EC 2251 ELECTRONIC CIRCUITS II

UNIT - I

FEEDBACK AMPLIFIERS

Feed back

In the process of feedback, a part of output is sampled and fed back to the input of the amplifier.

Input has two signals,

- Input signal
- part of the output which is fed back to the input.

Both the signals may be inphase or out of phase.

Positive Feedback:

When input signal and part of output signal are in-phase, the feedback is called positive feedback.

Negative Feedback:

When they are in out-of phase, the feedback is called negative feedback.

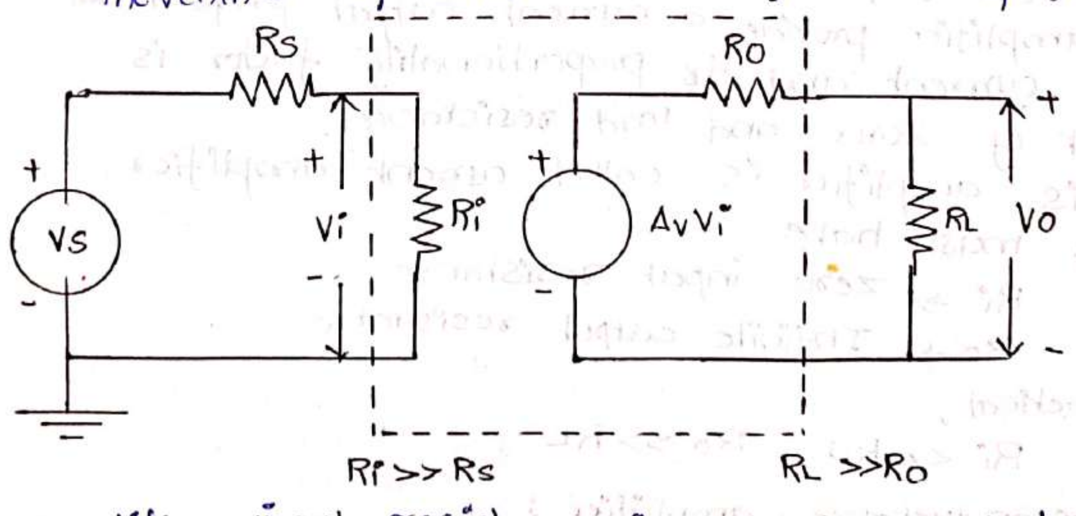
Classification of Amplifier:

Amplifiers can be classified as

1. Voltage
2. current
3. Transconductance
4. Transresistance

1. Voltage Amplifier:

Thevenin's equivalent circuit of an amplifier.



Amplifier input resistance is large compared with the source resistance ($R_i > R_s$)

then $V_i \simeq V_s$

If the external load resistance R_L is large compared with the output resistance R_o of the amplifier.

then
$\rightarrow$ $V_o = A_V V_i \simeq A_V V_s$

Such amplifier circuit provide a voltage output proportional to the voltage input and the proportionality factor does not depend on the magnitude of the source and load resistance.

Hence this amplifier is called Voltage amplifier

Voltage amplifier must have

$R_i \rightarrow$ infinite input resistance ($I \downarrow$)

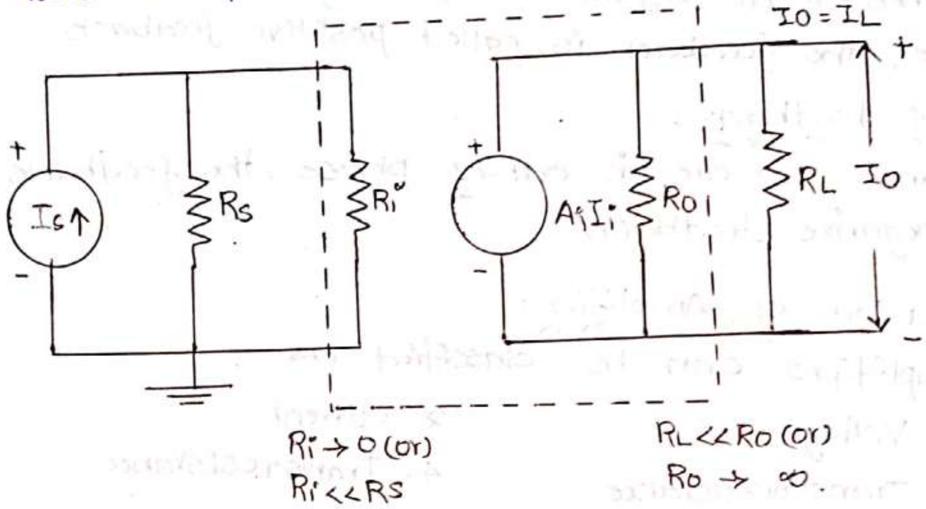
$R_o \rightarrow$ zero output resistance ($I \uparrow$)

Practical voltage amplifier

$R_i >> R_S$, $R_L >> R_o$

2. Current amplifier :

Norton's equivalent circuit of a current amp.



If $R_i \rightarrow 0$, then $I_i \simeq I_S$

If $R_O \rightarrow \infty$, then $I_L \simeq A_i I_i$

Such amplifier provide a current output proportional to the signal current and the proportionality factor is independent of source and load resistances.

This amplifier is called current amplifier.

It must have

$R_i \rightarrow$ zero input resistance

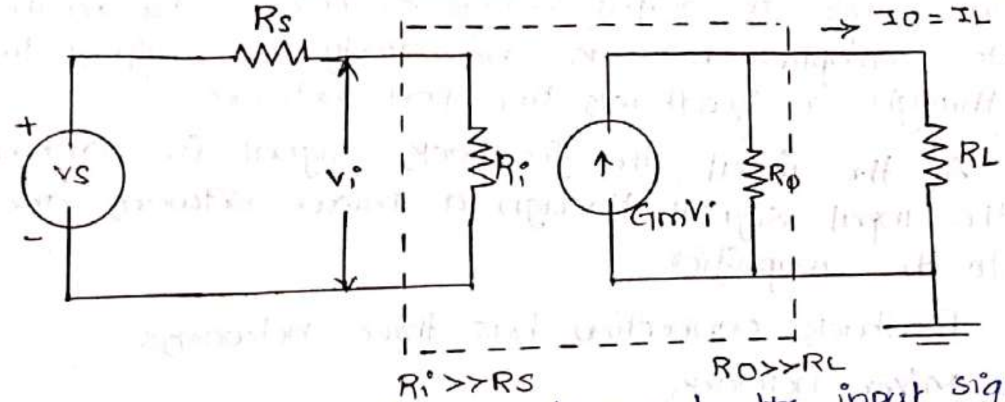
$R_o \rightarrow$ Infinite output resistance.

In practical,

$R_i << R_S$, $R_o >> R_L$

3. Transconductance amplifier :

Thevenin's equivalent circuit of an amplifier in its input circuit and Norton's equivalent in its output circuit.



The output current is proportional to the input signal voltage and the proportionality factor is independent of the magnitude of the source and load resistances.

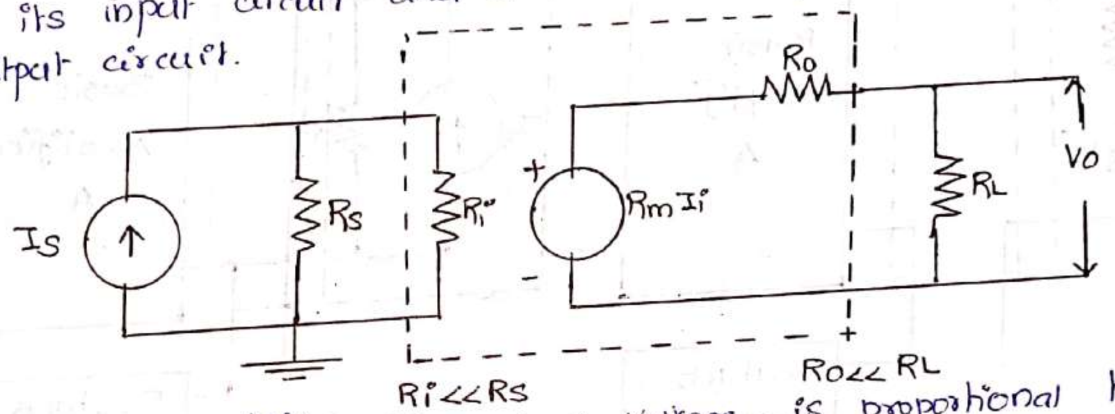
Amplifier must have

$R_i \rightarrow$ Infinite input resistance

$R_o \rightarrow$ Infinite output resistance

4. Transresistance Amplifier :

Transresistance amplifier with a Norton's equivalent in its input circuit and a 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 resistance.

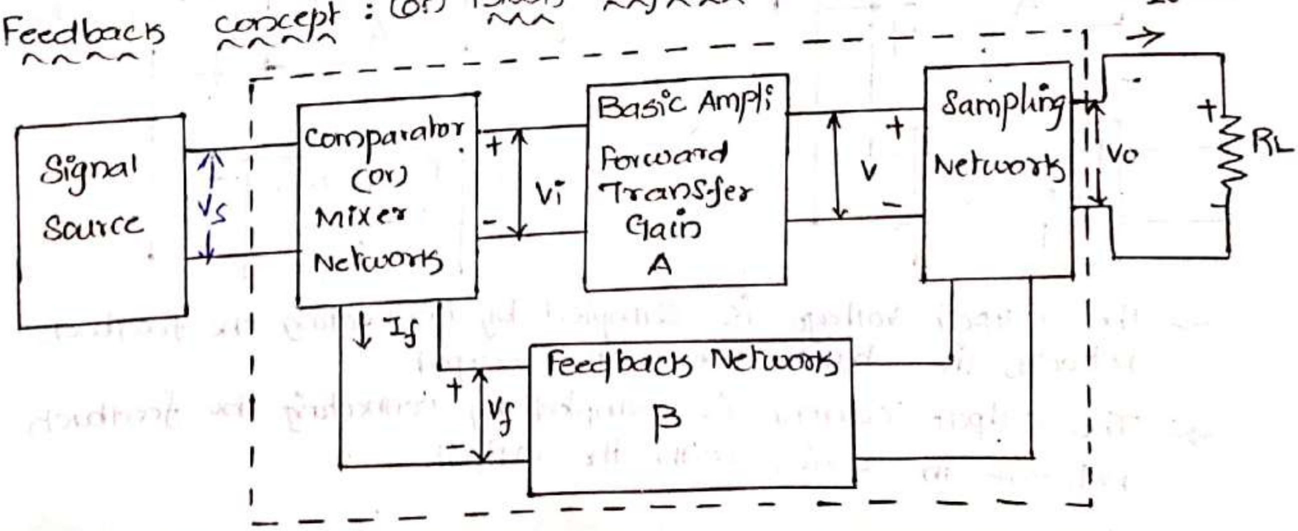
It must have,

$R_i \rightarrow$ Zero Input resistance

$R_o \rightarrow$ zero output resistance

$R_i << R_s$ and $R_o << R_L$

Feedback concept : (or) Block Diagram :



We can sample the output voltage or current by means of suitable sampling network and apply this signal to the input through a feedback two port network.

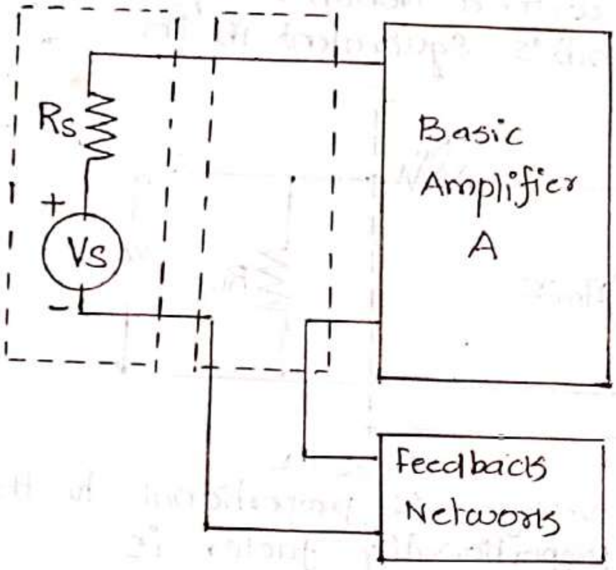
At the input, the feedback signal is combined with the input signal through a mixer network and is fed into the amplifier.

Feedback connection has three networks

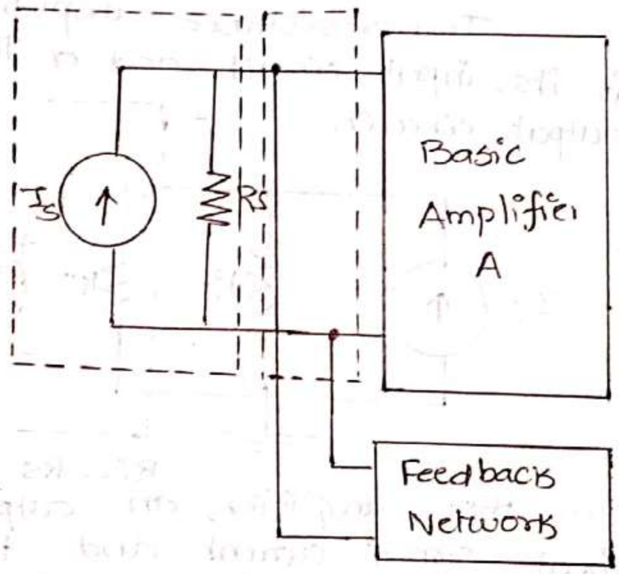
- Mixer network
- Feedback network
- Sampling network

Mixer Networks:

- Series Input connection
- Shunt Input connection

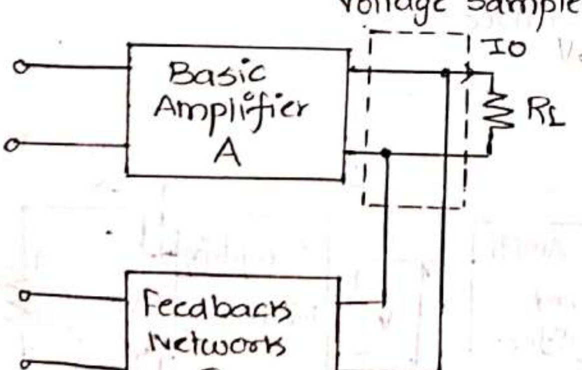


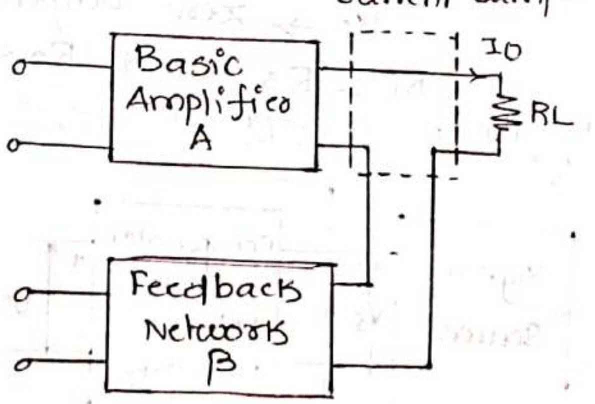
Series Mixing



Shunt Mixing

Sampling Networks:





- The output voltage is sampled by connecting the feedback network in shunt across the output.
- The output current is sampled by connecting the feedback network in series with the output.

Feedback Network :

It consists of resistors, capacitors and Inductors. often it is a resistive configuration.

It provides reduced portion of the output as feedback signal to the input mixer network.

It is given as

$\rightarrow V_f = \beta V_o$

β - feedback factor or feedback ratio and it lies between 0 and 1.

Transfer ratio (or) Gain :

The ratio of the output signal to the input signal of the basic amplifier is represented by the symbol A.

$\frac{V_o}{V_i} = A_V$ = Voltage gain ; $\frac{I_o}{V_i} = G_m$ = Transconductance

$\frac{I_o}{I_i} = A_I$ = Current gain ; $\frac{V_o}{I_i} = R_m$ = Transresistance

A_V, A_I, R_m, G_m are the transfer gain of the basic amplifier without feedback.

The transfer gain with feedback is represented by the symbol A_f.

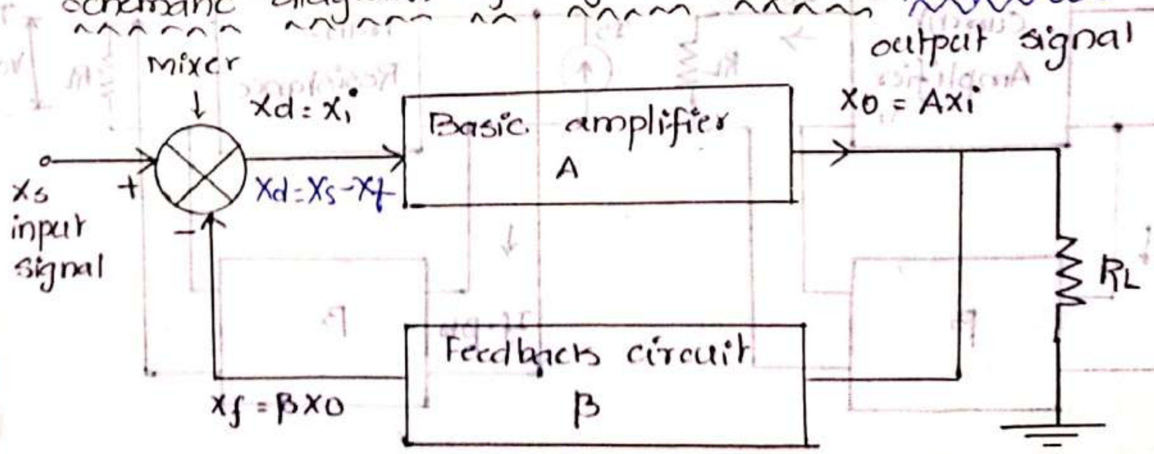
$\rightarrow \frac{V_o}{V_s} = A_{Vf}$ = Voltage gain with feedback

$\rightarrow \frac{I_o}{I_s} = A_{If}$ = Current gain with feedback

$\rightarrow \frac{I_o}{V_s} = G_{Mf}$ = Transconductance with feedback

$\rightarrow \frac{V_o}{I_s} = R_{Mf}$ = Transresistance with feedback

Schematic diagram of Negative Feedback Amplifier

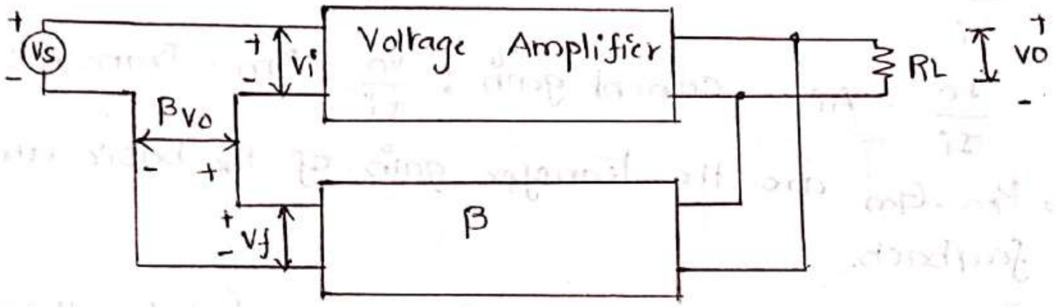


Advantages of Negative Feedback :

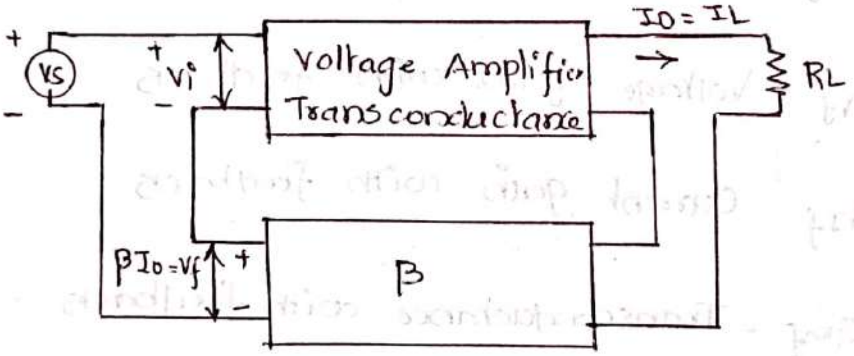
1. Normally high input resistance of a voltage amplifier can be made higher.
2. Normally low output resistance of a Voltage amplifier can be lowered.
3. The proper use of Negative feedback improves frequency response of the amplifier.

Four Basic Feedback Topologies :
(or)
Ways of Introducing negative feedback in amplifier :

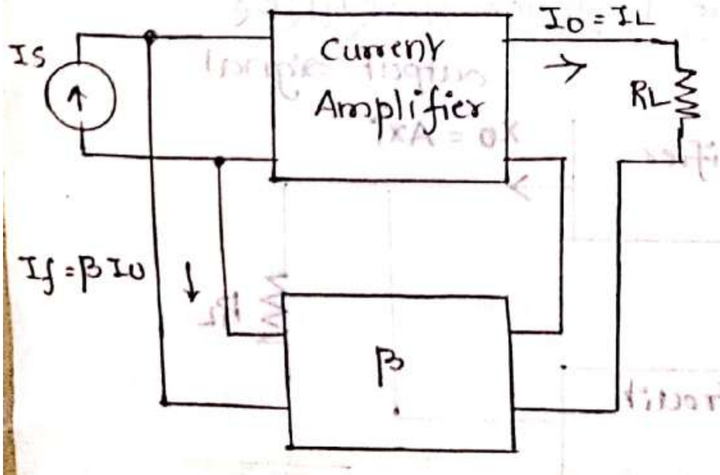
1. Voltage amplifier with Voltage series feedback :



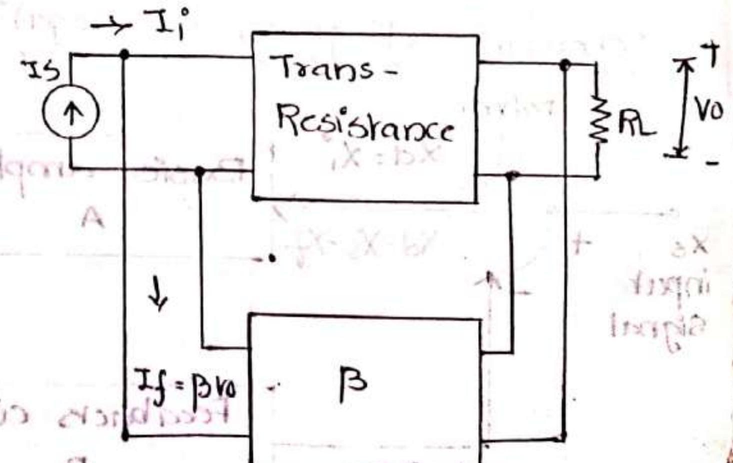
2. Transconductance amplifier with current series feedback :



3. current amplifier with current shunt feedback :



4. Transresistance amplifier with Voltage shunt feedback.



Effects of Negative Feedback :

1. Gain With Feedback {or} Loop gain {or} Transfer gain :-

The symbol A is used to represent transfer gain of the basic amplifier without feedback and the symbol A_f is used to represent transfer gain of the basic amplifier with feedback. and it is given as

$$\rightarrow A = \frac{X_o}{X_i} \quad \text{and} \quad A_f = \frac{X_o}{X_s}$$

Where X_o = output voltage or current
X_i = input voltage or current
X_s = Source voltage or current

As it is a negative feedback the relation between X_i and X_s is given as

$$\rightarrow X_i = X_s + (-X_f) \quad \& \quad X_s = X_i + X_f$$

Where X_f = feedback voltage or current

$$\rightarrow A_f = \frac{X_o}{X_s} = \frac{X_o}{X_i + X_f}$$

÷ by X_i to N & D, we get.

$$\rightarrow A_f = \frac{X_o/X_i}{\frac{X_i+X_f}{X_i}} \qquad \because A = X_o/X_i$$

$$\rightarrow A_f = \frac{A}{1+\frac{X_f}{X_i}} = \frac{A}{1+\left(\frac{X_f}{X_o}\right)\left(\frac{X_o}{X_i}\right)}$$

$\because$ w.k.t, $V_f = \beta V_o$

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

$X_f = \beta X_o$

$\beta = X_f/X_o$

So, we conclude that the gain without feedback (A) is always greater than gain with feedback $(A/1+\beta A)$ and it decreases with increase in β.

For Voltage Amplifier, gain with negative feedback as

$$\rightarrow A_{vf} = \frac{A_v}{1+A_v\beta}.$$

Where A_v = open loop gain, β = feedback factor.

2. Loop gain (or) Return Ratio :

In Negative feedback, the difference signal, X_d is multiplied by A in passing through the amplifier. is multiplied by β in transmission through the feedback network, and is multiplied by -1, in the mixing network. A path of a signal from input terminals through basic amplifier, through the feedback network and back to the input terminal forms a loop.

Upper cutoff frequency :

W.k.t, the relation between gain at high and mid frequency is given as,

$$\rightarrow \frac{A_{high}}{A_{mid}} = \frac{1}{1-j(f/f_H)}$$

$$\rightarrow A_{high} = \frac{A_{mid}}{1-j(f/f_H)}$$

sub A_{high} in A_f high

$$\rightarrow A_f\,high = \frac{\frac{A_{mid}}{1+(-j)\,f/f_H}}{1+\beta\left[\frac{A_{mid}}{1-j(f/f_H)}\right]} = \frac{\frac{A_{mid}}{1-j(f/f_H)}}{1+\beta\left[\frac{A_{mid}}{1-j(f/f_H)}\right]}$$

$$\rightarrow A_f\,high = \frac{\frac{A_{mid}}{1-j(f/f_H)}}{\frac{1-j(f/f_H)+\beta A_{mid}}{1-j(f/f_H)}} = \frac{A_{mid}}{1-j(f/f_H)+\beta A_{mid}}$$

$\div$ N & D by $(1+A_{mid}\beta)$,

$$\rightarrow A_f\,high = \frac{A_{mid}/1+(A_{mid}\beta)}{\frac{1+\beta A_{mid}-j(f/f_H)}{1+A_{mid}\beta}} = \frac{A_{mid}/1+A_{mid}\beta}{1-\frac{j(f/f_H)}{1+A_{mid}\beta}}$$

$$\rightarrow A_f\,high = \frac{A_f(mid)}{1-j(f/f_{Hf})} \qquad \because A_{fmid} = A_{mid}/1+A_{mid}\beta$$

$$\rightarrow \frac{A_f\,high}{A_f(mid)} = \frac{1}{1-j(f/f_{Hf})} \qquad f_{Hf} = (1+A_{mid}\beta)f_H$$

Where f_{Hf} = upper cutoff frequency with feedback = $f_H(1+A_{mid}\beta)$

The upper cutoff frequency with feedback is greater than upper cut off frequency without feedback by factor $(1+A_{mid}\beta)$

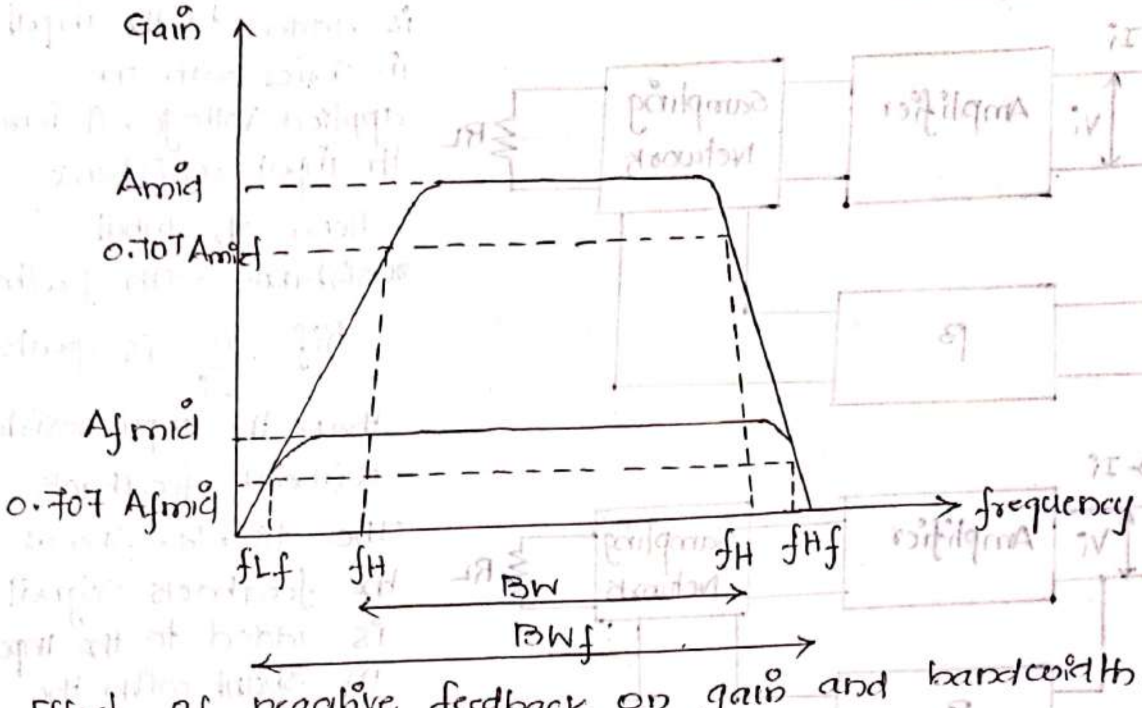
5. Bandwidth :

The Bandwidth of the amplifier is given as

$\rightarrow$ BW = upper cutoff frequency − lower cutoff frequency

Bandwidth of the amplifier with feedback is given as

$\Rightarrow BW = f_{Hf} - f_{Lf} = (1 + A_{mid}\beta) f_H - \frac{f_L}{1 + A_{mid}\beta}$



Effect of negative feedback on gain and bandwidth.

Hence bandwidth of amplifier with feedback is greater than bandwidth of amplifier without feedback.

6. Distortion :

1. Frequency Distortion :

If the feedback network does not contain reactive elements, the overall gain is not a function of frequency, so the frequency and phase distortion is reduced.

If β is made of reactive components, the reactance of these components will change with frequency, change the β. As a result gain will also change with frequency.

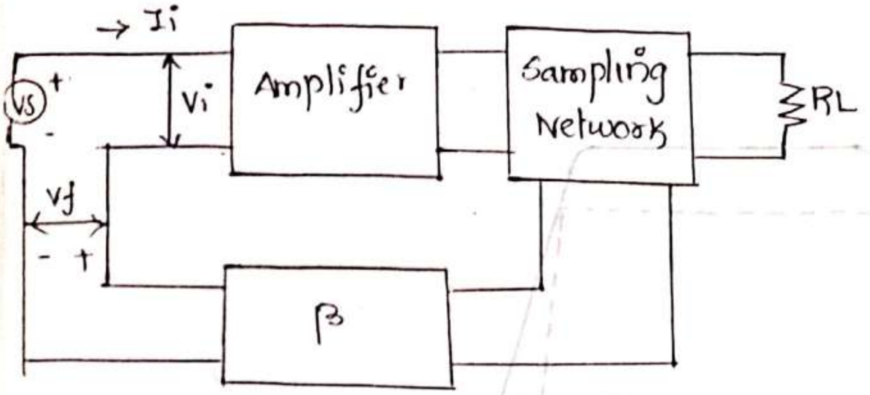
This is used in tuned amplifiers. In tuned amplifiers feedback network is designed such that at tuned frequency $\beta \to 0$ and at other frequency $\beta \to \infty$. As a result amplifier provides high gain for signals at tuned frequency and reject all other frequency.

2. Noise and Non-linear distortion :

Signal feedback reduces the amount of noise signal and non linear distortion. The factor $(1 + \beta A)$ reduces both input noise and resulting non-linear distortion.

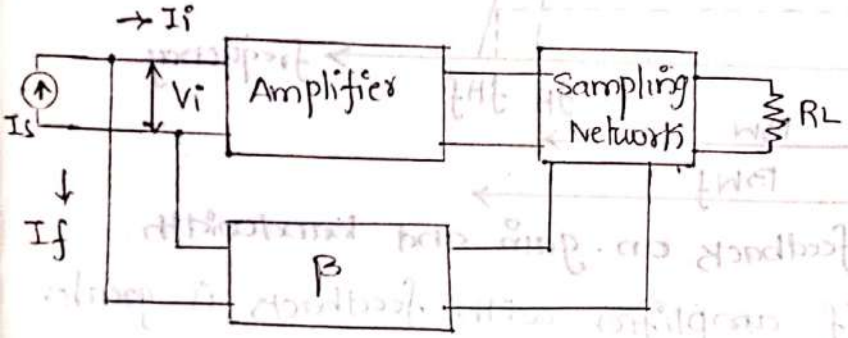
Input Impedance and output Impedance with feedback:

Input resistance:



If the feedback signal is added to the input in series with the applied Voltage. it increas the input resistance. Hence, the input resistance with feedback

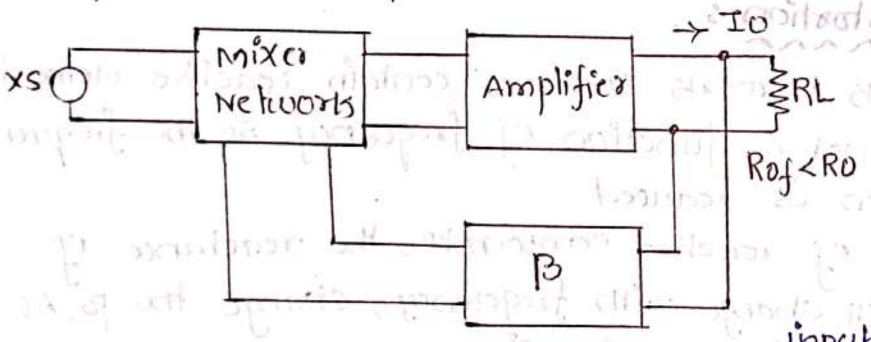
$$R_{if} = \frac{V_s}{I_i}$$ is greater than the input resistance without feedback.



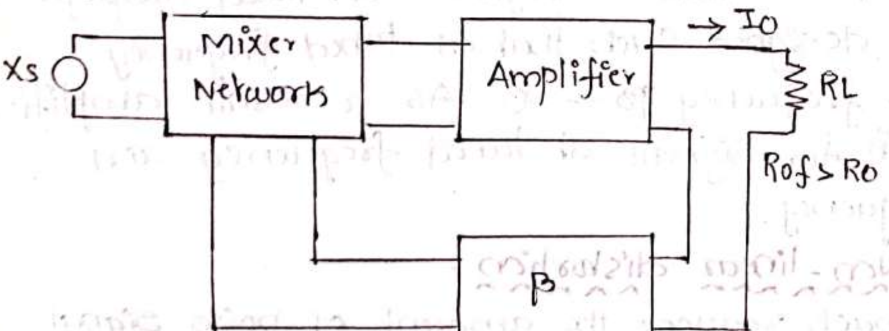
The dia. 2. Shows the feedback signal is added to the input in Shunt with the applied Voltage, it decrea the input resistance. Since, $I_s = I_i + I_f$

Hence, the input resistance with feedback, $R_{if} = \frac{V_i}{I_s}$ is decreased

output resistance:



The negative feedback which samples the output Voltage, regardless of how the output signal is returned to the input, tends to decrease the output resistance.

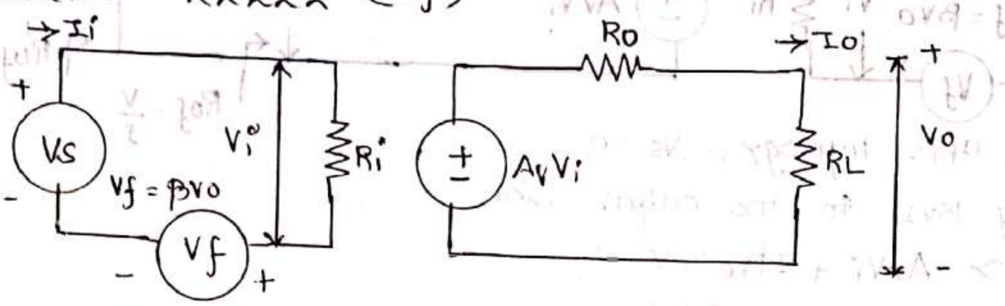


The dia. 2 Shows the negative feedback which Samples the output current, regardless of how the output signal is returned to the output, tends to increase the output resistance.

Four Types of Negative Feedback connections :

1. Voltage series Feedback : (Series - shunt)

Input Resistance :(R_{if})



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

The above dia. shows Voltage series feedback topology with amplifier is replaced by thevenin's model.

From the dia,

$\rightarrow \quad R_{if} = \frac{V_S}{I_i} \rightarrow (1)$

Apply KVL to the Input Side,

$\rightarrow \quad V_S - I_i R_i - V_f = 0 \quad \because V_f = \beta V_O.$

$\rightarrow \quad V_S = I_i R_i + \beta V_O \rightarrow (2)$

The output Voltage is given as,

$\rightarrow \quad V_o = \frac{A_V V_i R_L}{R_O + R_L} \quad \because \; AV = \frac{A_V R_L}{R_O + R_L}$

$\rightarrow \quad V_o = AV\,R V_i = AV V_i \rightarrow (3)$

Sub V_o in (2)

$\rightarrow \quad V_S = I_i R_i + \beta V_O$

$\rightarrow \quad V_S = I_i R_i + \beta\, AV V_i \quad \because V_i = R_i I_i$

$\rightarrow \quad V_S = I_i R_i + \beta\, AV R_i I_i$

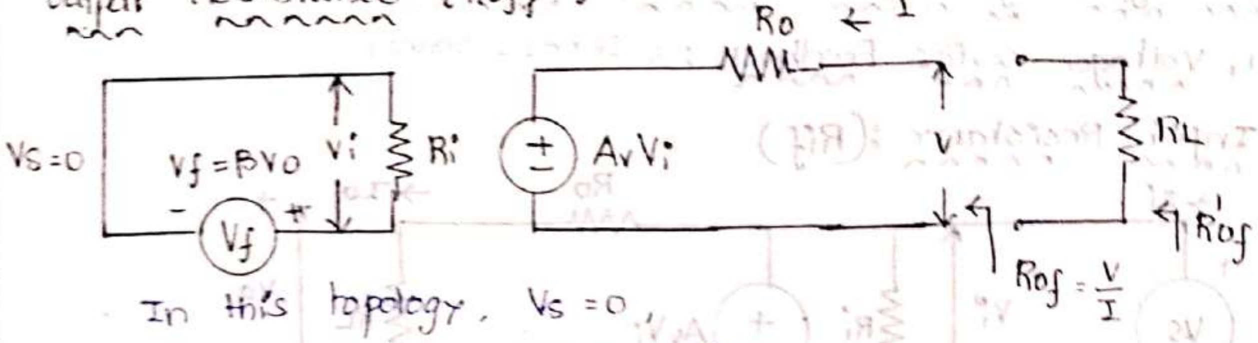
$\rightarrow \quad V_S = I_i \{R_i + \beta\, AV R_i\}$

$\rightarrow \quad \frac{V_S}{I_i} = R_i + \beta\, AV R_i$

$\rightarrow \quad R_{if} = R_i + \beta\, AV R_i \quad \because R_{if} = \frac{V_S}{I_i}$

$\rightarrow \quad R_{if} = R_i \{1 + \beta AV\} \rightarrow (4)$

Output Resistance (R_{of}) :



In this topology, $V_s = 0$,

Apply KVL to the output side,

$\rightarrow A_V V_i + I R_o - V = 0$

$\rightarrow I R_o = V - A_V V_i$

$\rightarrow I = \dfrac{V - A_V V_i}{R_o} \rightarrow (5)$

The Input voltage is given as

$\rightarrow V_i = -V_f = -\beta V \rightarrow (6)$

Sub Eq (6) in (5)

$$\rightarrow I = \frac{V - A_V V_i}{R_o} = \frac{V + A_V \beta V}{R_o} = \frac{V(1+\beta A_V)}{R_o}$$

$$\rightarrow R_{of} = \frac{V}{I} = \frac{R_o}{1+\beta A_V} \rightarrow (7)$$

$\rightarrow$ Now

$$\rightarrow R'_{of} = R_{of} \parallel R_L = \frac{R_{of} R_L}{R_{of} + R_L} = \frac{\frac{R_o}{1+\beta A_V} \cdot R_L}{\frac{R_o}{1+\beta A_V} + R_L}$$

$$\rightarrow R'_{of} = \frac{\frac{R_o R_L}{1+\beta A_V}}{\frac{R_o + R_L(1+\beta A_V)}{1+\beta A_V}} = \frac{R_o R_L}{R_o + R_L + R_L \beta A_V}$$

$\div$ N & D by $R_o + R_L$

$$\rightarrow R'_{of} = \frac{R_o R_L / R_o + R_L}{\frac{R_o + R_L + R_L \beta A_V}{R_o + R_L}} = \frac{R_o R_L / R_o + R_L}{1 + \frac{R_L \beta A_V}{R_o + R_L}}$$

$$\rightarrow R'_{of} = \frac{R'_o}{1 + \beta A'_V}$$

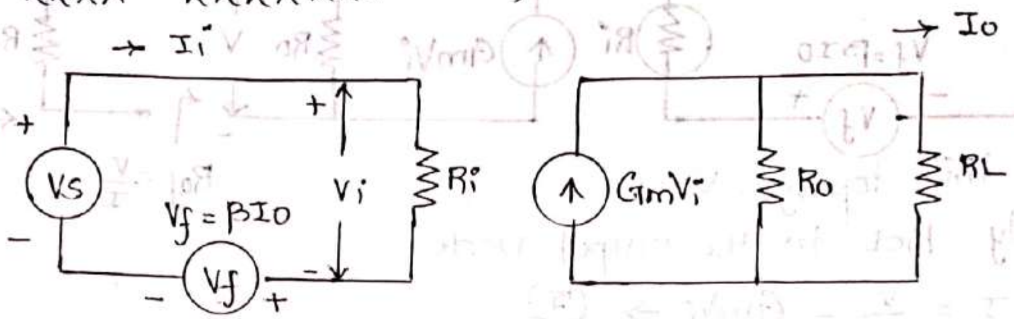
$\because R'_o = \dfrac{R_o R_L}{R_o + R_L}$

$A'_V = \dfrac{A_V R_L}{R_o + R_L}$

Derive i/p and o/p impedance of an current series feedback

2. Current Series Feedback : (Series - Series)

Input Resistance : {R_{if}}



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

The above dia, the amplifier input circuit is Thevenin's model and output circuit is Norton's Model.

From the dia,

$$\Rightarrow R_{if} = \frac{V_s}{I_i} \rightarrow (1)$$

Apply KVL to the Input side,

$$\Rightarrow V_s - R_i I_i - V_f = 0 \quad \because V_f = \beta I_O$$

$$\Rightarrow V_s = R_i I_i + V_f$$

$$\Rightarrow V_s = R_i I_i + \beta I_O \rightarrow (2)$$

The output current, I_O is given as

$$\Rightarrow I_O = \frac{G_m V_i R_o}{R_o + R_L} \quad \because G_M = \frac{G_m R_o}{R_o + R_L}$$

$$\Rightarrow I_O = G_M V_i \rightarrow (3)$$

Sub Eq. (3) in (2)

$$\Rightarrow V_s = R_i I_i + \beta I_O$$

$$\Rightarrow V_s = R_i I_i + \beta G_M V_i \quad \because V_i = R_i I_i$$

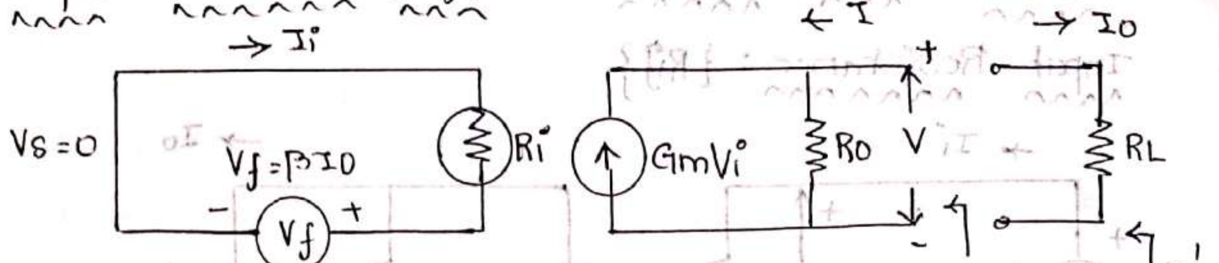
$$\Rightarrow V_s = R_i I_i + \beta G_M R_i I_i$$

$$\Rightarrow V_s = I_i \{ R_i + \beta G_M R_i \} \quad \because R_{if} = \frac{V_s}{I_i}$$

$$\Rightarrow R_{if} = \frac{V_s}{I_i} = R_i + \beta G_M R_i$$

$$\Rightarrow R_{if} = R_i \{1 + \beta G_M\} \rightarrow (4)$$

Output Resistance {R_{of}}:



In this topology, $V_S = 0$.

Apply KCL to the output node,

$$\Rightarrow I = \frac{V}{R_o} - G_m V_i \rightarrow (5)$$

The Input Voltage is

$$\rightarrow V_i = -V_f = -\beta I_0 \qquad \because I_0 = -I$$

$$\rightarrow V_i = \beta I \rightarrow (6)$$

Sub Eq. (6) in (5)

$$\Rightarrow I = \frac{V}{R_o} - G_m V_i = \frac{V}{R_o} - G_m \beta I$$

$$\Rightarrow I + G_m \beta I = \frac{V}{R_o}$$

$$\Rightarrow I\{1 + G_m \beta\} = \frac{V}{R_o}$$

$$\Rightarrow I\{1 + G_m \beta\} R_o = V$$

$$\rightarrow (1 + G_m \beta) R_o = \frac{V}{I}. \qquad \because R_{of} = \frac{V}{I}.$$

$$\Rightarrow R_{of} = \frac{V}{I} = (1 + G_m \beta) R_o \rightarrow (7)$$

$$\Rightarrow R'_{of} = R_{of} \parallel R_L = \frac{R_{of} R_L}{R_{of} + R_L} = \frac{(1 + G_m \beta) R_o R_L}{(1 + G_m \beta) R_o + R_L}$$

$$\Rightarrow R'_{of} = \frac{R_o R_L (1 + G_m \beta)}{R_o + R_o G_m \beta + R_L} = \frac{R_o R_L (1 + G_m \beta)}{R_o + R_L + R_o G_m \beta}$$

$\div$ N & D by $R_o + R_L$,

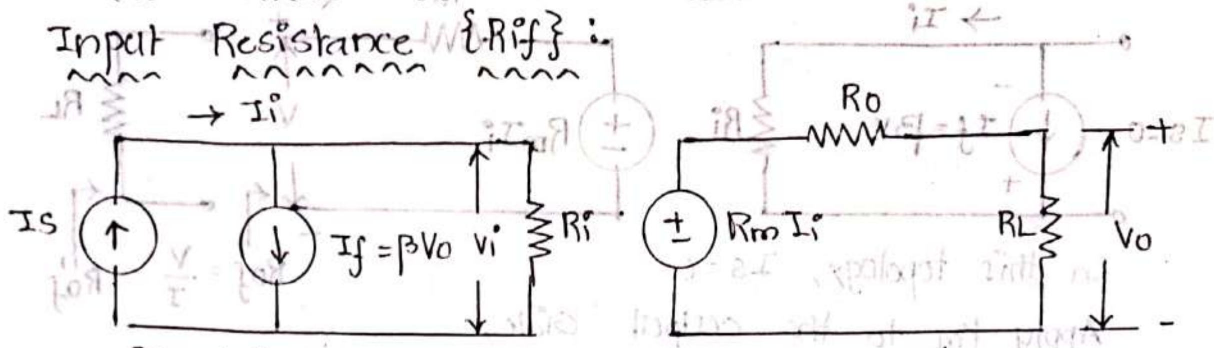
$$\rightarrow R'_{of} = \frac{R_o R_L (1 + G_m \beta) / R_o + R_L}{R_o + R_L + R_o G_m \beta / R_o + R_L} = \frac{R_o R_L (1 + G_m \beta) / R_o + R_L}{1 + \frac{R_o G_m \beta}{R_o + R_L}}$$

$$\rightarrow R'_{of} = \frac{R'_o (1 + G_m \beta)}{1 + \beta G_M} \qquad \because R'_o = \frac{R_o R_L}{R_o + R_L}$$

$$G_M = \frac{G_m R_o}{R_o + R_L}$$

3. Voltage Shunt Feedback : (Shunt - Shunt)

Input Resistance {R_{if}} :



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

The above dia. the amplifier circuit (input) is Norton's model and output circuit is Thevenin's model.

Apply KCL at input node :

→ $I_S = I_i + I_f$ $\quad \because I_f = \beta V_o$

→ $I_S = I_i + \beta V_o \rightarrow (1)$

The output voltage is

→ $V_o = \dfrac{R_m I_i R_o}{R_o + R_L}$ $\quad \because R_M = \dfrac{R_m R_o}{R_o + R_L}$

→ $V_o = R_M I_i \rightarrow (2)$

Sub Eq. (2) in (1)

→ $I_S = I_i + \beta V_o$

→ $I_S = I_i + \beta R_M I_i$

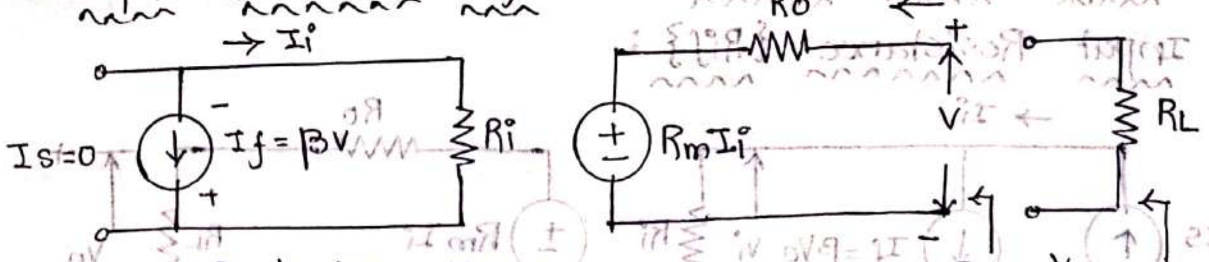
→ $I_S = I_i (1 + \beta R_M)$

WKT, $R_{if} = \dfrac{V_i}{I_S} = \dfrac{V_i}{I_i (1 + \beta R_M)}$ $\quad \because R_i = \dfrac{V_i}{I_i}$

→ $R_{if} = \dfrac{R_i}{1 + \beta R_M} \rightarrow (3)$

→ $R_{if} = \dfrac{R_i}{1 + \beta R_M}$

Output Resistance $\{R_{of}\}$:



In this topology, $I_s=0$

Apply KVL to the output side,

$\rightarrow R_m I_i + I R_o - V = 0$

$\rightarrow I R_o = V - R_m I_i$

$\rightarrow I = \dfrac{V - R_m I_i}{R_o} \rightarrow (4)$

The Input current is

$\rightarrow I_i = -I_f = -\beta V \rightarrow (5)$

Sub Eq. (5) in (4)

$\rightarrow I = \dfrac{V - R_m I_i}{R_o} = \dfrac{V + R_m \beta V}{R_o} = \dfrac{V\{1 + R_m \beta\}}{R_o}$

$\rightarrow I R_o = V(1 + R_m \beta)$

$\rightarrow R_{of} = \dfrac{V}{I} = R_o/(1 + R_m\beta) = \dfrac{R_o}{1+R_m\beta} \rightarrow (6)$

$\rightarrow R'_{of} = R_{of} \parallel R_L = \dfrac{R_{of} R_L}{R_{of} + R_L} = \dfrac{\dfrac{R_o}{1+R_m\beta} \cdot R_L}{\dfrac{R_o}{1+R_m\beta} + R_L}$

$\rightarrow R'_{of} = \dfrac{\dfrac{R_o R_L}{1+R_m\beta}}{\dfrac{R_o + R_L(1+R_m\beta)}{1+R_m\beta}} = \dfrac{R_o R_L}{R_o + R_L + R_L R_m \beta}$

$\div$ N & D by $R_o + R_L$

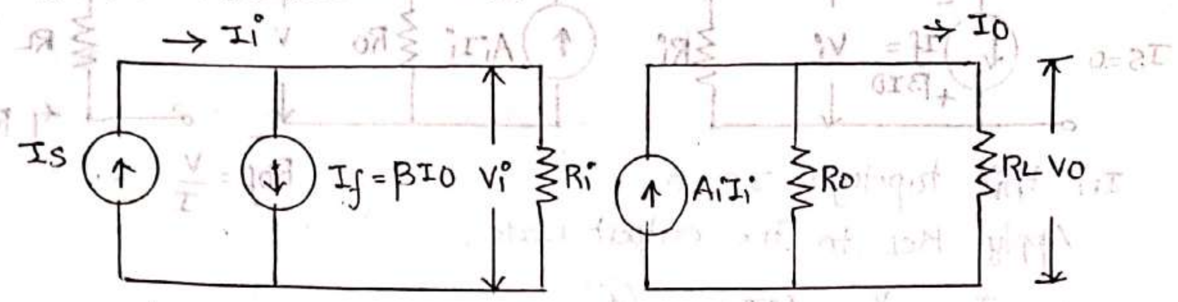
$\rightarrow R'_{of} = \dfrac{R_o R_L / R_o + R_L}{\dfrac{R_o + R_L + R_L R_m \beta}{R_o + R_L}} = \dfrac{R_o R_L / R_o + R_L}{1 + \dfrac{R_L R_m \beta}{R_o + R_L}}$

$\rightarrow R'_{of} = \dfrac{R'_o}{1 + \beta R_M} \rightarrow (7)$ $\quad \because R'_o = \dfrac{R_o R_L}{R_o + R_L}$

$R_M = \dfrac{R_m R_L}{R_o + R_L}$

4. Current Shunt Feedback : (Shunt - Series)

Input Resistance {Rif} :



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

The above diagram, the amplifier input circuit is Norton's Model and output circuit is also Norton's Model.

Apply KCL to input side,

$\Rightarrow I_s = I_i + I_f \quad \because I_f = \beta I_o$.

$\rightarrow I_s = I_i + \beta I_o. \rightarrow (1)$

The output current is

$\rightarrow I_o = \frac{A_i I_i R_o}{R_o + R_L} \qquad \because A_I = \frac{A_i R_o}{R_o + R_L}$

$\rightarrow I_o = A_I I_i \rightarrow (2)$

Sub Eq. (2) in (1)

$\rightarrow I_s = I_i + \beta I_o$

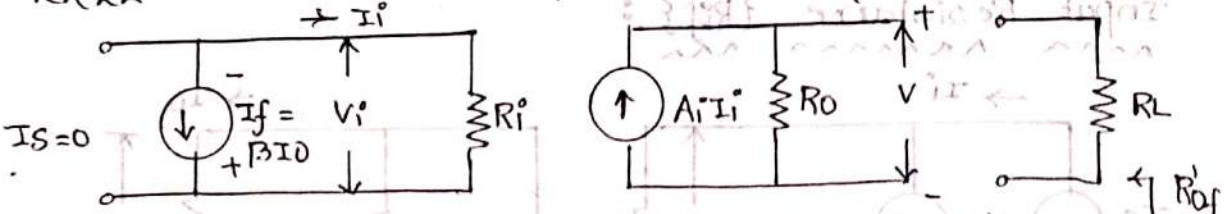
$\rightarrow I_s = I_i + \beta A_I I_i$

$\rightarrow I_s = I_i \{1 + \beta A_I\}$

$\rightarrow R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i\{1+\beta A_I\}} \qquad \because R_i = \frac{V_i}{I_i}$

$\rightarrow R_{if} = \frac{R_i}{(1+\beta A_I)}$

Output Resistance : $\{R_{of}\}$



In this topology, $I_S = 0$,

Apply KCL to the output node,

$\Rightarrow I = \frac{V}{R_o} - A_i I_i \rightarrow (4)$

The input current is

$\Rightarrow I_i = -I_f = -\beta I_O \quad \because I = -I_O$

$\Rightarrow I_i = \beta I \rightarrow (5)$

Sub Eq. (5) in (4)

$\Rightarrow I = \frac{V}{R_o} - A_i I_i = \frac{V}{R_o} - A_i \beta I$

$\Rightarrow I + A_i \beta I = \frac{V}{R_o}$

$\Rightarrow I(1 + A_i \beta) = \frac{V}{R_o}$

$\Rightarrow R_{of} = \frac{V}{I} = R_o \{1 + \beta A_i\} \rightarrow (6)$

$\Rightarrow R'_{of} = R_{of} \parallel R_L = \frac{R_{of} R_L}{R_{of} + R_L} = \frac{R_o (1 + \beta A_i) R_L}{R_o (1 + \beta A_i) + R_L}$

$\Rightarrow R'_{of} = \frac{R_o R_L (1 + \beta A_i)}{R_o + R_o \beta A_i + R_L} = \frac{R_o R_L (1 + \beta A_i)}{R_o + R_L + R_o \beta A_i}$

$\div$ N & D by $R_o + R_L$

$\Rightarrow R'_{of} = \frac{R_o R_L (1 + \beta A_i) / R_o + R_L}{R_o + R_L + R_o \beta A_i / R_o + R_L} = \frac{\frac{R_o R_L (1 + \beta A_i)}{R_o + R_L}}{1 + \frac{R_o \beta A_i}{R_o + R_L}}$

$\Rightarrow R'_{of} = \frac{R'_o (1 + \beta A_i)}{1 + \beta A_I} \rightarrow (7) \quad \because R'_o = \frac{R_o R_L}{R_o + R_L}$

$\Rightarrow R'_{of} = \frac{R'_o (1 + \beta A_i)}{1 + \beta A_I} \qquad A_I = \frac{A_i R_o}{R_o + R_L}$

Compare and contrast on Vtg shunt and Vtg series feedback

Advantages of Negative Feedback:

1. Increases Stability
2. Increased Bandwidth & frequency Response
3. Reduced Frequency Distortion
4. Reduced Noise & Non-linear distortion.

Disadvantages:

It Reduces the Gain of the amplifier.

Effects of Negative Feedback on Amplifiers:

Parameter	Voltage Series	current Series	Voltage Shunt	current Shunt
Gain With Feedback	$A_{vf} = \frac{A_v}{1+\beta A_v}$ decreases	$G_{mf} = \frac{G_m}{1+\beta G_m}$ decreases	$A_{if} = \frac{A_i}{1+\beta A_i}$ decreases	$R_{mf} = \frac{R_m}{1+\beta R_{mf}}$ decreases
Stability	Improves	Improves	Improves	Improves
Frequency Response	Improves	Improves	Improves	Improves
Frequency Distortion	Reduces	Reduces	Reduces	Reduces
Noise and Non-linear Distortion	Reduces	Reduces	Reduces	Reduces
Input Resistance	$R_{if} = R_i(1+\beta A_v)$ increase	$R_{if} = R_i(1+\beta G_m)$ increase	$R_{if} = R_i/_{1+\beta R_m}$ decrease	$R_{if} = R_i/_{1+\beta A_i}$ decrease
output Resistance	$R_{of} = R_o/_{1+\beta A_v}$ decrease	$R_{of} = R_o(1+\beta G_m)$ increase	$R_{of} = R_o/_{1+\beta R_m}$ decrease	$R_{of} = R_o(1+\beta A_i)$ Increase

Method of Identifying Feedback topology and Feedback Factor:

To analyse the feedback amplifier, it is necessary to go through the following steps.

Step 1: Identify Topology

a. To find the type of sampling network:

1. By shorting the o/p ie) $V_o = 0$, if feedback signal (x_f) becomes zero, then it is Voltage Sampling.
2. By opening the o/p loop ie) $I_o = 0$, if feedback signal (x_f) becomes zero, then it is current sampling.

b. To find the type of Mixing network:

1. If the feedback signal is subtracted from the externally applied signal as a Voltage in i/p loop, it is Series Mixing.

2. If the feedback signal is subtracted from the externally applied signal as a current in the i/p loop, it is shunt mixing.

For example,

If amplifier uses a voltage sampling and series mixing, then it is voltage series amplifier.

Step 2: Find Input circuit:

1. For voltage sampling make $V_o = 0$ by shorting the output.
2. For current sampling make $I_o = 0$ by opening the output loop.

Step 3: Find output circuit:

1. For series mixing make $I_i = 0$ by opening the input loop.
2. For shunt mixing make $V_i = 0$ by shorting the input.

Step 4: optional. Replace each active device by its h-parameter model at low frequency.

Step 5: Find the open loop gain (gain without feedback), A of the amplifier.

Step 6: Indicate X_f and X_o on the circuit and Evaluate $\beta = X_f / X_o$.

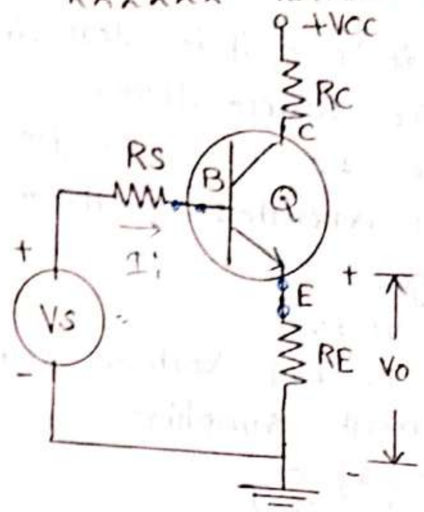
Step 7: From A & B, find D, A_f, R_{if}, R_{of} and R'_{of}.

characteristics	Topology			
	Voltage series	current series	current shunt	voltage shunt
Sampling signal	Voltage	Voltage	current	current
Mixing signal	Voltage	current	current	Voltage
To find Input loop, set	$V_o = 0$	$I_o = 0$	$I_o = 0$	$V_o = 0$
To find output loop, set	$I_i = 0$	$I_i = 0$	$V_i = 0$	$V_i = 0$
Signal Source	Thevenin	Thevenin	Norton	Norton

Analysis of Feedback Amplifiers :

1. Voltage series feedback :

Transistor Emitter Follower :



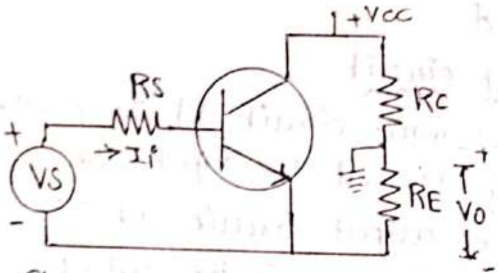
Step 1 : Identify Topology :

By shorting output voltage ($V_o = 0$), feedback signal becomes zero and hence it is voltage sampling.

The feedback signal V_f is subtracted from the externally applied signal V_s and hence it is series mixing.

combining two conclusion, then it is voltage series feedback amplifier.

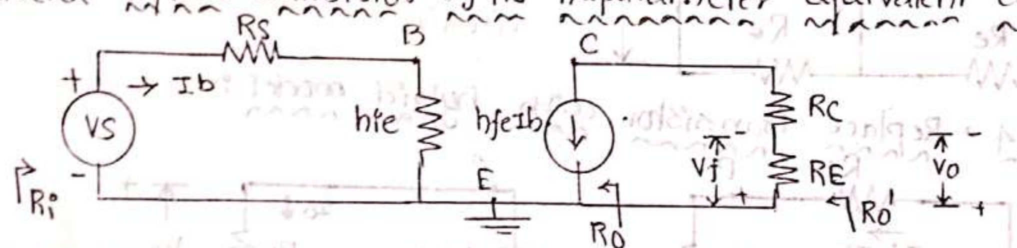
Step 2 and 3 : Find Input and output circuit :



To find Input circuit, set $V_o = 0$, and hence Vs in series with Rs appears between B and E.

To find output circuit, set $I_i = 0$ or $I_b = 0$ and hence RE appears only in the output loop.

Step 4 : Replace transistor by its h-parameter equivalent circuit :



Step 5 : Find open loop gain :

$\rightarrow A_V = \frac{V_o}{V_s} = \frac{h_{fe} I_b R_e}{V_s}$

Apply KVL to I/p loop,

$\rightarrow V_s = I_b (R_S + h_{ie})$

Sub Vs in Av

$\rightarrow A_V = \frac{V_o}{V_s} = \frac{h_{fe} I_b R_e}{I_b (R_S + h_{ie})}$

$\rightarrow A_V = \frac{h_{fe} R_e}{R_S + h_{ie}}$

Step 6 : Indicate Vo & Vf and calculate β :

$\rightarrow \beta = \frac{V_f}{V_o} = 1$ $\because$ Both Voltage present across Re.

Step 7 :

$\rightarrow D = 1 + \beta A_V$

$\rightarrow A_{Vf} = \frac{A_V}{1 + \beta A_V} = \frac{A_V}{D}$

$\rightarrow R_i = R_S + h_{ie}$

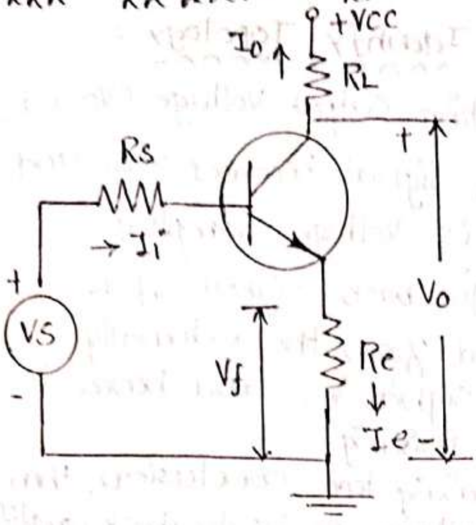
$\rightarrow R_{if} = R_i D$

$\rightarrow R_o = \infty = R_{of}$

$\rightarrow R_{of}' = \frac{R_o'}{D} = \frac{R_e}{D}$ $\because R_o' = R_e$

R. Current series feedback:

Common Emitter configuration with unbypassed R_e:



Step 1: Identify topology.

For Voltage sampling, $V_o = 0$ and $V_f = 0$

But, if $V_o = 0$, it is short circuited so I_c current flows.

$\because I_c = I_e$, I_e also flows

So it generates feedback voltage.

$V_f \neq 0$

$\therefore$ It is not Voltage sampling

For current sampling,

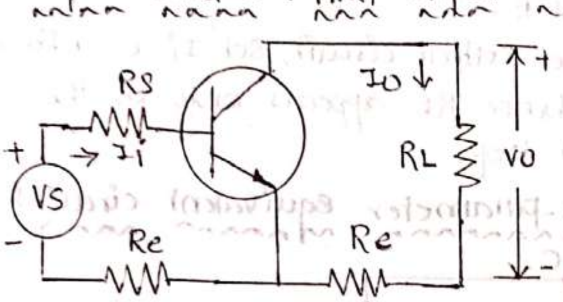
$I_o = 0$, $I_f = 0$

For $I_o = 0$, collector circuit is open circuited.

$I_c = 0$, $I_E = 0$ and $I_f = 0$

The condition is satisfied

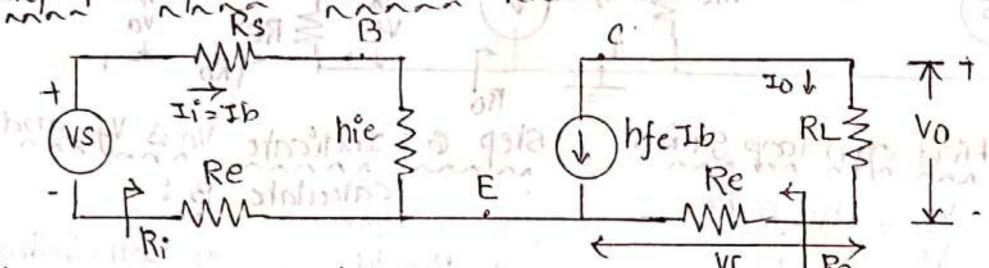
Step 2 and 3: Find input and output circuit



To find input circuit set $I_o = 0$ then R_e appears at the input side.

To find output circuit set $I_i = 0$, then R_e appears in the output circuit.

Step 4: Replace transistor with hybrid model:



Step 5: Find open loop Transfer Gain:

$\rightarrow G_m = \frac{I_o}{V_i} = \frac{-h_{fe} I_b}{V_S} = \frac{-h_{fe} I_b}{I_b(R_S + h_{ie} + R_e)}$

$\rightarrow G_m = \frac{-h_{fe}}{R_S + h_{ie} + R_e}$

Step 6: Calculate β:

$\rightarrow \beta = \frac{V_f}{I_o} = \frac{I_e R_e}{I_o}$

$\rightarrow \beta = \frac{-I_o R_e}{I_o} = -R_e$

$\because I_e = -I_o$

Step 7:

$\rightarrow D = 1 + \beta G_M$

$\rightarrow G_{Mf} = \frac{G_M}{D}$

$\rightarrow R_i = R_s + h_{ie} + R_e$

$\rightarrow R_{if} = R_i D$

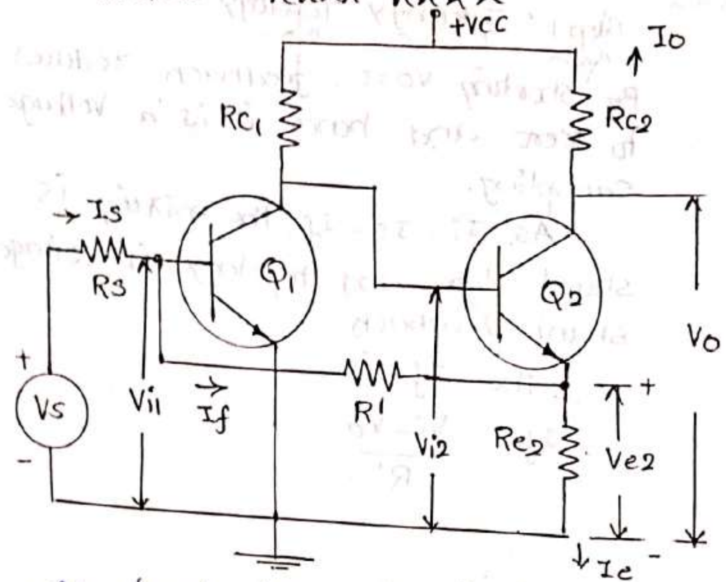
$\rightarrow A_{vf} = \frac{V_o}{V_s} = \frac{I_o R_L}{V_s}$ $\quad \because G_{Mf} = \frac{I_o}{V_s}$

$\rightarrow A_{vf} = G_{Mf} R_L$

$\rightarrow R_o = \infty$ & $R_{of} = R_{oD} = \infty$

$\rightarrow R'_{of} = R_{of} \parallel R_L = R_L$

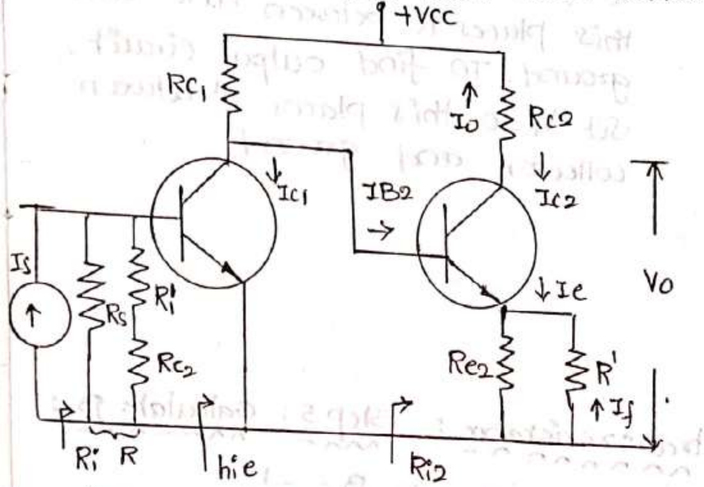
3. Current Shunt Feedback :



Step 1 : Identify Topology

By shorting output voltage ($V_o = 0$), feedback signal does not become zero and hence it is not voltage sampling. By opening the output loop ($I_o = 0$), feedback signal becomes zero and hence it is a current feedback. The feedback signal appears in shunt with input, hence the topology is current shunt feedback amplifier.

Step 2 and 3 Find Input and output circuit :



The input circuit without feedback is obtained by opening the output loop of the emitter of Q2 ($I_o = 0$). This place R' in series with Re from base to emitter of Q_1. The output circuit is found by shorting the input node (base of Q_1) i.e. making $V_i = 0$. This places R' in parallel with Re.

Step 4 : Find open loop Transfer gain :

$\Rightarrow A_I = \dfrac{-I_{C2}}{I_S} = \dfrac{-I_{C2}}{I_{b2}} \cdot \dfrac{I_{b2}}{I_{C1}} \cdot \dfrac{I_{C1}}{I_{b1}} \cdot \dfrac{I_{b1}}{I_S}$

wkt, $\dfrac{-I_{C2}}{I_{b2}} = -h_{fe}$, $\dfrac{+I_{C1}}{I_{b1}} = +h_{fe}$

$\Rightarrow \dfrac{I_{b2}}{I_{C1}} = \dfrac{-R_{C1}}{R_{C1}+R_{i2}}$ $\because R_{i2} = h_{ie} + (1+h_{fe})(R_{e2} \| R')$

$\Rightarrow \dfrac{I_{b1}}{I_S} = \dfrac{R}{R + h_{ie}}$ $\because R = R_s \| (R' + R_e)$

Step 5: Calculate β :

$\Rightarrow \beta = \dfrac{I_f}{I_o}$

$\because I_f = -I_e R_{e2} / R_e + R'$

$= \dfrac{-I_C R_{e2}}{R_e + R_1}$ $\because I_C \simeq I_E$

$= \dfrac{I_o R_{e2}}{R_{e2} + R'}$ $\because I_o = -I_C$

$\Rightarrow \beta = \dfrac{I_f}{I_o} = \dfrac{R_{e2}}{R_{e2} + R'}$

Step 6

$\Rightarrow D = 1 + \beta A_I$

$\Rightarrow A_{If} = \dfrac{A_I}{D}$

$\Rightarrow A_{vf} = \dfrac{V_o}{V_S} = \dfrac{-I_{C2} R_{C2}}{I_S \, R_S}$

$\Rightarrow A_{vf} = \dfrac{A_{If} \cdot R_{C2}}{R_S}$ $\because \dfrac{I_{C2}}{I_S} = A_{If}$

$\Rightarrow R_i = R \| h_{ie} =$

$\Rightarrow R_{if} = \dfrac{R_i}{D}$

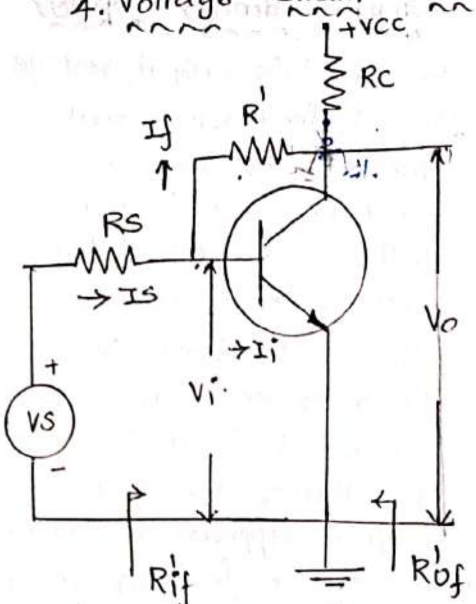
$\Rightarrow R_o = \infty$

$\Rightarrow R_{of} = R_{oD} = \infty$

$\Rightarrow R_o' = R_o \| R_{C2} = \infty \| R_{C2}$

$\Rightarrow R_{of}' = R_o' \dfrac{1+\beta A_I}{1+\beta A_I} = R_o' = R_{C2}$

4. Voltage shunt Feedback :



Step 1 : Identify Topology :

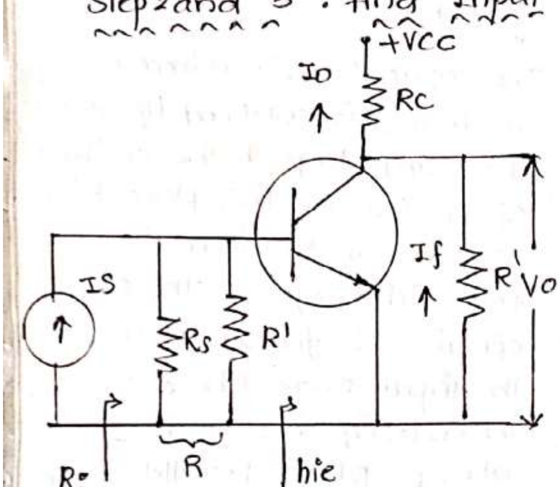
By shorting $V_o = 0$, feedback reduces to zero and hence it is a voltage sampling.

As $I_i = I_s - I_f$, the mixing is shunt type and topology is voltage shunt feedback.

The I_f is

$$I_f = \frac{V_i - V_o}{R'}$$

Step 2 and 3 : find Input and output circuit :



To find input circuit, set $V_o = 0$, this places R' between base and ground. To find output circuit, set $V_i = 0$, this places R' between collector and ground.

Step 4 : Find open circuit transresistance :

$$\rightarrow R_M = \frac{V_o}{I_s} = \frac{I_o R'_c}{I_s} = \frac{-I_c R'_c}{I_s}$$

$$\because R'_c = R_c \| R', \quad \frac{-I_c}{I_s} = \frac{-I_c}{I_b} \cdot \frac{I_b}{I_s}$$

$$\because \frac{-I_c}{I_b} = A_i = -h_{fe}, \quad \frac{I_b}{I_s} = \frac{R}{R + h_{ie}}$$

$$\because R = R_s \| R'$$

Step 5 : Calculate β :

$$\Rightarrow \beta = \frac{-1}{R'}$$

Step 6 :

$$\rightarrow D = 1 + \beta R_M$$

$$\rightarrow R_{Mf} = \frac{R_M}{D}$$

$$\rightarrow A_{Vf} = \frac{V_o}{V_s} = \frac{V_o}{I_s R_s} = \frac{R_{Mf}}{R_s}$$

$$\rightarrow R_i = R \| h_{ie} = \frac{R h_{ie}}{R + h_{ie}}$$

$$\rightarrow R_{if} = \frac{R_i}{D}$$

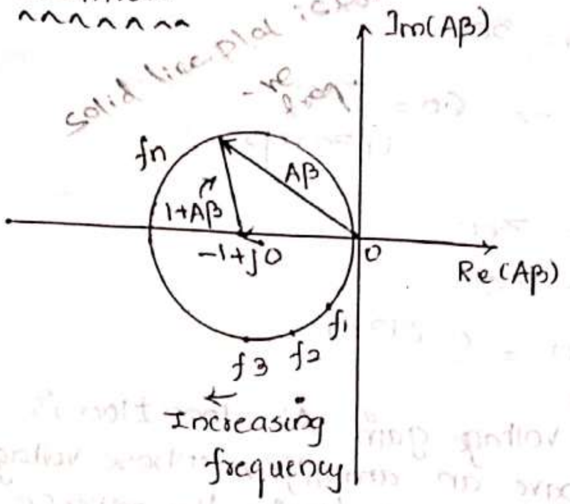
$$\rightarrow R_o = \infty$$

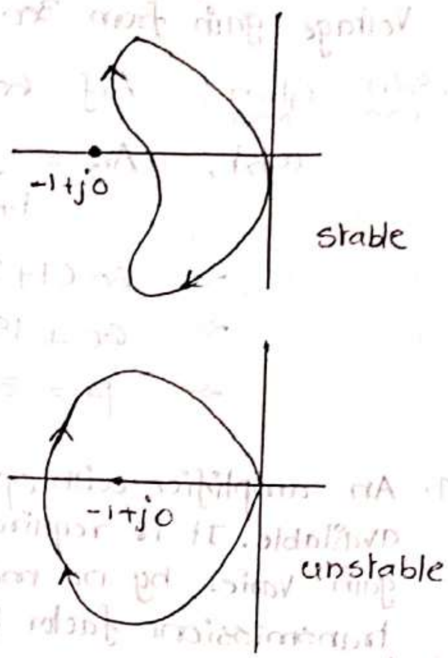
$$\rightarrow R_{of} = \frac{\infty}{D} = \infty$$

$$\rightarrow R'_o = R_o \| R'_c$$

$$\rightarrow R'_{of} = \frac{R'_o}{D}$$

Nyquist criterion for stability of feedback Amplifier :
(or)
condition





→ $A_{vf} = \frac{A_v}{1+A_v\beta}$ = $\frac{\text{poles}}{\text{zeros}}$.

→ $|1+A\beta| = 1$. - Polar form

→ $-1+j0$ = Rect. form.

$A\beta$ - radial distance
ϕ - Phase angle.

Problems :

1. A feedback Amplifier has an open loop gain of 600 and feedback factor $\beta = 0.01$. Find the closed loop gain with negative feedback.

Soln Given $A_v = 600$, $\beta = 0.01$.

$$\rightarrow A_{vf} = \frac{A_v}{1+\beta A_v} = \frac{600}{1+0.01\times600} = 85.714$$

2. The distortion in an amplifier is found to be 3%, when the feedback ratio of negative feedback amplifier is 0.04. When the feedback is removed, the distortion becomes 15%. Find the open loop and closed loop gain

Soln Given Distortion with feedback = 3%,
Distortion without feedback = 15%.
$\beta = 0.04$.

→ D = Distortion = Desensitivity = $\frac{15}{3} = 5$

→ $D = 1+A\beta = 5$

$5 = 1 + A_v \times 0.04$

→ $A_v = 100$, = open loop gain

→ $A_{vf} = \frac{A_v}{D} = \frac{100}{5} = 20$ = closed loop gain.

3. Addition of negative feedbacks to an amplifier reduces its Voltage gain from 300 to 60. Determine the feedback factor.

Soln Given : $A_{vf} = 60$, $A_v = 300$

wkt, $A_{vf} = \frac{A_v}{1+A_v\beta}$ $\rightarrow 60 = \frac{300}{1+300\times\beta}$

$\rightarrow 60(1+300\beta) = 300$

$\rightarrow 60 + 18000\beta = 300$

$\rightarrow \beta = \frac{300-60}{18000} = 0.01333.$

4. An amplifier with open loop voltage gain $A_v = 1000 \pm 100$ is available. It is required to have an amplifier whose voltage gain varies by no more than $\pm$ 0.1%. Find the reverse transmission factor β.

Soln Given.

wkt, $\frac{dA_f}{A_f} = \frac{1}{1+\beta A} \frac{dA}{A}$

$\rightarrow \frac{0.1}{100} = \frac{1}{1+\beta A} \times \frac{100}{1000}$ $\rightarrow 1+\beta A = \frac{10\cancel{0}}{1\cancel{0}\cancel{00}} \times \frac{1\cancel{00}}{0.1}$

$\rightarrow 1+\beta A = 100$

$\rightarrow \beta A = 99$

$\rightarrow \beta = \frac{99}{A} = \frac{99}{1000} = 0.099$

$\rightarrow \beta = 9.9\%$

5. An RC coupled amplifier has a voltage gain of 1000, f_1 = 50 Hz, f_2 = 200 kHz and a distortion of 5% with feedback. Find A_{vf}, BW_f, when $\beta = 0.01$ is applied.

Soln Given $A_v = 1000$, $f_1 = 50Hz$, $f_2 = 200kHz$, $\beta = 0.01$

Distortion with feedback = 5%.

1. $A_{vf} = \frac{A_v}{1+\beta A_v} = \frac{1000}{1+0.01\times1000} = 90.9$

2. $f_{1f} = \frac{f_1}{1+\beta A_v} = \frac{50}{1+0.01\times1000} = 4.545Hz$

3. $f_{2f} = f_2(1+\beta A_v) = 200k(1+0.01\times1000) = 2.2 MHz$

4. $BW_f = f_{2f} - f_{1f} = 2.2M - 4.545Hz = 2.2MHz.$

6. The gain and distortion of an amplifier are 100 and 4% respectively. If a negative feedback with $\beta = 0.3$ is applied, find the new distortion in the system.

Soln Given $D = 4\%$, $A_V = 100$, $\beta = 0.3$

$\rightarrow$ $D = 1 + \beta A_V = 1 + 0.3 \times 100 = 31$

$\rightarrow$ New Distortion $= \frac{4\%}{31} = 0.129\%$

7. A voltage series feedback amplifier has a voltage gain with feedback as 83.33 and feedback ratio as 0.01. Calculate the voltage gain of the amplifier without feedback.

Soln Given $A_{vf} = 83.33$, $\beta = 0.01$

WKT, $A_{vf} = \frac{A_v}{1 + A_v\beta} \Rightarrow 83.33 = \frac{A_v}{1 + A_v(0.01)}$

$\rightarrow$ $83.33 + 0.8333 A_v = A_v$

$\rightarrow$ $A_v = 500$.

8. An amplifier has a voltage gain of 1000, with negative feedback. The voltage gain reduces to 10. Calculate the fraction of the output, that is feedback to the input.

Soln Given $A_v = 1000$, $A_{vf} = 10$.

$\rightarrow A_{vf} = \frac{A_v}{1 + A_v\beta} \Rightarrow 10 = \frac{1000}{1 + 1000 \times \beta}$

$\rightarrow$ $10(1 + 1000\beta) = 1000$

$\rightarrow$ $10 + 10000\beta = 1000$

$\rightarrow$ $10000\beta = 1000 - 10$

$\rightarrow$ $\beta = \frac{1000 - 10}{10000} = 0.099$.

$\rightarrow$ $\beta = 0.099$

9. An amplifier has a midband gain of 125 and a bandwidth 250 KHz.

1. If 4% negative feedback is introduced, find the new bandwidth and gain.
2. If the bandwidth is to be restricted to 1 MHz find the feedback ratio.

Soln Given $A_v = 125$, $BW = 250\,KHz$, $\beta = 0.04$.

1. Gain, $A_{vf} = \frac{A_v}{1 + \beta A_v} = \frac{125}{1 + 0.04 \times 125} = 20.83$.

$\rightarrow$ $BW_f = BW \times (1 + \beta A_v) = 250K\,(1 + 0.04 \times 125) = 1.5\,MHz$.

2. $BW_f = 1\,MHz$

$\rightarrow$ $1\,MHz = BW \times (1 + \beta A_v)$

$\rightarrow$ $1\,MHz = 250K\,(1 + \beta \times 125)$

$\rightarrow$ $B = 0.024$.

10. An amplifier with negative feedback give an output of 12.5 V with an input of 1.5 V. When feedback is removed, it requires 0.25 V input for the same output. Find

1. The value of voltage gain without feedback.
2. Value of feedback β, if the input and output are in phase and β is real.

Soln Given $V_{of} = 12.5V$, $V_{inf} = 1.5V$, $V_{in} = 0.25V$

$$A_{vf} = \frac{V_{of}}{V_{inf}} = \frac{12.5}{1.5} = 8.333$$

1. $A_v = \frac{V_o}{V_{in}} = \frac{12.5}{0.25} = 50$

2. $A_{vf} = \frac{A_v}{1+A_v\beta} \rightarrow 8.333 = \frac{50}{1+50\beta} \rightarrow \beta = 0.1$

$\rightarrow \beta = 0.1$

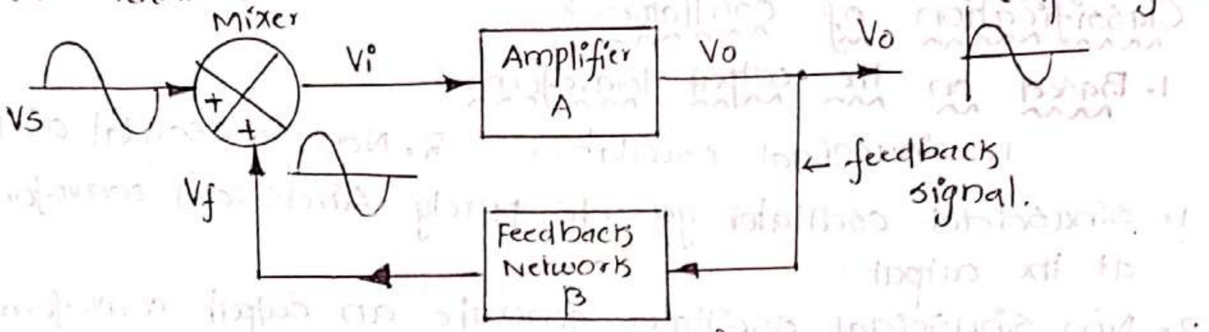
UNIT - II

OSCILLATORS

Oscillator :

An oscillator is an amplifier, which uses a positive feedback and without any external input signal generates an output waveform at a desired frequency.

Basic theory of oscillators :



Concept of positive feedback

Consider a non-inverting amplifier with the voltage gain A. From the dia, Vs is applied since the amplifier is non inverting, the output voltage is in phase with the input signal Vs. (+ve feedback).

open loop gain :

The amplifier gain is A ie) it amplifies its input V_i, A times to produce output V_o.

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

Closed loop gain (Gain with feedback) :

The ratio of output to input Vs, considering effect of feedback is called closed loop gain.

$$\Rightarrow A_f = \frac{V_o}{V_s}$$

The feedback is positive and voltage V_f is added to V_s to generate input of amplifier, V_i, and it is given by

$$\rightarrow V_i = V_s + V_f \rightarrow (1)$$

The feedback voltage V_f depends on the feedback element gain β, we can write,

$$\rightarrow V_f = \beta V_o \rightarrow (2)$$

sub Eq (2) in (1)

$$\rightarrow V_i = V_s + \beta V_o$$

$$\rightarrow V_s = V_i - \beta V_o \rightarrow (3)$$

UNIT – II

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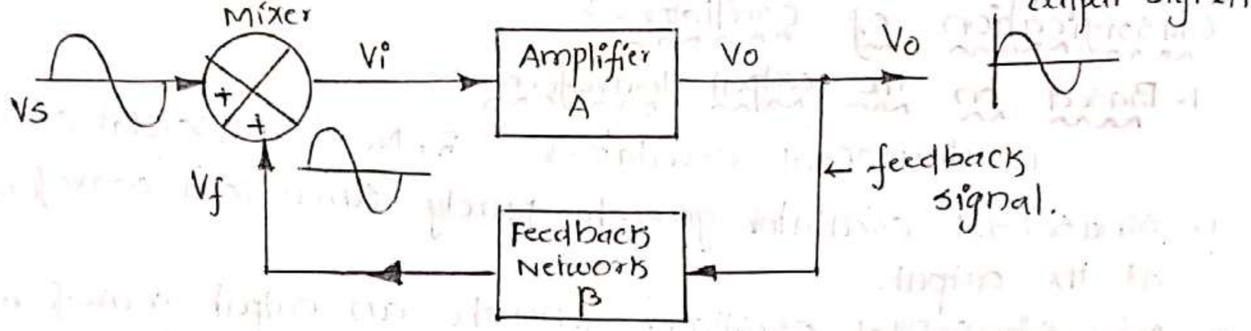

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sub Eq (2) in (1)

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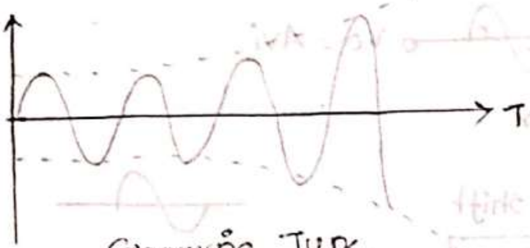
$$\rightarrow \quad V_s = V_i - \beta V_o \quad \rightarrow (3)$$

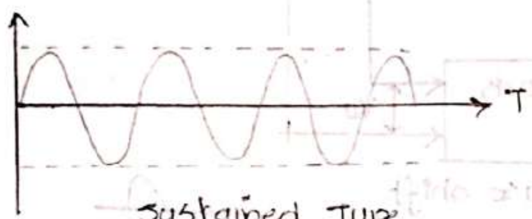
Case 1 : $|A\beta| > 1$



Growing Type.

When the total shift around a loop is 0° or 360°, then the output oscillates but the oscillations are of Growing type. The amplitude of oscillations goes on Increasing.

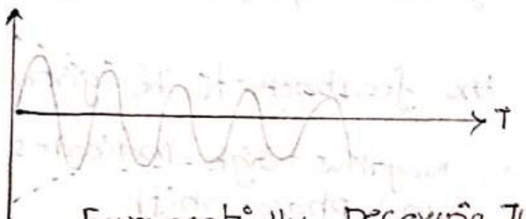
Case 2 : $|A\beta| = 1$



Sustained Type

When total phase shift around a loop is 0° or 360° ensuring positive feedback and $|A\beta| = 1$ then the oscillations are with constant frequency and amplitude called sustained oscillations.

Case 3 : $|A\beta| < 1$



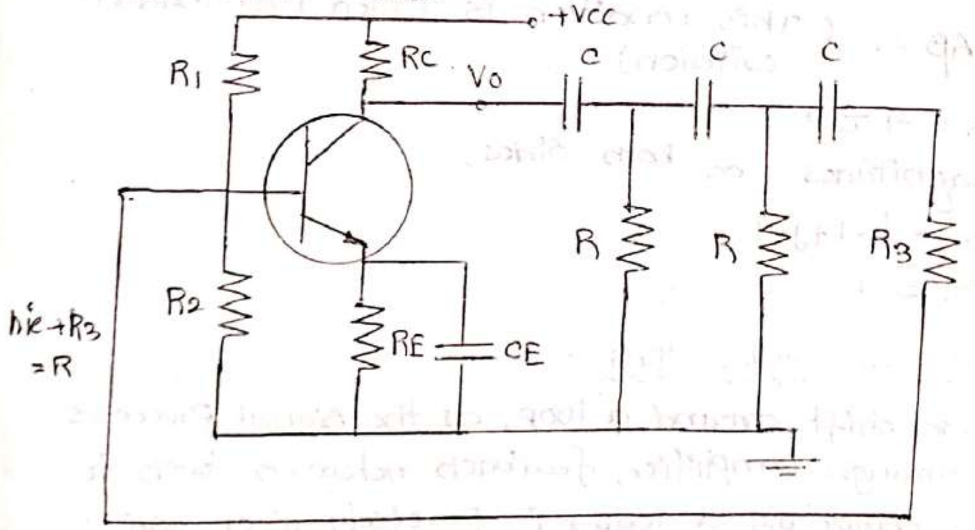
Exponentially Decaying Type

When total phase shift around a loop is 0° or 360° but $|A\beta| < 1$ then the oscillations are of decaying type ie such oscillation amplitude decreases exponentially.

General Form of an oscillators:

Rc Oscillators :

Rc phase shift oscillator :



Transistorised RC Phase shift oscillator.

The Rc Phase shift which uses a CE amplifier and a phase shifting networks consisting of three Rc sections. All the resistances and Capacitances Values are Same, So that for a particular frequency, each section of R and c produces a phase shift of 60°. This networks is called Ladder network.

The feedback network, output of the amplifier is given as an input to the feedback network. While the output of the feedback network is given as an input to the amplifier. Thus amplifier supplies its own input, through the feedback network.

Neglecting R_1 & R_2, these are sufficiently large, we get

$\rightarrow \quad h_{ie} + R_3 = R \rightarrow (1) \qquad \because R_i' = h_{ie}$

and If R_1 & R_2 are not neglected

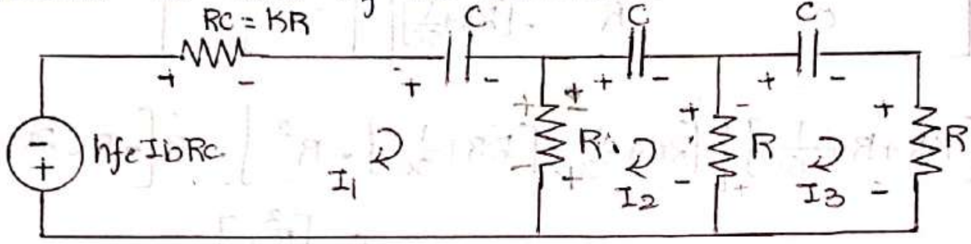
$R_i' = R_1 || R_2 || h_{ie} \rightarrow (2)$

In such a case, the value of R_3 is

$\rightarrow \quad R_i' + R_3 = R \rightarrow (3)$

Equivalent circuit using h-parameter model :

Assume the ratio of the resistance R_C to R be K ie) $K = \frac{R_C}{R}$



Apply KVL,

For Loop 1 :

$\rightarrow -h_{fe} I_b R_C - R_C I_1 - \frac{1}{j\omega C} I_1 - (I_1 - I_2) R = 0$

$\rightarrow -h_{fe} I_b R_C - R_C I_1 - \frac{1}{j\omega C} I_1 - I_1 R + I_2 R = 0$

Replace R_C by KR, and $j\omega$ by s,

$\rightarrow -I_1 \{ KR + \frac{1}{sC} + R \} + I_2 R = h_{fe} I_b K R \rightarrow (4)$

For Loop 2 :

$\rightarrow -(I_2 - I_1) R - I_2 \frac{1}{j\omega C} - (I_2 - I_3) R = 0$

$\rightarrow (I_2 - I_1) R + \frac{I_2}{j\omega C} + (I_2 - I_3) R = 0$

Replace R_C by KR, $j\omega$ by s,

$\rightarrow (I_2 - I_1) R + \frac{I_2}{sC} + (I_2 - I_3) R = 0$

$\rightarrow I_2 R - I_1 R + \frac{I_2}{sC} + I_2 R - I_3 R = 0$

$\rightarrow -I_1 R - I_3 R + I_2 (2R + \frac{1}{sC}) = 0$

$\rightarrow I_1 R - I_2 (2R + \frac{1}{sC}) + I_3 R = 0 \rightarrow (5)$

For Loop 3 :

$$\rightarrow -(I_3 - I_2)R - \frac{I_3}{j\omega C} - I_3 R = 0$$

$$\rightarrow -I_3R + I_2R - \frac{I_3}{j\omega C} - I_3R = 0$$

Replace $s = j\omega$, $RC = KR$

$$\rightarrow -I_3R + I_2R - \frac{I_3}{sC} - I_3R = 0$$

$$\rightarrow I_2R - I_3\left(2R + \frac{1}{sC}\right) = 0 \rightarrow (6)$$

Using Cramer's rule to solve for I_3,

$$\rightarrow D = \begin{bmatrix} -\left[KR + R + \frac{1}{sC}\right] & R & 0 \\ R & -\left[2R + \frac{1}{sC}\right] & R \\ 0 & R & -\left[2R + \frac{1}{sC}\right] \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} = \begin{bmatrix} h_{fe} I_b KR \\ 0 \\ 0 \end{bmatrix}$$

$$\rightarrow D = -\left[KR + R + \frac{1}{sC}\right]\left[\left[2R + \frac{1}{sC}\right]\left[2R + \frac{1}{sC}\right] - R^2\right] - R\left[-R\left(2R + \frac{1}{sC}\right) - 0\right] - 0\left[R^2 - 0\right]$$

$$\rightarrow D = -\left[KR + R + \frac{1}{sC}\right]\left[4R^2 + \frac{2R}{sC} + \frac{2R}{sC} + \frac{1}{s^2c^2} - R^2\right] - R\left[-2R^2 - \frac{R}{sC}\right]$$

$$\rightarrow D = -\left[\frac{KRsC + RsC + 1}{sC}\right]\left[\frac{4R^2s^2c^2 + 2RsC + 2RsC + 1 - R^2s^2c^2}{s^2c^2}\right] - R\left[\frac{-2R^2sC - R}{sC}\right]$$

$$\rightarrow D = -\left[\frac{KRsC + RsC + 1}{sC}\right]\left[\frac{3R^2s^2c^2 + 4RsC + 1}{s^2c^2}\right] + \frac{2R^3sC}{sC} + \frac{R^2}{sC}$$

$$\rightarrow D = -\left[\frac{3KR^3s^3c^3 + 4KR^2s^2c^2 + KRsC}{s^3c^3} + \frac{3R^3s^3c^3 + 4R^2s^2c^2 + RsC}{s^3c^3} + \frac{3R^2c^2s^2 + 4RsC + 1}{s^3c^3}\right] + 2R^3 + \frac{R^2}{sC}$$

$$\rightarrow D = -\left[3KR^3 + \frac{4KR^2}{sC} + \frac{KR}{s^2c^2} + 3R^3 + \frac{4R^2}{sC} + \frac{R}{s^2c^2} + \frac{3R^2}{sC} + \frac{4R}{s^2c^2} + \frac{1}{s^3c^3}\right] + 2R^3 + \frac{R^2}{sC}$$

$$\rightarrow D = -\left[3KR^3 + \frac{4KR^2}{sC} + \frac{KR}{s^2c^2} + 3R^3 + \frac{7R^2}{sC} + \frac{5R}{s^2c^2} + \frac{1}{s^3c^3}\right] + 2R^3 + \frac{R^2}{sC}$$

$$\rightarrow D = -3KR^3 - \frac{4KR^2}{sC} - \frac{KR}{s^2c^2} - 3R^3 - \frac{7R^2}{sC} - \frac{5R}{s^2c^2} - \frac{1}{s^3c^3} + 2R^3 + \frac{R^2}{sC}$$

(X) s^3c^3 N & D,

$$\rightarrow D = -\frac{3KR^3s^3c^3}{s^3c^3} - \frac{4KR^2s^2c^2}{s^3c^3} - \frac{KRsc}{s^3c^3} - \frac{3R^3s^3c^3}{s^3c^3} - \frac{7R^2s^2c^2}{s^3c^3} - \frac{5Rsc}{s^3c^3}$$
$$(1) \quad (2) \quad (3) \quad (1) \quad (2) \quad (3)$$
$$-\frac{1}{s^3c^3} + \frac{2R^3s^3c^3}{s^3c^3} + \frac{R^2s^2c^2}{s^3c^3}$$
$$(1) \quad (2)$$

$$\rightarrow D = -\frac{R^3s^3c^3}{s^3c^3}\left[3K+1\right] - \frac{R^2s^2c^2}{s^3c^3}\left[4K+6\right] - \frac{Rsc}{s^3c^3}\left[K+5\right] - \frac{1}{s^3c^3} = 0$$

$$\rightarrow D = \frac{R^3s^3c^3\,[3K+1] + R^2s^2c^2\,[4K+6] + Rsc\,[K+5] + 1}{s^3c^3} \rightarrow (7)$$

Now,

$$\rightarrow D_3 = \begin{bmatrix} -\left[KR+R+\frac{1}{sc}\right] & R & h_{fe}I_bKR \\ R & -\left[2R+\frac{1}{sc}\right] & 0 \\ 0 & R & 0 \end{bmatrix}$$

$$\rightarrow D_3 = -\left[KR+R+\frac{1}{sc}\right]\left[0-0\right] - R\left[0-0\right] + h_{fe}I_bKR\left[R^2-0\right]$$

$$\rightarrow D_3 = +h_{fe}I_bKR^3 \rightarrow (8)$$

$$\rightarrow I_3 = \frac{D_3}{D} = \frac{+h_{fe}I_bKR^3s^3c^3}{R^3s^3c^3[3K+1] + R^2s^2c^2[4K+6] + Rsc\,[K+5]+1}$$

Where I_3 = output current , I_b = Input current

$I_C = h_{fe}I_b$ = Input current of feedback.

$$\beta = \frac{\text{output of Feedback circuit}}{\text{Input to Feedback circuit}} = \frac{I_3}{h_{fe}I_b}$$

$$A = \frac{\text{output of Amplifier circuit}}{\text{Input to Amplifier circuit}} = \frac{I_3}{I_b} = h_{fe}$$

Now $\quad A\beta = h_{fe} \times \frac{I_3}{h_{fe}I_b} = \frac{I_3}{I_b} \rightarrow (9)$

$$\rightarrow A\beta = \frac{I_3}{I_b} = \frac{+h_{fe}KR^3s^3c^3}{R^3s^3c^3\,[3K+1] + R^2s^2c^2[4K+6] + Rsc\,[K+5]+1} \rightarrow (10)$$

sub $s = j\omega$, $s^2 = j^2\omega^2 = -\omega^2$, $s^3 = j^3\omega^3 = -j\omega^3$

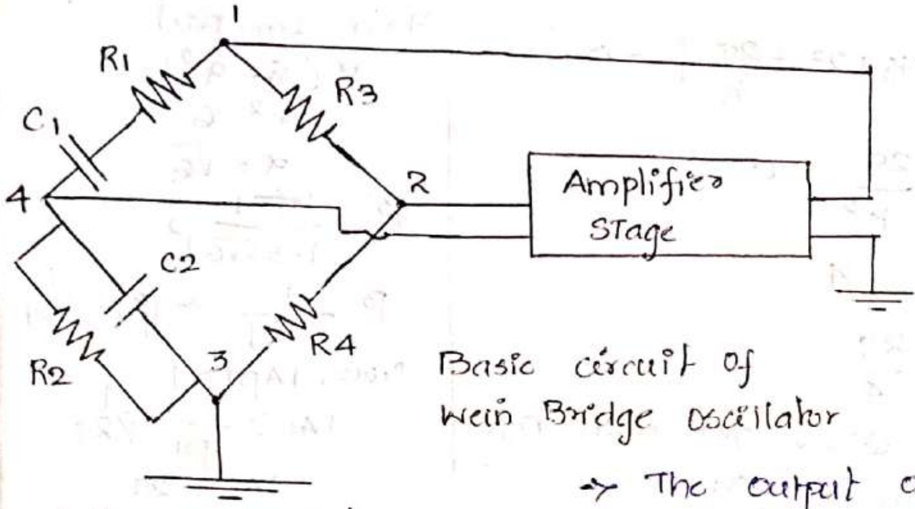
$$\rightarrow A\beta = \frac{-h_{fe}Kj\omega^3R^3c^3}{-j\omega^3R^3c^3[3K+1] - \omega^2R^2c^2[4K+6] + j\omega Rc\,(K+5]+1}$$

Separate real and Imaginary part in Denominator,

$$\rightarrow A\beta = \frac{-j\omega^3R^3c^3Kh_{fe}}{[1-4K\omega^2R^2c^2 - 6\omega^2R^2c^2] - j\omega[3K\omega^2R^3c^3 + \omega^2R^3c^3 - KRc - 5Rc]}$$

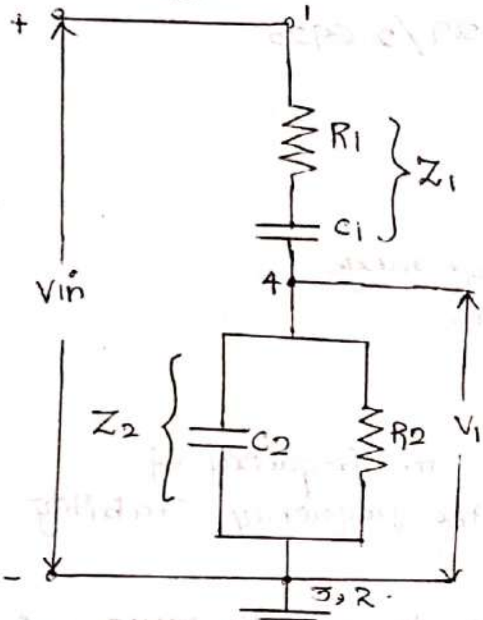
wkt → $R_3 = R - R_1' = 12 - 2 = 10\,k\Omega$.

Wein Bridge Oscillator :



Basic circuit of Wein Bridge oscillator

Wein Bridge Oscillator uses a non-inverting amplifier and hence does not provide any phase shift during amplifier stage. so no phase shift is necessary through feedback.



Feedback Network

- → The output of the amplifier is applied between 1 and 3.
- → The amplifier input is supplied from the dialog terminals 2 & 4.
- → Thus amplifier supplies its own input through the wein bridge as a feedback network
- → R_1, C_1 in series, R_2, C_2 in parallel } Frequency sensitive arms.
- → Vin to the feedback network is between is 1 and 3 while output V_f of the feedback network is between 2 and 4. Such a feedback network is called lead-lag network.

From the feedback network,

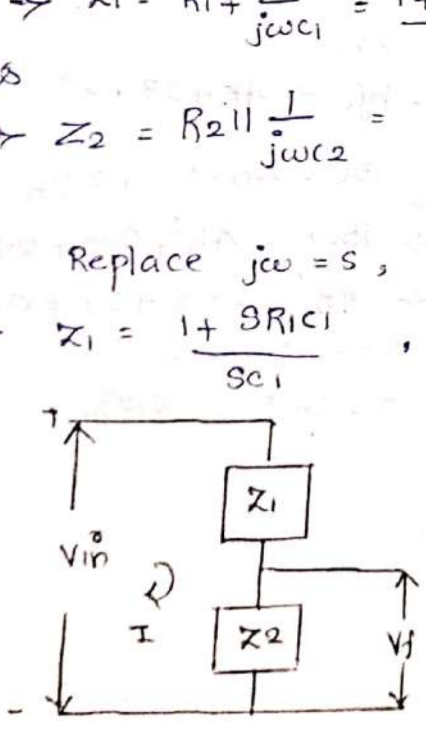
$$\rightarrow Z_1 = R_1 + \frac{1}{j\omega C_1} = \frac{1 + j\omega R_1 C_1}{j\omega C_1} \rightarrow (1)$$

&

$$\rightarrow Z_2 = R_2 \parallel \frac{1}{j\omega C_2} = \frac{R_2 \times \frac{1}{j\omega C_2}}{R_2 + \frac{1}{j\omega C_2}} = \frac{R_2}{R_2 j\omega C_2 + 1} \rightarrow (2)$$

Replace $j\omega = s$,

$$\rightarrow Z_1 = \frac{1 + sR_1C_1}{sC_1}, \quad Z_2 = \frac{R_2}{1 + sR_2C_2}$$



From the simplified ckt,

$$\rightarrow I = \frac{V_{in}}{Z_1 + Z_2}$$

&

$$\rightarrow V_f = I Z_2 = \frac{V_{in}}{Z_1 + Z_2} \cdot Z_2$$

WKT, $\beta = \frac{V_f}{V_{in}} = \frac{Z_2}{Z_1+Z_2} \rightarrow (3)$

Sub Z_1, Z_2 in (3)

$$\rightarrow \beta = \frac{R_2/1+SR_2C_2}{\left[\frac{1+SR_1C_1}{SC_1}\right]+\left[\frac{R_2}{1+SR_2C_2}\right]} = \frac{\frac{R_2}{1+S\cancel{R_2}C_2}}{\frac{(1+SR_1C_1)(1+SR_2C_2)+R_2SC_1}{SC_1(1+S\cancel{R_2}C_2)}}$$

$$\rightarrow \beta = \frac{R_2SC_1}{(1+SR_1C_1)(1+SR_2C_2)+R_2SC_1} = \frac{SR_2C_1}{1+SR_2C_2+S^2R_1R_2C_1C_2+SR_1C_1+SR_2C_1}$$

$$\rightarrow \beta = \frac{SR_2C_1}{1+S(R_1C_1+R_2C_2+R_2C_1)+S^2R_1R_2C_1C_2}$$

Replace $S=j\omega$, $S^2=j^2\omega^2=-\omega^2$, $S^3=j^3\omega^3=-j\omega^3$.

$$\rightarrow \beta = \frac{j\omega R_2C_1}{1+j\omega(R_1C_1+R_2C_2+R_2C_1)-\omega^2R_1R_2C_1C_2}$$

$$\rightarrow \beta = \frac{j\omega R_2C_1}{(1-\omega^2R_1R_2C_1C_2)+j\omega(R_1C_1+R_2C_2+R_2C_1)} \rightarrow (4)$$

Rationalising the Expression,

$$\rightarrow \beta = \frac{j\omega R_2C_1\left((1-\omega^2R_1R_2C_1C_2)-j\omega(R_1C_1+R_2C_2+R_2C_1)\right)}{\left((1-\omega^2R_1R_2C_1C_2)^2+\omega^2(R_1C_1+R_2C_2+R_2C_1)^2\right)}$$

$$\rightarrow \beta = \frac{j\omega R_2C_1-j\omega^3R_1R_2^2C_1^2C_2+\omega^2R_1R_2C_1^2+\omega^2R_2^2C_1C_2+\omega^2R_2^2C_1^2}{(1-\omega^2R_1R_2C_1C_2)^2+\omega^2(R_1C_1+R_2C_2+R_2C_1)^2}$$

$$\rightarrow \beta = \frac{\omega^2C_1R_2(C_1R_1+C_2R_2+C_1R_2)+j\omega R_2C_1(1-\omega^2R_1R_2C_1C_2)}{(1-\omega^2R_1R_2C_1C_2)^2+\omega^2(R_1C_1+R_2C_2+R_2C_1)^2} \rightarrow (5)$$

To have zero phase shift of feedback network, its Im.part is zero

$\rightarrow \omega R_2C_1(1-\omega^2R_1R_2C_1C_2)=0$

$\rightarrow 1-\omega^2R_1R_2C_1C_2=0$

$\rightarrow -\omega^2R_1R_2C_1C_2=-1$

$\rightarrow \omega^2=\frac{1}{R_1R_2C_1C_2}$

$\rightarrow \omega=\frac{1}{\sqrt{R_1R_2C_1C_2}}$ $\quad \because \omega=2\pi f$

$\rightarrow 2\pi f=\frac{1}{\sqrt{R_1R_2C_1C_2}}$

$\rightarrow f=\frac{1}{2\pi\sqrt{R_1R_2C_1C_2}} \rightarrow (6)$ $\quad \because$ This is Frequency of oscillator.

In practical, $R_1=R_2=R$, $C_1=C_2=C$,

$\rightarrow f=1/2\pi\sqrt{R^2C^2} \rightarrow f=1/2\pi RC \rightarrow (7)$

At $R_1 = R_2 = R$, $C_1 = C_2 = C$, in (5). $\omega = \frac{1}{RC}$

$$\rightarrow \beta = \frac{\omega^2 RC\{RC + RC + RC\} + j\omega RC\left(1 - \frac{1}{R^2C^2}R^2C^2\right)}{\left(1 - \frac{1}{R^2C^2}R^2C^2\right)^2 + \frac{1}{R^2C^2}(9R^2C^2)}$$

$$\rightarrow \beta = \frac{\frac{1}{R^2C^2}RC(3RC) + 0}{0 + 9} = \frac{3}{9} = \frac{1}{3}$$

$$\rightarrow \beta = \frac{1}{3} \rightarrow (8)$$

The Barkhausen criterion for sustained oscillations, we get.

$\rightarrow |A\beta| \geq 1$

$\rightarrow |A| = \geq \frac{1}{|\beta|} \geq \frac{1}{1/3}$.

$\rightarrow |A| \geq 3$.

If $R_1 \neq R_2$, $C_1 \neq C_2$ then Eq (5) becomes, $\omega = \frac{1}{\sqrt{R_1R_2C_1C_2}}$

$$\rightarrow \beta = \frac{\frac{1}{R_1R_2C_1C_2}C_1R_2(C_1R_1 + R_2C_2 + C_1R_2) + j\frac{1}{\sqrt{R_1R_2C_1C_2}}C_1R_2\left(1 - \frac{1}{R_1R_2C_1C_2}R_1R_2C_1C_2\right)}{\left(1 - \frac{1}{R_1R_2C_1C_2}R_1R_2C_1C_2\right)^2 + \frac{1}{R_1R_2C_1C_2}(R_1C_1 + R_2C_2 + R_2C_1)^2}$$

$$\rightarrow \beta = \frac{C_1R_2}{R_1C_1 + R_2C_2 + C_1R_2} \rightarrow (9)$$

$\rightarrow |A\beta| \geq 1$

$\rightarrow |A| \geq 1/|\beta|$. $\Rightarrow |A| \geq \frac{1}{|\beta|} \rightarrow$

$$\rightarrow |A| \geq \frac{R_1C_1 + R_2C_2 + C_1R_2}{C_1R_2} \rightarrow (10)$$

Advantages:

By Varying the two Capacitor Values simultaneously, different frequency ranges can be provided.

1. A Wein bridge is used for operation at 10KHZ. If the Value of the resistance R is 100KΩ, what is the Value C required?

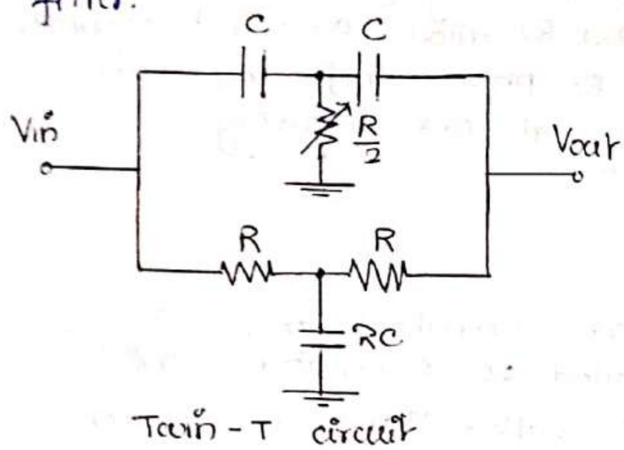
Soln Given $f = 10K$, $R = 100K$

$$\rightarrow f = \frac{1}{2\pi RC} \rightarrow 10\times10^3 = \frac{1}{2\pi\times100\times10^3\times C}$$

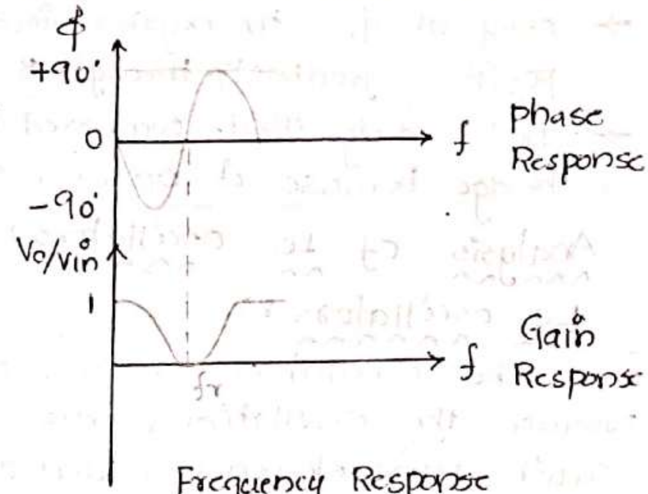
$$\rightarrow C = 159.155\ pF.$$

Twin-T Oscillator :

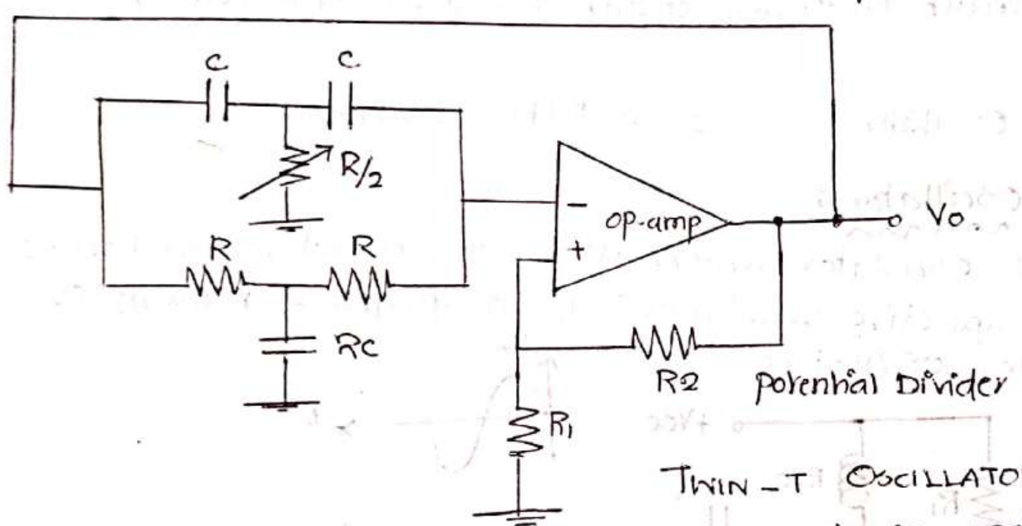
This is another type of RC oscillator, which uses a typical circuit consisting of R and C called Twin-T network (or) Twin-T filter.



Twin-T circuit



Frequency Response



TWIN-T OSCILLATOR

- It is basically lead-lag circuit whose phase angle varies between +90° and -90° against the frequency.
- At $f = f_r$, its phase angle is 0° and it does not introduce any phase shift.
- Gain is 1 at low and high frequencies and at $f = f_r$, the gain reduces to zero.
- The resonant frequency is given as
  - $f_r = \frac{1}{2\pi RC}$
- It is a combination of high pass and low pass filter. The combined parallel combination is called Twin-T filter.
- Twin-T circuit used to obtain Twin-T oscillator.
- +ve feedback given to non inverting terminal through R_1 & R_2
- -ve feedback given to inverting terminal through Twin-T network.
- When power is on, the R_2 is low, so the positive feedback is max. and the oscillation grow.
- As oscillation grow, the R_2 ↑ and ↓ the +ve feedback maintaining sustained oscillation.

→ Oscillation cannot occur in any frequency other than f_r.

→ Because at the frequency, other than f_r, there is a negative feedback, which is not allowed for oscillation

→ only at f_r, the negative feedback is negligible and hence positive feedback through R_1 and R_2 allows circuit to oscillate.

→ It is rarely used compared to RC phase shift and wein bridge because it operated only at one frequency.

Analysis of LC oscillators:

LC oscillators:

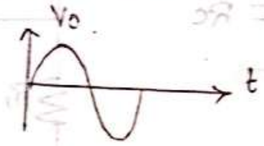
The oscillators which use the elements L and C to produce the oscillations are called LC oscillators. The circuit using elements L and C is called Tank circuit (or) oscillatory circuit (or) Tuned circuit (or) Resonating circuit.

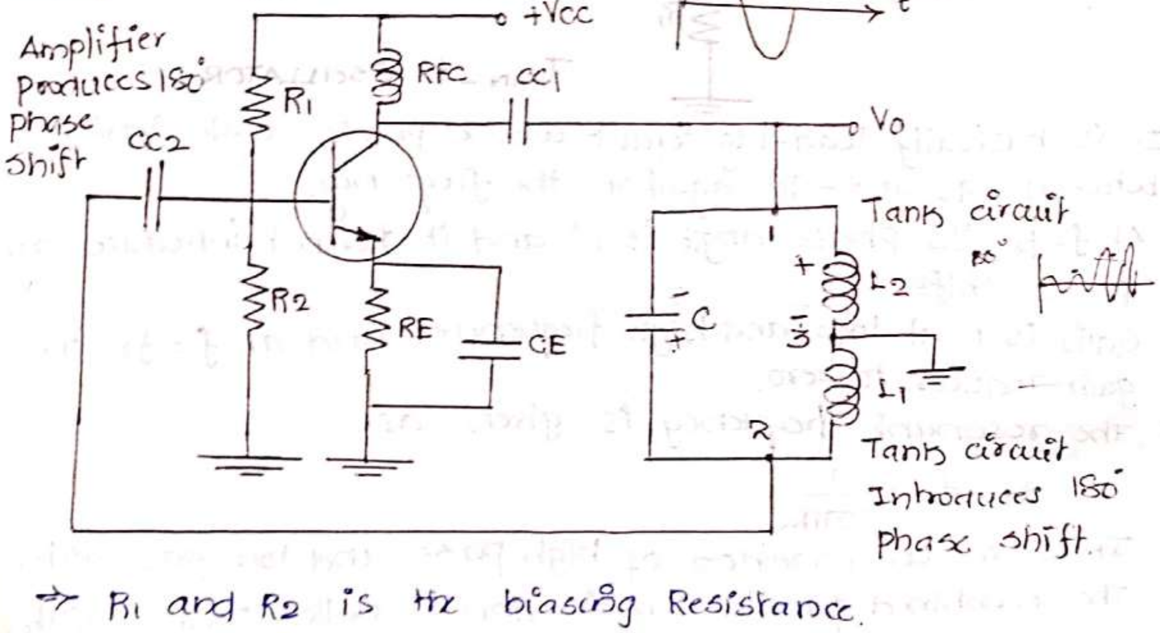
Types

1. Hartley Oscillator 2. Colpitts oscillator.

1. Hartley oscillator:

A LC oscillator which uses two Inductive reactances and one capacitive reactances in a feedback network is called Hartley oscillators.





→ R_1 and R_2 is the biasing Resistance.

→ RE is Biasing circuit resistance

→ CE is Emitter bypass capacitor.

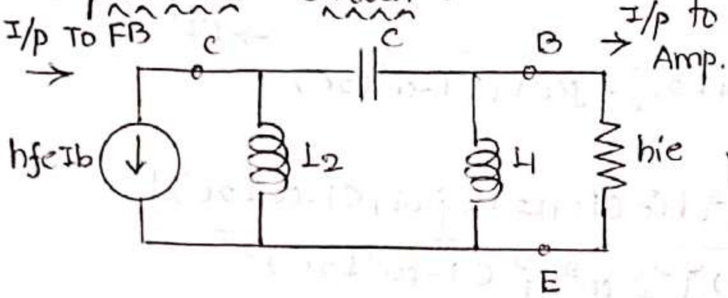
→ CC_1 and CC_2 are coupling capacitors.

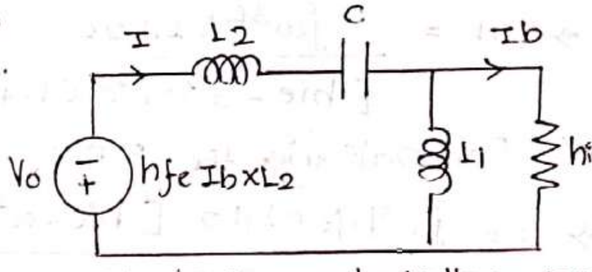
→ RFC is Radio frequency choke. Its reactance Value is Very high for high frequency and it act as a open circuit. For DC condition, the reactance is zero and hence causes no problem for dc capacitor.

→ Due to RFC, isolation between AC and DC operation is achieved.

→ The CE amplifier provides 180° phase shift and LC feedback network gives an additional phase shift of 180° to satisfy the oscillation conditions.

Equivalent circuit :





current source to voltage source

From the circuit,

$$\rightarrow V_o = h_{fe} I_b X_{L2} = h_{fe} I_b j\omega L_2 \rightarrow (1)$$

∴ -ve sign indicates current shown in opposite to Vo.

&

$$\rightarrow I = \frac{-V_o}{(X_{L2}+X_C)+(X_{L1} \parallel h_{ie})} \rightarrow (2)$$

$$\rightarrow I = \frac{-V_o}{\left[j\omega L_2 + \frac{1}{j\omega C}\right] + \frac{j\omega L_1 h_{ie}}{j\omega L_1 + h_{ie}}} = \frac{-h_{fe} I_b j\omega L_2}{\left[j\omega L_2 + \frac{1}{j\omega C}\right] + \frac{j\omega L_1 h_{ie}}{j\omega L_1 + h_{ie}}} \rightarrow (3)$$

Replace $j\omega$ by s.

$$\rightarrow I = \frac{-s h_{fe} I_b L_2}{\left[sL_2 + \frac{1}{sC}\right] + \left[\frac{sL_1 h_{ie}}{sL_1 + h_{ie}}\right]} = \frac{-s h_{fe} I_b L_2}{\left[\frac{1+s^2 L_2 C}{sC} + \frac{sL_1 h_{ie}}{sL_1 + h_{ie}}\right]}$$

$$\rightarrow I = \frac{-s h_{fe} I_b L_2 (sC)(sL_1 + h_{ie})}{[(1+s^2 L_2 C)(sL_1 + h_{ie}) + sC(sL_1 h_{ie})]}$$

$$\rightarrow I = \frac{-s^2 h_{fe} I_b L_2 C (sL_1 + h_{ie})}{sL_1 + h_{ie} + s^3 L_1 L_2 C + s^2 L_2 C h_{ie} + s^2 C L_1 h_{ie}}$$

$$\rightarrow I = \frac{-s^2 h_{fe} I_b L_2 C (sL_1 + h_{ie})}{sL_1 + h_{ie} + s^2 C h_{ie} (L_1 + L_2) + s^3 L_1 L_2 C} \rightarrow (4)$$

According to current division in parallel ckt,

$$\rightarrow I_b = I \times \frac{X_{L1}}{X_{L1} + h_{ie}} = I \times \frac{j\omega L_1}{j\omega L_1 + h_{ie}} = I \times \frac{sL_1}{sL_1 + h_{ie}} \rightarrow (5)$$

Sub (4) in (5)

$$\rightarrow I_b = \frac{-s^2 h_{fe} I_b L_2 C \cancel{(sL_1 + h_{ie})}}{sL_1 + h_{ie} + s^2 C h_{ie} (L_1 + L_2) + s^3 L_1 L_2 C} \times \frac{sL_1}{\cancel{sL_1 + h_{ie}}}$$

$$\rightarrow I_b = \frac{-s^3 h_{fe} I_b L_1 L_2 C}{sL_1 + h_{ie} + s^2 h_{ie} C (L_1 + L_2) + s^3 L_1 L_2 C}$$

$$\rightarrow 1 = \frac{-s^3 h_{fe} L_1 L_2 C}{sL_1 + h_{ie} + s^2 h_{ie} C (L_1 + L_2) + s^3 L_1 L_2 C} \rightarrow (6)$$

Sub $j\omega = s$, $s^2 = -\omega^2$, $s^3 = -j\omega^3$

$$\rightarrow 1 = \frac{j\omega^3 h_{fe} C L_1 L_2}{j\omega L_1 + h_{ie} + \omega^2 C h_{ie} (L_1 + L_2) - j\omega^3 L_1 L_2 C}$$

$$\rightarrow 1 = \frac{j\omega^3 h_{fe} L_1 L_2 C}{[h_{ie} - \omega^2 C h_{ie} (L_1+L_2)] + j\omega L_1 (1 - \omega^2 L_2 C)} \rightarrow (7)$$

Rationalising the R.H.S,

$$\rightarrow 1 = \frac{j\omega^3 h_{fe} C L_1 L_2 \, [h_{ie} - \omega^2 C h_{ie} (L_1+L_2) - j\omega L_1 (1-\omega^2 L_2 C)]}{[h_{ie} - \omega^2 h_{ie} C (L_1+L_2)]^2 + \omega^2 L_1^2 (1-\omega^2 L_2 C)^2}$$

$$\rightarrow 1 = \frac{j\omega^3 h_{fe} C L_1 L_2 [h_{ie} - \omega^2 C h_{ie} (L_1+L_2)) + \omega^4 h_{fe} L_1^2 L_2 C [1-\omega^2 L_2 C]}{[h_{ie} - \omega^2 h_{ie} C [L_1+L_2]]^2 + \omega^2 L_1^2 (1-\omega^2 L_2 C)^2} \rightarrow (8)$$

Imaginary part of R.H.S must be zero,

$\rightarrow \omega^3 h_{fe} C L_1 L_2 (h_{ie} - \omega^2 C h_{ie} (L_1 + L_2)) = 0$

$\rightarrow \omega^3 h_{fe} h_{ie} C L_1 L_2 (1 - \omega^2 C (L_1+L_2)) = 0$

$\rightarrow 1 - \omega^2 C (L_1 + L_2) = 0$

$\rightarrow \omega^2 = \frac{1}{C(L_1+L_2)} \rightarrow \omega = \frac{1}{\sqrt{C(L_1+L_2)}}$

$\rightarrow f = \frac{1}{2\pi \sqrt{C(L_1+L_2)}} \rightarrow (9)$

At $\omega = 1/\sqrt{C(L_1+L_2)}$ in Eq (8),

$$\rightarrow 1 = \frac{\omega^{\cancel{4}} h_{fe} L_1^{\cancel{2}} L_2 C (1-\omega^2 L_2 C)}{0 + \cancel{\omega^2} \cancel{L_1^2} (1-\omega^2 L_2 C)^{\cancel{2}}}$$

$$\rightarrow 1 = \frac{\omega^2 h_{fe} L_2 C}{1 - \omega^2 L_2 C} = \frac{\frac{1}{\cancel{C}(L_1+L_2)} h_{fe} L_2 \cancel{C}}{1 - \frac{L_2 \cancel{C}}{\cancel{C}(L_1+L_2)}} = \frac{\frac{h_{fe} L_2}{L_1 + \cancel{L_2}}}{\frac{L_1 + \cancel{L_2} - \cancel{L_2}}{L_1 + \cancel{L_2}}}$$

$\rightarrow 1 = \frac{h_{fe} L_2}{L_1}$

$\rightarrow L_1 = h_{fe} L_2$

$\rightarrow h_{fe} = L_1 / L_2 \rightarrow (10)$ = Voltage gain

For a mutual Inductance of M,

$\rightarrow h_{fe} = \frac{L_1 + M}{L_2 + M} \rightarrow (11)$

Now L_1 and L_2 connected in series denoted as

$\rightarrow L_{eq} = L_1 + L_2 \rightarrow (12)$, In mutual Inductance, $L_{eq} = L_1 + L_2 + 2M$.

Sub (12) in (9)

$\rightarrow f = \frac{1}{2\pi \sqrt{C L_{eq}}} \rightarrow (13)$

1. If $L_1 = 1\,mH$, $L_2 = 2\,mH$ and $C = 0.1\,nF$, what is the frequency of oscillation of the Hartley Oscillator ?

Soln Given $L_1 = 1\,mH$, $L_2 = 2\,mH$, $C = 0.1\,nF$

$$\rightarrow f = \frac{1}{2\pi\sqrt{C(L_1+L_2)}} = \frac{1}{2\pi\sqrt{(0.1\times10^{-9}(1\times10^{-3}+2\times10^{-3})}} = 290.5\,KHz$$

2. In a Hartley oscillator, $L_1 = 15\,mH$ and $C = 50\,pF$. Calculate L_2 for a frequency of 168 KHz. The mutual inductance between L_1 and L_2 is $5\,\mu H$. Also find the required gain of the transistor to be used for the oscillations.

Soln Given $L_1 = 15\,mH$, $C = 50\,pF$, $f = 168\,KHz$, $2M = 5\,\mu H = M$

$$\rightarrow f = \frac{1}{2\pi\sqrt{L_{eq}C}} \Rightarrow 168\times10^{3} = \frac{1}{2\pi\sqrt{L_{eq}\times50\times10^{-12}}}$$

$$168\,10^{3}\times2\pi\sqrt{L_{eq}\times50\times10^{-12}} = 1$$

squaring

$$(2\pi\,168\times10^{3})^2 L_{eq}\,50\times10^{-12} = 1$$

$$L_{eq} = \frac{1}{(2\pi\,168\times10^{3})^2\,50\times10^{-12}}$$

$\rightarrow L_{eq} = 17.95\,mH$

where $L_{eq} = L_1 + L_2 + 2M$

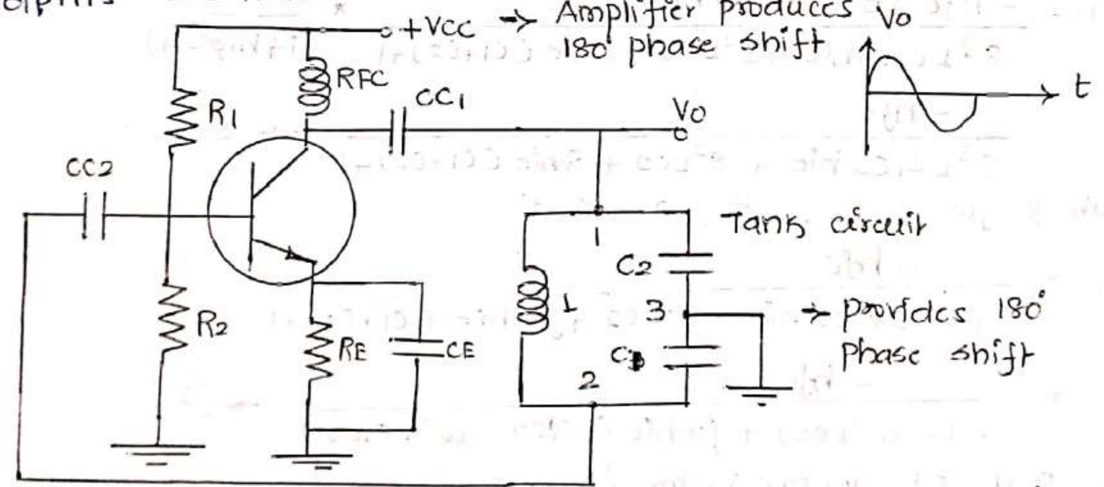
$\rightarrow 17.95\times10^{-3} = 15\times10^{-3} + L_2 + 5\times10^{-6}$

$\rightarrow L_2 = 2.945\,mH$

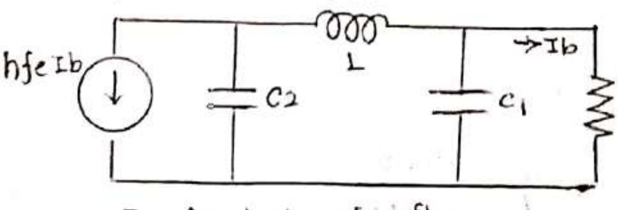
$$\rightarrow h_{fe} = \frac{L_1+M}{L_2+M} = \frac{15\times10^{-3}+5\times10^{-6}}{2.945\times10^{-3}+5\times10^{-6}} = 5.08$$

Colpitts Oscillator :

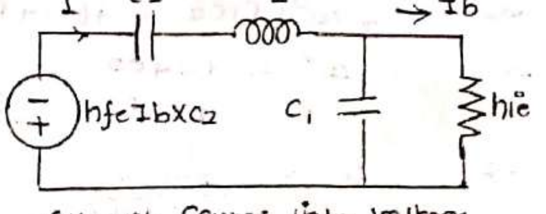
An LC oscillator which uses two capacitive reactances and one inductive reactance in the feedback network is called Colpitts Oscillator.



The amplifier stage produces 180° phase shift and tank circuit gives an additional phase shift of 180° to satisfy the oscillation conditions.



Equivalent circuit.



current source into voltage source

From the circuit,

$$\rightarrow V_o = h_{fe} I_b X_{C2} = h_{fe} I_b \frac{1}{j\omega C_2} \rightarrow (1)$$

$$\text{Ey} \rightarrow I = \frac{-V_o}{(X_{C2}+X_L)+(X_{C1} \| h_{ie})} \rightarrow (2)$$

∴ Neg. sign indicates current direction in opposite.

$$\rightarrow I = \frac{-h_{fe} I_b \frac{1}{j\omega C_2}}{\left[\frac{1}{j\omega C_2} + j\omega L\right] + \left[\frac{h_{ie}/j\omega C_1}{h_{ie}+\frac{1}{j\omega C_1}}\right]} \rightarrow (3)$$

Replace $j\omega$ by s,

$$\rightarrow I = \frac{-h_{fe} I_b \left(\frac{1}{sC_2}\right)}{\left[\frac{1}{sC_2} + sL\right] + \left[\frac{h_{ie}/sC_1}{h_{ie}+1/sC_1}\right]} = \frac{\frac{-h_{fe} I_b}{sC_2}}{\left[\frac{1+s^2LC_2}{sC_2}\right] + \left[\frac{h_{ie}}{1+h_{ie}sC_1}\right]}$$

$$\rightarrow I = \frac{-\frac{h_{fe} I_b}{s\cancel{C_2}} (s\cancel{C_2})(1+h_{ie}sC_1)}{(1+s^2LC_2)(1+sC_1h_{ie})+h_{ie}sC_2} = \frac{-h_{fe} I_b (1+h_{ie}sC_1)}{1+sC_1h_{ie}+s^2LC_2+s^3LC_1C_2h_{ie}+h_{ie}sC_2}$$

$$\rightarrow I = \frac{-h_{fe} I_b (1+sC_1h_{ie})}{s^3LC_1C_2h_{ie}+s^2LC_2+sh_{ie}(C_1+C_2)+1} \rightarrow (4)$$

According to current division in parallel circuit,

$$\rightarrow I_b = I \frac{X_{C1}}{X_{C1}+h_{ie}} = \frac{I \times \frac{1}{j\omega C_1}}{\frac{1}{j\omega C_1}+h_{ie}} = \frac{I}{1+h_{ie}j\omega C_1} = \frac{I}{1+h_{ie}sC_1} \rightarrow (5)$$

sub (4) in (5)

$$\rightarrow I_b = \frac{-h_{fe} I_b (1+\cancel{sC_1h_{ie}})}{s^3LC_1C_2h_{ie}+s^2LC_2+sh_{ie}(C_1+C_2)+1} \times \frac{1}{(1+\cancel{h_{ie}sC_1})}$$

$$\rightarrow 1 = \frac{-h_{fe}}{s^3LC_1C_2h_{ie}+s^2LC_2+sh_{ie}(C_1+C_2)+1} \rightarrow (6)$$

Sub $s = j\omega$, $s^2 = -\omega^2$, $s^3 = -j\omega^3$.

$$\rightarrow 1 = \frac{-h_{fe}}{-j\omega^3LC_1C_2h_{ie}-\omega^2LC_2+j\omega h_{ie}(C_1+C_2)+1}$$

$$\rightarrow 1 = \frac{-h_{fe}}{(1-\omega^2LC_2)+j\omega h_{ie}(C_1+C_2-\omega^2LC_1C_2)} \rightarrow (7)$$

Im. part of denominator is zero,

$$\rightarrow \omega h_{ie}(C_1+C_2-\omega^2LC_1C_2) = 0$$

$$\rightarrow C_1+C_2-\omega^2LC_1C_2 = 0$$

$$\rightarrow +\omega^2LC_1C_2 = +(C_1+C_2)$$

$$\rightarrow \omega^2 = \frac{C_1+C_2}{LC_1C_2} = \frac{1}{L\left(\frac{C_1C_2}{C_1+C_2}\right)}$$

$$\rightarrow \omega = \frac{1}{\sqrt{L\left(\frac{C_1C_2}{C_1+C_2}\right)}} \quad \because C_{eq} = \frac{C_1C_2}{C_1+C_2}$$

$\rightarrow \omega = \frac{1}{\sqrt{Lceq}}$ $\because \omega = 2\pi f$

$\rightarrow f = \frac{1}{2\pi\sqrt{Lceq}} \rightarrow (8)$

At $\omega = \sqrt{\frac{C_1+C_2}{L(C_1C_2)}}$ in (7)

$\rightarrow 1 = \frac{-hfe}{1-\omega^2 L C_2} \Rightarrow 1 = \frac{-hfe}{1-\left(\frac{C_1+C_2}{\cancel{L}C_1\cancel{C_2}}\right)\cancel{L}C_2}$

$\rightarrow 1 = \frac{-hfe}{1-\frac{C_1+C_2}{C_1}} \Rightarrow 1 = \frac{+hfe}{\frac{\cancel{C_1}-\cancel{C_1}+C_2}{C_1}}$

$\rightarrow hfe = \frac{C_2}{C_1} \rightarrow (9)$ = Voltage gain.

1. In a colpitt's oscillator $C_1 = 0.001\mu F$, $C_2 = 0.01\mu F$ and $L = 10\mu H$. Find the frequency of oscillation, feedback factor and voltage gain.

Soln Given $C_1 = 0.001\mu F$, $C_2 = 0.01\mu F$, $L = 10\mu H$

$\rightarrow f = \frac{1}{2\pi\sqrt{Lceq}}$ $\because ceq = \frac{C_1C_2}{C_1+C_2} = \frac{0.001\times10^{-6}\times0.01\times10^{-6}}{0.001\times10^{-6}+0.01\times10^{-6}}$

$ceq = 9.09\times10^{-10}F$

$\rightarrow f = \frac{1}{2\pi\sqrt{10\times10^{-6}\times9.09\times10^{-10}}} = 1.6692$ MHz.

$\rightarrow A\beta = 1$ $\because A = \frac{C_2}{C_1} = \frac{0.01\times10^{-6}}{0.001\times10^{-6}} = 10$

$\beta = \frac{1}{A} = \frac{1}{10} = 0.1$

2. If C_1 and C_2 are 200pF and 50pF respectively. calculate the value of inductance for producing oscillations at 1MHz in the colpitts oscillator circuit.

Soln Given $C_1 = 200\times10^{-12}$, $C_2 = 50\times10^{-12}$, $f = 1\times10^6$

$\rightarrow f = \frac{1}{2\pi\sqrt{Lceq}}$ $\because ceq = \frac{C_1C_2}{C_1+C_2} = 40pF$

$\rightarrow f = \frac{1}{2\pi\sqrt{Lceq}} \Rightarrow 1\times10^6 = \frac{1}{2\pi\sqrt{L\times40\times10^{-12}}}$

$\rightarrow L = 0.6332$ mH.

3. In a colpitts oscillator, the values of the inductors and capacitors in the tank circuit are $L = 40mH$, $C_1 = 100pF$, $C_2 = 500pF$

find 1. Frequency of oscillation
2. If the output voltage is 10V, find the feedback voltage
3. Find the minimum gain, if the frequency is changed by changing 'L' alone.
4. Find the value of C_1 for a gain of 10.
5. Also, find the new frequency of oscillation.

Soln Given $L = 40mH$, $C_1 = 100pF$, $C_2 = 500pF$

1. $f = \frac{1}{2\pi\sqrt{LC_{eq}}}$ $\because C_{eq} = \frac{C_1C_2}{C_1+C_2} = 83.33pF$ $\rightarrow f = 87.17KHz$

2. $V_o = 10V$, V_f $\because \beta = h_{fe} = \frac{C_2}{C_1} = \frac{500p}{100p} = 5$

WKT, $\beta = \frac{V_o}{V_f}$

$\rightarrow V_f = \frac{V_o}{\beta} = \frac{10}{5} = 2V.$

3. Min. Gain,

Min. Gain $= \frac{C_2}{C_1} = \frac{500p}{100p} = 5$ (or) $h_{fe(min)} = 5.$

4. C_1 for a gain of 10.

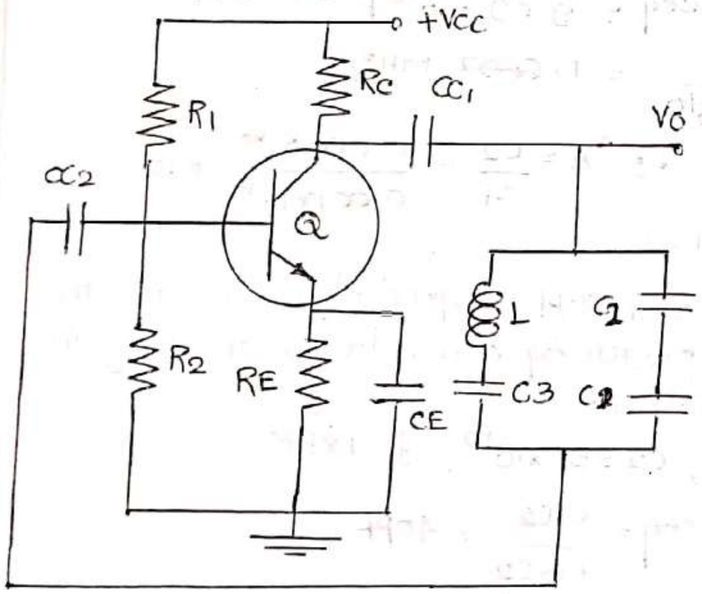
$\rightarrow$ Gain $= 10 = \frac{C_2}{C_1} \rightarrow 10 = \frac{500pF}{C_1} \rightarrow C_1 = 50pF$

$\rightarrow C_1 = 50pF$

5. $C_1 = 50pF$, $C_2 = 500pF$

$\rightarrow f = \frac{1}{2\pi\sqrt{LC_{eq}}} = 118.032KHz$

Clapp Oscillator :



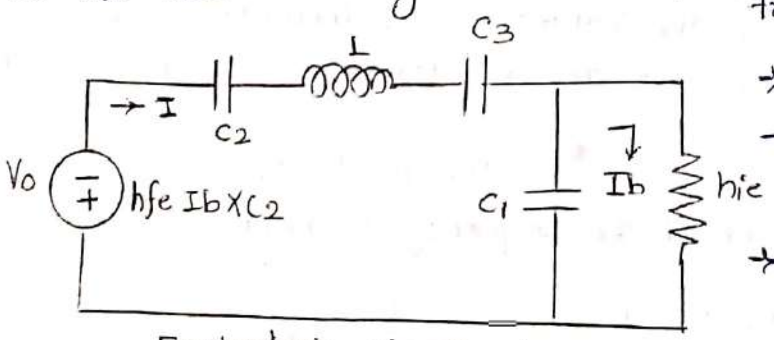
The dia. shows clapp oscillator.

To achieve frequency stability, colpitts oscillator is slightly modified in practice c's called clapp oscillator.

The tank circuit of clapp oscillator introduces one more capacitor C_3 in series with the Inductance.

The capacitor C_3 gives good stability is achieved.

Another advantage of C_3 is that it can be kept variable. As frequency is dependent on C_3, the frequency can be varied in the desired range.



Equivalent circuit

From the circuit,

$\rightarrow V_o = h_{fe} I_b X_{C2}$

$\rightarrow V_o = h_{fe} I_b \cdot \frac{1}{j\omega C_2} \rightarrow (1)$

and

$\rightarrow I = \frac{-V_o}{[X_{C2}+X_{C3}+X_L]+[X_{C1} \parallel h_{ie}]} \rightarrow (2)$

$$\Rightarrow I = \frac{-h_{fe} I_b \cdot \frac{1}{j\omega C_2}}{\frac{1}{j\omega C_2} + \frac{1}{j\omega C_3} + j\omega L + \left[\frac{\frac{1}{j\omega C_1} h_{ie}}{\frac{1}{j\omega C_1} + h_{ie}}\right]} \rightarrow (3)$$

Replace $j\omega$ by s

$$\Rightarrow I = \frac{-h_{fe} I_b \frac{1}{sC_2}}{\frac{1}{sC_2} + \frac{1}{sC_3} + sL + \left[\frac{h_{ie}/sC_1}{\frac{1}{sC_1} + h_{ie}}\right]} \rightarrow (4)$$

(x) by sC_2 in eq (4)

$$\Rightarrow I = \frac{-h_{fe} I_b}{1 + \frac{C_2}{C_3} + s^2 L C_2 + \frac{\frac{h_{ie}}{s\cancel{C_1}}}{\frac{1 + h_{ie} s C_1}{s\cancel{C_1}}} \times sC_2} = \frac{-h_{fe} I_b}{1 + \frac{C_2}{C_3} + s^2 L C_2 + \frac{h_{ie} sC_2}{1 + h_{ie} sC_1}}$$

(x) N & D by C_3

$$\Rightarrow I = \frac{-h_{fe} I_b C_3}{C_3 + C_2 + s^2 L C_2 C_3 + \frac{sC_2 C_3 h_{ie}}{1 + h_{ie} sC_1}} = \frac{-h_{fe} I_b C_3 (1 + h_{ie} sC_1)}{C_3 + sC_1 C_3 h_{ie} + C_2 + sC_1 C_2 h_{ie} + s^2 L C_2 C_3 + s^3 L C_1 C_2 C_3 h_{ie} + sC_2 C_3 h_{ie}}$$

$$\Rightarrow I = \frac{-h_{fe} I_b C_3 (1 + sC_1 h_{ie})}{s^3 L C_1 C_2 C_3 h_{ie} + s^2 L C_2 C_3 + s h_{ie}[C_2 C_3 + C_1 C_2 + C_1 C_3] + C_2 + C_3} \rightarrow (5)$$

From the circuit,

$$\Rightarrow I_b = I \times \frac{X_{C1}}{X_{C1} + h_{ie}} = \frac{I \times \frac{1}{j\omega C_1}}{\frac{1}{j\omega C_1} + h_{ie}} = \frac{I \times \frac{1}{sC_1}}{\frac{1}{sC_1} + h_{ie}} = \frac{\frac{I}{sC_1}}{\frac{1 + h_{ie} sC_1}{sC_1}} = \frac{I}{1 + h_{ie} sC_1} \rightarrow (6)$$

Sub (5) in (6)

$$\Rightarrow \cancel{I_b} = \frac{-h_{fe} \cancel{I_b} C_3 (1 + \cancel{sC_1 h_{ie}})}{s^3 L C_1 C_2 C_3 h_{ie} + s^2 L C_2 C_3 + s h_{ie}[C_2 C_3 + C_1 C_2 + C_1 C_3] + C_2 + C_3} \times \frac{1}{1 + h_{ie} sC_1}$$

$$\Rightarrow 1 = \frac{-h_{fe} C_3}{s^3 L C_1 C_2 C_3 h_{ie} + s^2 L C_2 C_3 + s h_{ie}[C_1 C_2 + C_2 C_3 + C_1 C_3] + C_2 + C_3}$$

Sub $s = j\omega$, $s^2 = -\omega^2$, $s^3 = -j\omega^3$.

$$\Rightarrow 1 = \frac{-h_{fe} C_3}{-j\omega^3 L C_1 C_2 C_3 h_{ie} + \omega^2 L C_2 C_3 + j\omega h_{ie}[C_1 C_2 + C_2 C_3 + C_1 C_3] + C_2 + C_3}$$

$$\Rightarrow 1 = \frac{-h_{fe} C_3}{C_2 + C_3 - \omega^2 L C_2 C_3 + j\omega h_{ie}[(C_1 C_2 + C_2 C_3 + C_3 C_1) - \omega^2 L C_1 C_2 C_3]} \rightarrow (7)$$

To satisfy Barkhausen, imaginary part in deno is zero,

$$\Rightarrow \omega h_{ie}[C_1 C_2 + C_2 C_3 + C_3 C_1 - \omega^2 L C_1 C_2 C_3] = 0$$

$$\Rightarrow C_1 C_2 + C_2 C_3 + C_3 C_1 - \omega^2 L C_1 C_2 C_3 = 0$$

$$\Rightarrow \omega^2 = \frac{C_1 C_2 + C_2 C_3 + C_3 C_1}{L C_1 C_2 C_3} = \frac{\frac{1}{C_3} + \frac{1}{C_1} + \frac{1}{C_2}}{L} = \frac{\frac{1}{C_{eq}}}{L}$$

$$\Rightarrow \omega^2 = \frac{1}{L C_{eq}} \rightarrow (8)$$

$$\Rightarrow \omega = 1/\sqrt{L C_{eq}}$$

$\because \omega = 2\pi f$, $f = \dfrac{1}{2\pi\sqrt{LC_{eq}}}$ → (9)

$\because \dfrac{1}{C_{eq}} = \dfrac{1}{C_1} + \dfrac{1}{C_2} + \dfrac{1}{C_3}$.

But $C_3 << C_1$ and C_2, $C_{eq} = C_3$

$\Rightarrow f = \dfrac{1}{2\pi\sqrt{LC_3}}$ → (10)

Advantages:

1. The frequency is stable and accurate
2. The good frequency stability
3. Keeping C_3 variable, frequency can be varied in desired range.

problem:

1. Calculate the frequency of oscillation for the clapp oscillator with $C_1 = 0.1\mu F$, $C_2 = 1\mu F$, $C_3 = 100pF$ and $L = 470\mu H$.

soln Given $C_1 = 0.1\mu F$, $C_2 = 1\mu F$, $C_3 = 100pF$, $L = 470\mu H$

$\Rightarrow \dfrac{1}{C_{eq}} = \dfrac{1}{C_1} + \dfrac{1}{C_2} + \dfrac{1}{C_3} = \dfrac{1}{0.1\mu} + \dfrac{1}{1\mu} + \dfrac{1}{100p}$.

$\Rightarrow C_{eq} = 99.89\ pF$

$\Rightarrow F = \dfrac{1}{2\pi\sqrt{LC_{eq}}} = \dfrac{1}{2\pi\sqrt{470\mu \times 99.89p}} = 734.53\ KHz$.

2. Calculate the Inductance value to produce 734.5 KHz frequency of oscillation in clapp oscillator having $C_1 = 0.1\mu F$, $C_2 = 1\mu F$, $C_3 = 100\mu F$

soln Given $f = 734.5K$, $C_1 = 0.1\mu$, $C_2 = 1\mu$, $C_3 = 100\mu$

$\Rightarrow \dfrac{1}{C_{eq}} = \dfrac{1}{C_1} + \dfrac{1}{C_2} + \dfrac{1}{C_3} \Rightarrow C_{eq} = 9.0826 \times 10^{-8} F$

$\Rightarrow f = \dfrac{1}{2\pi\sqrt{LC_{eq}}} \Rightarrow 734.5K = \dfrac{1}{2\pi\sqrt{L \times 9.0826 \times 10^{-8}}}$

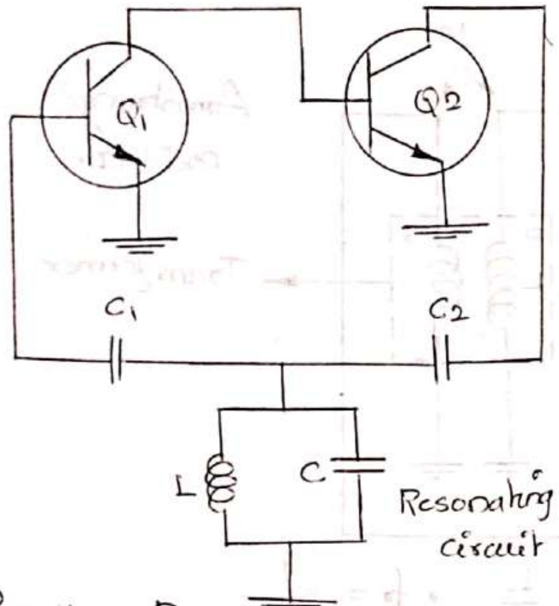
$\Rightarrow L = 0.5169\ \mu H$.

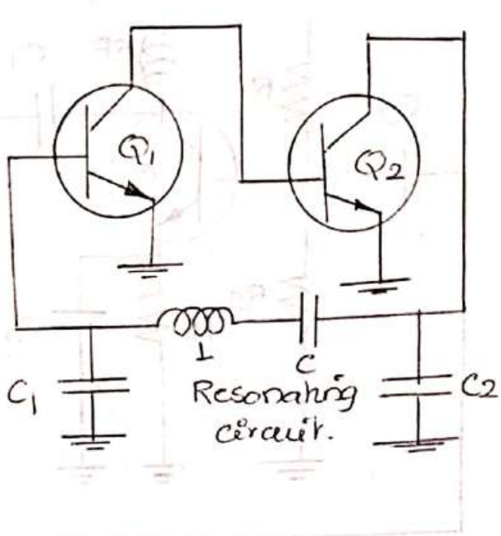
Franklin oscillator:

The Franklin oscillator uses two transistor stages with some common terminal which is usually an Emitter. Each stage provides a phase shift of 180° and hence total phase shift is 360°, as required by an oscillator. Both stages provide amplification as well as phase inversion.

Features:

1. It is used with MOSFET devices.
2. Used to achieve frequency stability.
3. The frequency is less affected by the changes in the active device parameters.

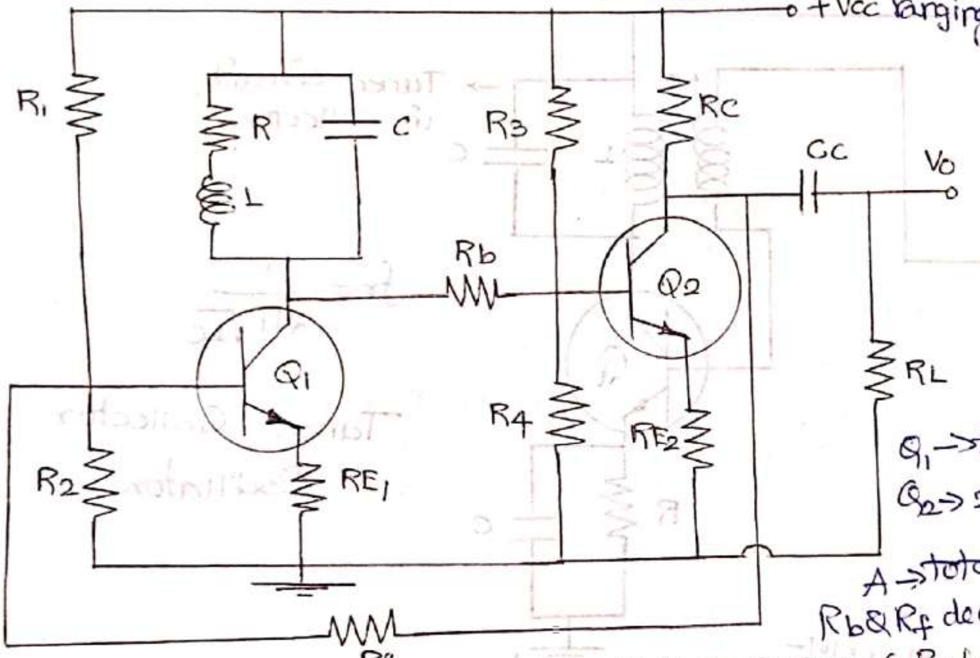


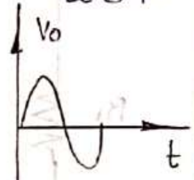


series Resonating circuit.

Parallel Resonating circuit → isolation from the active device by using the capacitors C_1 & C_2

Practical Parallel Resonating circuit : → It is used to achieve the frequencies ranging from 100MHz to 3GHz





Q_1 → forms 1^{st} stage
Q_2 → 2^{nd} stage

A → total gain
R_b & R_f decides the loop gain $A\beta$

Note : Series. (Refer Bakshi) → proper selection is achieved $A\beta > 1$

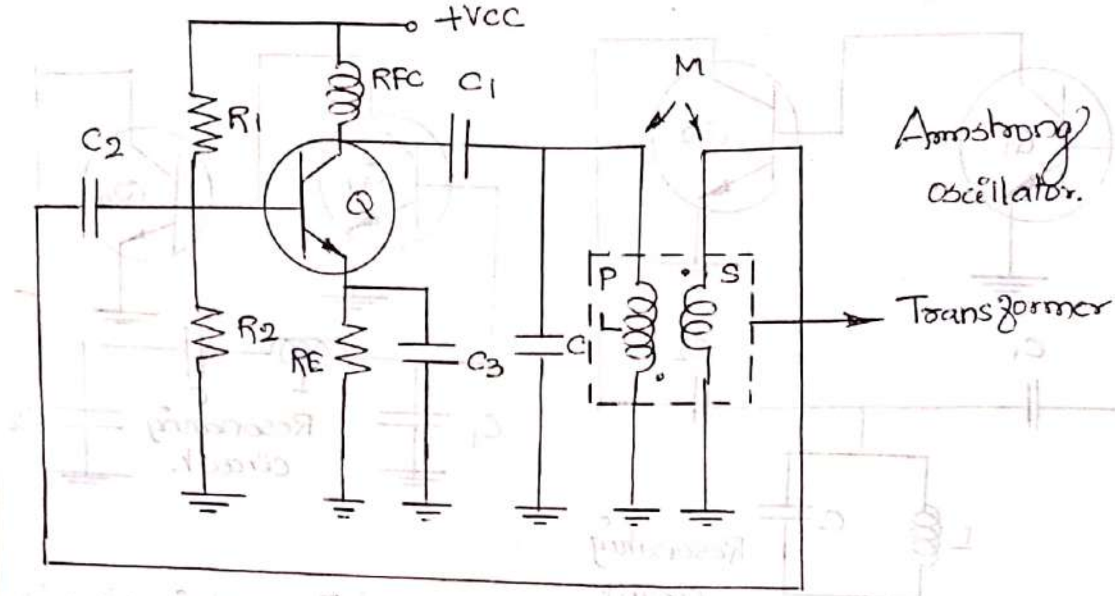
$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

Applications :

1. Used for RF oscillator
2. Many laboratory devices.
3. Precision Frequency Meters

Armstrong Oscillator :

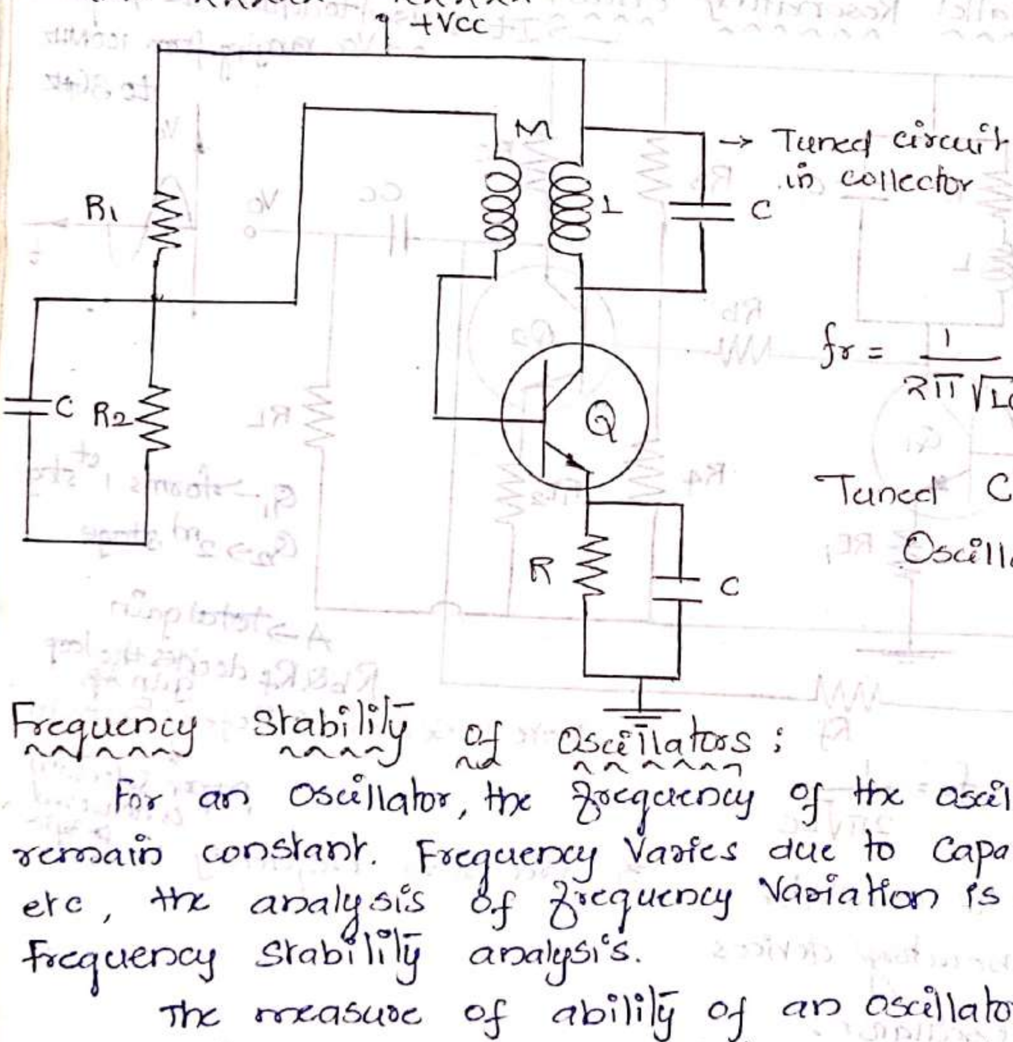
In This Oscillator, a transformer is used, whose Primary acts as L in the circuit while the Voltage across secondary is used as a feedback. This is another type of LC oscillator.



Armstrong oscillator.

$\rightarrow \beta = \frac{M}{L}$, $A = \frac{1}{\beta} = \frac{L}{M}$, $f_r = \frac{1}{2\pi\sqrt{LC}}$

Tuned Collector Oscillator :



$f_r = \frac{1}{2\pi\sqrt{LC}}$

Tuned Collector Oscillator.

Frequency Stability of Oscillators :

For an Oscillator, the frequency of the oscillations must remain constant. Frequency Varies due to Capacitance, temp etc, the analysis of frequency variation is called 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.

The Transistors are temperature sensitive. As temp changes, the oscillating frequency also changes and no longer remains stable. hence pratically the circuits Cannot provide Stable frequency.

Factors affecting the frequency stability :

1. Due to the change in temp, the capacitor and Inductor Value also changes.
2. Due to changes in temp, the parameters like BJT, FET gets affected which in terms of frequency.
3. The Variation in the power supply in another factor affecting the frequency.
4. The changes in atmospheric conditions, ageing and unstable transistor parameters affect the frequency.
5. The changes in the load connected, affect the Effective resistance of tank circuit.
6. The Capacitive effect in transistor affect the capacitance of tank circuit and hence the frequency.

. Variation of frequency with temp is given by factor denoted as

→ $S_{\omega,T} = \frac{\Delta\omega/\omega_r}{\Delta T/T_r}$ parts per million per °c ∴ $\Delta\omega$ = change in freq.

ω_r = Desired freq.

ΔT = Charg. in Tc

T_r = openg Temp

The frequency stability is defined as

→ $S_{\omega} = \frac{d\theta}{d\omega}$

The freq. stability can be Improved by,

1. Enclosing the circuit in a constant temperature chamber.
2. Maintaining constant voltage by using 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 impedance.

Crystal Oscillators : (or) Quartz crystal construction :

→ Crystals occurs naturally or can be manufactured.

→ It exhibits piezo electric effect.

Piezo Electric effect :

Under the influence of 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 ac voltage, it vibrates causing mechanical distortion in the crystal shape.

Every crystal has its own resonating freq. depending on its shape.

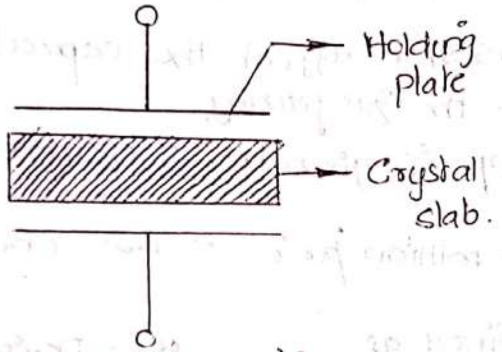
→ So under the influence of mechanical vibrations, the crystal generates an electrical signal of very constant freq.

The main substances exhibiting the piezo electric effect are

1. Quartz 2. Rochelle salt 3. Tourmaline.

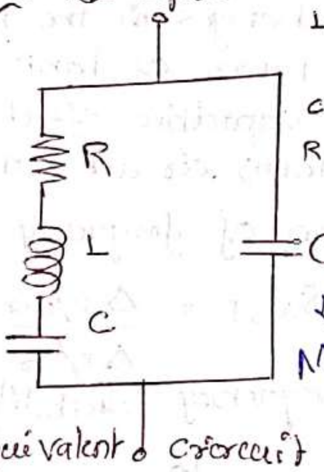
1. Rochelle salts have the greatest piezo electric activity but mechanically weak. → microphone associated with LS, Tape recorders, headsets
2. Tourmaline shows least piezo electric effect but mechanically strongest. It is most expensive.
3. Quartz is a compromise between the piezo electric activity of the Rochelle salts and strength of tourmaline. Quartz is inexpensive.

Construction Details :



Symbol Representation

Electrical Ac Equivalent Circuit :



L = Eq. to mass of vib crys.
C = Mec. compliance
R = Electrical Eq. of Mc. fric.

Ac Equivalent circuit of a crystal.

The resonating frequency is

$$\rightarrow f_r = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{Q^2}{1+Q^2}}, \quad Q = \frac{\omega L}{R} \quad \because Q = \text{Quality factor.}$$

When $\sqrt{\frac{Q^2}{1+Q^2}}$ approaches to unity.

$$\rightarrow f_r = \frac{1}{2\pi\sqrt{LC}} = f \propto \frac{1}{t}$$ → thickness of the crystal

Hence practically are used upto 200 to 300 KHZ only

Series and parallel Resonance :

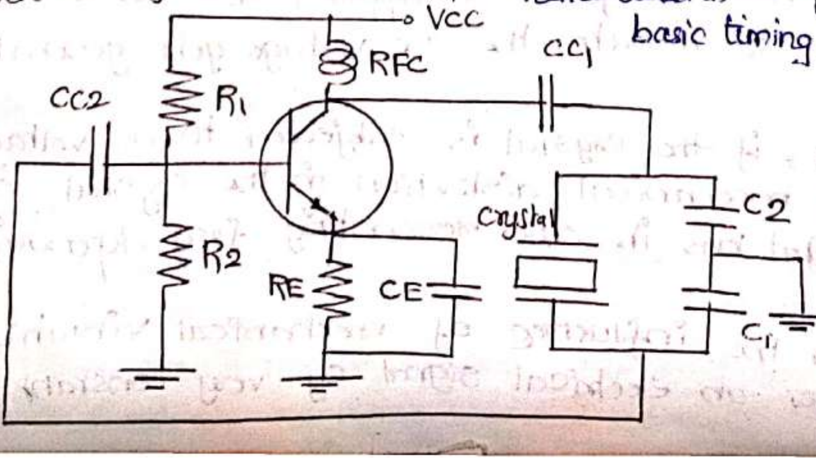
$$\rightarrow f_s = \frac{1}{2\pi\sqrt{LC}} \qquad \rightarrow f_p = \frac{1}{2\pi\sqrt{LC_{eq}}} \qquad \because C_{eq} = \frac{C_M C}{C_M + C}$$

Crystal stability :

1. Temperature stability → change in freq to change in temp
2. Long term stability → aging 2×10^{-8}
3. Short term stability.

crystal has good freq stability hence used in counters, computers, basic timing devices in elec const watches etc.

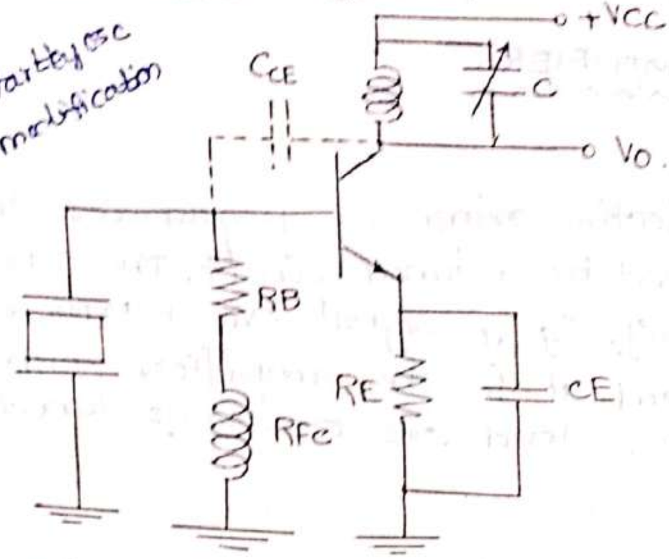
1. **Pierce crystal Oscillator :**

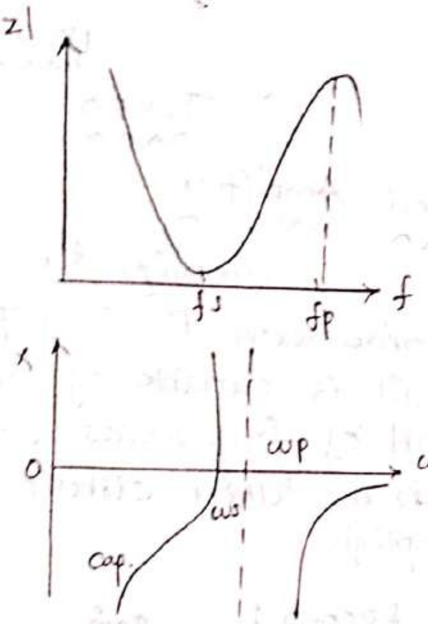


Colpitt osc modification.

2. Miller Crystal Oscillator :

Hartley osc modification





Problem :

1. A crystal has $L = 0.33H$, $C = 0.065\,PF$ and $C_M = 1\,PF$ with $R = 5.5k$ find 1. Series resonant frequency 2. parallel resonant frequency 3. By what percent does the parallel resonant frequency exceeds the series resonant frequency ? 4. find Q factor of the crystal.

Soln Given $L = 0.33H$, $C = 0.065\,PF$, $C_M = 1\,PF$, $R = 5.5k$.

1. $f_s = \frac{1}{2\pi\sqrt{LC}} = 1.087\ MHz$ 2. $f_p = \frac{1}{2\pi\sqrt{LC_{eq}}} = 1.121\ MHz$

3. % increase $= \frac{1.121 - 1.087}{1.087} \times 100$ $C_{eq} = \frac{C C_M}{C + C_M} = 0.061\,PF$

$= 3.127\,\%$

4. $Q = \frac{\omega L}{R} = \frac{2\pi f_s L}{R} = 409.789.$

2. A quartz crystal has $L = 3H$, $C_s = 0.05\,pF$, $R = 2000\,\Omega$ and $C_M = 10\,pF$. Calculate the series and parallel resonant frequencies f_s and f_p of the crystal.

Soln Given $L = 3H$, $C_s = 0.05\,pF$ $(= C)$, $R = 2000\,\Omega$, $C_M = 10\,pF$.

→ 1. $f_s = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{3H \times 0.05p}} = 410.9362\ KHz.$

→ 2. $f_p = \frac{1}{2\pi\sqrt{LC_{eq}}} = 411.9623\ KHz$ $C_{eq} = \frac{C_M C}{C_M + C} = 4.9 \times 10^{-14} F$

3. A crystal has the following parameters : $L = 0.5H$, $C_s = 0.06\,pF$, $C_p = 1\,pF$ and $R = 5k$. find Series and parallel resonant frequencies and corresponding Q-factor of the crystal.

Soln Given : $L = 0.5H$, $C_s = 0.06p$, $C_p = 1\,pF$, $R = 5k$

1. $f_s = \frac{1}{\sqrt{LC_s}\,2\pi} = 918.8\ KHz$ 2. $f_p = \frac{1}{2\pi\sqrt{LC_{eq}}} = 946.05k$

3. $Q = \frac{\omega_s L}{R} = 577.35$ $C_{eq} = \frac{C_s C_p}{C_s + C_p} = 5.66 \times 10^{-14} F$

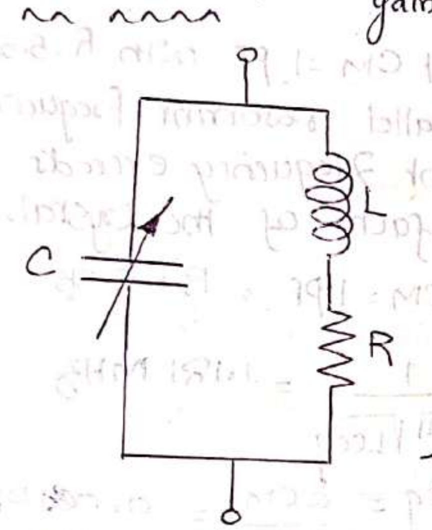
UNIT - III

TUNED AMPLIFIERS

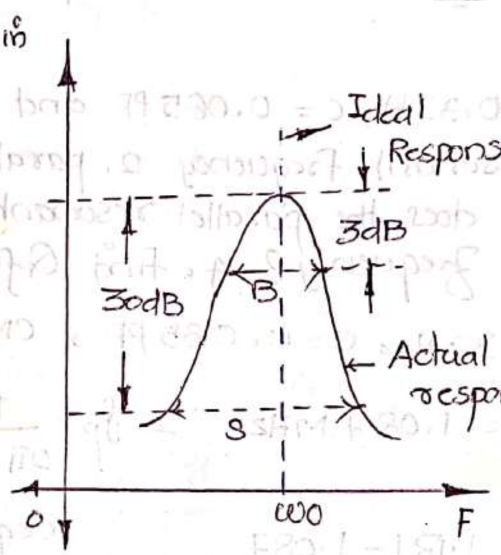
Tuned Amplifier :

To amplify the selective range of frequencies, the resistive load R_c is replaced by a tuned circuit. The tuned circuit is capable of amplifying a signal over a narrow band of frequencies centered at f_r. The amplifiers with such a tuned circuit as a load are known as tuned amplifier.

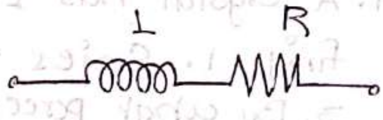
Coil Losses :



: Tuned circuit



Frequency Response



Inductor with leakage Resistance

→ $f_r = \frac{1}{2\pi\sqrt{LC}}$

→ $Z_r = \frac{L}{CR}$

→ Skirt Selectivity

The tuned circuit consist of coil. Practically, coil is not purely inductive. It consist of few losses represented in the form of leakage resistance in series with the Inductor.

The Losses are copper Losses, Eddy current Loss and hysteresis loss.

1. The copper loss at low frequency is equivalent to the d.c resistance of the coil. and it is inversely proportional to frequency. As frequency increases, the copper loss decrease.
2. Eddy current is a loss due to heating within the Inductor. and it is directly proportional to frequency.
3. Hysteresis loss is proportional to the area enclosed by the hysteresis loop and to the rate at which this loop is transversed. It is a function of signal level and increase with frequency. and it is independent of frequency.

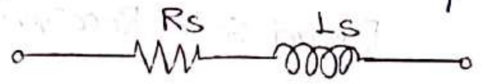
Quality factor (Q factor):

The Q factor is the ratio of reactance to resistance. It can also be defined as the measure of efficiency with which inductor can store the energy.

$$Q = 2\pi \frac{\text{Max energy stored per cycle}}{\text{Energy dissipate per cycle}}$$

Dissipation factor: (D)

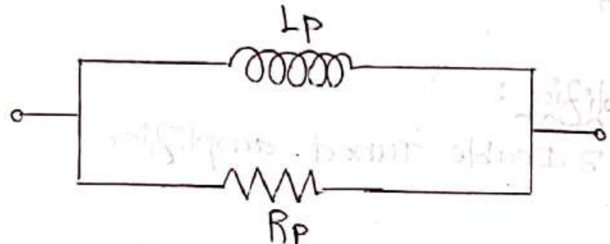
The dissipation factor (D) that can be referred to as the total loss within a component is defined as $1/Q$.



Series circuit

Inductive Impedance

$$\frac{\omega L_S}{R_S}$$



Parallel circuit

Inductive admittance

$$\frac{R_P}{\omega L_P}$$

$$Q = \frac{1}{D} = \frac{\omega L_S}{R_S} = \frac{R_P}{\omega L_P}$$

Quality factor Equations.

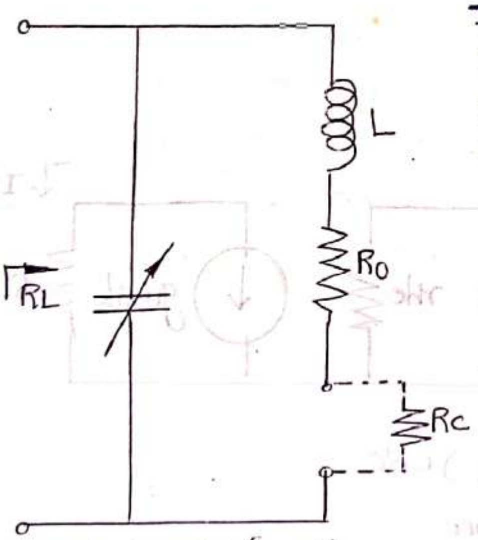
Unloaded and Loaded Q: of tank circuits:

Unloaded Q:

It is the ratio of stored energy to dissipated energy in a reactor or resonator.

Loaded Q (Q_L):

The loaded Q of a resonator is determined by how tightly the resonator is coupled to its terminations.



Tuned load circuit.

From the circuit, L and C represents tank circuit. The internal circuit losses of inductor are represented by R_O and R_C represents the coupled in load.

From the circuit, we write,

$\rightarrow R_O = \frac{\omega_0 L}{Q_u}$, $R_C = \frac{\omega_0 L}{Q_L}$

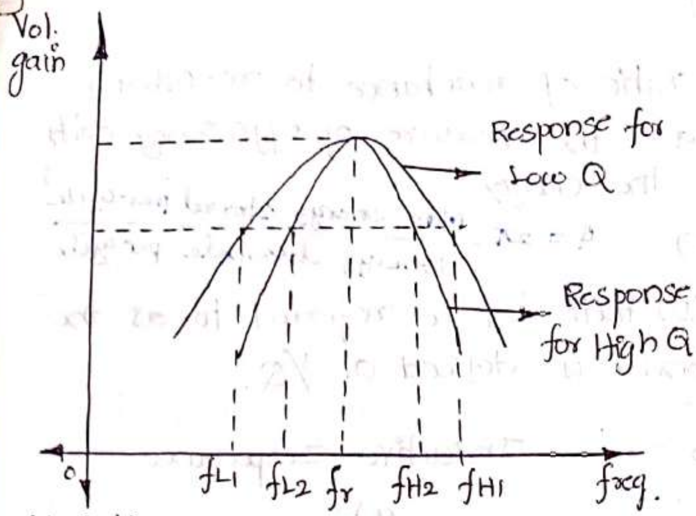
where Q_u is Unloaded Q, Q_L is Loaded Q.

$\rightarrow$ Efficiency of tank circuit is

$$\rightarrow \eta = \frac{Q_u}{Q_u + Q_L} = \frac{I^2 R_C}{I^2 (R_C + R_O)} \times 100\%$$

The quality factor Q_L determines the 3dB bandwidth for the resonating circuit. The 3dB bandwidth for resonant circuit is

$\rightarrow BW = f_r / Q_L$ ∵ f_r = Resonant centre frequency.



Variation of 3dB bandwidth with Variation in Quality factor.

If Q is large, BW is small.
for small Q, BW is high.
Thus in tuned amplifier Q is kept as high as possible to get better selectivity.

Application:
Communication,
Broad Cast Receivers.

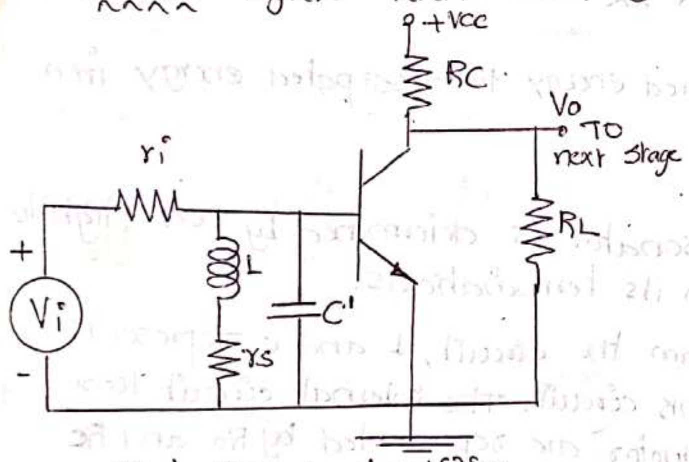
Classification of Tuned Amplifier:

1. Single Tuned amplifier 2. Double Tuned amplifier
3. Stagger Tuned amplifier.

These amplifier are further classified according to coupling,

1. Capacitive 2. Inductive 3. Transformer coupled.

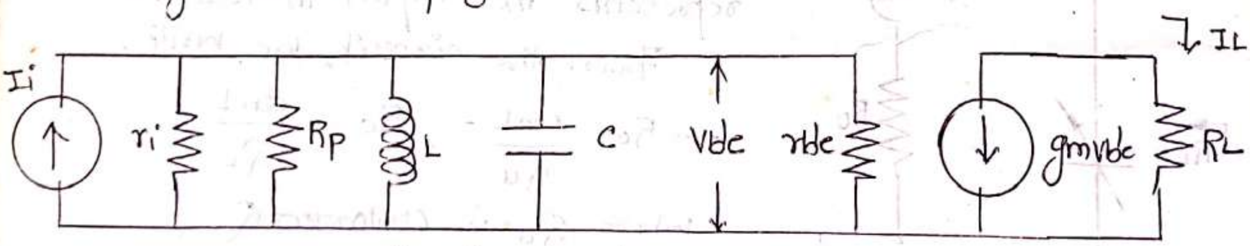
1. Small Signal Tuned Amplifier:



Single Tuned Amplifier.

Assumptions:

1. $R_L << R_C$
2. $r_{bb'} = 0$.



Equivalent circuit.

Where $C_{eq} = C' + C_{b'e} + (1 + g_m R_L) C_{b'c}$

C' = External capacitance

$C(1 + g_m R_L) C_{b'c}$ = Miller Capacitance.

r_s = Losses in coil.

Assuming coil losses are low and Q is large, then

$$Q_C = \frac{\omega L}{r_C} >> 1. \rightarrow (1)$$

From single tuned circuit, the admittance is

$$\Rightarrow Y_1 = \frac{1}{z_1} = \frac{1}{r_c + j\omega L} = \frac{r_c - j\omega L}{r_c^2 + \omega^2 L^2} = \frac{r_c}{r_c^2 + \omega^2 L^2} - \frac{j\omega L}{r_c^2 + \omega^2 L^2}$$

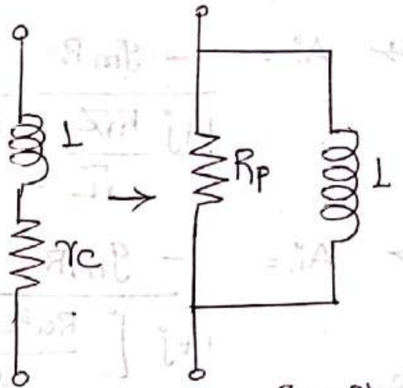
$$\Rightarrow Y_1 = \frac{r_c}{\omega^2 L^2} + \frac{\omega L}{j\omega^2 L^2} \qquad \because \omega L >> r_c \text{ from Eq } 1$$

$$Y_1 = \frac{r_c}{\omega^2 L^2} + \frac{1}{j\omega L} \qquad \because \frac{1}{j} = -j$$

$$\rightarrow Y_2 = \frac{1}{R_p} + \frac{1}{j\omega L} \rightarrow \text{Parallel ckt.}$$

ie)

$$Y_2 = \frac{1}{Z_2} = \frac{1}{\frac{R_p j\omega L}{R_p + j\omega L}} = \frac{R_p + j\omega L}{R_p j\omega L}$$



$$\rightarrow Y_2 = \frac{R_p}{R_p j\omega L} + \frac{j\omega L}{R_p j\omega L}$$

Equivalent circuits.

$$\rightarrow Y_2 = \frac{1}{j\omega L} + \frac{1}{R_p}$$

Now Equating Y_1 and Y_2,

$$\rightarrow \frac{r_c}{\omega^2 L^2} + \cancel{\frac{1}{j\omega L}} = \frac{1}{R_p} + \cancel{\frac{1}{j\omega L}}$$

$$\rightarrow \frac{1}{R_p} = \frac{r_c}{\omega^2 L^2} \times \frac{r_c}{r_c} \rightarrow \frac{r_c^2}{r_c \omega^2 L^2}$$

$$\rightarrow \frac{1}{R_p} = \frac{r_c^2}{r_c \omega^2 L^2} \qquad \because Q_c = \frac{\omega L}{r_c}$$

$$\rightarrow \frac{1}{R_p} = \frac{1}{r_c Q_c^2} \qquad \because \omega L = Q_c r_c$$

$$\rightarrow R_p = r_c Q_c^2 = \omega L Q_c \rightarrow 2$$

From the equivalent circuit,

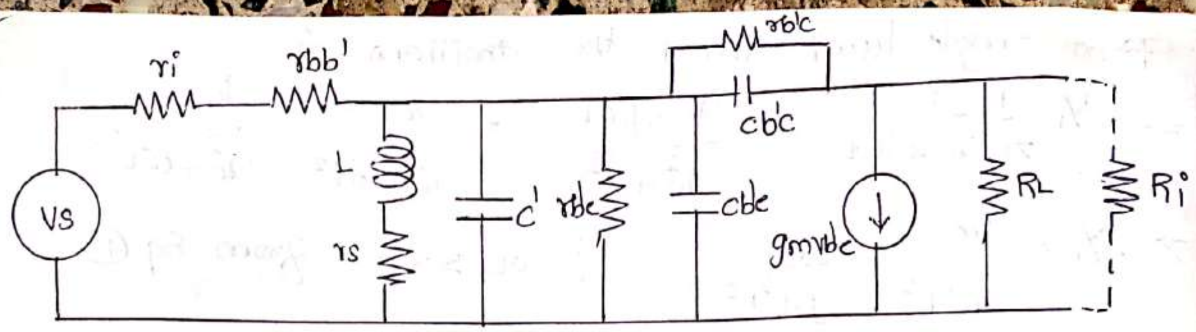
$$\rightarrow R = r_i \,||\, R_p \,||\, r_{b'e} \rightarrow 3$$

The current gain of the amplifier is then.

$$\rightarrow A_i = \frac{-g_m R}{1 + j(\omega R C - R/\omega L)} = \frac{-g_m R}{1 + j\omega_0 R C(\omega/\omega_0 - \omega_0/\omega)} \rightarrow 4$$

where $\omega_0^2 = \frac{1}{LC}$

$$\rightarrow A_i = \frac{-g_m R}{1 + j\frac{1}{\sqrt{LC}} \cdot RC\left[\frac{\omega}{\frac{1}{\sqrt{LC}}} - \frac{\frac{1}{\sqrt{LC}}}{\omega}\right]} = \frac{-g_m R}{1 + j\frac{R\sqrt{C}}{\sqrt{L}}\left[\omega\sqrt{LC} - \frac{1}{\omega\sqrt{LC}}\right]}$$



Hybrid π model.

$$\rightarrow A_i = \frac{-g_m R}{1+j\frac{R\sqrt{C}}{\sqrt{L}}\left[\frac{\omega^2 LC - 1}{\omega\sqrt{LC}}\right]} = \frac{-g_m R}{1+j\left[\frac{R(\omega^2 LC-1)}{\omega L}\right]}$$

$$\rightarrow A_i = \frac{-g_m R}{1+j\left[\frac{R\omega^2 LC - R}{\omega L}\right]} = \frac{-g_m R}{1+j\left[R\omega C - \frac{R}{\omega L}\right]}$$

Now,

Quality factor of tuned circuit at resonant frequency ω_0.

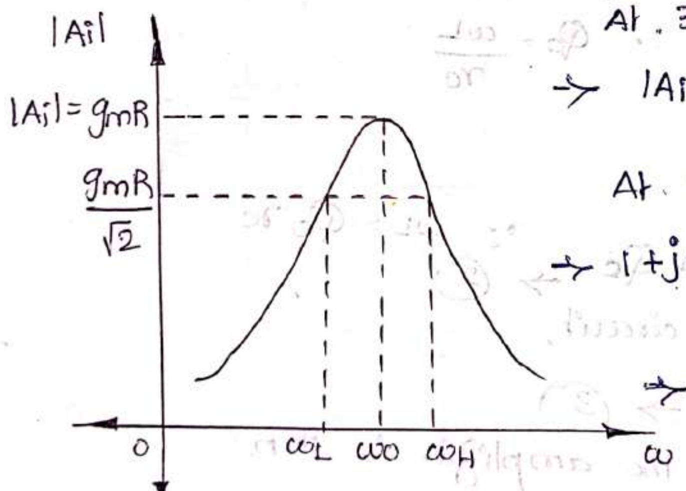
$$\rightarrow Q = \frac{R}{\omega_0 L} = \omega_0 RC \rightarrow (5)$$

Sub (5) in (4)

$$\rightarrow A_i = \frac{-g_m R}{1+jQ(\omega/\omega_0 - \omega_0/\omega)}$$

At $\omega = \omega_0$, gain is max and it is given as

$$\rightarrow A_i = -g_m R \rightarrow (6)$$



Gain Vs Frequency for single tuned Amp.

At 3 dB frequency,

$$\rightarrow |A_i| = \frac{g_m R}{\sqrt{2}} \rightarrow (7)$$

At 3 dB frequency,

$$\rightarrow 1+jQ[\omega/\omega_0 - \omega_0/\omega] = \sqrt{2}$$

Take Square

$$\rightarrow 1 + Q^2\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]^2 = 2 \rightarrow (8)$$

From the response,

$$\rightarrow BW = f_H - f_L = \frac{f_r}{Q_L} = \frac{\omega_0}{2\pi Q} = \frac{\omega_0}{2\pi \omega_0 RC}$$

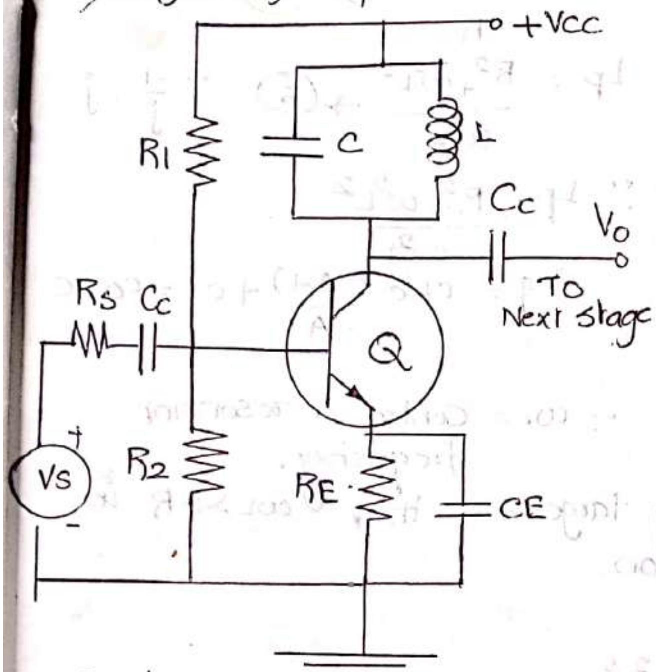
$$\rightarrow BW = \frac{1}{2\pi RC} \rightarrow (9)$$

$\because Q = \omega_0 RC$ in (5)

$\because \omega = 2\pi f_r$

$f_r = \frac{\omega}{2\pi}$

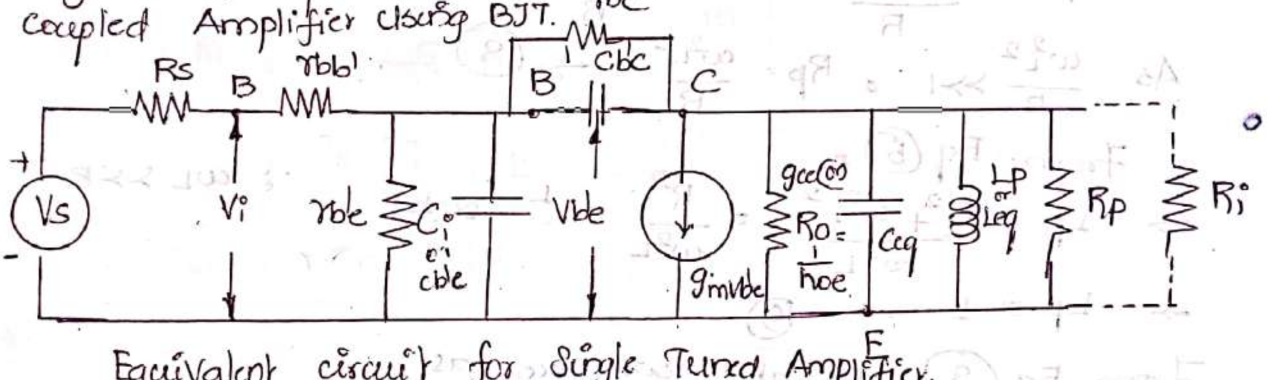
Analysis of capacitive coupled single tuned Amplifier:



The dia. shows single tuned CE amplifier.

Single tuned amplifier circuit uses one parallel tuned circuit as a load in each stage with tuned circuits in all stage tuned to the same frequency.

Single Tuned Capacitive coupled Amplifier using BJT.



Equivalent circuit for Single Tuned Amplifier.

From Eq. (15),

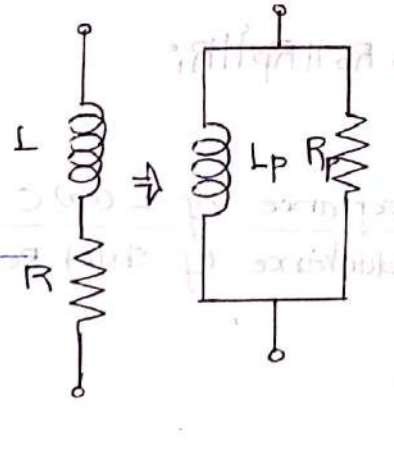
$\Rightarrow C_i$ = Input capacitance $= C_{b'e} + C_{b'c}(1-A)$ → (1)

&

$\Rightarrow C_{eq}$ = output capacitance $= C_{b'c}\left(\frac{A-1}{A}\right) + C$ → (2)

&

$\Rightarrow g_{ce} = \frac{1}{r_{ce}} = h_{oe} - g_m h_{re} \simeq h_{oe} = \frac{1}{R_o}$ → (3)



Admittance of series of RL is

$\Rightarrow Y = \frac{1}{Z} = \frac{1}{R + j\omega L}$

$\Rightarrow Y = \frac{1}{R + j\omega L} \times \frac{R - j\omega L}{R - j\omega L} = \frac{R - j\omega L}{R^2 + \omega^2 L^2}$

$\Rightarrow Y = \frac{R - j\omega L}{R^2 + \omega^2 L^2} = \frac{R}{R^2 + \omega^2 L^2} - \frac{j\omega L}{R^2 + \omega^2 L^2}$

$\Rightarrow Y = \frac{R}{R^2 + \omega^2 L^2} - \frac{j\omega L}{R^2 + \omega^2 L^2} \times \frac{\omega}{\omega}$

$\Rightarrow Y = \frac{R}{R^2 + \omega^2 L^2} - \frac{j\omega^2 L}{\omega (R^2 + \omega^2 L^2)}$

$$\Rightarrow Y = \frac{1}{R_p} + \frac{1}{j\omega L_p} \quad \text{where } R_p = \frac{R^2+\omega^2L^2}{R} \rightarrow (4)$$

$$L_p = \frac{R^2+\omega^2L^2}{\omega^2 L} \rightarrow (5) \quad \because \frac{-1}{j} = j$$

1. Centre Frequency:

It is given as

$$\rightarrow f_r = \frac{1}{2\pi\sqrt{L_p C_{eq}}} \rightarrow (6) \quad \because L_p = \frac{R^2+\omega^2L^2}{\omega^2 L},$$

$$C_{eq} = C_{b'c}\left(\frac{A-1}{A}\right) + C = C_o + C$$

2. Quality Factor Q:

It is given by

$$\rightarrow Q_r = \frac{\omega_r L}{R} \rightarrow (7) \quad \because \omega_r = \text{Centre or resonant frequency.}$$

The Q of the coil is usually large so that $\omega L >> R$ in the frequency range of operation.

From Eq (4),

$$\Rightarrow R_p = \frac{R^2+\omega^2L^2}{R} = R + \frac{\omega^2L^2}{R}$$

$$\text{As } \frac{\omega^2L^2}{R} >> 1, \quad R_p = \frac{\omega^2L^2}{R} \rightarrow (8)$$

From Eq (5),

$$\rightarrow L_p = \frac{R^2+\omega^2L^2}{\omega^2 L} = \frac{R^2}{\omega^2 L} + L \simeq L \quad \because \omega L >> R$$

$$\rightarrow L_p \simeq L \rightarrow (9)$$

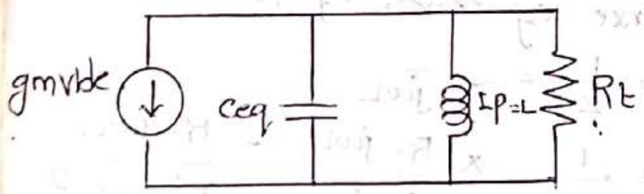
From Eq (8), R_p at resonance as

$$\rightarrow R_p = \frac{\omega_r^2 L^2}{R} = \omega_r Q_r L \quad \because Q_r = \frac{\omega_r L}{R} \rightarrow (10)$$

From (10),

$$Q_r = \frac{R_p}{\omega_r L} \rightarrow (11)$$

The effective quality factor including load can be calculated at the Equivalent circuit



$$\because R_t = R_o \| R_p \| R_i.$$

$$\rightarrow \text{Effective Quality factor, } Q_{eff} = \frac{\text{susceptance of } L \text{ (or) } C}{\text{Conductance of shunt Res, } R_t}$$

$$\rightarrow Q_{eff} = \frac{R_t}{\omega_r L} \text{ (or) } \omega_r C_{eq} R_t.$$

3. Voltage Gain :

It is given by

$\rightarrow A_V = \frac{V_o}{V_i}$. where $V_o = -g_m V_{b'e} (\underbrace{C_{eq} \| L_p \| R_t})$

$\because Z = C_{eq} \| R_t \| L_p$.

$\because V_o = -g_m V_{b'e} Z \rightarrow (12)$

Now, $Y = \frac{1}{Z} = \frac{1}{R_t} + \frac{1}{j\omega L} + j\omega C_{eq}$ $\quad \because \omega L >> RC$

$\rightarrow Y = \frac{1}{R_t}\left[1 + \frac{R_t}{j\omega L} + j\omega C_{eq} R_t\right]$

(x) N & D by ω_r,

$\rightarrow Y = \frac{1}{\omega_r R_t}\left[1 + \frac{\omega_r R_t}{j\omega L \omega_r} + j\omega \frac{C_{eq} R_t \omega_r}{\omega_r}\right]$ $\quad \because Q_{eff} = \frac{R_t}{\omega_r L}$
$= \omega_r C_{eq} R_t$.

$\rightarrow Y = \frac{1}{R_t}\left[1 + \frac{Q_{eff}\,\omega_r}{j\omega} + \frac{Q_{eff}\, j\omega}{\omega_r}\right]$

$\rightarrow Y = \frac{1}{R_t}\left[1 + jQ_{eff}\left[-\frac{\omega_r}{\omega} + \frac{\omega}{\omega_r}\right]\right]$ $\quad \because \frac{-1}{j} = j$

$\rightarrow Y = \frac{1}{R_t}\left[1 + jQ_{eff}\left(\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right)\right]$

$\rightarrow Z = \frac{1}{R_t} = \frac{R_t}{1 + jQ_{eff}\left(\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right)} \rightarrow (13)$

Let δ be the fractional variable in f_r,

$\rightarrow \delta = \frac{\omega - \omega_r}{\omega_r} = \frac{\omega}{\omega_r} - 1$

$\rightarrow \frac{\omega}{\omega_r} = \delta + 1 \rightarrow (14)$

Sub (14) in (13)

$\rightarrow Z = \frac{R_t}{1 + jQ_{eff}\left[\frac{\delta+1}{1} - \frac{1}{\delta+1}\right]} = \frac{R_t}{1 + jQ_{eff}\left[\frac{(\delta+1)^2 - 1}{\delta+1}\right]}$

$\rightarrow Z = \frac{R_t}{1 + jQ_{eff}\left[\frac{\delta^2 + 2\delta + \cancel{1} - \cancel{1}}{\delta+1}\right]} = \frac{R_t}{1 + jQ_{eff}\, 2\delta \left(\frac{\delta/2 + 1}{\delta+1}\right)} \rightarrow (15)$

$\because \delta << 1$, the (15) Eqn. becomes,

$\rightarrow Z = \frac{R_t}{1 + j2\delta Q_{eff}} \rightarrow (16)$

At resonance, $\omega = \omega_r$, the Eqn (14) becomes,

$\rightarrow \frac{\omega}{\omega_r} = \delta + 1 \rightarrow \frac{\omega_r}{\omega_r} = \delta + 1 \rightarrow 1 = \delta + 1 \Rightarrow \delta = 0.$

Sub $\delta = 0$ in (16)

$\rightarrow Z = \frac{R_t}{1 + j2(0)Q_{eff}} = R_t$

$\rightarrow$ Sub $Z = R_t$ in (12) $\quad \because V_{b'e} = \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \cdot V_i$

$\rightarrow V_o = -g_m V_{b'e} R_t$

sub (17) in Vo and find V_o/V_i, $\rightarrow$ (17)

$\rightarrow \frac{V_o}{V_i} = \frac{-g_m R_t r_{b'e}}{r_{bb'} + r_{b'e}} = A_{v\,(Resonance)} \rightarrow$ (18)

If it is not at Resonance,

wkt, $V_o = -g_m V_{b'e} Z \quad \because V_{b'e} = \frac{r_{b'e}}{r_{bb'} + r_{b'e}} V_i$

$\rightarrow \frac{V_o}{V_i} = -g_m Z \frac{r_{b'e}}{r_{bb'} + r_{b'e}}$

sub (16) in V_o/V_i

$\rightarrow \frac{V_o}{V_i} = -g_m \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \cdot \frac{R_t}{1 + j2\delta Q_{eff}} \rightarrow$ (19) $= A_v.$

Taking ratio of (18) and (19),

$\Rightarrow \frac{A_v}{A_{v(res)}} = \frac{1}{1 + 2j\delta Q_{eff}}$

$\rightarrow \left| \frac{A_v}{A_{v(res)}} \right| = \frac{1}{\sqrt{1 + (2\delta Q_{eff})^2}} \rightarrow$ (20)

$\rightarrow$ Phase angle, $\phi = -\tan^{-1}(2\delta Q_{eff}) \rightarrow$ (21)

4. 3dB Bandwidth:

At f_L, transistor gain drops by 3dB, we have,

$\rightarrow \left| \frac{A_v}{A_{v(res)}} \right| = \frac{1}{\sqrt{1 + (2\delta Q_{eff})^2}} = \frac{1}{\sqrt{2}}$

$\rightarrow 1 + (2\delta Q_{eff})^2 = 2$

$\rightarrow (2\delta Q_{eff})^2 = 1^2 \Rightarrow 2\delta Q_{eff} = 1$

$\rightarrow \delta = \frac{1}{2Q_{eff}} = f_L$ and also f_H. $\rightarrow$ (22)

The $\Delta\omega$ can be given as

$\rightarrow \Delta\omega = (\omega_2 - \omega_r) + (\omega_r - \omega_1) \rightarrow (23)$

(x) N & D by ω_r,

$\rightarrow \Delta\omega = \left[\frac{(\omega_2 - \omega_r) + (\omega_r - \omega_1)}{\omega_r}\right]\omega_r = \left[\frac{\omega_2 - \omega_r}{\omega_r} + \frac{\omega_r - \omega_1}{\omega_r}\right]\omega_r \rightarrow (24)$

wkt, $\delta = \frac{\omega - \omega_r}{\omega_r}$.

$\rightarrow \Delta\omega = [\delta + \delta]\omega_r$ at $\omega_2 = \omega_1 = \omega$.

$\rightarrow \Delta\omega = 2\delta\omega_r$. $\because \delta = \frac{1}{2Q_{eff}}$.

$\rightarrow \Delta\omega = \frac{2\omega_r}{2Q_{eff}} = \frac{\omega_r}{Q_{eff}}$ $\because Q_{eff} = \frac{R_t}{\omega_r L} = R_t \omega_r C_{eq}$.

$\rightarrow \Delta\omega = \frac{\omega_r}{\frac{R_t}{\omega_r L}} = \frac{\omega_r^2 L}{R_t} \rightarrow (25)$ & $\Delta\omega = \frac{\omega_r}{R_t \omega_r C_{eq}} = \frac{1}{R_t C_{eq}} \rightarrow (26)$

wkt, $\Delta\omega = 2\pi\Delta f$ $\because (\omega = 2\pi f)$

$\rightarrow \Delta f = \frac{\Delta\omega}{2\pi}$

From (25) From (26)

$\rightarrow \Delta f = \frac{\omega_r^2 L}{R_t 2\pi} \rightarrow (27)$ $\rightarrow \Delta f = \frac{1}{2\pi R_t C_{eq}} \rightarrow (28)$

From (27).

$\rightarrow \Delta f = \frac{\omega_r^2 L}{R_t 2\pi} = \frac{\omega_r}{2\pi} \cdot \frac{\omega_r L}{R_t} = \frac{\omega_r}{2\pi Q_{eff}}$ $\because Q_{eff} = \frac{R_t}{\omega_r L}$

$\rightarrow \Delta f = \frac{2\pi f_r}{2\pi Q_{eff}} = \frac{f_r}{Q_{eff}}$

$\rightarrow \Delta f = \frac{f_r}{Q_{eff}} \rightarrow (29)$

Advantages:

1. They amplify defined frequencies.
2. signal to noise ratio at output is good.
3. Well suited for radio transmission and receiver

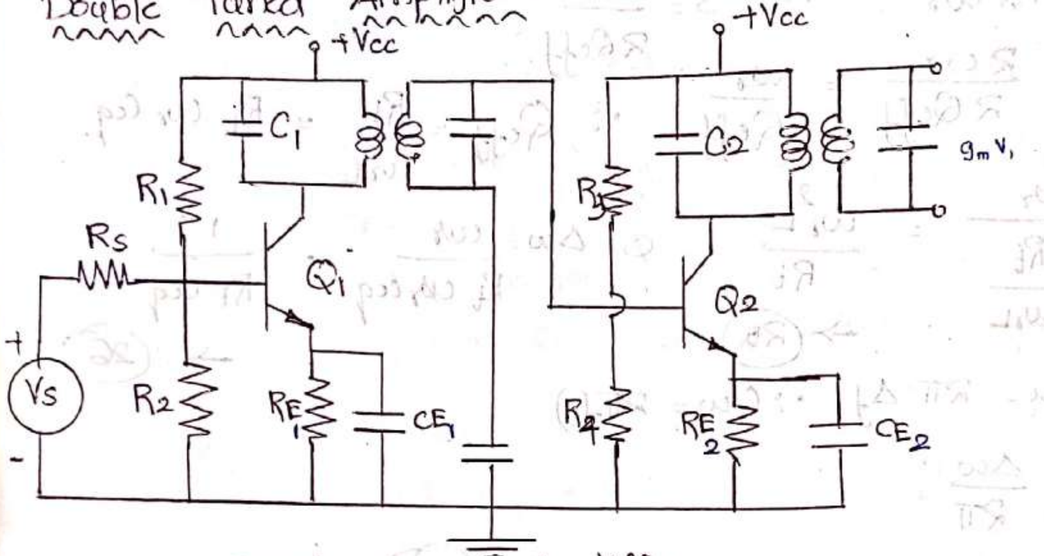
Disadvantages:

1. Because of Inductor and capacitor, circuit become bulky and costly.
2. Not suitable to amplify audio frequency.
3. If frequency is increased, design becomes complex.

Applications:

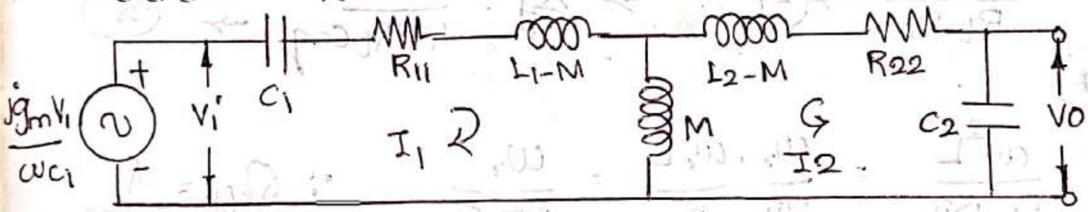
1. Used in active filters like Low pass, High pass and band pass to allow amplify a signal only in desired narrow band.
2. Used in radio receiver
3. Tuned class B and class C are used an output RF amplifier in radio transmitters to increase the output efficiency and to reduce the harmonics.

Double Tuned Amplifier:



Double Tuned Amplifier.

EQuivalent circuit:

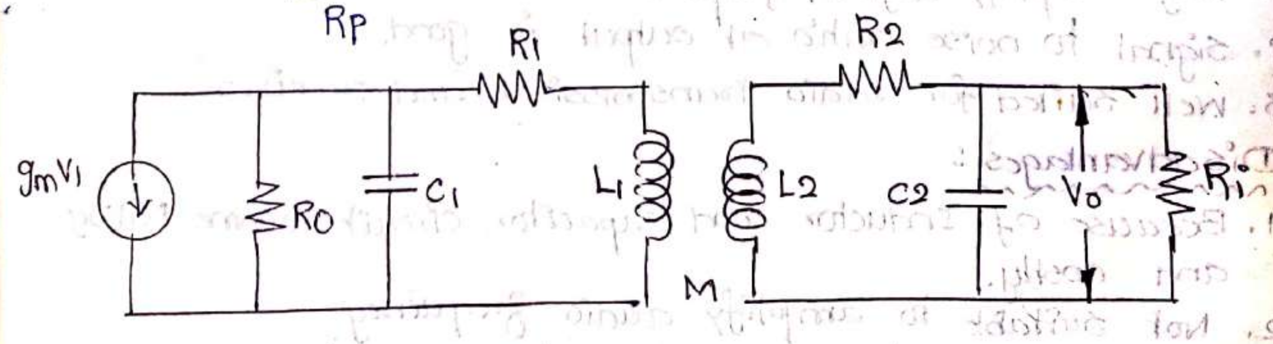


The dia. shows double Tuned RF amplifier in CE configuration. Here Voltage developed across tuned circuits is coupled inductively to another tuned circuit.

In the circuit,

$\rightarrow R_p = \frac{\omega^2 L^2}{R_p} = \frac{\omega^2 L^2}{R}$; R represents Series resistance, Rp represents parallel resistance

& $R = \frac{\omega^2 L^2}{R_p}$



From Eq. ckt,

$$R_{11} = \frac{\omega_0^2 L_1^2}{R_0} + R_1 \quad \& \quad R_{22} = \frac{\omega_0^2 L_2^2}{R_i} + R_2$$

wkt,

$$Q = \frac{\omega_r L}{R} \rightarrow Q_1 = \frac{\omega_r L_1}{R_{11}} \quad \text{and} \quad Q_2 = \frac{\omega_r L_2}{R_{22}}$$

But Q-factor of Both tuned circuit are same,

$$\rightarrow Q = Q_1 = Q_2$$

$$\rightarrow \frac{\omega_r L_1}{R_{11}} = \frac{\omega_r L_2}{R_{22}} \; ; \; \frac{L_1}{R_{11}} = \frac{L_2}{R_{22}}.$$

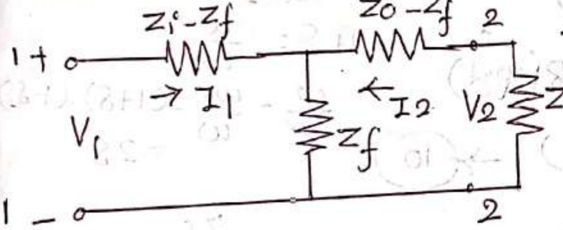
When capacitance in Effect,

$$\rightarrow \frac{\omega_r L_1}{C_1} = \frac{\omega_r L_2}{C_2} \; ; \; \frac{L_1}{C_1} = \frac{L_2}{C_2}.$$

From Sim. Eq. CKt,

$$\rightarrow V_0 = \frac{-j}{\omega_r C_2} \cdot I_2 \rightarrow (1)$$

Let us consider the circuit, Find transfer admittance, Y_t.



$$\rightarrow Y_T = \frac{I_2}{V_1} = \frac{I_2}{I_1 Z_{11}} = \frac{A_i}{Z_{11}} \rightarrow (2)$$

$\because V_1 = I_1 Z_{11}$

$A_i = \frac{I_2}{I_1}$

$$\therefore Z_{11} = \frac{V_1}{I_1}$$

$$\therefore V_1 = Z_i I_1 + \frac{Z_f^2}{Z_0 + Z_L} I_2 \qquad \because I_2 = -I_1$$

$$V_1 = Z_i I_1 - \frac{Z_f^2}{Z_0 + Z_L} I_1.$$

$$\rightarrow Z_{11} = \frac{V_1}{I_1} = Z_i - \frac{Z_f^2}{Z_0 + Z_L} \rightarrow (3)$$

and

$$\rightarrow A_i = \frac{I_2}{I_1} = -\frac{Z_f}{Z_0 + Z_L} \rightarrow (4)$$

and

$$\rightarrow Z_f = j\omega_r M \rightarrow (5)$$

$$\rightarrow Z_i = R_{11} + j\omega L_1 + \frac{1}{j\omega C_1} \qquad \because \frac{1}{j} = -j$$

$$Z_i = R_{11} + j\omega L_1 - \frac{j}{\omega C_1}$$

and

$$\rightarrow Z_0 + Z_L = R_{22} + j\omega L_2 + \frac{1}{j\omega C_2} \qquad Z_i = R_{11} + j\left(\omega L_1 - \frac{1}{\omega C_1}\right) \rightarrow (6)$$

$$\rightarrow Z_0 + Z_L = R_{22} + j\left(\omega L_2 - \frac{1}{\omega C_2}\right) \rightarrow (7)$$

The Eqn for Z_f, Z_i, $Z_0 + Z_L$ can be further simplified as

$$\rightarrow \underline{Z_f}: \quad Z_f = j\omega_r M = j\omega_r K\sqrt{L_1 L_2} \rightarrow (8)$$

$\because M = K\sqrt{L_1 L_2}$

K = coupling co-eff.

and

$$\rightarrow Z_{11} = R_{11} + j\left(\omega L_1 - \frac{1}{\omega C_1}\right)$$

× and ÷ both N & D by $\omega_r L_1$

$$\rightarrow Z_i = \frac{R_{11}\,\omega_r L_1}{\omega_r L_1} + j\omega_r L_1\left(\frac{\omega L_1}{\omega_r L_1} - \frac{1}{\omega C_1 \omega_r L_1}\right)$$

$$\rightarrow Z_i = \frac{\omega_r L_1}{Q} + j\omega_r L_1\left\{\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right\} \quad \because Q = \frac{\omega_r L}{R_1} = \frac{\omega_r L_1}{R_{11}}$$

$$\& \ \frac{1}{\omega_r L_1} = \omega_r C_1$$

$$\rightarrow Z_i = \frac{\omega_r L_1}{Q} + j\omega_r L_1 (2\delta) \quad \because \delta = \frac{\omega - \omega_r}{\omega_r} = \frac{\omega}{\omega_r} - 1$$

$$\because j\omega_r L_1 2\delta = 1 + 2jQ\delta. \qquad \delta + 1 = \frac{\omega}{\omega_r}$$

$$\rightarrow Z_i = \frac{\omega_r L_1}{Q} * (1 + j2Q\delta) \rightarrow (9) \quad \because (\delta+1) - (\delta-1)$$

$$(1+\delta) - (1-\delta) = 2\delta.$$

$Z_0 + Z_L$:

$$\rightarrow Z_0 + Z_L = R_{22} + j\left(\omega L_2 - \frac{1}{\omega C_2}\right)$$

(x) and ($\div$) N & D by $\omega_r L_2$

$$\rightarrow Z_0 + Z_L = \frac{R_{22}\,\omega_r L_2}{\omega_r L_2} + j\omega_r L_2\left(\frac{\omega L_2}{\omega_r L_2} - \frac{1}{\omega C_2 \omega_r L_2}\right)$$

$$\rightarrow Z_0 + Z_L = \frac{\omega_r L_2}{Q} + j\omega_r L_2\left(\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right) \quad \because Q = \frac{\omega_r L}{R} = \frac{\omega_r L_2}{R_{22}}$$

$$\& \ \frac{1}{\omega_r L_2} = \omega_r C_2$$

$$\rightarrow Z_0 + Z_L = \frac{\omega_r L_2}{Q} + j\omega_r L_2 (2\delta) \qquad \& \ 1 + \delta = \frac{\omega}{\omega_r}$$

$$\because j\omega_r L_2 2\delta = (1 + 2jQ\delta) \qquad \frac{\omega}{\omega_r} - \frac{\omega_r}{\omega} = (1+\delta) - (1-\delta) = 2\delta.$$

$$\rightarrow Z_0 + Z_L = \frac{\omega_r L_2}{Q} * (1 + j2Q\delta) \rightarrow (10)$$

$$\text{Then, } Y_T = \frac{A_i}{Z_{11}} = \frac{-Z_f}{Z_0 + Z_L} \times \frac{Z_0 + Z_L}{Z_i(Z_0 + Z_L) - Z_f^2} = \frac{Z_f}{Z_f^2 - Z_i(Z_0 + Z_L)}$$

$$\rightarrow Y_T = \frac{1}{Z_f - \frac{Z_i}{Z_f}(Z_0 + Z_L)} = \frac{1}{j\omega_r K\sqrt{L_1 L_2} - \left[\frac{\frac{\omega_r L_1}{Q} * (1 + j2Q\delta)\left[\frac{\omega_r L_2}{Q} * (1 + j2Q\delta)\right]}{j\omega_r K\sqrt{L_1 L_2}}\right]}$$

$$\rightarrow Y_T = \frac{1}{\frac{(j\omega_r K\sqrt{L_1 L_2})^2 - \left[\frac{\omega_r L_1}{Q} + (1 + j2Q\delta)\right]\left[\frac{\omega_r L_2}{Q} + (1 + j2Q\delta)\right]}{j\omega_r K\sqrt{L_1 L_2}}}$$

$$\rightarrow Y_T = \frac{1}{\frac{(j\omega_r K\sqrt{L_1 L_2})^2 - \frac{\omega_r^2 L_1 L_2}{Q^2} - \frac{\omega_r L_1}{Q}[1 + j2Q\delta] + \frac{\omega_r L_2}{Q}(1 + j2Q\delta) + (1 + j2Q\delta)^2}{j\omega_r K\sqrt{L_1 L_2}}}$$

$$\rightarrow Y_T = \frac{1}{\frac{(j\omega_r K\sqrt{L_1 L_2})^2 - \frac{\omega_r^2 L_1 L_2}{Q^2} - \frac{\omega_r L_1}{Q} - \frac{\omega_r L_1 j2\cancel{Q}\delta}{\cancel{Q}} + \frac{\omega_r L_2}{Q} + \frac{\omega_r L_2 j2\cancel{Q}\delta}{\cancel{Q}} + (1 + j2Q\delta)^2}{j\omega_r K\sqrt{L_1 L_2}}}$$

$$\rightarrow Y_T = \frac{1}{\dfrac{j\omega_r K\sqrt{L_1L_2} - \left[\dfrac{\omega_r L_1}{Q}(1+j2Q\delta)\left[\dfrac{\omega_r L_2}{Q}(1+j2Q\delta)\right]\right]}{j\omega_r K\sqrt{L_1L_2}}}$$

$$\rightarrow Y_T = \frac{j\omega_r K\sqrt{L_1L_2}}{-\omega_r^2K^2L_1L_2 - \left[\dfrac{\omega_r L_1}{Q}(1+j2Q\delta)\left[\dfrac{\omega_r L_2}{Q}(1+j2Q\delta)\right]\right]}$$

$$\rightarrow Y_T = \frac{j\omega_r K\sqrt{L_1L_2}}{-\omega_r^2K^2L_1L_2 - \left[\dfrac{\omega_r L_1}{Q} + \dfrac{j2Q\delta\omega_r L_1}{Q}\left\{\dfrac{\omega_r L_2}{Q} + \dfrac{j2Q\delta\omega_r L_2}{Q}\right\}\right]}$$

$$\rightarrow Y_T = \frac{j\omega_r K\sqrt{L_1L_2}}{-\omega_r^2K^2L_1L_2 - \left[\dfrac{\omega_r^2L_1L_2}{Q^2} + \dfrac{j2Q\delta\omega_r^2L_1L_2}{Q^2} + \dfrac{j2Q\delta\omega_r^2L_1L_2}{Q^2} - \dfrac{4Q^2\delta^2\omega_r^2L_1L_2}{Q^2}\right]}$$

$$\rightarrow Y_T = \frac{+j\omega_r K\sqrt{L_1L_2}}{-\omega_r^2K^2L_1L_2 - \left[\dfrac{\omega_r^2L_1L_2}{Q^2} + \dfrac{4jQ\delta\omega_r^2L_1L_2}{Q^2} - \dfrac{4Q^2\delta^2\omega_r^2L_1L_2}{Q^2}\right]}$$

$$\rightarrow Y_T = \frac{j\cancel{\omega_r}K\cancel{\sqrt{L_1L_2}}Q^2}{-\omega_r^{\cancel{2}}K^2Q^2\cancel{L_1L_2} - \omega_r^{\cancel{2}}\cancel{L_1L_2} - 4jQ\delta\omega_r^{\cancel{2}}\cancel{L_1L_2} - 4Q^2\delta^2\omega_r^{\cancel{2}}\cancel{L_1L_2}}$$

$$\rightarrow Y_T = \frac{KQ^2}{j\omega_r K^2Q^2\sqrt{L_1L_2} + j\omega_r\sqrt{L_1L_2} + 4j^2Q\delta\omega_r\sqrt{L_1L_2} - j4Q^2\delta^2\omega_r\sqrt{L_1L_2}}$$

$$\rightarrow Y_T = \frac{KQ^2}{\omega_r\sqrt{L_1L_2}\left[jK^2Q^2 + j + 4j^2Q\delta - j4Q^2\delta^2\right]}$$

$$\rightarrow Y_T = \frac{KQ^2}{\omega_r\sqrt{L_1L_2}\left[jK^2Q^2 + j - 4Q\delta - j4Q^2\delta^2\right]}$$

$$\rightarrow Y_T = \frac{KQ^2}{\omega_r\sqrt{L_1L_2}\left[-4Q\delta + j\left[1 + K^2Q^2 - 4Q^2\delta^2\right]\right]}$$

$$\rightarrow Y_T = \frac{KQ^2}{+\omega_r\sqrt{L_1L_2}\left[4Q\delta - j\left[1 + K^2Q^2 - 4Q^2\delta^2\right]\right]} \rightarrow (11)$$

Now, w.r.t, $Y_T = \dfrac{I_2}{V_i}$

$\rightarrow I_2 = V_i\, Y_T$

From the Eq. circuit,

$\rightarrow V_i = \dfrac{jg_m V_1}{\omega C_1}$

$$\rightarrow I_2 = V_i\, Y_T = \frac{jg_m V_1}{\omega C_1} \times \frac{KQ^2}{\omega_r\sqrt{L_1L_2}\left[4Q\delta - j\left[1 + K^2Q^2 - 4Q^2\delta^2\right]\right]}$$

Sub I_2 in (1)

$\rightarrow V_0 = \dfrac{-j}{\omega_r C_2}\, I_2$

$$\rightarrow V_0 = \frac{-j}{\omega_r C_2} \cdot \frac{j g_m V_1}{\omega_r C_1} \times \frac{KQ^2}{\omega_r\sqrt{L_1L_2}\,[4Q\delta - j[1+K^2Q^2 - 4Q^2\delta^2]]}$$

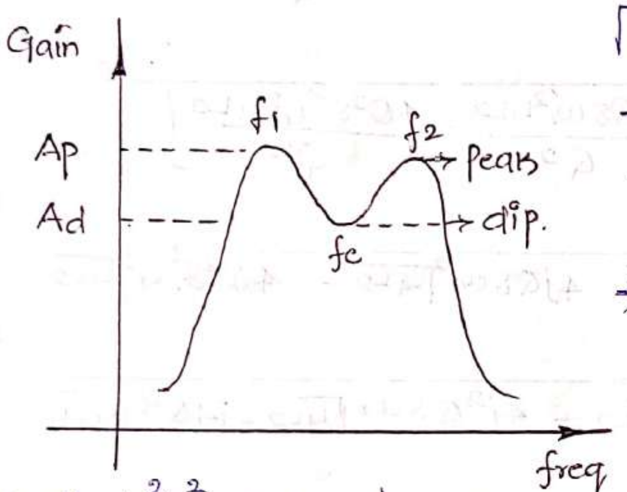
$$\rightarrow \frac{V_0}{V_1} = \frac{g_m}{\omega_r^2 C_1 C_2}\left[\frac{KQ^2}{\omega_r\sqrt{L_1L_2}\,[4Q\delta - j[1+K^2Q^2 - 4Q^2\delta^2]]}\right]$$

$$\rightarrow A_V = g_m\, \omega_r^{\cancel{2}} L_1 L_2 \left[\frac{KQ^2}{\cancel{\omega_r}\sqrt{L_1L_2}\,[4Q\delta - j[1+K^2Q^2 - 4Q^2\delta^2]]}\right] \quad \because \frac{1}{\omega_r C_1} = \omega_r L_1,\ \because \frac{1}{\omega_r C_2} = \omega_r L_2$$

$$\rightarrow A_V = g_m\, \omega_r \sqrt{L_1L_2}\, Q \left[\frac{KQ}{[4Q\delta - j[1+K^2Q^2 - 4Q^2\delta^2)]}\right] \rightarrow (12)$$

Taking magnitude, $A+jB = |\sqrt{A^2+B^2}|$

$$\rightarrow |A_V| = g_m \omega_r \sqrt{L_1L_2}\, Q \frac{KQ}{\sqrt{(4Q\delta)^2 - j(1+K^2Q^2 - 4Q^2\delta^2)^2}} \rightarrow (13)$$



Two Gain peaks of frequency response in double Tuned Amplifier is

$\rightarrow f_1 = f_r\left(1 - \frac{1}{2Q}\sqrt{K^2Q^2-1}\right)$ and

$\rightarrow f_2 = f_r\left(1 + \frac{1}{2Q}\sqrt{K^2Q^2-1}\right) \rightarrow (14)$

1. At $K^2Q^2 = 1$, i.e. $K = \frac{1}{Q}$, $f_1 = f_2 = f_r$, This condition is critical coupling.
2. At $K < 1/Q$, Peak gain is less than Maximum gain and coupling is poor.
3. At $K > \frac{1}{Q}$, circuit is over coupled and response shows double peak.

The Gain Magnitude at Peak is given as,

$$\Rightarrow |A_p| = \frac{g_m \cdot \omega_r \sqrt{L_1L_2}\, KQ}{2} \rightarrow (15)$$

The Gain at the dip at $\delta = 0$ is

$$\rightarrow |A_d| = |A_p| \frac{2KQ}{1+K^2Q^2} \rightarrow (16)$$

The ratio of peak gain and dip gain is denoted as γ.

$$\Rightarrow \gamma = \left|\frac{A_p}{A_d}\right| = \frac{1+K^2Q^2}{2KQ}.$$

$$\rightarrow \gamma = \frac{1+K^2Q^2}{2KQ}$$

$$\rightarrow 2KQ\gamma = 1+K^2Q^2 \rightarrow K^2Q^2 - 2KQ\gamma + 1 = 0.$$

This is of the form Quadratic Equation.

$$\rightarrow \frac{-(-2KQ\gamma) \pm \sqrt{4\gamma^2K^2Q^2 - 4(KQ)^2}}{2(KQ)^2} \qquad \because x = \frac{-b \pm \sqrt{b^2-4ac}}{2a}$$

$$\rightarrow KQ = \frac{\cancel{2}KQ\gamma \pm \cancel{2}K\cancel{Q}\sqrt{\gamma^2-1}}{\cancel{2}(KQ)^{\cancel{2}}} \rightarrow KQ = \gamma + \sqrt{\gamma^2-1} \rightarrow$$

Bandwidth is given as,

$$\rightarrow BW = \sqrt{2}\,(f_2 - f_1)$$

At 3dB Bandwidth, $\gamma = \sqrt{2}$.

$$\rightarrow KQ = \gamma + \sqrt{\gamma^2-1} = \sqrt{2} + \sqrt{\sqrt{2}^2-1} = 2.414.$$

3dB Bandwidth $= \sqrt{2}\,(f_2 - f_1)$

$$\rightarrow BW = \sqrt{2}\left[f_r\left(1+\frac{1}{2Q}\sqrt{K^2Q^2-1}\right) - f_r\left(1-\frac{1}{2Q}\sqrt{K^2Q^2-1}\right)\right]$$

$$\rightarrow BW = \sqrt{2}\left[\cancel{f_r} + \frac{f_r}{2Q}\sqrt{K^2Q^2-1} - \cancel{f_r} + \frac{f_r}{2Q}(\sqrt{K^2Q^2-1})\right]$$

$$\rightarrow BW = \sqrt{2}\,\frac{\cancel{2}f_r}{\cancel{2}Q}\sqrt{K^2Q^2-1} = \sqrt{2}\,\frac{f_r}{Q}\sqrt{K^2Q^2-1}$$

$$\rightarrow BW = \sqrt{2}\,\frac{f_r}{Q}\sqrt{(2.414)^2-1} = \frac{3.1 f_r}{Q} \qquad \because KQ = 2.414$$

Advantages:

1. Possesses a flatter response having steeper sides
2. provides larger 3 dB bandwidth
3. provides large gain-bandwidth product.

Effect of cascading single tuned and double tuned amplifiers on bandwidth:

1. In order to obtain a high over all gain, consider n stages of single tuned amplifier connected in cascade.

wkt,

$$\rightarrow \left|\frac{A_v}{A_{v(res)}}\right| = \frac{1}{\sqrt{1+(2\delta Q_{eff})^2}}$$

$$\rightarrow \left|\frac{A_v}{A_{v(res)}}\right|^n = \left[\frac{1}{\sqrt{1+(2\delta Q_{eff})^2}}\right]^n = \frac{1}{[1+(2\delta Q_{eff})^2]^{n/2}}$$

The 3dB frequencies for n stage cascade amp,

$$\rightarrow \left|\frac{A_v}{A_{v(res)}}\right|^n = \frac{1}{[1+(2\delta Q_{eff})^2]^{n/2}} = \frac{1}{\sqrt{2}}$$

$$\Rightarrow \frac{1}{[1+(2\delta Q_{eff})^2]^{n/2}} = \frac{1}{\sqrt{2}} \Rightarrow \frac{1}{[1+(2\delta Q_{eff})^2]^{n/2}} = \frac{1}{2^{1/2}}$$

$$\Rightarrow [1+(2\delta Q_{eff})^2]^n = 2 \Rightarrow [1+(2\delta Q_{eff})^2] = 2^{1/n}$$

$$\Rightarrow 2\delta Q_{eff} = \pm\sqrt{2^{1/n}-1} \quad \because \delta = \frac{\omega-\omega_r}{\omega_r} = \frac{f-f_r}{f_r}$$

$$\Rightarrow 2\left(\frac{f-f_r}{f_r}\right)Q_{eff} = \pm\sqrt{2^{\frac{1}{n}}-1}$$

$$\Rightarrow 2(f-f_r)\,Q_{eff} = \pm f_r\sqrt{2^{\frac{1}{n}}-1}$$

$$\Rightarrow f-f_r = \pm\frac{f_r}{2Q_{eff}}\sqrt{2^{1/n}-1}$$

Assume f_1 and f_2 are lower and upper 3dB freq.

$\Rightarrow$ we have

$$(f_2-f_r) = \pm\frac{f_r}{2Q_{eff}}\sqrt{2^{\frac{1}{n}}-1} \quad \& \quad (f_r-f_1) = \pm\frac{f_r}{2Q_{eff}}\sqrt{2^{\frac{1}{n}}-1}$$

The bandwidth of n stage single tuned amplifier is

$$\Rightarrow BW_n = f_2-f_1 = (f_2-f_r)+(f_r-f_1)$$

$$\Rightarrow BW_n = \frac{f_r}{2Q_{eff}}\sqrt{2^{\frac{1}{n}}-1} + \frac{f_r}{2Q_{eff}}\sqrt{2^{\frac{1}{n}}-1} = \frac{f_r}{Q_{eff}}\sqrt{2^{\frac{1}{n}}-1}$$

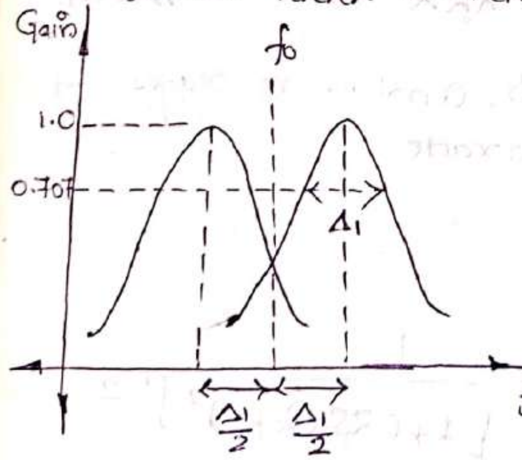
$$\Rightarrow BW_n = BW_1\sqrt{2^{\frac{1}{n}}-1}$$

2. In order to obtain overall bandwidth of the system is narrowed and steepness of sides of response is increased consider n stages double tuned amplifier.

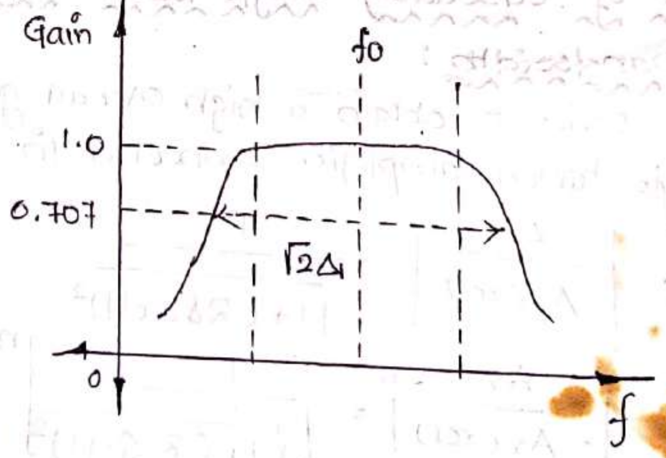
The 3dB bandwidth of n stages doubled tuned amplifier is $= \Delta_2 \times (2^{\frac{1}{n}}-1)^{1/4}$.

where Δ_2 = 3 dB BW of single stage.

Staggered Tuned Amplifier:



Response of Individual stages.



Overall Response of staggered pair

$\Rightarrow$ Removing the pair of cascading single tuned amplifier.

Gain of single tuned amplifier,

wkt, $\frac{A_v}{A_v(res)} = \frac{1}{1 + 2jQ_{eff}\delta} = \frac{1}{1+jx}$. $\because X = 2Q_{eff}\delta$

Assume the f_r of one stage is tuned to the frequency $f_r + \delta$ and other stage is tuned to frequency $f_r - \delta$,

$\rightarrow f_{r1} = f_r + \delta$ and $\rightarrow f_{r2} = f_r - \delta$.

Using these

$\rightarrow \frac{A_v}{A_v(res)_1} = \frac{1}{1+j(x+1)}$ and $\frac{A_v}{A_v(res)_2} = \frac{1}{1+j(x-1)}$

The overall gain of these two stages is product of individual gains of two stages.

$$\rightarrow \frac{A_v}{A_v(res)\,cas} = \frac{A_v}{A_v(res)_1} \times \frac{A_v}{A_v(res)_2} = \frac{1}{1+j(x+1)} \times \frac{1}{1+j(x-1)}$$

$$\Rightarrow \frac{A_v}{A_v(res)} = \frac{1}{(1+jx+j)(1+jx-j)} = \frac{1}{1+jx-\cancel{j}+jx-x^2+\cancel{x}+\cancel{j}-\cancel{x}+1}$$

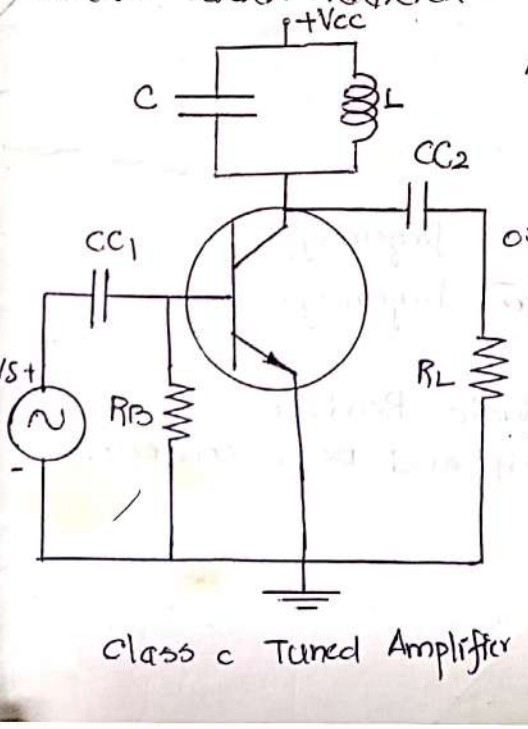
$$\Rightarrow \frac{A_v}{A_v(res)} = \frac{1}{2+2jx-x^2} = \frac{1}{(2-x^2)+2jx}$$

$$\rightarrow \left|\frac{A_v}{A_v(res)}\right| = \frac{1}{\sqrt{(2-x^2)^2+(2x)^2}} = \frac{1}{\sqrt{4-\cancel{4x^2}+x^4+\cancel{4x^2}}} = \frac{1}{\sqrt{4+x^4}}$$

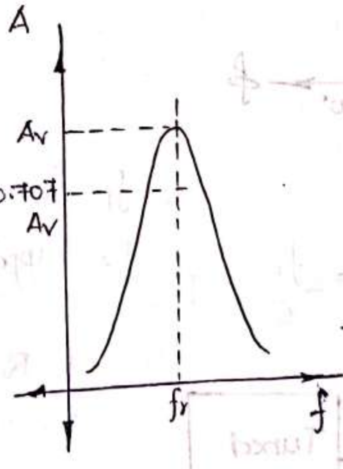
$$\Rightarrow \left|\frac{A_v}{A_v(res)}\right| = \frac{1}{\sqrt{4+(2Q_{eff}\delta)^4}} = \frac{1}{\sqrt{4+16Q_{eff}^4\delta^4}} = \frac{1}{2\sqrt{1+4Q_{eff}^4\delta^4}}$$

Large signal Tuned Amplifier:

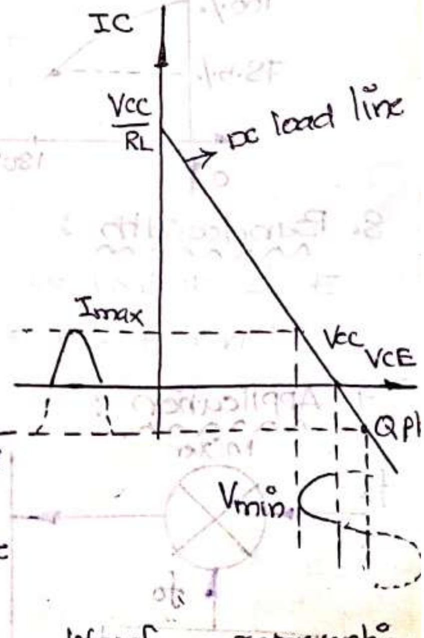
Class C Tuned Amplifier:



Class c Tuned Amplifier



Frequency Response



Waveform representing class c operation

1. Resonant Frequency :

The resonant frequency is given as

$$\rightarrow f_r = \frac{1}{2\pi\sqrt{LC}}$$

2. Power Gain :

The power gain is defined as

$$\rightarrow G = \frac{P_{out}}{P_{in}}$$

3. output power : (P_{AC})

It is given as

$$\rightarrow P_{out} = \frac{V_{rms}^2}{R_L}$$

$$\because V_{rms} = V_{PP}/2\sqrt{2}$$

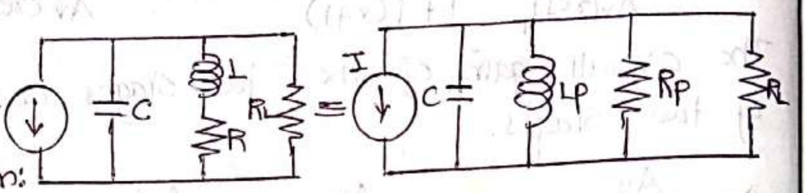
$$\rightarrow P_{out} = \frac{V_{PP}^2}{8R_L}$$

4. DC input power (P_{dc})

It is given as

$$P_{dc} = V_{cc}\, I_{dc}.$$

5. Ac collector resistance :

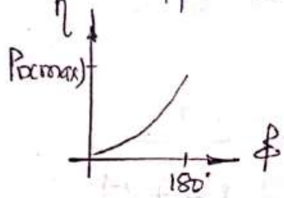


6. Transistor power Dissipation:

It is given as

$$\rightarrow P_{D(max)} = \frac{V_{PP(max)}^2}{40\, r_c}$$

where $V_{PP} = 2V_{cc}$.



The Q is defined as

$$\rightarrow Q_L = \frac{\omega_r L}{R} \text{ - In series.}$$

and

$$\rightarrow R_P = Q_L\, \omega_r L \text{ - In Parallel.}$$

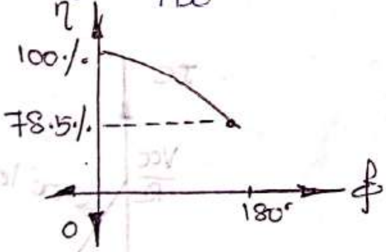
The overall of Q is given by

$$\rightarrow Q = \frac{r_c}{\omega_r L}$$

7. Efficiency :

It is defined as

$$\rightarrow \%\ \eta = \frac{P_{AC}}{P_{DC}} \times 100 = \frac{V_{PP}^2}{8R_L \times V_{CC} I_{DC}} \times 100 = 78.5\%$$



8. Bandwidth :

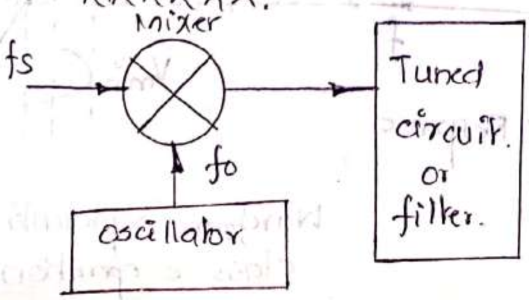
It is defined as

$$\rightarrow BW = f_2 - f_1 = \frac{f_r}{Q}$$

$\because f_1$ = Lower frequency

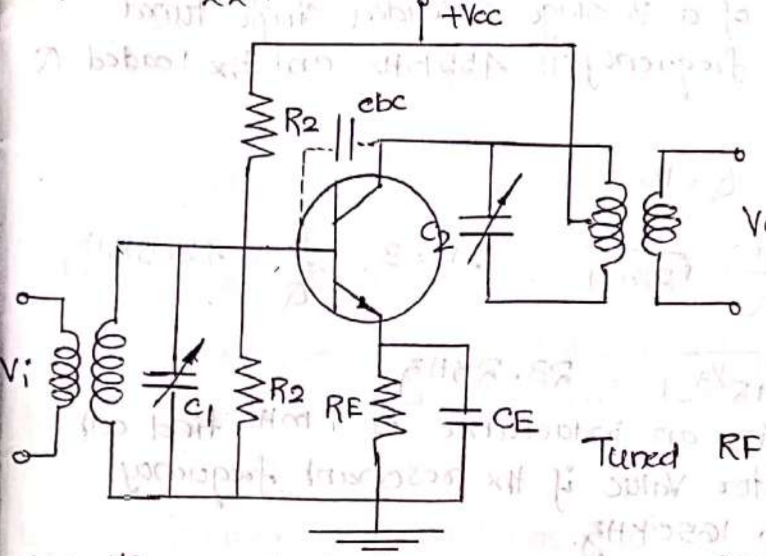
f_2 = upper frequency.

9. Application :



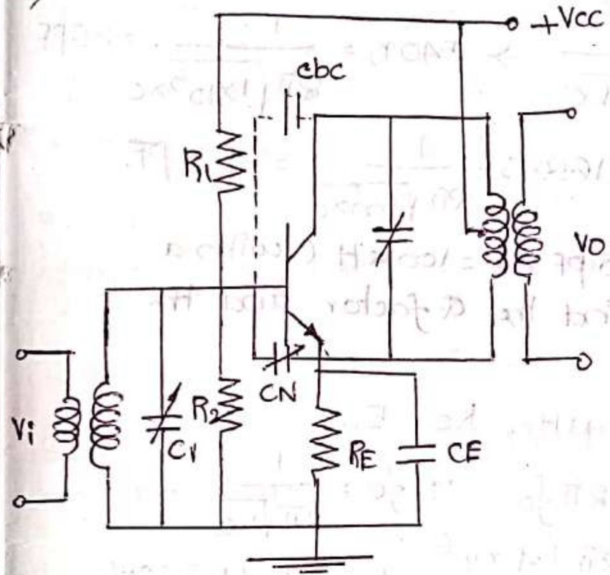
1. Radio Receivers.
2. up and down converters.

Stability of Tuned Amplifiers :

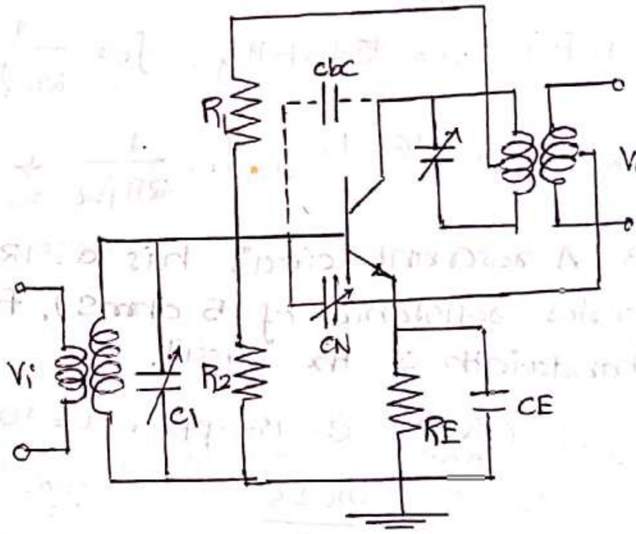


Tuned RF stage.

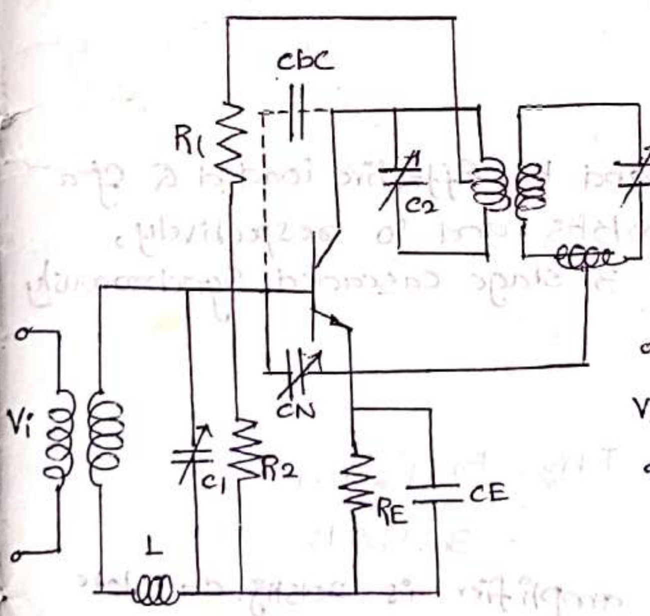
Hazelline Neutralization :



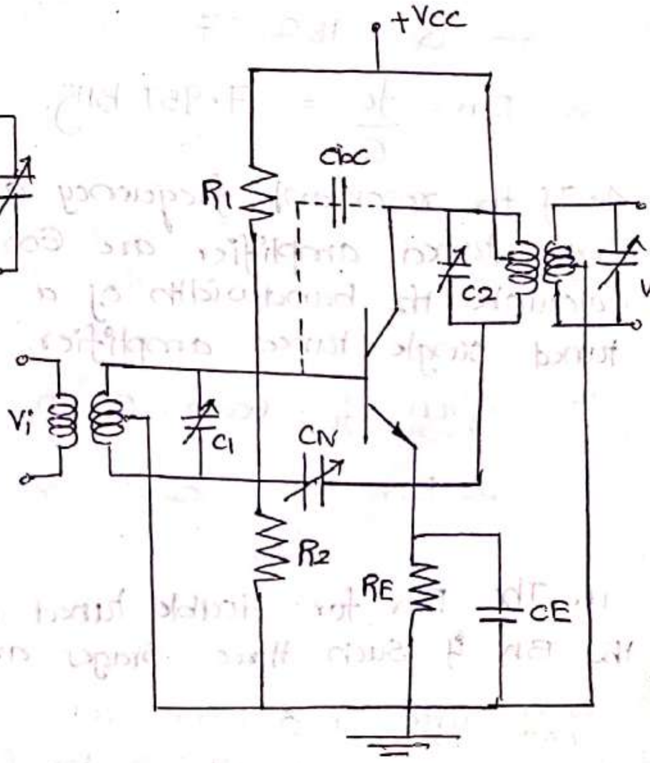
Neutrodyne Neutralization :



Neutralization using coil :



Rice Neutralization :



Problems:-

1. Determine the bandwidth of a 3 stage cascaded single tuned amplifier if the resonant frequency is 455KHz and the loaded Q of each stage is 10.

Soln Given $f_r = 455K$, $Q = 10$

wkt, $BW_n = BW\left(\frac{f_r}{Q}\right)\sqrt{2^{1/n}-1}$ ∴ $n=3$, $\frac{f_r}{Q} = 45.5KHz$.

$\rightarrow BW_3 = \frac{455K}{10}\sqrt{2^{1/3}-1} = 23.2KHz$.

2. A tank circuit contains an inductance of 1 mH. Find out the range of tuning capacitor value if the resonant frequency ranges from 540KHz to 1650KHz.

Soln Given $f_r = 540 - 1650KHz$, $L = 1mH$.

1. For $f_r = 540KHz$. $f_r = \frac{1}{2\pi\sqrt{LC}} \rightarrow 540K = \frac{1}{2\pi\sqrt{1\times10^{-3}\times C}} = 86PF$

2. For $f_r = 1650KHz$, $f_r = \frac{1}{2\pi\sqrt{LC}} \rightarrow 1650K = \frac{1}{2\pi\sqrt{1m\times C}} = 9.3PF$.

3. A resonant circuit has C = 120 pF, L = 100μH (with a series resistance of 5 ohms). Find the Q factor and the bandwidth of the circuit.

Soln Given $C = 120pF$, $L = 100\mu H$, $R_s = 5$

1. $\rightarrow Q = \frac{\omega_0 L_s}{R_s}$ ∵ $\omega_0 = 2\pi f_0$ ∵ $f_0 = \frac{1}{2\pi\sqrt{LC}}$

$\omega_0 = 2\pi\,1.4\times10^6$ $f_0 = 1.452MHz$

$\rightarrow Q = \frac{9.12\times10^6\times100\times10^{-6}}{5}$ $\omega_0 = 9.12\times10^6$

$\rightarrow Q = 182.57$

2. $BW = \frac{f_0}{Q} = 7.957KHz$.

4. If the resonant frequency and the Effective loaded Q of a single tuned amplifier are 600KHz and 10 respectively, calculate the bandwidth of a 3 stage cascaded synchronously tuned single tuned amplifier.

Soln Given $f_0 = 600K$, $Q = 10$

$\rightarrow BW = \frac{f_0}{Q} = 60K$ $\rightarrow BW_3 = BW\sqrt{2^{1/n}-1}$

$= 30.589K$.

11. The BW for double tuned amplifier is 20KHz. Calculate the BW if such three stages are cascaded.

Soln Given $n=3$, $BW = 20K$

$\rightarrow BW^n = BW(2^{1/n}-1)^{1/4} = 14.78KHz$.

5. A single tuned amplifier using n channel JFET with $g_m = 5mA/V$ and $r_d = 20k\Omega$, has tank circuit with $L = 1mH$, series resistance of the coil $R_s = 25\Omega$ and $C = 1nF$. Calculate the voltage gain at resonance if $R_L = 32k\Omega$.

Soln Given $g_m = 5mA/V$, $r_d = 20k$, $L = 1mH$, $R_s = 25\Omega$, $C = 1nF$
$R_L = 32k$

→ 1. $A_V = g_m R$ ∵ $R = r_d || R_p || R_L$ ∵ $R_p = \frac{\omega^2 L^2}{R_s}$

→ $A_V = 5mA \times 9.4k$ $R = 20k || 40k || 32k$ $= \frac{(2\pi \times 159.1k)^2 (1m)^2}{25}$

→ $A_V = 47$ $R = 9.4k$ $= 40k$

∵ $f_0 = \frac{1}{2\pi\sqrt{LC}} = 159.1k$

6. An inductor of $250\mu H$ has $Q = 300$ at 1M. Determine R_s and R_p of Inductor.

Soln Given $L = 250\mu$, $Q = 300$, $f = 1MHz$.

1. $R_p = \omega_0 L Q = 2\pi f_0 L Q = 2\pi \times 1\times10^6 \times 250\mu \times 300 = 471.2 k\Omega$

2. $R_s = \frac{\omega_0 L}{Q} = 5.235\Omega$

7. Derive the bandwidth of a synchronous tuning system with three single tuned amplifiers. Assume bandwidth of individual stage is 10kHz.

Soln Given $BW_n = BW\sqrt{2^{1/n}-1} = 10k\sqrt{2^{1/3}-1} = 5.09kHz$.

8. A tuned Amplifier has its maximum gain at a frequency of 2MHz and has a bandwidth of 50kHz. calculate the Q factor.

Soln Given $f_r = 2MHz$, $BW = 50kHz$.

→ $BW = \frac{f_r}{Q}$ → $Q = \frac{f_r}{BW} = 40$.

9. A parallel resonant circuit has an inductance of $150\mu H$ and a capacitance of 100pF. Find the resonant frequency.

Soln Given $L = 150\mu H$; $C = 100pF$

→ $f_0 = \frac{1}{2\pi\sqrt{LC}} = 1299.4 kHz$

10. A parallel resonant circuit has a capacitor of 100pF and an inductor of 100 micro H. The inductor has a resistance of 5ohms. Find the value of frequency at which the circuit resonates and the circuit impedance at resonance.

Soln Given $C = 100pF$, $L = 100\times10^{-6} H$., $R_s = 5$ ohms.

1. $f_0 = \frac{1}{2\pi\sqrt{LC}} = 1.591 MHz$

2. Impedance at Resonance.

→ $R_p = \frac{\omega^2 L^2}{R_s}$

→ $R_p = \frac{(2\pi \; 1.59M)^2 (100\mu)^2}{5}$

→ $R_p = 200k\Omega$.

UNIT- IV

WAVESHAPING AND MULTIVIBRATOR CIRCUITS

Linear wave shaping:

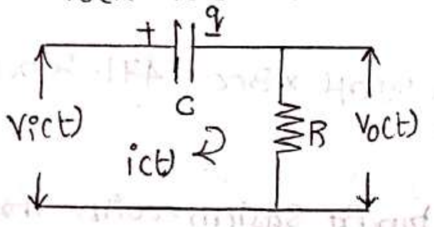
The process by which the shape of a nonsinusoidal signal is changed by passing the signal through the networks consisting of linear elements is called Linear wave shaping.

RC & RL Integrator and Differentiator circuits:

RC circuit can be divided into two,

1. High pass RC circuit 2. Low pass RC circuit.

1. High pass RC circuit:



The dia. shows high pass RC circuit. The circuit obstructs the low frequencies while it allows high frequencies to reach the output is called High pass RC circuit. At zero frequency, the reactance becomes infinite and hence offers open circuit.

At very high frequency, the capacitor acts as a short circuit and all the input appears at the output.

The capacitive reactance is given by

$$\rightarrow X_C = \frac{1}{2\pi f C} \rightarrow (1)$$

The gain is given by (Transfer Gain (or) Amplification)

$$\rightarrow |A| = \frac{1}{\sqrt{1+\left(\frac{f_1}{f}\right)^2}} \rightarrow (2) \qquad \because f_1 = \frac{1}{2\pi RC}$$

f = Input frequency.

1. step Input Voltage:

Consider the step input voltage of magnitude A Volts is applied as an input to the high pass RC circuit.

Let V_o is of the form,

$$\Rightarrow V_o(t) = B_1 + B_2 e^{-t/\tau} \rightarrow (3)$$

Where B_1, B_2 = constant

t = time constant

$\tau = RC$

Case 1:

Now, $t \to \infty$, be 0.

$$\lim_{t\to\infty} V_o(t) = \lim_{t\to\infty} (B_1 + B_2 e^{-t/\tau})$$

$$0 = B_1 + 0.$$

$$0 = B_1$$

Let consider, $B_1 = V_f \rightarrow (4)$

Case 2:

The output voltage at $t=0$ be V_i, we get

$\rightarrow V_o(t)\big|_{t=0} = B_1 + B_2 e^{-t/\tau} = V_i$

$\rightarrow V_i = V_f + B_2 \quad \because B_1 = V_f$

$\rightarrow B_2 = V_i - V_f \rightarrow (5)$

sub B_1 and B_2 in (3)

$\rightarrow V_o = V_f + (V_i - V_f)e^{-t/\tau} \rightarrow (6)$

This is the general expression for the output voltage.

wkt, At zero frequency, output voltage is zero,

$\rightarrow V_f = 0V$

At high frequency, $t=0$,

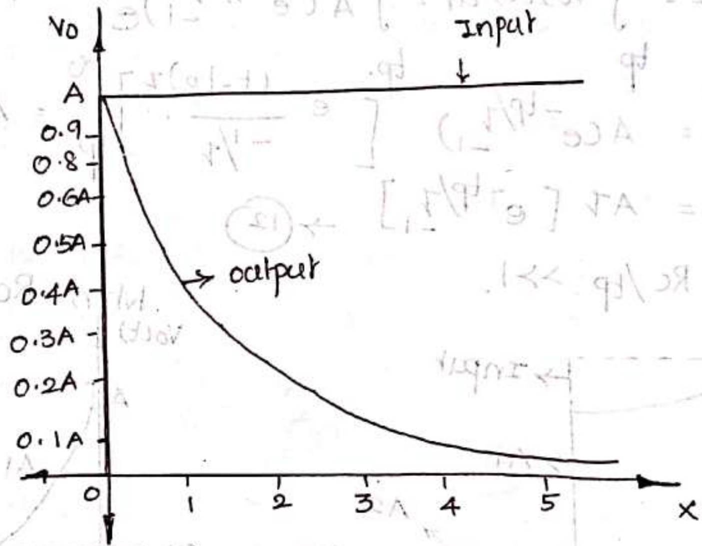
$\rightarrow V_i = A$ volts

Now, $V_o(t) = A e^{-t/\tau} \rightarrow (7)$

Let $+t/\tau = x$

$\rightarrow V_o(t) = A e^{-x}$

x	V_o
0	A
0.5	0.606A
1	0.367A
2	0.135A
3	0.05A
4	0.018A
5	0.007A



step response.

2. Pulse Input Voltage:

Consider the pulse input voltage is applied to the input to High pass RC circuit. and pulse is the sum of two step voltages.

Consider, the output voltage, for first pulse,

$\rightarrow V_{o1}(t) = A e^{-t/RC}$ for $0 < t < t_p$

At $t = t_p$, $V_{o1}(t_p) = A e^{-t_p/RC} = V_p \rightarrow (8)$

For second pulse,

$\rightarrow t = t_p \rightarrow V_{o2}(t_p) = A(e^{-t_p/RC} - 1) = V_i \rightarrow (9)$

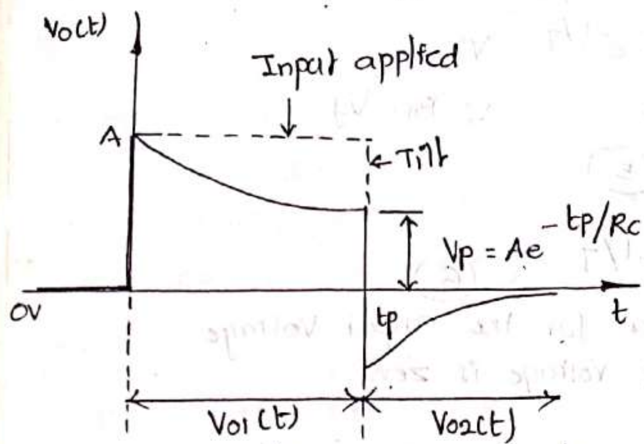
The final value of output voltage using general Expression is

$\rightarrow V_{o2}(t) = V_f + (V_i - V_f)e^{-t/\tau}$

$\because V_f = 0,\ V_i = A(e^{-t_p/RC} - 1),\ t - t_p = t,\ \tau = 1/RC$

$\rightarrow V_{o2}(t) = 0 + A(e^{-t_p/RC} - 1)e^{-(t-t_p)/RC}$ —

$$\rightarrow V_{o2}(t) = A\left(e^{-t_p/RC} - 1\right) e^{-(t-t_p)/RC} \rightarrow (10)$$



The are two types of distortion resulted the above response.

1. A tilt at the top of the pulse
2. An undershoot at the end of the pulse.

To minimize the distortion

Let $A_1 = \int_0^{t_p} V_{o1}(t)\,dt = \int_0^{t_p} A e^{-t/\tau} dt \quad \because \tau = RC$

$$\rightarrow A_1 = A\left[\frac{e^{-t/\tau}}{-1/\tau}\right]_0^{t_p} = A\tau\left[-e^{-t_p/\tau} + 1\right] \rightarrow (11)$$

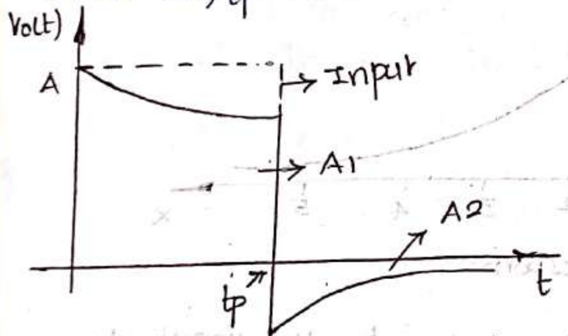
and

$$\rightarrow A_2 = \int_{t_p}^{\infty} V_{o2}(t)\,dt = \int_{t_p}^{\infty} A\left(e^{-t_p/\tau} - 1\right) e^{-(t-t_p)/\tau} dt$$

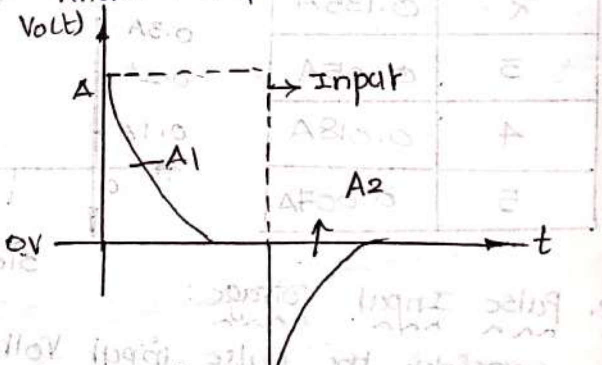
$$\rightarrow A_2 = A\left(e^{-t_p/\tau} - 1\right)\left[\frac{e^{-(t-t_p)/\tau}}{-1/\tau}\right]_{t_p}^{\infty} = A\left(e^{-t_p/\tau} - 1\right)\left[0 - \left(-\frac{1}{1/\tau}\right)\right]$$

$$\rightarrow A_2 = A\tau\left[e^{-t_p/\tau} - 1\right] \rightarrow (12)$$

When $RC/t_p >> 1$.



When $RC/t_p << 1$



3. Square Wave Input Voltage :

Consider the square wave Input is applied to high pass RC circuit.

Let V_i = Input Voltage

V_c = Voltage across capacitor

V_o = output Voltage

q = charge on Capacitor.

Apply KVL, $\quad \because V_c = \frac{q}{c}$

$$\rightarrow V_i = V_c + V_o$$

$$\rightarrow V_i = \frac{q}{c} + V_o$$

Diff with res. to dt,

$$\rightarrow \frac{dV_i}{dt} = \frac{1}{c}\frac{dq}{dt} + \frac{dV_o}{dt} \quad \text{but } \frac{dq}{dt} = i$$

$\rightarrow \frac{dv_i}{dt} = \frac{i}{c} + \frac{dv_o}{dt}$

$\because V_o = iR, \; i = \frac{V_o}{R}$

$\rightarrow \frac{dv_i}{dt} = \frac{V_o}{Rc} + \frac{dv_o}{dt}$

(X) by dt.

$\rightarrow dv_i = \frac{V_o}{Rc} dt + dv_o.$

$\int$ ing from 0 to T,

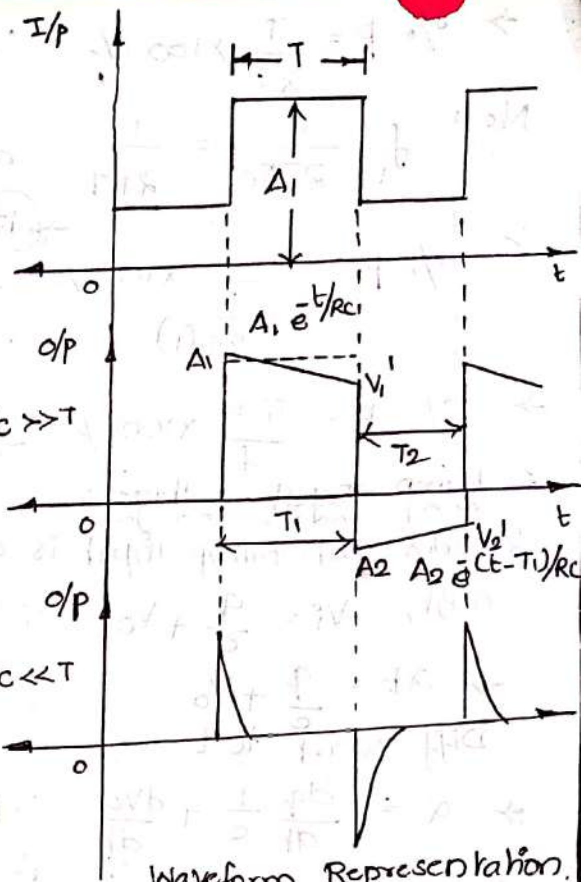
$\rightarrow \int_0^T dv_i = \int_0^T \frac{V_o}{Rc} dt + \int_0^T dv_o$

$\rightarrow [v_i]_0^T = \frac{1}{Rc}\int_0^T V_o dt + [v_o]_0^T$

$\rightarrow V_i(T) - V_i(0) = \frac{1}{Rc}\int_0^T V_o dt + V_o(T) - V_o(0)$

Where $V_i(T) = V_i(0)$, $V_o(T) = V_o(0)$

$\rightarrow \frac{1}{Rc}\int_0^T V_o dt = 0 \rightarrow (13)$



Waveform Representation.

From the waveform,

$\rightarrow T = T_1 + T_2 \quad \& \quad T_1 = T_2 = T/2$

and $A_1 = +A_2$ and $V_1' = V_2'$

Where $A_1 = \frac{A}{1+e^{-T/2RC}} \rightarrow (14)$ and $V_1' = \frac{A}{1+e^{+T/2RC}} \rightarrow (15)$

Now, Percentage tilt is defined as

$\rightarrow P = \frac{A_1 - V_1'}{\text{input Amplitude } (A/2)} \times 100 = \frac{\frac{A}{1+e^{-T/2RC}} - \frac{A}{1+e^{T/2RC}}}{\frac{A}{2}} \times 100$

$\rightarrow P = \frac{(\cancel{1} + e^{T/2RC} - \cancel{1} - e^{-T/2RC})}{(1+e^{-T/2RC})(1+e^{T/2RC})} \times 2 \times 100$

Now $e^{T/2RC} - e^{-T/2RC} = (1-e^{-T/2RC})(1+e^{T/2RC})$

$\rightarrow P_1 = \frac{(1-e^{-T/2RC})(1+e^{T/2RC})}{(1+e^{-T/2RC})(1+e^{T/2RC})} \times 200 = \frac{(1-e^{-T/2RC})}{(1+e^{-T/2RC})} \times 200.$

Now $e^{-T/2RC} \simeq 1 - T/2RC$

$\rightarrow \% P = \frac{1-(1-T/2RC)}{1+(1-T/2RC)} \times 200 = \frac{T/2RC}{2-T/2RC} \times 200 \quad \because \frac{T}{2RC} << 2$

$\rightarrow \% P = \frac{T}{4RC} \times 200 = \frac{T}{2RC} \times 100 \%.$

$\rightarrow \% P = \frac{T}{2\tau} \times 100\%$ $\quad \because \tau = RC \rightarrow (16)$

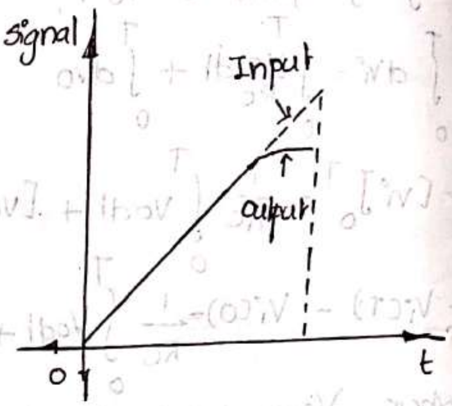
Now $f_1 = \frac{1}{2\pi RC} = \frac{1}{2\pi\tau} \rightarrow (17)$ and RC or $\tau = \frac{1}{2\pi f_1}$

$$\rightarrow \% P = \frac{T}{2\left(\frac{1}{2\pi f_1}\right)} \times 100\% = \pi T f_1 \times 100\% = \frac{\pi f_1}{f} \times 100\%$$

$$\rightarrow \% P = \frac{\pi f_1}{f} \times 100\% \rightarrow (18)$$

4. Ramp Input Voltage:

Consider that Ramp input is applied.



WKT, $V_i = \frac{q}{c} + V_o \quad \because V_i = \alpha t$

$\rightarrow \alpha t = \frac{q}{c} + V_o$

Diff w.r.t to t.

$\rightarrow \alpha = \frac{dq}{dt}\frac{1}{c} + \frac{dV_o}{dt} \quad \because \frac{dq}{dt} = i$

$\rightarrow \alpha = \frac{i}{C} + \frac{dV_o}{dt} \quad \because V_o = iR$

$\because i = V_o/R$

$\rightarrow \alpha = \frac{V_o}{RC} + \frac{dV_o}{dt} \rightarrow (19)$

The diff. eqn has a soln,

$\rightarrow V_o = \alpha RC(1 - e^{-t/RC}) \rightarrow (20)$

$\rightarrow V_o = \alpha RC(\cancel{1} - (\cancel{1} - \frac{t/RC}{1!} + \frac{(t/RC)^2}{2!} + \dots))$

$\rightarrow V_o = \alpha RC\left(t/RC - \frac{t^2}{2R^2C^2} + \dots\right) = \alpha t\left(1 - \frac{t}{2RC}\right) \rightarrow (20)$

Now,

The Error at time T can be obtained as

$$\rightarrow e_t = \left.\frac{V_i - V_o}{V_i}\right|_{t=T} = \frac{\alpha t - \alpha t(1 - t/2RC)}{\alpha t}$$

$$\rightarrow \% e_t = \frac{\cancel{\alpha t} - \cancel{\alpha t} + \frac{\alpha t^2}{2RC}}{\alpha t} \times 100 = \frac{T}{2RC}$$

$\rightarrow \% e_t = \pi T f_1 \rightarrow (21) \quad \because \% P = \pi T f_1$

5. High pass RC act as a Differentiator:

For a high pass RC, if time constant is very small as compared to the time required by input signal to make a change, the circuit acts as a differentiator.

The drop across R is negligible compared to drop across C. Thus entire input Vi can be assumed to be appearing across C.

The current i is given by

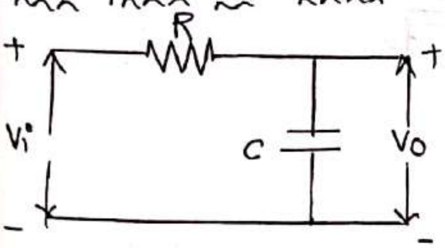
$$\rightarrow i = C\frac{dV_c}{dt} = C\frac{dV_i}{dt} \rightarrow (22)$$

wkt, $V_0 = iR \quad \because i = C\frac{dv_i}{dt}$

$\rightarrow V_0 = RC\frac{dv_i}{dt} \rightarrow (23)$

The Equation shows that the output is differentiation of the input. and hence the circuit is called differentiator.

2. Low pass RC circuit :



The dia. shows low pass RC.

At high frequencies, the capacitor acts as a Virtual short circuit and hence output falls to zero.

The circuit passes the low frequencie readily, hence called low pass RC circuit.

The transfer function is given by

$$\rightarrow |A| = \frac{1}{\sqrt{1+\left(\frac{f}{f_2}\right)^2}} \quad \because f_2 = \frac{1}{2\pi RC}$$

$\rightarrow (1) \quad f$ = Input frequency.

1. Step Input Voltage :

Consider that the step input voltage is applied to low pass RC.

Let $V_i = 0V$ and $V_f = AV$

The general expression for the output as

$\rightarrow V_o(t) = V_f + (V_i - V_f)e^{-t/\tau} \rightarrow (2)$

Sub $V_i = 0$, $V_f = AV$

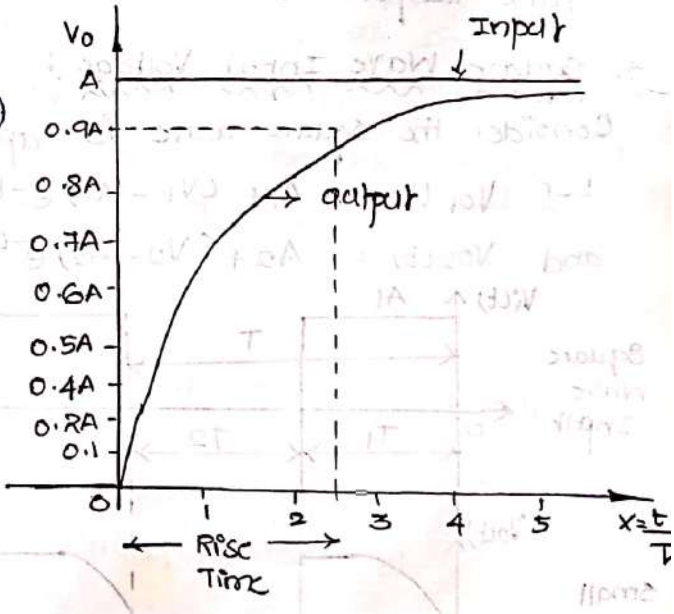
$\rightarrow V_o(t) = A + (0-A)e^{-t/\tau}$

$\rightarrow V_o(t) = A(1-e^{-t/\tau}) \quad \because \tau = RC$

Now Let $t/\tau = x$

$\rightarrow V_o(t) = A(1-e^{-x}) \rightarrow (3)$

X	V_o
0	0
0.5	0.39A
1	0.63A
2	0.86A
3	0.95A
4	0.98A
5	0.99A
6	1.02A



Rise time :

It is the time required by output response to rise from 10% to 90% of its final steady state value.

Now, Time required for the output to achieve 10%

$\rightarrow V_o(t) = A(1-e^{-t/RC})$

$\rightarrow 0.1\cancel{A} = \cancel{A}(1-e^{-t/RC})$

$\rightarrow e^{-t/RC} = 0.9$ and III^{ly} To achieve 90%.

$\rightarrow -t/RC = \ln(0.9) \rightarrow t = 2.3\,RC \rightarrow (5)$

$\rightarrow t = 0.1\,RC \rightarrow (4)$

From Rise time,

$\rightarrow t_r = 2.3RC - 0.1RC = 2.2RC \simeq 2.2\tau \rightarrow (6)$

But $f_2 = \dfrac{1}{2\pi RC} = \dfrac{1}{2\pi\tau}$ and $RC = \dfrac{1}{2\pi f_2}$

$\rightarrow t_r = \dfrac{2.2}{2\pi f_2} = \dfrac{0.35}{f_2} \rightarrow (7)$

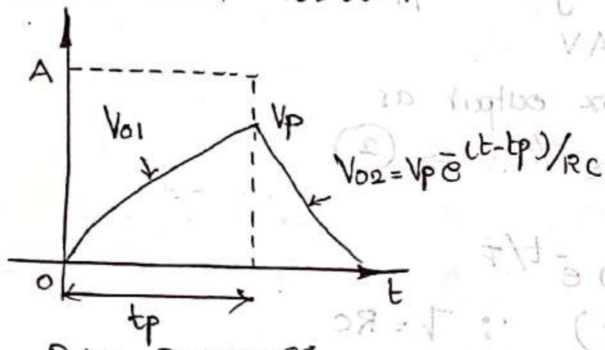
2. Pulse Input Voltage :

Consider a pulse input Voltage is applied as input to low pass RC.

Let $V_{o1}(t = t_p) = A(1 - e^{-t_p/RC}) = V_p \qquad 0 < t < t_p \rightarrow (8)$

and $V_{o2}(t) = V_p\, e^{-(t-t_p)/RC} \qquad t > t_p \rightarrow (9)$

Here over all response of low pass RC in addition of $V_{o1}(t)$ and $V_{o2}(t)$.

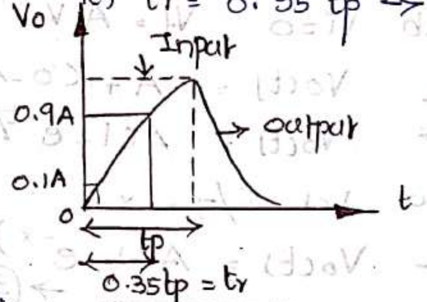


Pulse Response

To minimize the distortion, the t_r must be small compared with pulse width t_p.

Let $f_2 = \dfrac{1}{t_p} \rightarrow (10)$

ie) $t_r = 0.35\, t_p \rightarrow (11)$

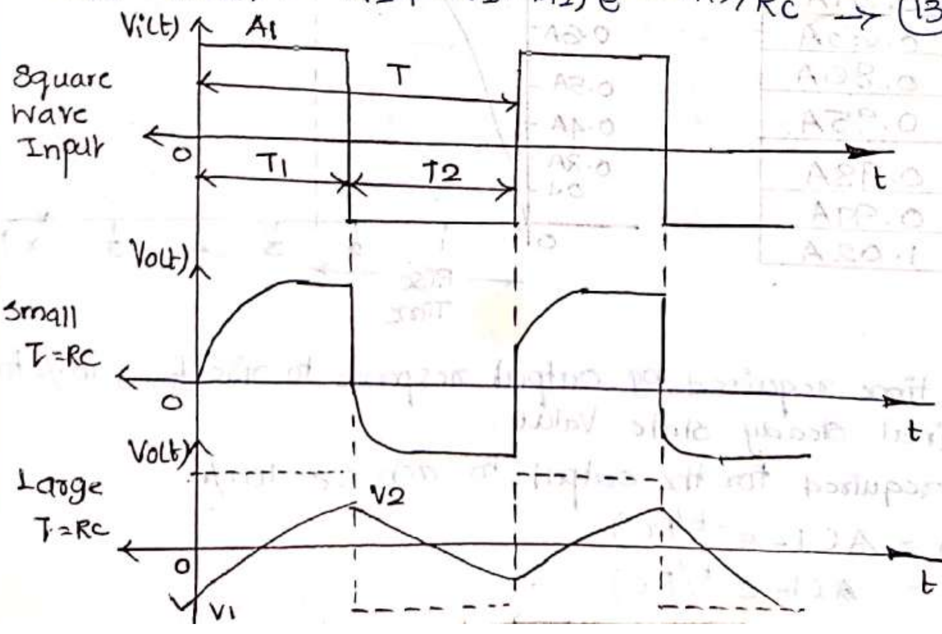


3. Square Wave Input Voltage :

Consider the square wave is applied as input to low pass RC

Let $V_{o1}(t) = A_1 + (V_1 - A_1)\, e^{-t/RC} \rightarrow (12)$

and $V_{o2}(t) = A_2 + (V_2 - A_2)\, e^{-(t-T_1)/RC} \rightarrow (13)$



4. Ramp Input Voltage:

Consider ramp input is applied as input to low pass RC.

Apply KVL,

$\rightarrow V_i = V_R + V_C \rightarrow (14)$

w.k.t, $V_o = \alpha RC\,(1 - e^{-t/RC})$ using these,

$\rightarrow V_R = \alpha RC\,(1 - e^{-t/RC}) \rightarrow (15)$

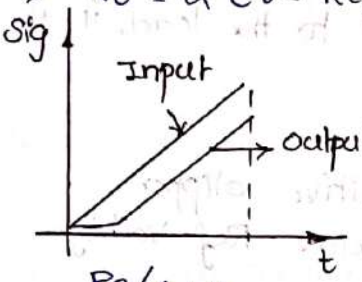
Sub 15 in 14

$\rightarrow V_C = V_i - V_R = V_i - \alpha RC\,(1 - e^{-t/RC}) = \alpha t - \alpha RC + \alpha RC\, e^{-t/RC}$ $\quad \because V_i = \alpha t$

$\rightarrow V_C = \alpha(t - RC) + \alpha RC\, e^{-t/RC} \rightarrow (16)$

Let $V_o = V_C$

$\rightarrow V_o = \alpha(t - RC) + \alpha RC\, e^{-t/RC} \rightarrow (17)$



$RC/T \ll 1$

The error at time,

$\rightarrow e_t = \dfrac{V_i - V_o}{V_i}$ $\quad \because V_i = \alpha T,\; V_o = \alpha(T - RC) + \alpha RC\, e^{-T/RC}$

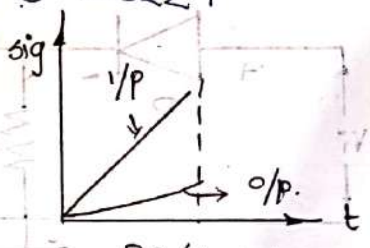
$\rightarrow e_t = \dfrac{\alpha T - [\alpha(T - RC) + \alpha RC\, e^{-T/RC}]}{\alpha T}$

$\rightarrow e_t = \dfrac{\alpha T - \alpha T + \alpha RC + \alpha RC\, e^{-T/RC}}{\alpha T}$

$\rightarrow e_t = \dfrac{\alpha RC\,(1 - e^{-T/RC})}{\alpha T} = \dfrac{RC}{T} \rightarrow (18)$ $\quad \because e^{-T/RC} \ll 1$

But $f_2 = \dfrac{1}{2\pi RC}$ $\quad \therefore RC = \dfrac{1}{2\pi f_2}$

$\rightarrow e_t = \dfrac{1}{2\pi f_2 T} \rightarrow (19)$



$RC/T \gg 1.$

5. Low Pass RC as an Integrator:

For low pass RC, if time constant is very large as compared to the time required by input signal to make an appreciable change, the circuit acts as an Integrator.

Under this case, the drop across C is negligible compared to drop across R. Then current i is given by

$\rightarrow i = \dfrac{V_i}{R} \rightarrow (20)$

The V_o is

$\rightarrow V_o = \dfrac{1}{C}\int i\,dt = \dfrac{1}{C}\int \dfrac{V_i}{R}\,dt = \dfrac{1}{RC}\int V_i(t)\,dt.$

$\rightarrow (21)$

The Equation (21) shows that the output is integration of the input. and the circuit is called Integrator.

RL circuits :

1. High pass RL circuit : 2. Low pass RL circuit

High pass RL – Integrator   Low pass RL – Differentiator

Diode clippers or Limiters or Slicers :

The circuits which are used to clip off unwanted portion of the waveform, without distorting the remaining part of the waveform are called clipper.

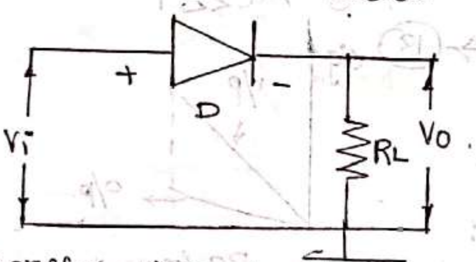
Two Types :

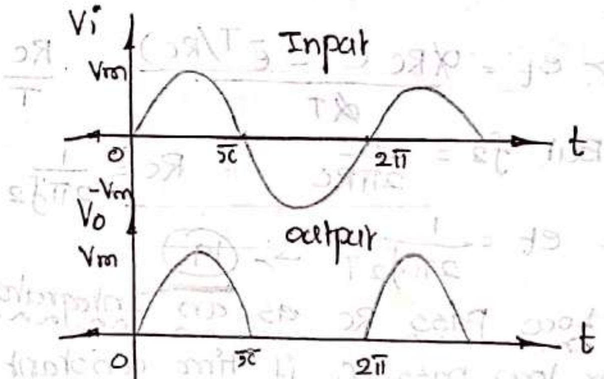
1. When the diode is connected in series with the load, such circuit is called series clipper.
2. When the diode is connected in parallel to the load, it is called parallel clipper.

1. Series clippers:

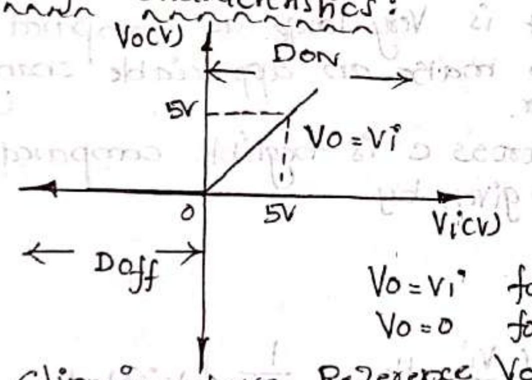
1. Series Negative clipper   2. Series positive clipper
3. Clipping above Ref voltage   4. Clipping Below Ref Voltage
5. Additional DC supply in series with Diode.

1. Series Negative clipper :





Transfer characteristics :



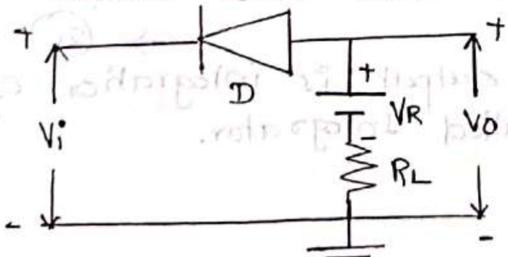
$V_o = V_i - 0.7$ $V_i > 0.7V$

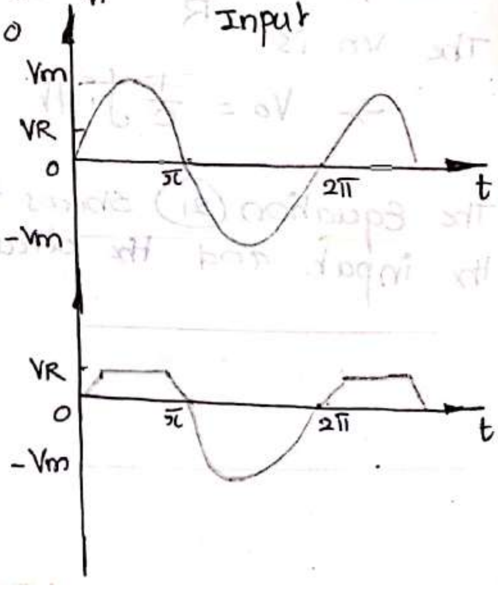
$V_o = 0$ $V_i \leq 0.7V$

$V_o = V_i$ for $V_i \geq 0$

$V_o = 0$ for $V_i \leq 0$

3. Clipping above Reference Voltage :

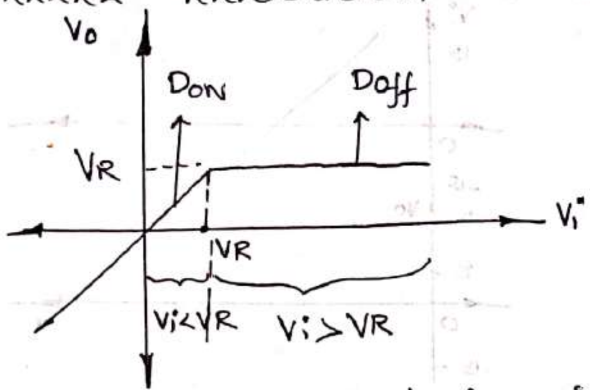




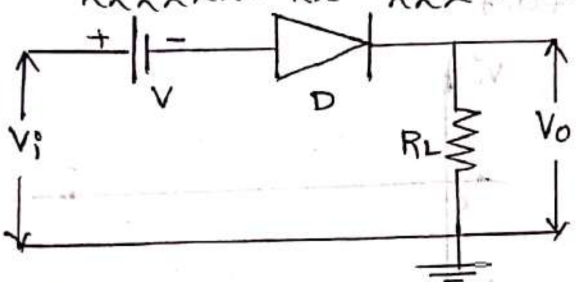
$V_o = V_i$ for $V_i < V_R$

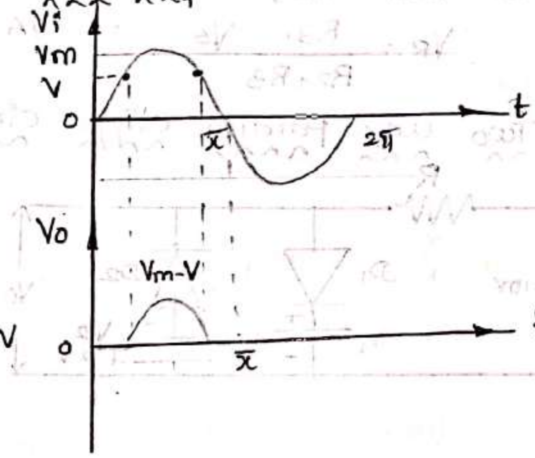
$V_o = V_R$ for $V_i > V_R$

Transfer characteristics :



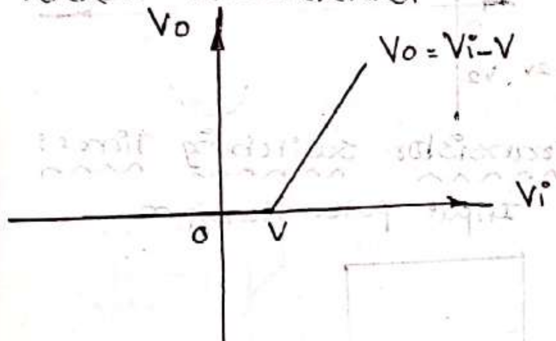
5. Additional D.C supply in series with Diode :





→ $V_i > V$ start conduction $V_o = V_i - V$

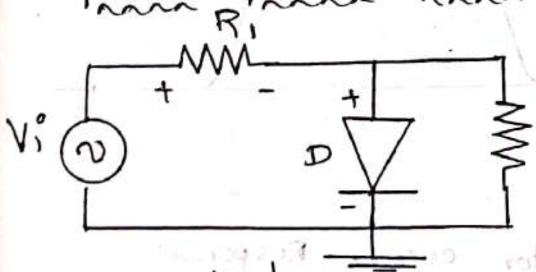
Transfer characteristics :

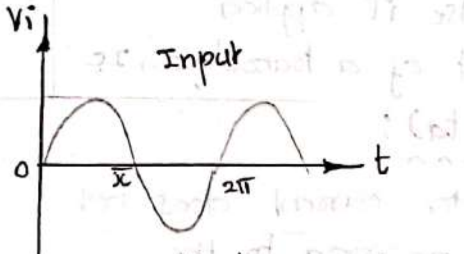


2. Parallel Clippers :

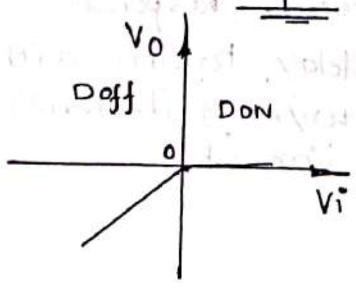
1. Parallel positive clipper
2. parallel Negative clipper
3. Effect of cutin Voltage of diode
4. Parallel +v clipper with Ref Voltage
5. Parallel Neg. clipper with VR
6. parallel Clipper with Vol. divider
7. Two way parallel clipper circuit.

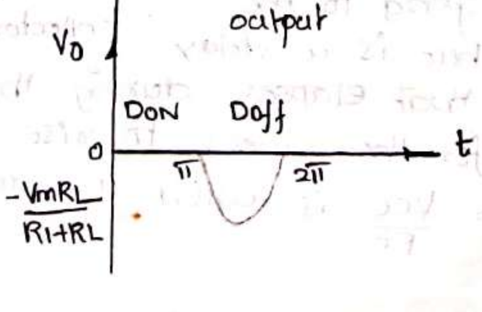
1. parallel positive clipper :



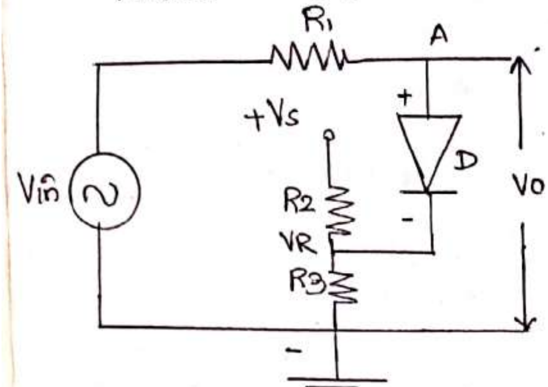


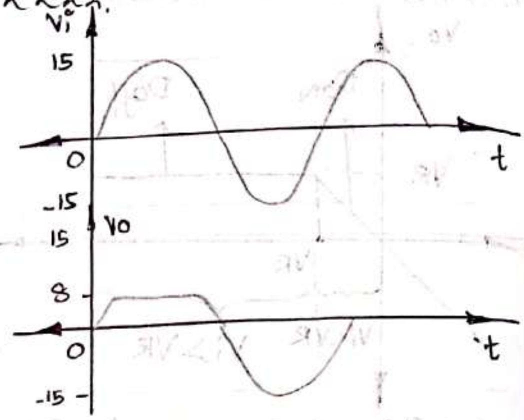
$V_o = \frac{V_i R_L}{R_L + R_1}$





6. Parallel clipper with Voltage divider :

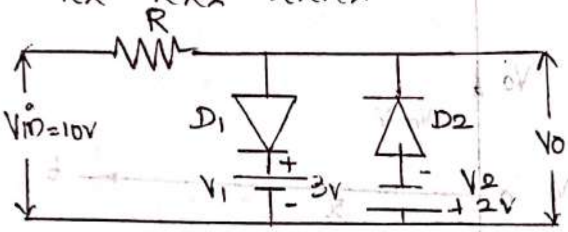


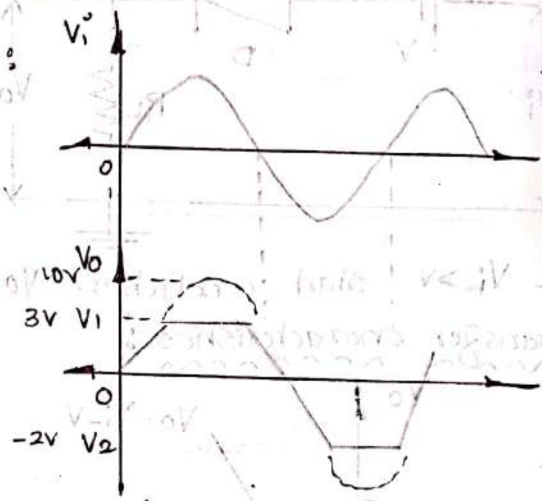


Vol. Div. Rule,

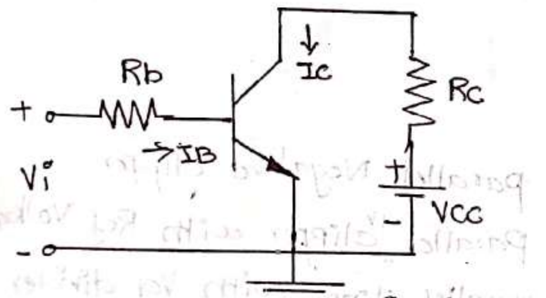
$$V_R = \frac{R_3}{R_2+R_3} V_s \quad \therefore V_A = V_R + 0.7$$

7. Two way parallel clipper circuit :

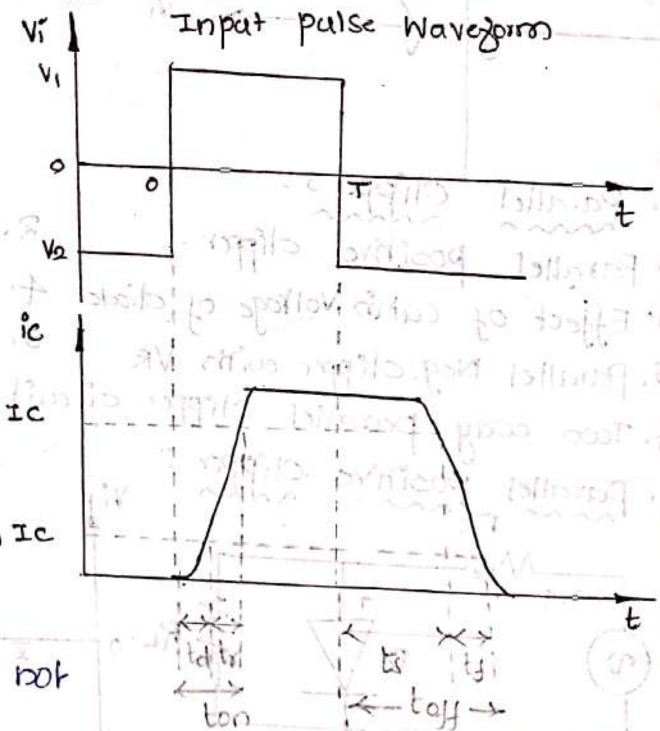




Storage, Delay and Calculation of Transistor switching Times :



Transistor switching circuit.



From the above dia.
when a pulse is applied to the input of a transistor

Collector current Response.

Delay time (td) :

The collector current does not immediately respond to the input pulse. There is a delay and the time that elapses during this delay, together with time needed for the current to rise to 10% of its maximum value i.e, $I_c = \frac{V_{cc}}{R_c}$ is called the delay time (td)

Rise time (tr):

The time required for the collector current to rise from 10% to 90% of the maximum value.

Turn-ON time (tON):

The sum of the delay and rise time is called Turn-on

$\rightarrow t_{ON} = t_d + t_r$.

Storage time (ts):

When the input signal returns back to its initial state at t=T, the collector current again fails to respond Immediately. The interval which elapses between the transition of the input voltage waveform and the time when collector current dropped to 90% of its max. value.

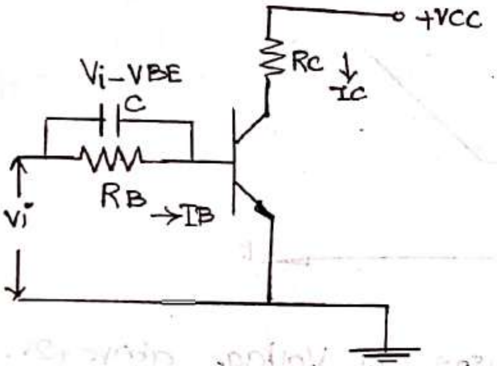
Fall time (tf):

The time required for the collector current to fall from 90% to 10% of its maximum value.

Turn-off time (toff):

The sum of the storage and fall time is called toff.

Speed-up (or) Transpose (or) commutating capacitor:



Transistor circuit with speed up capacitor.

From the dia, the capacitor C is called speed-up or commutating capacitor.

The speed-up capacitor tends to reduce td and ts as well as tr and tf. However, if c is made very small, it becomes completely changed within the delay time and hence it will not have any effect on rise time. Similarly if C is totally discharging during the storage time, it will not result in a significant improvement of the fall time.

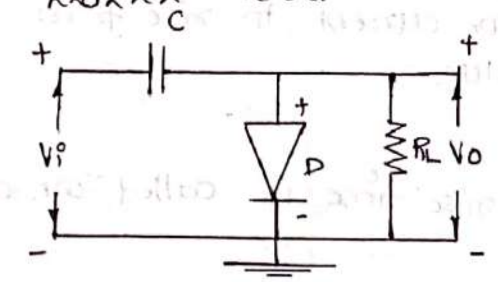
Clampers: (Dc restorer) (Dc Inserter):

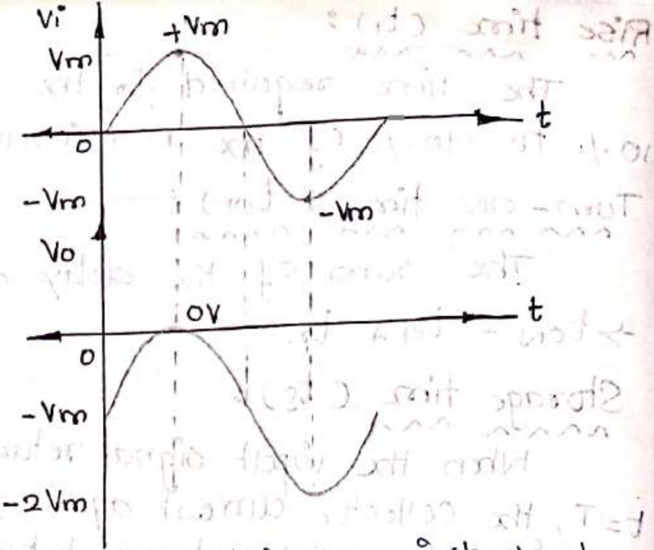
Clamping networks shifts (clamps) a signal to a different dc level i.e, it introduces a dc level to an ac signal. Hence, the clamping networks is also known as dc restorer.

Types:

1. Negative clamper
2. positive clamper.

1. Negative Clamper:



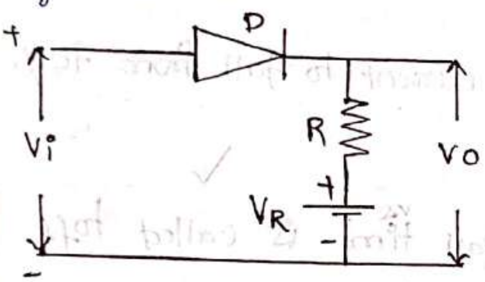


→ $V_i = V_m$, $V_o = 0$

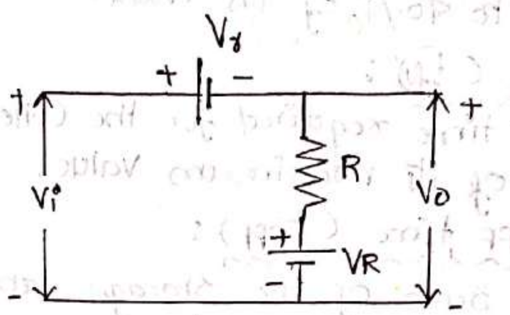
→ $V_i = -V_m$, $V_o = -2V_m$.

Diode comparator:

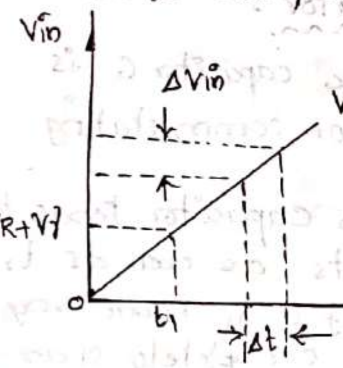
A comparator circuit is used to identify the instant at which the arbitrary input waveform attains a particular reference level.

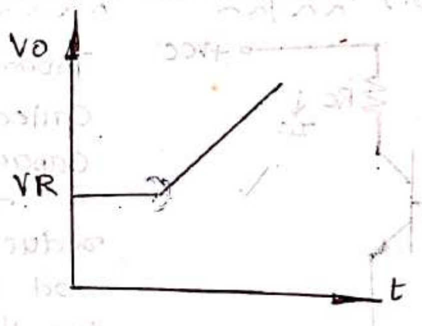


Diode comparator



Equivalent circuit

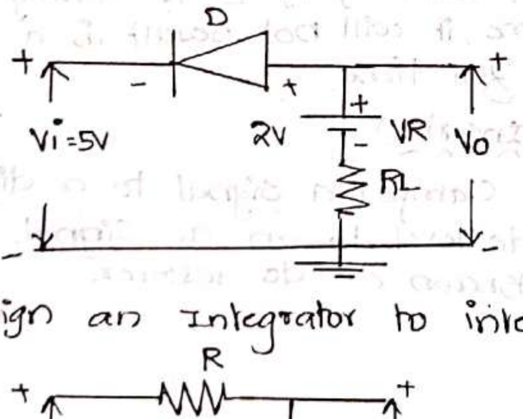


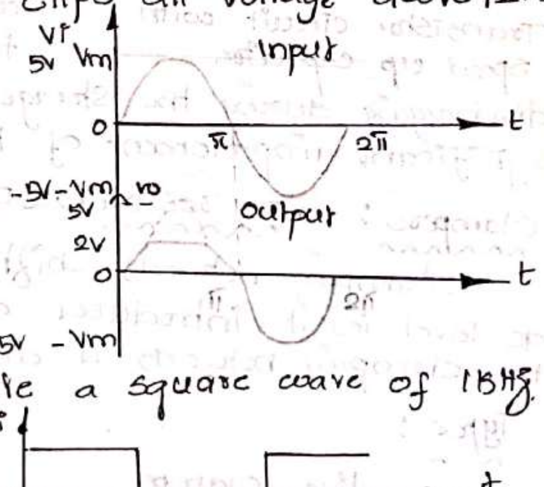


Problem:

1. Draw a clipper circuit which clips all voltage above +2V.

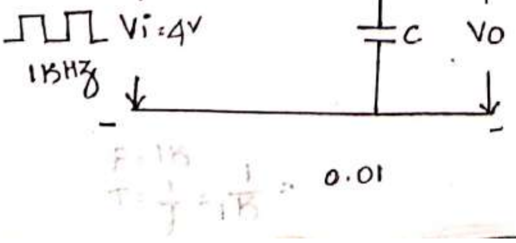
Soln

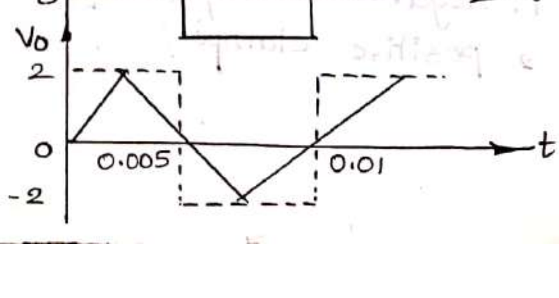




2. Design an Integrator to integrate a square wave of 1kHz.

Soln

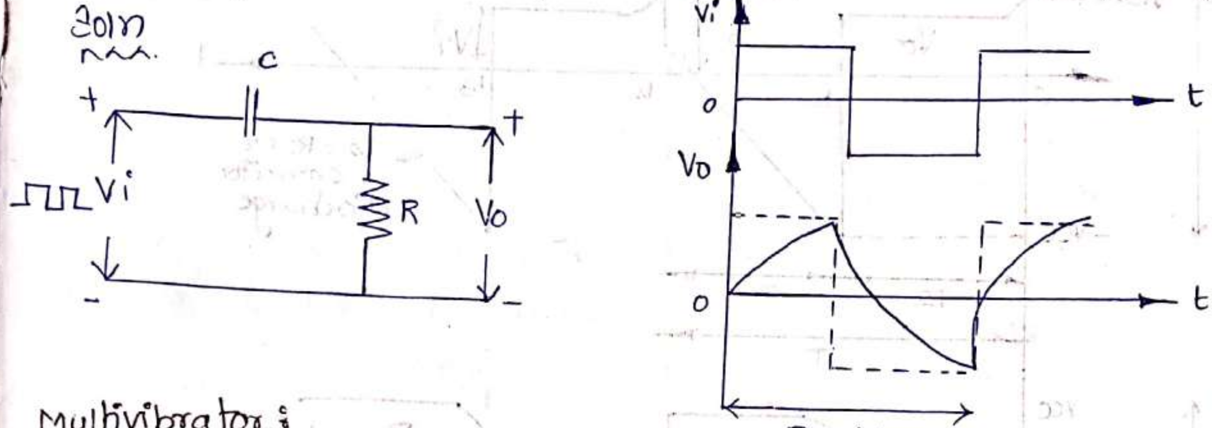




0.01

3. Sketch the output waveforms of a differentiating circuit with square wave input for $RC = PW$ and $RC = PW/10$.

Soln



$RC = PW$.

Multivibrator:

The electronic circuits which are used to generate non-sinusoidal waveforms are called Multivibrators.

Types:

a) Astable Multivibrator
b) Monostable Multivibrator
c) Bistable Multivibrator.

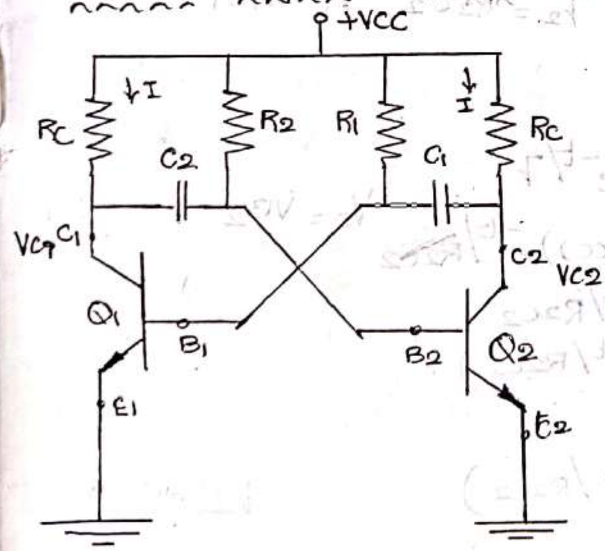
a) Astable Multivibrator: (Free running Multivibrator):

Astable multivibrator does not require any external pulse for the transition, is called free running Multivibrator.

It has both the states as quasi-stable states. Due to this, the multivibrator automatically makes the successive transitions from one quasi stable to other without any external triggering pulse.

It is used as the generator of square waves.

Collector coupled Astable Multivibrator:



Assume

Q_2 - ON

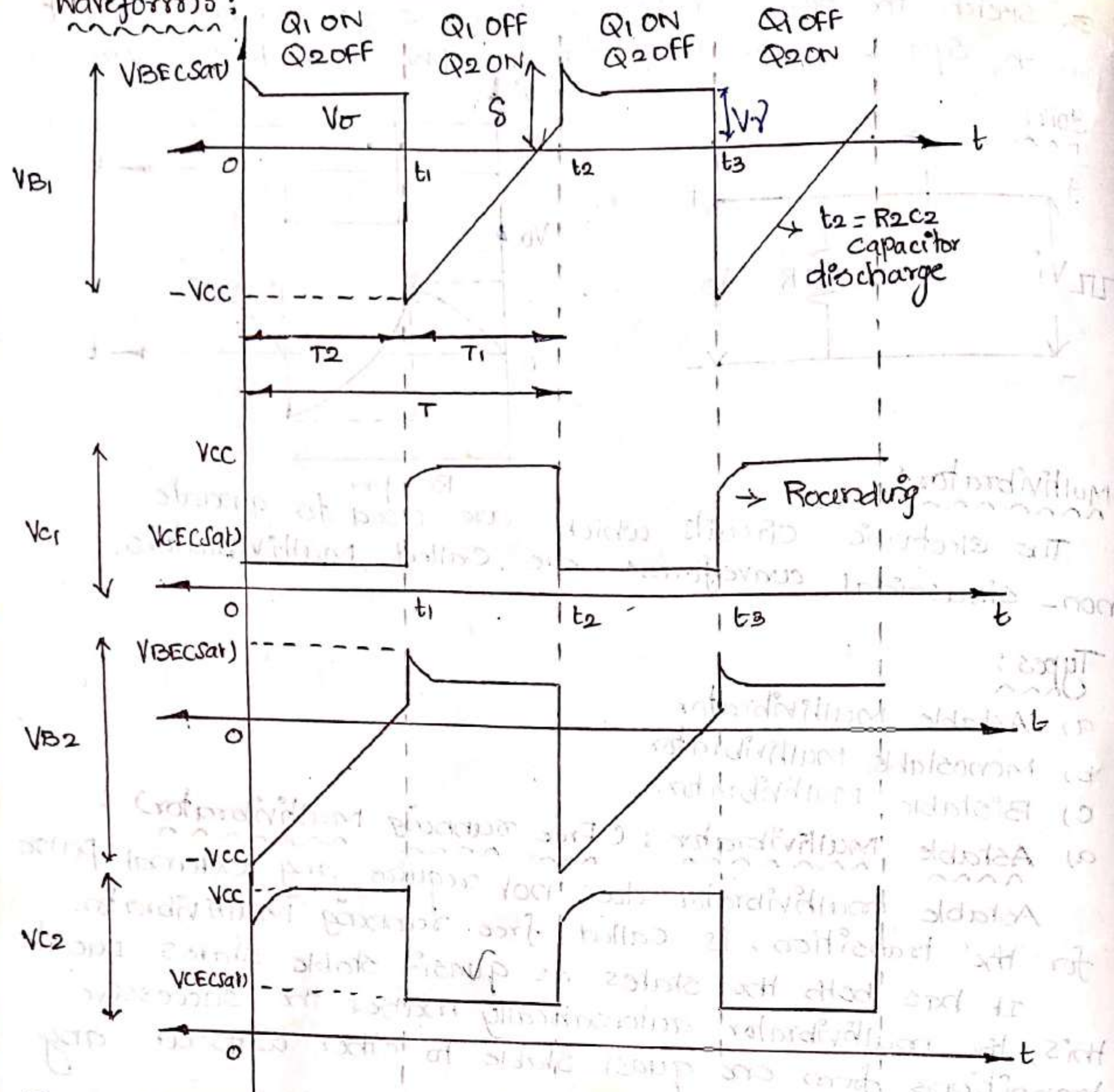
Q_1 - OFF

C_2 charges - C_1 discharges

C_2 gives charging and discharging input to base of Q_2. Using C_2 input, Q_2 gives square wave output.

Now C_1 discharges makes a Q_1 is ON. C_2 start discharge, and C_1 start charging. C_1 input comes base of Q_1. Q_1 gives one more square wave output.

Waveforms:



From the waveform, Derivation of Time Period:

$V_0 = V_{B2} = V_{BE}(sat)$, $V_{C2} = V_{CE}(sat)$, $V_{C1} = V_{CC}$

$V_i = -V_{CC}$, $V_f = +V_{CC}$, $\tau_2 = R_2C_2$, $\tau_1 = R_1C_1$

For the capacitor,

wkt, Basic Equation.

V_F - Final value

V_i - initial value

$\Rightarrow V_0 = V_f - (V_f - V_i)\, e^{-t/\tau}$

$\Rightarrow V_{B2} = V_{CC} - (V_{CC} - (-V_{CC}))\, e^{-t/R_2C_2}$ $\quad \because V_0 = V_{B2}$

$\Rightarrow V_{B2} = V_{CC} - 2V_{CC}\, e^{-t/R_2C_2}$

$\Rightarrow V_{B2} = V_{CC}(1 - 2e^{-t/R_2C_2})$

Now $t = T_2$, $V_{B2} = V_\gamma$

$\Rightarrow V_\gamma = V_{CC}(1 - 2e^{-T_2/R_2C_2})$

Let $V_\gamma = 0V$,

$\Rightarrow 0 = V_{CC}(1 - 2e^{-T_2/R_2C_2})$

$\Rightarrow 1 - 2e^{-T_2/R_2C_2} = 0$

$\Rightarrow e^{-T_2/R_2C_2} = 0.5$

Take log.

$\rightarrow \frac{-T_2}{R_2C_2} = \ln(0.5) \rightarrow \frac{-T_2}{R_2C_2} = -0.69 \rightarrow T_2 = 0.69 R_2C_2$

$|||^{ly}$ $T_1 = 0.69\ R_1C_1$

$\rightarrow T = T_1 + T_2$

$\rightarrow T = 0.69(R_1C_1 + R_2C_2)$

If $R_1 = R_2 = R$, $C_1 = C_2 = C$,

$\rightarrow T = 0.69(2)RC$

$\rightarrow T = 1.38\ RC$

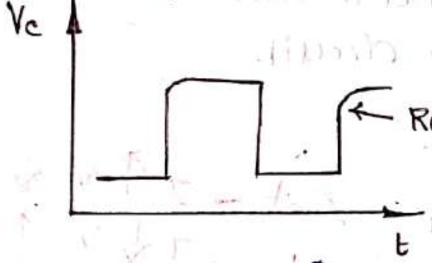
$I_2 = I_{C2} = \frac{V_{CC} - V_{C2}}{R_C}$

$I_{B2} = 1.5\ I_{B2}(min)$

$I_{B2}(min) = \frac{I_{C2}}{h_{fe}}$

$I_{B2} = \frac{V_{CC} - V_{B2}}{R_2}$

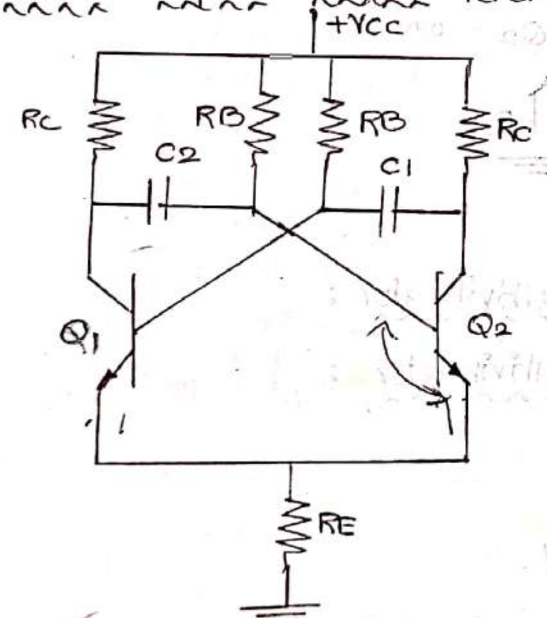
Rounding :



There is a certain distortion present in the output as a square wave of collector coupled Astable multivibrator is called rounding.

Such a rounding can be eliminated to obtain the vertical edges square wave by adding two collector diodes and two resistors.

Emitter coupled Astable Multivibrator :



→ Waveform
→ Derivation of Time period.

$RE = \frac{V_{CC}}{2I_C}$

(Same to collector coupled)

Application :

1. Used as square wave Generator
2. Clocks for Binary logic signals.
3. Digital Voltmeter and SMPS
4. Generate audio and radio freq.

Monostable Multivibrator :

The monostable has one stable state and one quasi stable state.

When an external trigger pulse is applied to the circuit, the circuit goes in to the quasi-stable state from its normal stable state. After some time interval, the circuit automatically returns to its stable state. The circuit does not require any external pulse to change from quasi to stable state.

It is also known as one shot, single shot, single cycle, single swing, single step multivibrator and univibrator., Gating or delay circuit.

It is used to generate rectangular waveform.

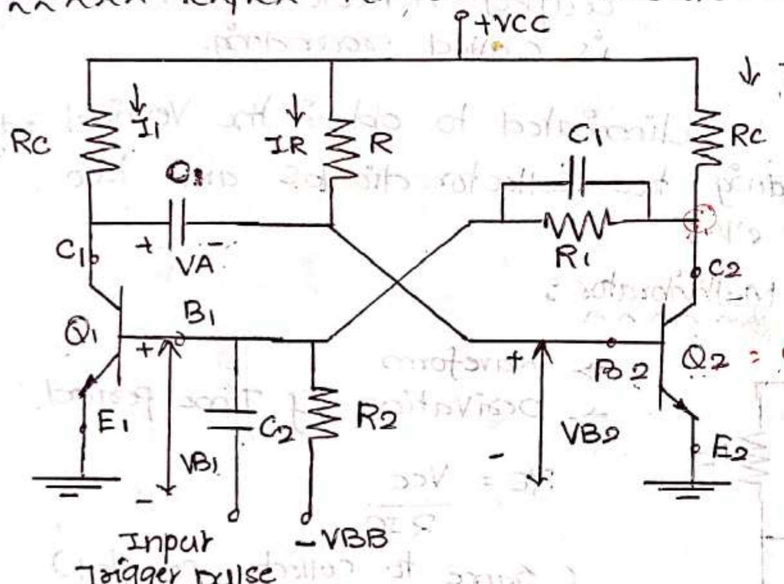
Gating Circuit :

The circuit is used to generate the rectangular waveform and can be used to gate other circuits is called gating circuit.

Delay Circuit

The time between the transition from quasi stable to stable state can be predetermined and it can be used to introduce time delay is called delay circuit.

Collector coupled Monostable Multivibrator :



Sat - I ↑ V = ↓

Cut - I ↓ V = ↑

operation:

From Vcc

- Q2 - ON, Q1 - off Because of -VBB make rev Bias of Base, Q1 - off (hole reg +)

Q2 ON, V_{C2} ↓.

Q_1 off V_{C1} ↑, C charges

charge cap o/p apt to i/p of B of Q2

Now

Apply pulse to B_1 of Q_1 → Q_1 - on

Q_1 on, V_{C1} ↓, cap ↓

Q_2 off, V_{C2} ↑, C.

Emitter coupled Monostable Multivibrator :

Triggering of Monostable Multivibrator :

→ Waveform

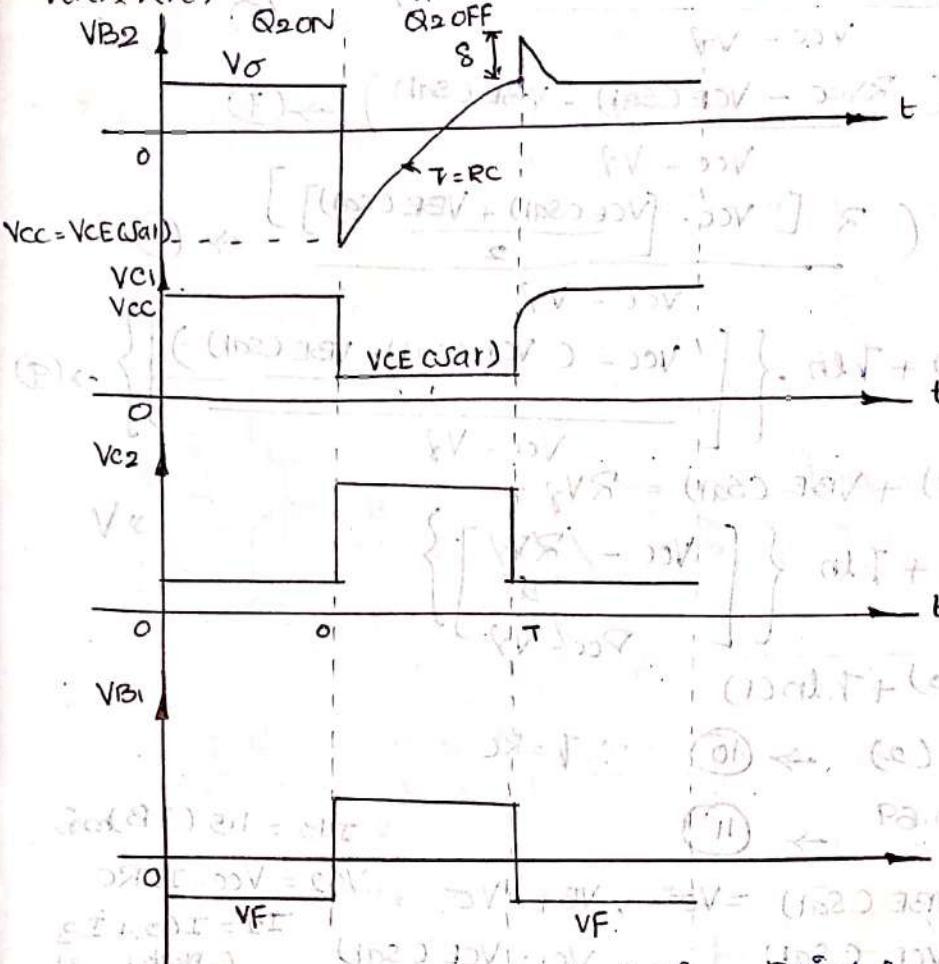
→ Time period

same for collector coupled.

Applications :

1. Used to generate adjustable pulse width generator
2. Generate uniform width pulse from variable width input pulse train
3. Used as time delay unit
4. Generate clean and sharp pulses from distorted pulse

Waveforms :



Derivation of pulse width and Time period :

From the waveform,

Let $V_i = V_\sigma - I_1 R_C$, $V_f = V_{cc}$, $V_o = V_c$,

wkt, Basic Equation,

$\rightarrow V_c = V_f + (V_i - V_f)\, e^{-t/\tau}$

Now, $V_c = V_{B2}$,

$\rightarrow V_{B2} = V_{cc} + (V_{cc} - V_\sigma + I_1 R_C)\, e^{-t/\tau} \rightarrow (1)$

at $t = T$, $V_{B2} = V_\gamma$

$\rightarrow V_\gamma = V_{cc} + (V_{cc} - V_\sigma + I_1 R_C)\, e^{-t/\tau}$

$\rightarrow V_\gamma - V_{cc} = (V_{cc} - V_\sigma + I_1 R_C)\, e^{-t/\tau}$

$\rightarrow$ Take log

$\rightarrow \ln(-V_{cc} + V_\gamma) = (V_{cc} - V_\sigma + I_1 R_C) - T/\tau$

$\rightarrow T = \tau \ln\left[\dfrac{V_{cc} + I_1 R_C - V_\sigma}{(V_{cc} - V_\gamma)}\right] \rightarrow (2)$ ∴ $V_\sigma = 0.3V$ for ge, $0.7V$ for Si

When Q_1 is satu,

$\rightarrow V_{C1} = V_{CE(sat)} \rightarrow (3)$

$\rightarrow I_1 R_C = V_{cc} - V_{CE(sat)} \rightarrow (4)$

$\rightarrow V_\sigma = V_{BE(sat)} \rightarrow (5)$

Sub (3), (4), (5) in (2)

$$\Rightarrow T = \tau \ln\left(\frac{V_{CC} + V_{CC} - V_{CE(sat)} - V_{BE(sat)}}{V_{CC} - V_\gamma}\right) \rightarrow (6)$$

$$\Rightarrow T = \tau \ln\left(\frac{2V_{CC} - V_{CE(sat)} - V_{BE(sat)}}{V_{CC} - V_\gamma}\right) \rightarrow (7)$$

$$\Rightarrow T = \tau \ln\left(\frac{2\left[V_{CC} - \left[\frac{V_{CE(sat)} + V_{BE(sat)}}{2}\right]\right]}{V_{CC} - V_\gamma}\right) \rightarrow (8)$$

$$\Rightarrow T = \tau \ln(2) + \tau \ln\left\{\left[\frac{V_{CC} - \left(\frac{V_{CE(sat)} + V_{BE(sat)}}{2}\right)}{V_{CC} - V_\gamma}\right]\right\} \rightarrow (9)$$

But $V_{CE(sat)} + V_{BE(sat)} = 2V_\gamma$

$$\Rightarrow T = \tau \ln(2) + \tau \ln\left\{\left[\frac{V_{CC} - \cancel{2}V_\gamma / \cancel{2}}{V_{CC} - V_\gamma}\right]\right\}$$

$$\Rightarrow T = \tau \ln(2) + \tau \ln(1)$$

$$\Rightarrow T = \tau \ln(2) \rightarrow (10) \quad \because \tau = RC$$

$$\Rightarrow T = RC\, 0.69 \rightarrow (11)$$

Now $V_{B2} = V_{BE(sat)} = V_\sigma$, $V_{B1} = V_\sigma$, 1. $V_{C2} = V_{CC} - I_2 R_C$

$V_{C2} = V_{CE(sat)}$, $V_{C1} = V_{CE(sat)}$ 2. $I_2 = I_{C2} + I_3$ (Neglect I_3)

$V_{C1} = V_{CC}$

and $I_{C2} = I_{C(sat)} = \dfrac{V_{CC} - V_{CE(sat)}}{R_C}$ $\because V_{CE(sat)} = 0.3V$

$V_\sigma = V_{BE(sat)} = 0.7V$

5. $I_B = 1.5\,(I_B)_{min}$

$\Rightarrow$ 4. $I_{B2\,sat} = \dfrac{I_{C(sat)}}{h_{fe}} = I_B$ Assume $I_{B1} = 0.3mA = I_4$

and 10. $\because I_3 = \dfrac{V_{CC} - V_\sigma}{R_1}$

$\Rightarrow$ 3. $I_{B2} = \dfrac{V_{CC} - V_{BE(sat)}}{R}$

$\Rightarrow$ 2. $I_3 = \dfrac{V_{CC} - V_\sigma}{R_1 + R_C}$, $I_4 = \dfrac{V_\sigma - V_{BB}}{R_2}$ 9. $\because I_3 = I_{B1} + I_4$

To find C_1,

6.1. $I_3 = \dfrac{V_{C2} - (-V_{BB})}{R_1 + R_2} = \dfrac{V_{CE(sat)} - V_{BE(cut)}}{R_1}$

$R_1 C_1 = 1\,\mu s$

Bistable Multivibrator:

The Bistable multivibrator has two stable states. It requires an external trigger pulse to change from one stable state to another. The circuit remains in one stable state unless an external trigger pulse is applied.

It is also known as Eccles - Jordan, trigger, scale-of-2 toggle, flip flop and binary.

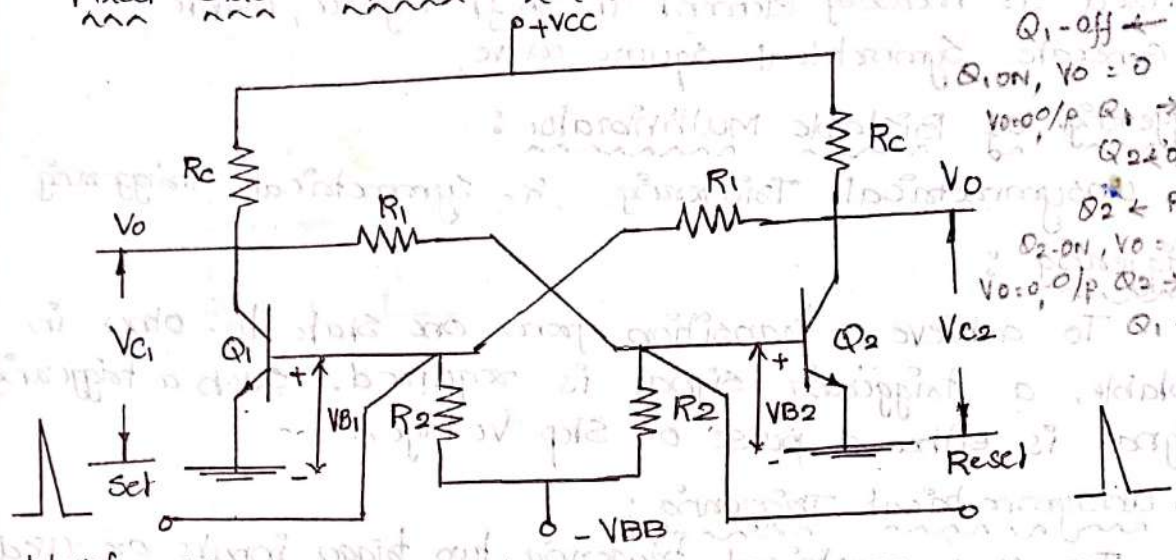
Used to generate pulse type waveforms.

Two Types:

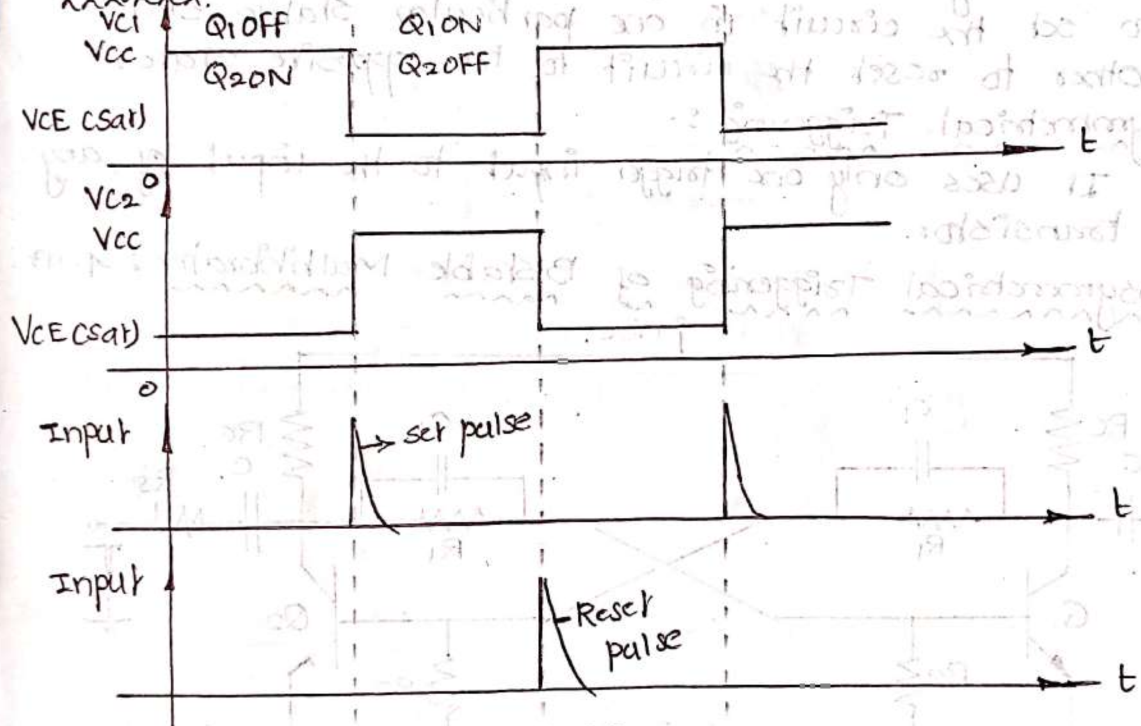
1. Fixed Bias 2. Self Bias.

1. Fixed Bias Bistable Multivibrator:

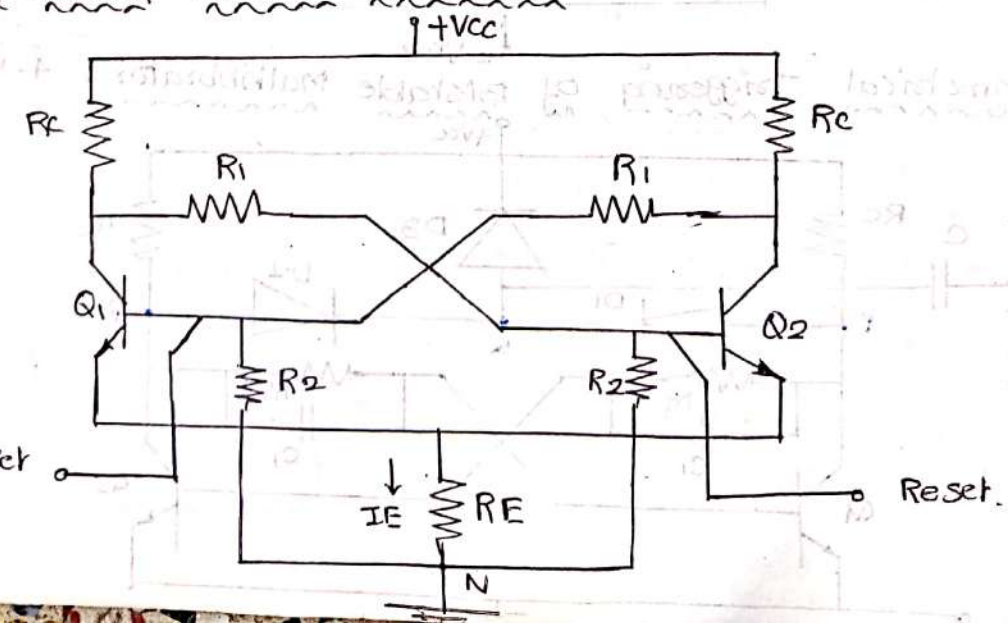
Assume
Q_2 ON, Q_1 off
Q_1 - off ← set
Q_1 ON, $V_o = 0$
$V_o = 0$ o/p Q_1 → i/p Q_2
Q_2 off
Q_2 ← Reset
Q_2-ON, $V_o = 0$
$V_o = 0$, o/p Q_2 → i/p Q_1
Q_1 - off



Waveforms:



2. Self Biased Bistable Multivibrator:



Applications:

1. Used as a frequency divider
2. Used in processing of pulse type waveform.
3. used as memory element in shift register, counter.
4. Generate symmetrical square wave.

Triggering of Bistable Multivibrator:

1. Unsymmetrical Triggering 2. Symmetrical Triggering

Triggering:

To achieve a transition from one state to other in Bistable, a triggering signal is required. Such a triggering signal is either a pulse or step voltage.

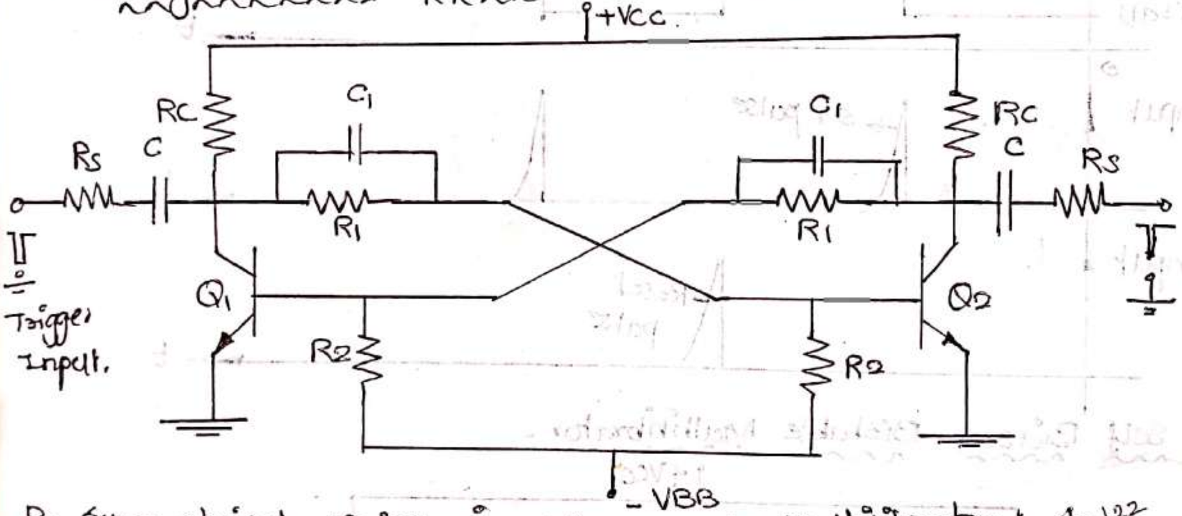
1. Unsymmetrical Triggering:

In unsymmetrical triggering, two trigger inputs are used. one to set the circuit in one particular stable state and other to reset the circuit to the opposite state.

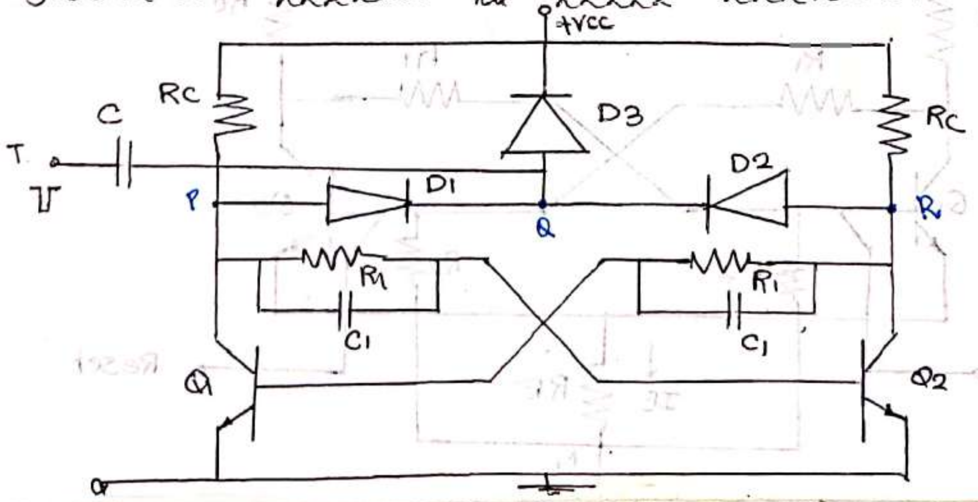
2. Symmetrical Triggering:

It uses only one trigger input to the input of any one transistor.

1. Unsymmetrical Triggering of Bistable Multivibrator: 4-118.



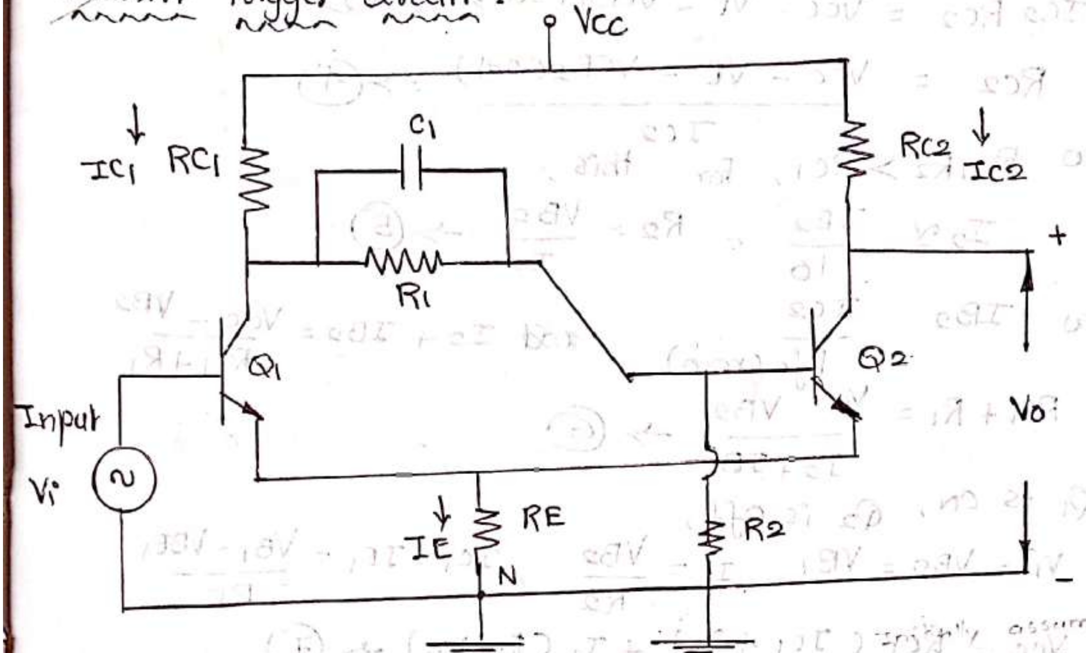
2. Symmetrical Triggering of Bistable Multivibrator: 4-122

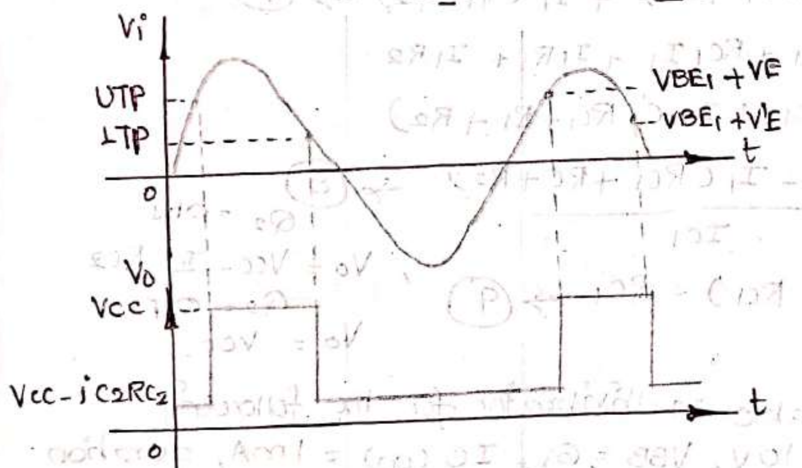


Complementary Multivibrator:

Two transistor multivibrator circuits it is possible to replace one of the transistors by its Complement. i.e. turning half the circuit upside down. so one transistor is npn while other is pnp is called complementary multivibrator.

Schmitt Trigger circuit:



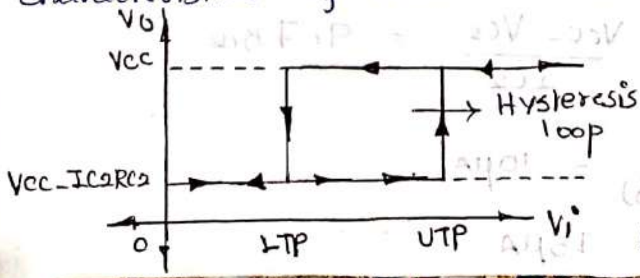


Initially assume Vi = 0. Vcc is on. Q2 is ON. Q1 is off i.e. Vi = 0 so Vo decreases. Now Vi ↑ Q1 - ON. Q2 goes to off. Vi ↑ when Q1 - ON, Q2 - off generate voltage UTP i.e. VBE1 + VE. After some time, Vi ↓ VCE ↑ if VCE ↑ it makes Q2 - ON. that time Q1 - off Vi ↓, Q2 - ON generates LTP i.e. VBE1 + V'E. Again Q2 ON. Vc is again decreases.

Hysteresis: (Dead band or Dead Zone).

The graph of output Voltage against input Voltage is Called transfer characteristics of schmitt trigger.

once o/p changes its state, it remains there indefinitely until the input voltage crosses any of the threshold voltage level. So when output changes its State from low to high at UTP. it remains there till input crosses LTP and vice Versa. This characteristics of schmitt trigger is called hysteresis.



It is also called dead band or Zone as there is no change in output in this zone though input changes. is called hysteresis loop.

The diff between UTP & LTP is called width of hysteresis

Design of schmitt trigger :

UTP must be equal to VB2 when Q2 is ON.

$\to V_i = V_{B2} = UTP \to (1)$

$\therefore V_E = V_i - V_{BE1} = V_{B2} - V_{BE2}$, $I_{E2} = I_{C2}$, $R_E = \frac{V_E}{I_{E2}} \to (2)$

Now, $V_{CE2} = 0.2V$ when Q_2 is ON,

$\to I_{C2}R_{C2} = V_{CC} - V_E - V_{CE2}(sat) \to (3)$

$\to R_{C2} = \frac{V_{CC} - V_E - V_{CE2}(sat)}{I_{C2}} \to (4)$

Now $R_1, R_2 > R_{C1}$, for this,

$I_2 \simeq \frac{I_{E2}}{10}$, $R_2 = \frac{V_{B2}}{I_2} \to (5)$

Now $I_{B2} = \frac{I_{C2}}{h_{fe}(min)}$ and $I_2 + I_{B2} = \frac{V_{CC} - V_{B2}}{R_{C1} + R_1}$

$\therefore R_{C1} + R_1 = \frac{V_{CC} - V_{B2}}{I_2 + I_{B2}} \to (6)$

Now Q_1 is ON, Q_2 is off,

$\to V_i = V_{B2} = V_{B1}$, $I_1 = \frac{V_{B2}}{R_2}$, $I_{C1} = I_{E1} = \frac{V_{B1} - V_{BE1}}{R_E}$

$\to V_{CC} = R_{C1}(I_{C1} + I_1) + I_1(R_1 + R_2) \to (7)$

$\to V_{CC} = R_{C1}I_{C1} + R_{C1}I_1 + I_1R_1 + I_1R_2$

$\to V_{CC} = R_{C1}I_{C1} + I_1(R_{C1} + R_1 + R_2)$

$\to R_{C1} = \frac{V_{CC} - I_1(R_{C1} + R_C + R_2)}{I_{C1}} \to (8)$

and

$\to R_1 = (R_1 + R_{C1}) - R_{C1} \to (9)$

$Q_2 = ON$, $V_o = V_{CC} - I_{C2}R_{C2}$

$Q_2 = OFF$, $V_o = V_{CC}$.

Problem :

1. Design the monostable multivibrator for the following specifications : $V_{CC} = 10V$, $V_{BB} = 6V$, $I_C(on) = 1mA$, duration of output pulse = 14 msec, $h_{fe}(min) = 100$, $I_{CBO} = 0$, $V_{BE}(off) = -0.5V$.

Soln Given : $V_{CC} = 10V$, $-V_{BB} = -6V$, $I_C(ON) = 1mA = I_{C2}$, $T = 14ms$, $h_{fe} = 100$, $I_{CBO} = 0$, $V_{BE}(off) = -0.5V$

Assume $V_{CE}(sat) = 0.3V$, $V_{BE}(sat) = 0.7V$.

(V_{C2})

$\to V_{C2} = V_{CC} - I_2R_C$

$\to V_{C2} - V_{CC} = -I_2R_C$

$\to R_C = \frac{V_{CC} - V_{C2}}{I_2} = \frac{V_{CC} - V_{C2}}{I_{C2}} = 9.7k\Omega$

and

$\to I_{B2}(min) = \frac{I_C}{h_{fe}(min)} = 10\mu A$

$\to I_{B2} = 1.5(I_{B2})_{min} = 15\mu A$

cont. $I_{B2} = \frac{V_{CC} - V_{BE(sat)}}{R} \Rightarrow 15\mu = \frac{10 - 0.7}{R}$

$\Rightarrow R = 620 k\Omega$

and

$T = 0.69 RC$

and $C = 32.72 nF$

$I_3 = \frac{V_{C2} - (-V_{BB})}{R_1 + R_2} = \frac{V_{CE(sat)} - V_{BE(cutoff)}}{R_1} = \frac{0.3+6}{R_1+R_2} = \frac{0.3-(0.}{R_1}$

$\rightarrow 6.3R_1 = 0.8R_1 + 0.8R_2$

$\rightarrow 6.3R_1 - 0.8R_1 = 0.8R_2$

$\rightarrow R_1 = R_2 \frac{0.8}{5.5} \rightarrow R_1 = 0.1454 R_2$

$I_3 = \frac{V_{CC} - V_O}{R_1} = \frac{9.3}{R_1}$

$I_4 = 6.7/R_2$

$I_4 = 3\times10^{-4} = I_{B1(sat)}$

$R_2 = \frac{6.7}{3\times10^{-4}} = 2.23\times10^{4}$

$I_3 = I_{B1} + I_4$

$I_3 = 6\times10^{-4}$

$R_1 = \frac{9.3}{I_3} = \frac{9.3}{6\times10^{-4}}$

and $I_3 = \frac{V_{CC} - V_{BE(sat)}}{R_1 + R_C} = \frac{9.3}{R_1 + 9.7K}$

& $I_4 = \frac{V_{BE(sat)} - (-V_{BB})}{R_2} = \frac{0.7+6}{R_2} = \frac{6.7}{R_2}$

& $I_{B1} = I_3 - I_4 = \frac{9.3}{R_1 + 9.7k} - \frac{6.7}{R_2}$

$\rightarrow 15\times10^{-6} = \frac{9.3}{R_1+9.7k} - \frac{6.7}{R_2}$

$\rightarrow 15\times10^{-6}[R_2][R_1 + 9.7k] = 9.3R_2 - 6.7(R_1+9.7k)$

$\rightarrow 15\times10^{-6}R_1R_2 + 0.1455R_2 = 9.3R_2 - 6.7R_1 - 64990$

$\rightarrow 15\times10^{-6}\times0.1454R_2^2 + 0.1455R_2 = 9.3R_2 - 6.7\times0.1454R_2 - 64990$

$\rightarrow 2.181\times10^{-6}R_2^2 - 8.1803R_2 + 64990 = 0$

$\rightarrow R_2 = \frac{8.1803 \pm \sqrt{(8.1803)^2 - 4\times2.181\times10^{-6}\times64990}}{2\times2.181\times10^{-6}}$

$\rightarrow R_2 = 3.7427 M$ (or) $7.9615 k\Omega$.

and

$R_1 = 0.1454 R_2 = 0.1454 \times 3.7427 M$

$R_1 = 544.19 k\Omega$.

and $R_1C_1 = 1\mu s$.

$C_1 = \frac{1\mu s}{R_1} \qquad C_1 = 1.8 pF$

R. Design a collector coupled astable multivibrator for the following specifications: output voltage 10V peak, $I_{C(ON)} = 1mA$, $h_{fe(min)} = 100$, $I_{CBO} = 0$, output to a positive pulse, the duration of which is 20 μsec, the time between the pulses to be 10μ sec.

Soln Given $V_{CC} = 10V$, $I_{C(ON)} = 1mA$, $h_{fe(min)} = 100$, $I_{CBO} = 0$

$T_1 = 10\mu sec$ $T_2 = 20\mu s$.

Assume, $V_{CE(sat)} = V_{C2} = 0.3V$, $V_{BE(sat)} = V_{B2} = 0.7V$

$I_2 = I_{C2} = I_{C(ON)} = 1mA$

$\rightarrow I_2 = \frac{V_{CC} - V_{C2}}{R_C} \rightarrow 1mA = \frac{10 - 0.3V}{R_C}$

$\rightarrow R_C = 9.7k\Omega$

Now, $I_{B2} = 1.5 (I_{B2(min)}) = 1.5 \frac{I_{C2}}{h_{fe}} = \frac{1.5 \times 1m}{100} = 15\mu A$

$\rightarrow I_{B2} = \frac{V_{CC} - V_{B2}}{R_2} = \frac{10 - 0.7}{R_2}$

$\rightarrow R_2 = 620k\Omega$

$\rightarrow T = 0.69 RC$, $T_1 = 0.69 R_1C_1$, $T_2 = 0.69 R_2C_2$

$\rightarrow T_1 = 10\mu s$, $T_1 = 0.69 R_1 C_1 \Rightarrow 10\mu = 0.69 R_1 \times 46.75p$,

$R_1 = 310k\Omega$.

$\rightarrow T_2 = 20\mu s$, $T_2 = 0.69 R_2C_2 \Rightarrow 20\mu = 0.69 \times 620k\ C_2$

$C_2 = 46.75\ pF = C_1 = C$.

3. Calculate the component values of a monostable multivibrator developing an output pulse of $140\mu s$ duration. Assume $h_{fe(min)} = 20$, $I_{C(sat)} = 6mA$, $V_{CC} = 6V$, $V_{BB} = -1.5V$. Draw the circuit for monostable multivibrator with calculated values

Soln Given $I_{C2} = I_{C(sat)} = 6mA$, $T = 140\mu s$, $h_{fe(min)} = 20$

$V_{CC} = 6V$, $V_{BB} = -1.5V$.

Assume, $V_{CE(sat)} = 0.3V$, $V_{BE(sat)} = 0.7V = V_\sigma$
(V_{C2})

$\rightarrow V_{C2} = V_{CC} - I_2 R_C$

$\rightarrow R_C = 950\Omega$

$\rightarrow I_{B2(min)} = \frac{I_C}{h_{fe(min)}} = 3\times10^{-4}A$,

$\rightarrow I_{B2} = \frac{V_{CC} - V_{BE(sat)}}{R}$

$\rightarrow R = 17.67k\Omega$

$\rightarrow T = 0.69 RC$

$\rightarrow C = 11.484\ nF$.

$\rightarrow I_3 = \frac{V_{CC} - V_\sigma}{R_1} = \frac{5.3}{R_1}$, $I_4 = \frac{V_\sigma - V_{BB}}{R_2} = \frac{2.2}{R_2}$

Assume $I_4 = I_{B1(sat)} = 3\times10^{-4}A$

$\rightarrow I_4 = \frac{2.2}{R_2} \rightarrow R_2 = 7.33k\Omega$

Now, $I_3 = I_{B1} + I_4 = 6\times10^{-4}$.

$\rightarrow I_3 = \frac{5.3}{R_1} \rightarrow R_1 = 8.833 k\Omega$ (28)

$\rightarrow R_1 C_1 = 1 \mu s$

$\rightarrow C_1 = 113.21\ pF.$

4. Design a schmitt trigger circuit for $V_{CC} = 10V$, $UTP = 5V$, $LTP = 3V$, Assume $h_{fe(min)} = 100$ and $I_{C(ON)} = 1mA$.

Soln Given $V_{CC} = 10V$, $UTP = 5V$, $LTP = 3V$, $h_{fe(min)} = 100$, $I_{C(ON)} = 1mA$

Ass $V_{BE2} = 0.7V$, $V_{CE2} = 0.2V$

1. $V_i = V_{B2} = UTP = 5V$

2. $V_E = V_i - V_{BE1} = V_{B2} - V_{BE2} = 5 - 0.7 = 4.3V$

3. $I_{C2} = I_{E2} = 1mA$

4. $R_E = \frac{V_E}{I_{E2}} = 4.3 k\Omega$

5. $I_{C2} R_{C2} = V_{CC} - V_E - V_{CE2(sat)}$

$R_{C2} = 5.5K$

6. $I_2 = \frac{I_{E2}}{10} = 1 \times 10^{-4} A$

7. $R_2 = \frac{V_{B2}}{I_2} = 50 k\Omega$

8. $I_{B2} = \frac{I_{C2}}{h_{fe(min)}} = 10 \mu A.$

9. $R_{C1} + R_1 = \frac{V_{CC} - V_{B2}}{I_2 + I_{B2}} = 45.45 \times 10^3$

9. $V_i = \cancel{V_{B2}} = V_{B1} = LTP = 3V,$

10. $I_1 = \frac{V_{B2}}{R_2} = 60 \mu A.$

11. $I_{C1} = I_{E1} = \frac{V_{B1} - V_{BE1}}{R_E} = \frac{3 - 0.7}{4.3K} = 5.3 \times 10^{-4}$

12. $V_{CC} = R_{C1}(I_{C1} + I_1) + I_1(R_1 + R_2)$

$R_{C1} = \frac{V_{CC} - I_1(R_{C1} + R_C + R_2)}{I_{C1}} = 7.98 k\Omega.$

$\because V_{CC} = I_{C1} R_{C1} + I_1 (R_1 + R_{C1}) + I_1 R_2$

$R_{C1} = 7.98 k\Omega$

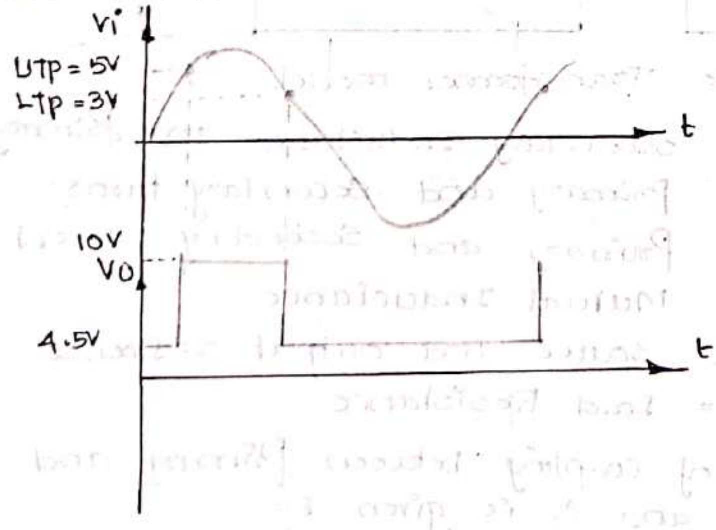
$\because R_{C1} + R_1 = 45.45 \times 10^3$

$R_1 = 37.465K$

13. $R_1 = (R_1 + R_{C1}) - R_{C1} = 37.465 k\Omega.$

14. Q_2 ON, $V_o = V_{CC} - I_{C2} R_{C2} = 4.5V$

15. Q_2 OFF, $V_o = V_{CC} = 10V.$



UNIT - V

BLOCKING OSCILLATORS AND TIME BASE GENERATORS:

UJT Sawtooth Waveform Generator : (Bakshi) Relaxation oscillator.(UJT)

Blocking oscillator:

A special type of wave generator which is used to produce a single narrow pulse or train of pulses is called a blocking oscillator.

Two important elements of blocking oscillator are

1. an active devices like transistor

2. a pulse transformer.

Applications:

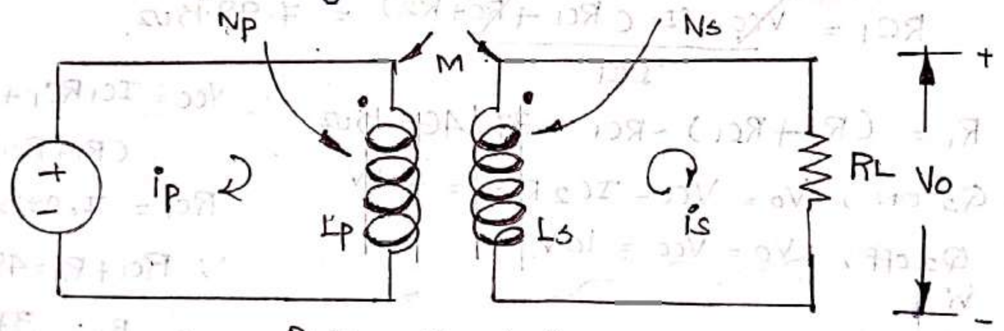
Frequency divider and counter circuits

Pulse transformer:

A pulse transformer is used to couple output of the transistor back to the input. The nature of such feedback through pulse transformer is controlled by relative winding polarities of a pulse transformer.

Characteristics

1. Iron cored and small in size
2. Leakage Inductance is minimum.
3. Cores have high permeability.
4. Have high magnetising inductance.



Ideal Pulse Transformer model.

Let
L_p and L_s = secondary Inductance and Primary Ind

N_p and N_s = Primary and secondary turns.

i_p and i_s = Primary and secondary current.

M = Mutual Inductance

V_i and V_o = source and output response.

R_L = Load Resistance

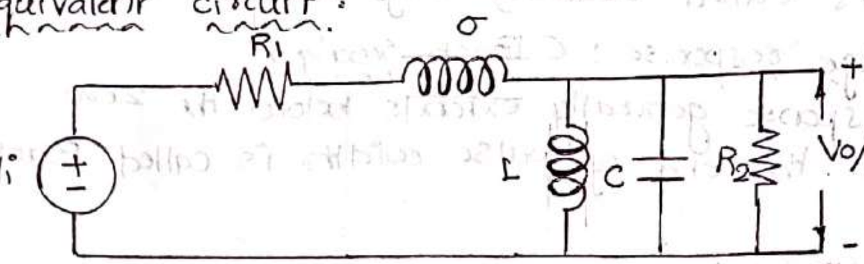
The co-efficient of coupling between primary and secondary is K. and it is given by

$\Rightarrow K = \frac{M}{\sqrt{L_P L_S}}$ $\because K = 1$

Let $n = N_S / N_P$ = transformer ratio.

and $\frac{V_o}{V_i} = \frac{i_P}{i_S} = \frac{N_S}{N_P} = n = \sqrt{\frac{L_S}{L_P}}$

Equivalent circuit:



Equivalent circuit of pulse Transformer.

In Equivalent circuit, transformer winding are removed and various values and parameters on secondary side are reflected to primary side.

Reflected output voltage on primary side is V_o/n and is given by

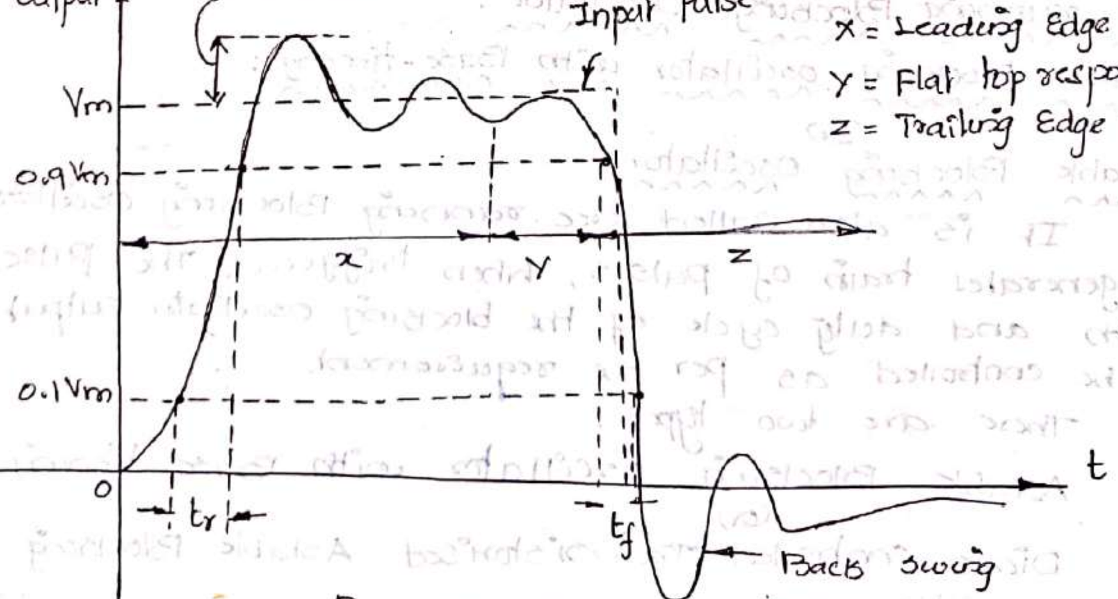
$\Rightarrow R_2 = \frac{R_2' + R_L}{n^2}$ $\because$ R_L = Load Resistance
R_2' = secondary winding
R_2 = Total resistance
n = Turn ratio.

The total Effective shunt capacitance reflected on primary side is denoted as c.

The total Leakage inductance is denoted as series inductance σ. and Magnetizing inductance is shunt inductance L.

The Resistance R_1 is the combined effect of primary winding resistance and source impedance which is resistive.

Pulse Response characteristics:



Pulse Transformer response

The distorted output response of a pulse transformer is shown above.

1. Leading edge response :

At start there is an overshoot and then the pulse settles down. The response till it settles down after the overshoot is called Leading Edge response.

2. Trailing edge response : (Back Swing)

The response generally extends below the zero amplitude after the end of pulse width is called Back swing.

3. Flat top response :

The portion of the response between the trailing edge and Leading edge is called flat top response.

4. Droop (or) tilt :

The displacement of the pulse amplitude during its flat response.

Applications of pulse Transformer :

1. used as Blocking oscillator
2. For coupling the stages of pulse transformer.
3. used in digital signal transmission.
4. To change the amplitude and Impedance level of pulse
5. Used in Transmission Lines
6. To provide dc isolation between source and load.
7. To differentiate a pulse.

Blocking Oscillator :

Free running Blocking Oscillator :

Astable Blocking oscillator with Base-timing :

(or)

Astable Blocking oscillator :

It is also called free running Blocking oscillator. It generates train of pulses, when triggered. The pulse width and duty cycle of the blocking oscillator output can be controlled as per the requirement.

There are two types :

1. Astable Blocking oscillator with Base timing

(or)

Diode controlled Transistorised Astable Blocking oscillator.

Power Amp & DC Convertr

UNIT- IV

LARGE SIGNAL AMPLIFIERS

Large Signal Amplifier : (Power Amplifier)

Features and Applications :

In multistage amplifier, we connected many stages in cascade. The input is a sound signal of human speaker and output is given to the Loud speaker.

The output last stage capable of handling large voltages or current or signals. It develops and gives sufficient power to the load like Loudspeaker, servomotor, handling large signals is called Large signal amplifier or power Amplifier.

Applications are radio receivers, tape players, TV receivers, cathode ray tube, Driving servomotor in industrial control system.

Classification of Large signal or power Amplifiers :

The position of the Q-point on the load line decides the class of operation of the power amplifier.

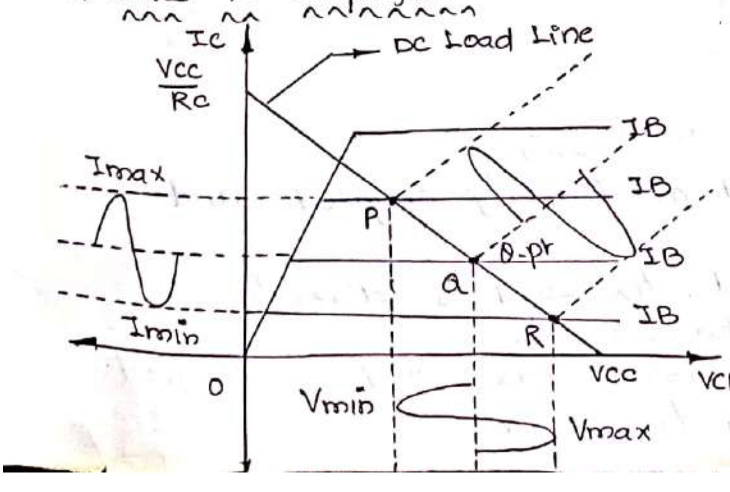
The various classes of power amplifier are

i. class A ii) Class B iii. Class C iv. Class AB v. class D

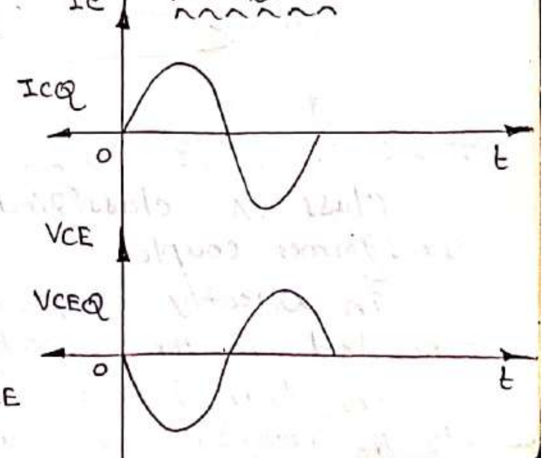
Comparison of power amplifiers :

Class	A	B	C	AB
operating cycle	360°	180°	Less than 180°	180° or 360°
Position OF Q-Point	Centre of Load Line	on x-axis	Below x-axis	Above x-axis but below the center of load line
Efficiency	25 %. to 50 %.	Better 78.5 %.	High	Higher than A but less than B 50 - 78.5 %.
Distortion	Absent No distortion	Present	Highest	Present

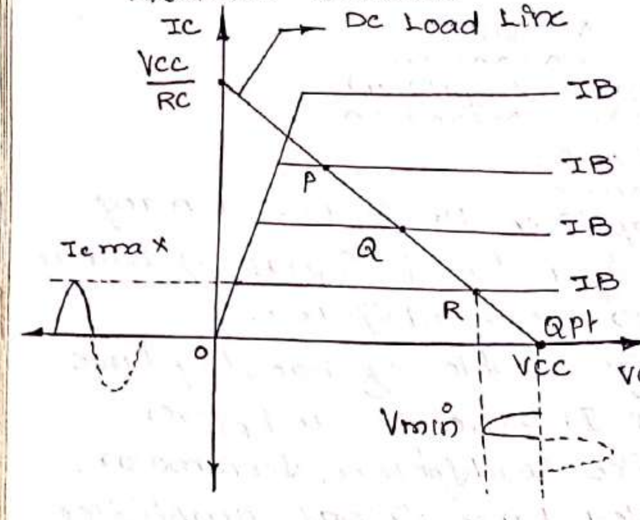
1. class A amplifier :

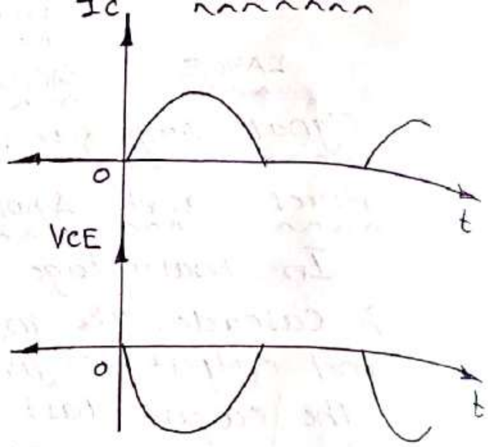


Waveforms :-



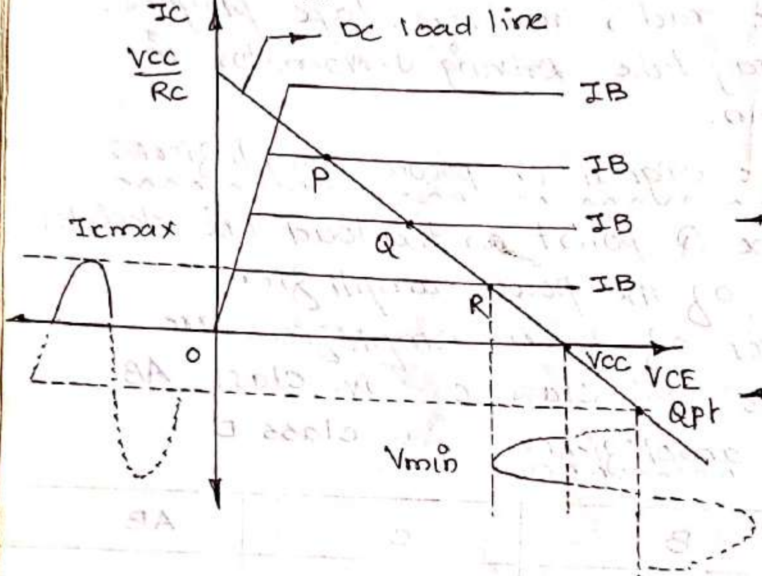
2. Class B amplifier:

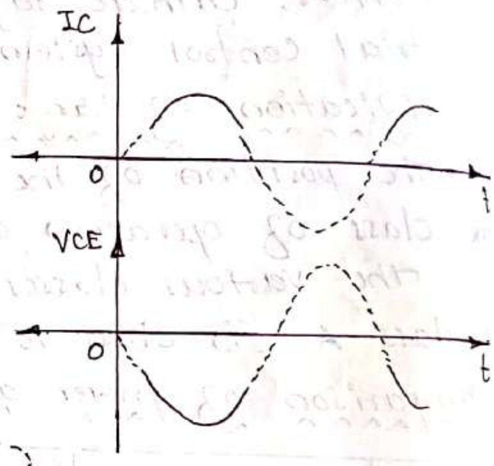




3. Class C amplifier:

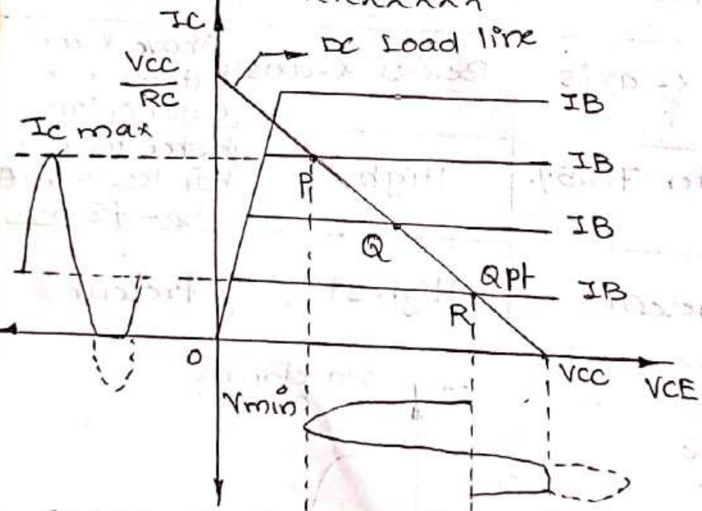
Waveforms:

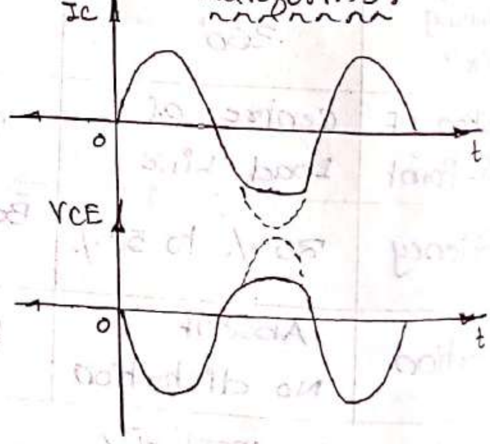




4. Class AB amplifier:

Waveforms:





CLASS A Large signal amplifiers:

Class A classified as Directly coupled and Transformer coupled.

In Directly coupled, the Load is directly connected in the collector circuits.

In transformer coupled, the load is coupled to the collector using a transformer.

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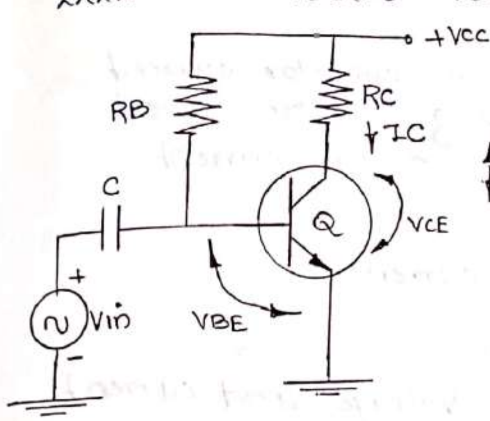
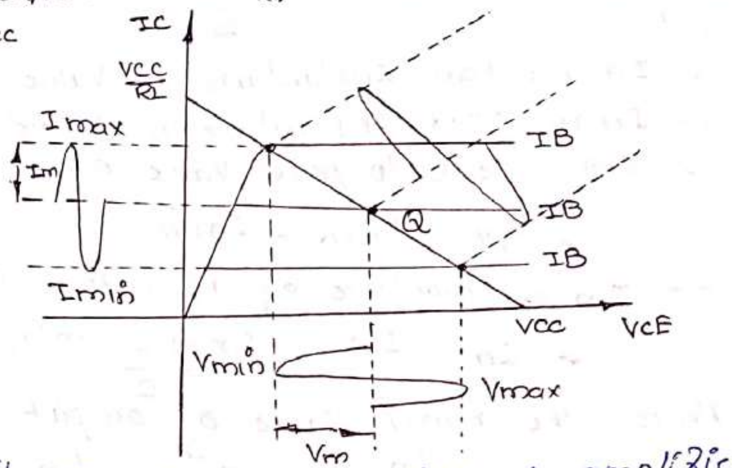
Series Fed, Directly Coupled class A amplifier : CRC coupled.





A Fixed Bias circuit can be used as class A amplifier. The Value of RB is selected in such a way that Q-point lies at the centre of dc load line. Here RL is directly connected in collector circuit.

Apply KVL to the CE Loop,

$\rightarrow$ $V_{CC} - I_C R_L - V_{CE} \rightarrow$ (1)

$\rightarrow$ $V_{CC} = I_C R_L + V_{CE}$

$\rightarrow$ $I_C = \left[\frac{V_{CC}}{R_L} - \frac{V_{CE}}{R_L} \right]$

1. D.C. operation :

Apply KVL to BE Loop,

$\rightarrow$ $V_{CC} - I_B R_B - V_{BE} = 0$

$\rightarrow$ $-I_B R_B = -V_{CC} + V_{BE}$

$\rightarrow$ $I_B = \frac{V_{CC} - V_{BE}}{R_B} \rightarrow$ (2)

wkt, $I_C = \beta I_B \rightarrow$ (3)

From (1)

$\rightarrow$ $V_{CE} = V_{CC} - I_C R_L \rightarrow$ (4)

Q point are

$\rightarrow$ $I_{CQ} = \beta I_B$

$\rightarrow$ $V_{CEQ} = V_{CC} - I_C R_L$

2. D.C Power Input (P_{dc}) :

The d.c power Input is provided by the supply with no ac input signal, the dc current drawn is the collector bias current I_{CQ}. Hence dc power input is

$\rightarrow$ $P_{dc} = V_{CC} \cdot I_{CQ} \rightarrow$ (5)

3. A.C operation :

The output current i.e) collector current varies around its quiescent value while the output voltage i.e) collector to Emitter voltage varies around its quiescent value. The Varying output voltage and current deliver an a.c power to the load.

4. A.c power Input :

For an alternating output voltage and current shown in the waveform. We can write

V_{min} = Min. Instantaneous Value of collector voltage

V_{max} = Max. Instantaneous Value of Collector Voltage

V_{PP} = peak to peak Value of ac o/p voltage across the load

$\rightarrow$ $V_{PP} = V_{max} - V_{min}$

V_m = Amplitude of ac output Voltage

$\Rightarrow V_m = \frac{V_{PP}}{2} = \frac{V_{max} - V_{min}}{2}$

III^{ly}

$\rightarrow I_{min}$ = Min. Instantaneous Value of collector current

$\rightarrow I_{max}$ = Max. Instantaneous Value of Collector current

$\rightarrow I_{PP}$ = peak to peak Value of ac output current

$\rightarrow I_{PP} = I_{max} - I_{min}$

$\rightarrow I_m$ = Amplitude of ac output current

$\rightarrow I_m = \frac{I_{PP}}{2} = \frac{I_{max} - I_{min}}{2}$

Hence, the r.m.s value of output voltage and current is $V_{rms} = \frac{V_m}{\sqrt{2}}$, $I_{rms} = \frac{I_m}{\sqrt{2}}$

$\rightarrow V_{rms} = I_{rms} R_L$ and $I_{rms} = V_{rms}/R_L$

$\rightarrow V_m = I_m R_L$ and $I_m = V_m/R_L$

The ac power delivered to the load is

① using r.m.s Value

$\rightarrow P_{ac} = V_{rms} . I_{rms}$.

or $\rightarrow P_{ac} = I^2_{rms} . R_L$ $\left\{ \because V_{rms} = I_{rms} . R_L \right.$

or $\rightarrow P_{ac} = V^2_{rms}/R_L$

② using peak value

$\rightarrow P_{ac} = V_{rms} . I_{rms}$

$\rightarrow P_{ac} = \frac{V_m}{\sqrt{2}} . \frac{I_m}{\sqrt{2}} = \frac{V_m I_m}{2}$

$\rightarrow P_{ac} = \frac{V_m^2}{2R_L} = \frac{I_m^2 R_L}{2}$

③ using peak to peak Value :

$\rightarrow P_{ac} = \frac{V_m I_m}{2} = \frac{\frac{V_{PP}}{2} \frac{I_{PP}}{2}}{2} = \frac{V_{PP} I_{PP}}{8}$

$\rightarrow P_{ac} = \frac{V^2_{PP}}{8R_L}$ or $\frac{I^2_{PP} R_L}{8}$ $\because \left\{ \begin{array}{l} V_{PP} = V_{max} - V_{min} \\ I_{PP} = I_{max} - I_{min} \end{array} \right.$

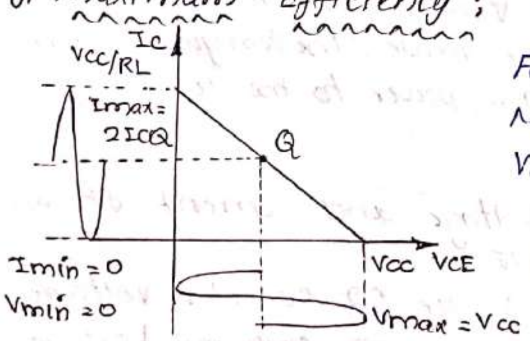
$\Rightarrow P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}$

5. Efficiency (η) :

Ratio of a.c power delivered to the load to the d.c power Input.

$\% \eta = \frac{P_{ac}}{P_{dc}} \times 100 = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8 V_{CC} I_{CQ}} \times 100.$

6. Maximum Efficiency ;



For Maximum Efficiency, Assume Maximum and Minimum output Voltage and current are

$V_{max} = V_{CC}$, $V_{min} = 0$

$I_{max} = 2I_{CQ}$, $I_{min} = 0$

$\rightarrow \% \eta_{max} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8 V_{CC} I_{CQ}} \times 100$

$\rightarrow \% \eta_{max} = \frac{(V_{CC} - 0)(2I_{CQ} - 0)}{8 V_{CC} I_{CQ}} \times 100 = 25\%$

7. Power Dissipation (P_d):

It is the difference between dc power input and ac power delivered to the load.

$\rightarrow P_D = P_{DC} - P_{ac}$

and

$\rightarrow P_{D(max)} = V_{CC} I_{CQ}$

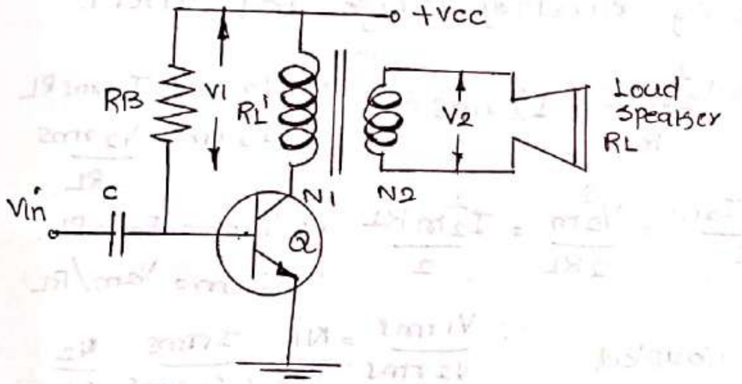
8. Advantages:

1. Circuit is simple
2. Easy to design and Implement.
3. Less Number of components is required
4. Circuit is cheaper.

9. Disadvantages:

1. Efficiency is very poor.
2. Power Dissipation is more.

Transformer coupled Class A audio power amplifier:



Turn Ratio, $n = \frac{N_2}{N_1}$

1. DC operation:

Assume winding Resistances are zero ohms. There is no dc voltage drop across primary winding. The slope of the dc load line is reciprocal of dc resistance in collector circuit, which is zero. So the dc load line is a vertically straight line.

Apply KVL to CE,

$\Rightarrow V_{CC} - V_{CE} = 0$

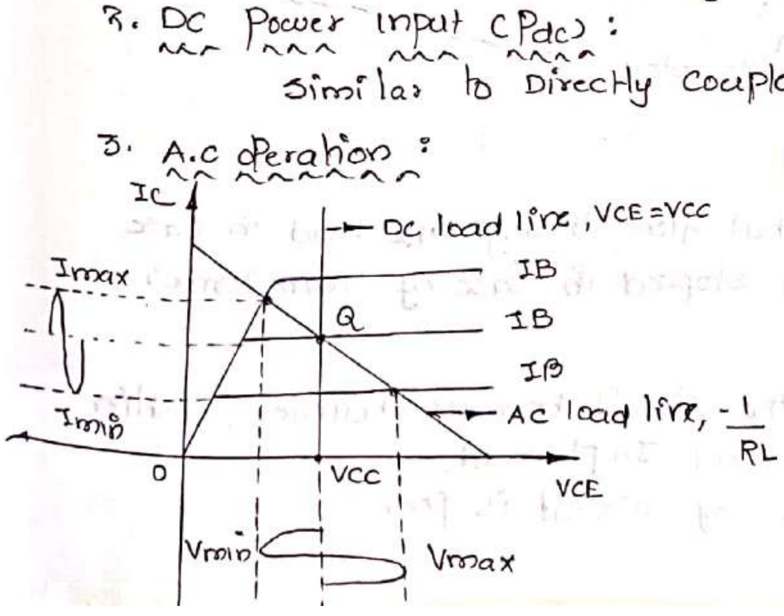
$\Rightarrow V_{CC} = V_{CE}$

Hence, dc load line is a vertical straight line passing through a voltage point on x-axis which is $V_{CEQ} = V_{CC}$.

2. DC Power input (P_{dc}):

Similar to Directly coupled.

3. A.C operation:



For ac purpose, the secondary is RL and primary is RL'. So ac load line is $-1/R_L$ slope passing through Q-point. When ac input is applied the Ic varies around its Q value I_{CQ} and also output voltage varies sinusoidally around its Q-value V_{CEQ}, which is Vcc in case.

4. AC output Power : P_{ac} :

In ac output power, Primary side of transformer consider

V_{1m} = Peak Value of Primary Voltage

V_{1rms} = R.M.S Value of Primary Voltage

I_{1m} = Peak Value of Primary current

I_{1rms} = R.M.S Value of Primary Current

hence ac power developed on primary is

$$\rightarrow P_{ac} = V_{1rms} I_{1rms} = \frac{V_{1rms}^2}{R_L'} = I_{1rms}^2 R_L' \qquad \because V_{1rms} = I_{1rms} R_L' \quad \because I_{1rms} = \frac{V_{1rms}}{R_L'}$$

(or)

$$\rightarrow P_{ac} = \frac{V_{1m}}{\sqrt{2}} \cdot \frac{I_{1m}}{\sqrt{2}} = \frac{V_{1m} I_{1m}}{2} = \frac{V_{1m}^2}{2R_L'} = \frac{I_{1m}^2 R_L'}{2} \qquad \because R_L' = R_L/n^2$$

$$\because V_{1m} = I_{1m} R_L' \quad \because I_{1m} = \frac{V_{1m}}{R_L'}$$

III'ly ac power developed on secondary is consider

$V_{2m} = I_{2m}$ = Peak Value of Secondary Voltage and current

$V_{2rms} = I_{2rms}$ = Rms Value of Secondary Voltage and current.

$$\rightarrow P_{ac} = V_{2rms} I_{2rms} = \frac{V_{2rms}^2}{R_L} = I_{2rms}^2 R_L \qquad \because V_{2rms} = I_{2rms} R_L \quad \because I_{2rms} = \frac{V_{2rms}}{R_L}$$

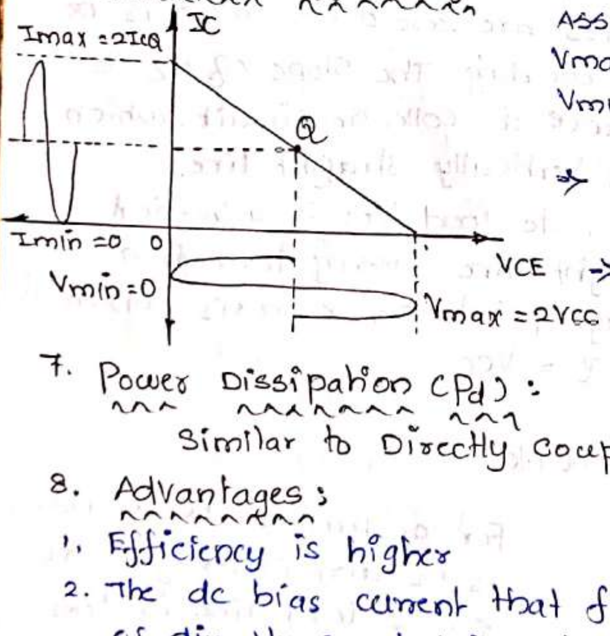
(or)

$$\rightarrow P_{ac} = \frac{V_{2m}}{\sqrt{2}} \cdot \frac{I_{2m}}{\sqrt{2}} = \frac{V_{2m} I_{2m}}{2} = \frac{V_{2m}^2}{2R_L} = \frac{I_{2m}^2 R_L}{2} \qquad \because V_{2m} = I_{2m} R_L \quad \because I_{2m} = V_{2m}/R_L$$

5. Efficiency (η) :

Similar to Directly Coupled. $\quad \because \frac{V_{1rms}}{V_{2rms}} = \frac{N_1}{N_2}, \; \frac{I_{1rms}}{I_{2rms}} = \frac{N_2}{N_1}$

6. Maximum Efficiency :



Assume

$V_{max} = 2V_{CC}$, $I_{max} = 2I_{CQ}$

$V_{min} = 0$, $I_{min} = 0$

$$\rightarrow \% \eta_{max} = \frac{(2V_{CC} - 0)(2I_{CQ} - 0)}{8 I_{CQ} V_{CC}}$$

$$\rightarrow \% \eta_{max} = 50\%$$

7. Power Dissipation (P_d) :

Similar to Directly Coupled

8. Advantages :

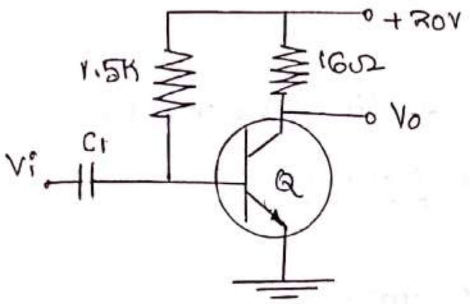
1. Efficiency is higher
2. The dc bias current that flow through the load in case of directly coupled is stopped in case of Transformer.

9. Disadvantages :

1. Due to transformer, the circuit becomes bulkier, costlier.
2. Difficult to design and Implement.
3. Frequency response of circuit is poor.

Problems:

1. A series fed class A amplifier shown in dia. operates from DC source and applied sinusoidal input signal generates peak base current 9mA amplifier. calculate 1. Quiescent voltage QVCE 2. Quiescent current ICQ. 3. DC power input, PDC 4. AC output power, PAC 5. Efficiency.



Soln $I_{b\,peak} = 9\,mA$

Assume $\beta = 50$, $V_{BE} = 0.7V$

1. Quiescent Voltage V_{CEQ}:

$\rightarrow V_{CC} - I_C R_L - V_{CE} = 0$

$\rightarrow V_{CE} = V_{CC} - I_C R_L = 9.70V.$

2. Quiescent current I_{CQ}:

$\rightarrow I_{CQ} = \beta I_B$

$\rightarrow V_{CC} - I_B R_B - V_{BE} = 0.$

$\rightarrow I_B = \frac{V_{CC} - V_{BE}}{R_B} = 12.8\,mA$

and

$\rightarrow I_{CQ} = \beta I_B = 643.50\,mA$

3. DC power input P_{DC}:

$\rightarrow P_{DC} = V_{CC}\, I_{CQ}$

$\rightarrow P_{DC} = 12.87\,W$

4. AC Power output P_{AC}:

$\because I_{b\,peak} = 9\,mA$

$\rightarrow P_{ac} = I_{rms}^2 R_L$

$\because I_{c\,peak} = \beta I_{b\,peak} = 450\,mA.$

$\rightarrow i_{c\,rms} = \frac{i_{c\,peak}}{\sqrt{2}} = \frac{450\,mA}{\sqrt{2}} = 318.1\,mA = I_{rms}.$

$\rightarrow P_{ac} = 318.1\,mA^2 \times 16\Omega$

$\rightarrow P_{ac} = 1.619\,W.$

5. Efficiency (η):

$\rightarrow \eta = \frac{P_{ac}}{P_{dc}} \times 100$

$\rightarrow \eta = 12.58\,\%$

2. The Loud speaker of 8Ω is connected to the secondary of the output transformer of a class A amplifier circuit. The Quiescent Collector current is 140mA. The turns ratio of the transformer is 3:1. The collector supply voltage is 10V. If ac power delivered to the loudspeaker is 0.48W. Assuming ideal transformer calculate

1. Ac power developed across primary
2. R.m.s Value of Load Voltage
3. R.m.s Value of primary voltage
4. R.m.s Value of Load current
5. R.m.s Value of primary current
6. Dc power input 7. Efficiency 8. power dissipation.

Soln Given $R_L = 8\Omega$, $I_{CQ} = 140\,mA$, $V_{CC} = 10V$, $P_{ac} = 0.48W$

$\because n = \frac{N_2}{N_1} = \frac{1}{3} = 0.333$ $\quad\because R_L' = \frac{R_L}{n^2} = 72\Omega$

1. Ac power developed across primary:

As the transformer, whatever is the power delivered to the load same is the power developed across primary.

$\rightarrow P_{ac}\,(primary) = 0.418\,W$

2. R.M.S Value of Load Voltage (V_{2rms}):

w.k.t, $\frac{V_{1rms}}{V_{2rms}} = \frac{N_1}{N_2} = \frac{3}{1}$

$\Rightarrow V_{2rms} \times 3 = V_{1rms} \times 1$

$\Rightarrow V_{2rms} = \frac{V_{1rms}}{3} = \frac{5.87}{3} = 1.9595V$

Find V_{1rms},

w.k.t, $P_{ac} = \frac{V_{1rms}^2}{R_L'} \Rightarrow 0.48 = \frac{V_{1rms}^2}{72\Omega} \Rightarrow V_{1rms} = 5.87V.$

3. R.M.S Value of Primary Voltage (V_{1rms}):

w.k.t, above $V_{1rms} = 5.87V$

4. R.M.S Value of Load Current (I_{2rms}):

$\Rightarrow P_{ac} = I_{2rms}^2 R_L \Rightarrow 0.48 = I_{2rms}^2 \times 8\Omega$

$\Rightarrow I_{2rms}^2 = 0.06$

$\Rightarrow I_{2rms} = 0.2449 A$

5. R.M.S Value of Primary current (I_{1rms})

$\Rightarrow \frac{I_{1rms}}{I_{2rms}} = \frac{N_2}{N_1} = \frac{1}{3}$

$\Rightarrow I_{2rms} \times 1 = I_{1rms} \times 3$

$\Rightarrow I_{1rms} = \frac{0.2449}{3} = 81.64 mA$

6. DC Power input (P_{DC}):

$\Rightarrow P_{DC} = V_{CC} I_{CQ} = 1.4W$

7. Efficiency (η):

$\Rightarrow \% \eta = \frac{P_{ac}}{P_{dc}} \times 100 = 34.28\%$

8. Power Dissipation (P_d):

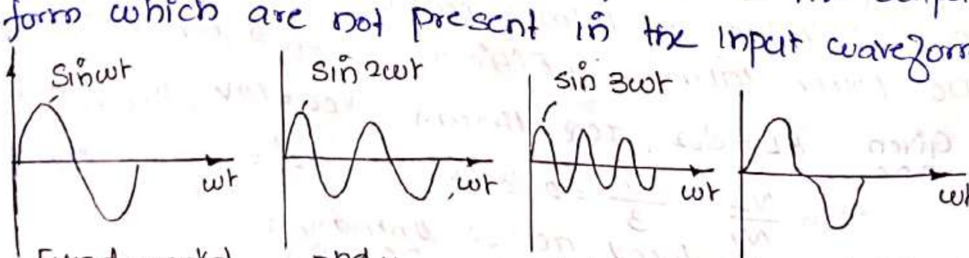
$\Rightarrow P_d = P_{dc} - P_{ac} = 0.92W$

Distortion in class A:

In class A, we get Harmonic distortion (or) amplitude distortion (or) Non linear distortion.

Harmonic Distortion:

The Present of frequency component in the output waveform which are not present in the input waveform.



Fundamental — 2nd Harmonic — 3rd Harmonic — Distorted waveform

Total Harmonic Distortion,

$\Rightarrow \% D = \sqrt{D_2^2 + D_3^2 + \dots + D_n^2} \times 100$

Scanned with CamScanner

Where D = Total Harmonic Distortion

$$\% D_2 = \frac{|B_2|}{|B_1|} \times 100 , \quad \% D_3 = \frac{|B_3|}{|B_1|} \times 100 , \quad \% D_n = \frac{|B_n|}{|B_1|} \times 100$$

∵ B_1 = fundamental Frequency component of amplitude
B_n = nth harmonic component of amplitude

Second Harmonic Distortion : (Three Point Method) :

Let the mathematical expression for base current due to second harmonic is

$$\rightarrow i_b = I_{Bm} \cos \omega t \rightarrow (1)$$

and Collector current is given as

$$\rightarrow i_c = G_1 i_b + G_2 i_b^2 = G_1 I_{Bm} \cos \omega t + G_2 I_{Bm}^2 \cos^2 \omega t.$$

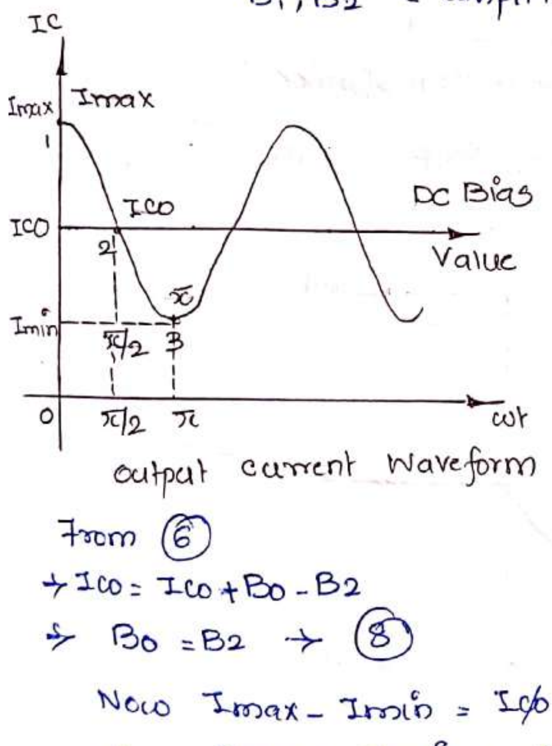
Using Trigonometric operation, we get $\rightarrow (2)$

$$\rightarrow i_c = B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t \rightarrow (3)$$

The total collector current including d.c Bias can be written as

$$\rightarrow i_c = I_{C0} + B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t \rightarrow (4)$$

where $(I_{C0} + B_0)$ = DC. component
B_1, B_2 = amplitudes of fundament , second harmonic



output current waveform

Consider the Various instants 1, 2, 3 shown in fig

At point 1, $\omega t = 0$, $i_c = I_{max}$.

$$\rightarrow i_c = I_{C0} + B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t$$

$$\rightarrow I_{max} = I_{C0} + B_0 + B_1 + B_2 \rightarrow (5)$$

At Point 2, $\omega t = \frac{\pi}{2}$, $i_C = I_{C0}$

$$\rightarrow i_c = I_{C0} + B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t$$

$$\rightarrow I_{C0} = I_{C0} + B_0 - B_2 \rightarrow (6)$$

$\cos 90 = 0$
$\cos 180 = -1$

At Point 3, $\omega t = \pi$, $i_C = I_{min}$

$$\rightarrow i_c = I_{C0} + B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t$$

$$\rightarrow I_{min} = I_{C0} + B_0 - B_1 + B_2 \rightarrow (7)$$

From (6)

$$\rightarrow I_{C0} = I_{C0} + B_0 - B_2$$

$$\rightarrow B_0 = B_2 \rightarrow (8)$$

Now $I_{max} - I_{min} = \cancel{I_{C0}} + \cancel{B_0} + B_1 + \cancel{B_2} - \cancel{I_{C0}} - \cancel{B_0} + B_1 - \cancel{B_2}$

$$\rightarrow I_{max} - I_{min} = 2B_1$$

$$\rightarrow B_1 = \frac{I_{max} - I_{min}}{2} \rightarrow (9)$$

Now

$$I_{max} + I_{min} = 2I_{C0} + 2B_0 + 2B_2 \qquad ∵ B_0 = B_2$$

$$\rightarrow I_{max} + I_{min} = 2I_{C0} + 4B_2$$

$$\rightarrow B_2 = \frac{I_{max} + I_{min} - 2I_{C0}}{4} \rightarrow (10)$$

As the amplitudes of fundamental, second harmonic are known the second harmonic distortion can be calculated as

$$\rightarrow \% D_2 = \frac{|B_2|}{|B_1|} \times 100 \rightarrow (11)$$

Higher order harmonic distortion

→ Similar analysis of second order method.

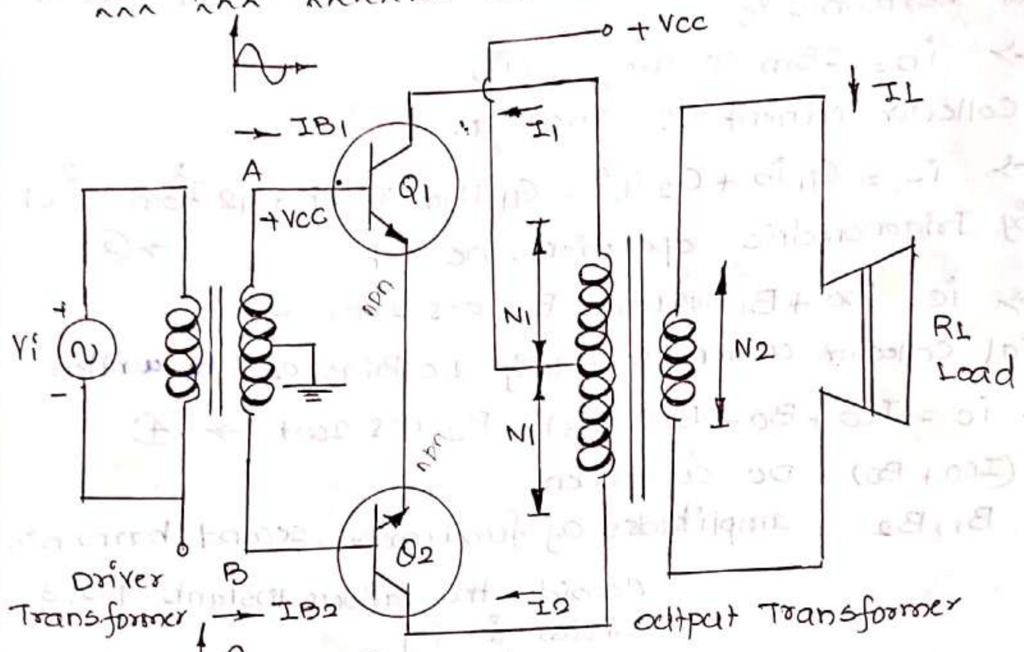
Class B amplifier :

Two Types :

1. Push pull class B

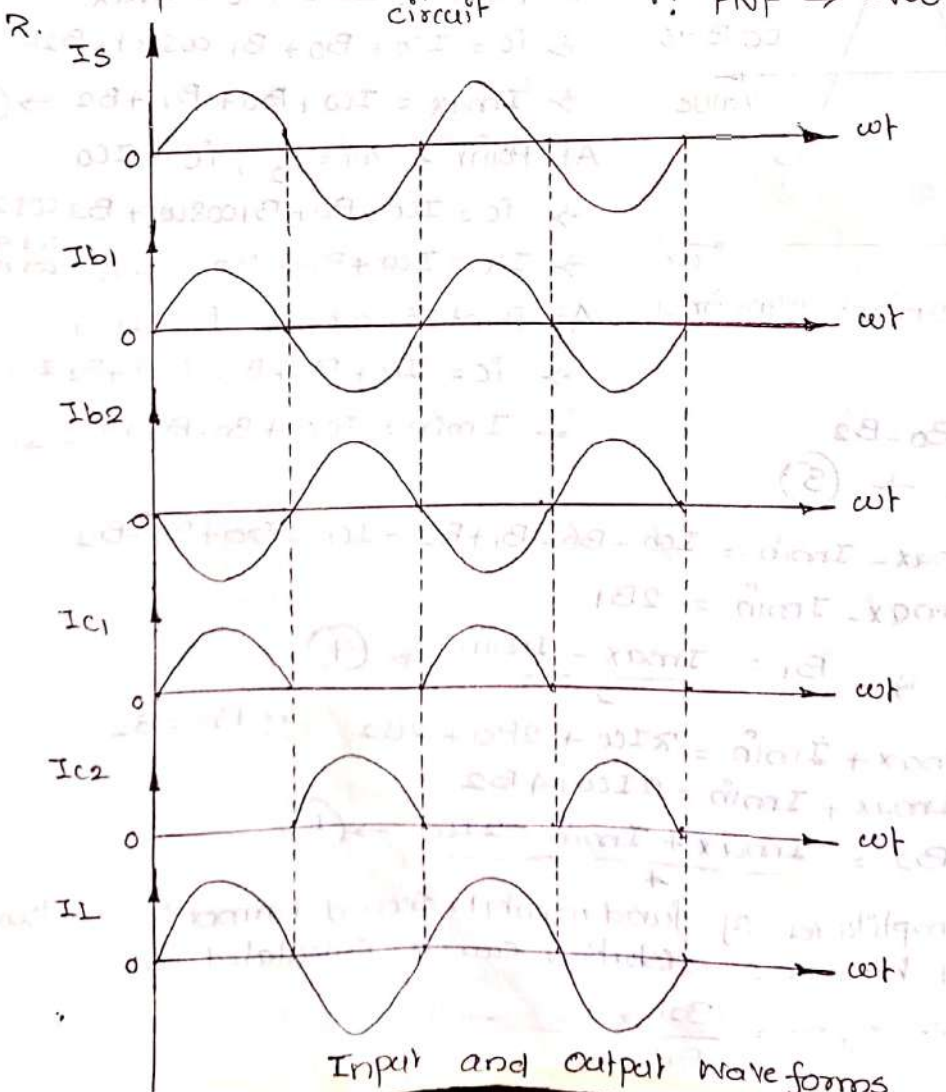
2. complementary symmetry class B

1. Push pull class B amplifier :



Push pull circuit

∴ PNP → -VCC



Input and output wave forms

3. operation ;

4. DC operation :

The dc biasing point i.e Q-point is adjusted on the x-axis such that VCE = Vcc and ICQ is zero.

hence $I_{CQ} = 0$

$V_{CEQ} = V_{CC}$

5. DC Power Input (Pdc) :

Each transistor output is in the form of Half rectified waveform. So, the output current of each transistor is $\frac{I_m}{\pi}$ due to half rectified waveform.

$\rightarrow I_{dc} = \frac{I_m}{\pi} + \frac{I_m}{\pi} = \frac{2I_m}{\pi}$ → Two Transistor

Total DC power Input is

$\rightarrow P_{dc} = V_{CC} \times I_{dc} = \frac{V_{CC} \cdot 2I_m}{\pi}$

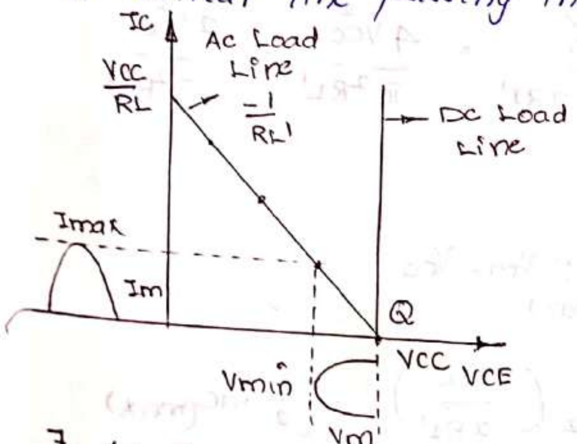
6. AC operation :

When ac signal is applied to the driver Transformer, the positive half cycle Q_1 conducts. the lower half of primary does not carry any current. when Q_2 conducts, upper half of primary does not carry any current.

Hence the RL on primary can

$\rightarrow R_L' = \frac{R_L}{n^2}$ where $n = \frac{N_2}{N_1}$

The slope of ac load line is $-1/R_L'$ when dc load line is vertical line passing through Q on the x-axis.



The sloap of the ac load line is represented in terms of Vm and Im as

$\rightarrow \frac{+1}{R_L'} = \frac{I_m}{V_m}$

$\therefore R_L' = \frac{V_m}{I_m}$

7. AC Power output (Pac) :

It is given as

$\rightarrow V_{rms} = \frac{V_m}{\sqrt{2}}, \quad I_{rms} = \frac{I_m}{\sqrt{2}}$

∵ Vm and Im are Peak value of Voltage and current.

Ac power output is expressed as

$\rightarrow P_{ac} = V_{rms} I_{rms} = \frac{V_{rms}^2}{R_L'} = I_{rms}^2 R_L'$

$\because V_{rms} = I_{rms} R_L'$, $I_{rms} = \frac{V_{rms}}{R_L'}$

using peak value,

$\rightarrow P_{ac} = \frac{V_m I_m}{2} = \frac{V_m^2}{2R_L'} = \frac{I_m^2 R_L'}{2}$

$\because V_m = I_m R_L'$, $I_m = V_m / R_L'$

8. Efficiency (η):

Ratio of ac power delivered to the load to the dc power

$$\Rightarrow \% \eta = \frac{P_{ac}}{P_{dc}} \times 100 = \frac{\frac{V_m I_m}{2}}{\frac{2 V_{cc} I_m}{\pi}} \times 100 = \frac{\pi V_m}{4 V_{cc}} \times 100$$

9. Maximum Efficiency:

Assume $V_m = V_{cc}$

$$\Rightarrow \% \eta_{max} = \frac{\pi V_m}{4 V_{cc}} \times 100 = \frac{\pi V_{cc}}{4 V_{cc}} \times 100 = 78.5\%$$

10. Power dissipation (P_d):

It is the difference between ac power output and DC power output.

$$\Rightarrow P_d = P_{dc} - P_{ac} = \frac{2 V_{cc} I_m}{\pi} - \frac{V_m I_m}{2}$$

$$\Rightarrow P_d = \frac{2 V_{cc} V_m}{\pi R_L'} - \frac{V_m^2}{2 R_L'} \quad \because I_m = \frac{V_m}{R_L'}$$

11. Maximum power dissipation (P_{dmax}):

wkt, $P_d = \frac{2 V_{cc} V_m}{\pi R_L'} - \frac{V_m^2}{2 R_L'}$

Diff w.r. to V_m, and Equating it to zero

$$\Rightarrow \frac{\partial P_d}{d V_m} = \frac{2 V_{cc}}{\pi R_L'} - \frac{2 V_m}{2 R_L'} = 0$$

$$\Rightarrow \frac{2 V_{cc}}{\pi} = V_m$$

Now $P_d = \frac{2 V_{cc}}{\pi R_L'} \cdot \frac{2 V_{cc}}{\pi} - \frac{4 V_{cc}^2}{\pi^2 2 R_L'} = \frac{4 V_{cc}^2}{\pi^2 R_L'} - \frac{2 V_{cc}^2}{\pi^2 R_L'}$

$$\Rightarrow P_d = \frac{2 V_{cc}^2}{\pi^2 R_L'}$$

wkt, $P_{ac} = \frac{V_m^2}{2 R_L'} = \frac{V_{cc}^2}{2 R_L'} = P_{ac\,(max)} \because V_m = V_{cc}$

Now

$$\Rightarrow P_{d(max)} = \frac{2 V_{cc}^2}{\pi^2 R_L'} = \frac{4}{\pi^2} \left(\frac{V_{cc}^2}{2 R_L'} \right) = \frac{4}{\pi^2} P_{ac(max)}$$

$\Rightarrow P_{d(max)} = \frac{4}{\pi^2} P_{ac}$ for Both transistor.

Consider, one transistor, $\frac{P_{dmax}}{2} = \frac{2}{\pi^2} P_{ac}$

2. Advantages:

1. Efficiency is higher
2. Due to transformer, Impedance matching is possible.

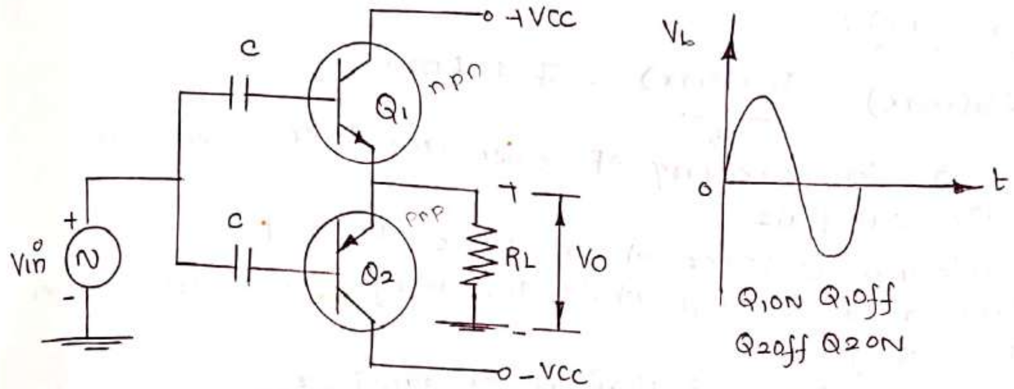
13. Disadvantages:

1. Two Transformers are necessary
2. Transformer makes circuit become bulky and costly.

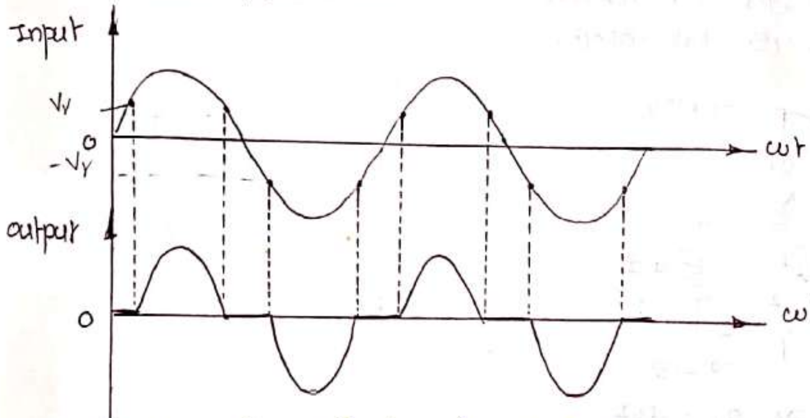
2. Complementary symmetry class B amplifier :

1. When both the transistors are of same type either npn or pnp, then the circuit is called push pull class B.
2. When two transistor form a complementary pair ie, one npn and other pnp then the circuit is called complementary symmetry class B.

Complementary symmetry class B amplifier :



Cross over Distortion :



Derivation is similar to push pull.

Problems :

1. A class B push pull amplifier supplies power to a resistive load of 12Ω. The output transformer has a turns ratio of 3:1 and Efficiency of 78.5%. obtain
   - i) Maximum power output
   - ii) Maximum power dissipation in each transistor
   - iii) Maximum Base and collector current for each Transistor.

Assume $h_{fe} = 25$, $V_{cc} = 20V$.

Soln Given : $R_L = 12\Omega$, $h_{fe} = 25$, $V_{CC} = 20V$, $\eta = 78.5\%$, $n = \frac{N_2}{N_1} = \frac{1}{3}$

$n = 0.333$

1. Maximum Power output : $P_{ac(max)}$

→ wkt, $P_{ac(max)} = \frac{V_{cc}^2}{2R_L'} = 1.8518W$ ∵ $R_L' = \frac{R_L}{n^2} = 108\Omega$

2. Maximum Power Dissipation in each Transistor : (P_{dmax})

→ wkt, $P_{d(max)} = \frac{2V_{cc}^2}{\pi^2 R_L'} = 0.7505W$

→ $P_{d(max)}$ Per Transistor $= \frac{0.7505}{2} = 0.3752W$

3. Maximum Base and Collector current For each transistor :

$I_{c(max)}$ and $I_{b(max)}$

1. $I_{c(max)} = I_m$:

$\Rightarrow P_{ac(max)} = V_{rms} \cdot I_{rms} = \dfrac{V_m I_m}{2} = \dfrac{V_{cc} \cdot I_m}{2}$ $\because V_m = V_{cc}$

$\Rightarrow I_m = 0.1851\,A = i_{c\,max}$.

and

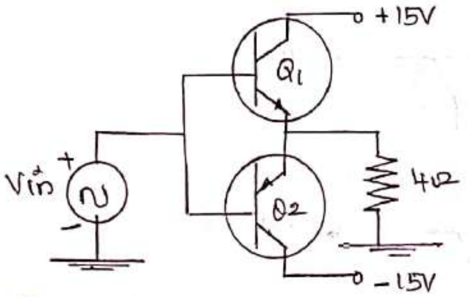
2. $I_{b(max)}$:

$\Rightarrow I_{b(max)} = \dfrac{I_{c(max)}}{\beta} = 7.407\,mA.$

2. Class B Complementary AF power amplifier shown in below dia. calculate

1. Maximum ac power which can be developed
2. Collector dissipation while developing Maximum ac power
3. Efficiency
4. Maximum power dissipation per transistor
5. Efficiency under Maximum power dissipation Condition

Assume pure sinusoidal input.



Soln Given $V_{cc} = 15V$, $R_L = 4\Omega$

1. Maximum Ac Power which can be developed : ($P_{ac\,max}$) :

$\Rightarrow P_{ac(max)} = \dfrac{V_{cc}^2}{2R_L} = 28.125\,W$ $\because R_L' = R_L$

2. Collector dissipation while developing Maximum ac Power : (P_d

$\Rightarrow P_d = P_{dc} - P_{ac}$

$\because P_{dc} = \dfrac{2 V_{cc} I_m}{\pi}$ $\because I_m = \dfrac{V_m}{R_L}$

$\Rightarrow P_d = 35.809 - 28.125$

$P_{dc(max)} = \dfrac{2 V_{cc}^2}{\pi^2 R_L'}$ = $I_m = 3.75$

$\Rightarrow P_d = 7.684\,W$

$\Rightarrow P_{dc} = \dfrac{2 V_{cc} I_m}{\pi} = \dfrac{2 \times 15 \times 3.75}{\pi}$

$\Rightarrow P_{dc} = 35.809\,W$

3. Efficiency (η) :

$\Rightarrow \%\eta = \dfrac{P_{ac}}{P_{dc}} \times 100 = 78.5\%$

4. Maximum power dissipation per Transistor :

$\Rightarrow$ WKT $P_{d(max)} = \dfrac{2}{\pi^2} P_{ac(max)} = 5.699\,W$ Per Transistor

5. Efficiency under Maximum Power dissipation condition :

$\rightarrow \%\eta = \frac{P_{ac} \text{ under } P_{dmax}}{P_{dc} \text{ under } P_{dmax}} \times 100$

$\because P_{d(max)} = \frac{2}{\pi} V_{CC} \cdot I_m$

$\because I_m = \frac{V_m}{R_L} = \frac{9.54}{4} = 2.38$

$\because V_m = \frac{2}{\pi} V_{CC} = 9.54$

$\rightarrow P_{d(max)} = \frac{2}{\pi} V_{CC} \cdot I_m = 22.797$

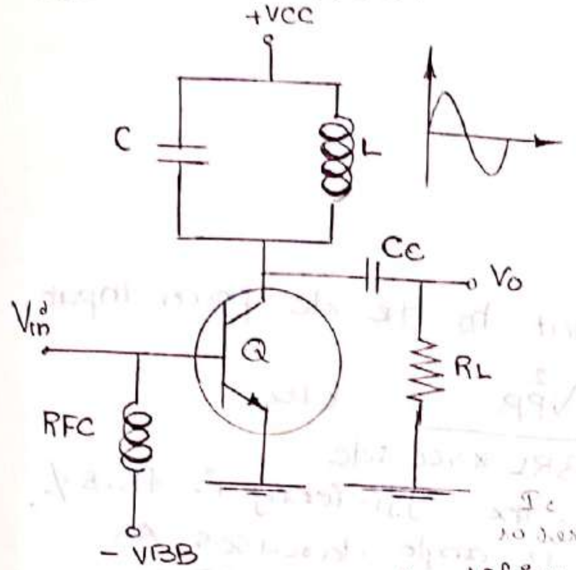
and

$\because P_{ac} = \frac{1}{2} \frac{V_m^2}{R_L} = \frac{1}{2} \cdot \frac{9.54^2}{4} = 11.398$

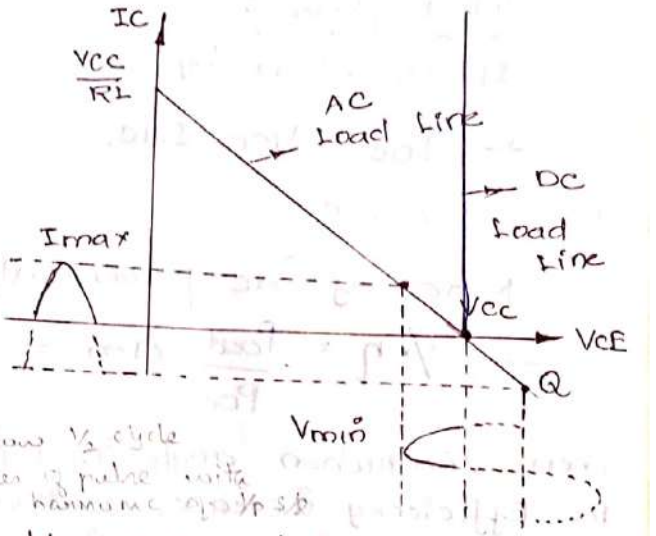
$\rightarrow \%\eta = \frac{11.398}{22.797} \times 100$

$\rightarrow \%\eta = 50\%$

Class C Amplifier :

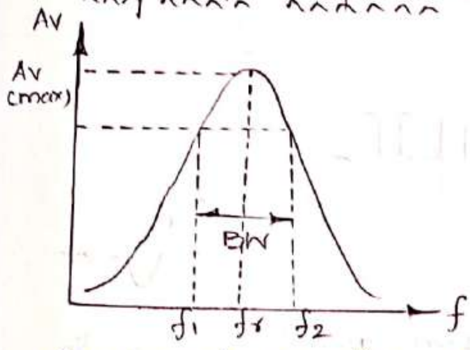


Class C Tuned Amplifier



Wave Representing class C operation

Frequency Response :



Operation :

The dia. shows class C tuned amplifier. A parallel tuned circuit acting as a load is tuned to the input frequency. Thus it filters the harmonic frequencies and Produce a sine wave output Voltage consisting of the fundament component of the signal.

The Resonant frequency is given by

$$\rightarrow f_r = \frac{1}{2\pi\sqrt{LC}} \text{ Hz}$$

Generally, the tank circuit, to amplify only narrow band of frequencies.

Quality Factor :

It is given by

$$\rightarrow Q = \frac{r_C}{\omega_r L}$$

Bandwidth :

It is the difference between upper and lower frequency.

$\rightarrow BW = f_2 - f_1$

$\because f_1$ = Lower half power (3dB) freq.

f_2 = upper half power (3dB) freq.

The bandwidth of class C Tuned amplifier is given by

$\rightarrow$ $BW = \frac{f_r}{Q}$ $\quad$ ∵ Q = Quality factor, f_r = Resonant Frequency.

Output power : (P_{out}) :

$\rightarrow$ $P_{out} = \frac{V_{PP}^2}{8R_L}$ $\quad$ ∵ $V_{PP} = 2V_m = 2\sqrt{2}\,V_{rms}$.

Transistor Dissipation :

The Maximum power dissipation is given by

$\rightarrow$ $P_{D(max)} = \frac{V_{PP(max)}^2}{40\,r_c}$ $\quad$ ∵ $V_{PP(max)} = 2V_{cc}$

DC Input Power :

It is given by

$\rightarrow$ $P_{DC} = V_{cc}\,I_{dc}$.

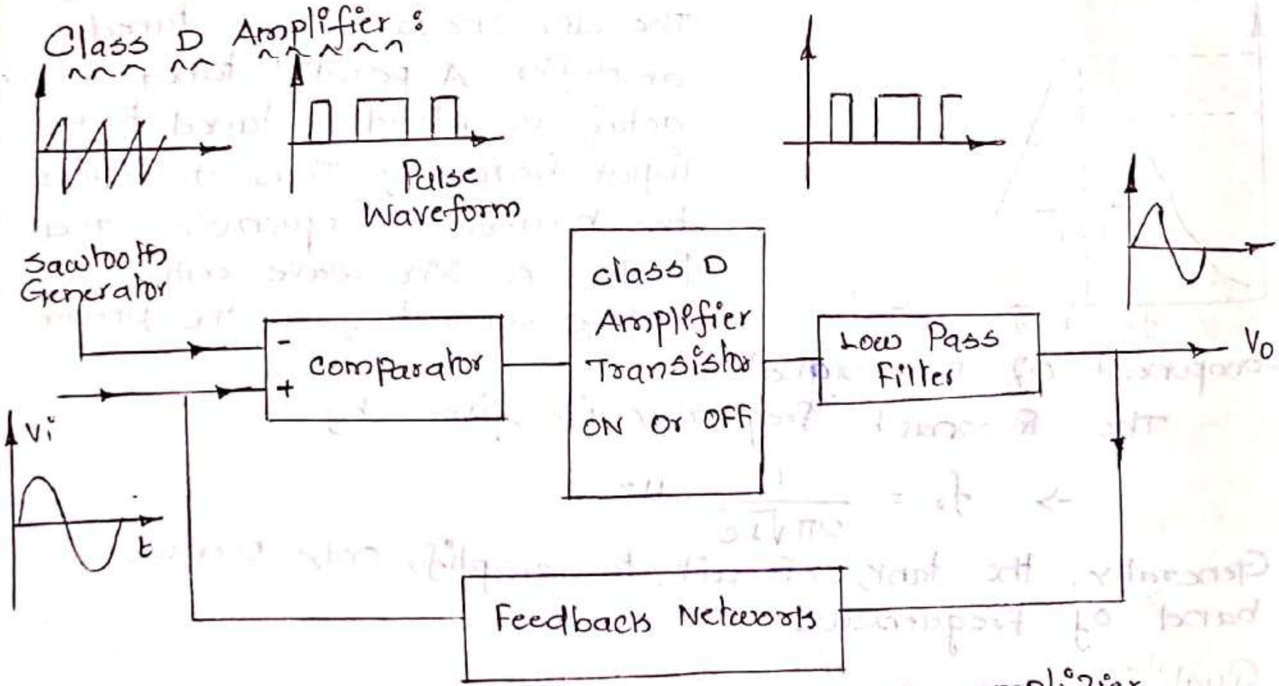
Efficiency : η :

Ratio of ac power output to the dc power input

$\rightarrow$ $\%\eta = \frac{P_{out}}{P_{DC}} \times 100 = \frac{V_{PP}^2}{8R_L \times V_{CC} I_{dc}} \times 100$.

when conduction angle is 180°, the Efficiency is 78.5%. The Efficiency increase, conduction angle decreases. As indicated, class C amplifier has maximum efficiency of 100%.

Class D Amplifier :



Block diagram of class D amplifier