ELECTRONIC CIRCUIT ANALYSIS

UNIT- 1 MULTISTAGE AMPLIFIER

Syllabus : Introduction,
Classification of amplifiers.
Analysis of Cascaded amplifiers
Different coupling Schemes used in Amplifiers
Analysis of two stage RC coupled Amplifier
High input resistance transistor amplifiers -
(1) Darlington Pair Amplifier.
(2) Boot strap Emitter Follower
(3) Cascode amplifier
(4) Differential Amplifier.
Analysis of multi stage amplifiers using FET.

Introduction

Amplifier Electronic circuit that increases small AC into large AC with DC biasing voltage

(or)

An electronic amplifier circuit is one, which modifies the characteristics of the input signal, when delivered the output side. The modification in the characteristics of the input signal can be with respect to voltage, current, power or phase. Any one or all these characteristics power, or phase may be changed by the amplifier circuit. (or)

→ Device used to increase the level of the input ac

Or

Strengthen the weak signal i.e either voltage (or) current (or) power signal.

Classification of Amplifiers.

Based on

- ↳ Active devices used
- ↳ Operating frequencies
- ↳ Primary function
- ↳ Type of feedbacks
- ↳ No of stages
- ↳ Period of conduction
- ↳ Type of load
- ↳ Based on Band width

↳ Based on Active devices used

- BJT Amplifier
  - CB Amplifiers
  - CE Amplifiers
  - CC Amplifiers
- FET Amplifier
  - CG Amplifier
  - CS Amplifier
  - CD Amplifier.

(2)

↳ Based on Operating Frequencies
- (Sound) Audio frequency Amplifier <20 kHz — 20 Hz to 20,000 Hz
- Radio frequency Amplifier 10 kHz to 100,000 M — 30 Hz to 300 GHz
- Video frequency Amplifier — 30 MHz to 300 M
- Ultra high frequency Amplifier

AF - <20 kHz
b/w RF — 10 kHz — 100,000 MHz
VF → 2 — 100 MHz
UHF → 1 MHz to 500 MHz

↳ Based on Primary function
- Voltage Amplifier ($A_v = \frac{V_o}{V_i}$)
- Current Amplifier ($A_i = \frac{I_o}{I_i}$)
- Transresistance Amplifier ($R_m = \frac{V_o}{I_i}$ Ω)
- ~~Transconductance~~ Transconductance Amplifier ($G_m = \frac{I_o}{V_i}$ ℧)

↳ Based on Type of feedback
- positive feed back (oscillators)
- Negative feed back (Amplifiers)
  - Voltage series fb amp
  - Current series fb amp
  - voltage shunt fb amp
  - Current shunt fb amp

↳ Based on no. of stages
- Single stage amplifier
- Multi stage amplifier
  - Direct coupled amplifier
  - RC coupled amplifier
  - Transformed coupled amplifier
  - Impedance coupled amplifier.

↳ Based on period of conduction.
- Class A power amplifier (In which active device conducts for the full 360° so $\theta = 360°$
- class B power amplifier (conduction for 180°. $\theta = 180°$)
- class AB power amplifier ($180° < \theta < 360°$
- class C power amplifier ($90° < \theta < 180°$)

↳ Based on type of load — Tuned amplifiers (reactive load) (at uses one or more L - C circuits as load impedance) / Untuned amplifiers (resistive load) pure resistor or complex impedance

↳ Based on Band width — Narrow band Amplifier (normally RF amp) / Wide band amplifier (normally video amplifier)

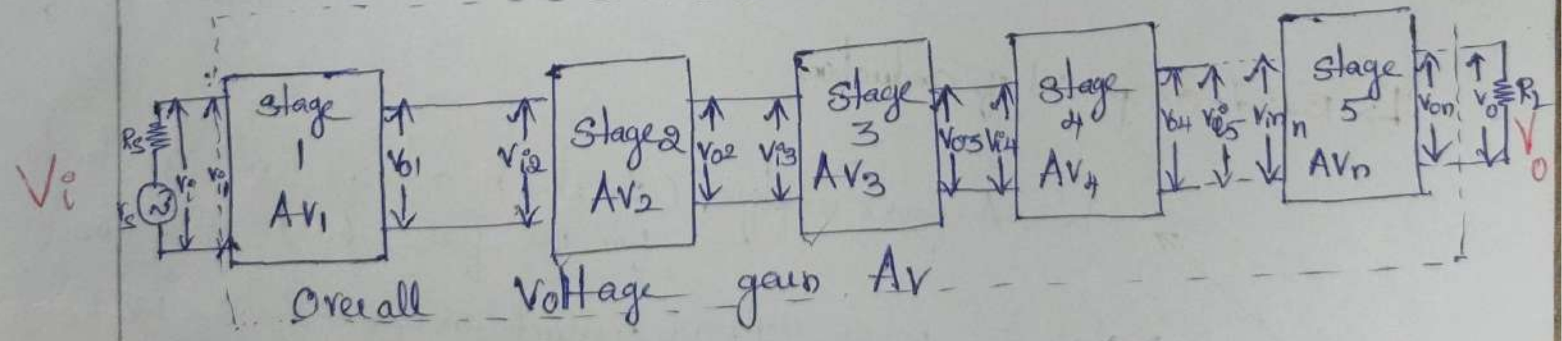
~~MULTISTAGE A~~

MULTISTAGE AMPLIFIER / CASCADED AMPLIFIERS

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

In Multi-stage amplifiers, the output of first stage is coupled to the input of next stage directly or any coupling devices. These coupling devices can be usually be a capacitor or a transformer.

The process of joining two amplifier stages using a is called as Cascading



$V_S \rightarrow$ Source voltage

$R_S \rightarrow$ Source resistance

$V_O \rightarrow$ output voltage

$R_L \rightarrow$ Load resistance

Where A_V = overall gain

$A_{V_1} \rightarrow$ Voltage gain of Ist stage

$A_{V_2} \rightarrow$ " " " IInd stage

$A_{V_3} \rightarrow$ " " " IIIrd stage

$A_{V_4} \rightarrow$ " " " IVth stage

$A_{V_n} \rightarrow$ " " " nth stage

The overall input $V_i = V_{i1}$, overall output $V_o = V_{on}$

Cascade connection $V_{o1} = V_{i2}$; $V_{o2} = V_{i3}$, $V_{o3} = V_{i4}$, $V_{o4} = V_{i5}$ ----

---- $V_{on-1} = V_{in}$

By definition of voltage gain, for all stages

$A_{V1} = \frac{V_{o1}}{V_{i1}}$, $A_{V2} = \frac{V_{o2}}{V_{i2}}$, $A_{V3} = \frac{V_{o3}}{V_{i3}}$, $A_{V4} = \frac{V_{o4}}{V_{i4}}$, --- $A_{Vn} = \frac{V_{on}}{V_{in}}$

Over all voltage gain $A_V = \frac{V_o}{V_i}$

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

$A_V = A_{V1} \times A_{V2} \times A_{V3} \times A_{V4} \cdots A_{Vn} = \frac{V_{o1}}{V_{i1}} \times \frac{V_{o2}}{V_{i2}} \times \frac{V_{o3}}{V_{i3}} \times \frac{V_{o4}}{V_{i4}} \times \cdots \times \frac{V_{on}}{V_{in}}$

$= \frac{V_{on}}{V_{i1}} = \frac{V_o}{V_i} = A_V$

$$\boxed{\therefore A_V = A_{V1} \cdot A_{V2} \cdot A_{V3} \times A_{V4} \cdots A_{Vn}}$$

Taking $20\log_{10}$ on both sides

$20\log_{10} A_V = 20\log_{10}[A_{V1} \cdot A_{V2} \cdot A_{V3} \cdot A_{V4} \cdots A_{Vn}]$

$= 20\log_{10} A_{V1} + 20\log_{10} A_{V2} + 20\log_{10} A_{V3} + \cdots + 20\log_{10} A_{Vn}$

Hence

$$\boxed{A_V(dB) = A_{V1}(dB) + A_{V2}(dB) + A_{V3}(dB) + \cdots + A_{Vn}(dB)}$$

Let $\theta_1, \theta_2, \theta_3, \theta_4$ -- θ_n & θ be the phase shift b/w the input and output voltages of stage 1, stage 2, stage 3, stage 4, --- stage n and multistage amplifier respectively.

$$\boxed{\text{Then } \theta = \theta_1 + \theta_2 + \theta_3 + \theta_4 + \cdots + \theta_n}$$

NOTE!- if all 'n' stages are identical with midband voltage gain A_m

Then $A_v = A_m \cdot A_m \cdot A_m \cdots A_m$

$\boxed{A_v = A_m^2}$ — overall voltage gain increases.

Problems !-

(1) Three identical stages are cascaded and have an overall an upper 3dB frequency of 20KHz and overall

(1) Individual voltage gains of 4 stage cascade amplifier are 30, 15, 20 and 40 respectively. Find overall voltage gain and also overall voltage gain in dB

$A_{v1} = 30, \quad A_{v2} = 15, \quad A_{v3} = 20, \quad A_{v4} = 40, \quad n = 4$

To find

overall voltage gain $A_v = ?$

$A_v = A_{v1} \cdot A_{v2} \cdot A_{v3} \cdot A_{v4} = 30 \times 15 \times 20 \times 40 \Rightarrow \boxed{A_v = 360000}$

overall gain in dB

$20\log_{10} A_v = 20\log_{10} A_{v1} + 20\log_{10} A_{v2} + 20\log_{10} A_{v3} + 20\log_{10} A_{v4}$

$= 29.54 + 23.52 + 26.02 + 32.04$

$A_v(dB) = 111.12$ (dB)

(2) Individual voltage gains in dB of a 5 stage cascaded amplifier are 10, 20, 30, 40, 50, respectively. Find overall voltage gain in dB and also overall voltage gain

$n = 5$

$A_{v1}(dB) = 10, \ A_{v2}(dB) = 20 \ A_{v3}(dB) = 30 \ A_{v4}(dB) = 40, \ A_{v5}(dB) = 50$

→ overall voltage gain in dB is

$A_v(dB) = 10 + 20 + 30 + 40 + 50 = 150$

→ overall voltage gain $A_v = ?$

W.K.T $20\log_{10} A_v = A_v(dB)$

$20\log_{10} A_v = 150$

$\log_{10} A_v = 150/20$

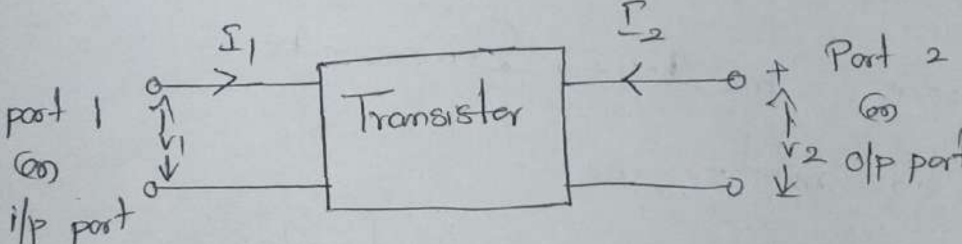
$\log_{10} A_v = 7.5$

$A_v = 10^{7.5} = 3,16,22,776.6$

Two Port Network Devices & its Parameters

A two port network is a four-terminal circuit in which the terminals are paired to form an input port and an output port. Different models for two port networks are

Model Name	Express	In terms of	Defining equations
(1) Impedance	V_1, V_2	I_1, I_2	$V_1 = Z_{11} I_1 + Z_{12} I_2$ & $V_2 = Z_{21} I_1 + Z_{22} I_2$
(2) Admittance	I_1, I_2	V_1, V_2	$I_1 = Y_{11} V_1 + Y_{12} V_2$ & $I_2 = Y_{21} V_1 + Y_{22} V_2$
(3) Hybrid	V_1, I_2	I_1, V_2	$V_1 = h_{11} I_1 + h_{12} V_2$ & $I_2 = h_{21} I_1 + h_{22} V_2$
(4) Transmission	V_1, I_1	$V_2, -I_2$	$V_1 = A V_2 - B I_2$ & $I_1 = C V_2 - D I_2$



A transistor can be treated as a two port network. The terminal behaviour of any two port network can be specified by the terminal voltages V_1 & V_2 at port 1 & port 2 respectively & the current i_1 & i_2 entering port 1 & 2 respective.

Hybrid Parameters: (or) h-Parameters

From the advance circuit theory voltages and currents in figure can be related by the following set of equations: Here i_1 & V_2 are taken as independent variables & V_1 & i_2 dependent variables.

$$V_1 = h_{11} i_1 + h_{12} v_2 \longrightarrow (1)$$
$$i_2 = h_{21} i_1 + h_{22} v_2 \longrightarrow (2)$$

$\cancel{V_1 = h_{11} i_1 + h v_2}$ $h_{11} = \left[\frac{V_1}{i_1}\right]_{v_2 = 0}$ (i/p impedance with Port 2 short circuit) $\rightarrow \Omega$

$h_{21} = \left[\frac{i_2}{i_1}\right]_{v_2 = 0}$ [forward current gain with output port short circuit] $\rightarrow$ Dimension less

$h_{12} = \left[\frac{V_1}{V_2}\right]_{i_1 = 0}$ [Reverse voltage transfer ratio with input port open circuit] $\rightarrow$ Dimension less.

$h_{22} = \left[\frac{i_2}{v_2}\right]_{i_1 = 0}$ [output admittance with input port open circuit] $\longrightarrow \Omega^{-}$ or $\mho$

The dimensions of h-parameters are

$h_{11} = \Omega$; $h_{22}\ \mho$; h_{12} $h_{21} \rightarrow$ Dimension less

IEEE recommended notations are

$i = 11 =$ input ; $r = 12 =$ reverse transfer
$o = 22 =$ output ; $f = 21 =$ forward transfer

hybrid model for two port network

Based on the definition of hybrid parameters the mathematical model for two-port networks known as h-parameter model can be developed

$$V_1 = h_i i_1 + h_r v_2 \rightarrow (3)$$
$$i_2 = h_f i_1 + h_o v_2 \rightarrow (4)$$

$V_1 = \underbrace{h_i i}_{V} + \overbrace{h_r v_2}^{\text{vlg gain}}$ $\hookrightarrow$ unit less

$i_2 = \underbrace{h_f i}_{\text{unitless } A} + h_o v_2$ $\hookrightarrow$ $ohm^{-1} \times V_2 = V/R \rightarrow$ Amp

The proposed model shown in figure below should satisfy these two equations and it can be readily verified by writing Kirchhoff's voltage law equations in the input loop and Kirchhoff's current law equations for the output node. It is to be noted that the input circuit have a dependent voltage generator and the output circuit contains a dependent current generator.

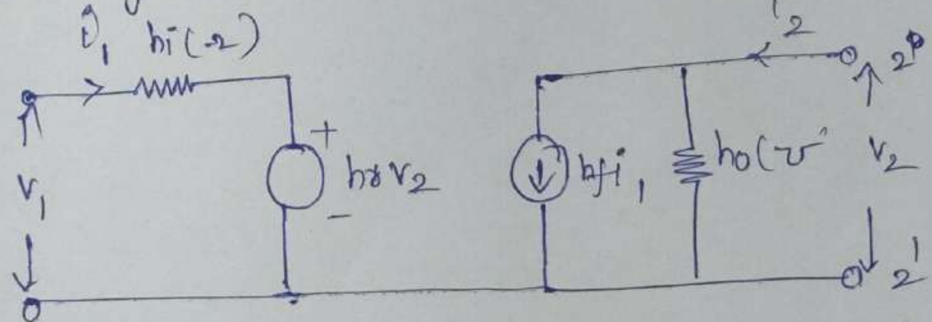
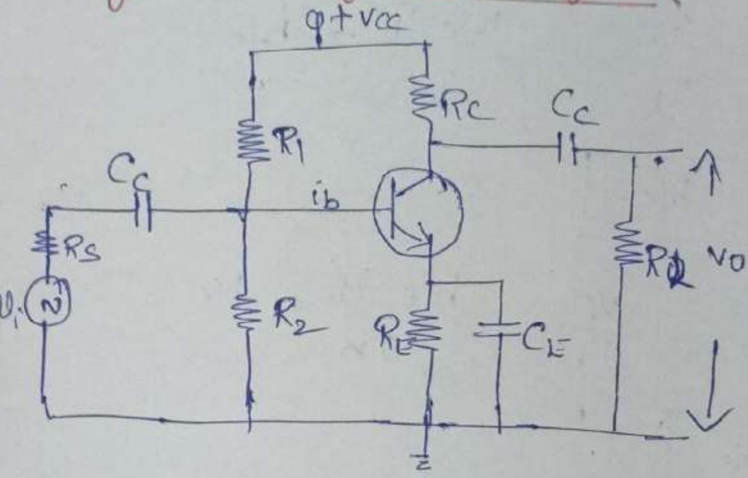
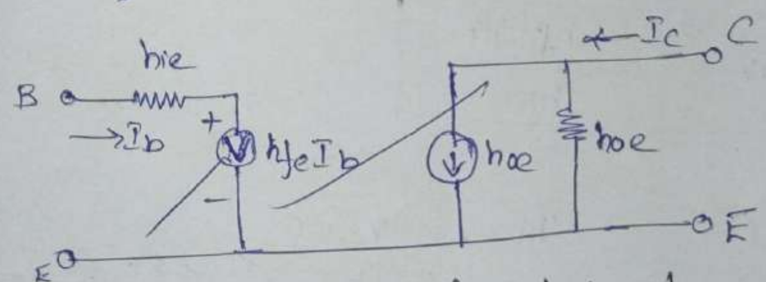


fig:- Hybrid model for a two port Network

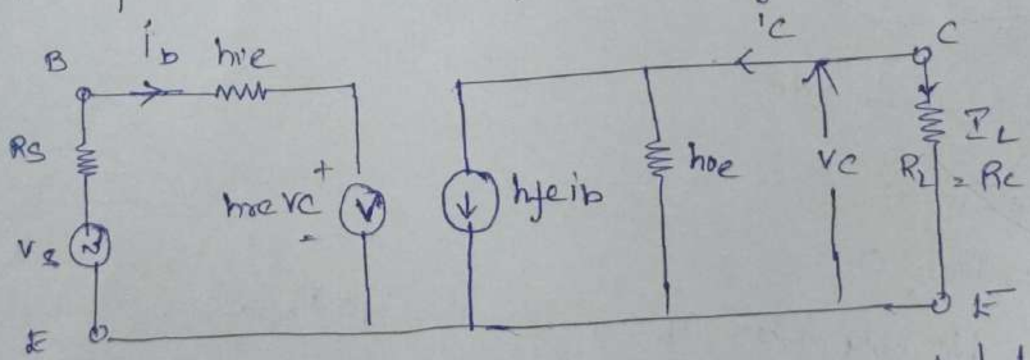
Single stage Amplifier.



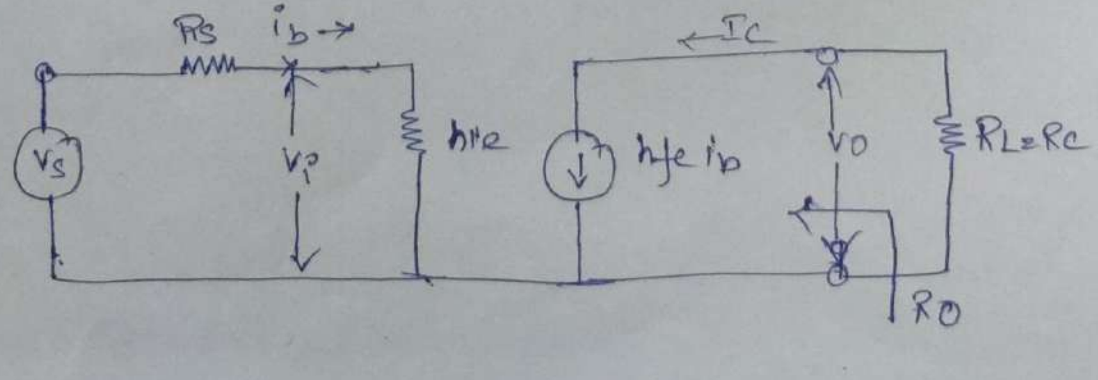
[Simplified/ Common Emitter (CE)/ hybrid model/ of a two port network.



Approximate hybrid model / Simplified/ Common Emitter (CE) hybrid model of a Transistor two port network amplifier.



hre & hoe is very small they can be neglected.



Parameters of Single stage amplifier

1) Current gain $(A_I) = \frac{I_c}{I_b} = \frac{-h_{fe} \cancel{I_b}}{\cancel{I_b}} = -h_{fe}$

$$\boxed{A_I = -h_{fe}}$$

(2) Input impedance $(R_i) = \frac{V_i}{I_b} = \frac{h_{ie} \cancel{I_b}}{\cancel{I_b}} = h_{ie}$

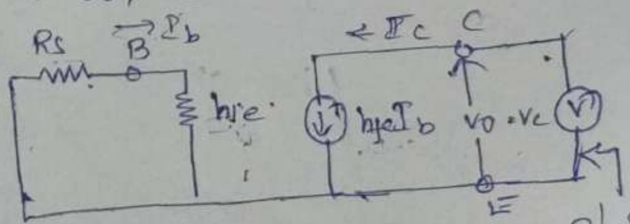
$$\boxed{R_i = h_{ie}}$$

(3) Voltage gain $(A_V) = \frac{V_0}{V_i} = \frac{I_c R_L}{h_{ie} I_b} = \frac{A_I R_L}{h_{ie}}$ $\quad \frac{I_c}{I_b} = A_i$

$$\boxed{A_V = \frac{-h_{fe} R_L}{h_{ie}}}$$

(4) Output impedance (R_o)

$R_L = \infty, \ V_S = 0$



$R_o = \frac{V}{I}$

$R_o = \frac{V_c}{I_c}$

$R_o' = R_o \| R_L$
$= R_L = R_C$

Apply KVL @ input side

$R_S I_b + h_{ie} I_b = 0$

$= I_b(R_S + h_{ie}) = 0$

It is possible only when $I_b = 0 \Rightarrow I_c = h_{fe} I_b = 0$

$R_o = \infty$

Frequency response of Single stage Amplifier

At lower frequencies

$$A_V = A_l = \frac{A_m}{1 - j(f_1/f)}$$

where f_1 is lower cutoff (or) lower 3 dB frequency.

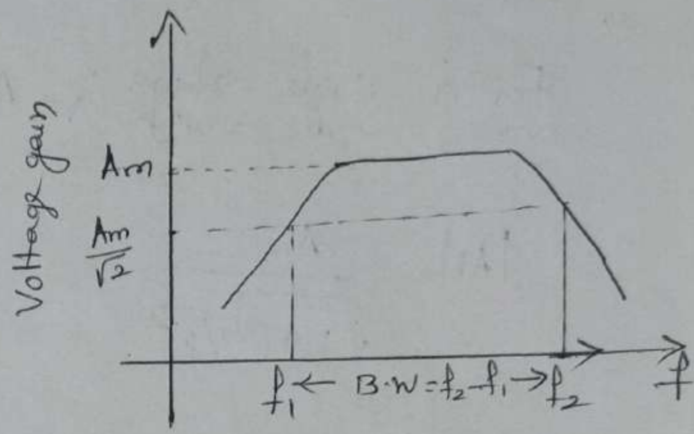


fig:- Frequency response of Single stage Amplifier

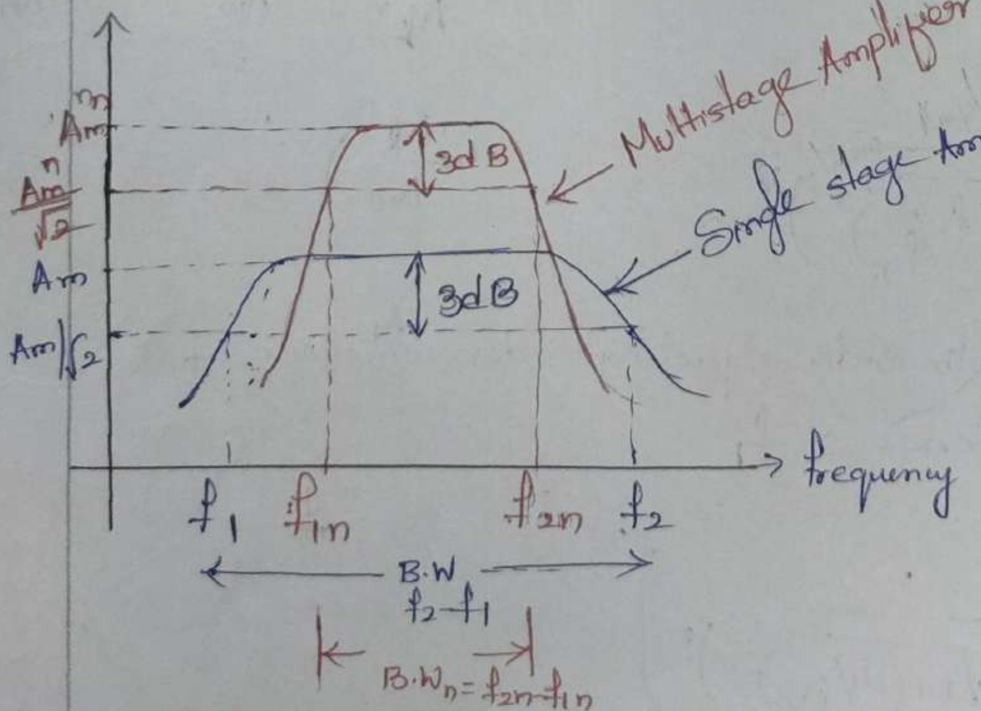
At Mid frequencies

$A_V = A_m$

At Higher Frequencies

$$A_V = A_h = \frac{A_m}{1 + j(f/f_2)}$$

where f_2 is upper cutoff (or) upper 3dB frequency.



frequency response of Single stage & Multistage amplifiers

$B.W = f_2 - f_1$;

Overall lower cutoff frequency (f_{1n})

For a single stage: $A = A_l = \dfrac{A_m}{1 - j(f_1/f)}$

$$|A_l| = \frac{A_m}{\sqrt{1 + (f_1/f)^2}}$$

$\therefore f_{1n} > f_1$

For 'n' stages

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

$$\left|\frac{A_l}{A_m}\right|^n = \frac{1}{\left(\sqrt{1 + (f_1/f)^2}\right)^n}$$

For single stage Amplifier
At $f = f_{1n}$ or $f = f_{2n}$
$A_l = \dfrac{A_m}{\sqrt{2}}$ (or) $A_h = \dfrac{A_m}{\sqrt{2}}$

For multistage,
$A_l^n = \dfrac{A_m^n}{\sqrt{2}}$ (or) $A_H^n = \dfrac{A_m^n}{\sqrt{2}}$

$$\frac{A_l^n}{A_m^n} = \frac{1}{\sqrt{2}}$$

At $f = f_{1n}$

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

squaring on both sides equaling denominators for solving f_{1n} interms of f_1

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

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

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

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

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

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

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

for $n \geq 2, 3, 4, \ldots \; f_{1n} > f_1$

i.e overall lower cutoff frequency increases.

<u>Overall upper cutoff frequency (f_{2n})</u>

For a single stage :. $A_h = \dfrac{A_m}{1 + j(f/f_2)}$

$$|A_h| = \frac{A_m}{\sqrt{1 + (f/f_2)^2}}$$

For 'n' stages

$$|A_h|^n = \left[\frac{A_m}{\sqrt{1 + (f/f_2)^2}}\right]^n$$

$$\left|\frac{A_h}{A_m}\right|^n = \frac{1}{\left[\sqrt{1 + (f/f_2)^2}\right]^n} \rightarrow (a)$$

At $f = f_{2n}$

at $f_2 = f_{2n}$

$A_H^n = \frac{A_m^n}{\sqrt{2}}$

Sub A_h^n value in a

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

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

Squaring on both sides equating denominators for solving f_{2n} interms of f_2

$$2 = \left[1+\left(\frac{f_{2n}}{f_2}\right)^2\right]^n$$

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

$$2^{1/n} - 1 = \left(\frac{f_{2n}}{f_2}\right)^2$$

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

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

for $n \geq 2, 3, 4 \ldots$ $f_{2n} < f_2$

i.e over all upper cutoff frequency decreases.

Over all Bandwidth ($B \cdot W_n$)

By definition of Bandwidth

For a single stage amplifier $B \cdot W = f_2 - f_1$

for n stage amplifier $B \cdot W_n = f_{2n} - f_{1n}$

As $f_{2n} < f_2$ & $f_{1n} > f_1$ $\boxed{B \cdot W_n < B \cdot W}$

i.e overall Bandwidth decreases.

Conclusions: When compared to single stage amplifier for multistage amplifier

* Overall voltage gain increases
* Overall lower cutoff frequency increases
* Overall upper cutoff frequency decreases.
* Overall Bandwidth decreases.

Problems

(1) For a given single stage amplifier f_1 is 100 Hz and f_2 is 100 kHz. If eight such stages are cascaded then determine overall lower and overall upper cutoff frequencies

Given

Lower cutoff frequency of a single stage amplifier is $f_1 = 100 Hz$

Upper cutoff frequency of a single stage amplifier is $f_2 = 100 kHz$

No. of stages $n = 8$.

find f_{1n} & f_{2n}

⇒ The overall lower cutoff frequency f_{1n}

$$f_{1n} = \frac{f_1}{\sqrt{2^{1/n}-1}} = f_{1n} = \frac{100}{\sqrt{2^{1/8}-1}} \rightarrow f_{1n} = 332.39 Hz$$

$f_{2n} =$

→ The overall upper cutoff frequency f_{2n}

$$f_{2n} = f_2\left(\sqrt{2^{1/n}-1}\right) \Rightarrow f_{2n} = 100\times10^3\left(\sqrt{2^{1/8}-1}\right) = 30.084 kHz$$

(2) Three identical stages are cascaded and have an overall upper 3dB frequency of 20kHz and overall lower 3dB frequency of 20Hz what are f_l & f_h

Given

Overall upper 3dB frequency $f_{2n} = 20\,kHz$
overall lower 3dB frequency $f_{1n} = 20\,Hz$

No. of stages $n=3$

To find

$f_l = ?$ & $f_h = ?$

$f_l = f_1 \quad f_h = f_2$

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

$$20\,Hz = \frac{f_l}{\sqrt{2^{1/3}-1}}$$

$$f_l = 20 \times \sqrt{2^{1/3}-1}$$

$$\boxed{f_l = 10.19\,Hz}$$

$$\rightarrow f_{2n} = f_2\sqrt{2^{1/n}-1} \Rightarrow 20\times10^3 = f_2\sqrt{2^{1/3}-1} \Rightarrow f_2 = \frac{20\times10^3}{\sqrt{2^{1/3}-1}}$$

$$f_2 = 39229\,Hz$$

$$\boxed{f_2 = 39.229\,kHz}$$

(3) If $f_2 = 40KHz$, $f_{2n} = 20KHz$ then find 'n' i.e (10)
no of stages

$$f_{2n} = f_2\sqrt{2^{1/n}-1} \Rightarrow 20\times10^3 = 40\times10^3\sqrt{2^{1/n}-1}$$

$$\frac{20}{40} = \sqrt{2^{1/n}-1}$$

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

squaring on both sides

$$\frac{1}{4} = 2^{1/n}-1$$

$$\frac{1}{4}+1 = 2^{1/n}$$

$$\frac{5}{4} = 2^{1/n}$$

$$1.25 = 2^{1/3}$$

$$1.25 = 1.25$$

No of stages $n=3$

Purpose of Coupling device:

- → To transfer the AC from the output of one stage to the input of next stage.
- → To block the DC to pass from the output of one stage to the input of next stage, which means to isolate the DC condition

Types of Coupling :-

Joining one amplifier stage with the other in cascade, using coupling devices from a Multi stage amplifier circuit. There are 4 basic methods of coupling, using these coupling devices such as resistors, capacitors, transformers etc. Let us have an idea about them.

Resistance - Capacitance Coupling :-

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

The Coupling Capacitor passes the AC from the output of one stage to the input of its next stage. While blocking the DC components from DC bias voltages to effect the next stage. Let us get into the details of this method of coupling in the coming chapters.

Impedance Coupling :-

The Coupling network that uses inductance and Capacitance as coupling elements can be called as impedance Coupling network.

In this impedance Coupling method, the impedance of Coupling coil depends on its inductance and signal frequency which is jwL.

Transformer Coupling :-

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

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

Direct Coupling :-

If the previous amplifier stage is connected to the next amplifier stage directly, it is called as direct coupling. The individual amplifier stage bias conditions are so designed that the stages can be directly connected without DC isolation.

The direct coupling method is mostly used when the load is connected in series, with the output terminal of the active circuit element. For example, head phones, loud speakers etc.

Role of Capacitors in Amplifiers :-

Other than the coupling purpose, there are other purposes for which few capacitors are especially employed in amplifiers. To understand this, let us know about the role of capacitors in Amplifiers.

1. The input Capacitor C_{in} :-

The input capacitor C_{in} present at the initial stage of the amplifier, couples AC signal to the base of the transistor. This capacitor C_{in} if not present, the signal source will be in parallel to resistor R_2 and the bias voltage of the transistor base will be changed.

Hence C_{in} allows the AC signal from source to flow into input circuit, without affecting the bias conditions.

2. The Emitter By-Pass Capacitor C_e :-

The emitter by-pass capacitor C_e is connected in parallel to the emitter resistor. It offers a low reactance path to the amplified AC signal.

In the absence of the capacitor, the voltage developed across R_E will feedback to the i/p side thereby reducing the output voltage. Thus in the presence of C_e the amplified AC will pass through this

Coupling Capacitor C_C:-

The Capacitor C_C is the Coupling Capacitor that connects two stages and prevents DC interference between the stages and controls the operating point from shifting. This is also called as blocking Capacitor because it does not allow the DC voltage to pass through it

In the absence of this Capacitor, R_C will come in parallel with the resistance R_1 of the biasing network of the next stage and thereby changing the biasing conditions of the next stage.

Amplifier Consideration:-

For an amplifier circuit, the overall gain of the amplifier is an important consideration. To achieve maximum voltage gain, let us find the most suitable transistor configuration for Cascading

1. CC amplifier:-
   1. Its voltage gain is less than unity
   2. It is not suitable for intermediate stages.

2. CB Amplifier:-
   1. Its voltage gain is less than unity
   2. it is not suitable for intermediate stages cascading

3. CE Amplifier:-
   1. Its voltage gain is greater than unity.
   2. Voltage gain is further increased by cascading

Coupling Networks

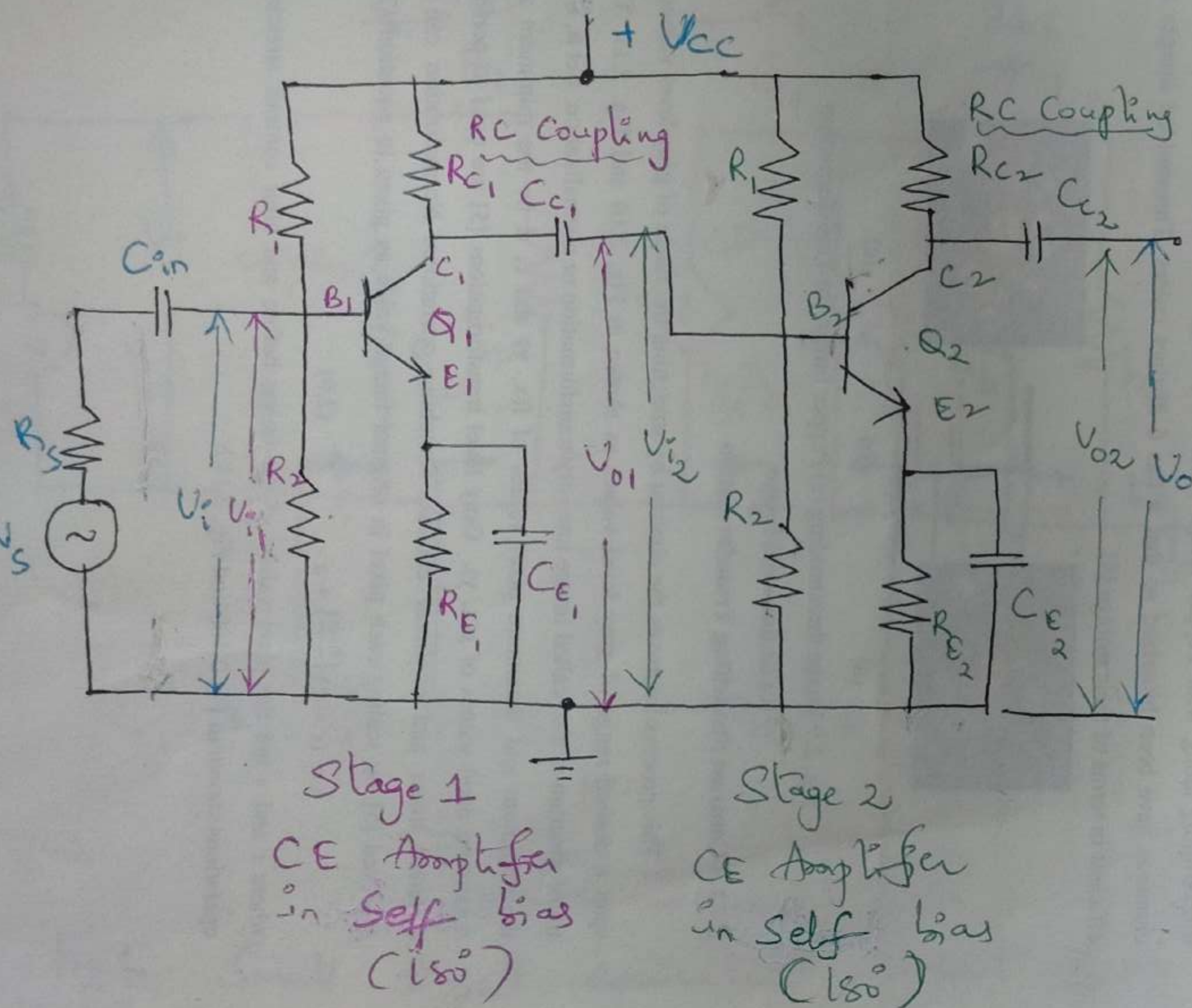
Multistage amplifier is formed by joining output of one stage to input of next stage i.e cascade connection by using coupling network.

→ Coupling network is used to couple AC from output of one stage to input of next stage. & it also blocks DC.

→ It isolates DC conditions of one stage to next stage.

→ It is necessary to prevent the shifting of Q-point

* Two stage RC Coupled Amplifier Analysis



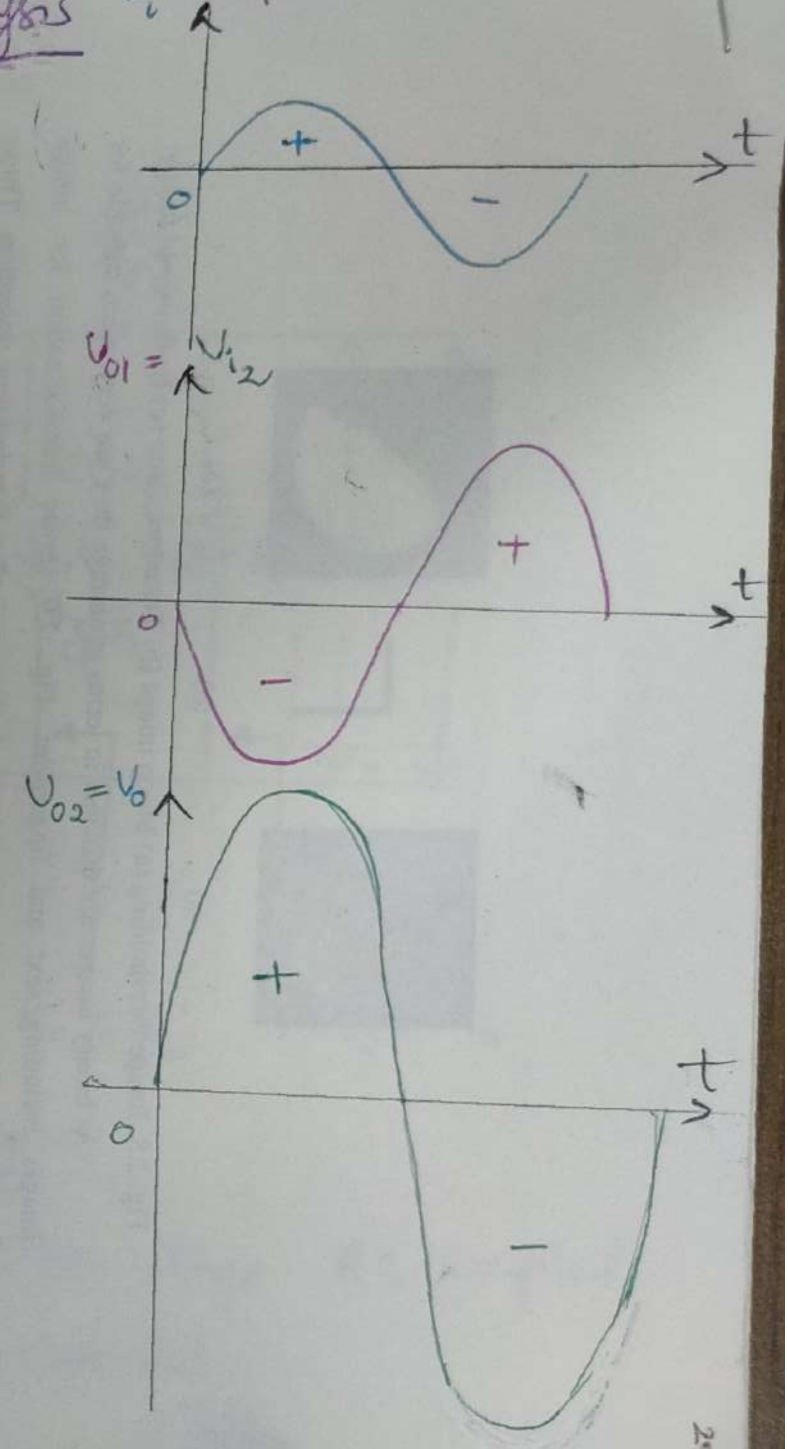


Fig a: Two stage RC Coupled Amplifier

Analysis

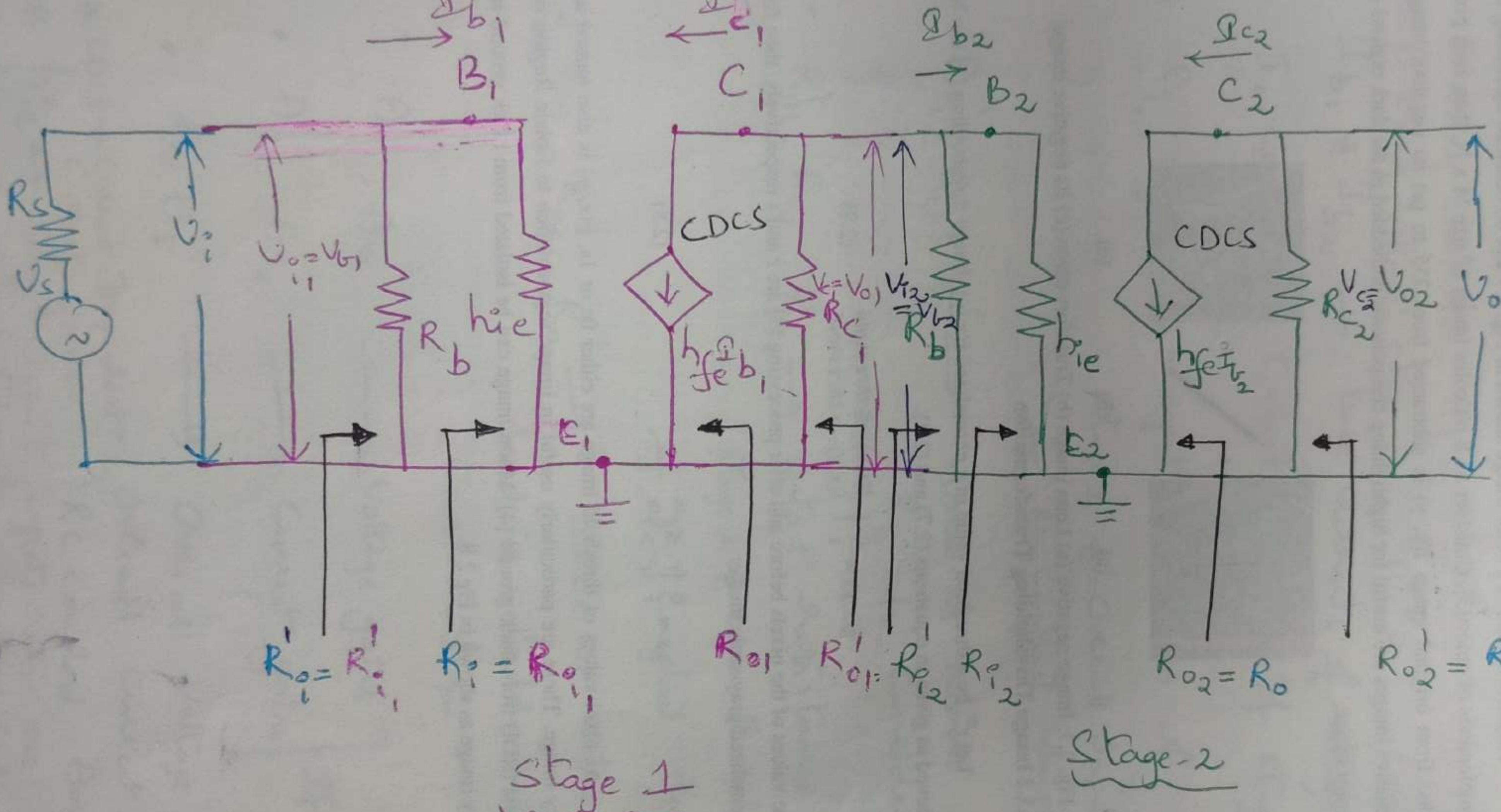


Fig b: Small signal equivalent circuit of a two-stage RC coupled amplified by using Approximate hybrid model.

Different representations used in fig b are as follows: Assume Q_1 & Q_2 are identical

- Biasing resistor $R_b = \frac{R_1 \cdot R_2}{R_1 + R_2}$ ∵ R_1 & R_2 are in parallel
- dc Biasing Voltage: $+V_{cc}$
- I_{b_1} & I_{b_2} → base currents of stage 1 & stage 2
- I_{c_1} & I_{c_2} → Collector currents of stage 1 & stage 2
- R_{i_1}, R_{i_2}, R_i → input resistances of stage 1, stage 2 & overall input resistance of RC coupled Amplifiers.
- R'_{i_1}, R'_{i_2}, R'_i → Same as above including R_b into consideration.
- R_{o1}, R_{o2}, R_o → output resistances of stage 1, stage 2 & overall output resistance of RC coupled Amplifier
- R'_{o1}, R'_{o2}, R'_o → Same as above including load ie collector resistor into consideration.
- A_{V_1}, A_{V_2} → Voltage gains } of stage 1 & stage 2
- A_{I_1}, A_{I_2} → Current gains } of stage 1 & stage 2
- A_V, A_I → Overall Voltage gain & Overall current gain of RC Coupled Amplifier

* CDCS → Current Dependent Current Source

* h_{ie} → Short Circuit input resistance (Ω)

* h_{fe} → Short Circuit forward current gain

* The Parameters of Stage-2 CE Amplifier are evaluated and then parameters of stage 1 CE Amplifier are evaluated to find Overall parameters of Two stage RC Coupled Amplifier.
* The order of the four parameters of amplifier to be evaluated is as follows
  * Current gain (output current / input current)
  * input resistance (input voltage / input current) Ω
  * Voltage gain (output voltage / input voltage)
  * output resistance (output voltage / output current keeping Source $\to 0$ & Load $\to \infty$) Ω
* By definitions & from fig b

For Stage 2 CE Amplifier

$$A_{I_2} = \frac{-I_{c_2}}{I_{b_2}} = \frac{-h_{fe} \cancel{I_{b_2}}}{\cancel{I_{b_2}}} = -h_{fe}$$

$$R_{i_2} = \frac{V_{b_2}}{I_{b_2}} = \frac{h_{ie} \cancel{I_{b_2}}}{\cancel{I_{b_2}}} = h_{ie} (\Omega)$$

$$R_{i_2}' = R_{i_2} \| R_b = \frac{R_{i_2} \cdot R_b}{R_{i_2} + R_b} \; (\Omega)$$

$$A_{V_2} = \frac{V_{o2}}{V_{i_2}} = \frac{-I_{c_2} \cdot R_{C_2}}{I_{b_2} \cdot R_{i_2}} = A_{I_2} \frac{R_{C_2}}{h_{ie}}$$

$$Y_{o2} = 0 \mho \Rightarrow R_{o2} = \infty \, \Omega, \quad R_{o2}' = R_{o2} \| R_{L_2} = R_{C_2} \, \Omega$$

For stage 1 : $A_{I_1} = \frac{-I_{c_1}}{I_{b_1}} = \frac{-h_{fe}\cancel{I_{b_1}}}{\cancel{I_{b_1}}} = -h_{fe}$

CE Amplifier

$$R_{i_1} = \frac{V_{b_1}}{I_{b_1}} = \frac{h_{ie}\cancel{I_{b_1}}}{\cancel{I_{b_1}}} = h_{ie}\,(\Omega)$$

$$R_{i_1}' = R_{i_1} \| R_b = \frac{R_{i_1} \cdot R_b}{R_{i_1} + R_b}\ (\Omega)$$

$$A_{V_1} = \frac{V_{o_1}}{V_{i_1}} = \frac{-I_{c_1} \cdot R_{L_1}}{I_{b_1} \cdot R_{i_1}} = \frac{A_{I_1}\left(R_{c_1} \| R_{i_2}'\right)}{h_{ie}}$$

$$= \frac{A_{I_1}\left(\frac{R_{c_1} \cdot R_{i_2}'}{R_{c_1} + R_{i_2}'}\right)}{h_{ie}}$$

$$Y_{o_1} = 0\,\mho \Rightarrow R_{o_1} = \infty\,\Omega,\ R_{o_1}' = R_{o_1} \| R_{L_1}$$

$$= R_{L_1} = \frac{R_{c_1} \cdot R_{i_2}'}{R_{c_1} + R_{i_2}'}\ (\Omega)$$

For Two stage RC Coupled Amplifier

$$A_I = \frac{-I_{c_2}}{I_{b_1}} = \frac{-I_{c_1}}{I_{b_1}}\left(\frac{-I_{b_2}}{I_{c_1}}\right)\frac{-I_{c_2}}{I_{b_2}}$$

$$= A_{I_1}\left(\frac{R_{c_1}}{R_{c_1} + R_{i_2}'}\right) A_{I_2}$$

$$R_i = R_{i_1} = h_{ie}\,(\Omega),\ R_i' = R_{i_1}' = h_{ie} \| R_b$$

$$= \frac{h_{ie} R_b}{h_{ie} + R_b}\ (\Omega)$$

$$A_V = \frac{V_{o2}}{V_{i_1}} = \frac{V_{o1}}{V_{i_1}} \times \frac{V_{o2}}{V_{i2}} = A_{V_1} \cdot A_{V_2}$$

$$R_o = R_{o2} = \infty \, \Omega, \quad R_o' = R_{o2}' = R_{L_2} = R_{C_2} \, (\Omega)$$

Conclusions:

* Overall current gain is equal to the product of individual current gains and a factor $\left(\frac{R_{C_1}}{R_{C_1} + R_{i_2}'}\right)$ is also multiplied.

* Overall input resistance is input resistance of stage 1

* Overall output resistance is output resistance of stage 2

* Overall voltage gain is product of individual voltage gains.

P. Find overall parameters of a two stage RC Coupled amplifier if $h_{ie} = 1k\Omega$, $h_{fe} = 50$, $R_{C_1} = R_{C_2} = 2k\Omega$, $R_1 = 5.2k\Omega$ and $R_2 = 1.24k\Omega$ by using approximate hybrid analysis.

frequency response:

The frequency response of the amplifier is a plot of gain vs frequency. The frequency response curve is shown in figure. The frequency response curve is shown in figure. The behaviour is briefly explained below.

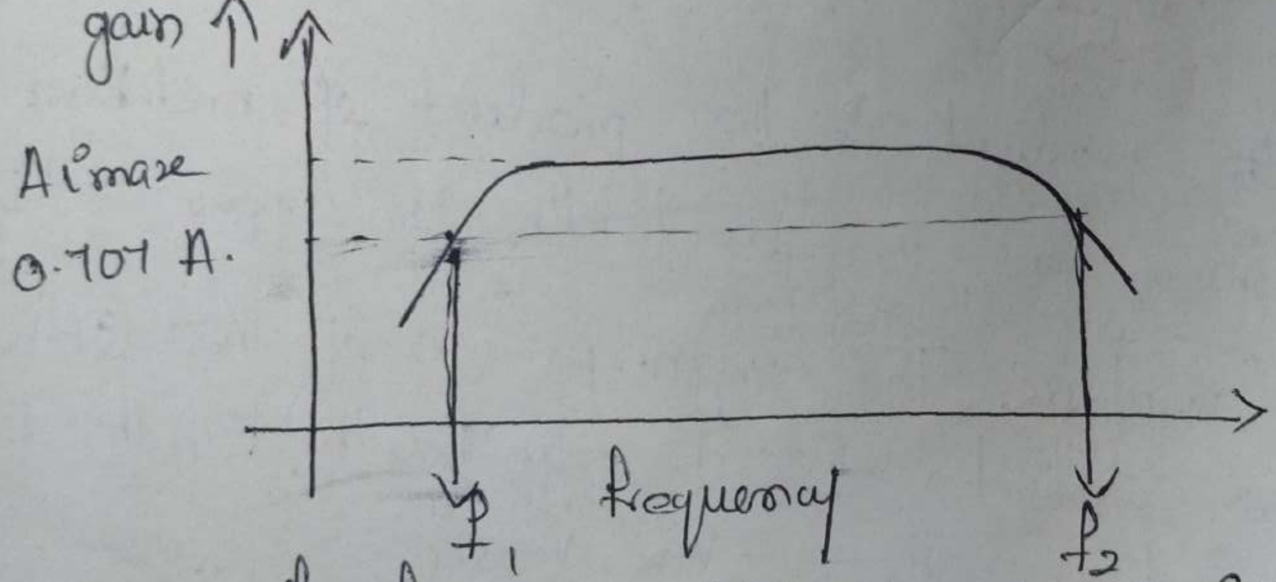


fig frequency response of RC coupled amplifier

(i) At low frequencies :-

At low frequencies the reactance of Coupling Capacitor C_c is quite high and a very small part of the signal passes from one stage to next stage. This can be compensated by taking capacitor of high value. Secondly, at low frequencies, the reactance is comparable to Re and hence ac signal flows through the emitter Re. This in return decreases output voltage. This effect of C_c reduces the gain at low frequencies.

(ii) At high frequencies :-

The reactance of C_o at high frequencies is small such that it is a short circuit. This increases the loading effect on the next stage and reduces voltage gain.

(iii) At mid frequencies :-

The voltage gain of the amplifier is constant in this frequency range. when frequency increases in this range reactance of C_c decreases which tends to increase the gain. At the same time, lower the reactance means higher loading of the first stage over the second stage and hence lower gain. These two factors almost cancel each other resulting in a uniform gain at mid frequencies.

Advantages:-

1. It is small, light and inexpensive because it requires no expensive or bulky components
2. it has excellent frequency response and the gain is constant over audio frequency range.
3. it has minimum possible nonlinear distortion because it does not use any coils or transformers which might pick up undesirable signals.

Disadvantages

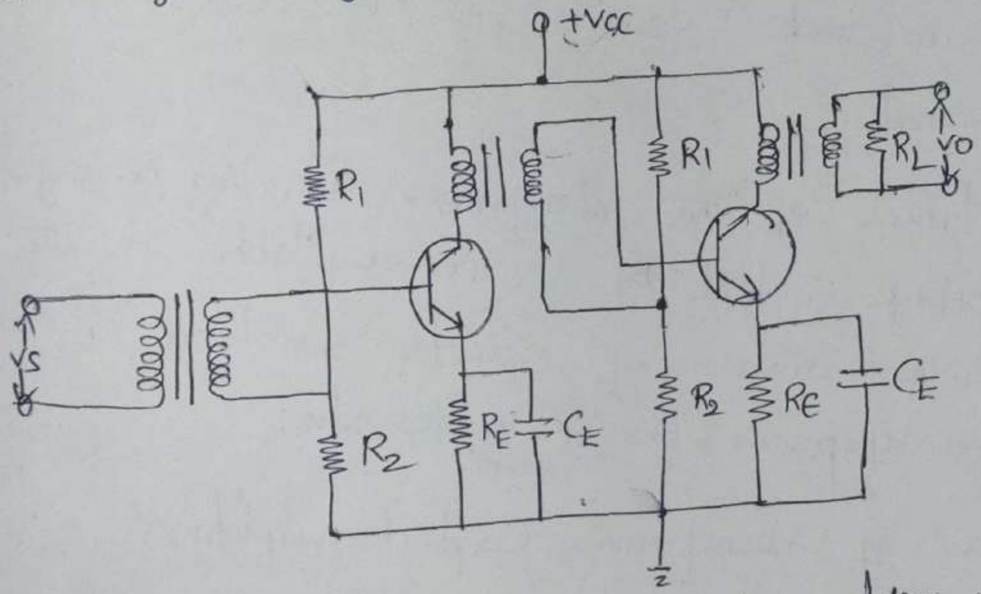
1. The gain of the RC coupled amplifier is comparatively small because of the loading effect of next stage.
2. They have a tendency to become noisy in most climates
3. Impedance matching is poor as the output impedance is several hundred ohm while of a speaker is only few ohms.

Applications

1. public address system.

Transformer Coupled Amplifier:

Fig shows transformer coupled amplifier using transistors. The output signal of first stage is coupled to the input of the next stage through an impedance matching transformer.



Two stage transformer coupled amplifier using transistors.

This type of coupling is used to match the impedance between output and input cascaded stage. Usually, it is used to match the larger output resistance of AF power amplifier to a low impedance load like loud speaker. As we know, transformer blocks dc, providing dc isolation between the two stages. Therefore, transformer coupling does not affect the quiescent point of the next stage.

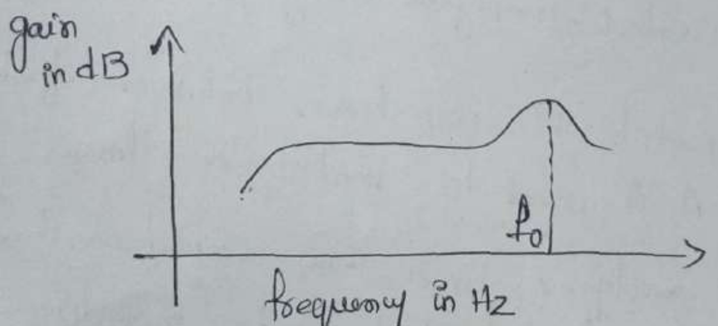
Frequency response of transformer coupled amplifier is poor in comparison with that of an RC coupled amplifier. Its leakage inductance and interwinding capacitance does not allow amplifier to amplify the signals of different frequencies equally well. interwinding capacitance of the transformer coupled may give rise resonance at certain frequency which makes amplifier to give very high gain at that frequency.

By putting shunting capacitors across each winding of the transformer, we can get resonance at any desired RF frequency. Such amplifiers are called "tuned voltage amplifiers"

There provide high gain at the desired of frequency, i.e they amplify selective frequencies. For this reason, the transformer coupled amplifiers are used in radio and TV receivers for amplifying RF signals.

As dc resistance of the transformer winding is very low almost all dc voltage applied by Vcc is available at the collector. Due to the absence of collector resistance it also eliminates unnecessary power loss in the resistor.

Frequency response of transformer Coupled amplifier:-



Advantages:-

1. There is no loss of signal power in collector or base resistors.
2. it provides a higher voltage gain
3. it provides an excellent impedance matching between input & output

Disadvantages:-

1. it is costly and bulky particularly at audio frequencies because of its heavy iron core
2. it has poor frequency response
3. This introduces hum in the output
4. At radio frequencies, the inductance and winding Capacitance produce a lot of problems
5. Frequency distortion is higher ie low frequency signals are less amplified as compared to the high frequency signals.

(i) Direct Coupled Amplifier:-

Stray Capacitance → considerable at high frequen
negligible at low frequen

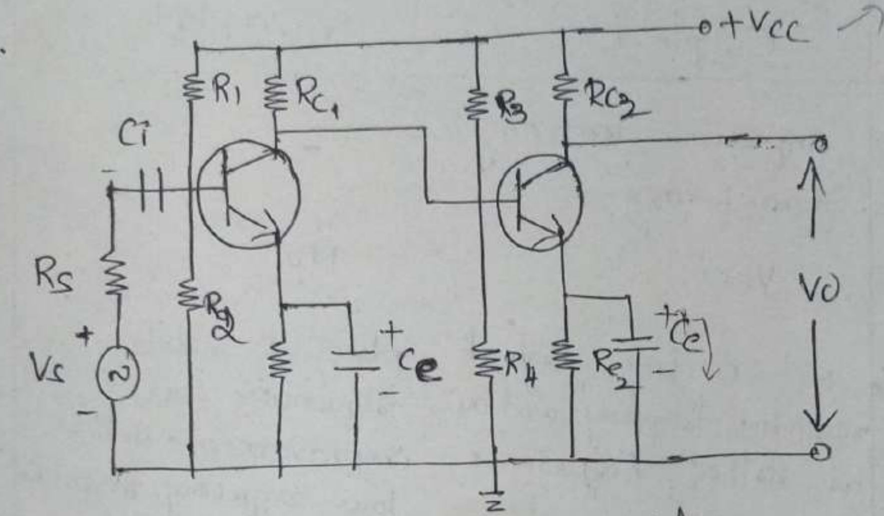


fig (a) Direct Coupled Amplifier

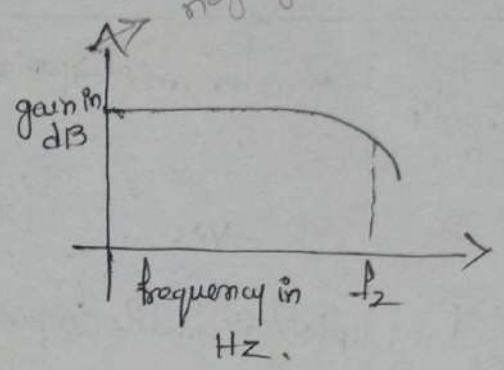


fig (b):- Frequency response of Direct Coupled Amplifier.

$X_C = 1/2\pi fC$

Fig shows direct coupled amplifier using transistors. The output signal of first stage is directly connected to the input of the next stage. This direct coupling allows the quiescent dc collector current of first stage to pass through base of the next stage, affecting its biasing conditions.

Due to absence of ~~Cc~~ Cc its low frequency response is good but at higher frequencies shunting capacitors such as stray capacitances reduces the gain of the amplifier. → present b/w windings in a coil or b/w conductors

The transistor parameters such as V_{BE} and β change with temperature causing the collector current and voltage to change. Because of direct coupling these changes appears at the base of the next stage, and hence in the output. Such an unwanted change in the output is called drift and it is serious problem in the direct coupled amplifiers.

V_{BE} → $I_C\uparrow$ $I_C = \beta I_B$

tmp β

Comparision between Various Cascading Methods

Parameter	RC Coupled	Transformer Coupled	Direct Coupled
Coupling components	Resistor and Capacitor	impedance matching transformer	-
Block DC	Yes	Yes	NO
frequency response	flat at middle frequencies	Not uniform, high at resonant frequency and low at other frequencies	flat at middle frequencies and improvements in the low frequency response
impedance matching	Not achieved	Achieved	Not achieved
DC amplification	NO	NO	Yes
weight	Light	Bulky & heavy	.
Drift	Not present	Not present	Present

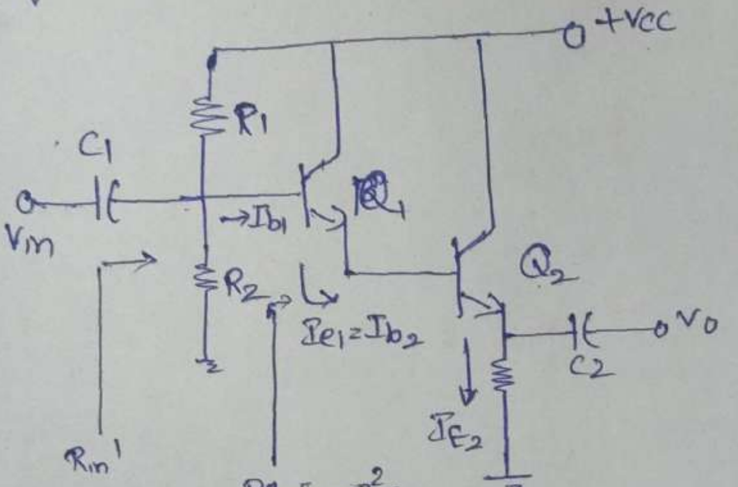
High Input Resistance Transistor Amplifier

1. Darlington Amplifier
2. Boot strapped emitter follower
3. Cascode Amplifier
4. Differential Amplifier

Darlington Amplifier!.

It is a combination of two cascaded emitter followers.

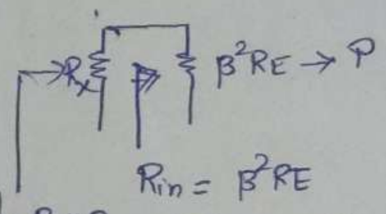
- The darlington amplifier has a high input resistance, low output resistance and high current gain. These characteristics make it useful as a current amplifier. The voltage gain of a Darlington amplifier is less than unity



$R_{in} = [1+\beta]^2 R_E$

$\boxed{R_{in} = \beta^2 R_E}$ ***

[for ac analysis $V_{cc} = 0$ so
$R_1 = 0$, $R_2 = 0$ R_1 & R_2 comes in parallel
$R_1 \| R_2 = R_x$.



$R_{in} = \beta^2 R_E$

$R_{in}' = R_x \| R_{in} \rightarrow$
where $R_x = R_1 \| R_2$

$R_x \ll R_{in}$ so R_{in}' is small bcz of $R_x \| R_{in}$

draw back → overall resistance ↓ses

Here current gain at stage 1 β_1 is for Q1

$$I_{e1} = \beta_1 \cdot I_{b1}$$

Illy Q_2 is $I_{E2} = \beta_2 \cdot I_{b2} = \beta_2 \cdot I_{E1}$

$$I_{E2} = \beta_1 \cdot \beta_2 \cdot I_{b1}$$

$\therefore$ overall current gain $\dfrac{I_{E2}}{I_{b1}} = A_I = \beta_1 \cdot \beta_2$

For identical $A_I = \beta^2$

Typically the i/p impedance range is 200kΩ to 300kΩ.

→ However the impedance of circuit can be improved by direct coupling of two stages of emitter follower amplifier.

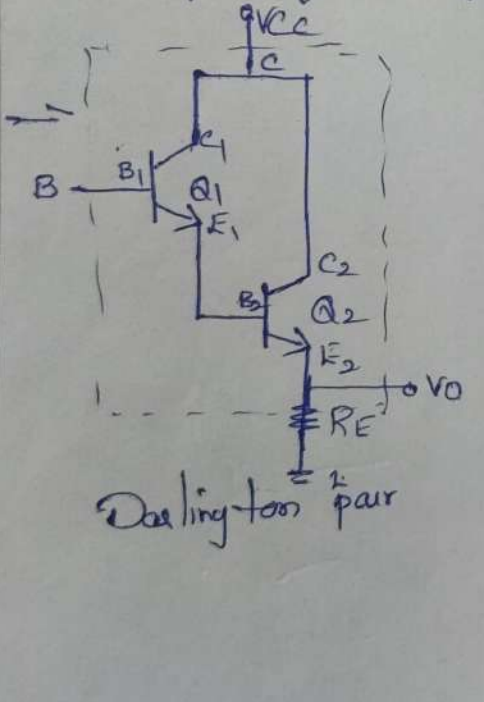
→ The input impedance can be increased by 2 techniques.

1) Direct Coupling [Darlington connection]
2) Using boot strap technique.

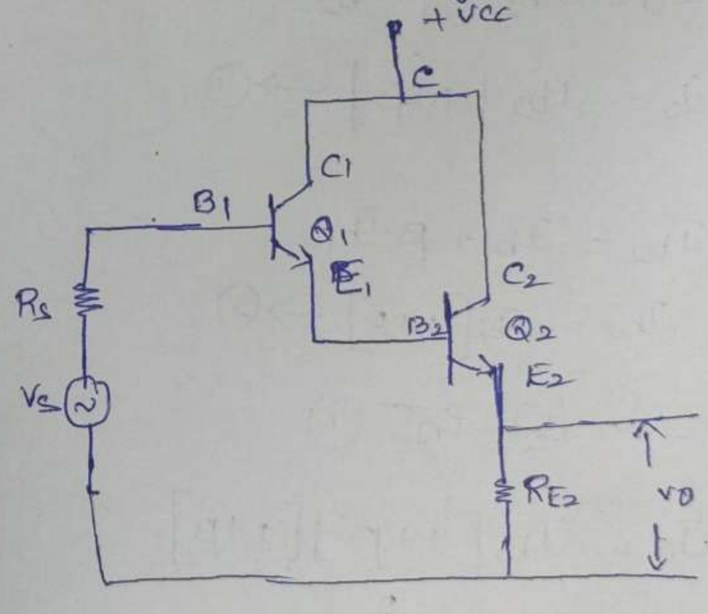
① Darlington Emitter follower (or) Darlington Amplifier (or) Super Beta(β) Amplifier (or) Two stage CC-CC amplifier

↳ in some application, high input resistance is need for an amplifier. For input resistance about less than 500kΩ emitter follower (common collector amplifier) is satisfactory to achieve high input resistance Darlington connection is used.

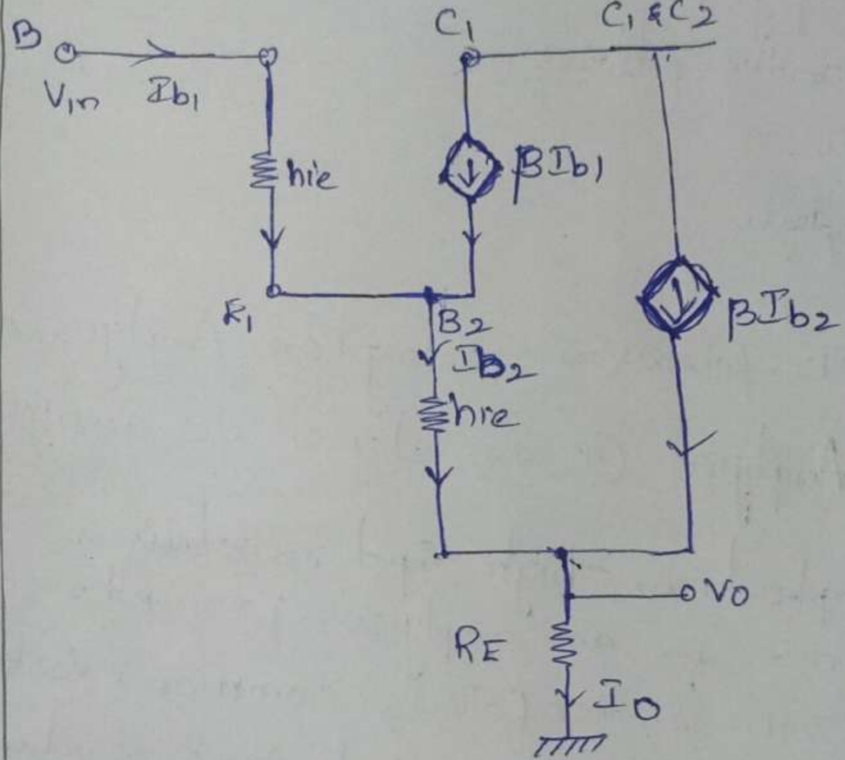
→ If two common collector configurations are connected in cascade (CC-CC) i.e output of stage 1 is connected to the input of stage-2 is called as Darlington pair



Darlington pair



Analysis of Darlington Amplifier:-



1. Current gain (AI)

$$A_I = \frac{o/p \text{ current}}{i/p \text{ current}} = \frac{I_o}{I_{in}} = \frac{I_o}{I_{b1}}$$

$$I_o = I_{b2} + \beta I_{b2}$$

$$I_o = I_{b2}[1+\beta] \rightarrow (1)$$

$$I_{b2} = I_{b1} + \beta I_{b1}$$

$$I_{b2} = I_{b1}[1+\beta] \rightarrow (2)$$

Sub (2) in (1)

$$I_o = I_{b1}[1+\beta][1+\beta]$$

$$\frac{I_o}{I_{b1}} = [1+\beta]^2$$

$1 << \beta$, so

$$\boxed{I_o/I_{b1} = A_I = \beta^2} \Rightarrow \quad 50\times50 =$$

(2) Input Resistance (R_{in}) (or) Input impedance (Z_{in})

$$R_{in} \text{ (or) } Z_{in} = \frac{V_{in}}{I_{in}}$$

Applying KVL @ i/p side

$V_{in} = I_{b1} h_{ie} + h_{ie} I_{b2} + I_0 R_E$

Sub I_0 & I_{b2} values in the above equation

$V_{in} = I_{b1} h_{ie} + \cancel{h_{ie}} (1+\beta) I_{b1}$

$V_{in} = I_{b1} h_{ie} + I_{b1}[1+\beta] h_{ie} + I_{b1}[1+\beta]^2 R_E$

$V_{in} = I_{b1}\left[h_{ie} + (1+\beta)h_{ie} + (1+\beta)^2 R_E\right]$

$\frac{V_{in}}{I_{b1}} = \left[h_{ie} + (1+\beta)h_{ie} + (1+\beta)^2 R_E\right]$

h_{ie} is small so neglect first two terms

$$\boxed{\frac{V_{in}}{I_{b1}} = [1+\beta]^2 R_E}$$

$1 << \beta$

$R_{in} = \beta^2 R_E$

$R_{in} = A_I R_E$

(or)

$$\boxed{\begin{aligned} R_{in} &= \beta^2 R_E \\ R_{in} &= A_I R_E \end{aligned}}$$

$\therefore \beta^2 = A_I$

(3) Voltage gain (A_V)

$$A_V = \frac{\text{out voltage}}{\text{input voltage}} = \frac{V_o}{V_{in}}$$

$V_o = I_0 R_E$

$I_0 = I_{b1}[1+\beta]^2$

Sub I_0 value in V_o

$$V_0 = I_{b_1}[1+\beta]^2 R_E$$

$$V_{in} = I_{b_1} h_{ie} + I_{b_1}[1+\beta] h_{ie} + I_{b_1}[1+\beta]^2 R_E$$

$$A_V = \frac{V_0}{V_{in}} = \frac{\cancel{I}_{b_1}[1+\beta]^2 R_E}{\cancel{I}_{b_1}[h_{ie} + (1+\beta)h_{ie} + \underbrace{[1+\beta]^2 R_E}_{\text{large value compared with first two terms}}}$$

$$A_V = \frac{[1+\beta]^2 R_E}{[1+\beta]^2 R_E}$$

$$\boxed{A_V = 1}$$

output impedance (R_o)

$$R_o = \frac{V_o}{I_o}\Bigg|_{V_s = 0} \rightarrow R_o = \infty$$

for $R_o' \approx R_E$

Conclusion:-

when compared to a single stage emitter follower, Darlington emitter follower has

→ higher current gain
→ higher input resistance
→ Voltage gain lessthan unity or ≈ unity
→ lower output resistance

Limitations:-

→ Assumed that Q_1 & Q_2 transistors are having identical h-parameters, but practically they depend on quiescent conditions

→ Leakage current of first transistor is amplified by second stage. hence Darlington connection of 3 or more stages is usually impractical.

Analysis of Darlington Amplifier

Stage 2 CC Amplifier

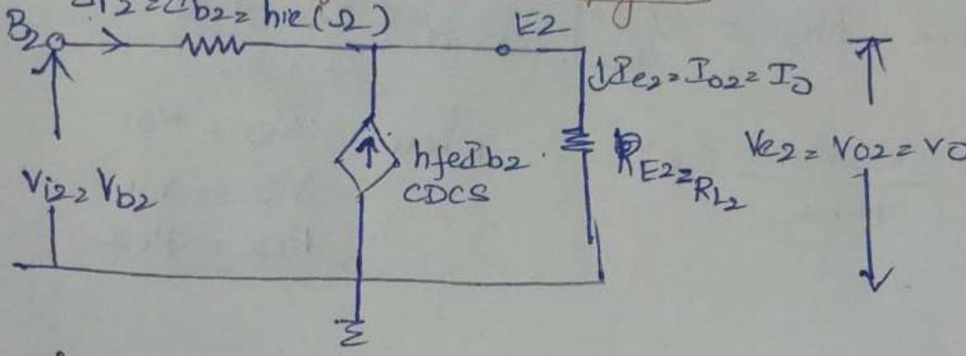


fig:- Small signal equivalent ckt of stage 2 CC amplifier using approximate hybrid model.

(1) Current gain $(A_{I_2}) = \frac{I_{o2}}{I_{i2}} = \frac{I_{e2}}{I_{b2}} = \frac{I_{b2} + h_{fe} I_{b2}}{I_{b2}} = \frac{\cancel{I_{b2}}(1+h_{fe})}{\cancel{I_{b2}}} = 1 + h_{fe}$

(2) Input impedance $(R_{I_2}) = \frac{V_{i2}}{I_{i2}} = \frac{V_{b2}}{I_{b2}} = \frac{h_{ie} I_{b2} + R_{E2}(I_{b2} + h_{fe} I_{b2})}{I_{b2}}$

$$= \frac{\cancel{I_{b2}}\left[h_{ie} + (1+h_{fe})R_{E2}\right]}{\cancel{I_{b2}}}$$

$$= h_{ie} + (1+h_{fe})R_{E2}$$

$$\simeq (1+h_{fe})R_{E2} \quad \therefore h_{ie} << (1+h_{fe})R_{E2}$$

(3) Voltage gain (A_{V_2}) :- $\frac{V_{o2}}{V_{i2}} = \frac{V_{e2}}{V_{b2}} = \frac{I_{e2} R_{E2}}{I_{b2} R_{i2}} = A_{i2} \cdot \frac{R_{E2}}{R_{i2}}$

But from basic definition of R_i

$$R_{i2} = h_{ic} + A_{I2} h_{rc} \cdot R_{L2}$$

$$\simeq h_{ie} + A_{I2}(1) R_{E2} \quad \because h_{ic} = h_{ie}$$

$h_{rc} = 1$

÷ with R_{i2} on both sides $\quad R_{L2} = R_{E2}$

$$\frac{R_{i2}}{R_{i2}} = \frac{h_{ie} + A_{i2} R_{E2}}{R_{i2}}$$

$$1 = \frac{h_{ie}}{R_{i2}} + A_{i2} \cdot \frac{R_{E2}}{R_{i2}}$$

$$1 - \frac{h_{ie}}{R_{i2}} = A_{i2} \cdot \frac{R_{E2}}{R_{i2}}$$

Hence $A_{V2} = 1 - \frac{h_{ie}}{R_{i2}}$

4 Output resistance $R_{o2} = \frac{V_{o2}}{I_{o2}}\Big|_{V_{S2}=0}$ & $R_{L2} \to \infty$

$$= \frac{R_{S2} + h_{ie}}{1 + h_{ie}} \ (\Omega)$$

where $R_{S2} = R_{o1}$

$V_{S2} = V_{o1}$

$R_{L2} = R_{E2}$

Stage 1 CC Amplifier :-

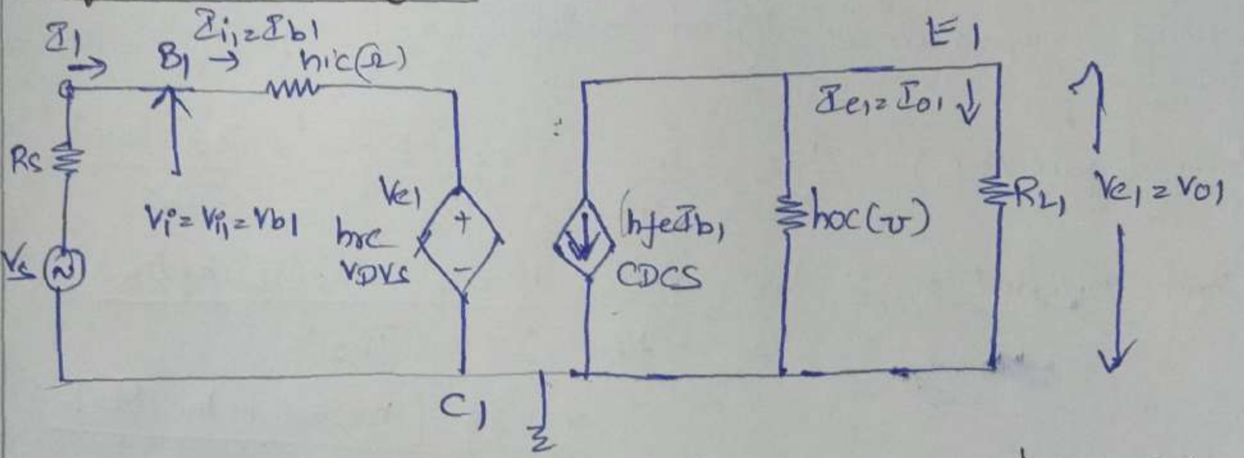


fig :- Small signal equivalent circuit of stage 1 CC Amplifier by using exact hybrid model.

Current gain $(A_{I1}) = \frac{I_{o1}}{I_{i1}} = \frac{I_{e1}}{I_{b1}} = \frac{-h_{fc}}{1 + h_{oc} R_{L1}} = \frac{1 + h_{fe}}{1 + h_{oe} R_{i2}}$

$$= \frac{-h_{fc}}{1 + h_{oc} R_{L1}} = \frac{1 + h_{fe}}{1 + h_{oe} R_{i2}}$$

$\therefore h_{fc} = 1 + h_{fe}$

$h_{oc} \simeq h_{oe}$

$R_{L1} = R_{i2}$

$$= \frac{1 + h_{fe}}{1 + h_{oe}(1 + h_{fe}) R_{E2}}$$

$$= \frac{1 + h_{fe}}{1 + h_{oe} h_{fe} R_{E2}}$$

$\therefore h_{oe} R_{E2} << h_{oe} h_{fe} R_{E2}$

input Resistance $(R_{i1}) = \frac{V_{i1}}{I_{i1}} = \frac{V_{b1}}{I_{b1}}$

$$= h_{ic} + A_{I1} h_{rc} R_{L1}$$

$$= h_{ic} + A_{I1} R_{i2}$$

$\therefore h_{rc} \simeq 1$

$R_{L1} = R_{i2}$

$= A_{I_1} R_{i2} \quad \because h_{ic} \ll A_{I_1} R_{i2}$

$$= \frac{1+h_{fe}}{1+h_{oe}h_{fe}R_{E2}} \times (1+h_{fe})R_{E2}$$

$$= \frac{(1+h_{fe})^2 R_{E2}}{1+h_{oe}h_{fe}R_{E2}}$$

Voltage gain (A_{V1}) :.

$$\frac{V_{o1}}{V_{i_1}} = \frac{V_{e1}}{V_{b1}} = 1 - \frac{h_{ie}}{R_{i1}} \quad \text{(similar to stage 2)}$$

output resistance (R_{o1})

$$R_{o1} = \left.\frac{V_{o1}}{I_{o1}}\right|_{V_S \to 0} \quad \& \quad R_{L_1} \to \infty$$

Hence Now $R_{o2} = \dfrac{\left[\dfrac{R_s + h_{ie}}{1+h_{fe}}\right] + h_{ie}}{1+h_{fe}}$ (Ω)

For Darlington Amplifier

(1) $A_I = \dfrac{I_o}{I_i} = \dfrac{I_{o2}}{I_{i_1}} = \dfrac{I_{o1}}{I_{i_1}} \times \dfrac{I_{o2}}{I_{i2}} = A_{I_1} \times A_{I_2}$

$$= \frac{1+h_{fe}}{1+h_{oe}h_{fe}R_{E2}} \times (1+h_{fe})$$

$$= \frac{(1+h_{fe})^2}{1+h_{oe}h_{fe}R_{E^2}}$$

(2) $R_i = \dfrac{V_i}{I_i} = \dfrac{V_{i1}}{I_{i1}} = R_{i1} =$

$$\frac{(1+h_{fe})^2 \times R_{E2}}{1+h_{oe}h_{fe}R_{E2}} = A_I R_{E2} (\Omega)$$

(3) $A_V = \dfrac{V_o}{V_i} = \dfrac{V_{o2}}{V_{i1}} = \dfrac{V_{o1}}{V_{i1}} \times \dfrac{V_{o2}}{V_{i2}}$

$= A_{V1} \cdot A_{V2} = \left[1 - \dfrac{h_{ie}}{R_{i1}}\right]\left[1 - \dfrac{h_{ie}}{R_{i2}}\right]$

$= 1 - \dfrac{h_{ie}}{R_{i2}} - \dfrac{h_{ie}}{R_{i1}} - \dfrac{h_{ie} \cdot h_{ie}}{R_{i1} \cdot R_{i2}}$

$\simeq 1 - \dfrac{h_{ie}}{R_{i2}} = A_{V2}$

(4)

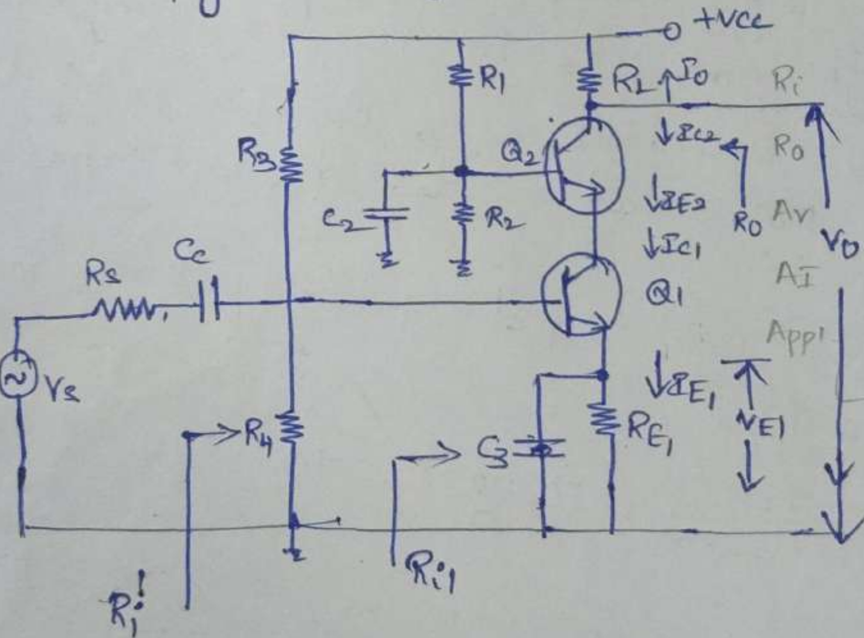
$R_o = \dfrac{V_o}{I_o}\Bigg|_{V_S \to 0 \;\&\; R_L \to \infty}$

$= \dfrac{\left[\dfrac{R_S + h_{ie}}{1 + h_{fe}}\right] + h_{ie}}{1 + h_{fe}} \quad (\Omega)$

3. ## Cascode Amplifier:-

Cascode amplifier is a composite amplifier with a large band width used for RF applications and as a video amplifier. It consists of a CE stage followed by a CB stage directly coupled to each other and combines some of the features of both the amplifiers.

For high frequency applications, CB configuration has the most desirable characteristics. However, it suffers from low input impedance ($Z_i = h_{ib}$). The cascode configuration is designed to have the input impedance essentially that of CB amplifier and good isolation between the input and output



	CB	CE	CC
Ri	100Ω	1kΩ 750Ω	500kΩ 750 kΩ
Ro	1M 450kΩ	40kΩ 45kΩ	50Ω
Av	150	500	1
AI	<1	High β 20 to few hundreds	High appreciable (20 to few hundreds)
Appl	for high freq	For AF	For impedance match

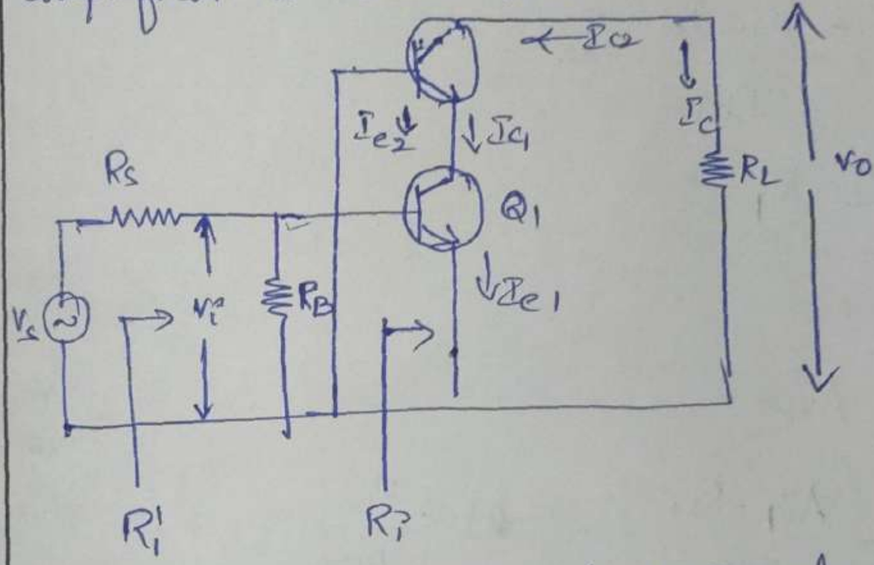
fig:- Cascode amplifier

The a.c equivalent

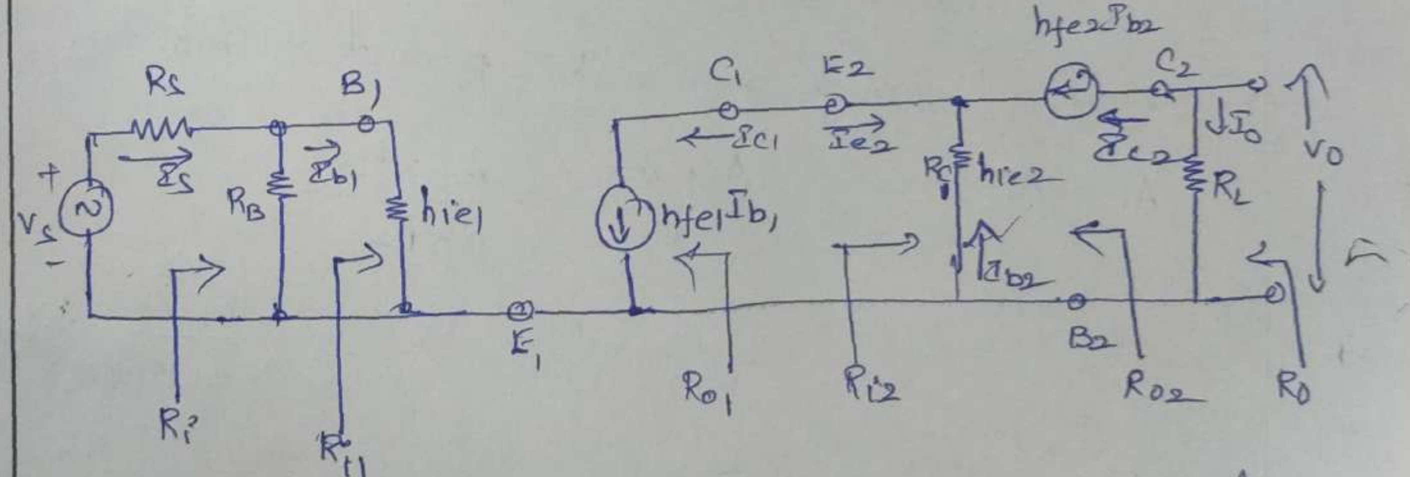
→ The cascode configuration is designed to have high input impedance bcz of CE amplifier, high current gain of CE and high voltage gain of CB amp & good isolation b/w i/p & o/p

→ AC equivalent circuit for cascode amplifier is drawn by shorting the d.c supply and coupling capacitors

→ The simplified h-parameter equivalent circuit for cascode amplifier is drawn below



AC equivalent circuit of Cascode amplifier



Simplified h-parameter equivalent circuit of Cascode amplifier

Analysis of second stage (CB) amplifier

Current gain $A_{I2} = \dfrac{I_{c2}}{I_{e2}} = \dfrac{-h_{fe}I_{b2}}{-(I_{b2}+h_{fe}I_{b2})} = \dfrac{h_{fe}\cancel{I_{b2}}}{\cancel{I_{b2}}[1+h_{fe}]} = \dfrac{h_{fe}}{1+h_{fe}}$

Input resistance $R_{i2} = \dfrac{V_{e2}}{I_{e2}} = \dfrac{-I_{b2}h_{ie}}{-(I_{b2}+h_{fe}I_{b2})} = \dfrac{-\cancel{I_{b2}}h_{ie}}{-\cancel{I_{b2}}[1+h_{fe}]} = \dfrac{h_{ie}}{1+h_{fe}}$

3. Voltage gain $A_{V_2} = \frac{V_{ce_2}}{V_{e_2}} = \frac{V_{o2}}{V_{e2}} = \frac{-h_{fe} \cancel{I_{b_2}} R_L}{-\cancel{I_{b_2}} h_{ie}} = \frac{h_{fe} R_L}{h_{ie}}$.

Analysis of first stage CE amplifier

(1) Current gain A_{I_1} =

$$A_{i_1} = \frac{I_{C1}}{I_{b1}} = \frac{-h_{fe} \cancel{I_{b1}}}{\cancel{I_{b1}}} = -h_{fe}$$

(2) Input impedance R_{i} =

$$R_i = h_{ie}$$

(3) Voltage gain A_{V_1} =

$$A_{V_1} = \frac{V_{C1}}{V_{b1}} = \frac{A_{i_1} \cdot R_{L_1}}{R_{i_1}} = -h_{fe} \cdot \frac{R_{L_1}}{h_{ie}}$$

$\because V_{C1} = I_{C1} R_{L1}$

$\because V_{b1} = I_{b1} R_{i1}$

$\therefore A_{i1} = \frac{I_{C1}}{I_{b1}}$

Here $R_{L1} = R_{i2}$

→ The overall voltage gain is the product of individual gains

$$A_V = A_{V_1} \times A_{V_2} = \frac{A_{i1} \times R_{L_1}}{R_{i_1}} \times \frac{A_{i2} \times R_{L2}}{R_{i2}}$$

$$= \frac{-h_{fe}}{h_{ie}} \times \frac{h_{fe}}{1+h_{fe}} \cdot R_{L_2} = \frac{-h_{fe}^2}{h_{ie}(1+h_{fe})} R_{L2}$$

→ The overall i/p resistance (R_i)

$$R_i = R_{i_1} \| R_B = R_{i_1} \| R_3 \| R_4$$

→ Overall voltage gain (A_{VS}) by considering source

$$A_{VS} = \frac{V_o}{V_S} = \frac{V_o}{V_i} \times \frac{V_i}{V_S} = A_V \times \frac{R_i}{R_i + R_S}$$

$$V_i = \frac{V_S R_i}{R_i + R_S}$$

→ Overall current gain (A_{is}) = $\frac{I_o}{I_S}$ =

$$\frac{I_o}{I_s} = \frac{I_o}{I_{C2}} \times \frac{I_{C2}}{I_{e2}} \times \frac{I_{e2}}{I_{C1}} \times \frac{I_{C1}}{I_{b1}} \times \frac{I_{b1}}{I_s}$$

$\frac{I_o}{I_{c2}} = -1, \quad \frac{I_{c2}}{I_{e2}} = -A_{i2}, \quad \frac{I_{e2}}{I_{c1}} = -1, \quad \frac{I_{c1}}{I_{b1}} = -A_{i1},$

$\frac{I_{b1}}{I_s} = \frac{R_B}{R_B + R_{i1}}$

$\therefore A_i = A_{i2} \times A_{i1} \times \frac{R_B}{R_B + R_{i1}}$

→ Output resistance R_0

$R_0 = R_{01} = R_{02} = \infty, \quad R_0 = R_{02} \| R_L \Rightarrow R_L$

Total current gain $A_I = \frac{I_o}{I_i} = \frac{I_L}{I_i}$

$I_L = -h_{fe} I_{b2}$

I_{b2} → apply KCL at output

$I_{b2} + h_{fe} i_{b2} = h_{fe} i_{b1}$

$i_{b2} = \frac{h_{fe} i_{b1}}{1 + h_{fe}}$

At port (1) current division

$i_{b1} = \frac{I_i \times R_b}{R_b + h_{ie}}$

Branch current = $\frac{\text{total } I \times \text{opposite } R}{\text{Total } R}$

$\therefore$

$A_I = \frac{-h_{fe} h_{fe} \cdot \cancel{I_i} \times \frac{R_b}{R_b + h_{ie}}}{\cancel{I_i}} = \frac{-h_{fe}^2}{R_b + h_{ie}} \times R_b$

$A_v = \frac{v_L}{v_i} = \quad v_L = i_L \cdot R_L$

$v_i = I_i [R_1 \| R_2 \| h_{ie}]$

$A_v = \frac{I_L \cdot R_L}{I_i (R_b \| h_{ie})}$

$\boxed{A_v = A_I \cdot \frac{R_L}{R_b \| h_{ie}}}$

Advantages:-

1. High Band width, B.W is high due to elemination of Miller effect
2. High Gain
3. High stability
4. High input impedance

Disadvantages

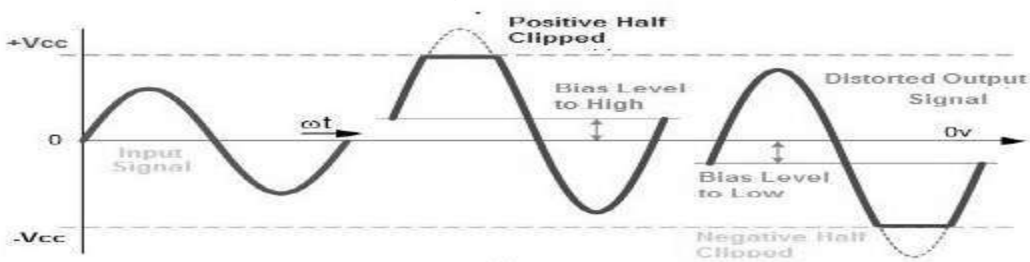
1. Requires High voltage gain

Applications

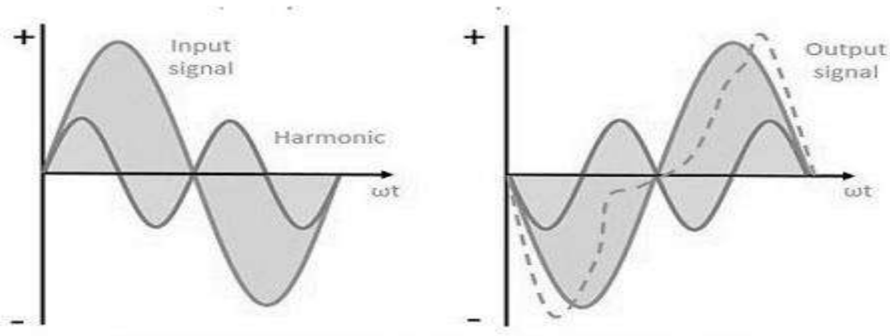
1. Used in Tuned amplifiers
2. Used in RF application as video amplifier

DISTORTION IN AMPLIFIERS:
Amplifier distortion can be defined as; any difference from the input signal of an amplifier that occurs throughout the amplification process and gives a changed output signal in terms of magnitude, shape, frequency content, etc. It occurs due to many factors like; non-linearity within the components of the amplifier, improper biasing, or amplifier overloading. The amplifier distortion is undesirable because it degrades the value of the amplified signal.

Types of Amplifier Distortion
1. Non-linear Distortion
Non-linear distortion mainly happens in an amplifier whenever the input signal applied is large & the active device is driven into a non-linear area of its characteristics. This distortion is used to describe a non-linear relationship between the input & output signals of an amplifier. So this distortion results from systems wherever the output signal is not proportional precisely to the input signal & intermodulation products or harmonics are generated.
2. Amplitude Distortion
Amplitude distortion is a type of nonlinear distortion that takes place because of the attenuation within the crest value of the signal. The shift within the Q point & amplification for below 360° of the signal leads mainly to distortion in amplitude. This distortion mainly occurs because of clipping & incorrect biasing. We know that that if the transistor's biasing point is correct, the output is similar to the input within the amplified shape. This can be understood through the following cases.
Suppose inadequate biasing is provided to the amplifier, then the Q-point will lie close to the minor half of the load line. So in this condition, the input signal's negative half is clipped & we acquire a distorted output signal of the amplifier.
If we provide an additional bias potential, then the Q-point will be at the higher side of the load line. So this condition provides an output that will be cut off at the positive half of the waveform.
Proper biasing can also lead to distortion sometimes within the output in case the input signal is large because this input signal is amplified through the amplifier's gain. So both the positive & negative half of the waveform will get clipped at some part which is called clipping distortion.



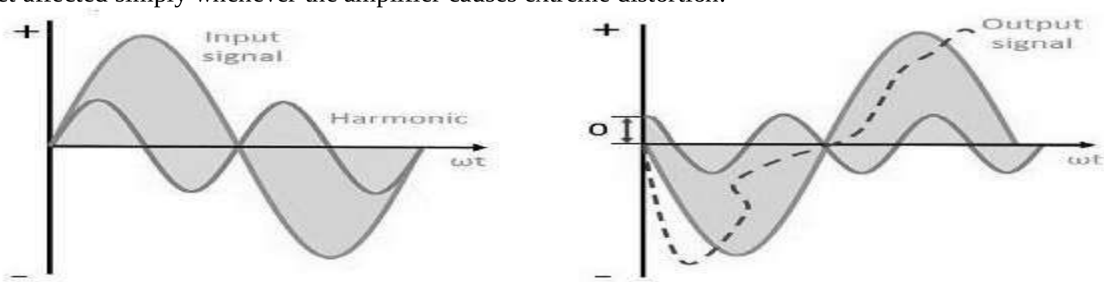
3. Linear Distortion
Linear distortion mainly occurs whenever the input signal applied to drive the device is small & functions in the linear section of its characteristics. So this distortion mainly happens because of active devices' frequency-dependent characteristics.
4. Frequency Distortion
In this type of distortion, the amplification level changes in frequency. The input signal during amplification in a realistic amplifier includes fundamental frequency with different frequency components which are called harmonics. The harmonic amplitude (HA) after amplification is fairly a fraction of the basic amplitude. It doesn't cause any severe cause to the output waveform. If the HA after amplification goes to a high value, its effect cannot be avoided because it is visible at the output.
Here the input has fundamental frequency including harmonics. So the combination of the two on amplification provides a distorted signal at the output. It happens either because of the occurrence of reactive elements (or) through the amplifier circuit's electrode capacitances.



Frequency Type

5. Phase Distortion

Phase Distortion is also called delay distortion in the amplifier because whenever there is a time delay between the input & output signal then it is said to be phase distorted signal. This distortion mainly occurs because of electrical reactance. Earlier we have discussed that a signal includes different frequency components thus, whenever different frequencies experience different phase shifts then phase distortion occurs. This type of distortion has no practical importance in audio amplifiers because the human ear is insensible to phase shift. The type & quantity of distortion that is bearable or unbearable mainly depends on the amplifier’s application. Usually, the working of the system will get affected simply whenever the amplifier causes extreme distortion.



1. Differential Amplifier:

1. The differential amplifier is a basic building block of an op-amp.
2. The main purpose of the differential amplifier stage is to provide high gain to the difference mode signal & cancel the common mode signal.
3. The differential amplifier amplifies the difference between two input voltage signals. Hence it is called difference amplifier (or differential amplifier)

→ V_1 & V_2 are the two input signals while V_0 is the single ended output.

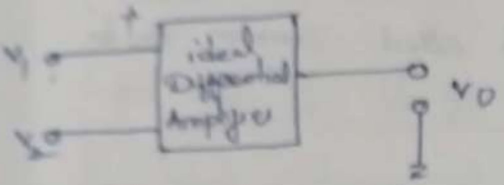
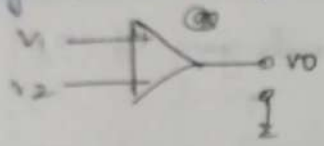


fig: ideal differential Amplifier



→ In an ideal differential amplifier, the output voltage V_0 is proportional to the difference b/w the two input signals.

Hence we can write,

$$\boxed{V_0 \propto (V_1 - V_2)} \rightarrow (1)$$

Differential Gain (Ad)

From equation (1) we can write,

$$\boxed{V_0 = A_d (V_1 - V_2)} \rightarrow (2)$$

Ad is the gain with which differential amplifier amplifies the difference between two input signals. Hence it is called differential gain of the differential amplifier.

A_d = differential gain

→ The difference b/w the two inputs $(V_1 - V_2)$ is generally called difference voltage and denoted as V_d.

$$\boxed{V_0 = A_d V_d} \rightarrow (3)$$

→ Hence the differential gain can be expressed as,

$$\boxed{A_d = \frac{V_0}{V_d}} \rightarrow (4)$$

→ Generally the differential gain is expressed in its decibel (dB) value as.

$$\boxed{A_d = 20 \log_{10} (A_d) \text{ in dB}}$$

Common Mode gain (Ac):

Let's consider if we apply two input voltages which are equal i.e $V_1 = V_2$ to a differential amplifier. Then ideally the output must be equal to zero.

But the output voltage of a practical differential amplifier not only depends on the difference voltage, but also depends on the average of two inputs.

The average of two input signals is called common mode signal, which is denoted as V_C.

$$\boxed{V_C = \frac{V_1 + V_2}{2}}$$

. The gain with which op-amp amplifies the common mode signal to produce the output is called common mode gain of an op-amp & is denoted as Ac

$$\boxed{V_O = A_C V_C}$$ ← There is some finite output for $V_1 = V_2$ in case of practical op-amps.

So the total output of any differential amplifier can be expressed as

$$\boxed{V_O = A_d V_d + A_c V_c}$$

Key point: For an ideal differential amplifier, the differential gain Ad must be infinite while the common mode gain must be zero.

Common Mode Rejection Ratio [CMRR]:

When the same voltage is applied to both the inputs, the differential amplifier is said to be operated in common mode configuration. Many disturbance signals, noise signals appear as a common input signal to both the input terminals of the differential amplifier. Such a common signal should be rejected by the differential amplifier.

" The ability of a differential amplifier to reject a common mode signal is expressed by a ratio called common mode rejection ratio denoted as CMRR.

→ It is defined as the ratio of the differential voltage gain Ad to common mode voltage gain Ac

$$\boxed{CMRR = \rho = \left|\frac{A_d}{A_c}\right|}$$

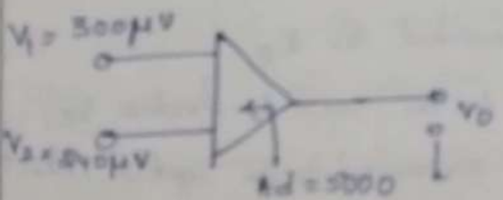
→ Key Point :- Ideally the common mode voltage gain is zero, hence the ideal value of CMRR is infinite.

For a practical differential amplifier Ad is large & Ac is small hence the value of CMRR is also very large.

Features of Differential Amplifier :-

1. High differential voltage gain
2. Low common mode gain
3. High CMRR
4. High input impedance
5. Large bandwidth
6. Low offset voltages & currents
7. Low output impedance

Problem Determine the output voltage of a differential amplifier for the input voltages of 300μV & 240μV. The differential gain of the amplifier is 5000 and the value of the CMRR is (i) 100 & (ii) 10^5



Given

$V_1 = 300\mu V$, $V_2 = 240\mu V$.

$A_d = 5000$

CMRR is (i) 100, (ii) 10^5

$V_o = A_d V_d + A_c V_c$

(i) CMRR is 100

$V_d = V_1 - V_2 = 300\mu V - 240\mu V = 60\mu V$

$V_c = \frac{V_1 + V_2}{2} = \frac{300\mu V + 240\mu V}{2} = 270\mu V$

$CMRR = \frac{A_d}{A_c} = 100 = \frac{5000}{A_c} \Rightarrow A_c = \frac{5000}{100} = 50$

$V_o = A_d V_d + A_c V_c = 5000 \times 60 + 50 \times 270 = 313500\ \mu V = 313.5\ mV$

(ii) $CMRR = 10^5$ $\quad A_c = \frac{A_d}{CMRR} = \frac{5000}{10^5} = 0.05$

$V_o = A_d V_d + A_c V_c = 5000 \times 60 + 0.05 \times 270 = 300013.5\ \mu V = 3000.135\ mV$

Differential Amplifier (or) Transistorised Differential Amplifier

The transistorised differential amplifier basically uses the emitter biased circuit which are identical characteristics.

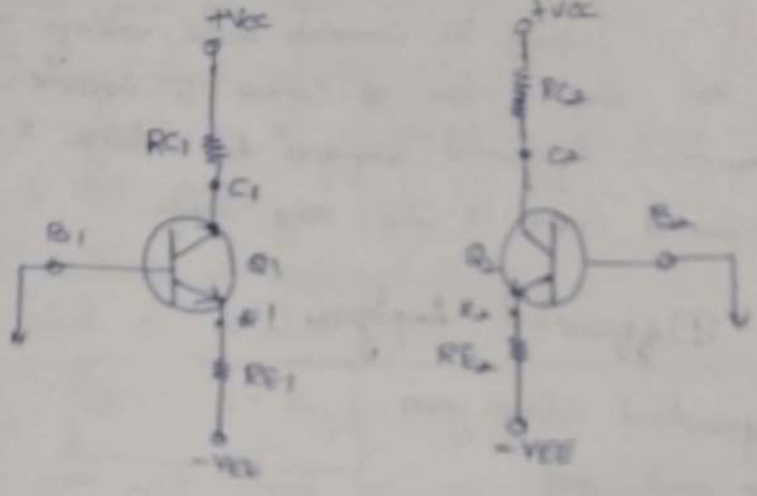


fig:- Emitter Biased Circuit.

The two transistor Q_1 & Q_2 have exactly matched characteristics. The two collector resistances R_{C1} & R_{C2} are equal, the two emitter resistances R_{E1} & R_{E2} are also equal.

→ The magnitudes of $+V_{CC}$ & $-V_{EE}$ are also same. The differential amplifier can be obtained by using such two emitter biased circuits. This is achieved by connecting emitter E_1 of Q_1 to the emitter E_2 of Q_2.

→ R_{E1} appears in parallel with R_{E2} and the combination can be replaced by a single resistance denoted as R_E.

→ The base B_1 of Q_1 is connected to the input 1 which is V_{S1} while the base of B_2 of Q_2 is connected to the input 2 which is V_{S2}.

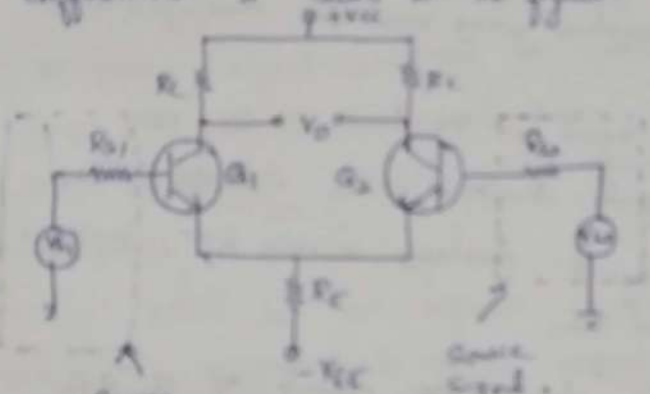
→ The balanced output is taken between the collector C_1 of Q_1 and the collector C_2 of Q_2. Such an amplifier is called emitter coupled differential amplifier. The two collector resistances are same hence can be denoted as R_C.

→ The o/p can be taken b/w two collectors or in b/w one of the two collectors and the ground.

→ When the output is taken between the two collectors, none of them is grounded then it is called balanced output, double ended output or floating output.

→ When the output is taken b/w any of the collectors and the ground, it is called unbalanced output or single ended output.

→ The complete circuit diagram of such a dual input balanced output differential is shown in the figure.



Dual input balanced output differential amplifier

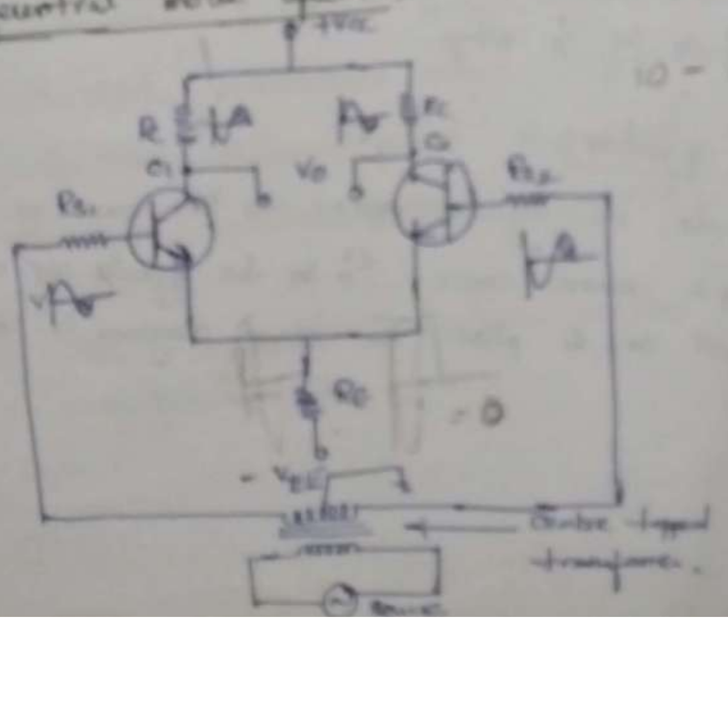
As the o/p is taken b/w

Let us study the circuit operation in the two modes namely.

i) Differential mode operation.

ii) Common mode operation.

Differential mode operation:



In the differential mode, the two input signals are different from each other. Consider the two input signals which are same in magnitude but 180° out of phase.

These signals with opposite phase can be obtained from the center tap transformer.

Assume that the sine wave on the base of Q_1 is positive going while on the base of Q_2 is negative going. With a positive going signal on the base of Q_1, an amplified negative going signal develops on the collector of Q_1.

Due to positive going signal, current through R_E also increases and hence a positive going wave is developed across R_E.

Due to negative going signal on the base of Q_2, an amplified positive going signal develops on the collector of Q_2. And a negative going signal develops across R_E, because of emitter follower action of Q_2.

So voltage across R_E, due to the effect of Q_1 & Q_2 are equal in magnitude and 180° out of phase. Hence these two signals cancel each other and there is no signal across the emitter resistance. Hence no ac signal at R_E. R_E in this case does not introduce negative feedback.

V_0 is the output taken across collector of Q_1 and collector of Q_2. The two outputs on collector 1 & 2 are equal in magnitude but opposite in polarity. if o/p voltage at $C_1 = 10V$

$V_0 = V_1 - V_2$

Ex. $V_0 = +10 - (-10) = 20$. Hence the difference output V_0 is twice as large as the i/p voltage.

Common mode operation:-

In this mode, the signals applied to the base of Q_1 & Q_2 are derived from the same source. So the two signals are equal in magnitude as well as in phase. This circuit diagram is shown in the figure.

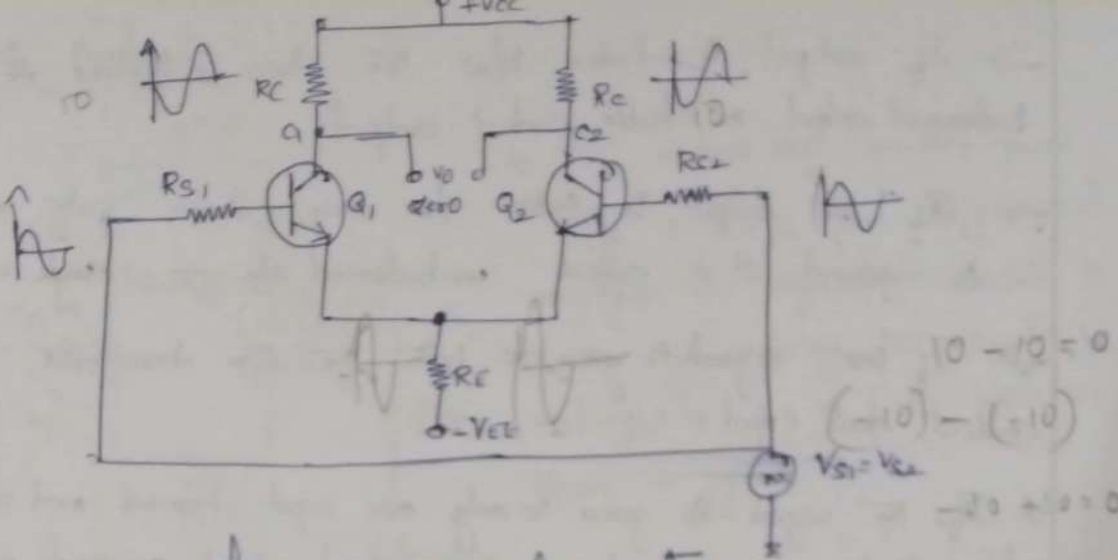


fig: Common mode operation

In phase signal voltages at the bases of Q_1 & Q_2 causes in phase signal voltages to appear across R_E, which add together. Hence R_E carries a signal current and provides a negative feedback. This feedback reduces the common mode gain of differential amplifier,

While the two signals causes in phase signal voltages of equal magnitude to appear across the two collectors of Q_1 & Q_2. Now the o/p voltage is the difference b/w the two collector voltages, which are equal and also same in phase, e.g. $(10)-(10)=0$. Thus the difference output V_o is almost zero. negligibly small. Ideally it should be zero.

Types of Differential Amplifier:- (or) Differential Amplifier Configuration

1. Dual input, balanced output differential amplifier.
2. Dual input, unbalanced output differential amplifier
3. Single input, balanced output differential amplifier.
4. Single input, unbalanced output differential amplifier

The differential amplifier uses two transistors in common emitter configuration. ;

→ If output is taken b/w the two collectors it is called balanced output or double ended output.

→ If the output is taken b/w one collector with respect to ground it is called unbalanced o/p or single ended o/p.

→ If the signal is given to both the i/p terminals it is called dual input.

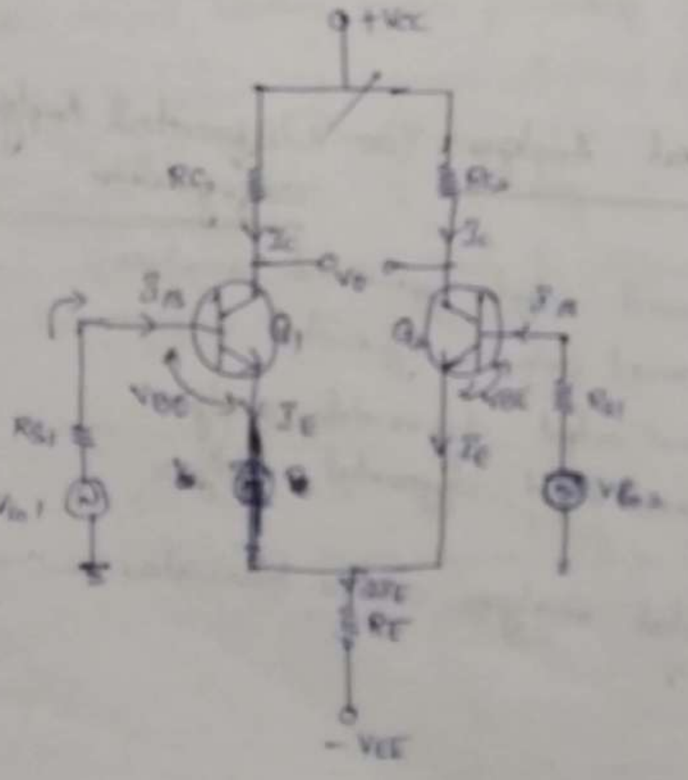
→ If the signal is given to only one input terminal and other terminal is grounded it is called single input or single ended i/p.

→ Out of these four configurations the dual i/p, balanced o/p is the basic differential amplifier configuration.

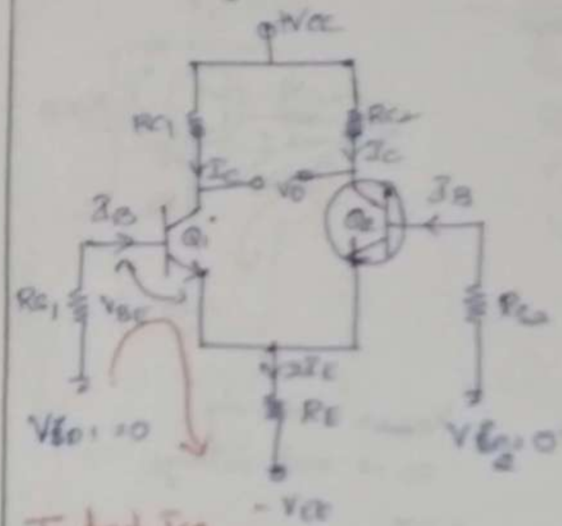
1. (i) Dual input Balanced output Differential amplifier [DIBO Differential amplifier]

→ The two input signals V_{in1} & V_{in2} are applied to the bases B_1 and B_2 of transistors Q_1 and Q_2. The

→ The output is measured b/w the two collectors C_1 & C_2 which are at the same dc potential.



DC Analysis:-



→ To obtain the operating point I_{CQ} & V_{CEQ} for differential amplifier dc equivalent circuit is drawn by reducing the input voltages V_1 & V_2 to zero as shown in figure.

[DC analysis for only half ckt because two transistor are matched.]

To find I_{CQ}

→ Apply KVL at i/p side to find I_{CQ}

$$I_B R_S + V_{BE} + 2I_E R_E - V_{EE} = 0 \rightarrow ①$$

$$I_B R_S + V_{BE} + 2I_E R_E = V_{EE} \rightarrow ②$$

$I_C = I_E$

$I_C = \beta I_B$

$I_E = \beta I_B$

$I_B = \frac{I_E}{\beta}$

Sub I_B in eqn ②

$$\frac{I_E}{\beta} R_S + 2I_E R_E = V_{EE} - V_{BE}$$

$$I_E \left[\frac{R_S}{\beta} + 2R_E \right] = V_{EE} - V_{BE}$$

$$I_E = \frac{V_{EE} - V_{BE}}{\frac{R_S}{\beta} + 2R_E} \rightarrow ③$$

$\beta \rightarrow$ Gain of the transistor is very high $\therefore \frac{R_S}{\beta}$ is << $2R_E$

so the above equation becomes.

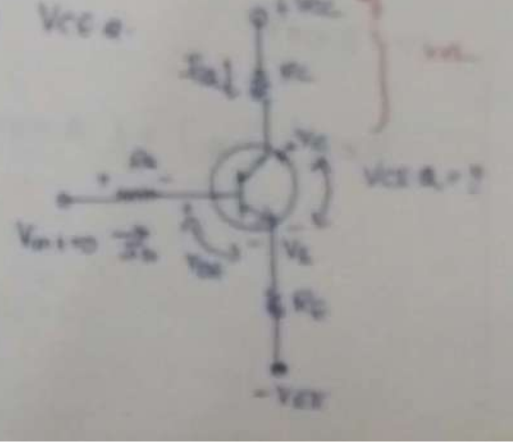
$$\boxed{I_E = I_C = I_{CQ} = \frac{V_{EE} - V_{BE}}{2R_E}}$$

∴ R_S/β can be neglected in two condition

1. If R_S is not present
2. If β is large value.

To Find V_{CEQ}: Consider the same half circuit

→ Apply KVL at o/p side to find V_{CEQ}



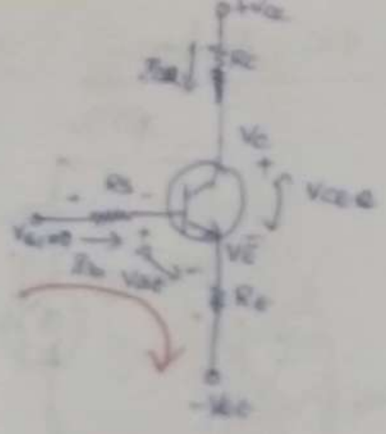
$$V_{CE} = V_C - V_E \rightarrow ①$$

$$V_{CC} - I_{CQ} R_C - V_C = 0$$

$$V_C = V_{CC} - I_{CQ} R_C \rightarrow ②$$

Sub ② in ①

$V_{CE} = V_{CC} - I_{CQ}R_C - V_E \rightarrow (3)$



to find V_E apply kvl at i/p loop

at input AC signal is zero,

$I_B R_S + V_{BE} + V_E = 0$

Transistor gain is very high
so we can ignore the $I_B R_S$

So

$\boxed{V_E = -V_{BE}} \rightarrow (4)$

$V_{BE} \rightarrow$ cut-in vtg
$V_E \rightarrow$ node voltage

sub (4) in (3)

$$\boxed{V_{CEQ} = V_{CC} - I_{CQ}R_C + V_{BE}}$$

So the operating point (1) $I_{CQ} = \dfrac{V_{EE} - V_{BE}}{2R_E}$

(2) $V_{CEQ} = V_{CC} - I_{CQ}R_C + V_{BE}$

The DC analysis is same for all differential amplifier configurations.

Problems:

(1) $R_C = 2.2k\Omega$, $R_E = 4.7k\Omega$, $R_1 = R_2 = 50\Omega$, $+V_{CC} = 10V$, $-V_{EE} = -10V$, $\beta_{dc} = 100$ and $V_{BE} = 0.715V$. Determine the operating points (I_{CQ} and V_{CEQ}) of the two transistors

$I_C = I_E = \dfrac{V_{EE} - V_{BE}}{2R_E}$

$= \dfrac{10 - 0.715}{2 \times 4.7k}$

$I_{CQ} = 0.988mA$

$I_C \approx I_E$

$V_{CEQ} = V_{CC} + V_{BE} - I_C R_C$

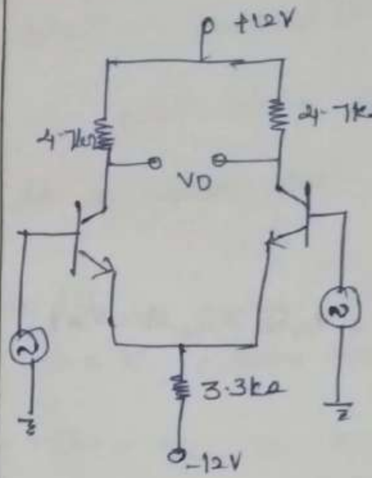
$V_{CEQ} = 10V + 0.715V - (0.988mA)(2.2k)$

$V_{CE} = 10 + 0.715 - (0.988 \times 10^{-3})(2.2 \times 10^{3})$

$= 8.54V$

$(V_{CEQ}, I_{CQ}) = (8.54V, 0.988mA)$

2 Calculate operating point values for the circuit shown in figure.



From the figure

$R_{C1} = R_{C2} = R_C = 4.7k\Omega$

$R_E = 3.3k\Omega$

$V_{CC} = +12V$

$V_{EE} = -12V$

W.K.T

① $I_{EQ} = I_{CQ} = \dfrac{V_{EE} - V_{BE}}{2R_E}$

$= \dfrac{12 - 0.7}{2 \times 3.3k} = 1.712mA$

② $V_{CEQ} = V_{CC} + V_{BE} - I_{CQ} R_C$

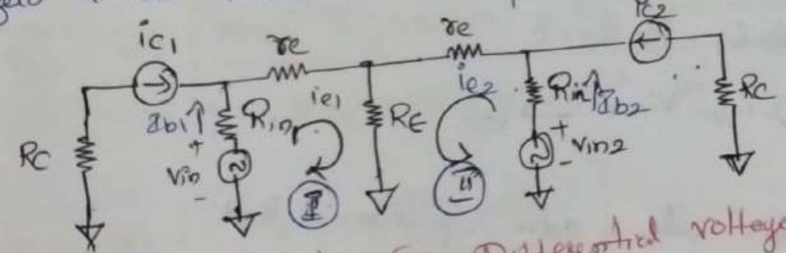
$= 12 + 0.7 - 1.712 \times 10^{-3} \times 4.7 \times 10^{3}$

$= 4.653V$

∴ Q point is [1.712mA, 4.653V]

Ac Analysis

To obtain the small signal Ac analysis let us consider an equivalent circuit by making the dc voltages (Vcc & VEE) equal to zero & both transistors are replaced by its equivalent 're' model.



① Voltage gain (Ad) :- (or) Differential voltage gain (Ad)

Current flowing to loop 1 is i_{e1} &

Current flowing to loop 2 is I_{e2}

By applying KVL in loop ① & ②

For loop ①

$V_{in1} - i_{b1} R_{in1} - i_{e1} r_e - R_E(i_{e1} + i_{e2}) = 0 \rightarrow ①$

$i_{b1} = i_{e1}/\beta_{ac} \rightarrow$

By applying i_{b1} value in equ ①.

$V_{in1} - \dfrac{R_{in1} \cdot i_{e1}}{\beta_{ac}} - i_{e1} r_e - (i_{e1} + i_{e2}) \cdot R_E = 0$

Since $\frac{R_{in1}}{\beta_{ac}}$ is very small, we should neglect the term,

~~likewise.~~

~~$V_{in1} = i_{e1} r_e + (i_{e1}+i_{e2}) R_E$~~

~~V_{in1} ie.~~

$i_{e1} \cdot r_e + (i_{e1}+i_{e2}) R_E = V_{in1} \Rightarrow i_{e1} \cdot r_e + i_{e1} R_E + i_{e2} R_E = V_{in1}$

$(r_e + R_E) \cdot i_{e1} + R_E \cdot i_{e2} = V_{in1} \rightarrow ①$

Apply KVL @ loop ②

$V_{in2} - i_{b2} \cdot R_{in2} - i_{e2} \cdot r_e - R_E (i_{e1} + i_{e2}) = 0 \rightarrow ②$

Since $i_{b2} = i_{e2} / \beta_{ac}$

~~$V_{in2} - \frac{i_{e2}}{\beta_{ac}} * R_{in2} -$~~

Sub i_{b2} value in the above equation ②

$$V_{in2} - \frac{i_{e2}}{\beta_{ac}} \cdot R_{in2} - i_{e2} \cdot r_e - R_E (i_{e1} + i_{e2}) = 0$$

Since R_{in2}/β_{ac} is small so neglect it

then the above equation becomes

$V_{in2} - i_{e2} \cdot r_e - R_E (i_{e1} + i_{e2}) = 0$

$V_{in2} - i_{e2} \cdot r_e - R_E i_{e1} - R_E i_{e2} = 0$ $\quad V_{in2} = i_{e2} r_e + R_E i_{e1} + R_E i_{e2}$

$R_E i_{e1} + (r_e + R_E) \cdot i_{e2} = V_{in2} \rightarrow ②$ $\quad V_{in2} = R_E i_{e1} + i_{e2}(r_e + R_E)$

By using Crammer's rule solve equation 1 & 2

$(r_e + R_E) I_{e1} + R_E \cdot i_{e2} = V_{in1} \rightarrow ①$

$R_E \cdot i_{e1} + (r_e + R_E) \cdot i_{e2} = V_{in2} \rightarrow ②$

$ax + by = e$

$cx + dy = f$

$x = \frac{\begin{vmatrix} e & b \\ f & d \end{vmatrix}}{\begin{vmatrix} a & b \\ c & d \end{vmatrix}} \quad y = \frac{\begin{vmatrix} a & e \\ c & f \end{vmatrix}}{\begin{vmatrix} a & b \\ c & d \end{vmatrix}}$

$$i_{e1} = \frac{\begin{vmatrix} V_{in1} & R_E \\ V_{in2} & r_e + R_E \end{vmatrix}}{\begin{vmatrix} r_e + R_E & R_E \\ R_E & r_e + R_E \end{vmatrix}} = \frac{(r_e + R_E) V_{in1} - R_E V_{in2}}{(r_e + R_E)^2 - (R_E)^2} \rightarrow ③$$

Similarly for

$$i_{e2} = \frac{\begin{vmatrix} r_e + R_E & V_{in1} \\ R_E & V_{in2} \end{vmatrix}}{\begin{vmatrix} r_e + R_E & R_E \\ R_E & r_e + R_E \end{vmatrix}} = \frac{(r_e + R_E) V_{in2} - R_E \cdot V_{in1}}{(r_e + R_E)^2 - (R_E)^2}$$

→ Since it is a dual input balanced output differential amplifier since the output is measured at both the collector terminal.

→ So output voltage V_0 is the difference b/w the two input voltages available at the collector terminals.

let us assume that the

→ let us assume that the collector terminal '1' is at higher potential w.r.to collector terminal '2' then V_0 is

$$V_0 = V_{C2} - V_{C1}$$

$$= -i_{C_2} \cdot R_C - (-i_{C1} R_C)$$

$$= -i_{C2} \cdot R_C + i_{C1} R_C$$

$$= R_C (i_{C1} - i_{C2})$$

$$= R_C (i_{e1} - i_{e2}) \quad (\because i_e \simeq i_C)$$

Substituting i_{e1} & i_{e2} values in the above equation

$$V_0 = R_C \left[\frac{(r_e + R_E) V_{in1} - R_E V_{in2} - (r_e + R_E) V_{in2} + R_E V_{in1}}{(r_e + R_E)^2 - (R_E)^2} \right]$$

collecting similar term together since we can

$$V_0 = R_C \left[\frac{(r_e + R_E)(V_{in} - V_{in2}) + R_E [V_{in1} - V_{in2}]}{r_e^2 + 2 r_e \cdot R_E + \cancel{R_E}^2 - \cancel{R_E}^2} \right]$$

$$V_0 = R_C \left[\frac{(V_{in1} - V_{in2}) \cancel{[r_e + 2R_E)}}{r_e \cancel{(r_e + 2R_E)}} \right]$$

$$V_0 = \frac{R_C}{r_e} V_{id} \qquad \therefore V_{in1} - V_{in2}$$

$$\text{voltage gain } A_d = \frac{V_0}{V_{id}} = \frac{R_C}{r_e}$$

$$\boxed{A_d = \frac{R_C}{r_e}}$$

4

feb-10-20

(2) Differential input resistance (R_i):

The differential input resistance at any one input terminal is defined as the equivalent resistance that would be measured at that terminal, while the other terminal is grounded.

$R_i = R_{in1}$ (or) R_{in2}

$\therefore R_{in1} = \left.\frac{V_{in1}}{i_{b1}}\right|_{V_2=0} \Rightarrow R_{in1} = \left.\frac{V_{in1}}{i_{e1}/\beta}\right|_{V_2=0}$

$\therefore I_c = \beta i_b$

$i_b = \frac{I_c}{\beta}$

substitute the value of i_{e1} in the above equation.

$$R_{in1} = \frac{V_{in1}\beta}{i_{e1}}$$

substitute the value of i_{e1} in the above equation

$$R_{in1} = \left.\frac{V_{in1}\beta}{\frac{V_{in1}(r_e+R_E) - V_2 R_E}{r_e^2+2r_eR_E}}\right|_{V_2=0}$$

$$R_{in1} = \frac{\cancel{V_{in1}}\beta(r_e^2+2r_eR_E)}{\cancel{V_{in1}}(r_e+R_E)}$$

But in practice $R_E \gg r_e$. neglect r_e

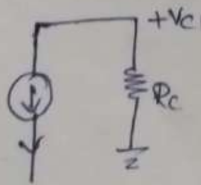
$R_{in1} = \cancel{V_{in1}}\beta$

$$R_{in1} = \frac{\beta\, 2r_e \cancel{R_E}}{\cancel{R_E}}$$

$$\boxed{R_{in1} = \beta 2r_e}$$

(3) Output resistance (R_o):

Output resistance is defined as the equivalent resistance that would be measured at either of the collector terminals with respect to ground.



From the above AC equivalent circuit the output resistance at C_1 or C_2 is

$R_o = R_{C1}$ (or) R_{C2}

$$\boxed{R_o = R_C}$$

[Current source offers high resistance → R_C having small resistance. And these two resistance are in parallel combination so gives small value $\simeq R_C$.

Scanned with OKEN Scanner

UNIT – II

HIGH FREQUENCY TRANSISTOR AMPLIFIERS – BJT & FET

BJT

→ Transistor at High frequencies

→ Hybrid – π Common Emitter transistor model

→ Validity of hybrid π model,

→ Determination of high - frequency parameters in terms of low frequency parameters.

→ Single stage CE Amplifier frequency response with short circuit load and resistive load, gain cutoff frequencies, Gain bandwidth product,

→ Emitter follower at higher frequencies.

→ Problems

FET

→ FET at High frequencies

→ High frequencies FET model

→ Analysis of Common source amplifier at High frequencies

→ Analysis of common drain amplifier at high frequencies

Introduction

The transistors are analysed at various frequencies

i.e) 1) at low frequencies → h-parameter model, re model
2) at high frequencies → Hybrid π model, y-parameter model.

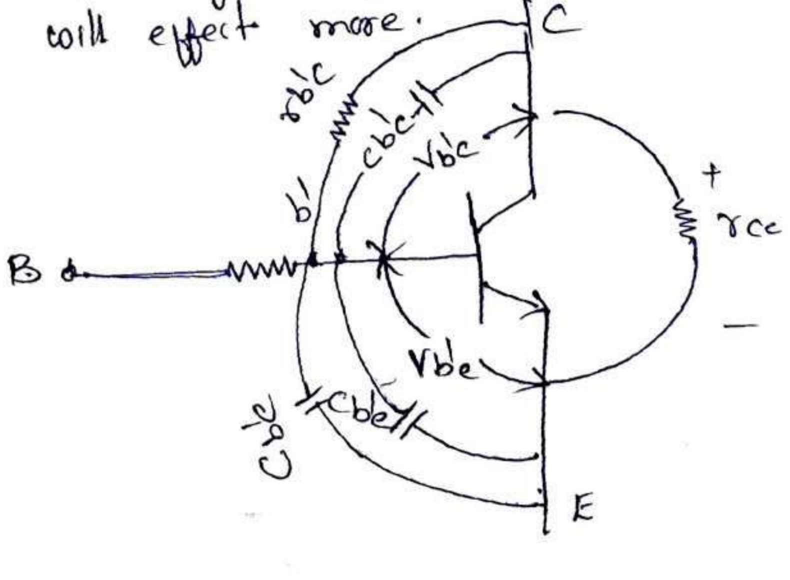
To analyse the transistor at lower frequencies we can replace the transistor with h-parameter model to find the parameter such as input impedance (R_i), output impedance (R_o), Voltage gain (A_V) & Current gain (A_I)

At high frequencies the transistor cannot be replaced by h-parameter model, because,

1. h-parameters are not constant at high frequencies
2. Not able to analyse transistor at each and every frequencies.
3. h-parameters are complex in nature at high frequencies

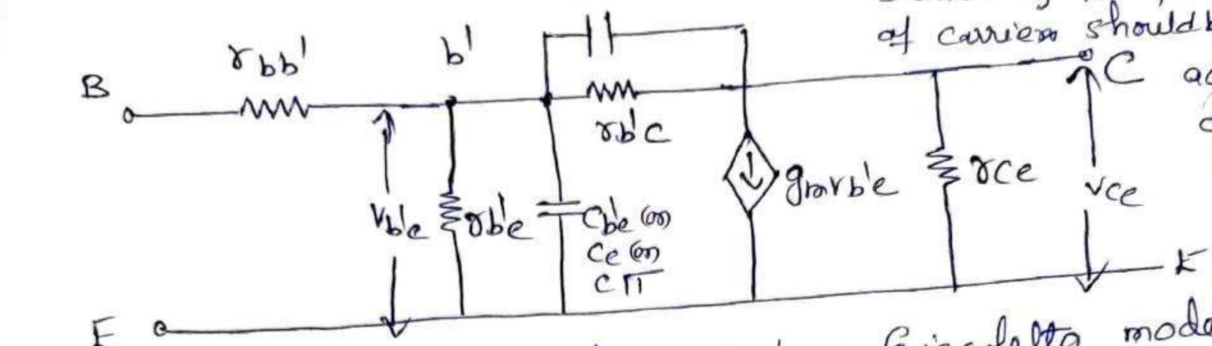
To analyze the transistor at high frequencies using Hybrid-π model.

→ At high frequencies the internal parameters of transistor will effect more.



Hybrid - π - model of BJT :-

1) At high frequencies, the capacitive effects of the transistor junctions and the delay in response of the transistor caused by the processes of diffusion of carriers should be taken into account for determining high frequencies.



Here → frequency π is also called as Giacoletto model.

$r_{bb'}$:- The resistance b/w external node and internal node of base terminal is $r_{bb'}$ it is also called "Base spreading resistance"

$r_{b'e}$:- The resistance between internal node of base terminal and emitter terminal.

$r_{b'c}$:- The resistance between internal node of base and collector terminal

r_{ce} :- The resistance between internal node of collector and emitter terminal.

$C_{b'e}$:- The internal capacitance between internal node of base and emitter terminals

→ It exists when base emitter junction in forward bias condition.

→ It is also called "diffusion (or) depletion capacitance.

$C_{b'c}$:- The internal capacitance between internal node of base and collector terminal.

→ It exists when collector to base junction is in reverse bias condition.

→ It is also called transition capacitance.

Determination of High frequency parameters in terms of low frequency parameters

(or)

Derivation of Hybrid-π parameters in terms of Hybrid parameters (or) H-parameters

The hybrid π model for the CE transistor at low frequencies is shown in fig (a). The h-parameter model for the same shown in fig (b).

→ The hybrid π model is drawn for low frequencies, the capacitive elements are considered as open circuit, because the capacitive reactance is very high.

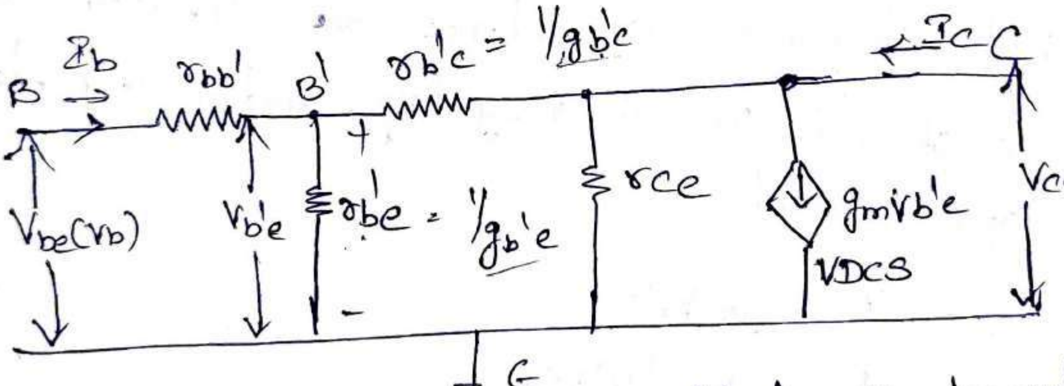


fig (a):- Hybrid π model for CE transistor at low frequency

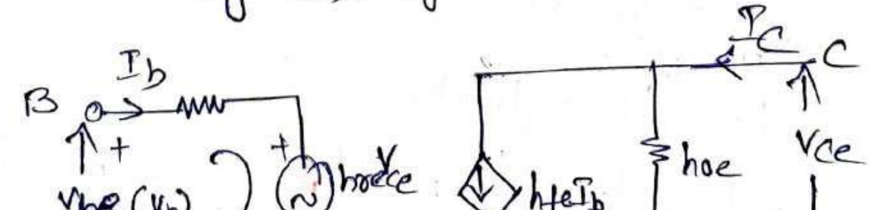


fig (b) h-parameter model for a common emitter transistor at low frequency.

By applying KVL to i/p & KCL to o/p

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

$I_c = h_{fe}I_b + h_{oe}V_c$

(1) short circuit input impedance

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

(2) open ckt reverse vtg gain $h_{re} = \frac{V_b}{V_c}\Big|_{I_b=0}$

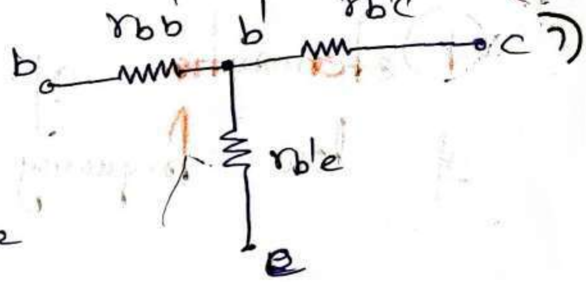
(3) short circuit forward current gain h_{fe}

$h_{fe} = \frac{I_c}{I_b}\Big|_{V_c=0}$

(4) open circuit output conductance ⇒ $h_{oe} = \frac{I_c}{V_c}\Big|_{I_b=0}$ (℧)

A · Hybrid π - Conductances

(1) Base spread resistance ($r_{bb'}$) :-



The value of input resistance ~~is~~ R_i is equal to h_{ie}, i.e when $V_{ce}=0$ i.e when o/p terminals are short circuited

$$Z_i \Big|_{V_{ce}=0} = r_{bb'} + r_{b'e} \| r_{b'c}$$

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

As $r_{b'c} \gg r_{b'e}$, the above equation can be written as

$$\boxed{h_{ie} = r_{bb'} + r_{b'e}} \quad (or) \quad \boxed{r_{bb'} = h_{ie} - r_{b'e}}$$

$$\Rightarrow g_{bb'} = \frac{1}{r_{bb'}} = \frac{1}{h_{ie} - r_{b'e}}$$

(2) Resistance between virtual base (b') & emitter (e) (or) $r_{b'e}$:-

(or)

~~consider short circuit~~

Conductance between virtual base (b') & emitter (e) (or) ($g_{b'e}$)

At high frequencies capacitance are replaced with short circuit and r_{ce} is very large and it is replaced with open circuit

$$I_c = g_m V_{b'e}$$

$$I_c = g_m I_b r_{b'e}$$

$$\frac{I_c}{I_b} = g_m r_{b'e}$$

$$\beta \text{ (or) } h_{fe} = g_m r_{b'e}$$

$$\boxed{r_{b'e} = \frac{h_{fe}}{g_m}}$$ resistance b/w b' & e

$$g_{b'e} = \frac{1}{r_{b'e}} \Rightarrow g_{b'e} = \frac{1}{\frac{h_{fe}}{g_m}} \Rightarrow \boxed{g_{b'e} = \frac{g_m}{h_{fe}}}$$

conductance b/w $b'e$

1) Transconductance (g_m) → repeated

The transconductance is nothing but the ratio of change in collector current due to small changes in the voltage V_{be} across emitter junction.

$$I_C = I_{CO}\left[e^{V_{b'e}/\eta V_T} - 1\right]$$

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

$\eta = 1$ for high freq

V_{be} = base emitter voltage

I_{CO} = Reverse saturation current

Thermal voltage $V_T = \frac{kT}{q} = \frac{T}{11,600} = 26\,mV$

$$I_C = I_{CO}\left[e^{V_{b'e}/V_T} - 1\right]$$

$$I_C = I_{CO}\, e^{V_{b'e}/V_T} - I_{CO}$$

$$I_C + I_{CO} = I_{CO}\, e^{V_{b'e}/V_T}$$

$I_{CO} << I_C$

$$I_C = I_{CO}\, e^{V_{b'e}/V_T}$$

diff w.r.to $V_{b'e}$

$$\frac{\partial I_C}{\partial V_{b'e}} = I_{CO}\, e^{V_{b'e}/V_T} \cdot \frac{1}{V_T}$$

$\therefore \frac{d}{dt} e^{at} = e^{at} \cdot \frac{d\,at}{dt}$

$a \cdot e^{at}$

$$g_m = \frac{I_C + I_{CO}}{V_T} = \frac{I_C}{V_T}$$

$$\therefore g_m = \frac{I_C}{V_T}$$

We know that dynamic emitter resistance

$$r_e = \frac{V_T}{I_e} = \frac{V_T}{I_C}$$

$$\boxed{g_m = \frac{1}{r_e}} \Rightarrow \boxed{g_m = \frac{I_C}{V_T}}$$

(Or)

(3) Trans conductance (g_m) = The transconductance is nothing but the ratio of change in I_c due to small changes in the voltage V_{be} across emitter junction.

$$g_m = \frac{o/p \text{ current}}{i/p \text{ } \cancel{\text{current}} \text{ voltage}} \Big|_{V_{out}=0}$$

in CE ~~C~~onfiguration $\rightarrow g_m = \frac{I_c}{V_{be}} \Big|_{V_{ce}=0}$

$$g_m = \frac{I_c}{V_T} = \frac{I_c}{26mV}$$

$$g_m = \frac{I_c}{26mV}$$

$$\boxed{g_m = \frac{1}{r_e}}$$

$\because r_e = \frac{26mV}{I_e}$

$I_e \simeq I_c$

$r_e = \frac{26mV}{I_c}$

(4) Conductance b/w b' & c (or) feed back conductance:

When i/p terminals are open circuited, then the reverse voltage gain hre

$$h_{re} = \frac{i/p\ vtg}{o/p\ vtg}$$

$$h_{re} = \frac{V_{b'e}}{V_{ce}} = \frac{\cancel{V_{ce}} \times \frac{r_{b'e}}{r_{b'e}+r_{b'c}}}{\cancel{V_{ce}}}$$ → Voltage division rule

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

$$h_{re}[r_{b'e}+r_{b'c}] = r_{b'e}$$

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

re arranging the above equation we get

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

$$h_{re}\, r_{b'c} = r_{b'e}[1-h_{re}]$$

The value of hre is in the range of 10^{-4}, the above equation can be approximated by

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

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

$$g_{b'c} = \frac{1}{r_{b'c}} \Rightarrow \frac{1}{\frac{r_{b'e}}{h_{re}}} = \frac{h_{re}}{r_{b'e}}$$

$$g_{b'c} = \frac{h_{re}}{r_{b'e}} \qquad \because 1/r_{b'e} = g_{b'e}$$

$$\boxed{g_{b'c} = h_{re}\, g_{b'e}}$$

(9) (5)

Conductance between terminals C & E (g_{ce}) :-

From fig a we can say $V_{b'e} = h_{re} V_{ce}$

For determining g_{ce}, input side $I_b = 0$ and I_C → by applying KCL

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

We know $h_{oe} = \left.\frac{I_c}{V_{ce}}\right|_{I_b=0} = \dfrac{\frac{V_{ce}}{r_{ce}} + \frac{V_{ce}}{r_{b'e}+r_{b'c}} + g_m V_{b'e}}{V_{ce}}$

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

We know $V_{b'e} = \dfrac{V_{ce} \cdot r_{b'e}}{r_{b'c} + r_{b'e}}$

$$h_{oe} = g_{ce} + \frac{1}{r_{b'e}+r_{b'c}} + \frac{g_m}{\cancel{V_{ce}}} \frac{\cancel{V_{ce}} \cdot r_{b'e}}{r_{b'c}+r_{b'e}}$$

$$= g_{ce} + \frac{g_m r_{b'e}}{r_{b'c}+r_{b'e}} + \frac{1}{r_{b'e}+r_{b'c}} \quad \text{—(a)}$$

Here compare circuit a & b

$I_C = g_m V_{b'e}$

$= g_m I_b \cdot r_{b'e}$

$\frac{I_c}{I_b} = g_m r_{b'e}$

$\therefore I_c | I_b = h_{fe}$

$h_{fe} = g_m r_{b'e}$ → sub h_{fe} value in eq (a)

$h_{oe} = \dfrac{g_{ce} + [h_{fe}+1]}{\cancel{r_{b'c} + r_{b'e}}}$

$h_{oe} = g_{ce} + \dfrac{h_{fe}+1}{r_{b'c}+r_{b'e}}$

$r_{b'c} \gg r_{b'e}$

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

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

$$\boxed{g_{ce} = h_{oe} - g_{b'c} h_{fe}}$$

$r_{b'c}$ is the o/p resistance
$r_{b'e}$ is the i/p resistance
o/p resistance (or) reverse resistance
is >> than the i/p resistance

Determination of Hybrid π Capacitance parameters.

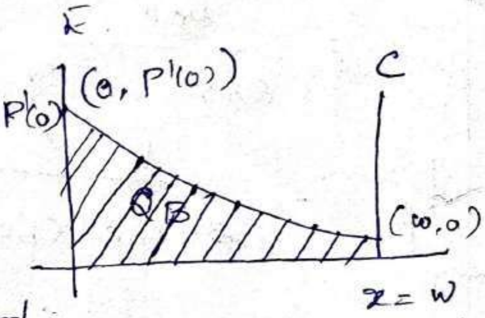
$C_{b'e}$ or C_D or

→ The capacitance parameters of hybrid π model are $C_{b'e}$ and $C_{b'c}$.

→ The value of $C_{b'c}$ is provided by the manufacturer

$C_{b'e}$ or C_D or C_e

→ The charge in the base region is given by



$$Q_B = \frac{1}{2} \times P'(0) A q w \rightarrow (1)$$

$P'(0)$ = minority charge carrier concentration

$W \rightarrow$ width of base region

$Q_B \rightarrow$ charge in base region

Diffusion current in the base region is expressed as

$$I = -Aq D_B \frac{dP'(0)}{dx}$$

$D_B \rightarrow$ Diffusion constant

$$I = -Aq D_B \times \frac{(0 - P'(0))}{W - 0}$$

$$I = \frac{Aq D_B \times P'(0)}{W} - (2)$$

from equ (1)

$$P'(0) Aq = \frac{2 Q_B}{W} \rightarrow (3)$$

Sub eq (3) in eq (2)

$$I = \frac{2 Q_B \times D_B}{W^2}$$

$$Q_B = \frac{I W^2}{2 D_B}$$

Diff w.r.to 'V'

$$\frac{dQ_B}{dV} = \left(\frac{W^2}{2D_B}\right)\frac{dI}{dV}$$

$Q = CV$. $C = \frac{Q}{V}$

$$\boxed{C_D = C_{b'e} = g_m \frac{W^2}{2D_B}}$$

3B Hybrid π Capacitances

In the hybrid π model, there are two capacitances namely (1) collector - junction capacitance or C_c or C_T [Transition Capacitance]
(2) Emitter - junction capacitance or C_e or C_D [Diffusion Capacitance]

1) Transition Capacitance (C_T) or C_c or Collector Junction Capacitance

It is measured between reverse bias collector base junction in CB configuration & is generally given by manufacturer has $\boxed{C_{ob} \text{ or } C_C \text{ or } C_T}$ and it varies as $[V_{CB}]^{-n}$, where n is 1/2 for abrupt junction and 1/3 for a graded junction

2) Diffusion Capacitance or C_D or Emitter junction Capacitance (C_e)

It is measured between forward bias emitter base junction in CB configuration and is given by C_e or C_d

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

where f_T is the frequency at which short circuit current gain falls to unity [when $h_{fe} = 1$]

Conclusion:

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

(2) $r_{b'e} = \frac{h_{fe}}{g_m}$ or $g_{b'e} = \frac{g_m}{h_{fe}}$

(3) $g_m = \frac{1}{r_e}$ or $\frac{I_C}{V_T}$

(4) $g_{b'c} = h_{re}\, g_{b'e}$

(5) $g_{ce} = h_{oe} - g_{b'c} h_{fe}$

(6) C_c or $C_T = C_{ob}$

(7) C_e or $C_D = g_m / 2\pi f_T$

Typical Values of Hybrid-π Parameters

Typical values are estimated at $I_C = 1.3mA$ at room temperature

$r_{bb'} = 100\Omega$

$r_{b'e} = 1K\Omega$

$r_{b'c} = 4M\Omega$

$r_{ce} = 80K\Omega$

$g_m = 50mA/V$

$C_{b'e} = 100PF$

$C_{b'c} = 3PF$

Proofs for Typical values :-

$\because I_C = 1.3mA$

$V_T = \frac{KT}{q} = 26mV$

$h_{ie} = 1.1K\Omega$

$h_{fe} = 50$

$h_{re} = 2.5\times10^{-4}$

$h_{oe} = 25\mu A/V$

(1) $g_m = \frac{I_C}{V_T}$

$g_m = \frac{1.3\times10^{-3}}{26\times10^{-3}} = \frac{1}{20} = 0.05$

$g_m = 50\,mA/V$

(2) $r_{b'e} = \frac{h_{fe}}{g_m} = \frac{50}{50\times10^{-3}} = 1K\Omega$

$\frac{I}{I/V} = R \quad I\times\frac{V}{I} =$

(3) $r_{bb'} = h_{ie} - r_{b'e} = 1100 - 1000 \Rightarrow r_{bb'} = 100\Omega$

(4) $r_{b'c} = \frac{1}{g_{b'c}} = \frac{r_{b'e}}{h_{re}} = \frac{1\times10^{3}}{2.5\times10^{-4}} = 0.4\times10^{7} = 4\times10^{6} \Rightarrow r_{b'c} = 4M\Omega$

(5) $r_{ce} = \frac{1}{g_{ce}} = \frac{1}{h_{oe} - h_{fe}\,g_{b'c}} = \frac{1}{25\times10^{-6} - 50\times0.25\times10^{-6}}$

$\because g_{b'c} = \frac{1}{r_{b'c}}$

$g_{b'c} = \frac{1}{4\times10^{6}}$

$g_{b'c} = 0.25\times10^{-6}$

$r_{ce} = \frac{1}{12.5\times10^{-6}}$

$r_{ce} = 0.08\times10^{6}$

$r_{ce} = 80K\Omega$

(6) C_C (or) $C_T = C_{ob}$

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

Problems

1) The following low frequency parameters for a given transistor $I_C = 5mA$, $V_{ce} = 10V$ at room temperature $h_{ie} = 600\Omega$ $h_{fe} = 100$, $h_{re} = 10^{-4}$, $h_{oe} = 20\mu℧$ at the same operating point $f_T = 500MHz$, $C_{ob} = 3pf$. Calculate the values of hybrid parameters

Given:
collector current $(I_c) = 5mA$
voltage b/w collector & emitter $(V_{ce}) = 10V$
short circuit input impedance $(h_{ie}) = 600\Omega$
short circuit forward current gain $(h_{fe}) = 100\Omega$
open circuit reverse voltage gain $(h_{re}) = 10^{-4}$
Open circuit output conductance $(h_{oe}) = 20\mu℧$
$f_T = 500MHz$
Transition Capacitance $C_{ob} = 3pf$

1) $g_m = \frac{I_C}{V_T} = \frac{5\times10^{-3}}{26mV} = 0.192℧ = 192m℧$

2) $r_{b'e} = \frac{h_{fe}}{g_m} = \frac{100}{0.192} = 520.83\Omega$

3) $r_{b'c} = \frac{r_{b'e}}{h_{re}} = \frac{520.83}{10^{-4}} = 520.83\times10^{4}\Omega$

4) $r_{bb'} = h_{ie} - r_{b'e} = 600 - 520.83 = 79.19\Omega$

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

$g_{b'c} = \frac{1}{r_{b'c}} = \frac{1}{520.83\times10^{4}} = 1.92\times10^{-7}$

$ho\ g_{ce} = 20\times10^{-6} - (1+100)(1.92\times10^{-7})$

$g_{ce} = 20\times10^{-6} - (101)(1.92\times10^{-7})$

$g_{ce} = (200 - 193.92)\times10^{-7}$

$g_{ce} = 6.08\times10^{-7}$

$$r_{ce} = \frac{1}{g_{ce}} = \frac{1}{6.08\times10^{-7}} = 1.644\times10^{6} = 1.644\ M\Omega$$

6) $C_T = C_{ob} = 3pf$

7) $$C_e = \frac{g_m}{2\pi f_T} = \frac{0.192}{2\times3.14\times500\times10^{6}} = \frac{0.192}{2\times3.14\times500\times10^{6}}$$

$$= 6.114\times10^{-11} = 61.14\ PF$$

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

$I_C = 5mA$, $h_{ie} = 1k\Omega$, $h_{re} = 10^{-4}$, $f_T = 10\ MHz$

$V_{ce} = 8V$, $h_{fe} = 100$, $h_{oe} = 4\times10^{-5}$, $C_{ob} = 2pf$

1) $$g_m = \frac{I_C}{V_T} = \frac{5\times10^{-3}}{26\times10^{-3}} = 0.192\ \mho$$

2) $$r_{b'e} = \frac{h_{fe}}{g_m} = \frac{100}{0.192} = 520.8\ \Omega$$

3) Resistance b/w b' & C $r_{b'c}$

$$r_{b'c} = \frac{r_{b'e}}{h_{re}} = \frac{521}{10^{-4}} = 521\times10^{4} = 5.2M\Omega$$

4) $r_{bb'} = h_{ie} - r_{b'e} = 1000 - 521 = 479\ \Omega$

5) $$r_{ce} = \frac{1}{g_{ce}} = \frac{1}{h_{oe} - [1+h_{fe}]g_{b'c}} \Rightarrow g_{b'c} = \frac{1}{r_{b'c}} = \frac{1}{521\times10^{4}} = 1.92\times10^{-7}$$

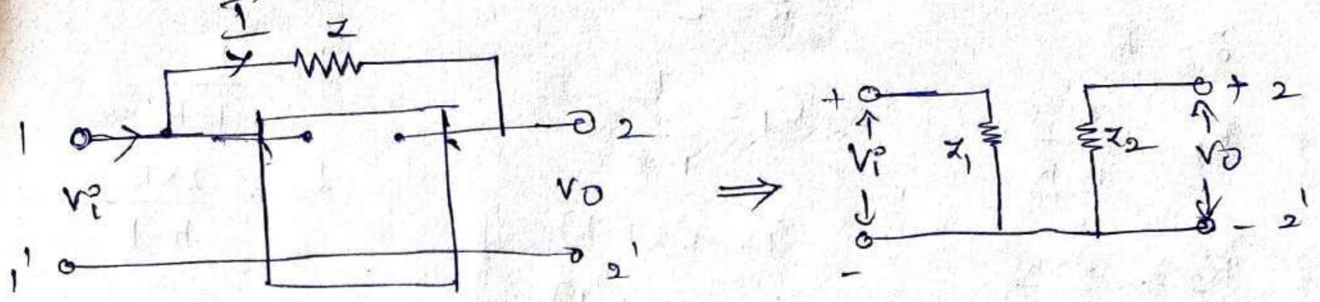
$g_{ce} = (4\times10^{-5}) - (1+100)(1.92\times10^{-7}) = 2.0608\times10^{-5}$

$$r_{ce} = \frac{1}{g_{ce}} = \frac{1}{2.0608} = 0.485\ M\Omega$$

6) $C_T = C_{ob} = 2Pf$

7) $$C_D = C_e = g_m/2\pi f_T = \frac{0.192}{2\times3.14\times10\times10^{6}} = 0.00305\ Pf$$

Miller's Theorem:



→ It is used to convert one form of configuration into another configuration

→ Miller theorem establishes that in a linear circuit, if there exists a branch with impedance Z, connecting two nodes with nodal voltages V_1 and V_2,

→ It asserts that a floating impedance element, supplied by two voltage sources connected in series, may be split in to two grounded elements with corresponding impedances.

$$I = \frac{V_i - V_0}{Z}$$

$$I = \frac{V_i\left[1 - \frac{V_0}{V_i}\right]}{Z}$$

$$I = \frac{V_i[1 - A_V]}{Z}$$

$$Z_1 = \frac{V_i}{I} \qquad A_V = K$$

$$\frac{V_i}{I} = \frac{Z}{1 - A_V}$$

$$\boxed{Z_1 = \frac{Z}{1 - A_V}}$$

$$I = \frac{V_0 - V_i}{Z}$$

$$I = \frac{V_0\left[1 - \frac{V_i}{V_0}\right]}{Z}$$

$$I = \frac{V_0\left[1 - \frac{1}{A_V}\right]}{Z}$$

$$Z_2 = \frac{V_0}{I} \qquad A_V = K$$

$$Z_2 = \frac{V_0}{I} = \frac{Z}{1 - \frac{1}{A_V}} = \frac{Z A_V}{A_V - 1}$$

$$Z_2 = \frac{Z \times K}{(K-1)}$$

Case (i) :

$Z = R$

$R_1 = \frac{R}{1-K}$

$R_2 = \frac{R \times K}{K-1}$

Case (ii)

$Z = C$

$C_1 = \frac{1/C}{1-K} =$

$C_1 = C(1-K)$

$C_2 = \frac{C(K-1)}{K}$

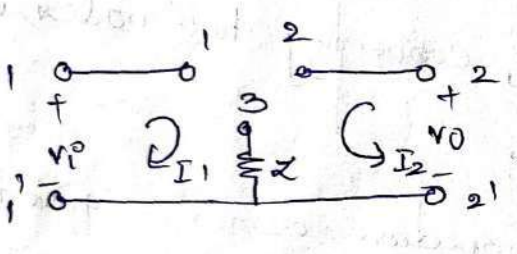
Case (iii)

$Z = L$

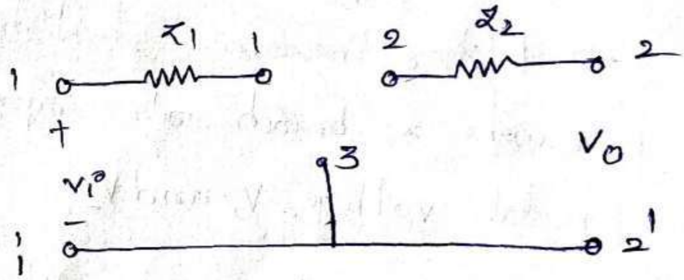
$L_1 = \frac{L}{1-K}$

$L_2 = \frac{L \times K}{K-1}$

Dual Miller Theorem :



$\Longrightarrow$



$V_i = Z[I_1 + I_2]$

$V_i = Z I_1 \left[1 + \frac{I_2}{I_1}\right]$

$\frac{V_i}{I_1} = Z\left[1 + \frac{I_2}{I_1}\right]$

$$\boxed{Z_1 = Z[1 - A_I]}$$

$\therefore \frac{-I_2}{I_1} = A_I$

$V_0 = Z[I_1 + I_2]$

$V_0 = Z I_2 \left[\frac{I_1}{I_2} + 1\right]$

$\frac{V_0}{I_2} = Z\left[\frac{-1}{A_I} + 1\right]$

$$\boxed{Z_2 = Z\left[\frac{A_I - 1}{A_I}\right]}$$

Single stage CE Amplifier with short circuit load. (18)

CE short circuit current gain

→ Consider a single stage CE transistor amplifier with load resistor R_L

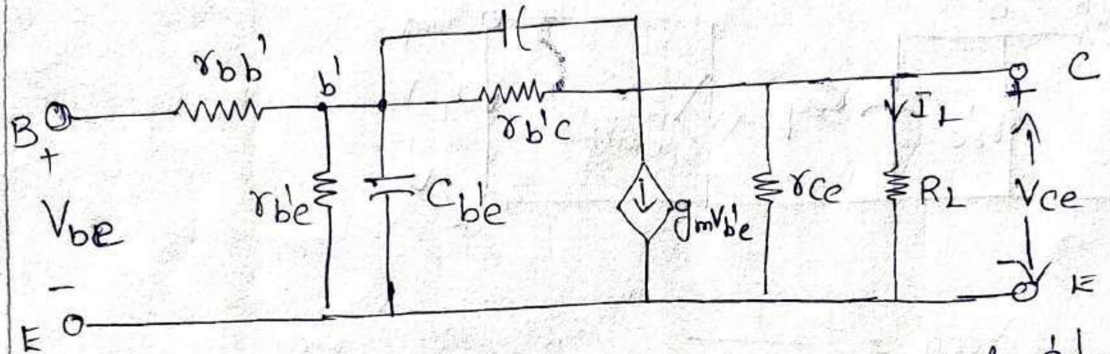
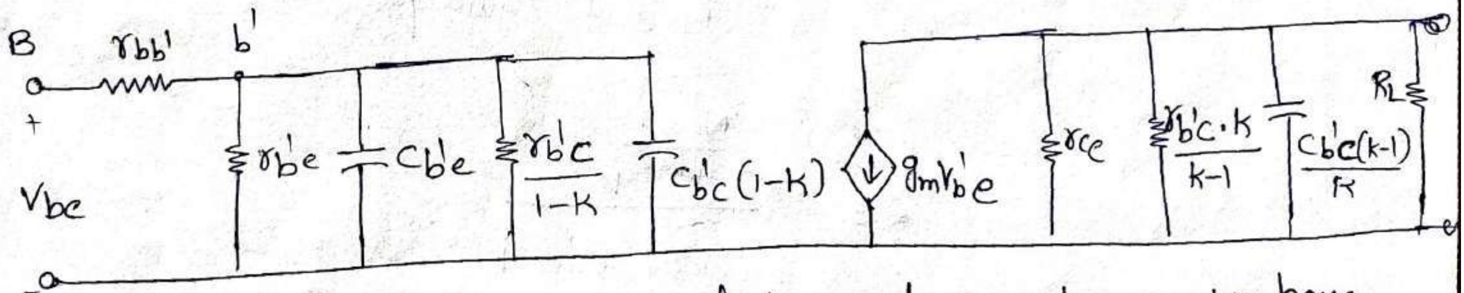


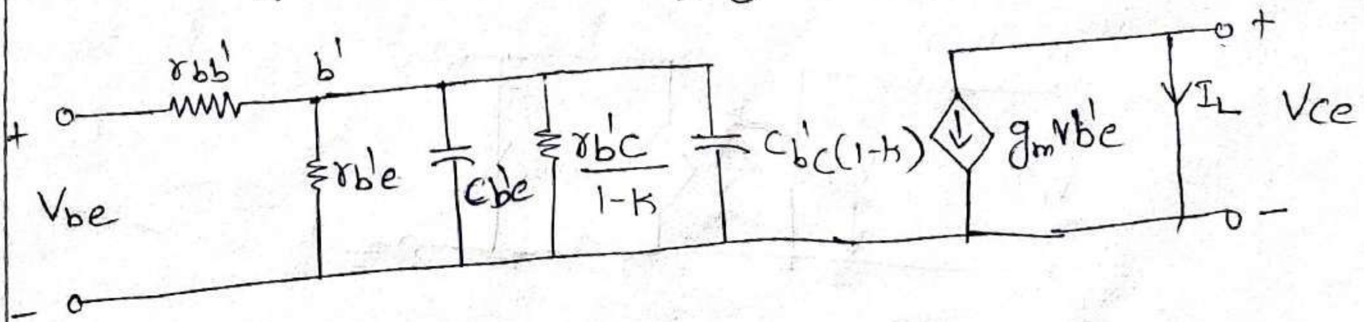
fig:. The hybrid-π circuit for a single stage CE transistor with a resistive load R_L

step 1:- According to miller theorem convert $C_{b'c}$ & $r_{b'c}$ in to two sub impedances



step-2: For the analysis of short circuit current gain we have to assume $R_L = 0$, it becomes short then

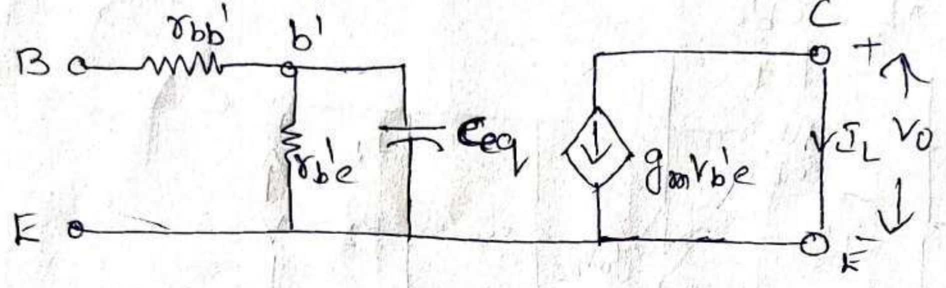
$$R_L \parallel \frac{C_{b'c}(K-1)}{K} \parallel \frac{r_{b'c}\cdot K}{K-1} \parallel r_{ce} = 0 \parallel \frac{C_{b'c}(K-1)}{K} \parallel \frac{r_{b'c}\cdot K}{K-1} \parallel r_{ce}$$
$$= 0$$



Step 3 Here $r_{b'e} \parallel \frac{r_{b'c}}{1-K} \simeq r_{b'e}$

Let $C_{b'e} = C_e$ and $C_{b'c}(1-K) = C_C$

$C_{eq} = C_e + C_C$



Step 4

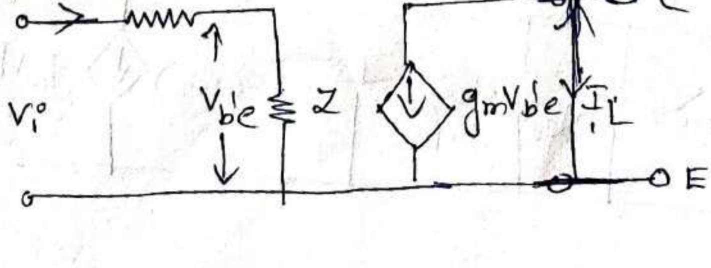
$$X_{ceq} = \frac{1}{j\omega C_{eq}}$$

impedance $Z = r_{b'e} \parallel C_{eq} = r_{b'e} \parallel X_{ceq}$

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

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

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



Step 5

$$A_I = \frac{I_L}{I_b} \rightarrow (1)$$

$$I_L = -g_m V_{b'e} \rightarrow (2)$$

We know that $V_{b'e} = Z \cdot I_b$

$$Z = \frac{V_{b'e}}{I_b} \rightarrow (3)$$

Sub 2 in (1)

$$A_I = \frac{-g_m V_{b'e}}{I_b},$$

$$A_I = -g_m \cdot Z \qquad \therefore Z = \frac{V_{b'e}}{I_b}$$

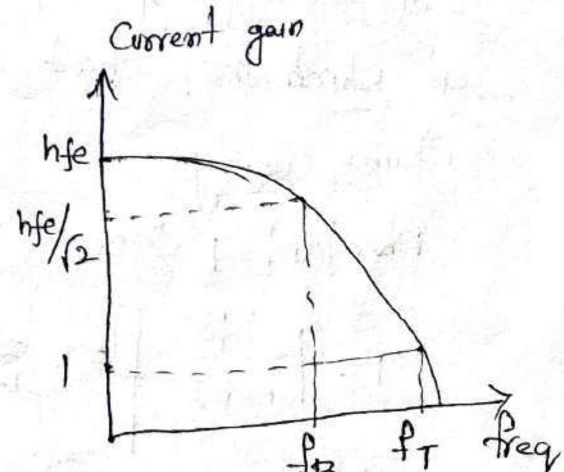
$$A_I = -g_m \times \frac{r_{b'e}}{1 + j\omega r_{b'e} C_{eq}}$$

We know that $r_{b'e} = \frac{h_{fe}}{g_m} \Rightarrow h_{fe} = g_m r_{b'e}$

$$\boxed{A_I = \frac{-h_{fe}}{1 + j\omega r_{b'e} C_{eq}}}$$

(or)

$$\boxed{\therefore A_I = \frac{-h_{fe}}{1 + j\omega r_{b'e} [C_e + C_c]}}$$



From this we can say current I is not constant, it depends on frequency. When f is small ie at low frequency $A_i = -h_{fe}$,

→ If frequency increases, the current gain A_i decreases.

Bandwidth (or) f_β :-

Where the frequency at which the CE short circuit current gain falls by 3dB is given by from its final gain is given by

Let $f_\beta = \frac{1}{2\pi r_{b'e}(C_e+C_c)} = \frac{g_{b'e}}{2\pi(C_e+C_c)}$ $\because g_{b'e} = \frac{1}{r_{b'e}}$

where $A_i = \frac{-h_{fe}}{1+j2\pi f \cdot r_{b'e}(C_e+C_c)}$

$\frac{-h_{fe}}{1+f}$

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

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

If $f = f_\beta$

The frequency range up to f_β is referred to as the bandwidth of the circuit.

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

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

Cutoff frequency f_T :- It is defined as the frequency at which the short-circuit common-emitter current gain attains unity.

when $|A_I| = 1$ then $f = f_T$

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

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

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

Here $f_T/f_\beta \gg 1$

$$1 = \frac{h_{fe}}{\sqrt{(f_T/f_\beta)^2}} \Rightarrow h_{fe} = f_T/f_\beta$$

$$f_T = f_\beta h_{fe} = \frac{h_{fe}}{2\pi r_{b'e} C_{eq}} \quad \because r_{b'e} = \frac{h_{fe}}{g_m}$$

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

$$f_T = \frac{g_m}{2\pi C_{eq}}$$

$$f_T = \frac{g_m}{2\pi (C_e + C_c)} \quad \because C_e \gg C_c$$

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

NOTE !

$f_T = f_\beta h_{fe}$

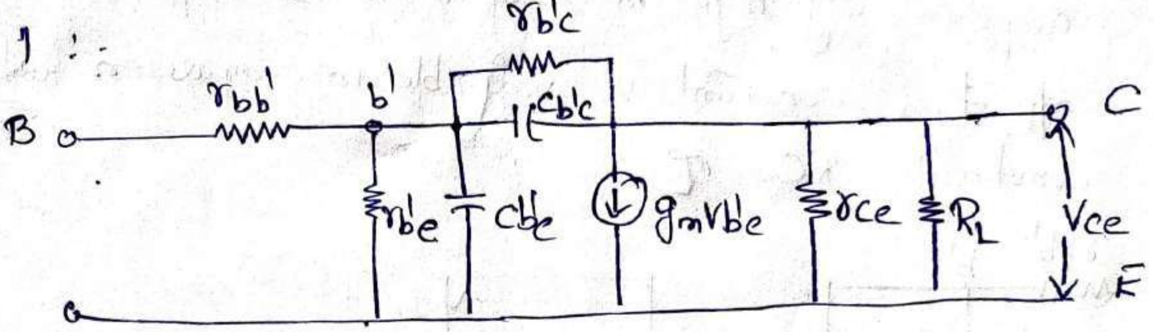
$h_{fe} \rightarrow$ current gain

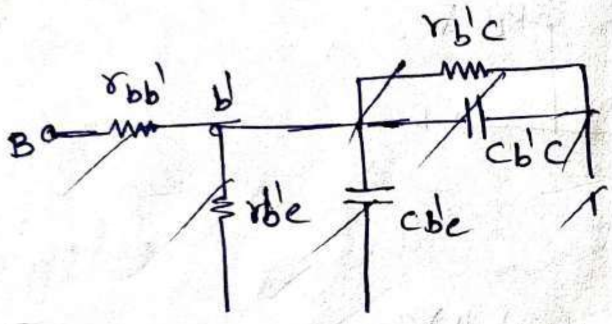
$f_\beta \rightarrow$ Bandwidth.

Since $f_T = h_{fe} f_\beta$, this parameter may be given as short circuit current gain bandwidth product.

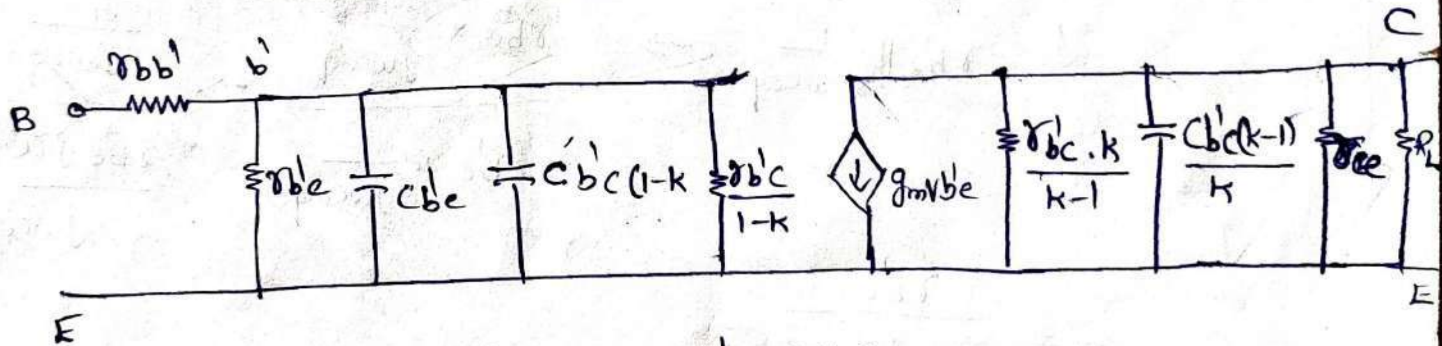
Current gain of CE with resistive load i.e $R_L \neq 0$

Step 1 :-





Step 2 :- Using Miller theorem



Step 3 :- W.K.T $r_{b'e} = 1k\Omega$, $r_{b'c} = 4M\Omega$

$r_{b'e} \parallel \frac{r_{b'c}}{1-k} \simeq r_{b'e}$

$V_0 = I_L R_L$

$V_0 = -g_m V_{b'e} R_L$

$\frac{V_0}{V_{b'e}} = -g_m R_L$

$\therefore \frac{V_0}{V_{b'e}} = A_v = -g_m R_L$

$\therefore A_v = K = -g_m R_L$

Let $C_{b'e} = C_e$

$C_{b'c}(1-K) = C_c$

$C_{b'c}[1 + g_m R_L) = C_c$

$C_{eq} = C_e + C_c$

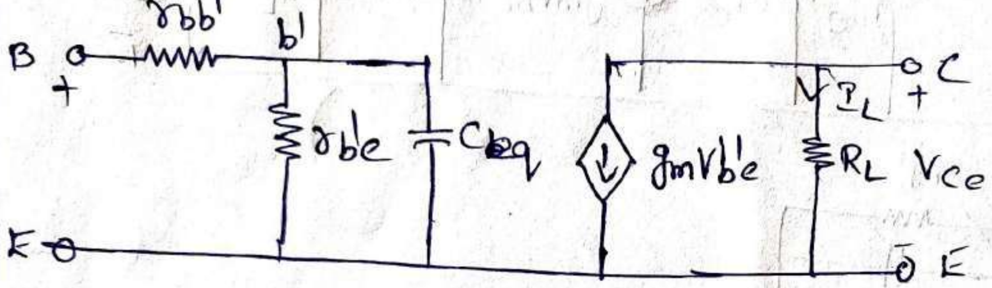
$C_{eq} = C_e + C_{b'c}[1 + g_m R_L]$

We know $r_{ce} \gg R_L$, $r_{ce} \| R_L = R_L$

$$r_{b'c}\left[\frac{K}{K-1}\right] \| R_L \simeq R_L$$

Here C_{eq} is very high in comparison to o/p $C_C \simeq C_{b'c}\frac{K-1}{K}$

Hence o/p time constant is negligible in comparison with i/p time constant $RC = \tau$

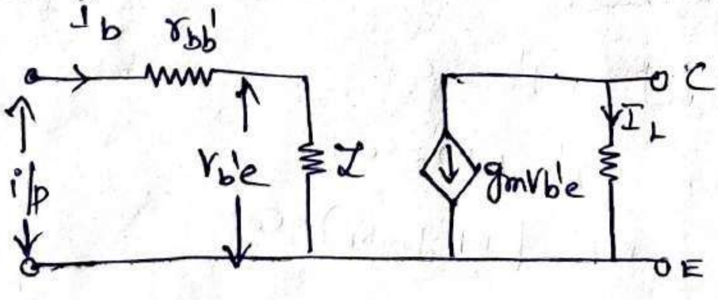


<u>Step 4</u>

$$X_{eq} = \frac{1}{j\omega C_{eq}} \qquad Z = r_{b'e} \| X_{C_{eq}}$$

$$Z = r_{b'e} \| \frac{1}{j\omega C_{eq}} \Rightarrow \frac{r_{b'e} \times \frac{1}{j\omega C_{eq}}}{r_{b'e} + \frac{1}{j\omega C_{eq}}} \Rightarrow \frac{\frac{r_{b'e}}{\cancel{j\omega C_{eq}}}}{\frac{r_{b'e}\, j\omega C_{eq} + 1}{\cancel{j\omega C_{eq}}}}$$

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



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

<u>Step 5</u> Current gain (A_i)

$$A_i = \frac{\text{o/p current}}{\text{i/p current}} = \frac{I_c}{I_b} = \frac{-I_L}{I_b} = \frac{-g_m V_{b'e}}{I_b} = -g_m Z$$

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

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

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

$\omega = 2\pi f$

$C_{eq} = \left[C_e + C_c\left(1+g_m R_L\right)\right]$

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

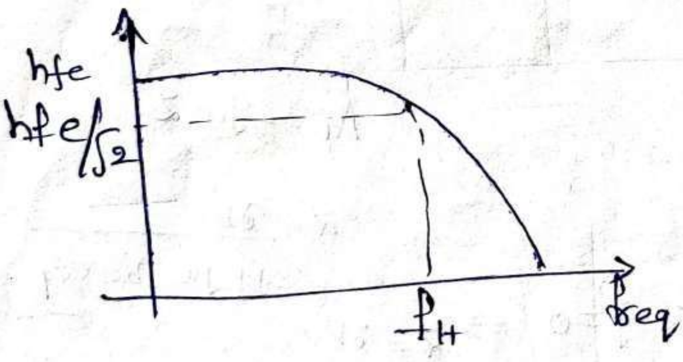
$$A_i = \frac{-h_{fe}}{1 + j\, f/f_H}$$

at $f = f_H$

$$A_i = \frac{-h_{fe}}{1 + j\, f_H/f_H} = \frac{-h_{fe}}{1+j}$$

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

Frequency Response



The f_H is the frequency at which transistor gain drops by 3 dB or $1/\sqrt{2}$ times from its final value.

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

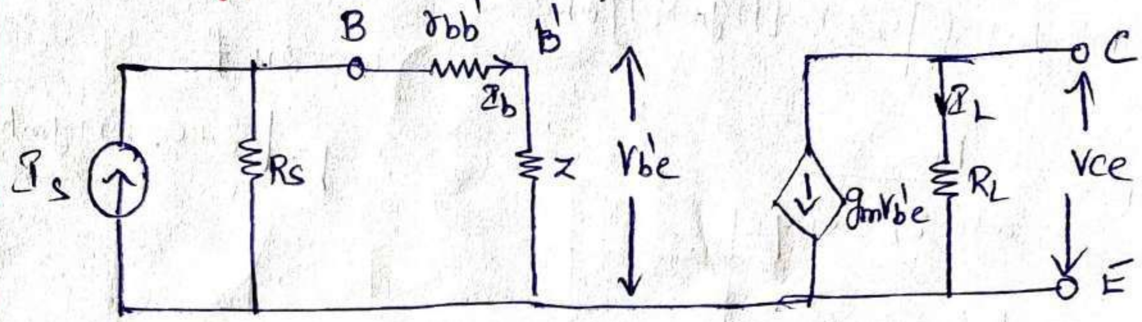
$$f_H = \frac{1}{2\pi r_{b'e}\left[C_e + C_c\left[1+g_m R_L\right]\right]}$$

$R_L = 0$

$$f_H = \frac{1}{2\pi r_{b'e}\left[C_e + C_c\right]} = f_\beta$$

From the above eqn we can say that the maximum possible value for f_H is f_β. As R_L ↑ then C_{eq} ↑ increases & ↓ decreases in f_H

Current gain including source resistance (R_s)



$A_i \rightarrow$ Current gain without source

$A_{is} \rightarrow$ Current gain including source

$$A_{is} = \frac{I_L}{I_s} = \frac{I_L}{I_b} \times \frac{I_b}{I_s} \Rightarrow A_i \cdot \frac{I_b}{I_s}$$

$$\frac{I_b}{I_s} = ?$$

$$I_b = \frac{R_s\, I_s}{R_s + r_{bb'} + z} \Rightarrow \frac{I_b}{I_s} = \frac{R_s}{R_s + r_{bb'} + z}$$

$$\therefore A_{is} = \frac{A_i R_s}{R_s + r_{bb'} + z}$$

$$= -\frac{g_m z\, R_s}{R_s + r_{bb'} + z},$$

$\therefore A_v = -g_m z$

$\therefore z = \dfrac{r_{b'e}}{1 + j\omega r_{b'e} C_{eq}}$

At low frequencies $f = 0 \Rightarrow z = r_{b'e}$

$$\therefore A_{is} \text{ at low freq} = \frac{-g_m r_{b'e} R_s}{R_s + r_{bb'} + r_{b'e}}$$

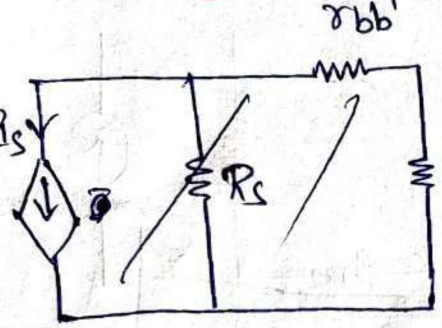
$$= \frac{-h_{fe} R_s}{R_s + r_{bb'} + r_{b'e}}$$

$$\boxed{A_{is}\left(\text{at low freq}\right) = \frac{-h_{fe} R_s}{R_s + h_{ie}}}$$

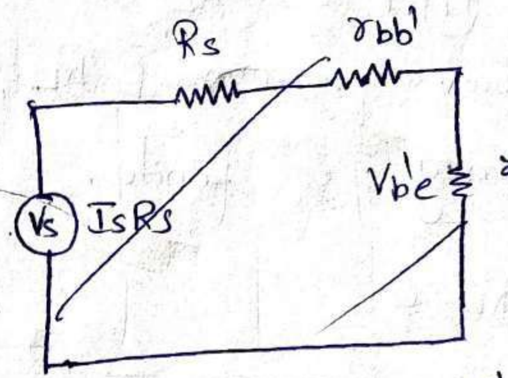
$\therefore h_{ie} = r_{bb'} + r_{b'e}$

A_{is} is independent of R_L

Voltage gain including Rs



$\simeq$



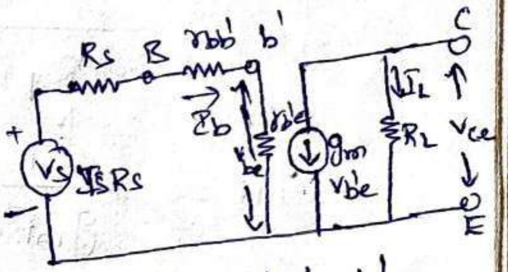


fig: equivalent ckt assuming V_g source

$A_v \rightarrow$ voltage gain without source

$A_{vs} \rightarrow$ Voltage gain including source

$$A_{vs} = \frac{V_o}{V_s} = \frac{I_L R_L}{I_s R_s} \qquad A_{is} \times \frac{R_L}{R_s}$$

$$\frac{V_{ce}}{V_{b'e}} \times \frac{V_{b'e}}{V_s}$$

$$A_{vs} = \frac{-h_{fe} \cancel{R_s}}{R_s + h_{ie}} \times \frac{R_L}{\cancel{R_s}}$$ ~~at~~ low freq

$$\boxed{A_{vs}(\text{low}) = \frac{-h_{fe} R_L}{R_s + h_{ie}}}$$

$A_{vs}(\text{low})$ increases linearly with R_L

Cutoff frequency including ~~Rs~~ Source resistance (R_s)

$$A_{vs}(\text{high}) = \frac{A_{vs}}{1 + j\, f/f_H}$$

$$A_{is}(\text{high}) = \frac{A_{is}}{1 + j(f/f_H)}$$

$f_H \rightarrow$ cutoff frequency

$$f_H \rightarrow \frac{1}{2\pi R_{eq} C_{eq}}$$

$$R_{eq} = r_{b'e} \parallel (r_{bb'} + R_s)$$

$$C_{eq} = C_{b'e} + C_{b'c}(1 + g_m R_L)$$

Here $f_H \uparrow$ as R_L is decreased because C is a linear function of R_L, At $R_L = 0$, the 3dB frequency is finite.

for $R_L = 0$

$$f_H = \frac{1}{2\pi R_{eq}(C_e + C_c)} \qquad \left[f_T = \frac{g_m}{2\pi(C_e + C_c)} \right]$$

$$f_H = \frac{f_T}{g_m R_{eq}} \qquad \therefore f_T = h_{fe} f_\beta$$

$$f_H = \frac{h_{fe} f_\beta}{g_m R_{eq}} \qquad g_{b'e} = g_m / h_{fe}$$

$$f_H = \frac{f_\beta}{g_{b'e} R_{eq}}$$

Gain Bandwidth Product for voltage

(1) Gain Bandwidth product for voltage

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

$$= \frac{-h_{fe} R_L}{R_s + h_{ie}} \times \frac{1}{2\pi\, r_{b'e} \| (r_{bb'} + R_s)\, C_{eq}}$$

$$\rightarrow r_{b'e} \| (r_{bb'} + R_s) = \frac{(r_{bb'} + R_s) r_{b'e}}{R_s + \underbrace{r_{bb'} + r_{b'e}}_{h_{ie}}} \Rightarrow \frac{(r_{bb'} + R_s) r_{b'e}}{R_s + h_{ie}}$$

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

Avslow

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

$$= \frac{-g_m \cancel{r_{b'e}} R_L}{2\pi C_{eq} \cancel{r_{b'e}} [r_{bb'} + R_s]}$$

$$= \frac{-g_m R_L}{2\pi C_{eq} [r_{bb'} + R_s]}$$

This equation can be further simplified as follows

$$= \frac{-R_L}{r_{bb'} + R_s} \times \frac{g_m}{2\pi [C_e + C_c (1 + g_m R_L))} \qquad g_m R_L >> 1$$

$$= \frac{-R_L}{r_{bb'} + R_s} \times \frac{g_m}{2\pi (C_e + C_c g_m R_L)} \qquad \therefore f_T = \frac{g_m}{2\pi C_e}$$

$$g_m = 2\pi f_T C_e$$

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

$$= \frac{-R_L}{r_{bb'} + R_s} \times$$

$$= \frac{-R_L}{r_{bb'} + R_s} \times \frac{2\pi f_T \cdot C_e}{2\pi [C_e + C_c \cdot 2\pi f_T \cdot C_e \cdot R_L]}$$

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

$$\boxed{Avslow \times f_H = \frac{-R_L}{r_{bb'} + R_s} \times \frac{f_T}{1 + 2\pi f_T C_c R_L}}$$

Gain Bandwidth product for current :

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

$$= \frac{-h_{fe} R_s}{R_s + h_{ie}} \times \frac{1}{2\pi [r_{b'e} \| r_{bb'} + R_s] C_{eq}}$$

$$\therefore r_{b'e} \| [r_{b'b} + R_s] = \frac{(r_{bb'} + R_s) r_{b'e}}{R_s + h_{ie}}$$

$$A_{is}low \times f_H = \frac{-h_{fe} R_s}{\cancel{R_s + h_{ie}}} \times \frac{\cancel{R_s + h_{ie}}}{2\pi \, r_{b'e} (r_{bb'} + R_s) \times C_{eq}}$$

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

$$= \frac{-g_m \cancel{r_{b'e}} \cdot R_s}{2\pi \cancel{r_{b'e}} (r_{bb'} + R_s) \times C_{eq}}$$

$$= \frac{-g_m R_s}{2\pi (r_{bb'} + R_s) C_{eq}}$$

$$= \frac{-R_s}{r_{bb'} + R_s} \times \frac{g_m}{2\pi (C_e + C_c (g_m R_L + 1))}$$

$$= \frac{-R_s}{r_{bb'} + R_s} \times \frac{g_m}{2\pi [C_e + C_c g_m R_L)}$$

$\therefore g_m R_L >> 1$

$$= \cancel{\frac{-R_s}{r_{bb'} + R_s}} \times \cancel{\frac{g_m}{2\pi C_e [}}$$

$g_m = \cancel{2\pi f_T}$

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

$$g_m = f_T 2\pi C_e$$

$$= \frac{-R_S}{r_{bb'}+R_S} \times \frac{2\pi f_T C_e}{2\pi\left[C_e + C_c 2\pi f_T C_e R_L\right]}$$

$$= \frac{-R_S}{r_{bb'}+R_S} \times \frac{\cancel{2\pi} f_T \cancel{C_e}}{\cancel{2\pi} \cancel{C_e}\left[1 + C_c 2\pi f_T R_L\right]}$$

$$\boxed{A_{is\,low} \times f_H = \frac{-R_S}{r_{bb'}+R_S} \times \frac{f_T}{1+2\pi f_T C_c R_L}}$$

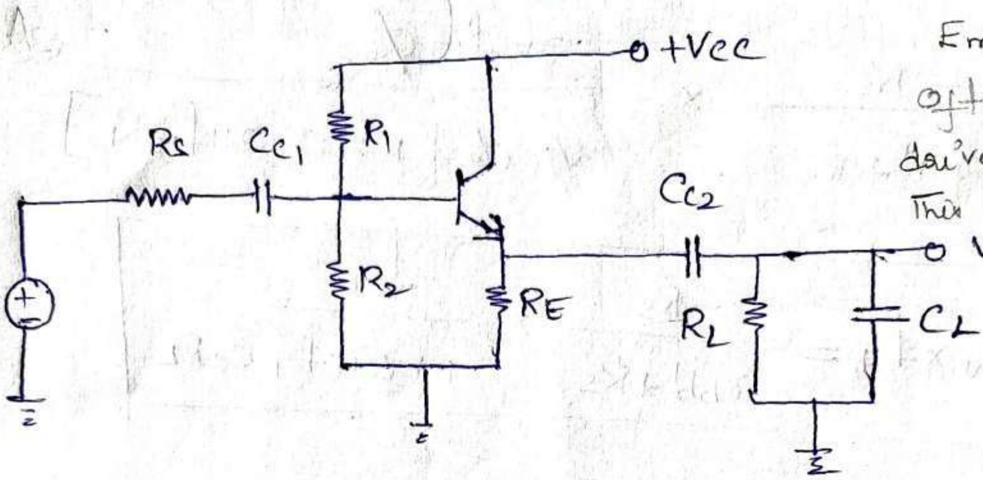
The quantities f_H, $A_{is}(low)$, $A_{vs}(low)$ will characterized the transistor stage, depending upon both R_L & R_S.

Here voltage gain bandwidth product ↑ses with increase in R_L & ↓ses with increase R_S.

∴ Gain band width product is not constant, depends on R_L & R_S.

Emitter Follower at High frequencies (CC)

Common Collector at High frequencies



Emitter follower is oftenly used to drive capacitive loads. This is bcz of its small output impedance

Emitter follower circuit.

An emitter follower circuit is shown in figure.

→ The output signal at the emitter being capacitively coupled to the load.

→ The high frequency small-signal equivalent circuit is shown in the below figure.

→ The coupling capacitors acting effectively as short circuits.

→ It is observed that capacitor $C_{b'c}$ is tied to ground potential and also that r_{ce} is in parallel with R_E & R_L. Thus we have $R_L' = R_E \| R_L \| r_{ce}$

→ In the analysis, the effect of C_L can be neglected.

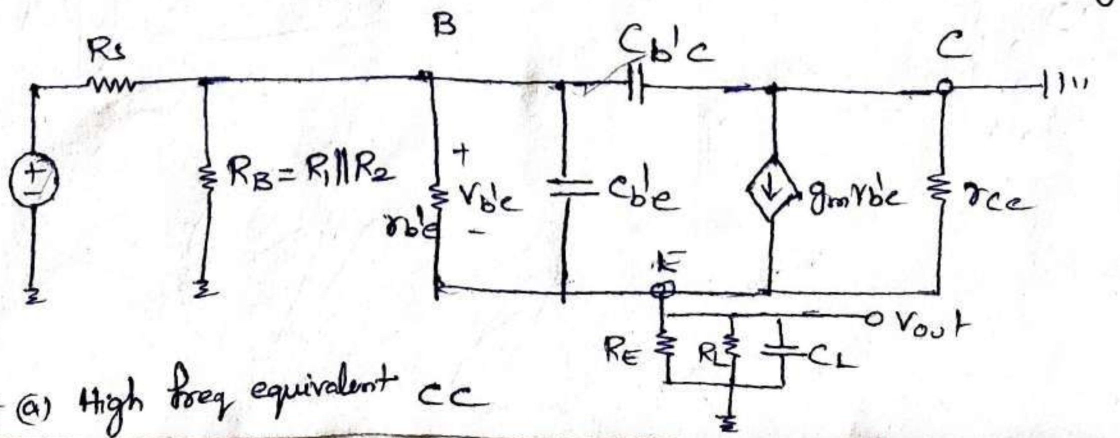


fig (a) High freq equivalent CC

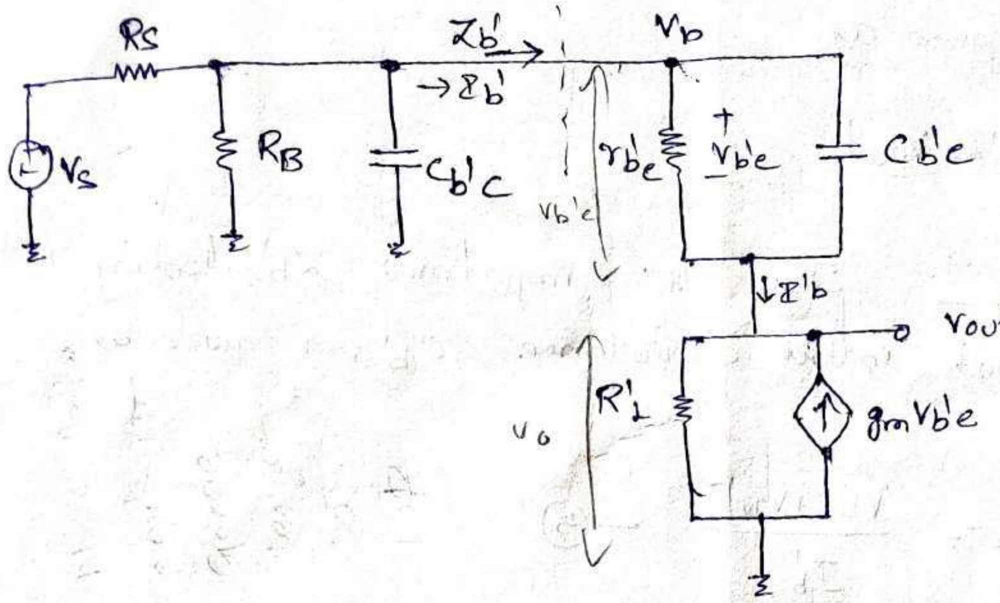


fig 1-(b) Rearranged High frequency equivalent circuit of emitter follower (or) CC

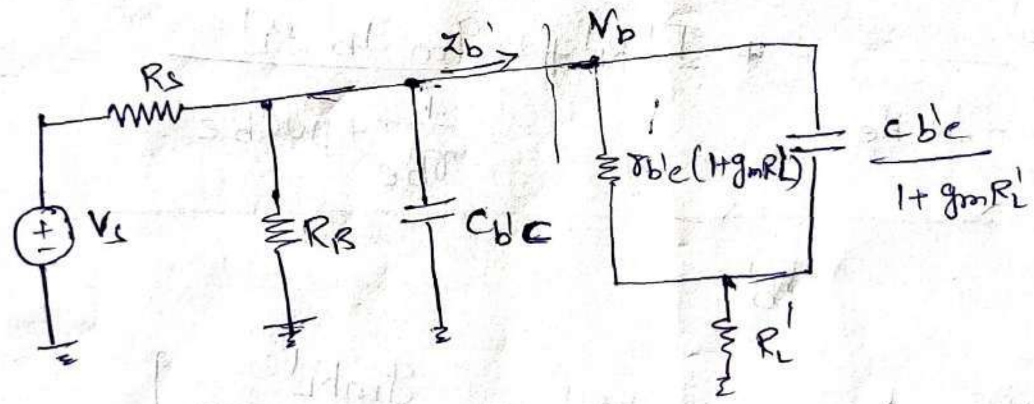


fig 1(c) High frequency equivalent circuit of Emitter follower with effective input base impedance

→ The current I_b' entering the parallel combination of $r_{b'e}$ and $C_{b'e}$ is the same as that coming out of the combination. So output voltage is given as

$$V_{out} = [I_b' + g_m V_{b'e}] R_L' \rightarrow (1)$$

$V = IR$

$V = \frac{I}{Y}$

$$V_{b'e} = \frac{I_b'}{\frac{1}{r_{b'e}} + j\omega C_{b'e}} \rightarrow \text{from figure} \quad (2)$$

$Y = \frac{1}{r_{b'e}} + C_{b'e}$

Voltage V_b is given as

$$V_b = V_{b'e} + V_{out}$$

Input impedance:- Thus the impedance Z_b' looking into the base without ~~incl~~ capacitance $C_{b'c}$ is given as.

$$\therefore Z_b' = \frac{V_b}{I_b'} = \frac{V_{b'e} + V_{out}}{I_b'} \quad — (3)$$

Combining equation 1, 2, 3

$$Z_b' = \frac{\dfrac{I_b'}{\dfrac{1}{r_{b'e}} + j\omega C_{b'e}} + I_b' R_L' + \dfrac{g_m I_b' R_L'}{\dfrac{1}{r_{b'e}} + j\omega C_{b'e}}}{I_b'}$$

$$= \frac{\cancel{I_b'}\left[\dfrac{1}{\dfrac{1}{r_{b'e}} + j\omega C_{b'e}} + R_L' + \dfrac{g_m R_L'}{\dfrac{1}{r_{b'e}} + j\omega C_{b'e}}\right]}{\cancel{I_b'}}$$

$$= \frac{1}{\dfrac{1}{r_{b'e}} + j\omega C_{b'e}} \times \left[1 + g_m R_L'\right] + R_L'$$

$$Z_b' = \frac{1}{\dfrac{1}{r_{b'e}[1 + g_m R_L']} + \dfrac{j\omega C_{b'e}}{[1 + g_m R_L']}} + R_L' \rightarrow (4)$$

The impedance Z_b' is shown in the equivalent circuit of figure (C). The above equations shows that the effect of capacitance $C_{b'c}$ is reduced in this configuration

From equation 1 & 2 we have

$$V_{out} = [I_b' + g_m V_{b'e}] R_L' \rightarrow (5)$$

$$I_b' = V_{b'e}\left[\frac{1}{r_{b'e}} + jwC_{b'e}\right]$$

Sub I_b value in eq (5)

$$V_{out} = \left[V_{b'e}\left[\frac{1}{r_{b'e}} + jwC_{b'e}\right] + g_m V_{b'e}\right] R_L'$$

$$\boxed{V_{out} = V_{b'e}\left[\left(\frac{1}{r_{b'e}} + jwC_{b'e}\right) + g_m\right] R_L'}$$

eq (4) shows the effect of capacitance $C_{b'c}$ is reduced in this config

~~Vout~~

which yields zero when $\frac{1}{r_{b'e}} + jwC_{b'e} + g_m = 0$

$$\frac{1 + jwC_{b'e}r_{b'e} + g_m r_{b'e}}{r_{b'e}} = 0$$

$$1 + jwC_{b'e}r_{b'e} + g_m r_{b'e} = 0$$

$$1 + j2\pi f \cdot C_{b'e}r_{b'e} + g_m r_{b'e} = 0 \qquad \therefore g_m r_{b'e} = h_{fe} = \beta$$

$$1 + \beta + j2\pi f \cdot C_{b'e} \cdot r_{b'e} = 0$$

$$1 + \beta = -j2\pi f \cdot C_{b'e} \cdot r_{b'e}$$

By taking modulus on both sides

$$f_0 = \frac{1+\beta}{2\pi C_{b'e} r_{b'e}}$$

$$\boxed{f_0 = \frac{1}{2\pi C_{b'e}\left[\frac{r_{b'e}}{1+\beta}\right]}}$$

Here frequency f_0 is usually very high because of very small value of $\frac{r_{b'e}}{1+\beta}$

→ An approximate value of one pole can be determined by making a simplifying assumption.

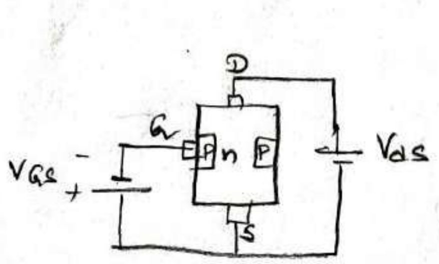
→ In many applications, the impedance of $r_{b'e}[1+g_m R_L')$ in parallel with $C_{b'e}/(1+g_m R_L')$ is very large in comparison to R_L', so neglecting R_L', the time constant is given as

$$\tau_p = \left[R_S \| R_B \| (1+g_m R_L')\, r_{b'e}\right] \left[C_{b'c} + \frac{C_{b'e}}{1+g_m R_L'}\right] \rightarrow (6)$$

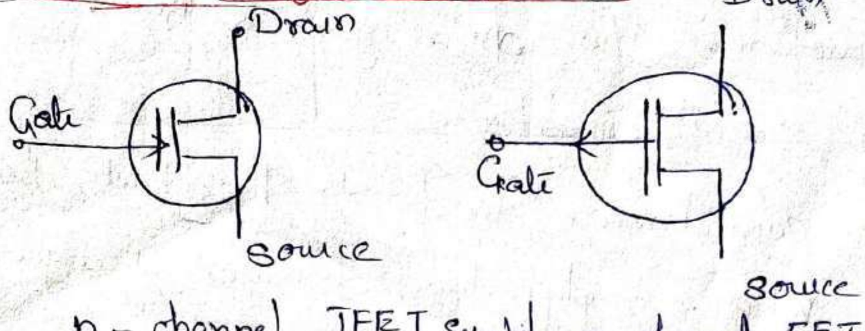
The upper cutoff [or 3dB] frequency is given as

$$\boxed{f_2 = \frac{1}{2\pi \tau_p}}$$

FET :- In small signal analysis equivalent circuit gate terminal is represented with open circuit, i.e. between Gate & Source open circuit, bcz. for normal operation of JFET, Gate is always reverse bias, the current in the reverse bias circuit is ideally equal to zero, so that the Gate terminal is in open circuit.

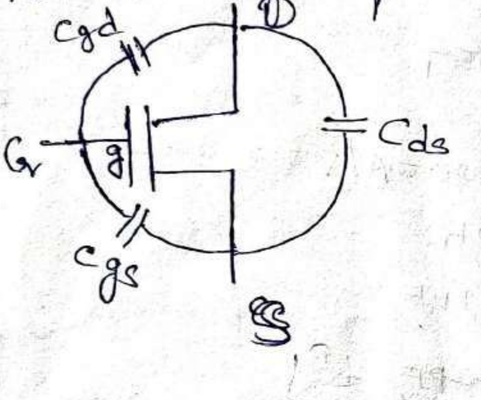


FET at high frequencies



n – channel JFET symbol p– channel FET symbol

FET at high frequencies



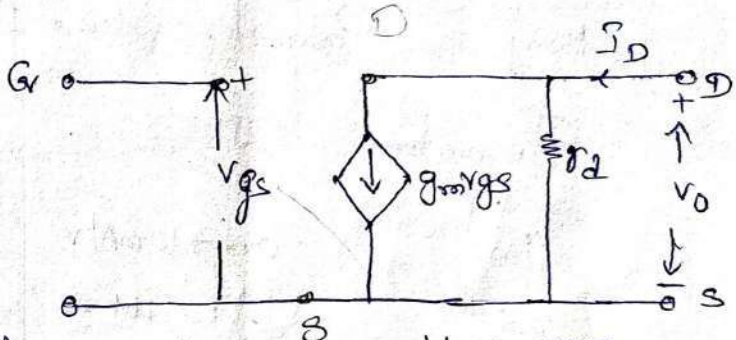


fig: Small signal model of FET

We know that $V_0 = I_d \cdot r_d$

$$V_0 = -g_m V_{gs} \cdot r_d$$

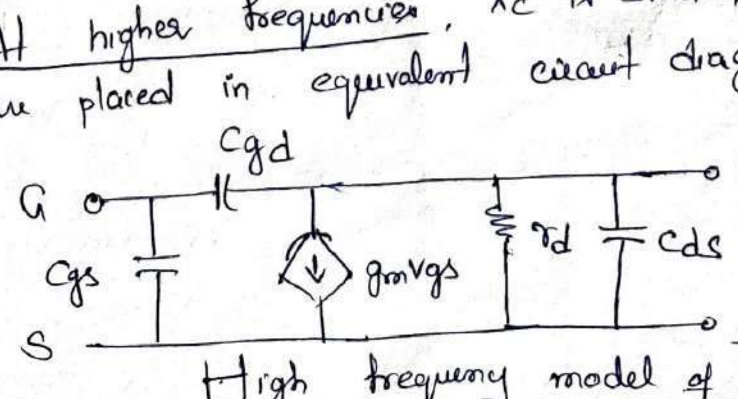
$$\frac{V_0}{V_{gs}} = -g_m r_d$$

$$K = \frac{V_0}{V_{gs}} = A_V = -g_m r_d = -\mu$$

→ At low frequencies

f is small → X_C is large ∴ $X_C = \frac{1}{2\pi f C}$ → so the Capacitive reactance can be represented as open circuit.

→ At higher frequencies, X_C is small, so the internal capacitances are placed in equivalent circuit diagram.



High frequency model of FET

C_{gs} → represents barrier capacitance b/w gate and source

C_{gd} → barrier capacitance b/w gate & drain

C_{ds} → Drain to source capacitance

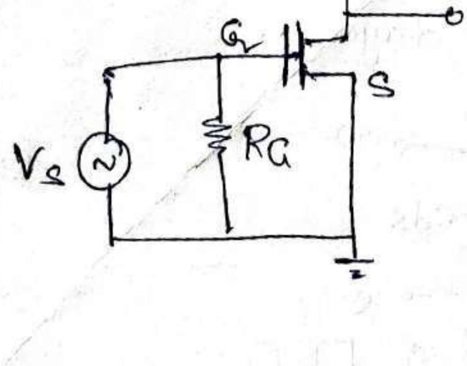
These internal capacitances leads to feedback from o/p to input and voltage amplification decreases at higher frequencies

Range of Parameter values for an FET

Parameter	JFET	MOSFET
g_m	0.1 – 10 mA/V	0.1 – 20 mA/V or more
r_d	0.1 – 1M Ω	1 – 50 K
C_{ds}	0.1 – 1PF	0.1 – 1PF
C_{gs}, C_{gd}	1 – 10 PF	1 – 10 PF
r_{gs}	$> 10^8$ Ω	
r_{gd}	$> 10^8$ Ω	

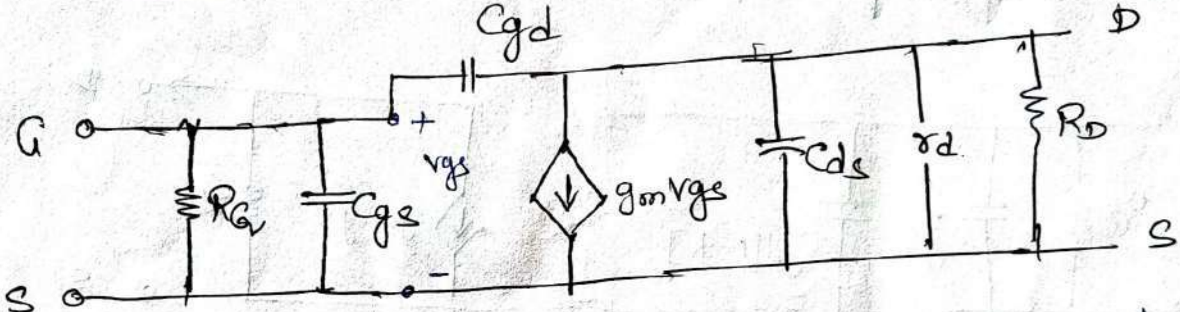
High frequency Analysis of Common Source FET

In common source amplifier the input signal is applied at gate terminal & o/p is obtained from drain terminals. The circuit diagram of CS Amplifier is shown in figure.



Procedure :-

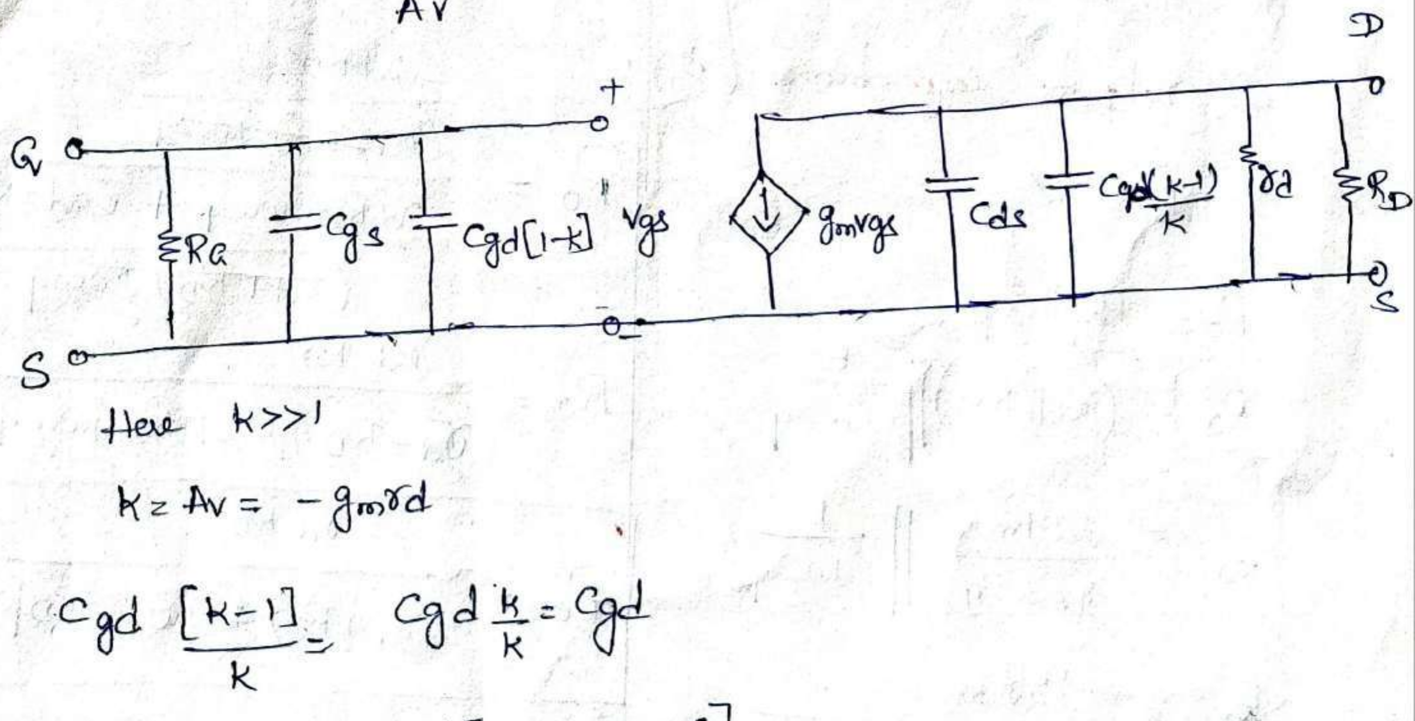
1. Replace field effect transistor with small signal model.
2. Connect the external components to the circuit diagram
3. Connect all internal components
4. ~~Conne~~ Calculate A_v, R_i & R_o



According to miller theorem the above circuit is simplified as

$C_1 = C_{gd}[1-A_v]$ (or) $C_1 = C_{gd} * [1-k]$

$C_2 = C_{gd}\frac{[A_v - 1]}{A_v}$ (or) $C_2 = C_{gd} * \left[\frac{k-1}{k}\right]$



Here $k >> 1$

$k = A_v = -g_m r_d$

$C_{gd}\frac{[k-1]}{k} \simeq C_{gd}\frac{k}{k} = C_{gd}$

$C_{gd}(1-k) = C_{gd}[1-(-g_m r_d)]$

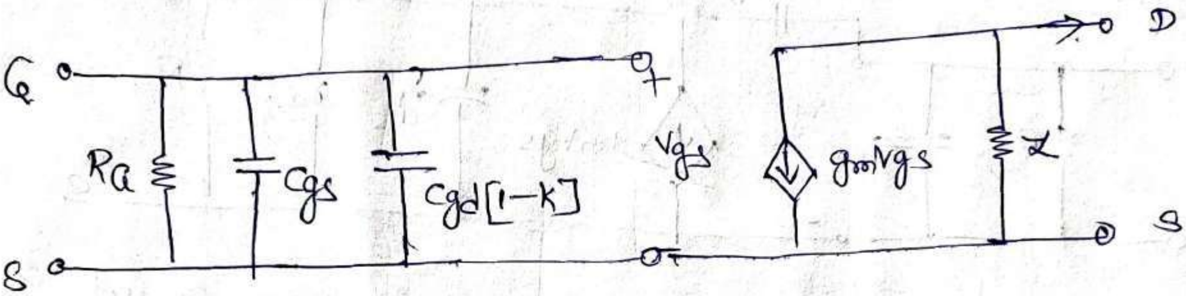
$C_{gd}(1-k) = C_{gd}[1+g_m r_d]$

$R_{eq} = r_d \| R_D$

$C_{eq} = C_{ds} + C_{gd}$

$X_{ceq} = \dfrac{1}{j\omega C_{eq}}$

$Z = X_{Ceq} \| R_{eq}$



(1) voltage gain (A_v) :- $V_0 = IZ$

$$V_0 = -g_m V_{gs} Z$$

$$\boxed{A_v = \frac{V_0}{V_{gs}} = -g_m Z} \longrightarrow \text{Voltage gain}$$

(2) Output impedance (R_0) :-

$R_0 = Z$

$R_0 = R_{eq} \| C_{eq}$

$R_0 = (r_d \| R_D) \| \dfrac{1}{j\omega C_{eq}}$

$R_0 = \dfrac{r_d R_D}{r_d + R_D} \| \dfrac{1}{j\omega C_{eq}}$

$$R_0 = \frac{\dfrac{r_d R_D}{r_d + R_D} \times \dfrac{1}{j\omega C_{eq}}}{\dfrac{r_d R_D}{r_d + R_D} + \dfrac{1}{j\omega C_{eq}}}$$

$$R_0 = \frac{\dfrac{r_d R_D}{\cancel{(r_d + R_D)\, j\omega C_{eq}}}}{\dfrac{r_d R_D\, j\omega C_{eq} + (r_d + R_D)}{\cancel{(r_d + R_D)\, j\omega C_{eq}}}}$$

$$R_0 = \frac{r_d R_D}{(r_d + R_D) + r_d R_D\, j\omega C_{eq}}$$

$$\boxed{R_0 = \frac{r_d R_D}{(r_d + R_D) + j\omega\, r_d R_D [C_{ds} + C_{gd}]}}$$

(3) Input impedance (R_i)

$$R_i = R_G \parallel \frac{1}{j\omega C_{eq}} \qquad C_{eq} = C_{gs} + C_{gd}[1+g_m r_d]$$

$$X_{ceq} = \frac{1}{j\omega C_{eq}}$$

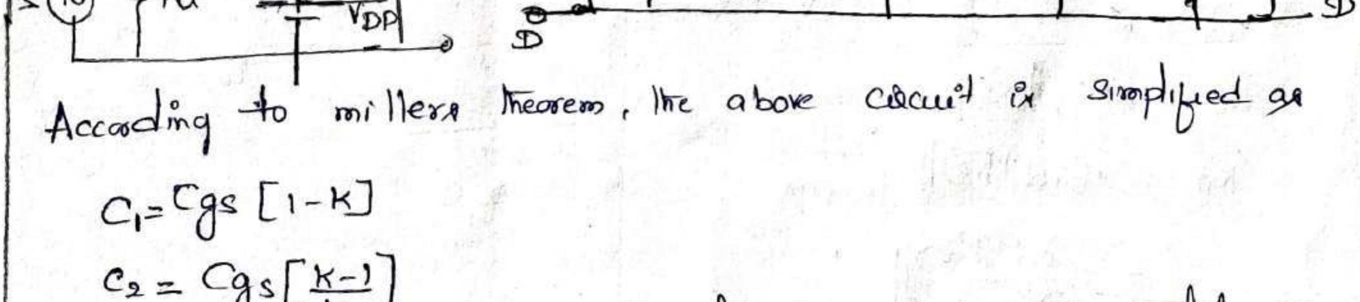
$$R_i = \frac{R_G \times \frac{1}{j\omega C_{eq}}}{R_G + \frac{1}{j\omega C_{eq}}}$$

$$R_i = \frac{\frac{R_G}{j\omega C_{eq}}}{\frac{j\omega C_{eq} R_G + 1}{j\omega C_{eq}}}$$

$$\boxed{R_i = \frac{R_G}{1 + j\omega C_{eq} R_G}}$$

High frequency analysis of Common drain Amplifier:

In common drain amplifier, the drain terminal acts as common between input and output ports. The input signal is applied at gate terminal & output is obtained across source.

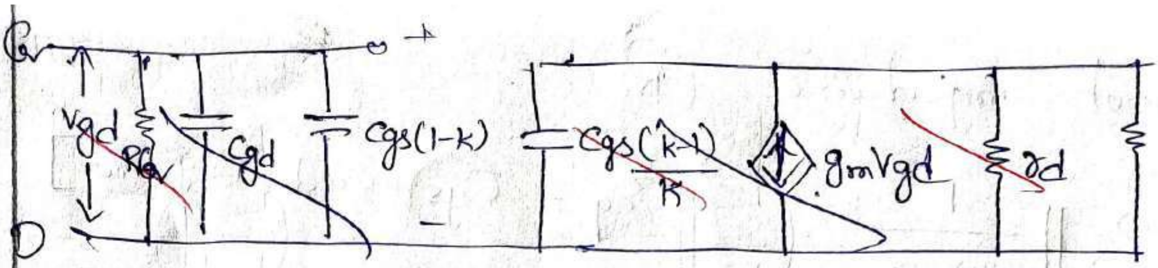


According to millers theorem, the above circuit is simplified as

$C_1 = C_{gs}[1-K]$

$C_2 = C_{gs}\left[\frac{K-1}{K}\right]$

we can also redraw the high frequency FET CD model as follows



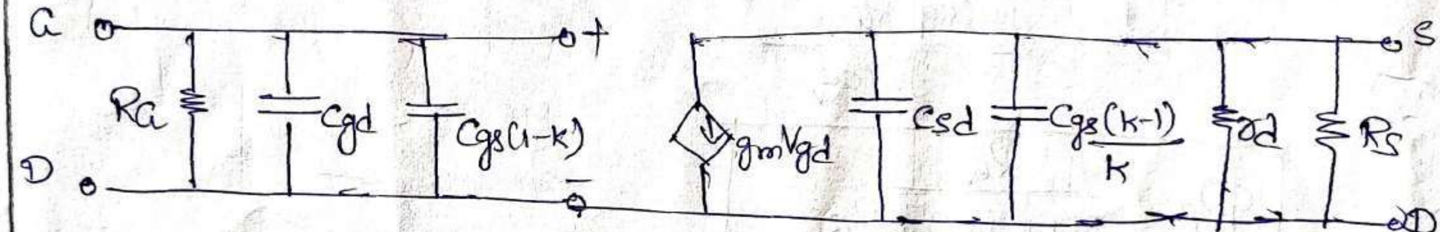
Voltage gain:

~~$Z = R_{eq} \parallel X_{ceq}$~~

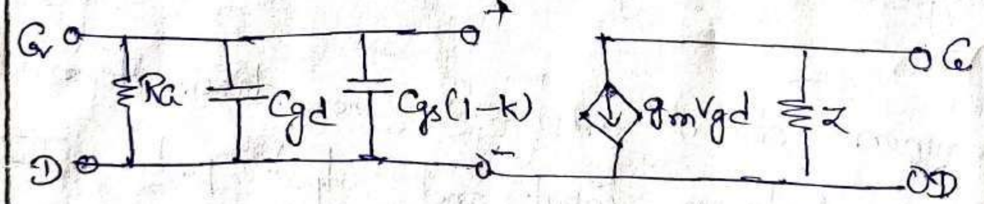
$K >> 1$

~~$R_{eq} = r_d \parallel R_L$~~

~~$C_{eq} = C$~~



(i) Voltage gain:



$$\left\{\begin{array}{l} C_{gs}\dfrac{(K-1)}{K} \quad K>>1 \\ C_{gs}\left[\dfrac{K}{K}\right] = C_{gs} \end{array}\right.$$

$V_0 = IZ$

$= -g_m V_{gd} Z$

$$\boxed{A_V = \frac{V_0}{V_{gd}} = -g_m Z} \Rightarrow \boxed{A_V = -g_m(R_{eq} \parallel C_{eq})}$$

$Z = R_{eq} \parallel X_{ceq}$

$K >> 1$

$R_{eq} = r_d \parallel R_S$

$C_{eq} = C_{sd} + C_{gs}$

$X_{ceq} = \dfrac{1}{j\omega C_{eq}}$

(ii) Output impedance (R_0):

$R_0 = Z$

$R_0 = R_{eq} \parallel C_{eq}$

$R_0 = (r_d \parallel R_S) \parallel \dfrac{1}{j\omega C_{eq}}$

$= \dfrac{r_d \cdot R_S}{r_d + R_S} \parallel \dfrac{1}{j\omega C_{eq}}$

$$R_o = \frac{\frac{r_d R_s}{r_d + R_s} \times \frac{1}{j\omega C_{eq}}}{\frac{r_d R_s}{r_d + R_s} + \frac{1}{j\omega C_{eq}}}$$

$$R_o = \frac{\frac{r_d R_s}{(r_d + R_s)\, j\omega C_{eq}}}{\frac{r_d R_s}{r_d + R_s} + \frac{1}{j\omega C_{eq}}}$$

$$R_o = \frac{\frac{r_d R_s}{\cancel{r_d + R_s\,(j\omega C_{eq})}}}{\frac{r_d R_s (j\omega C_{eq}) + (r_d + R_s)}{\cancel{r_d + R_s\,(j\omega C_{eq})}}}$$

$$R_o = \frac{r_d R_s}{r_d R_s (j\omega C_{eq}) + (r_d + R_s)}$$

$$R_o = \frac{r_d R_s}{r_d R_s j\omega (C_{gs} + C_{ds}) + (r_d + R_s)}$$

(iii) Input impedance (R_i):

$$R_i = R_G \parallel \frac{1}{j\omega C_{eqi}}$$

$$C_{eq} = C_{gd} + C_{gs}[1 + g_m r_d]$$

$$X_{eq} = \frac{1}{j\omega C_{eq}}$$

$$R_i = \frac{R_G \times \frac{1}{j\omega C_{eq}}}{R_G + \frac{1}{j\omega C_{eq}}}$$

$$R_i = \frac{R_G}{1 + j\omega R_G C_{eq}}$$

$$\boxed{R_i = \frac{R_G}{1 + j\omega R_G (C_{gd} + C_{ds}(1 + g_m r_d))}}$$

f_α:

f_α is the frequency at which the gain of common base amplifier is reduced to $1/\sqrt{2}$ times of maximum gain. Procedure to define f_α is similar to f_β.

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

$h_{fb} \rightarrow$

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

if $f = f_\alpha$

$$A_i = \frac{-h_{fb}}{1+j\left(\frac{f_\alpha}{f_\alpha}\right)} = -\frac{h_{fb}}{1+j}$$

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

f_α interms of Common Emitter Amplifier

$$h_{fb} = \frac{-h_{fe}}{1+h_{fe}}$$

$$f_\alpha = \frac{1}{2\pi r_{b'e}\left[1-\frac{h_{fe}}{1+h_{fe}}\right]C_{eq}}$$

$$f_\alpha = \frac{1}{2\pi r_{b'e}\left[\frac{1+\cancel{h_{fe}}-\cancel{h_{fe}}}{1+h_{fe}}\right]C_{eq}}$$

$$f_\alpha = \frac{1+h_{fe}}{2\pi r_{b'e}C_{eq}}$$

$\because h_{fe} >> 1$

$$\boxed{f_\alpha = \frac{h_{fe}}{2\pi r_{b'e}C_{eq}}}$$

Previous year Problems!-

3). Given $I_C = 5mA$, $V_{ce} = 10V$, $h_{fe} = 100$, $h_{ie} = 600\Omega$ - $C_e = 3pf$

$A_{ie} = 10$ at $10MHz$, find f_β & f_T

We know $|A_i| = \dfrac{h_{fe}}{\sqrt{1+(f/f_\beta)^2}}$

Given $|A_i| = 10$ at $f = 10MHz$

$h_{fe} = 100$

$f_\beta = ?$

$$\cancel{10} = \frac{10\cancel{0}}{\sqrt{1+\left[\frac{10\times10^6}{f_\beta}\right]^2}} \neq \cancel{10} = \frac{100}{\sqrt{1+}}$$

$$\cancel{100} = \frac{\cancel{10}}{\cancel{1+\left[\frac{10+10^6}{f_\beta}\right]^2}}$$

$\cancel{100} = \cancel{100}$

$$1 + \left[\frac{10\times10^6}{f_\beta}\right]^2 = 10^2$$

$$\left[\frac{10\times10^6}{f_\beta}\right]^2 = 100-1$$

$$\frac{10\times10^6}{f_\beta} = \sqrt{99}$$

$$\frac{10^7}{f_\beta} = \sqrt{99}$$

$f_\beta = 1MHz$

$f_T = h_{fe} f_\beta = 100\times1\times10^6$

$f_T = 100MHz$

A high frequency amplifier uses a transistor which is driven from a source with $R_S = 0$. Calculate value of f_H, if $R_L = 0$ & $R_L = 1\,k\Omega$. Assume typical values of hybrid – π parameter.

(1) Case (i) $R_L = 0$

$$f_H = \frac{1}{2\pi r_{b'e}[C_e + C_C]}$$

Typical values of

$r_{b'e} = 1\,k\Omega$

$C_e = 100\,pf$

$C_C = 3\,pf$

$$f_H = \frac{1}{2\pi \times 1\times10^3[100\times10^{-12} + 3\times10^{-12}]}$$

$$f_H = \frac{1}{2\pi \times 1\times10^3[103\times10^{-12}]}$$

$$f_H = 1.545\times10^6$$

(2) Case (ii) $R_L = 1\,k\Omega$

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

Typical values of

$r_{b'e} = 1\,k\Omega$, $C_e = 100\,pf$, $C_C = 3\,pf$

$g_m = 50\,mA/V$

$$f_H = \frac{1}{2\pi\times1\times10^3\left[100\times10^{-12} + 3\times10^{-12}[1 + 50\times10^{-3}\times1\times10^3]\right]}$$

$$f_H = \frac{1}{2\pi\times10^3\left[103\times10^{-12}[1 + 50\times10^{-3}\times10^3]\right]}$$

$$f_H = \frac{1}{2\pi\times10^3[103\times10^{-12}[51]]}$$

$$f_H = \frac{1}{2\pi\times103\times10^{-12}\times51\times10^3}$$

$= 629\,kHz$

$f_H = \cancel{30297.9\times10^{12}\,Hz}$

$f_H = \cancel{30291.9\,Hz = 0.030\,MHz}$

6.29×10^{11}

$= 0.6290\,MHz$

A high frequency amplifier uses a transistor which is driven from a source with $R_S = 1\,k\Omega$, Calculate f_H, A_V low and A_{Vs} high, if $R_L = 0$, $R_L = 1\,k\Omega$, Assume typical values of π parameter.

(1) For $R_L = 0$

We have $f_H = \frac{1}{2\pi R_{eq} C_{eq}}$

$R_{eq} = r_{b'e} \| (r_{bb'} + R_s)$

$C_{eq} = C_e + C_c[1 + g_m R_L]$

$r_{b'e} = 1\,K\Omega$, $r_{bb'} = 100\,\Omega$, $C_e = 100\,pf$, $C_c = 3\,pf$, $g_m = 50\,mA/V$

$$f_H = \frac{1}{2\pi \left[r_{b'e} \| [r_{bb'} + R_s]\right] [C_e + C_c]}$$

$$= \frac{1}{2\pi \times \left[\frac{1 \times 10^3 \times [100 + 1 \times 10^3]}{1 \times 10^3 + [100 + 1 \times 10^3]}\right] \left[100 \times 10^{-12} + 3 \times 10^{-12}\right]}$$

$f_H = 29.5\ MHz$

→ $A_{vs}\ low = \frac{-h_{fe} R_L}{R_s + h_{ie}} = 0 \quad \because R_L = 0$

→ $A_{is}\ low = \frac{-g_m r_{b'e} R_s}{R_s + r_{bb'} + r_{b'e}} = -23.8$

For $R_L = 1K\Omega$

We have $f_H = \frac{1}{2\pi R_{eq} C_{eq}}$

$R_{eq} = r_{b'e} \| (r_{bb'} + R_s)$

$C_{eq} = C_e + C_c(1 + g_m R_L)$

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

$f_H = 1.2 MHz$

$A_{is} low = \frac{-g_m r_{b'e} R_s}{r_{b'e} + R_s + r_{bb'}}$

$= -23.8$

$A_{vs}\ low = \frac{-h_{fe} R_L}{R_s + h_{ie}}$

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

Previous years Problem

(42)

A transistor amplifier in CE configuration is operated at high frequency with $f_T = 6MHz$, $g_m = 0.04$, $h_{fe} = 50$, $r_{bb'} = 100\Omega$, $R_S = 500\Omega$, $C_{b'c} = 10pf$, $R_L = 100\Omega$. Compute A_v, f_H, gain bandwidth product.

→ Voltage gain including $R_S = \frac{-h_{fe} \cdot R_L}{R_S + h_{ie}}$

$$A_{vs\,low} = \frac{-50 \times 100}{500 + h_{ie}} = \frac{-50 \times 100}{500 + 1.35k\Omega} = \frac{-5000}{1850}$$

$$\boxed{A_{vs} = -2.70}$$

$$r_{b'e} = \frac{h_{fe}}{g_m} = \frac{50}{0.04} = 1.25k\Omega$$

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

$$= 100\Omega + 1250\Omega$$

$$h_{ie} = 1.35k\Omega$$

→ $f_H = \frac{1}{2\pi R_{eq} C_{eq}} = 373888.1767 \Rightarrow 0.373 MHz = 373.888 kHz$

$R_{eq} = r_{b'e} \| (r_{bb'} + R_S) = 405\Omega$

$C_{eq} = [C_e + C_c(1 + g_m R_L)] = [1nf + 10pf[1 + 0.04 \times 100\Omega]] = 1.05 \times 10^{-9} f$

$$C_e = \frac{g_m}{2\pi f_T} = \frac{0.04}{2\pi \times 6 \times 10^6} = 1nf$$

→ Gain band width product for voltage

$= A_{vs\,low} \times f_H$ (or). $A_{vs\,low} \times f_H = \frac{-R_L}{r_{bb'} + R_S} \times \frac{f_T}{1 + 2\pi f_T C_c R_L}$

$= -999000 \text{ Hz}$

$= -0.9 \text{ MHz}$

$\simeq -1 \text{ MHz}$

Given $g_m = 38mV/A$, $r_{b'e} = 5.9k\Omega$, $h_{ie} = 6k\Omega$, $r_{bb'} = 100\Omega$, $C_{b'c} = 12pf$, $C_{b'e} = 63pf$, $h_{fe} = 224$ at $1kHz$. Calculate f_α, f_β, & f_T

$$\frac{224}{2\pi \times 5.9\times10^3 \times 7.5} \quad C_{eq} = 7.5\times10^{-11}$$

$$f_\alpha = \frac{h_{fe}}{2\pi r_{b'e} C_{b'e}} = 95.9MHz$$

$$80.56MHz$$

$$= \frac{\cancel{224}}{2\pi \times}$$

$$f_\beta = \frac{1}{2\pi r_{b'e}[C_{b'e}+C_{b'c}]} = 0.359MHz$$

$$f_T = \frac{g_m}{2\pi[C_{b'e}+C_{b'c}]} = 80.63MHz$$

Determine all hybrid π parameters of a transistor at collector current of $I_C = 2mA$, $V_{CE}=20V$, $I_B = 20\mu A$, the transistor specifications are $\beta_0 = 100$, unity gain frequency $f_T = 50MHz$, $h_{oe} = 25\mu\mho$, $T = 300°K$, $C_{b'c} = 3pf$, $h_{ie} = 1.4k\Omega$, $h_{re} = 2.5\times10^{-4}$

$\because V_T = 26mV$ @ room temp

1. $g_m = \frac{I_C}{V_T} = \frac{2\times10^{-3}}{26\times10^{-3}} = 0.076 = 76mA$

2. $r_{b'e} = \frac{h_{fe}}{g_m} = \frac{100}{76\times10^{-3}} = 1.31k\Omega$

3. $r_{bb'} = h_{ie} - r_{b'e} = 1.4\times10^3 - 1.31\times10^3 = 0.09\times10^3 = 90\Omega$

4). $r_{b'c} = \frac{r_{b'e}}{h_{re}} = \frac{1.31\times10^3}{2.5\times10^{-4}} = 0.52\times10^7 = 5.2\times10^6 = 5.2M\Omega$

5). $r_{ce} = \frac{1}{h_{oe} - h_{fe}g_{b'c}} = \frac{1}{25\times10^{-6} - \frac{100}{5.2\times10^6}} = 170k\Omega$ $\quad \because g_{b'c} = \frac{1}{r_{b'c}}$

6) $C_{b'e} = \frac{g_m}{2\pi f_T} = \frac{76\times10^{-3}}{2\pi\times50\times10^6} = 0.24\times10^{-9} = 0.24nf$

Unit - 3 Oscillators

Introduction:

Any circuit which is used to generate a periodic voltage without an a.c. input signal is called oscillator. To generate the periodic voltage, the circuit is supplied with energy from a d.c. source. If the output voltage is a sine wave function of time, the oscillator is called a "sinusoidal" or "Harmonic" oscillator. Positive feedback and negative resistance oscillators belong to this category. There is another category of oscillators which generate non-sinusoidal wave forms such as square, rectangular, triangular, or sawtooth values. This chap surveys methods of generating the sinusoidal waveforms.

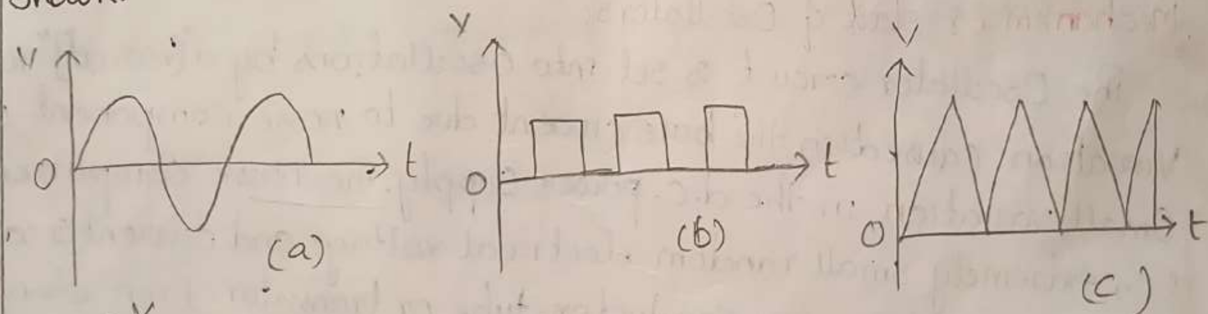
Classification of oscillators:

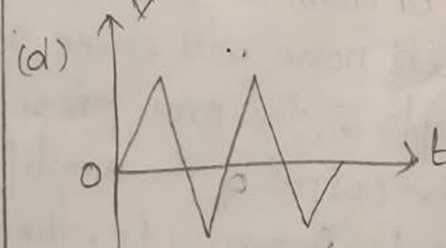
Oscillators are classified in the following different ways.

1. According to the waveforms generated:

(a) Sinusoidal oscillator (b) Relaxation Oscillator.

Sinusoidal Oscillator generates sinusoidal voltages (or) currents as shown.





Waveforms generated by oscillators: (a) Sinusoidal (b) Square (c) Sawtooth and (d) Triangular:

Relaxation Oscillator generates voltages or currents which vary abruptly one or more times in a cyclic of oscillation as shown in fig (b) to (d)

2. According to the fundamental mechanisms involved:

(a) Negative resistance Oscillators

(b) Feedback Oscillators.

Negative resistance Oscillator uses negative resistance of the amplifying device to neutralize the positive resistance of the Oscill

Feedback Oscillator uses positive feedback in the feedback amplifie to satisfy the Barkhausen criterion.

3. According to the frequency generated:

(a) Audio frequency oscillator (AFO): up to 20 kHz.

(b) Radio frequency Oscillator (RFO): 20 kHz to 30 MHz.

(c) very high frequency (VHF) oscillator: 30 MHz to 300 MHz.

(d) Ultra high frequency (UHF) oscillator: 300 MHz to 3 GHz.

(e) Microwave frequency Oscillator: above 3 GHz.

4. According to the type of circuit used, sine-wave Oscillators may be classified as.

(a) LC tuned Oscillator.

(b) RC phase shift Oscillator.

Conditions for Oscillation (Barkhausen Criterion)

Mechanism for start of Oscillators:

The Oscillator circuit is set into oscillations by a "variation" rand The Oscillator circuit is set into oscillations by a random variation caused in the base current due to noise component or variation caused in the d.c. power supply. The noise components small variation in the electrical voltage and currents are i.e., extremely small random electrical voltage and currents are always present in any conductor, tube or transistor. Even when no external signal is applied, the ever-present noise will cause some small signal at the output signal is applied, the ever-present small signal at the output of the amplifier noise will cause some small signal at a particular frequency fo, the a When the amplifier is tuned at a particular frequency fo, the a signal caused by noise signals will be predominantly at fo. If a small fraction (β) of the output signal is fed back to the input wi proper phase relation, then this feedback signal will be amplified the amplifier. If the amplifier has a gain or more than 1/β, then output increases and thereby the feedback signal becomes larger

This process continues and the output goes on increasing. But as the signal level increases, the gain of the amplifier [by the amplifier] decreases and at a particular value of output, the gain of the amplifier is reduced exactly equal to $1/\beta$. Then the output voltage remains constant at frequency f_0, called frequency of oscillation.

The essential conditions for maintains oscillations are:

1. $|A\beta| = 1$, i.e., the magnitude of loop gain must be unity.
2. The total phase shift around the closed loop is zero or 360 degrees.

Practical considerations

The condition $|A\beta| = 1$ gives a single and precise value of $A\beta$ which should be set through out the operation of the oscillator circuit. But in practise, as transistor characteristics and performance of other circuit components change with time, $|A\beta|$ will become greater or less than unity. Hence, in all practical circuits $|A\beta|$ should be set greater than unity so that the amplitude of oscillations will continue to increase without limit but such an increase in amplitude of oscillation is limited by the onset of the nonlinearity of operation in the active devices associated with the amplifier as shown in fig below. In this circuit, $A\beta$ is larger than unity for positive feedback. This onset of nonlinearity is an essential feature of all practical oscillators.

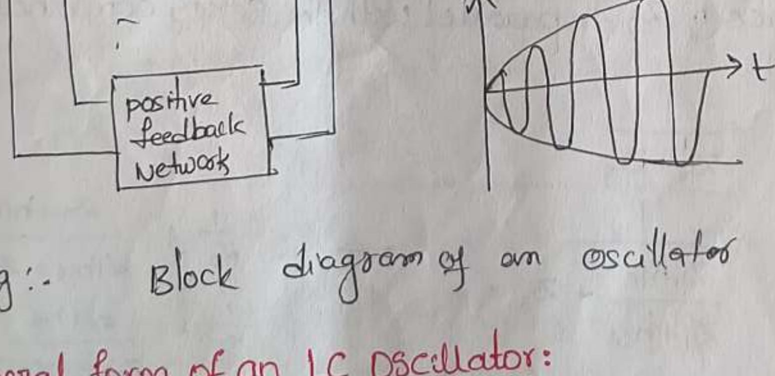


fig :- Block diagram of an oscillator

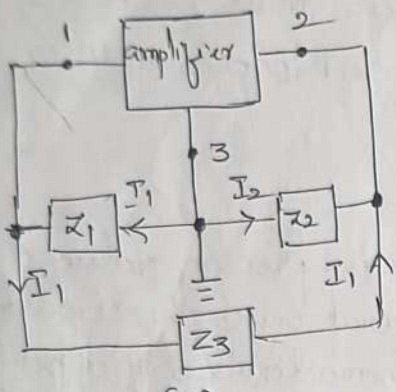
General form of an LC oscillator:

In the general form of oscillator shown in Fig 4.3 (a), any of the active devices such as vaccum tube, Transistor, FET and Operational amplifier may be used in the amplifier section. Z_1, Z_2 and Z_3 are reactive elements

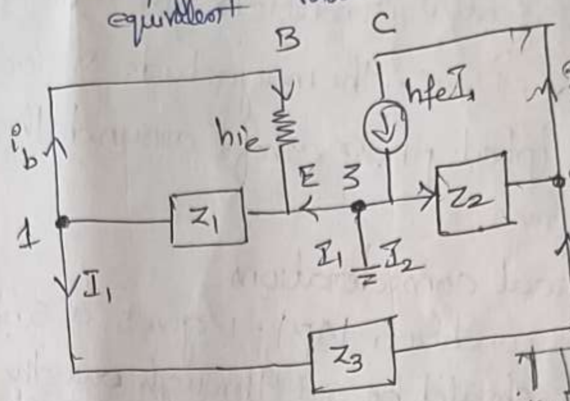
Constituting the feedback tank circuit which determining the frequency of oscillation. Here, z_1 and z_2 serve as an a.c. voltage divider for the output voltage and feedback signal. Therefore, the voltage across z_1 is the feedback signal. The frequency of oscillation of the LC oscillator is

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

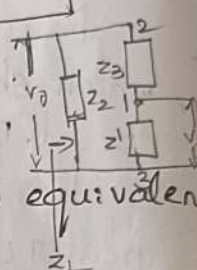
amplifier is replaced by hybrid equivalent model.



(a)



(b)



LC

Fig 4.3 (a) General form of an oscillator and (b) its equivalent circuit.

The inductive or capacitive reactances are represented by z_1, z_2 and Z_3. In Fig 4.3, the output terminals are 2 and 3, and input terminals are 1 and 3. Figure 4.3 (b) gives the equivalent circuit of Fig 4.3 (a).

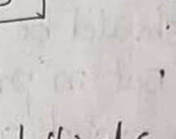
Load impedance Since z_1 and the input resistance h_{ie} of the transistor are in parallel, their equivalent impedance, z' is given by

$$\frac{1}{z'} = \frac{1}{z_1} + \frac{1}{h_{ie}}$$

{load impedance z_L by looking b/w the terminals 2 & 3

$$\frac{1}{z'} = \frac{z_1 + h_{ie}}{z_1 h_{ie}}$$

$$z' = \frac{z_1 h_{ie}}{z_1 + h_{ie}}$$



From this equation, we get

$$z' = \frac{z_1 h_{ie}}{z_1 + h_{ie}}$$

Now the load impedance z_L between the output terminals 2 and 3 is equivalent impedance of z_2 in parallel with the series combination of z' and z_3.

Therefore,
$$\frac{1}{z_L} = \frac{1}{z_2} + \frac{1}{z' + z_3}$$

$$= \frac{1}{z_2} + \frac{1}{\frac{z_1 h_{ie}}{z_1 + h_{ie}} + z_3}$$

$$\longrightarrow \frac{1}{z_2} + \frac{z_1 + h_{ie}}{z_1 h_{ie} + z_3(z_1 + h_{ie})}$$

$$\frac{1}{z_2} + \frac{z_1 + h_{ie}}{z_1 h_{ie} + z_3 z_1 + z_3 h_{ie}}$$

$$\frac{1}{z_2} + \frac{z_1 + h_{ie}}{h_{ie}[z_1 + z_3] + z_1 z_3}$$

$$= \frac{1}{z_2} + \frac{z_1 + h_{ie}}{h_{ie}(z_1 + z_3) + z_1 z_3}$$

$$\frac{h_{ie}[z_1 + z_3] + z_1 z_3 + z_2(z_1 + h_{ie})}{z_2[h_{ie}(z_1 + z_3) + z_1 z_3]}$$

$$= \frac{h_{ie}(z_1 + z_3) + z_1 z_3 + z_1 z_2 + z_2 h_{ie}}{z_2[h_{ie}(z_1 + z_3) + z_1 z_3]}$$ by taking LCM

i/p is applied b/w 1 & 3, o/p is collected b/w 2 & 3

Clapp oscillator: z_1 & z_2 – capacitor, z_3 – inductor series with a capacitor

Colpitts: z_1 & z_2 – capacitor, z_3 – inductor

depending on the type of reactance elements: Hartley: z_1 & z_2 – inductor, z_3 – capacitor

Therefore,
$$Z_L = \frac{Z_2\left[h_{ie}[Z_1+Z_3] + Z_1 Z_3\right]}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3}$$

Voltage gain without feedback This is given by. The voltage gain of CE amplifier without feedback for approximate hybrid model is $A_V = \frac{-h_{fe} Z_L}{h_{ie}}$

-ve sign indicates

$$A_{ve} = \frac{-h_{fe} Z_L}{h_{ie}}$$

Feedback Fraction β The output voltage between the terminals 3 and 2 interms of the current I_1 is given by

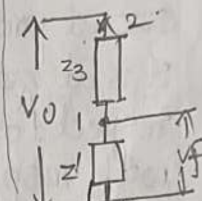
find the voltage gain A & feedback factor β to $A\beta = 1$ equivalent $A\beta = 1$

$\beta = \frac{V_f}{V_O}$

The o/p vltg is appeared across 2 & 3, & fb is applied b/w the terminals 1 & 3

$$V_O = -I_1(Z' + Z_3) = \left[\frac{Z_1 h_{ie}}{Z_1+h_{ie}} + Z_3\right] - I_1$$

$$= -I_1\left[\frac{h_{ie}(Z_1+Z_3)+Z_1Z_3}{Z_1+h_{ie}}\right]$$



The voltage fedback to the input terminals 3 and 1 is given by

$$V_{fb} = -I_1 Z' = -I_1\left[\frac{Z_1 h_{ie}}{Z_1+h_{ie}}\right]$$

Therefore, the fedback ratio β is given by

$$\beta = \frac{V_{fb}}{V_O} = \frac{I_1}{I_1}\left[\frac{Z_1 h_{ie}}{Z_1+h_{ie}}\right]\left[\frac{Z_1+h_{ie}}{h_{ie}(Z_1+Z_3)+Z_1Z_3}\right]$$

$$\beta = \frac{Z_1 h_{ie}}{h_{ie}(Z_1+Z_3)+Z_1Z_3}$$

Equation for oscillator, For oscillation, we must have,

$$A_{ve}\beta = 1$$

Substituting the values of A_{ve} and β, we get

$$\left[\frac{-h_{fe} Z_L}{\cancel{h_{ie}}}\right]\left[\frac{Z_1 \cancel{h_{ie}}}{h_{ie}(Z_1+Z_3)+Z_1Z_3}\right] = 1$$

$$\Rightarrow \left[\frac{h_{fe} Z_2 \cancel{[h_{ie}(Z_1+Z_3)+Z_1Z_3)}}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3}\right]\left[\frac{Z_1}{\cancel{h_{ie}(Z_1+Z_3)+Z_1Z_3}}\right] = -1$$

$$\Rightarrow \frac{h_{fe} Z_1 Z_2}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3} = -1$$

$$\Rightarrow h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3 = -h_{fe}Z_1Z_2$$

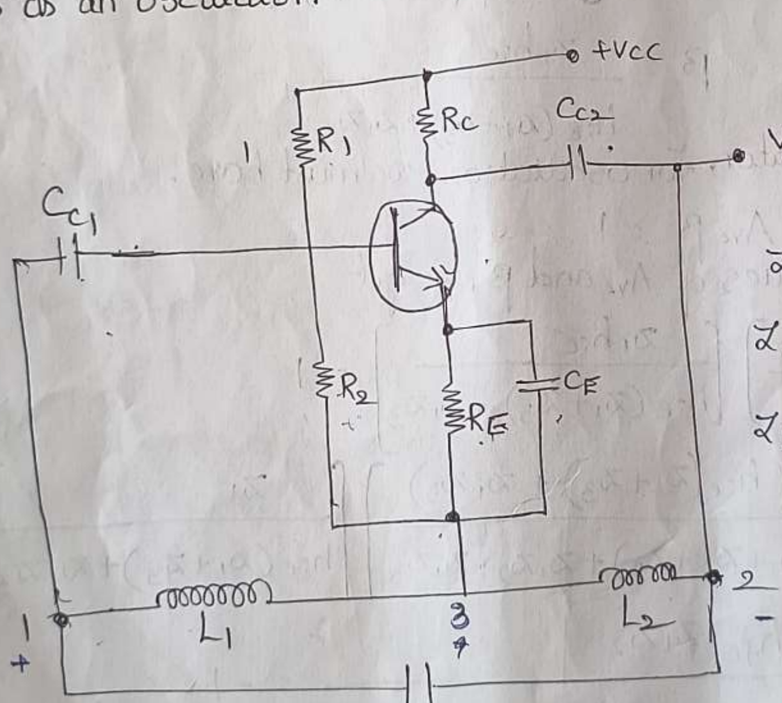
$$\Rightarrow \boxed{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2(1+h_{fe})+Z_1Z_3 = 0}$$

This is the general equation for the LC oscillator. to find the frequency of oscillation & condition for oscillation for all 3 LC oscillator, hearlty colpitts & Clapp oscillator

HARTLEY OSCILLATOR

In the Hartley oscillator shown in Fig 4.4, Z_1 and Z_2 are ind and Z_3 is a capacitor. Resistors R_1, R_2 and R_E provide the necessa d.c. bias to the transistor. C_E is a bypass capacitor. C_{C1} and C_{C2} are co capacitors. The feedback network consisting of inductors L_1 and L_2, a capacitor C determines the frequency of the oscillator.

When the supply voltage $+V_{CC}$ is switched ON, a transient curren produced in the tank circuit and consequently, damped harmonic oscillations are set up in the circuit. The oscillatory current in the tan circuit produces a.c. voltages across L_1 and L_2. As terminal 3 is earth it is at zero potential. If the terminal 1 is at a positive potential with respect to 3 at any instant, terminal 2 will be at a negative potentia with respect to 3 at the same instant. Thus the phase difference bet 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. The the total phase shift is 360°. Thus, at the frequency determined for th tank circuit, the necessary condition for sustained oscillations is satisfied. If the feedback is adjusted so that the loop gain $A\beta = 1$, the ci acts as an oscillator.



$Z_1 = j\omega L_1 + j\omega M$

$Z_2 = j\omega L_2 + j\omega M$

$Z_3 = \frac{1}{j\omega C} = \frac{-j}{\omega C}$

The frequency of oscillation is $f_0 = \frac{1}{2\pi\sqrt{LC}}$, where $L = L_1 + L_2$ and M is the value of mutual inductance between coils L_1 and L_2. The condition for sustained oscillation is

$$h_{fe} \geq \frac{L_1 + M}{L_2 + M}$$

Derivation for frequency of oscillations.

General equation for frequency of oscillations

$$h_{ie}(z_1+z_2+z_3)+z_1z_2(1+h_{fe})+z_1z_3=0 \longrightarrow (1)$$

$z_1 = j\omega L_1 + j\omega M$; $z_2 = j\omega L_2 + j\omega M$; $z_3 = \frac{1}{j\omega C} = \frac{-j}{\omega C}$

Substituting the values of z_1, z_3, z_2 in the equation (1)

$$h_{ie}\left(j\omega L_1 + j\omega M + j\omega L_2 + j\omega M - \frac{j}{\omega C}\right) + (j\omega L_1 + j\omega M)(j\omega L_2 + j\omega M)$$

$$(1+h_{fe}) + (j\omega L_1 + j\omega M)\left(\frac{-j}{\omega C}\right) = 0$$

$$j\omega h_{ie}\left(L_1+M+L_2+M-\frac{1}{\omega^2 C}\right) + j\omega(L_1+M)\,j\omega(L_2+M)(1+h_{fe}) +$$

$$j\omega(L_1+M) \times -j\left[\frac{1}{\omega C}\right] = 0$$

$$j\omega h_{ie}\left(L_1+2M+L_2-\frac{1}{\omega^2 C}\right) - \omega^2(L_1+M)(L_2+M)(1+h_{fe}) + \frac{L_1+M}{C} = 0$$

To find, frequency imaginary part $= 0$.

$$j\omega h_{ie}\left(L_1+L_2+2M-\frac{1}{\omega^2 C}\right) = \cancel{\omega^2(L_1+M)(L_2+M)}\ 0$$

$$L_1+L_2+2M = \frac{1}{\omega^2 C}$$

$$\omega^2 = \frac{1}{C(L_1+L_2+2M)}$$

$$(2\pi f)^2 = \frac{1}{C(L_1+L_2+2M)} \qquad \therefore L_1+L_2+2M = L_{eq}$$

$\frac{1}{2\pi\sqrt{L_{eq}C}}$

$$f = \frac{1}{2\pi\sqrt{(L_1+L_2+2M)C}} = f = \frac{1}{2\pi\sqrt{L_{eq}C}}$$

Real part $= 0$ [To find h_{fe}]

$$W^2(\cancel{L_1+M})(L_2+M)(1+h_{fe}) = \frac{\cancel{L_1+M}}{C}$$

L_1+L_2+2M

$$W^2(L_2+M)(1+h_{fe}) = \frac{1}{C}$$

$$(L_2+M)(1+h_{fe}) = \frac{1}{\omega^2 C}$$

$$1+h_{fe} = \frac{1}{\omega^2 C(L_2+M)}$$

$h_{fe} = \frac{1}{\omega^2 C} \times \frac{1}{L_2+M}$

$h_{fe} = L_1+L_2+2M \times \frac{1}{L_2+M}$

$h_{fe} = (L_1+M)+(\cancel{L_2+M}) \times \frac{1}{\cancel{L_2+M}}$

$\therefore 1+h_{fe} = 1+\frac{L_1+M}{L_2+M}$

$1+h_{fe} = \frac{1}{\omega^2 C} \times \frac{1}{(L_2+M)}$

$1+h_{fe} = (L_1+L_2+2M) \times \frac{1}{L_2+M}$

$1+h_{fe} = (L_1+M)+[\cancel{L_2+M}] \times \frac{1}{\cancel{L_2+M}}$

$1+h_{fe} = 1+\frac{L_1+M}{L_2+M}$

$h_{fe} = \frac{L_1+M}{L_2+M}$

$$1 + h_{fe} = 1 + \frac{L_1 + M}{L_2 + M}$$

$$h_{fe} = \frac{L_1 + M}{L_2 + M}$$

Disadvantages:

→ It cannot be used as a low frequency oscillator since the val of inductors becomes large and the size of the inductors beco large.

→ The harmonic content in the output of this oscillator is very hig and hence it is not suitable for the applications which require pure sine wave.

Advantages:

→ Very few components are needed including fixed inductors or tapped coil.

→ By using a variable capacitor or by varying the inductance, freq of oscillation can be varied.

→ Instead of two seperate coils L1 and L2, a single coil of bare wi can be used.

→ The amplitude of the output remains constant over the working frequ range.

Applications:

→ used as local oscillator in radio receivers

→ It is suitable for oscillations in RF range up to 30MHz.

COLPITTS OSCILLATOR :

In the colpitts oscillator shown below Z_1 and Z_2 are capacitors and Z_3 is an inductor. The resistors R_1, R_2 and ~~R_3~~ R_E provide necessary d.c. 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 Capacitors C_1 and C_2 and an L determines the frequency of the oscillator.

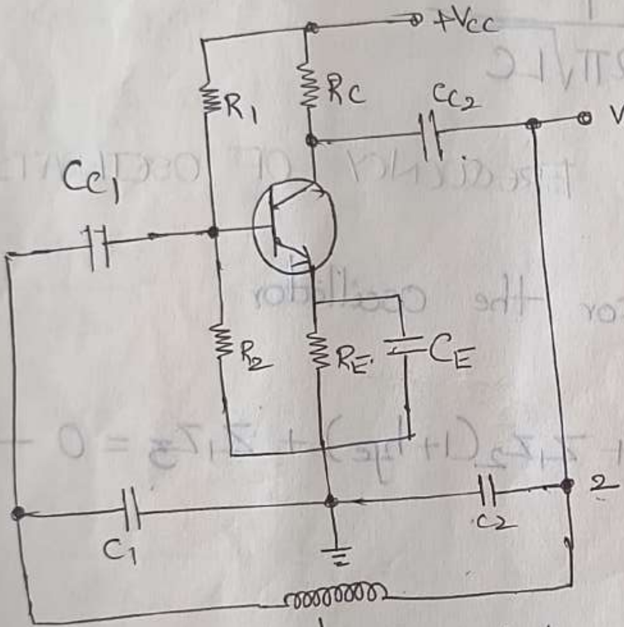


fig : Colpitts oscillator

When the supply voltage $+V_{CC}$ is switched ON, a transient current is produced in the tank circuit Consequently, damped harmonic Oscillations are set up in the circuit. The Oscillatory current in the tank circuit produces a.c. voltages C_1 and C_2. As terminal 3 is earthened it will be at zero potential. If terminal 1 is at a positive potential with respect to 3 at any instant, terminal will be at a negative potential with respect to 3 at any instant, Same intant. Thus the phase difference between the terminals 1 and 2 is always 180°. In the CE mode, the transistor provides the phase difference of 180° betw the input and output. Therefore, the total phase shift is 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. The frequency of oscillation is

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

DERIVATIONS FOR FREQUENCY OF OSCILLATIONS

Gentral equation for the oscillator

$$h_{ie}(Z_1+Z_2+Z_3) + Z_1Z_2(1+h_{fe}) + Z_1Z_3 = 0 \quad \text{—— (1)}$$

$$Z_1 = \frac{1}{j\omega C_1}$$

$$Z_2 = \frac{1}{j\omega C_2}$$

$$Z_3 = j\omega L$$

Substitute Z_1, Z_2, Z_3 in equation (1)

$$h_{ie}\left[-j/\omega C_1 - j/\omega C_2 + j\omega L\right] + \left[-\frac{j}{\omega C_1}\right]\left[-\frac{j}{\omega C_2}\right]\left[1+h_{fe}\right] + \left(\frac{-j}{\omega C_1}\right)\left[j\omega L\right] =$$

$$h_{ie}\left[\frac{-j}{\omega C_1} - \frac{j}{\omega C_2} + j\omega L\right] - \frac{1}{\omega^2 C_1 C_2}(1+h_{fe}) + \frac{L}{C_1} = 0$$

$$h_{ie}\frac{j}{\omega}\left[\omega^2 L - \frac{j}{C_1} - \frac{j}{C_2}\right] +$$

imaginary part is equal to zero

$$h_{ie}\frac{i}{\omega}\left[\omega^2 L-\frac{1}{C_1}-\frac{1}{C_2}\right]=0$$

$$\omega^2 L=\frac{1}{C_1}+\frac{1}{C_2}$$

$$\omega^2 L=\frac{C_1+C_2}{C_1C_2}$$

$$\omega=\sqrt{\frac{C_1+C_2}{C_1C_2L}}$$

$\omega = 2\pi f$

$$f=\frac{1}{2\pi\sqrt{C_{eq}L}}$$

$$C_{eq}=\frac{C_1C_2}{C_1+C_2}$$

$$\omega^2=\frac{C_1+C_2}{C_1C_2L}$$

$$\omega^2=\sqrt{\frac{C_1+C_2}{C_1C_2L}}$$

$$f=\frac{}{2\pi\, C_{eq}L}$$

Real part $=0$

$$-\frac{1}{\omega^2C_1C_2}(1+h_{fe})+\frac{L}{C_1}=0$$

$$\frac{-(1+h_{fe})}{\omega^2C_1C_2}+\frac{L}{C_1}=0$$

$$\boxed{\therefore h_{fe}=\frac{C_2}{C_1}}$$

$$\frac{-(1+h_{fe})}{\omega^2C_1C_2}+\frac{L}{C_1}=0$$

$$\frac{1+h_{fe}}{\omega^2C_1C_2}=\frac{L}{C_1}$$

$$\omega^2L=\left[\frac{\cancel{C_1}+C_2}{C_1\cancel{C_2}}\right]C_2$$

$$1+h_{fe}=\frac{L\omega^2\cancel{C_1}C_2}{\cancel{C_1}}$$

$$1+h_{fe}=\left(\frac{C_1+C_2}{C_1\cancel{C_2}}\right)\cancel{C_2}$$

$$h_{fe}=\frac{C_1+C_2}{C_1}-1$$

$$h_{fe}=\frac{\cancel{C_1}+C_2-\cancel{C_1}}{C_1}$$

$$\Rightarrow \boxed{h_{fe}=\frac{C_2}{C_1}}$$

ADVANTAGES

→ The Colpitts Oscillator can be used in high freq to produce pure sinusoidal waveform because of low impendance paths of the Capacitors at high frequencies.

→ It has wide operation range from 1 to 60 MHz

DISADVANTAGES

→ It is difficult to design

→ It has poor isolation property

→ Because the circuit is Complicated, the Cost of to Construct is high

APPLICATIONS

→ It is used for the development of mobile and rad Communications

→ It has many applications used for the Commercial purpo

→ It is used for generation of sinusoidal output signals with very high frequencies.

→ Used for applications in which undamped and Continu oscillations are desired for functioning.

Clapp Oscillator: To achieve the frequency stability, Colpitts oscillator circuit is slightly modified in practice, called Clapp oscillator. The basic tank circuit with two capacitive reactances and one inductive reactance remains same. But the modification in the tank circuit is that one more capacitor C_3 is introduced in series with the inductance as shown in fig below.

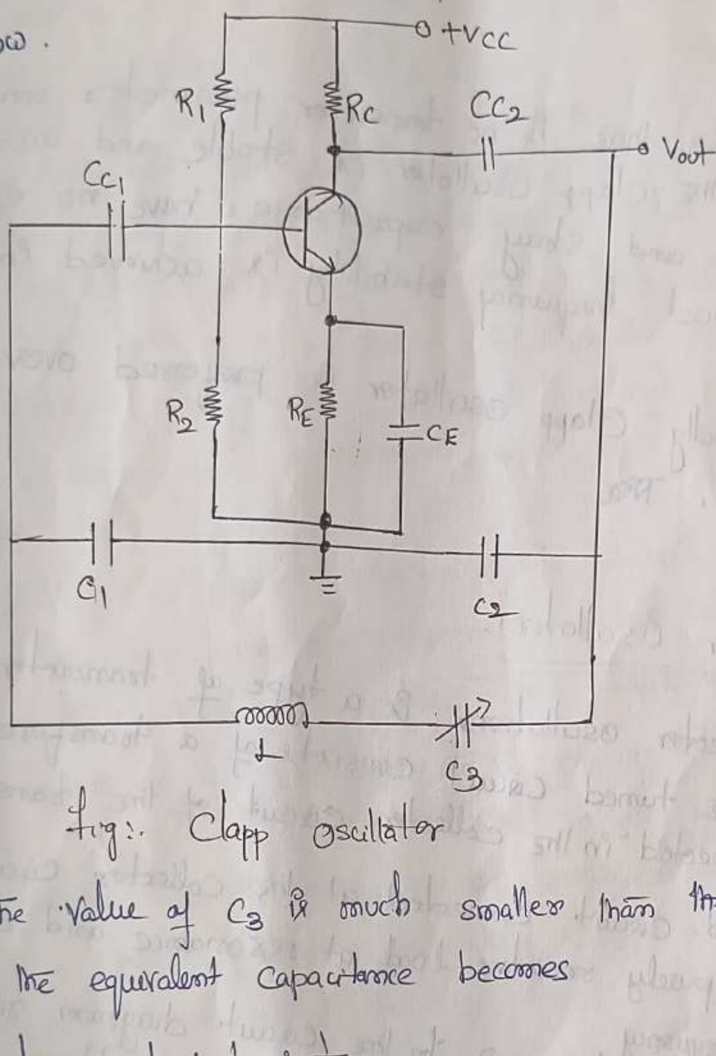


fig: Clapp oscillator

The value of C_3 is much smaller than the values of C_1 & C_2. Now the equivalent capacitance becomes

$$\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

while the oscillator frequency is given by the same expression as $f = \frac{1}{2\pi\sqrt{LC_{eq}}}$

if C_1 and C_2 are neglected then $C_{eq} = C_3$

$$f = \frac{1}{2\pi\sqrt{LC_3}}$$

The frequencies are almost same, hence in practice the C_1, C_2 values are neglected and C_3 is assumed to be C_{eq}.

Hence the frequency is given by $\boxed{f = \frac{1}{2\pi\sqrt{LC_3}}}$

Advantages :-

1. The frequency is stable and accurate
2. The good frequency stability
3. The stray capacitances have no effect on C_3 which decide the frequency
4. keeping C_3 variable, frequency can be varied in the desired range.

→ Now across C_3, there is no transistor parameter and hence the frequency of the clapp oscillator is stable and accurate.

→ The transistor and stray capacitances have no effect on C_3 hence good frequency stability is achieved in Clapp Oscillator.

→ Hence practically Clapp oscillator is preferred over Colpitts oscillator. ~~The~~

Tuned Collector Oscillator :-

Tuned collector oscillators is a type of transistor LC oscillator where the tuned circuit consists of a transformer and capacitor is connected in the collector circuit of the transistor.

The tuned circuit connected at the collector circuit behaves like a purely resistive load at resonance and determines the oscillator frequency.

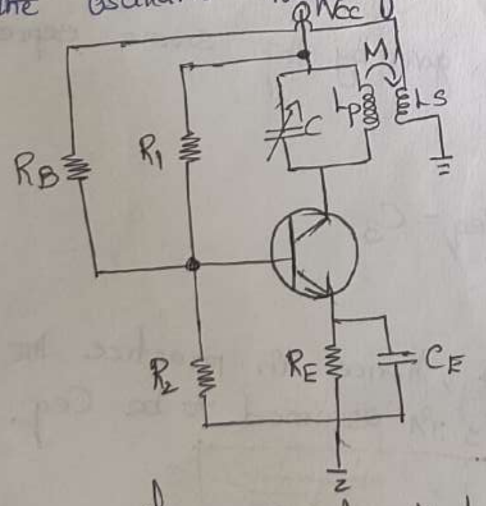


fig: Tuned Collector oscillator

→ In the circuit diagram resistor R_1 and R_2 forms a voltage divider bias for the transistor. R_E is the emitter resistor which is meant for thermal stability.

→ It also limits the collector current of the transistor. C_E is the emitter by pass capacitor.

→ The job of C_E is to by-pass the amplified oscillations

→ If C_E is not there, the amplified AC oscillations will drop across R_E and will add on to the base-emitter voltage (V_{BE}) of the transistor and this will alter the DC biasing conditions.

Working:- When the power supply is switched ON, the transistor starts conducting and the capacitor C_p starts charging. When the capacitor is fully charged, it starts discharging through the primary coil L_p. When the capacitor is fully discharged, the energy stored in the capacitor as electrostatic field will be moved to the inductor as electromagnetic field. Now there will be no more voltage across the capacitor to keep the current through the coil starts to collapse. In order to oppose this the coil (L_p) generates a back emf (by electromagnetic induction) and this back emf charges the capacitor again, then capacitor discharges through the coil and the cycle is repeated. This charging and discharging sets up a series of oscillations in the tank circuit.

The oscillations produced in the tank circuit is fed back to the base of transistor Q_1 by the secondary coil by inductive coupling. The amount of feedback can be adjusted by varying the turns ratio of the transformer. The winding direction of the secondary coil (L_2) is in such a way that the voltage across it will be 180° phase opposite to that of the voltage across primary (L_p). Thus the feedback circuit produces a phase shift of 180° and the transistor alone produces a phase shift of another 180°. As a result total phase shift of 360° obtained b/w i/p and o/p.

The collector current of the transistor compensates the energy lost in the tank circuit. This is done by taking a small amount of voltage from the tank circuit, amplifying it and applying it back to the tank circuit. Capacitor C_1 can be made variable in variable frequency applications.

The frequency of oscillations of the tank circuit can be expressed by

$$\boxed{f_0 = \frac{1}{2\pi\sqrt{L_p C}}}$$

Crystal Oscillators :-

The crystals are either naturally occuring or synthetically manufactured, exhibiting the piezoelectric effect.

Piezoelectric effect :- The piezoelectric effect means under the influence of the mechanical pressure, the voltage gets generated across the opposite faces of the crystal. If the mechanical force is applied in such a way to force the crystal to vibrate, the a.c voltage gets generated across it. Conversely, if the crystal is subjected to a.c voltage, it vibrates causing mechanical distortion in the crystal shape.

Every crystal has its own resonating frequency depending on its cut. So under the influence of the mechanical vibration, the crystal generates an electrical signal of very constant frequency. The crystal has a greater stability in holding the constant frequency.

The main substances exhibiting the piezoelectric eff[ect] quartz, Rochelle salt, and tourmalene. Rochelle salts ha[ve] greatest piezoelectric activity.

Constructional Details :-

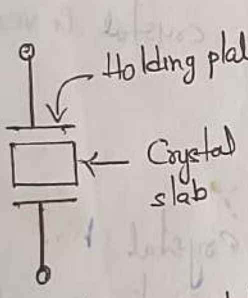


fig :- Symbolic representation of a crystal

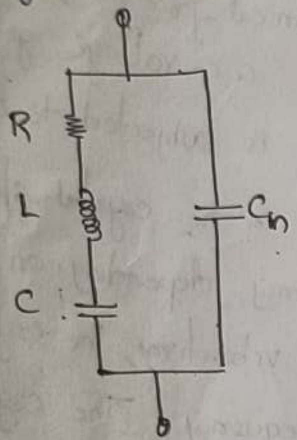
The natural shape of a quartz is a hexagonal prism. But for its practical use, it is cut to the rectangular slab. This slab is then mounted between the two metal plates. The metal plates are called holding plates, as they hold the crystal slab in between them.

A.C Equivalent Circuit :-

When the crystal is not vibrating, it is equivalent to a capacitance due to the mechanical mounting of the crystal. Such a capacitance existing due to the two metal plates seperated by a dielectric like crystal slab, is called mounting capacitance denoted as C_M or C

A.C. Equivalent Circuit :-

When the crystal is not vibrating, it is equivalent to a capacitance due to the mechanical mounting of the crystal. Such a capacitance existing due to the two metal plates seperated by a dielectric like crystal slab, is called mounting capacitance denoted as C_M or C'



When it is vibrating, there are internal frictional losses which are denoted by a resistance R. While the mass of the crystal, which is indication of its inertia is represented by an inductance L. In vibrating condition, it is having some stiffness, which is represented by a capacitor C. The mounting capacitance is a shunt capacitance. And hence the overall equivalent circuit of a crystal can be shown

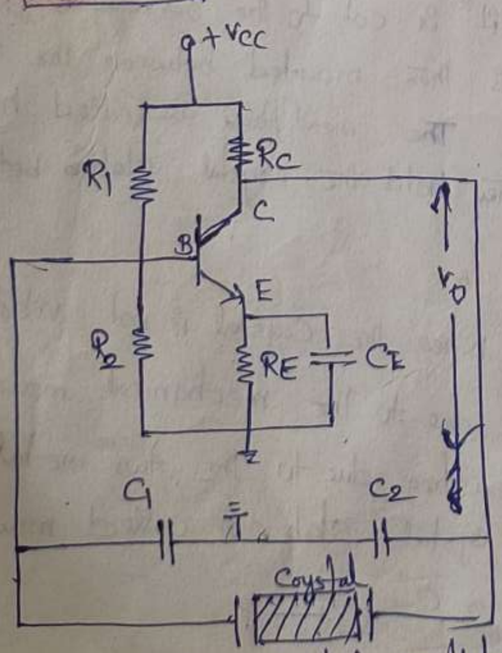
RLC forms a resonating circuit. The expression for the resonating frequency f_r is

$$\boxed{f_r = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{Q^2}{1+Q^2}}}$$

Quality factor of crystal

$$\boxed{Q = \frac{\omega L}{R}}$$

The Q factor of the crystal is very high.



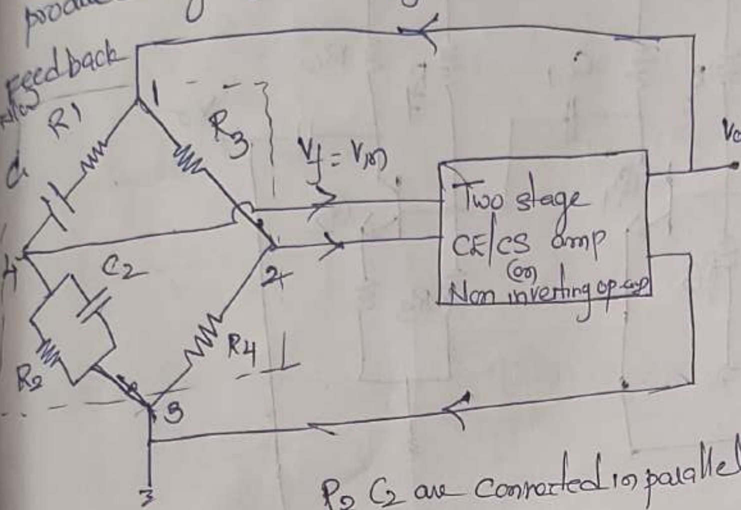
Ckt diagram of a crystal oscillator using BJT

→ Crystal has series resonance frequency $f = \frac{1}{2\pi\sqrt{LC}}$

→ Parallel resonance frequency of crystal, $f = \frac{1}{2\pi\sqrt{L\left[\frac{C_s C_p}{C_s + C_p}\right]}}$

Wien Bridge Oscillator:

- It is a type of RC oscillator which consists of two stage amplifier circuit and a feedback network [frequency determining network]
- Feedback n/w is a balanced wien bridge circuit which does not produce any phase shift.



Adv:- The frequency can be varied by varying the 'C' of the feedback n/w

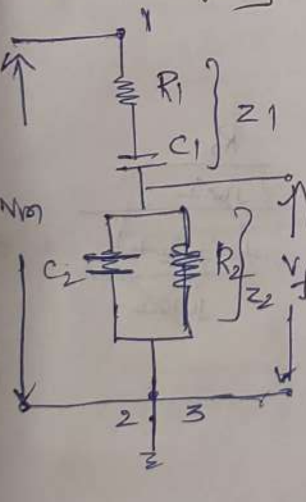
R_2 C_2 are connected in parallel arm)

(R_1C_1) One series arm, one parallel arm

→ The feedback n/w consists of R_1C_1 in series, R_2C_2 in shunt and R_3 & R_4 as voltage divider circuit.

→ The two arms of the ween bridge [Z_1 & Z_2) are called as frequency sensitive arms, bcz it is used to determine the frequency range.

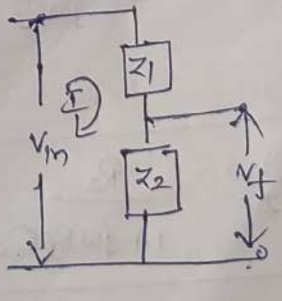
[R_1C_1 & $R_2 \| C_2$ are connected in series]



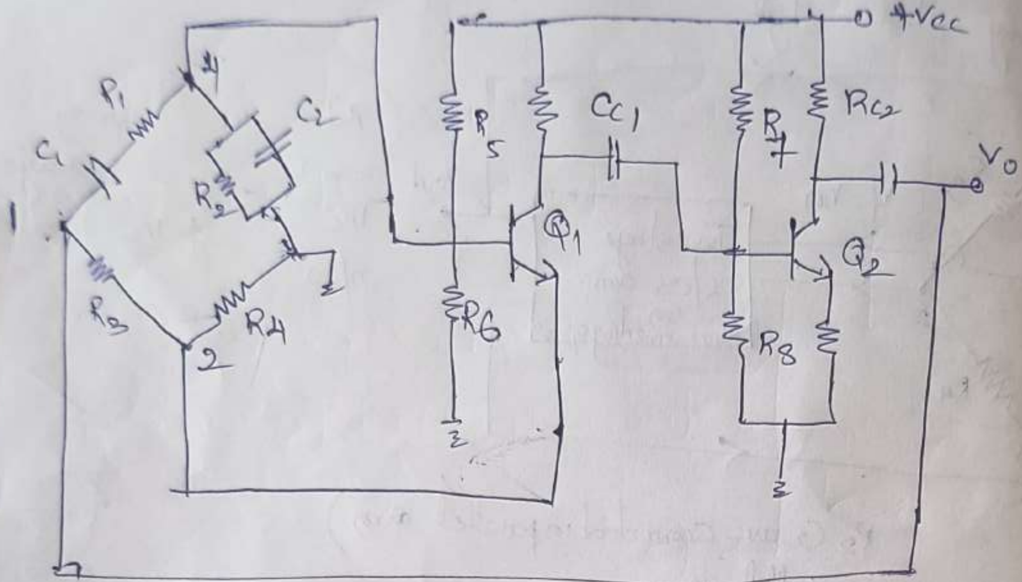
$Z_1 = R_1 \cdot \frac{1}{j\omega C_1} =$

$\cancel{Z_2}$

$\Longrightarrow$



Lead Lag N/w :- The Wien



The series impedance is given as

$$Z_1 = R_1 + \frac{1}{j\omega C_1} \qquad X_C = \frac{1}{j\omega C}$$

$$\boxed{Z_1 = \frac{j\omega R_1 C_1 + 1}{j\omega C_1}} \longrightarrow (1)$$

The parallel impedance is given as

$$Z_2 = R_2 \parallel \frac{1}{j\omega C_2} = \frac{\frac{R_2}{j\omega C_2}}{R_2 + \frac{1}{j\omega C_2}} = \frac{\frac{R_2}{\cancel{j\omega C_2}}}{\frac{j\omega C_2 R_2 + 1}{\cancel{j\omega C_2}}}$$

$$\boxed{Z_2 = \frac{R_2}{1 + j\omega R_2 C_2}} \longrightarrow (2)$$

The current in the feedback n/w is

$$I = \frac{V_{in}}{Z_1 + Z_2} \longrightarrow (3)$$

The feedback voltage is given as

$$V_f = I Z_2 \longrightarrow (4)$$

$$V_f = \frac{V_{in}}{Z_1 + Z_2} \times Z_2$$

The feedback gain of the Wein bridge oscillator,

$$\beta = \frac{V_f}{V_{in}}$$

$$\beta = \frac{\frac{V_{in}}{Z_1+Z_2} \times Z_2}{V_{in}} \Rightarrow \frac{\cancel{V_{in}} Z_2}{Z_1 + Z_2} \times \frac{1}{\cancel{V_{in}}} = \frac{Z_2}{Z_1 + Z_2}$$

$$\beta = \frac{\frac{R_2}{1 + j\omega R_2 C_2}}{\left(\frac{1 + j\omega R_1 C_1}{j\omega C_1}\right) + \left[\frac{R_2}{1 + j\omega R_2 C_2}\right]}$$

further simplify the denominator by taking LCM

$$\beta = \frac{\frac{R_2}{1 + j\omega R_2 C_2}}{\frac{(1 + j\omega R_1 C_1)(1 + j\omega R_2 C_2) + R_2 (j\omega C_1)}{(j\omega C_1)(1 + j\omega R_2 C_2)}}$$

$$\beta = \frac{R_2 \, j\omega C_1}{(1 + j\omega R_1 C_1)(1 + j\omega R_2 C_2) + R_2 (j\omega C_1)}$$

$\therefore j^2 = 1$

$$\beta = \frac{R_2 \, j\omega C_1}{1 + j\omega R_2 C_2 + j\omega R_1 C_1 - \omega^2 R_1 C_1 R_2 C_2 + R_2 j\omega C_1}$$

group real term

$$\beta = \frac{j\omega R_2 C_1}{1 - \omega^2 R_1 C_1 R_2 C_2 + j\omega [R_1 C_1 + R_2 C_2 + R_2 C_1]}$$

The above equation can be simplified by rationalizing the expression

$$= \frac{j\omega R_2 C_1}{[1-\omega^2 R_1 R_2 C_1 C_2] + j\omega[R_1C_1+R_2C_2+R_2C_1]} \times \frac{[1-\omega^2 R_1 R_2 C_1 C_2] - j\omega[R_1C_1 + R_2 C_2 + R_2 C_1]}{[1-\omega^2 R_1 R_2 C_1 C_2] - j\omega[R_1 C_1 + R_2 C_2 + R_2 C_1]}$$

$$\beta = \frac{j\omega R_2 C_1[1-\omega^2 R_1 R_2 C_1 C_2] - j^2\omega^2 R_2 C_1 [R_1C_1+R_2C_2+R_2C_1]}{[1-\omega^2 R_1 R_2 C_1 C_2]^2 + \omega^2[R_1C_1+R_2C_2+R_2C_1]^2}$$

$$\boxed{\beta = \frac{\omega^2 R_2 C_1 [R_1C_1 + R_2C_2 + R_2C_1] + j\omega R_2 C_1 [1-\omega^2 R_1 R_2 C_1 C_2]}{[1-\omega^2 R_1 R_2 C_1 C_2]^2 + \omega^2[R_1C_1+R_2C_2+R_2C_1]^2}} \quad (5)$$

→ At resonant condition $X_L = X_C$ i.e there is no imaginary term.

→ In order to find the frequency of oscillation the imaginary term of equ (5) is equal to zero

$$\rightarrow \frac{\omega R_2 C_1[1-\omega^2 R_1 R_2 C_1 C_2]}{[1-\omega^2 R_1 R_2 C_1 C_2]^2 + \omega^2[R_1C_1+R_2C_2+R_2C_1]^2} = 0 \times ? = 0$$

$$\omega R_2 C_1 [1-\omega^2 R_1 R_2 C_1 C_2] = 0$$

$$[1-\omega^2 R_1 R_2 C_1 C_2] = \frac{0}{\omega R_2 C_1}$$

$$1-\omega^2 R_1 R_2 C_1 C_2 = 0$$

$$\omega^2 R_1 R_2 C_1 C_2 = 1$$

$$\omega^2 = \frac{1}{R_1 R_2 C_1 C_2} = \frac{1}{R^2 C^2}$$

$$\omega = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$f = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

under the matched/balanced condition

$$R_1 = R_2 = R \;\&\; C_1 = C_2 = C$$

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

→ Gain of the feedback network

Equate the real part of equation (5) to gain

$$\beta = \frac{\omega^2 R_2 C_1 [R_1 C_1 + R_2 C_2 + R_2 C_1]}{[1 - \omega^2 R_1 R_2 C_1 C_2]^2 + \omega^2 [R_1 C_1 + R_2 C_2 + R_2 C_1]^2}$$

if $R_1 = R_2 = R$ & $C_1 = C_2 = C$ then

$$\beta = \frac{\omega^2 RC [RC + RC + RC]}{[1 - \omega^2 R^2 C^2]^2 + \omega^2 (3RC)^2} = \frac{3\omega^2 R^2 C^2}{[1 - \omega^2 R^2 C^2]^2 + 9\omega^2 R^2 C^2}$$

Sub equ (6) in the above equation

i.e. $\omega^2 = \frac{1}{R^2 C^2}$

$$\beta = \frac{3 \frac{1}{R^2C^2} \times \cancel{R^2C^2}}{\left[1 - \frac{1}{\cancel{R^2C^2}} \cancel{R^2C^2}\right]^2 + 9 \frac{1}{\cancel{R^2C^2}} \cancel{R^2C^2}}$$

$$\beta = 3/9 = 1/3$$

$$\boxed{\beta = 1/3}$$

$|A\beta| = 1$

$|A| = 1/\beta = \frac{1}{1/3} = 3$

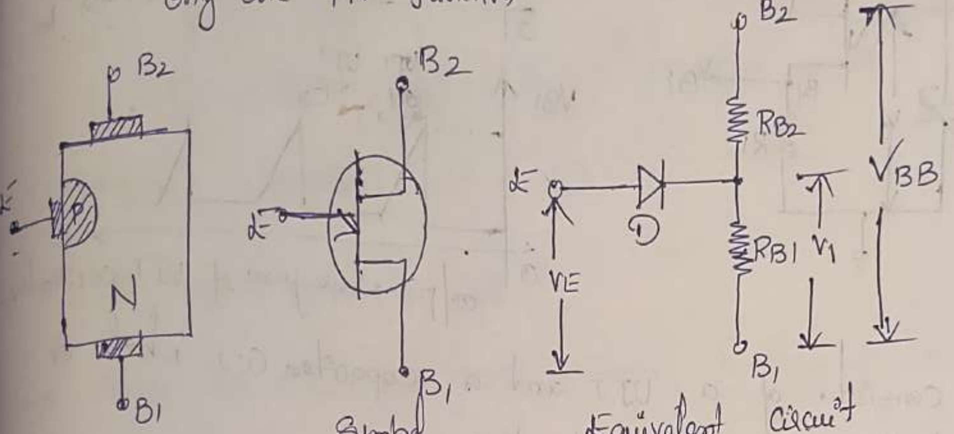
$\boxed{|A| \geqslant 3}$

UJT Oscillator

UJT [Uni Junction Transistor]

↳ UJT is a 3 terminal semiconductor switching device.

↳ It has 3 terminals [Base 1, Base 2 and Emitter] and only one PN Junction



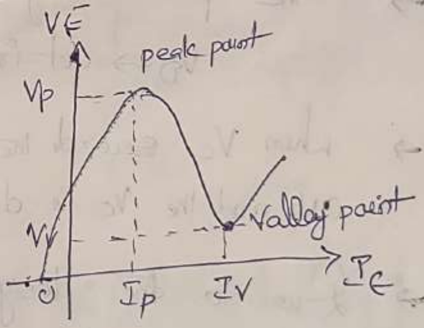
→ It consists of a lightly doped N-type Si material with a heavily doped p-type material closer to B2

* Interbase resistance : $R_{BB} = R_{B1} + R_{B2}$

* Intrinsic stand off ratio :

$$\boxed{\eta = \frac{R_{B1}}{R_{B1} + R_{B2}}}$$

$\eta = 0.56$ to 0.75



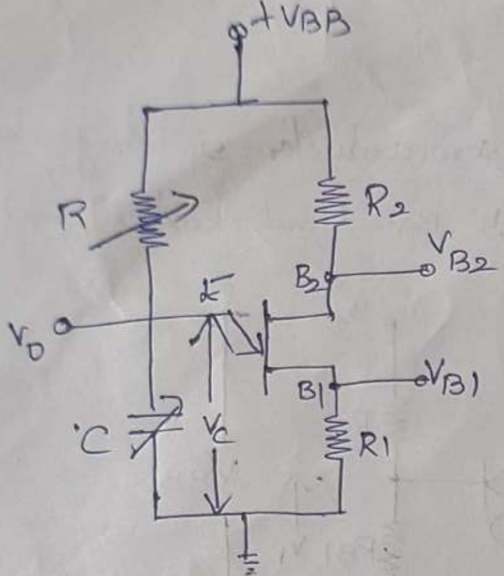
* Voltage across RB1 is : $V_1 = \eta V_{BB}$

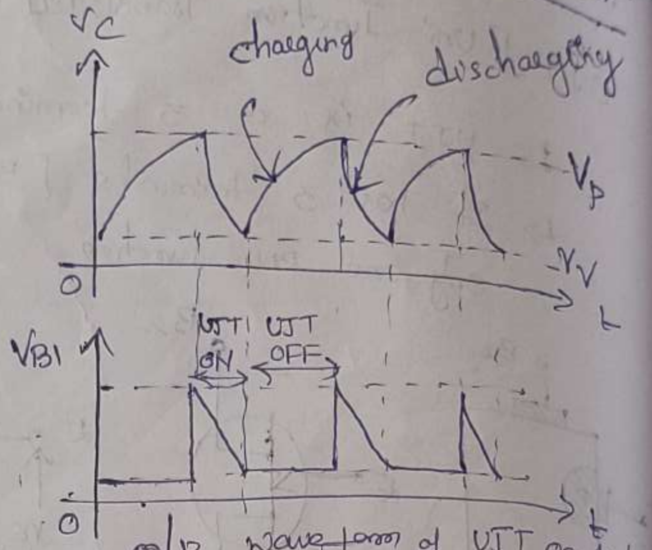
(UJT turns on when $V_E > V_1 + V_D$)

Applications of UJT

→ Sawtooth wave form generator
→ pulse generator
→ Switching circuits
→ Time and phase control circuits.

UJT oscillator (or) UJT Relaxation oscillator (or) UJT Sawtooth Generator.





o/p waveform of UJT oscillator.

→ It consists of a UJT and a capacitor (C) which is charged through 'R' towards VBB.

→ When the supply voltage VBB is switched 'ON', the voltage across the capacitor increases exponentially

→ The peak voltage of UJT : $\boxed{V_P = \eta V_{BB} + V_D}$ → ①

V_D → cut-in voltage of diode.

→ When V_C exceeds the peak voltage (V_P), the UJT is switched ON and the V_C is discharged through $E - B_1$ and R_1

→ Due to the design of R_1, the discharge of V_C produces a pulse across R_1

→ When V_E falls below the valley voltage, the UJT is switched OFF. Then the capacitor starts charging again.

→ The frequency of the output waveform can be changed by changing the values of R and C.

Time constant $\boxed{\tau = RC}$

Frequency of Oscillation:

~~In~~ UJT oscillator, the charging equation of the capacitor is given as

$$V_C(t) = V_V + V_{BB}\left[1 - e^{-t/RC}\right]$$

At $t = T$

$$V_C(t) = V_P$$

$$V_P = V_V + V_{BB}\left[1 - e^{-T/RC}\right] \rightarrow (2)$$

compare equations (1) & (2)

$$\eta V_{BB} + V_D = V_V + V_{BB}\left[1 - e^{-T/RC}\right]$$

→ Since V_D & V_V are small compared to V_{BB} so neglect the V_D & V_V

$$\eta \cancel{V_{BB}} = \cancel{V_{BB}}\left[1 - e^{-T/RC}\right]$$

$$\eta = 1 - e^{-T/RC}$$

$$e^{-T/RC} = 1 - \eta$$

$$\boxed{T = RC \ln\left[\frac{1}{1-\eta}\right]}$$

[To remove exponential term use logarithmic

$$\frac{1}{e^{-T/RC}} = \frac{1}{1-\eta}$$

$$e^{+T/RC} = \frac{1}{1-\eta}$$

$$\ln\left(e^{T/RC}\right] = \ln\left[\frac{1}{1-\eta}\right]$$

$$\frac{T}{RC} = \ln\left[\frac{1}{1-\eta}\right]$$

→ The frequency of oscillation is

$$\boxed{f = \frac{1}{T} = \frac{1}{RC\ln\left[\frac{1}{1-\eta}\right]}}$$

Frequency stability of Oscillator

For an oscillator, the frequency of the oscillations must remain Constant. The analysis of the dependence of the oscillating frequency on the various factors like stray Capacitance, temperature etc. is called as the frequency stability analysis.

The measure of ability of an oscillator to maintain the desired frequency as precisely as possible for as long a time as possible is called frequency stability of an Oscillator.

In a transistorised Colpitts oscillator or Hartley Oscillator, the base-Collector junction is reverse biased and there exists an internal Capacitine which is dominant at high frequencies. This Capacitance affects the value of Capacitine in the tank circuit and hence the Oscillating frequency.

Similarly the transistor parameters are temperatures Sensitive. Hence as temperature changes, the Oscillating frequency also changes and no longer remains stable. Hence practically the circuits Cannot provide stable frequency.

Factors affecting the frequency stability:

In general following are the factors which affect the frequency stability of an oscillator:

→ Due to the changes in temperature, the values of the Components of tank circuit get affected. So changes in the values of inductors and Capacitors due to the changes in the temperatures is the main Cause due to which frequency does not remain stable

→ Due to the changes in temperatures, the parameters of the active device used like BJT, FET get affected which intum affect the frequency.

→ The variation is the power supply is another factor affecting the frequency

→ The changes in the atmospheric Conditions, aging and unstable transistor parameters affect the frequency

→ The changes in the load Connected, affect the effective resistance of the tank circuit.

→ The Capacitive effect in transistor and stray Capacitances, affect the Capacitive Capacitance of the tank Circuit and hence the frequency

The variation of frequency with temperatures is given by the factor denoted as S.

$$\boxed{S_{\omega,T} = \frac{\Delta\omega/\omega_r}{\Delta T/T_r} \text{ parts per million per } ^\circ C} \quad \cdots (1)$$

where ω_r = Desired frequency

T_r = Operating temperature

$\Delta\omega$ = change in frequency

ΔT = change in temperature.

The frequency stability is defined as,

$$\boxed{S_\omega = \frac{d\theta}{d\omega}}$$

where $d\theta$ = Phase shift introduced for a small frequency change in desired frequency f_r.

Key Point: Larger the value of $d\theta/d\omega$, more stable is the Oscillator.

The frequency stability can be improved by the following modifications:

1) Enclosing the circuit in a constant temperature chamber.
2) Maintaining constant voltage by using the zener diodes.
3) The load effect is reduced by coupling the oscillator to the load loosely or with the help of a circuit having high input impedance and low output impedence

Need of Amplitude Limiting

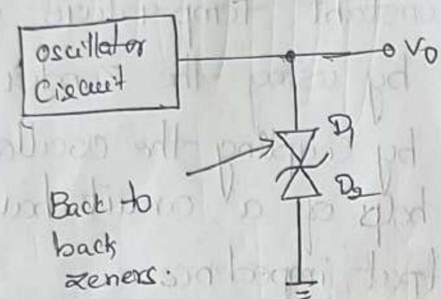
The oscillator output amplitude, if not limited, attains the extreme levels of saturation. This can cause the distortion in the output waveform due to either clipping off some part of the waveform or it may drive amplifier into saturation. The circuit may stop functioning as oscillator.

Keypoint: Thus to get undistored sinusoidal output with a stable oscillator circuit, it is necessary to limit the output amplitude in oscillators.

The circuits used to in the oscillators for this purpose are called amplitude limiting circuits. These circuits make the oscillations damped if amplitude increases beyond limit and ensures that the oscillations will not sustain in such undesirable conditions

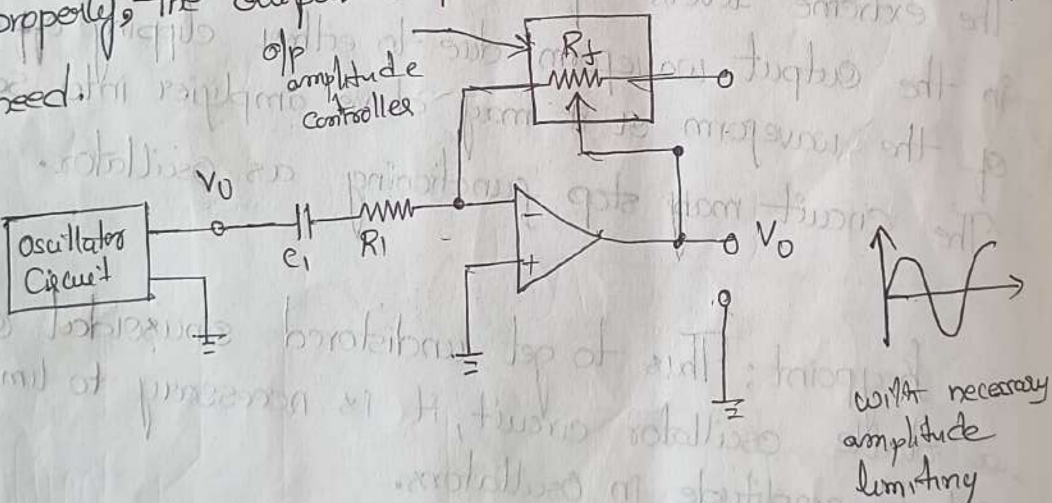
One simple way of limiting the oscillator output amplitude is to provide back to back zener diodes Connected at the output terminals. This is shown below. The voltage across back to back zeners remains Constant and limits the value of output amplitude. The output amplitude can be set by selecting proper zener diodes

fig:- limiting o/p amplitude of oscillator



→ The output amplitude adjustment can also be achieved by an operational amplifier circuit with variable feedback resistor as shown in below fig.

→ The gain of the op-amp circuit is $\left|\frac{R_f}{R_1}\right|$. Thus when $R_f = R_1$ the output amplitude remains same as oscillator output expect phase reversal. By Controlling the value of R_f properly, the output amplitude can be Controlled as per the need.



Hit pls// bit.ly/neot-apply

UNIT-

FEEDBACK AMPLIFIERS

Introduction:- In this chapter we introduce the concept of feedback and show how to modify the characteristics of an amplifier by combining a portion of the output signal with the external signal. Many advantages are to be gained from the use of negative feedback, and these are studied. It is possible for the feedback to be positive, and the circuit may then oscillate.

Classification of Amplifiers

→ Before proceeding with the concept of feedback, it is useful to classify amplifiers into four broad categories.

1. Voltage amplifier
2. Current amplifier
3. Transconductance amplifier → converts an i/p vltg into an o/p current
4. Transresistance amplifier → converts an i/p of current to an o/p of vlg

This classification is based on the magnitudes of the input and output impedances of an amplifier relative to the source and load impedances respectively.

→ Based on configuration

(a) Common base Amplifier
(b) Common Emitter amplifier
(c) common Collector amplifier

→ Based on the period of Conduction.

(a) Class-A
(b) class-B
(c) class-C
(d) Class-'D'
(e) Class AB
(f) Class S
(g) FET

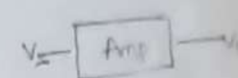
→ Based on the frequency

(a) High frequency amplifiers
(b) Low frequency amplifiers
(c) RF amplifiers
(d) Microwave amplifiers.

key points

1) R_{in} = very high, ideal = ∞
2) R_0 = very low ideal = 0

$V_i = \frac{R_i}{R_i + R_S} \times V_S$



if R_s is small

$V_0 = \frac{R_L}{R_0 + R_L} A_v V_s$

If $R_i >> R_S$ then $V_i = V_S$ & if $R_0 << R_L$ then $V_0 = A_v V_i = A_v V_S$ hence $A_v = V_0/V_i$ with $R_L = \infty$ represents the open circuit voltage gain

Voltage Amplifiers:

Figure 'a' shows Thevenin's equivalent circuit of a two-port network which represents an amplifier.

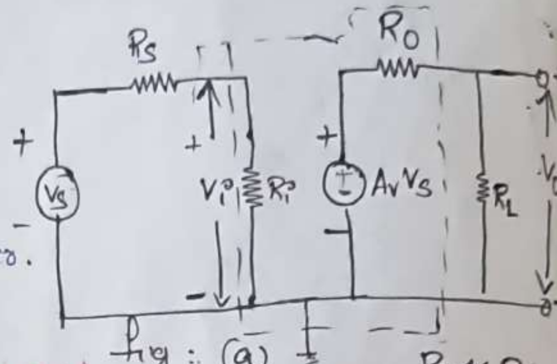


fig : (a)

$R_0 << R_L$

If the amplifier input resistance R_i is large compared with the source resistance R_S, then $V_i \simeq V_S$. $R_i >> R_S$ If the external load resistance R_L is large compared with the output resistance R_0 of the amplifier, then $V_0 \simeq A_v V_i \simeq A_v V_S$. This amplifier provides a voltage output proportional to the voltage input, and the proportionality factor is independent of the magnitudes of the source and load resistances. Such a circuit is called a voltage amplifier.

An ideal voltage amplifier must have infinite input resistance R_i and zero output resistance R_0. The symbol A_v in fig 'a' represents V_0/V_i with $R_L = \infty$, and hence represents the open-circuit voltage amplification or gain.

If $R_i << R_S$ then $I_i \simeq I_S$ & $R_0 >> R_L$ then $I_0 = A_i I_i \simeq A_i I_S$ hence $A_i = \frac{I_0}{I_i}$ with $R_L = 0$, short circuit current gain

$R_i << R_S$

$R_0 >> R_L$

$i_0 = i_L$

Current amplifier:

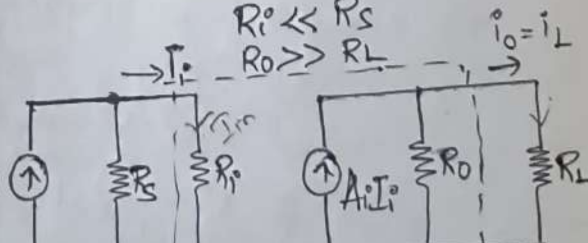


fig (b)

An ideal current amplifier is defined as an amplifier which provides an output current proportional to the signal current, and the proportionality factor is independent of R_S and R_L.

An ideal current amplifier must have zero input resistance R_i and infinite output resistance R_o. In practice, the amplifier has low input resistance and high output resistance. It drives a low resistance load ($R_o >> R_L$), and is driven by a high resistance source ($R_i << R_s$).

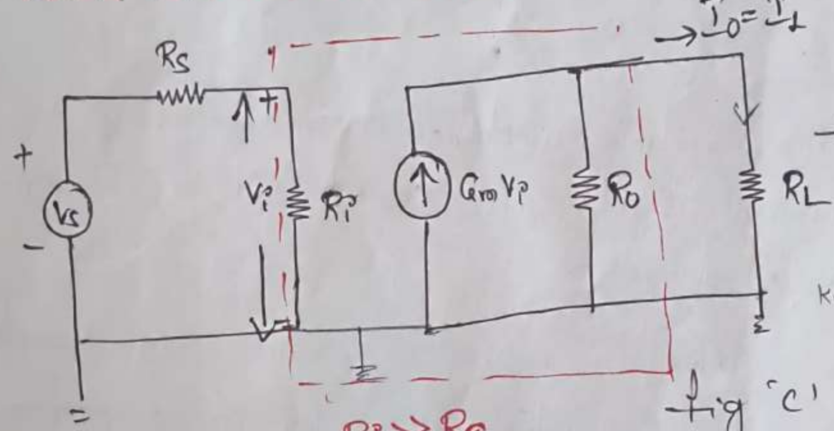
Fig (b) shows Norton's equivalent circuit of a current amplifier. Note that $A_i = I_L / I_i$ with $R_L = 0$, representing the short circuit current amplification, or gain. We see that if $R_i << R_s$, $I_i \approx I_s$ and if $R_o >> R_L$, $I_L \simeq A_i I_i \simeq A_i I_s$. Hence the output current is proportional to the signal current. Key points: 1) R_{in} = very low ideal = 0

2) R_o = very high ideal = ∞

$I_{in} = \frac{R_s}{R_s + R_i} \times I_s$

$I_L = \frac{R_o}{R_L + R_o} \times A_i I_i$

Transconductance Amplifier:



$V_{in} = \frac{R_i}{R_i + R_s} \times V_s \Rightarrow V_{in} \simeq V_s$

$I_L = \frac{R_o}{R_o + R_L} \times G_m V_i \quad I_L = G_m V_i$

Key → R_i = very large = ideal ∞

R_o = very large = ideal ∞

$R_i >> R_S$

$R_o >> R_L$

fig 'c'

fig (c) A transconductance amplifier is represented by a Thevenin's equivalent in the input circuit and a Norton's equivalent in the output circuit.

The ideal transconductance amplifier supplies an output current which is proportional to the signal voltage, independently of the magnitudes of R_s and R_L. This amplifier must have an infinite input resistance R_i and infinite output resistance R_o.

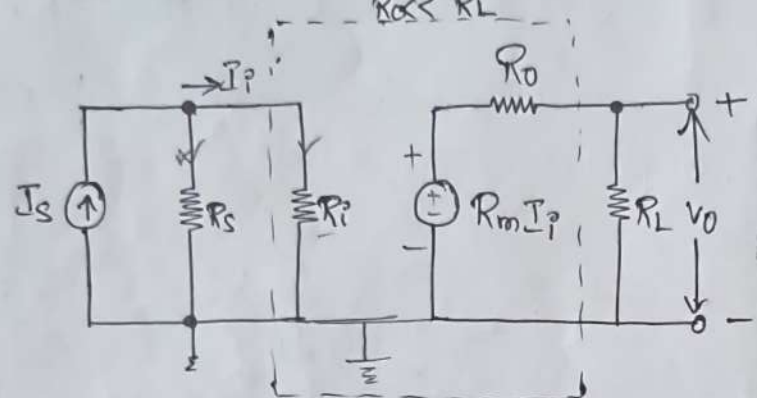
if $R_i >> R_s$ then $V_i = V_s$ & if $R_o >> R_L$ then $I_o = G_m V_i = G_m V_s$ with $R_L = 0$ → represents the short circuit mutual or transconductance hence $G_m = \frac{I_o}{V_s}$

A practical transconductance amplifier has a large input resistance ($R_i >> R_s$) and hence must be driven by a low-resistance source. It presents a high output resistance ($R_o >> R_L$) and hence drives a low-resistance load. The equivalent circuit of a transconductance amplifier is shown fig 'c'

TRANSRESISTANCE AMPLIFIER

$R_i << R_s$

$R_o << R_L$



$I_{in} = \frac{R_s}{R_s+R_i} \times I_s \Rightarrow I_{in} \simeq I_s$

$V_0 = \frac{R_L}{R_0+R_L} \times R_m I_i \Rightarrow V_0 = R_m I_i$

key → R_{in} very low → ideal $0\,\Omega$

→ R_0 very low → ideal $0\,\Omega$

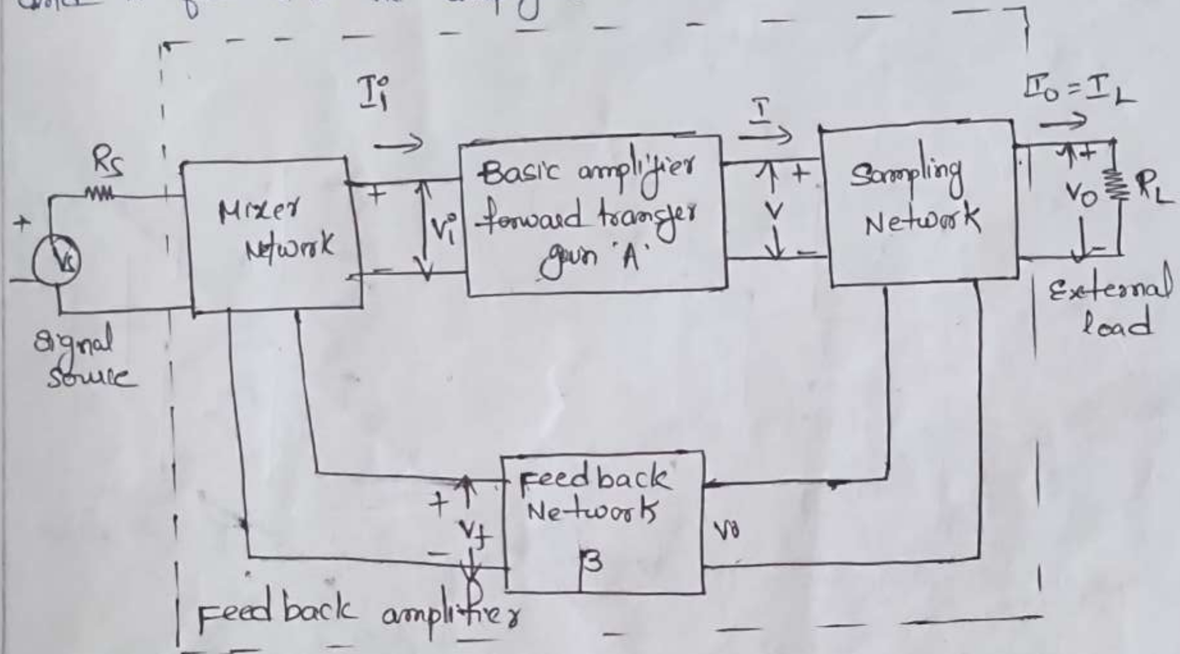
fig (d) Transresistance amplifier is represented by a Norton's equivalent in its input circuit and a Thevenin's equivalent in its output.

Fig (d) shows the equivalent circuit of an amplifier which ideally supplies an output voltage V_0 in proportion to the signal current I_s independently of R_s and R_L. This amplifier is called a transresistance amplifier. For a practical transresistance amplifier we must have $R_i << R_s$ and $R_o << R_L$. Hence the input and output resistances are low relative to the source and load resistances. From fig 'd' we see that if $R_s >> R_i$, $I_i \simeq I_s$ and if $R_o << R_L$, $V_o \simeq R_m I_i$, $V_0 \simeq R_m I_i \equiv R_m I_s$. Note that $R_m \cong V_0 / I_i$ with $R_L = \infty$. In other words, R_m is the open circuit mutual (or) transfer resistance.

→ $R_i << R_s$ then $I_{in} = I_s$, if $R_L >> R_0$ then $V_0 = R_m I_s = R_m I_{in}$, $R_m = \frac{V_0}{I_i}$

The Feedback Concept:-

In the preceding section we summarize the four basic amplifier types. In each one of these circuits we may sample the output voltage or current by means of a suitable sampling network and apply this signal to the input through a feedback two-port network, as shown in fig below. At the input the feedback signal is combined with the external (source) signal through a mixer network and is fed into the amplifier



Feedback Network:- This block is usually a passive two-port network which may contains resistors, capacitors and inductors. Very often it is simply a resistive configuration. it provides reduced portion of the output as

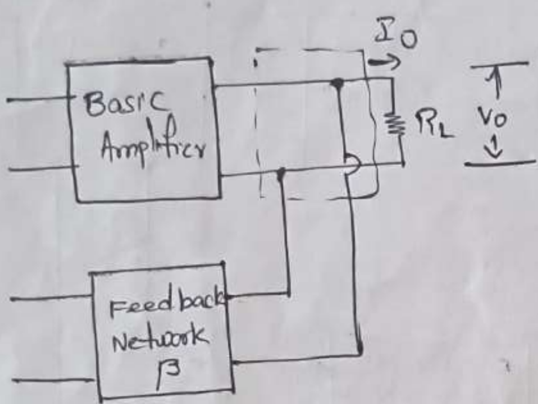
Sampling Network:- Several sampling blocks are shown in fig. feedback signal to the input mixer network. It is given as

$$\boxed{V_f = \beta V_o}$$

where β is a feedback factor or feedback ratio. The symbol β used in feedback circuits represents feed back factor which always lies between 0 and 1. It is totally different from β symbol used to represent current gain in common emitter amplifier, which is greater than 1.

Sampling Network!:

There are two ways to sample the output, according to the sampling parameter, either voltage or current. The output voltage is sampled by connecting the feedback network in shunt across the output, as shown in the above below fig(a) This type of connection is referred to as voltage or node sampling. The output current is sampled by connecting the feedback network in series with the output as shown in the fig (b). This type of connection is referred to as current or loop sampling.



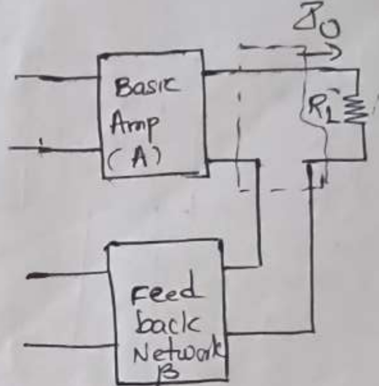


fig (a) Sampling the output voltage   fig(b) Sampling the output current.

Mixing Networks-

Like sampling, there are two ways of mixing feedback signal with the input signal. There are; series input connection and shunt input connection. The fig (a) and (b) show the simple and very common series (loop) input and shunt (node) input connections, respectively.

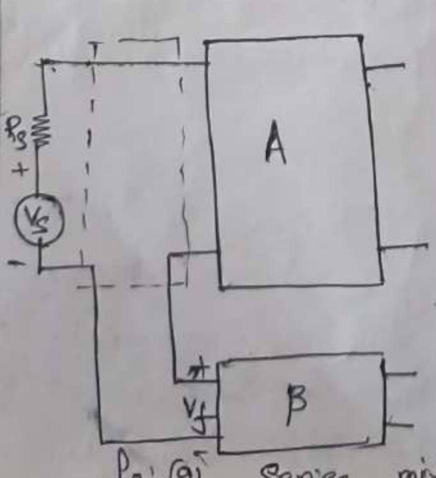


fig: (a) Series mixing

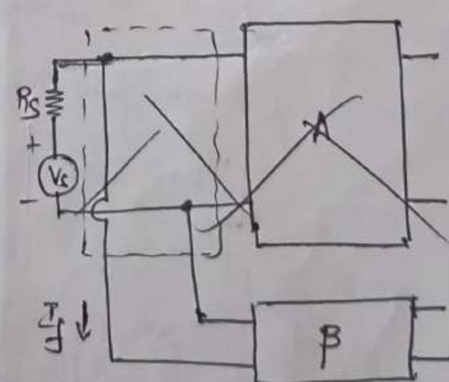


fig: (b)

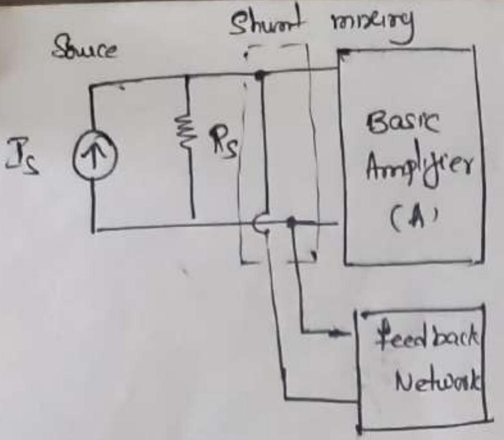


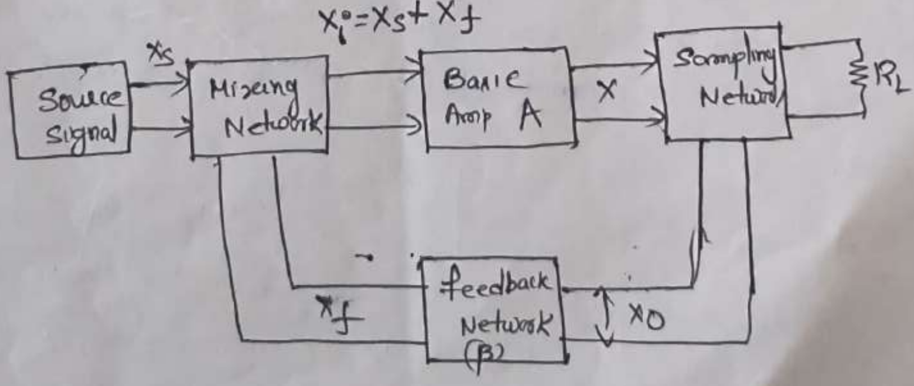
fig (b) shunt mixing.

Types of feed back:-

1. Positive feed back (or) direct (or) regenerative
2. Negative feed back.

1. Positive feed back:- if the feed back signal X_f is in phase with input signal X_s, then the net effect of the feed back will increase the input signal given to the amplifier, i.e $X_i = X_s + X_f$. Hence, the input voltage applied to the basic amplifier is increased there by increasing X_o exponentially. This type of feedback is said to be positive (or) regenerative feedback.

Gain of the feedback amplifier



Gain of amplifier $A = \frac{x_o}{x_i} \rightarrow x_o = A x_i$

Feed back factor $\beta = \frac{x_f}{x_o} \rightarrow x_f = \beta x_o$

$$x_o = A x_i$$

$$x_o = A(x_s + x_f) \qquad \because x_i = x_s + x_f$$

$$x_o = A x_s + A x_f$$

$$x_o = A x_s + A\beta x_o$$

$$x_o - A\beta x_o = A x_s$$

$$x_o[1 - A\beta] = A x_s$$

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

Here A is open loop gain of amplifier

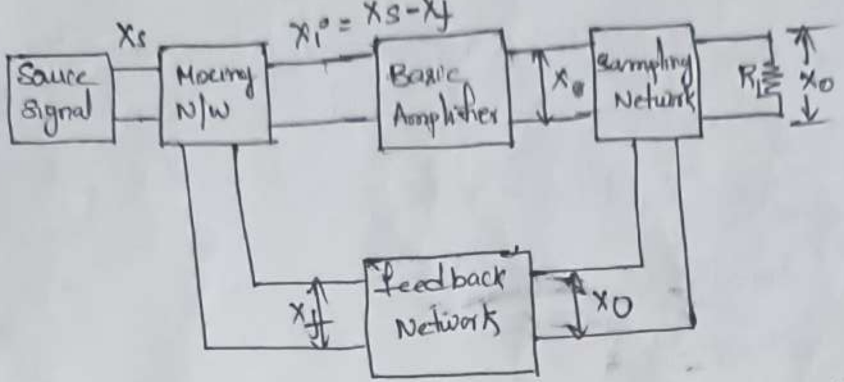
A_f is gain of feed back amplifier.

Here $A_f > A$. The product of the open loop gain and the feedback factor is called the loop gain. i.e loop gain $= A\beta$ if $|A\beta| = 1$ then $A_f = \infty$. Hence the gain of the amplifier with positive feed back is infinite and the amplifier gives an a.c output without a.c input signal. Thus the amplifier acts as an oscillator.

The positive feedback increases the instability of an amplifier, reduces the band width and increases the distortion and noise. The property of the positive feedback is utilized in oscillators.

2. Negative feed back Amplifiers : if the feed back signal x_f is out of phase with the input signal x_s, then $x_i = x_s - x_f$. So the input voltage applied to the basic amplifier is decreased and correspondingly the output is decreased. Hence the voltage gain is reduced. This type of feedback is known as negative or degenerative feedback.

Gain of negative feedback amplifier:



Gain of amplifier $A = \frac{x_o}{x_i} \Rightarrow A x_i = x_o$

feed back factor $\beta = \frac{x_f}{x_o} \Rightarrow x_f = \beta x_o$

$x_o = A x_i$

$x_o = A(x_s - x_f)$

$x_o = A x_s - A x_f$

$x_o = A x_s - A\beta x_o$

$x_o + A\beta x_o = A x_s$

$x_o[1 + A\beta] = A x_s$

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

Gain with out feed back is always greater than gain with feedback

Here, $|A_f| < |A|$. if $|A\beta| \gg 1$, then $A_f \simeq \frac{A}{A\beta} \Rightarrow A_f \simeq \frac{1}{\beta}$

where β is a feedback ratio. Hence the gain depends less on the operating potentials and the characteristics of the transistor. The gain may be made to depend entirely on the feedback network. If the feedback network contains only stable passive elements. The gain of the amplifier using negative feedback is also stable.

The stabilisation of the dc operating point of a transistor amplifier is accomplished by the use of negative feed back as far as the dc potential is concerned and the operating point is kept constant in case of change in temperature or a change in the hfe or β of a transistor. Negative feedback is used to improve the performance of electronic amplifiers. Negative feedback always helps to increase the bandwidth, decrease distortion and noise, modify input and output resistances as desired. All the above advantages are obtained at the expense of reduction in voltage gain.

EFFECT OF NEGATIVE FEEDBACK ON AMPLIFIER CHARACTERISTICS

(or)

GENERAL CHARACTERISTICS OF NEGATIVE FEEDBACK AMPLIFIERS.

Since negative feedback reduces the transfer gain, why is it used? The answer to this question is that it is used because, many desirable characteristics are obtained for the price of gain reduction. We now examine some of the advantages of negative feed back.

The negative feedback in amplifier circuit results in decreased voltage gain, noise and distortion, but there will be an increase in bandwidth. In addition to these characteristics, input and output impedances get raised according to feedback connection.

Although there is a reduction in overall voltage gain, there are some improvements in using negative feedback in amplifier

1. Better stabilized voltage gain : The variation due to aging, temperature, replacement etc. of the circuit components and transistor or tube characteristics is reflected in corresponding lack of stability of the amplifier transfer gain.

The transfer gain of the amplifier is not constant as it depends on the factors such as operating point, temperature etc. The lack of stability in amplifiers can be reduced by introducing negative feedback

Since A represents either A_v, G_m, A_I or R_m and A_f represents the corresponding transfer gains with feed back, either A_{vf}, G_{mf}, A_{if} or R_{mf} the equation signifies that :

(6)

→ For voltage & series feedback

$$\boxed{A_{vf} = \frac{1}{\beta}}$$ Voltage gain is stabilized

→ For Current series feedback

$$\boxed{G_{mf} = \frac{1}{\beta}}$$ transconductance gain is stabilized

→ For voltage shunt feedback

$$\boxed{R_{mf} = \frac{1}{\beta}}$$ transresistance gain is stabilized.

→ For current shunt feedback

$$\boxed{A_{if} = \frac{1}{\beta}}$$ Current gain is stabilized.

Problems :-

Q) An Amplifier has an open loop gain of 1000 and a feedback ratio of 0.04. If the open loop gain changes by 10% due to temperature, find the percentage change in gain of the amplifier with feedback.

Solution :- Given $A = 1000$, $\beta = 0.04$ and $\frac{dA}{A} = 10$

we know that the percentage change in gain of the amplifier with feed back is

$$\frac{dA_f}{A_f} = \frac{dA}{A} \frac{1}{1+A\beta}$$

$$= 10 \times \frac{1}{1+1000 \times 0.04} = 0.25\%$$

Q2) An amplifier has voltage gain with feedback of 100. If the gain without feedback changes by 20% and the gain with feedback should not vary more than 2%, determine the values of open loop gain A and feedback ratio β.

We know that

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

Differentiating both sides with respect to A we get,

$$\frac{dA_f}{dA} = \frac{(1+\beta A)1 - \beta A}{(1+\beta A)^2} = \frac{1}{(1+\beta A)^2} \left[\because \frac{d}{dt}\left(\frac{a}{b}\right) = \frac{\frac{d}{dt}a \cdot b - a \cdot \frac{d}{dt}b}{b^2} \right]$$

$$dA_f = \frac{dA}{(1+\beta A)^2}$$

Dividing both sides by A_f we get,

$$\frac{dA_f}{A_f} = \frac{dA}{(1+\beta A)^2} \times \frac{1}{A_f}$$

$$\frac{dA_f}{A_f} = \frac{dA}{(1+\beta A)^2} \times \frac{(1+\beta A)}{A} \qquad \text{Since } A_f = \frac{A}{1+\beta A}$$

$$\frac{dA_f}{A_f} = \frac{dA}{A}\,\frac{1}{(1+\beta A)} \longrightarrow (1)$$

where $\frac{dA_f}{A_f}$ = Fractional change in amplification with feedback

$\frac{dA}{A}$ = fractional change in amplification without feedback

Looking at equation (1) we can say that change in the gain with feedback is less than the change in gain without feed back by factor $(1+\beta A)$. The fractional change in amplification with feed back divided by the fractional change with out feedback is called the Sensitivity of the transfer gain $(1/(1/(1+\beta A)))$.

The reciprocal of the sensitivity is called the desensitivity $D(1+\beta A)$

Therefore, stability of the amplifier increases with increase in desensitivity.

if $\beta A >> 1$, then

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

and the gain is dependent only on the feedback network

Solution:- Given $A_f = 100$,

$$\frac{dA_f}{A_f} = 2\% = 0.02 \text{ and } \frac{dA}{A} = 20\% = 0.2$$

W.K.T $$\frac{dA_f}{A_f} = \frac{dA}{A} \frac{1}{1+A\beta}$$

$$0.02 = 0.2 \times \frac{1}{1+A\beta}$$

$$\therefore (1+A\beta) = \frac{0.2}{0.02} = 10$$

Also we know that the gain with feedback is

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

$$100 = \frac{A}{10}$$

$$A = 1000$$

$$\therefore 1+A\beta = 10; \text{ ie } A\beta = 9$$

$$\beta = \frac{9}{1000} = 0.09$$

Q3):- The gain (A_f) of an amplifier with feedback is to be nominally 20, and a variation of 5% is permissible. If the magnitude of the return ratio ($A\beta$) must be at least 1000 (so that $A\beta \gg 1$), then determine the minimum value of the open loop gain (A) and the maximum permissible variation in it.

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

if $A\beta \ggg 1$ then $A_f = \frac{A'}{A\beta} \Rightarrow A_f = \frac{1}{\beta}$

$A_f = 20$, we get $\beta = \frac{1}{20} = 0.05$. Then for $A\beta = 1000$, we must have $A \geq 1000/0.05 = 20000$

$$\left|\frac{dA_f}{A_f}\right| = \frac{1}{|1+A\beta|} \left|\frac{dA}{A}\right|$$

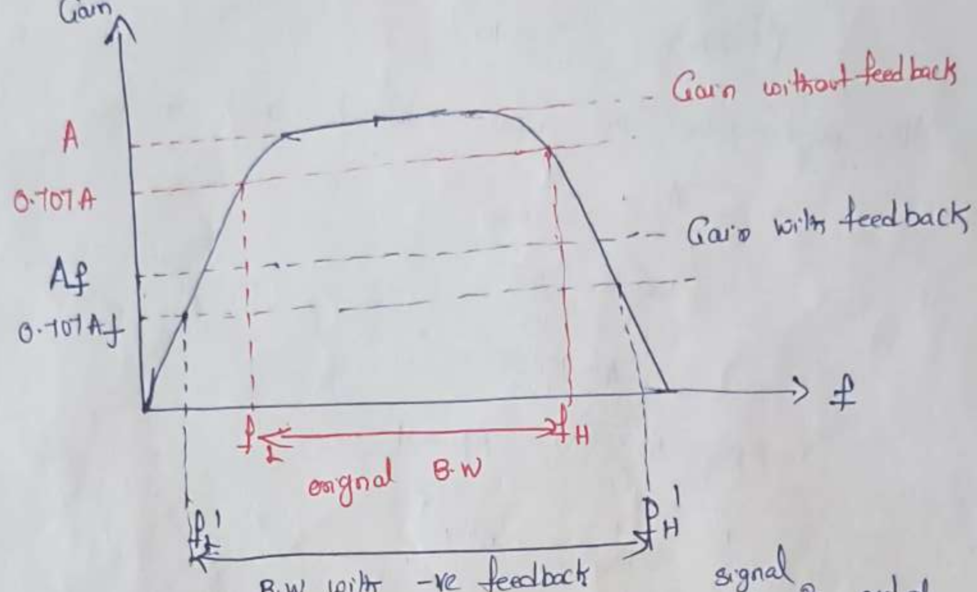
$$0.05 = \frac{1}{1000} \left|\frac{dA}{A}\right|$$

$$\left|\frac{dA}{A}\right| = 50 \rightarrow$$ This shows that the open-loop gain can vary by a factor of 50, while the gain with feedback varies only by 5%

Q. Enhanced frequency response (or) increased band width :- (or) increase of cut off frequencies

The bandwidth of an amplifier is the difference between the upper cut-off frequency f_2 and the lower cut-off frequency f_1

$$B.W = f_H - f_L$$



In negative feed back amplifiers the feed back signal is out of phase w.r. to the incoming signal so $V_i = V_s - V_f$. The incoming signal V_i at the basic amplifier input decreases. So gain also decreases. The gain of the a (-ve) feed back amplifier reduces by the factor $\overline{(1+A\beta)}$ by, it is seen that the lower cutoff frequency is also lowered by this factor $(1+A\beta)$ and upper cutoff frequency is raised by the same factor. As a result the difference between the frequencies means bandwidth is increased.

The product of voltage gain and bandwidth of an amplifier without feedback and with feedback remains the same. i.e $A_f \times BW_f = A \times BW$. As the voltage gain of a feedback

Due to the negative feedback in the amplifier, the upper cut off frequency f_H' is increased by the factor $(1+A\beta)$ and the lower cut-off frequency f_L' is decreased by the same factor $(1+A\beta)$.

Q2. An RC capled amplifier has a mid frequency gain of 200 and a frequency response from 100Hz to 20kHz. A negative feedback network with $\beta = 0.02$ is incorporated into the amplifier circuit. Determine the new system performance.

Solution:

$$A_f = \frac{A}{1+A\beta} = \frac{200}{1+200\times 0.02} = 40$$

$$f_1' = \frac{f_1}{1+A\beta} = \frac{100}{1+200\times 0.02} = 20\,Hz$$

$$f_2' = f_2(1+A\beta) = 20\times 10^3 \times (1+200\times 0.02) = 100\,kHz$$

BW with feed back is

$$BW_f = f_2' - f_1' = 100\times 10^3 - 20 \simeq 100\,kHz$$

$$A_f \times BW_f = 40\times 100\times 10^3 = 4000\,kHz$$ → Gain Band width product with feedback

$$BW = f_2 - f_1 = 20\times 10^3 - 100 \simeq 20\,kHz$$

$$A\times BW = 200\times 20\times 10^3 = 4000\,kHz$$ → Gain Band width product without feed back.

This shows that the gain bandwidth product of the amplifier with negative feedback is same as that of the gain-band width product of the amplifier without feed back.

3. Increase in input impedance

An amplifier should have high input impedance (resistance). So that it will not load the preceding stage or the input voltage source. Such a desirable characteristics can be achieved with the help of negative ~~se~~ voltage series negative feedbacks. The input impedance with feedback is given by

$$\boxed{R_{if} = R_i(1+A\beta)}$$

Thus the input impedance is increased by a factor of $(1+A\beta)$. The proof for increase in input impedance is obtained in voltage series feedback amplifier.

There upper and lower 3dB frequencies of an amplifier with negative feedback are given by the relations

$$f_2' = f_2(1+A\beta) \text{ and } f_1' = \frac{f_1}{1+A\beta}$$

→ As the voltage gain of a feedback amplifier reduces by the factor $1/(1+A\beta)$, its bandwidth would be increased by $(1+A\beta)$ i.e $BW_f = BW(1+A\beta)$, where A is the mid band gain without feedback

Q1. An amplifier has a mid band gain of 125 and a bandwidth of 250KHz (a) if 4% negative feedback is introduced, find the new bandwidth and gain (b) if the bandwidth is to be restricted to 1MHz, find the feedback ratio. Given $A = 125$, $BW = 250\,KHz$ & $\beta = 4\% = 0.04$

(a) We know that $BW_f = (1+A\beta)BW$

$$= (1 + 125 \times 0.04) \times 250 \times 10^3 = 1.5\,MHz$$

(b) Gain with feedback $A_f = \frac{A}{1+A\beta} = \frac{125}{1+125\times0.04} = \frac{125}{6} = 20.83$

(b) $BW_f = (1+A\beta)BW$

$$1\times10^6 = (1+125\beta') \times 250\times10^3$$

$$\therefore (1+125\beta') = \frac{1\times10^6}{250\times10^3} = 4$$

$$\text{i.e } \beta' = \frac{3}{125} = 0.024 = 2.4\%$$

3. increase ~~in~~ input impedance!

An amplifier ~~should~~ have high input impedance (resistance) so that

4. Decrease in output impedance :-

An amplifier with low output impedance (resistance) is capable of delivering power (or voltage) to the load without much loss. Such a desirable characteristic is achieved by employing negative series voltage feed back in an amplifier. The output impedance with feedback is expressed by

$$Z_{of} = \frac{Z_o}{A+A\beta}$$

$$\boxed{R_{of} = \frac{R_o}{1+A\beta}}$$

Here the output impedance is reduced by a factor $(1+A\beta)$. The proof for decrease in output impedance is given in voltage series feedback amplifiers.

(5) Decreased Distortion :-

Consider an amplifier with an open loop voltage gain and a total harmonic distortion D. Then with the introduction of negative feedback with the feed back ratio, β, the distortion will reduce to

$$D_f = \frac{D}{1+A\beta}$$

Q1) An amplifier has a voltage gain of 400, $f_1 = 50Hz$, $f_2 = 200kHz$ and a distortion of 10% without feed back. Determine the amplifier voltage gain f_1' & f_2' and D_f when a negative feed back is applied with feedback ratio of 0.01.

Given $A = 400$, $f_1 = 50Hz$, $f_2 = 200kHz$, $D = 10\%$ and $\beta = 0.01$

We know that voltage gain with feed back

$$A_f = \frac{A}{1+A\beta} = \frac{400}{1+400\times0.01} = 80$$

New lower 3dB frequency

$$f_1' = \frac{f_1}{1+A\beta} = \frac{50}{1+400\times0.01} = 10Hz$$

New upper 3dB frequency

$$f_{2f} = f_2' = f_2 \times (1+A\beta)$$

$$= 200\times10^3 \times (1 + 400\times0.01) = 1\,MHz$$

→ Distortion with feedback

$$D_f = \frac{D}{1+A\beta} = \frac{10}{5} = 2\%$$

6 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. Thus the noise with feedback is given by —

$$\boxed{N_f = \frac{N}{1+A\beta}}$$

7. Desensitivity:

Sensitivity: The fractional change in amplification with feedback divided by the fractional change in amplification without feedback is called the sensitivity of the transfer gain.

$$\text{Sensitivity} = \frac{\frac{dA_f}{A_f}}{\frac{dA}{A}} = \frac{1}{1+A\beta}$$

Desensitivity is defined as the reciprocal of sensitivity. It indicates the factor by which the voltage gain has been reduced due to feedback network. Desensitivity factor $[D] = 1+A\beta$

Disadvantages of Negative feed back

Gain with feedback decreases when compared to gain without feedback $A_f = \frac{A}{1+A\beta}$,

The gain of the feedback amplifier reduces by a factor of $(1+A\beta)$

Classification of Amplifiers:

→ Based on transistor configuration

A) CE Amplifier
B) CB Amplifier
c) CC Amplifier

→ Based on the period of conduction

A> Class - A amplifier
B> Class - B amplifier
C) Class - C amplifier
D> Class - AB amplifier
E> Class - D amplifier
F> Class - S amplifier
G> MOSFET amplifier.

→ Based on the Coupling elements

A> RC coupled Amplifier
B> Transformer Coupled Amplifier
C> Direct Coupled Amplifier.

→ Based on stages.

A> Single stage amplifier
B> Multi stage amplifier

→ Based on input signal

A> Small signal Amplifiers
B> Large signal Amplifiers [power Amplifiers]

→ Based on its o/p

A> Voltage amplifiers
B> power amplifiers

→ Based on frequency

A> Audio frequency amplifiers.
B> Radio frequency amplifiers
C> Microwave frequency amplifiers
D> Ultra high frequency amplifiers

Generalised Analysis of Negative feedback Amplifiers:

1 Voltage Amplifier (or) Voltage Series feedback Amplifier:

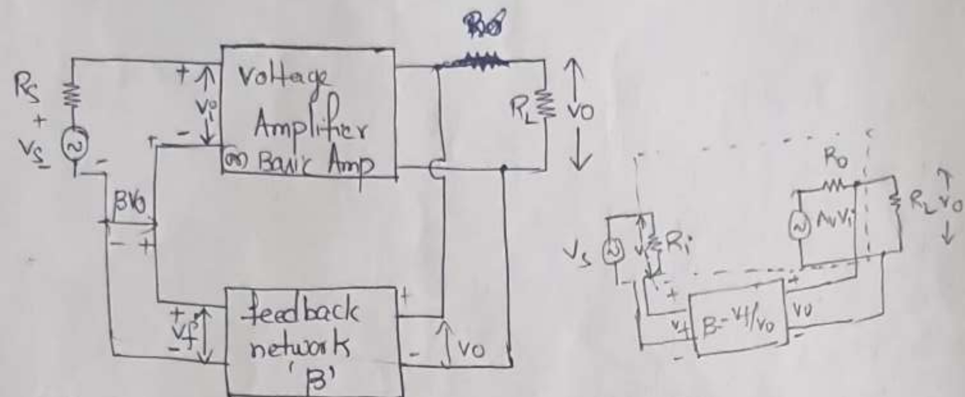


fig (a) voltage amplifier with voltage series feedback

The voltage series feedback topology shown in fig (b). with amplifier is replaced by Thevenin's model. Here, A_V represents the open-circuit voltage gain taking R_S in to account. Since throughout the discussion of feedback amplifiers we will consider R_S to be part of the amplifier and we will drop the subscript on the transfer gain and input resistance. (

Equivalent Circuit:

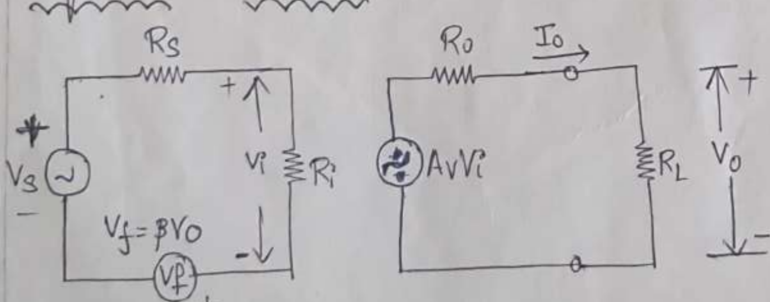
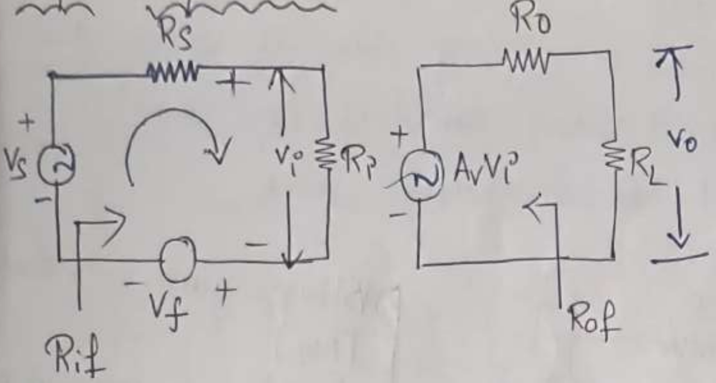


fig: Thevenins equivalent circuit for voltage Amplifier.

→ If R_i is large compared to R_S $[R_i >>> R_S]$ then V_i is similarly equal to V_S $[V_i \simeq V_S]$

→ If R_L is large compared with o/p resistance R_O $[R_L >>> R_O]$, V_O is equals to $V_O = A_V V_i$ or $V_O = A_V V_S$

Input impedance:



from fig we have

without feedback $R_i = \frac{V_i}{I_i}$

with feedback $R_{if} = \frac{V_S}{I_i}$

From the circuit input impedance increases and output impedances decreases.

→ Apply KVL to the i/p ckt (or) when the feedback is present $V_S - V_f = V_i$

$$V_S - V_f = V_i$$

We know that

$$\beta = \frac{V_f}{V_O} \Rightarrow V_f = \beta V_O$$

$$V_S - \beta V_O = V_i \longrightarrow ①$$

From output circuit

if $R_L >>> R_O$, then $V_O = A_V V_i \rightarrow ②$

sub eq ② in eq ①

$$V_S - \beta A_V V_i = V_i$$

$$V_S = V_i [1 + \beta A_V]$$

$$\frac{V_S}{V_i} = 1 + \beta A_V$$

$\therefore V_i = I_i R_i$

$$\boxed{V_S - \beta A_V V_i = V_i}$$

$$V_S = V_i + \beta A_V V_i$$

$$V_S = V_i [1 + \beta A_V]$$

$$V_S = I_i R_i [1 + \beta A_V]$$

$$\frac{V_S}{I_i} = R_i [1 + \beta A_V]$$

$$\boxed{R_{if} = R_i [1 + \beta A_V]}$$

where

R_{if} = i/p impedance with feedback

R_i = i/p impedance without f/b

$A_V \rightarrow$ vtg gain without f/b

$\beta \rightarrow$ feedback factor $(\beta) = \frac{V_f}{V_O}$

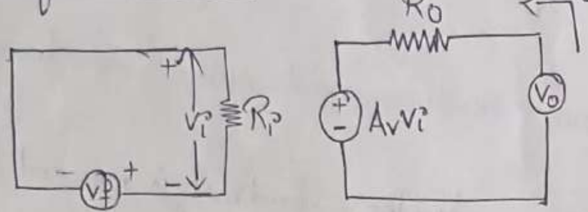
Output impedance: in this topology, the output resistance can be measured by shorting the input source $V_S = 0$ and looking in to output terminals with R_L disconnected as shown in the figure i.e

(i) Remove the load resistance

ii) At i/p side source must be zero, if source is voltage then it is short circuit, if source is current then it is open circuit.

iii) We need to connect external voltage source

<u>Modified Circuit</u>



from the o/p side circuit

$$V_0 = I_0 R_0 + A_V V_i \rightarrow (*)$$

$$V_0 - A_V V_i = I_0 R_0 \longrightarrow (1) \qquad \beta = \frac{V_f}{V_0}$$

from i/p side circuit if $V_s = 0$, $V_i = -V_f$

$$-V_f = V_i \longrightarrow (2)$$

sub eq (2) in eq (1)

$$V_0 + A_V V_f = I_0 R_0 \qquad [V_f = \beta V_0]$$

$$V_0 + A_V \beta V_0 = I_0 R_0$$

$$V_0 + A_V \beta V_0 = I_0 R_0$$

$$V_0 [1 + A_V \beta] = I_0 R_0$$

$$\boxed{\frac{V_0}{I_0} = \frac{R_0}{1 + A_V \beta}} \Longrightarrow \boxed{R_{of} = \frac{R_0}{1 + A_V \beta}}$$

(3) Voltage gain with feedback $[A_{vf}]$

→ from figure we can find if there is no feedback ($V_f = 0$),

→ The vtg gain of the Amplifier

$$\boxed{A_V = \frac{V_0}{V_i}}$$

→ if there is feedback $[V_s = V_i + V_f]$

The vtg gain $\boxed{A_{vf} = \frac{V_0}{V_s}}$

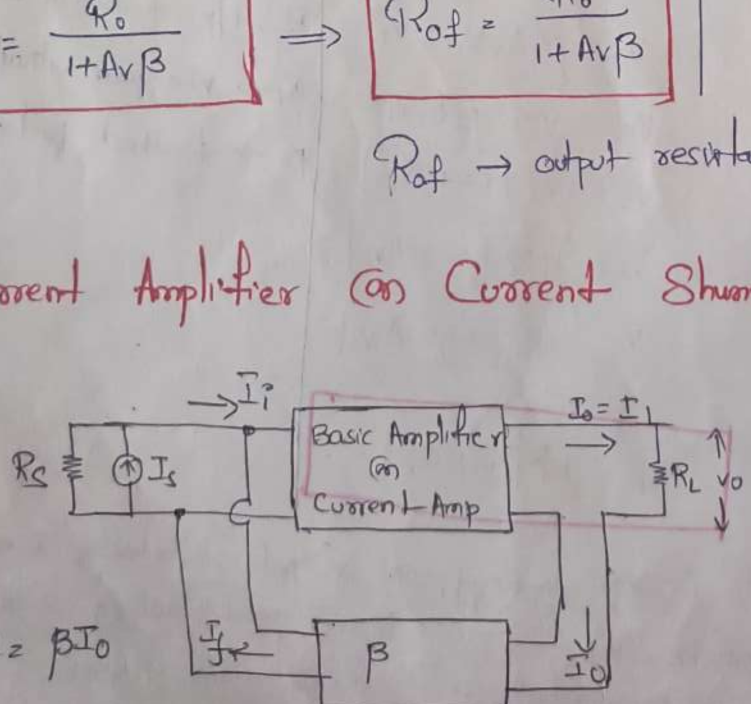
$$A_{vf} = \frac{V_0}{V_i + V_f}$$

$$A_{vf} = \frac{V_0 / V_i}{V_i/V_i + V_f/V_i} = \frac{A_V}{1 + \frac{V_f}{V_0} \times \frac{V_0}{V_i}}$$

$$\boxed{A_{vf} = \frac{A_V}{1 + \beta A_V}}$$

R_{of} → output resistance with feedback

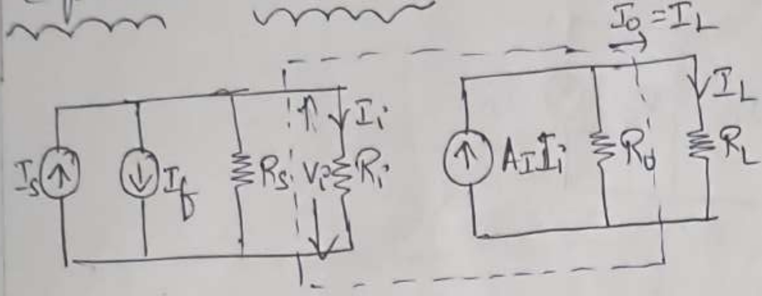
Current Amplifier (or) Current Shunt amplifier.



$I_f = \beta I_0$

A current amplifier is defined as an amplifier which provides an output current proportional to the input current and the proportionality factor is independent of the magnitudes of the source resistance R_s and load resistance R_L

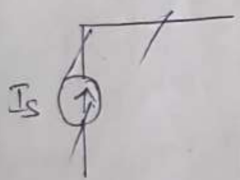
Equivalent Circuit :-



fig(b) Norton's equivalent circuit for current amplifier

Fig(b) shows Norton's equivalent circuit of a current amplifier. if amplifier input resistance $R_i \to 0$, then $I_i \simeq I_s$. if amplifier output resistance $R_o \to \infty$, then $I_L = A_i I_i$. Such amplifier provides an output current.

i> Input impedance :-



from the input circuit

$$I_s = I_f + I_i \qquad [\because I_f = \beta I_o]$$

$$I_s = \beta I_o + I_i \rightarrow (1)$$

from the output circuit

$$I_o = A_i I_i \rightarrow (2) \qquad \therefore [R_L <<< R_o]$$

Sub eq (2) in eq (1)

$$I_s = \beta A_i I_i + I_i$$

$$I_s = I_i [1 + \beta A_i]$$

$$\frac{I_s}{I_i} = [1 + \beta A_i] \Rightarrow I_s = \frac{V_i}{R_i}[1 + \beta A_i] \Rightarrow$$

$$\frac{I_s}{V_i} = \frac{1 + \beta A_i}{R_i}$$

$$\frac{V_i}{I_s} = \frac{R_i}{1 + \beta A_i} \Longrightarrow \boxed{R_{if} = \frac{R_i}{1 + A_i\beta}}$$

Output impedance :

In this topology, the output resistance can be measured by :

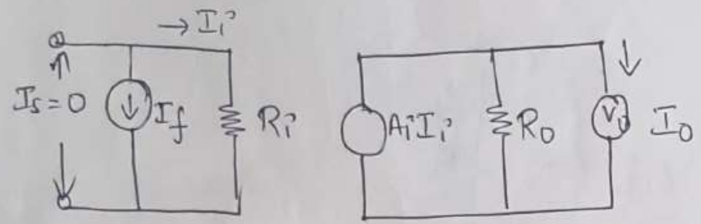
(1) open co

(1) open circuiting the input source $I_S = 0$

(2) R_L is removed

(3) External source is added at the output

Voltag



from the output ckt

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

from the input circuit

$$I_S = 0, \quad I_i = -I_f \rightarrow (2)$$

sub eq (2) in eq (1) $\quad [\because I_f = \beta I_o]$

$$I_o = \frac{V_o}{R_o} + A_i(-I_f)$$

$$I_o = \frac{V_o}{R_o} - A_i \beta I_o$$

$$I_o [1 + A_i \beta] = \frac{V_o}{R_o}$$

$$\frac{V_o}{I_o} = R_o [1 + A_i \beta]$$

$$\boxed{R_{of} = R_o [1 + A_i \beta]}$$

Transfer gain (A_{if}) :

→ The current gain without feedback $A_i = I_o / I_i$

→ The current gain with feedback $A_{if} = I_o / I_s$

→ From the i/p current circuit = $\boxed{I_S = I_i + I_f}$

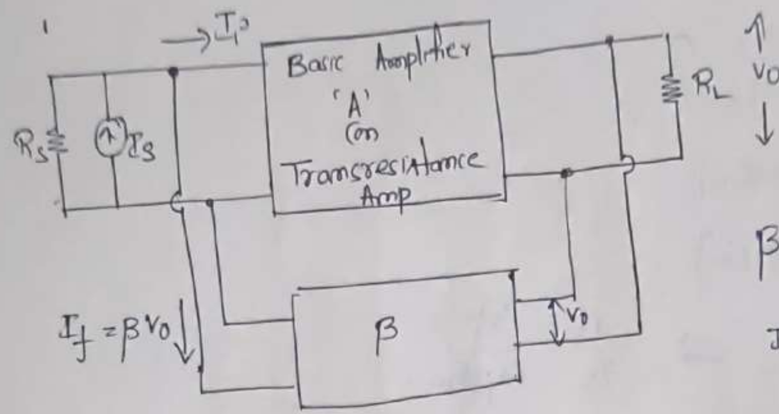
→ $A_{if} = \dfrac{I_o}{I_i + I_f}$, $A_{if} = \dfrac{I_o / I_i}{I_i / I_i + \beta I_o / I_i}$

$\beta = I_f / I_o$

$I_f = \beta I_o$

$$\boxed{A_{if} = \frac{A_i}{1 + \beta A_i}}$$

3) Voltage shunt feedback Amplifier (or) Transresistance Amplifier:

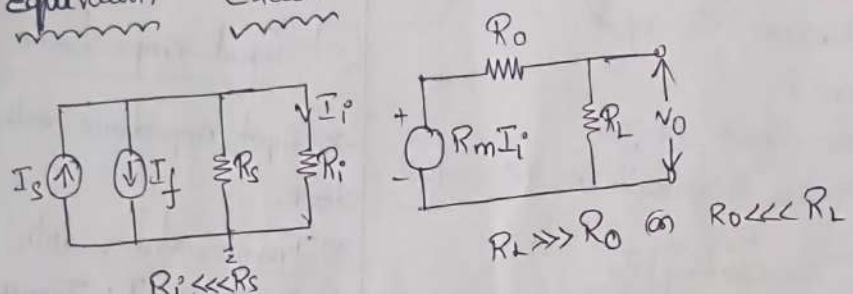


$$\beta = \frac{I_f}{V_O}$$

$$I_f = \beta V_O$$

fig:- Transresistance amplifier with voltage shunt feedback

Equivalent Circuit



★ fig shows a transresistance amplifier with a Norton's equivalent in its input circuit and Thevenin's equivalent in its output circuit. In this amplifier an output voltage is proportional to the input signal current and the proportionality factor is independent on the source and load resistances. Ideally, the amplifier must have zero input resistance R_i and zero output resistance R_O. For practical transresistance amplifier we must have $R_i \ll R_S$ and $R_O \ll R_L$

① **Input impedance:-**

from the input circuit

$$I_S = I_f + I_i \qquad\qquad I_f = \beta V_O$$

$$= \beta V_O + I_i \rightarrow ①$$

from o/p circuit

$$V_O = R_m I_i \rightarrow ② \quad [\text{if } R_L \ggg R_O]$$

Sub. eq ② in eq ①

$I_f = \beta V_0$

$I_s = \beta V_0 + I_i$

$= \beta [R_m I_i] + I_i$

$I_s = I_i [1 + \beta R_m]$

$I_s = \frac{V_i}{R_i} [1 + \beta R_m]$

$\frac{V_i}{\cancel{I_s}} = \frac{\cancel{R_m}}{\cancel{1+\beta R_m}} \Rightarrow \frac{V_i}{I_s} = \frac{R_i}{1+\beta R_m}$

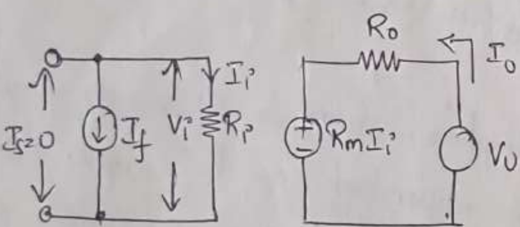
$\boxed{\cancel{R_{mf}} = \frac{R_{\cancel{m}}}{1\cancel{+}\beta R_m}}$ $R_{inf} = \frac{R_i}{1+\beta R_m}$

② **Output impedance:** In this topology, the output resistance can be measured by shorting the input.

i> R_L is removed

ii> Source at the input is zero

iii> External source is added at the output.



from the o/p circuit apply KVL

$V_0 = R_m I_i + I_0 R_0$

$V_0 - R_m I_i = I_0 R_0 \rightarrow ①$

from i/p circuit apply KCL

$I_i = -I_f \rightarrow ②$

sub eq ② in eq ①

$V_0 + R_m I_f = I_0 R_0$

$V_0 + R_m \beta V_0 = I_0 R_0$

$V_0 [1 + R_m \beta] = I_0 R_0$

$\frac{V_0}{I_0} = \frac{R_0}{1 + R_m \beta}$

$\boxed{R_{of} = \frac{R_0}{1 + R_m \beta}}$

③ ~~Input impedance~~

→ ~~input impedance without feed.~~

(3) Transresistance with feedback [R_{mf}]: Transfer gain with feedback.

→ Transresistance without feedback $R_m = V_0 / V_i$

→ Transresistance with feedback $R_{mf} = \frac{V_0}{I_s}$

→ from the i/p circuit

$I_s = I_i + I_f$

$R_{mf} = \frac{V_0}{I_i + I_f}$ $\beta = I_f / V_0$

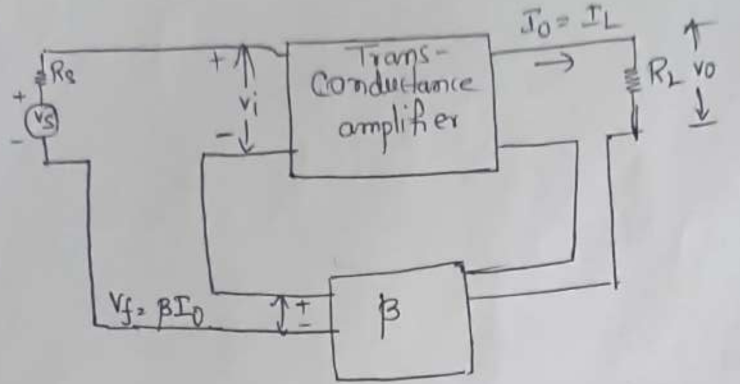
$I_f = \beta V_0$

$R_{mf} = \frac{V_0}{I_i + \beta V_0}$

$R_{mf} = \frac{V_0 / I_i}{I_i / I_i + \beta \, V_0 / I_i}$

$\boxed{R_{mf} = \frac{R_m}{1 + \beta R_m}}$

Current Series Amplifier (or) Transconductance amplifier :-



Transconductance amplifier with current series feedback.

Equivalent Circuit :-

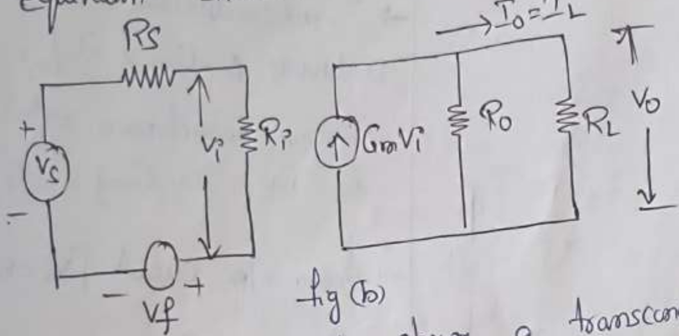


fig (b)

The above fig (b) shows a transconductance amplifier with a Thevenin's equivalent in its input circuit and Norton's equivalent in its output circuit. In this amplifier, an output current is proportional to the input signal voltage and the proportionality factor is independent of the magnitudes of the source and load resistances. ideally this amplifier must have an infinite input resistance R_i and infinite output resistance R_o. For practical transconductance amplifier we must have $R_i \gg R_s$ and ~~$R_o \ll R_L$~~ $R_o \gg R_L$

Input impedance :-

from input side

$$V_s - V_f = V_i \rightarrow (1)$$

from output circuit

$$I_o = G_m V_i \rightarrow (2)$$

Sub eq ② in eq ①

$$\beta = \frac{V_f}{I_o}$$

$$V_s - \beta I_o = V_i$$

$$V_s - \beta (G_m V_i) = V_i$$

$$V_s = \beta G_m V_i + V_i$$

$$V_s = V_i [1 + \beta G_m]$$

$$V_s = I_i R_i [1 + \beta G_m]$$

$$\frac{V_s}{I_i} = R_i [1 + \beta G_m]$$

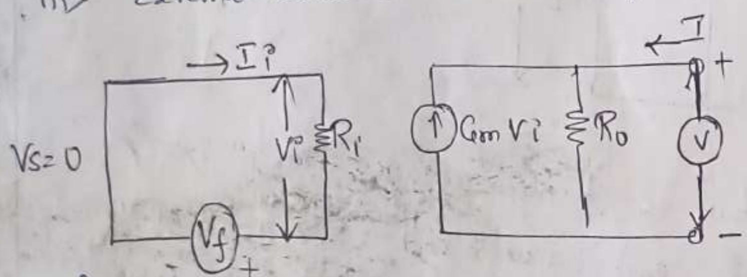
$$\boxed{R_{if} = R_i [1 + \beta G_m]}$$

Output impedance :- In this topology the output resistance can be measured by

i> R_L is zero

ii> Source is zero at i/p

iii> External source is added at o/p



from the output circuit

$$I_o = \frac{V_o}{R_o} + G_m V_i \rightarrow ①$$

from the i/p circuit.

$$V_i = -V_f \rightarrow ②$$

Sub eq ② in eq ①

$$\beta = \frac{V_f}{I_o}$$

$$V_f = \beta I_o$$

$$I_o = \frac{V_o}{R_o} + G_m(-V_f)$$

$$I_o = \frac{V_o}{R_o} - G_m V_f$$

$$I_o = \frac{V_o}{R_o} - G_m \beta I_o$$

$$I_o + G_m \beta I_o = \frac{V_o}{R_o}$$

$$I_o [1 + \beta G_m] = \frac{V_o}{R_o}$$

$$\frac{V_o}{I_o} = R_o [1 + \beta G_m]$$

$$\boxed{R_{of} = R_o [1 + \beta G_m]}$$

(3) Transfer Gain $[G_{mf}]$:-

→ Transconductance without feedback is $G_m = I_o / V_i$

→ Transconductance with feedback is $G_{mf} = I_o / V_s$

→ from i/p circuit $[V_s = V_i + V_f]$

$$G_{mf} = \frac{I_o}{V_i + V_f}$$

$$G_{mf} = \frac{I_o}{V_i + \beta I_o}$$

$$G_{mf} = \frac{I_o / V_i}{V_i / V_i + \beta I_o / V_i}$$

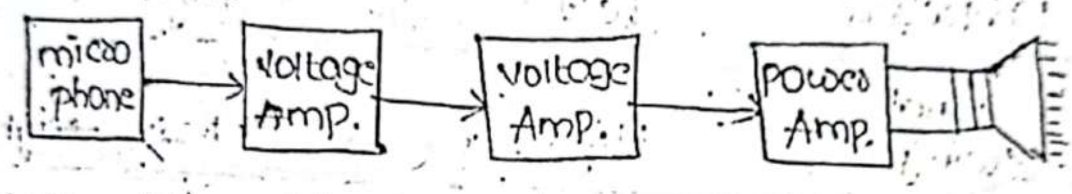
$$\boxed{G_{mf} = \frac{G_m}{1 + \beta G_m}}$$

Power Amplifiers and Tuned Amplifiers

Power Amplifiers:-

It is an electronic circuit which is used to strengthing the power level of any signal without changing its shape.

The basic block diagram of power amplifier consisting of microphone followed by voltage amplifier and power amplifier then the output of power amplifier is connected to the load speaker



Microphone is used to convert voice signal into electrical signal

Voltage amplifier is used to amplify the output signal of microphone. Here we are using multistage amplifier to produce large signal to the power amplifier circuit.

power amplifier converts DC power into AC power, which is required for the input of load speaker.

Classification of power Amplifiers:-

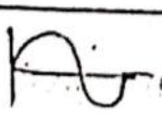
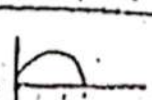
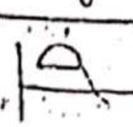
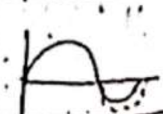
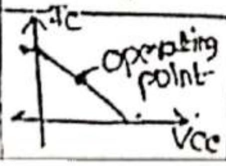
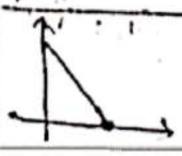
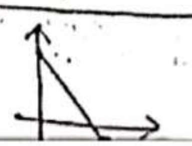
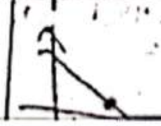
Based on location of operating point and conduction angle the power amplifiers are classified into different types and they are

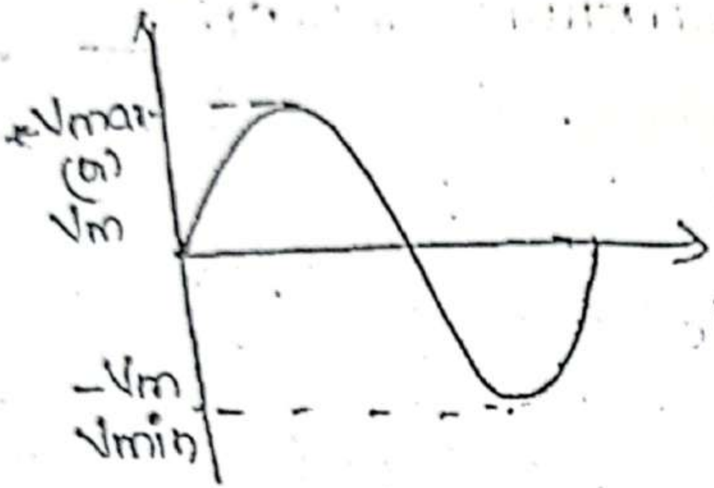
1. class A power Amplifier
2. class B power Amplifier
3. class C power Amplifier
4. Class AB power Amplifier

The comparison between small signal Amplifier and large signal Amplifier

Parameter	Small Signal Amplifier	Large signal Amp.
1. Name of the amp.	voltage Amplifier	power Amplifier
2. Transistor	Normal transistor	power transistor
3. Analysis model	h-parameter	Graphical
4. β current gain	Large	Small
5. power handling capability	less	high
6. Harmonic Distortion	less	High
7. Signal Swing	~~High~~ less	less ~~High~~
8. Size	less	High
9. Heak Sinks	not Required	Required

Comparison among power Amplifiers:-

Parameter	Class A	class B	class C	class AB
1. conduction angle	360°	180°	<180°	180°-360°
2. location of operating point	middle	X-axis	below X-axis	Between middle and X-axis
3. efficiency	25-50%	78.5%	High	25-78%
4. Distortion	Absent	present	present	High
5. power dissipation	very high	low	very low	High
6. output waveform				
7. DC load line				



$V_{pp} = V_{max} - V_{min}$

$= V_m - (-V_m)$

$\boxed{V_{pp} = 2V_m}$

power = V × I

$\boxed{P_{dc} = \text{D.C power} = V_{dc} \times I_{dc}}$

$P_{ac} = \text{ac power} = V_{ac} \times I_{ac}$

$= V_{rms} \times I_{rms}$

$V_{rms} = \frac{V_m}{\sqrt{2}}$, $I_{rms} = \frac{I_m}{\sqrt{2}}$

$P_{ac} = \frac{V_m}{\sqrt{2}} \times \frac{I_m}{\sqrt{2}} = \frac{V_m I_m}{2}$

$P_{ac} = \frac{\frac{V_{pp}}{2}}{\sqrt{2}} \times \frac{\frac{I_{pp}}{2}}{\sqrt{2}}$

$P_{ac} = \frac{V_{pp}}{2\sqrt{2}} \times \frac{I_{pp}}{2\sqrt{2}}$

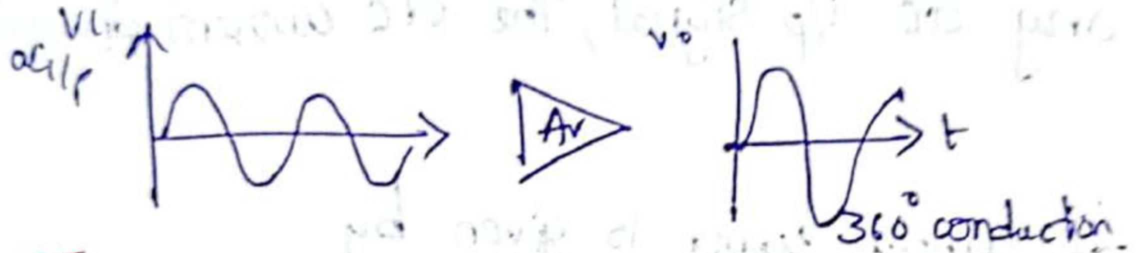
$\boxed{P_{ac} = \frac{V_{pp} \times I_{pp}}{8}}$

$\boxed{P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}$

123

Class - A - Amplifier

→ It is a large signal amplifier that produces 360° amplified o/p signal. i.e. The transistor is on for the full cycle of the i/p signal.

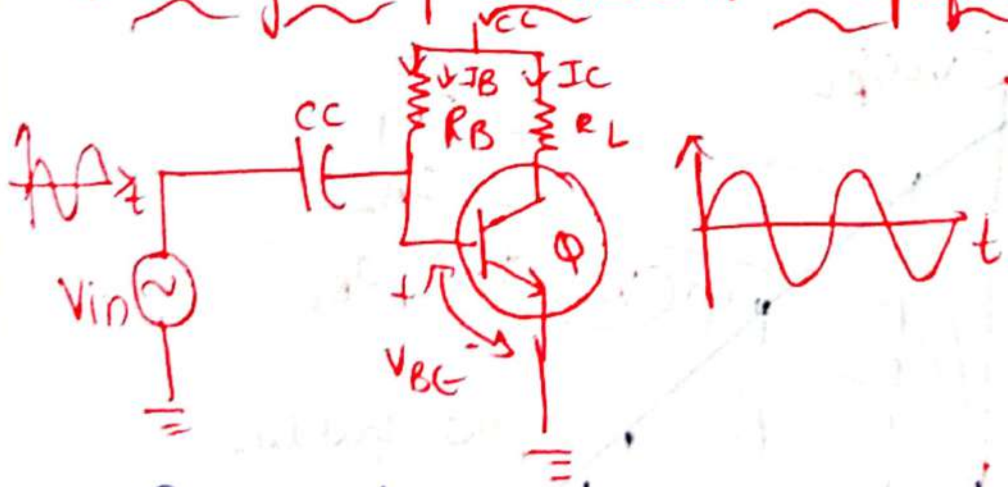


Types:-

(i) Directly coupled Class - A - Amplifier.

ii Transformer Coupled class A - Amplifier:-

Directly Coupled Class A Amplifier or Series fed.



DC - power
$P_{dc} = V_{CC} \cdot I_{CQ}$.

→ Power transistor is used to amplify the large power signal.

→ The load resistor is directly connected to the collector circuit, is known as directly coupled or series fed class A power amplifier.

Analysis of Class-A Amplifier:-

DC power:-

→ The DC power is provided by the supply voltage (V_{CC})

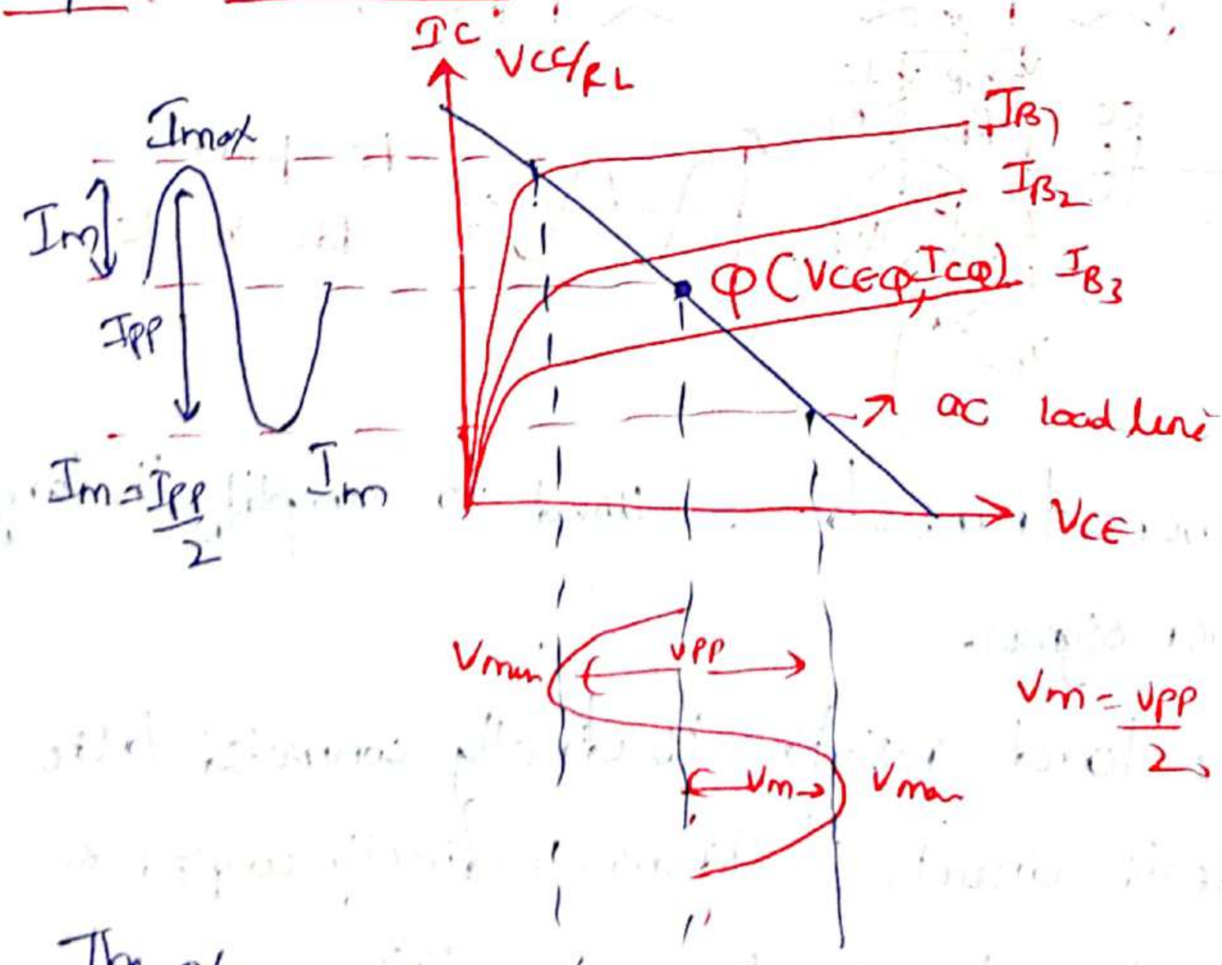
→ Without any ac i/p signal, the dc current drawn is I_{CQ}.

Then the dc supply power is given by

$$P_{dc} = V_{CC} \cdot I_{CQ} \longrightarrow ①$$

AC power:-

O/p characteristics:-



The o/p power is given as

$P = V_c I_c = I_c^2 R_L = \frac{V_c^2}{R_L}$

∴ The Rms values of the o/p current & voltage are given as

$$V_{rms} = \frac{V_m}{\sqrt{2}} = \frac{V_{max} - V_{min}}{2\sqrt{2}}$$

$$I_{rms} = \frac{I_m}{\sqrt{2}} = \frac{I_{max} - I_{min}}{2\sqrt{2}}.$$

∴ The ac power delivered to the load line is

$$P_{ac} = V_{rms} \cdot I_{rms}$$

$$= \frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}} = \boxed{\frac{V_m I_m}{2} = P_{ac}} \rightarrow (2)$$

$$P_{ac} = \frac{V_m I_m}{2} = \frac{I_m^2 R_L}{2} = \frac{V_m^2}{2R_L}.$$

In terms of max & min values

$$P_{ac} = \frac{\left\{\left(\frac{V_{max} - V_{min}}{2}\right)\left(\frac{I_{max} - I_{min}}{2}\right)\right\}}{2}$$

$$P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8} \rightarrow (3)$$

Efficiency :-

→ It represents the amount of ac power delivered to the load from the dc source.

$$\eta\% = \frac{\text{ac power delivered to the load}}{\text{dc supply power}} \times 100$$

→ It is also called as conversion efficiency.

$$\eta\% = \frac{P_{ac}}{P_{dc}} \times 100$$

$$\eta\% = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8 V_{CC} I_{CQ}} \times 100 \quad \text{— (4)}$$

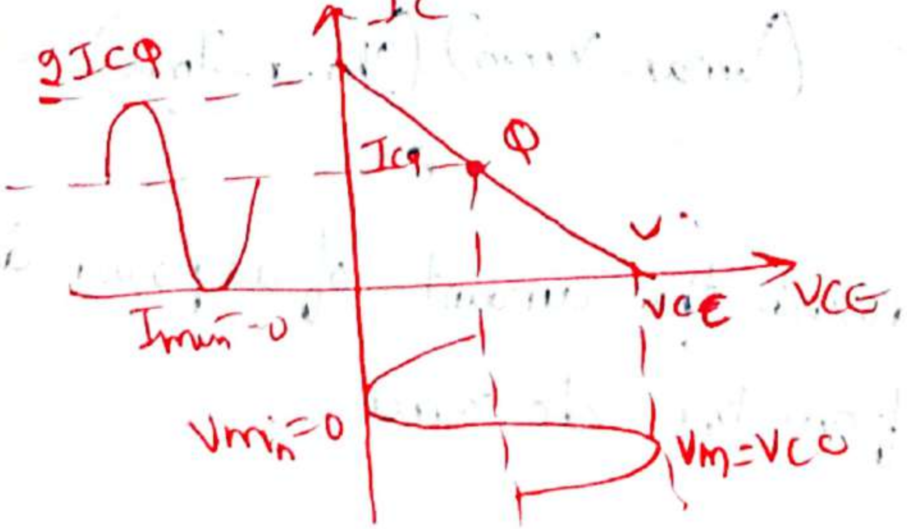
In terms of RMS values.

$$\eta\% = \frac{V_m I_m}{2 \times V_{CC} I_{CQ}} \times 100$$

$$\boxed{\eta\% = 50 \times \frac{V_m I_m}{V_{CC} I_{CQ}} \%} \longrightarrow (5)$$

* Maximum Efficiency :-

For max efficiency, the max swings of both o/p voltage & current are considered.



$\therefore V_{max} = V_{CC}, \quad I_{max} = 2 I_{CQ}$.

Sub in eqn (4)

$$\eta\% = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8 V_{CC} I_{CQ}} \times 100$$

$$= \frac{V_{CC} \cdot 2 I_{CQ}}{8 V_{CC} I_{CQ}} \times 100$$

$$\eta\% = 25\%.$$

In practical, η is of the order of 10% to 15%.

Power Dissipation :-

It is the difference between the dc i/p power & the ac o/p power.

$$\boxed{P_d = P_{dc} - P_{ac}.}$$

The amount of power dissipated by the transistor in the form of heat.

Maximum power Dissipation :-

→ The max power dissipation occurs when there is no ac i/p signal i.e the dc power without ac i/p signal is the max power dissipation.

$$\boxed{P_{dmax} = V_{CC} \cdot I_{CQ}}$$

* Advantages:

→ Simple circuit.

→ Transformer less circuit → cheaper.

→ Less no. of components required as load is directly connected

Disadvantages:-

→ More power dissipation

→ Poor efficiency

→ It cannot be used for low impedance loads (i.e) loud-speakers (4-20Ω)

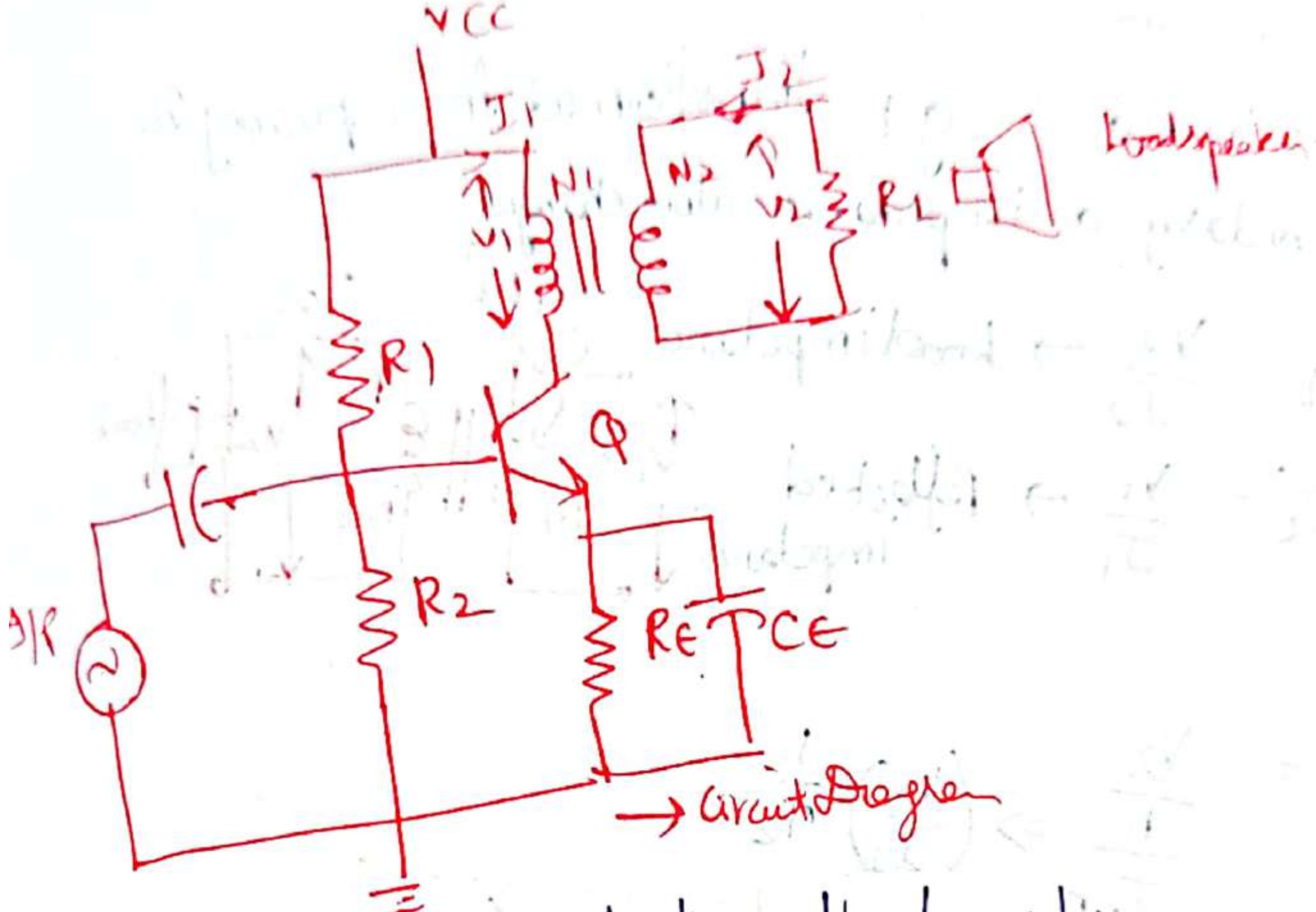
→ The load directly connected is collector circuit causes large amount of power wastage.

Transformer coupled class A power amplifier

→ In order to improve the efficiency, Impedance matching is required for maximum power transfer to the load.

→ For low impedance loads (4 to 20Ω) like loud speakers, it is difficult to match the impedance in directly coupled class-A Amplifier.

→ This problem can be eliminated by using o/p transformer to deliver max power to the load.



→ Circuit Diagram

→ R_1 & R_2 are used to bias the transistor.

→ The transformer transforms the voltage applied on one side to other side proportional to N_2/N_1.

Stepdown transformer is used to match the low impedance load.

→ Voltage transformation ratio

$$\eta = \frac{N_2}{N_1} = \frac{\text{No. of secondary turns}}{\text{No. of primary turns}}$$

$$\boxed{\frac{V_2}{V_1} = \frac{N_2}{N_1} = n.}$$

→ Current transformation

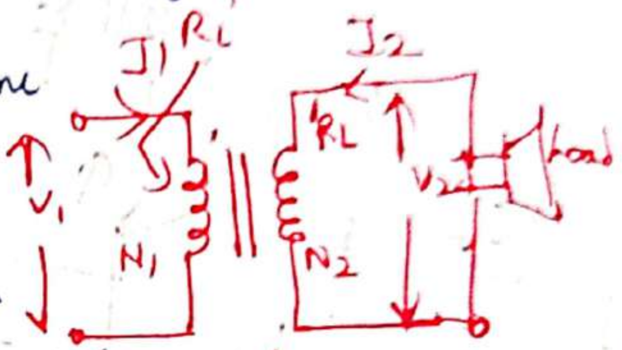
$$\boxed{\frac{I_2}{I_1} = \frac{N_1}{N_2} = 1/n}$$

Impedance transformation

As V and I get transformed from primary to secondary, as impedance also changes.

$R_L = \frac{V_2}{I_2}$ → Load impedance

$R_L' = \frac{V_1}{I_1}$ → reflected impedance



$$= \frac{\frac{V_2}{n}}{I_2 n} \Rightarrow \left(\frac{V_2}{I_2}\right)\frac{1}{n^2}$$

$\therefore$ $\boxed{R_L' = \frac{R_L}{n^2}}$ or $\boxed{R_L' = R_L\left(\frac{N_1}{N_2}\right)^2}$

→ Ensure always $R_L' >> R_L$ for a step down Transformer

→ n is much less than unity $n << 1$.

DC i/p power

→ Without any ac i/p signal, the dc power is provided by supply voltage V_{cc} & dc current is collector I_{CQ}

$$\boxed{P_{dc} = V_{cc} I_{CQ}}$$

→ **AC o/p power:**

The ac power delivered to the load is on the secondary side of transformer

$$P_{ac} = \frac{V_{rms} I_{rms}}{R} = \frac{V_m}{\sqrt{2}} \frac{I_m}{\sqrt{2}} = \frac{V_m I_m}{2}$$

In terms of max & min values

$$P_{ac} = \frac{(V_{PP}/2)(I_{PP}/2)}{2}$$

$$\boxed{P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}}$$

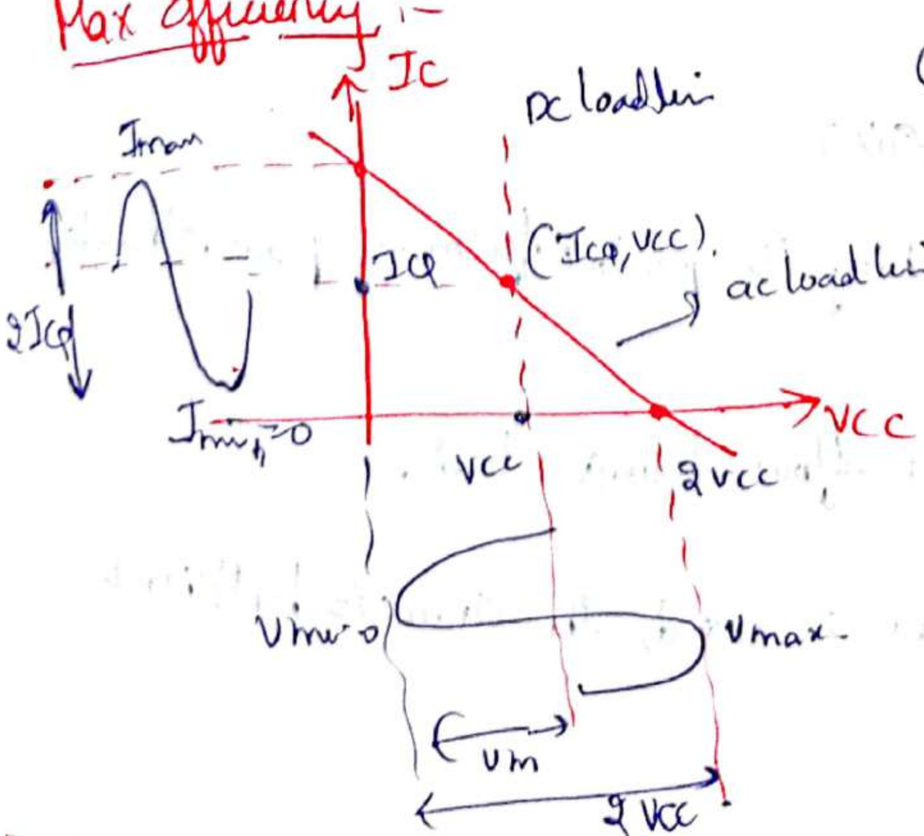
Efficiency :-

$$\eta = \frac{\text{ac power delivered to the load}}{\text{dc power (i/p)}}$$

$$\eta\% = \frac{P_{ac}}{P_{dc}} \times 100$$

$$\eta\% = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8 V_{CC} I_{CQ}} \times 100$$

Max Efficiency :-



Q point is exactly at the centre of load line

$V_{max} = 2V_{CC}$

$V_{min} = 0$

$I_{max} = 2I_{CQ}$

$I_{min} = 0.$

$$\eta\% = \frac{(2V_{CC}-0)(2I_{CQ}-0)}{8V_{CC}I_{CQ}} \times 100 = \frac{4I_{CQ}V_{CC}}{8V_{CC}I_{CQ}} \times 100$$

$$\eta_{max} = 50\%.$$

Theoritical effeciency = 50%.

Practical effeciency = 30-40%.

* **Power dissipation:-**

$$P_d = P_{dc} - P_{ac}.$$

It is differen. b/w P_{dc} & P_{ac}.

→ When the ac i/p given, max power is delivered to the load with less power dissipation.

→ Without any input, the dc i/p power is dissipated as heat (i.e) the max power dissipation

$$P_{d\,max} = V_{CC} \cdot I_{CQ}$$

Advantages:-

→ Higher efficiency (50%)

→ Impedance matching is possible for many power transfer to load.

→ No dc current flow through load.

Dis:-

→ Since transformer is used, the circuit is bulkier & costlier.

→ Poor frequency response
→ ~~Complefi~~ complicated to design

Class-B power Amplifier:-

Two circuit configurations (depending on the type of transistors)

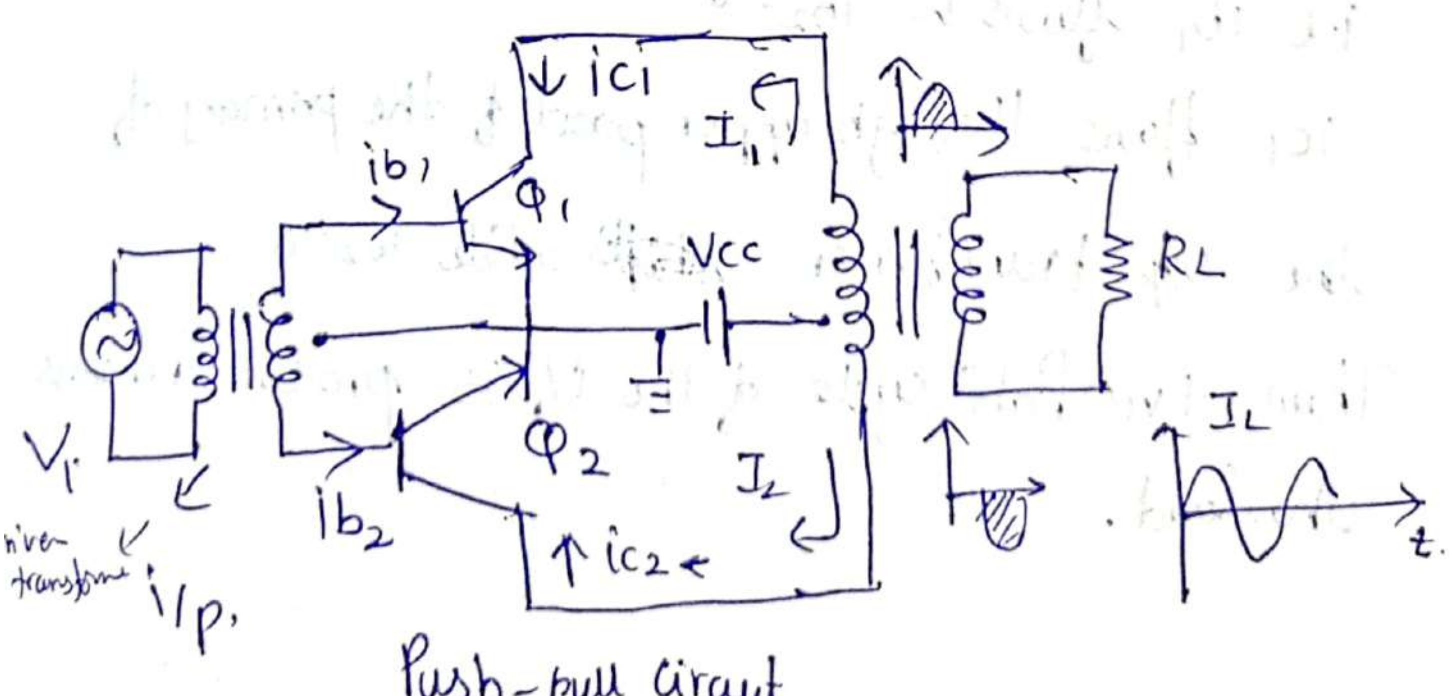
(i) Push-pull - class B Amplifiers.

(Both the transistors are of same type either pnp (or) npn.

(ii) Complementary symmetry class B Amplifier:-

→ Two transistors form a complementary pair (or) one pnp and other npn.

Push-pull Class B Amplifiers:-



Push-pull Circuit

→ Push-Pull circuit consists of two transistors as same type i.e (either) npn or pnp.

→ It consists of two centre-tapped transformers.

(i) Input Transformer (Driver transformer)

→ The i/p signal is given to the primary.

→ Due to the centre tap, it produces two signals which are 180° out of phase with each other.

→ These two signals drive the transistor.

(ii) Output Transformer:-

→ It couples the ac o/p signal from the collector to load.

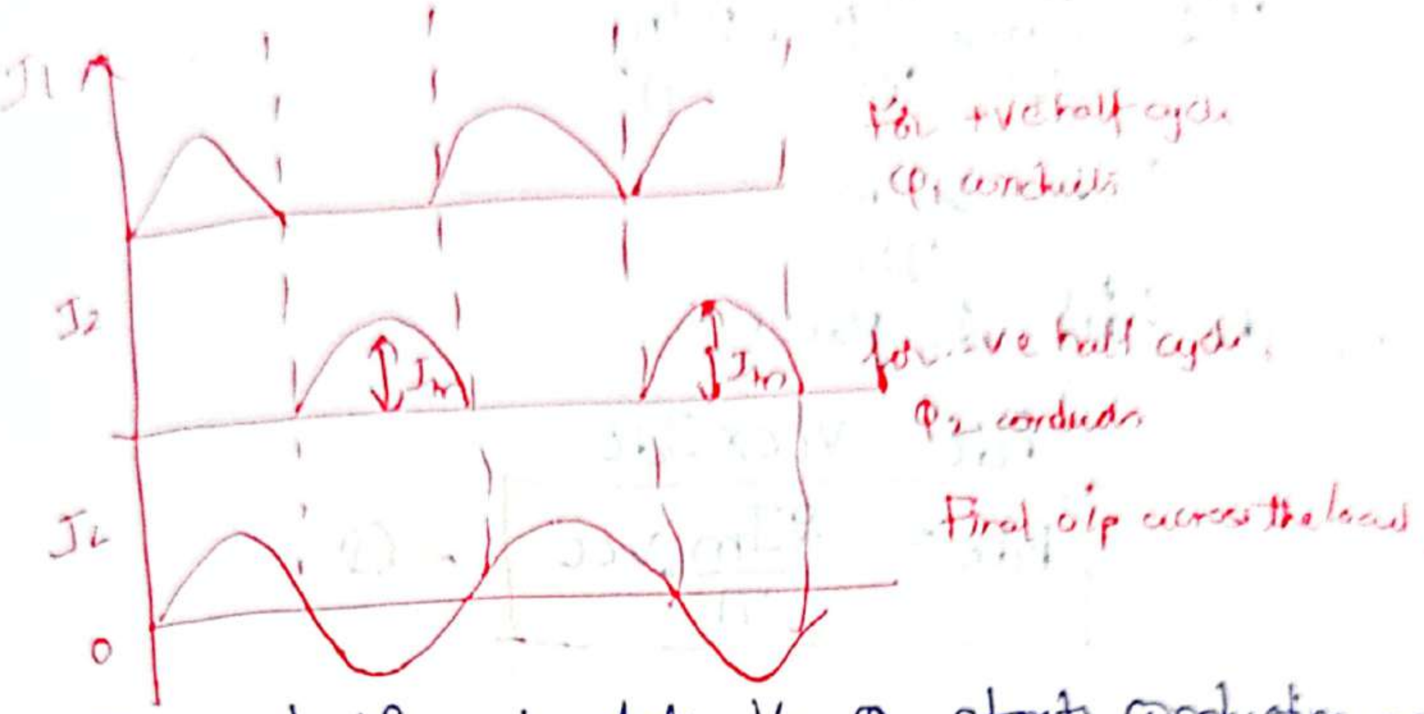
Operation:-

→ For +ve half cycle of the i/p, Q_1 starts conducting and Q_2 is in OFF condition.

i.e i_{b1} flows & $i_{b2} = 0$.

i_{c1} flows through upper part of the primary of the o/p transformer ~~while~~ while $i_{c2} = 0$.

Thus +ve half cycle of the i/p is produced across the load.



→ For -ve half cycle of the i/p, Q_2 starts conducting and Q_1 is in off state.

i.e i_{b2} flows & $i_{b1} = 0$

→ i_{c2} flows through lower part of the primary of the o/p transformer.

→ Thus the negative half cycle is produced across the load.

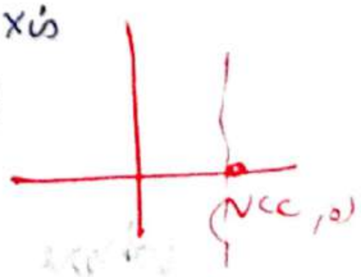
Full cycle is obtained across the load.

DC operation:

The Q-point is adjusted on the X-axis

$V_{CEQ} = V_{CC}$

$I_{CQ} = 0$ $\quad [V_{CC}, 0]$



No DC base bias voltage.

→ The two currents (I_1 & I_2) by the transistors are in same direction $\left[I_1 = I_2 = \frac{I_m}{\pi} \right]$

$$I_{dc} = \frac{I_m}{\pi} + \frac{I_m}{\pi} = \frac{2I_m}{\pi}$$

$$I_{dc} = \frac{2I_m}{\pi}.$$

$\therefore$ The total dc power is

$$P_{dc} = V_{CC} \times I_{dc}$$

$$\boxed{P_{dc} = \frac{2I_m}{\pi} \cdot V_{CC}} \quad — (1)$$

AC operation:-

(i) AC o/p power (P_{ac}):-

The o/p current & voltage peak values are I_m & V_m.

The Rms values of current & voltage are

$$V_{rms} = \frac{V_m}{\sqrt{2}} \quad \& \quad I_{rms} = \frac{I_m}{\sqrt{2}}$$

The ac o/p power is

$$P_{ac} = \frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}} = \frac{V_m I_m}{2}$$

$$P_{ac} = \frac{V_m^2}{2R_L'} \text{ (or) } \frac{I_m^2 R_L'}{2} \quad — (2)$$

where

$$R_L' = \frac{V_m}{I_m}$$

↳ Slope of ac load line

Efficiency :-

The efficiency of the Class B amplifier is

$$\eta\% = \frac{P_{ac}}{P_{dc}} \times 100$$

$$= \frac{V_m I_m/2}{\frac{2 I_m}{\pi} V_{cc}} \times 100$$

, V_{m}

$$\boxed{\eta\% = \frac{\pi}{4} \cdot \frac{V_m}{V_{cc}} \times 100}$$

Maximum Efficiency:-

For max efficiency $V_m = V_{cc}$

$$\eta_{max}\% = \frac{\pi}{4} \times \frac{V_{cc}}{V_{cc}} \times 100$$

$$\boxed{\eta_{max}\% = 78.5\%}$$

Power dissipation:-

The difference btw ac o/p power & dc o/p power is known as power dissipation

$$P_d = P_{dc} - P_{ac}$$

$$\left[\because R_L' = \frac{V_m}{I_m}\right]$$

$$= \frac{2 I_m}{\pi} V_{cc} - \frac{V_m I_m}{2}$$

$$\boxed{P_d = \frac{2}{\pi} V_{cc} I_m - \frac{V_m^2}{2R_L'}}$$

Maximum Power dissipation

$$\frac{dP_d}{dV_m} = 0$$

$$\frac{2V_{cc}}{\pi R_L'} - \frac{2V_m}{2R_L'} = 0$$

$$\boxed{V_m = \frac{2}{\pi} V_{cc}} \rightarrow \text{condition}$$

$$(P_d)_{max} = \frac{2}{\pi^2} \frac{V_{cc}^2}{R_L'}$$

Advantages:-

→ η = 78.5% higher efficiency than class-A

→ less power distortion

→ The effects of ripple voltages can be reduced.

→ Reduced harmonic distortion

→ As the dc currents flow in opposite direction through the primary winding, no dc saturation of the core.

Dis:

→ Two centre-tape transformers are used.

→ Bulky & costlier → (cross-over distortion).

→ Poor frequency response.

→ Applications

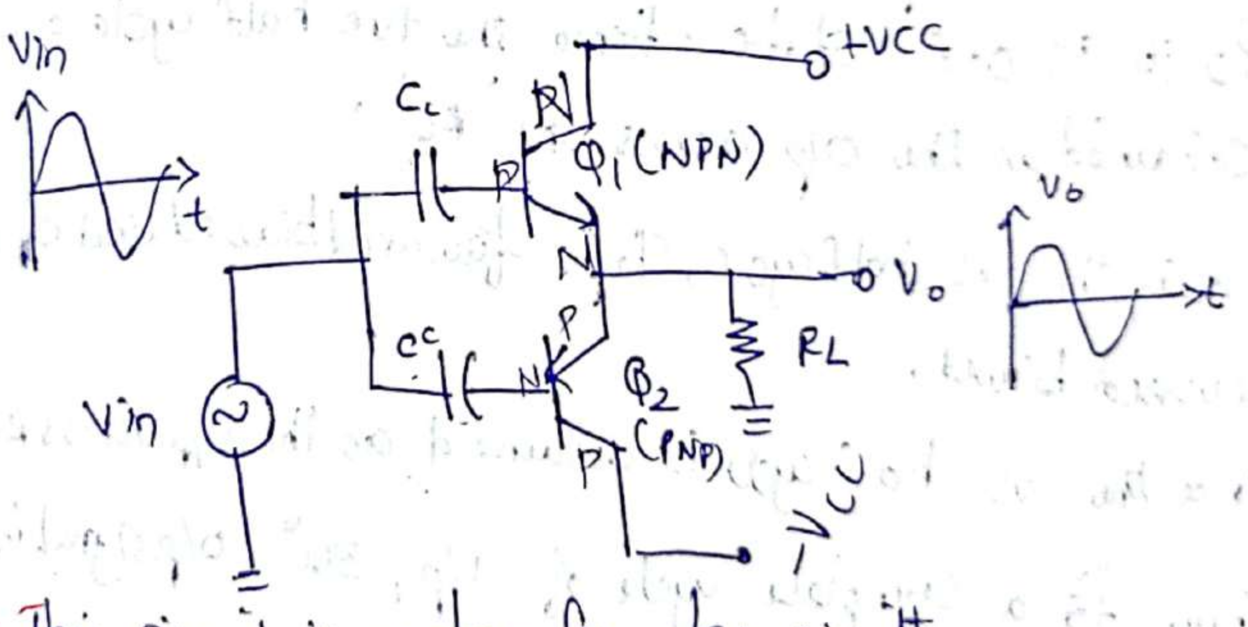
→ Used in high power applications such as audio power amplifiers.

→ Used in PA System.

Complementary Symmetry class B Amplifier

→ In this amplifier, a complementary pair of two transistors, is used to get 360° o/p.

"one NPN and other PNP."



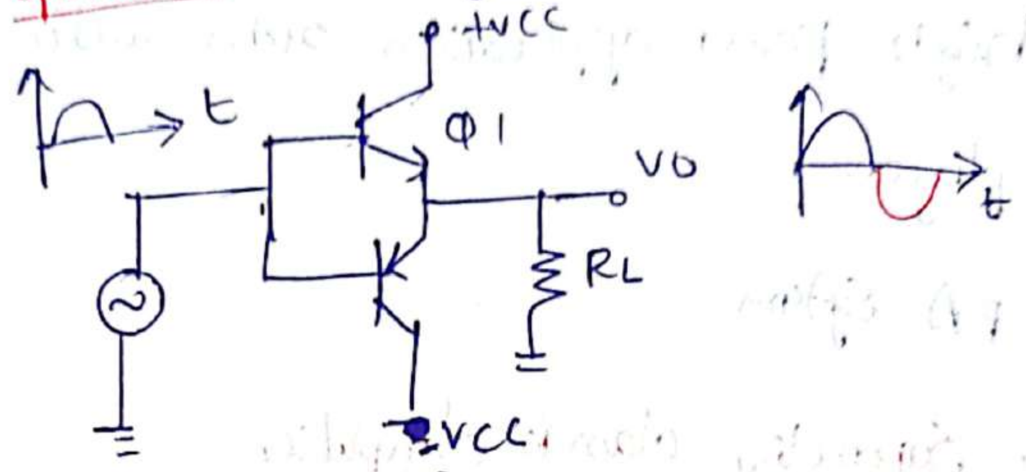
→ This circuit is a transformerless circuit.

* $Q_1 \rightarrow$ NPN ; $Q_2 \rightarrow$ PNP

→ The common collector configuration is used to provide Impedance matching for maximum power transfer.

→ The circuit is derived from a dual power supply of ±Vcc.

Operation :- i/p is given to both Q1 and Q2



→ During the +ve half cycle of i/p signal, the B-E junction of Q_1 is forward biased (i.e) driven into active region and starts conducting (Q_1–ON)

↳ As Q_2 is complementary type, Q_2 is reverse biased (ie) Q_2 is in off state. Hence the +ve half cycle can be obtained as the o/p across the 'RL'.

→ During the -ve half cycle, Q_2 is forward biased and Q_1 is reversed biased.

Hence the -ve half cycle is obtained as the o/p across RL.

Thus for a complete cycle of i/p, 360°, o/p signal is obtained.

Efficiency -

o/p ac power is $P_{ac} = \frac{V_{CC}^2}{2R_L}$

→ The i/p dc power is $P_{dc} = \frac{2V_{CC}^2}{\pi R_L}$

$\frac{2 I_m}{\pi} V_{CC}$ $= 2 \frac{V_m}{R_L} V_{CC}$ $\frac{2V_{CC}^2}{\pi R_L}$

when $V_m = V_{CC}$, the o/p power is max.

$$\text{Efficiency } (\eta\%) = \frac{P_{ac}}{P_{dc}} \times 100$$

$$= \frac{V_{CC}^2}{2R_L} \times \frac{\pi R_L}{2V_{CC}^2} \times 100$$

$$\eta\% = 78.5\%$$

Advantages:-

→ Transformerless circuit

→ less weight & low cost

→ Improved frequency response

→ Common collector configuration is used to provide impedance matching for maximum power transfer.

→ High efficiency 78.5%.

Disadvantages

→ Two seperate supply voltages ($\pm V_{CC}$) are required.

→ Cross-over distortion

Comparision :-

Push Pull	Complementary
* Both the transistors are of same type (NPN or PNP	Two transistors form a complementary pair on NPN other PNP
* Transformer is used	It is transformerless circuit

* Bulky & costlier	less weight & low cost
* Impedance matching is possible due to transformer	Due to common collector configuration
* Dual power supply is not required (+Vcc)	Dual power supply is required (±Vcc)
* Poor frequency response	Improved frequency response

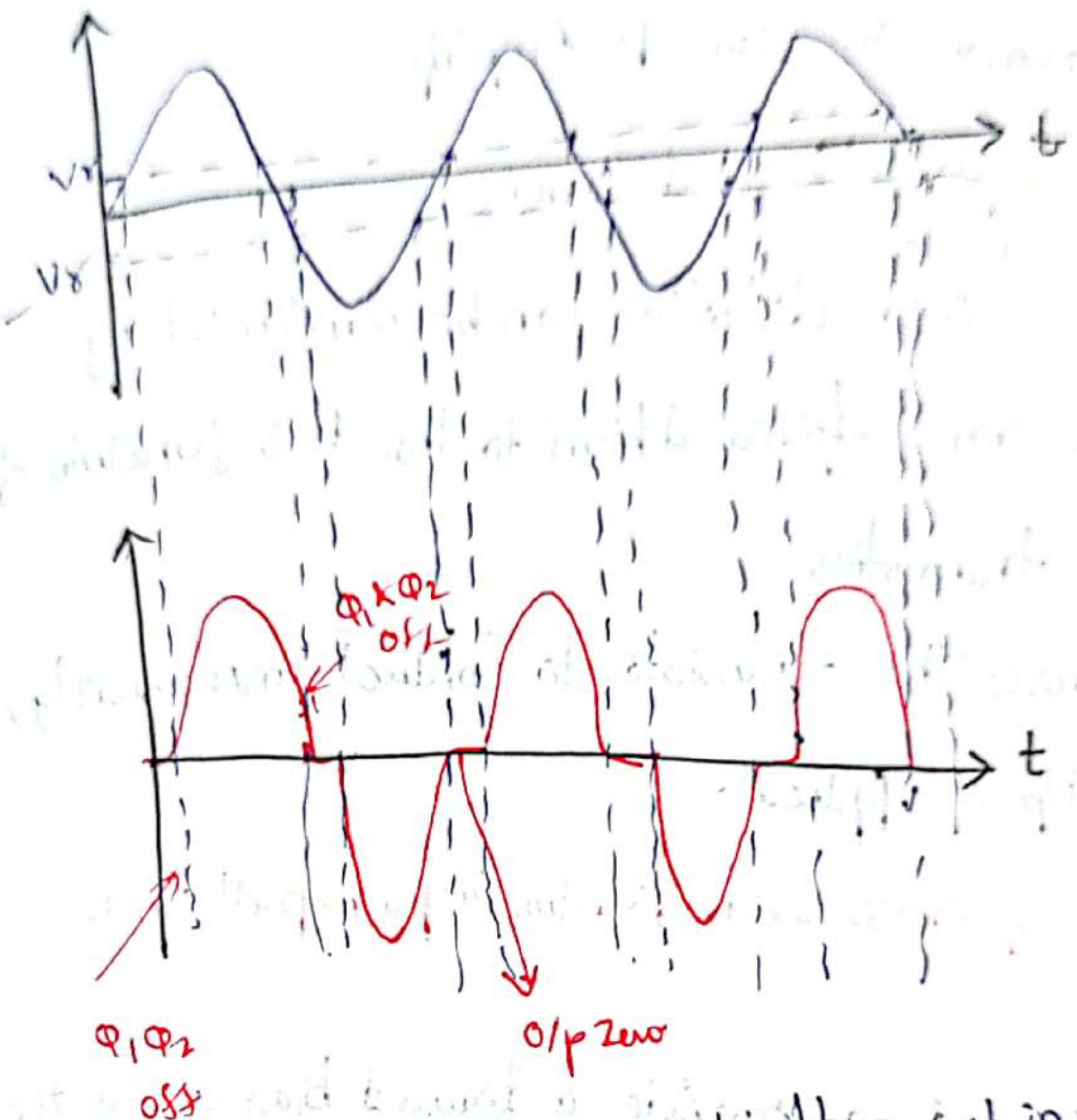
Cross-over distortion

The difference b/w the o/p and i/p of an amplifier is known as distortion.

→ The non-linearity of the i/p characteristics of the transistor causes cross-over distortion.

→ Cut-in voltage of the transistor is the basic reason for cross-over distortion.

Ge → 0.3V & Si = 0.7V



→ When the i/p voltage is greater than cut in voltage of the transistor, transistor starts conducting (i.e) forward biased

$V_{in} > V_\gamma$.

→ When $V_i < V_\gamma$, there is a time period btw the crossing of the half cycles for which the o/p is zero.

→ Hence the o/p signal gets distorted, hence called as cross-over distortion.

→ Due to cross-over distortion each transistor conducts for less than a half cycle rather than the

Complete half cycle.

→ It is common in class-B Amplifier.

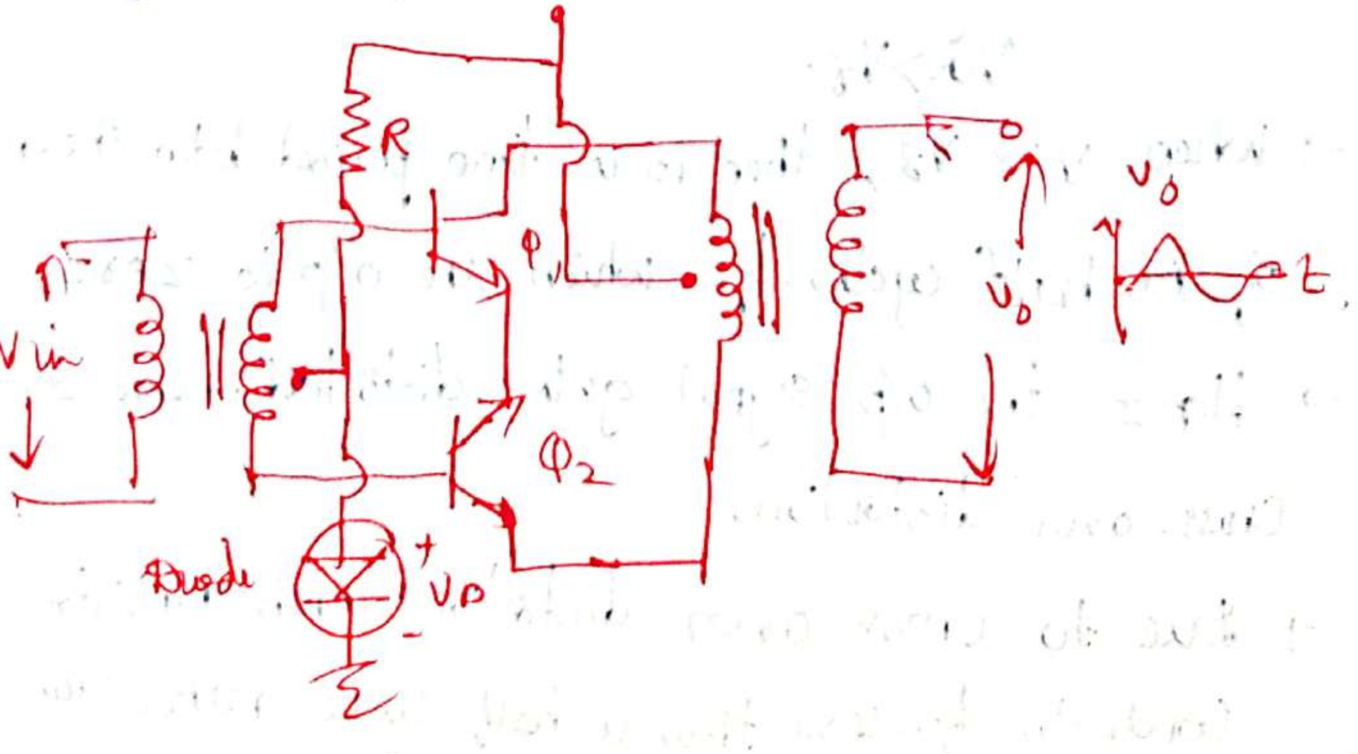
Elimination of Cross-over distortion

→ The cross-over distortion can be eliminated by applying a small forward bias to the B-E junction of both the transistors.

→ It causes the transistor to conduct immediately, when the i/p is applied.

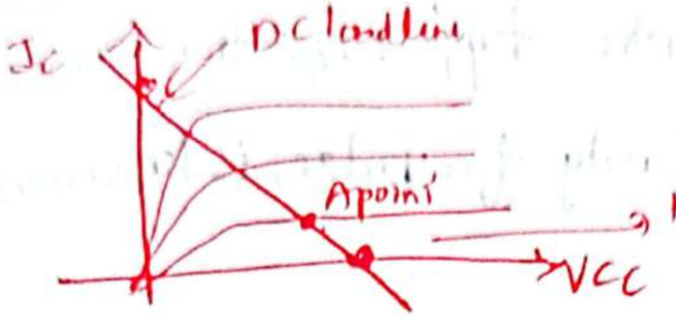
Elimination of cross-over distortion in push-pull class B Amplifier

Diode is used to provide a forward bias across the B-E junction of the transistor.

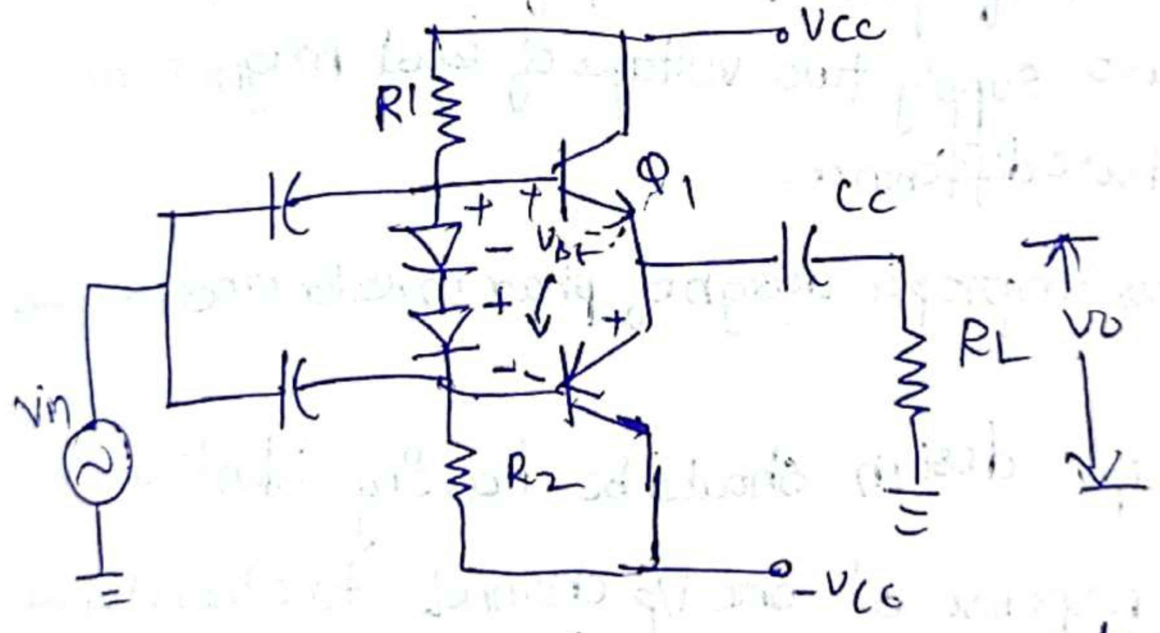


→ The voltage drop across the diode D, VD = Vγ → cut-in voltage of the BE-junction

→ Due to the forward biased diode, the Q-point shifts upwards the loadline.



elimination of Cross over distortions in Complementary symmetry Class-B Amplifier



Diodes are used to provide a required temperature stability.

i.e Maintain the necessary biasing to overcome the cross-over distortion when temperature changed.

Phase Inverters

→ A circuit that changes the phase of a signal by 180° as required for feeding a push-pull amplifier stage without using a centre-tapped i/p transformer or for changing the polarity of a pulse, is known as phase inverter.

→ The primary function of the phase inverter circuit is to change the phase of a signal by 180° and it is mostly used a the i/p of push-pull amplifier. So the phase inverter must supply two voltages of equal magnitude with 180° phase difference.

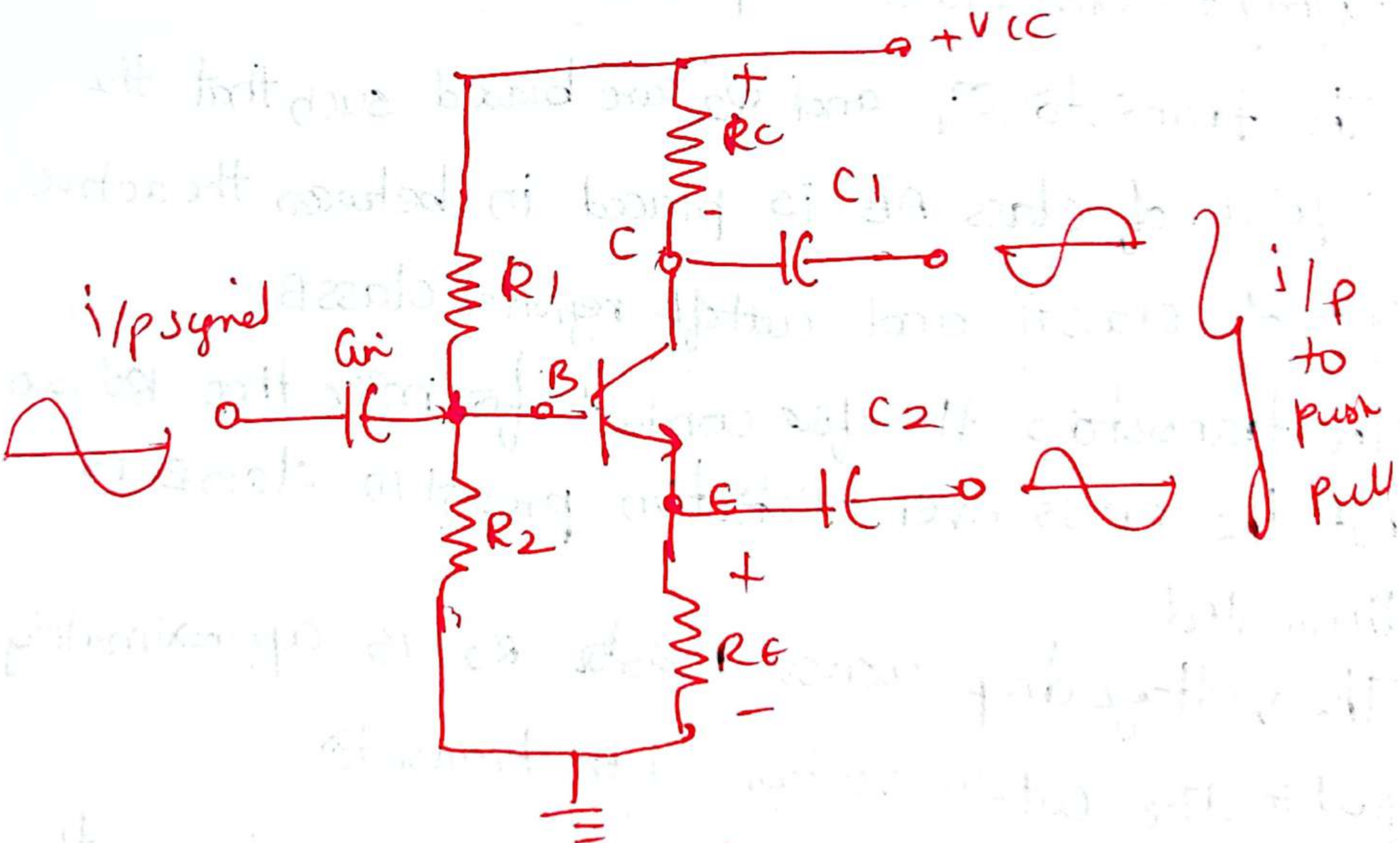
→ This causes improper design of phase inverter and push-pull amplifier.

→ The principle design should be having identical frequency response of one i/p channel to other channel of push-pull amplifier.

→ One simplest method of using phase-inverter is a transformer with centre-tapped used at pull push amplifier i/p stage.

→ We should see that the voltage across secondary windings must be equal.

Phase Inverter Circuit Diagram



Phase- Splitter Circuit for Push Pull Amplifier.

→ The transformer forms a good inverter which supplies power to i/p of Push-Pull Amplifier, but has several disadvantages i.e cost, space (Bulkier)

→ Other simple form is phase inverter with transistor that overcomes drawbacks of transformer ~~i~~ phase inverter.

→ Transistor phase inverter circuit also provides two equal o/p signals with 180° of phase shift.

→ Here R_3 & R_4 resistance & capacitors C_1 & C_2 are equal, while signal V_{S2} at the emitter terminal is in phase with i/p signal V_S while other signal V_{S1} at the collector terminal is out of phase with V_S.

→ Approximately same current flows through R_3 & R_4 hence and hence if R_3 & R_4 are equal, then ac o/p voltages from the collector and from emitter are equal in magnitude and 180° out of phase.

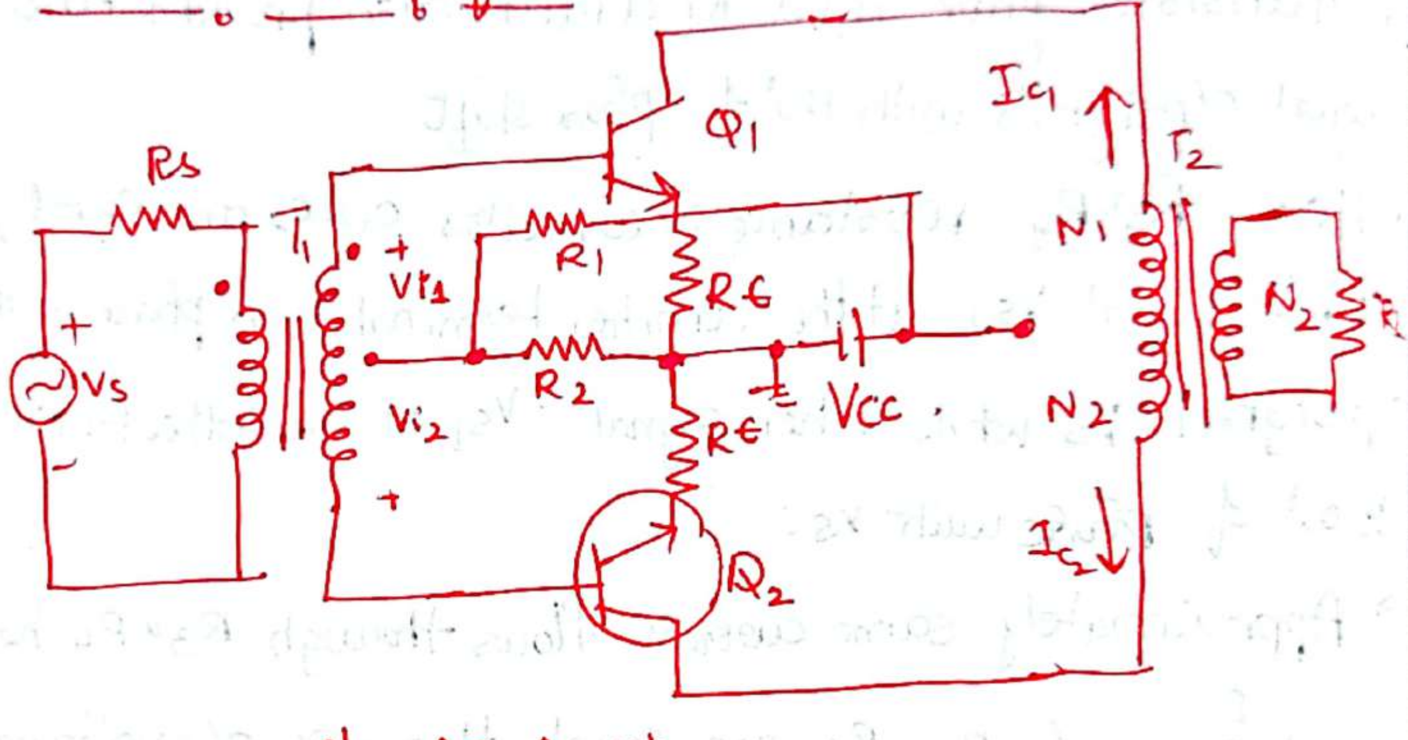
Limitations:-

(i) There is no voltage gain as the circuit is of emitter follower type in which the gain is unity.

(ii) The o/p voltages V_{S1} & V_{S2} will result in imbalance when the emitter & collector currents are not equal.

(iii) This phase inverter could be used only to drive class A push-Pull circuit in which the load is constant on each O/P whereas I could not be used in class B push-Pull amplifier, since o/p voltages are V_{S_1} & V_{S_2} are equal only if it drives a resistive load over the full cycle.

Class AB Amplifier



Class AB Amplifier

→ class AB amplifier overcomes the problem of crossover distortion present in class B amplifiers, in which a small current flows even at zero input signal level.

→ The circuit of a class AB push-pull amplifier is shown above. The circuit, which is essentially the same as that of class B amplifier, has additional RE resistors referred to as the emitter-stabilizing resistors.

→ This biases the transistor away from class B slightly towards class A operation.

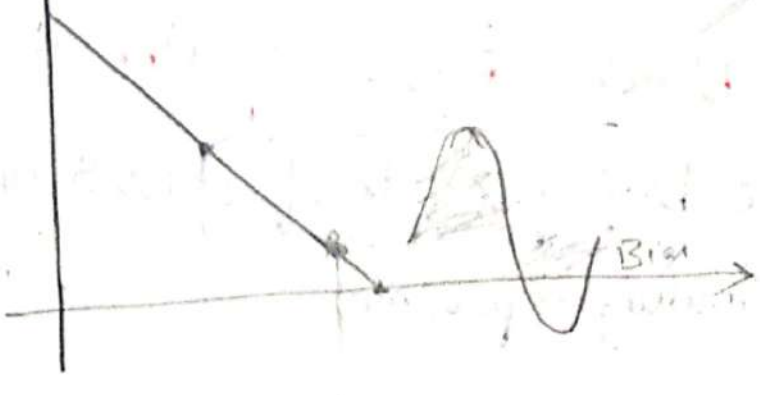
→ The transistors Q_1 and Q_2 are biased such that the Q point of Class AB is placed in between the active region of class A and cutoff region of class B.

→ The transistors therefore conduct for more than 180°, so that the cross over distortion present in class B is eliminated.

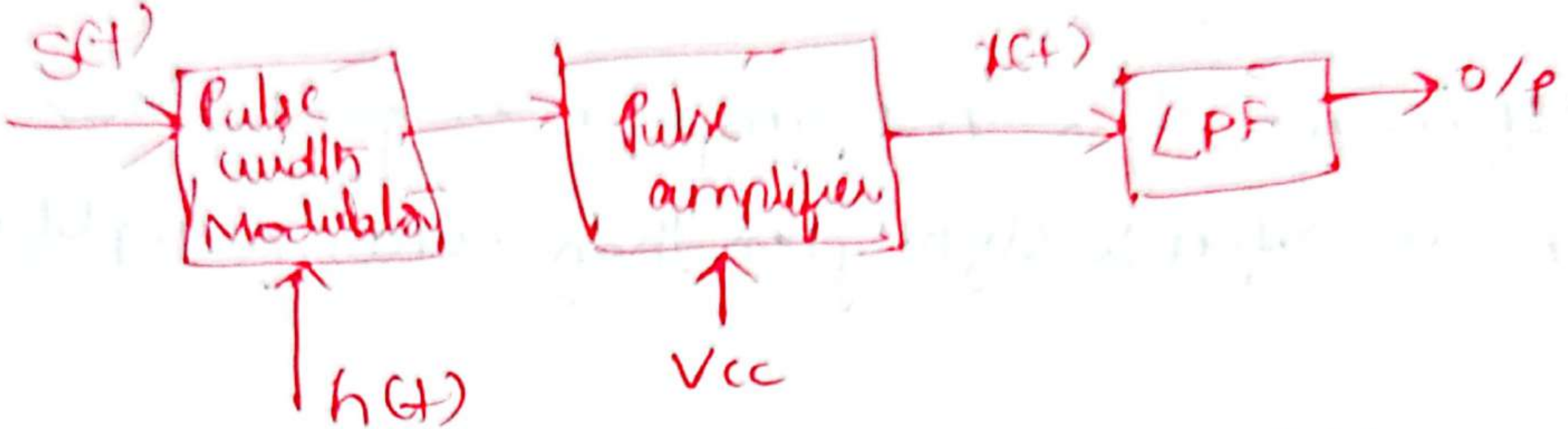
→ The voltage drop across resistor R_2 is approximately equal to the cut-in voltage of the transistor.

→ When the ac signal is applied to the base, the collector current starts flowing immediately, But there will be decrease in the o/p power due to the -ve feedback effect.

→ The efficiency of class A B amplifier is greater than class A amplifier & slightly less than class B amplifier



Class S power Amplifier :



Block diagram of Class S amplifier

→ The basic principle of class S amplifier includes the following operations

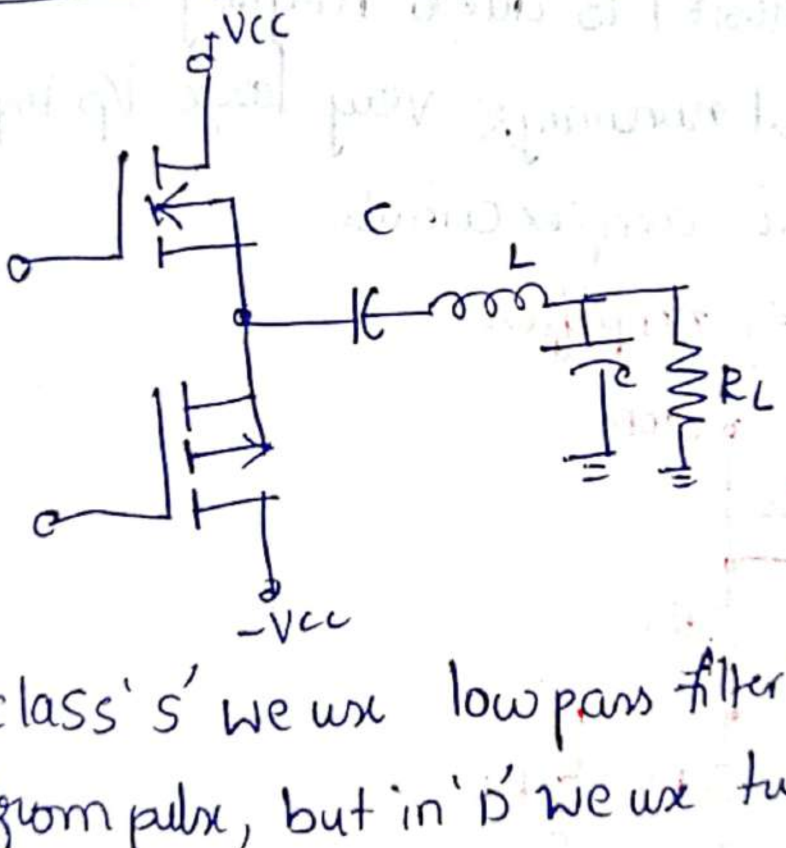
(1) The i/p is PWM signal.

(2) The pulses are amplified by high efficiency pulse amplifier like class D amplifier.

(3) The demodulation using LPF of the amplified signal.

(4) Difference b/w class S & D is no feedback circuit.

Simplified class S amplifier



class 'S' we use low pass filter to demodulate the signal from pulse, but in 'D' we use tuned circuit.

→ Efficiency is approximately 100%.

Entire operation is similar to 'D' except lowpass filter

L & C is lowpass, O/p across RL.

MOSFET Power Amplifiers

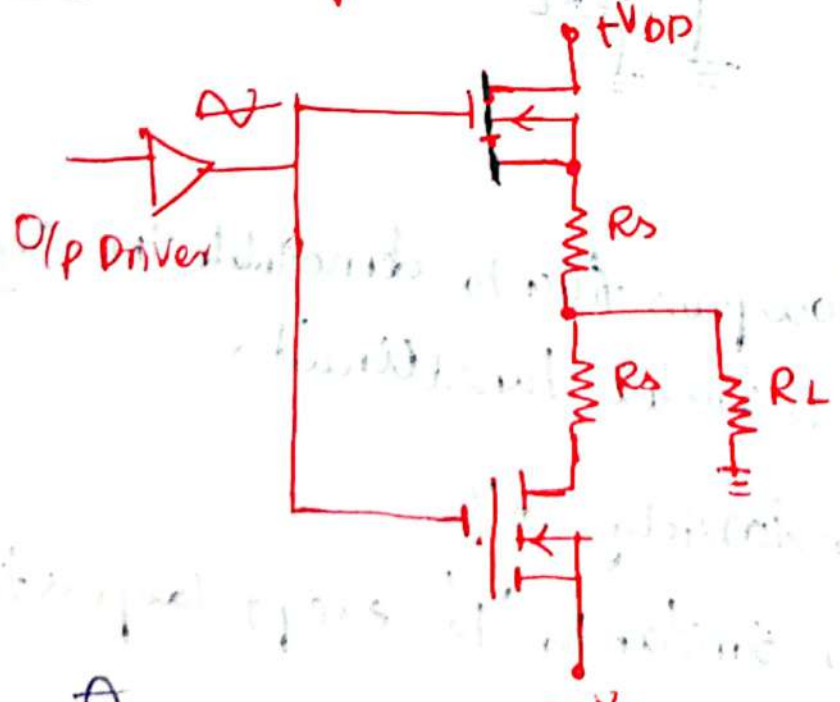
→ Power Amplifiers are designed to switch large currents ON & OFF using MOSFET devices.

→ MOSFET based Class-D amplifiers are commonly used.

→ The advantage of MOSFET is switching to turn off time is not delayed by minority-carrier storage, as it is in a BJT.

→ Further current in MOSFET is due to majority carriers only and no thermal runaway & very large i/p impedance which can drive complex circuits.

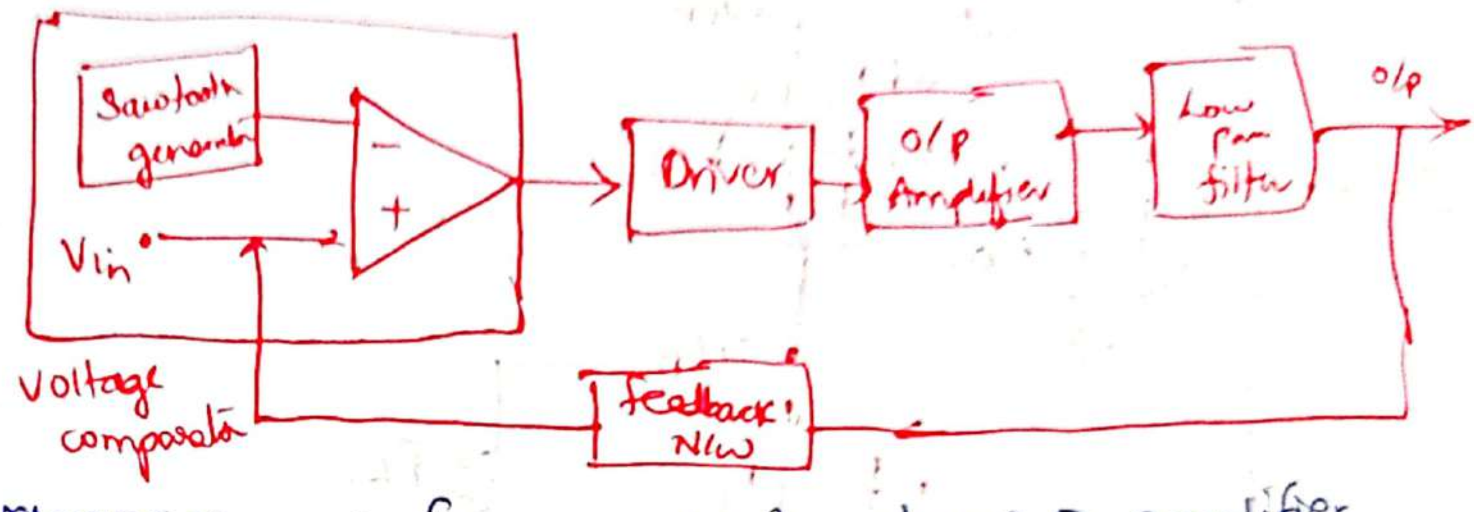
Complementary MOSFET Amplifiers



A complementary MOSFET Amplifier

MOSFET Amplifiers using complementary o/p devices for class B operation is shown in above fig where availability of complementary pairs makes the design simpler as the need for transformer is eliminated.

MOSFET based Class-D Amplifier



MOSFET amplifiers are used in class ~~A~~ D amplifier, here the active device is switched "ON" & "OFF" so that it is held in the linear range for zero timing of each cycle of i/p sine wave. For BJT we use complementary symmetry BJT.

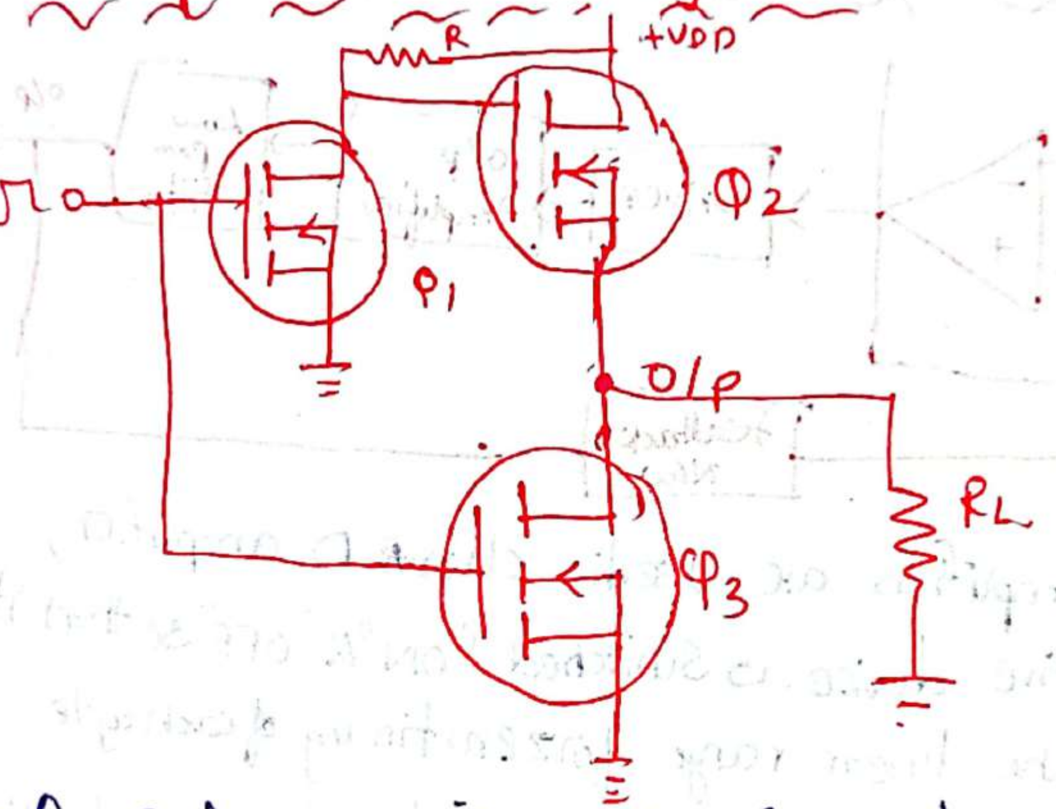
From block diagram of D-Amplifier, analog signal modulates the saw tooth waveform so that a pulse width modulated o/p is obtained, which drives the class D o/p amplifier, causing it to switch ON & OFF as the pulses switch btw high & low.

→ Here class D amplifier must have a filter circuit to extract the signal to be amplified from the pulsed waveform.

→ As signal has many frequency components, a LPF having cuttoff frequency near to highest signal frequency is better choice.

→ LPF filter supress high freq components & o/p is exact replica of i/p.

Totem pole MOSFET switching device



→ A Switching circuit called Totem pole with MOSFET as switching device as shown in above fig.

→ The Totem pole inverts the i/p, that is the o/p is high when i/p is low & vice versa.

→ when i/p high Q_1 & Q_3 are ON acting as closed switches & Q_2 is OFF acting as open switch, o/p is low.

→ when i/p is low, ϕ_1 & ϕ_3 are off & V_{DD} is available at the ϕ_2 switching it ON, Hence o/p is high.

Advantages:-

(1) A High efficiency (100% due to D amplifier)

(2) Minimum power dissipation

Drawbacks:-

→ Filters with sharp cuttoff frequenigs are complex in design.

→ High speed switching of large currents generates noise through electromagnetic coupling called electromagnetic interference

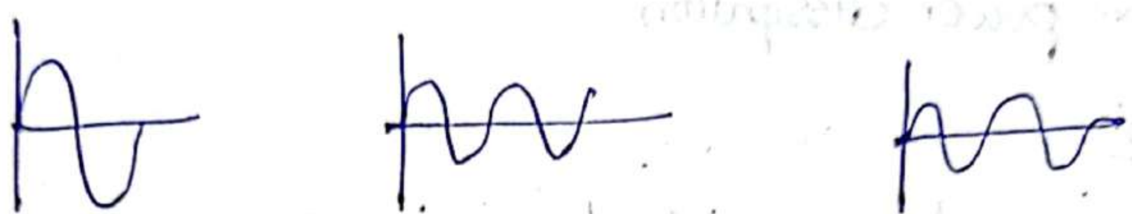
Harmonic Distortion:-

Harmonic distortion is defined as the presence of frequency components in the o/p signal which are not their in i/p signal.

If the frequency components in o/p signal is same as frequency component in the i/p signal, such type of frequency component is called fundamental frequency component

The harmonic frequency components are integer multiples of fundamental frequency components.

Let us assume 'f' is a fundamental frequency component then harmonic frequency components are $f, 2f, 3f, 4f \dots$



sin ωt — 1st fundamental
sin 2ωt — 1st harmonic
sin 3ωt — 2nd harmonic

frequency components.

→ The amplitude of frequency component is decreased if no. of harmonic components are increasing in the O/p signal.

The % of nth harmonic component distortion is defined as

$$\% D_n = \frac{B_n}{B_1} \times 100$$

B_1 = Amplitude of fundamental frequency component

B_n = Amplitude of nth harmonic freq component

$$\% D_2 = \frac{B_2}{B_1} \times 100$$

$$\% D_3 = \frac{B_3}{B_1} \times 100\%.$$

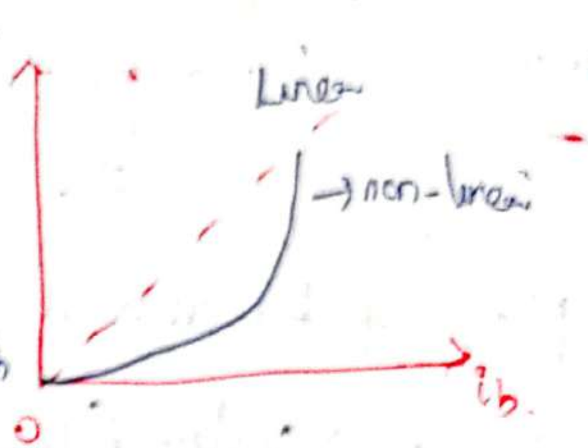
$$\% D_n = B_n / B_1 \times 100$$

Total harmonic distortion is defined as

$$D = \sqrt{D_2^2 + D_3^2 + D_4^2 + \cdots D_n^2}$$

Determination of 2nd harmonic Distortion (3 point method)

The second harmonic distortion is determined from the dynamic transfer curve using the 3 point method for small signals, which is parabolic (non-linear) in nature.



The relationship between alternating current i_c & the i/p excitation i_b is given as follows.

$$i_c = K_1 i_b + K_2 i_b^2 \quad \text{where } K_1 \,\&\, K_2 \text{ constants}$$

Substitute $i_b = I_{bm} \cos\omega t$

$$i_c = K_1 I_{bm} \cos\omega t + K_2 I_{bm}^2 \cos^2\omega t$$

where $\cos 2\omega t = 1 - 2\cos^2\omega t$.

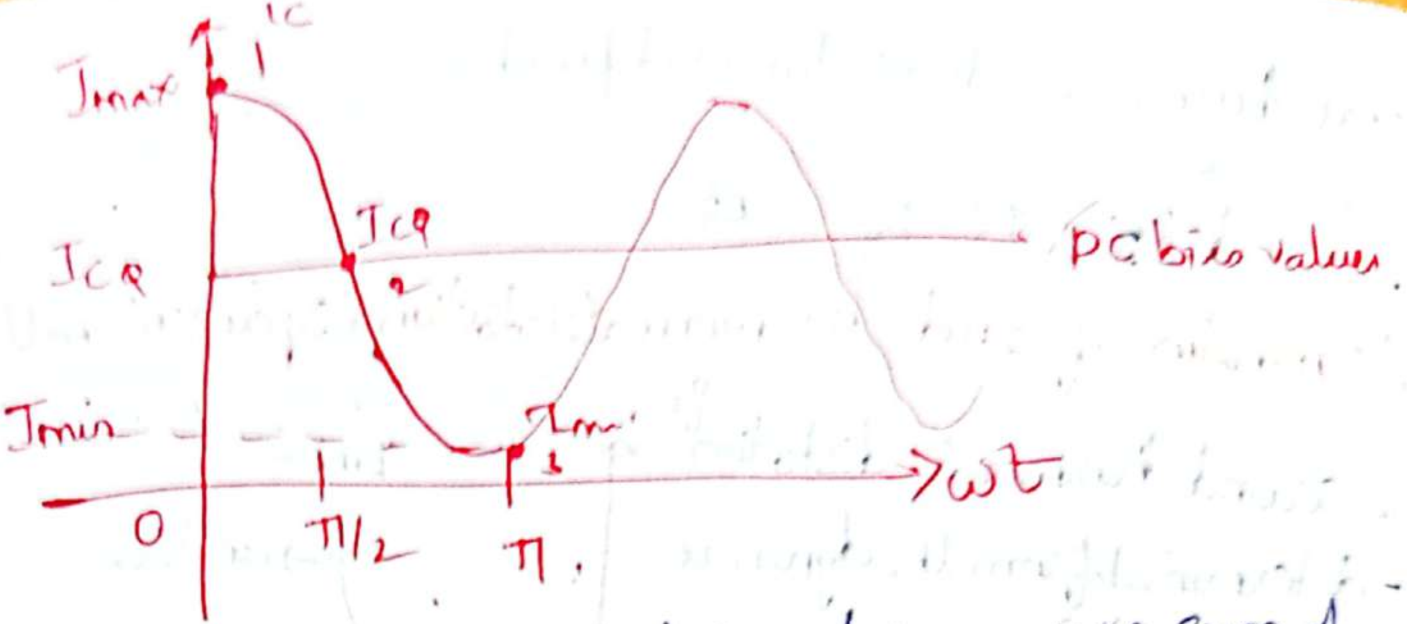
$$i_c = K_1 I_{bm} \cos\omega t + K_2 I_{bm}^2 \left(\frac{1 + \cos 2\omega t}{2}\right)$$

$$= \frac{K_2 I_{bm}^2}{2} + K_1 I_{bm} \cos\omega t + \frac{K_2 I_{bm}^2}{2} \cos 2\omega t$$

let represent i_c in terms of harmonic distortion

$$i_c = B_0 + B_1 \cos\omega t + B_2 \cos 2\omega t \quad \text{—} (1)$$

Last in eqn (1) represents secondary harmonic component



Here total current i_c can be represent as sum of d.c bias values (I_{CQ}), B_0 extra d.c component due to rectification signal, $B_1 \rightarrow$ amplitude of fundamental frequency, $B_2 \rightarrow$ amplitude of 2nd harmonic frequency.

$$\therefore \boxed{i_c = I_C + B_0 + B_1 \cos\omega t + B_2 \cos 2\omega t}$$

when $\omega t = 0$, $i_c = I_{max}$

$$I_{max} = I_{CQ} + B_0 + B_1 + B_2$$

$\omega t = \pi/2$, $i_c = I_{CQ}$.

$$\cancel{I_{CQ}} = \cancel{I_{CQ}} + B_0 + 0 + (-B_2)$$

$$\boxed{B_0 = B_2}$$

$\omega t = \pi$, $i_c = I_{min}$

$$I_{min} = I_{CQ} + B_0 - B_1 + B_2$$

$$\frac{I_{max} - I_{min}}{2} = B_1$$

$$B_2 = \frac{I_{max} + I_{min} - I_{cq}}{4}.$$

I_{max}, I_{min}, I_{cq} obtained by dynamic transfer curve of transistor,

Secondary harmonic distortion D_2 in %

$$D_2 = \frac{|B_2|}{|B_1|} \times 100\%.$$

if $I_{cq} = \frac{I_{max} - I_{min}}{2}$ then B_2 & $B_2 = 0$, no distortion.

Higher order harmonic Distortion generation

As non linearity present in dynamic characteristics ↑, the order of harmonic distortion also increases!

ic due to higher ~~orders~~ order harmonics

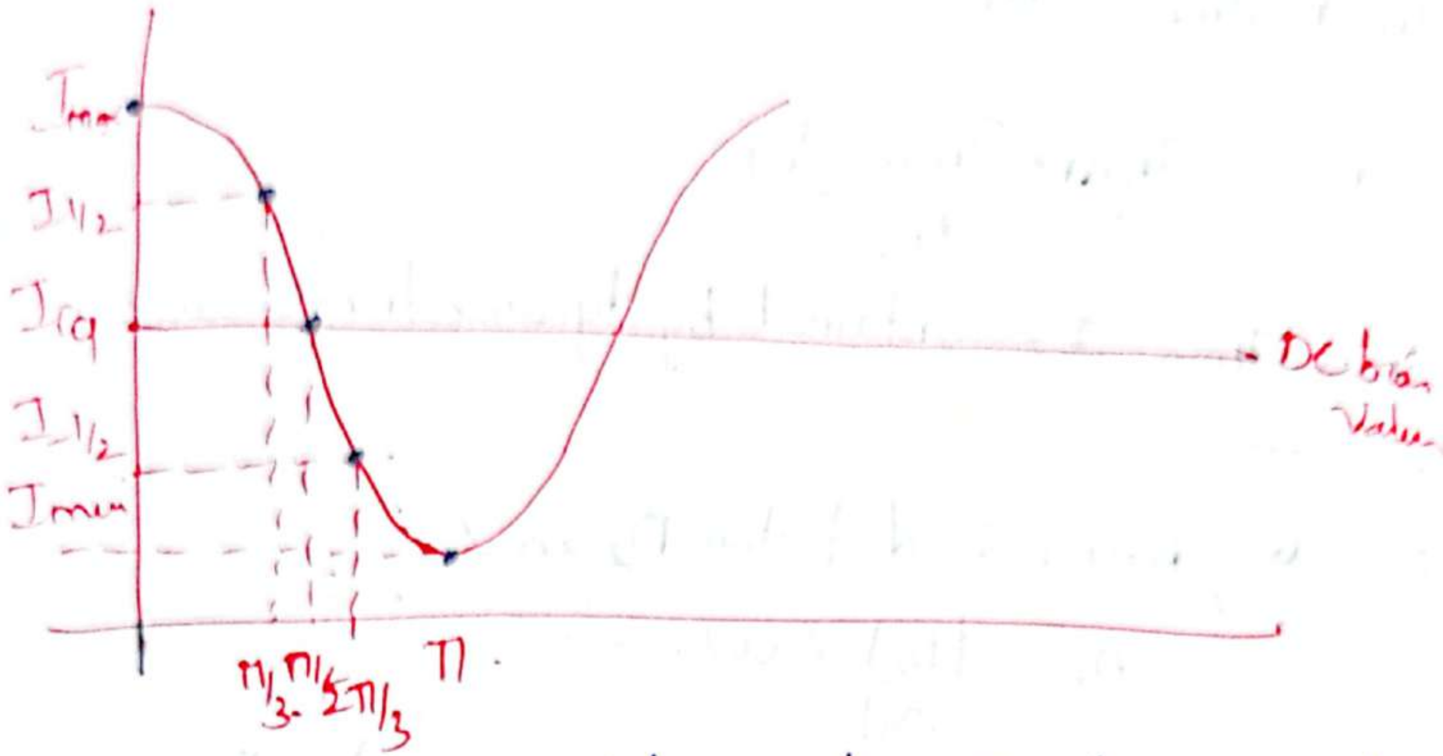
$$i_c = K_1 i_b + K_2 i_b^2 + K_3 i_b^3 + K_4 i_b^4$$

Sub $i_b = I_{Bm} \cos\omega t$

$$i_c = K_1 I_{Bm} \cos\omega t + K_2 I_{Bm}^2 \cos^2\omega t + K_3 I_{Bm}^3 \cos^3\omega t + K_3 I_{Bm}^4 \cos^4\omega t$$

Substitute $\cos^2\omega t, \cos^3\omega t, \cos^4\omega t$ doing trigonometric operations

$$i_c = B_0 + B_1 \cos\omega t + B_2 \cos 2\omega t + B_3 \cos 3\omega t + B_4 \cos 4\omega t$$



∴ i_c including D.C bias values, B_0, harmonics, fundamental harmonics

$$i_c = I_{CQ} + B_0 + B_1\cos\omega t + B_2\cos 2\omega t + B_3\cos 3\omega t + B_4\cos 4\omega t \quad — (1)$$

$(I_{CQ} + B_0) \rightarrow$ d.c component.

(1) at $\omega t = 0$, $i_c = I_{max}$.

$$I_{max} = I_{CQ} + B_0 + B_1 + B_2 + B_3 + B_4.$$

at $\omega t = \pi/3$, $i_c = I_{1/2}$

$$I_{1/2} = I_{CQ} + B_0 + 0.5B_1 - 0.5B_2 - B_3 - 0.5B_4.$$

at $\omega t = \pi/2$, $i_c = I_{CQ}$

$$I_{CQ} = I_{CQ} + B_0 - B_2 + B_4.$$

at $\omega t = 2\pi/3$, $i_c = I_{-1/2}$

$$I_{-1/2} = I_{CQ} + B_0 - 0.5B_1 - 0.5B_2 + B_3 - 0.5B_4$$

at $\omega t = \pi$, $i_c = I_{min}$

$$I_{min} = I_{CQ} + B_0 - B_1 + B_2 - B_3 + B_4.$$

By solving we get

$$B_0 = \frac{1}{6}\left(I_{max} + 2I_{1/2} + 2I_{-1/2} + I_{min}\right]$$

$$B_1 = \frac{1}{3}\left[I_{max} + I_{1/2} - I_{-1/2} - I_{min}\right]$$

$$B_2 = \frac{1}{4}\left[I_{max} - 2I_{CQ} + I_{min}\right]$$

$$B_3 = \frac{1}{6}\left[I_{max} - 2I_{1/2} + 2I_{-1/2} - I_{min}\right)$$

$$B_4 = \frac{1}{12}\left[I_{max} - 4I_{1/2} + 6I_{CQ} - 4I_{-1/2} + I_{min}\right).$$

$$\therefore \quad D_n = \frac{|B_n|}{B_1}).$$

Thermal Resistance & Thermal stability

→ The transistor is a temperature dependent device. In order to keep the temperature within the limits, the heat generated must be dissipated to the surroundings.

→ the heat within the transistor is produced at the collector junction. If the temperature exceeds the limit, junction is destroyed. for Si transistor, the temperature is in the range of 150° to 225°C.

→ For Ge transistor, the temperature is 60°c to 100°c.

Let T_A°c be the Ambient temperature i.e, temperature of the surrounding air.

T_j°c be the temperature of collector-Base junction.

$$T_j - T_A = \Theta P_D$$

Θ → Thermal resistance,

Θ_{j-A} → total thermal resistance from transistor junction to the ambient temperature.

Thermal stability:-

→ For self bias circuit, assume the transistor is biased in active region. Then power is given by

$$P_C = V_{CB} I_{CB} - V_{CE} I_{CE}$$

Assume I_C & I_E currents are in eq n. then

$$P_C = I_C V_{CC} - I_C^2 (R_C + R_E)$$

here $V_{CC} I_C$ is power drawn by supply; Then diff P_C w.r.t I_C is

$$\frac{\partial P_C}{\partial I_C} = V_{CC} - 2 I_C (R_C + R_E)$$

Condition to avoid thermal runaway is

$$\frac{\partial P_c}{\partial T_j} = \frac{\partial P_c}{\partial I_c} \times \frac{\partial I_c}{\partial T_j} < \frac{1}{\theta_{j-A}} \quad \text{---(1)}$$

where $S = \frac{\partial I_C}{\partial I_{CO}} = \frac{\partial I_C}{\partial T_j} \times \frac{\partial T_j}{\partial I_{CO}}$

or $\frac{\partial I_C}{\partial T_j} = S \times \frac{\partial I_{CO}}{\partial T_j}$

Sub $\frac{\partial P_C}{\partial I_C}$ & $\frac{\partial I_C}{\partial T_j}$ in eq(1).

$$\frac{\partial P_C}{\partial T_j} = \left(V_{CC} - 2I_C(R_C+R_E)\right) \times S \times \frac{\partial I_{CO}}{\partial T_j} \quad \left(\because \frac{\partial I_{CO}}{\partial T_j} = 0.07 I_{CO}\right)$$

$$= V_{CC} - 2I_C(R_C+R_E) \times S \times 0.07 I_{CO}$$

So the condition for thermal stability is

$$V_{CC} - 2I_C(R_C+R_E) \times S \times 0.07 I_{CO} < \frac{1}{\theta_{j-A}}$$

in selfbias $V_{CE} < \frac{1}{2} V_{CC}$, then stability is automatically ensured.

Hence condition for thermal stability is

$$\frac{\partial P_c}{\partial T_j} < \frac{\partial P_d}{\partial T_j}$$

max power dissipated

heat released at collector

Heat sinks:-

→ Heat sink is basically a large metallic heat conducting device, which is placed near a transistor, it cooling increases over effective surface area.

Requirements of heat sinks:-

For transistor operating at 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 temperature of the transistors controlled by preventing thermal breakdown.

→ Increase in I_C increase I_{CO}, which results in increased power dissipation & temperature ↑, which is a cummutative process, due to this transistor break downs. In order to prevent this, heat sinks are used, which maintain low temperatures & helps low power dissipation.

→ If heat sinks are used, the heat is transfered from surface of package & from it to heat sinks and from heat sink to the ambient (surroundings).

→ Heat sink fastens the power dissipation & prevents breakdown of the device.

Types of heat sinks:-

(1) Low power transistor type

(2) High power transistor type

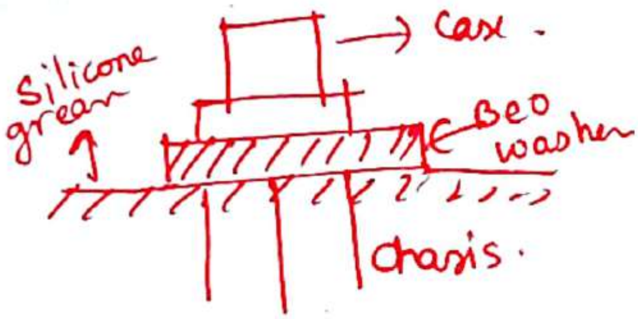
Low power transistor type:-

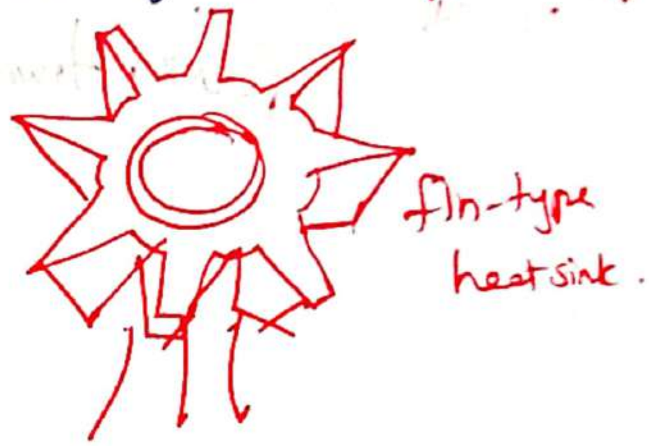
→ Low power transistor can be mounted directly on the metal chassis to increase the heat dissipation capacity.

→ The casing ~~of~~ on the transistor must be insulated from metal chassis to prevent short circuit.

→ Beryllium oxide, insulating washers are used for insulating casing from the chassis which passes good thermal conductivity.

→ Zinc oxide film ~~btw~~ using silicon compound btw washer & chassis improve the heat transform from semiconductor device to ~~case~~ the chassis.



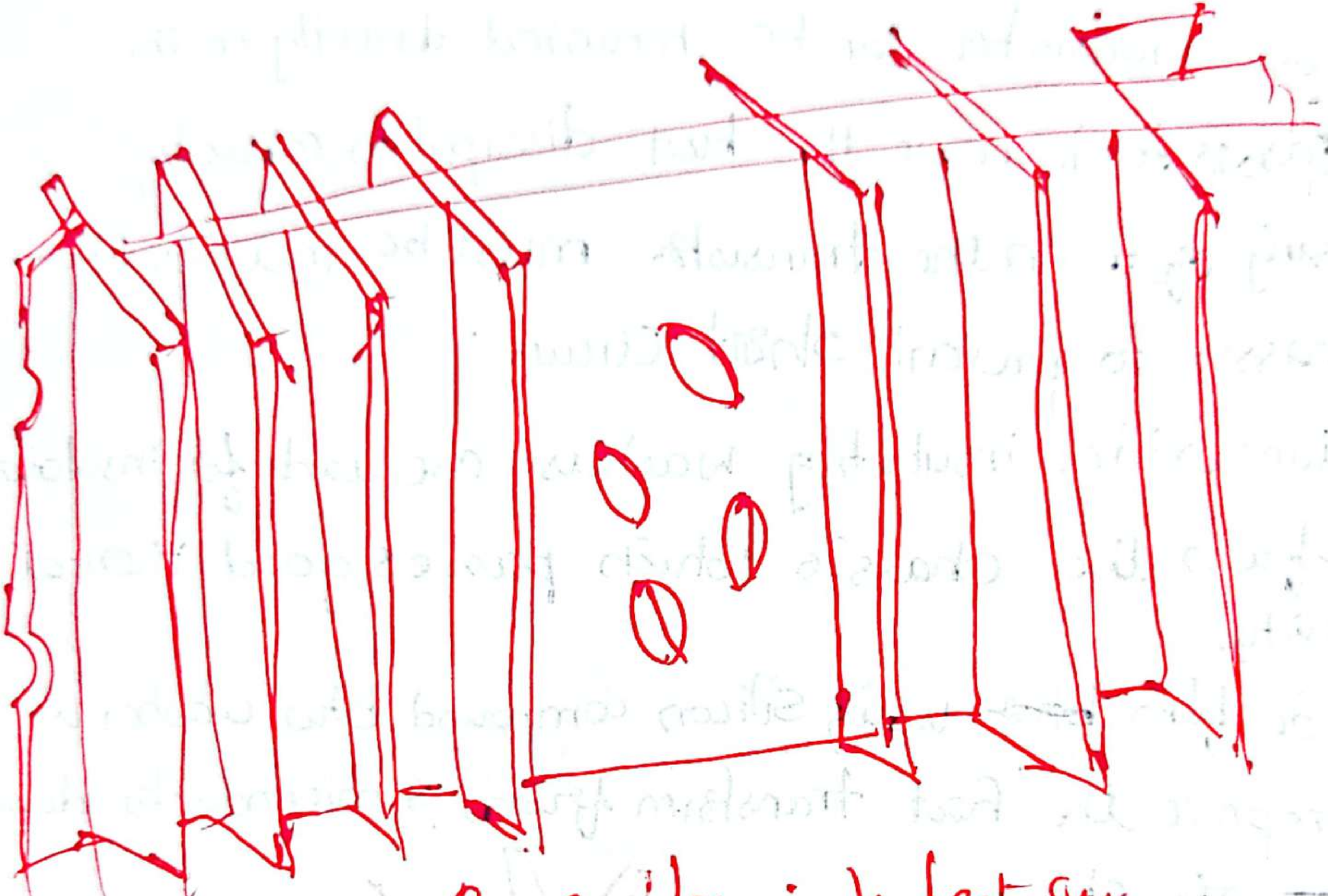


High power transistor type:-

The diamond shaped TO-3, TO-66 types are popular mounting packages of power transistors having power dissipation of 100W.

These have two leads for E & B, but the case or mounting flange of the case is collector terminal

TO-3 provides cooling by conduction, convection, radiation.



Power transistor heat sink

V Unit

Tuned Amplifier:-

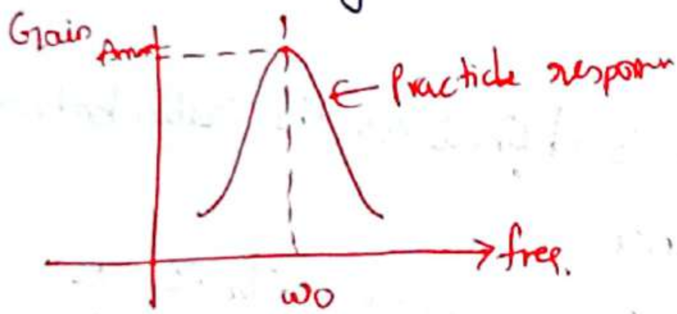
It is an electronic circuit which is used to improve the strength of the signal at desired frequencies.

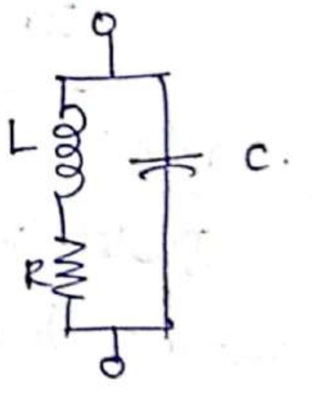
Tuned circuits are designed by replacing resistive load with tuned circuit.

Tuned Circuit:-

It consisting of basic elements such as inductor, capacitor and resistor. Sometimes it is also called as tank circuit.

The frequency response of tuned circuit is shown below with circuit diagram.





Resonant frequency:- (f_r)

It is the frequency at which rectance of inductor is equal to reactance of capacito.

$$|X_L| = |X_C|$$

$$|j\omega L| = \left|\frac{1}{j\omega C}\right|$$

$$|\omega^2| = \frac{1}{|j^2 LC|} \Rightarrow \omega^2 = \frac{1}{LC}$$

$$\omega = \frac{1}{\sqrt{LC}}$$

$$2\pi f_r = \frac{1}{\sqrt{LC}} \Rightarrow \boxed{f_r = \frac{1}{2\pi\sqrt{LC}}}$$

Resonant Impedance (Z_r)

$$Z = (R+j\omega L) \,||\, \frac{1}{j\omega C}$$

$$= \frac{(R+j\omega L)\left(\frac{1}{j\omega C}\right)}{R+j\omega L+\frac{1}{j\omega C}} = \frac{R+j\omega L}{((R+j\omega L)(j\omega C)+1)}$$

$$= \frac{R+j\omega L}{Rj\omega C - \omega^2 LC + 1} = \frac{R+j\omega L}{1-\omega^2 LC + j\omega RC}$$

We know at $\omega^2 = \frac{1}{LC}$

$$Z_r = \frac{R+j\omega L}{j\omega RC}, \quad R << j\omega L$$

$$Z_r = \frac{j\omega L}{j\omega RC}$$

$$\boxed{Z_r = \frac{L}{RC}}$$

Q factor:-

Quality (or) Q factor is defined as the ratio between reactance to the resistance.

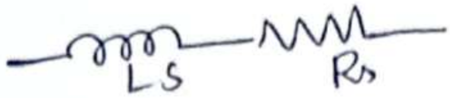
$$Q = \frac{X_L}{R} = \frac{\text{Voltage across L or C}}{\text{Applied voltage}} = \frac{I_m X_L}{I_m R} = \frac{X_L}{R}.$$

where X is the reactance of circuit.

Quality factor is also defined as it is the ratio btw total energy stored by the coil per cycle to the total power dissipated per cycle.

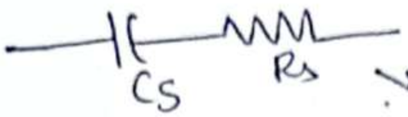
$$Q = \frac{2\pi \times \text{Total energy stored by coil per cycle}}{\text{Total power dissipated per cycle.}}$$

$$= 2\pi \times \frac{V_m^2/2}{V_m^2/2R \times T} \quad T = 1/f_0$$



$$\varphi = \frac{j\omega L_s}{R_s}$$

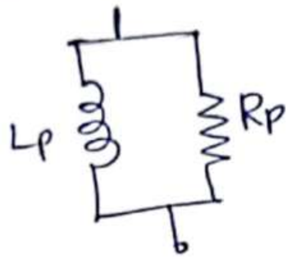
$$|\varphi| = \frac{\omega L_s}{R_s}.$$



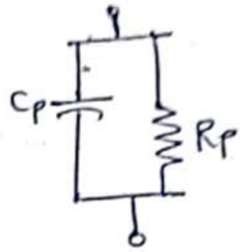
→ Capacitive $\varphi = \frac{1}{j\omega C_s \times R_s}$

$$|\varphi| = \frac{1}{\omega R_s C_s}.$$

For parallel circuit



$$\boxed{\varphi = \frac{R_p}{\omega L_p}}$$



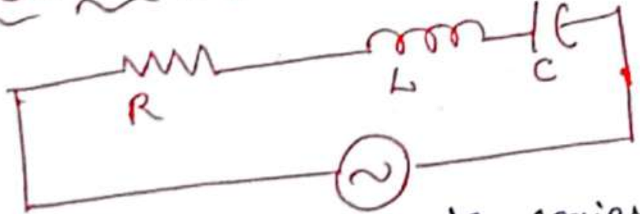
$$\boxed{\varphi = \omega R_p C_p.}$$

Quality factor measures the quality of goodness in inductor.

→ If the value of quality factor is high, then inductor is good i.e, No losses.

→ If the value of quality factor is low, the inductor is poor, measures coil losses.

Series Resonance



We know impedance for series $Z = R + j\omega L + \frac{1}{j\omega C}.$

$$Z = R + j\left(\omega L - \frac{1}{\omega C}\right)$$

$$I = \frac{V}{Z} = \frac{V}{R + j(\omega L - 1/\omega C)}$$

At resonance $X_L = X_C$, $f = f_r$ & $I = V/R$.

$$\Rightarrow \omega L = \frac{1}{\omega C}$$

$$\omega^2 = \frac{1}{LC}$$

$$\omega = \frac{1}{\sqrt{LC}}$$

$$\Rightarrow 2\pi f_r = \frac{1}{\sqrt{LC}} \Rightarrow f_r = \boxed{\frac{1}{2\pi\sqrt{LC}}}$$

Hence Q- factor: $Q = \frac{X}{R} = \frac{\omega_0 L}{R} = \frac{1}{\omega_0 CR} = \frac{1}{R}\sqrt{\frac{L}{C}}$

$$\boxed{\therefore Q = \frac{1}{R}\sqrt{\frac{L}{C}}}$$

[We know $\omega = \frac{1}{\sqrt{LC}}$]

Transformation of series resistor & Inductor:-

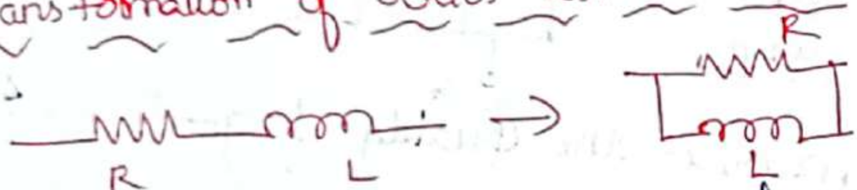


fig:- Series to parallel transformation

For series connection:

$$Z_S = R_S + j\omega L$$

$$Y_S = \frac{1}{R_S + j\omega L}$$

$$= \frac{1}{R_S + j\omega L} \times \frac{R_S - j\omega L}{R_S - j\omega L} = \frac{R_S - j\omega L}{R_S^2 + \omega^2 L^2} \quad (\omega^2 L^2 >> R_S)$$

$$Y_S = \frac{R_S - j\omega L}{\omega^2 L^2} \Rightarrow \frac{R_S}{\omega^2 L^2} - j\left(\frac{1}{\omega L}\right) \rightarrow (1)$$

For Parallel Connection:-

$$Y_P = \frac{1}{R_P} + \frac{1}{j\omega L}$$

$$Y_P = \frac{1}{R_P} - \frac{j}{\omega L} \rightarrow (2)$$

Compare eqn (1) & (2) Then

$$\boxed{R_p = \frac{\omega^2 L^2}{R_s} (\Omega)}$$

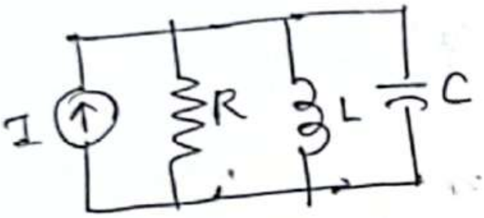
(or)

eqn ② - ①

$$\frac{1}{R_p} - \cancel{j}\left(\frac{1}{\omega L}\right) - \frac{R_s}{\omega^2 L^2} + \frac{\cancel{j}}{\cancel{\omega L}} = 0$$

$$\frac{1}{R_p} = \frac{R_s}{\omega^2 L^2} \Rightarrow \boxed{R_p = \frac{\omega^2 L^2}{R_s}}$$

Parallel Resonance:



In parallel resonance, admittance Y is

$$Y = \frac{1}{R} + \frac{1}{j\omega L} + j\omega C.$$

We know $I = \frac{V}{Z} = V \cdot Y = V\left[\frac{1}{R} + j\left(\omega C - \frac{1}{\omega L}\right)\right].$

at resonance $X_L = X_C$, $\omega = \omega_r$, $f = f_r$, $I = \frac{V}{R}$.

$$\omega_r C = \frac{1}{\omega_r L}$$

$$\omega_r^2 = \frac{1}{LC} \Rightarrow \boxed{f_r = \frac{1}{2\pi\sqrt{LC}}}$$

or $\omega_0 = \frac{1}{\sqrt{LC}}$ or $2\pi f_0 = \frac{1}{\sqrt{LC}} \Rightarrow \boxed{f_0 = \frac{1}{2\pi\sqrt{LC}}}$

Quality factor for parallel is

$$Q = \frac{\frac{1}{X}}{\frac{1}{R}} = \frac{R}{X} = \frac{R}{X_C} = \frac{R}{X_L}$$

$$Q_p = 2\pi f_r RC = \frac{R}{2\pi f_r L}.$$

$$\boxed{\therefore Q_p = R\sqrt{\frac{C}{L}}}$$

Impedance variation of resonance.

$$Y = \frac{1}{R} + \frac{1}{j\omega L} + j\omega C.$$

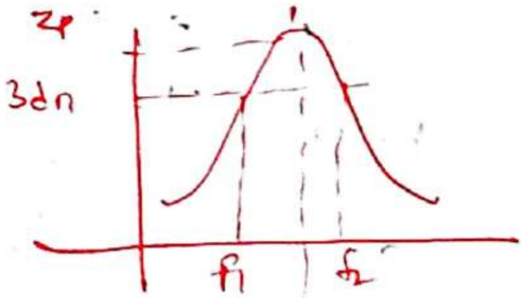
$$= \frac{1}{R} + j\left(\omega C - \frac{1}{\omega L}\right)$$

$$= 1 + jR\left(\omega C - 1/\omega L\right)/R = \frac{1 + j\omega RC\left(1 - \frac{1}{\omega^2 LC}\right)}{R}$$

$$= \frac{1 + j\omega CR\left[1 - \left(\frac{\omega r}{\omega}\right)^2\right]}{R}.$$

$$\therefore Z = \frac{1}{Y} = \frac{R}{1 + j\omega RC\left(1 - \left(\frac{\omega r}{\omega}\right)^2\right)}$$

Bandwidth of Resonance



$$BW = f_2 - f_1 = \Delta f.$$

B.W is measured at $1/\sqrt{2}$ or 3dB of max value of impedance at resonance.

Sharpness of Resonance

It is ratio of bandwidth and Resonant frequency

$$= \frac{BW}{f_0} = \frac{f_2 - f_1}{f_0} = \frac{1}{Q_0}.$$

where Q is quality factor at resonance.

Extra

Parallel resonance derivation

$$\phi = \frac{2\pi \times \text{Max energy stored by coil}}{\text{Power dissipated}}$$

$$= \frac{2\pi \times \frac{1}{2} \times (LI^2)}{\frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}} \times t} = \frac{2\pi \times \frac{1}{2} \times L \left(\frac{V}{\omega L}\right)^2}{\frac{I_m^2}{2} \cdot R \times \frac{1}{f}}$$

$$\left(I_m = \frac{V_m}{R}\right)$$

$$= \frac{2\pi \times \frac{1}{2} L \times \frac{V_m^2}{\omega^2 L^2}}{\frac{V_m^2}{2} \cdot \frac{1}{R} \times \frac{1}{f}} = \frac{2\pi f \cdot R}{\omega^2 \cdot L} = \frac{R}{\omega L} = \left|\frac{R}{X}\right|$$

Classification of Tuned Amplifiers:-

Tuned Amplifiers are classified into two types based on the type of signals there are.

Tuned Amplifier
- small signal tuned Amplifier
  - ① small signal tuned amplifier
  - ② Signal tuned capacitive Amplifier & Tansform coupled
  - ③ Double tuned Amplifier
  - ④ staggered tuned Amplifier
- Large signal tuned Amplifier

With +ve solutions ω_H & ω_L, we get

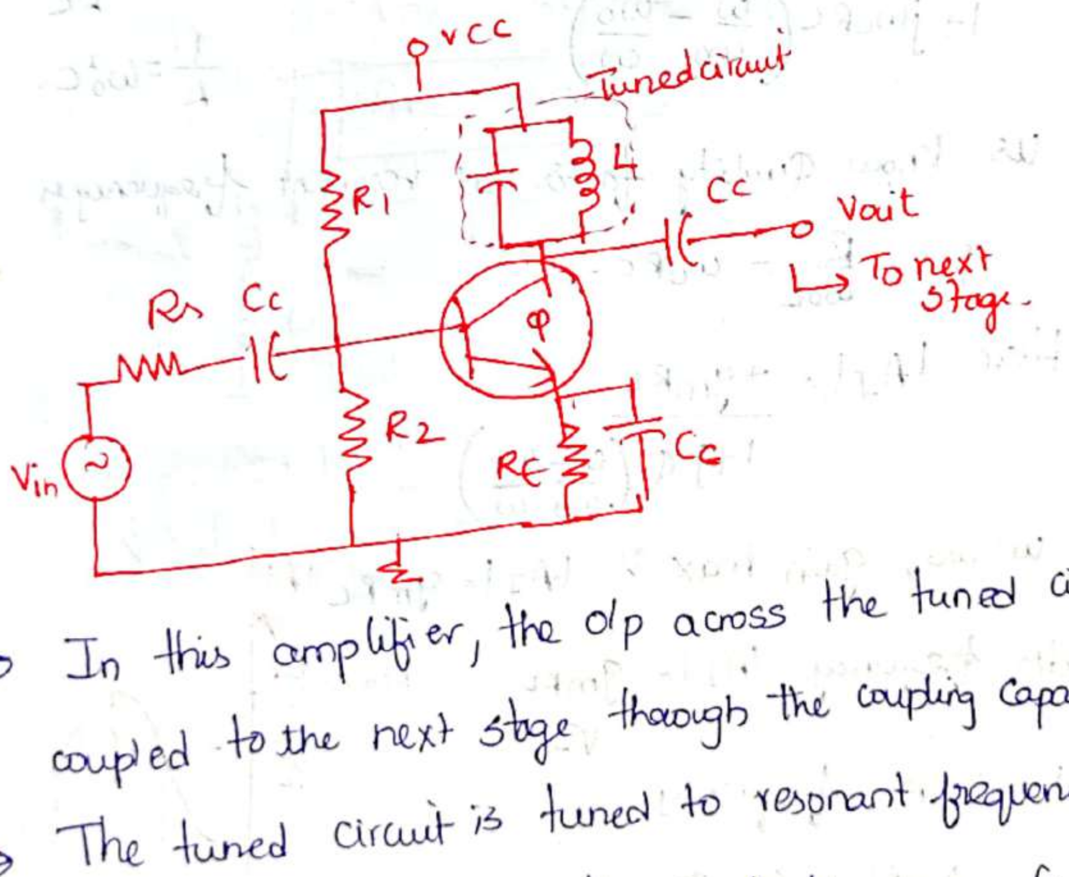
$$\text{Bandwidth} = f_H - f_L = \frac{f_0}{Q} = \frac{1}{2\pi RC}$$

From these we can construct the signal tuned amplifier.

Capacitance Coupled Single Tuned Amplifier

→ Single tuned amplifier uses one parallel resonant circuit as the load impedance in each stage of amplifier.

→ All tuned circuits are tuned to same frequency.

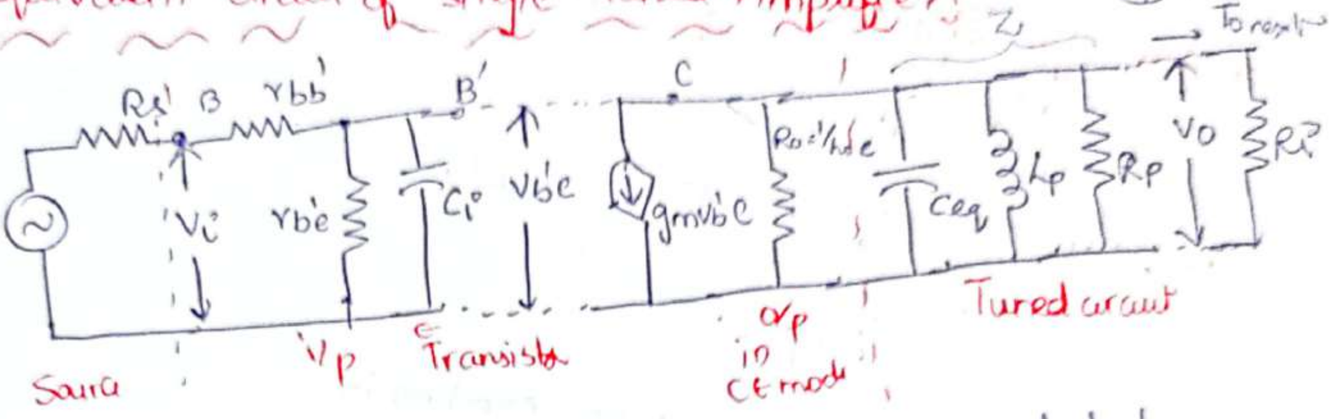


→ In this amplifier, the o/p across the tuned circuit is coupled to the next stage through the coupling capacitor C_c.

→ The tuned circuit is tuned to resonant frequency.

→ At resonance condition, the circuit offers very high impedance to amplify only the detected narrow band of frequencies.

Equivalent circuit of single Tuned Amplifier:



→ Ci → Miller form of reactance impedance included.

$$C_i = C_{b'e} + C_{b'c}(1-K).$$

K → voltage gain of the amplifier.

Ceq → o/p circuit capacitance.

$$C_{eq} = C_{b'c}\left(\frac{K-1}{K}\right) + C.$$, C → tuned circuit capacitance

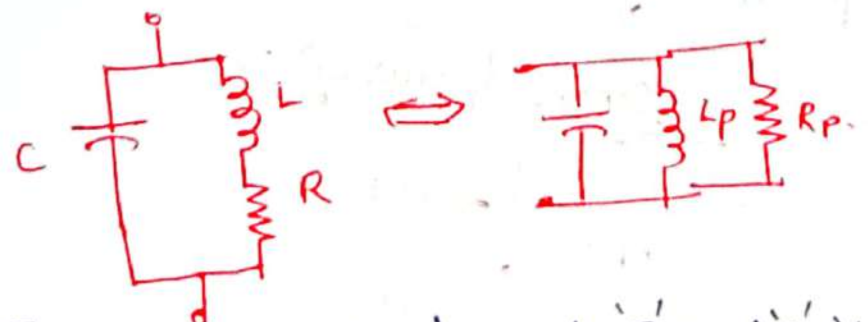
→ Ro → o/p resistance $R_o = 1/h_{oe}$.

→ The output voltage is given as,

$$V_o = -g_m V_{b'e} \cdot Z$$

Z → impedance of C, L and R in parallel.

Analysis of single Tuned Amplifier:-



→ The series combination of 'R' and 'L' is used to represent the actual inductance in the tuned circuit.

→ For analysis, the parallel resonant circuit is used because of its high gain & selectivity.

Condition for equivalence of series & parallel circuit

$$Y_s = Y_p$$

The admittance of the series circuit is

$$Y_s = \frac{1}{R+j\omega L}$$

→ Rationalizing the above equation is

$$Y_s = \frac{1}{R+j\omega L} \times \frac{R-j\omega L}{R-j\omega L}$$

here $\omega L >> R$.

$$Y_s = \frac{R-j\omega L}{R^2+\omega^2 L^2}$$

$$\boxed{Y_s = \frac{R}{\omega^2 L^2} - j\left(1/\omega L\right)} \rightarrow (1).$$

→ The admittance of the parallel circuit is

$$Y_p = \frac{1}{R_p} + \frac{1}{j\omega L_p} = \frac{1}{R_p} - j\left(1/\omega L_p\right) \quad (2)$$

→ Comparing (1) & (2),

$$\boxed{\frac{1}{R_p} = \frac{R}{\omega^2 L^2}} \quad , \quad \cancel{\omega L_p = \omega L} \quad \boxed{L_p = L}$$

Resonant frequency [Centre frequency] :-

$$\boxed{f_r = \frac{1}{2\pi\sqrt{L_p \cdot C_{eq}}}} \rightarrow (5)$$

Where $L_p = L$ & $C_{eq} = C_0 + C$.

Quality factor [Q] :-

The quality factor 'Q' of the coil at resonance is given as

$$\boxed{Q_r = \frac{\omega_r \cdot L_p}{R_p}} \rightarrow \text{unloaded Q- factor} \quad (6)$$

→ But in practical case, R_o and the i/p resistance of the next stage act as a load.

At resonant frequency

$$Q_r = \frac{\omega_r . L_p}{R_p} = \frac{\omega_r . L}{\frac{\omega_r^2 . L^2}{R}} = \frac{R}{\omega_r . L}$$

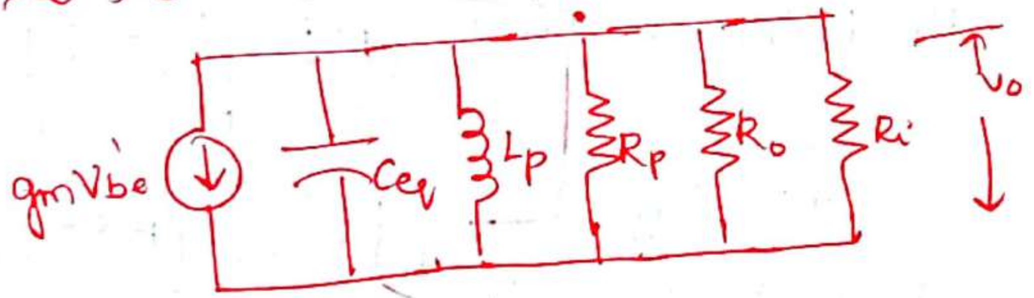
∴ For parallel circuit

$$\boxed{Q_r = \frac{R_s}{\omega_r . L}} \quad \text{— (7)}$$

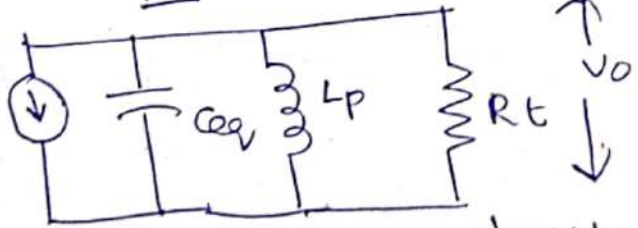
Effective Quality factor [Q_{eff}]:-

It includes the load (sum of R_o & R_i of the next stage].

Equivalent circuit of the o/p circuit:-



≃



$$\frac{1}{R_t} = \frac{1}{R_o} + \frac{1}{R_p} + \frac{1}{R_i}$$

For the parallel resonance circuit,

$$Q_{eff} = \frac{\text{Susceptance of inductance (or) Capacitance}}{\text{conductance of shunt resistance } R_t}$$

Susceptance of $L_p = 1/X_L = \frac{1}{\omega_r L_p}$ $= \omega_r C_{eq}$

Susceptance of $C_{eq} = 1/X_C = 1/\omega_r C_{eq}$

$$\because Y = \frac{1}{Z} = \frac{1}{R_t} + \frac{1}{j\omega L} + j\omega C.$$

$$= \frac{1}{R_t}\left[1 + \frac{R_t}{j\omega L} + R_t j\omega C\right]$$

multiply Dr & Nr by ω_r.

$$Y = \frac{1}{R_t}\left[1 + \frac{R_t \omega_r}{j\omega \omega_r L} + \frac{R_t j\omega \omega_r C}{\omega_r}\right]$$

$$\therefore \frac{R_t}{\omega_r L} = \omega_r R_t C = Q_{eff}$$

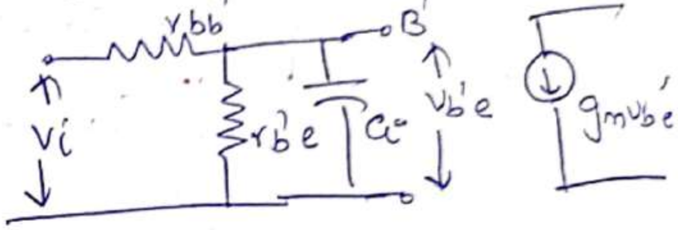
conductance of $[R_t] = 1/R_t = 1/R_t$.

$$\boxed{Q_{eff} = \frac{R_t}{\omega_r L_p} = \omega_r C_{eq} R_t} \quad (8)$$

Voltage gain :-

The voltage gain (A_v) for single tuned amplifier is

$$A_v = \frac{v_o}{v_i}$$



For equivalent circuit,

$$v_o = -g_m v_{b'e} \cdot Z$$

$$V_{b'e} = \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \cdot v_i \quad \therefore \quad \boxed{A_v = -g_m\left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}}\right) Z} \quad (9)$$

$$Z = R_t = R_o \| R_p \| R_i$$

$$\Rightarrow Z = \frac{R_t}{1 + 2\delta Q_{eff}}$$

where $\delta \rightarrow$ fractional variation in 'f_r'

$$\delta = \frac{\omega - \omega_r}{\omega_r} = \frac{\omega}{\omega_r} - \frac{\omega_r}{\omega_r}$$

$$\boxed{\frac{\omega}{\omega_r} = 1 + \delta} \quad (10)$$

$$Y = \frac{1}{Z} = \frac{\left[1 + jQ_e\left[\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right]\right]}{R_t} \quad \text{sub eqn (10)}$$

$$Z = \frac{R_t}{1 + jQ_e\left[1 + \delta - \frac{1}{1+\delta}\right]}$$

$$Z = \frac{R_t}{1 + jQ_e\left[\frac{1 + \delta^2 + 2\delta - 1}{1+\delta}\right]}$$

$$= \frac{R_t}{1 + jQ_e\left[\frac{2\delta(1/2 + \delta)}{1+\delta}\right]}$$

app eqn.

$$\boxed{Z = \frac{R_t}{1 + 2j\delta Q_{eff}}}$$

$$A_v = -g_m\left[\frac{r_{b'e}}{r_{bb'} + r_{b'e}}\right]\left[\frac{R_t}{1 + j2\delta Q_{eff}}\right] \quad (11)$$

Voltage gain at resonance frequency, $\delta = 0$

$$A_{v(res)} = -g_m\left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}}\right) R_t \quad (12)$$

Relative gain of single tuned amplifier

$$\frac{A_v}{A_{v(res)}} = \frac{-g_m\left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}}\right)\left(\frac{R_t}{1 + j2\delta Q_{eff}}\right)}{-g_m\left(r_{b'e}/r_{bb'} + r_{b'e}\right) R_t}$$

$$\boxed{\therefore \frac{A_v}{A_{v(res)}} = \frac{1}{1+j2\delta Q_{eff}}}$$

The magnitude of relative gain of tuned amplifier

$$\left|\frac{A_v}{A_{v(res)}}\right| = \left|\frac{1}{1+j2\delta Q_{eff}}\right|$$

$$\therefore \boxed{\frac{A_v}{A_{v(res)}} = \frac{1}{\sqrt{1+(2\delta Q_{eff})^2}}}$$

3dB B.W $\rightarrow \Delta f = \dfrac{1}{2\pi R_t C_{eq}}$

Effect of Cascading single Tuned Amplifiers on B.W

In order to obtain high overall gain, several identical stages or tuned amplifiers can be used in cascade. At same time the high voltage gain is obtained by a narrower bandwidth than for a single stage.

→ Consider n stages of single tuned direct coupled amplifiers connected in cascade. Then determine overall gain & b.w of such amplififier. we know

$$\left|\frac{A}{A_{res}}\right| = \frac{1}{\sqrt{1+(2\delta Q_e)^2}} \qquad \left[\because \delta = \frac{\omega-\omega_r}{\omega_r}\right]$$

Now at n stage cascaded amplifier becomes

$$\left|\frac{A}{A_{res}}\right|^n = \left|\frac{1}{\sqrt{1+(2\delta Q_{eff})^2}}\right|^n = \frac{1}{\left[1+(2\delta Q_{eff})^2\right]^{n/2}}$$

The 3dB frequencies for the n-stage cascaded amplifier can be found by equating $\left|\frac{A}{A_{res}}\right|^n = \frac{1}{\sqrt{2}}$

$$\left|\frac{A}{A_{res}}\right|^n = \frac{1}{\left[\sqrt{1+(2\delta Q_{eff})^2}\right]^n} = \frac{1}{\sqrt{2}}$$

$$\Rightarrow \left[\sqrt{1+(2\delta Q_{eff})^2}\right]^n = \sqrt{2}$$

$$1+(2\delta Q_{eff})^2 = 2^{1/n}$$

$$2\delta Q_{eff} = \pm\sqrt{2^{1/n}-1}$$

sub δ in above eqn $\delta = \frac{\omega-\omega_r}{\omega_r} = \frac{f-f_r}{f_r}$

$$2\left(\frac{f-f_r}{f_r}\right)Q_{eff} = \pm\sqrt{2^{1/n}-1}$$

$$2(f-f_r)\,Q_{eff} = \pm f_r\sqrt{2^{1/n}-1}$$

Now $f_2-f_r = \frac{f_r}{2Q_e}\sqrt{2^{1/n}-1}$ —①

&

$f_r-f_1 = \frac{f_r}{2Q_e}\sqrt{2^{1/n}-1}.$ —②

Then b.w of n^{th} stage identical amplifier is

$B_{1n} = f_2-f_1 = (f_2-f_r)+(f_r-f_1).$ [add ①&②]

$$= \frac{f_r}{2Q_e}\sqrt{2^{1/n}-1} + \frac{f_0}{2Q_e}\sqrt{2^{1/n}-1}$$

$$= \frac{f_0}{Q_e}\sqrt{2^{1/n}-1} \qquad \left[\because B.W = \frac{f_r}{Q}\right]$$

$$= B_1\sqrt{2^{1/n}-1}$$

Here B_{1n} is B.W of n-stages of cascade amplifier & B_1 is the B.W for single stage B.W of n stages.

when $n=2$; $\sqrt{2^{1/n}-1} = 0.643$

$n=3$; $\sqrt{2^{1/n}-1} = 0.510$

$\therefore$ B.W is reduced to 64.3% for two stages & 51% for 3 stages of cascade amplifier. In order to maintain a prescribed 3dB B.W, Q of the tuned circuit should be reduced.

Single Tuned Transformer Coupled or Inductive Coupled Amplifier

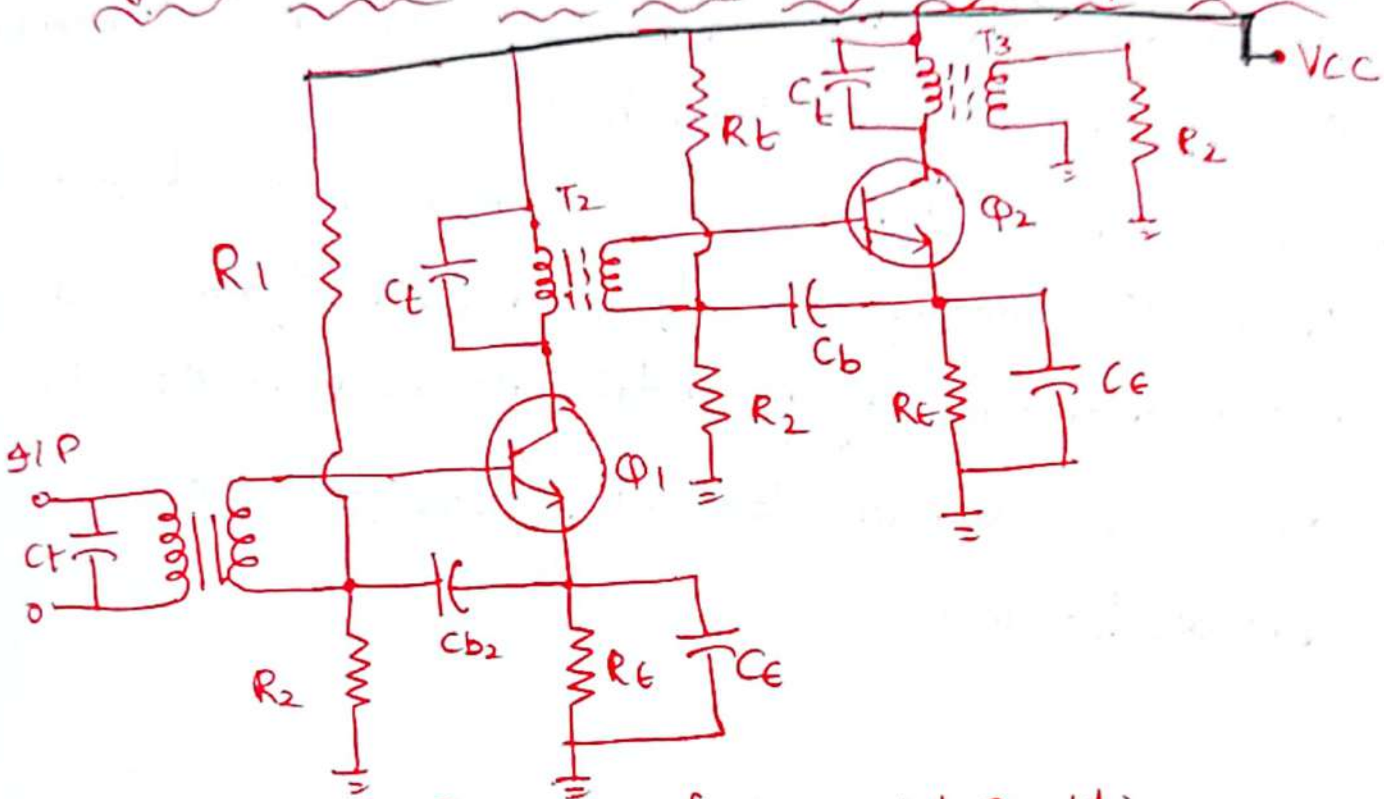


fig: Tuned transformer coupled Amplifier

We know that 3dB B.W of tuned amplifier is

$$B.W_{(3dB)} = \frac{f_r}{Q_{eff}} \text{ or } \frac{f_0}{Q_{eff}}$$

and $Q_{eff} = \frac{R_t}{\omega_r L}$ or $\omega_r C_{eq} R_t$ $[X_L = X_C]$.

→ The value of R_t is likely to be fairly low, especially in base-biased versions then Q value decreases & B.W increases.

→ The B.W is not only broad but not defined properly because of uncertainty in the value of R_t.

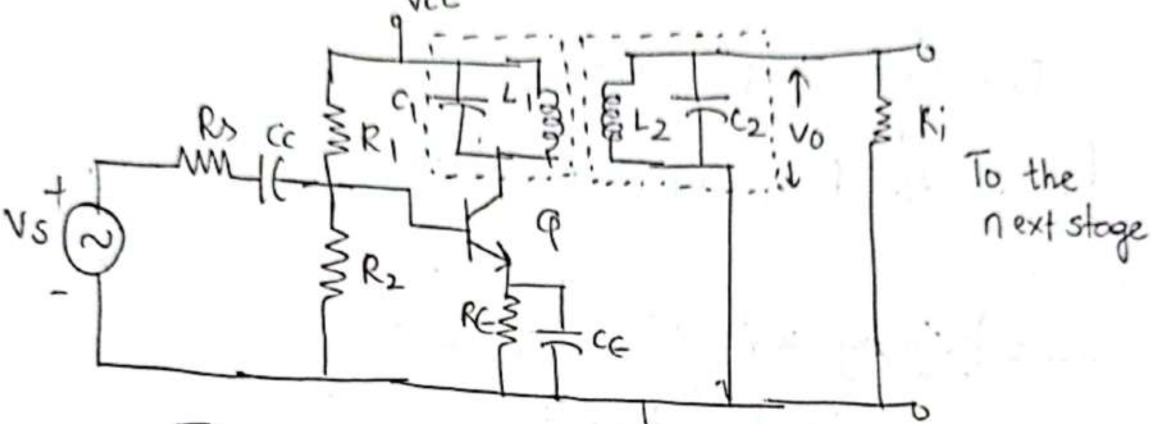
→ As a result a tuned capacitor circuit coupled is not very often used.

→ A better approach is to use transformer coupling where the value of R_t reflected into the tuned primary can be made high enough to allow high values of Q and narrow bandwidth. The above fig shows the tuned transformer coupled amplifier with base bias.

→ Even if a broad-band circuit is required, it is best to shunt the primary with a fixed resistor of any desired value to get a very predictable bandwidth, rather than depend upon the uncertain value of R_t to arrive at a broader bandwidth.

Double Tuned Amplifier

Double tuned amplifiers used two inductively coupled tuned circuits per stage, both the tuned circuits being tuned to the same frequency.



It is mainly used to obtain wider bandwidth. Here Rc is replaced with L, C1 tuned, it is inductively coupled to another tuned circuit L_2, C_2. The o/p single from stage is inductively coupled thorough L_2, C_2. The equivalent circuit of double tuned amplifier across o/p side.

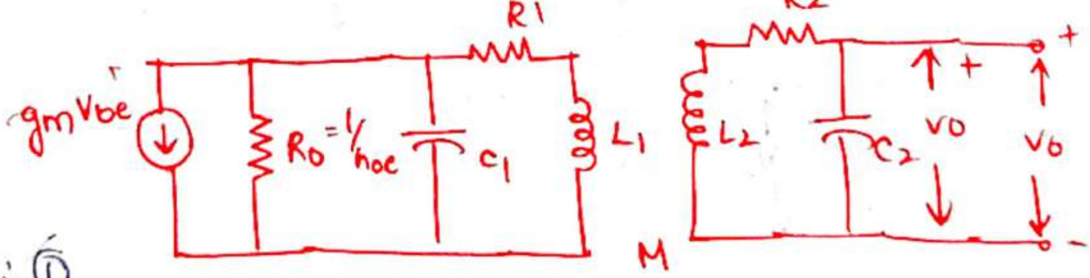


fig ①

The parallel elements of $L_1 C_1$ replaced as series elements.

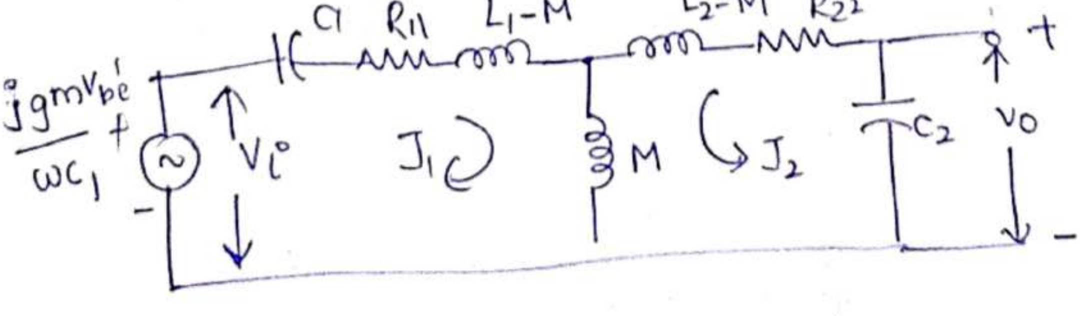


fig ②

We know for parallel to series vice versa conversion.

$$R_p = \frac{\omega_r^2 L^2}{R}, \quad R = \frac{\omega_r^2 L^2}{R_p}$$

We can write $R_{11} = \frac{\omega_0^2 L_1^2}{R_0} + R_1$ and $R_{22} = \frac{\omega_0^2 L_2^2}{R_e} + R_2$

We know $Q = \frac{\omega_r L}{R}$

$$Q_1 = \frac{\omega_r L_1}{R_{11}} \ \& \ Q_2 = \frac{\omega_r L_2}{R_{22}}$$

The Q factors for both circuits are to be same.

$\therefore Q_1 = Q_2 = Q$;

means resonant frequency $\omega_r^2 = \omega_0^2 = \frac{1}{L_1 C_1} = \frac{1}{L_2 C_2}$.

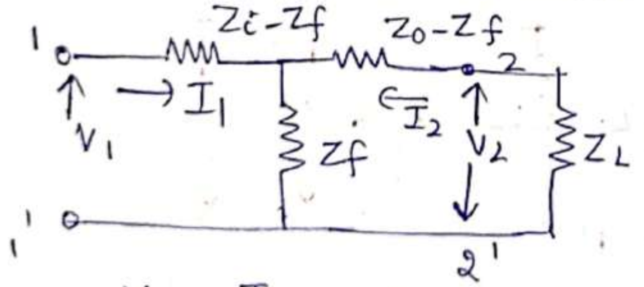
From the simplified circuit from loop (2)

$$V_0 = I_2 \cdot \frac{1}{j\omega_r C_2} = \frac{-j I_2}{\omega_r C_2}$$

For this I_2 is represents in terms of V_1

$$V_0 = -j\,(V_1 \cdot Y_2)/\omega_r C_2$$

The transfer admittance can be calculated by using the following figure.



$$Y_T = \frac{I_2}{V_1} = \frac{I_2}{I_1 Z_{11}} = \frac{A_i}{Z_{11}} = \frac{Z_f}{Z_f^2 - Z_i(Z_o + Z_L)}$$

From loop eqn's $V_1 = I_1(Z_i - Z_f) + Z_f(I_1 + I_2)$

$$= I_1 Z_i + I_2 Z_f.$$

$$\frac{V_1}{I_1} = Z_i + \left(\frac{I_2}{I_1}\right) Z_f = Z_i + A_i Z_f.$$

lly for o/p side

$$0 = (I_1+I_2)(Z_f) + I_2(Z_0+Z_f) + I_2 Z_L$$

$$= I_1 Z_f + I_2(Z_0+Z_L)$$

$$\frac{I_2}{I_1} = A_i = \frac{-Z_f}{Z_0+Z_L}$$ sub A_i is Z_{11} Then

$$Z_{11} = Z_i - \left(\frac{Z_f}{Z_0+Z_L}\right) Z_f = \frac{Z_i(Z_0+Z_L) - Z_f^2}{Z_0+Z_L}.$$

$$\therefore Y_T = \frac{A_i}{Z_{11}} = \frac{-Z_f}{Z_0+Z_L} \cdot \frac{Z_0+Z_L}{Z_i(Z_0+Z_L) - Z_f^2} = \frac{+Z_f}{Z_f^2 - Z_i(Z_0+Z_L)}.$$

$$\cancel{Z_{11}} = \frac{\cancel{V_1}}{\cancel{I_1}} = \frac{1}{\cancel{Y_T}} = Z_i\cancel{f^2}$$

$$\therefore Z_{11} = \frac{V_1}{I_1} = Z_i - \frac{Z_f^2}{Z_0+Z_L} \quad \& \quad A_i = \frac{I_2}{I_1} = \frac{-Z_f}{Z_0+Z_L}$$

Here $Z_f = j\omega_r M$

$$Z_i = R_{11} + j\left[\omega L_1 - \frac{1}{\omega C_1}\right]$$ from fig (2)

$$Z_0 + Z_L = R_{22} + j\left(\omega L_2 - \frac{1}{\omega C_2}\right].$$

From the above equations Z_f, Z_i and Z_0+Z_L we can further simplify $Z_f = j\omega_r M = j\omega_r k\sqrt{L_1 L_2}$, k is coupling coefficient.

In Z_i eqn, multiply Nr & Dr by $\omega_r L_1$, we get.

$$Z_i = R_{11}\frac{\omega_r L_1}{\omega_r L_1} + j\omega_r L_1\left[\frac{\omega L_1}{\omega_r L_1} - \frac{1}{\omega\omega_r L_1 C_2}\right]$$

$$= \frac{\omega_r L_1}{Q_1} + j\omega_r L_1\left[\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right] \quad \left[\because \omega_r^2 = \frac{1}{LC}\right]$$

$$= \frac{\omega_r L_1}{Q_1} + j\omega_r L_1(2\delta) \qquad \left[1+\delta - \frac{1}{1+\delta}\right] = 2\delta.$$

$$= \frac{\omega_r L_1}{\varphi}\left[1 + j2\varphi\delta\right]$$

Illy $z_0 + Z_l = R_{22} + j\left[\omega L_2 - \frac{1}{\omega C_2}\right]$

We can write similarly $= \frac{\omega_r L_2}{\varphi}\left[1 + j2\varphi\delta\right]$.

We know $Y_T = \frac{Z_f}{Z_f^2 - Z_i(z_0 + Z_l)} = \frac{1}{Z_f - Z_i\left(\frac{Z_0 + Z_l}{Z_f}\right)}$

Sub all above values of Z_f, Z_i & $Z_0 + Z_L$

$$Y_T = \frac{1}{j\omega_r K\sqrt{L_1 L_2} - \dfrac{\frac{\omega_r L_1}{\varphi}\left[1 + j2\varphi\delta\right] \cdot \frac{\omega_r L_2}{\varphi}(1 + j2\delta\varphi)}{j\omega_r K\sqrt{L_1 L_2}}}$$

$$= \frac{K\varphi^2}{\omega_r\sqrt{L_1 L_2}\left[4\varphi\delta - j\left(1 + K^2\varphi^2 - 4\varphi^2\delta^2\right)\right]}$$

Sub the value of I_2 in terms of v_i. V_T is v_0.

$$v_0 = \frac{-j}{\omega_r C_2} \cdot \frac{j g_m v_{b'e}}{\omega_r C_1} \cdot \frac{K\varphi^2}{\omega_r\sqrt{L_1 L_2}\left[4\varphi\delta - j\left(1 + K^2\varphi^2 - 4\varphi^2\delta^2\right)\right]}$$

since $v_{b'e} = v_i = \frac{j g_m v_1}{\omega C_1}$

$$A_v = \frac{v_0}{v_1} = g_m \omega_r^2 L_1 L_2 \cdot \frac{K\varphi^2}{\omega_r\sqrt{L_1 L_2}\left[4\varphi\delta - j\left(1 + K^2\varphi^2 - 4\varphi^2\delta^2\right)\right]}$$

since $\omega_r^2 = \frac{1}{LC}$.

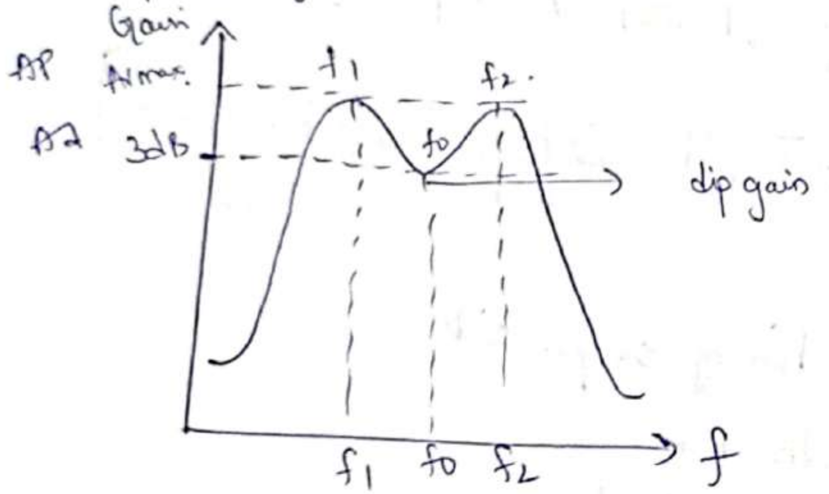
$$\therefore A_v = \left[\frac{g_m \omega_r\sqrt{L_1 L_2}\, K\varphi^2}{4\varphi\delta - j\left(1 + K^2\varphi^2 - 4\varphi^2\delta^2\right)}\right]$$

$$|A_v| = g_m \omega r \sqrt{L_1 L_2}\, Q \cdot \frac{kQ}{\sqrt{(1+k^2Q^2-4Q^2\delta^2)^2+16Q^2\delta^2}}$$

From above eqn the frequency deviation δ at which gain peak occurs can be found by making

$$4Q\delta - j(1+k^2Q^2-4Q^2\delta^2)=0.$$

From gain vs frequency figure, the two gain peaks in the frequency response of the double tuned amplifier is given by



$$f_1 = f_r\left[1-\frac{1}{2Q}\sqrt{k^2Q^2-1}\right] \;\&\; f_2 = f_r\left[1+\frac{1}{2Q}\sqrt{k^2Q^2-1}\right]$$

At $k^2Q^2=1 \Rightarrow k=\frac{1}{Q}$, $f_1=f_2=f_r$

This is known as critical coupling

If $k<1/Q$, the peak gain is less than max gain & the coupling is poor.

If $k>1/Q$, the circuit is over coupled & the response shows the double peak. This is useful when more bw is required.

$$|A_p| = \frac{g_m \omega r \sqrt{L_1 L_2}\, Q}{2}$$

The gain at dip $\delta=0$ is given as

$$|A_d| = |A_p| \cdot \frac{2kQ}{1+k^2Q^2}$$

where γ is the magnitude of ripple in the gain curve

$$\gamma = \left|\frac{A_P}{A_d}\right| = \frac{1+k^2\varphi^2}{2k\varphi}$$

solve we get $k\varphi = \gamma + \sqrt{\gamma^2-1}$

The B.W at frequencies gain $|A_d|$ is useful

At 3dB gain $\gamma = \sqrt{2}$

$$k\varphi = \gamma + \sqrt{\gamma^2-1} = 2.414$$

$$B.W = \sqrt{2}(f_2-f_1)$$

$$= \sqrt{2}\left[f_r\left(1+\frac{1}{2\varphi}\sqrt{k^2\varphi^2-1}\right) - f_r\left(1-\frac{1}{2\varphi}\sqrt{k^2\varphi^2-1}\right)\right]$$

$$= \sqrt{2}\left(\frac{f_r}{\varphi}\sqrt{k^2\varphi^2-1}\right] = \sqrt{2}\left[\frac{f_r}{\varphi}\sqrt{2.414^2-1}\right] = \boxed{\frac{3.1 f_r}{\varphi}}$$

Advantages:-

(1) It has flatter response having steeper sides.

(2) It provides larger 3dB B.W

(3) Provides larger gain b.w product

Effect of cascading double tuned amplifier on B.W:-

→ When identical double tuned amplifier stages are connected in cascade, the overall B.W of the system becomes narrow & the steepness of the sides of the response is increased just as when single tuned stages are cascaded.

→ The relationship btw 3dB B.W of n identical double tuned circuit critically coupled stages are compared with the B.W of a single stage double tuned amplifier as n stages double tuned amplifiers = $\Delta_2 [2^{1/n} - 1]^{1/4}$ or

$$B_{2n} = B_2 \left[2^{1/n} - 1\right]^{1/4}.$$

Where B_2 is the 3dB B.W of single stage double tuned amplifier

Stagger Tuned Amplifier:-

Need for stagger Tuned Amplifier:-

→ The double tuned amplifier gives greater 3dB bandwidth having steeper sides and flat top. But alignment of double tune amplifier is difficult.

→ To over come this problem two single tuned amplifiers are Cascaded.

→ In stagger tuned circuit two single tuned cascade amplifier having a certain b.w are taken

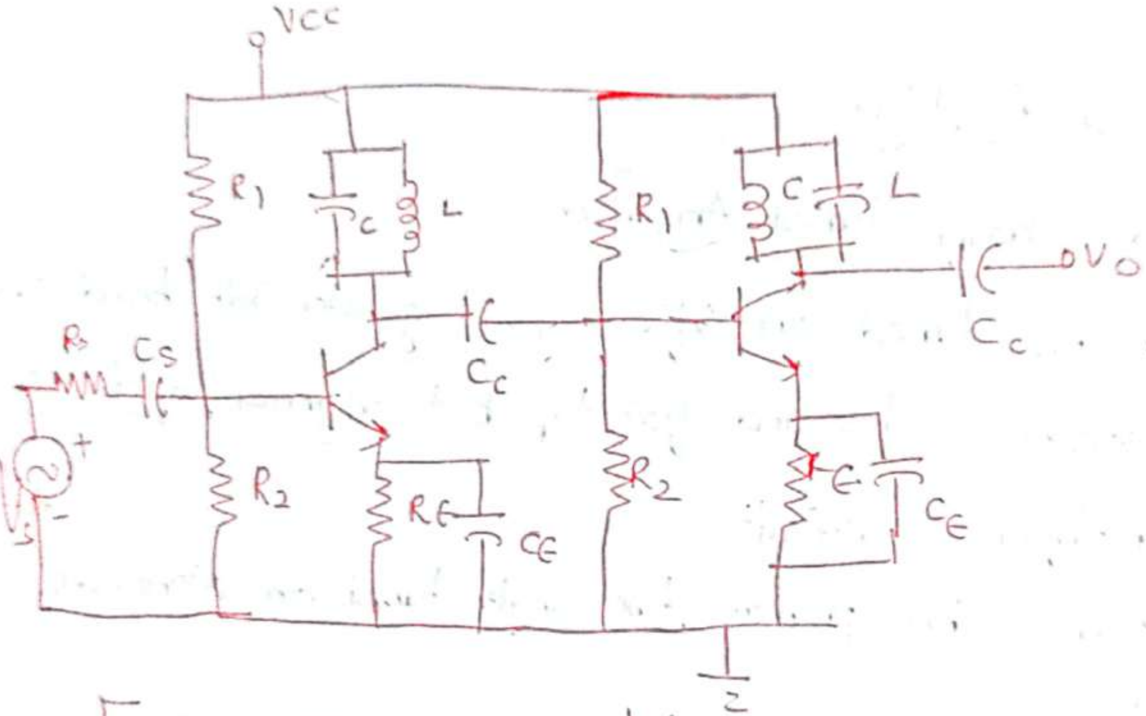
→ The resonant frequencies of the two tuned circuit are so adjusted that they are separated by an amount equal to the bandwidth of each stage.

→ Since the resonant frequencies are displaced or staggered, they are known as stagger tuned circuits.

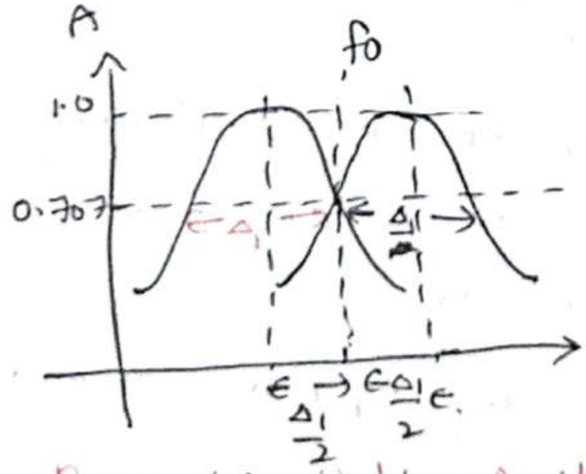
Def:- It is a circuit in which two single tuned cascaded amplifiers having certain bandwidth are taken and their resonant frequencies are adjusted that they are sepctated by an amount equal to the bandwidth of each stage.

since resonant frequencies are displaced it is called stagger tuned amplifier.

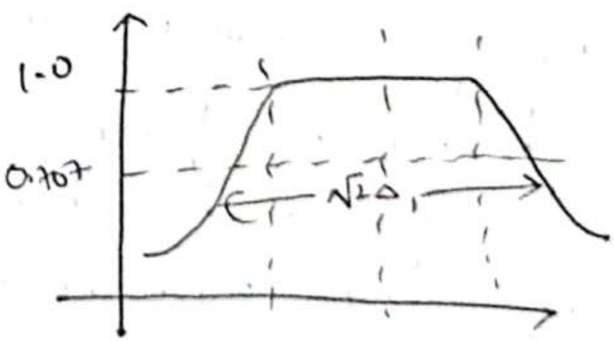
Circuit of tuned stagger Amplifier



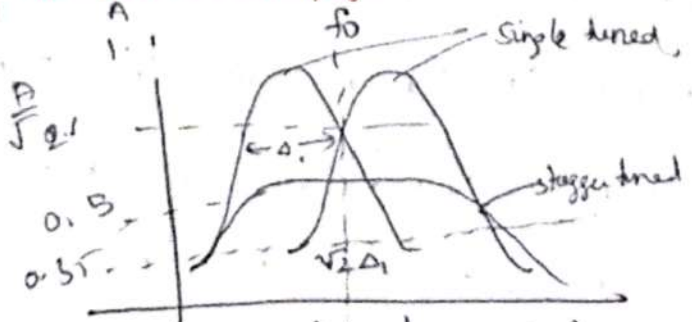
Frequency Response of stagger tuned amplifier



Response of individual tuned Amplifier



Overall response of stagger tuned amplifier



Comparison of singletuned and stagger tuned amplifier.

Analysis of stagger tuned circuit

The gain of the single tuned amplifier is

$$\frac{A_v}{A_{v(res)}} = \frac{1}{1+2Q_{eff}\delta} = \frac{1}{1+jX}; \quad X = 2Q_{eff}\delta$$

stagger tuned → two single tuned.

Let $\frac{\Delta_1}{2} = \delta$ then one stage is tuned to $f_r + \delta$ and second stage tuned to $f_r - \delta$

$\therefore f_{r_1} = f_r + 1 \quad f_{r_2} = f_r - \delta$.

$$\therefore \frac{A_v}{A_{v(reson)_1}} = \frac{1}{1+j(X+1)} \text{ and } \frac{A_v}{A_{v(res)_2}} = \frac{1}{1+j(X-1)}$$

$\therefore$ The overall gain of these two stages is.

$$\frac{A_v}{A_{v(res)cascade}} = \frac{1}{1+j(X+1)} \cdot \frac{1}{1+j(X-1)} = \frac{1}{2+2jX-X^2}$$

$$= \frac{1}{(2-X)^2+2jX}$$

$$\left|\frac{A_v}{A_{v(res)cas}}\right| = \frac{1}{\sqrt{(2-X)^2+(2X)^2}} = \frac{1}{\sqrt{4+X^4-4X^2+4X^2}} = \frac{1}{\sqrt{4+X^4}}$$

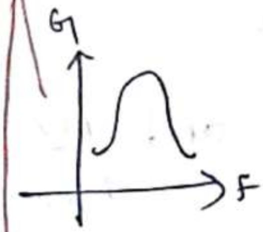
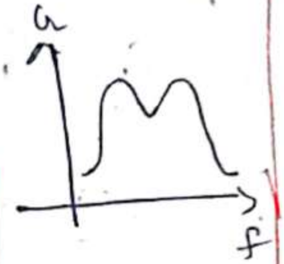
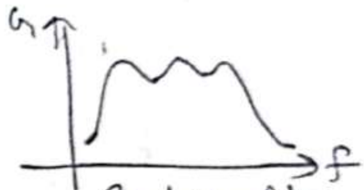
We know $X = 2Q_{eff}\delta$.

$$\therefore \left|\frac{A_v}{A_{v(res)cas}}\right| = \frac{1}{\sqrt{4+X^4}} = \frac{1}{\sqrt{4+(2Q_{eff}\delta)^4}} = \frac{1}{\sqrt{4+16Q_{eff}^4\delta^4}}$$

$$\therefore = \frac{1}{2\sqrt{1+4Q_{eff}^4\delta^4}}$$

$\therefore$ This relative gain of staggered tuned amplifier; gain is less than remaining single tuned amplifier.

Comparision of Tuned Amplifier

S.No	Parameter	Single tuned circuit	Double tuned circuit	Stagger tuned circuit
1.	No. of tuned circuits	one	two	More than two
2.	Q factor	high	high	Moderate low
3.	Selectivity	very high	Moderate	Moderate low
4.	B.W	Small	Moderate	High
5.	Frequency response vs gain			 ~~Band pass filter~~
6.	Application	Rf Amplifier stage in radio receiver	Intermediate frequency stage in radio receiver	Band pass filter

Stability of tuned amplifier :-

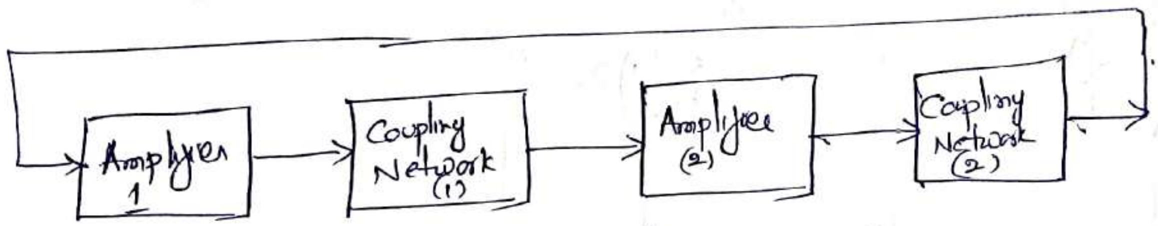
1. At high frequencies, because of the inter junction capacitance between base and collector (C_{bc}), there will be feedback from o/p to i/p.
2. At high frequency reactance is low and signal comes from o/p. to i/p
3. As a resultant circuit becomes unstable and it will stop working as an amplifier and works as oscillator because of +ve feedback.

To overcome this problem we use different techniques using

Multivibrators

A multivibrator produces non-sinusoidal voltages, that are very rich in harmonics. It produces pulsating output waveform of square, rectangular, ramp voltage as required in the different applications. There are three types of multivibrator, They are

1. The astable multivibrator (or) free running multivibrator
2. The mono stable multivibrator (or) one-shot multivibrator.
3. The bi-stable multivibrator (or) two shot multivibrator (or) flipflop.



Block diagram of Multivibrator.

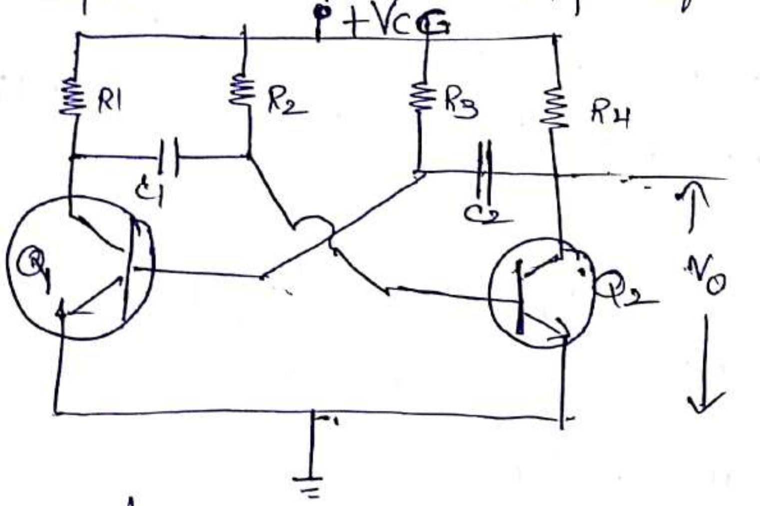
→ Figure shows the general configuration of a multivibrator. There are two amplifiers and two coupling networks. The two amplifiers must be connected regeneratively. At any instant, one amplifier is ON and the other is OFF. After certain time one amplifier is OFF and the other is ON. In this way the output will be square wave.

Astable Multivibrator:

An astable multivibrator has no stable states. Once the Multivibrator is ON, it just changes its states on its own after a certain time period which is determined by the RC time constants..

Construction of Astable Multivibrator

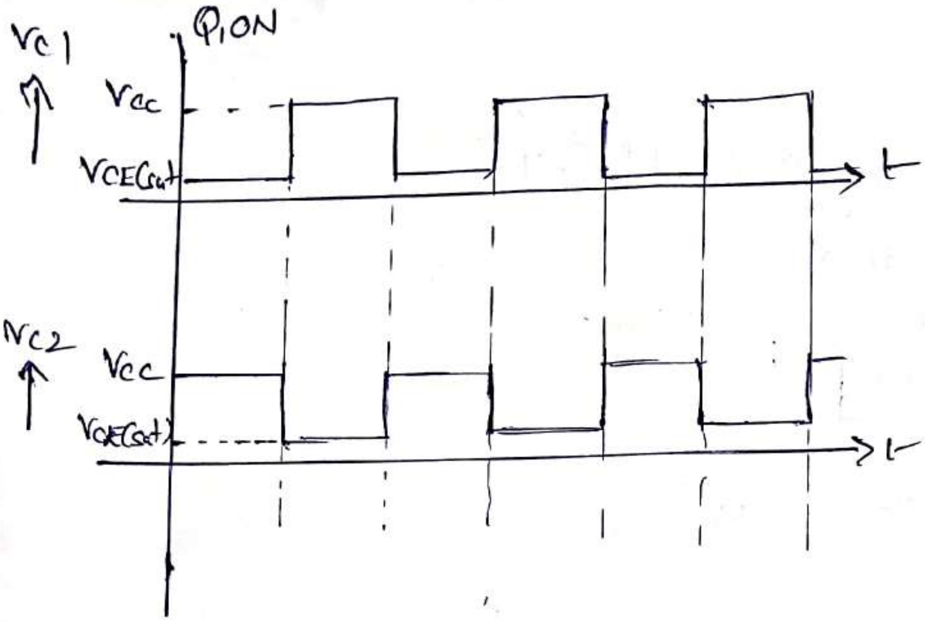
Two transistors named Q_1 & Q_2 are connected in feedback to one another i.e cross coupled connection. The collector of transistor Q_1 is connected to the base of transistor Q_2 through the capacitor C1 and vice versa. The Emitter of both the transistors are connected to ground. The collector load resistors R_1 & R_4 and the biasing resistors R_2 & R_3 are of equal values. The capacitors C_1 & C_2 are of equal values.



Operation :-

When Vcc is applied, both the transistors are tried to conduct. As no transistor characteristics are alike, one of the two transistors say Q_1 has its collector current increases and thus conducts. The collector of Q_1 is applied to the base of Q_2 through C_1. Now Q_2 starts conducts. The voltage at collector of Q_2 is nearly less value. This is given to the base of Q_1, so Q_1 comes in OFF state.

Two transistors, Q_1 & Q_2, alternately switch between saturation and cutoff due to feedback through capacitors C_1 & C_2. When Q_1 conducts, its collector voltage drops, turning Q_2 off. As C_1 charges, it eventually turns Q_2 on, which in turn drives Q_1 off through C2. This cycle continues, producing a square wave output with amplitude Vcc. The time period depends on the RC values of the circuits.



Frequency of Oscillation

$$V_C(t) = V_C(\infty) + [V_C(0) + V_C(\infty)]e^{-t/RC}$$

$$V_C(0) = -V_{CC} + V_{BE}$$

C_1 is fully charged

$$V_C(\infty) = V_{CC}$$

at $t = T_1 \Rightarrow V_C(T_1) = V_{BE}$

$$V_{BE} = V_{CC} + [-V_{CC} + V_{BE} - V_{CC}]e^{-t/RC}$$

$R = R_{B2}$, $C_1 = C$

$$e^{-t/R_{B2} \times C_1} = \frac{V_{BE} - V_{CC}}{V_{BE} - 2V_{CC}}$$

$$e^{-t/R_{B2} \times C_1} = \frac{-V_{CC}}{-2V_{CC}}$$

$$e^{-t/R_{B2} \times C_1} = 1/2$$

By taking natural logarithmic

$$T_1 = -R_{B2} \times C_1 \times \ln[1/2]$$

$$= R_{B2} \times C_1 \times \ln[2]$$

$$T_2 = R_{B1} \times C_2 \times \ln[2]$$

$$T = T_1 + T_2$$

$$T = T_1 + T_2$$

$$T = [R_{B2} \times C_1 + R_{B1} \times C_2] \ln[2]$$

if $= R_{B2} = R_{B1} = R$

$C_1 = C_2 = C$

$$\boxed{T = 1.386RC}$$

$$f = \frac{1}{1.386RC} = \frac{0.72}{RC}$$

$$\boxed{f = \frac{0.72}{RC}}$$

Advantages:

1. No external triggering required
2. Circuit design is simple
3. Inexepensive

Disadvantages:

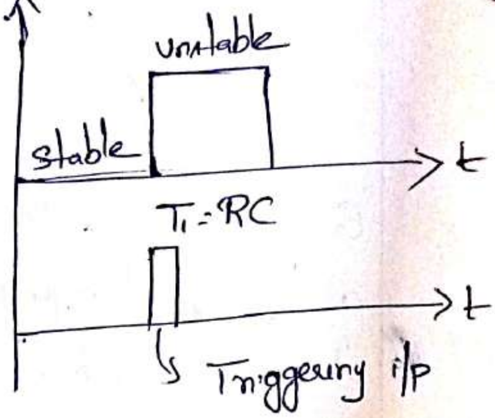
1) Energy absorption is more within the circuit
2) Output signal is of low energy
3) Duty cycle less than or equal to 50% cant be achived

Applications

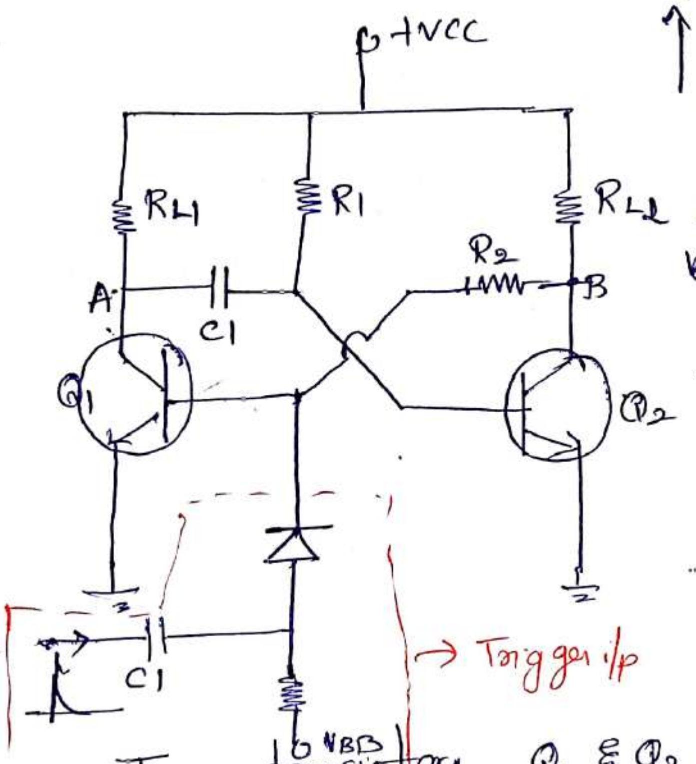
Astable multivibrators are used in many applications such as

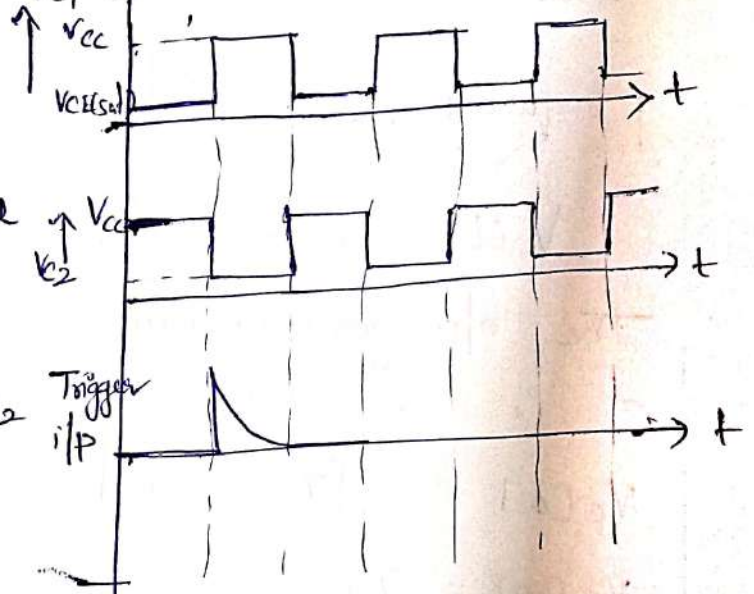
-> Radio equipment
-> Morse code generators
-> Timer circuits
-> Analog circuits & TV systems.

Mono stable Multivibrator: On One short multivibrator



Mono stable multivibrator consists of one stable state and one unstable state. Mono stable multivibrator changes its state from stable to unstable After application of a triggering input. The time duration of unstable state depends upon the values of R & C.





Two transistors, Q_1 & Q_2 are connected in a feedback configuration. Q1's collector connects to Q_2's base via capacitor C_1, and Q_1's base connects to Q_2's collector through resistor R_2 and capacitor C. A DC bias VBB is applied to Q_1's base through resistor R_3. A trigger pulse is given to Q_1's base through capacitor C_2 to change its state. R_{L1} and R_{L2} are the load resistors for Q_1 & Q_2. When one transistor is in a stable state, an external trigger shifts it to a quasi-stable [meta-stable] state for a time defined by the RC constant, after which it returns to the stable state.

Operation:

When the circuit is powered ~~on~~, on, Q_1 is OFF and Q_2 is ON. This is the stable state. With Q_1 OFF point A is at Vcc, charging capacitor C1. A trigger pulse at Q_1's base switches it ON, lowering its collector voltage and turning Q_2 OFF. C_1 begins discharging, keeping Q_1 ON - this is the quasi stable state. ~~Q~~ Once C_1 fully discharges, Q_2 turns ON again switching Q_1 OFF and returning the circuit to its original stable state.

Frequency of Oscillation:-

$$V_C(t) = V_C(\infty) + [V_C(0) - V_C(\infty)]\, e^{-t/RC}$$

The o/p goes unstable state but before voltage across C is $V_{CC} - V_{BE}$

$$V_C[0] = -[V_{CC} - V_{BE}]$$

$$V_C(\infty) = V_{CC}$$

at time T_1 voltage across the 'C' is V_{BE}

$$t = T_1$$

$$V_C(T_1) = V_{BE}$$

$$V_{BE} = V_C(\infty) + [V_C(0) - V_C(\infty)]\, e^{-T_1/R_2C}$$

$$V_{BE} = V_{CC} + [V_{BE} - V_{CC} - V_{CC}]\, e^{-T_1/R_2C}$$

$$e^{-T_1/R_2C} = \frac{V_{BE} - V_{CC}}{V_{BE} - 2V_{CC}}$$

$$V_{CC} \gg V_{BE}$$

$$e^{-T_1/R_2C} = \frac{-V_{CC}}{-2V_{CC}}$$

and

$$e^{-T_1/R_2 C} = 1/2$$

By natural log

$$-T_1/R_2 C = \ln[1/2]$$

$$T_1 = -\ln[1/2] R_2 C$$

$$T_1 = R_2 C \ln[2]$$

$$\boxed{T_1 = 0.693 R_2 C}$$

$$F = \frac{1}{T_1} = \frac{1}{0.693 R_2 C} \Rightarrow \boxed{F = \frac{1.44}{R_2 C}}$$

Advantages

- One trigger pulse is enough.
- Circuit design is simple
- Inexpensive

Disadvantages

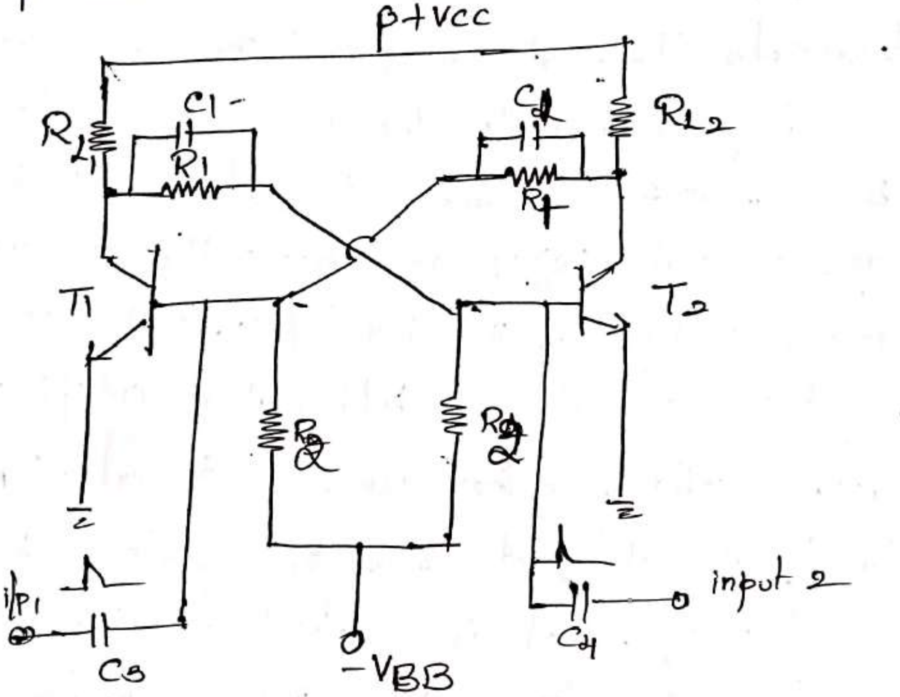
The major drawback of using a mono stable multivibrator is that the time between the application of trigger pulse than to be greater than the RC time constant of the circuit.

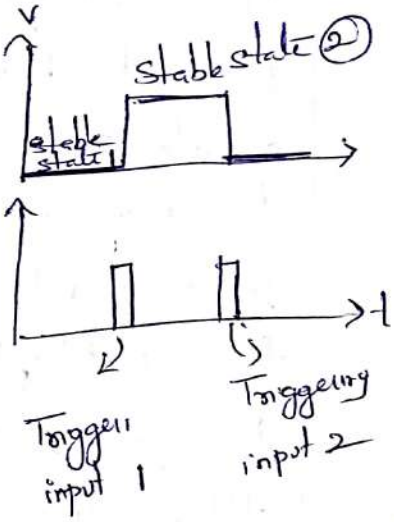
Application:

- Used in application such as television circuits and control system circuits
- pulse generation
- Time delay circuits
- Debouncing switches
- frequency divider
- Missing pulse detector

Bistable Multivibrator:

The bistable multivibrator has two stable states. The circuit will remain in any of its two states until a trigger pulse causes it to switch to the other state.





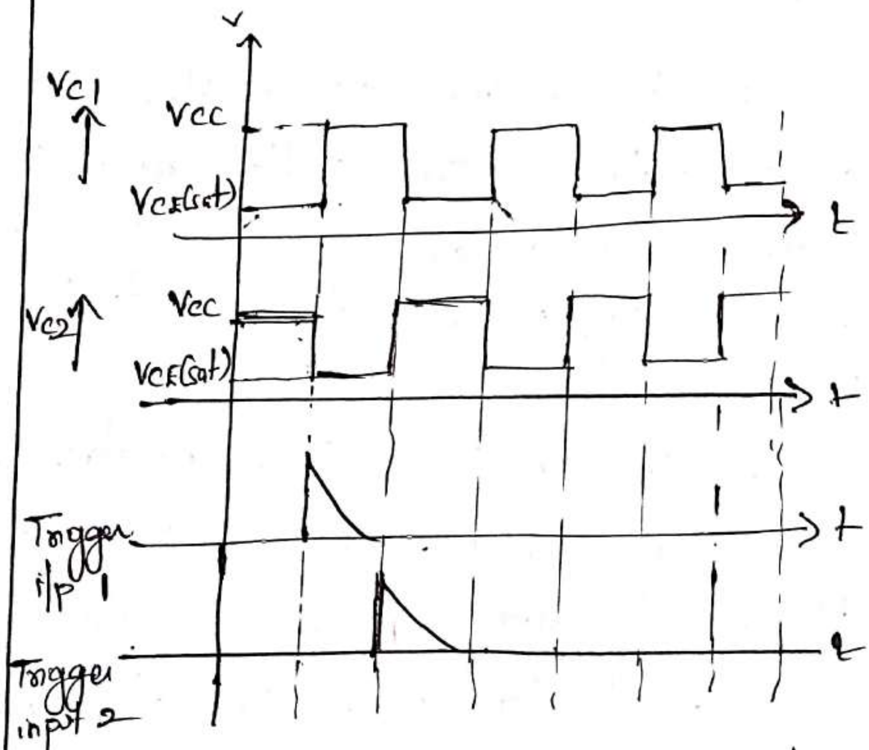


fig:. The o/p waveform at the collector of Q_1 and Q_2 along with trigger inputs given at the base of Q_1 & Q_2

Operation:

Transistors Q_1 and Q_2 are identical transistors. Assume that Q_1 is OFF and Q_2 is ON. Hence collector to emitter voltage of Q_1 is +Vcc. Then +Vcc becomes the input voltage of Q_2, and the former is of the polarity to forward bias transistor T2. Assuming that the two switch circuits have been designed properly, the forward bias applied to transistor T_2 will be sufficient to cause transistor T_2 in feedback on the input voltage to switch 1(Q_1). If no input voltage is applied to switch 1(Q_1) the negative voltage on the base of T1, with reference to ground, reverse biases transistor T2, holding it at cutoff or OFF. This negative voltage at the base of transistor T_1 with reference to ground, is obtained from the voltage drop across R_1. The voltage drop across R_1 is supplied from the voltage source VBB and in turn, from the resistive voltage divider, which consists of R_1 and R_2.

~~Switch 1~~ Q_1 will remain in this OFF condition and Q_2 remain in ON condition. An external voltage must be applied to change these stable states. When the forward bias of the transistor T_2 is removed, it will start to cutoff. As this occurs, the collector - emitter voltage will start to rise from $V_{CE\,sat}$ towards Vcc. This positive rising voltage, from the collector to emitter of T_2, is the input voltage to transistor T_1. Hence, transistor T_1 will start to conduct in turn, the output voltage will drop from +Vcc toward V_{CEsat}. Zero input voltage to Q_2 will cause transistor T_2 to be cutoff. The two [Q_1 & Q_2] will remain in this stable state namely Q_1 is ON & Q_2 is OFF. Because the circuit will remain in either of the two stable states indefinitely, this circuit is called a bistable multivibrator

Advantages:-

→ Stores the previous output unless disturbed.
→ Circuit design is simple
→ stable and reliable: Once in a stable state, it remains there until triggered
→ Versatile: Used in various digital applications

Disadvantages:-

→ Two kinds of trigger pulses are required
→ A bit costlier than other multivibrators.
→ Potential for Noise sensitivity: May be susceptible to noise in some applications

Applications:-

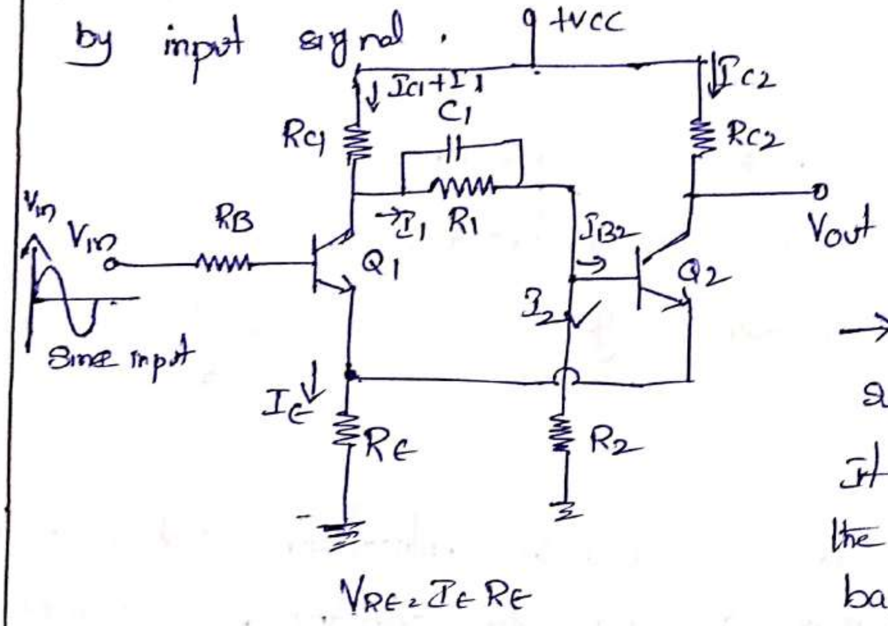
Bistable multivibrators are used in applications such as pulse generation and digital operations like counting and storing of binary information.

→ Switching
→ Counters & registers: Counting pulses or events
→ Frequency Dividers: Dividing the frequency of a signal
→ Timers: Generating precise time intervals
→ Flip Flops.

Schmitt Trigger

Schmitt trigger is a wave shaping circuit used to generate a square wave from any periodic signal, (i.e) sine, triangular (or) noisy square signals as input.

→ It is a bistable circuit whose output states are controlled by input signal.



→ Schmitt trigger uses regenerative feed back. It implies feedback where the output signal is added back to the input signal, resulting a positive feedback loop.

Construction:

→ It consists of two identical transistors Q_1 & Q_2 which are connected regeneratively through R_E.

→ R_1 & R_2 provides voltage divider bias

Operation:

1) Without any input voltage signal

When the supply voltage V_{CC} is switched ON the transistor Q_2' starts conducting the current which causes a voltage drop across R_E.

$$V_{RE} = I_E R_E$$

→ This voltage provides a reverse bias across Emitter Base junction of Q_1 and drives it in to cut-off state.

Q_1 is in OFF state and the voltage at collector rises to V_{CC}. (7)

→ This will increase the forward bias of Q_2.

→ Q_2 is in to saturation, $\boxed{\begin{aligned} V_{C1} &= V_{CC} \\ V_{C2} &= V_{CE}(sat) + V_{RE} \end{aligned}}$

(2) <u>with an input signal</u>

→ when the input is applied to Q_1, it remains in the OFF state until the voltage crosses the upper trigger potential [UTP]

∴ when $V_i \geq V_{RE} + V_{BE1}$, Q_1 conducts

→ The point at which Q_1 starts conducting is known as upper trigger point [UTP]

→ As Q_1 conducts, Q_2 is driven to cut-off region.

$$V_{C1} = V_{CE}(sat) + V_{RE}$$

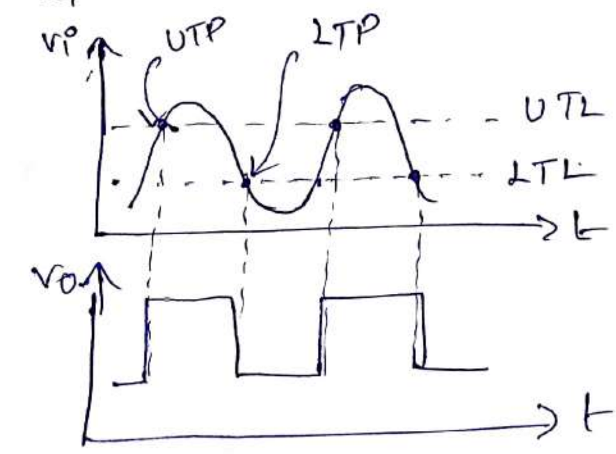
$$V_{C2} = V_{CC}$$

→ Q_1 conducts till the input voltage crosses the lower trigger potential.

ie $V_{in} < V_{RE} + V_{BE1}$, Q_1 is OFF and Q_2 starts conducting

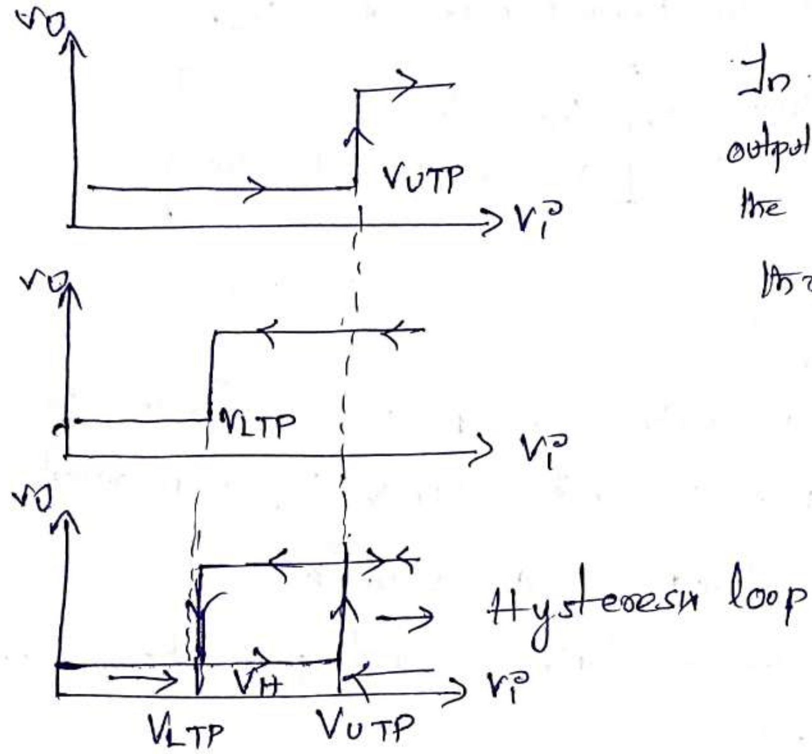
→ The point at which Q_2 starts conducting is called lower Triggering point (LTP]

$V_{C1} = V_{CC}$; $V_{C2} = V_{CE}(sat) + V_{RE}$



Hysteresis of the schmitt trigger

In schmitt trigger, the output state remains same until the Vi crosses any of the threshold level [UTL & LTL]



→ The difference between UTP and LTP is known as Hysteresis voltage 'VH'

→ It is also known as dead zone of the schmitt trigger.

Applications

→ It is used as a wave shaping circuit

→ It can be used as a voltage comparator.

→ Generation of rectangular wave form (or) Square wave from any periodic wave form.

→ Hysteresis concept is an important for conditioning the noisy signals in digital circuits