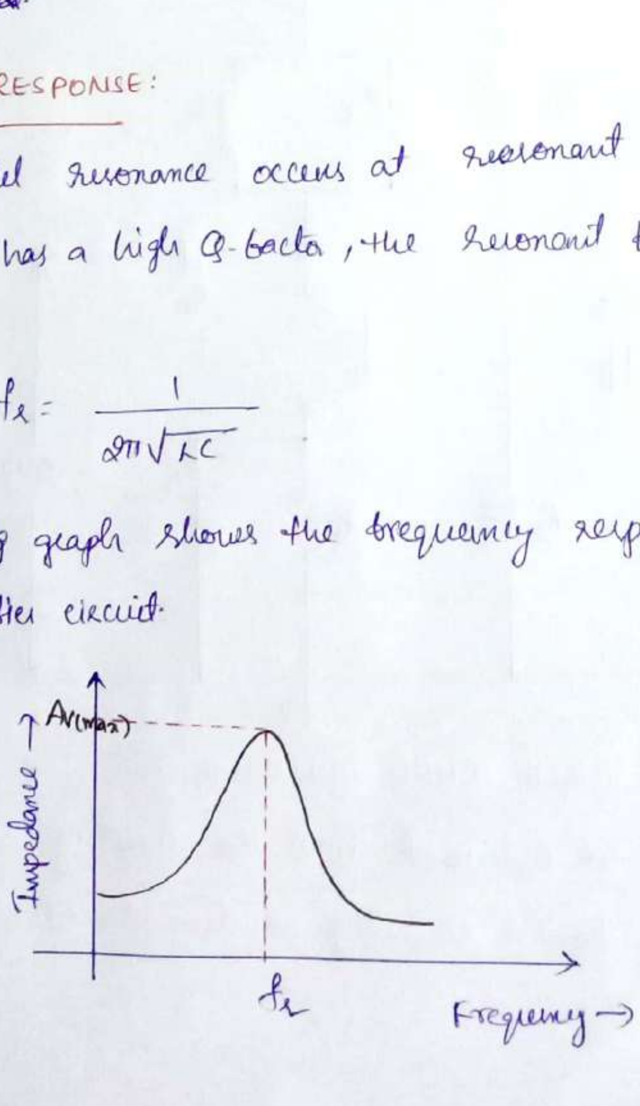
* The high frequency signal that has to be amplified is applied at the input of the amplifier.
* The resonant frequency of the parallel tuned circuit is made equal to the frequency of the signal applied by altering the Capacitance value of the Capacitor C, in the tuned circuit.
* At this stage, the tuned circuit offers high impedance to the signal frequency, which helps to offer high output across the tuned circuit.
* As high impedance is offered only for the tuned frequency, all the other frequencies which get lower impedance are rejected by the tuned circuit.
* Hence the tuned amplifier selects and amplifies the desired frequency signal.

FREQUENCY RESPONSE:

* The parallel resonance occurs at resonant frequency f_r when the circuit has a high Q-factor, the resonant frequency f_r, is given by,

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

* The following graph shows the frequency response of a single tuned amplifier circuit.



* At resonant frequency f_r, is the impedance of parallel tuned circuit is very high and is purely resistive.

* The voltage across R_L is therefore maximum, when the circuit is tuned to resonant frequency.
* Hence the voltage gain is maximum at resonant frequency and drops off above and below resonant frequency.
* The higher the Q, the narrower will the curve.

ANALYSIS:

* The tuned amplifiers are used for high frequency signals so hybrid-π model is used for analysis.

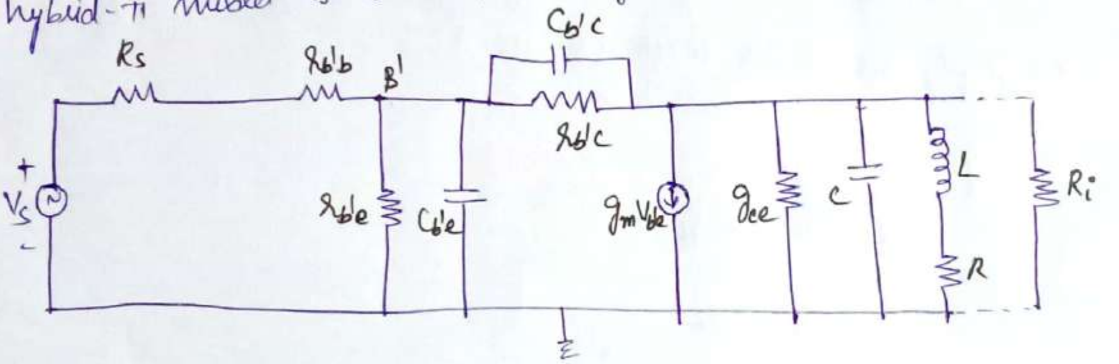
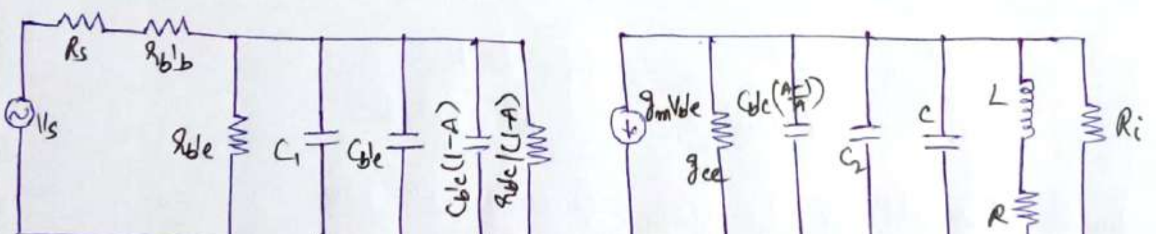
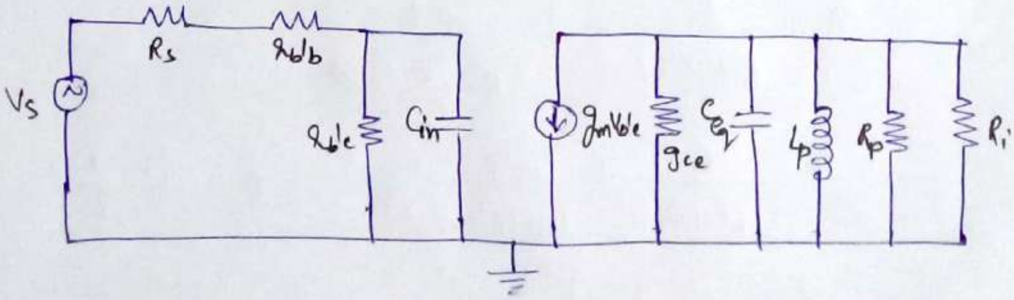


FIG: HYBRID π MODEL FOR SINGLE TUNED AMPLIFIER

* Apply miller's theorem then the circuit will be



* By neglecting Ct, Cb
* Upon approximations the modified circuit will be,



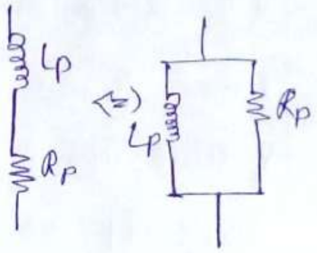
From the Circuit

1. $g_{ce} = \frac{1}{r_{ce}} = h_{oe} - g_m h_{re} = h_{oe}$

$$h_{oe} = \frac{1}{R_0}$$

2. $C_{in} = C_{b'e} + C_{b'c}(1-A)$

3. $C_{eq} = C_{b'c}\left(\frac{A-1}{A}\right) + C$



where C = tuned circuit capacitance

ADMITTANCE:

$$Y = \frac{1}{R + j\omega L}$$

multiply and divide with $(R - j\omega L)$

$$Y = \frac{R - j\omega L}{(R+j\omega L)(R - j\omega L)}$$

$$= \frac{R - j\omega L}{R^2 - (j\omega L)^2}$$

$$= \frac{R - j\omega L}{R^2 + \omega^2 L^2}$$

$$Y = \frac{R}{R^2+\omega^2 L^2} - j \frac{\omega L}{R^2+\omega^2 L^2}$$

divide and multiply the second term with ω

$$Y = \frac{R}{R^2+\omega^2 L^2} - j \cdot \frac{(\omega L)(\omega)}{(R^2+\omega^2 L^2)\omega}$$

$$Y = \frac{R}{R^2+\omega^2 L^2} - j \cdot \frac{\omega^2 L}{(R^2+\omega^2 L^2)\omega}$$

$$Y = \frac{R}{R^2+\omega^2 L^2} + \frac{\omega^2 L}{j\omega(R^2+\omega^2 L^2)} \quad \left(\because -j = \frac{1}{j}\right)$$

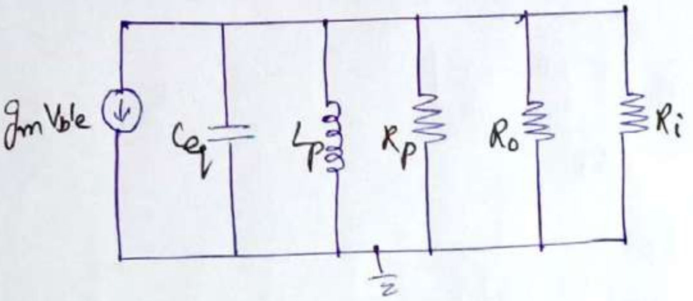
$$Y = \frac{1}{R_P} + \frac{1}{j\omega L_P}$$

where $R_p = \frac{R^2 + \omega^2 L^2}{R}$ and $L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L}$

$$R_p = \frac{R^2+\omega^2L^2}{R} = \frac{R^2}{R} + \frac{\omega^2L^2}{R} = \frac{\omega^2L^2}{R} \quad (\because \omega L >> R)$$

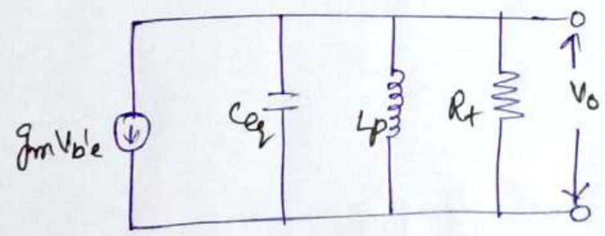
$$L_p = \frac{R^2+\omega^2L^2}{\omega^2 L} = \frac{R^2}{\omega^2 L} + \frac{\omega^2L^2}{\omega^2L} = L \quad \left(\because \omega L >> R, \ \frac{R^2}{\omega^2 L} << L\right)$$

* From the output circuit



$$\frac{1}{R_t} = \frac{1}{R_p} + \frac{1}{R_o} + \frac{1}{R_i}$$

where R_t is the total resistance.



* The effective Q-factor of the output circuit (loaded Q-factor) is defined as $Q_{effective}$.

$$Q_{eff} = \frac{\text{Susceptance of inductance (or) capacitance}}{\text{Conductance of shunt resistance } (R_t)}$$

Q_{eff} for inductor $L = \frac{R_t}{\omega L}$ $\quad$ Q_{eff} for capacitor $C = \omega_0 C R_t$

$$Y = \frac{1}{Z} = \frac{1}{R_t} + \frac{1}{j\omega L} + j\omega C$$

$$= \frac{1}{R_t}\left[1 + \frac{R_t}{j\omega L} + j\omega C R_t\right]$$

multiply numerator and denominator by ω_0

$$Y = \frac{\omega_0}{R_t\omega_0}\left[\frac{1}{R_t} 1 + \frac{R_t}{j\omega L} + j\omega C R_t\right]$$

$$= \frac{1}{R_t}\left[1 + \frac{\omega_0 R_t}{\omega_0 j\omega L} + \frac{j\omega\omega_0 C R_t}{\omega_0}\right]$$

$$= \frac{1}{R_t}\left[1 + \frac{Q_{eff}\,\omega_0}{j\omega} + \frac{jQ_{eff}\,\omega}{\omega_0}\right] \quad \left[\because Q_{eff} \text{ for inductor} = \frac{R_t}{\omega_0 L}\right.$$

Q_{eff} for capacitor $= \omega_0 C R_t]$

$$= \frac{1}{R_t}\left[1 + jQ_{eff}\left[-\frac{\omega_0}{\omega} + \frac{\omega}{\omega_0}\right]\right]$$

$$Y = \frac{1 + jQ_{eff}\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]}{R_t}$$

$$Z = \frac{1}{Y} = \frac{R_t}{1 + jQ_{eff}\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]}$$

* δ(delta) is the fractional change of the resonant frequency ω_0.

$$\delta = \frac{\omega - \omega_0}{\omega_0} = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega_0} = \frac{\omega}{\omega_0} - 1$$

$$\delta = \frac{\omega}{\omega_0} - 1$$

$$\frac{\omega}{\omega_0} = \delta + 1 \quad \text{and} \quad \frac{\omega_0}{\omega} = \frac{1}{\delta + 1}$$

Now, $$Z = \frac{R_t}{1 + jQ_{eff}\left[\delta + 1 - \frac{1}{\delta + 1}\right]}$$

$$= \frac{R_t}{1 + jQ_{eff}\left[\frac{(\delta+1)^2 - 1}{(\delta+1)}\right]}$$

$$= \frac{R_t}{1 + jQ_{eff}\left[\frac{\delta^2 + 2\delta + \cancel{1} - \cancel{1}}{\delta + 1}\right]}$$

$$= \frac{R_t}{1 + jQ_{eff}\left[\frac{\delta^2 + 2\delta}{\delta + 1}\right]} = \frac{R_t}{1 + jQ_{eff}2\delta\left[\frac{\delta/2 + 1}{\delta + 1}\right]}$$

* At any frequency, we close to the resonant frequency (ω_0)

$\delta << 1$ $(\because \omega - \omega_0 << 1)$

$$\boxed{Z = \frac{R_t}{1 + jQ_{eff} 2\delta}}$$

* From the circuit output voltage is given by,

$$V_o = -g_m V_{b'e} * Z = -g_m V_{b'e} * \frac{R_t}{1 + jQ_{eff} 2\delta}$$

$$V_{b'e} = V_i * \frac{r_{b'e}}{r_{b'e} + r_{b'b}}$$

$$A = \frac{V_o}{V_i} = \frac{-g_m * \cancel{V_i} * \frac{r_{b'e}}{r_{b'e} + r_{b'b}} * \frac{R_t}{1 + jQ_{eff} 2\delta}}{\cancel{V_i}}$$

$$\boxed{A = \frac{-g_m r_{b'e} R_t}{(r_{b'e} + r_{b'b})(1 + jQ_{eff} 2\delta)}}$$

→ This is the gain without resonance.

* The voltage gain at resonance is $\delta = 0$

$$A_{res} = \frac{-g_m \times r_{b'e} R_t}{(r_{b'e} + r_{b'b})(1 + jQ_{eff} 2(0))}$$

$$A_{res} = \frac{-g_m r_{b'e}}{r_{b'e} + r_{b'b}}$$

$$\frac{A}{A_{res}} = \frac{-g_m \cancel{r_{b'e} R_t} / (\cancel{r_{b'e} + r_{b'b}})(1 + jQ_{eff} 2\delta)}{-g_m \cancel{r_{b'e} R_t} / (\cancel{r_{b'e} + r_{b'b}})}$$

$$\frac{A}{A_{res}} = \frac{1}{1 + jQ_{eff} 2\delta}$$

$$\left|\frac{A}{A_{res}}\right| = \frac{1}{\sqrt{1 + Q_{eff}^2 2\delta^2}} \quad (\text{or}) \quad \boxed{\left|\frac{A}{A_{res}}\right| = \frac{1}{\sqrt{1 + (2\delta Q_{eff})^2}}}$$

EFFECT OF CASCADING SINGLE TUNED AMPLIFIER ON BANDWIDTH:

* When two or more stages are cascaded then overall gain will be increased and bandwidth will be decreased

$$Q = \frac{f_r}{BW}$$

* We know that $\left|\frac{A}{A_{res}}\right| = \frac{1}{\sqrt{1 + (2\delta Q_{eff})^2}}$

* The relative gain of n-stage cascaded amplifier is

$$\left|\frac{A}{A_{res}}\right|^n = \left[\frac{1}{\sqrt{1+(2\delta Q_{eff})^2}}\right]^n$$

The 3-dB frequency for cascaded amplifier is,

$$\left|\frac{A}{A_{res}}\right|^n = \frac{1}{\sqrt{1+2\delta}}$$

$$\left|\frac{A}{A_{res}}\right|^n = \frac{1}{\sqrt{2}}$$

$$\left[\frac{1}{\sqrt{1+(2\delta Q_{eff})^2}}\right]^n = \frac{1}{\sqrt{2}}$$

$$\left[\frac{1}{1+(2\delta Q_{eff})^2}\right]^{n \times 1/2} = \left(\frac{1}{2}\right)^{1/2}$$

$$\left[\frac{1}{1+(2\delta Q_{eff})^2}\right]^n = \frac{1}{2}$$

$$(1+(2\delta Q_{eff})^2)^n = 2$$

$$1+(2\delta Q_{eff})^2 = 2^{1/n}$$

$$(2\delta Q_{eff})^2 = 2^{1/n} - 1$$

$$2\delta Q_{eff} = \sqrt{2^{1/n} - 1}$$

$$\delta = \frac{\omega - \omega_0}{\omega_0} \text{ (or) } \frac{f - f_0}{f_0}$$

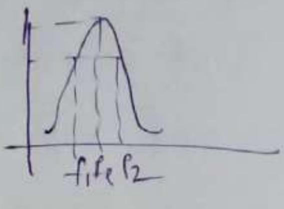
$$2\left(\frac{f-f_0}{f_0}\right) Q_{eff} = \pm\sqrt{2^{1/n}-1}$$

$$Q_{eff} = \pm \frac{f_0}{f-f_0}$$

$$(f-f_0) = \pm \frac{f_0}{2Q_{eff}}\sqrt{2^{1/n}-1}$$

* Let us assume f_1 and f_2 are upper and lower cut-off frequencies then

$$\text{Bandwidth} = f_2 - f_1$$



from the fig, $(f_2-f_1) = (f_2-f_0) - (f_0-f_1)$

$$f_2 - f_0 = +\frac{f_0}{2Q_{eff}} * \sqrt{2^{1/n}-1}$$

$$f_0 - f_1 = -\frac{f_0}{2Q_{eff}}\sqrt{2^{1/n}-1}$$

$$\therefore f_2 - f_1 = \frac{f_0}{2Q_{eff}} - \left(-\frac{f_0}{2Q_{eff}}\sqrt{2^{1/n}-1}\right)$$

$$= \left(\frac{f_0}{2Q_{eff}} + \frac{f_0}{2Q_{eff}}\right)\sqrt{2^{1/n}-1}$$

$$= \frac{\cancel{2} f_0}{\cancel{2} Q_{eff}}\sqrt{2^{1/n}-1}$$

$$f_2 - f_1 = \frac{f_0}{Q_{eff}}\sqrt{2^{1/n}-1}$$

$$\boxed{\therefore (B.W)_n = B.W\sqrt{2^{1/n}-1}} \qquad \left(\because BW = \frac{f_0}{Q_{eff}}\right)$$

ADVANTAGES:

1. The power loss is less due to the lack of collector resistance.
2. Selectivity is high
3. Voltage supply of the collector is small due to lack of R_c.

LIMITATIONS:

1. This tuned amplifier is required to be high selective. But high selectivity required a tuned circuit with a high Q-factor.

2. A high Q-factor circuit will give a high A_V but at the same time, it will give much reduced band width.

3. It means that tuned amplifier with reduced bandwidth may not be able to amplify equally the complete band of signals and result in poor reproduction. This is called "potential instability in tuned amplifier.

DOUBLE TUNED AMPLIFIER:

* An amplifier circuit with a double tuned circuits at the collector of the amplifier circuit is called as Double tuned amplifier circuit
* The problem of potential instability with a single tuned amplifiers overcome in a double tuned amplifier which consists of independently coupled two tuned circuit.

CONSTRUCTION:

* The circuit diagram for double tuned amplifier is as shown

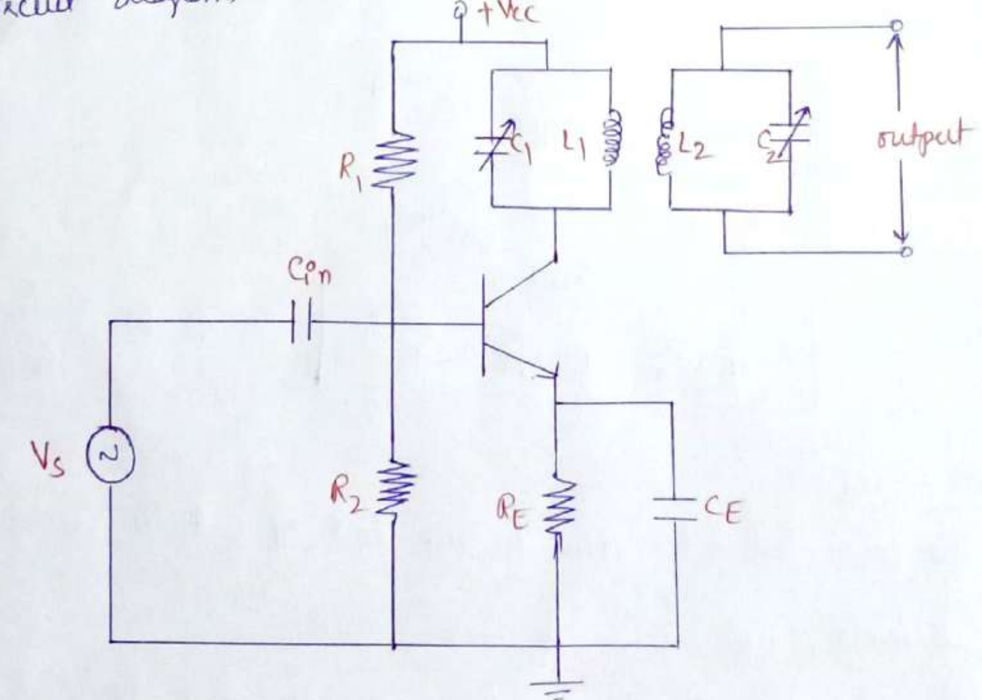


FIG: DOUBLE TUNED AMPLIFIER

* This circuit consists of two tuned circuits L_1C_1 and L_2C_2 in the collector section of the amplifier. 25

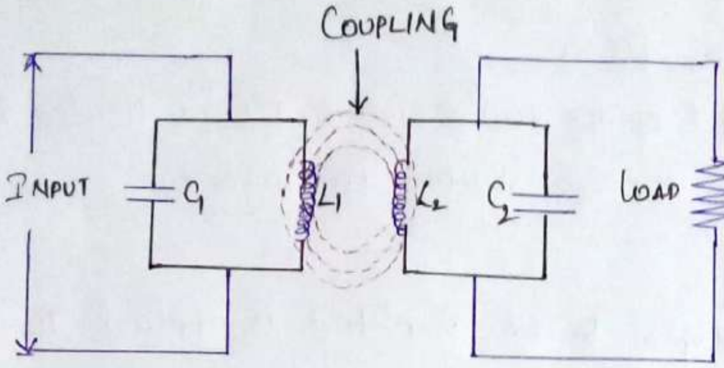
* The signal at the output of the tuned circuit L_1C_1 is coupled to the other tuned circuit L_2C_2 through mutual inductance.

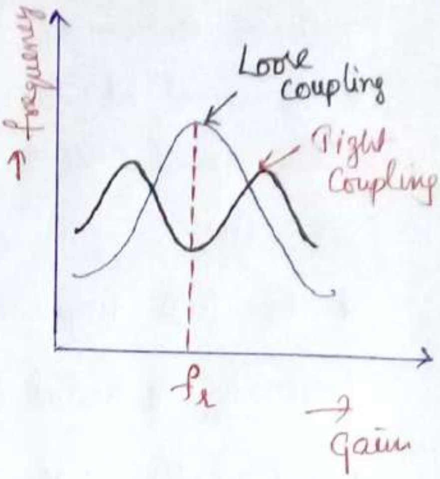
OPERATION:

* The high frequency signal to be amplified is applied. The resonant frequency of tuned circuit L_1C_1 is made equal to the signal frequency.
* Consequently large output appears across the tuned circuit L_1C_1. The o/p from this tuned circuit is transferred to the second tuned circuit L_2C_2 through mutual induction.
* Double tuned circuits are extensively used for coupling the various circuits of radio and television receivers.

FREQUENCY RESPONSE:

* The frequency response of a double tuned circuit depends upon the degree of coupling i.e., upon the amount of mutual inductance between the two tuned circuits.
* When coil L_2 is coupled to coil L_1, a portion of load resistance had been/added is coupled into the primary tank circuit L_1C_1 and affects the primary circuit.
* When the coils are spaced apart, all the primary coil L_1 flux will not link the secondary coil L_2. The coils are said to have "Loose Coupling".
* Under such conditions, the resistance reflected from the load i.e., secondary circuit is small.
* The resonance curve will be sharp and the circuit Q is high.
* When the primary and secondary coils are very close together, then it is called "Tight Coupling".
* Under such conditions, the reflected resistance will be large and the circuit Q is lower.





BANDWIDTH:

* From the frequency response of double tuned amplifier, it is clear that bandwidth increases with the degree of Coupling.
* obviously, the determining factor in a double tuned circuit is not Q-factor but the coupling. For a given frequency, the tighter the coupling, the greater is the bandwidth.

$$\boxed{B.W = K f_r}$$

Here, K is the Coefficient of Coupling.

Effect of Cascading double tuned amplifiers on Bandwidth:

* When a number of identical double tuned amplifiers are connected in cascade the overall bandwidth of a System is narrowed.

$$\boxed{B.W_n = BW(2^{1/n}-1)^{1/4}}$$

ADVANTAGES:

* Steeper sides in the Curve
* flat top The main advantage of a double-tuned amplifier is an amplifier including a tuned circuit on the i/p and o/p.
* It has narrow bandwidth.
* Impedance matching using previous phase
* 3dB Bandwidth is large.
* Selectivity is improved.

DISADVANTAGES:

* There are not suitable for amplifying audio frequencies.
* If the frequency band increases, then this design becomes complex.
* Two LC circuits tunes separately. the alignment is difficult.
* The design uses tuning elements like capacitors and inductors, then the circuit is costly and bulky.

STAGGER TUNED AMPLIFIERS:

* Stagger tuned amplifier is a cascaded stage single tuned amplifier designed to improve the total frequency response of the tuned amplifier
* The total frequency response of this amplifier can be achieved by adding up the separate response as one.
* When the different tuned circuits resonant frequencies are staggered otherwise displaced, then it is known as stagger tuned amplifier

WORKING:

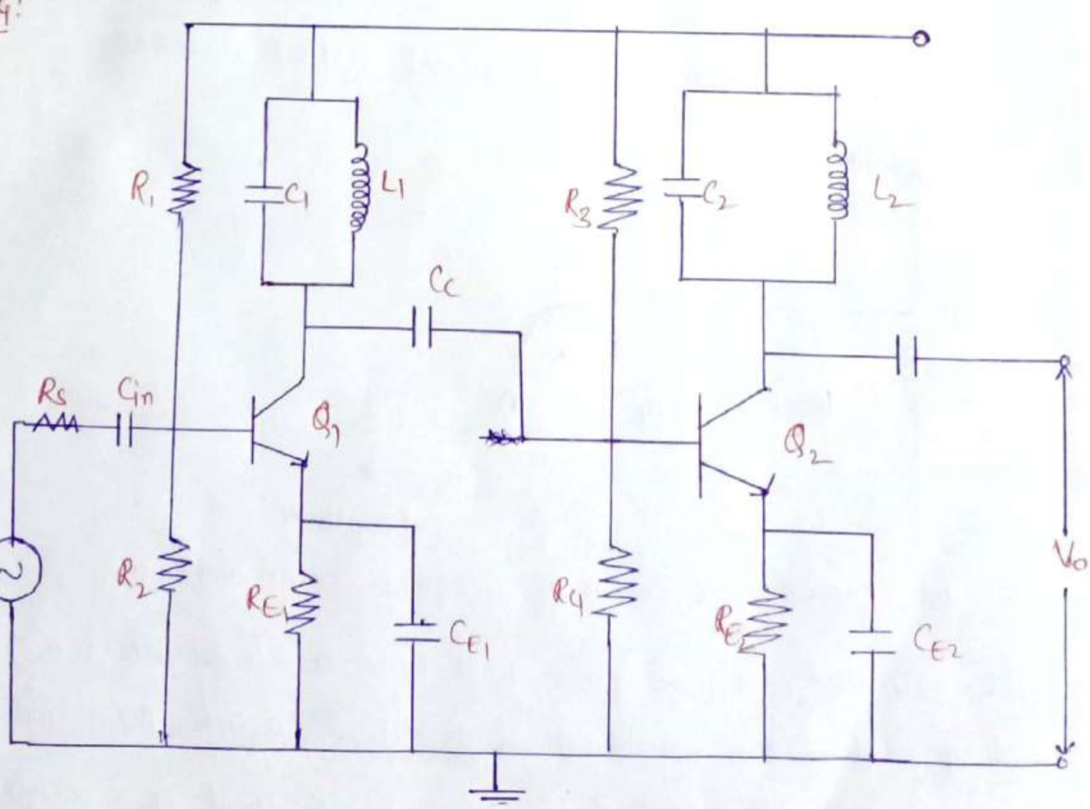


FIG: STAGGER TUNED AMPLIFIER

* In order to increase bandwidth, double tuned amplifiers are preferred. but alignment of double tuned amplifiers is difficult.
* In stagger tuned circuits, two single tuned cascade amplifiers having a certain bandwidth are taken.
* The resonant frequencies of the two tuned circuits are so adjusted that they are separated by an amount equal to the bandwidth of each stage.
* Since the resonant frequencies are displaced or staggered, they are known as stagger tuned circuits.

FREQUENCY RESPONSE:

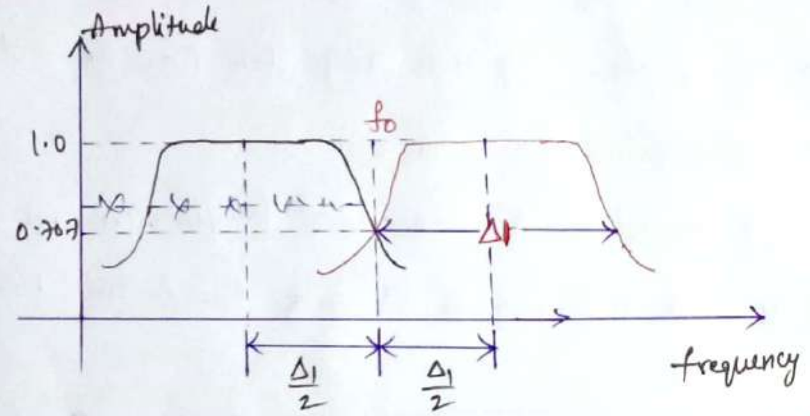


FIG: RESPONSE OF INDIVIDUAL TUNED AMPLIFIER.

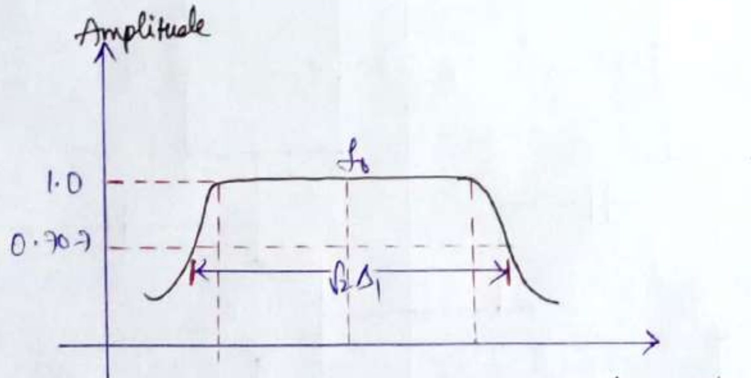


FIG: OVERALL RESPONSE OF STAGGER TUNED AMPLIFIER.

* The resultant staggered pair will have a bandwidth i.e., $\sqrt{2}$ times that of each of the individual single tuned circuits. The overall selectivity function will be identical in form with that of a single stage double tuned system.

* The relative gain of a single tuned direct coupled amplifier is given by,

$$\frac{A}{A_{res}} = \frac{1}{1+j2\delta Q_{eff}}$$

Let $$\frac{A}{A_{res}} = \frac{1}{1+jX} \quad \text{where } X = 2\delta Q_{eff}$$

* As Bandwidth is $B = \frac{f_0}{Q_{eff}}$ and under 3 dB frequency $\delta = \frac{1}{2Q_{eff}}$, the equation for Bandwidth can be written as $B = 2\delta f_0$.

* Since one stage is tuned to a frequency $\delta_0 f_0$ below f_0 and the other stage is tuned to a frequency $\delta_0 f_0$ above f_0.

* The corresponding selectivity function of the circuits are

$$\left(\frac{A}{A_{res}}\right)_1 = \frac{1}{1+j(X-1)} \quad \text{and} \quad \left(\frac{A}{A_{res}}\right)_2 = \frac{1}{1+j(X+1)}$$

* The overall gain function becomes,

$$\left(\frac{A}{A_{res}}\right)_{pair} = \left(\frac{A}{A_{res}}\right)_1 * \left(\frac{A}{A_{res}}\right)_2$$

$$= \left[\frac{1}{1+j(X-1)}\right] * \left[\frac{1}{1+j(X+1)}\right]$$

$$= \frac{1}{(1+j(X-1))(1+j(X+1))}$$

$$= \frac{1}{1+j(X+1)+j(X-1)+j^2(X+1)(X-1)}$$

$$= \frac{1}{1+j(X+1+X-1)+j^2(X^2-1)}$$

$$= \frac{1}{1+j2X-X^2+1}$$

$$= \frac{1}{2+j2X-X^2}$$

$$= \frac{1}{(-X^2+2)+j2X}$$

The magnitude of the resulting function is,

$$\left|\left(\frac{A}{A_{res}}\right)_{pair}\right| = \frac{1}{\sqrt{(2-x^2)^2+(2x)^2}}$$

$$= \frac{1}{\sqrt{4+x^4\cancel{-4x^2}+\cancel{4x^2}}}$$

$$= \frac{1}{\sqrt{4+x^4}}$$

$$= \frac{1}{\sqrt{4+(2\delta Q_{eff})^4}} \qquad (\because x = 2\delta Q_{eff})$$

$$= \frac{1}{\sqrt{4(1+4\delta Q_{eff})^4}}$$

$$\boxed{\left|\left(\frac{A}{A_{res}}\right)_{pair}\right| = \frac{1}{2}\,\frac{1}{\sqrt{1+4\delta^4 Q_{eff}^4}}}$$

ADVANTAGES:

* Increased Bandwidth compared to single tuned amplifier
* High gain Bandwidth product.
* Enhanced Stability

Stability of Tuned Amplifiers.

Stability → Frequency < Low: distortion will occur; High: because of non-linear transistor

Distortion occur in low frequency

1. Non-linear < Amplitude harmonic distortion; Frequency Harmonic distortion
2. Linear

⇒ we are used $C_{b'c}$ in radio frequency Tuned Circuit. From this we are provided positive (or) negative feedback. If we provided positive feed back we get instability.

⇒ To Avoid instability we are connected another Component from base to collected. It provides neutralization. The Neutralization process introduced by L.A. Hazeltine. There are ~~the four~~ types of Neutralization.

1. Hazeltine Neutralization.
2. Neutrodyne Neutralization
3. Neutralization with Coil.

~~4~~

8.9 NEUTRALIZATION

The technique used for the elimination of potential oscillations is called neutralization. *BJT* and *FET* are potentially unstable over some frequency range due to the feedback parameter (Y_N) present in them. If the feedback can be cancelled by an additional feedback signal that is equal in amplitude and opposite in sign, the transistor becomes unilateral from input to output the oscillations completely stop. This is achieved by Neutralization. The small-signal equivalent circuit of a BJT is shown in Fig. 8.28.

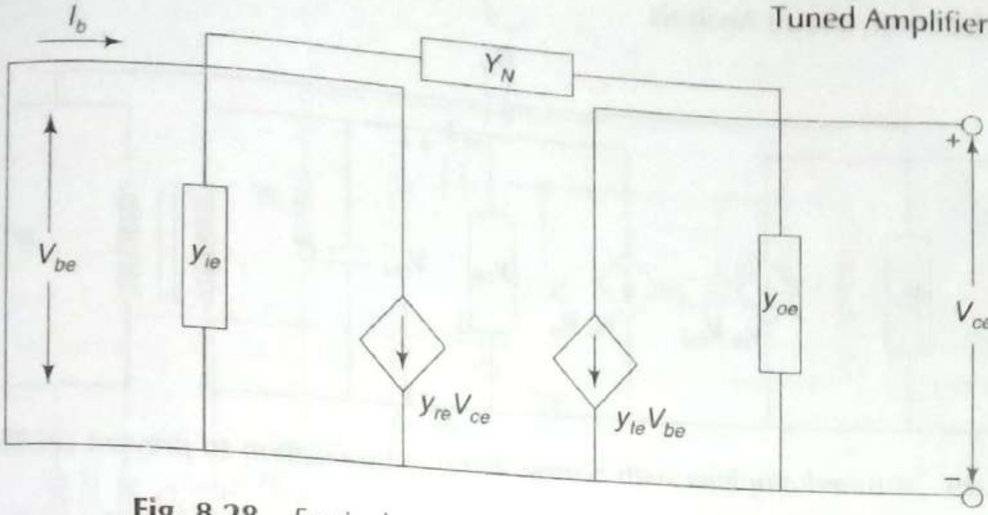


Fig. 8.28 *Equivalent circuit of a neutralized transistor*

From the definition of admittance parameters,

$$y'_{re} = \left.\frac{I_b}{V_{ce}}\right|_{V_{fe}=0}$$

and with the input terminal shorted

$$I_b = y_{re}V_{ce} - Y_N V_{ce}$$

Therefore,

$$I_b = V_{ce}[y_{re} - Y_N]$$

i.e.

$$\frac{I_b}{V_{ce}} = y_{re} - Y_N$$

Comparing the above equations, we get

$$y'_{re} = y_{re} - Y_N$$

For perfect neutralization, $y'_{re} = 0$. Therefore, $y_{re} = Y_N$. This indicates that oscillation does not exist if the designed circuit element matches y_{re} for all values of frequency and operating conditions. y_{re} is a nonlinear parameter which is given by

$$y_{re} = -j\omega C_{re}$$

which implies C_{re} is a slow varying function of both operating point and frequency. Hence the desired component that is used for neutralization is a *negative capacitor*. Since the fabrication of capacitor is complex, an inductor with negative susceptance, $B = -j(1/\omega L)$ is perferred, which has the inverse frequency dependence characteristics. Moreover, the inductor acts as a short circuit at dc condition and need not be considered. One practical approach to this problem is to use a fixed capacitance that is transformer coupled for 180°phase shift to produce neutralization over a limited frequency range. One such circuit is shown in Fig. 8.29. Here, perfect neutralization is not possible and the problems created by limited neutralization may exceed their benefits.

Hazeltine neutralization method This is a neutralization technique employed in tuned *RF* amplifiers to maintain stability. In the circuit shown in Fig. 8.30, the undesired effect of the collector to base capacitance of the transistor is neutralized

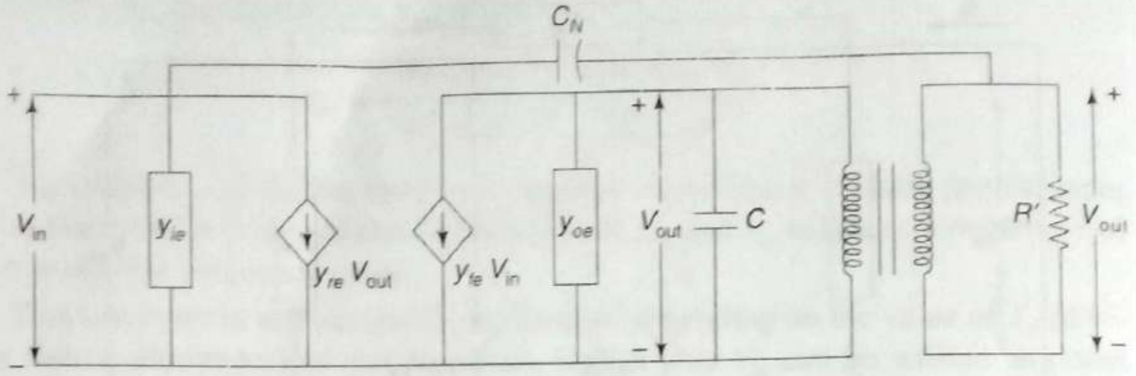


Fig. 8.29 *A tuned amplifier with narrow band neutralization to prevent oscillations*

by introducing a signal which cancels the signal coupled through the collector to base capacitance.

The Fig. 8.30 shows that a small variable capacitance C_N is connected from the bottom of the coil to the base of the transistor. The neutralization process is achieved by C_N. It introduces a signal to the base of the transistor such that is cancels out the signal fed to the base by C_{bc}. Generally a variable capacitor is used for neutralization as the value of C_{bc} changes with time. By properly adjusting C_N, exact neutralization is achieved.

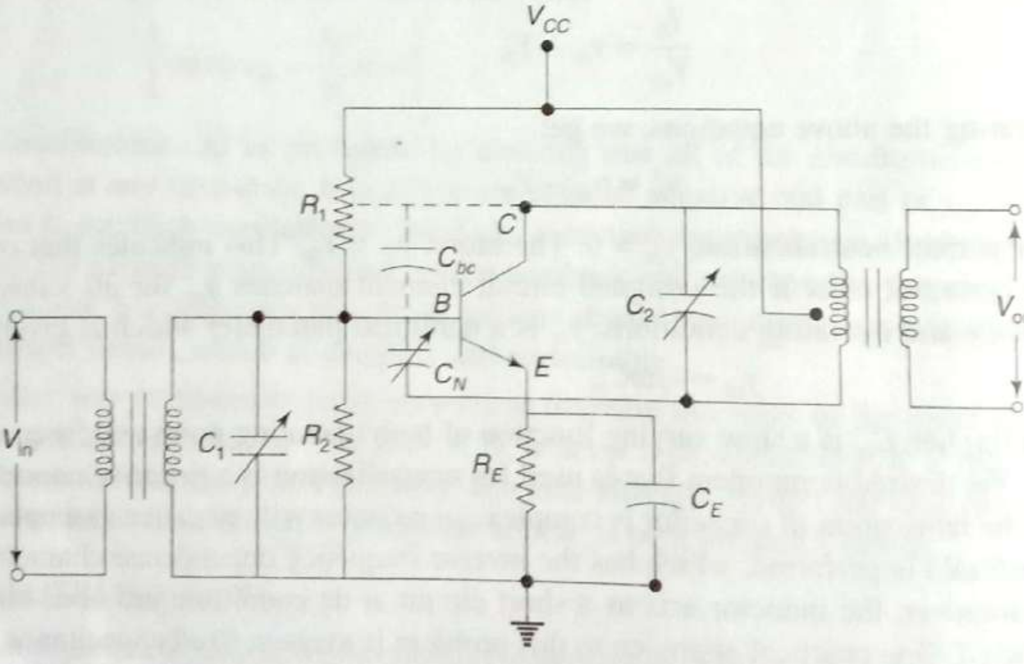


Fig. 8.30 *Tuned RF amplifier with Hazeltine neutralization (C_N = neutralization capacitance)*

In the modified version of Hazeltine neutralization called Neutrodyne neutralization technique, C_N is connected to the lower end of the secondary coil of the next stage. Hence it is connected with V_{CC} which ensures that, it is insensitive to any variation in the supply voltage V_{CC} and provides higher stabilization for the tuned amplifier. The circuit for the same is shown in Fig. 8.31.

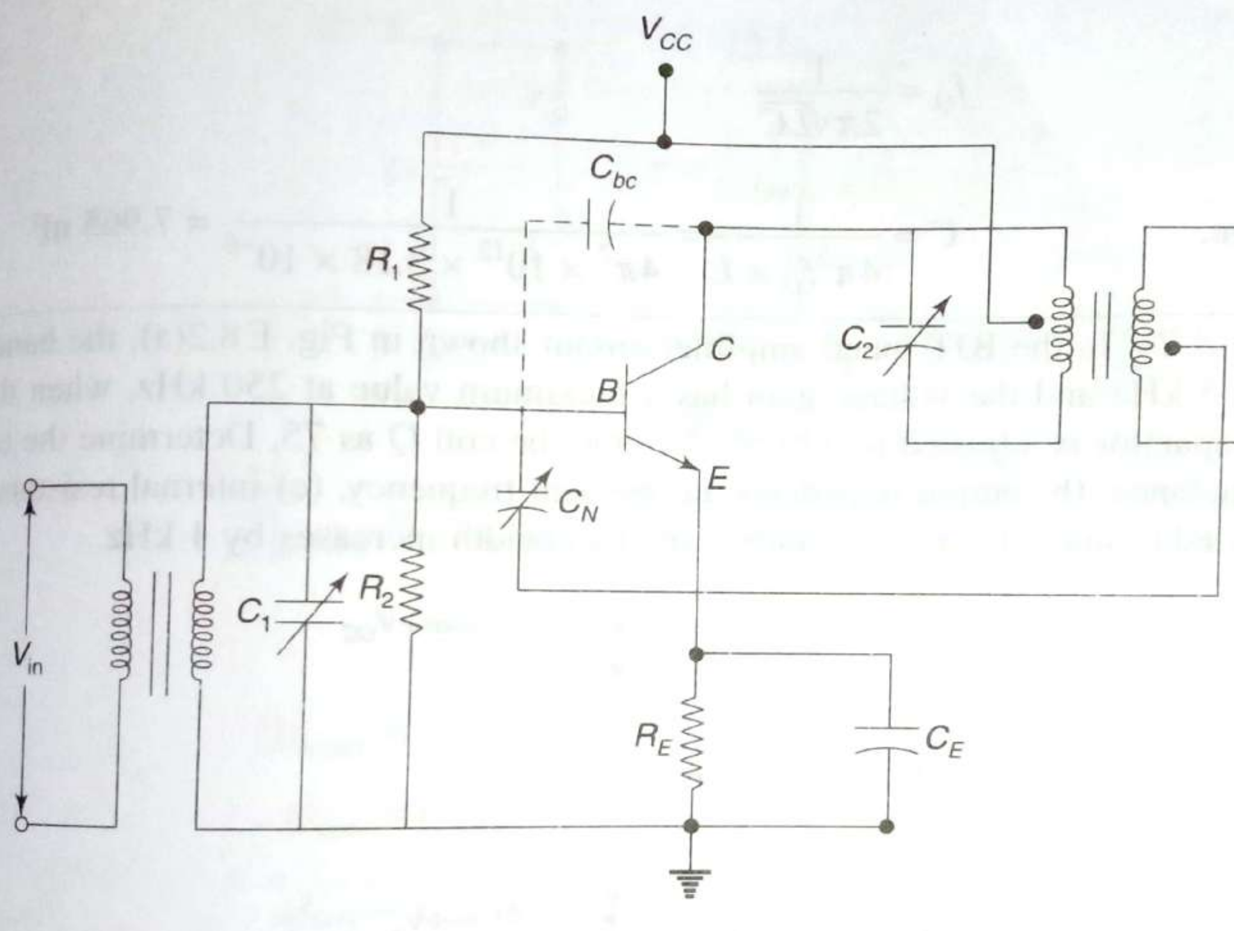


Fig. 8.31 *Modified Hazeltine—neutrodyne neutralization technique*