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-: HAND WRITTEN NOTES:-

OF

# ELECTRONICS & COMMUNICATION

## ENGINEERING

(1)

-: SUBJECT:-

# ANALOG ELECTRONICS





## Syllabus :-

(3)

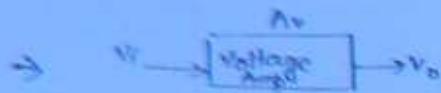
- 0 1) Op-Amp
- 0 2) Linear Wave Shaping circuit <Taub>
- 0 3) Schmitt Trigger
- 0 4) Waveform Generator
  - Multivibrators
    - Bistable Multivibrator.
    - Monostable "
    - Astable " (Square Wave Generator)
    - Triangular Wave Generator.
- 0 5) Diode Circuits
  - Rectifiers & filters
  - Precision Rectifiers
  - Clipper & clippers
  - Voltage Doubters.
- 0 6) Bipolar Junction Transistor.
  - Transistor Biasing & Stabilisation
  - Current Mirror Circuit
  - Voltage Regulator
  - Power Amplifiers
- 0 7) Multivibrator by using BJT. <Taub>
- 0 8) Amplifiers -
  - Low frequency Analysis of BJT.
  - High " "
  - Multistage Amp.
  - Feedback "
  - Low frequency analysis of FET
  - Oscillators (Sinusoidal)
  - 555 - Timer
  - Multivibrators

## Books

- Millman - Halkias - Yellow Book
- Pulse Digital & switching circuit  
- Millman & Taub

## Operational Amplifier

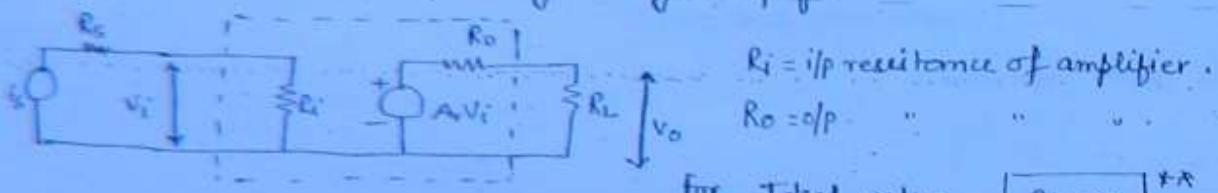
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→  $A_v = \frac{V_o}{V_i} = \text{Gain}$

→ Op-Amp is a VCVS.  $\rightarrow + \circlearrowleft A_v V_i$

→ Equivalent circuit for any voltage amplifier-



For Ideal value -

$R_i = \infty$	*R
$R_o = 0$	" "
$A_v = \infty$	" "
$BW = \infty$	" "
$G \cdot BW = \infty$	" "

→  $R_i \ggg R_s$ , so that  $V_o \approx V_i$

→  $R_o \lll R_L$ , so that  $V_o \approx A_v V_i$

\* To get  $A_v \rightarrow \infty$ , multistaging is done but the BW will  $\downarrow$ .

\* BW is defined as the freq. range for which gain is independent of frequency.

→ Gain of practical Op-Amp =  $10^6$ .

→ Op-Amp is a multistage amplifier.

→  $\boxed{\text{Gain} \times \text{BW} = \text{constant}}$ ; Ideally  $G \cdot BW$  should be  $\infty$ .

\* BW cannot be  $\infty$  due to the presence of  $C_T$  &  $C_d$  in multistage amplifiers. (internal capacitance)

→ For practical Op-Amp;  $G \cdot BW = 10^6$  Hz.

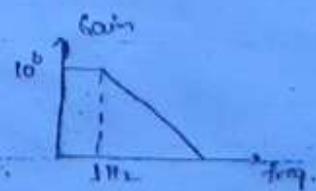
→ Max. possible BW =  $10^6$  Hz for gain = 1.

→ Negative feedback will  $\downarrow$  the gain of the system and  $\uparrow$  the BW and hence  $\uparrow$  the stability of the system.

(B)

### Op-Amp Imp. Points -

- (i) It is a monolithic IC or a semiconductor chip fabricated with VLSI by using epitaxial method.
- (ii) In epitaxial method, entire IC is fabricated on single crystal of Si.
- (iii) It is basically a voltage controlled device or voltage amplifier or VCVS.
- (iv) Popularly used op-amp is IC-741. For IC-741, maximum power supply is  $\pm 15V$ .
- (v) Op-amp is versatile, predictable and economic system building block as small size, high reliability, reduced cost, low offset voltage & current and low power consumption.
- (vi) It is originally invented to execute the mathematical operations, hence called op-amp.
- (vii) It is a direct coupled, high gain amplifier, i.e., open loop gain is very high, therefore frequency stability of the signal is less, and to compensate this, small amount of -ve feedback is added so that the gain is reduced & the frequency stability increases (since BW  $\uparrow$ ).
- (viii) Op-amps are generally operated under closed loop condition, i.e., by applying -ve feedback.
- (ix) In an op-amp, Gain  $\times$  BW = constant.



## Characteristic of Operational Amplifier.

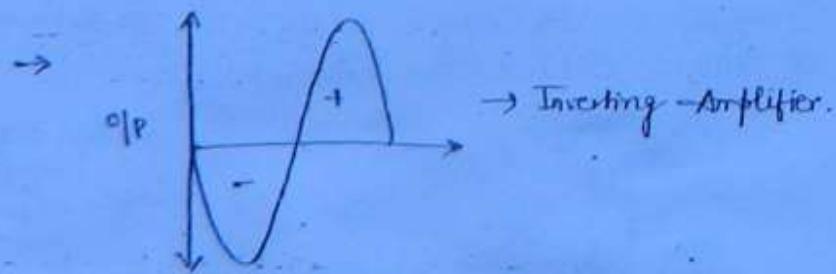
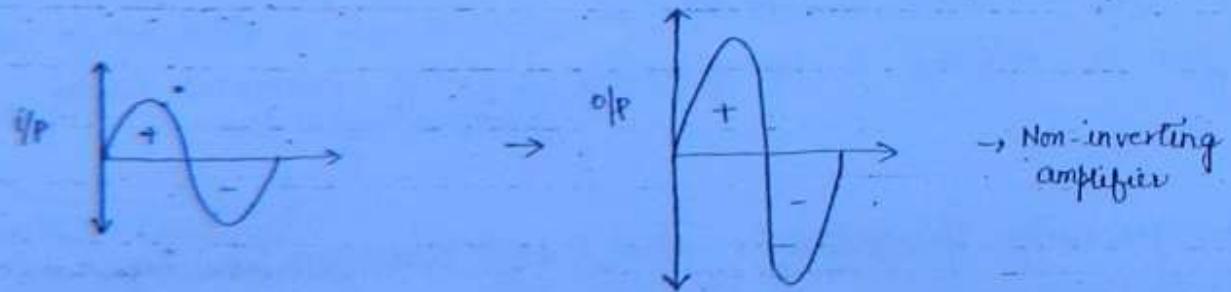
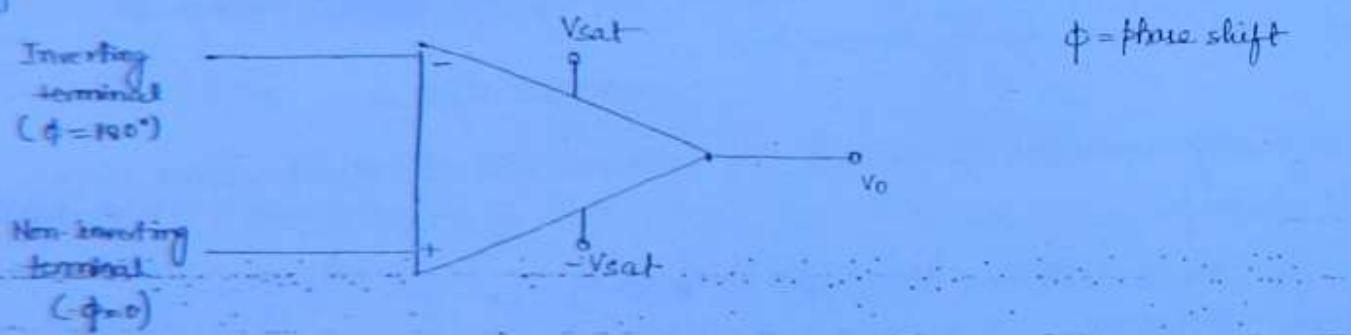
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### Characteristic

	<u>Ideal</u>	<u>Practical</u>
- Voltage Gain, $A_v$	$\infty$	$10^6$
- Input Resistance, $R_i$	$\infty$	$1M\Omega$
- Output Resistance, $R_o$	0	$10 - 100 \Omega$
- G. BN	$\infty$	$10^6 \text{ Hz}$
- BW	$\infty$	$10^6 \text{ Hz (for Gain=1)}$
- CMRR	$\infty$	$10^6 \text{ or } 100 \text{ dB}$
- Slew Rate [SR]	$\infty$	$80 \text{ V}/\mu\text{sec.}$

→ It is also referred as Basic linear Integrated Circuit.

### Symbol :-





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Case 1 :- When  $V_1 \neq 0, V_2 = 0$ , then  $V_0 > 0$

Case 2 :- When  $V_1 \neq 0, V_2 \neq 0$  and  $V_1 > V_2$ , then  $V_0 > 0$ .

Case 3 :- When  $V_1 = 0, V_2 \neq 0$ , then  $V_0 < 0$

Case 4 :- When  $V_1 \neq 0, V_2 \neq 0$  and  $V_2 > V_1$ , then  $V_0 < 0$ .

Representation of Gain →

$$|Av| = 10^6$$

Case 1 -

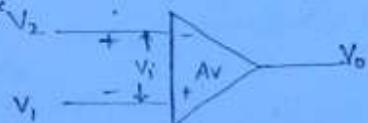


$$V_i' = V_1 - V_2 ; \quad V_0 = Av V_i' \Rightarrow \text{If we represent like this then} \\ = Av(V_1 - V_2) \quad Av = 10^6 \text{ Hz, i.e., } Av > 0$$

Case 2 -

$$V_i = V_2 - V_1$$

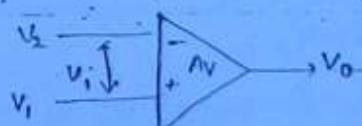
$$V_0 = Av V_i = Av(V_2 - V_1)$$



then for this representation,  $Av = -10^6$ , i.e.,  $Av < 0$

When  $Av \rightarrow \infty$ :

$$Av \rightarrow \infty$$



$$V_i' = V_1 - V_2$$

$$V_0 = \text{finite} \Rightarrow V_i = V_0/\infty = 0$$

$$\Rightarrow V_1 = V_2$$

→ There is finite o/p w/o any input.

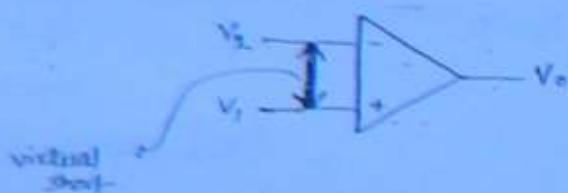
→ Now, if we attach any voltage source at  $V_1$ , the same will appear at  $V_2$  ( $\because V_1 = V_2$ ) but  $R_i = \infty$  (ideally) hence they should be 0.

but they are behaving as SC. This condition is called Virtual short, i.e., even though they are not physically short, they are behaving as short.

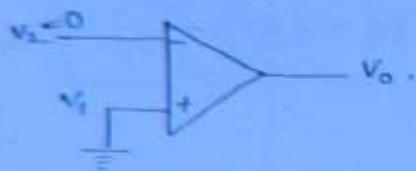


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Symbol for virtual short -

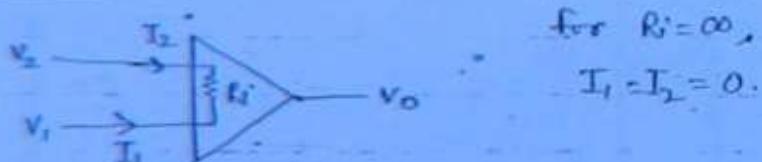


→ If we ground  $V_1$ , i.e., connect  $V_1 = 0$ , then  $V_2$  will also become 0V. This is called virtual ground. It is a special case of virtual short.



Virtual Ground Process

→ When  $R_i = \infty$  —



for  $R_i = \infty$ ,

$$I_1 = I_2 = 0.$$

→ Internal power consumption  $\approx 0$ .

Ex  $V_o = 5V$  and  $A_v = 10^6$ ,  $R_i = 10^6 \Omega$ .

$$\Rightarrow V_i = \frac{V_o}{A_v} = 5\mu V \approx 0$$

$$I_i = \frac{V_i}{R_i} = 5 \mu A \approx 0$$

→ If the gain in the given problem is ↑ (very high), then we can use the concept of virtual ground. It is an approximate concept.

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→  $V_o = V_i \rightarrow [Av] \rightarrow V_o$   $A_{OL} = Av \rightarrow$  open loop system.

$$A_{OL} = Av = \frac{V_o}{V_i} = \frac{V_o}{V_s} ; V_s = V_i$$

→  $V_s - V_i \xrightarrow{\pm} V_i$

$$V_i = V_s \pm \beta V_o$$

$$V_i = \begin{cases} V_s + \beta V_o & \rightarrow +ve \text{ feedback} \\ V_s - \beta V_o & \rightarrow -ve \text{ feedback.} \end{cases}$$

$$V_o = (V_s \pm \beta V_o) \cdot Av$$

For +ve feedback -

$$\boxed{A_{OL} = \frac{V_o}{V_s} = \frac{Av}{1 - \beta Av} > Av} \rightarrow \text{closed loop gain for +ve feedback}$$

for -ve feedback -

$$\boxed{A_{OL} = \frac{V_o}{V_s} = \frac{Av}{1 + \beta Av} < Av} \rightarrow \text{closed loop gain for -ve feedback}$$

→ +ve feedback is used in oscillators & -ve feedback is used in amplifier.

\* Op-Amp with -ve feedback

Op-Amp with +ve feedback.

$$\rightarrow |A_{OL}| \ll |A_{OL}|$$

$$\rightarrow |A_{OL}| \gg |A_{OL}|$$

We can assume  $A_{OL} = Av \rightarrow \infty$

We can assume  $A_{OL} = \infty$ , but we can't assume  $A_{OL} = 0$ ,

∴ Virtual Ground process is valid.

∴ Virtual ground process is invalid.

## Mode of Operation

### i) Inverting Mode -

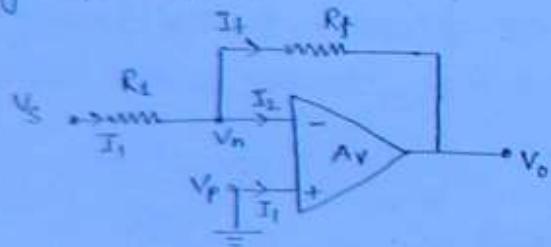
→ Phase shift =  $180^\circ$

### ii) Non-Inverting Mode $\rightarrow (\phi = 0^\circ)$

### iii) Differential Mode $\Rightarrow V_o \propto [V_1 - V_2]$

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## Inverting Op-Amp -



→ Negative feedback.

→  $|A_{OL}| \ll |A_{AV}| \Rightarrow$  we can apply V.G.P.  
 $\Rightarrow V_P = V_N = 0$ .

→  $R_i = \infty \Rightarrow I_2 = I_1 = 0$ .

→ Incoming current = outgoing current

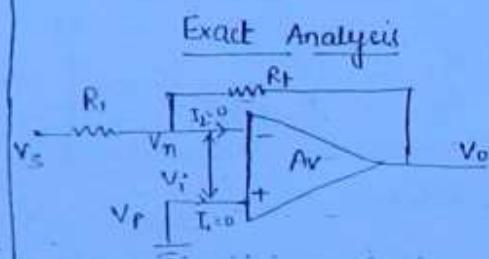
$$\Rightarrow I = I_f + I_2$$

$$\Rightarrow \frac{V_S - V_N}{R_i} = \frac{V_N - V_o}{R_f} + 0$$

$$\Rightarrow \frac{V_S}{R_i} = -\frac{V_o}{R_f}$$

$$\Rightarrow A_{OL} = \frac{V_o}{V_S} = -\frac{R_f}{R_i}$$

$$\Rightarrow \phi = 180^\circ$$



### Exact Analysis

$$V_i = V_P - V_N ; \Rightarrow \{ A_v > 0 \}$$

$$V_o = A_v [V_P - V_N] \quad \text{---(1)}$$

$$V_P = 0 \quad \text{---(2)}$$

$$\begin{aligned} R_f &\frac{V_o}{V_N} \\ R_i &\frac{V_N - V_o}{V_S} + \frac{V_N - V_P}{R_f} = 0 \\ &\Rightarrow V_N \left[ \frac{1}{R_i} + \frac{1}{R_f} \right] = \frac{V_o}{R_f} + \frac{V_S}{R_i} \end{aligned}$$

$$\Rightarrow V_N = \frac{V_o R_i + V_S R_f}{R_i + R_f}$$

L  
continued

$$I_f = I_1 + I_2 + I_3$$

$$\frac{V_o - V_0}{R_f} = \frac{V_1 - V_o}{R_1} + \frac{V_2 - V_o}{R_2} + \frac{V_3 - V_o}{R_3}$$

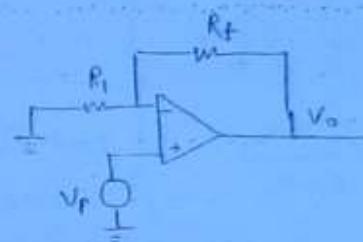
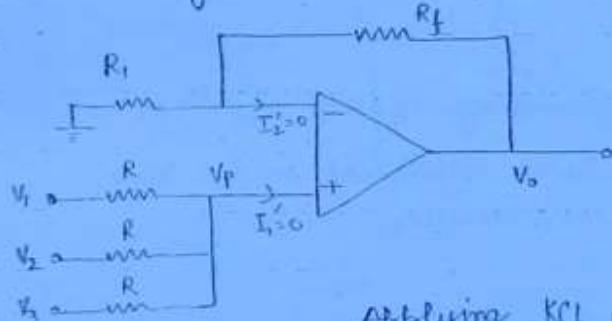
(11)

$$\Rightarrow V_o = -R_f \left[ \frac{V_1}{R_1} + \frac{V_2}{R_2} + \frac{V_3}{R_3} \right]$$

When  $R_f = R_1 = R_2 = R_3$  -

$$V_o = -[V_1 + V_2 + V_3] \quad (\phi = 180^\circ)$$

### Non-Inverting Summer



Applying KCL at non-inverting terminal -

$$\frac{V_p - V_1}{R_1} + \frac{V_p - V_2}{R_2} + \frac{V_p - V_3}{R_3} = 0$$

$$\Rightarrow V_p = \frac{V_1 + V_2 + V_3}{3}$$

$$\text{Now, } \frac{V_o}{V_p} = 1 + \frac{R_f}{R_1}$$

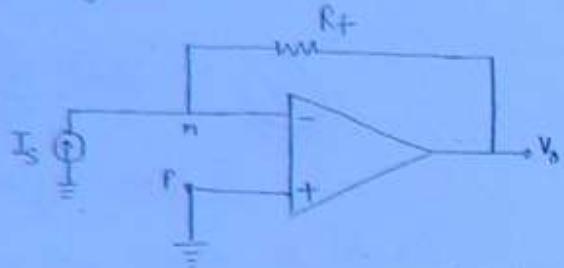
$$\Rightarrow V_o = \left(1 + \frac{R_f}{R_1}\right) \left(\frac{V_1 + V_2 + V_3}{3}\right) \quad (\phi = 0^\circ)$$

If  $R_f = 2R_1$  ;

$$V_o = [V_1 + V_2 + V_3]$$

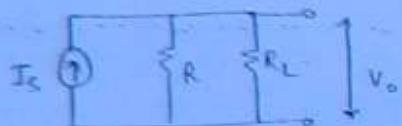
## Current to Voltage Converter :-

(12)



$$V_P = V_N = 0$$

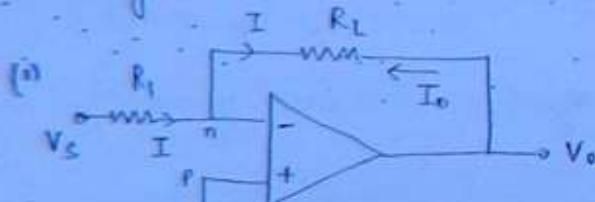
$$\frac{V_N - V_o}{R_f} = I_S \Rightarrow V_o = -I_S \cdot R_f \rightarrow \{ V_o \text{ is independent of } R_L \text{ and hence it is a converter}\}$$



$V_o = \frac{I_S \cdot R R_L}{R + R_L}$  but this is not converting  $I_S$  into a voltage source because  $V_o$  is dependent on  $R_L$ . Hence, given circuit is not a converter.

## Voltage to Current Converter :-

### (a) Floating load :-

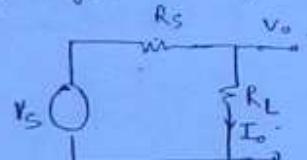


$$I = \frac{V_S - 0}{R_f} \Rightarrow I = \frac{V_S}{R_f}$$

$$I_o = -I = -\frac{V_S}{R_f}$$

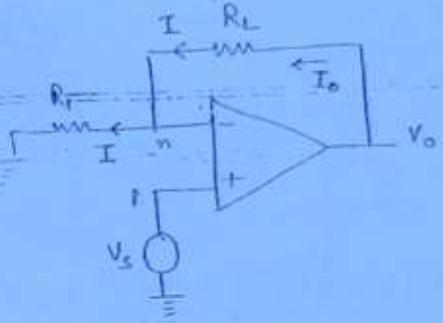
$\rightarrow I_o = \text{output current independent of } R_L$

{ It is standard convention to take load current  $I_o$  in direction away from output voltage  $V_o$ . i.e.,  $I_o$  leaving from  $V_o$  }



$$\frac{V_o}{R_L} = I_o = \frac{V_S}{R_f + R_L}$$

$I_o$  depends on  $R_L$ , hence not a converter.



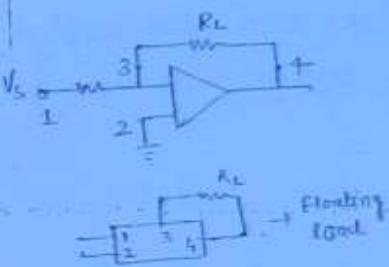
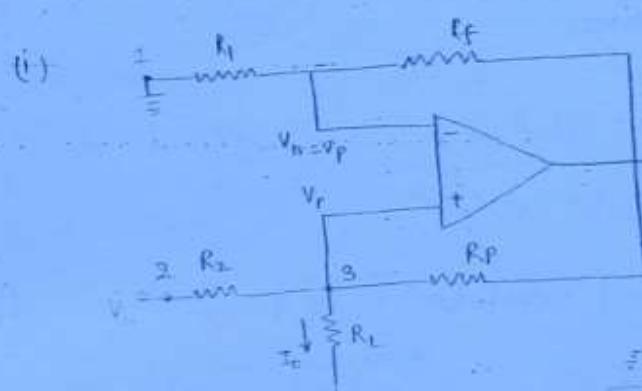
$$V_n = V_s \quad \text{--- (1)}$$

$$\frac{V_s - 0}{R_1} = I$$

$$\Rightarrow I_o = I = \frac{V_s}{R_1}$$

(13)

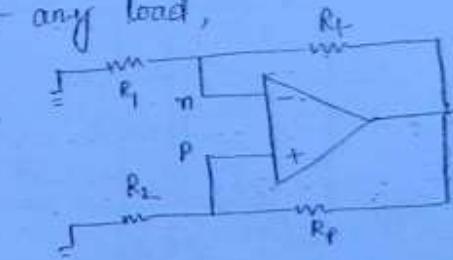
(b) Furnished Load :-



Circuit E

above all, as well as  $-ve$  feedback is present for the stability system should have  $-ve$  feedback and hence  $-ve$  feedback should be more than the feedback.

→ Without any load,



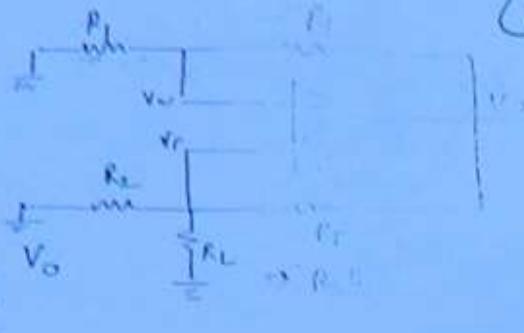
$$V_n = \frac{R_1}{R_1 + R_f} V_o$$

$$V_p = \frac{R_2}{R_2 + R_f} V_o$$

for stability,  $V_n \geq V_p$  ( $-ve$  feedback more than  $+ve$  feedback)

$$\Rightarrow \frac{R_1}{R_1 + R_f} \geq \frac{R_2}{R_2 + R_f}$$

(14)

After  $R_L$ ,

$$R_2' = R_2 \parallel R_L < R_2$$

$$V_p' = \frac{R_2'}{R_2 + R_p} V_o = \frac{1}{1 + R_p/R_2'} V_o$$

Now  $R_2' < R_2$  and hence  $V_p' < V_p$   $\Rightarrow$  +ve feedback  $\therefore$  which is favourable for stability.  
Hence, even after applying  $R_L$ , if original condition is violated, then the system will remain in -ve feedback.

Now, from circuit I —

$$I_D = \frac{V_p}{R_L} \quad \text{--- (1)}$$

$$V_F = V_n \quad \text{--- (2)} \quad \text{by concept of virtual short}$$

Applying KCL at inverting terminal —

$$\frac{V_p - V_S}{R_2} + \frac{V_p - 0}{R_L} + \frac{V_p - V_o}{R_p} = 0$$

$$\Rightarrow V_p \left[ \frac{1}{R_2} + \frac{1}{R_L} + \frac{1}{R_p} \right] - \frac{V_S}{R_2} - \frac{V_o}{R_p} = 0 \quad \text{--- (3)}$$

KCL at non-inverting terminal —

$$\frac{V_p - V_S}{R_2} + \frac{V_p - 0}{R_L} + \frac{V_p - V_o}{R_p} = 0$$

$$\Rightarrow V_p \left[ \frac{1}{R_2} + \frac{1}{R_L} + \frac{1}{R_p} \right] - \frac{V_S}{R_2} - \frac{V_o}{R_p} = 0 \quad \text{--- (4)}$$

from (3) &amp; (4) —

$$\Rightarrow V_p \left[ \frac{1}{R_2} + \frac{1}{R_L} + \frac{1}{R_p} \right] - \frac{1}{R_p} \left[ 1 + \frac{R_F}{R_1} \right] V_p = \frac{V_S}{R_2}$$

$$\Rightarrow V_p \left[ \frac{1}{R_2} + \frac{1}{R_L} + \frac{1}{R_P} - \frac{V_o}{R_P} - \frac{R_f}{R_1 R_P} \right] = \frac{V_s}{R_2} \quad (B)$$

$$\Rightarrow V_p \left( \frac{R_1 R_P R_L + R_1 R_2 R_P - R_f R_2 R_L}{R_1 R_P R_2 R_L} \right) = \frac{V_s}{R_2}$$

$$\frac{V_F}{R_L} \Rightarrow V_p = \frac{V_s R_1 R_P R_L}{R_L [R_1 R_P - R_2 R_f] + R_1 R_2 R_P}$$

$$I_o = \frac{V_p}{R_L} = \frac{V_s R_1 R_P}{R_L [R_1 R_P - R_2 R_f] + R_1 R_2 R_P}$$

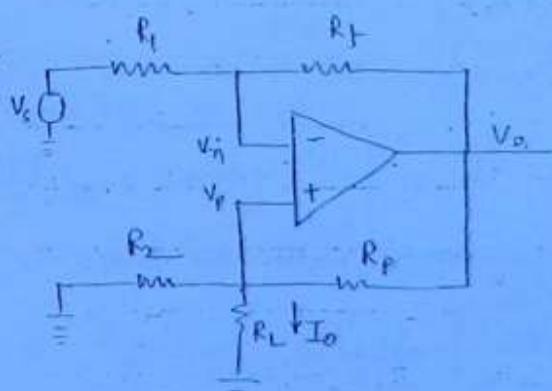
If  $R_1 R_P = R_2 R_f$

or  $\frac{R_P}{R_2} = \frac{R_f}{R_1}$  \*\*def  $\rightarrow$  Balanced Bridge condition

then,

$$I_o = \frac{V_s}{R_2}$$
 result

(ii)



Prove that if  $\frac{R_f}{R_1} = \frac{R_P}{R_2}$

then  $I_o = -\frac{V_s}{R_2}$

Applying KCL at  $V_p$  —

$$V_p \left[ \frac{1}{R_2} + \frac{1}{R_L R_P} \right] = \frac{V_o}{R_P} \quad (1)$$

Soln:  $V_n = V_p$  ;

$$\frac{V_F}{R_L} = I_o \quad (2)$$

Applying KCL at  $V_0$  -

$$\frac{V_p - V_L}{R_1} + \frac{V_p - V_0}{R_f} = 0 \quad \text{--- (3)}$$

$$\Rightarrow V_p \left[ \frac{1}{R_1} + \frac{1}{R_f} \right] = \frac{V_s}{R_1} + \frac{V_0}{R_f}$$

Putting  $V_0$  from eqn (1) -

$$\Rightarrow V_p \left[ \frac{R_f + R_1}{R_f R_1} \right] = \frac{V_s}{R_1} + \frac{R_f}{R_f} \cdot V_p \left[ \frac{1}{R_2} + \frac{1}{R_2 + R_f} \right]$$

On simplifying -

$$V_p = \frac{-R_2 R_L R_f R_f V_s}{R_1 R_2 R_f^2 + R_L R_f (R_f R_1 - R_2 R_f)} \quad \text{--- (4)}$$

from circuit -

$$I_o = \frac{V_p}{R_L} = \frac{-R_2 R_f R_f V_s}{R_1 R_2 R_f^2 + R_L R_f (R_f R_1 - R_2 R_f)}$$

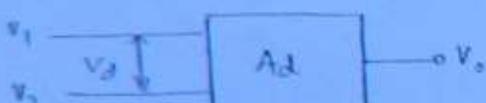
$$\text{If } R_f R_1 = R_2 R_f -$$

$$\text{or } \frac{R_f}{R_1} = \frac{R_f}{R_2} \Rightarrow \frac{1}{R_2} = \frac{R_f}{R_1 R_f}$$

$$I_o = -\frac{V_s \cdot R_f}{R_1 R_f} = -\frac{V_s}{R_2}$$

Differential Amplifier :-

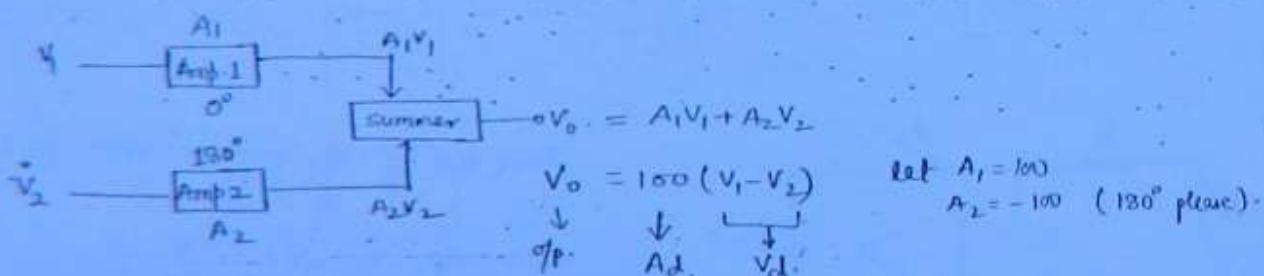
Ideal



$$V_d = A_d \cdot V_d$$

$$V_d = V_1 - V_2 = \text{difference voltage}$$

Ad = Difference Gain



→ To write the above eqn,  $A_1$  and  $A_2$  should be equal with  $180^\circ$  phase diff., but it is not possible to have identical amplifiers.

$$\text{Eq. } V_d = 100V_1 - 100V_2 = 100(V_1 - V_2) + (\text{Noise})$$

→ If there is some noise signal is present at both terminal and ideally it should cancel out but for unidentical amplifiers -

$$V_o = 100(V_1 + V_n) - 90(V_2 + V_n)$$

$$= 90(V_1 - V_2) + (10V_1 + 10V_n) \text{ Noise.}$$

(12)

For Practical Amplifier,

$$V_o = A_d V_d + A_c V_c \quad \text{--- (1)}$$

where

$$V_d = V_1 - V_2 \quad \text{--- (2)}$$

$$V_c = \frac{V_1 + V_2}{2} = \text{common mode signal} \quad \text{--- (3)}$$

$A_c$  = common mode gain

→ Ideally  $A_c \rightarrow 0$

Practically  $A_c \rightarrow$  very small.

→ Common Mode Rejection Ratio -

for Diff Amp 1 →  $A_c = 10$ ,  $A_d = 1000 \Rightarrow \frac{A_d}{A_c} = 100$

" " " 2 →  $A_c = 1$ ,  $A_d = 10 \Rightarrow \frac{A_d}{A_c} = 10$ .

→ Amp 1 is better than Amp 2.

$$\text{CMRR} = \beta = \frac{|A_d|}{|A_c|}$$

ideally  $\text{CMRR} = \infty$

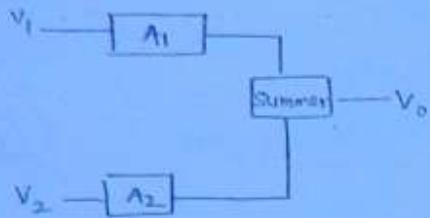
Practically  $\text{CMRR} = 10^6 = 120 \text{ dB}$

$$[\text{CMRR}]_{\text{dB}} = 20 \log \frac{|A_d|}{|A_c|}$$

→ CMRR is figure of merit of practical op-amp.

from diagram -

$$V_o = A_1 V_1 + A_2 V_2 \quad \text{--- (4)}$$



(18)

Adding (2) and (3) -

$$2V_C + V_d = 2V_1$$

$$\Rightarrow V_1 = V_C + \frac{V_d}{2} \quad \text{--- (5)}$$

Subtracting (2) from (3) -

$$2V_C - V_d = 2V_2$$

$$\Rightarrow V_2 = V_C - \frac{V_d}{2} \quad \text{--- (6)}$$

Putting

Putting 5 & 6 in (4) -

$$V_o = A_1 \left( V_C + \frac{V_d}{2} \right) + A_2 \left( V_C - \frac{V_d}{2} \right)$$

$$V_o = \left[ \frac{A_1 - A_2}{2} \right] V_d + (A_1 + A_2) V_C \quad \text{--- (7)}$$

Comparing (1) and (7) -

$$\boxed{A_d = \frac{A_1 - A_2}{2} \quad | \quad A_c = A_1 + A_2}$$

↳ Here  $A_2 = -ve$  due to  $180^\circ$  phase diff  
and hence  $A_d > A_c$ .

2nd Method :-

Calculation of  $A_c$  :-

$$\text{Put } V_1 = V_2 = V_s \Rightarrow V_C = V_s$$

$$\because V_d = 0, \Rightarrow V_o = A_d V_d + A_c V_C$$

$$\Rightarrow V_o = 0 + A_c V_s$$

$$\Rightarrow \boxed{A_c = \frac{V_o}{V_s}}$$

Calculation of  $A_d$  :-

$$\text{Put } V_1 = V_s/2 \text{ and } V_2 = -V_s/2$$

$$\Rightarrow V_d = V_s' \text{ and } V_C = 0$$

$$\Rightarrow \boxed{A_d = \frac{V_o}{V_s'}}$$

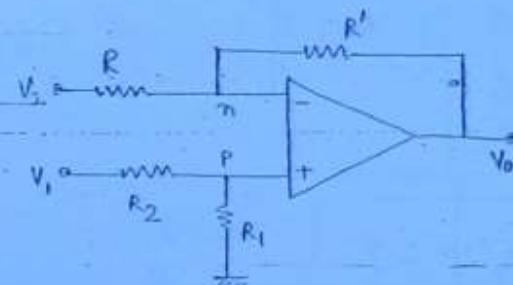
Ques: The circuit shown is a differential amplifier using an ideal op-amp.

(a) Find the opf voltage  $V_o$ .

(b) Find CMRR.

(c) Show that if  $CMRR = \infty$  if  $\frac{R'}{R} = \frac{R_1}{R_2}$ .

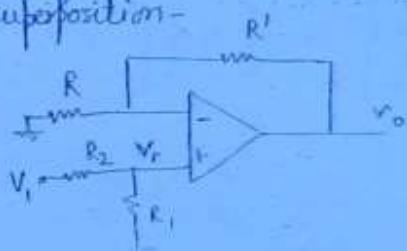
(19)



Soln: (a) By applying superposition -

(i) Taking  $V_2 = 0$  -

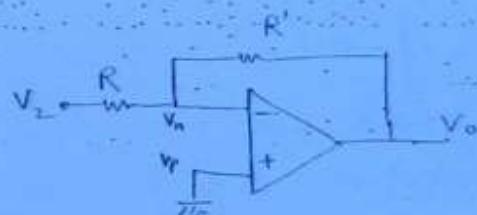
$$V_p = \frac{V_1 R_1}{R_1 + R_2}$$



$$V_{01} = \left[ 1 + \frac{R'}{R} \right] V_p \Rightarrow V_{01} = \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1 + R_2} \right] V_1$$

(ii) Taking  $V_1 = 0$  -

$$V_{02} = -\frac{R'}{R} V_2$$



$$\therefore V_0 = V_{01} + V_{02} = \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1 + R_2} \right] V_1 - \frac{R'}{R} V_2$$

$$(b) V_0 = A_d V_d + A_c V_c = A_1 V_1 + A_2 V_2$$

$$A_c = A_1 + A_2 \Rightarrow A_c = \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1 + R_2} \right] - \frac{R'}{R}$$

$$A_d = \frac{A_1 - A_2}{2} \Rightarrow A_d = \frac{1}{2} \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1 + R_2} \right] + \frac{R'}{2R}$$

$$CMRR = \frac{1}{2} \cdot \frac{\left[ \left\{ 1 + \frac{R'}{R} \right\} \left\{ \frac{R_1}{R_1+R_2} \right\} + \frac{R'}{R} \right]}{\left[ \left\{ 1 + \frac{R'}{R} \right\} \left\{ \frac{R_1}{R_1+R_2} \right\} - \frac{R'}{R} \right]}$$

(20)

(c) When  $\frac{R'}{R} = \frac{R_1}{R_2}$  —

$$CMRR = \frac{1}{2} \cdot \frac{\left[ \left\{ 1 + \frac{R_1}{R_2} \right\} \left\{ \frac{R_1}{R_1+R_2} \right\} + \frac{R_1}{R_2} \right]}{\left[ \left\{ 1 + \frac{R_1}{R_2} \right\} \left\{ \frac{R_1}{R_1+R_2} \right\} - \frac{R_1}{R_2} \right]}$$

$$\therefore CMRR = \frac{1}{2} \cdot \left[ \left\{ 2 \frac{R_1}{R_2} \right\} \div \left\{ 0 \right\} \right]$$

$$\Rightarrow CMRR = \infty$$

$$\therefore CMRR = \infty \Rightarrow |A_C| = 0 \text{ or } |A_{V1}| = |A_{V2}|$$

$$\Rightarrow \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1+R_2} \right] - \frac{R'}{R} = 0$$

$$\Rightarrow \frac{R_1}{R_1+R_2} + \frac{R'}{R} \left[ \frac{R_1}{R_1+R_2} - 1 \right] = 0$$

$$\Rightarrow \frac{R_1}{R_1+R_2} - \frac{R_1 \cdot R'}{R(R_1+R_2)} = 0$$

$$\Rightarrow \frac{R_1}{R_2} = \frac{R'}{R} \quad \text{Hence Proved}$$

3rd method:

$$\text{Put } V_1 = V_2 = V_C \Rightarrow V_{O1} = 0, V_C = V_S.$$

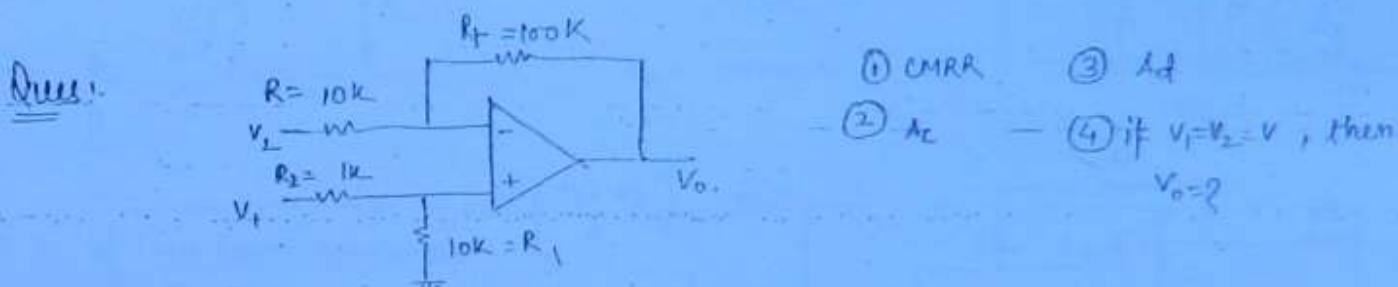
$$A_C = V_O/V_C \quad (1)$$

$$V_O = \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1+R_2} \right] V_S - \frac{R'}{R} V_S \Rightarrow A_C = \frac{V_O}{V_S} = \left[ 1 + \frac{R'}{R} \right] \left[ \frac{R_1}{R_1+R_2} \right] - \frac{R'}{R}$$

For  $A_d$ ,  $V_1 = \frac{V_s'}{2}$ ,  $V_2 = -\frac{V_s'}{2} \dots \Rightarrow A_d = \frac{V_o}{V_s'} \text{ and } A_c = 0$ .

$$V_o = \left[ 1 + \frac{R_f}{R} \right] \left[ \frac{R_1}{R_1 + R_2} \right] \cdot \frac{V_s'}{2} + \frac{R_f}{R} \frac{V_c'}{2}. \quad (2)$$

$$A_d = \frac{V_o}{V_s'} = \frac{1}{2} \left[ \left( 1 + \frac{R_f}{R} \right) \left[ \frac{R_1}{R_1 + R_2} \right] + \frac{R_f}{R} \right].$$



Soln:- → for objective, first check  $\frac{R_f}{R} = \frac{R_1}{R_2}$

$\Rightarrow \frac{100}{10} = \frac{10}{1} \Rightarrow$  Since ratio is equal  $\Rightarrow CMRR = \infty$   
 $\Rightarrow A_c = 0$

④  $\rightarrow V_o = A_d V_d + A_c V_c$

③  $A_d = ?$

$\Rightarrow V_o = A_d (V_1 - V_2)$

$\because A_c = 0 \Rightarrow |A_1| = |A_2|$

$\therefore V_o = 0$

$V_o = A_1 V_1 + A_2 V_2$

$\Rightarrow V_o = A_1 [V_1 - V_2]$

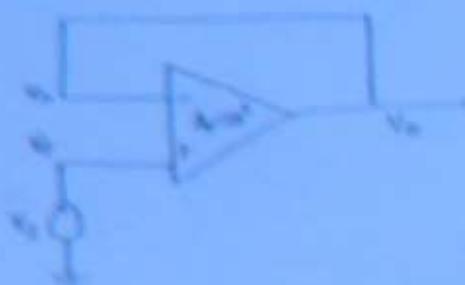
from  
previous ques

$A_2 = \frac{-R_f}{R} = -10 \Rightarrow A_1 = 10$

$\therefore A_d = \frac{A_1 - A_2}{2} = \frac{10 - (-10)}{2} = 10$

Aug 2011

## Voltage Followers



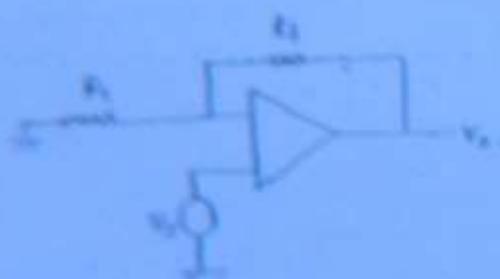
$$V_f = V_{in} + V_o$$

22

$$V_f = V_{in} + V_o$$

-s. Output follows the input hence called *voltage follower*.

$$\rightarrow A_{v_f} = \frac{V_o}{V_i} = 1 \quad f_f = 0$$



$$V_o = \left( 1 + \frac{R_f}{R_i} \right) V_i$$

when  $R_f \ll R_i$ ,

$$[V_o = V_i] - \text{ADM as voltage follower}$$

-s. Voltage follower is voltage series feedback.

-s. For voltage series feedback,  $f_f$  &  $R_o L$ .

-s. For voltage follower,  $[R_f = 10^6 \Omega]$  and  $[R_i = 50 \Omega = 0]$

-s. Because of -ve feedback,  $|A_{v_f}| = 1 \Rightarrow [f_f = 10^6 \text{ Hz} = 1MHz]$

## Application:-

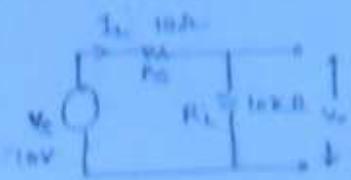
- It is used in designing of couple and half circuit.

- first order butterworth filter.

- as a buffer, ie., impedance matching device b/w high resistance and low resistance.

→ Application of a buffer -

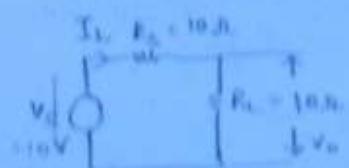
(23)



$$V_B = \frac{I_C}{R_L + R_E} \approx V_E \quad \because R_L \gg R_E$$

$$I_E = \frac{V_B}{R_E + R_L} \leq \frac{10}{10k\Omega} = 1mA$$

$R_E \uparrow$  and  $I_E \downarrow$   
there is low loading effect.



$$M_E = \frac{10}{10 + 10} = 1.5V$$

$$T_E = \text{given value}$$

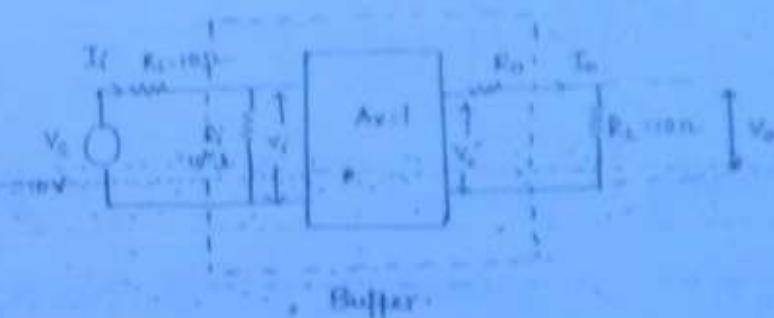
$R_E \downarrow$  &  $T_E \uparrow$   
there is high loading effect.

Low Load Resistance :-

→  $R_L = \text{very low}$   $\Rightarrow$  loading effect  
 $I_E = \text{very high}$   $\Rightarrow$  high.

Low Load :-

- a) Loading-loading effect
- b)  $I_E = \text{low}$
- c)  $R_E = \text{high}$



$$\because R_L \gg R_E \Rightarrow V_L \approx V_B$$

$$\because A_V=1, \quad V_{B'} \approx V_L \approx V_B, \quad \text{and} \quad \because R_E = 0 \Rightarrow V_0 = V_{B'} \approx V_B$$

$$I_E = \frac{10}{10 + 10} = 1mA \rightarrow \text{very small.}$$

$$I_E = \frac{10}{10} = 1A \rightarrow \text{this extra current is given by buffer.}$$

Other buffers -

→ Voltage follower by using op-amp  $\rightarrow$  VCVS  $\rightarrow$  Current follower using JFET  $\rightarrow$  JCVS

→ Emitter  $\rightarrow$  GFT  $\left\{ \begin{array}{l} \text{Common collector} \\ \text{config} \end{array} \right\}$  (emitter chain  $\rightarrow$  VCVS config).

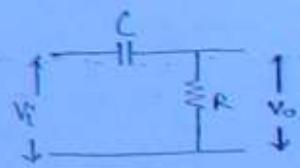
## Linear Wave Shaping circuit :-

(24)

- High Pass RC  $\rightarrow$  Differentiator.
- Low Pass RC  $\rightarrow$  Integrator

The process whereby form of a non-sinusoidal signal is altered by transmission through a linear filter is called linear wave shaping.

### 1) High Pass RC circuit -



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

$$V_o = \frac{R}{R + 1/\omega C} \cdot V_i$$

$$\Rightarrow \frac{V_o}{V_i} = \frac{1}{1 + \frac{1}{\omega RC}} \Rightarrow A = \frac{1}{1 + \frac{j}{\omega RC}} \quad \text{--- (1)}$$

$$\Rightarrow |A| = \frac{1}{\sqrt{1 + 1/\omega^2 R^2 C^2}}, \quad \phi \text{ shift} = -\tan^{-1}\left(\frac{1}{\omega RC}\right) \quad \text{--- (2)}$$

$\rightarrow$  Since  $\phi$  shift = +ve, it is called leading circuit.

$\rightarrow$  from eqn (2) — as  $\omega \uparrow$ , gain  $\uparrow$

$\rightarrow$  At  $\omega = 0$ ,  $|A| = 0$

$\rightarrow$  At  $\omega = \infty$ ,  $|A| = 1 \equiv A_{\max}$

$\rightarrow$  At  $\omega = \omega_L$ ,  $|A| = \frac{A_{\max}}{\sqrt{2}}$ ;  $\omega_L$  = cut-off frequency, freq at which gain reduces to  $\frac{1}{\sqrt{2}}$  of max. value (3dB).

$$\Rightarrow \frac{1}{\sqrt{2}} = \frac{1}{\sqrt{1 + 1/\omega_L^2 R^2 C^2}}$$

$$\Rightarrow \boxed{\omega_L = 2\pi f_L = 1/RC} \quad \text{--- (4)} \quad \omega_L = 3\text{dB frequency}$$

from (1) and (4) -

$$A = \frac{1}{1 - j/\omega_0 RC} \Rightarrow A = \frac{1}{1 - j\omega_0/\omega} = \frac{1}{1 - j\tau_0/f}$$

(25)

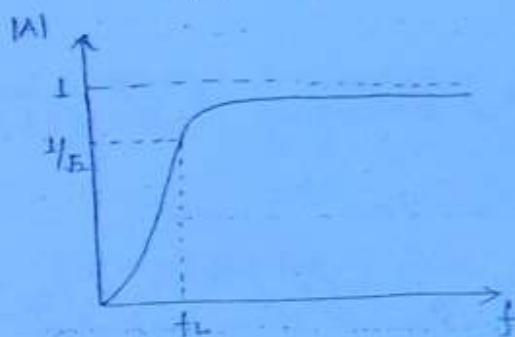
$$|A| = \sqrt{\frac{1}{1 + (\tau_0/f)^2}} \quad \text{where } f = \text{instantaneous frequency}$$

$$\Rightarrow f=0 ; |A|=0$$

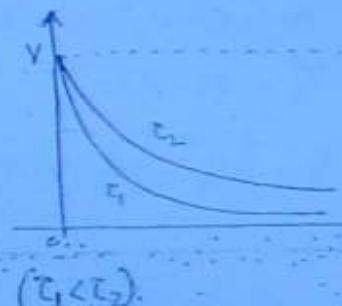
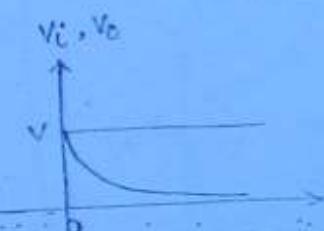
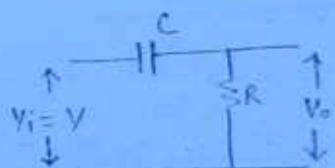
$$f=f_L ; |A| = 1/\sqrt{2}$$

$$f=\infty ; |A|=1$$

$$\Rightarrow \boxed{B.W. = \infty}$$



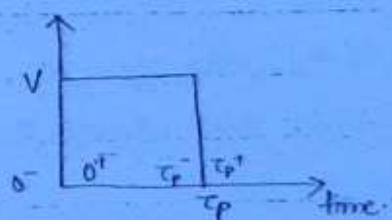
Step Response :



$$\Rightarrow V_o = V_R = V e^{-t/RC} \rightarrow \text{decays exponentially}$$

$$\Rightarrow V_c = V [1 - e^{-t/RC}] \rightarrow \text{decays exponentially}$$

Pulse Response



$$\text{Case 1} \quad \frac{\tau_p}{RC} \ll 1$$

$\tau_p$  = pulse width.

$$\text{Case 2} \quad \frac{\tau_p}{RC} \gg 1$$

$$\rightarrow V_i(0^-) = 0, \quad V_i(0^+) = V$$

$$V_i(\tau_p^-) = 0V, \quad V_i(\tau_p^+) = 0V$$

(26)

$$\rightarrow V_t(0^-) = 0 = V_t(0^+), \quad \Rightarrow V_R(0^+) = V$$

$$V_C = V [1 - e^{-t/RC}] ; \quad V_R = V e^{-t/RC}$$

Case 1  $RC \gg \tau_p$

$\rightarrow V_R$  will start discharging till  $0 < t < \tau_p$

$$\text{At } t = \tau_p^-, \quad -V_o = V e^{-\tau_p/RC} = V'$$

$$\therefore V_C = V - V' = V (1 - e^{-\tau_p/RC}) = V_C(\tau_p^-)$$

At  $t = \tau_p^+$  -

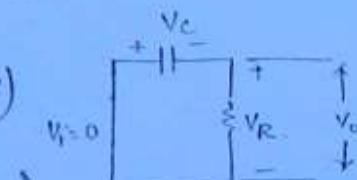
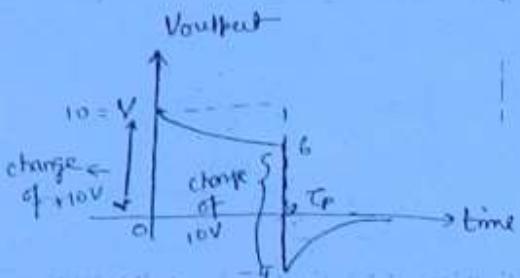
$$V_C(\tau_p^+) = V_C(\tau_p^-) = V (1 - e^{-\tau_p/RC}) = (V - V')$$

$\forall V_i = 0$  at  $\tau_p(0^+)$

$$V_R = 0 \rightarrow V_C' = -V (1 - e^{-t/RC}).$$

$$V_C = (V - V') e^{-t/RC}$$

$$V_o = -V_C = -(V - V') e^{-t/RC}, \quad (t = t - \tau_p)$$



Time	$V_o$
$t < 0$	0
$t = 0^+$	$V$
$0 < t < \tau_p$	$V_o = V e^{-t/RC}$
$t = \tau_p^-$	$V' = V e^{-\tau_p/RC}$
$t = \tau_p^+$	$-(V - V')$
$t > \tau_p$	$-(V - V') e^{-t/RC}$
	$t' = t - \tau_p$

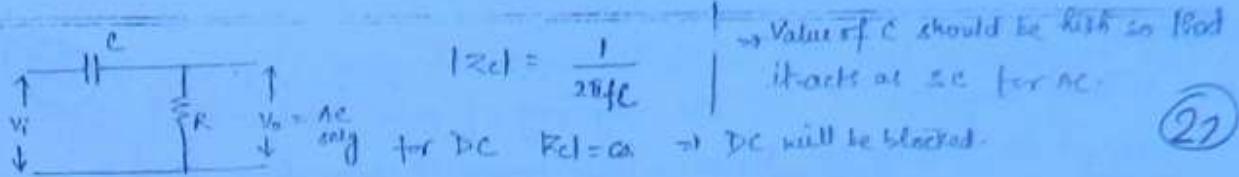
$\rightarrow$  When there is sudden change in  $i/p$ ,  
the same change will occur at the o/p.

$\rightarrow$  If the ~~o/p~~ input is maintaining some constant level, then output will tend towards zero (between 0 to  $\tau_p$ ) and from ( $\tau_p$  to  $\infty$ ), the o/p is tending towards 0.

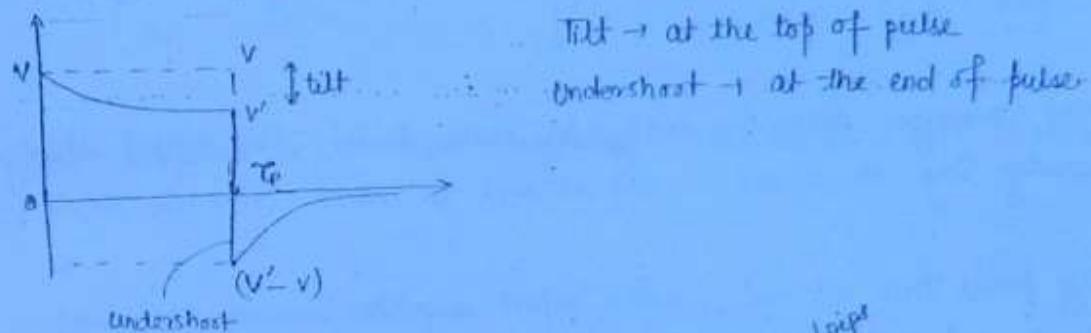
$\rightarrow$  Area of the pulse =  $V \tau_p = +ve$  = average DC level.

$$\text{For output, } A_+ \text{ area} = \int_0^{\tau_p} V e^{-t/RC} dt, \quad A_- = \int_{\tau_p}^{\infty} (V - V') e^{-t-\tau_p/RC} dt$$

$$A_+ = A_- = (\text{as charging} = \text{discharging})$$



$\rightarrow$  Avg. level of o/p in high pass RC signal is always 0 irrespective of the avg. level of i/p.



Case 2:  $R_C \ll \tau_p$

$$V_C(0^-) = 0 = V_C(0^+) \Rightarrow V_0 = V$$

$$V_R = V e^{-t/\tau_p} \text{ for } 0 < t < t_1$$

At  $t = t_1^-$ :

$$V_C = V \Rightarrow V_R = 0$$

At  $t = \tau_p^-$ ,

$$V_C = V$$

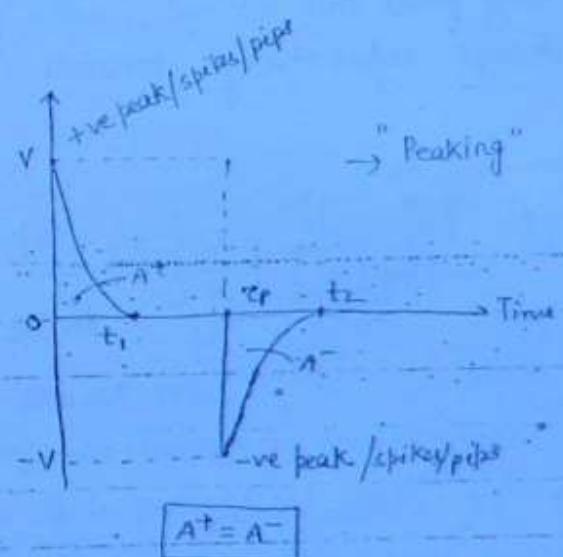
At  $t = \tau_p^+$ ,  $V_C = V$ ,  $V_R = 0$

$$V_i = 0 \Rightarrow V_R = -V$$

Now, a capacitor will start discharging,  $V_C = V e^{-t/\tau_p}$ , hence

$$V_R = V_0 = -V e^{-t/\tau_p} \Rightarrow (t' = t - \tau_p)$$

At  $t = t_2$ ,  $V_C = 0$  and  $\Rightarrow V_R = V_0 = 0$ .

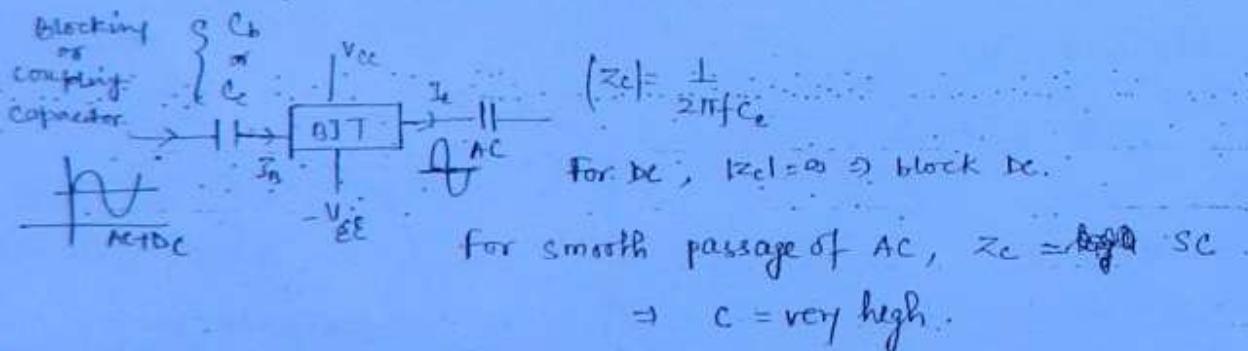


## Conclusion-

- Positive spike of amplitude 'V' at the beginning of pulse and -ve of same size at the ending of pulse. This process of converting pulse into spike by means of a high pass RC circuit of short time constant is called Peaking.

For HPF -

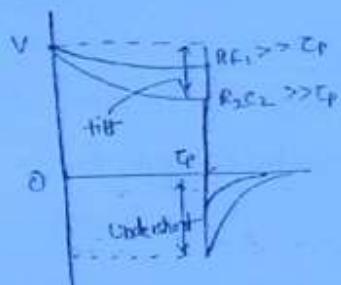
- Average level of o/p is always zero, independent of average level of i/p, i.e.,  $A_+ = A_-$
- When input changes discontinuously by amount 'V', The output changes discontinuously by an equal amount and in same direction.
- During any finite time interval, when input maintains a constant level, o/p decays exponentially towards '0' voltage.



But if AC  $\rightarrow$  20Hz - 20kHz,  
 the low freq. components will not pass smoothly as for low frequencies  
 $|Z_c| = \text{high}$ . Hence there will be distortion in the o/p. Whereas,  
 the high freq. component will be received accurately at the o/p.

- This capacitor provides DC isolation to the BJT. The DC is blocked as it will interfere with the biasing condition of the BJT.

→ For  $RC \ggg \tau_p$ , pulse distortion will ↓ as tilt and undershoot will be small. (29)

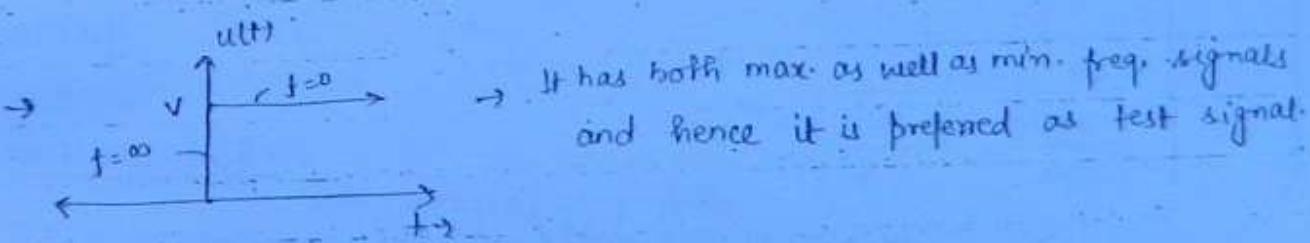
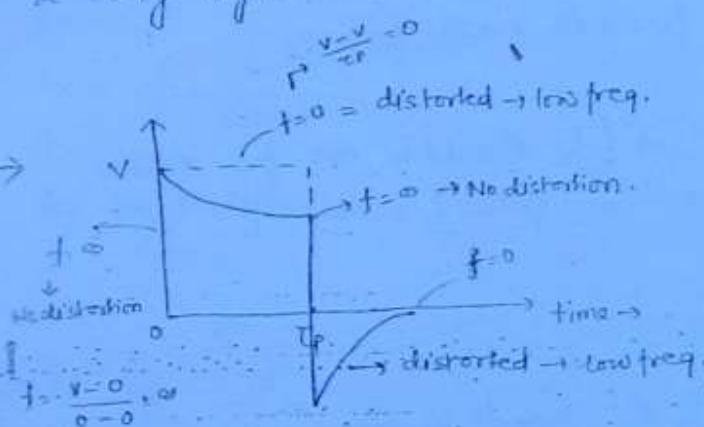


$R_1C_1 > R_2C_2$ . If we keep ↑ RC, distortion will keep ↓. Tilt and undershoot are distortions.

→ C should be high for minimum distortion. (same as previous discussion).

$C_c$  →  $R_i$  should also be high.  
Similar to HPF.

→ For better coupling or minimum distortion,  $R_i$  and  $C_c$  both should be very high.



### Tilt or Sags :-

→  $RC \ggg \tau_p$ .

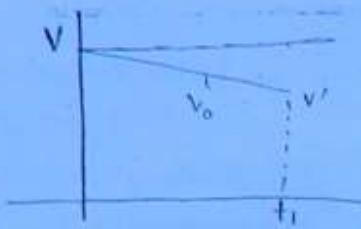
→  $V_0 = V e^{-t/RC}$

$$\chi = t/RC = \text{very small}$$

$$e^{-\chi} = 1 - \chi + \frac{\chi^2}{2!} - \frac{\chi^3}{3!}$$

$$\therefore e^{-\chi} = 1 - \chi$$

$$\Rightarrow V_o = V \left[ 1 - \frac{t}{RC} \right] \quad \textcircled{1}$$



$$\text{Tilt at } t=t_1 = V - V' \quad \textcircled{2}$$

$$\% \text{ tilt} = P\% = \frac{V - V'}{V} \times 100 \quad \textcircled{3}$$

$$V' = V \left[ 1 - \frac{t_1}{RC} \right] \Rightarrow V - V' = \frac{V t_1}{RC}$$

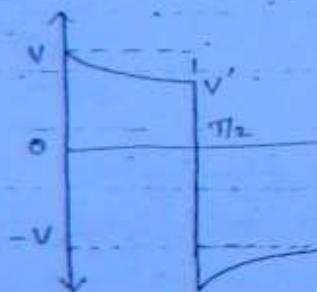
$$\Rightarrow \frac{V - V'}{V} = \frac{t_1}{RC}$$

$$\Rightarrow \boxed{P\% = \frac{t_1 \times 100}{RC}} \quad \text{**} \rightarrow (\text{When } RC \text{ is } \uparrow, \text{ tilt will } \downarrow)$$

Since,  $2\pi f_L = \omega_L = 1/RC = \{ t_1 = 3 \text{ dB cutoff frequency} \}$ .

$$\boxed{P\% = 2\pi f_L t_1 \times 100\%} \quad \text{**} \rightarrow (f_L \text{ should be low for smaller tilt})$$

= Tilt for symmetrical Square Wave —



$$f = \frac{1}{T} = \text{freq. of sq. wave}$$

$$t_1 = T/2 = \frac{1}{2f}$$

$$P\% = 2\pi f_L \left( \frac{1}{2f} \right) \times 100\%$$

$$\Rightarrow \boxed{P\% = \pi \left( \frac{f_L}{f} \right) \times 100\%} \quad \text{**}$$

$\hookrightarrow$  (for high frequency signal,  
 $\therefore P = \text{low.}$ ) and vice versa.

16<sup>th</sup> August, 2012

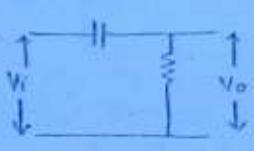
### High pass RC circuit as a differentiator :-

(31)

- When time constant,  $RC$ , is very very small as compared to time period of input signal, the circuit is called differentiator.

Criteria for good differentiator -

- Ideally,  $RC=0$ .



$$\frac{V_o}{V_i} = |A| = \frac{1}{1 + j\omega RC}$$

$$|A| = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}} ; \quad \phi = \tan^{-1} \left( \frac{1}{\omega RC} \right)$$

for  $V_i = V_{rms} \sin \omega t$

$$V_o = |A| \cdot V_{rms} \sin(\omega t + \phi)$$

$$= \frac{V_{rms}}{\sqrt{1 + \frac{1}{\omega^2 R^2 C^2}}} \cdot \sin(\omega t + \phi)$$

for ideal differentiator,  $\phi = 90^\circ \Rightarrow \omega RC = 0$ . (which is practically not possible).

	$\phi$	$\omega RC$
Ideal Diff.	$90^\circ$	0
Best Diff.	$89.4^\circ$	0.01
Good diff	$84.3^\circ$	0.1

for Best diff -

$$\omega RC = 0.01$$

$$\Rightarrow |A| = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}} \quad \text{but } \frac{1}{\omega^2 R^2 C^2} \ggg 1$$

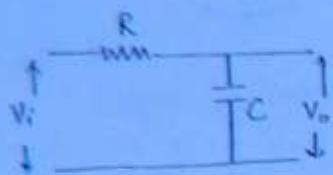
$\Rightarrow V_o = (0.01)V_{rms} \sin(\omega t + 89.4^\circ) \rightarrow$  Amplitude is very small effect differentiation.

To get amplitude,  $V_o$ , we have to follow it with a high gain amplifier

A high pass RC differentiator is always followed by a high gain amplifier

## Low Pass RC circuit :-

(3)



$$V_o = \frac{V_{CS}}{R + \frac{1}{\omega CS}} V_i \Rightarrow A = \frac{1}{1 + RCS}$$

$$\Rightarrow A = \frac{1}{1 + j\omega RC}$$

$$\rightarrow |A| = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}} ; \quad \phi = -\tan^{-1}(\omega RC) \rightarrow (\text{lagging})$$

$\phi = -ve$  circuit

$$\rightarrow \text{for } \omega = 0 ; \quad |A| = 1$$

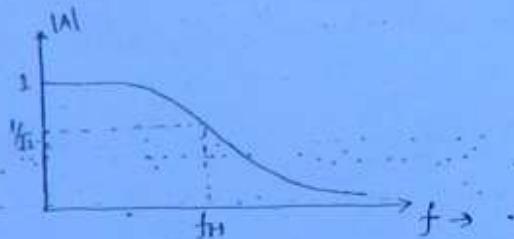
$$\omega = \infty ; \quad |A| = 0$$

$\omega \uparrow ; \quad |A| \downarrow \rightarrow \text{Low pass filter}$

$$\rightarrow \text{At } \omega = \omega_H \quad |A| = \frac{|A_{max}|}{\sqrt{2}} = \frac{1}{\sqrt{2}}$$

$$\Rightarrow \omega_H = \frac{1}{RC} \quad \text{or} \quad 2\pi f_H = \frac{1}{RC} \Rightarrow f_H = \frac{1}{2\pi RC} \rightarrow \text{3dB cutoff frequency}$$

$$\rightarrow A = \frac{1}{1 + j(f/f_H)}$$



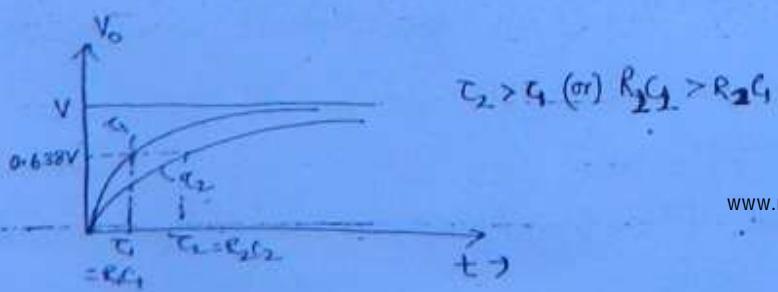
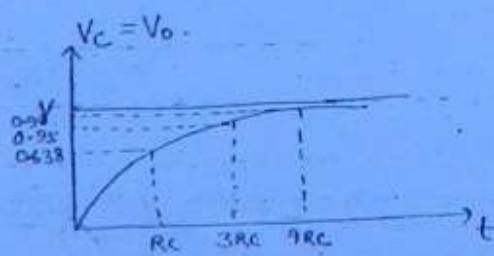
## Step Response :-

$$V_C(0^-) = V_C(0^+) = 0$$

$$\text{At } t=0^+, \quad V_C(0^+) = 0$$

$$\text{At } t=\infty, \quad V_C = V$$

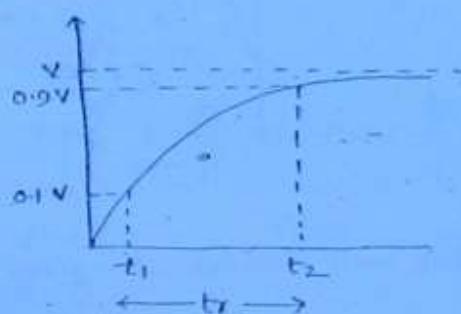
$$V_C = V [1 - e^{-t/RC}]$$



(33)

Rise Time :-  $(t_r)$  -

- It is the time taken by the signal to rise from 10% to 90% of its final value.



$$\text{At } t = t_1, \quad V_0 = 0.1V \Rightarrow 0.1V = V [1 - e^{-\frac{t_1}{RC}}]$$

$$\Rightarrow t_1 \approx 0.1RC$$

$$\text{At } t = t_2, \quad V_0 = 0.9V \Rightarrow 0.9V = V (1 - e^{-\frac{t_2}{RC}})$$

$$\Rightarrow t_2 \approx 2.3RC$$

Rise time =  $t_2 - t_1 = 2.2RC$

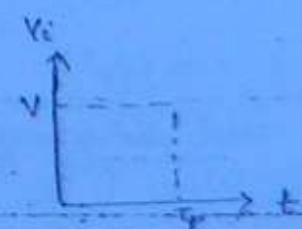
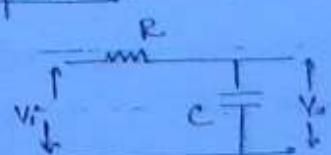
$$\therefore \omega_{HP} = 2\pi f_H = \frac{1}{RC}$$

$$\therefore t_r = \frac{2.2}{2\pi f_H} \Rightarrow t_r = \frac{0.35}{f_H}$$

→ Rise time of the circuit should be low for fast response.

→  $\Rightarrow f_H$  should be high

Pulse Response :-



$$V_c(0^-) = V_c(0^+) = 0V$$

$$\therefore V_o(0^+) = 0$$

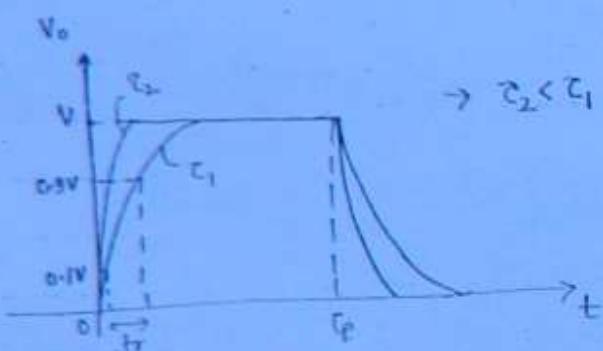
Case I :  $\frac{R_p}{RC} \gg 1$  (or)  $RC \ll \tau_p$

When  $RC$  is small, rate of charging is very fast.

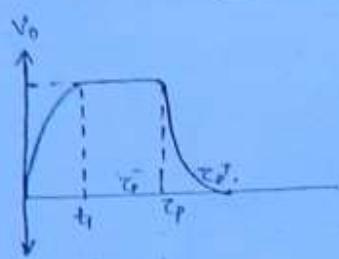
$$V_C = V \text{ and will remain } V \text{ till } t = \tau_p$$

$V_c(\tau_p^-) = V_c(\tau_p^+) = V$ , and now the capacitor will start discharging through  $R$ .

$$V_o = V_C = V e^{-(t-\tau_p)/RC} \quad \text{for } t > \tau_p$$



(34)



ideally (practically very small).

$\rightarrow \tau_2 < \tau_1$ , when  $RC=0$ , the capacitor will charge instantly and <sup>hence</sup> pulse shape will be preserved if

$$f_H \geq \frac{1}{\tau_p}$$

$$\rightarrow f_H \geq \frac{1}{\tau_p} \Rightarrow \frac{0.35}{\tau_r} > \frac{1}{\tau_p}$$

$$\Rightarrow \tau_r \leq 0.35 \tau_p$$

Case II :  $\frac{\tau_p}{RC} \ll 1$  or  $RC \gg \tau_p$

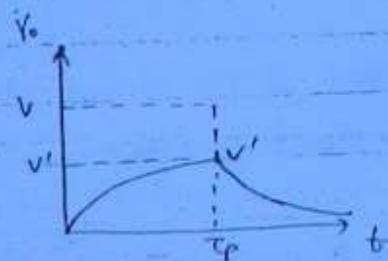
$RC$  is high, hence rate of charging is very slow.

At  $t = \tau_p$ ,

$$V' = V \left[ 1 - e^{-\tau_p/RC} \right] = V_c(\tau_p^+)$$

After  $t = \tau_p$ ,  $V_c$  will start discharging,

$$V_o = V' e^{-(t-\tau_p)/RC} \quad \text{for } t > \tau_p$$

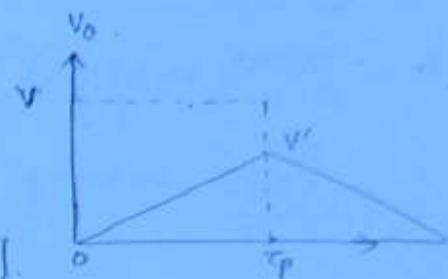


When  $RC$  is very high, then.

$$\alpha = \frac{t}{RC} \ll 1$$

$$\therefore e^{-\alpha} = 1 - \alpha + \frac{\alpha^2}{2!} - \frac{\alpha^3}{3!} \dots \Rightarrow e^{-\alpha} = [1 - \alpha]$$

(35)



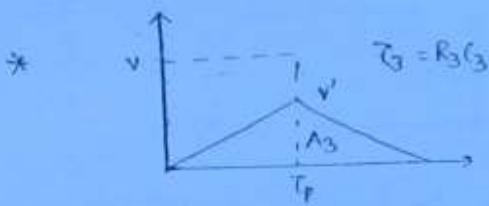
$$\therefore V_C = V \left[ 1 - \frac{t}{RC} \right] \Rightarrow V_C = \frac{V \cdot t}{RC} \quad (\text{linear equation}).$$

for discharging.

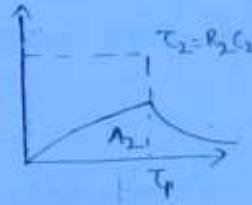
$$V_O = V'e^{-(t-t')/RC} = V'e^{-t'/RC}, \text{ for } RC \ggg t'$$

$$V_O = V' \left[ 1 - \frac{t'}{RC} \right] \quad (\text{linear eqn})$$

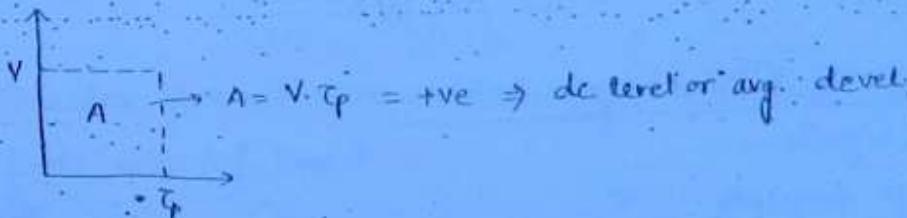
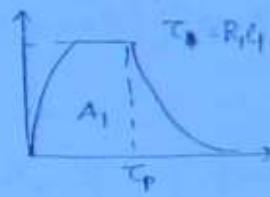
\* RC low pass circuit will act as an integrator when  $RC$  is very high.



$$\boxed{\tau_3 > \tau_2 > \tau_1}$$



$$\boxed{A = A_1 = A_2 = A_3}$$



\* for low pass RC circuit, dc level of output is always equal to dc level of input dc level.

### Low Pass RC as an integrator

→ When the time constant is very large as compared to time period of ip signal, the circuit is called integrator.

for  $V_i = V_m \sin \omega t$  —

$$V_o = |A| \cdot V_m \sin(\omega t + \phi)$$

$$V_o = \frac{V_m}{\sqrt{1+\omega^2 R^2 C^2}} \cdot \sin(\omega t + \phi)$$

$\phi$

wRC

(36)

Ideal  $\int$

-90°

$\infty$  (Practically  
not possible)

Best  $\int$

-89.4°

$R_C > 15.1T$

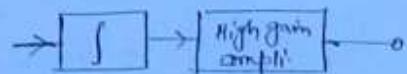
for Best  $\int$  —

$$-\tan^{-1}(wRC) = -89.4^\circ$$

$$\Rightarrow wRC = \tan(89.4^\circ)$$

$$\Rightarrow RC \times \frac{2\pi}{T} = \tan(89.4^\circ)$$

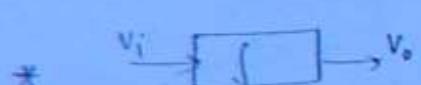
$$\Rightarrow RC = 15.1T$$



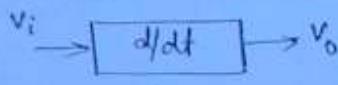
$$V_o = \frac{V_m}{wRC} \sin(\omega t - 89.4^\circ) \quad \text{for } wRC \gg 1$$

In this case also, amplitude is very low.

→ The op-amp is followed by a high gain amplifier.



$$\text{Laplace } \int \rightarrow \frac{1}{s} = \frac{1}{j2\pi f}$$



$$\text{Laplace } \frac{d}{dt} = s = j2\pi f$$

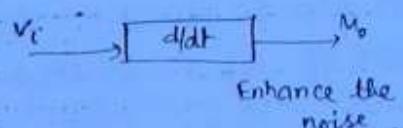
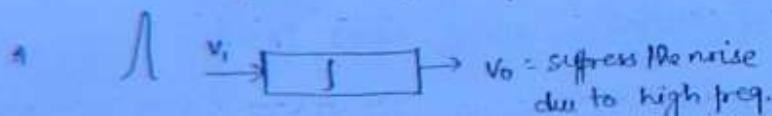
⇒  $f \uparrow$  then  $IN \downarrow$

⇒  $f \uparrow$  then  $IT \uparrow$

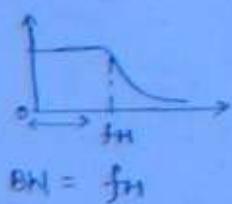
→ Integrator is preferred over differentiator because —

i) for spurious signals/Noise signals —

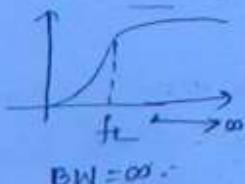
They are of high frequency.



2) for LPF,



for HPF,



→ Due to  $\propto$  BW, some unwanted signals will also come above the req. signal band.

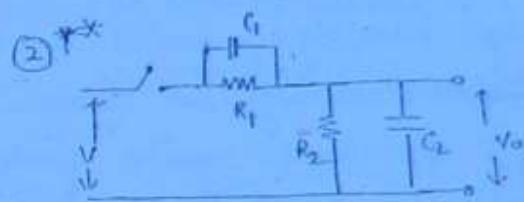
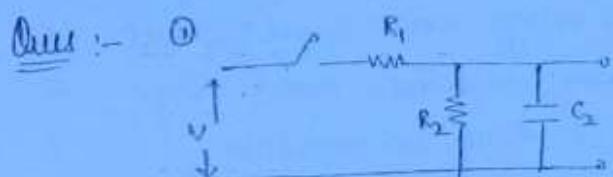
→ LPF is placed at the last stage of multi-stage amplifier so as to prevent the noise to reach the output.

(37)

→ Integrator is almost preferable over differentiator for following reasons-

① since gain of  $\int$   $\downarrow$  with  $f$ , whereas gain of  $d/dt \uparrow$  with  $f$   
therefore, it is easier to stabilize  $\int$  than  $d/dt$  w.r.t spurious oscillations (high freq. noise).

② As a result of its limited BW, an  $\int$  is less sensitive to ~~diff~~ noise voltage than a  $d/dt$ .



Switch is closed at  $t=0$ .  $V_o(0^+)$  = ?

Soln: ① At  $V_c(0^-) = 0 = V_c(0^+)$   $\Rightarrow V_o = 0$  at  $t = 0^+$  {  $I(0^+) = V/R_1 = \text{finite}$  }  
 $V_o(\infty) = VR_2/R_1 + R_2$

② Capacitor does not allow sudden change in voltage but only for finite value of current

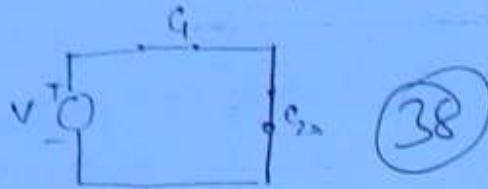
Wrong result:-  
 $V_{c_1}(0^-) = 0 = V_{c_1}(0^+)$   $\Rightarrow V_o = 0$   $\rightarrow$  all paths are short  
 $V_{c_2}(0^-) = 0 = V_{c_2}(0^+)$   $\Rightarrow V_o = 0$  from  $\partial/\partial_v = V_1 - V_2$   
 $\text{and } V_1 - V_2 = V \text{ from i/p.}$   
 $\text{KCL is violated.}$   
 $\therefore V_o = V_2 \neq 0$  is a wrong result.

Correct result:- Due to  $V_{c_1} = 0$ , the current in the circuit will be

for  $V_{c_1} = 0$  &  $V_{c_2} = 0$ ,  $V_o = I(0^+) = \infty$ . for  $I(0^+) = \infty$ , there will be a finite voltage in capacitors,

$\int_{0^-}^{0^+} I(t) dt = q(0^+) = \text{finite}$ . {  $I(t)$  will behave as impulsive current & hence it will allow sudden change of  $V_{c_1}$  }

$$q(0^+) = C_{eq} \cdot V = \frac{C_1 C_2}{C_1 + C_2} V.$$



(38)

$$V_o(t^+) = V_{C_2}(t^+) = \frac{q(0^+)}{C_2}$$

$$V_o(t^+) = \frac{C_1}{C_1 + C_2} V ; V_1(t^+) = \frac{q(0^+)}{C_1} = \frac{C_2 V}{C_1 + C_2}$$

→ voltages will be distributed b/w  $C_1$  and  $C_2$ .

At  $t=0$ ,  $C_1$  &  $C_2$  will be ac.

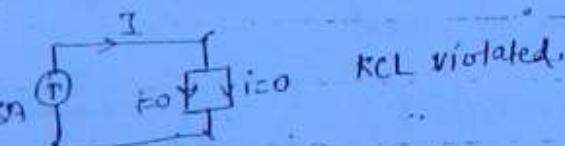
$$V_o = \frac{R_2}{R_1 + R_2} V.$$

Conclusion :-

at  $t=0^+$

→ As input changes abruptly by amount  $V$ , then voltage across  $C_1$  and  $C_2$  must also change discontinuously but voltage across capacitor cannot change instantaneously if current remains finite and hence an impulsive current must flow in the circuit

→ an infinite current exists for  $t=0^+$ , so that a finite charge  $q(0^+)$  is delivered to each capacitor and capacitor allows sudden change of voltage



RCL violated.

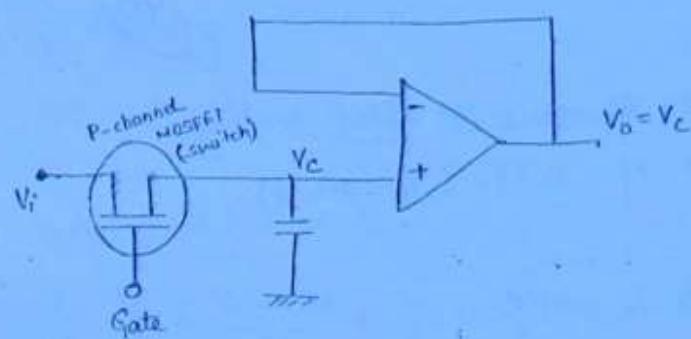
$$Z_L = \infty (\because i=0)$$

$$\Rightarrow V = 5 \times \infty = \infty = \text{impulsive voltage.}$$

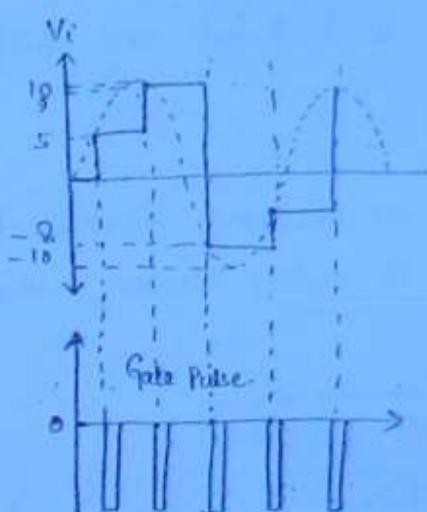
Hence it will allow sudden change in current.

$$I_{L_2} = \frac{5A}{L_1 + L_2} , I_{L_1} = \frac{5A}{L_1 + L_2}$$

## Sample and Hold circuit :-

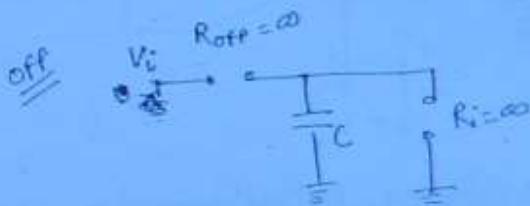
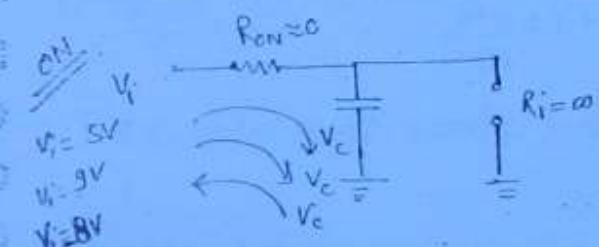


(Q4)



Switch       $R_{\text{switch}}$       Time constant

ON	$R_{\text{on}} \approx 0$	$R_{\text{on}}C \approx 0$	Capacitor will suddenly charge upto instantaneous value of $V_i$ .
OFF	$R_{\text{off}} \approx \infty$	$R_{\text{off}}C \approx \infty$	C will hold the value of $V_i$ .



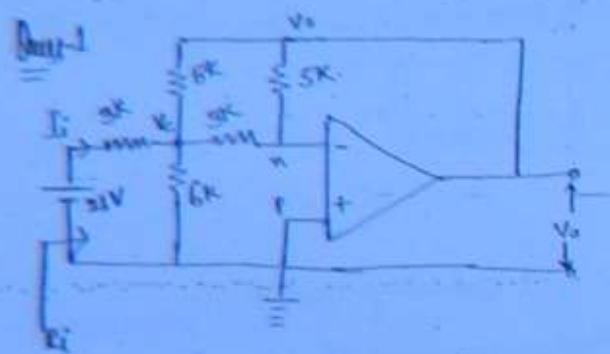
- \* When  $V_i$  is < the value held by capacitor.
- in that case C will discharge or C will charge towards value of  $V_i$ , which is smaller than its previous value.

\* -ve triggered P-MOSFET is used as trigger because -ve triggered pulses will not generate spikes/noise.

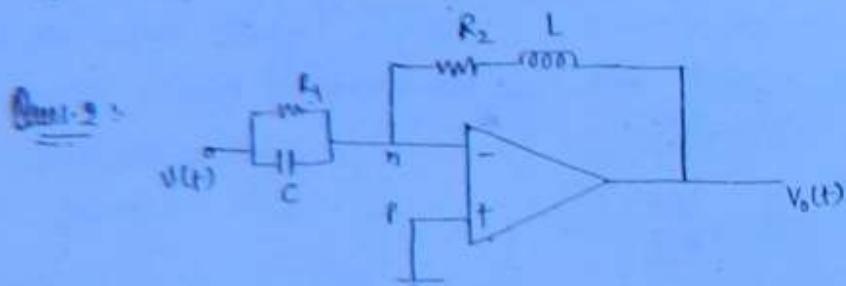
- The op-amp is used (voltage follower), because it will make the  $R_i = \infty$  which will help capacitor to hold the value, and C will not discharge through it. (if  $R_i$  = some finite value, the hold value will discharge through it).

- Also, MOSFET is preferred over BJT as a switch because of its zero offset voltage.
- MOSFET makes an excellent chopper (switch) because its offset voltage ( $\approx 5\mu V$ ) is much smaller than that of BJT. ( $V_T = 0.5V$ ).

(40)



Find  $V_C$ ,  $I_i$  and  $R_i$



$$-V_0 = \frac{R_2}{R_1} V + \left[ R_2 C + \frac{L}{R_1} \right] \frac{dV}{dt} + LC \frac{d^2V}{dt^2}$$

Sol(2) :- By applying virtual ground,  $V_p = V_n = 0$

Now applying KCL at  $V_n$  -

$$\frac{V_n - V}{Z_{R_1}C} + \frac{V_n - V_0}{Z_{R_2}L} = 0$$

$$\Rightarrow \frac{V}{Z_{R_1}C} + \frac{V_0}{Z_{R_2}L} = 0$$

$$\Rightarrow \frac{V}{\left( \frac{R_1 + R_1 CS}{R_1 + R_1 CS} \right)} + \frac{V_0}{R_2 + LS} = 0$$

$$\Rightarrow \frac{V(1 + R_1 CS)}{R_1} + \frac{V_0}{R_2 + LS} = 0$$

$$\Rightarrow V(1 + R_1 CS)(R_2 + LS) + V_0 R_1 = 0$$

$$\Rightarrow -V_0 R_1 = V \left[ R_2 + LS + R_1 R_2 C S + \frac{R_1 L C S^2}{R_1 + R_1 C S} \right]$$

$$\Rightarrow -V_0 = \frac{VR_2}{R_1} + \left[ \frac{L}{R_1} + R_2 C \right] SV + LC S^2 V$$

$$\Rightarrow -V_0 = \frac{R_2}{R_1} V + \left[ \frac{L}{R_1} + R_2 C \right] \frac{dV}{dt} + LC \frac{d^2V}{dt^2}$$

Soln by virtual ground method,  $V_P = V_N = 0$

(4)

Applying KCL at 'n' -

$$\frac{0 - V_C}{3} + \frac{0 - V_0}{5} = 0 \Rightarrow 5V_C + 3V_0 = 0 \quad \text{--- (1)}$$

Applying KCL at  $V_C$  -

$$\frac{V_C}{3} + \frac{V_C - 21}{3} + \frac{V_C}{6} + \frac{V_C - V_0}{8} = 0$$

$$\Rightarrow \frac{V_C}{3} + \frac{V_C - 21}{3} + \frac{V_C}{6} + \frac{V_C + 5V_0}{8} = 0 \quad \left\{ \text{from (1)} \right\}$$

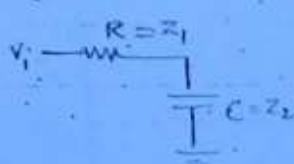
$$\therefore V_C = 6V$$

$$\therefore I_f = \frac{21 - V_C}{3k} = 5.8 \text{ mA} ; \quad V_0 = -\frac{5}{3} \times 6 = -10V ; \quad R_f = \frac{V_f}{I_f} = \frac{21}{5} = 4.2 \Omega$$

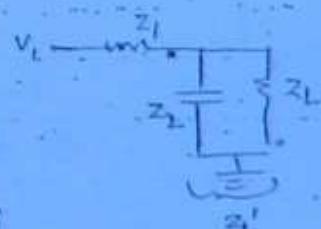
17<sup>th</sup> August, 2012

First order Butterworth Filter:

- Loading Effect -



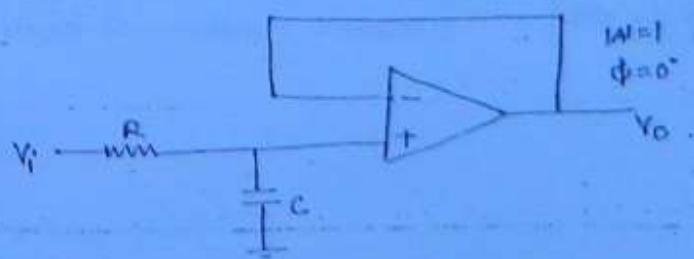
$$T = \frac{V_1}{z_1 + z_2}$$



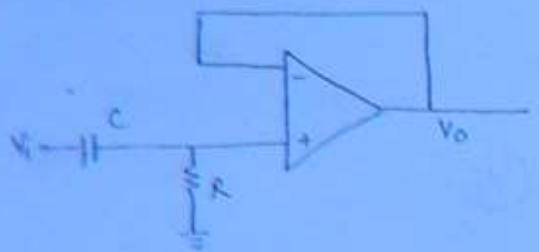
$$T' = \frac{V_1}{z_1 + z_L}$$

$$| T < T' |$$

→  $T'$  will keep on ↑ as  $z_L$  is ↑. Hence, there will be loading effect and parameters of filter will change. To avoid this, voltage follower circuit is used.



→ first order butterworth filter.  
LPF

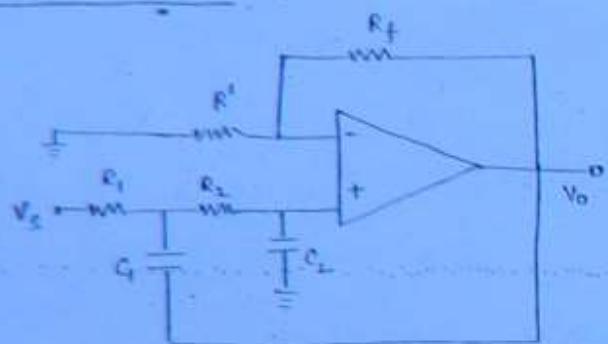


→ HPF - First order BWF.

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

(42)

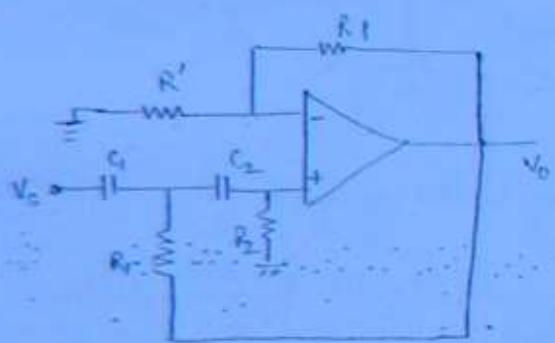
2<sup>nd</sup> order LP BWF -



$$f_c = \frac{1}{2\pi \sqrt{R_1 R_2 C_1 C_2}} = \frac{1}{2\pi R C}$$

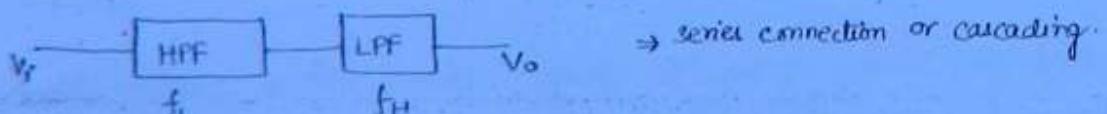
if  $R_1 = R_2 = R$   
&  $C_1 = C_2 = C$

2<sup>nd</sup> order HP BWF -



$f_c$  = same as above.

Band Pass filter :-



$f_H$  = high 3dB cutoff frequency [for LPF]

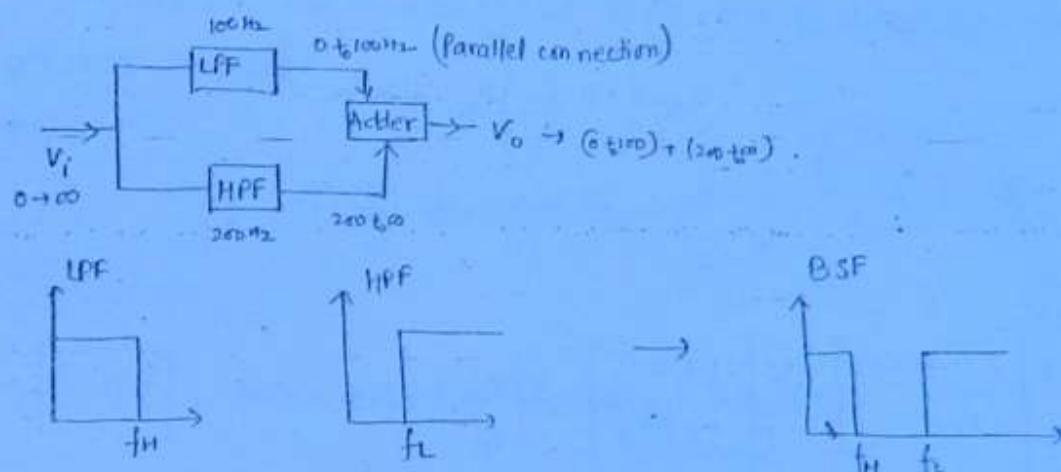
$f_L$  = low " " " " [for HPF]



→ for BPF,  $f_H > f_L$

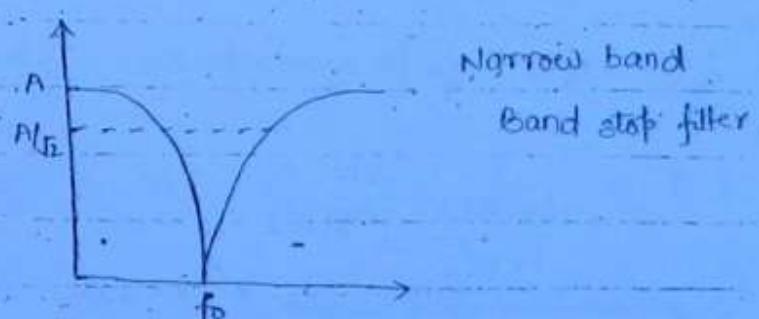
(43)

### Band Reject (or stop or Rejection) filter



→ for BSF,  $f_H < f_L$

### Notch filter :-



→ Used in communication systems to eliminate power supply noise.

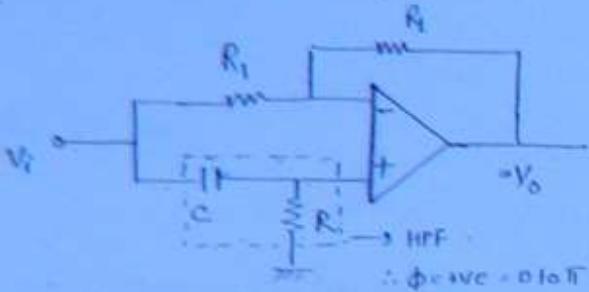
- Notch frequencies are 50Hz, 100Hz, 60Hz etc.

- Also used to remove harmonics

## All-Pass Filter

(44)

- It allows all input signal freq to pass w/o any amplification or attenuation.



Applying superposition -

for non-inverting terminal -

$$V_P = \frac{R}{R+L} \cdot V_i$$

for inverting terminal -

$$V_{D1} = \left(1 + \frac{R_1}{R_1}\right) V_P = 2V_P$$

$$V_{D2} = -\frac{R_1}{R_1} \cdot V_P = -V_P \quad \text{--- (2)}$$

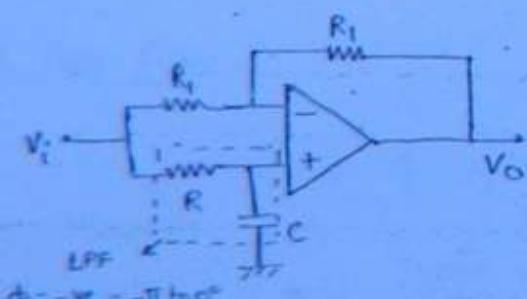
$$V_{D1} = -\frac{2RCs}{RCs+1} \cdot V_i \quad \text{--- (1)}$$

$$\therefore V_o = V_{D1} + V_{D2} = V_i \left[ \frac{2RCs}{1+RCs} - 1 \right] \Rightarrow \left[ \frac{-1+RCs}{1+RCs} \right] = \frac{V_o}{V_i}$$

$$\Rightarrow A = -\frac{1-RCs}{1+RCs}; |A|=1$$

$$\Rightarrow \boxed{\phi = 180 - 2\tan^{-1}(wRC)} \quad \begin{matrix} \text{xcept} \\ \rightarrow \end{matrix} \quad \begin{matrix} \text{for } \omega=0 & \phi = 180 \text{ or } \pi \\ \text{for } \omega=\infty & \phi = 0 \end{matrix}$$

$$\Rightarrow \text{Range of phase} \rightarrow \boxed{0 \leq \phi \leq 180^\circ}$$



$$A = \frac{1-RCs}{1+RCs}; |A|=1$$

$$\boxed{\phi = -2\tan^{-1}(wRC)} \quad \begin{matrix} \text{xcept} \\ \rightarrow \end{matrix}$$

$$\omega \rightarrow 0$$

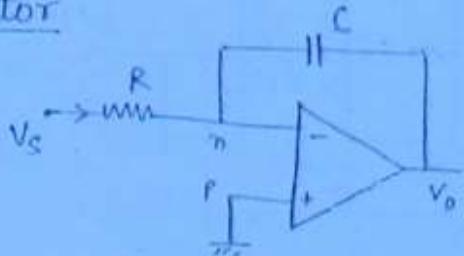
$$\phi = 0$$

$$\omega \rightarrow \infty$$

$$\phi = -180^\circ \text{ or } -\pi$$

$$\begin{matrix} ? \\ \rightarrow \end{matrix} \boxed{\text{Range of } \phi = -\pi \text{ to } 0^\circ}$$

## Integrator



$$C \frac{d(V_n - V_o)}{dt} = \frac{V_o - V_n}{R}$$

$$\Rightarrow \frac{V_S}{R} = -C \frac{dV_o}{dt}$$

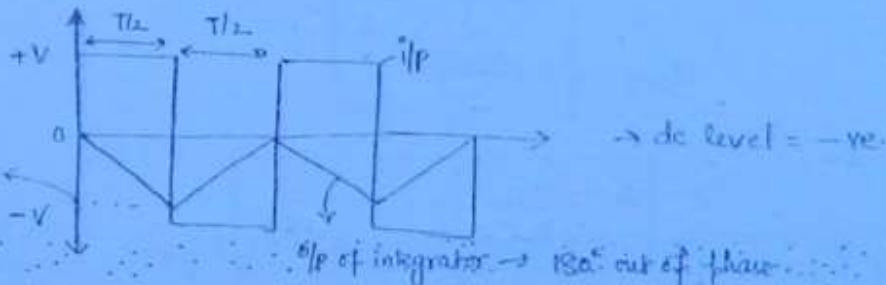
(45)

- \* Linear charging of capacitor is possible when we are providing constant current in the circuit and this can be achieved through current mirror circuit.

$$\Rightarrow \frac{dV_o}{dt} = -\frac{V_S}{RC} \Rightarrow V_o = -\frac{1}{RC} \int_0^t V_S dt + V_o(0^+)$$

Initial value

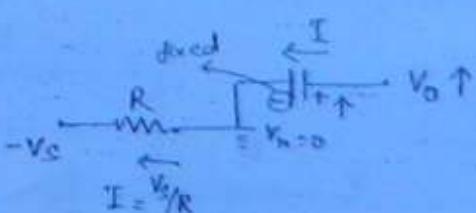
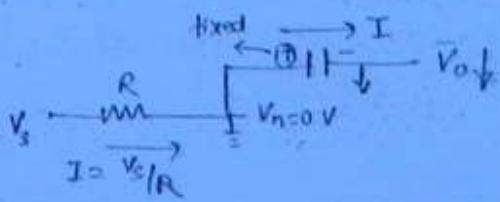
→  $\phi = 180^\circ$ , hence called as **Inverting Integrator**

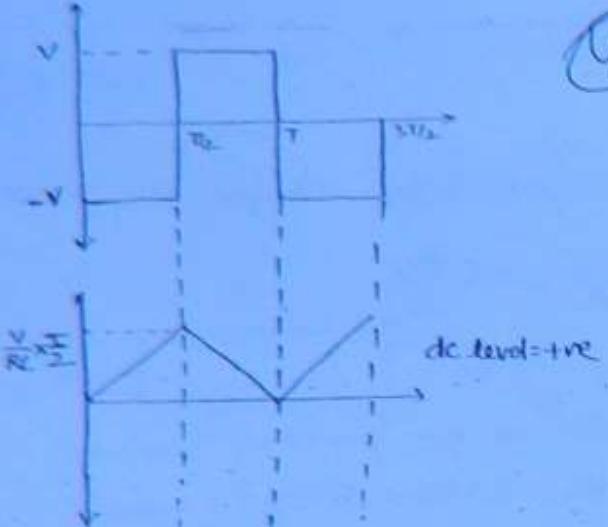


$$\frac{dV_o}{dt} = \text{rate of change of o/p} = \text{slope} = -\frac{V_S}{RC}$$

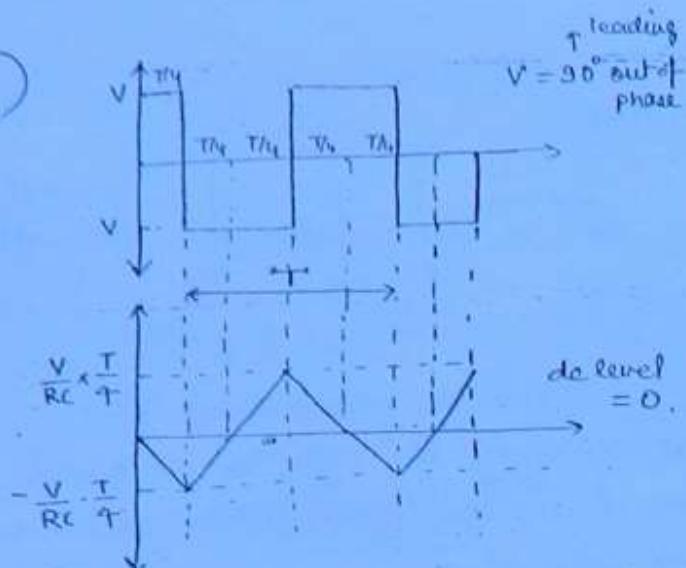
$$\text{for } V_S = +V \Rightarrow \frac{dV_o}{dt} = -\frac{V}{RC} \Rightarrow V_o \downarrow$$

$$\text{for } V_S = -V \Rightarrow \frac{dV_o}{dt} = \frac{V}{RC} \Rightarrow V_o \uparrow$$





Q6

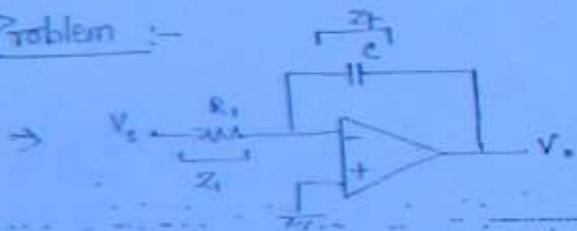


- Time period of output = time period of input in all the three cases.
- only change is in the dc levels of off.

\* → Swing =  $V_{max} - V_{min} = \frac{V}{RC} \times \frac{T}{2}$  is same in all the three cases.

$$\text{Swing} = \frac{V}{2RCf}$$

Problem :-



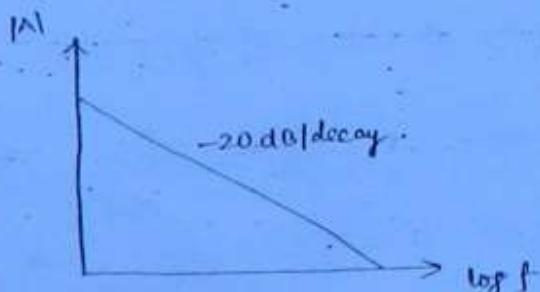
$$\text{Gain} \rightarrow A = -\frac{Z_f}{Z_1}$$

$$A = -\frac{1}{R_f C S} = -\frac{1}{j \omega R_f C}$$

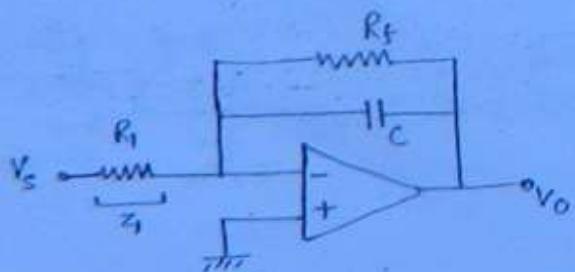
$$\rightarrow |A| = \frac{1}{\omega R_f C}$$

→ At  $\omega=0$ ,  $|A|=0$ .

→ Gain is not stable for entire freq range, hence no freq. stability. This is called Roll off problem.



Practical Integrator :-



$$Z_f = Z_C \parallel R_f = \frac{R_f \cdot Y_{CS}}{R_f + Y_{CS}} = \frac{R_f}{1 + R_f C_s}$$

(42)

$$Z_1 = R_1 ; \quad \therefore \text{Gain, } A = -\frac{I_f}{Z_1}$$

$$\Rightarrow A = -\frac{R_f / R_1}{1 + R_f C_s}$$

$$\Rightarrow A = \frac{-\left(R_f / R_1\right)}{1 + j\omega R_f C_s}$$

$$\rightarrow |A| = \frac{R_f / R_1}{\sqrt{1 + \omega^2 R_f^2 C^2}}$$

$$\rightarrow \text{At } \omega=0, \quad |A|_{\max} = \frac{R_f}{R_1}.$$

$$\rightarrow \text{3dB cutoff freq, } \omega_a = 2\pi f_a = \frac{1}{R_f C}$$

$$\Rightarrow f_a = \frac{1}{2\pi R_f C}$$

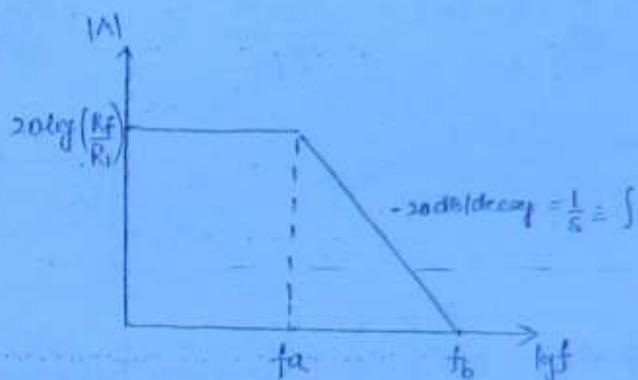
$$\rightarrow \text{At } \omega=\omega_b, \quad |A| = 1$$

$$\Rightarrow \frac{R_f^2}{R_1^2} = 1 + \omega_b^2 R_f^2 C^2 \Rightarrow \omega_b^2 R_f^2 C^2 \approx \frac{R_f^2}{R_1^2} \quad \left. \begin{array}{l} \text{Neglecting } 1 \\ \end{array} \right\}$$

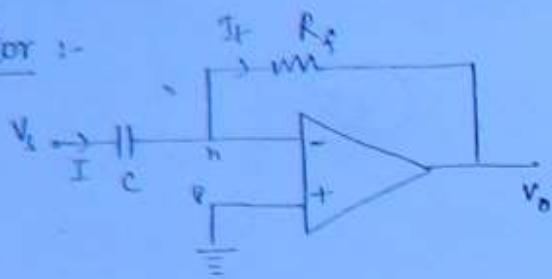
$$\Rightarrow \boxed{\omega_b = \frac{1}{R_1 C}} \quad \text{or} \quad \boxed{f_b = \frac{1}{2\pi R_1 C}} \quad \text{using} \quad \left. \begin{array}{l} \text{?} \\ f_b = \text{break freq.} \\ \text{at } |A|=1 \text{ at } ? \\ \text{this freq.} \end{array} \right\}$$

$\rightarrow$  Circuit acts as integrator between  $f_a$  and  $f_b$ .

$$\therefore f_a < f_b \Rightarrow \frac{1}{2\pi R_f C} < \frac{1}{2\pi R_1 C} \Rightarrow R_1 < R_f$$



Differentiator :-



$$V_P = V_N = 0$$

(48)

$$I = I_f$$

$$C \frac{d}{dt} (V_S - V_N) = \frac{V_N - V_O}{R_f} \Rightarrow \frac{CdV_S}{dt} = -\frac{V_O}{R_f}$$

$$\Rightarrow V_O = -R_f C \cdot \frac{dV_S}{dt}$$

$\rightarrow$   $\boxed{\phi \text{ shift} = 180^\circ}$   $\Rightarrow$  Inverting Differentiator

$$\rightarrow A = -\frac{R_f}{BZ_f} \Rightarrow A = -R_f C s = -j\omega R_f C$$

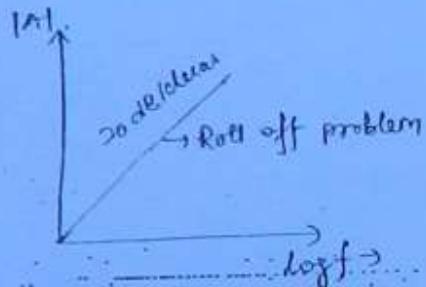
$$\rightarrow |A| = \omega R_f C$$

Problem -

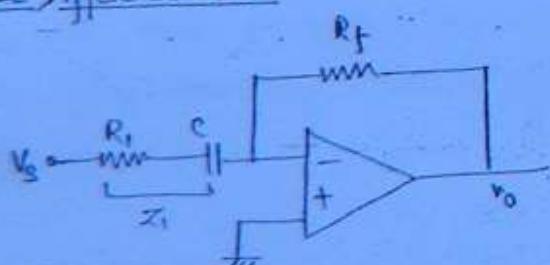
$$\rightarrow \text{At } \omega = \infty, |A| = \infty$$

$\rightarrow$  Frequency stability is less.

$\rightarrow$  Roll off problem.



Practical Differentiator :-



$$Z_f = R_f$$

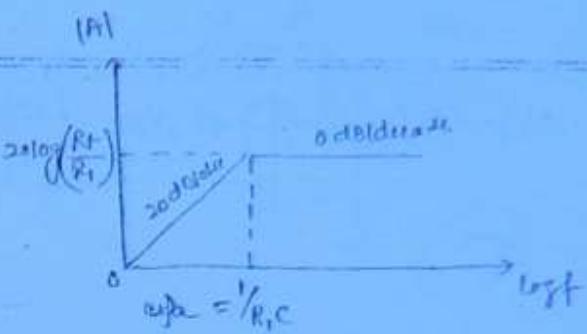
$$Z_1 = R_1 + j\omega C = \frac{R_1 C s + 1}{C s}$$

$$\therefore A = \frac{-R_f}{R_1 + j\omega C} = \frac{-R_f / R_1}{1 + \frac{j\omega C}{R_1}} = \frac{-R_f / R_1}{1 - \frac{j}{\omega R_1 C}}$$

$$\Rightarrow A = \frac{-R_f / R_1}{1 - \frac{j}{\omega R_1 C}}$$

$$\rightarrow |A| = \frac{R_f R_1}{1 + \frac{1}{\omega^2 R_1^2 C^2}}$$

(49)



$$\rightarrow \text{As } \omega \rightarrow \infty, |A|_{\text{max}} = \frac{R_f}{R_1}.$$

$\rightarrow$  Circuit is stable at high frequency.

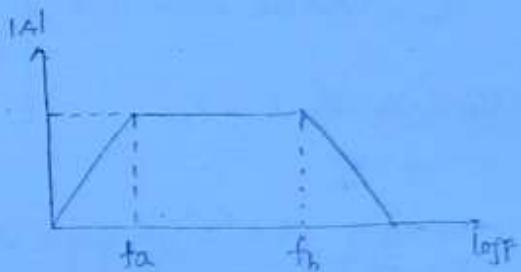
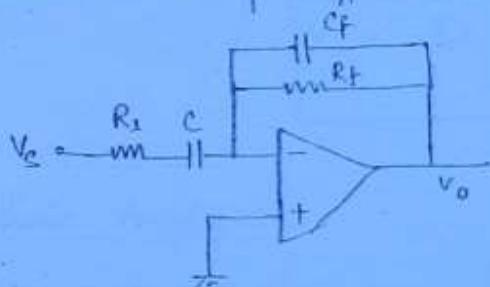
$$\rightarrow A = -\frac{sCR_f}{1+sCR_1} \rightarrow \text{Plotting Bode plot.}$$

$$f_a = \frac{1}{2\pi R_1 C}$$

$\rightarrow$  Circuit will act as differentiator if  $\omega \ll f_a$

$\rightarrow$  But,  $BW = \omega_a$ .

To limit the BW of Differentiator -

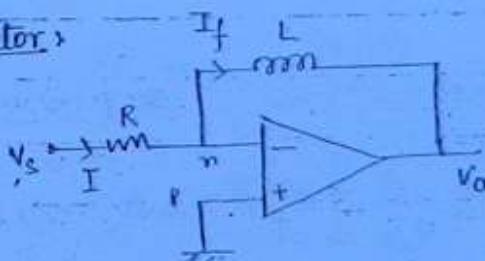


$\rightarrow C_f$  will limit the bandwidth of differentiator.

$\rightarrow$  BPF is also called as Practical differentiator.

$\rightarrow R_1$  is added to increase the frequency stability of the output.

Differentiator



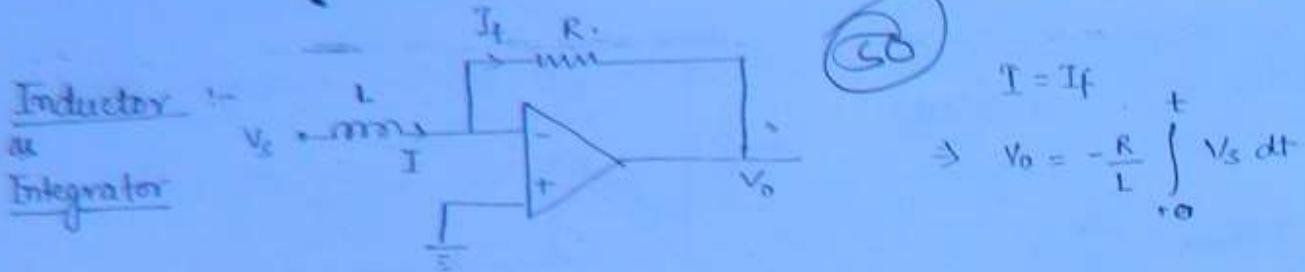
$$V_p = V_n = 0$$

$$I = I_f$$

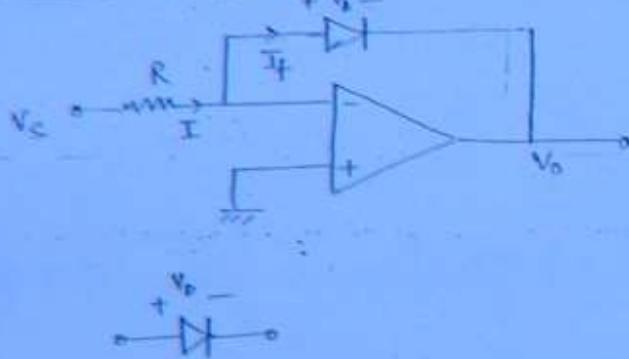
$$\frac{V_s}{R} = -\frac{1}{L} \int V_o dt$$

$$\Rightarrow \frac{-L}{R} \frac{dV_s}{dt} = V_o$$

$\rightarrow$  Bulky and Heavy due to L.



### Logarithmic Amplifier



$$V_P = V_N = 0$$

$$I = I_F = I_B$$

$$I_B = I_0 \left[ e^{\frac{V_D}{\eta V_T}} - 1 \right]$$

→



$I_0$  = reverse saturation current.

$$V_D = V_N - V_o = -V_o$$

$$I = I_B$$

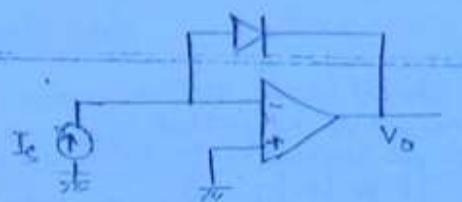
$$\Rightarrow \frac{V_s}{R} = I_0 \left[ e^{-\frac{V_o}{\eta V_T}} - 1 \right]$$

$$\Rightarrow \frac{V_s}{I_0 R} + 1 = e^{-\frac{V_o}{\eta V_T}} \Rightarrow \frac{V_o}{\eta V_T} = -\ln \left[ \frac{V_s}{I_0 R} + 1 \right]$$

$$\Rightarrow V_o = -\eta V_T \ln \left[ \frac{V_s}{I_0 R} + 1 \right]$$

⇒  $I_0$  is very small  $\Rightarrow \frac{V_s}{I_0 R} \gg 1$ .

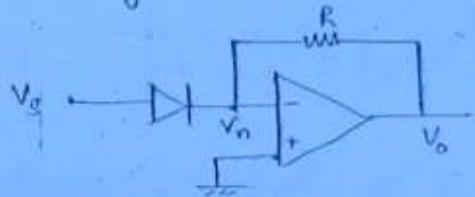
$$\therefore V_o = -\eta V_T \ln \left[ \frac{V_s}{I_0 R} \right] \xrightarrow{\text{neglect } 1}$$



$$V_0 = -\eta V_T \ln \left( \frac{I_c}{I_b} \right) \quad \left\{ \frac{V_0}{R} = \frac{I_c}{R} \right\}$$

(57)

### Anti-Logarithmic Amplifier



$$V_D = V_S - V_n = V_S \quad ; \quad I = I_B = I_D$$

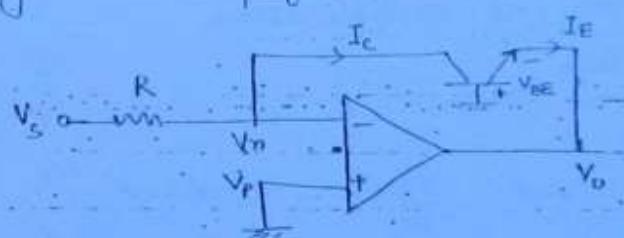
$$I_D \cdot \left[ e^{\frac{V_D}{\eta V_T}} - 1 \right] = \frac{V_n - V_D}{R}$$

$\Rightarrow \because V_S > 0$  and if  $e^{\frac{V_D}{\eta V_T}} \gg 1$

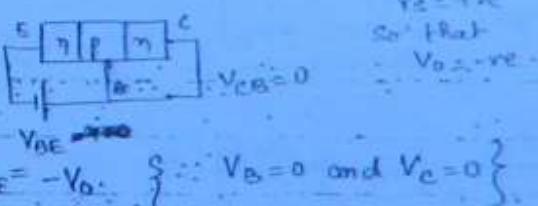
$$\Rightarrow V_0 = -I_D R e^{\frac{V_D}{\eta V_T}}$$

$$\Rightarrow \boxed{V_0 = -I_D R \text{ antilog} \left( \frac{V_S}{\eta V_T} \right)}$$

### Logarithmic Amplifier :-



Tr. should be in active region, ie,  
 $V_E < V_C$



$$\Rightarrow V_{BE} = -V_0 \quad ; \quad \left\{ V_B = 0 \text{ and } V_E = 0 \right\}$$

$$I_D = I_C \approx I_E = I_{C0} \left[ e^{\frac{V_{BE}}{\eta V_T}} - 1 \right] = \frac{V_S}{R}$$

$$\Rightarrow \frac{V_S}{I_{C0} R} + 1 = e^{\frac{V_{BE}}{\eta V_T}}$$

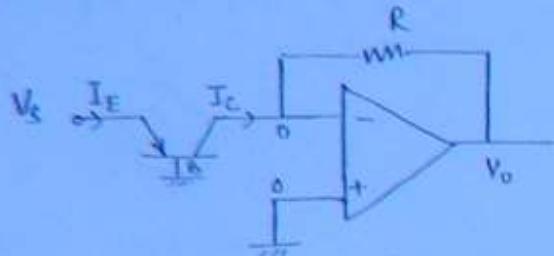
$$\Rightarrow V_0 = -V_T \ln \left[ \frac{V_S}{I_{C0} R} + 1 \right] \quad \text{for } \eta=1$$

$$\Rightarrow \boxed{V_0 = -V_T \ln \left[ \frac{V_S}{I_{C0} R} \right]}$$

## Anti-logarithmic Amplifier :-

(S2)

Tr. should be in active region,  
hence  $V_B = +ve$ , so that  $V_{EB} = +ve$ .



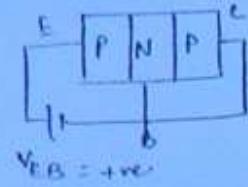
$$V_{EB} = V_s$$

$$I_E = I_C = I_f$$

$$\Rightarrow I_{co} \left[ e^{\frac{V_s}{\eta N_T}} - 1 \right] = \frac{V_o}{R}$$

$$\Rightarrow V_o = -I_{co} R \left[ e^{\frac{V_s}{\eta N_T}} - 1 \right]$$

$$I_b = I_c \approx I_E = I_{co} \left[ e^{\frac{V_s}{\eta N_T}} - 1 \right]$$



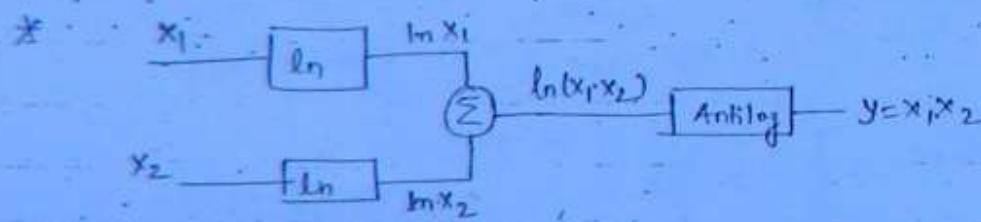
$$V_{CB} = 0, \rightarrow R_B$$

$$V_{EB} = +ve \rightarrow f.e.$$

$$V_o \cong -I_{co} R \text{ antilog} \left( \frac{V_s}{\eta N_T} \right)$$

## Applications

- Log and Antilog amplifiers are used in designing of multiplication, division, square root and squaring circuits.

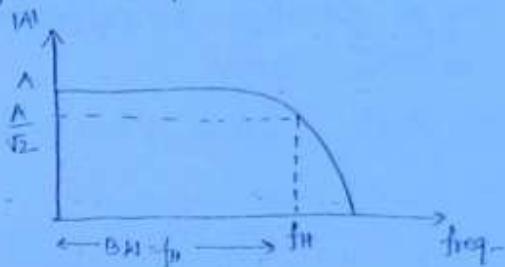


→ If the -ve sign is not set for op-amp, then pass o/p through a inverting amplifier with gain = 1.

20<sup>th</sup> August, 2012

(5)

### Frequency response of Practical op-amp -



$$|Z_d| = \frac{1}{2\pi f_c} ; f \uparrow, |Z_d| \downarrow$$

- Op-Amp is basically a dc amplifier. It can amplify a dc signal and also an ac signal in a wide band, extending from 0-1 MHz.

### Stair Rate (SR) -

- It is the time rate of change of closed loop amplifier op voltage under large signal condition. (Typical value =  $10^6$  V/μsec). Unit → V/μsec.

$$\rightarrow SR = \left. \frac{dV_o}{dt} \right|_{max} \Rightarrow SR = \frac{dV_o}{dV_i} \times \frac{dV_i}{dt}$$

$$SR = |A_{cl}| \times \left. \frac{dV_i}{dt} \right|_{max}$$

$$V_i = V_m \sin \omega_m t \quad \therefore \frac{dV_i}{dt} = V_m \omega_m \cos \omega_m t \quad \therefore SR = |A_{cl}| \times V_m \omega_m$$

$$\Rightarrow \left. \frac{dV_i}{dt} \right|_{max} = V_m \omega_m \quad \Rightarrow \quad \omega_m = 2\pi f_m = \frac{SR}{|A_{cl}| \cdot V_m} \text{ rad/sec}$$

$\omega_m$  or  $f_m$  → Max freq. of operation.

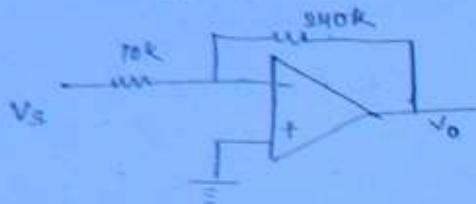
- for  $f \leq f_m \rightarrow$  op without distortion
- $f > f_m \rightarrow$  " with "

- Ques - for an operational amplifier having a SR of  $2V/\mu sec$ , for what is the max. closed loop voltage gain that can be used when if?

signal changes by 0.05V in 10 μsec?

(54)

Ans for the given circuit, determine the max freq. of operation in rad/sec  
that can be used by taking  $SR = 0.5V/\mu\text{sec}$  and  $V_m = 0.02V$ .



Soln (1)  $SR = 2V/\mu\text{sec}$   $\therefore SR = |A_{OL}| \times \frac{dV_i}{dt}$   
 $\frac{dV_i}{dt} = 0.05V/\mu\text{sec}$   $\Rightarrow 2 = |A_{OL}| \times 0.05$   
 $|A_{OL}| = 40$ .

Ans

(2)  $|A_{OL}| = \left| \frac{240}{10} \right| = +24$ .

$$\omega_m = \frac{240 \times 0.5 \times 10^6}{24 \times 0.02} = \frac{2.08 \text{ rad/sec} \times 10^6}{2} = 1.04 \text{ rad/sec.} \times 10^6$$

Note - If  $A_{OL}$  is not given, take  $A_{OL} = 1$ .

\* Slow rate is limited by internal capacitances of op-amp. hence it is not  $\infty$ .

Offset voltages and currents :-

→  $V_o = 0 \rightarrow$  Balanced Op-amp.

if  $V_o + 0 = V_{O0} \rightarrow$  output offset voltage.

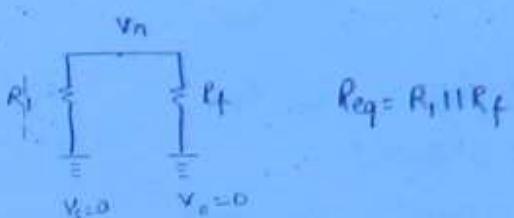
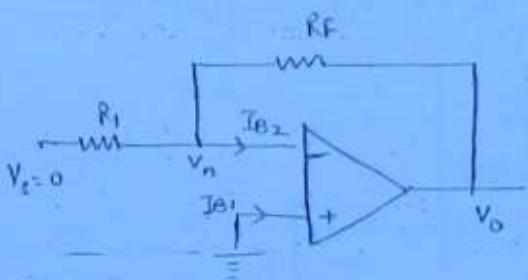
→  $V_o = 0 \rightarrow$  Now it is balanced.  
 $V_{o1}, V_{o2}$  - Biasing voltages.

→ Ifp bias current =  $I_B = \frac{I_{B1} + I_{B2}}{2}$ , when  $V_o = 0$ .

(SS)

→ Ifp offset current =  $I_{BO} = \frac{I_{B1} - I_{B2}}{2}$  when  $V_o = 0$ .

→ Ifp offset voltage =  $V_{IO} = I_{BO} \cdot R_i = V_{B1} - V_{B2}$  when  $V_o = 0$ .



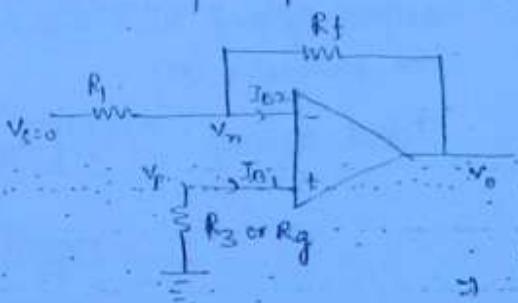
$$|V_n| = R_{eq} \cdot I_{B2} = I_{B2} [R_1 || R_f]$$

$$|V_p| = 0$$

∴ Amplifier is again unbalanced when  $V_s = 0$  (ie without signal) & feedback is connected.

→ Normally,  $I_{B1} \approx I_{B2}$ .

→ To balance the op-amp,  $R_3$  is connected to non-inverting terminal.



$$\text{Now, } |V_p| = I_{B1} \cdot R_3.$$

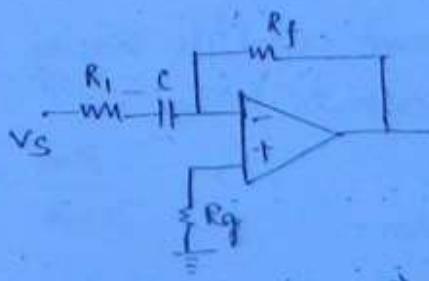
Hence, for balanced condition

$$|V_p| = |V_n|$$

$$\Rightarrow R_3 = [R_1 || R_f]$$

\* To minimise the effect of Ifp bias current, one should place in non-inverting terminal, a resistance equal to dc resistance seen by inverting terminal.

Ques What is  $R_g$ ?



$$\therefore R_g = [R_1 + z_c] || R_f.$$

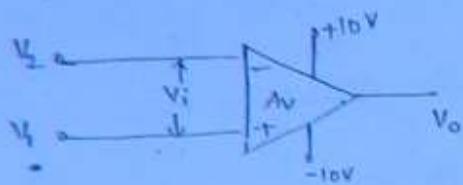
$$z_c = \frac{1}{2\pi f C} \Rightarrow z_c = \infty \text{ for DC.}$$

$$\Rightarrow R_g = [R_1 + \infty] || R_f = R_f. \text{ Ans}$$

## Transfer Characteristics of Op-Amp :-

(56)

### i) Practical Op-amp :-



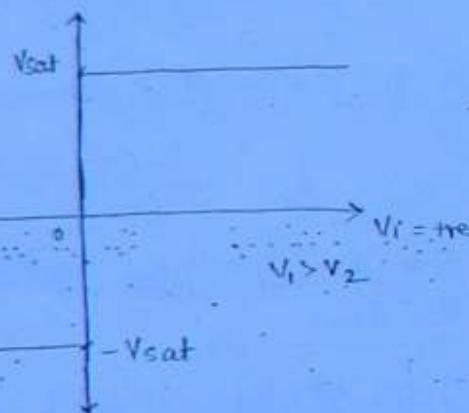
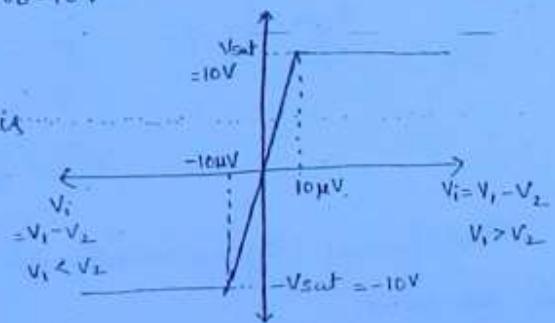
$$\text{Now, } (i) \quad V_1 = 10\mu\text{V}, \quad V_2 = 9\mu\text{V} \\ \Rightarrow V_o = 1\text{V}$$

$$(ii) \quad V_1 = 120\mu\text{V}, \quad V_2 = 100\mu\text{V} \\ \Rightarrow V_o = 20\text{V} \text{ but } > 10\mu\text{V} \rightarrow \text{Op-amp is} \\ \Rightarrow V_o = 20\mu\text{V}.$$

$$A_v = 10^6 \\ V_i = V_1 - V_2 \\ \Rightarrow V_o = 10^6(V_1 - V_2). \quad \text{--- (1)}$$

$$(i) \quad V_1 = 100\mu\text{V}, \quad V_2 = 90\mu\text{V} \\ \Rightarrow V_o = 10\text{V}$$

saturated.



Practical Op-amp with sufficient  
+ve feedback. (or)

Ideal Op-amp , i.e.,  $[A_v \approx \infty]$

→ As  $|A_v| \uparrow$ , for a very small i/p, output will shoot up to  $+V_{sat}$  or  $-V_{sat}$  depending on whether  $V_1 - V_2$  to be +ve or -ve.

→ In a practical op-amp, op voltage cannot exceed its biasing voltage, i.e., range of op voltage is from  $-V_{sat}$  to  $+V_{sat}$ .

→ Op-amp can enter into saturation when-

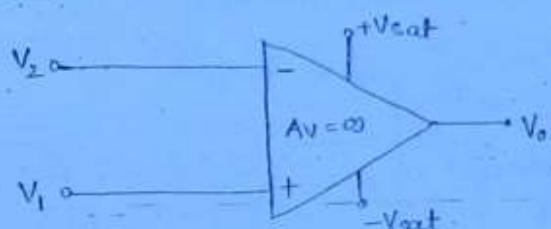
- (a) Large i/p signals are applied. ( $>$  than few  $\mu\text{V}$ )

- (i) When sufficient +ve feedback is provided.

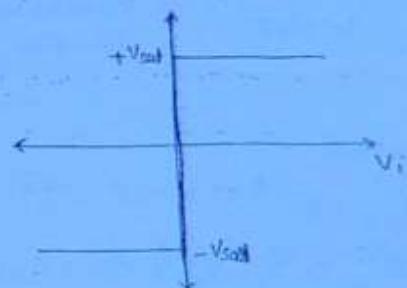
(S1)

### Comparator :-

#### Ideal comparator :-



	V0	
if $V_1 > V_2$	+Vsat	1
$\Rightarrow V_1 < V_2$	-Vsat	0

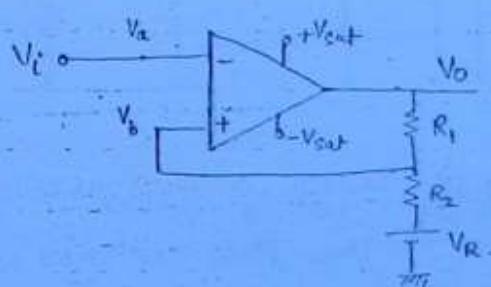


Characteristic

→ It is operated under open loop condition.

#### Practical Comparator :-

#### Schmitt Trigger :-



$|V_R| \rightarrow$  Reference voltage  $< |V_{sat}|$ .

	V0	
$V_b > V_a$	+Vsat	1
$V_b < V_a$	-Vsat	0

$$\Rightarrow V_b = \frac{V_o R_2}{R_1 + R_2} + \frac{V_R R_1}{R_1 + R_2}$$

(58)

$$\textcircled{1} \quad V_o = +V_{sat}$$

$$V_{b1} = \frac{V_{sat} \cdot R_2}{R_1 + R_2} + \frac{V_R \cdot R_1}{R_1 + R_2} = V_{um} \rightarrow \text{Upper threshold}$$

$$\textcircled{2} \quad V_o = -V_{sat}$$

$$V_{b2} = -\frac{V_{sat} \cdot R_2}{R_1 + R_2} + \frac{V_R \cdot R_1}{R_1 + R_2} = V_{LTH} \rightarrow \text{Lower threshold}$$

**Assumption:** let  $V_{um} = V_b = 6V$  These are set before applying  $V_i$ , on the basis of  $\pm V_{sat}$  &  $V_R$ .  
 $V_{LTH} = V_{b2} = 3V$ .  
 $V_m = 10V$ .

\textcircled{3} At  $t=0$ ,

$$\text{let } V_o = +V_{sat}$$

$$\therefore V_b = V_{b1} = 6V$$

$$V_i = V_a = 0 \Rightarrow V_b > V_a \Rightarrow V_o = +V_{sat}$$

Hence, our assumptions are right.

\textcircled{4} Let  $V_o = -V_{sat}$

$$V_b = V_{b2} = 3V$$

$$V_i = V_a = 0 \Rightarrow V_b > V_a \Rightarrow V_o = +V_{sat}$$

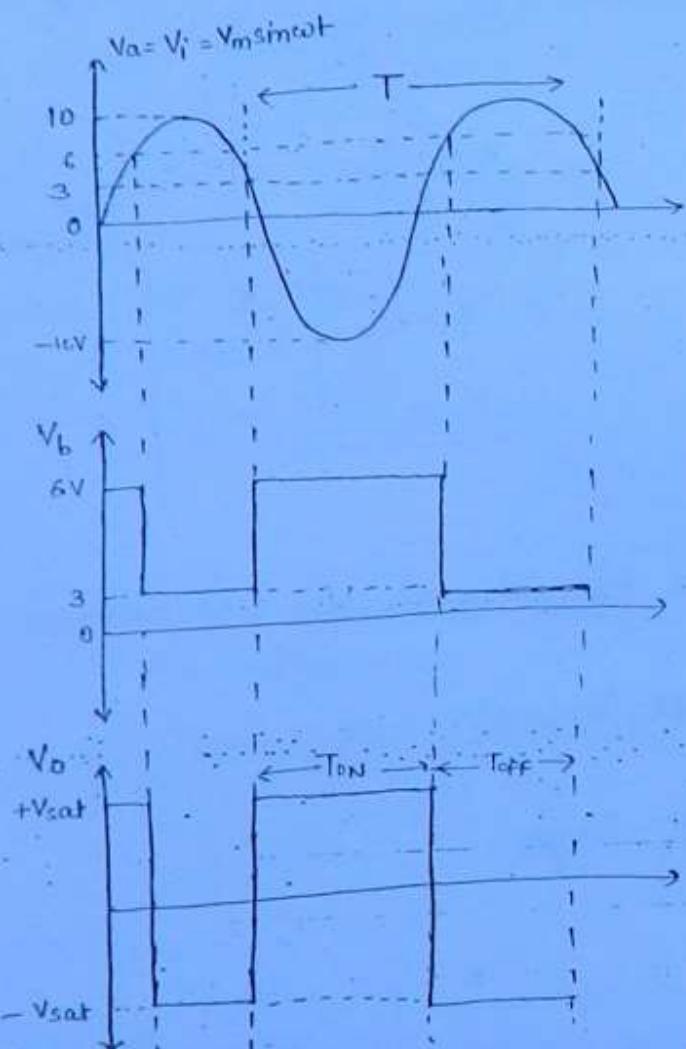
Hence our assumption was wrong.

$$V_o = +V_{sat}$$

\textcircled{5} If  $V_a \uparrow$  & when  $V_a \geq V_{b1} = 6V$ ,

then  $V_o$  switches from  $+V_{sat}$  to  $-V_{sat}$

and  $V_b$  " "  $+6V \rightarrow +3V$ .



Thus these two steps will be repeated.

\textcircled{6} If  $V_a \downarrow$  & when  $V_a \leq V_{b2} = 3V$ ,

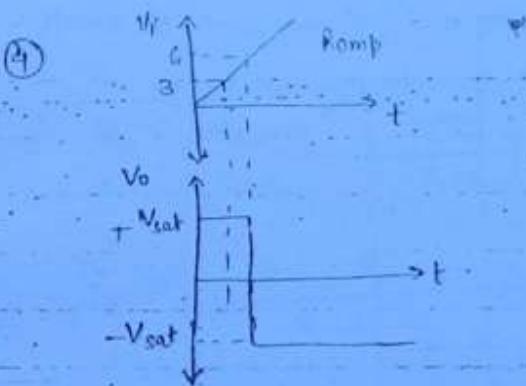
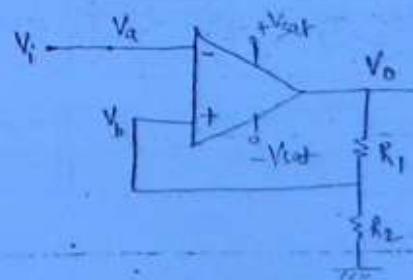
then  $V_o$  switches from  $-V_{sat}$  to  $+V_{sat}$ .

and  $V_b$  " "  $+3V \rightarrow +6V$ .

(59)

\* Necessary condition -

- (a)  $V_i$  should  $\uparrow$  and cross  $V_{TH}$ ; so that  $V_o$  switches from  $+V_{sat}$  to  $-V_{sat}$   
 (b)  $V_i$  should  $\downarrow$  and cross  $V_{TH}$ ; " " " "  $-V_o$  " "  $-V_{sat}$  "  $+V_{sat}$ .

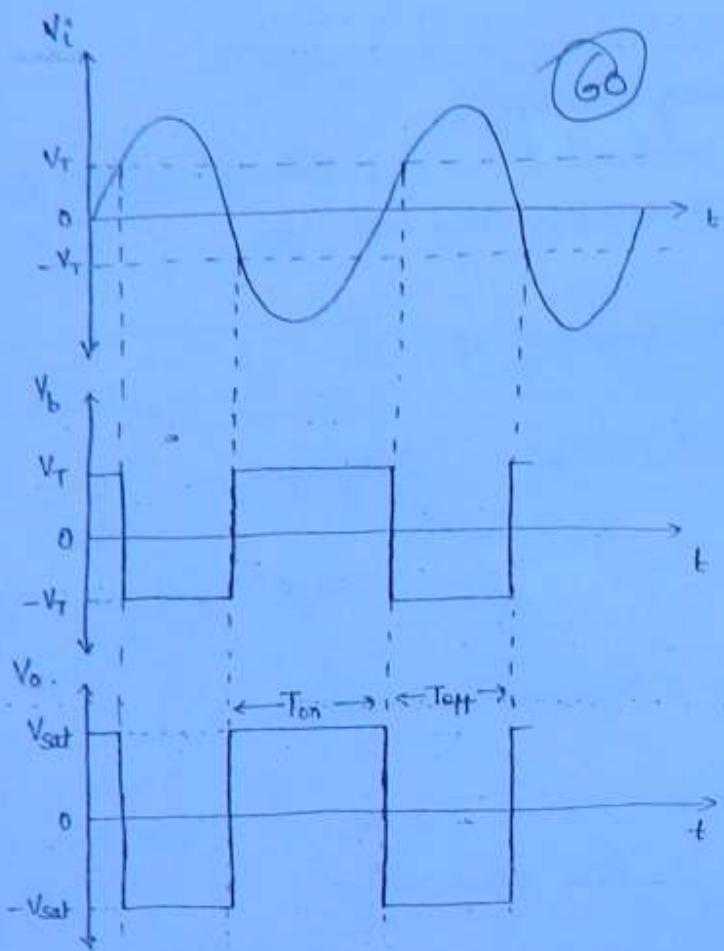
\* Time period of op  $\Rightarrow T_0 = T_{ON} + T_{OFF} = T$  time period of op. $\rightarrow \because T_{ON} > T_{OFF}; \text{ Duty cycle} = \frac{T_{ON}}{T_{ON} + T_{OFF}} \times 100\%$  $\Rightarrow D > 50\%$ ,  $\Rightarrow$  Asymmetrical square wave $\rightarrow$  It is a square wave converter.Eg - ①  $V_i = 25\sin\omega t$ . &  $V_{b1} = 6V, V_{b2} = 3V$  ②  $V_i = 4 \text{ to } 5V$  $\therefore V_o = +V_{sat}$  always. $V_o$  depends on initial condition  
if  $+V_{sat}$  then will remain  $+V_{sat}$   
 $" -V_{sat} " " -V_{sat}$ ③  $V_i = > 6 \text{ volts.}$  $\therefore V_o = -V_{sat}$  always.Duty cycle :- for  $D = 50\%, V_R = 0$ .

$$V_{Op} = \frac{+V_{sat} \cdot R_2}{R_1 + R_2} = +V_T$$

$$V_{O2} = \frac{-V_{sat} \cdot R_2}{R_1 + R_2} = -V_T$$

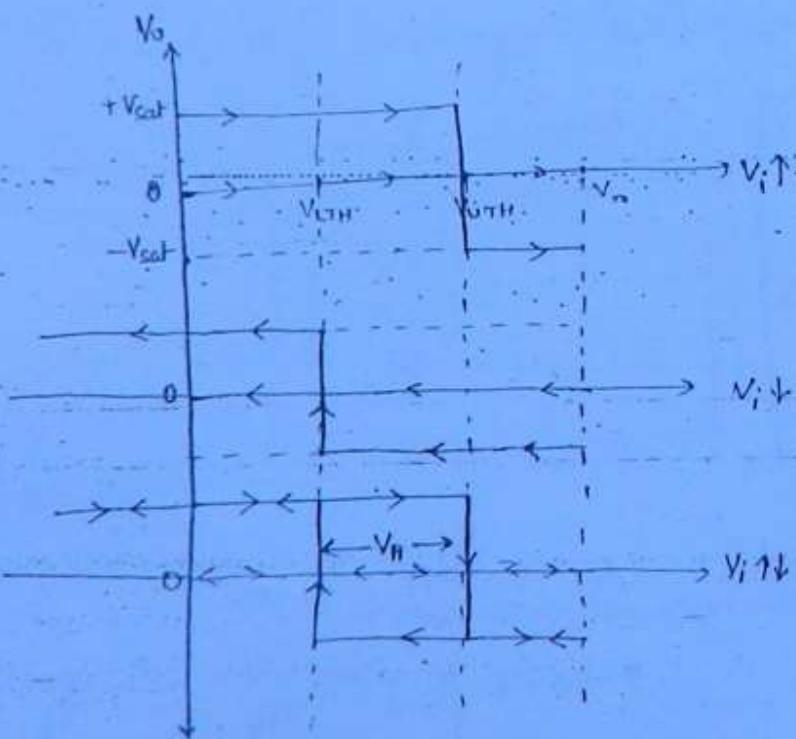
→ Time period of  $v_{op}$  = same as time period of  $v_{ip}$  and hence by changing  $V_R$  we cannot change the frequency of output, we can only change the duty cycle, in turn, the avg. dc level (the area) of  $v_{op}$  will change.

$V_R = +ve$	$D > 50\%$
$V_R = 0$	$D = 50\%$
$V_R = -ve$	$D < 50\%$



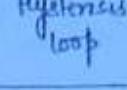
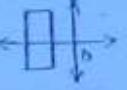
### Hysteresis Loop :-

- i) for Asymmetrical Wave →  
 $V_H$  = Hysteresis voltage  
 $= V_{UH} - V_{LH}$ .

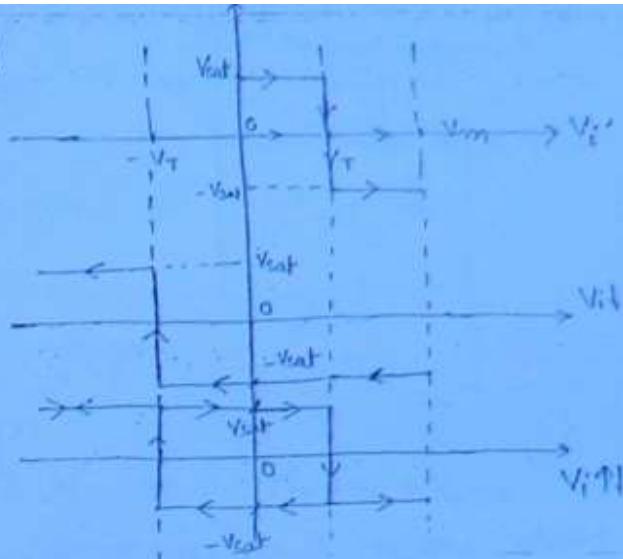


② For symmetrical wave—

$$V_H = 2V_T$$

$V_R$	Duty cycle	Avg. DC level	Hysteresis loop
$V_R > 0$	$D > 50\%$	+ve	
$V_R = 0$	$D = 50\%$	= 0	
$V_R < 0$	$D < 50\%$	<0 or -ve	

(6)



→ This table is valid only for the circuit discussed earlier.

$$\rightarrow V_R = \frac{R_2}{R_1+R_2} \cdot V_{sat} = \alpha \cdot V_{sat}; \quad \alpha = \frac{R_2}{R_1+R_2} < 1.$$

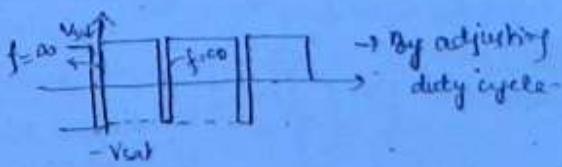
Hence, on non-inverting terminal, we are getting attenuated square wave.

$$\rightarrow V_H = \left( \frac{V_{sat} \cdot R_2}{R_1+R_2} + \frac{V_R \cdot R_1}{R_1+R_2} \right) - \left( \frac{-V_{sat} \cdot R_2}{R_1+R_2} + \frac{V_R \cdot R_1}{R_1+R_2} \right)$$

$$\Rightarrow V_H = \frac{2V_{sat} \cdot R_2}{R_1+R_2}$$

Hysteresis voltage is independent of  $V_R$ ; only the position of loop will change.

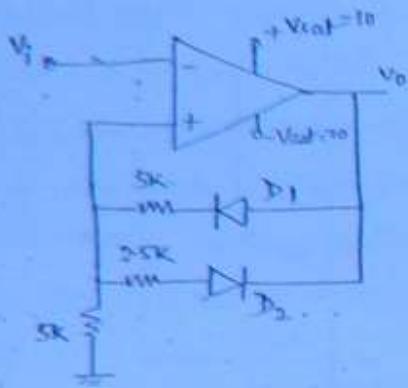
→ A slow moving waveform (sin $\omega t$ ) can be converted into a fast moving waveform (square wave) by using schmitt trigger.



→ By adjusting duty cycle

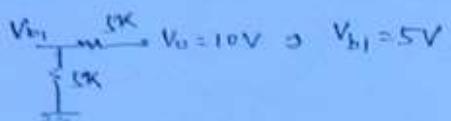
→ Slew rate should be high, so that the triggering pulse reaches  $+V_{sat}$  or  $-V_{sat}$  very fast.

Hence, SR ↑ for triggering op-amp..



Consider schmitt-trigger

Sol<sup>n</sup> For  $+V_{sat} = 10V$ ,  $\Rightarrow D_1 = ON, D_2 = OFF$

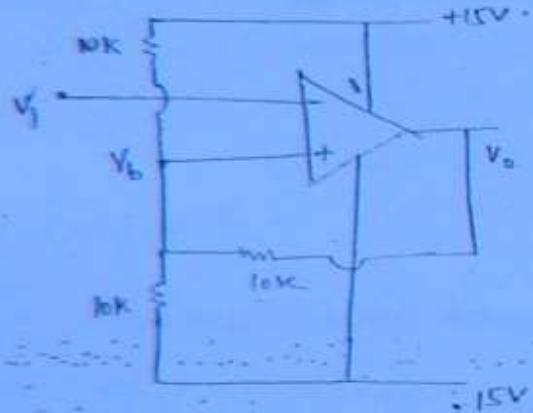


for  $V_{sat} = -10V$ ,  $\Rightarrow D_1 = OFF, D_2 = ON$

$$V_{oL} \xrightarrow[5K]{2.5K} -10V \therefore V_{b2} = -\frac{10 \times 5}{7.5} = -6.67V$$

: hysteresis loop :-

Ques Consider schmitt trigger ckt. A square wave which goes from  $-15V$  to  $+15V$  is applied to inverting i/p of op-amp. Assume that off swings from  $+15$  to  $-15V$ . The voltage at non-inverting i/p switches b/w.



Sol<sup>n</sup> KCL at  $V_b$  -

$$V_b \left[ \frac{1}{10} + \frac{1}{10} + \frac{1}{10} \right] - \frac{15}{10} + \frac{15}{10} = \frac{V_o}{10} \approx 0$$

$$\Rightarrow 3V_b = V_o$$

$$\Rightarrow V_b = V_o/3$$

$$V_b \approx +5V \text{ to } -5V$$

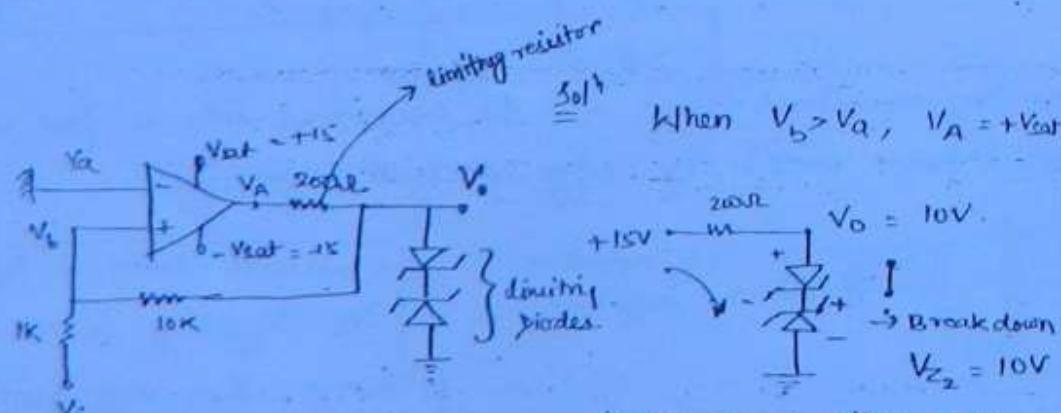
Find range -

- (a)  $-12$  to  $12V$  (b)  $-5$  to  $+5V$
- (c)  $-7.5$  to  $7.5V$  (d)  $0$  to  $5V$

21<sup>st</sup> August, 2012

workbook Pg. 62

Ques



Sol<sup>n</sup>

When  $V_b > V_A$ ,  $V_A = +V_{sat} = 15V$

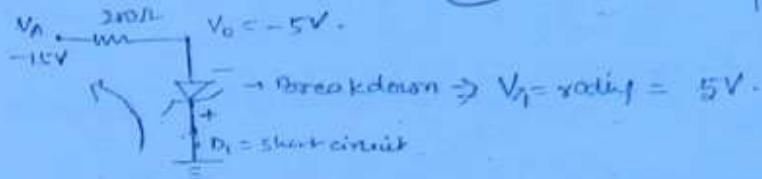
$$+15V \xrightarrow{200\Omega} V_o = 10V$$

$\therefore$  Breakdown  
 $V_{Z2} = 10V$

Hence, zener voltage rating  
 $= 10V$ .

When  $V_b < V_a$ ,  $V_o = -V_{sat} = -15V$ .

(63)



For practical diodes  
 $V_{ac} = 0.7V$   
 $-V_{ZT} = 4.3V$

→ 2kΩ is connected to dissipate the extra voltage.

Applying KCL at  $V_b$  -

$$V_b - V_i + \frac{V_b}{10} - \frac{V_o}{10} = 0$$

$$\Rightarrow V_b = \frac{V_o + 10V_i}{11}$$

$V_b = 0 \Rightarrow$  the o/p will switch from +10V to -5V.

$$\therefore V_b = \frac{10V_i + 10}{11} = 0 \Rightarrow V_i = -1V = V_{TH}$$

$$\text{At } t=0, V_i=0, V_b = \frac{V_o}{11}, \text{ but } V_o = -5V \Rightarrow V_b = -5/11$$

case I

when  $V_o = +10V \Rightarrow V_b > V_a \Rightarrow V_b > 0$

$$V_b = \frac{V_i \cdot 10 + 10}{11}$$

As  $V_i \uparrow$ ,  $V_b$  will also increase.

and when  $V_b > V_a \Rightarrow V_b > 0$

Case II

$V_o = -5V \Rightarrow V_b < V_a \Rightarrow V_b < 0$

$$\Rightarrow V_b = \frac{10V_i - 5}{11}$$

As  $V_i \uparrow$ ,  $V_b$  will also ↑.

and when  $V_b = 0$ ,  $V_o$  will switch

from -5V to +10V  $\Rightarrow 10V_i - 5 = 0$

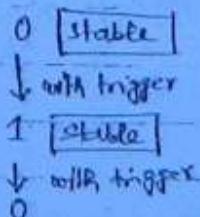
$$\Rightarrow V_i = 0.5V = V_{TH}$$

### Multivibrator

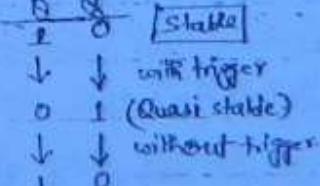
→ It's a device whose o/p vibrates b/w two levels, i.e., low level and high level (or 0 and 1).

→ These are of three types -

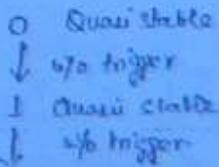
(i) Bistable



(ii) Monostable



(iii) Astable → free running Multivibrator



Eg - flip-flop, binary Schmitt trigger

One shot or Uni-vibrator

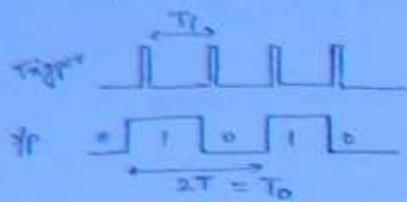
Square wave generator

with  $V_i$  not very high.

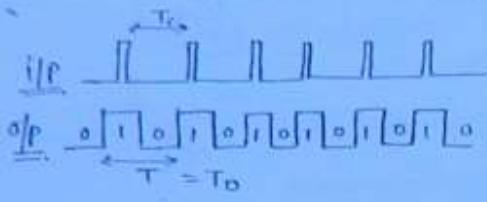
- \* when  $V_i \uparrow$  then  $V_o \downarrow$  and vice versa  $\rightarrow$  -ve feedback.
- $V_i \uparrow \Rightarrow V_o \uparrow$  and  $\downarrow \Rightarrow$  +ve feedback.

(64)

$\rightarrow$  Bistable :-



- Monostable :-



$\rightarrow$  Bistable and Monostable are simply converters and not square wave generators whereas for Astable Multivibrator, no need of an input / trigger to generate square wave.

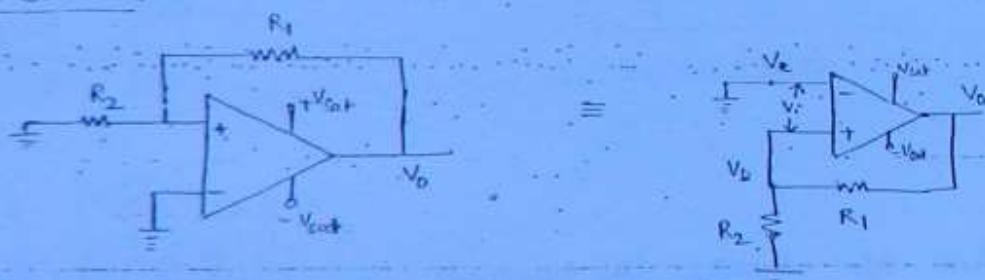
$\rightarrow$  Fastest off will be given by astable multivibrator as it is not limited by frequency of input trigger.

Note:

- Astable multivibrator is often used as fastest waveform generator.

### Multivibrator by using operational Amplifier :-

Bistable Multivibrator :-



$$V_b = \frac{R_2}{R_1 + R_2} V_{0+}$$

Initially,  $V_0 = 0$ ,  $V_b = 0$ ,  $V_a = 0$   
 $\Rightarrow V_i = 0 \Rightarrow V_b = 0$  (Ideally).

but because of noise:-

$$V_b \uparrow \Rightarrow V_i = V_b - V_a \uparrow \Rightarrow V_0 = \text{Av} V_i \uparrow$$

$\Rightarrow V_b$  again  $\uparrow$  due to  $V_0$  and it will keep on  $\uparrow$  till it reaches  $+V_{sat}$ . and then it will remain in one of the stable states.

→ Because of noise, op initially can be at  $+V_{sat}$  or  $-V_{sat}$  depending on initial noise effect.

(65)

Eg let  $R_1 = R_2$  and  $V_{sat} = 10V \Rightarrow V_b = 5V$ .

Now, to change the stable state of  $V_b$  from  $+V_{sat}$  to  $-V_{sat}$  -

(a) Positive trigger can be applied at  $V_a$  (err)

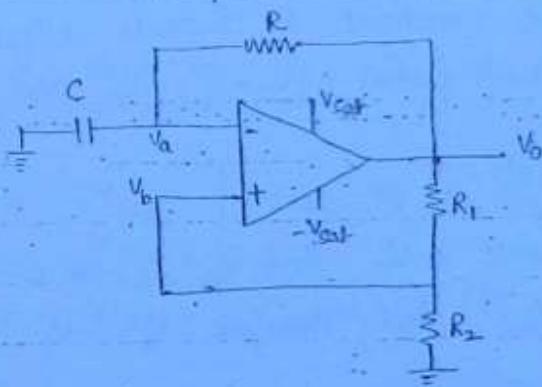
(b) Negative " " " " "  $V_b$  and trigger value should be more than the present  $V_b$  value: (ie, more than 5V)

i.e. if  $\boxed{-5.5V}$  at b or  $\boxed{+5V}$  at a and the circuit will

switch to its other stable state

- It has volatile memory i.e., memory is lost when power supply is interrupted.

### Astable Multivibrator / Square Wave Generator :-



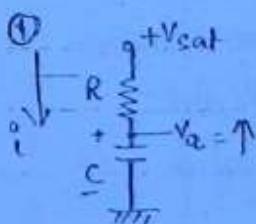
$$V_b = \frac{R_2}{R_1 + R_2} V_o \quad \text{--- (1)}$$

At  $t=0$ ,

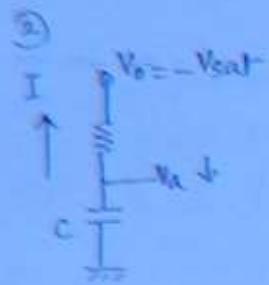
$$V_o = +V_{sat} \quad (\text{by noise}).$$

$$V_b = \frac{R_2 V_{sat}}{R_1 + R_2} = V_T$$

$$\therefore V_b > V_a \Rightarrow V_o = +V_{sat} \quad \Leftarrow \text{but } V_a = V_c = 0$$

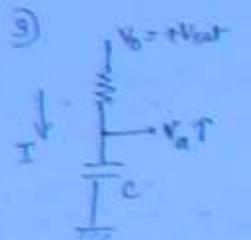


→ capacitor will start charging and the voltage of terminal  $V_a$  will start increasing. As soon as  $V_a$  reaches  $V_T$ ,  $V_o$  will switch to  $-V_{sat}$ .



$V_o = -V_{sat}$

→ capacitor will start discharging or start charging towards  $-V_{sat}$ .



$V_b = +V_{sat}$

→ capacitor will again start charging towards  $+V_{sat}$ .

→  $V_o \rightarrow$  square wave

$V_b \rightarrow$  attenuated square wave.

$V_a \rightarrow$  approximate triangular wave.

→ Swing of  $V_o = \tau V_{sat}$  to  $+V_{sat}$

" "  $V_a \& V_b = -V_T$  to  $+V_T$

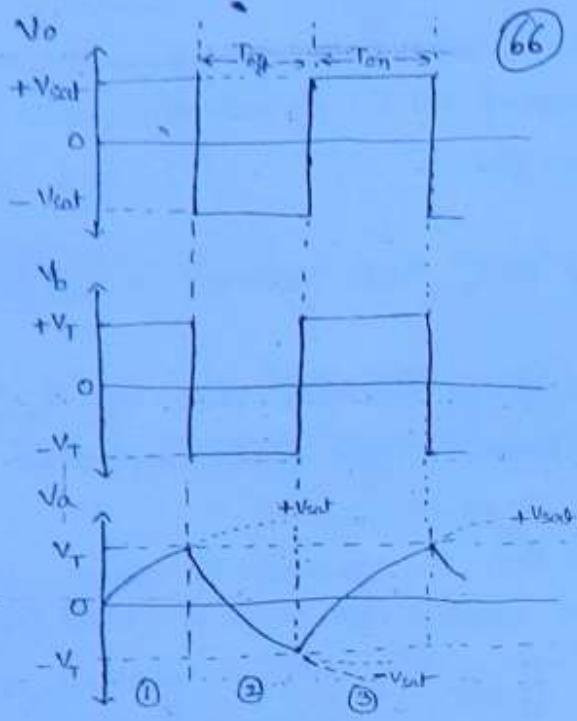
→ At initial condition, consider  $V_o = +V_{sat}$ ,  $V_b = +V_T$  and  $V_a = V_c = 0$ .

Now, capacitor will charge by time constant  $RC$  towards  $+V_{sat}$ .

Capacitor will charge upto  $+V_T$  at this point  $V_a = V_b = +V_T$  and op-amp comes out of saturation.

→ When the capacitor further charges above  $+V_T$  then  $V_a > V_b$  as a result of which  $V_o$  switch over to  $-V_{sat}$  and therefore  $V_b = -V_T$ .

→ Now, capacitor starts discharging from  $+V_T$  to  $-V_T$  towards  $-V_{sat}$  with time constant  $RC$ . Thus, when capacitor discharge upto  $-V_T$ , then  $V_a = V_b = -V_T$  and op-amp comes out of saturation. When the capacitor further discharges below  $-V_T$ , then  $V_a < V_b$ , as a result of which  $V_o$  will switch over to  $+V_{sat}$  and again  $V_b = +V_T$ . and thus, the cycle will repeat.



Derivation of  $T_{on}$ :

(67)

capacitor charges from  $-V_T$  to  $V_f$  in time  $T_{on}$ .

$$V_C = V_a = V_f - [V_f - V_i] e^{-t/RC}$$

$$V_i = -V_T \text{ at } t=0, \quad V_f = +V_{sat} \text{ at } t=\infty.$$

$$\therefore V_C \downarrow V_{sat} = [V_{sat} + V_T] e^{-t/RC} \rightarrow \text{charging eqn}$$

At  $t = T_{on}$ :

$$V_T = V_{sat} - [V_{sat} + V_T] e^{-T_{on}/RC}$$

$$\Rightarrow T_{on} = RC \ln \left[ \frac{V_{sat} + V_T}{V_{sat} - V_T} \right]$$

Derivation of  $T_{off}$ :

capacitor discharges from  $V_f$  to  $-V_T$  in time  $T_{off}$ .

$$V_i = V_f \text{ at } t=0, \quad V_f = -V_{sat} \text{ at } t=\infty.$$

$$\therefore V_C = -V_{sat} - [-V_{sat} - V_T] e^{-t/RC}$$

$$\Rightarrow V_C = -V_{sat} + [V_{sat} + V_T] e^{-t/RC} \rightarrow \text{Discharging eqn.}$$

At  $t = T_{off}$ ,  $V_C = -V_T$ ,

$$\therefore -V_T = -V_{sat} + [V_{sat} + V_T] e^{-T_{off}/RC}$$

$$\Rightarrow T_{off} = RC \ln \left[ \frac{V_{sat} + V_T}{V_{sat} - V_T} \right]$$

$\Rightarrow$

$$T_{off} = T_{on}$$

$$\Rightarrow \text{Duty cycle} = 50\%.$$

$\Rightarrow$  Square wave generator.

→ Astable multivibrator does not have any stable states due to continuous charging & discharging of C.

→ Time period,

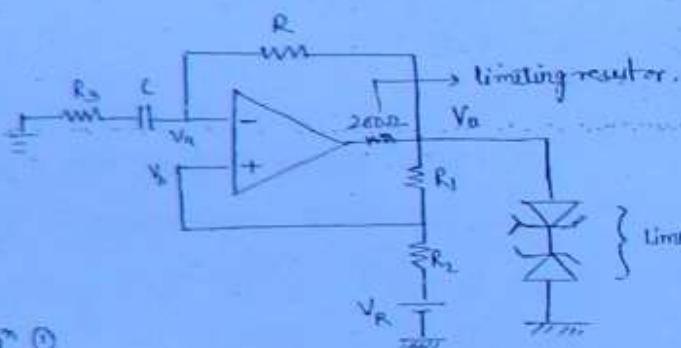
$$T = 2RC \ln \frac{V_{sat} + V_T}{V_{sat} - V_T} \quad \text{--- [Eqn ①]}$$

(68)

$$\therefore V_T = \frac{R_2}{R_1 + R_2} V_{sat}$$

$$T = 2RC \ln \left[ 1 + \frac{2R_2}{R_1} \right] \quad \text{Ans}$$

→  $f = \frac{1}{T}$  → frequency of square wave generated.



→ New time constant for charging/discharging of C -

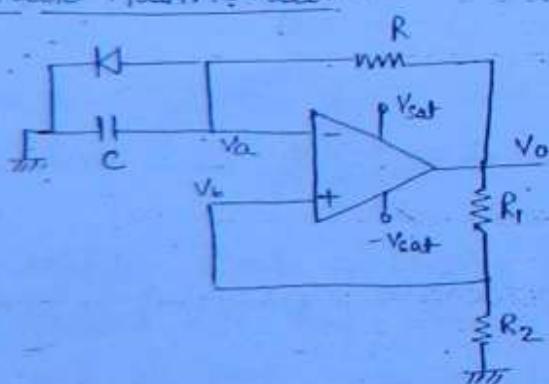
$$\tau = (R + R_3) \cdot C$$

→ Duty cycle can be altered by  $V_R$ .

$V_o = \pm V_{sat}$  can be altered using limiting diodes; i.e., final voltage states of charging & discharging of C can be changed.

\* Time period / charging of C is non linear, to make it linear, we can use a current mirror circuit in place of 'R'.

### Monostable Multivibrator



t < 0 → ckt is in stable state.

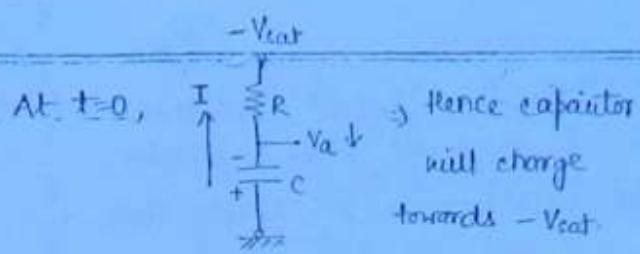
Let  $V_o = +V_{sat} \Rightarrow$  Diode → ON

$$\Rightarrow V_a = V_c = 0$$

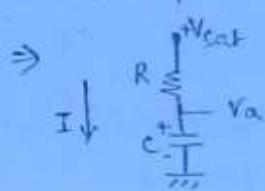
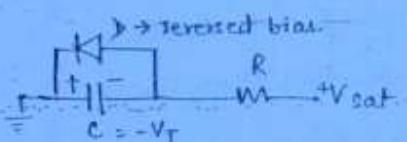
$$V_b = \frac{R_2}{R_1 + R_2} \cdot V_{sat} = +V_T \gg V_a = 0 \\ \Rightarrow \therefore V_o = V_{sat}$$

? At  $t = 0$ , If at b is given, so that  $V_b < V_a$

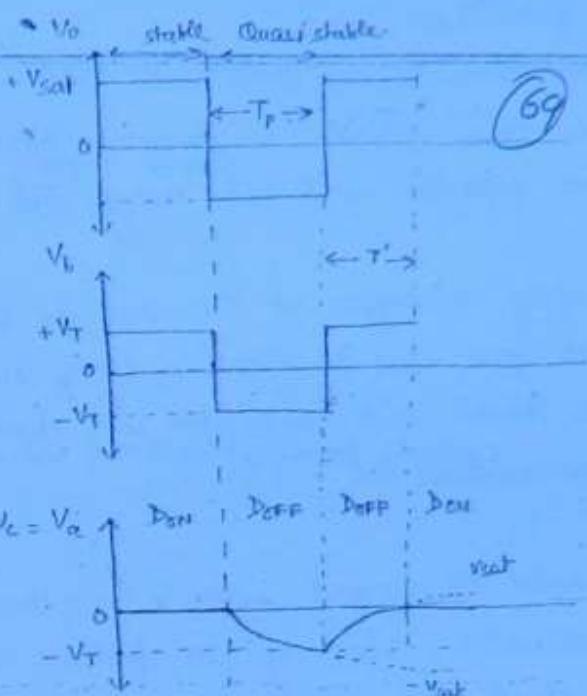
$$\Rightarrow V_o = -V_{sat} \text{ and Diode = OFF}$$



Once,  $V_c$  reaches  $-V_T$ ,  $V_o$  will switch to  $+V_{sat}$  but capacitor does not allow sudden change of voltage. and it will start charging towards  $+V_{sat}$ , since Diode will be off.



Now, as soon as  $V_c$  reaches  $0^+$ , then D will be on and it will start stop charging.



22<sup>nd</sup> August, 2012:-

Derivation of  $T_p$  - (Pulse width) :-

C discharges from 0 to  $-V_T$ .

$$V_c = V_a = -V_{sat} - [-V_{sat} - V_T] e^{-t/R_C}$$

$$\Rightarrow V_c = V_{sat} [1 + e^{-t/R_C}]$$

At  $t = T_p$ ,  $V_c = -V_T$

$$\Rightarrow -V_T = -V_{sat} [1 - e^{-T_p/R_C}]$$

$$\Rightarrow T_p = RC \ln \frac{V_{sat}}{V_{sat} - V_T}$$

$$\Rightarrow T_p = RC \ln \left[ 1 + \frac{R_2}{R_1} \right]^{**}$$

when  $R_2 = R_1$ ,

$$\text{then } T_p = RC \ln 2 = 0.63RC$$

Exp'n for  $T'$  :-

$$T' = RC \ln \frac{V_{sat} + V_T}{V_{sat}}$$

when  $R_1 = R_2$ ,

$$T' = RC \ln (3/2).$$

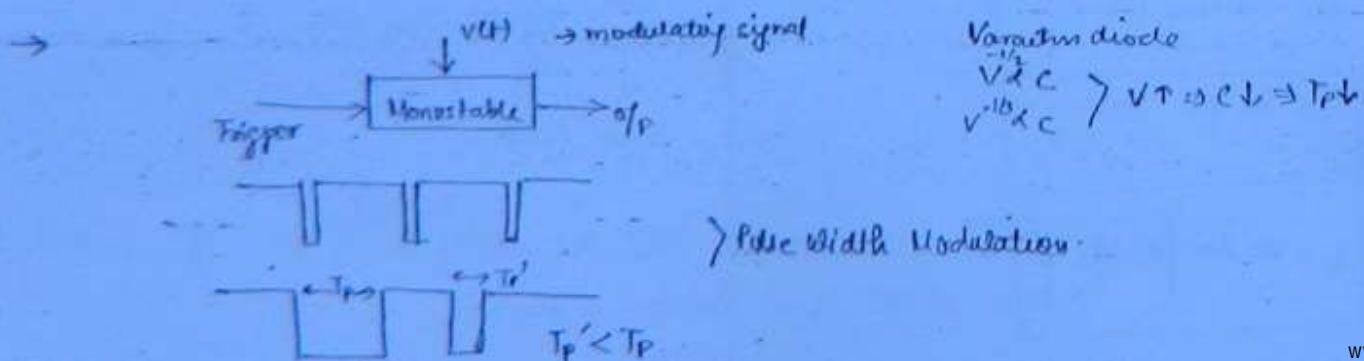
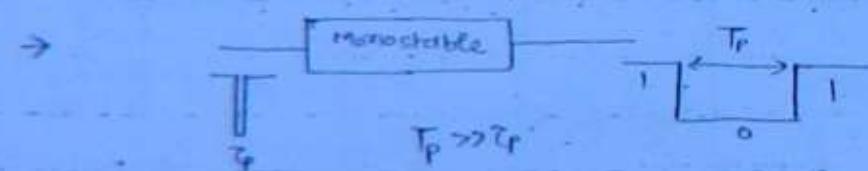
## Circuit Operation

(7)

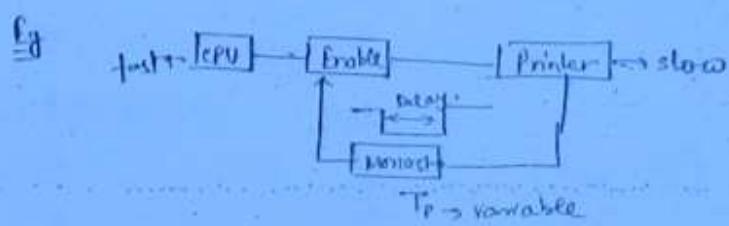
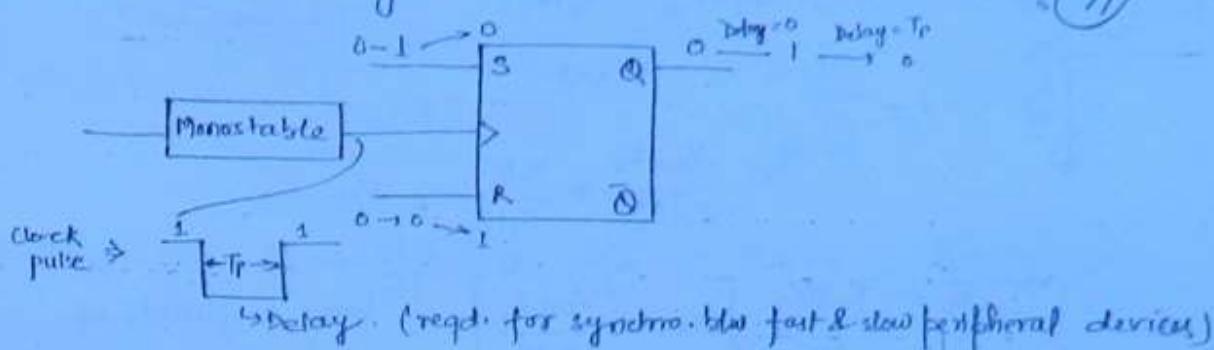
- For  $T_{CO}$ , circuit is in stable state with  $V_o = +V_{sat}$ ,  $V_b = +V_T$ ,  $V_C = V_A = 0$ .
- Since  $V_b = +V_{sat}$ , diode is forward biased & short ckt the capacitor, therefore capacitor will not charge and ckt will remain in stable state.
- Now we apply -ve trigger at  $t=0$  and for short interval,  $V_b < V_A$  and  $V_o$  will switch from  $+V_{sat}$  to  $-V_{sat}$  and  $V_b$  switch from  $+V_T$  to  $-V_T$ . Now diode is reverse biased and capacitor will discharge below 0 towards  $-V_{sat}$  with a time constant  $R.C$ .
- When capacitor discharged upto  $-V_T$ , then  $V_A = V_b = -V_T$  and op-amp comes out of saturation, when capacitor further discharge below  $-V_T$ , then  $V_A < V_b$  as a result of which  $V_o$  switch over to  $+V_{sat}$  and again  $V_b = +V_T$ .
- Now the capacitor will charge above  $-V_T$  towards  $+V_{sat}$  but capacitor can charge only upto 0 because when capacitor charge above 0, diode becomes forward biased and se the capacitor.

## Application :-

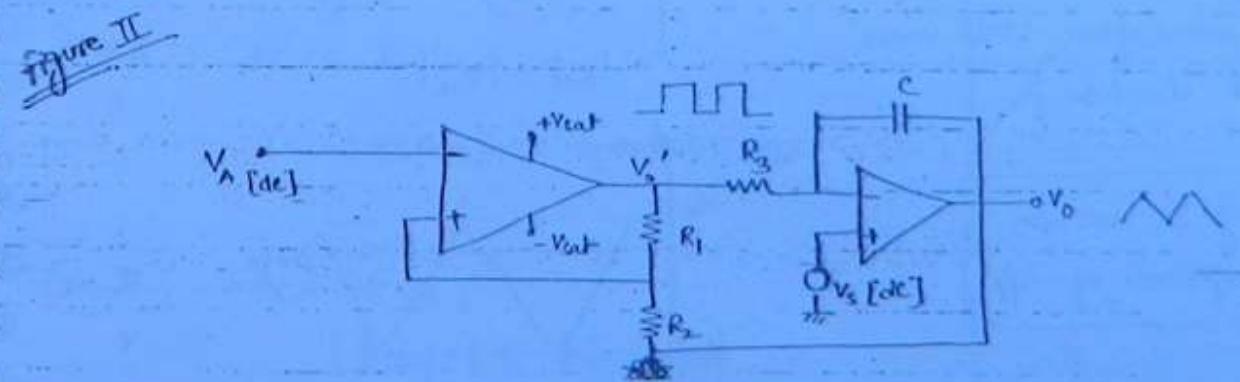
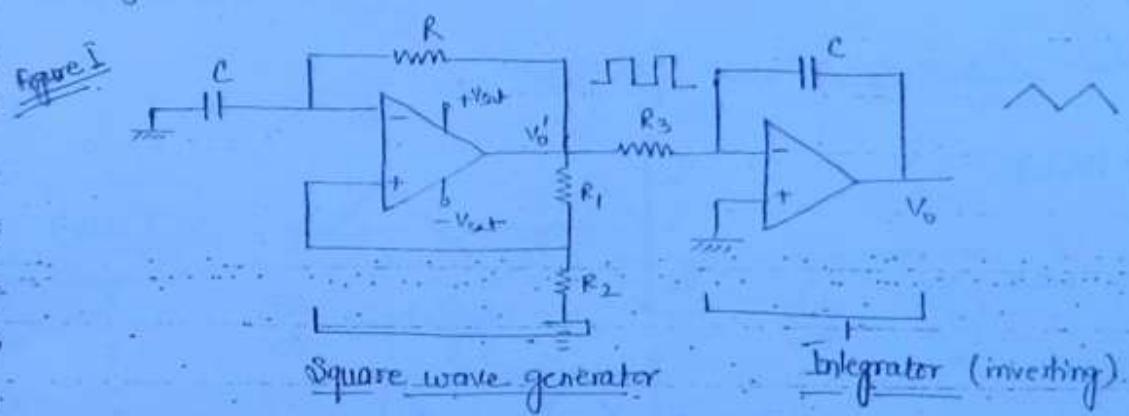
- It is used as Pulse stretcher circuit.



→ It is used as a delay element.

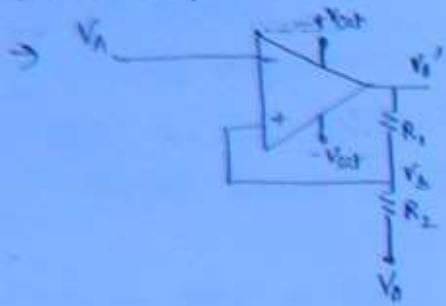


### Triangular Wave Generator :-



→ Figure 2 is better than fig. 1 due to less no. of components.

Variation of  $V_A$ :



(72)

$$V_b = \frac{V_o' R_2}{R_1 + R_2} + \frac{V_o R_1}{R_1 + R_2}$$

$$\textcircled{1} - V_o' = +V_{sat}$$

$$\Rightarrow V_b = \frac{V_{sat} R_2}{R_1 + R_2} + \frac{V_o R_1}{R_1 + R_2} = V_A$$

$$\Rightarrow V_o = \frac{R_1 + R_2}{R_1} \left[ V_A - \frac{V_{sat} R_2}{R_1 + R_2} \right] = \text{lower amplitude of } \Delta \text{ wave.}$$

$$\textcircled{2} - \frac{V_b - V_o}{V_o} = -V_{sat}$$

$$\Rightarrow V_{b2} = -\frac{V_{sat} R_2}{R_1 + R_2} + \frac{V_o R_1}{R_1 + R_2} = V_A$$

$$\Rightarrow V_o = \frac{R_1 + R_2}{R_1} \left[ V_A + \frac{V_{sat} R_2}{R_1 + R_2} \right] = \text{upper amplitude of } \Delta \text{ wave.}$$

Case I - when  $V_A = 0$ ,  $V_{oL} = \frac{R_2}{R_1} V_{sat}$ ,  $V_{OL} = -\frac{R_2}{R_1} V_{sat}$

$$|V_{OL}| = |V_{oL}|$$

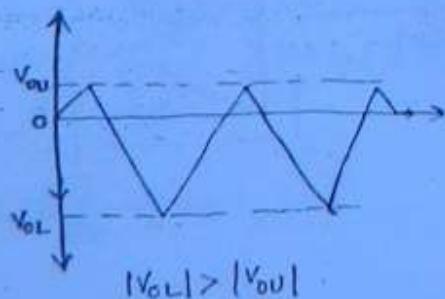
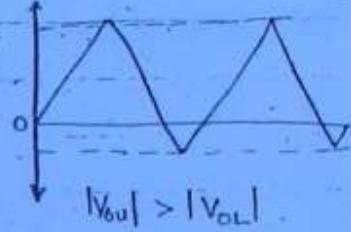
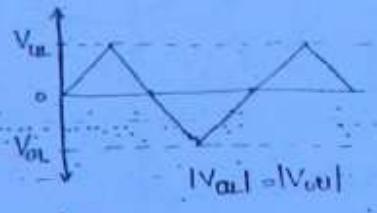
Case II

When  $V_A \uparrow$ , waveform will move

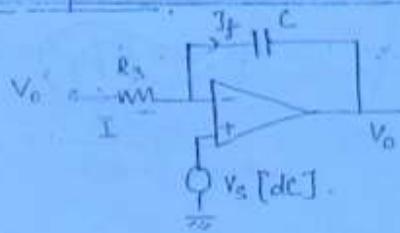
in upward direction.

Case III When  $V_A \downarrow$ , waveform will move in downward direction -

Hence, by changing  $V_A$ , we can control the amplitude of o/p.



### Variation of $V_o$



$$V_p = V_n - V_s$$

(23)

$$I_f = I$$

$$C \frac{d}{dt} (V_n - V_s) = \frac{V_o' - V_s}{R_3}$$

$$\Rightarrow \text{Since } V_s = \text{dc} \Rightarrow \frac{dV_s}{dt} = 0 \Rightarrow -C \frac{dV_o}{dt} = \frac{V_o' - V_s}{R_3}$$

$$\Rightarrow \frac{dV_o}{dt} = -\frac{[V_o' - V_s]}{R_3 C} \quad \text{--- (1)}$$

$\rightarrow$  When  $V_o' = +V_{sat}$ ,

$$\Rightarrow \frac{dV_o}{dt} = -\frac{[V_{sat} - V_s]}{R_3 C} \quad \text{--- (2)} \quad = \text{--ve slope} \quad (\because V_{sat} > V_s),$$

$= \text{constant.} \quad \left\{ \because V_{sat}, V_s, R_3, C = \text{constants} \right.$

$\Rightarrow V_o \downarrow \text{linearly}$

$\rightarrow$  When  $V_o' = -V_{sat}$ ,

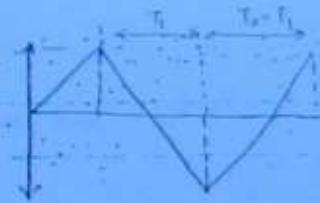
$$\Rightarrow \frac{dV_o}{dt} = \frac{V_{sat} + V_s}{R_3 C} \quad \text{--- (3)} \Rightarrow V_o \uparrow \text{linearly}$$

Case I :-  $V_s = 0$ , in (2) & (3) -

$$\frac{dV_o}{dt} = -\frac{V_{sat}}{R_3 C}$$

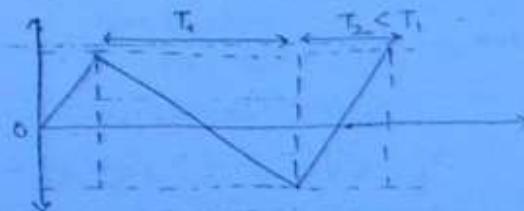
$$\frac{dV_o}{dt} = \frac{+V_{sat}}{R_3 C}$$

$$\Rightarrow |\downarrow \text{slope}| = |\uparrow \text{slope}|$$



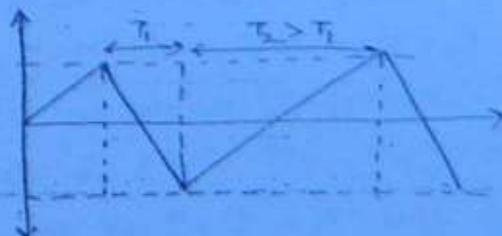
Case II :- when  $V_s > 0$ ,

$$|\uparrow \text{slope}| > |\downarrow \text{slope}|$$



Case III When  $V_s < 0$ ;

$$|\uparrow \text{slope}| < |\downarrow \text{slope}|$$

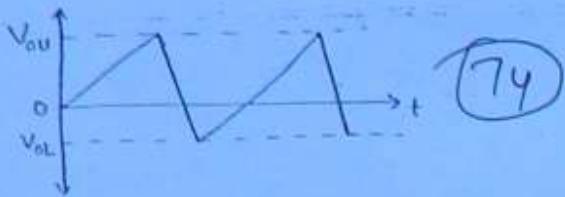


Mence by varying  $V_s$ , we can change the slope of off.

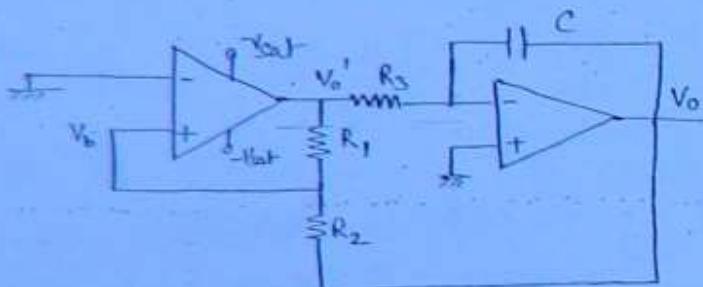
Ex for  $V_A = 2V$ ,  $V_S = -2V$ .

$\Rightarrow V_A = +ve \Rightarrow |V_{oL}| > |V_{oU}|$

$V_C = -ve \Rightarrow |\downarrow slope| > |\uparrow slope|$



Symmetrical Triangular Wave (with  $V_A=0$  and  $V_S=0$ ) :-



① If  $V_o' = +V_{sat}$ ,  $V_o$  will  $\downarrow$  with slope  $\frac{dV_o}{dt} = -\frac{V_{sat}}{R_3 C}$  upto  $V_{oL} = -\frac{R_2}{R_1} V_{sat}$ .

$$V_{oL} = -\frac{R_2}{R_1} V_{sat}$$

② When  $V_o' = -V_{sat}$ ,  $V_o$  will  $\uparrow$  with slope  $\frac{dV_o}{dt} = +\frac{V_{sat}}{R_3 C}$  upto  $V_{oU} = \frac{R_2}{R_1} V_{sat}$ .

Ex : let  $R_1=R_2$  and  $V_{sat} = 10V$ .

$$V_b = \frac{R_2}{R_1+R_2} V_o' + \frac{R_1}{R_1+R_2} V_o = \frac{V_o'}{2} + \frac{V_o}{2}$$

due to noise.

let at  $t=0$ ,  $V_o' = +V_{sat} = 10V$

$V_o = 0$  {due to  $\{ \}$ }

$\therefore V_b = 5V > V_a = 0 \Rightarrow V_o' = +V_{sat}$

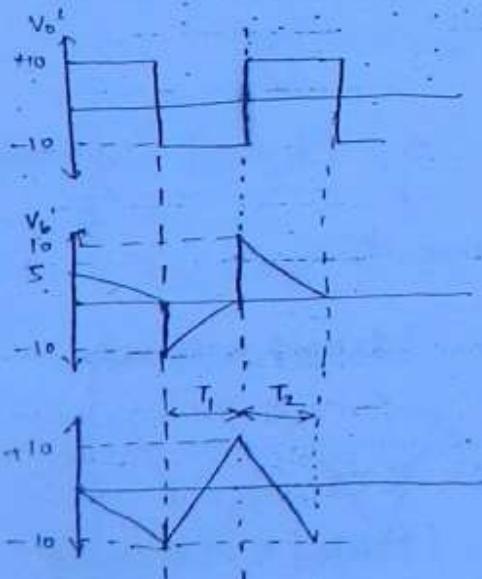
$\therefore V_o' = +V_{sat}$ ,  $V_o \downarrow$  and in turn  $V_b \downarrow$ .

When  $V_o = V_{oL} = -10V$ ,  $V_b = \frac{10-10}{2} = 0V$ .

Note,  $\because V_b \leq V_a$ ,  $V_o'$  switches from  $+V_{sat}$  to  $-V_{sat}$ .

$V_b = -\frac{10-10}{2} = -10V$  and  $\because V_o' = -V_{sat}$ ,  $\therefore V_o$  will  $\uparrow$  and  $V_o \uparrow$  and when  $V_o = V_{oU} = 10V$

then  $V_b = -\frac{10+10}{2} = 0V$  and  $\because V_b \geq V_a$ ;  $V_o'$  switches from  $-V_{sat}$  to  $+V_{sat}$  and cycle will be repeated.



Calculation of  $T_1$  and  $T_2$  :-

75

$$\text{Time} = \frac{\text{Change}}{\text{Rate of change}} = \frac{V_{\text{final}} - V_{\text{initial}}}{\text{slope}}$$

$$\Rightarrow T_1 = \frac{V_{OL} - V_{OU}}{dV_O/dt} = \frac{\frac{R_2}{R_1} V_{sat} - \left(-\frac{R_2}{R_1}\right) V_{sat}}{V_{sat}/R_3 C} \Rightarrow T_1 = \frac{2R_2 R_3 C}{R_1}$$

$$T_2 = \frac{V_{OL} - V_{OU}}{dV_O/dt} = \frac{-\left(\frac{R_2}{R_1}\right)V_{sat} - \left(\frac{R_2}{R_1}\right)V_{sat}}{-V_{sat}/R_3 C} \Rightarrow T_2 = \frac{2R_2 R_3 C}{R_1}$$

$\rightarrow \because T_2 = T_1 \therefore V_o = \text{symmetrical triangular wave.}$   
 $V_o' = \text{symmetrical square wave.}$

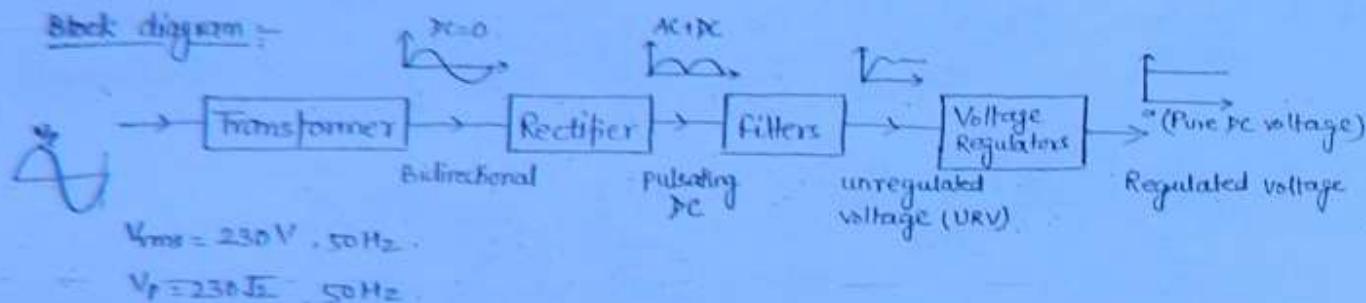
$$\rightarrow \text{Time period} = T = T_1 + T_2 \Rightarrow \boxed{\frac{4R_2 R_3 C}{R_1} = T} \quad \text{or} \quad \boxed{f = \frac{R_1}{4R_2 R_3 C}}^{**}$$

## Diode Circuit :-

(76)

### \* Rectifiers:-

Block diagram :-



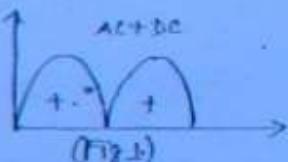
- Basic purpose of a rectifier is to convert a bidirectional voltage or current waveform into unidirectional voltage or current waveforms.

Important terms :-

- Average or DC level,  $I_{dc} = \frac{1}{2\pi} \int_0^{2\pi} I(t) dt$ .

- RMS value,  $I_{rms} = \left[ \frac{1}{2\pi} \int_0^{2\pi} I^2(t) dt \right]^{1/2}$ .

- Ripple Voltage :-



$$V = V_{dc} + V_{ac}$$

$$V_{dc} = \text{dc value of o/p}$$

$$V_{rms} = \text{RMS value of o/p.}$$

$$V_{acms} = \text{RMS value of ac component}$$

$$V_{rms} = \sqrt{V_{dc}^2 + V_{acms}^2}$$

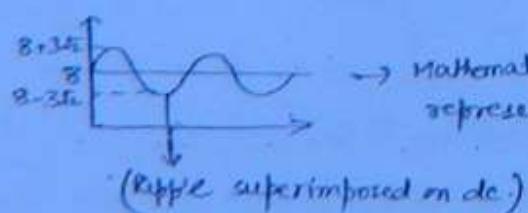
$$\Rightarrow V_{acms} = \sqrt{V_{rms}^2 - V_{dc}^2}$$

Ex let  $V_{dc} = 8V$  and  $V_{acms} = 3V \Rightarrow V_{acpm} = 3\sqrt{2}$

$$\therefore V = 8 + 3\sqrt{2} \sin(\omega t)$$

↓  
Ripple

(Variation of o/p voltage from pure dc)



- It is the deviation of o/p voltage from its dc value. The waveform after rectification is not pure dc. It has an ac component called Ripple superimposed on dc.

→ Ripple factor :-  $\tau = \frac{\text{rms value of ac component}}{\text{dc value}}$

$$\Rightarrow \left[ \tau = \frac{V_{\text{ac rms}}}{V_{\text{dc}}} \right] \quad ; \quad \text{ideally } V_{\text{ac rms}} = 0 \text{ or } \tau = 0$$

$$\Rightarrow \tau = \sqrt{V_{\text{rms}}^2 - V_{\text{dc}}^2} \quad \Rightarrow \tau = \sqrt{\left(\frac{V_{\text{rms}}}{V_{\text{dc}}}\right)^2 - 1}$$

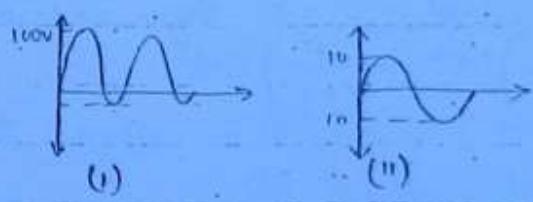
→ form factor :-

$$f = \frac{V_{\text{rms}}}{V_{\text{dc}}} \quad \Rightarrow \tau = \sqrt{f^2 - 1}$$

-  $\boxed{\text{Ideally, } f = 1}$

→ Crest factor :-  $C = \frac{\text{Peak value}}{\text{RMS value}}$

- It should be as low as possible



$\Rightarrow$  RMS is same for both then (ii) signal should be preferred since peak is & and circuit elements will have to be designed accordingly.

→ Peak Inverse Voltage (PIV)

- It is the max voltage across the diode in reverse direction, i.e., when the diode is reverse biased.

- Diode is selected on the basis of PIV rating.

- PIV should be as low as possible.

- We can ↑ PIV of ckt by cascading two or more diodes in series.

Rectifier Efficiency :-  $\eta = \frac{\text{o/p dc power}}{\text{i/p ac power}} \times 100\%$ .

(78)

Transformer Utilization factor (TUF) :-

- It indicates how much is the utilization of transformer in the circuit.
- It should be as  $\uparrow$  as possible.

Types of Rectifiers.

① Half wave rectifier    ② Full wave rectifier - (a) Center tapped transformer type  
 (b) Bridge Rectifier.

Workbook

Chap. 10.

$$1) (A_v)_{dB} = 20 \log A_v = 80 \\ \Rightarrow A_v = 10^4.$$

$$6) BW = A_v \cdot BW = 20 \times 10^4 = 200 \text{ kHz}.$$

Alternate  
 find the point  
 where 20dB/dec.  
 is intersecting freq.  
 axis.

$$3) V_m = V_p = 2V$$

$$\therefore I_E = \frac{10 - 2}{1K} = 8 \text{ mA}$$

$I_B = \text{negligible}$

$$2) \because \frac{10K}{1K} = \frac{10K}{1K} \Rightarrow CMRR = \infty \Rightarrow A_C = 0$$

$$\Rightarrow V_o = A_d (V_1 - V_2)$$

$$\Rightarrow V_o = 0$$

$$\Rightarrow I_E = I_C = 8 \text{ mA}$$

→ 3. (a).

$$4) V_o = \left(1 + \frac{4.14}{10}\right) \cdot V_p = \sqrt{2} \cdot V_p ; \quad V_p = \left(\frac{1/C_S}{R + 1/C_S}\right) \sin t = \left(\frac{1}{1+j}\right) \sin t$$

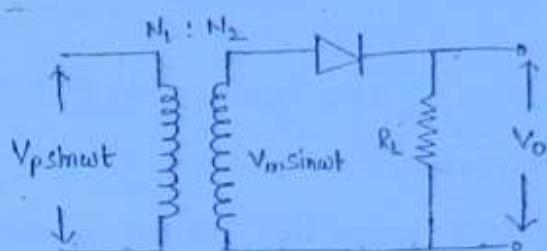
$$\Rightarrow V_o = \frac{\sqrt{2} \sin(t)}{(1+j)} \Rightarrow V_o = \sin(t - \pi/4).$$

$$5) (c). \quad 6) V_o = \left(1 + \frac{2R}{R}\right) \left[ \frac{\sin(100t) + 2 - 2}{2} \right] = \frac{3}{2} \cdot \sin(100t)$$

$$10) CMRR = \frac{A_d}{A_C}, \quad \% \text{ error} = \frac{A_C V_C \times 100}{A_d V_d} = \frac{1}{1000} \times \frac{10 \times 100}{1} = 1\%.$$

23/08/2012

### Half-Wave Rectifier :-



(79)

$$\frac{V_p}{V_m} = \frac{N_1}{N_2}$$

Assuming ideal circuit,

- ①  $V_i > 0$ ,  $\Rightarrow$  FB  $\rightarrow$  short ckt.
- ②  $V_i < 0$ ,  $\Rightarrow$  RB  $\rightarrow$  open ckt.

When D  $\rightarrow$  ON -  $I_L = \frac{V_m \sin \omega t}{R_L}$

$\rightarrow I_m = \frac{V_m}{R_L}$  = max. or peak current through  $R_L$

$\rightarrow V_o = I_L R_L = V_m \sin \omega t$  (ideal case).

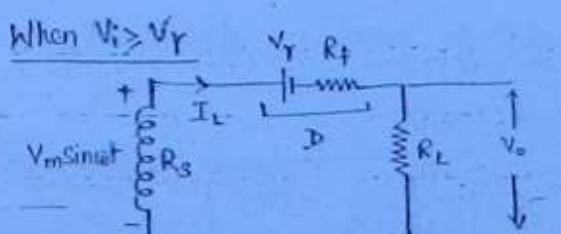
$\rightarrow$  When D  $\rightarrow$  OFF,  $I_L = 0 \Rightarrow V_o = 0$

$\rightarrow$  The opp. frequency or ripple frequency =  $f_r$  = supply frequency f

$\rightarrow$  Conduction angle  $\phi = \pi$  or  $180^\circ$ .

= Practical circuit - ①  $V_i \leq V_Y$ , D  $\rightarrow$  OFF  $\rightarrow$  RB

②  $V_i \geq V_Y$ , D  $\rightarrow$  ON  $\rightarrow$  FB

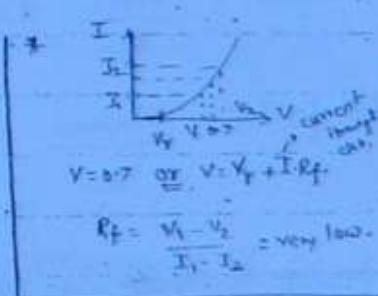
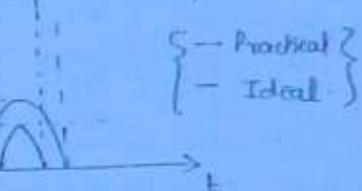


$$* V_D = V_Y + I_f R_f \approx 0.7V \text{ for SI}$$

$R_s$  = Resistance of secondary coil

$$I_L = \frac{V_m \sin \omega t - V_Y}{R_s + R_f + R_L}$$

$$I_m' = \text{max. current} = \frac{V_m - V_Y}{R_s + R_f + R_L} < I_m \text{ (ideal)}$$



When  $V_L < V_T$  —  $\Rightarrow \alpha_f = T_L = 0$

(8)

→ When  $\beta = \infty$ ,  $V_m^2 = I_m^2 \cdot R_L$  —

→ Ripple frequency will remain same as ideal case.

→ Conduction angle  $\left[ \phi = \pi - 2\theta \right] \left\{ \theta < 180^\circ \right\}$

$$V_m \sin \theta = V_T \Rightarrow \left[ \theta = \sin^{-1} \left( \frac{V_T}{V_m} \right) \right]$$

- Average or dc level — (For half wave)

$$I_{dc} = \frac{1}{2\pi} \int_0^{\pi} I_m \sin \omega t \, d(\omega t) \Rightarrow I_{dc} = \frac{I_m}{\pi} \quad \text{— for Ideal}$$

Similarly,

$$V_{dc} = \frac{V_m}{\pi} \quad \text{— for Ideal}$$

- RMS value — (For half wave).

$$I_{rms} = \frac{I_m}{2}$$

$$V_{rms} = \frac{V_m}{2}$$

- Form Factor =  $\frac{V_{rms}}{V_{dc}}$   $\Rightarrow F = 1.57$

- Ripple factor =  $\frac{V_{ac rms}}{V_{dc}} = \sqrt{F^2 - 1} \Rightarrow \tau = 1.21$

- Crest factor =  $\frac{V_{peak}}{V_{rms}} \Rightarrow C = 2$

-  $PIV = +V_m$  (Drop across  $R_L > 0$ ,  $\therefore I_L > 0$ )  $\Rightarrow \eta = \frac{4}{\pi^2} \cdot \frac{R_L}{R_L + R_F + R_S} \times 100\%$

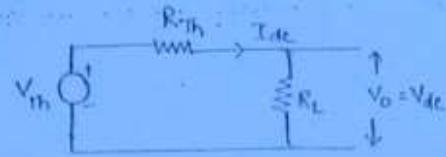
- Rectifier Efficiency,  $\eta = \frac{\text{eff de power}}{\text{imp ac power}} \times 100\% \Rightarrow \eta = \frac{I_{dc}^2 \cdot R_L}{I_{rms}^2 (R_L + R_S + R_F)} \times 100\%$

$$\Rightarrow \eta = 0.406 \times \frac{1}{\frac{R_s + R_f}{R_L} + 1} \times 100\% \quad (81) \quad \left\{ \begin{array}{l} \eta = 40\% \text{ means only } 40\% \\ \text{ac power is converted to dc} \end{array} \right.$$

If  $R_L \ggg R_s + R_f$ , then  $\boxed{\eta_{max} = 40.6\%}$

If the efficiency is 100%, it means that 100% of ac power is converted into dc and remaining 60% (approx.) power is in form of ripple (ac component at off).

### Thevenin's equivalent of Half Wave Rectifier



$$I_{dc} = \frac{V_m}{R_m + R_L} = 0$$

$$V_{dc} = I_{dc} R_L$$

$$I_{dc} = \frac{1}{2\pi} \int_0^{\pi/2} I_L \sin \omega t \, d\omega t ; \theta = \sin^{-1}\left(\frac{V_r}{V_m}\right); I_L = \frac{V_m \sin \omega t - V_r}{R_s + R_f + R_L}$$

$$\text{Let } V_r = 0, \Rightarrow \theta = 0, I_L = \frac{V_m \sin \omega t}{R_s + R_f + R_L} = I_m \sin \omega t.$$

$$\therefore I_{dc} = \frac{1}{2\pi} \int_0^{\pi/2} I_m \sin \omega t \, d\omega t = \frac{I_m}{\pi} = \frac{V_m}{\pi(R_s + R_f + R_L)} \quad (2)$$

Comparing (1) and (2) :-

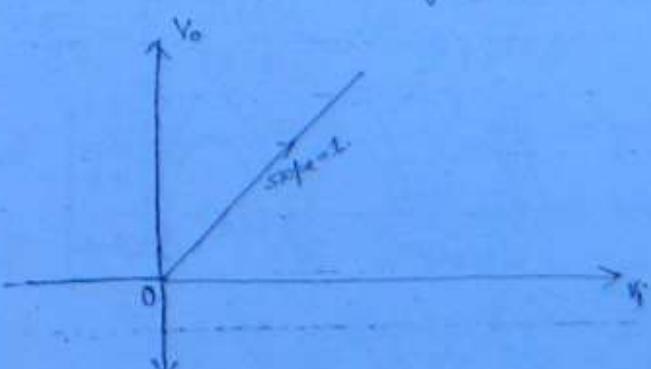
$$V_{Th} = \frac{V_m}{\pi}, \quad R_{Th} = R_s + R_f$$

\*  $R_{Th}$  is the off resistance of ckt & it represents the losses occurring at off.

### Transfer Curve:-

Ideal :

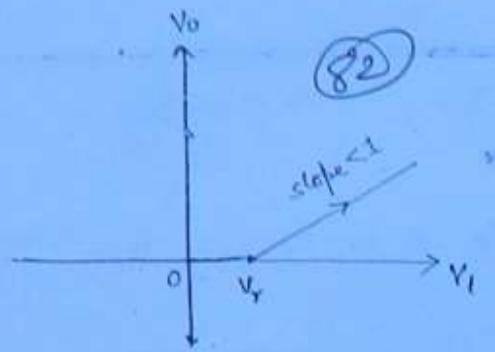
$V_i$	D	$V_o$
$V_i \leq 0$	off	0
$V_i > 0$	on	$V_i$



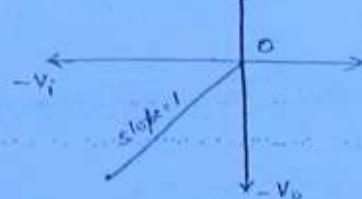
Practically

$V_i$	D	$V_o$
$V_i \leq V_r$	OFF	0
$V_i \geq V_r$	ON	$I_L R_L = \frac{(V_i - V_r)}{R_s + R_L + R_f} \times R_L$

$$\text{slope} = \frac{R_L}{R_s + R_L + R_f} < 1$$



\* If diode polarity is reversed, then charac will come into III quadrant.



→ Transfer Utilization factor-

$$\boxed{\text{TUF} = 0.286} \rightarrow (\text{very low})$$

Ques: A HVR is supplied by a 230V, 50Hz supply with a step down ratio of 3:1 to a resistive load  $R_L = 10\text{ k}\Omega$ . If  $R_f = 750\Omega$  and  $R_s = 10\Omega$ ,

Calculate—

- ① Max, average and rms value of current
- ② DC value of off voltage
- ③ Efficiency

$$\text{Soln} \quad V_m = \frac{230}{3} \text{ V} \Rightarrow V_m = \frac{230\sqrt{2}}{3} \text{ V.} = 108.4 \text{ V}$$

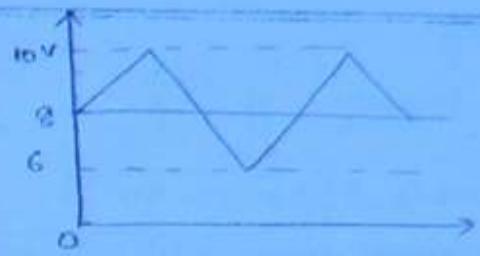
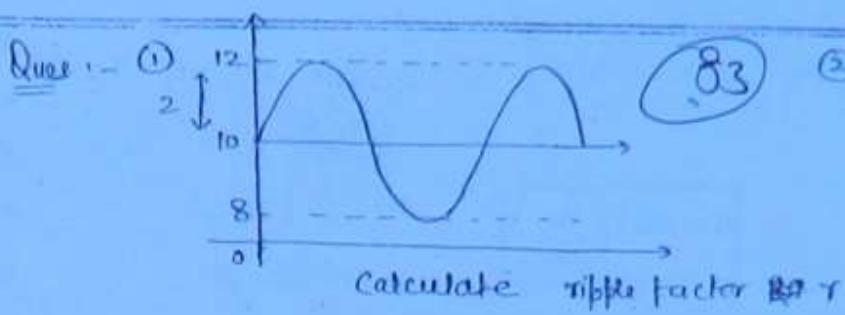
$\Rightarrow V_r = 0 \text{ V} \quad \{ \text{not mentioned} \}$

$$I_m = \frac{V_m - V_r}{R_f + R_L + R_s} \approx \frac{230\sqrt{2}}{3(10\text{k})} = \frac{23\sqrt{2}}{3} \text{ mA.} = 10.84 \text{ mA}$$

$$I_d = I_{avg} = \frac{I_m}{\pi} = \frac{23\sqrt{2}}{3\pi} \text{ mA} \approx 3.45 \text{ mA} \quad V_{dc} = I_{dc} \times R_L = \frac{23\sqrt{2} \times 10}{3\pi} \text{ V} \approx 19.8 \text{ V}$$

$$V_{rms} = \frac{230}{3 \times 10\text{k}} \times \frac{23\sqrt{2}}{3} \text{ mA} = 11.66 \text{ mA} \quad \eta = \frac{0.106 \times 10\text{k}}{10\text{k} + 9.5} \approx 40.6\%$$

$$I_{rms} = \frac{I_m}{2} = 5.42 \text{ mA.}$$



$$\textcircled{1} \quad F = \frac{V_{\text{rms}}}{V_{\text{dc}}} = \frac{\sqrt{10^2 + (2/\pi)^2}}{10} = \frac{\sqrt{102}}{10}$$

$$\therefore \gamma = \sqrt{F^2 - 1} = \frac{1}{5\pi}$$

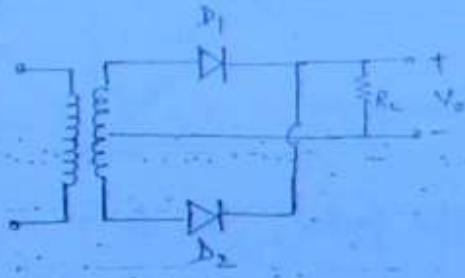
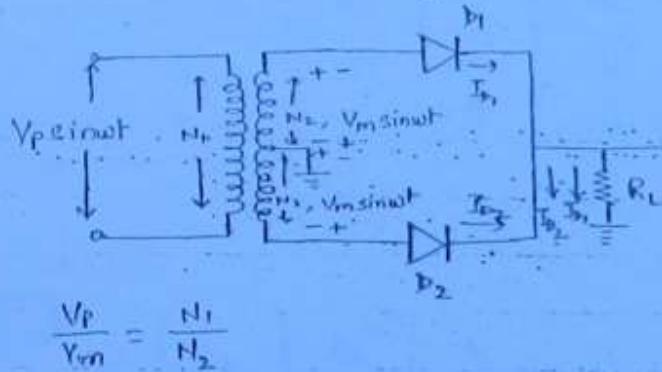
$$\textcircled{2} \quad \text{rf} = \frac{V_{\text{ac rms}}}{V_{\text{dc}}} = \frac{(2/\pi)}{10}$$

$$\gamma = \frac{1}{5\pi}$$

$\textcircled{3} \quad \text{rf} = \frac{V_{\text{ac rms}}}{V_{\text{dc}}} = \frac{(2/\pi)}{8} = \frac{1}{4\pi}$

Full Wave Rectifier :-

a) Center Tapped Transformer Type :-



Ideally:

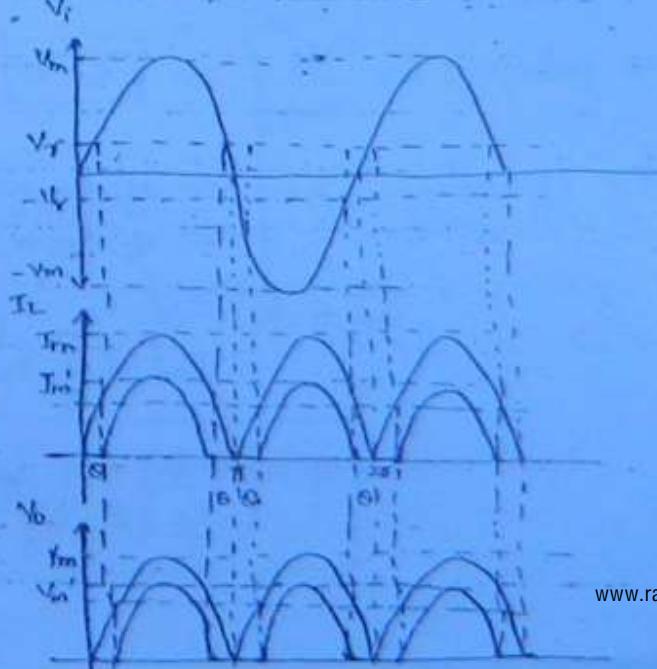
$$I_L = \frac{V_m \sin \omega t}{R_L} = I_m \sin \omega t; \quad I_m = \frac{V_m}{R_L}$$

$$V_o = I_L R_L = V_m \sin \omega t$$

$$\rightarrow \text{Ripple frequency} = [f_r = 2f]^{**}$$

$$\rightarrow \text{conduction Angle} = [\phi = 2\pi]^{**}$$

$$\text{for individual diode, } [\phi = \pi]^{**}$$



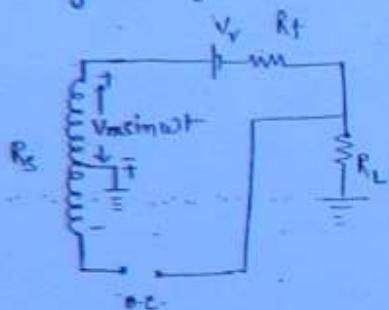
Practically:

→ Ripple frequency,  $f_r = 2f$ .

→ for circuit,  $\phi = 2\pi - 4\theta$ ;  $\theta = \sin^{-1}\left(\frac{V_r}{V_m}\right)$   
for individual diode,  $\phi = \pi - 2\theta$

(84)

→ During +ve cycle -



$$I_L = \frac{V_m \sin \omega t - V_r}{R_f + R_L + \frac{R_s}{2}} \quad (\text{Amp})$$

$$I_m' = \frac{V_m - V_r}{R_f + R_L + \frac{R_s}{2}} \quad (< I_m)$$

$$V_o = V_L \cdot I_L$$

$$\rightarrow I_{dc} = \frac{1}{\pi} \int_0^{\pi} I_m \sin \omega t \cdot d\omega t \Rightarrow I_{dc} = \frac{2I_m}{\pi}, \quad V_{dc} = \frac{2V_m}{\pi}$$

$$\rightarrow I_{m_{avg}} = \frac{I_m}{\sqrt{2}}, \quad V_{m_{avg}} = \frac{V_m}{\sqrt{2}}$$

$$\rightarrow \text{form factor, } f = \frac{V_m/\sqrt{2}}{2V_m/\pi} \Rightarrow F = 1.11$$

$$\rightarrow \text{Ripple factor, } \tau = \sqrt{F^2 - 1} \Rightarrow \tau = 0.48$$

$$\rightarrow \text{Growth factor, } c = \frac{V_m}{V_m/\sqrt{2}} \Rightarrow C = \sqrt{2}$$

$$\rightarrow \text{Rectifier Efficiency, } \eta = \frac{\text{dc o/p power}}{\text{ac i/p power}} \times 100\%$$

$$\Rightarrow \eta = \frac{I_{dc}^2 \cdot R_L}{I_{ans}^2 \left( \frac{R_s}{2} + R_f + R_L \right)} \times 100\%$$

$$\Rightarrow \eta = \left( \frac{0.812 \times R_L}{\frac{R_s}{2} + R_f + R_L} \right) \times 100\%$$

$$\Rightarrow \eta = \left( 1 + \frac{R_f + \frac{R_s}{2}}{R_L} \right)^{-1} \times 100\%$$

$$\text{If } R_L \gg R_f + \frac{R_s}{2}$$

$$\Rightarrow \eta_{max} = 81.2\%$$

→ Peak Inverse voltage  $\Rightarrow$   $PIV = 2V_m$

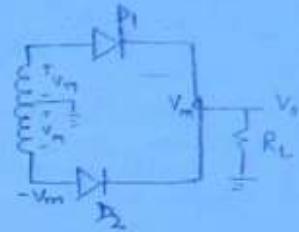
(85)



$$V_{D_1} = V_o - V_m = -V_m - V_m$$

$$\Rightarrow V_{D_1} = -2V_m$$

$$\Rightarrow PIV = 2V_m$$



$\left. \begin{array}{l} \text{When } V_D > 0 \quad D \rightarrow \text{FB} \\ V_D < 0 \quad D \rightarrow \text{RB} \end{array} \right\}$

Thevenin's Equivalent of FWR :-



$$R_{TH} = \frac{R_S + R_L}{2}$$

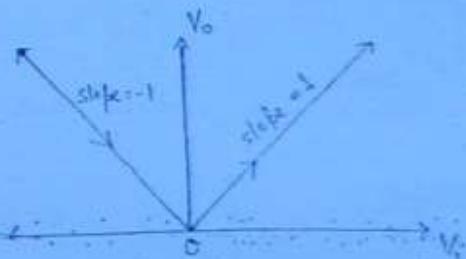
$$V_{TH} = \frac{2V_m}{\pi}$$

Transfer curve :-

Ideally :-

$V_i$	$D_1$	$D_2$	$V_o$
$V_i < 0$	off	on	$-V_i$
$V_i > 0$	on	off	$V_i$

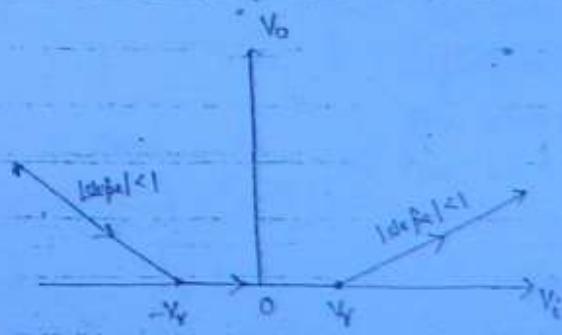
$$\therefore V_o = -V_i$$



Practically :-

$V_i$	$D_1$	$D_2$	$V_o$
$V_i \leq -V_Y$	off	on	$V_o'$
$-V_Y \leq V_i \leq V_Y$	off	off	0
$V_i > V_Y$	on	off	$V_o'$

$$V_o' = I_L R_L = \frac{V_i - V_Y}{\frac{R_S + R_f + R_L}{2}} \times R_L$$



$$\text{slope} = \frac{R_L}{\frac{R_S + R_f + R_L}{2}} \quad (< 1) \Rightarrow |(\text{slope})| < 1$$

→ If the polarity of diodes is reversed, the transfer curve will be present in III and IV quadrant.

(86)

→ In practical condition, it is not possible to rectify very small signals using centre tapped Transfer rectifier.

e.g.  $V_i = 5 \sin \omega t \text{ mV}$   $\Rightarrow V_m = 5\text{mV} = 0.005 \text{ V} \ll V_T$ , hence o/p will be 0.

→ TUF  $\sim$   $\boxed{\text{TUF} = 0.693}$

→ Workbook

Pg 57 (chap. 10).

(ii)  $\times$

24<sup>th</sup> August, 2012

Bridge Rectifier

$$\frac{V_p}{V_m} = \frac{N_1}{N_2}$$

for positive half-

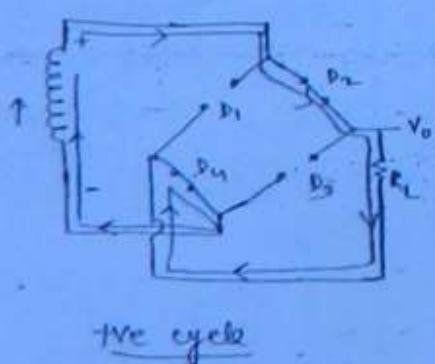
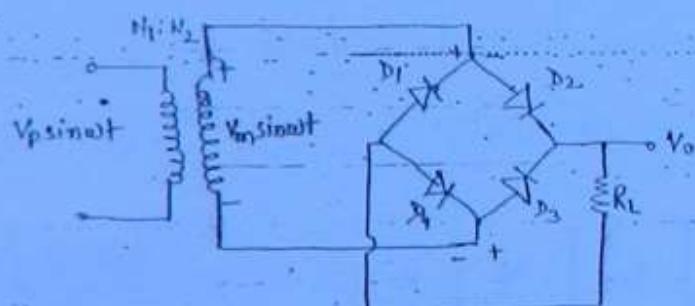
$D_1$  &  $D_3$  off.

$D_2$  &  $D_4$  on.

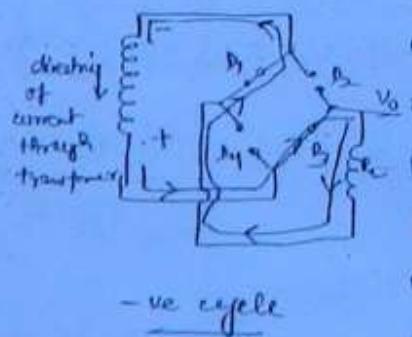
for -ve half-

$D_1$  &  $D_3$  on

$D_2$  &  $D_4$  off.



positive cycle



-ve cycle

- The current through transformer coil is bidirectional, hence avg. dc component is zero, which in turn results in minimum loss in  $\text{X}_{\text{per}}$ . (87)
- TUF is maximum for Bridge rectifier due to above mentioned reason.
- zero dc prevents the Eddy current, hysteresis losses and saturation of  $\chi_{\text{per}}$ .

### \* Ideally

$$\rightarrow V_{\text{dc}} = \frac{2V_m}{\pi}; I_{\text{dc}} = \frac{2I_m}{\pi}$$

$$\rightarrow I_{\text{rms}} = \frac{I_m}{\sqrt{2}}; V_{\text{rms}} = \frac{V_m}{\sqrt{2}}$$

$$\rightarrow r = 0.48$$

$$\rightarrow F = 1.11$$

$$\rightarrow C = \sqrt{2}$$

$$\rightarrow \phi = 2\pi$$

$\rightarrow$  Individual diode,  $\phi = \pi$

### Practically

$$\rightarrow I_L = \frac{V_m \sin \omega t - 2V_r}{R_s + 2R_f + R_L}$$

$$\rightarrow I_m' = \frac{V_m - 2V_r}{R_s + 2R_f + R_L}$$

$$\rightarrow \eta = 0.812 \times \frac{1}{\frac{R_s + 2R_f + 1}{R_L}} \times 100\%$$

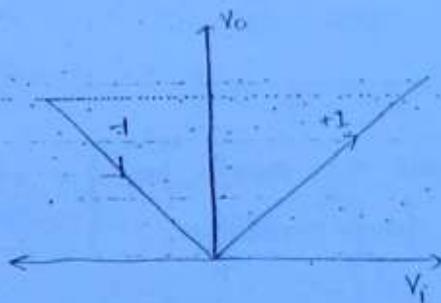
$$\rightarrow \eta_{\text{max}} = 81.2\%$$

$$\rightarrow \phi = 2\pi - 4\theta; \theta = \sin^{-1} \frac{2V_r}{V_m}$$

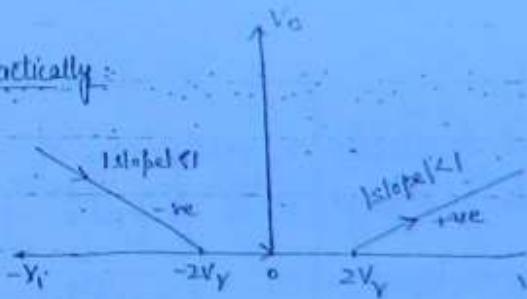
$\rightarrow$  Individual diode,  $\phi = \pi - 2\theta$ .

### Transfer Curve

#### Ideally:

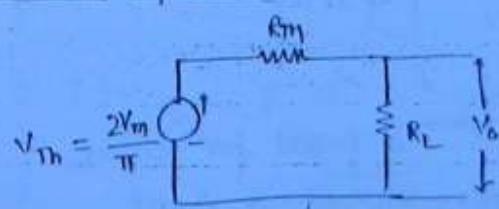


#### Practically:

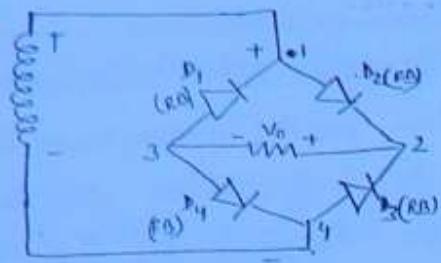
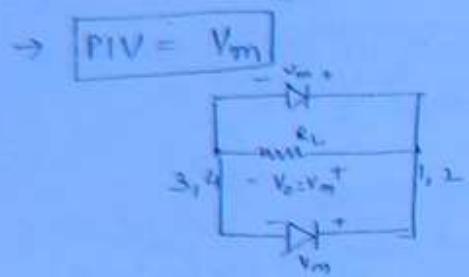


$$\text{Slope: } \frac{R_L}{R_s + 2R_f + R_L} (< 1)$$

### Thevenin's Equivalent:



$$R_{\text{Th}} = R_s + 2R_f$$



$\rightarrow \text{TUF} = 0.812$

### Advantages of Bridge Rectifier

- TUF is highest.
- $\rightarrow$  Transformer can be replaced by ac source if step up/down of voltage is not required.
- $\rightarrow$  PIV is smaller as compared to half-wave.
- $\rightarrow$  Voltage required to deliver same power is smaller ~~cost~~ half wave rectifier, hence to do so, no. of turns is more in HWR, hence the size of transformer used in Bridge rectifier is smallest.

### Disadvantage -

- $\rightarrow$  It cannot be used for rectification of small signals as cutoff freq. for response is  $2V_o$ , though it is preferable for high power ratings.

$V_m$	21hr	loss
2	1	50%
10V	1	10%
20V	1	5%

### By Sir :- advantages:-

- (i) The current in both primary & secondary of X<sup>mer</sup> is present for entire cycle and hence for a given power o/p, power X<sup>mer</sup> of a small size and less cost may be used.

→ No centre-tap is required in  $X^{\text{mer}}$  secondary, hence whenever possible, ac voltage can directly be applied to bridge.

(89)

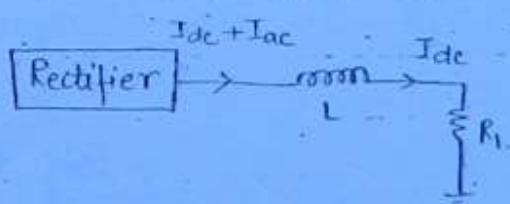
- The current in secondary of  $X^{\text{mer}}$  is in opposite direction in two half cycles and hence net dc component through  $X^{\text{mer}}$  coil is zero. Which reduces the losses and reduces the danger of saturation of  $X^{\text{mer}}$ .
- As two diodes conduct in series, in each half cycles, inverse voltage appearing across the diode get shared hence the circuit can be used for high voltage applications. (since PIV is less)

### Filter Circuits :-

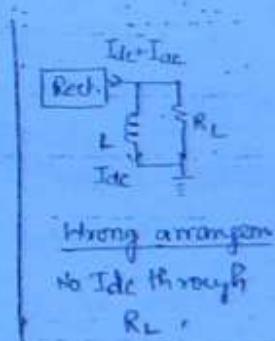
- To minimise ripple (ac component) at the o/p, filter circuits are used. We are using inductor & capacitor in filter circuits.

► Inductor :-  $|Z_L| = \omega L = 2\pi f L$

for dc,  $\therefore f=0 \Rightarrow Z_L=0 \Rightarrow L$  acts as SC for dc.



$|Z_L|$  should be very high so that it blocks ac  
 $L \rightarrow$  very high



→  $L \uparrow$ , and/or  $f \uparrow \Rightarrow |Z_L| \uparrow \Rightarrow$  ac at o/p  $\downarrow \Rightarrow$  Ripple  $\downarrow \Rightarrow \tau \downarrow$

$$\left( \tau \propto \frac{1}{fL} \right) - \textcircled{1}$$

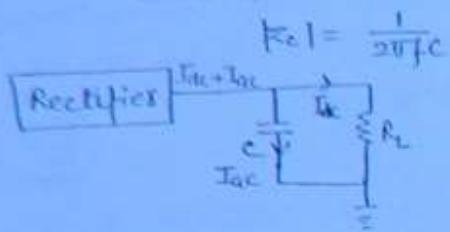
→  $\tau \uparrow \Rightarrow \frac{L}{R_L} \uparrow \Rightarrow L \uparrow$  and  $R_L \downarrow \Rightarrow$  variation in current  $V$ .  
 $\Rightarrow \tau \downarrow$

$$\therefore \left( \tau \propto \frac{1}{V} \right) - \textcircled{2}$$

from  $\textcircled{1} \& \textcircled{2}$  -

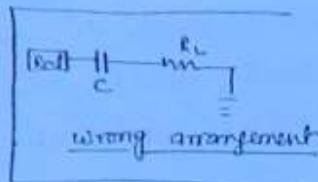
$$\tau \propto \frac{R_L}{fL}$$

2) Capacitance :-



90 - for dc,  $f=0 \Rightarrow Z_C = \infty$ .

→ Capacitor acts as open circuit for dc.



-  $|Z_{CL}|$  should be very high now for ac to bypass it

$$\Rightarrow [C \rightarrow \text{very high}]$$

-  $C \uparrow$  and/or  $f \uparrow \Rightarrow |Z_{CL}| \downarrow \Rightarrow$  ac through  $R_L \downarrow \Rightarrow$  ripple &  $v \downarrow$

$$\Rightarrow v \propto \frac{1}{C \cdot f}$$

$\Rightarrow T = R_L C \Rightarrow$  should be very  $\uparrow$   $\Rightarrow$  variation in  $v \downarrow \Rightarrow$  ripple &  $v \downarrow$ .

$$v \propto \frac{1}{C} \Rightarrow T = R_L C \Rightarrow [C \uparrow \text{ and } R_L \uparrow] \text{ for } v \downarrow$$

$$\Rightarrow [v \propto \frac{1}{f \cdot C \cdot R_L}]$$

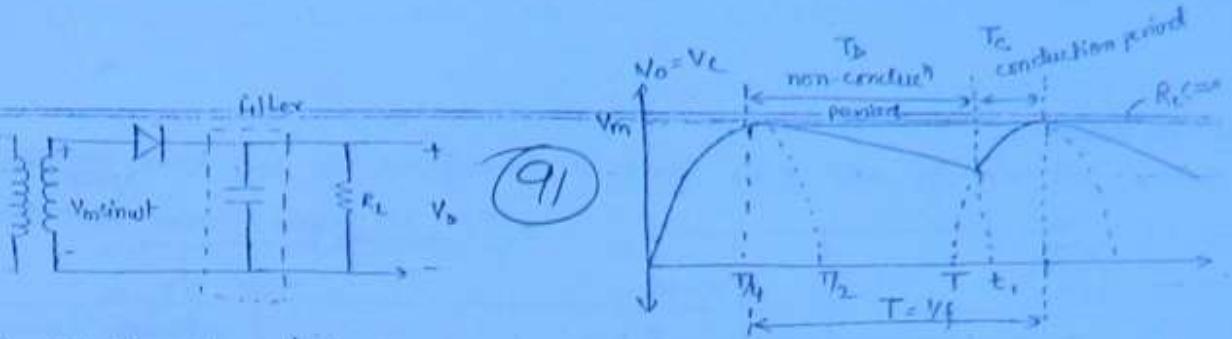
∴ Hence, for low load resistance, inductor is preferred and for high load resistance, capacitor is preferred.

Alternatively, for low load ( $R_L$  high) capacitor is preferred and for high load ( $R_L$  low) inductor is preferred.

Types of filter :-

- 1) Capacitor Filter
- 2) Choke or Inductor filter
- 3) L section or L-C filter
- 4) TI or CLC filter
- 5) TI or CRC filter for compact circuit.

→ Capacitor filter :- HWR with capacitor filter



- $V_c(0) = V_c(0^+) = 0V$ . Initially C acts as S.C and  $V_o = V_c = 0V$ .
- For the first half of  $V_i$ ,  $D \rightarrow FB \rightarrow ON$ .
- $\tau = (R_f \parallel R_L) \cdot C$   
 $= R_f \cdot C \quad (\because R_f \ll R_L)$
- $\tau$  should be such that  $\tau \ll T \Rightarrow$  rate of changing of C should be very high.
- At  $T = T/4$ ,  $V_i = V_m$  and  $V_c = V_o = V_m$ .

$$V_i = V_m \rightarrow D \rightarrow V_c = V_o = V_m$$

for  $T > T/4$ ,  $V_i \downarrow$  & when  $V_i < V_0$ ,  $\Rightarrow D \rightarrow RB \rightarrow OFF$ .

$\rightarrow$   $r' R_{LC}$  should be  $\gg T$ , so that rate of discharging of C is very slow.  $\Rightarrow V_c \downarrow$  exponentially.

$\rightarrow V_i$  will go up and then fall and when  $V_i \geq V_c \Rightarrow D \rightarrow FB \rightarrow ON$  and it will again charge C with  $\tau = R_f C$  upto  $V_m$ . Thus, the cycle repeats.

$\rightarrow$  from plot  $\rightarrow$

$$T_D + T_C = T = \text{time period of signal.}$$

$\rightarrow$  When we  $\uparrow R_{LC}$ ; then  $T_D \uparrow$ ,  $T_C \downarrow$ , variation  $\downarrow$ ,  $f \downarrow$ .

for best filter,  $T_D \gg T_C$

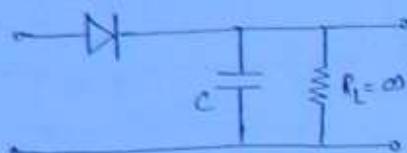
$$\text{or } T_D \approx T = V_f \quad \text{--- (1)}$$

Ideal condition:

$\rightarrow$  If  $R_{LC} = \infty$ ,  $V_o = V_m \rightarrow$  pure dc

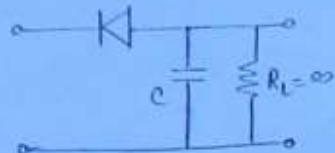
$$\Rightarrow \omega = 0, \quad f = 1; \quad C = 1 \dots$$

## Peak Detector :-

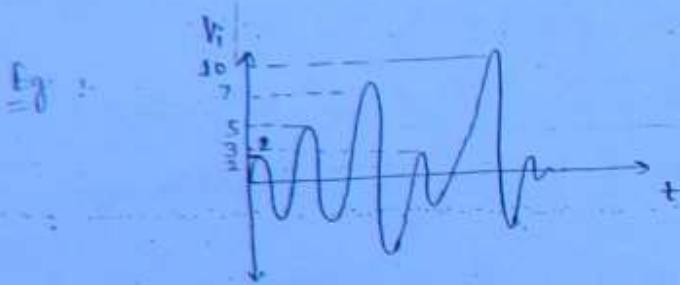


+ve Peak detector

(92)



-ve peak detector



$V_o$  = will charge upto

(i) 2V and hold  
Hence 8V and hold.

(ii) 5V .. "

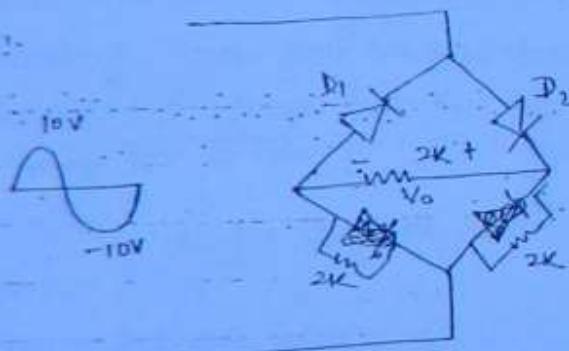
(iii) 7V .. "

(iv) 3V .. don't charge  $\rightarrow 7V$   
(diode will be RB).

(v) 10V  $\rightarrow$  charge upto 10V.

Hence, the op-amp will always hold the max. value of i/p.

Ques :-



Sol<sup>n</sup> for +ve half -

$$V_{24} = V_{234} = 10V$$

$$\therefore V_o = \frac{10}{2} = 5V$$

During -ve half

$$\text{again } V_o = \frac{10}{2} = 5V$$

in same direction.

Assume ideal diodes

① - Draw the o/p waveform.

② - find o/p dc level

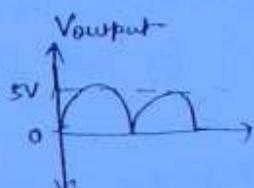
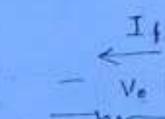
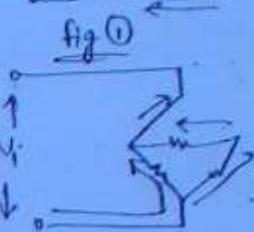
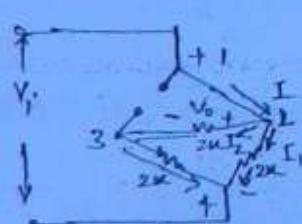
③ - find PIV.

$$\text{② o/p dc level} = \frac{2 \times V_m}{\pi} = \frac{10}{\pi}$$

$$\text{③ PIV} = \text{from fig ①} \quad PIV = V_1 - V_3$$

$$= 10 - 5 = 5V$$

Ans



\* The circuit of given que is comparable to HVR since the opf for the same ip would have been same as  $\frac{10V_m}{\pi}$ .

(93)

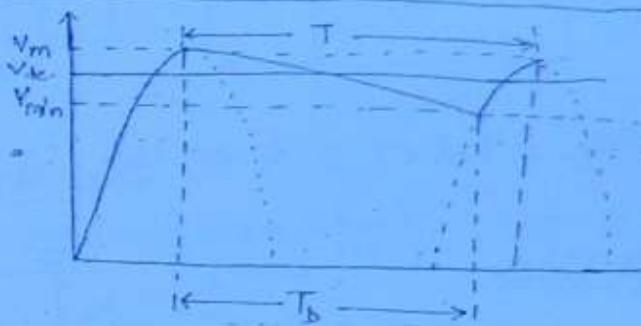
Approximate solution

-  $T_D \approx T = \frac{1}{f} \quad \text{--- (1)}$

-  $R_L C = \text{very high}$

- During  $T_D$ , C will discharge

$$V_o = V_c = V_m e^{-t/R_L C}$$



$$V_c \approx V_m \left[ 1 - \frac{t}{R_L C} \right] \quad \left\{ \because R_L C \text{ very high} \right\}$$

-  $V_{dc} = \frac{V_{m+V_{min}}}{2} \quad \text{--- (2)}$

-  $V_r = \text{peak to peak value of ripple voltage.}$

-  $\therefore V_{dc} = V_m - \frac{V_r}{2} \quad \text{--- (3)}$

-  $V_f = V_{min} - V_{max} = \text{change in } V_c \text{ during time } T_D.$

-  $V_f = \frac{\Theta(\text{discharge})}{C}$

-  $I_m = \frac{V_m}{R_L}, \quad T_{min} = \frac{V_{min}}{R_L}$

$$\frac{I_{dc}}{T} = \frac{V_{dc}}{R_L}$$

$\therefore T_{dc} = \frac{1}{2} \left( \frac{V_m + V_{min}}{R_L} \right) = \frac{V_{dc}}{R_L}$

$$\frac{I_{dc}}{T} = \frac{V_{dc}}{R_L}$$

$$\therefore \Theta(\text{discharge}) = \frac{V_{dc}}{R_L} \times T_D = I_{dc} \times T \quad \left\{ \because T = T_D \right\}$$

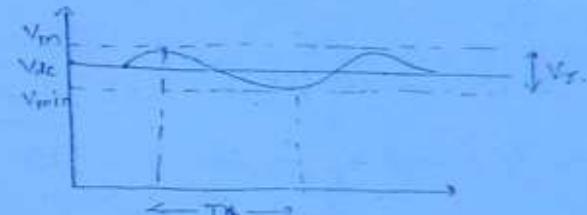
$$V_f = \frac{I_{dc}}{C \cdot f}$$

$$V_f = \frac{V_{dc}}{R_L \cdot C \cdot f}$$

from (3) -

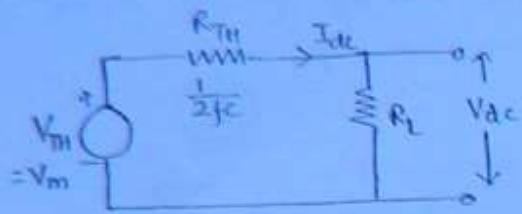
$$V_{dc} = V_m - \frac{I_{dc}}{2fC}$$

$$\text{or } \left\{ V_{dc} = \frac{I_{dc}}{R_L} \right\}$$



(94)

### Thevenin's Equivalent :-



$$V_{dc} = V_m - I_{dc}(R_{TH})$$

Comparing with last eqn-

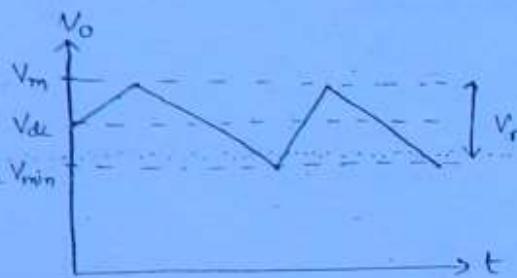
$$V_m = V_{m0} ; R_{TH} = \frac{1}{2fC}$$

\* When  $f \uparrow$  &  $C \uparrow$  or  $R_L \rightarrow \infty$  then  $V_{dc} \approx V_m$ . (Ideal case  $\rightarrow$  pure dc) in off

### Ripple factor :-

$$\begin{aligned} V_{acrms} &= \frac{V_p}{\sqrt{3}} = \frac{V_p}{2\sqrt{3}} \\ &= \frac{I_{dc}}{2\sqrt{3} f C} = \frac{V_{dc}}{2\sqrt{3} f C R_L} \end{aligned}$$

$$\left\{ \because V_r = V_{p-p} \text{ and } \frac{V_{p-p}}{2} = V_p \right\} \Rightarrow \tau = \frac{V_{acrms}}{V_{dc}} \Rightarrow \boxed{\tau = \frac{1}{2\sqrt{3} f C R_L}}$$



### HWR with C

-  $f_T$

- Ripple voltage,  $V_r$

-  $V_{dc}$

- Thevenin's equivalent

-  $\tau$

- Waveform.

→ 3φ rectifier

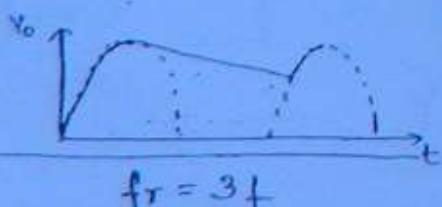
$f$

$I_{dc}/fC$

$$V_{dc} = V_m - \frac{I_{dc}}{2fC}$$

$$V_{m0} = V_m, R_{TH} = \frac{1}{2fC}$$

$$\tau = \frac{1}{2\sqrt{3} f C R_L}$$



### FWR with C (Bridge/center tapped)

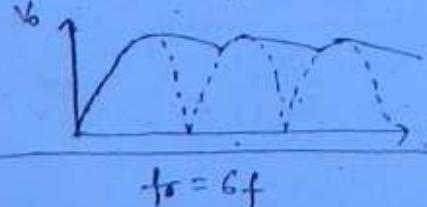
$2f$

$I_{dc}/2fC$

$$V_{dc} = V_m - \frac{I_{dc}}{4fC}$$

$$V_{m0} = V_m, R_{TH} = \frac{1}{4fC}$$

$$\tau = \frac{1}{4\sqrt{3} f C R_L}$$



## \* Peak Inverse Voltage with C -

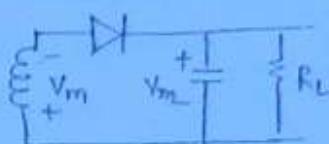
(95)

HWR

C → max. charged.

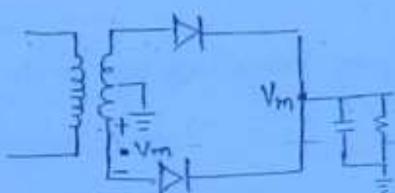
$$\Rightarrow V_C = V_m.$$

$$\therefore \text{PIV} = 2V_m$$



$$V_{D} = V_m - (-V_m) \\ = 2V_m.$$

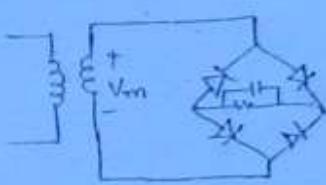
FWR



Center Tapped

$$\text{PIV} = 2V_m$$

{ same as before? }  
(w/o C)



Bridge

$$\text{PIV} = V_m$$

{ same as before, i.e., w/o C filter }

## Surge Current or Peak Diode Current :-

During  $T_D \rightarrow$  C. discharge

$$Q(\text{discharge}) = I_{dc} \cdot T_D.$$

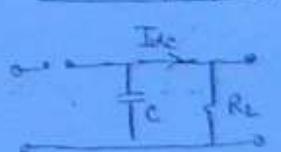
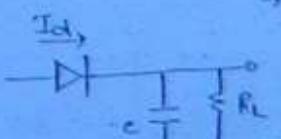
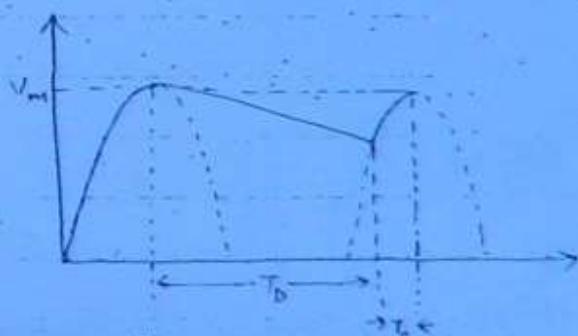
During  $T_C \rightarrow$  Diode ON  $\rightarrow$  C will charge

$$Q(\text{charge}) = I_d \cdot T_C.$$

According to law of conservation of Q -

$$I_D \cdot T_C = I_{dc} \cdot T_D$$

$$\therefore I_p = \frac{I_{dc} \cdot T_D}{T_C}$$



27/08/2012

Q6

\* If  $R_L \ll T$ , then  $T_D \approx T_C$  &  $\left\{ \begin{array}{l} T \downarrow \\ \text{(adv)} \end{array} \right.$  but  $I_D \uparrow$   $\left( \text{Diode} \right)$



for best filter,

$$T_D \gg T_C$$

$$\Rightarrow V_{dc} \approx V_m, \quad I_{dc} \approx I_m = \frac{V_m}{R_L} \quad \text{For } V_m = 10 \text{ V} \& R_L = 10 \text{ k}\Omega \Rightarrow I_m = 1 \text{ mA}$$

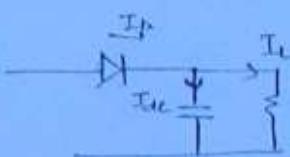
$$\text{For } f=50 \text{ Hz}, \quad T=1/f = 20 \text{ msec}, \quad T_D = 19.98 \text{ msec}, \quad T_C = 0.02 \text{ msec} \quad (\text{say})$$

$$I_D = \text{surge current} = \frac{I_{m\text{MAX}} 19.98}{0.02} \approx 1000 \text{ mA.} \quad (\text{very large}) \rightarrow \text{High power diss.} \\ \rightarrow \text{diode damage.}$$

Conduction Angle =

$$\phi = \omega T_C = \sqrt{\frac{2V_T}{V_m}}, \quad \omega = \frac{d\theta}{dt} \text{ rad/sec.}$$

$$\rightarrow V_T = \frac{V_{dc}}{fR_L C}$$



$$* I_{Dmax} = I_L \left[ 1 + 2\pi \sqrt{\frac{2V_m}{V_T}} \right]$$

$$\rightarrow V_{dc} = V_m - \frac{V_T}{2} ; \quad V_{dc} \approx V_m ; \quad I_L \approx I_{DC} \approx I_m.$$

$$\Rightarrow I_m = \frac{V_m}{R_L}, \quad I_{dc} = \frac{V_{dc}}{R_L}$$

for FWR :

$$\phi = \omega T_C = \sqrt{\frac{2V_T}{V_m}}$$

$$\rightarrow V_T = \frac{I_{dc}}{2fC}$$

$$\rightarrow I_{Dmax} = I_L \left[ 1 + 2\pi \sqrt{\frac{V_m}{2V_T}} \right]$$

$$\rightarrow V_m \approx V_{dc}; \quad I_L \approx I_{rm} = I_{dc}.$$

Pg.30 (Workbook)

Q1 (conventional) -

Given -  $V_{dc} = 30V$ ,  $\gamma \leq 0.01$ ,  $R_L = 500\Omega$ ,  $f = 50Hz$ .

$I_{max} = ?$ ,  $C = ?$

$$\because \gamma = \frac{1}{2\pi f C R_L} \leq 0.01 \Rightarrow C \geq 1154 \mu F$$

$$I_{max} = I_L \left[ 1 + 2\pi \sqrt{\frac{2V_m}{V_r}} \right]$$

$$V_r = \frac{V_{dc}}{f C R_L} = \frac{30}{50 \times 1154 \times 10^{-6} \times 500} = 1.02V$$

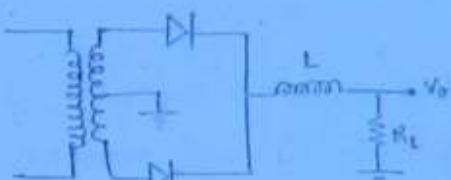
$$I_L \approx I_{dc} \approx \frac{V_{dc}}{R_L}$$

$$V_m = V_{dc} + \frac{V_r}{2} = 30.5V$$

Substitute,  $I_L$ ,  $V_r$  and  $V_m$  & calc.  $I_{max}$ .

Inductor Filter (or) Choke filter :-

$$\rightarrow \gamma = \frac{2}{3\pi^2} \cdot \frac{1}{\sqrt{1 + (\frac{X_L}{R_L})^2}}$$



$$X_L = \omega L \text{ for HWR}$$

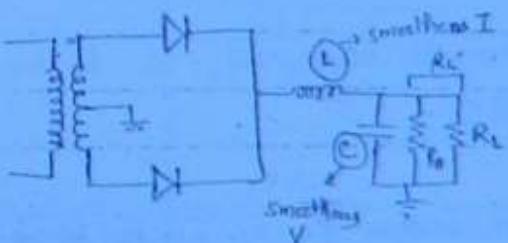
$$Y_L = 2\omega L \text{ for FWR}$$

$$\rightarrow \text{If } \left(\frac{X_L}{R_L}\right)^2 \gg 1 \Rightarrow \gamma \ll \frac{R_L}{X_L} \Rightarrow \boxed{\gamma \ll \frac{R_L}{f \cdot \pi^2 L}} = \boxed{\gamma \ll \frac{1}{f \cdot C}}$$

L-section or LC filter :-

$$\rightarrow \gamma = \frac{5}{3} \cdot \left( \frac{X_C}{X_L} \right) ; X_C = \frac{1}{\omega C} ; X_L = \omega L \text{ for HWR}$$

$$X_C = \frac{1}{2\omega C} ; X_L = 2\omega L \text{ for FWR}$$



$$\rightarrow \boxed{\gamma \ll \frac{1}{f^2}} \quad \therefore \gamma \text{ is very small.}$$

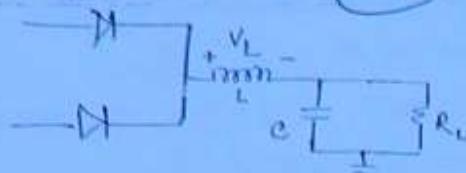
$$R_B = \text{Bleeder Resistance}$$

$$R'_L = R_B || R_L \geq R_L \left[ \left( \because f_B \gg R_L \right) \right]$$

$$\rightarrow \boxed{\gamma \text{ is independent of } R_L}$$

98

- \* When there is sudden change in current - then  $\frac{di}{dt}$  = large.



$$\therefore V_L = \frac{L}{R_L} \frac{di}{dt} = \text{very large.} \rightarrow \text{Back emf.}$$

- This  $V_L$  will act as reverse bias for both diodes.
- > This sudden change occurs when circuit is ON w/o  $R_L$ .

$$\therefore Z = \frac{L}{R_L} = 0 \quad (\because R_L \text{ off} = \infty) \quad (\text{O.C})$$

$$\therefore t=0, \text{ then } \frac{di}{dt} = \text{large} \Rightarrow V_L = \text{large.}$$

Hence,  $R_B$  is attached in the off. so that even if  $R_L = \infty$ . (O.C) effective resistance,  $R'_L = R_B \parallel \infty = R_B$  and hence  $Z = \frac{L}{R'_L}$  is never equal to 0. Therefore, no sudden change of current.

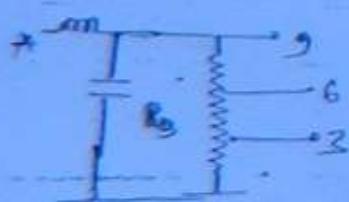
- > When  $R_B$  is attached across a capacitor, it helps C to discharge through it. when supply and  $R_L$  are removed.

By Sir

- \* The basic req. of this filter is the current through choke must be continuous. An interrupted current through choke may develop large back emf which may be in excess of PV rating of diode and/or max rating of capacitor.

- \* To eliminate back emf, a bleed resistor  $R_B$  is connected across off terminal.

- \* Another reason for  $R_B$  is to bleed off voltage stored in filter capacitor when supply is turned off.

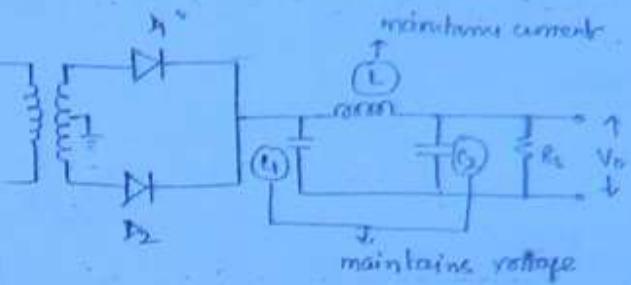


Potential Divider  $\rightarrow$  Different off's from diff. points from  $R_B$ .

(99)

### II or CLC Filter :-

$$\tau = \frac{\sqrt{2} \cdot X_C \cdot X_{C_2}}{R_L \cdot X_L} = \frac{\sqrt{2} \cdot (X_C)^2}{R_L \cdot X_L} \quad \left\{ C_1 = C_2 \right\}$$



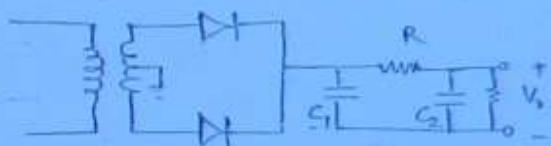
$$\rightarrow X_C = \frac{1}{\omega C} ; X_L = \omega L \quad \text{for HWR}$$

$$\rightarrow X_C = \frac{1}{2\omega C} ; X_L = 2\omega L \quad \text{for FWR.}$$

$$\rightarrow \tau \propto \frac{1}{\omega^2 C_1 C_2 L R_L} \Rightarrow \boxed{\tau \propto \frac{1}{f^2}} \quad \therefore \tau = \text{Very small.}$$

### II or CRC filter :-

$$\tau = \frac{\sqrt{2} \cdot X_C \cdot X_{C_2}}{R_L \cdot R} = \frac{\sqrt{2} (X_C)^2}{R_L \cdot R} \quad (\text{for } C_1 = C_2).$$



$$\rightarrow X_C = \frac{1}{\omega C} \quad \text{for HWR}$$

$$\rightarrow X_C = \frac{1}{2\omega C} \quad \text{for FWR.}$$

$$\rightarrow \tau \propto \frac{1}{\omega^2 C_1 C_2 R_L R} \Rightarrow \boxed{\tau \propto \frac{1}{f^2}} \quad \tau = \text{small (relatively more than CLC filter.)}$$

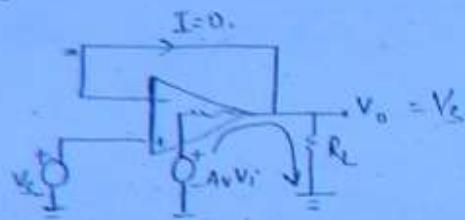
<u>*</u>	<u>C</u>	<u>L</u>	<u>LC</u>	<u>CLC</u>
$\rightarrow \tau \propto \frac{1}{f}$		$\tau \propto \frac{1}{f}$	$\tau \propto \frac{1}{f^2}$	$\tau \propto \frac{1}{f^3}$
$\rightarrow \tau \propto \frac{1}{\tau}$		$\tau \propto \frac{1}{\tau}$	$\tau \propto \frac{1}{\tau_1 \tau_2}$	$\tau \propto \frac{1}{\tau_1 \tau_2 \tau_3}$
$\rightarrow \tau = R_L C$		$\tau = \frac{1}{R_L}$	$\tau_1 \tau_2 = \frac{L \cdot R_L C}{R_L}$ $= LC$	$\tau_1 \tau_2 \tau_3 = (R_L C_1) (R_L C_2) \times \frac{L}{R_L}$ $= C_1 C_2 L R_L$
$\rightarrow \tau \propto \frac{1}{R_L C}$		$\tau \propto \frac{1}{R_L}$	$\therefore \tau \propto \frac{1}{LC}$	$\therefore \tau \propto \frac{1}{C_1 C_2 L R_L}$

## Precision Rectifiers :-

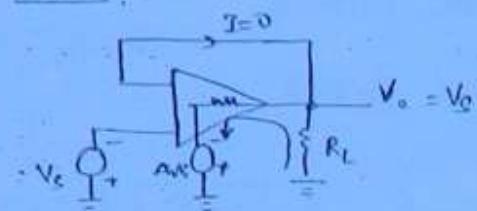
(100)

### - Voltage follower -

$V_s > 0$



$V_s < 0$



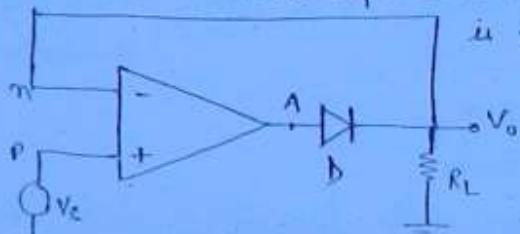
Hence, to maintain  $V_o$  at a level of input  $V_s$ , op-amp should provide a current as indicated above. If this current = 0, then  $V_o = 0$ .

### - Precision HWR :-

→ Assuming Ideal op-amp & practical diode.

$V_s$	$V_A$	D	$V_o$
$V_s > 0$ (very small) $\approx \mu V$	↑ towards $+V_{sat}$ Reaches till $V_A \approx 0.7V$	ON	$V_s$
$V_s < 0$	$V_{A(b)} (-V_{sat})$ Reaches to final value $= -V_{sat}$	OFF	0 (since there is no current due to D in $R_B$ )

(This circuit is also called superdiode since cut-in voltage is very small,  $\approx \mu V$ )



→ Till the time D is about to 'ON', the op-amp will act as open loop. When D is on, the due to -ve feedback, applying virtual ground, high gain. Hence,  $V_P = V_N = V_s = V_o$ . (see next page  $V_o$  vs  $V_i$  plot)

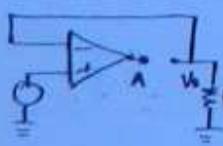
$$\therefore V_A = 0.7V + V_o$$

$$\Rightarrow V_A = 0.7V + V_s$$

$$\Rightarrow V_A \approx 0.7V$$

Hence, the diode D will avoid op-amp to go into positive saturation.

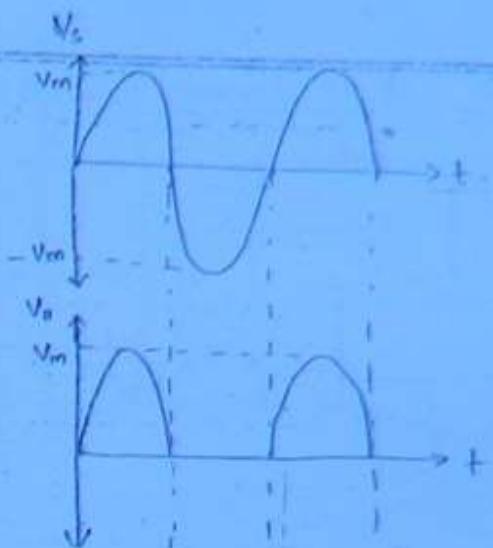
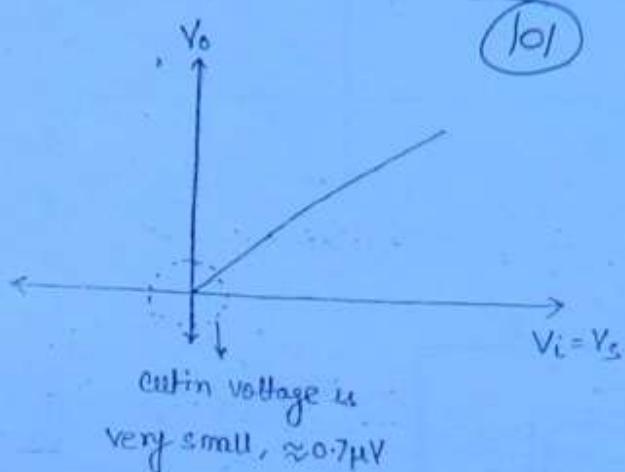
When  $V_s < 0$  —



Op-amp will behave as open loop and  $A_v = 10^6$

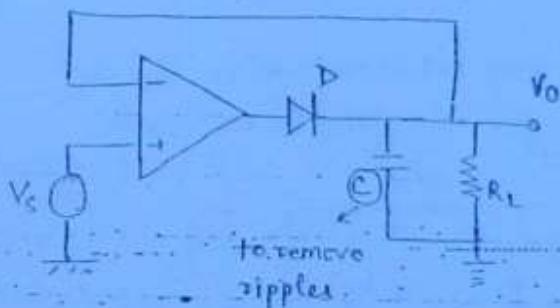
$$\therefore \text{for small } V_s, V_o = -V_{sat}$$

$\rightarrow$  PIV for the diode D is  $-V_{sat}$ .

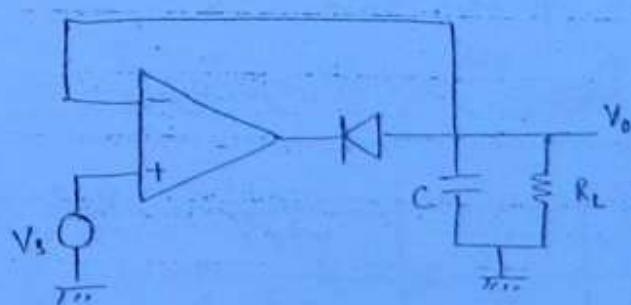


$$\left. \begin{aligned} \text{When } V_i = 0.7 \mu\text{V}, \quad V_A = A_v \cdot V_i = 10^6 \times 0.7 \mu\text{V} &= 0.7 \text{ V} \\ &= V_0 \end{aligned} \right\}$$

Drawback: PIV is very high.



\* If  $R_L C = \infty$ , then above circuit will act as +ve peak detector.

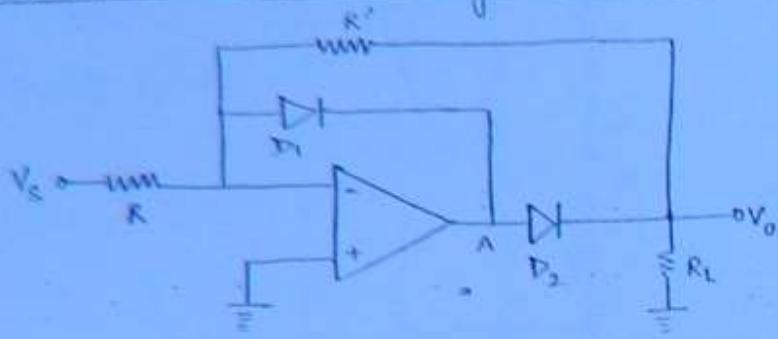


$\rightarrow$  This circuit will avoid -ve saturation for  $V_o < 0$ .

$\rightarrow$  If  $R_L C = \infty$ , then it will act as -ve peak detector.

Precision HWR (i/p at Inverting Terminal).

(162)



When  $V_c > 0$  -

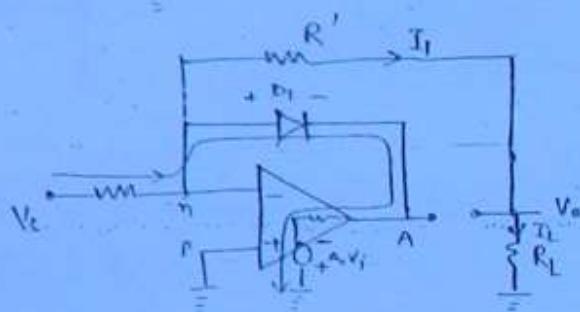
$$D_1 = \text{ON}, D_2 = \text{OFF}$$

$$V_m = V_p = 0$$

$$V_m = 0.7V \therefore V_A = -0.7V$$

$\therefore D_1$  will avoid negative saturation when i/p is +ve.

$$I_1 = I_2 = 0 \Rightarrow V_o = 0$$



$\rightarrow$

$$-0.7V \xrightarrow{D_1} 0 \Rightarrow \text{PIV} = 0.7V. \quad \left\{ \text{very less as compared to } V_{\text{sat}} \text{ as we were getting in last case} \right\}$$

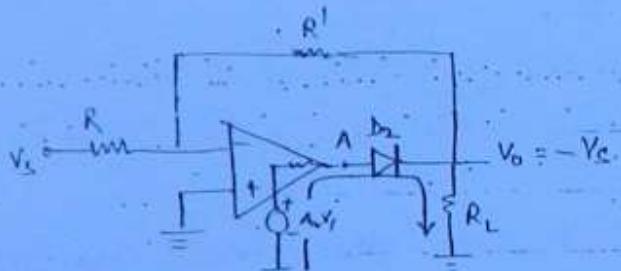
When  $V_c < 0$  -

$$D_1 = \text{OFF}, D_2 = \text{ON}$$

$$\rightarrow V_o = -\frac{R'}{R} V_s = -V_s \quad \left\{ \text{if } R = R' \right\}$$

$$\rightarrow V_A = 0.7 - V_s \approx 0.7V$$

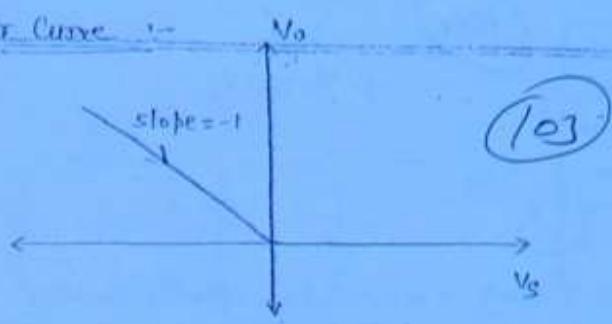
$\therefore D_2$  will avoid positive saturation when i/p is -ve.



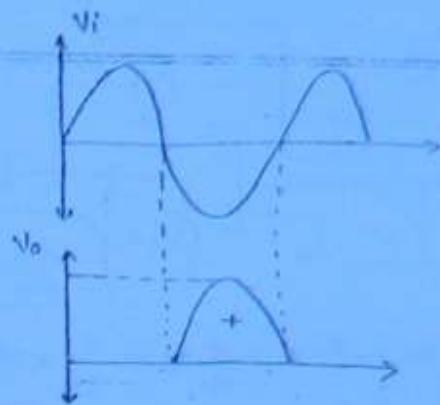
$$\rightarrow 0 \xrightarrow{D_1} -0.7 \Rightarrow \text{PIV} = 0.7V$$

$V_s$	$V_A$	$D_1$	$D_2$	$V_c$	PIV
$\rightarrow V_s > 0$	$\downarrow \text{towards } -V_{\text{sat}} (-0.7V)$	ON	OFF	0	0.7V for $D_2$
$\rightarrow V_s < 0$	$\uparrow \text{towards } +V_{\text{sat}} (0.7V)$	OFF	ON	$-V_s$	0.7V for $D_1$

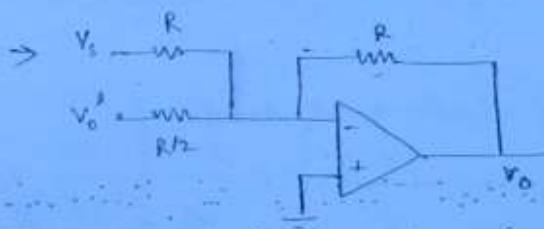
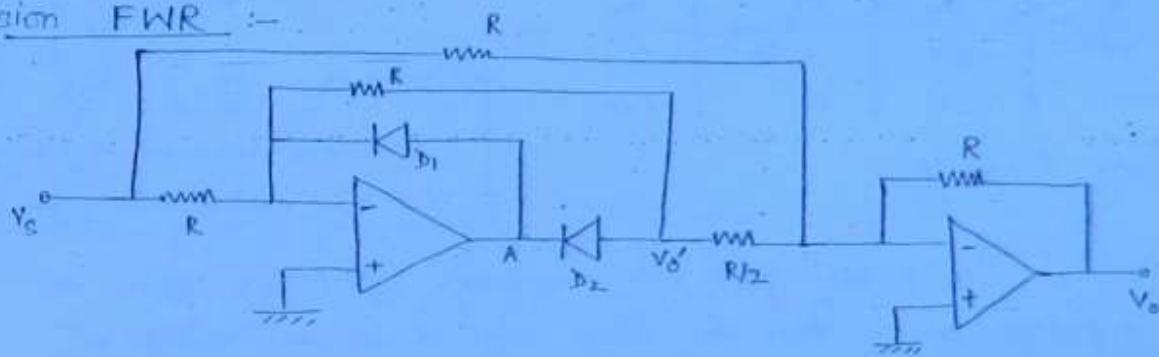
Transfer Curve :-



(Q3)



Precision FWR :-



$$V_o = \frac{-R}{R/2} \cdot V_o' - \frac{R}{R} V_S$$

$$\Rightarrow V_o = -2V_o' - V_S \quad \text{(1)}$$

	$V_S$	$V_A$	$D_1$	$D_2$	$V_o'$
$V_S > 0$		$\downarrow -V_{cat}$ (-0.7V)	OFF	ON	$-V_S$
$V_S < 0$		$\uparrow +V_{cat}$ (0.7V)	ON	OFF	0

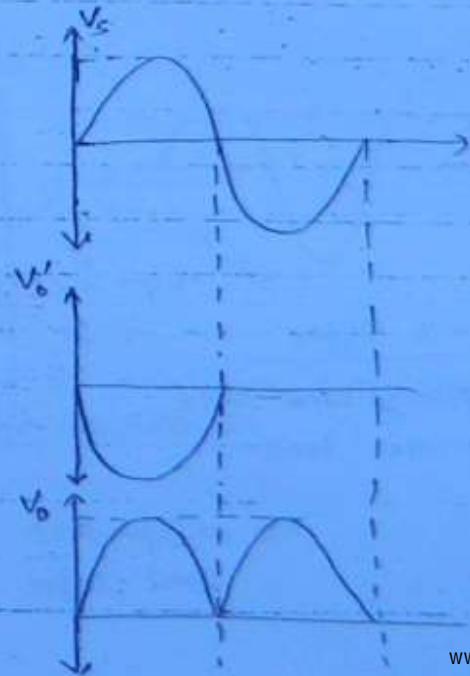
for +ve half -

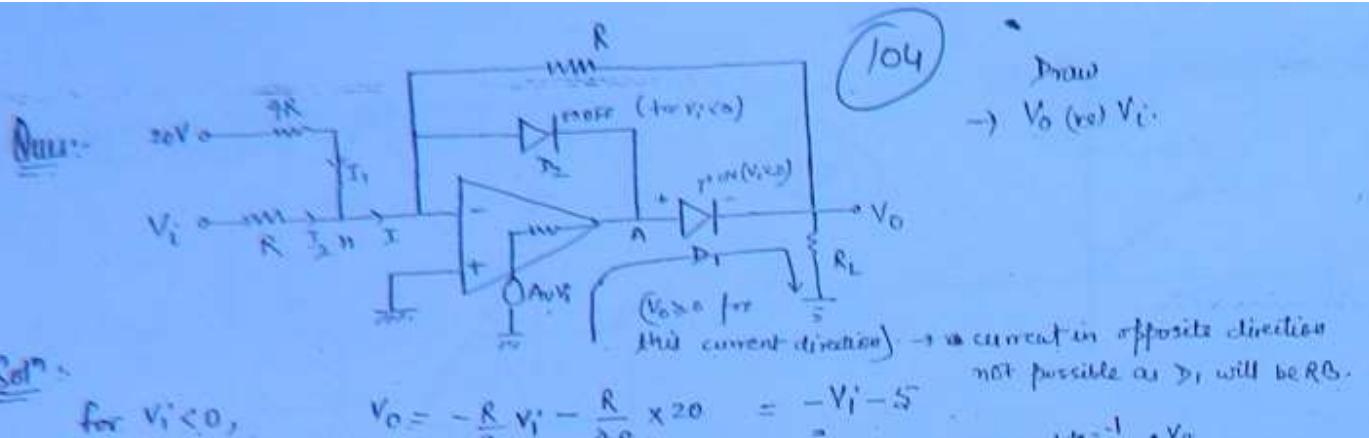
$$V_S > 0 = 5 \text{ (let)} \therefore V_o' = -5V$$

$$\therefore V_o = -2(-5) - 5 = 5V \quad (\text{from eqn (1)})$$

for -ve half -

$$V_S < 0, \quad V_o' = 0 \Rightarrow V_o = 5V$$





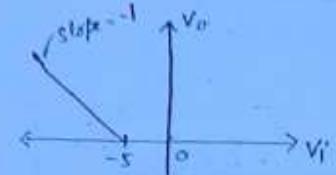
Soln :-

for  $V_i < 0$ ,  
( $D_1 = \text{ON}, D_2 = \text{OFF}$ )

$$V_o = -\frac{R}{R} V_i - \frac{R}{4R} \times 20 = -V_i - 5$$

but,  $V_o \geq 0 \Rightarrow -V_i - 5 \geq 0 \Rightarrow V_i \leq -5$

$V_i$	$D_1$	$D_2$	$V_o$	$V_m$
$V_i \leq -5$	OFF	ON	$-V_i - 5$	$\geq 0$
$V_i > -5$	ON	OFF	0	$< 0$



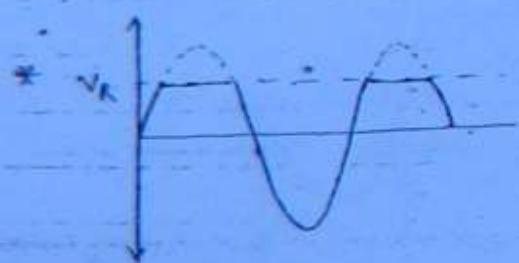
$$\begin{aligned} I &= I_1 + I_2 = \frac{20}{4R} + \frac{V_i}{R} = \frac{V_i + 5}{R} \\ I &> 0 \Rightarrow V_i = \text{effectively +ve} \Rightarrow V_o = \text{+ve} \\ I &< 0 \Rightarrow V_i = \text{effectively -ve} \Rightarrow V_o = \text{-ve} \end{aligned}$$

### Clippers / Limiting Circuits :-

- These are used to select that part of waveform which lies above or below some reference level. These are also referred to as voltage or current limiters, amplitude selectors or slicers.

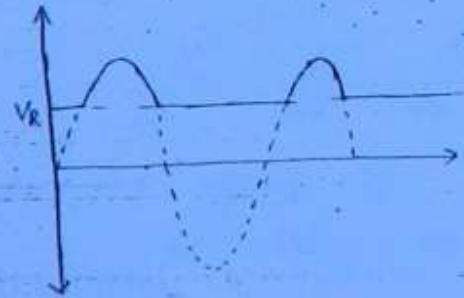
- There are of two types (according to the position of diode w.r.t. load) -

- (a) Series clipper
- (b) Shunt clipper



+ve clipper

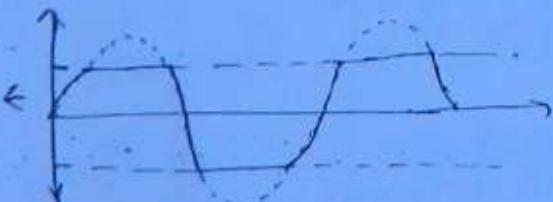
→ clipping above some reference level.



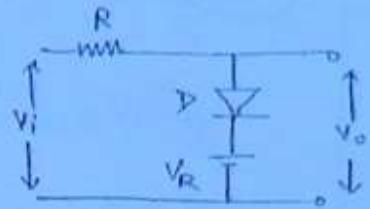
-ve clipper

→ clipping below some reference level.

Two independent level clipper

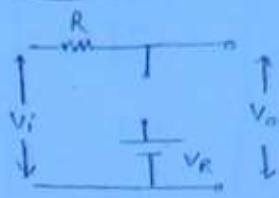


28<sup>th</sup> August, 2012

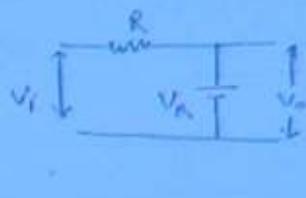


Shunt Clipper (Positive)

105

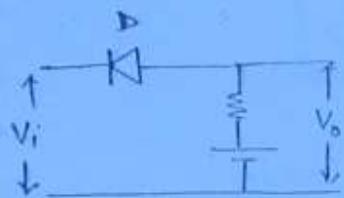


D → RB  
( $V_R > V_i$ )



D → FB  
( $V_R < V_i$ )

$V_i$	D	$V_o$
$V_i \leq V_R$	OFF	$V_i$
$V_i \geq V_R$	ON	$V_R$



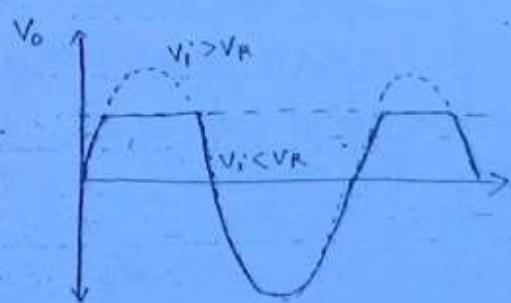
Series Clipper (positive)

D → RB  
( $V_i > V_R$ )

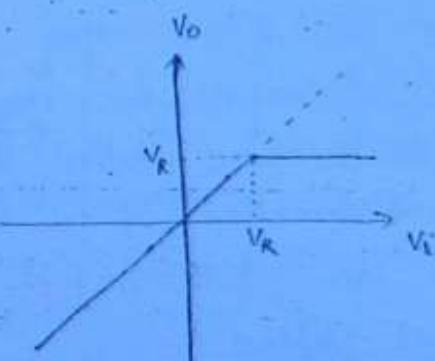


D → FB  
( $V_i < V_R$ )

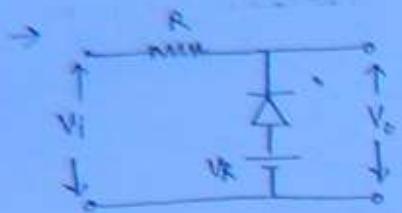
$V_i$	D	$V_o$
$V_i \leq V_R$	ON	$V_i$
$V_i \geq V_R$	OFF	$V_R$



O/P



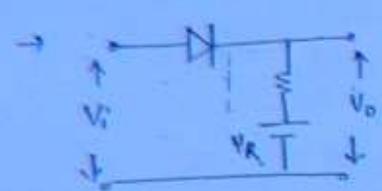
Transfer Characteristic



$V_i$	D	$V_o$
$V_i \geq V_R$	OFF	$V_i$
$V_i \leq V_R$	ON	$V_R$

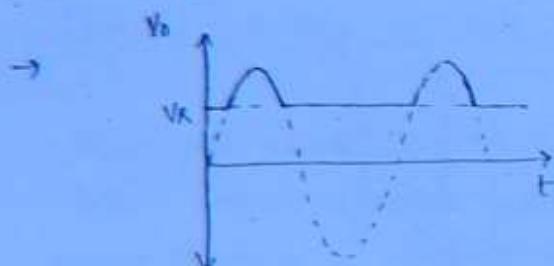
(106)

Shunt clipper (negative)

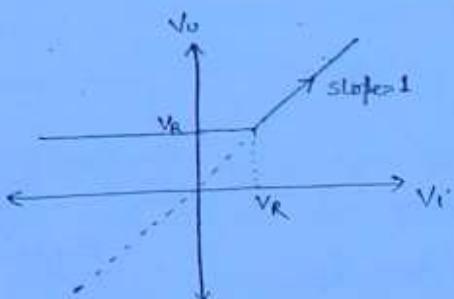


$V_i$	D	$V_o$
$V_i > V_R$	ON	$V_i$
$V_i \leq V_R$	OFF	$V_R$

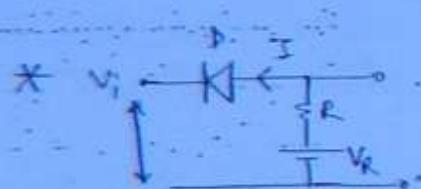
Series Negative Clipper



O/P curve



Transfer characteristic

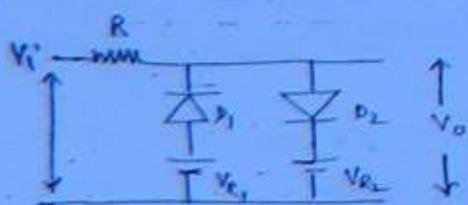


$$I = \frac{V_i - V_R}{R}$$

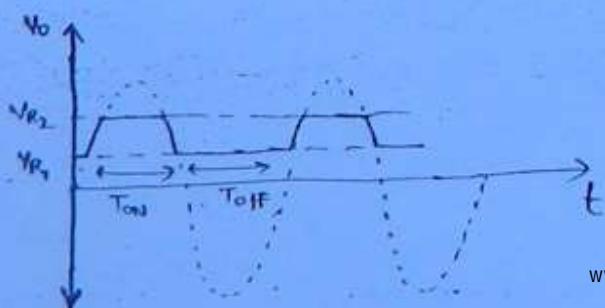
$I \geq 0 \rightarrow D \rightarrow \text{ON}$ ;  $I < 0 \rightarrow D \rightarrow \text{OFF}$   
 $\Rightarrow V_i \geq V_R \rightarrow D \rightarrow \text{ON}$        $\Rightarrow V_i < V_R \rightarrow D \rightarrow \text{OFF}$

→ It is easiest way to determine whether diode is ON or OFF. Calculate the current in forward direction of diode and apply the condition.

→ Two Independent level clipper :-



$(V_{R2} > V_{R1})$



→ Range of  $V_i$

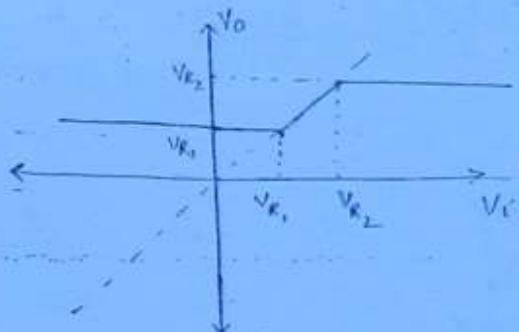
$$V_i \leq V_{R_1}$$

$$V_{R_1} \leq V_i \leq V_{R_2}$$

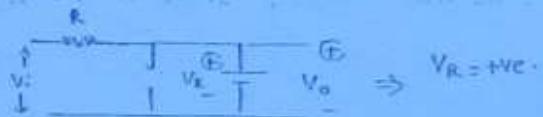
$$V_i \geq V_{R_2}$$

$V_i$	$D_1$	$D_2$	$V_o$
$V_i \leq V_{R_1}$	ON	OFF	$\frac{V_o}{V_{R_1}}$
$V_{R_1} \leq V_i \leq V_{R_2}$	OFF	OFF	$V_i$
$V_i \geq V_{R_2}$	OFF	ON	$\frac{V_o}{V_{R_2}}$

(107)



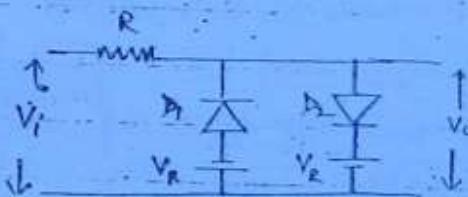
→ To conclude that  $V_{R_1}$  &  $V_{R_2}$  are +ve, check polarity at  $V_o$  whenever  $D_1$  is  $V_{R_1}$ . If same polarity then  $V_{R_1}$  = +ve, else -ve.



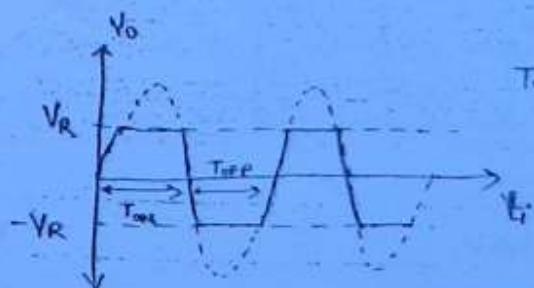
→ From op curve,  $T_{off} > T_{on} \Rightarrow D < 50\%$ . Output is an asymmetrical square wave.

→ This circuit is used as a means of converting a sinusoidal waveform into a square wave.

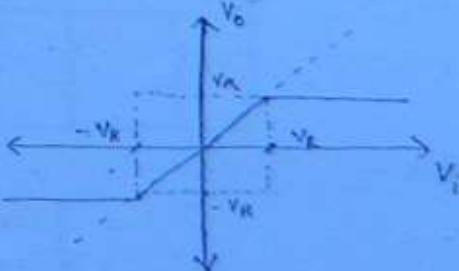
→ To generate a symmetrical square wave,  $V_{R_1}$  and  $V_{R_2}$  are adjusted to be numerically equal but are of opposite sign.



$V_i$	$D_1$	$D_2$	$V_o$
$V_i \leq -V_R$	ON	OFF	$-V_R$
$-V_R \leq V_i \leq V_R$	OFF	OFF	$V_i$
$V_i \geq V_R$	OFF	ON	$V_R$



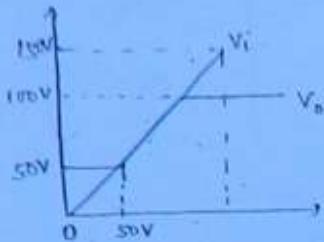
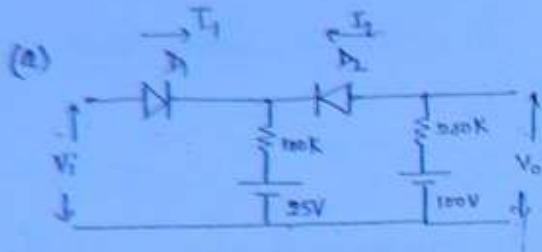
→ Symmetrical Sq. wave / Symmetrical Clipping



Transfer characteristic

Ques: (1) The i/p voltage  $V_i$  to the two level clipper shown in fig. varies linearly from 0 to 150V. Sketch the o/p voltage  $V_o$ . to the same time scale as the i/p voltage. Assume ideal diodes.

(108)



<u>Range of <math>V_i</math></u>	$D_1$	$D_2$	$V_o$
$0 \leq V_i \leq 50$	OFF	ON	50V
$50 \leq V_i \leq 100$	ON	ON	$V_i$
$V_i \geq 100$	ON	OFF	100V

$$\rightarrow I_1 = \frac{V_i - 25}{100k} + \frac{V_i - 100}{200k}$$

$$\text{For } D_1 = \text{ON}, \quad I_1 \geq 0 \Rightarrow \frac{V_i - 25}{100k} + \frac{V_i - 100}{200k} \geq 0$$

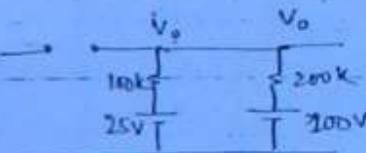
$$\Rightarrow V_i \geq 50$$

$$\rightarrow I_2 = \frac{-V_i + 100}{200k}, \text{ for } D_2 = \text{ON} -$$

$$I_2 \geq 0$$

$$\Rightarrow V_i \leq 100$$

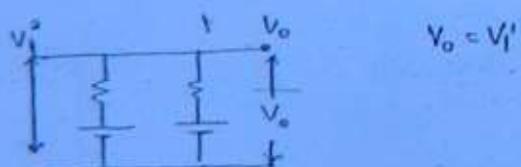
$\rightarrow$  When  $0 \leq V_i \leq 50$ ,



$$\frac{V_o - 100}{200k} + \frac{V_o - 25}{100k} = 0$$

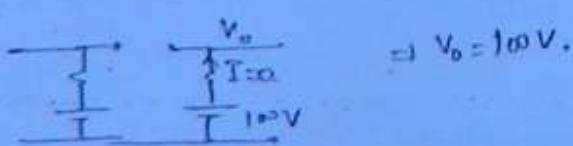
$$\Rightarrow V_o = 50V$$

$\rightarrow$  When  $50 \leq V_i \leq 100$

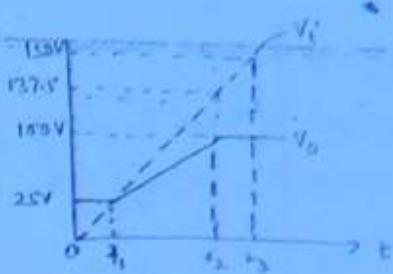
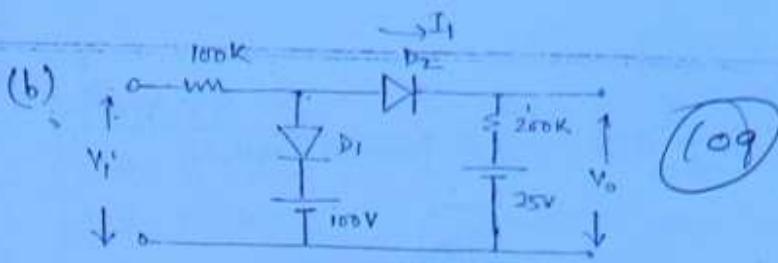


$$V_o = V_i$$

$\rightarrow$  When  $V_i \geq 100$  -



$$\Rightarrow V_o = 100V$$



Soln : Since voltage across  $D_1$  is very high (100V), then  $D_2$  will ON before  $D_1$ .

$$\therefore I_1 = \frac{V_i - 25}{300\text{K}} \quad \text{for } D_2 = \text{ON}$$

(Assuming  $D_2$  OFF)

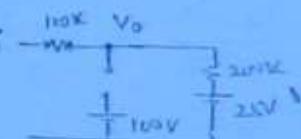
$$I_1 > 0 \Rightarrow V_i > 25 \rightarrow \text{(Breakpoint.)}$$

$V_i$	$D_1$	$D_2$	$V_o$
$0 < V_i \leq 25$	OFF	OFF	25V
$25 \leq V_i \leq 137.5$	OFF	ON	$\frac{2V_i + 25}{3}$
$V_i \geq 137.5$	ON	ON	100V

for  $V_i > 25$ ,  $V_o$  will be -

$$\frac{V_o - 25}{2.5\text{K}} + \frac{V_o - V_i}{100} = 0$$

$$\Rightarrow V_o = \frac{2V_i + 25}{3}$$



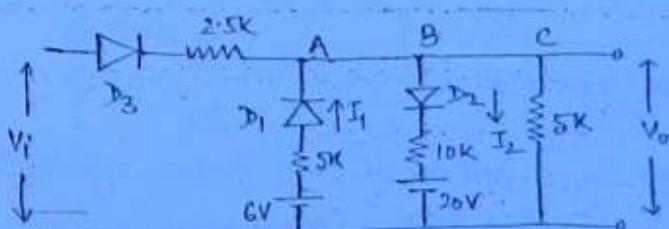
Now for  $D_2 = \text{ON}$  -

$$V_o \geq 100V$$

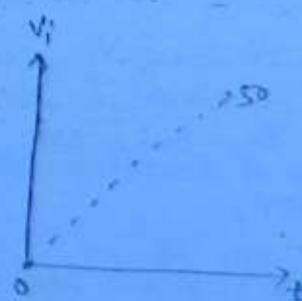
$$\Rightarrow \frac{2V_i + 25}{3} \geq 100 \Rightarrow V_i \geq 137.5V$$

L (Breakpoint)

Ques:- Assume that the diodes are ideal, make a plot of  $V_o$  vs  $V_i$  for the range of  $V_i$  from 0 to 50V. Indicate all slopes and voltage levels. Indicate for each region, which diodes are conducting.



Soln



Ques: When  $V_i = 0$ ,

voltage across  $D_2 = \infty$  is large ( $\approx 20V$ ) and  $A =$  forward biased.  
Due to  $D_1$ , current voltage at  $A$ ,

(110)

$$V_A = 3V.$$

Now, this  $V_A$  is making diode  $D_3$  RB.



Now, when  $V_i > 3V$  then  $D_1 = ON$ .  
↳ Break point.

$$\text{for } 0 \leq V_i \leq 3, V_o = V_A = 3V$$

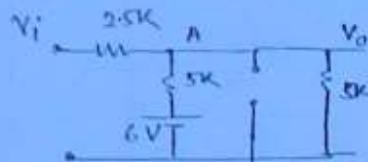
for  $V_i > 3$ ,

$D_1 = ON, D_2 = OFF, D_3 = ON$ .

$$\frac{V_A - 6}{2.5K} + \frac{V_A - 6}{5} + \frac{V_A}{5} = 0$$

$$\Rightarrow \frac{4V_A}{5} = \frac{2V_i}{5} + \frac{6}{5}$$

$$\Rightarrow V_A = \frac{V_i + 3}{2}$$



Now diode  $D_1$  will remain in FB till

$$\left. \begin{array}{l} V_A < 6 \\ \Rightarrow \frac{V_i + 3}{2} < 6 \end{array} \right\}$$

$$\Rightarrow V_i < 9V \rightarrow \text{Break point.}$$

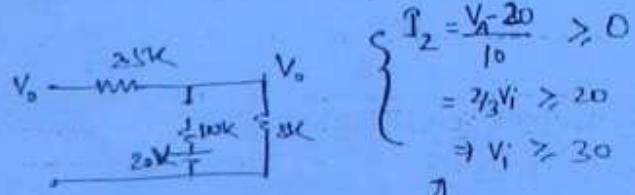
$$\left. \begin{array}{l} I_A = \frac{V_A - 6}{5} > 0 \text{ for } D_1 = ON \\ \Rightarrow V_A > 6 \end{array} \right\}$$

$$\text{for } 3 \leq V_i \leq 9; V_o = V_A = \frac{V_i + 3}{2}$$

for  $V_i > 9V$ ,

$D_1 = OFF, D_2 = OFF, D_3 = ON$ .

$$V_o = \frac{5}{7.5} V_i = \frac{2}{3} V_i$$



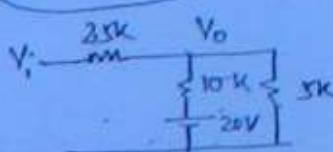
Now for  $D_2$  to be ON,

$$\left( \frac{2}{3} V_i \geq 20V \right) \Rightarrow V_i \geq 30V \rightarrow \text{Break point.}$$

for  $V_i \geq 30V$ ,

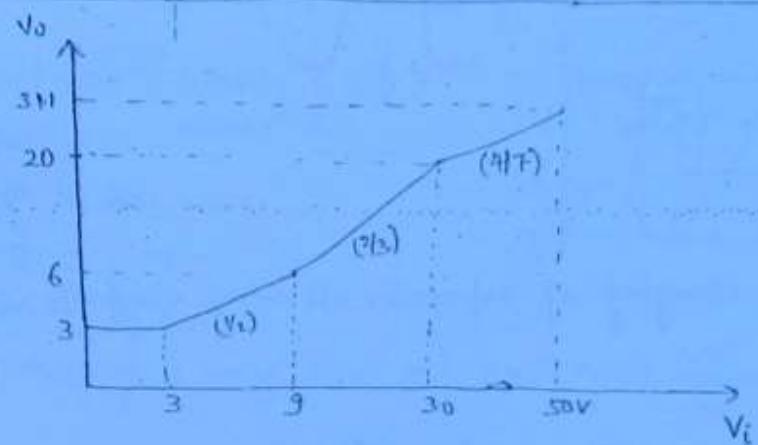
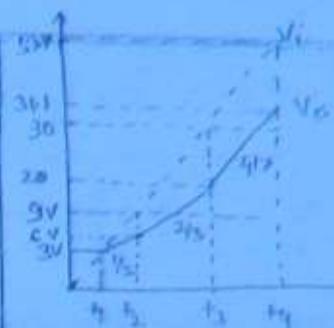
on applying KCL,

$$V_o = \frac{4V_i + 20}{7}$$

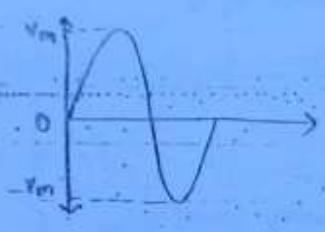


Range of  $V_i$

	$D_1$	$D_2$	$D_3$	$V_o$
$0 \leq V_i \leq 3$	ON	OFF	OFF	3V
$3 \leq V_i \leq 9$	ON	OFF	ON	$\frac{V_i+3}{2}$
$9 \leq V_i \leq 30$	OFF	OFF	ON	$\frac{2}{3}V_i$
$30 \leq V_i \leq 50$	OFF	ON	ON	$\frac{4}{7}V_i + \frac{20}{7}$



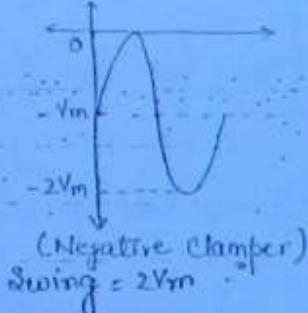
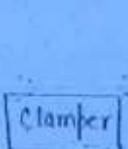
### Clamper Circuit :-



$$\text{swing} = 2V_m$$

$$f = 50 \text{ Hz}$$

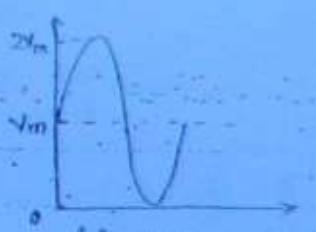
$$\text{dc level} = 0$$



(Negative Clamper)  
swing =  $2V_m$

$$f = 50 \text{ Hz}$$

$$\text{dc level} = -V_m$$



(Positive Clamper)  
swing =  $2V_m$

$$f = 50 \text{ Hz}$$

$$\text{dc level} = V_m$$

→ Clamper circuits are also called as dc translator, dc restorer, dc inserter.

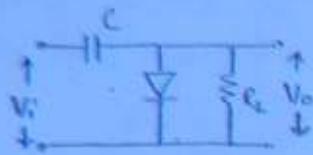
→ The circuit which are used to add a dc level as per the requirements to ac off signal are called clamper circuit.

→ These are of two types — Negative Clamper → adds  $-ve$  level to ac off signal  
Positive clamper → adds  $+ve$  level.

20<sup>th</sup> August, 2012

### Negative Clamper

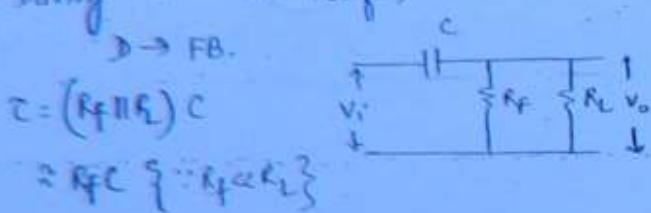
(112)



Initially,  $V_C(0^-) = V_C(0^+) = 0$ .

hence, C will act as S.C.

During 1<sup>st</sup> half,

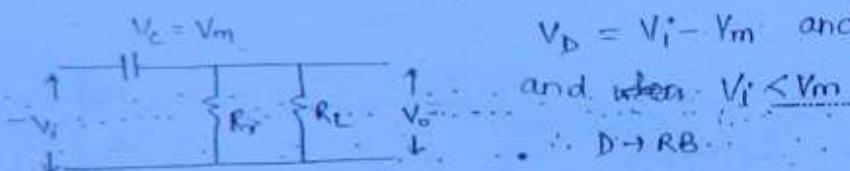


$\because R_F C \ll T$ . Hence, rate of charging of capacitor is very high. It will charge till maximum value  $V_m$ .

$\rightarrow$  At  $t = T/4$ ,  $V_i = V_m$  &  $V_C = V_m$ .

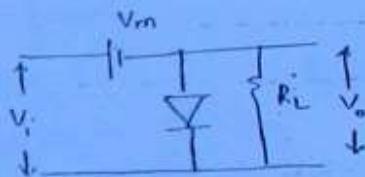
$\therefore D \rightarrow ON$ ,  $R_F \approx 0$  and hence,  $V_D \approx 0$ .  $\therefore V_o = V_D = 0$ .

$\rightarrow$  for  $t > T/4$ ,



$\tau = (R_F \parallel R_L) C \approx R_L C \gg T \Rightarrow$  Rate of discharging is very small  $\times 0$ .

$V_i$	$V_o = V_i - V_m$
$V_m$	0
0	$-V_m$
$-V_m$	$-2V_m$

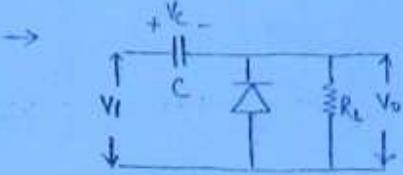


$\rightarrow$  Diode will remain off. after this instant.

- \* We cannot use a dc battery instead of capacitor as the value of battery will vary w.r.t the peak value of signal.

→ Once the capacitor is charged till  $V_m$ , it will act as battery of value  $V_m$

Positive Clamper: It adds positive dc level to ac off.



→  $V_C(0) = V_c(0^+) = 0V \rightarrow$  initially C will act as S.C.

During 1<sup>st</sup> +ve half-

at  $t=0$  S.C.

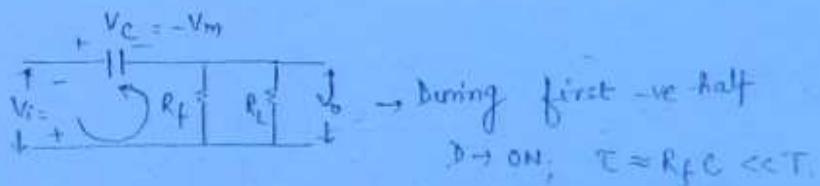
→ Diode D will be R.B.



$$\tau = R_L C \gg T$$

∴ Rate of charging is very low and C will remain uncharged till  $t=T/2$ .

→ for  $t > T/2$  -



→ During first -ve half  
D → ON,  $\tau \approx R_F C \ll T$ .

Now, the rate of charging is very high and will charge till  $t=3T/4$ .

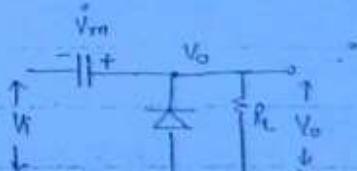
→ At  $t=3T/4$ ,  $V_i = -V_m$ , and  $V_C = -V_m$ .

→ for  $t \geq 3T/4$  -

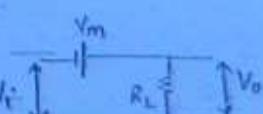
$V_i$	$V_o = V_i + V_m$
$-V_m$	0V
0	$V_m$
$V_m$	$2V_m$

$$V_o = V_i + V_m$$

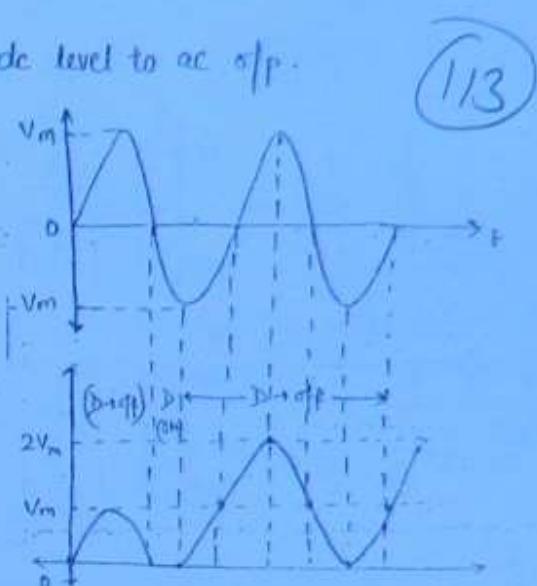
∴ Voltage across D  
is always -ve, hence D → off.  
( $V_o = +ve$ )



and ∵ D is off, C will never discharge.

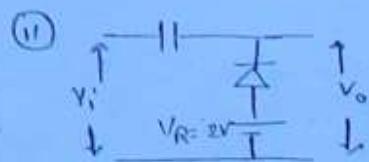
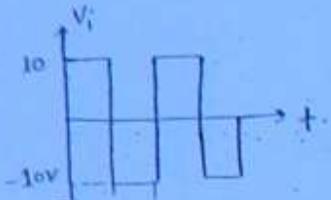
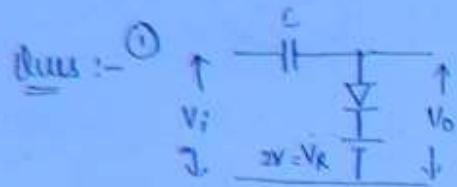


→  $R_{L\min} = \sqrt{R_F R_S}$  - for proper functioning of clamper



→ During first negative half, capacitor gets charged upto  $-V_m$  through FB diode D. The capacitor once charged to  $-V_m$ , will act as a battery of  $-V_m$  and therefore  $V_o = V_i + V_m$ .

(114)

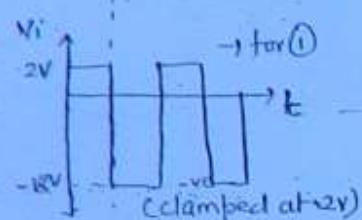


Soln

$$V_i = 10 \text{ V}$$

$$V_o = 10 - 2 = 8 \text{ V}$$

$$V_R = 2 \text{ V}$$



$$V_C = -10 - 2 = -12 \text{ V}$$

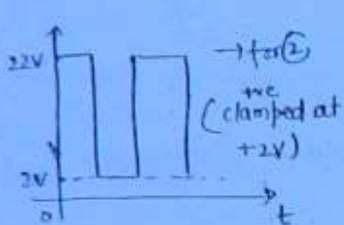
$$10 \text{ V}$$

$$2 \text{ V}$$

$$V_i = 8 \text{ V}$$

$$V_o = V_i - 8$$

$$V_R = 2 \text{ V}$$



$$V_C = -12 \text{ V}$$

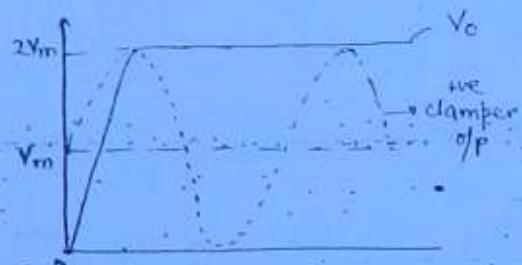
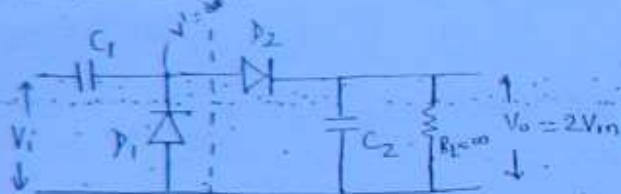
$$1 \text{ V}$$

$$-2 \text{ V}$$

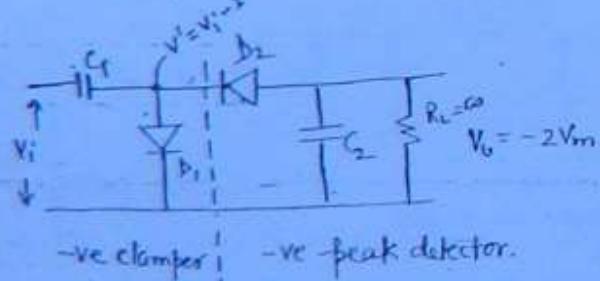
$$V_o = V_i - V_C$$

$$= V_i + 12$$

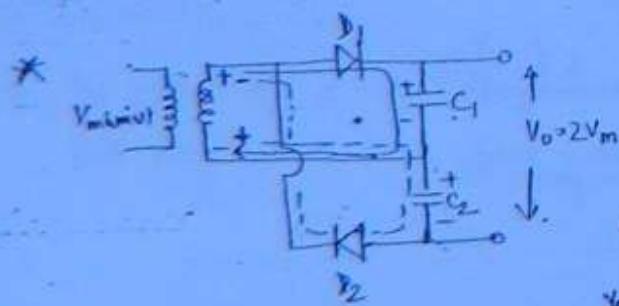
### Voltage Doubler :-



+ve clumper | +ve peak detector



-ve clumper | -ve peak detector.



$C_1$  = will charge till  $+V_m$ . during +ve half

$C_2$  = will charge till  $+V_m$  during -ve half.

$$V_p \rightarrow D_1 \rightarrow V_n \quad V_D = V_p - V_n = V_p - V_m \geq 0 \text{ for FB}$$

Not possible  $\Rightarrow V_p > V_m$  for FB of

$$V_p \rightarrow D_2 \rightarrow V_n \quad V_D = V_p - V_n \text{ hence } D_1, D_2 \text{ will remain off.}$$

$$= -V_m - V_n \geq 0 \Rightarrow V_n \leq -V_m \text{ for FB.}$$

## Workbook

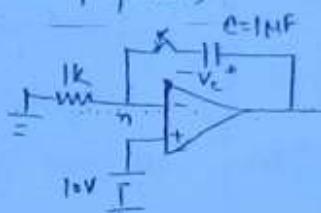
Pg. 27

(1)	$V_i$	$D_1$	$D_2$	$V_o$	(2) d
	$V_i < 0$	OFF	OFF	0	(10) b
	$0 \leq V_i \leq 20$	ON	OFF	$V_i/2$	(11) a
	$V_i > 20$	ON	ON	10V	(12) d

11B

## Chapter 10 Pg. 56

Q7 c, a (c preferred)



$$V_n = 10V$$

$$V_o = 10 + V_C$$

for  $V_C$  = values in option -

$$\therefore V_o = 10V + 0 = 10V$$

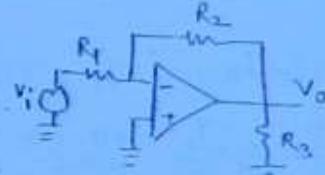
: Ans 0V  $\rightarrow$  (a).

$$\left. \begin{array}{l} V_o = 16.3V \\ V_o = 19.5V \\ V_o = 20V \end{array} \right\} \times \text{cannot exceed } 15V$$

Q18 C Q22 C Q23 b

Q25 (a)

$$\frac{0 - V_i}{R_1} + \frac{0 - V_o}{R_2} = 0 \Rightarrow V_o = -\frac{R_2}{R_1} V_i$$



## Conventional

Q2 When switch is ON, gain = 1; S  $\rightarrow$  OFF, gain = -2.

$$(3) R_1 = 3k\Omega, R_2 = 2k\Omega$$

R = dc resistance seen by inverting terminal;  $R = R_1 \parallel R_2 \parallel 6k = 1k$

30<sup>th</sup> August, 2012.

## Bipolar Junction Transistor :-

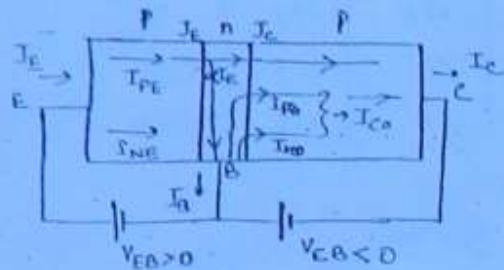
(116)

p-n-p transistor in active mode :-

$$I_E = I_{PE} + I_{NE} \approx I_{PE}$$

$\downarrow$  holes       $\downarrow e^-$

{ Minority carrier injection }  
(Diffusion)



$$I_{CO} = I_{PCO} + I_{NCO}$$

{ Majority carriers inf. }  
(Drift)

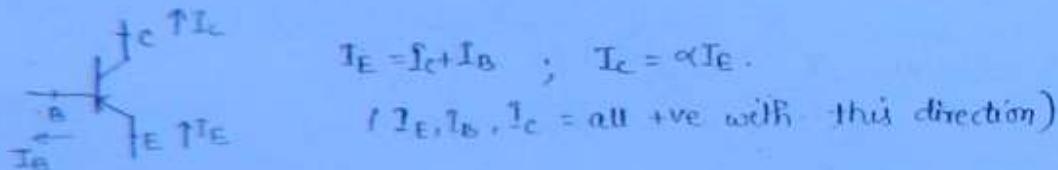
$$\bar{I}_C = \alpha (I_E + I_{CO})$$

;  $\alpha$  = large signal current gain or  $\alpha_{dc}$ .

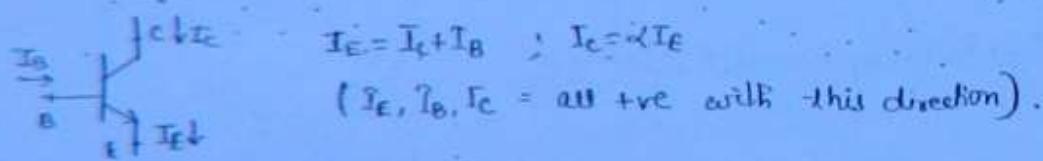
(valid only for active current region)

$$\alpha \text{ or } \alpha'_{dc} = \frac{I_C - I_{CO}}{I_E} \text{ or } \Rightarrow \boxed{\alpha \approx \frac{I_C}{I_E}}$$

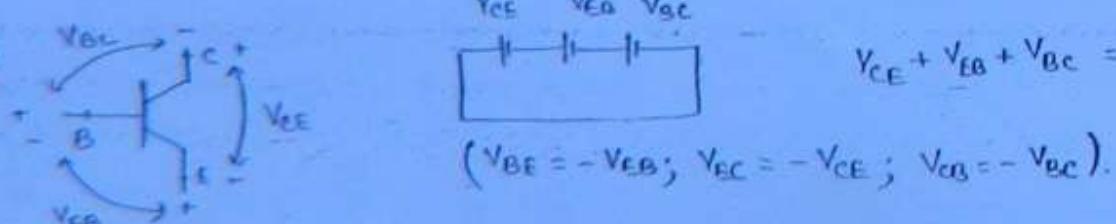
for CB configuration.



p-n-p



n-p-n



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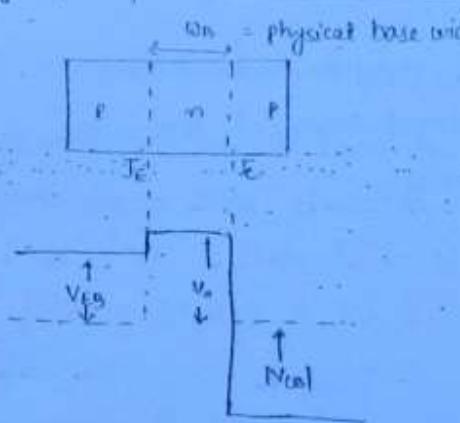
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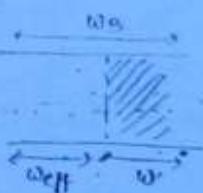
Note:- Guys Be Cool Dude I am here for help You 😊

	Common Emitter	(Emitter Follower) Common Collector	Common Base
1) Input Terminal	B	B	E
2) Output Terminal	C	E (117)	C
3) Common Terminal	E	C	B
4) Current Gain, $A_I$	high (moderate)	very high.	very low ( $< 1$ )
5) Voltage Gain, $A_V$	high (moderate)	very low ( $< 1$ )	very high
6) Input Resistance, $R_I$	high (moderate)	very high	very low
7) Output Resistance, $R_O$	high (moderate)	very low	very high
8) Power Gain $(A_P = A_V \cdot A_I)$	Highest	Moderate	Moderate
9) Phase Shift	$180^\circ$	$0^\circ$	$0^\circ$
10) Normally used as	Amplifier in multistage	Buffer (voltage).	High freq. applic. Buffer (current)

Early effect / Base width Modulation -



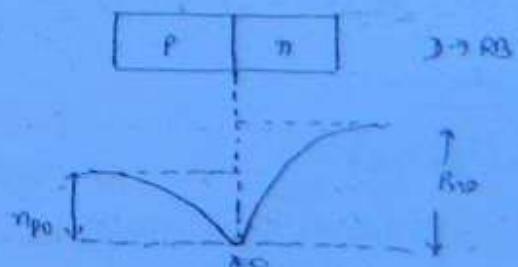
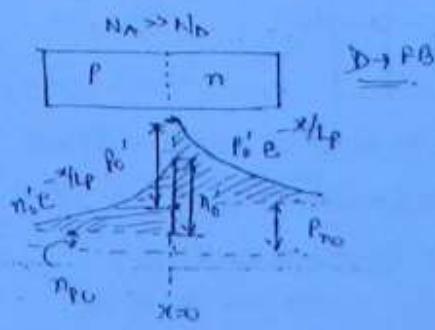
When collector junction is RB -



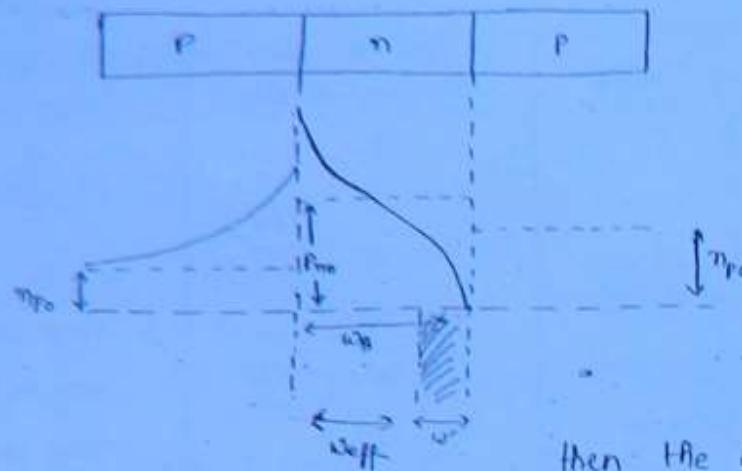
$$w_{eff} = \text{effective base width} = w_B - w'$$

$w'$  = depletion width.

As  $|V_{CB}| \uparrow$ ;  $J_E$  becomes more RB and  $w_{eff} \downarrow$ , therefore recombination and hence  $I_{B\downarrow}$  and  $\propto \uparrow$ .



→ 118

 $I_E = \text{diff current}$ 

$$I_E = q D_P \frac{dp}{dx}$$

$$\frac{dI_E}{dx} \propto \frac{dp}{w_{eff}}$$

Now, due to early effect,  $w_{eff} \downarrow$  and  $I_E \uparrow$ . When  $w_{eff} = 0$ , then the condition is called reach through or punch through.  $\Rightarrow I_E$  will be very large.

→ The variation of effective base width with  $|V_{CB}|$  is called base width modulation or early effect. This results in following-

- There is less chance of recombination in Base region as effective base width reduced. Therefore,  $\alpha \uparrow$  causing an  $\uparrow$  in collector current  $I_C$ .
- conc gradient of injected holes (minority carriers in base region) also  $\uparrow$  due to reduced base width. Since, diffusion current is directly proportional to conc gradient,  $I_E$  also  $\uparrow$ .
- for large value of  $V_{CB}$ , effective base width may be reduced to 0 causing extremely large  $I_E$ . This result in breakdown of transistor and is called punch through or reach through.

### Input and Output characteristics :-

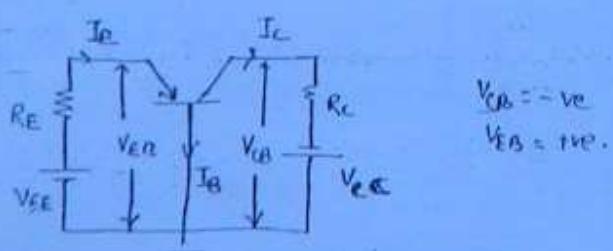
#### Common Base configuration :-

##### if characteristic —

$$V_{EB} = f_1(I_E, V_{CB})$$

##### of characteristic —

$$\nabla I_C = f_2(I_E, V_{CB})$$



$$V_{CB} = -ve \\ V_{EB} = +ve.$$

## Output characteristic

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

(119)

### Cutoff Region

$$I_E = 0; I_C = I_{CO}$$

$$\& I_C + I_B = I_E \Rightarrow I_B = -I_{CO}$$

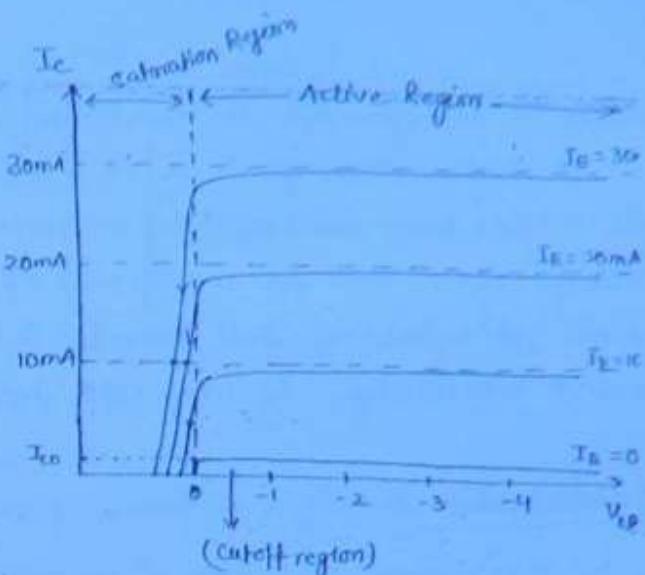
### Active Region

$$I_C = \alpha I_E \quad \& \quad \alpha < 1 \Rightarrow I_C < I_E$$

Now, as  $V_{BE} \uparrow$ , junction is more RB.

&  $\alpha \uparrow$  due to early effect. (but the effect is very small).

In this region,  $I_C \propto I_E$  and is almost constant of  $V_{CE}$  variation, hence transistor in this config can be used as CCBS.



### Saturation Region

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

$$\frac{dI_C}{dV_{CE}} = \frac{\alpha I_E}{I_{CO} [e^{V_{CE}/V_T} - 1]}$$

$V_{CE} = +ve$  = since collector junction is RB.

(1)  $\alpha I_E \approx$  almost constant ; Hence, as  $V_{CE} \uparrow$  (+ve value),  $I'$  will  $\uparrow$  and hence  $I_C$  will decrease as  $V_{CE} \uparrow$  towards +ve value for constant o/p current.

(2) When we are from circuit,  $V_{CE} = -V_{CC} + I_C R_C$ .

If we  $\uparrow I_E$ , then  $I_C \uparrow \Rightarrow I_C R_C \uparrow \Rightarrow V_{CE}$  will move towards +ve.

When  $|I_C R_C| > |V_{CC}|$ , then  $V_{CE} = +ve$ .  $\Rightarrow$  junction  $T_C \in FB$ .

After this,  $(\alpha I_E) \uparrow$

- when  $I_E \uparrow$ ,  $\Rightarrow I_C \uparrow \Rightarrow V_{CE} \uparrow \Rightarrow I' \uparrow$  and  $I_C = \alpha I_E + I' \approx$  constant hence o/p current will not change after this and the transistor will move into saturation.

## Important Points :-

- Cutoff  $\rightarrow$  Region below  $I_E = 0$ .

- Active  $\rightarrow$   $I_C$  is <sup>almost</sup> independent of o/p voltage  $V_{CE}$ .

- o/p characteristic of CCBS is called constant current characteristic.

- It is a CCBS.

- Saturation  $\rightarrow$  The region left to  $V_{CB} = 0$  and above  $T_E = 0$ .

(120)

$\rightarrow$  As the collector junction is pB, the holes flow from p-type collector towards n-type base and constitute a current  $I'$  in a direction opposite to direction of  $+I_C$ . Even for small value of  $+V_{CB}$ , large change in  $I_C$  take place and characteristics fall towards 0 as  $V_{CB}$  is made more & more +ve. Since  $I' \uparrow$  exponentially,  $I_C$  may even become -ve.

Input Characteristics :-

$$V_{EB} = f_1(I_E, V_{CB}).$$

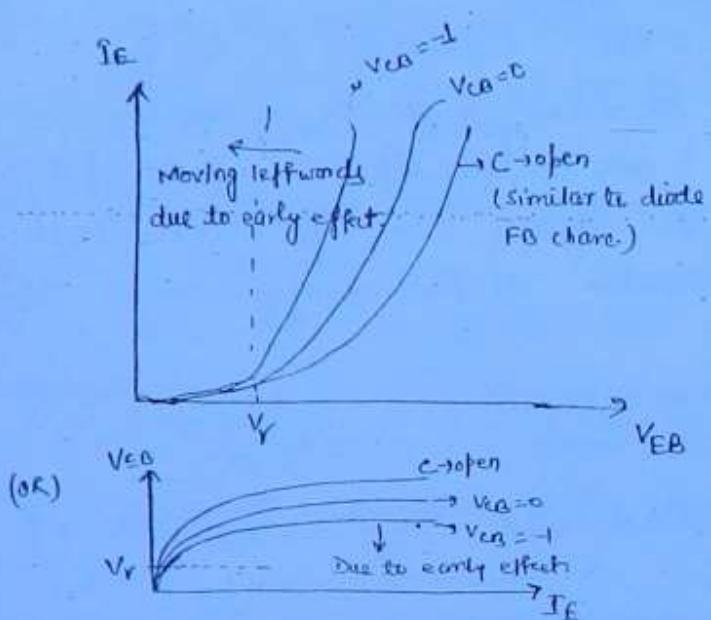
When  $V_{CB} = \text{ext. } \infty$ ; i.e., C-B  $\rightarrow$  O.C.

then there will be similar to diode.

When  $V_{CB} = 0$ ,  $I_C = \text{slightly } R_B$ .

Due to early effect,  $I_E \uparrow$  more rapidly.

Now, As  $V_{CB} \uparrow$ , more early effect



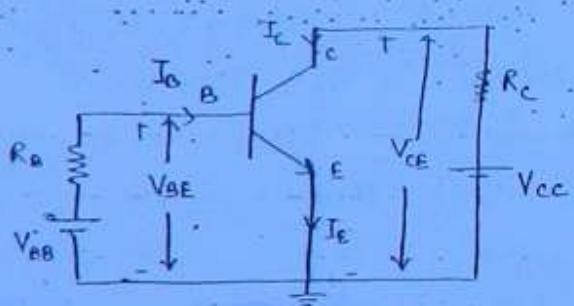
## II) COMMON Emitter Configuration :-

i/p charc -

$$V_{EE} = f_1(I_B, V_{CE}).$$

o/p charc -

$$\nabla I_C = f_2(I_B, V_{CE}).$$



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

$$\Rightarrow I_E = \alpha (I_C + I_B) + I_{CO}$$

$$\therefore I_E = \left( \frac{\alpha}{1-\alpha} \right) I_B + \frac{I_{CO}}{(1-\alpha)}$$

or

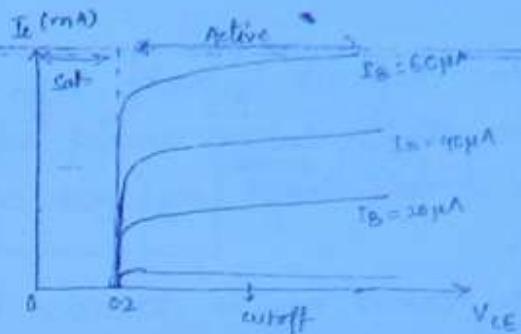
$$I_E = \beta I_B + (\beta + 1) I_{CO}$$

$$\beta = \frac{\alpha}{1-\alpha}$$

$\Rightarrow \beta \gg \alpha$

## Output Characteristic

(1/2)



## Active Region

- $I_c = \beta I_B + (\beta+1) I_{CO} \approx \beta I_B$ .

- $V_{CE} = V_C - V_E = V_C$ .

- As  $V_{CE} \uparrow \Rightarrow V_C \uparrow \rightarrow I_c$  more RB  $\Rightarrow$  width  $\downarrow$  &  $\alpha \uparrow$

- Now,  $\alpha \rightarrow 0.98 \rightarrow 0.985 \rightarrow 0.5\% \uparrow$

- $\beta \rightarrow 49 \rightarrow 63 \rightarrow 34\% \uparrow \rightarrow$  can't be neglected.

- $\rightarrow$  current gain is high, i.e., small change in  $I_B$  results in large change in  $I_c$

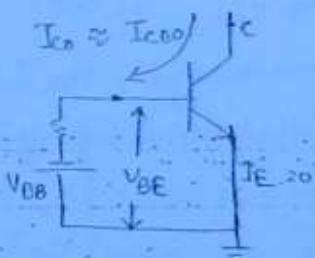
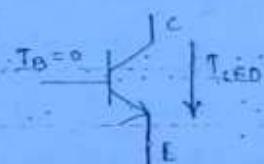
## Cutoff Region

- When  $I_B = 0$ ,  $I_c = (\beta+1) I_{CO}$   $\rightarrow I_c \neq I_{CO}$  and transistor is not in cut off.

- $I_{CEO} = (\beta+1) I_{CO} = \frac{I_{CO}}{\alpha-\alpha} \quad [I_{CEO} \gg I_{CO}], \because \beta \gg 1$ .

- Now, for  $I_c = I_{CO}$ ,  $I_E$  should be 0.

- When  $I_E = 0$



$$I_{CEO} \gg I_{CO} > I_{CO}$$

- The  $I_c$  in a physical (real) non-idealised device when  $I_E = 0$  is designated by symbol  $I_{CO}$ .

- Cutoff is defined as a condition where  $I_c = I_{CO}$  and  $I_E = 0$ . In order to cutoff transistor, it is not enough to reduce  $I_B \approx 0$ , instead it is necessary to reverse bias the emitter junction slightly, i.e.,  $V_{BE} = -V_E$ .

- $V_{BE} = -0.1$  for Ge ; 0.0 for Si.

- The actual  $I_c$  with collector junction RB & base open circuit is designated by symbol  $I_{CEO}$ .

- $I_{CEO} =$  reverse collector saturation current.

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- Two factors co-operate to make  $I_{CBO}$  larger than  $I_{CB}$ .
  - a) There exist a leakage current which flows ~~across~~ through junction, but around it and through surfaces and it is  $\propto$  to voltage across the junction.
  - b) Now carrier may be generated by collision in  $T_c$  transition region leading to avalanche multiplication of current.
- $I_{CBO} = nA$  for Ge  
 $nA$  for Si.
- $I_{CBO}$  approximately doubles for every  $10^\circ$  rise in temp for both Ge & Si and Si can be used upto about  $200^\circ\text{C}$  and Ge upto about  $150^\circ\text{C}$ .

31<sup>st</sup> August, 2013 :

### Saturation

 $\Rightarrow J_E \rightarrow PB$ ,  $J_C \rightarrow PB$ .

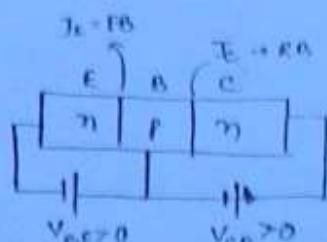
$$\Rightarrow J_E = \beta I_B \quad ; \quad V_{CE} = V_{CE} - T_c k_e \quad (\text{from eqn})$$

When we  $\uparrow V_{BE}$ , then  $J_B \uparrow$  (due to barrier lowering), then  $J_E \uparrow$  and  $V_{CE} \downarrow$ .

When  $V_{BE} = 0.8\text{V}$ ,  $V_{CE} = 0.2\text{V}$  and is constant... for further  $\uparrow$  in  $V_{BE}$ ,

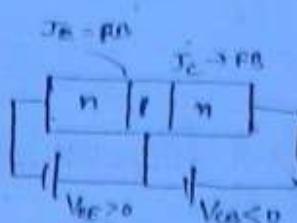
$$\therefore V_{CEsat} = 0.2\text{V}$$

$$I_{Csat} = \frac{V_{CE} - V_{CEsat}}{R_C}$$



$$\text{Now, } V_{CE} + V_{EB} + V_{BC} = 0 \Rightarrow V_{CE} - V_{BE} + V_{EB} = 0$$

$$\therefore V_{EB} = 0.2 - 0.8 = -0.6\text{V}$$



Now,  $J_E = PB$ , and a reverse current will start flowing which will oppose  $J_E$  and the diode transistor will go into saturation

$$\left\{ J_E = J_{E0} [e^{\frac{V_{BE}}{V_{Tsat}}}] \right\}$$

If  $V_{BE}$  is kept constant and  $V_{CE}$  is changed -

when  $V_{CE} \uparrow \Rightarrow V_{CE} \downarrow$

$V_{CE} = V_{CE} - V_{CE}$   $\Rightarrow V_{CE} \downarrow$  and when it is  $-ve$  then  $J_E \rightarrow PB$   
 hence, Transistor will be in saturation.

Checking Transistor for saturation:

\* Let  $\theta$  be in saturation.

Then  $V_{CE} = V_{CEsat}$  &  $V_{BE} = V_{BEsat}$

$$\therefore I_{Csat} = \frac{V_{CE} - V_{CEsat}}{R_C}$$

$$\text{and } \because I_C = \beta I_B \Rightarrow I_{Csat} = \beta I_{Bmin} \Rightarrow I_{Bmin} = \frac{I_{Csat}}{\beta}$$

Now, if  $I_B > I_{Bmin}$ , then transistor is in saturation.

$$I_B = \frac{V_{BE} - V_{BEsat}}{R_B}$$

To bring transistor in saturation —

- 1) Increase  $I_B$  by  $\Delta V_{BE}$  so that  $I_B > I_{Bmin}$ .
- 2) If  $I_B$  = constant, then  $\Delta T_B$  so that  $I_{Bmin} = I_B$  by  $\Delta V_{CE}$  and/or  $\Delta R_C$ .
- 3) If  $I_B$  = constant &  $I_{Csat} = \text{constant}$ , then  $\Delta \beta$  so that  $I_{Bmin} = I_B$ .

At  $\beta = \beta_{sat}$ ,

$$T_{Bmin} = \frac{I_{Csat}}{\beta_{sat}} - I_B \Rightarrow \beta_{sat} = \frac{I_{Csat}}{T_{Bmin} - I_B}$$

Important points for saturation—

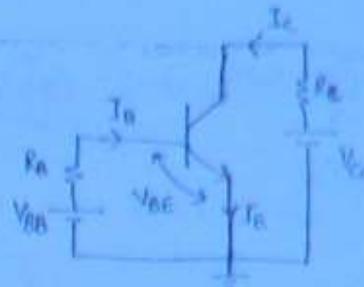
- $I_{Coff} = I_C$  &  $I_C$  are  $I_B$  by cutin voltage  $V_T$ .
- If a transistor has to be operated in saturation region, we should design the ckt, so that  $I_B > I_{Bmin}$  by a factor of 2 to 10.
- The ratio of  $I_{Csat}$  &  $I_B$  to ensure saturation is called fwd. p.

Input Characteristic :-

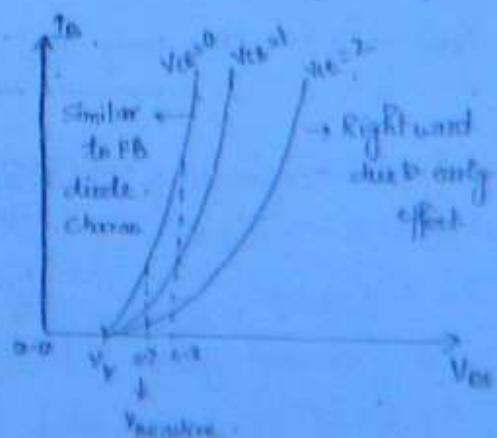
$$V_{BE} = f(I_B, V_{CE})$$

When  $V_{CE} = 0$ , off char similar to diode.

When  $V_{CE} \uparrow \Rightarrow V_C \uparrow \& V_E = 0$ , and due to early effect ( $V_{BE}$  more  $R_B$ ),  $I_B \downarrow$ .



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### III Common Collector Configuration

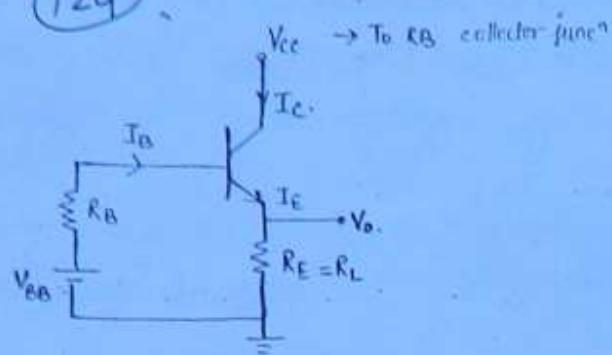
(124)

If char. :-

$$V_{CE} = f(I_E, V_{CE})$$

off char. :-

$$I_E = f(I_B, V_{CE})$$



Output characteristic :-

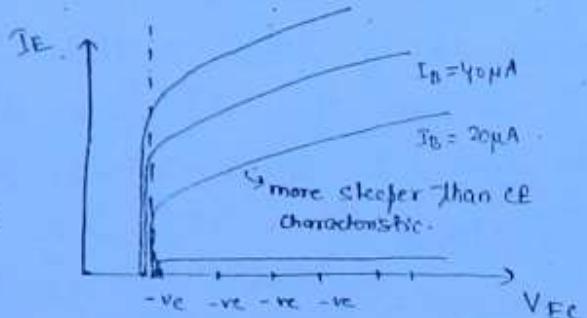
$$I_C = \beta I_B + (1+\beta) I_{CO}$$

$$\Rightarrow I_E - I_B = \beta I_B + (1+\beta) I_{CO} \Rightarrow I_E = (1+\beta) I_B + (1+\beta) I_{CO} \approx (1+\beta) I_B$$

- When  $V_{CE} \downarrow (-ve)$ ,  $V_{CE} \uparrow$  &  $I_C$  is more  $R_B$ .

& due to early effect  $\alpha \uparrow$  and  $\beta \uparrow$ .

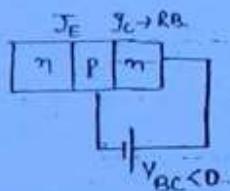
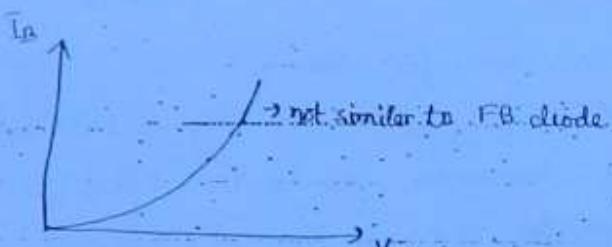
- Curve is more steeper than CE config. since  $(\beta+1)$  variation is  $>$   $\beta$  variation.



Input characteristic :-

→ If characteristic is not similar to forward biased diode charac. since it is taken across  $R_B$  junction.

→  $V_{BC} \uparrow \Rightarrow V_{CB} \downarrow \Rightarrow I_C \downarrow$  in  $R_B$ , early effect  $\downarrow$  &  $I_B \uparrow$ .



Important points regarding CC config. :-

- Highest  $\beta$  (50K $\beta$  - 500K $\beta$ )

- Lowest  $R_o$  ( $< 100 \Omega$ )

- Highest  $A_I$  (current gain) ;  $|A_I| = \frac{I_E}{I_B} = (1+\beta)$  } lowest for  $CB = \alpha$  ?

- Lowest  $A_V (< 1)$  ; typical value - 0.98. Max.  $A_V = 1$  (ideal condn), hence it

-  $R_o$  is also called emitter follower.

- It is basically CCVS.

- Emitter follower is analogous to voltage follower in op-amp and source follower in FET.

Voltage follower & source follower are VCVS. (125)

lowest Power gain; typical value = 48.

Phase shift =  $0^\circ$ .

Application -

- i) Highest i/p resistance device.
- ii) As a buffer amplifier, i.e., an impedance matching device b/w high resistance & low resistance device.
- iii) As an audio freq. power amplifier.

Important Points for CB configuration:

lowest  $R_i$  ( $< 100\Omega$ )

Highest  $R_o$  ( $> 1M\Omega$ )

lowest  $A_I$  ( $= 1$ )

Highest  $A_V$

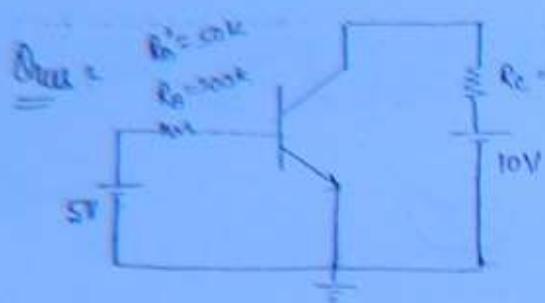
Moderate  $A_P$ ; typical value = 68

Phase shift =  $0^\circ$

CB amplifier will offer largest bandwidth & hence more suitable for high freq. applications.

Application -

- (i) As a constant current source
- ii) As an non-inverting voltage amplifier
- iii) As a high frequency amplifier
- iv) As an impedance matching device b/w low resistance & high resistance.



$\beta = 100$ ,  $I_{CO} = 20\text{nA}$ , Si transistor.

(126)

\* Typical Junc<sup>n</sup> voltages for npn transistor at 25°C -

	Si	Ge
$V_{BEsat}$	0.2	0.1
$V_{BEsat} = V_B$	0.8	0.3
$V_{BEactive}$	0.7	0.2
$V_{BEcutoff} = V_T$	0.5	0.1
$V_{BECutoff}$	0.0	-0.1
$V_{CEsat}$ (for preg.)	> 0.2	> 0.1

\* For pnp transistor, sign of all the entries should be reversed.

Sol<sup>n</sup> - ①  $R_B = 200\text{k}\Omega$ .

$$\because V_{BE} = +ve = 5V \Rightarrow V_{BE} = +ve \Rightarrow I_E = \beta I_B.$$

at ① is in active region -

$$I_E = \beta I_B + (1+\beta) I_{CO}$$

$$I_B = \frac{V_{BE} - V_{BE}}{R_B} = \frac{5 - 0.7}{200k} = 0.0215 \text{ mA.}$$

$$I_E \approx \beta I_B = 2.15 \text{ mA.}$$

$$V_{CE} = 10 - 3k \cdot I_E \Rightarrow V_{CE} = 10 - 3 \times 2.15 = 3.55 \text{ V} \gg 0.2 \text{ V hence, transistor is in active region.}$$

Definitely,  $V_{CB} = V_{CE} - V_{BE}$

$$= 3.55 - 0.7$$

$$= 2.85 \text{ V} \Rightarrow \text{junction is definitely } R_B.$$

②  $R_E = 50\text{k}\Omega$ .

$$I_B = \frac{5 - 0.7}{50} = 0.086 \text{ mA}$$

$$\therefore I_C = 8.6 \text{ mA} \Rightarrow V_{CE} = 10 - 3k(8.6) = -15.8 < 0.2 \Rightarrow \text{Tr} \rightarrow \uparrow \text{saturation.}$$

↓ our assumption was wrong.

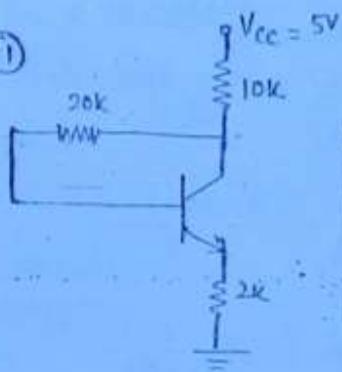
$$* \text{Overdrive factor} = \frac{I_B}{I_{B\min}}$$

Eg  $I_{B\min} = \frac{3.27 \text{ mA}}{100} = 0.0327 \text{ mA}$

(127)

Overdrive factor =  $\frac{0.084}{0.0327} = 2.5 \text{ times} \rightarrow \text{Hence, transistor is well in saturation}$

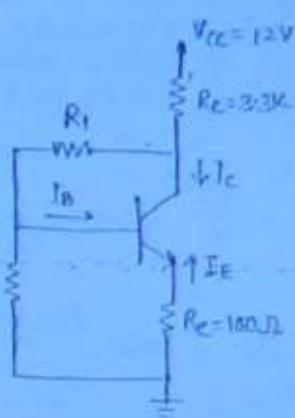
Ques:-



$\beta = 75$ , if does find  $V_C$ ? (2)

If  $\alpha = 0.98$  and  $V_{BE} = 0.7 \text{ V}$   
find  $R_1$  in circuit for an  
emitter current  $I_E = -2 \text{ mA}$   
Neglect reverse sat. current.

Ans:-  $I_B = 4.6 \mu\text{A}$   
 $V_C \approx 1.49 \text{ V}$



Soln:

## Early Voltage:

(28)

Ques In the given ckt, determine the value of  $R_1$ ,  $R_2$  and  $R_L$  so that collector current through the transistor is 1mA.  $V_B = 3V$ ,  $V_{C2} = 6V$ . Take  $V_{BE} = 0.7V$  and let  $\beta$  of transistors are very high.

$$\text{Soln. } V_1 = V_{BE} + I_c R_C$$

$$V_1 = 0.7 + 0.2 \times 1 = 0.9V$$

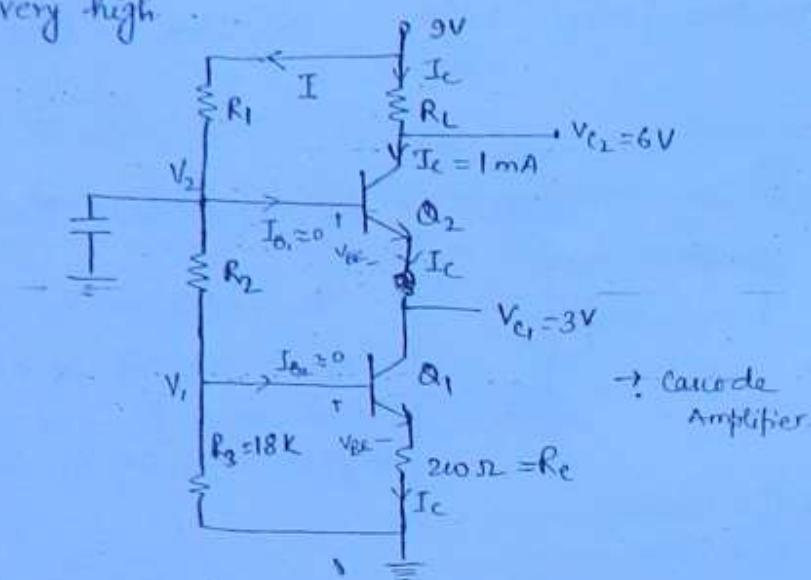
$$V_2 = V_{BE} + V_{CE1} = 0.7 + 3 = 3.7V$$

$$I_B = \frac{V_1}{R_3} = \frac{0.9}{18} = 0.05 \text{ mA}$$

$$\therefore R_1 = \frac{9 - V_2}{I} = 10K$$

$$R_2 = \frac{V_2 - V_1}{I} = 56K$$

$$\therefore R_L = \frac{9 - 6}{1 \text{ mA}} = 3K$$



$\because \beta = \text{very high} \Rightarrow \alpha \approx 1$

$\Rightarrow I_C \approx I_E \text{ & } I_B \approx 0$

? Cenode Amplifier

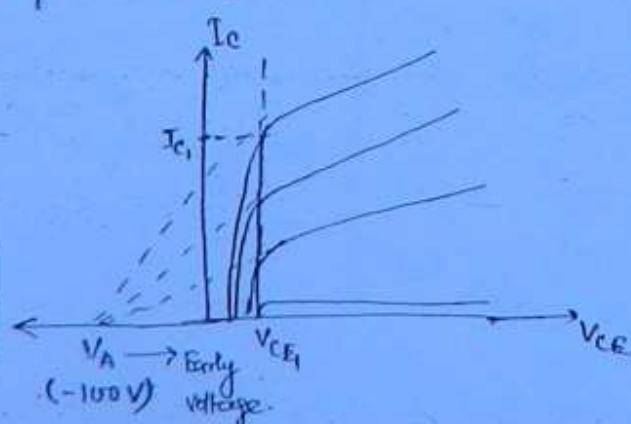
## Early Voltage:

- It helps in finding o/p resistance of transistor.

$$\text{Slope} = \frac{I_C - 0}{V_{CE1} - (-V_A)} = \frac{I_C}{V_{CE1} + V_A} \approx \frac{I_C}{V_A} = \frac{1}{r_o}$$

$\rightarrow V_A \gg V_{CE1}$

$$\therefore r_o = \frac{V_A}{I_C} = \text{o/p resistance of ckt}$$



- $V_A$  = very high for CB.  $\Rightarrow r_o$  = very high.  $\approx \infty$ .
- $V_A$  for CC is slightly less than  $V_A$  of CE.  $\Rightarrow r_{OCC} < r_{OCE}$
- $V_{ACB} \gg V_{ACE} > V_{ACC}$  or  $r_{OCE} \gg r_{OCC} > r_{OCC}$

03<sup>rd</sup> September, 2012.

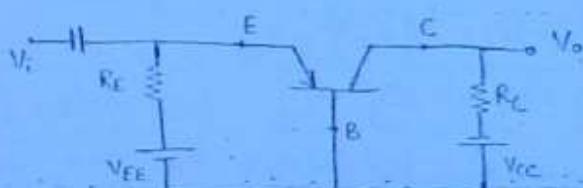
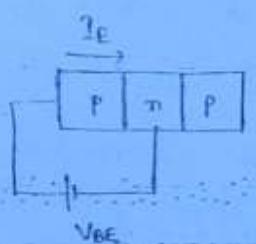
(129)

Ques: In a CE-transistor, at  $V_{CE}=1V$ ,  $V_{EE}$  is adjusted to give a collector current of 1mA. Keeping  $V_{EE}$  constant,  $V_{CE}$  is  $\uparrow$  to 11V. Find new value of  $I_c$  if  $V_A = 150V$ .

$$\text{Soln: } \frac{0 - 1\text{mA}}{-100 - 1} = \frac{11 - 1}{11 - 1}$$

$$\Rightarrow \frac{+10}{101} = \alpha - 1 \Rightarrow \alpha = \frac{11}{101} = 1.09 \text{ mA.}$$

Transistor as an Amplifier :-



$\Rightarrow$  Transistor is in active region

$J_C \rightarrow F_B, J_E \rightarrow R_B$

$$I_C = \alpha I_E + I_{Co} \approx \alpha I_E$$

AC analysis :-

On applying signal at  $V_i$ . If  $V_i \uparrow$  by  $\Delta V_i$ , then  $I_E \uparrow$  by  $\Delta I_E$  &  $I_C$  also increases.

$$I_C + \Delta I_C = \alpha [I_E + \Delta I_E]$$

$$\Rightarrow \boxed{\Delta I_C = \alpha \Delta I_E}$$

Now,  $V_o$  will also  $\uparrow$ .  $\Rightarrow \Delta V_o = \Delta I_C \cdot R_C \Rightarrow \Delta V_o = \alpha \Delta I_E \cdot R_C$

Change in i/p.  $\Delta V_i = \Delta I_E \cdot R_E$

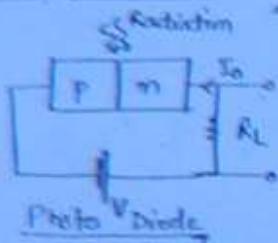
$\therefore$  Voltage gain,  $A_V = \frac{\Delta V_o}{\Delta V_i} \Rightarrow \boxed{A_V = \frac{\alpha R_C}{R_E} \approx \frac{R_C}{R_E}}$  {Invert}

$\Rightarrow A_V \gg 1 \rightarrow$  Hence, Amplifier

→ Transistor provides power gain as well as voltage or current amplification. Current in low resistance JFET is transferred to high resistance of ckt. The word transistor which originated as a contraction of transfer resistor is based upon above physical picture of device.

(130)

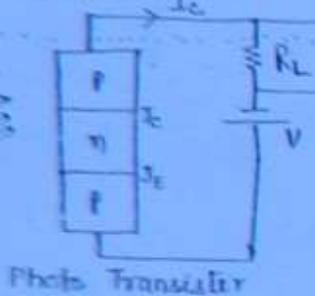
### Photo Transistor :- (Photo-diode)



$I_{0s}$  = Reverse saturation current

Due to radiation  $T \uparrow$ ,  $I_{0s}$  increases by  $\Delta I_{0s}$

$V_o$  also increases by  $\Delta V_o$   $\therefore \Delta V_o = \Delta I_{0s} \cdot R_L$



$J_E \rightarrow F_B$ ,  $J_C \rightarrow R_B$   $\Rightarrow$  Photo transistor is in active region

$I_c = \beta I_B + (1+\beta) I_{0s}$  but  $I_B = 0$ , since base is open

$$\therefore I_c = (1+\beta) I_{0s}$$

Now, due to radiation,  $T \uparrow$ ,  $I_{0s} \uparrow$ ;  $I_c \uparrow$  by  $\Delta I_c$ .

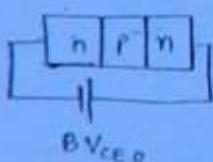
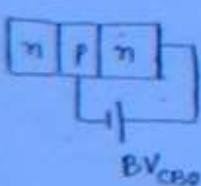
$$\Delta I_c = (1+\beta) \Delta I_{0s}$$

Now,  $V_o$  increases by  $\Delta V_o = \Delta I_c \cdot R_L \Rightarrow \Delta V_o = (1+\beta) \Delta I_{0s} \cdot R_L$

Therefore, Photo transistor is more sensitive than photo diode by a factor  $(1+\beta)$ .

### Maximum Voltage rating of Transistor :-

#### Avalanche Multiplication :-



$$BV_{CEO} > BV_{CBO}$$

- $BV_{CEO} \rightarrow$  maximum reverse biasing voltage which may be applied before breakdown b/w c & e of transistor, keeping E open., i.e.,  $I_E = 0$ .

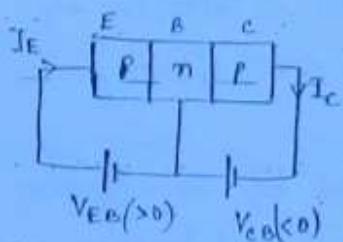
•  $-BV_{CEO} \rightarrow$  for CE configuration, collector to emitter breakdown voltage with open circuit base.

Note:

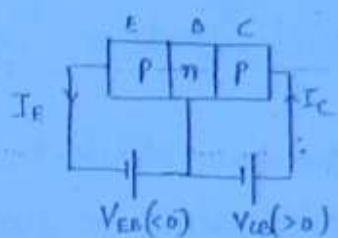
(13)

• In a particular transistor, voltage limit is determined by punch-through or breakdown (due to avalanche multiplication) whichever occurs at the lower voltage.

Ebers Moll Model :-



(Forward or Normal Active mode)  
( $\alpha_F$  or  $\alpha_N$ )



(Reverse or Inverse Active Mode)  
( $\alpha_I$ )

$I_E \rightarrow FB$ ,  $I_C \rightarrow RB$

$I_E \rightarrow RB$ ,  $I_C \rightarrow FB$ .

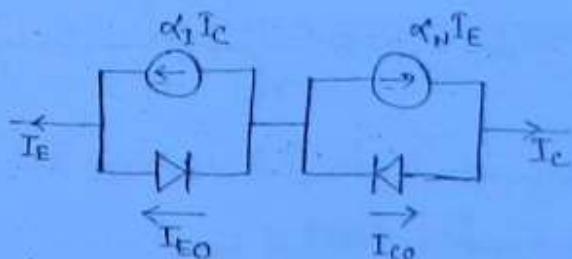
$$I_C = \alpha_N I_E + I_{CO} \quad \text{--- (1)}$$

$$I_E = \alpha_I I_C + I_{EO} \quad \text{--- (2)}$$

→ parameters  $\alpha_N, \alpha_I, I_{CO}, I_{EO}$  are not independent (experimentally).

They depend on each other as  $\alpha_N \cdot I_{EO} = \alpha_I \cdot I_{CO}$

→  $I_{EO} = 0.5 I_{CO}$  to  $I_{CO}$ . { i.e.,  $I_{EO} < I_{CO}$  as conc. of E  $>$  conc. of C  
 $\Rightarrow$  conc. of minority in E  $<$  conc. of minority in C  
 $\Rightarrow I_{EO} \leq I_{CO}$  }



Now, if  $\alpha_I = \alpha_N = 0$  ;  
then



Ebers-Moll Model

- Model involves two ideal diodes placed back to back with reverse saturation current  $I_{eo}$  &  $I_{bo}$  and two dependent CCCS shunting ideal diodes. (132)
- observe from the figure that, dependent current source can be eliminated from this figure provided  $\alpha_f = \alpha_N = 0$ . for eg., by making base width much larger than diffusion length of minority carrier in base, then all minority carriers will recombine in base and none will survive to reach collector. for this case current gain  $\alpha$  will be 0. Under this condition transistor action ceases and we simply have two diodes placed back to back.
- This discussion shows why it is impossible to construct a transistor by simply connecting two separate or isolated diode in series opposing.
- A cascade of two p-n diode exhibits transistor properties like amplification only if carriers injected across one junction diffuse across 2nd junction.

### Cutoff Mode :

$$I_E \& I_C \rightarrow R_B \Rightarrow I_E = 0.$$

from eqn ① -  $I_C = I_{co}$

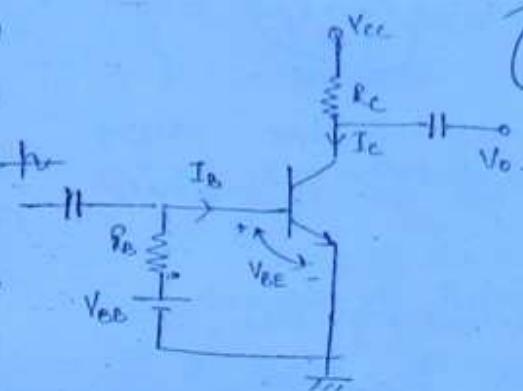
### Saturation

$$I_E \& I_C \rightarrow F_B.$$

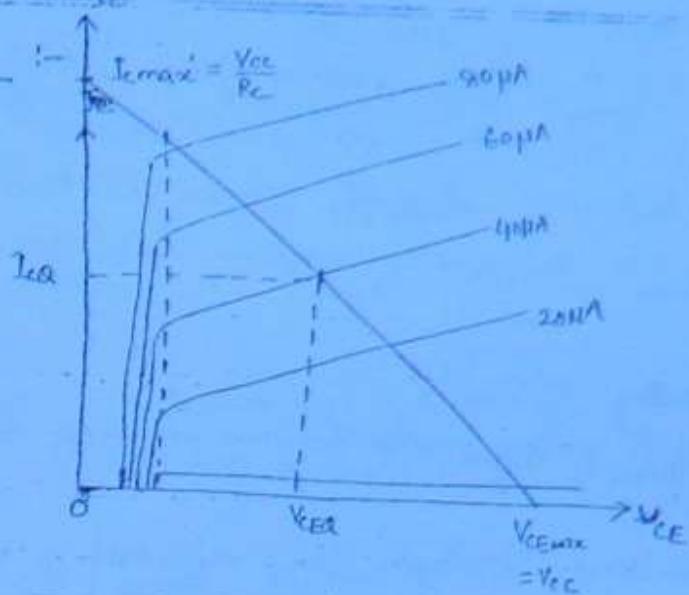
from eqn ① -

$$I_C = \alpha_N I_E - I_{co} \left[ e^{\frac{V_{ce}}{V_T}} - 1 \right]$$

## Transistor Biasing and Stabilization :-



(133)



-  $I_E \rightarrow R_B$ ,  $I_C \rightarrow R_E$ .

$$I_B = \frac{V_{BB} - V_{BE}}{R_B} \quad \text{--- (1)}$$

$$I_C = \beta I_B + (1+\beta) I_{C0} \quad \text{--- (2)}$$

$$V_{CE} = V_{CC} - I_C R_C \Rightarrow I_C R_C = V_{CC} - V_{CE} \Rightarrow I_C = -\frac{V_{CE}}{R_C} + \frac{V_{CC}}{R_C} \quad \text{--- (3)}$$

↳ DC load line.

Eqn (3) is similar to  $y = mx + c$ ; slope  $= -1/R_C$ .

$$\text{When } V_{CE} = 0, I_{Cmax} = \frac{V_{CC}}{R_C}$$

$$\text{When } I_C = 0, V_{CEmin} = V_{CC}$$

from eqn (1); set  $I_B = 40\mu A$ , then

$$\boxed{\beta = \frac{I_{CQ}}{I_B}} \quad \begin{aligned} &\rightarrow Q = \text{Quiescent Point / operating point.} \\ &\rightarrow Q = f(I_B, I_C, V_{CE}) \end{aligned}$$

On application of input -

$$I_b = I_B + i_b \quad \Rightarrow \quad I'_c = \beta I_b = \beta (I_B + i_b) = I_c + i_c$$

(DC) (AC)

Now, let  $I_B = 40\mu A$  &  $i_b = 20 \sin \omega t \mu A$ , then  $I_b = 40 + 20 \sin \omega t$

$$\therefore I_{bmax} = 60 \mu A, I_{bmin} = 20 \mu A$$

Hence, Q point lies well within the active region. Therefore, no distortion in the output.

Ex 2  $I_B = 40\mu A$ ,  $i_b = 40 \sin \omega t$   
 $I_{bmax} = 80\mu A$ ,  $I_{bmin} = 0$

Hence, transistor is just in active region.

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Ex 3  $I_B = 40\mu A$ ,  $i_b = 50 \sin \omega t$

$I_{bmax} = 90\mu A$ ,  $I_{bmin} = -10$ , Now, there will be distortion in the opf.

Ex 4 Adjusting  $I_B = 20\mu A$ ,  $i_b = 20 \sin \omega t$   
 $I_{bmax} = 40\mu A$ ,  $I_{bmin} = 0 A$

Ex 5  $I_B = 20\mu A$ ,  $i_b = 40 \sin \omega t$   
 $I_{bmax} = 60$ ,  $I_{bmin} = -20$   
 $\downarrow$   
 $\omega_{cutoff}$

Ex 6  $I_B = 60\mu A$ ,  $i_b = 30 \sin \omega t \Rightarrow I_{bmax} = 90$ ,  $I_{bmin} = 30$   
 $\downarrow$   
saturation

### Important Points

- The collector characteristic or opf characteristic of transistor is divided into saturation, cutoff and active regions.
- Transistor can work as a switch when operated in saturation and cutoff region, ie, extreme ends of the characteristics.
- Procedure to plot dc load line & Q point
  - 1) Identify the value of  $V_{ce}$  &  $I_{cmox}$  of the circuit & locate this point on given charac.
  - 2) Draw a straight line joining  $I_{cmox}$  &  $V_{ce}$  & this straight line is called dc load line.
  - 3) find the operating values  $I_B$ ,  $I_c$  &  $V_{ce}$  for the given ckt & locate these values on given charac.
  - 4) Project three operating values on dc load line & the intercepting point is called Q-point.
- The transistor is said to be under quiescent cond'n when zero i/p signal is applied.

→ Transistor can work as an amplifier if Q point is within active region but Q point is temp sensitive, i.e., as  $T \uparrow$ ,  $I_C \uparrow$  and  $V_{CE} \downarrow$ , so that Q point will be moving towards saturation region and if entered into saturation region, transistor will stop working as an amp.

(135)

- Trans. will provide more power gain or amplification, when Q point is in middle of dc load line.
- for a given trans., Q point is plotted to get faithful reproduction of op signal.
- If shape of op signal differs from shape of ip signal, it is said to be distorted.
- for a stable circuit, the variation in Q-point due to temp. must be small.

### Bias Stability:-

- Stability is effected due to -

#### 1) Temp Instability

- (a)  $I_{CO}$  → It doubles for every  $10^\circ$  rise in temp.  
 $T \uparrow \Rightarrow I_{CO} \uparrow \Rightarrow I_C \uparrow \Rightarrow$  Q point shift towards saturation.

#### (b) $V_{BE}$ →

$$\frac{dV_B}{dt} = -2.5 \text{ mV}/^\circ\text{C}, \text{ & } \frac{dV_{BE}}{dt} = -2.3 \text{ mV}/^\circ\text{C}.$$

$$I_B = \frac{V_{BE} - V_{BE}}{R_B} \quad \text{As } T \uparrow, V_{BE} \downarrow, I_B \uparrow \Rightarrow I_C \uparrow \text{ & } V_{CE} \downarrow \\ \Rightarrow Q \rightarrow \text{saturation.}$$

Note -  $\beta \uparrow$  with temp but change is negligible.

2) Replacement of Transistor -  $\beta$  is highly affected due to replacement of transistor.

(∴ Since small change in  $\alpha$  results in large change in  $\beta$ )

## Stabilisation Techniques



### Biassing Techniques

- (i) C-B Biassing.
- (ii) Self-Biassing.



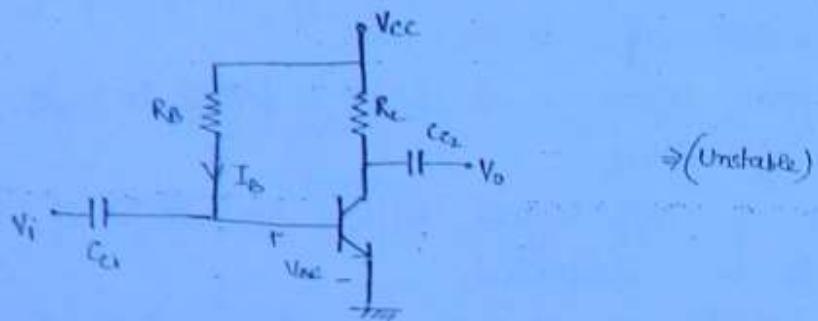
### Compensation Techniques

- (i) Diode compensation
- (ii) Emitter & Transistor compensation.
- (iii) Transistor compensation

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## Fixed Biased circuit :

$$\rightarrow I_B = \frac{V_{CC} - V_{BE}}{R_B} = \text{constant}$$



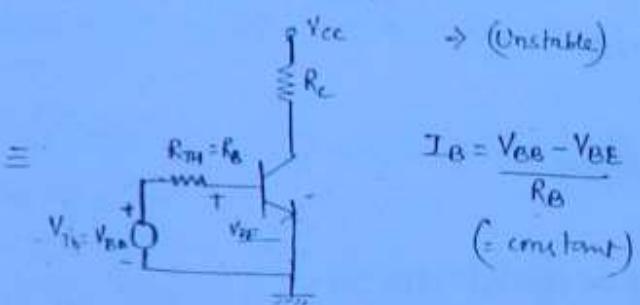
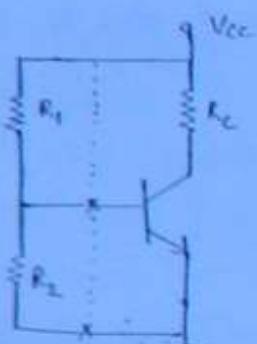
$$\rightarrow V_{BE} = \frac{R_2}{R_1 + R_2} \cdot V_{CC}$$

$$R_{TH} = R_1 \parallel R_2$$

Normally,

$$R_1 \approx 10R_2$$

∴  $R_1 \gg R_2$



$$I_B = \frac{V_{BE} - V_{BE}}{R_B}$$

(= constant)

## Stability factors :

$$\rightarrow S = \left. \frac{dI_c}{dI_{Co}} \right|_{V_{BE} \Delta \beta = \text{constant}}$$

$$\rightarrow S' = S_\beta = \left. \frac{dI_c}{d\beta} \right|_{I_{Co} \Delta V_{BE} = \text{constant}}$$

$$\rightarrow S' = \left. \frac{dI_c}{dV_{BE}} \right|_{I_{Co} \Delta \beta = \text{constant}} = S_\alpha$$

\* As  $T \uparrow$ ,  $I_{Co} \uparrow$ ,  $I_c \uparrow \Rightarrow S = +ve$

\* As  $T \uparrow$ ,  $V_{BE} \downarrow$ ,  $I_c \uparrow \Rightarrow S' = S_\alpha = -ve$

\* As  $\beta \uparrow$ ,  $\beta \uparrow$ ,  $I_c \uparrow \Rightarrow S'' = +ve$

(because of replacement)  
no change in temp

Stability factor,  $S$

$$S = \left| \frac{dI_c}{dT_{c_0}} \right|_{I_B, \beta = \text{constant}}$$

In active region,

(137)

$$I_c = \beta I_B + (1+\beta)T_{c_0}$$

$$\Rightarrow \frac{dI_c}{dT_{c_0}} = \beta \cdot \frac{dI_B}{dT_{c_0}} + (1+\beta)$$

$$\Rightarrow \frac{dI_c}{dT_{c_0}} = \beta \cdot \frac{dI_B}{dT_c} \cdot \frac{dT_c}{dT_{c_0}} + (1+\beta)$$

$$\Rightarrow \frac{dI_c}{dT_{c_0}} \left[ 1 - \beta \cdot \frac{dI_B}{dT_c} \right] = (1+\beta)$$

$$\Rightarrow S = \frac{1+\beta}{1 - \beta \cdot \frac{dI_B}{dT_c}}$$

→ If  $I_B$  = constant,  $\frac{dI_B}{dT_c} = 0 \rightarrow [S = (1+\beta)] \rightarrow$  circuit is unstable  $\rightarrow$  fixed bias

→ If  $I_c \uparrow$  then  $I_B$  should  $\downarrow$ , i.e.,  $dI_B/dT_c < 0$ . In ideal case,  
 $\frac{dI_B}{dT_c} = -1 \rightarrow [S = 1] \rightarrow$  highly stable.

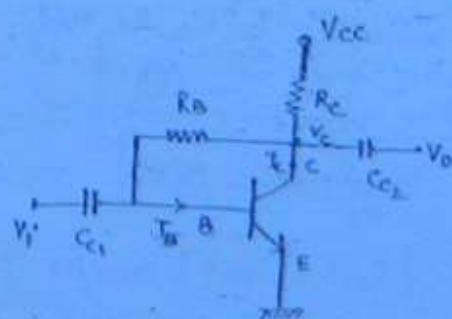
For  $[S < 1+\beta]$ , circuit is stable.

Range of  $S \rightarrow [1 < S < (1+\beta)]$

Techniques :-

1) Collector-Bias Bias -

→ During DC analysis,  $C_{c_1}$  &  $C_{c_2}$  will act as open circuit.



When transistor is in active region-

$$I_C = \beta I_B + (1+\beta) I_{CO} \quad \text{--- (1)}$$

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On manipulating (1) -

$$S = \frac{dI_C}{dI_{CO}} = \frac{1+\beta}{1-\beta \cdot \frac{dI_B}{dI_C}} \quad \text{--- (2).} \quad (\text{Keeping } V_{BE} \text{ & } \beta \text{ constant})$$

$$\text{Applying KVL at i/p} - V_{CC} = (I_C + I_B) R_C + I_B R_B + V_{BE}$$

$$\text{Differentiating w.r.t } I_C - 0 = (R_C + R_B) \frac{dI_B}{dI_C} + R_C + 0$$

$$\Rightarrow \frac{dI_B}{dI_C} = \frac{-R_C}{R_C + R_B} \quad \text{--- (3)}$$

Substituting (3) in (2) -

$$S = \frac{1+\beta}{1 + \frac{\beta R_C}{R_C + R_B}} < (1+\beta) \quad \text{Hence circuit is stable.}$$

$$S = (1+\beta) \cdot \frac{R_C + R_B}{R_B + R_C (1+\beta)}$$

If  $R_C (1+\beta) \gg R_B$ , then

$$S = 1 + \frac{R_C}{R_C} \quad \text{obj}$$

→ If  $R_C \uparrow$  &  $R_B \downarrow$  then  $S \downarrow$ ; Hence S depends on load resistance.

→ It is voltage shunt feedback, hence  $R_i \downarrow$  and  $R_o \downarrow$ .

→ There is unnecessary -ve feedback, circuit is not preferable.

### Theoretical Analysis -

From circuit,  $V_C = V_{CC} - (I_C + I_B) R_C$  and  $I_C = \beta I_B + (1+\beta) I_{CO} \approx \beta I_B$

$$\therefore V_C \approx V_{CC} - I_C R_C \quad \text{--- (1)} \quad \because \frac{I_C}{\beta} = I_B \Rightarrow I_C \gg I_B$$

$$I_B = \frac{V_C - V_{BE}}{R_B} \quad \text{--- (2)}$$

From (1) & (2) - If  $T \uparrow$ ,  $[I_{CO} \uparrow \text{ & } V_{BE} \downarrow]$  and/or  $\beta \uparrow$ , then  $I_C \uparrow$   
 Then  $V_C \uparrow$  but  $I_B \downarrow$  with  $V_C \uparrow \Rightarrow I_C \downarrow$  → -ve feedback.

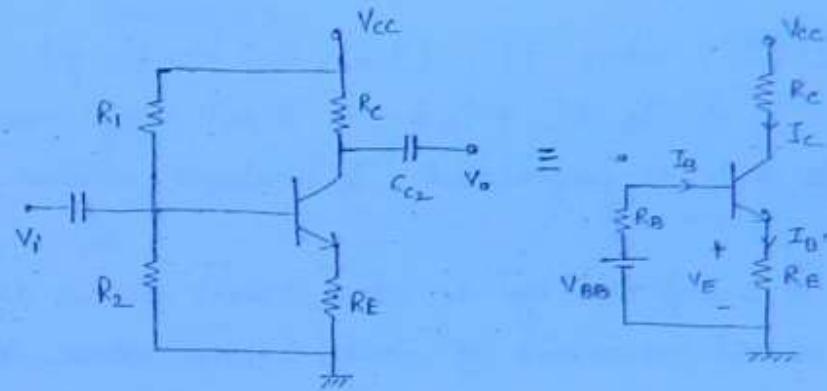
Therefore, rise in  $I_c$  is compensated and  $I_c$  is almost constant. 139  
 - This circuit will compensate for all type of variations, i.e.,  $T_{ao}$ ,  $V_{BE}$  or  $\beta$ .

### Self-Biased Circuit :

$$\rightarrow V_{BB} = V_{Th} = \frac{R_2}{R_1 + R_2} V_{CC}$$

$$\rightarrow R_{Th} = R_E = R_1 \parallel R_2$$

$$\rightarrow I_c = \beta I_B + (1+\beta) I_{Co} \quad \text{--- (1)}$$



$$\rightarrow S = \frac{1 + \beta}{1 - \beta \frac{dI_E}{dI_C}} \quad \text{--- (2)}$$

Writing KVL at i/p —

$$V_{BB} = I_B R_E + V_{BE} + (I_B + I_C) R_E$$

$$\rightarrow \text{Differentiating wrt } I_C \Rightarrow 0 = (R_E + R_E) \frac{dI_E}{dI_C} + R_E$$

$$\Rightarrow \frac{dI_B}{dI_C} = -\frac{R_E}{R_E + R_E} \quad \text{--- (3)}$$

$\rightarrow$  Substituting in eqn (2) —

$$S = \frac{\frac{1 + \beta}{1 + \beta R_E}}{R_E + R_E} < (1 + \beta) \rightarrow \text{Hence it is stable}$$

$$\Rightarrow S = (1 + \beta) \cdot \frac{R_E + R_E}{R_E + (1 + \beta) R_E}$$

$\rightarrow$  If  $(1 + \beta) R_E \gg R_E$ , then,

$$S = 1 + \frac{R_E}{R_E}$$

$\xrightarrow{\text{Advantage}}$   $\left\{ \begin{array}{l} \text{Independent of load resistance} \\ R_E \end{array} \right\}$

$\rightarrow$  Ideally,  $R_E = \infty$ ,  $S = 1$ , Hence,

$R_E \uparrow$  and/or  $R_E \downarrow$ , then  $S \uparrow$

\* Input & o/p resistance will increase.

\* It is current series feedback.

Explanation :-

$$- V_E = (I_B + I_C) R_E \approx I_C R_E \quad \left\{ \because I_C \gg I_B \right\} \quad \text{--- (1)}$$

$$- I_B = \frac{V_{BE} - V_{BE0} - V_E}{R_B} \quad \text{--- (2)}$$

(140)

From (1) & (2), when  $T \uparrow$ , [ $I_{C0} \uparrow$ ,  $V_{BE0}$ ] and/or  $\beta \uparrow$ , then  $I_C \uparrow \Rightarrow V_E \uparrow$ .

$\Rightarrow I_B \downarrow \Rightarrow T_c \downarrow \rightarrow$  It will control variation due to all factors.

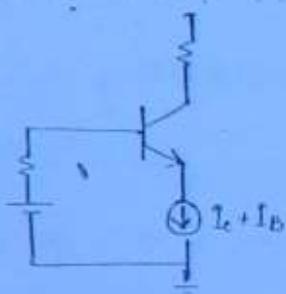
A rise in  $I_C$  is compensated,  $I_C$  is almost constant.

\* If  $R_E$  is replaced by an ideal current source, then  $S$  will become 1.  
(as internal resistance of active (current) source is very high, ideally  $\infty$ )

Ideal current source,

$$R_S = R_E = \infty$$

$$\rightarrow S = 1$$



Practical current source,

$$R_E = \text{very high.}$$

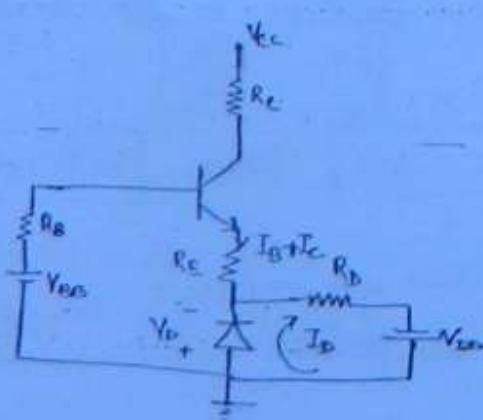
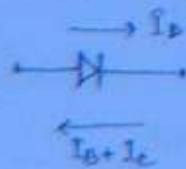
\* Self-biased circuit is also called as voltage divider / Potential divider or emitter bias circuit.

### Compensation Techniques :- (Bias Compensation)

→ Compensation techniques refers to use of temp-sensitive devices like diode, thermistor, sensistor etc.

#### y Diode Compensation

a) for  $V_{BE}$  :-



(14)

$$\rightarrow I = I_B + (I_C + I_E) \quad \left. \begin{array}{l} \text{From eqn, for diode to be FB} \\ \text{---} \end{array} \right\} \quad I > 0 \rightarrow I_B > I_B + I_C.$$

Hence, if we set  $I_B$  based on eqn(2) then we can make D forward biased.

$\rightarrow$  When  $D \rightarrow FB$ , then

$\rightarrow$  Transistor is in active region

$$\rightarrow I_C = \beta I_B + (1+\beta) I_{CO} \quad \dots \quad (3)$$

$$\text{and, } \frac{I_C - (1+\beta) I_{CO}}{\beta} = I_B \quad \dots \quad (2)$$

$$\text{from circuit, } V_{EB} = I_B R_B + V_{BE} + (I_C + I_B) R_E - V_D \quad \dots \quad (3).$$

from (2) & (3)  $\therefore$

$$I_C = \frac{\beta [V_{EB} - (V_{BE} - V_D)] + (R_B + R_E)(1+\beta) I_{CO}}{R_B + R_E(1+\beta)} \quad \dots \quad (4)$$

If, transistor & diode are of similar materials -

$$\therefore \frac{dV_{BE}}{dT} = \frac{dV_D}{dT} = -2.5 \text{ mV/}^{\circ}\text{C}$$

$I_C$  depends on  $(V_{BE} - V_D)$

$$\text{for } 1^{\circ}\text{C change in } T, \quad (V_{BE} - 2.5) - (V_D - 2.5) = (V_{BE} - V_D),$$

Hence,  $I_C$  will remain constant, even if  $V_{BE}$  is changing.

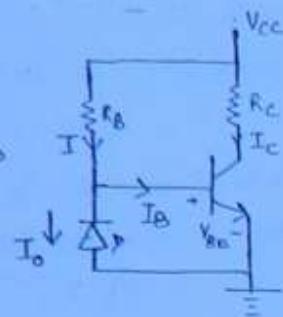
$\rightarrow$  The change of  $V_{BE}$  with temp contribute significantly to change in  $I_C$  of Silicon transistor, therefore circuit is useful for stabilising Si transistor.

$\rightarrow$  The diode is kept forward biased by source  $V_D$  and resistance  $R_D$ .

If the diode is of same material & type, voltage  $V_D$  across diode will have same temp. coeff as  $V_{BE}$ . then from eqn (4), it is clear that  $I_C$  will be insensitive to variation in  $V_{BE}$ .

(b) for  $I_{CO}$ :

→ For Ge,  $V_{BE} = 0.2V$  = voltage across  
(diode) since  
they are in parallel



(142)

$$\rightarrow I_B = I - I_o.$$

$$\rightarrow I = \frac{V_{CC} - V_{BE}}{R_B} = \text{constant} \quad \left. \begin{array}{l} \text{considering } V_{BE} = \text{constant} \end{array} \right\}$$

$$\rightarrow I_c = \beta I_B + (1 + \beta) I_{CO}$$

$$\Rightarrow I_c = \beta [I - I_o] + (1 + \beta) I_{CO} \Rightarrow I_c = \beta I - \beta I_o + \beta I_{CO} \quad \left. \begin{array}{l} \text{if } \beta \gg 1 \end{array} \right\}$$

$$\Rightarrow I_c = \beta I + \beta (I_{CO} - I_o) \quad \downarrow \text{constant}$$

→ for Ge transistor, change in  $I_{CO}$  with temp. play more important role in collector current stability, therefore, this circuit is useful for stabilising Ge Tr.

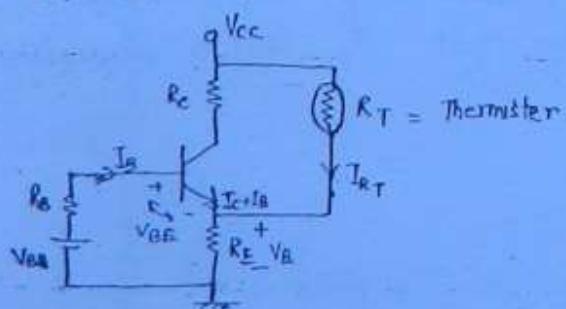
→ If the diode & Tr. are of same type, then  $I_o$  of diode will vary with T at same rate as  $I_{CO}$ . Therefore,  $I_c$  will be insensitive to variation in  $I_{CO}$ .

## 2) Thermister and Sensistor compensator:

→ Thermister → NTC of resistivity ;  $T \uparrow \sigma \downarrow$   
(lightly doped)

→ Sensistor → PTC of resistivity ;  $T \uparrow, \sigma \downarrow$   
(highly doped)

$$\Rightarrow I_B = \frac{V_{BB} - V_{BE} - V_E}{R_B}$$

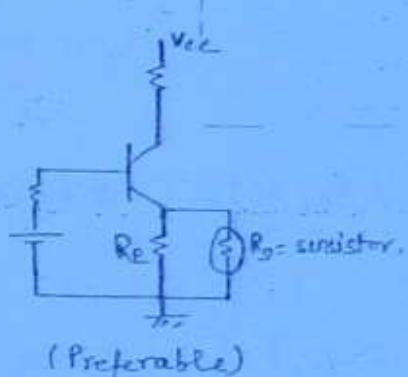
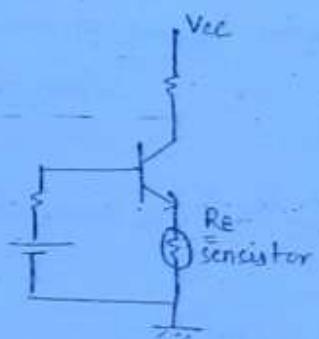


$$\rightarrow V_E = (I_B + I_C + I_{R_T}) R_E \approx (I_C + I_{R_T}) R_E$$

(143)

Now, When  $T \uparrow$ , ( $I_{CO} \uparrow, V_{BE} \downarrow$ ), then  $I_C \uparrow, R_T \downarrow \Rightarrow I_{R_T} \uparrow \Rightarrow V_E \uparrow$   
 $\Rightarrow I_B \downarrow \Rightarrow I_C \downarrow$   
Hence, rise in  $I_C$  is compensated.

By using sensistor :-



\*  $R_E$  replaced by a sensistor or we can place a sensistor parallel to  $R_E$ .

$$T \uparrow \quad (1) R_E = 1\text{ k}\Omega \quad T \uparrow \quad (1) R_E = 2\text{ k}\Omega \quad \Delta R_E = 100\%$$

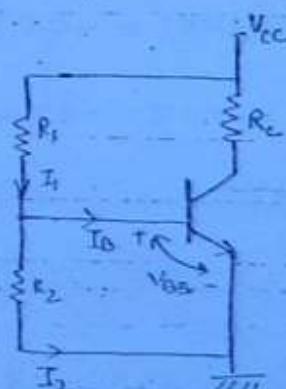
$$2\text{k}\Omega \quad (1) 2\text{k}\Omega \equiv \frac{R_{eq}}{1\text{k}\Omega} \quad T \uparrow \quad \frac{1}{2\text{k}} \quad (1) 4\text{k}\Omega \equiv R_{eq} = 1.33\text{k}\Omega, \Delta R_E = 33\%$$

→ Hence, controlled feedback by using sensistor.

$$I_B = I_1 - I_2$$

$$I_1 = \frac{V_{cc} - V_{BE}}{R_1}$$

$$I_2 = \frac{V_{BE}}{R_2}$$



Now, when  $T \uparrow$ , ( $I_{CO} \uparrow, V_{BE} \downarrow$ )  
then  $I_C \uparrow$ , then to compensate  
we want  $I_B \downarrow$ . or  
 $\Rightarrow I_1 \uparrow$  and/or  $I_2 \uparrow$   
 $\Rightarrow R_1 \uparrow$  and/or  $R_2 \downarrow$

Hence,  $R_1$  can be replaced by sensistor &  $R_2$  can be replaced by Thermistor. or  $R_1$  can be replaced by sensistor in II &  $R_2$  with thermistor in II.

Ques: In two stage ckt, assume  $\beta = 100$  for each transistor

(a) Determine  $R$  so that

quiescent conditions are

$$V_{CE_1} = -4V, \quad V_{CE_2} = -6V$$

(b) Explain how Q-point stabilization is obtained.

Take  $V_{BE} = 0.2V$ .

Sol: Since  $\beta \gg 1, I_{B_2} \ll I_{C_1}$  &  $I_{B_1} \ll I_{C_1}$ .  $\Rightarrow$  we will neglect  $I_{B_1}$  &  $I_{B_2}$ .

By KVL—

$$-24 - 17.8(I_{C_1}) - V_{CE_1} - 2.2(I_{C_1}) = 0 \Rightarrow -24 - (-4) = 17.8I_{C_1} + 2.2I_{C_1}$$

$$\Rightarrow -20I_{C_1} = 20$$

$$\Rightarrow I_{C_1} = -1mA$$

By KVL—

$$-24 - 8I_{C_2} - V_{CE_2} - 1(I_{C_2}) - 3(I_{C_2}) = 0$$

$$\Rightarrow I_{C_2} = -4.5mA$$

Now,

$$R = \frac{V_A - V_B}{I_{B_1}} ; \quad V_A = 3K \times (-1.5) = -4.5V$$

$$V_B = V_{BE} + I_{C_1} \times 2.2 = -0.2 - 1 \times 2.2$$

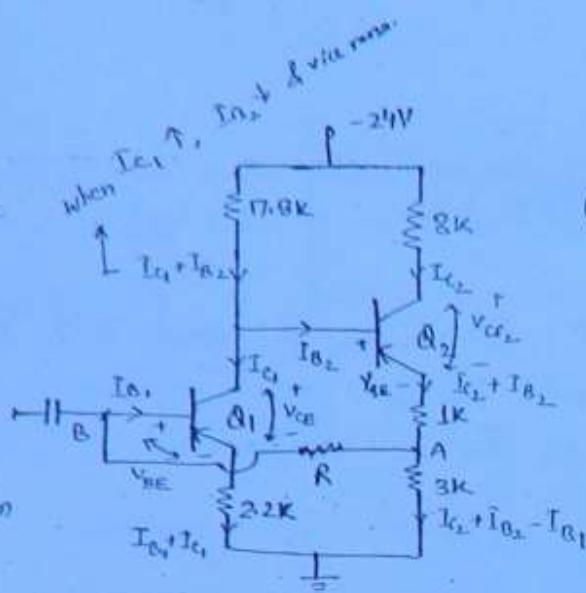
$$\therefore V_B = -2.4V$$

$$\Rightarrow I_{B_1} = \frac{I_{C_1}}{\beta} = -0.01mA$$

$$\therefore R = \frac{-4.5 - (-2.4)}{-0.01} = 210K\Omega$$

(b) When  $T \uparrow, |I_{C_2}| \uparrow, |V_A| \uparrow, |I_{B_1}| \uparrow, |I_{C_1}| \uparrow, |I_{B_2}| \downarrow, |I_{C_2}| \downarrow \rightarrow$  compensated.

When  $T \uparrow, |I_{C_1}| \uparrow, |I_{B_2}| \downarrow, |I_{C_2}| \downarrow, |V_A| \downarrow, |I_{B_1}| \downarrow, |I_{C_1}| \downarrow \rightarrow$  " .



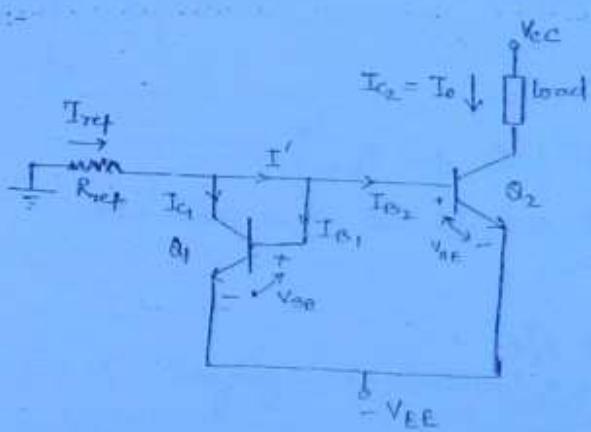
04<sup>th</sup> September, 2012

## Current Mirror Circuit :-

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- The output current is forced to equal the i<sub>p</sub> current, i.e., o/p current is a mirror image of i<sub>p</sub> current.
- They are widely used in designing of differential amplifiers & etc.
- Their major advantages are -
  - Simplicity in circuit design.
  - Easy to fabricate.
  - Minimum no. of components are reqd.
  - Low cost.

## Basic Diagram :-



### Reqd. conditions :-

- Both Tr. are in active region
- Both Tr. are identical, i.e.,  $\beta_1 = \beta_2 = \beta$  &  $V_{BE1} = V_{BE2} = V_{GE}$ .
- $\beta$  should be very large.

Writing KVL -

$$0 - I_{ref} \cdot R_{ref} - V_{BE} = -V_{EE}$$

$$\Rightarrow I_{ref} = \frac{V_{EE} - V_{BE}}{R_{ref}} \rightarrow \text{independent of load.}$$

from figure -

$$I_{B1} = I_{B2} \quad \{ \because I' \text{ is divided among two identical paths} \}$$

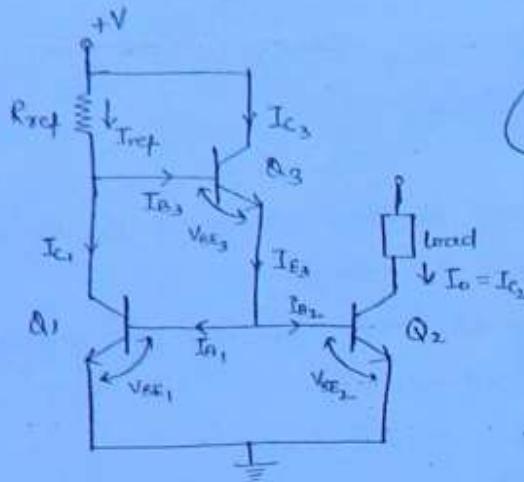
$$\therefore \beta_1 = \beta_2 \Rightarrow I_{C1} = I_{C2}$$

$$\text{Since, } I_{ref} = I_{C1} + I' \Rightarrow I_{ref} = I_{C2} + 2I_{B2}$$

$$\Rightarrow I_{ref} = I_{C2} + \frac{2I_{C2}}{\beta}$$

$$\Rightarrow I_{ref} \approx I_{C2} \quad \text{if } \beta \text{ is very large.}$$

- $\theta_1, \theta_2, \theta_3$  are in active region
  - $\theta_1, \theta_2, \theta_3$  should be identical,  
i.e.,  $\beta_1 = \beta_2 = \beta_3 = \beta$  &  $V_{BE_1} = V_{BE_2} = V_{BE_3} = V_{BE}$
  - $\rightarrow$



By applying KVL -

$$V_T = I_{ref} \cdot R_{ref} + V_{BE_3} + V_{BE_2}$$

$$\Rightarrow I_{ref} = \frac{V_f - 2V_{BE}}{R_{ref}} \quad \text{--- (1).} \quad \{ \text{Independent of load} \}.$$

Head,

$$T_{ref} = T_{C_1} + T_{B_3} \quad \text{--- (2)}$$

from fig.,  $T_{B_1} = T_{B_2}$  &  $\therefore$  identical paths for  $T_{E_3}$  }.

$$\Rightarrow \boxed{\mathbb{P}_1 = \mathbb{P}_{\mathcal{C}_2}} \quad \left\{ \because \beta_1 = \beta_2 \right\} \quad \text{--- (3)}$$

$$\rightarrow T_{B_3} + T_{C_3} = T_{E_3} \Rightarrow T_{E_3} = T_{B_3} + \beta_3 T_{B_3}$$

$$\Rightarrow \hat{I}_{E_3} = (\hat{B}_3 + I) \hat{I}_{B_3}$$

$$\Rightarrow 2\Gamma_{B_2} = (\Gamma_{B_3} + 1) \Gamma_{B_3}$$

$$\Rightarrow \frac{2\mathfrak{I}_{C_2}}{\beta} = (\beta_3 + 1) \mathbb{I}_{B_3}.$$

$$\Rightarrow T_{B_3} = \frac{\alpha T_{C_2}}{\beta(1+\beta_3)} \quad \text{--- (4)}$$

From (3) & (4) in (2) -

$$\therefore T_{ref} = T_0 + \frac{2T_0}{\beta(1+\beta)} \Rightarrow T_0 = \frac{T_{ref}}{1 + \frac{2}{\beta(1+\beta)}} \quad \left\{ \because T_0 \in T_{G_2} \right\}$$

Since  $\beta \gg 1$  -

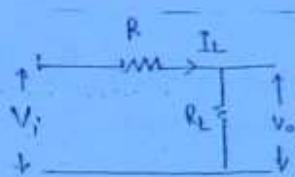
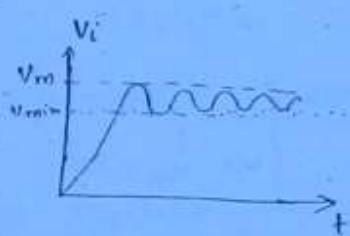
$$I_0 = \frac{T_{ref}}{1 + \beta^2} \Rightarrow \boxed{I_0 \approx T_{ref}} \quad \Rightarrow \quad \boxed{I_0 = \left( \frac{\beta^2 + \beta}{\beta^2 + \beta + 2} \right) \cdot T_{ref}}$$

$\rightarrow$  It is necessary that  $\Omega_1 \neq \Omega_2$  are identical. If  $\Omega_3$  is not identical then...

$$T_0 = \frac{T_{ret}}{1 + \frac{2}{\beta(1+\beta_3)}} \quad \left\{ \beta_3 \rightarrow \text{for } q_3 \right\}. \quad (147)$$

In this circuit, it is not required to have very high  $\beta$ , since a term of  $\beta^2$  is appearing in denominator which will be very large.

### Voltage Regulator circuit :-



→ Line variation :- Variation in  $V_o$  due to variation in line voltage  $V_L$ .

→ load variation :-  $R_L$

Line Regulation :-  $V_i = \text{varying}$ ,  $R_L = \text{constant}$ ,  $V_o$  should be constant.

$V_o = I_L R_L$ , hence  $I_L$  should be constant for  $V_o$  to be constant.

Hence in line regulation, line voltage is varying but load current remains constant.

Load Regulation:  $V_i = \text{constant}$ ,  $R_L = \text{varying}$ ,

for  $V_o$  to be constant, when  $R_L \uparrow$ ,  $I_L$  should  $\downarrow$  and vice versa.

Hence, in load regulation,  $R_L$  varying but  $V_o$  remains constant.

## Voltage Regulator:

- It regulates load voltage.

- In regulator circuit, load voltage  $V_o$  will be maintained almost constant.

irrespective of load variations & i/p voltage variations (the varant).

- Performance of a regulator circuit is analysed by its regulation, i.e.,

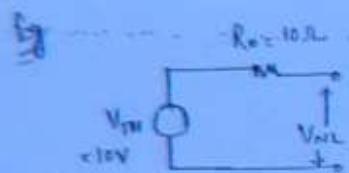
$$\% \text{ Regulation} = \frac{V_{NL} - V_{FL}}{V_{FL}} \times 100\%$$

(148)

$\rightarrow V_{NL}$  = No load voltage ,  $I_L \rightarrow 0$  or  $R_L \rightarrow \infty$ .

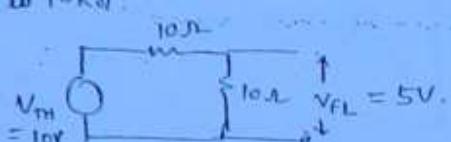
$V_{FL}$  = Full load voltage ,  $I_L \rightarrow I_{L\max}$  or  $R_L \rightarrow R_{L\min}$ .

$\rightarrow$  Ideally,  $V_{NL} = V_{FL}$  &  $\% \text{ Regulation} = 0\%$



$$V_{NL} = 10V$$

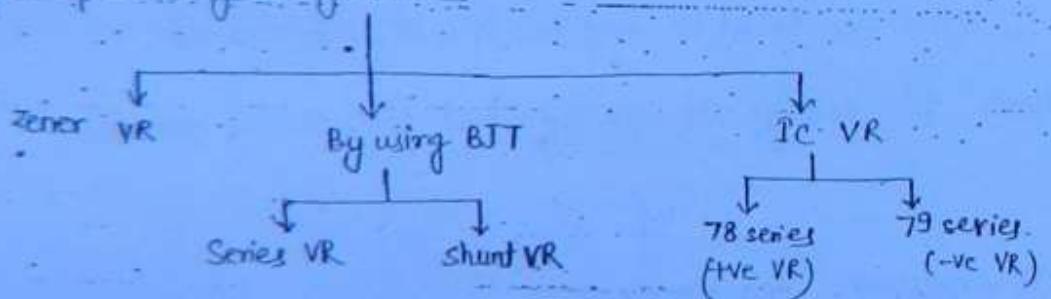
$R_L = 10\Omega \text{ to } 10K\Omega$



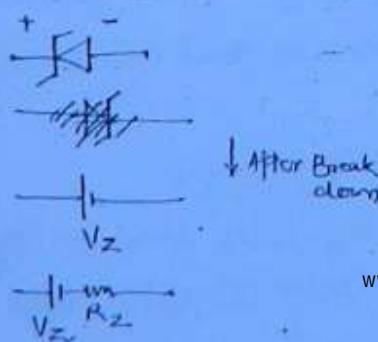
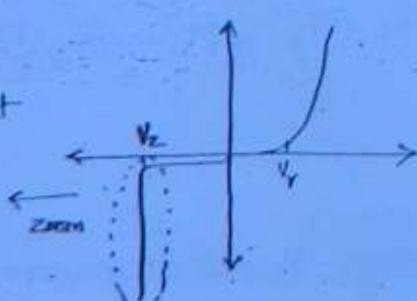
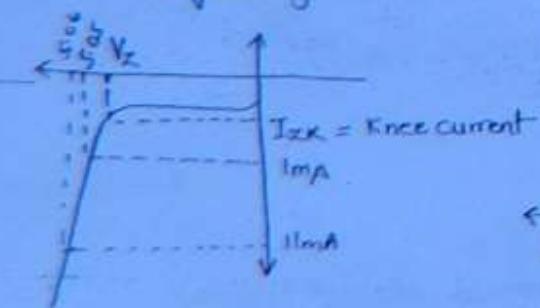
$$\% \text{ Regulation} = \frac{10-5}{5} \times 100 = 100\% \rightarrow \text{very poor.}$$

Note: For better performance of ekt, % regulation should be as low as possible.

### Types of Voltage Regulator



### Zener Voltage Regulator



\*  $I_{zK}$  = knee current or minimum current reqd. for zener diode to go in breakdown

$$\rightarrow P_{zmin} = I_{zK} \times V_z$$

(14g)

$I_{zmax}$  = maximum current across zener diode without damaging it.

$$\rightarrow P_{zmax} = I_{zmax} \times V_z \Rightarrow I_{zmax} = \frac{P_{zmax}}{V_z}$$

$V_z$  is almost constant but not exactly constant.

From plot, Slope =  $\frac{1}{R_2} = \frac{11-1}{5.05 - 5.05} = 1000 \text{ mA/V} = 1 \text{ A/V}$

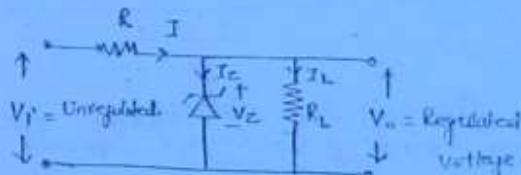
$$\Rightarrow R_2 = 1 \Omega$$

Hence, exact representation of zener diode BD is battery followed by  $R_2$ .

### - Zener Voltage Regulator -

$V_i$  = unregulated voltage

$V_o$  = Regulated "



→ for voltage regulation,

zener diode should be in BD for entire range of  $V_i$  ( $V_{min}$  to  $V_{max}$ ).

$$\therefore V_o = V_z$$

$$\rightarrow \because I = \frac{V_i - V_o}{R} \quad \text{and} \quad I = I_z + I_L$$

$$\rightarrow I_L = \frac{V_o}{R_L} = \frac{V_z}{R_L} \quad \text{Case If } R_L = \text{constant, then } I_L = \text{constant}$$

Now,  $V_i \rightarrow$  varying then  $I \rightarrow$  varying &  $I_L = \text{constant}$

$\therefore I_z = \text{varying}$

Hence, for satisfactory performance of ckt

$$I \geq I_{zK} + I_L$$

{ " Range of  $I_z$ ,  $I_{zK} \leq I_z \leq I_{zmax}$  }  
to be working in BD

$$\rightarrow \therefore I_{min} = \frac{V_{imin} - V_z}{R}, \quad I_{max} = \frac{V_{imax} - V_z}{R}$$

$$\therefore I_{min} \geq I_{zK} + I_L$$

Case 1.  $V_i = \text{constant}$ ,  $R_L = \text{varying}$

$$\Rightarrow I = \text{constant} = \frac{V_i - V_z}{R}$$

$I_L = \text{variable}$

$$\rightarrow I = I_x + I_L$$

$$\Rightarrow I_{L\max} = \frac{V_i - V_z}{R_{L\min}} \quad \text{If } R_L = R_{L\min}.$$

$$\Rightarrow I = I_{z\min} + I_{L\max} \quad \& \quad I_{z\min} \geq I_{zK}$$

$$\Rightarrow \text{When } I_{L\min} = \frac{V_i - V_z}{R_{L\max}}$$

$$\Rightarrow I = I_{z\max} + I_{L\min} \quad \& \text{ hence } \quad I_{z\max} \leq \frac{P_{z\max}}{R_z}$$

$\rightarrow$  Combining both cases, the eqn for satisfying operation of regulator circuit -

$$\frac{V_{i\min} - V_z}{R} \geq I_{zK} + \frac{V_z}{R_{L\min}}$$

$\rightarrow$  zener diode to be in B.D.

$\rightarrow$  Power dissipation across zener diode  $\leq P_{\max}$ , hence the following condition should be satisfied -

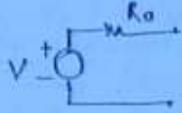
$$\frac{V_{i\max} - V_z}{R} \leq I_{z\max} + \frac{V_z}{R_{L\max}} \quad \rightarrow \text{ZD not to burn}$$

Workbook - Chap.1. Pg.29.

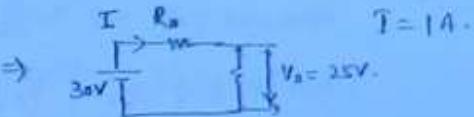
(15)

Ques.13 :-  $V_{NL} = 30V$ ,  $V_{FL} = 25V$   $\eta \cdot \text{Regulation} = \frac{30-25}{25} \times 100 = 20\%$ .

(a)



$$V = V_{NL} = 30V.$$

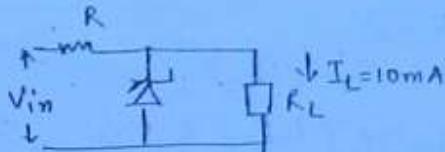


$$T = 1A.$$

$$\therefore R_{L\min} = \frac{25}{1} = 25\Omega, \text{ op resistance} \\ = R_0 = \frac{5V}{1A} = 5\Omega.$$

Ques.15 :-

(a)



$$V_2 = V_o = 10mA.$$

$$V_{in} = 30 \text{ to } 50V. \text{ for satisfactory opp} - I \geq I_{ZK} + I_L$$

$$\Rightarrow \frac{V_{min} - V_2}{R} \geq (I_{ZK} + I_L) \Rightarrow \frac{30 - 10}{R} \geq 11mA \\ \Rightarrow R \leq 1818\Omega$$

Ques.16 :-  $I_L \rightarrow 100 \text{ to } 500mA$ .

(d)  $V_{in} = 12V$

$$I_{ZK} \approx 0.$$

$$\text{When } \frac{V_{in} - V_2}{R} \geq I_{ZK} + I_{max}$$

$$\left\{ I_{max} = \frac{V_2}{R_{min}} \right.$$

$$\Rightarrow \frac{12 - 5}{R} = 0 + 500mA \Rightarrow R = 14\Omega.$$

Ques.17 :-  $V_i \rightarrow 20 \text{ to } 30V$ .

(c) load current max.  $\Rightarrow$  min zener current.

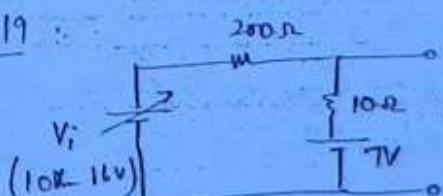
$$\frac{V_{in\min} - V_2}{R} \geq I_{ZK} + I_{max} \Rightarrow \frac{20 - 5.8}{1k\Omega} \geq 0.5mA + I_{max}$$

$$\Rightarrow I_{max} \leq 14.2 - 0.5$$

$$\Rightarrow I_{max} \leq 13.7mA.$$

Ques.19 :-

(c)



when  $V_i = 10V$  -

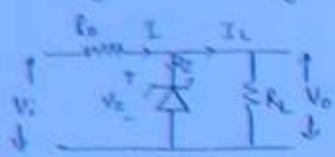
$$i = \frac{10 - 7}{210} = 1/70 A$$

$$V_o = 7 + \frac{1}{70} \times 10 = 7.14V$$

when  $V_i = 16V$  -

$$i = \frac{16 - 7}{210} = 3/70 A \Rightarrow V_o = 7 + \frac{3}{70} \times 10 = 7.14V$$

- Line Regulation using Zener diode -



$V_i \rightarrow \text{varying}, R_L = \text{constant}$

$$I = \frac{V_i - V_z}{R} , \quad I = I_z + I_L \Rightarrow I_L = I - I_z$$

When  $V_i \uparrow, I \uparrow, V_z \uparrow (\text{slightly}), I_z \uparrow \uparrow, I_L \text{ remains constant}$

When  $V_i \downarrow, I \downarrow, V_z \downarrow (\text{slightly}), I_z \downarrow \downarrow, I_L \text{ remains constant}$

(IS2)

- Load Regulation using Zener diode -

$$V_i = \text{constant}, R_L = \text{varying}. \quad I \rightarrow \text{constant} \Rightarrow I_L = I - I_z.$$

When  $R_L \uparrow, V_o \uparrow (\text{slightly}), V_z \uparrow (\text{slightly}), I_z \uparrow \uparrow, I_L \downarrow \downarrow$

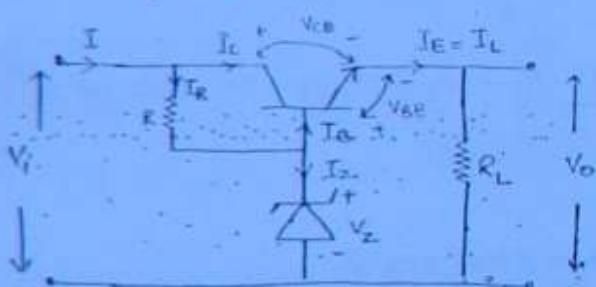
$$\therefore V_o = I_L R_L = \text{constant}$$

When  $R_L \downarrow, V_o \downarrow (\text{slightly}), V_z \downarrow (\text{slightly}), I_z \downarrow \downarrow, I_L \uparrow \uparrow$

$$\therefore V_o = I_L R_L = \text{constant}.$$

Voltage Regulation by using BJT :-

Series Voltage Regulator



→ BJT should be in active region & zener diode in breakdown region for full range of  $V_i$  from  $V_{min}$  to  $V_{max}$ .

from ohm's law -

$$\rightarrow V_o = V_z - V_{BE}. \quad (\Rightarrow \text{Regulated Voltage})$$

$$\rightarrow I_R = \frac{V_i - V_z}{R}$$

$$\rightarrow I_R = I_E + I_z.$$

$$\rightarrow I_C = \beta I_B \quad \text{and} \quad I_L = I_E = I_C + I_B \Rightarrow I_E = I_L = (1 + \beta) I_B.$$

$$\rightarrow V_E = V_i - V_o = URV - RV$$

power dissipation :-

Across Zener diode :-

Across BJT :-

$$P_z = V_z \cdot I_z \leq P_{z\max}$$

$$P_T = I_C \cdot V_{CE}$$

## Line Regulation -

$$V_i \rightarrow V_{\text{af}} \quad R_L = \text{constant} \quad I_B = I_R - I_Z.$$

(IS3)

$V_i \uparrow \Rightarrow I \uparrow, I_R \uparrow$  ( $I_C$  is not controlled by  $V_i$ , it is controlled by  $I_B$ ),  
then  $V_Z \uparrow$  (slightly)  $I_Z \uparrow \Rightarrow I_B = \text{constant} \Rightarrow I_E = \text{constant}$

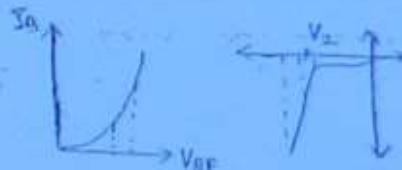
$I_E = I_L = \text{almost constant}$ , therefore  $V_o = \text{constant}$ .

## Load Regulation :

$$V_i = \text{constant}, R_L = V_{\text{af}} \quad V_o = V_Z - V_{BE}, \quad S_R = I_E + I_B \downarrow = \text{constant}.$$

$R_L \uparrow, V_o \uparrow, \left( \begin{array}{l} V_Z \uparrow \quad I_Z \uparrow \\ V_{BE} \downarrow \quad I_B \end{array} \right), I_E \downarrow \left\{ \because I_B \downarrow \right\}, I_L \downarrow$

$$\therefore V_o = I_L R_L = \text{constant}.$$



\* The circuit is in common collector configuration and hence this regulator is also called emitter follower VR.

Note:

\* Let  $I_Z$  variation if  $\Delta I_Z = 1 \text{ to } 11 \text{ mA} \Rightarrow \Delta I_Z = 10 \text{ mA}$ .

$$\left. \begin{aligned} \Delta I_L &= 10 \text{ mA} \\ \Delta R_L &= \frac{V_Z}{\Delta I_L} \end{aligned} \right\} \text{for zener diode ckt}$$

$$\left. \begin{aligned} \Delta I_Z &= 10 \text{ mA} = \Delta I_B \\ \Delta I_E &= (\mu\beta) \Delta I_B = 1000 \text{ mA} \end{aligned} \right\} \text{for BJT ckt}$$

for  $\mu = 99$

Hence, BJT ckt regulation can bear more variations in  $R_L$  as compared to zener ckt.

But, for  $V_i$  variation, same problem is present in both.

for BJT, As  $V_i \uparrow, I_R \uparrow, I_Z \uparrow$  hence for large  $V_i$  variation,  $I_Z$  will vary to  $I_{Z\max}$  and  $P_Z$  will cross  $P_{Z\max}$ .

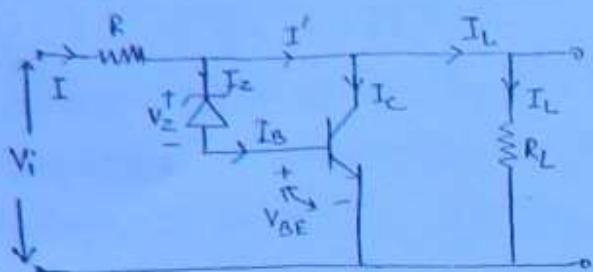
## Shunt Regulator :-

$$\Rightarrow V_o = V_z + V_{BE} \rightarrow RV$$

$$\rightarrow I = \frac{V_i - V_o}{R}$$

$$\left\{ \begin{array}{l} = \frac{URV - RV}{R} \\ \hookrightarrow \text{limiting resistance} \end{array} \right.$$

$$\rightarrow P_T = V_{CE} \cdot I_C$$



Tr  $\rightarrow$  Active  
Vz  $\rightarrow$  BD

(154)

\* Transistor is in common emitter configuration.

$$\rightarrow I_B = I_Z$$

$$\rightarrow P_Z = V_Z \cdot I_Z$$

$$\rightarrow I_C = \beta I_B$$

$$\rightarrow I = I_Z + I_C + I_L \quad \text{--- (1)}$$

Line Regulation :-  $V_i = \text{vary}$ ,  $R_L = \text{constant}$

When  $V_i \uparrow$ ,  $I \uparrow$ ,  $\left\{ \begin{array}{l} V_Z \uparrow \quad I_Z \uparrow \\ V_{BE} \uparrow \quad I_B \uparrow \end{array} \right\}$ ,  $I_C \uparrow$  {due to  $I_B$ },  $I_L$  (constant).

Ex :- let  $\Delta I = 1000 \mu A$ , then  $\Delta I_Z = \Delta I_B = 10 \mu A$ ,

$$\therefore \Delta I_C = \beta \Delta I_B = 990 \mu A$$

Since the total change is distributed b/w  $I_Z$  &  $I_C$ . {from (1)}

$$\& I_L = \text{constant}.$$

Load Regulation :-  $V_i = \text{constant}$ ,  $R_L = \text{vary}$ .

$\rightarrow V_i = \text{constant} \Rightarrow I = \text{constant}$ .

+ when  $R_L \uparrow$ ,  $V_o \uparrow$ ,  $\left\{ \begin{array}{l} V_Z \uparrow \quad I_Z \uparrow \\ V_{BE} \uparrow \quad I_B \uparrow \end{array} \right\}$ ,  $I_C \uparrow$ ,  $I_L \downarrow$ .  $\left\{ I = I_C + I_L + I_Z \text{ constant} \right\}$ .

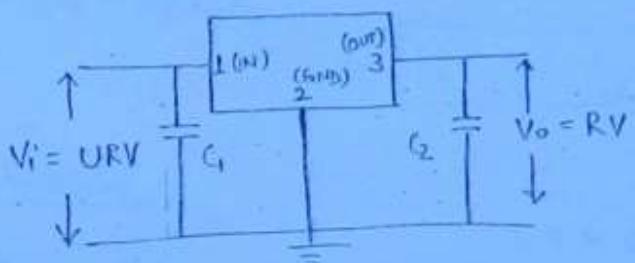
$$\therefore R_L \cdot I_L = V_o = \text{constant}.$$

$\rightarrow$  This circuit is suitable for high variation of  $R_L$  as well as  $V_i$ .

## Regulator:

(155)

- II three terminal voltage regulator, IN, OUT and GROUND.



$C_1$  &  $C_2$  is connected to bypass high frequency noise.

78 series  
(+ve o/p voltage)

	$\frac{V_o}{V_i}$
7805	+5V
7810	+10V
7812	+12V
7815	+15V
7824	+24V

79 series  
(-ve output voltage)

	$\frac{V_o}{V_i}$
7905	-5V
7910	-10V
7912	-12V
7915	-15V
7924	-24V

## Low frequency Analysis of BJT :-

(156)

h-parameters :-

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

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

$$\rightarrow h_{11} = \left. \frac{V_1}{I_1} \right|_{V_2=0} = \text{i/p impedance when o/p is sc.}$$

$$\rightarrow h_{12} = \left. \frac{V_1}{V_2} \right|_{I_1=0} = \text{Reverse voltage gain when i/p is o.c.}$$

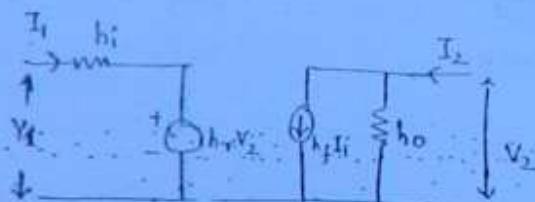
$$\rightarrow h_{21} = \left. \frac{I_2}{I_1} \right|_{V_2=0} = \text{forward current gain when o/p is s.c.}$$

$$\rightarrow h_{22} = \left. \frac{I_2}{V_2} \right|_{I_1=0} = \text{o/p admittance with i/p o.c.}$$

$h_{11} = h_i$	$h_{12} = h_r$
$h_{21} = h_f$	$h_{22} = h_o$

Hence,  $V_1 = h_i I_1 + h_r V_2$

$$I_2 = h_f I_1 + h_o V_2$$



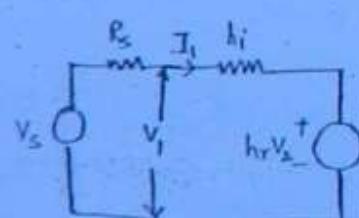
Derivation of  $A_I$ ,  $R_i$ ,  $A_V$ ,  $A_{VS}$ ,  $R_o$  :-

Current Gain  $A_I$

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

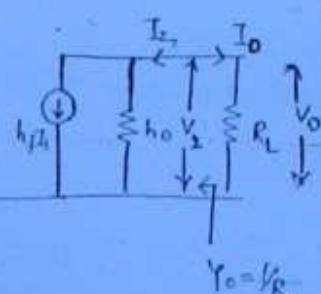
$$I_2 = h_f I_1 + h_o V_2 \quad \text{--- (1)}$$

$$V_2 = I_2 R_L = -I_2 R_L \quad \text{--- (2)}$$



from (1) & (2) -

$$I_2 (1 + h_o R_L) = h_f I_1$$



$$\Rightarrow A_I = \frac{-h_f}{1 + h_o R_L}$$

Input Resistance,  $R_i$  :-

$$\rightarrow R_i = \frac{V_1}{I_1}$$

(157)

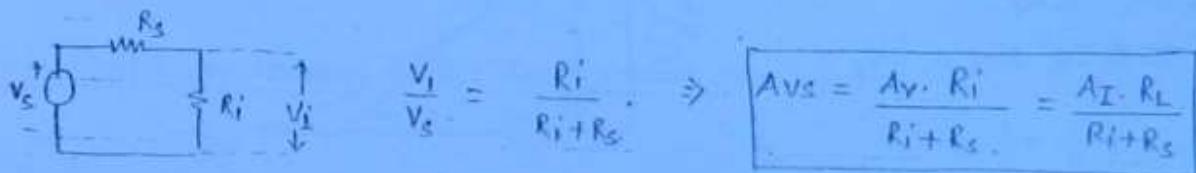
$$\begin{aligned} \rightarrow V_1 &= h_i I_1 + h_r V_2 \\ \rightarrow V_2 &= -T_2 R_L = A_T I_1 R_L \end{aligned} \quad \left\{ \begin{array}{l} V_1 = h_i I_1 + A_T I_1 R_L h_r \\ \Rightarrow R_i = h_i + h_r A_T \cdot R_L \end{array} \right.$$

Voltage Gain,  $A_v$  :-

$$A_v = \frac{V_2}{V_1} = \frac{-T_2 R_L}{R_i R_L} \Rightarrow A_v = \frac{A_T R_L}{R_i} \quad \text{or} \quad A_v R_i = A_T R_L$$

Overall voltage Gain,  $A_{vS}$  :-

$$A_{vS} = \frac{V_o}{V_S} = \frac{V_2}{V_S} = \frac{V_2}{V_1} \times \frac{V_1}{V_S} = A_v \cdot \frac{V_1}{V_S}$$



\* If current source is present instead of  $V_S$  -

$$I_S \xrightarrow{R_i} I_1 \quad A_{IS} = \frac{I_0}{I_S} = -\frac{T_2}{I_S} = -\frac{T_2}{I_1 + I_S} \cdot \frac{I_1}{T_2} = A_T \cdot \frac{I_1}{T_2}$$

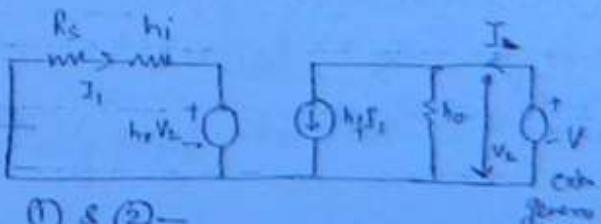
$$\frac{I_1}{I_S} = \frac{R_S}{R_S + R_i} \quad \therefore A_{IS} = A_T \cdot \frac{R_S}{R_S + R_i}$$

Output Resistance,  $R_o$  :-

$$\rightarrow I = h_f I_1 + h_o V \quad \text{--- (1)}$$

$$\rightarrow \text{KVL at i/p} \rightarrow -h_r V_2 = I_1 \quad \text{--- (2)}$$

$$\rightarrow R_o' = R_o \parallel R_L$$



from (1) & (2) -

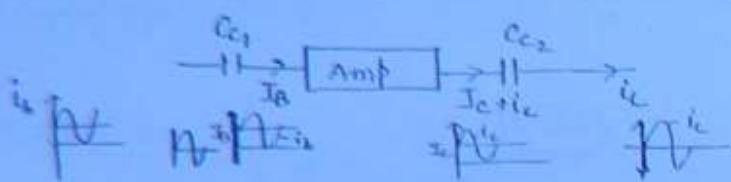
$$Y_o = \frac{I}{V} = h_o - \frac{h_f \cdot h_r}{R_S + h_i} = 1/R_o$$

05/09/2012

$$I_c + i_c = \beta I_B + \beta_{ac} i_b$$

$$|\Delta I| = \frac{i_c}{i_b}$$

(158)



Tr in active region.

\* We can neglect DC sources in ac analysis as long as they are keeping ~~active~~ in active region.

\* During AC analysis -

- All DC sources = 0, i.e., Voltage source = S.C., current source = D.C.
- Coupling capacitors  $C_{c1}$  &  $C_{c2}$  ( $C_{b1}$  &  $C_{b2}$ ) & bypass capacitor acts as S.C.

$$\beta_{dc} = \frac{|I_{cl}|}{|I_{B1}|} = h_{FE} ; \quad \beta_{ac} = \frac{|I_{ib}|}{|I_{cl}|} = h_{f\bar{e}}$$

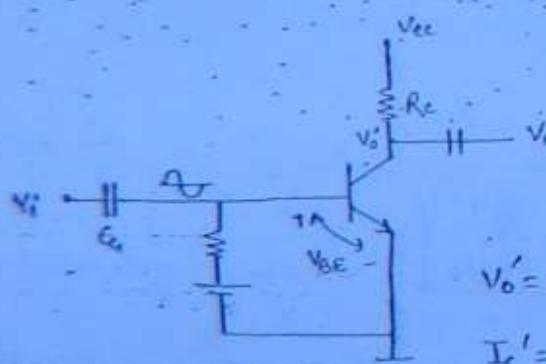
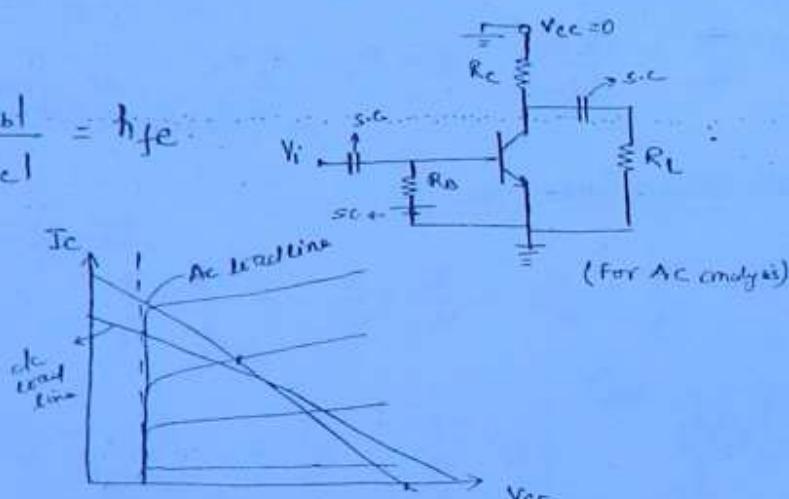
AC load line -

\* Slope of dc load line =  $-1/R_C$

$$R_L' = R_L // R_L < R_C$$

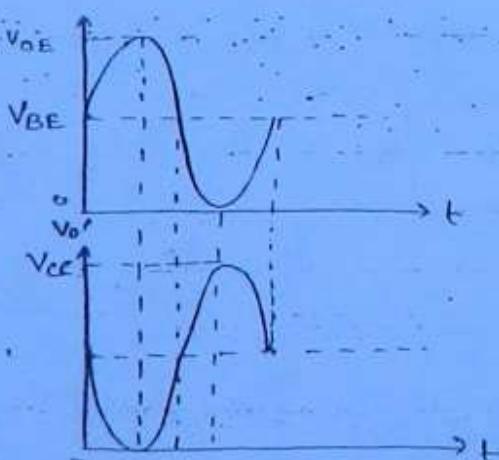
Slope of ac load line =  $-1/R_L'$

$$-\frac{1}{R_L'} = -\left(\frac{1}{R_C} + \frac{1}{R_L}\right)$$



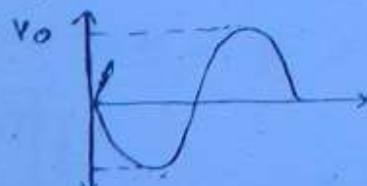
$$V_o' = V_{cc} - I_c' R_C$$

$$I_c' = I_{ct} + i_c$$



for +ve half cycle,  $V_{cc} \uparrow, I_B \uparrow \Rightarrow I_c \uparrow \Rightarrow V_o \downarrow$

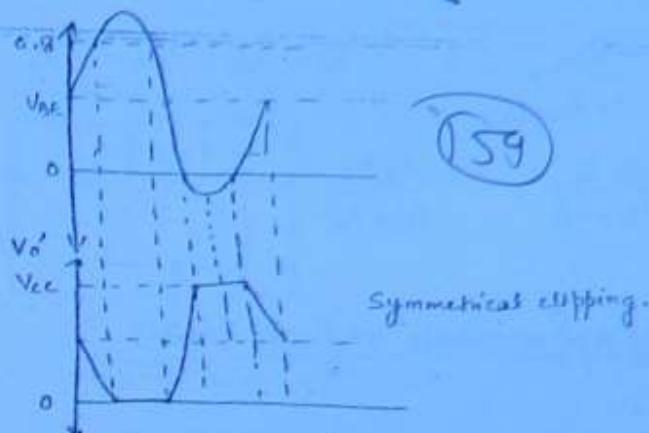
for -ve " ",  $V_{BE} \downarrow, I_B \downarrow \Rightarrow I_{ct} = V_o' \uparrow$ .



Symmetrical clipping :-

Ideally, voltage swing =  $V_{ce}$

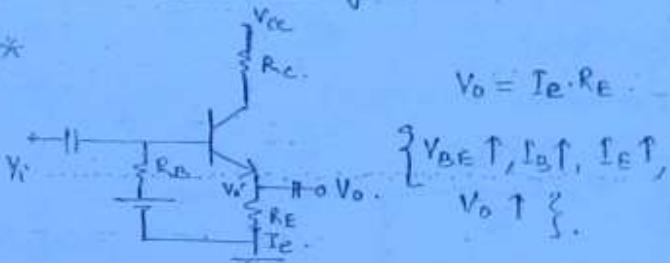
Practically, " =  $V_{cc} - V_{cesat}$



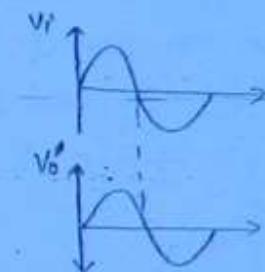
(159)

Symmetrical clipping.

Common Collector Config. :-



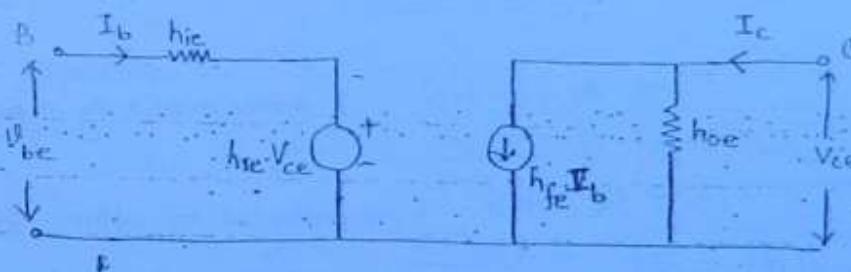
Common collector



Hence, for cc configuration,

phase shift =  $180^\circ$

Hybrid Model for Common Emitter Configuration :-



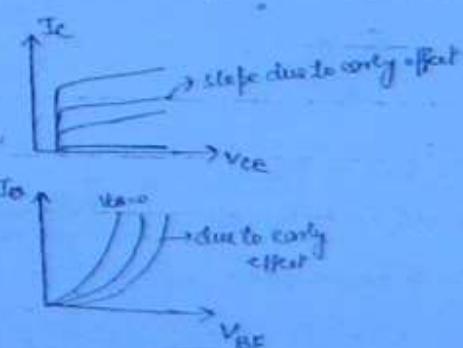
$$V_{be} = h_{ie} I_b + h_{re} V_{ce} \quad \text{--- (1)}$$

↓ due to early effect.

$$I_c = h_{fe} I_b + h_{oe} V_{ce} \quad \text{--- (2)}$$

↓ due to early effect

$$h_{oe} = \frac{1}{r_o} \Rightarrow r_o = \frac{V_A}{I_C} ; \quad V_A = \text{early voltage.}$$



\* Typical values -  $h_{ie} = 1.1 \text{ K}\Omega$ ,  $h_{re} = 2.5 \times 10^{-4}$

$h_{fe} \approx 50$ ,  $h_{oe} = 1/r_o \text{ K.}$

$$\rightarrow A_I = -\frac{h_{fe}}{1+h_{oc}R_L}; \quad R_i = h_{ie} + h_{re}A_I \cdot R_L; \quad Av = \frac{A_I R_L}{R_i}$$

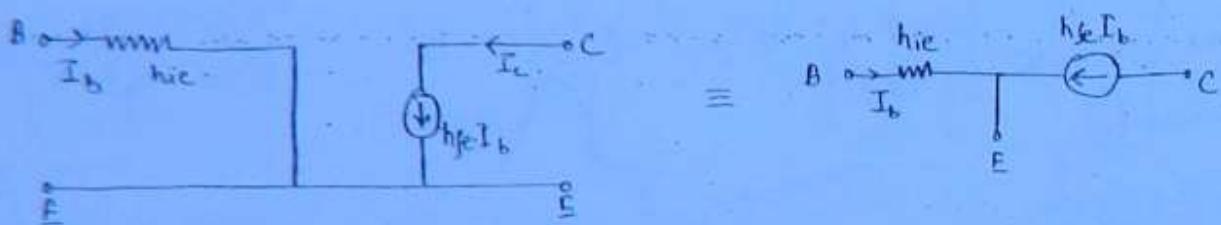
(166)

$$Av_s = \frac{Av R_i}{R_s + R_i} = \frac{A_I R_L}{R_s + R_i}; \quad Y_0 = \frac{1}{R_o} = h_{oe} - \frac{h_{re} h_{fe}}{R_s + h_{ie}}$$

### Simplified/ Approximate Hybrid Model -

→ If  $h_{oc}R_L < 0.1$ , then error in approx calculation  $\leq 10\%$ , therefore we can use approximate model, ie. we can neglect early effect.

$h_{oc} = 0, h_{oe} = 0$  ( $\Rightarrow$  admittance = 0  $\Rightarrow$  resistance =  $\infty$   $\Rightarrow$  open)



→ It is valid for CE, CC & CB configuration and for npn as well as pnp Tr.

#### \* Exact

for  $h_{oc} = 0.1$

$$\rightarrow A_I = \frac{-h_{fe}}{1+1}$$

$$\rightarrow R_i = h_{ie} + h_{re} \underbrace{A_I R_L}_{-ve}$$

$$\rightarrow Av = \frac{A_I R_L}{R_i}$$

$$\rightarrow Y_0 = h_{oe} - \frac{h_{re} h_{fe}}{R_s + h_{ie}} \approx \frac{1}{40k}$$

$$\Rightarrow R_o = 40k$$

#### Approximate

$\rightarrow A_I = -h_{fe}$   $\rightarrow$  overestimated by approx. 10%

$\rightarrow R_i = h_{ie}$   $\rightarrow$  overestimated by approx. 5%

$\rightarrow Av$  is overestimated by 5%.

$$\rightarrow Y_0 = 0$$

$\rightarrow R_o = \infty$   $\rightarrow$  overestimated (but not large)

Miller's Theorem:

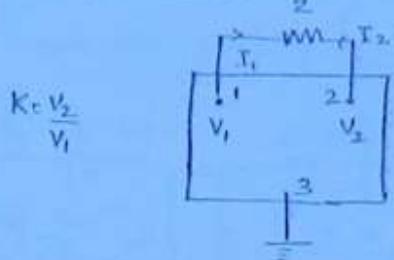


Fig. 1.

(16)

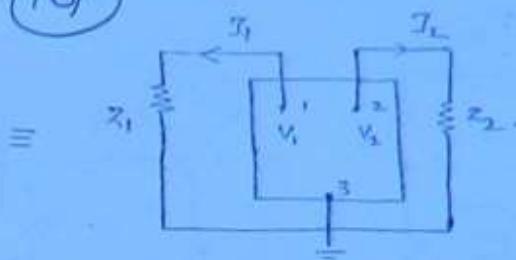


Fig. 2.

from fig. 1 —

$$I_1 = \frac{V_1 - V_2}{Z} = \frac{V_1}{Z_1} \quad (\text{from fig. 2})$$

$$\Rightarrow Z_1 = \frac{V_1 Z}{V_1 - V_2} \Rightarrow Z_1 = \frac{Z}{\frac{V_2 - V_1}{V_1}} \Rightarrow Z_1 = \frac{Z}{1-K}$$

Similarly,

$$I_2 = \frac{V_2 - V_1}{Z} = \frac{V_2}{Z_2}$$

$$\Rightarrow Z_2 = \frac{KZ}{K-1}$$

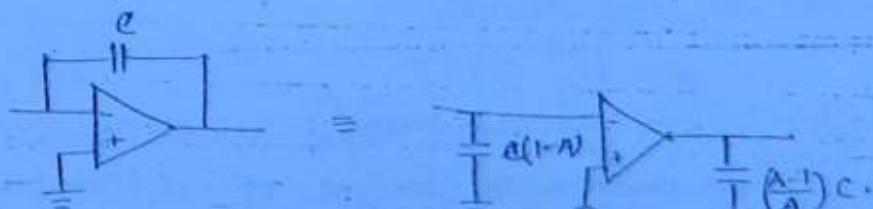
When  $Z = \text{capacitor}$  —

$$Z_1 = \frac{Z}{1-K} \Rightarrow \frac{1}{\omega C_1} = \frac{1/\omega C}{1-K} \Rightarrow C_1 = (1-K)C$$

$$Z_2 = \frac{KZ}{K-1} \Rightarrow \frac{1}{\omega C_2} = \frac{K(1/\omega C)}{K-1} \Rightarrow C_2 = \left(\frac{K-1}{K}\right)C$$

Workbook

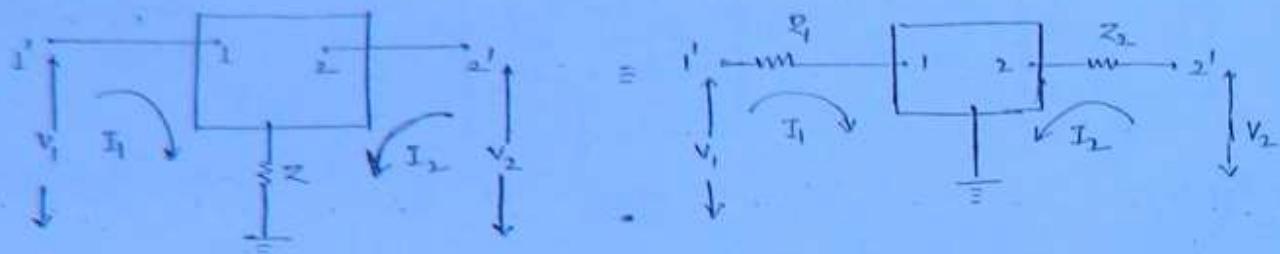
Chap 10 Q. 23



- ∴ Hence i/p & o/p capacitances increases and impedance will decrease.  
 {parallel cap.}
- 2 Due to this capacitance, i/p path will be short (low impedance) & i/p to o/p op-amp will be low  $\therefore$  overall gain will be low.

## Dual of Miller's Theorem:-

(162)



$$A_T = -\frac{I_2}{I_1}$$

from fig ① & ② -

$$V_1 = (I_1 + I_2)Z_1 = I_1 Z_1 \\ \Rightarrow Z_1 = \left[ 1 + \frac{I_2}{I_1} \right] Z_1 \Rightarrow Z_1 = (1 - A_T)Z_1$$

Similarly,  $V_2 = (I_1 + I_2)Z_2 = I_2 Z_2$

$$\Rightarrow Z_2 = \left( 1 - \frac{1}{A_T} \right) Z_2 \Rightarrow Z_2 = \left( \frac{A_T - 1}{A_T} \right) Z_2$$

## Advantage of h parameters :-

- 1) They are real nos at low frequency.
- 2) They are graphically obtained from i/p & o/p characteristics of transistor.

## Disadvantages -

- 3) All four h-parameters are temp. sensitive.

## Application -

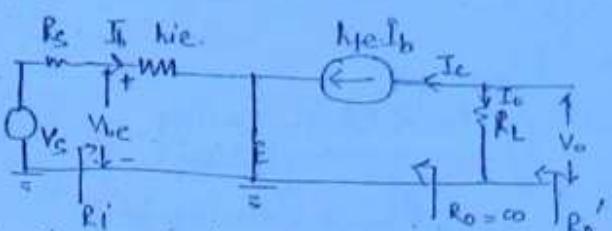
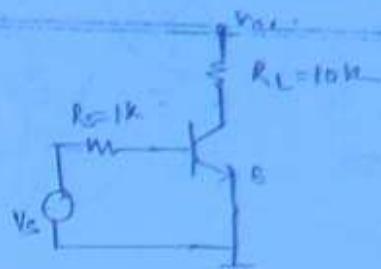
- 1) They are obtained only for small signal analysis of a transistor amplifier.

Ques If  $R_L = 10\text{ k}\Omega$ ,  $R_S = 1\text{ k}\Omega$ . find the various gains & i/p & o/p impedances.

$$h_{ie} = 11\text{ k}\Omega, h_{fe} = 50, h_{re} = h_{oe} = 0,$$

Soln: Since  $h_{oe} = h_{re} = 0$  then we can use simplified model.

(163)



$$\text{Current gain } A_I = \frac{I_o}{I_b} = -\frac{I_c}{I_b} = -\frac{h_{fe} I_b}{I_b}$$

$$\Rightarrow A_I = -h_{fe} = 50$$

O/p Resistances:

$$R_i = \frac{V_{be}}{I_b} = h_{ie} = 1.1 \text{ k}\Omega$$

$$\text{Internal voltage gain } A_v = \frac{V_o}{V_{be}} = -\frac{h_{fe} \cdot R_L \cdot I_b}{V_{be}} = -\frac{h_{fe} \cdot R_L}{R_{in}}$$

$$\Rightarrow A_v = -454$$

Voltage gain

$$A_{vs} = \frac{V_o}{V_{vs}} \quad A_{vs} = \frac{V_o}{V_s} = \frac{V_o}{V_{be}} \times \frac{V_{be}}{V_s}$$

$$\Rightarrow A_{vs} = A_v \cdot \frac{R_i}{R_i + R_o} \Rightarrow A_{vs} = -2.37$$

O/p resistances:

$$R'_o = R_o || R_L = \infty || 10k = 10k$$

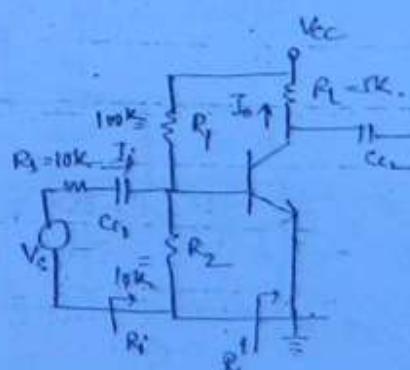
Ques: Given  $h_{fe} = 50$

$$h_{ie} = 1.1k$$

$$h_{re} = h_{oe} = 0$$

Soln:

—mA

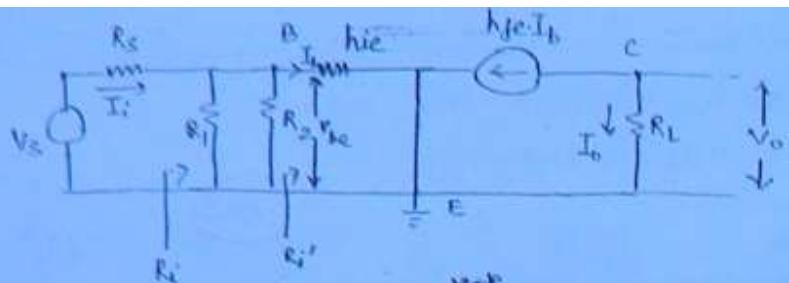


Find  
 $A_I = I_o / I_i$   
 $A_I'$ ,  $R_i'$ ,  $A_{vs}$ ,  $A_{vs}'$

$$\rightarrow A_I' = \frac{I_o}{I_b} = -h_{fe} \cdot I_b$$

$$A_I' = -h_{fe} = -50$$

(164)



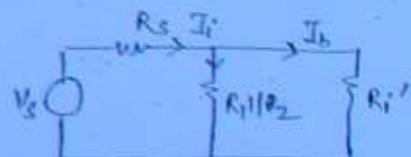
$$\rightarrow R_i' = \frac{V_{be}}{I_b} = h_{ie} = 11k\Omega$$

$$\rightarrow R_i = R_1 \parallel R_2 \parallel R_3 = (10k\Omega) \parallel (10k\Omega) \parallel (1.1k\Omega) = 980\Omega$$

$R_i' > R_i \rightarrow$  Biasing problem

$R_1 \parallel R_2$  is reducing i/p resistance

$$\begin{aligned} \rightarrow A_I &= \frac{I_o}{I_i} \\ &= -h_{fe} \frac{I_b}{I_i} \end{aligned}$$



$$I_b = \frac{R_s + R_1 \parallel R_2}{R_i' + (R_1 \parallel R_2)} \cdot I_i \Rightarrow \frac{I_b}{I_i} = \frac{9.09}{1.1 + 9.09}$$

$$\therefore A_I = -50 \times \frac{9.09}{10.19} \approx -45$$

$$\rightarrow A_V = \frac{V_o}{V_{be}} = \frac{-I_o \cdot R_L}{I_b \cdot R_i'} = -\frac{I_b \cdot h_{fe} \cdot R_L}{I_b \cdot R_i'} = -227.3$$

$$\rightarrow A_{VS} = \frac{V_o}{V_s} = \frac{V_o}{V_{be}} \times \frac{V_{be}}{V_s} = A_V \cdot \frac{V_{be}}{V_s} = A_V \cdot \frac{R_i}{R_i + R_s} = -20.5$$

every small change  
to low i/p resistance

Given :-  $h_{ie} = 11k\Omega$

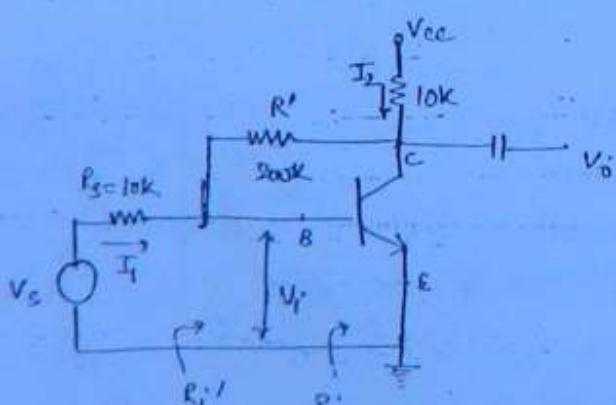
$h_{fe} = 50$

$h_{re} = h_{oe} = 0$

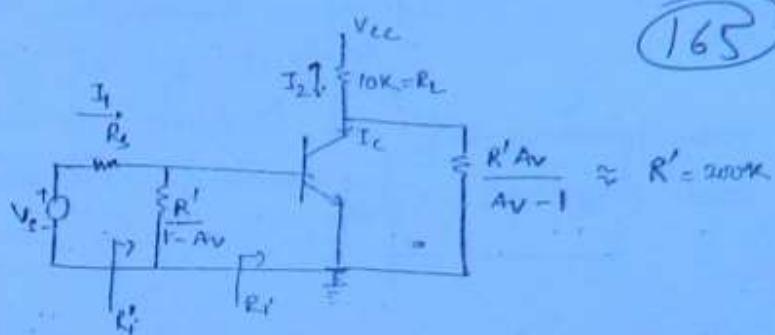
Calculate

$R_i, R_i', A_I,$

$A_I' = \frac{I_o}{I_i}, A_V, A_S.$

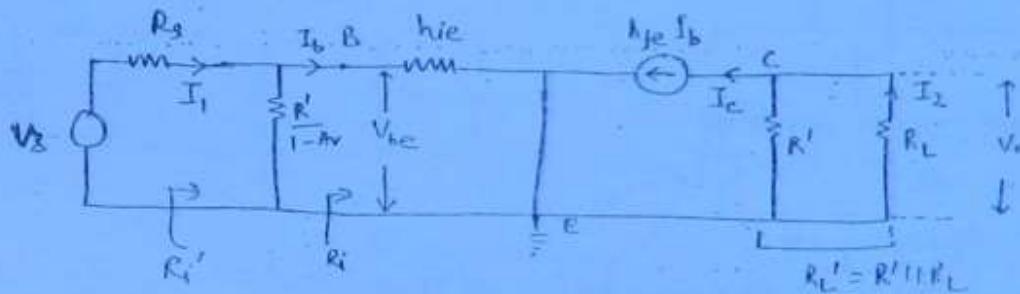


Soln :  $|Av| > 1$  for CE configuration.  
Apply Miller's theorem -



(165)

Using approximate model -



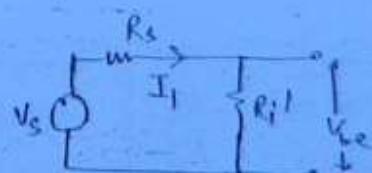
$$Av = \frac{V_o}{V_{be}} = \frac{V_o R_c}{I_b h_{fe}} = \frac{(R' R)}{(R' + R)} h_{fe} = \frac{(h_{fe} I_b) \cdot R'}{I_b \cdot h_{fe}} = \frac{h_{fe} R'}{h_{fe}} = -433$$

$$\Delta I = \frac{I_c}{I_b} = h_{fe} = 50$$

$$R_i' = \frac{V_{be}}{I_b} = h_{fe} = 11k$$

$$\frac{R'}{1 - Av} = \frac{200}{1 - (-433)} = 0.46 k\Omega$$

$$\left. \begin{aligned} R_i' &= R_i \parallel \left( \frac{R'}{1 - Av} \right) = 11 \parallel 0.46 = 0.30 k\Omega. \end{aligned} \right\}$$



$$\begin{aligned} Av_s &= \frac{V_o}{V_s} = \frac{V_o}{V_{be}} \times \frac{V_{be}}{V_s} \\ &= -433 \times \frac{R'_i}{R_i + R'_i} \end{aligned}$$

$$\rightarrow \Delta I' = \frac{V_o}{I_1} = \frac{V_o / R_L}{R'_i + R_i / (1 - Av)}$$

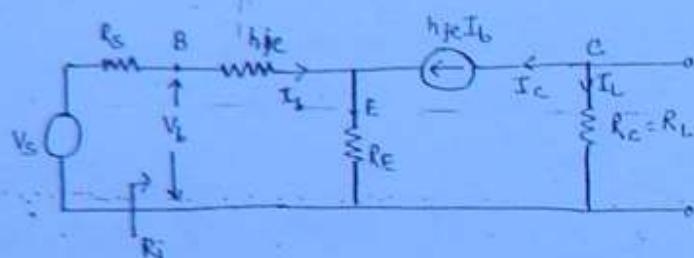
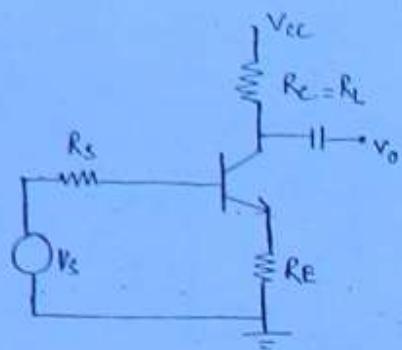
$$Av_s = -12.6 \text{ due.}$$

$$\therefore \Delta I' = \frac{V_o}{V_s} \left( \frac{R'_i + R_i / (1 - Av)}{R_L} \right) = (-12.6) \left( \frac{10 + 0.3}{10} \right) \approx -12.99 \text{. Okay}$$

## Common Emitter with unbypassed emitter resistor, $R_E$ -

(166)

→ If  $h_{FE}(R_E + R_L) \leq 0.1$  then we can use approximate model.



$$\rightarrow A_I = -\frac{h_{FE}I_b}{I_b} = -h_{FE} \Rightarrow \text{it will remain unaffected}$$

$$\rightarrow \text{if resistance } R_i = \frac{V_b}{I_b} -$$

Applying KVL -

$$V_b = I_b \cdot h_{ie} + R_E (1+h_{fe}) I_b$$

$$\therefore \boxed{R_i = h_{ie} + R_E (1+h_{fe})} \Rightarrow R_i \text{ increases}$$

$$\rightarrow A_V = \frac{V_o}{V_b} = \frac{-h_{FE} I_b \cdot R_L}{I_b [h_{ie} + R_E (1+h_{fe})]} \Rightarrow \boxed{A_V = \frac{-h_{FE} R_L}{h_{ie} + (1+h_{fe}) R_E}}$$

$A_V \downarrow$  due to -ve feedback

$$\text{If } (1+h_{fe}) R_E \gg h_{ie} \text{ & } h_{fe} \gg 1, \text{ then } \boxed{A_V = -\frac{R_L}{R_E} = -\frac{R_L}{R_E}} \text{ (approx),}$$

Due to -ve feedback, gain is highly stable as it is independent of Transistor parameters (which in turn depends on temp.).

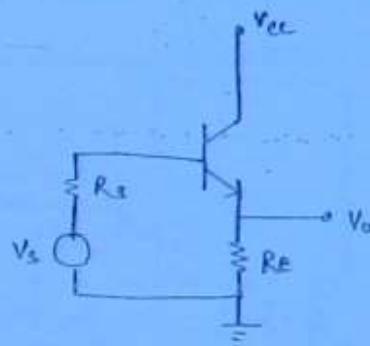
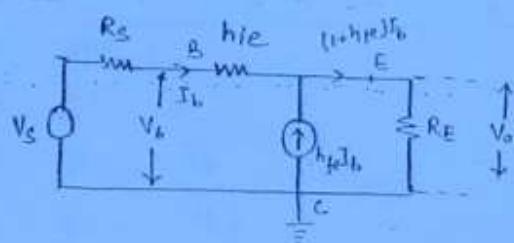
$$\rightarrow A_{Vs} = A_V \cdot \frac{R_i}{(R_i + R_S)} ; \text{ if } R_i \gg R_S \text{ then } \boxed{A_{Vs} \approx A_V \approx -\frac{R_L}{R_E}} \text{ (approx),}$$

Effect of using  $R_E$ —

- Current gain will remain unaffected.
- O/p resistance  $\text{Ro}$  by  $(1+h_{fe})R_E$ .
- Voltage gain is stabilized, i.e.,  $A_v$  is independent of any Tr. parameters.
- O/p resistance  $\text{Ro}$ . (current series feedback, check dual of Miller a/c).

(16)

Common Collector or Emitter Follower—



$$\rightarrow A_I = \frac{I_o}{I_b} = \frac{(1+h_{fe})I_b}{I_b} \Rightarrow [1+h_{fe} = A_I] \quad , \phi = 0^\circ$$

$$\rightarrow \text{O/p resistance } R_i = \frac{V_b}{I_b} \quad \text{By applying KVL}$$

$$V_b = hie I_b + (1+h_{fe}) I_b R_E$$

$$\Rightarrow R_i = hie + (1+h_{fe}) R_E. \quad (\text{high due to } R_E)$$

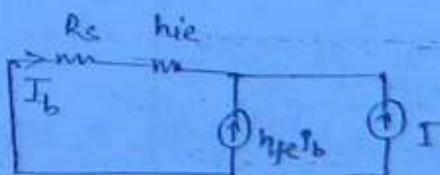
$$\rightarrow \text{Voltage Gain} : A_v = \frac{V_o}{V_b} = \frac{(1+h_{fe}) R_E \cdot I_b}{R_i I_b} \Rightarrow A_v = \frac{(1+h_{fe}) R_E}{hie + (1+h_{fe}) R_E} \quad (< 1)$$

$$\text{If } (1+h_{fe}) R_E \gg hie, \quad [A_v = 1]$$

O/p resistance :  $R_o$  —

$$V = (R_s + hie)(I + h_{fe} I_b)$$

$$I_b + h_{fe} I_b + I = 0 \\ \Rightarrow (1+h_{fe}) I_b = -I$$

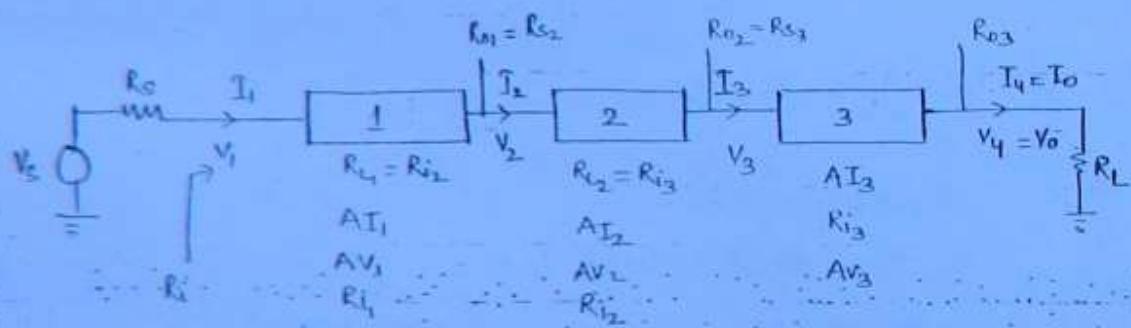


$$\therefore V = (R_s + hie)\left(I - \frac{I}{1+h_{fe}}\right)$$

$$\Rightarrow R_o = \frac{(R_s + hie)}{(1+h_{fe})}$$

	CE	CE with RE	CC
A <sub>I</sub>	-h <sub>fe</sub>	-h <sub>fe</sub>	(1+h <sub>fe</sub> )
$\Delta R_i$	h <sub>ie</sub>	$h_{ie} + (1+h_{fe})R_E$	$h_{ie} + (1+h_{fe})R_E$
A <sub>V</sub>	$\frac{A_I \cdot R_L}{R_i}$	$\frac{-h_{fe} \cdot R_C}{h_{ie} + (1+h_{fe})R_E}$	$\frac{(1+h_{fe})R_E}{h_{ie} + (1+h_{fe})R_E}$
R <sub>O</sub>	$\infty$	$\infty$	$\frac{h_{ie} + R_S}{1 + h_{fe}}$
R <sub>O'</sub>	$R_O    R_L = R_L$	$R_O    R_L = R_L$	$R_O    R_L$

### Cascaded Amplifier :-



$$\rightarrow A_V = \frac{V_o}{V_1} = \frac{V_o}{V_3} \times \frac{V_3}{V_2} \times \frac{V_2}{V_1} = A_{V_3} \cdot A_{V_2} \cdot A_{V_1}$$

$$\rightarrow 20 \log A_V = 20 \log A_{V_1} + 20 \log A_{V_2} + 20 \log A_{V_3}$$

$$\rightarrow A_I = \frac{I_o}{I_1} = \frac{I_o}{I_3} \times \frac{I_3}{I_2} \times \frac{I_2}{I_1} = A_{I_3} \cdot A_{I_2} \cdot A_{I_1}$$

$$\rightarrow 20 \log A_I = 20 \log A_{I_3} + 20 \log A_{I_2} + 20 \log A_{I_1}$$

$$\rightarrow A_p = A_V \cdot A_I$$

\*

$$= A_{V_1} \times \left( \frac{R_f}{R_f + R_1} \right) \times A_{V_2} \times \left( \frac{R_f}{R_f + R_2} \right)$$

(If there is no proper impedance matching)

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- \* If source is voltage source, then input stage should be common emf current.
- \* " " " " " " " " " " " " common base
- \* If p is delivering voltage, then last stage should be common collector current, " " " " " " " " " " " " base.
- \* All the intermediate stages should be common emitter config.
- \* CC is used in first & last stage due to high  $R_i$  & low  $R_o$  respect.
- \* CB " " " " " " " " " " " " low  $R_i$  & high  $R_o$ .

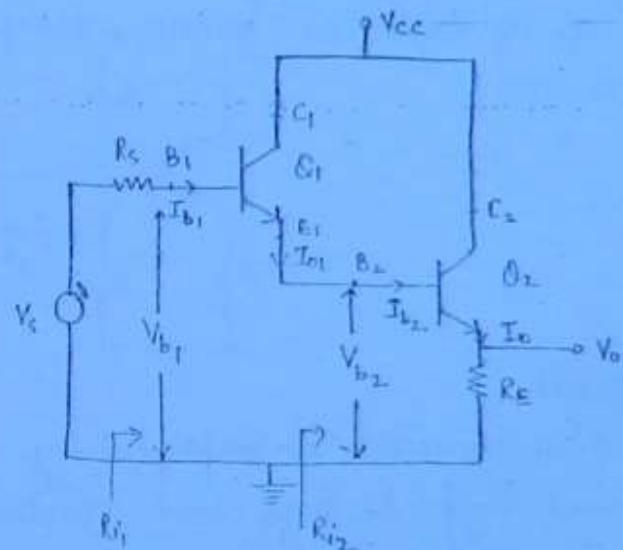
### Darlington Pair (cc-cc)

2<sup>nd</sup> stage - CC

$$A_{I_2} = \frac{I_o}{I_{b_2}} = (1 + h_{fe})$$

$$R_{i_2} = \frac{V_{b_2}}{I_{b_2}} = h_{ie} + (1 + h_{fe}) R_E$$

$R_{i_2} \approx (1 + h_{fe}) R_E$



$$A_{V_2} = \frac{A_{I_2} \cdot R_L}{R_{i_2}} \Rightarrow A_{V_2} \ll 1$$

1<sup>st</sup> stage - CC

$$\rightarrow R_{L1} = R_{i_2} = (1 + h_{fe}) R_E$$

$$\rightarrow A_{I_1} = \frac{I_{o1}}{I_{b_1}} = (1 + h_{fe})$$

$$\rightarrow R_{i_1} = \frac{V_{b_1}}{I_{b_1}} = h_{ie} + (1 + h_{fe}) R_{L1}$$

$$\Rightarrow R_{i_1} = h_{ie} + (1 + h_{fe})^2 R_E$$

$$\Rightarrow [R_{i_1} \approx (1 + h_{fe})^2 \cdot R_E] \rightarrow \text{Very large.}$$

$$\rightarrow A_{V_1} \leq 1$$

Overall current gain

$$A_I = \frac{I_o}{I_{b_1}} = \frac{I_o}{I_{b_2}} \times \frac{I_{b_2}}{I_{b_1}}$$

$$A_I = A_{I_1} \times A_{I_2}$$

$$\Rightarrow [A_I = (1 + h_{fe})^2]$$

$$\rightarrow A_V = A_{V_1} \cdot A_{V_2} \leq 1$$

For  $n$ -stages in cascade:-

Assuming  $h_{ie} = h_{re} = h_{oe} = 0$ ,

$$R_i = (1 + h_{fe})^n \cdot R_E$$

$$A_I = (1 + h_{fe})^n$$

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Advantage:-

- Very high current gain. Darlington integrated transistor pairs are commercially available with  $h_{fe}$  as high as 30,000, therefore this is also called super  $\beta$  transistor.
- Very large i/p resistance.

Disadvantage:-

- Highly expensive circuit.
- leakage current of first transistor is amplified by second, hence the overall leakage current may be high and darlington connection of 3 or more transistor is usually impractical.

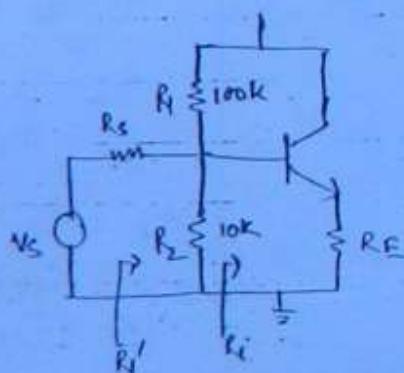
Biasing Problem :-

for CC -

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

Let  $h_{ie} = 1k$ ,  $h_{fe} = 99$ ,  $R_E = 22k$

$\therefore R_i \approx 200k$ .



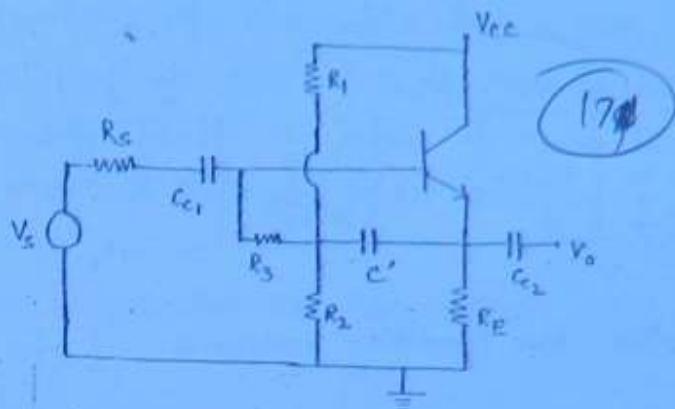
Now  ~~$R_i'$~~   $R_i' = R_1 || R_2 || R_E$  and Resultant  $R_i' < 10k$ . But we need  $R_i'$  to be high so that whole V\_S is transferred to i/p

→ Even if a darlington pair is attached with  $R_i = 2.5M\Omega$ , then also  $R_i' < 10k$ . This is called biasing problem.

## Boot Strapping :-

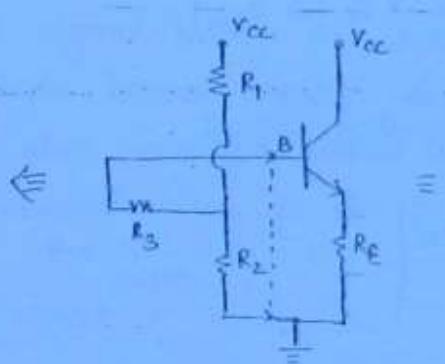
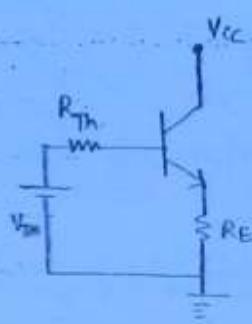
→ Value of  $C'$  should be very high so that it acts as s.c. for A<sub>c</sub>.

→ For dc analysis,  $C_C$ ,  $C_E$  &  $C'$  will act as o.c.



(17)

## DC analysis :

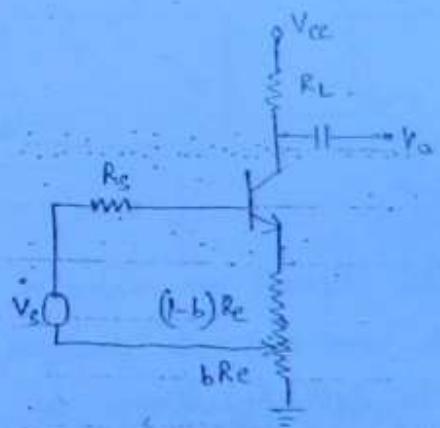
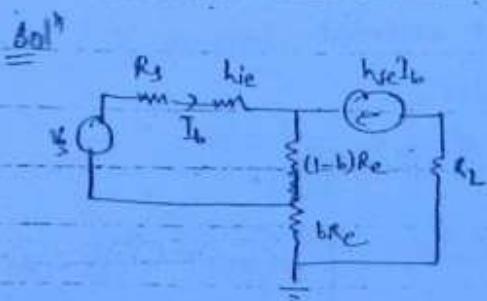


$$R_{TH} = \frac{R_2}{R_1 + R_2}, V_{cc}$$

$$R_{TH} = (R_1 || R_2) + R_3$$

Ques 1 - Calculate  $A_{Vc} = \frac{V_o}{V_s}$

$$R_i = \frac{V_s}{I_b}$$



$$V_s - I_b (R_S + h_{ie}) - (1-b) R_E (1+h_{fe}) I_b = 0$$

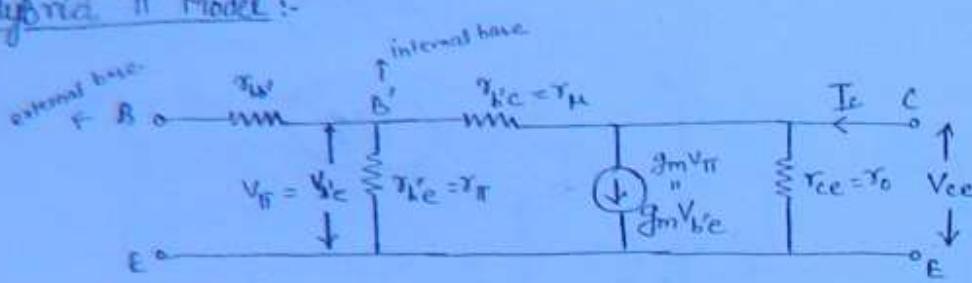
$$\therefore \frac{V_s}{I_b} = (R_S + h_{ie}) + (1-b)(1+h_{fe}) R_E = R_i$$

$$V_o = -h_{fe} I_b \cdot R_L$$

If  $b \ll 1$ , then  $R_i$  &  $R_o$  gain  $\propto$ .

$$A_{Vc} \doteq \frac{V_o}{V_s} = \frac{V_o}{I_b} \times \frac{I_b}{V_s} \Rightarrow A_{Vc} = \frac{-h_{fe} R_L}{(R_S + h_{ie}) + (1-b)(1+h_{fe}) R_E}$$

## Hybrid T Model :-



(172)

→  $r_{bb'}$  or  $r_b$  = ohmic base spreading resistance (small A,  $R \uparrow$  for base).

→  $r_{ce} = r_o$  → early effect.

→  $g_m V_{be}$  → shows dependence of  $I_C$  on  $V_B$  (or  $V_{BE}$ ).

→  $r_{be}$  → forward junction resistance.

→  $r_{be}$  → shows early effect for  $J_c$  junction  $\rightarrow$  high.

$$\rightarrow g_m = \text{Transconductance} \Rightarrow \boxed{g_m = \frac{I_C}{V_T}} ; \quad V_T = \frac{T}{11600} \text{ volt}$$

$$\rightarrow r_{be} = r_{ff} = \frac{h_{fe}}{g_m} = \frac{\beta}{g_m}$$

$$\rightarrow I_C = g_m V_{ff} + \frac{V_{ce}}{r_o}$$

early effect.

→  $g_m$  and  $r_{ff}$  in model depends on value of dc quiescent current  $I_{CQ}$  and hence provide more accurate analysis of transistor.

→ Model is applicable to both pnp & npn in w/o change of polarities.

→  $r_o$  is represented as a VCCS.

→  $r_{bb'}$  = Base region of  $T_o$  is very thin compared to emitter & collector region & its resistance lies  $10\text{ to }400\Omega$ . The ohmic resistance of E & C is usually of order of  $10\Omega$  and can be neglected in comparison to that of base region.

- $r_{TF}$  → Incremental resistance of E-B diode which is FB in active region.
- $r_p$  → It accounts for feedback from o/p to i/p due to base width modulation or early effect. The value of  $r_p$  is usually very high (several M $\Omega$ ) and will be neglected in analysis. (173)
- $r_o$  → o/p resistance and is also due to Early effect.

$$r_o = \frac{V_A}{I_{col}}$$

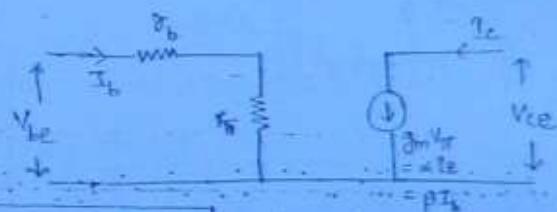
→  $g_m V_T$  → any small signal voltage  $V_{be}$  at emitter junction results in a signal collector current  $g_m V_T$  when  $V_{ce} = 0$ . BJT is represented as a VCCS when controlled current is  $g_m V_T$  & controlling voltage is  $V_{be}$ .  $g_m$  represents transconductance of Tr.

### Simplified / Approximate Model :-

$$g_m = \frac{I_{col}}{V_T}, r_{TF} = \frac{g_m}{\alpha}$$

$$r_{TF} = \frac{h_{fe}}{g_m} = \frac{\beta}{g_m}$$

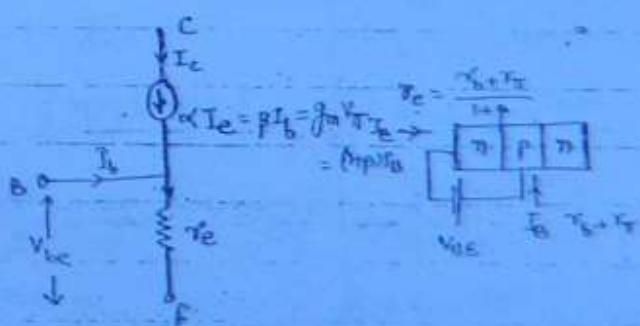
$$R_i = \frac{V_{be}}{I_b} = r_b + r_{TF}$$



### $r_e$ or T-Model :

$$r_e = \frac{V_T}{I_{col}}$$

$$\frac{V_{be}}{I_b} = R_i = (1+\beta)r_e$$



$$r_b + r_{TF} = (1+\beta)r_e = h_{ie}$$

$$g_m V_T = \beta I_b = \alpha I_e$$

$$r_b \ll r_{TF}, \quad \beta \gg 1$$

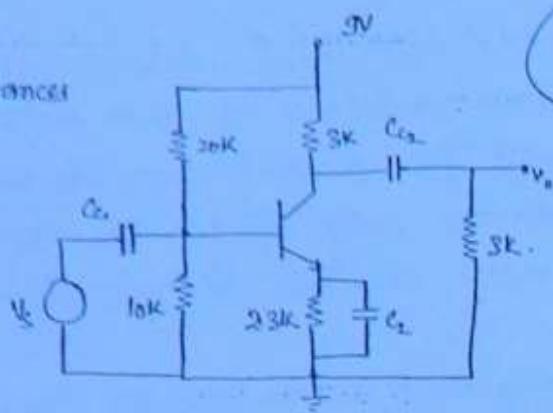
$$r_{TF} = \beta r_e = h_{ie}$$

Given  $V_{BE} = 0.7V$ ,  $\beta$  & all capacitances are very large.

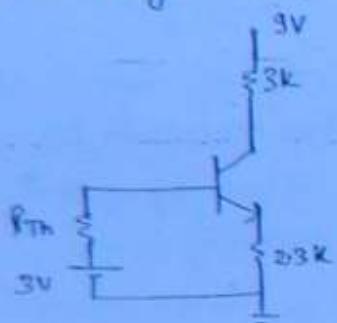
$$V_E = \frac{25mV}{10k}$$

- Find dc biasing current  $I_E$
- Find midband voltage gain.

(174)



Soln Method I  
DC analysis -



$$I_B \approx 0$$

$$3 - 0.7 = 2.3k I_E$$

$$\Rightarrow I_E = 1mA$$

$$\therefore V_E = \frac{25mV}{1mA} = 25\Omega$$

AC analysis -

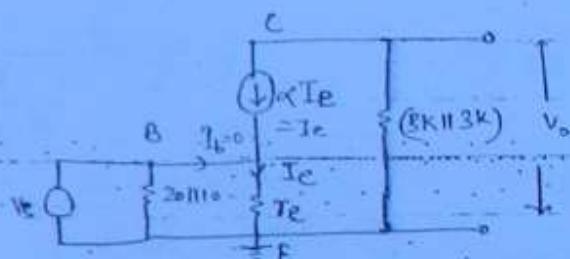
$C_1, C_2, C_3$  will act as short.

$$I_b \propto 0, \alpha = 1 \quad \left\{ \because \beta = \text{very high} \right\}$$

$$V_E = I_E V_E$$

$$V_o = (-1.5K) \propto I_E$$

$$\therefore \frac{V_o}{V_S} = -\frac{-1.5 \times 10^3}{25} = -60$$

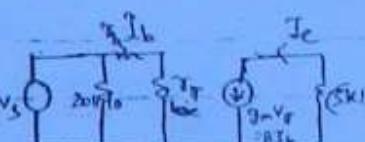


Method II

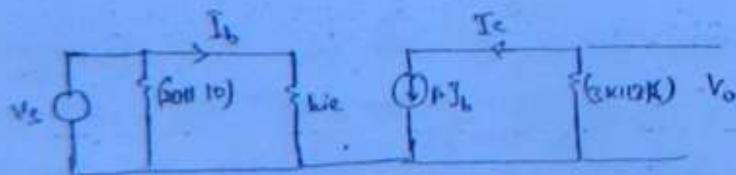
$$r_e = \frac{V_T}{|I_E|} = \frac{V_T}{|I_C|} \quad \left\{ \because \beta = \text{large} \right\}$$

$$\therefore g_m = \frac{1}{25}$$

$$\frac{V_o}{V_S} = \frac{g_m r_h (1.5K)}{r_h + r_o} = \frac{-\beta I_b (1.5K)}{I_b (\beta + 1)} = -60$$



Method III



$$\frac{V_o}{V_S} = \frac{-P I_b (1.5K)}{I_b h_{ie}} = \frac{-1.5K}{r_C} = -60$$

$$h_{ie} = \beta r_e$$

$$h_{ie}/\beta = r_e$$

Note

\* If  $R_E$  is unbypassed -

$$\frac{V_O}{V_S} = \frac{-R_E}{R_E + r_E} \approx \frac{-R_E}{R_E}$$

$\rightarrow$  (overestimated)

\* If  $R_E$  is bypassed -

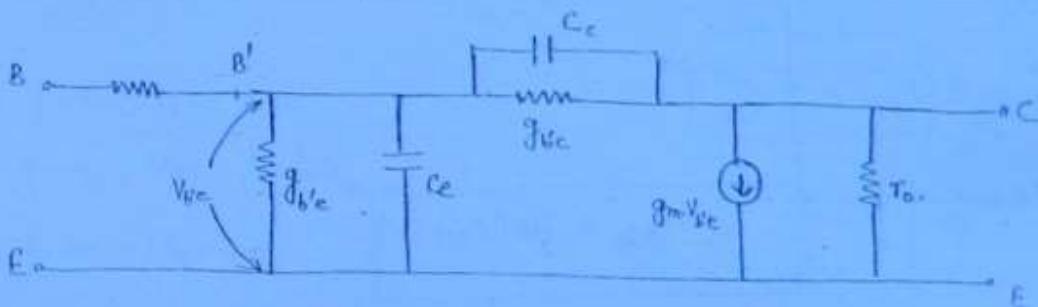
$$\frac{V_O}{V_S} = -\frac{R_E}{r_E}$$

(175)

\* If an extra  $R_L$  is present in o/p -

$$\begin{aligned} R_E \text{ unbypassed} &= \frac{-(R_E \parallel R_L)}{R_E} \\ R_E \text{ bypassed} &= \frac{-(R_E \parallel R_L)}{r_E} \end{aligned}$$

## High Frequency Analysis of BJT :-



Giacchetto Model

$$\rightarrow g_{BE} = \frac{1}{r_{BE}} = \frac{\beta_m}{h_{FE}} \quad \left. \right\} \quad \because r_{BE} \gg r_{BC} \quad \Rightarrow g_{BC} \ll g_{BE}$$

$$\rightarrow g_{EC} = \frac{1}{r_{BC}}$$

$$\rightarrow g_m = \frac{|I_C|}{V_T}$$

$$\rightarrow C_{IT} = C_E = C_D \rightarrow \text{Diffusion capacitance}$$

$$\rightarrow C_E = C_I = C_{ob} = C_{pi} \rightarrow \text{Transition capacitance}$$

Typical Values

$$\rightarrow g_m = 50 \text{ mA/V}$$

$$\rightarrow r_{BE} = r_{BB} = 100 \Omega$$

$$\rightarrow r_{BC} = r_{II} = 1K$$

$$\rightarrow r_{EC} = 4 \text{ M}\Omega$$

$$\rightarrow r_o = r_{EE} = 80K$$

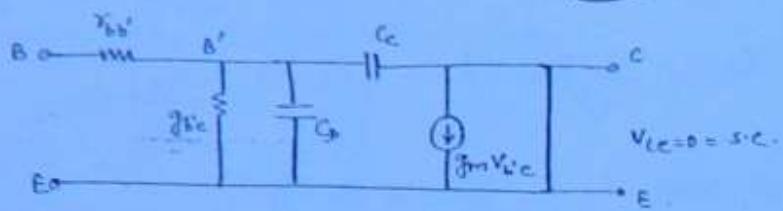
$$\rightarrow C_C = 5 \text{ pF}$$

$$\rightarrow C_b = C_E = 100 \text{ pF}$$

## CE short circuit current gain -

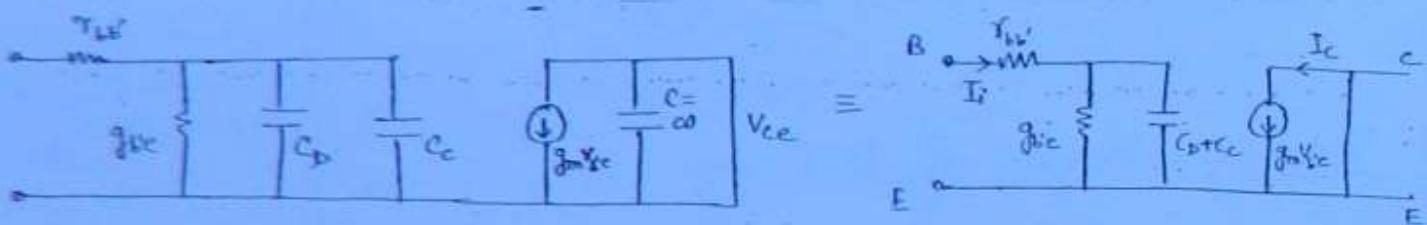
(176)

$$\rightarrow A_V = \frac{V_{CE}}{V_{BE}} = 0.$$



→ Now applying Miller's theorem -

$$C_1 = C_C (1 - A_V) = C_C ; \quad C_2 = \frac{C_C (A_V - 1)}{A_V} = \infty \text{ (short).}$$



i/p capacitance  $\therefore [C_i = C_B + C_C]^{**}$

i/p conductance  $\therefore \left[ Y_i = \frac{I_i}{V_{BE}} = g_{BE} + j\omega(C_B + C_C) \right]^{**}$

Current gain -

$$A_I = \frac{I_o}{I_i} = -\frac{g_m V_{BE}}{V_{BE} Y_i} \therefore A_I = -\frac{g_m}{Y_i}$$

$$\therefore \boxed{A_I = \frac{-g_m}{g_{BE} + j\omega(C_B + C_C)}}$$

→ This will act as LPF at higher frequency.

Rearranging,  $A_I = \frac{-g_m / g_{BE}}{1 + j\omega \frac{(C_B + C_C)}{g_{BE}}} , \text{ Now } \because \gamma_{BE} = \frac{1}{g_{BE}} = \frac{h_{FE}}{g_m} \therefore -g_m / g_{BE} = h_{FE}.$

$$\therefore \boxed{A_I = \frac{-h_{FE}}{1 + j\omega \frac{(C_B + C_C)}{g_{BE}}}}^{**}$$

$$\rightarrow |A| = \frac{h_{fe}}{\sqrt{1 + \left[ \frac{w(c_b + c_c)}{g_{be}} \right]^2}}, \quad |A|_{max} = h_{fe} \text{ at } w=0. \quad (177)$$

$$\rightarrow \text{At } w=w_p, |A|= |A|_{max}/\sqrt{2} \Rightarrow \frac{h_{fe}}{\sqrt{2}} = \frac{h_{fe}}{\sqrt{1 + \left[ \frac{w_p(c_b + c_c)}{g_{be}} \right]^2}}$$

On solving, -

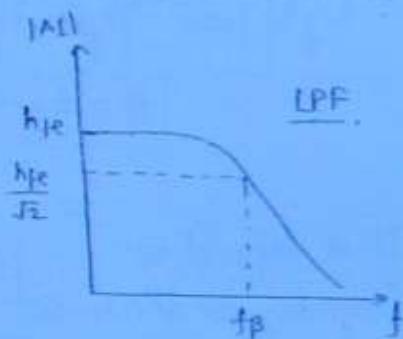
$$\omega_p = 2\pi f_p = \frac{g_{be}}{c_p + c_c}$$

$$f_p = \frac{1}{2\pi R_{be} \cdot (c_p + c_c)}$$

$f_p = 3\text{dB}$   
cutoff freq.

Hence,

$$|A_1| = \frac{-h_{fe}}{1 + j(w/w_p)} = \frac{-h_{fe}}{1 + j(f/f_p)}$$



$$\rightarrow \text{At } f=0, |A_1|=h_{fe}$$

$$\text{At } f=f_p, |A_1|=h_{fe}/\sqrt{2}$$

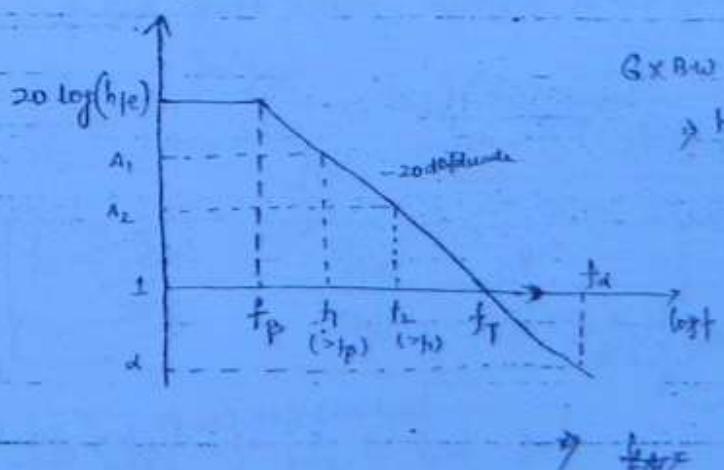
$$\text{At } f=\infty, |A_1|=0$$

$\rightarrow$  At  $f=f_T$ ,  $|A_1|=1$   $\rightarrow$  frequency till transistor will act as amplifier

$$1 = \frac{h_{fe}}{\sqrt{1 + \left( \frac{f_T}{f_p} \right)^2}} \Rightarrow \left( \frac{f_T}{f_p} \right)^2 = h_{fe}^2 - 1 \Rightarrow \left( \frac{f_T}{f_p} \right)^2 \approx h_{fe}^2$$

$$\Rightarrow f_T = h_{fe} \cdot f_p \quad \Rightarrow f_T \gg f_p$$

Bode Plot:



$$G \times B.W = \text{constant}$$

$$\Rightarrow h_{fe} \cdot f_p = A_1 \cdot f_1 = A_2 \cdot f_2 = f_T = \alpha \cdot f_c$$

$$f_T = \text{unity gain bandwidth product}$$

$$f_T = h_{fe} \cdot f_p = \frac{h_{fe} \cdot g_{be}}{2\pi (c_b + c_c)}$$

20V

$$V_B = \frac{R_1}{R_1 + R_2} \times 20 = \frac{10}{20+10} \times 20 = 6.67V (\approx 10V)$$

Hence, zero bias is not on BD.

R<sub>ce</sub>, I<sub>C</sub> = 0, P<sub>C</sub> = 0, V<sub>O</sub> = 6.67V ( $\neq$  10V)

$$f_T = \frac{g_m}{2\pi(C_S + C_C)}$$

(178)

$$f_P << f_T < f_A$$

; f<sub>A</sub> = frequency at which gain  $< 1/\mu$ , gain of CB = 1

$$h_{FE} \cdot f_P = \frac{h_{FE}}{1+h_{FE}} \cdot f_A \Rightarrow f_A = (1+h_{FE}) f_P$$

$$\therefore f_T = h_{FE} f_P$$

$f_P$   $\Rightarrow$  'P' cutoff frequency and also called as bandwidth of CE at high freq. Typical value -

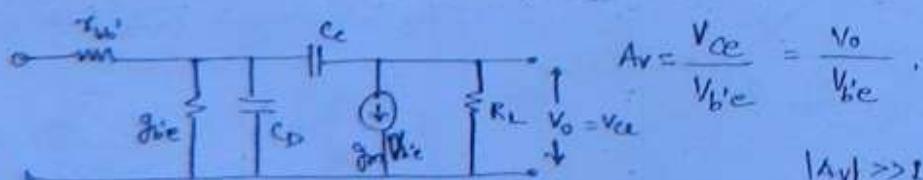
$$f_P = 1.6 \text{ MHz.}$$

$f_T \Rightarrow$  It is defined as

- $\Rightarrow$  frequency at which SC CE current gain attains unit magnitude.
- $\Rightarrow$  highest freq. upto which CE tr. will be working as an amplifier.
- $\Rightarrow$  freq. where CE tr.  $\beta$  reduces to unity.
- $\Rightarrow$  Unity gain bandwidth product of CE tr. and thus ( $g \times BQ$ ) is limited by junction capacitance.

$f_A \Rightarrow$  'A' cutoff frequency. It is also called BW of CB transistor at high frequency. The BW of CB is always greater than BW of CE or BW of CC. tr.

Common Emitter with Resistive load R<sub>L</sub> :-



|Av| > 1 for CE Tr.

Apply Miller's theorem -

$$C_2 = C_C (1 + A_V)$$

$$C_1 = \frac{C_C (A_V - 1)}{A_V} \approx C_C \quad \left\{ \text{if } A_V \gg 1 \right\}$$

$$Z_{C_1} = \frac{1}{2\pi f C_1} \quad : C_1 \equiv \text{pf} \quad \text{if } f \equiv \text{MHz} \Rightarrow Z_C \approx 10^6 \Omega \Rightarrow Z_C \parallel R_L = R_L$$

Therefore, approximate model -

Now,

$$A_V = -g_m V_{BE} \cdot R_L$$

$$\Rightarrow A_V = -g_m R_L$$

$$\therefore C_2 = C_C (1 + g_m R_L)$$

$\rightarrow$  Input capacitance :-

$$C_i = C_B + C_C (1 + g_m R_L) \quad \text{using}$$

$\rightarrow$  Input conductance :-

$$Y_i = g_{BE} + j\omega C_i \quad \text{using}$$

$\rightarrow$  Due to Miller's effect,  $C_i \uparrow$ ,  $Y_i \downarrow$ ,  $Z_f \downarrow \Rightarrow$  Gain  $\downarrow$ .

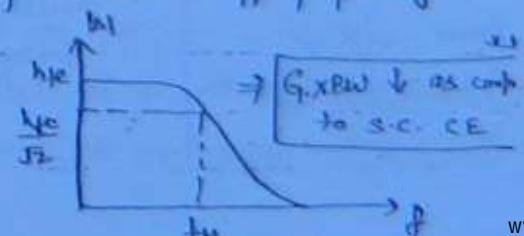
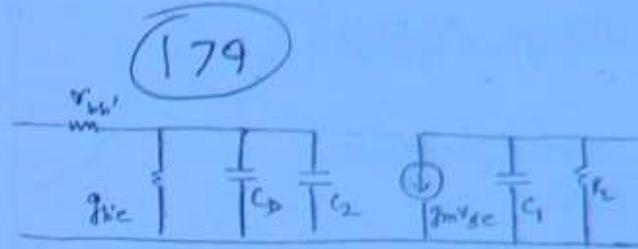
$$\rightarrow \begin{cases} I_o = -g_m V_{BE} \\ I_i = Y_i, V_{BE} \end{cases} \quad \left\{ \because A_I = \frac{I_o}{I_i} = \frac{-g_m}{Y_i} \Rightarrow A_I = \frac{-g_m}{g_{BE} + j\omega C_i} \right.$$

$$\therefore A_I = \frac{-g_m / g_{BE}}{1 + j\omega C_i / g_{BE}} \Rightarrow A_I = \frac{-h_{FE}}{1 + j(\omega / \omega_H)} = \frac{-h_{FE}}{1 + j(f_H / f_H)}$$

$$\rightarrow \boxed{f_H = \frac{g_{BE}}{2\pi C_i} = \frac{g_{BE}}{2\pi (C_B + C_C (1 + g_m R_L))}} \quad ; \quad f_H = 3\text{dB cut-off frequency}$$

$$\rightarrow \boxed{f_H < f_P} \quad ; \quad \boxed{f_P = \lim_{R_L \rightarrow 0} f_H}$$

$$\rightarrow \boxed{G_{BW_{RL}} < G_{BW_{SC}}}$$



## Multistage Amplifiers :-

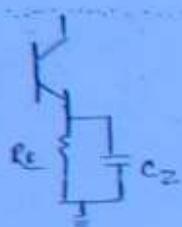
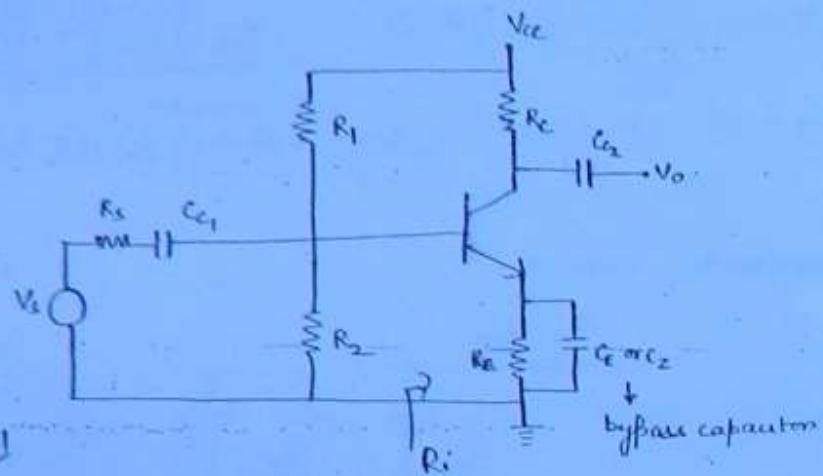
(180)

### RC Coupled Amplifier :-

#### i) Single stage -

→ Audio freq. amplifier  
(20Hz - 20KHz).

→ CE configuration, i.e.,  
180° phase shift



$$R_i = h_{ie} + (1+\beta)R_E$$

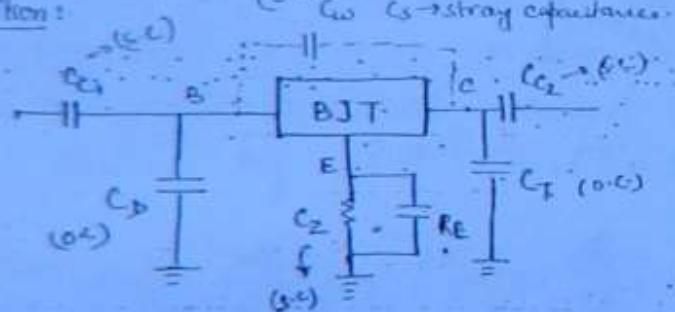
$$R_o = R_L + k_{fP}$$

$R_i \uparrow, A_v \downarrow$

$A_v \uparrow, R_i \downarrow$

\*  $C_2$  is bypass capacitor & its value should be high so that it will act as short for AC.

Ideal condition:



$C_{C1}, C_{C2}, C_2 \rightarrow$  very high

$C_P, C_T \rightarrow pF$

$C_S, C_W \rightarrow 10^{-14} F$

$$Z_C = \frac{1}{2\pi f C}$$

#### Low Frequency

- As  $f \downarrow, Z_C \uparrow$ .
- All capacitive impedances  $\rightarrow \infty$ , and they will act as O.C. Gain will  $\downarrow$  due to  $C_1, C_2$  &  $C_2$ .

#### Mid frequency

- $Z_C$  is not decided by  $f$ , decided by the value of  $C$ . Hence, ideal condn achieved. and gain is independent of freq.

#### High frequency

- As  $f \uparrow, Z_C \downarrow$ ,
- All capacitive Impedance  $\rightarrow 0$  and they will act as short. Gain will  $\downarrow$  due to  $C_D, C_T, C_S, C_W$ .

→ frequency Response curve :-

Important Point :-

→ It is audio freq. amplifier.

→ Single stage RC couple introduce a phase shift of  $180^\circ$  & two stages introduce  $360^\circ$  or  $0^\circ$ .

→ Coupling capacitors ( $C_4$  &  $C_2$ ) are also called dc blocking capacitors ( $C_b$ ,  $C_{b1}$ ) and are used to couple ac signals and simultaneously block dc current or biasing current.

→ By using emitter resistor, w/o a bypass capacitor, there will be a -ve feedback across  $R_E$  and this reduces the voltage gain and is ip resistance of amplifier.

→ Bypass capacitor ( $C_2$  or  $C_b$ ) is used to bypass ac signal current through it for dc current or biasing current,  $C_2$  is open.  
By using  $C_2$ , -ve feedback (due to ac signal) across  $R_E$  is eliminated, therefore voltage gain  $\uparrow$  and ip resistance  $\downarrow$ .

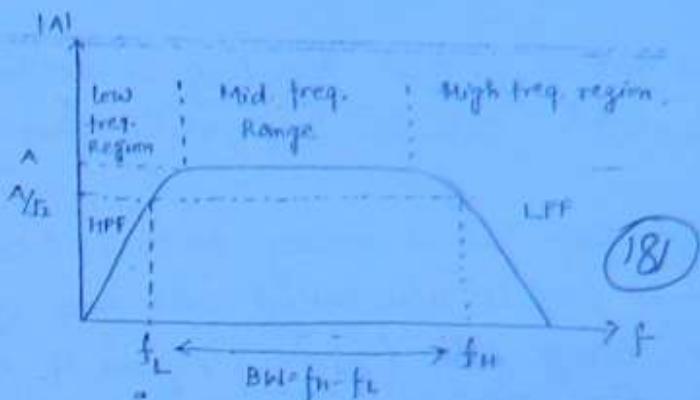
→ In an amplifier, for better performance,  $R_E$  &  $C_E$  combination is used.

freq. Response curve :-

→ The fall of gain in low freq. region is due to effect of  $C_C$ ,  $C_{C2}$  and  $C_E$ .

→ In mid frequency region, all coupling & bypass capacitor will be treated as ac short. All junc<sup>n</sup> capacitors ( $C_4$  &  $C_7$ ), cutting capacitor ( $C_W$ ) & stray capacitor ( $C_S$ ) will be treated as open.

→ The gain of amp is more & almost independent of f. at mid freq. and hence the amplifier analysis is generally done at mid freq. range.



→ We fall of gain at high freq is due to the effect of junction capacitor ( $C_J$ ,  $C_T$ ) &  $C_{w.c.}$  and early effect.

(182)

→ High freq. fall is mainly due to  $C_J$  &  $C_{w.c.}$ .

→ Cutoff freq. is also called 3dB freq. or half power freq.

→ At cutoff freq. ( $f_H$  or  $f_L$ ), gain of amp reduces to 70.7% of peak value, i.e.,  $|A_{mid}|/\sqrt{2}$ . and o/p power of amplifier is reduced to 50% of peak value.

→ At cutoff frequency, the relative gain of amp is reduced by 3dB from its peak value.

Bandwidth :-

→ It is the band of i/p signal frequencies where the gain of amp. is almost constant.

$$\boxed{BW = f_H - f_L}$$

→ Larger BW indicates better reproduction of i/p signal.

→ In an amplifier, gain-bandwidth product is always constant, i.e., when one rises, other falls & vice versa.

Disadvantage :- Smaller gain  $\times$  BW.

Note - Amplifiers are connected in cascade to get larger gain  $\times$  BW product.

Calculation of  $f_L$  :-

(183)

→  $f_L$  due to  $C_C$  :- Assume  $C_E \& C_O \rightarrow \infty$  & acts as  $\infty$ .

→ We can replace transistor with its input resistance.

$$\rightarrow R_i = h_{ie} = \beta r_e = T_{IR} + r_b$$



$$\rightarrow R_i' = R_i \| R_2 \| R_1 \approx R_i$$

$$\therefore f_L = \frac{1}{2\pi R_{eq} C} \Rightarrow \boxed{f_L = \frac{1}{2\pi (R_S + R_i') \cdot C_c}}$$

→ for high BW,  $C_c$  &  $R_i'$  values should be as high as possible.  
( $R_S$  should not be high as it will ↑ posse in i/p)

→  $f_L$  due to  $C_O$  :- Assume  $C_E \& C_C \rightarrow \infty$  & acts as  $\infty$ .



$$\boxed{f_L = \frac{1}{2\pi (R_C + R_L) \cdot C_o}}$$

\* If  $f_L$  due to  $C_C$  &  $C_O$  is different then take the bigger value.

\* If  $f_H$  due to  $C_W/C_S$  or  $C_D$ , then " " " smaller value.

### Low frequency Analysis

amplifier

→ RC coupled acts as HPF for low freq.

$$\rightarrow \text{Total phase shift} = 180^\circ + \tan^{-1} \left( \frac{f_L}{f} \right) \xrightarrow{\substack{\downarrow \text{due to E} \\ \text{config}}} \Rightarrow \boxed{\text{At } f=f_L, \phi_T = 225^\circ}$$

$$\rightarrow H = \frac{1}{1 - j(\frac{1}{f}f_T)}$$

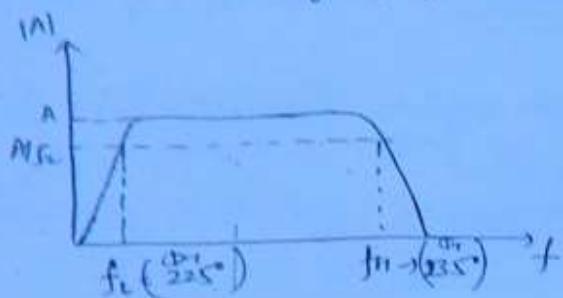
## High freq. analysis:-

(184)

$\rightarrow A = \frac{1}{1+j(\frac{f}{f_H})}$  : RC coupled amplifier acts as LPF at high freq.

$\rightarrow \phi_T = 180 - \tan^{-1} \left( \frac{f}{f_H} \right)$ .

$\rightarrow [At f=f_H, \phi_T = 135^\circ]$



## Cascaded Amplifier / Multistage Amplifier :-

Amplifiers are connected in cascade to get larger gain × BW.

When amplifiers are connected such that o/p of one is given to i/p to other, they are said to be cascaded.

When amplifiers are cascaded, proper impedance matching must be provided in b/w stages so that —

1) o/p will not be distorted.

2) More power will be transferred from one to another stage.

~~Note~~ If mismatch is more in amplifier, o/p will be highly distorted.

## Different types of coupling -

i) RC coupling → (for voltage amplifiers)

ii) Transformer coupling → (for power amplifiers)

iii) Direct coupled → (basically used for dc amplification).

In a multistage amplifier,  $G \times BW = \text{constant}$ .

~~Note~~ →  $G \times BW$  of two stage amplifier is greater than that of single stage amp.

→ In multistage amp, BW reduces.

## (185<sup>th</sup>)

### Bandwidth of Multistage Amplifier :-

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- $BW^* = f_H^* - f_L^*$       }       $BW^* \rightarrow$  Bandwidth of multistage Amp.
- $BW = f_H - f_L$       }       $BW \rightarrow$  "      " single stage amp.
- $f_H/f_L^* \rightarrow$  High/Low 3dB cutoff freq. of multistage A
- $f_H/f_L \rightarrow$  "      "      "      " single stage

Case-  $n$ - identical non-interacting (proper impedance matching) stages in cascade

#### Derivation of $f_H^*$ :-

Gain for individual stage ,  $|A| = \frac{1}{\sqrt{1 + (f/f_H)^2}}$

for  $n$ - such stages,  $|A^*| = \left[ \frac{1}{\sqrt{1 + (f/f_H)^2}} \right]^n$

→ At  $f=0$ ,  $|A^*_{max}| = 1$ .

→ At  $f = f_H^*$ ,  $|A^*| = 1/\sqrt{2} \Rightarrow \frac{1}{\sqrt{2}} = \left[ \frac{1}{1 + (f^*/f_H)^2} \right]^{n/2}$

$$\Rightarrow f_H^* = f_H \left[ \sqrt[2]{2^{n/2}-1} \right]$$

$$\rightarrow f_H^* < f_H, \quad n=2, \quad f_H^* = 0.64 f_H$$

$$n=3, \quad f_H^* = 0.51 f_H$$

#### Derivation of $f_L^*$ :-

Gain for individual stage,  $|A| = \frac{1}{\sqrt{1 + (f/f_L)^2}}$

for  $n$ - such stages,  $|A^*| = \left[ \frac{1}{\sqrt{1 + (f/f_L)^2}} \right]^n$

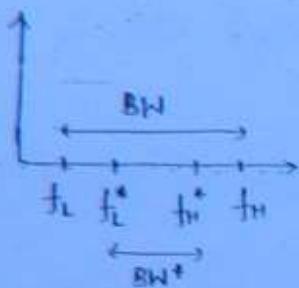
$\rightarrow$  At  $f = \omega$ ,  $|A^*|_{\max} = 1$ .

(186)

$$\rightarrow \text{At } f = f_L^*, |A^*| = 1/\sqrt{2} \Rightarrow \frac{1}{\sqrt{2}} = \left[ \frac{1}{\sqrt{1 + (f_L/f_L^*)^2}} \right]^n$$

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

$$\rightarrow f_L^* > f_L ; \begin{cases} n=2 & f_L^* = 1.56 \\ n=3 & f_L^* = 1.96 \end{cases}$$



$BW^* < BW$   $\rightarrow$  There is shrinkage in BW.

$$\text{Approximate Bandwidth: } \rightarrow BW = f_H - f_L \approx f_H \quad (f_H \gg f_L)$$

$$\rightarrow BW^* = f_H^* - f_L^* \approx f_H^* \quad (f_H^* \gg f_L^*)$$

$$\Rightarrow BW^* = (\sqrt{2^{1/n} - 1}) \cdot BW ; (BW^* < BW)$$

Case :- n - non-identical interacting (ie, no proper impedance matching).  
Stages in cascade.

$$\rightarrow \frac{1}{f_H^*} = 1.1 \times \sqrt{\frac{1}{f_{H_1}^2} + \frac{1}{f_{H_2}^2} + \dots + \frac{1}{f_{H_n}^2}}$$

Note. Disadvantages

$\rightarrow BW \downarrow$

$\rightarrow$  Rise time  $\uparrow$

(due to multi-stage)  
 $\therefore$  (slow response.)

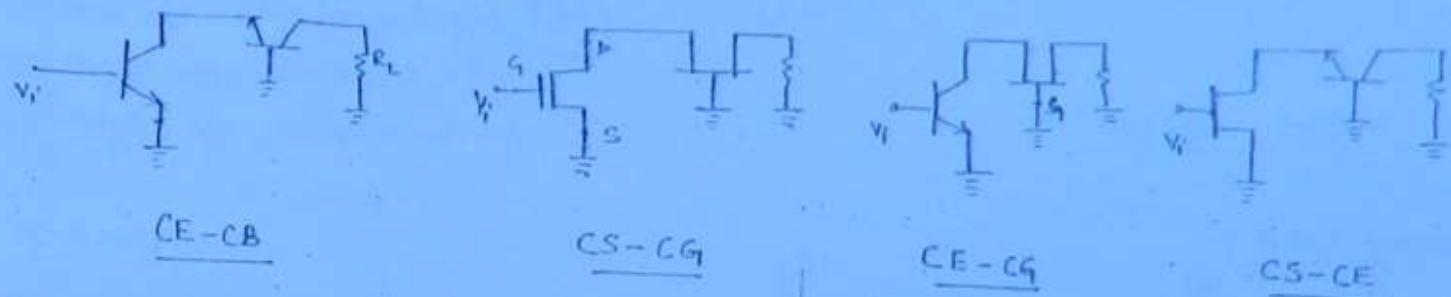
$$\rightarrow \text{Rise time, } t_r^* = 1.1 \times \sqrt{t_{r_1}^2 + t_{r_2}^2 + \dots + t_{r_n}^2}$$

$$\rightarrow \text{If } t_{r_0} = \text{rise time of signal} , \quad t_r^* = 1.1 \times \sqrt{t_{r_0}^2 + t_{r_1}^2 + t_{r_2}^2 + \dots + t_{r_n}^2}$$

07/05/2012

## Cascade Amplifier:

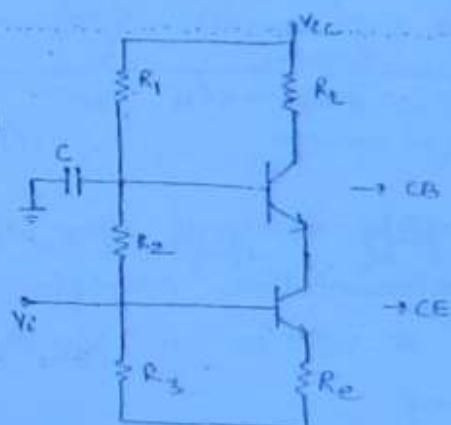
(187)



→ These all are series connections.

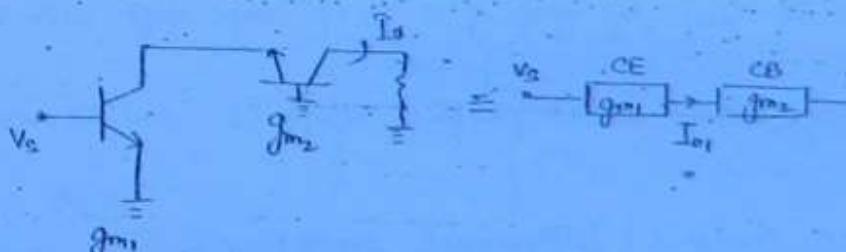
## Basic Diagram:

- C → Bypass capacitor; its value should be very low so that it charges quickly.
- Purpose of this 'C' is to maintain CB in active region.



## Transconductance:

$$g_{m1} = \frac{I_{o1}}{V_s}$$



Ans for CB,  $A_I \approx 1$ , it acts as buffer. for current.

$$\therefore g_m = \frac{I_o}{V_s} = \frac{\alpha I_{o1}}{V_s} \Rightarrow \boxed{g_m = \frac{\beta}{1+\beta} \cdot g_{m1}} \quad \text{--- (exact)} \quad g_m \approx g_{m1} \quad \text{if } \beta \gg 1$$

## Imp. Points:

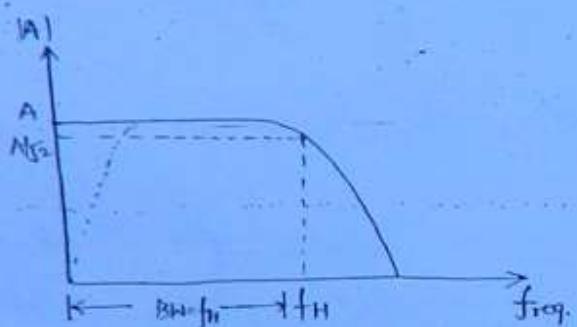
- It is specially designed multistage amplifier the type of coupling provide is direct coupled, therefore suitable to amplify ac & dc signal and major application is as a high freq. amplifier.

→ The input resistance is equal to  $y_{ip}$  resistance CE & output resistance is decided by  $y_{op}$  resistance of CB.

(188)

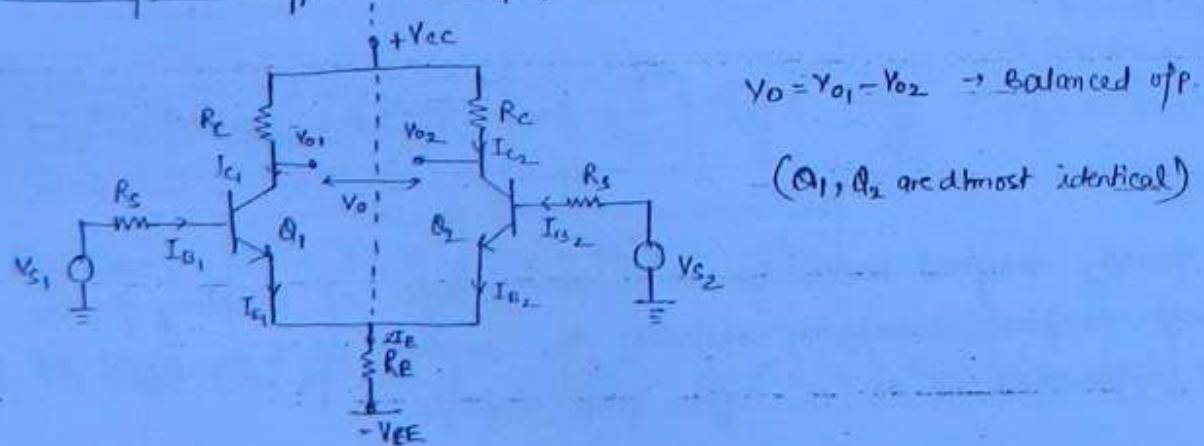
### Direct Coupled Amplifier:

Frequency curve :-



- It is suitable to amplify dc signal along with a wideband of ac signals.
- Widely used as instrumentation amplifier.
- There is no proper dc isolation in b/w the stages, therefore stability is less.
- Any dc amplifier suffers from drift problem. Drift problem is mainly due to  $I_{CO}$ . { gain or  $y_{op}$  of amp drift with temp as  $I_{CO}$  changes }
- Popularly used direct coupled amp is emitter coupled differential amp.

### Emitter Coupled Differential Amplifier:



Mode of operation:-

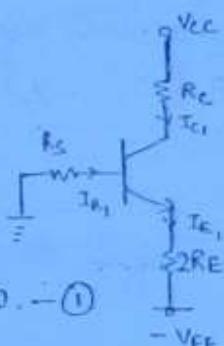
- 1) Dual i/p balanced o/p.
- 2) Dual i/p unbalanced o/p.
- 3) Single i/p balanced o/p.
- 4) Single i/p unbalanced o/p.

(189)

DC Analysis :-

$$\rightarrow V_{S1} = V_{S2} = 0$$

Applying KVL -



→ (because of feedback)  
(potential, should be equal to  $2I_E R_E$   
but  $I_E$  cannot be doubled, hence  
resistance is doubled)

$$I_B1 R_S + V_{BE} + (1+\beta) I_B1 \cdot 2R_E - V_{EE} = 0. \quad \text{--- (1)}$$

$$V_{CE} = I_C1 R_C + V_{CE} + (1+\beta) I_B1 \cdot 2R_E - V_{EE}. \quad \text{--- (2)}$$

$$I_E = \beta I_B1. \quad \text{--- (3)}$$

Now, if  $V_{CE} > 0.2$ , transistor is in active region.

AC analysis :-  $V_{EE} = V_{CC} = 0.$ ,  $V_{S1} = V_{S2} = V_S$ .

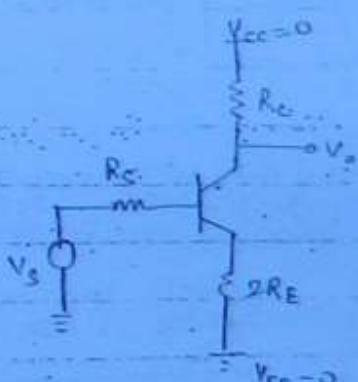
→  $A_e \rightarrow$  common mode gain.

→  $A_d \rightarrow$  differential

Now, for  $A_c$  -  $V_{S1} = V_{S2} = V_S$   
 $V_o = A_c V_c + A_d V_d$ .

$$V_d = V_{S1} - V_{S2} = 0. \rightarrow V_o = A_c V_s$$

$$V_c = \frac{V_{S1} + V_{S2}}{2} = V_S. \rightarrow A_c = \frac{V_o}{V_s}$$



L,  
Common emitter with  
emitter resistance  $2R_E$ .

From circuit,

$$A_I = -h_{fe}, R_i = h_{ie} + (1+h_{fe})(2R_E)$$

$$A_{VS} = \frac{V_o}{V_s} = \frac{A_I R_L}{R_S + R_i} \Rightarrow A_c = \frac{-h_{fe} \cdot R_c}{h_{ie} + (1+h_{fe})2R_E}$$

$$\text{Approximate value, } A_c = \frac{-R_c}{2R_E} \rightarrow \text{when } h_{fe} \gg 1.$$

→ Ideally,  $A_C = 0$ ,  $\Rightarrow R_E \rightarrow \infty$ . (Possible with current mirror)

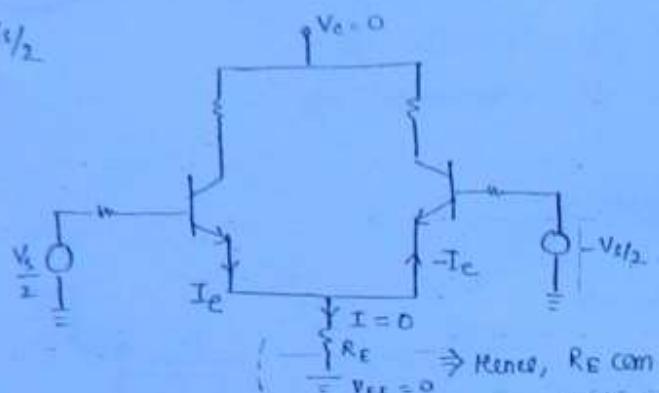
→ As  $R_E \uparrow$ ,  $A_C \downarrow$ .

(190)

for  $A_d$  —  $V_{S2} - V_{S1} = -V_S/2 \Rightarrow V_{S1} = V_S/2$

$\therefore V_d = V_S ; V_C = 0$

$$A_d = \frac{V_o}{V_S}$$



from fig ① —

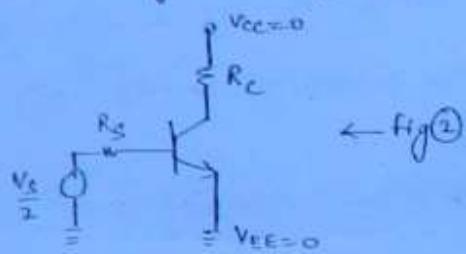
$$A_1 = -h_{fe} ; R_i = h_{ie}$$

$$A_{V_2} = \frac{V_o}{(V_S/2)} = \frac{-h_{fe} \cdot R_E}{R_E + h_{ie}}$$

$$\Rightarrow 2 \cdot A_d = \frac{-h_{fe} \cdot R_E}{R_E + h_{ie}}$$

$$\Rightarrow A_d = \boxed{\frac{-h_{fe} \cdot R_E}{2(R_E + h_{ie})}}$$

Now, dividing the ckt.



← fig ②

→ (It does not depend on  $R_E$ )

$$\Rightarrow CMRR = \frac{|A_d|}{|A_C|} \Rightarrow CMRR = \frac{R_E + h_{ie} + [1 + h_{fe}] \cdot 2R_E}{2[R_E + h_{ie}]}$$

if  $(1+h_{fe})2R_E \gg R_E + h_{ie}$ , then

$$CMRR = \boxed{\frac{[1+h_{fe}] \cdot R_E}{R_E + h_{ie}}} \quad \text{**}$$

(As  $R_E \rightarrow \infty$ ,  $CMRR \rightarrow \infty$ )  
(Ideal value)

Effect of increasing  $R_E$  :

\*  $g_m = \frac{|I_c|}{V_T} \rightarrow$  as  $R_E \uparrow$ ,  $V_{BE} \uparrow \rightarrow$  feedback  $\uparrow$   
 $\rightarrow I_B \downarrow \Rightarrow I_c \downarrow$ .

1) Negative feedback across  $R_E$  Tcs.

2) ~~As~~ yes, CMRR Tcs.,  $g_m \downarrow$ , gain  $\downarrow$  { gain  $\propto g_m$  }

3)  $R_E \uparrow$

## Application:

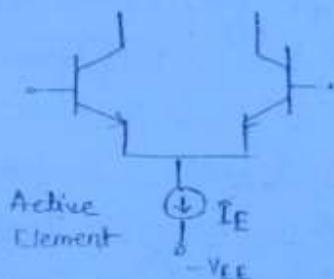
(19)

- It is used as the first internal stage in op-amp.
- As an instrumentation amp.
- As a very good clipper.
- As a linear amplifier, i.e. we can apply superposition theorem.
- It is used in designing of AVC (Automatic voltage control) or AGC (auto. gain control).

→ Any  $\rightarrow$

### Note:

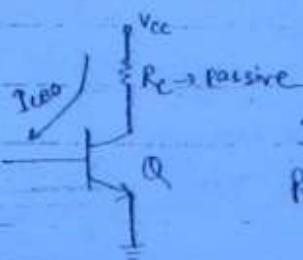
- \* Any ideal diff. amplifier can be designed by connecting an ideal current source in place of  $R_E$ .



Ideal source  
source resistance =  $\infty = R_E$   
 $A_C = 0$   
 $CMRR = \infty$

Practicality  
 $R_E =$  Very high  
Very low  
Very high.

- In a practical diff. amplifier, active load is connected to get best performance. In place of passive load  $R_E$ , pnp transistor is used to get minimum peak-to-peak o/p voltage or maximum swing.



$$\text{Ideal swing} = V_{cc}$$

$$\text{Prac. swing} = V_{cc} - \frac{I_{ceo}R_c}{\text{High}}$$

$\theta$	Ideal $V_o$	Practical $V_o$
$V_i > 0$	saturation	0
$V_i < 0$	cutoff	$V_{cc} - I_{ceo}R_c$

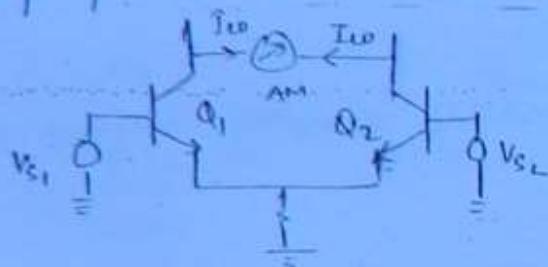
$V_i$	$\theta_1$	$\theta_2$	$R_{o2}$	$V_o \text{ (Prac.)}$
$V_i \gg 0$	on	off	$\approx 0$	$V_{ceo}$
$V_i \ll 0$	off	on	$\approx 0$	$\approx V_{cc}$

: swing  $\uparrow$  →  $I_{ceo}$  is still present but  $R_{o2} \approx 0$

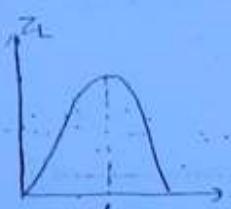
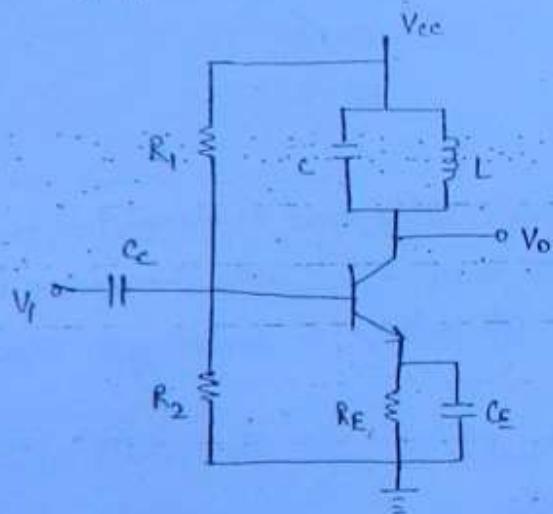
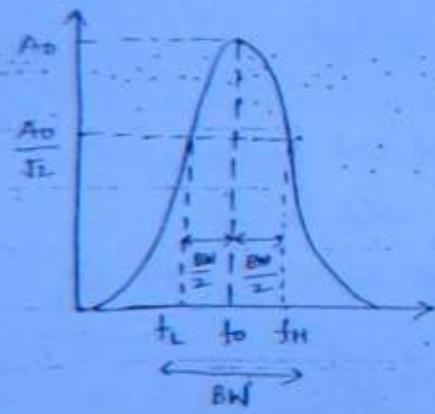
→ Differential amplifier is often used in dc application. It is difficult to design dc amplifier using  $T_f$  because of shift due to variations of  $V_{BE}$ ,  $V_{GE}$  &  $I_{CBO}$  with temp.

(192)

→ With  $Q_1$  and  $Q_2$  having almost identical characteristic, any parameter change due to temp will cancel out and o/p will not vary.  
for eg., leakage current of  $Q_1$  &  $Q_2$  are equal in magnitude but flowing in opposite direction into ammeter & they get cancelled, and hence drift problem is eliminated in emitter coupled diff. amp.



### Tuned Amplifier (Class C amplifiers):-



$$A_v = \frac{A_I \cdot Z_L}{R_L}$$

$Z_L \rightarrow$  LC tank circuit.

$$\rightarrow \text{Resonance freq. : } f_0 = \frac{1}{2\pi\sqrt{LC}}$$

$$\rightarrow BW = \frac{f_0}{Q} ; Q \rightarrow \text{quality factor}$$

$$\rightarrow [A_o \uparrow, BW \downarrow] \Rightarrow [\text{selectivity} \propto Q]$$

{ To change the BW,  $Q$  should be changed and not  $f_0$ , as changing  $f_0$  will change the centre frequency.

$$f_L = f_0 + \frac{BW}{2}$$

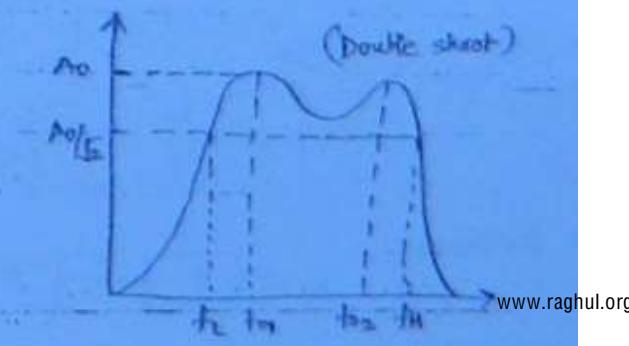
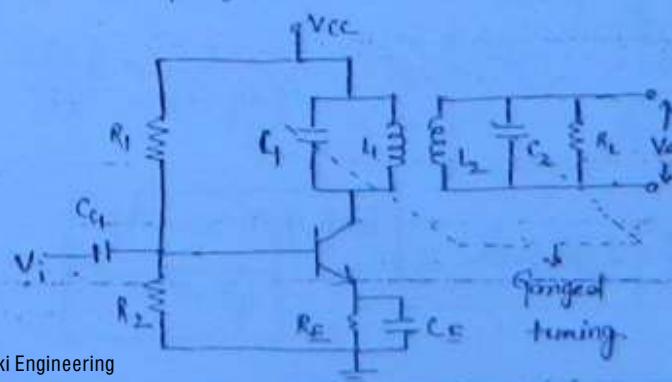
$$f_L = f_0 - \frac{BW}{2}$$

(93)

- It is also called tuned voltage amplifier.
- If signal freq. range = 30KHz to 300KHz. (RF band, hence also called RF amplifier).
- Working principle is parallel resonance.
- Ability of amplifier to reject unwanted frequencies is called selectivity.
  - It has ability to select a particular station signal for amplification by rejecting all other unwanted station signals, i.e., selectivity is very high.
- Front end selectivity of receiver is done by RF amplifier, therefore tuned amplifier is first stage in superheterodyne receiver.
- Tuned amp is class C amplifier and it is a non-linear amp.
- For a tank circuit, Q is very large (100-500).
- BW is very small and this is due to -
  - i) Larger Q.
  - ii) larger gain ( $\because \text{Gain} \times \text{BW} = \text{constant}$ ).
- It is also called Narrow Band amplifier.

Disadvantage: Narrow BW. (with  $\downarrow$  in quality, BW requirement  $\uparrow$  but with  $\uparrow$  BW, gain  $\downarrow$ ).

### Double-Tuned Amplifier



$$\rightarrow f_{01} = \frac{1}{2\pi\sqrt{L_1 C_1}}, \quad f_{02} = \frac{1}{2\pi\sqrt{L_2 C_2}}$$

(194)

- In double tuned amplifier, two tank circuits which are tuned to resonant freq. are inductively coupled and placed in collector ckt.
- BW can be↑ed w/o reducing gain of amp, hence  $\text{gain} \times \text{BW}$  is not a constant

Advantage:-

- A larger BW when compared to a single tuned voltage amp.

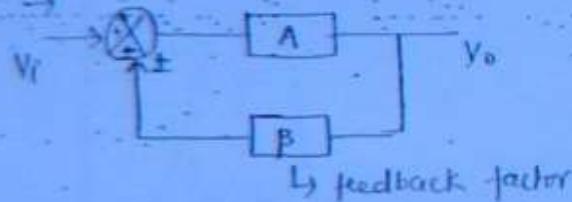
### FEEDBACK AMPLIFIERS

$$\rightarrow V_i \rightarrow [A] \rightarrow V_o \quad A_{OL} = \frac{V_o}{V_i} = A$$

→ (Regenerative feedback)

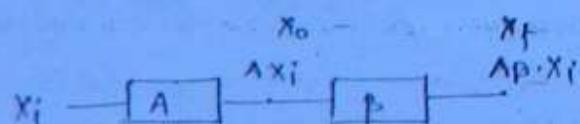
$$\text{for +ve feedback} \dots A_f = \frac{A}{1 + \beta A} (> 1)$$

$$\text{For -ve feedback} \dots A_f = \frac{A}{1 + \beta A} (< A) \quad (\text{Degenerative FB})$$



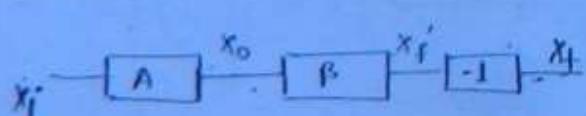
→ Loop Gain (Return Ratio):-      } OLTf } <sup>\* (by me)</sup>

→ +ve feedback -



$$\boxed{\text{loop gain} = \frac{X_o}{X_i} = A \beta}$$

→ -ve feedback:-



Return Difference :-  $D = 1 - \text{loop gain}$

→ for +ve feedback -  $D = 1 - AB$

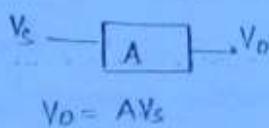
→ for -ve feedback -  $D = 1 + AB$ .

(195)

Advantage of Negative feedback :-

→ Stability of transfer gain increases.

(a) Desensitivity of transfer gain

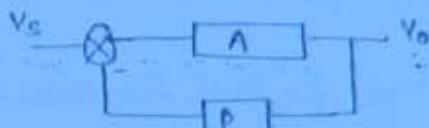


$$V_o = AV_s$$

$$V_o \rightarrow [A + dA] \rightarrow V_o + \Delta V_o$$

$$\Delta V_o = dA \cdot V_s$$

$\frac{dA}{A}$  = fractional variation in A  
w/o feedback.



$$V_o = \frac{A}{1 + BA} \cdot V_s \Leftrightarrow A_f = \frac{A}{1 + BA}$$

$\frac{dA_f}{A_f}$  = fractional variation with  
feedback.

→ if  $\frac{dA_f}{A_f} < \frac{dA}{A}$ , then gain after feedback is stable.

Sensitivity -  $S = \frac{dA_f/A_f}{dA/A}$

→ For stability ;  $|S| < 1$

→ Desensitivity ,  $|D = 1/S|$  ;  $|D| > 1$  for stability after feedback.

Now,

$$\frac{dA_f}{dA} = \frac{1}{(1 + BA)^2} \Rightarrow \frac{dA_f}{A_f} = \frac{dA/A}{(1 + BA)}$$

$$\rightarrow S = \frac{1}{1 + AB} ; D = 1 + AB$$

(b) If feedback n/w contains only stable passive elements then there is improvement in stability

$$A_f = \frac{A}{1+AB} = \frac{1}{B} \text{ if } AB \gg 1.$$

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Hence,  $B$  should consist of stable passive elements.

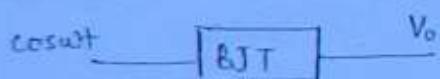
### Reduction in Frequency Distortion :-

Frequency Distortion :- Variation in magnitude of gain with frequency.

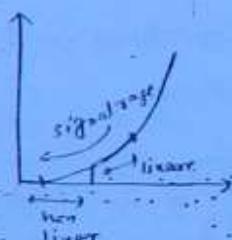
Phase Distortion :- Variation in phase of gain with freq.

→ If  $A_f = 1/B$  and feedback n/w does not contain reactive element, then overall gain is not a func<sup>n</sup> of freq., and there is reduction in frequency & phase distortion.

### Reduction in non-linear distortion :-



$$V_o = B_0 + B_1 \cos \omega t + B_2 \underbrace{\cos 2\omega t}_{\text{dc, desired (fundamental) component}} + B_3 \cos 3\omega t + \dots$$



→  $\omega \uparrow$ , Amplitude  $\downarrow$ ,  $B_2 \gg B_3 \gg B_4 \dots$

$$\rightarrow D_2 = 2^{\text{nd}} \text{ Harmonic Distortion} = \frac{|B_2|}{|B_1|}$$

$$D_3 = 3^{\text{rd}} \dots = \frac{|B_3|}{|B_1|}$$

$$D_4 = 4^{\text{th}} \dots = \frac{|B_4|}{|B_1|}.$$

→ After -ve feedback -

$$D_{2f} = \frac{D_2}{1+AB} \quad \text{**}$$

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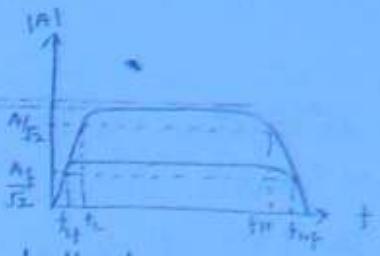
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Note:- Guys Be Cool Dude I am here for help You 😊

→ Bandwidth Increases -

$$B_{wp} = BW [1 + AP]$$



→ Since  $G \times BW$  constant & gain ↓ by  $(1+AP)$  after feedback.

→ Reduction in Noise -

$$N_{of} = \frac{N_o}{1 + AP}$$

(197)

other advantages -

→ It modifies i/p & o/p resistance.

→ It ↑s thermal stability & freq. stability of o/p signal.

Disadvantage -

- It reduces gain.

Application -

-ve feedback is widely used in designing of amp. ckt and control system.

Positive feedback :-

→ Advantage - ↑es gain of amp.

→ Disadvantage -

→ Reduces BW, hence reproduction of i/p signal is very bad.

- It ↑ noise & harmonic distortion at the o/p.

- It reduces stability of amp.

Application

- In designing of oscillator circuits.

Ques An amplifier w/o feedback gives a fundamental o/p of 36V with 7% 2nd harmonic distortion when i/p is 0.028V.

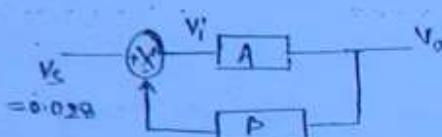
(198)

- If 1.2% of o/p is feedback into i/p in a -ve voltage series feedback ckt, what is o/p voltage.
- If fundamental o/p is maintained at 36V, but the 2<sup>nd</sup> harmonic distortion is reduced to 1%, what is i/p voltage.

Soln)  $V_i = V_s = 0.028 \xrightarrow{A} 36 + D_2 = 7\%$

$$\therefore A = \frac{36}{0.028} = 1285$$

a)  $B = 1.2\% = 0.012$



$$A_f = \frac{V_o}{V_s} = \frac{A}{1+AB}$$

$$A_f = \frac{1285}{1+1285 \times 0.012} = 78.2, \therefore V_o = 78.2 \times 0.028 = 2.19 \text{ V}_o$$

b)

$$D_{2f} = \frac{\lambda_2}{1+AB}$$



$$\Rightarrow 1 = \frac{7}{1+AB} \Rightarrow 1+AB = 7$$

$$\therefore A_f = \frac{A}{1+AB} = \frac{A}{7}, \quad \therefore \frac{V_o}{V_{S1}} = A, \quad \frac{V_o}{V_{S2}} = A_f$$

$$\Rightarrow \frac{V_{S2}}{V_{S1}} = \frac{A}{A_f} \Rightarrow V_{S2} = 7 \times 0.028 = 0.196 \text{ V}_{S2}$$

→ Feedback is often expressed in dB.

$$N_{dB} = 20 \log \left| \frac{A_f}{A} \right|$$

- for +ve feedback,

$$N_{dB} = 20 \log \left( \frac{1}{1+A\beta} \right)$$

199  $\Rightarrow N_{dB} > 0$  or +ve.

- for -ve feedback,

$$N_{dB} = 20 \log \left( \frac{1}{1+A\beta} \right)$$

$\Rightarrow N_{dB} < 0$  or -ve.

Ques: An amp with open loop voltage gain of 1000 delivers 10W of o/p power at 10% 2<sup>nd</sup> harmonic distortion, when i<sub>pp</sub> is 10mV.

If 40 dB -ve voltage series feedback is applied and o/p power is to remain at 10W, determine

- Required i/p signal.
- 2<sup>nd</sup> harmonic distortion.

Sol:  $-40 = 20 \log \left( \frac{1}{1+A\beta} \right) \Rightarrow 1+A\beta = 10^0$   
 $\Rightarrow \beta = \frac{99}{1000}$ .

$$\rightarrow A_f = \frac{A}{1+A\beta} \Rightarrow A_f = 10$$

$$\rightarrow D_f = \frac{D_2}{1+A\beta} = \frac{10}{10^0} = 0.1\%$$

$$\rightarrow V_i' = \frac{10^0}{10} V_i = 100 \times 10mV = 1V$$

### Classification of Amplifiers:

1> Voltage Amplifiers

2> Current " "

3> Transconductance.

4> Trans resistance.

## Voltage Amplifiers :-

→  $R_i \gg R_s$ ,  $R_i = \infty$  (ideally)

→  $R_o \ll R_L$ ,  $R_o = 0$  (ideally).

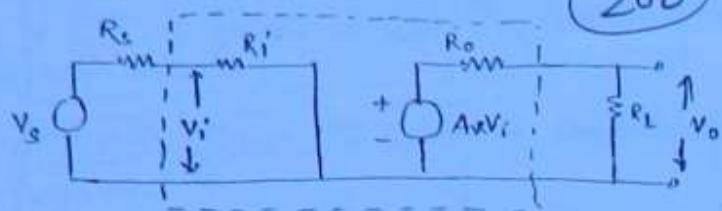
→  $A_v \rightarrow$  internal gain,  $A_{v\text{ext}}$  external gain.

$$\rightarrow V_o = \frac{A_v V_i \cdot R_L}{R_o + R_L}$$

$$A_v = \frac{A_v \cdot R_L}{R_o + R_L}$$

$$A_{v\text{ext}} = \lim_{R_L \rightarrow \infty} A_v$$

When  $R_L = \infty$ ,  
external gain =  
internal gain



## Current Amplifier :-

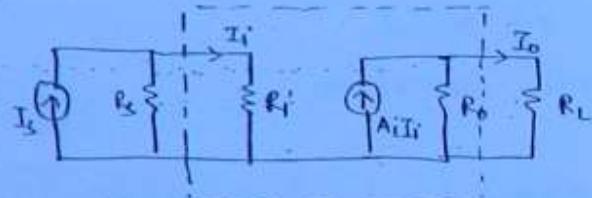
→  $R_i \ll R_s$ , i.e.,  $R_i = 0$  (ideally)

so that whole current passes through  $R_i$ .

→  $R_o \gg R_L$ , i.e.,  $R_o = \infty$  (ideally)

so that max current is delivered to load.

→  $A_I = \text{ext. gain}$ ,  $A_i = \text{internal gain}$ .



$$A_I = \frac{I_o}{I_s} = \frac{A_i R_o}{R_o + R_L}$$

$$A_I^* = \lim_{R_L \rightarrow 0} A_I$$

## Transconductance :-

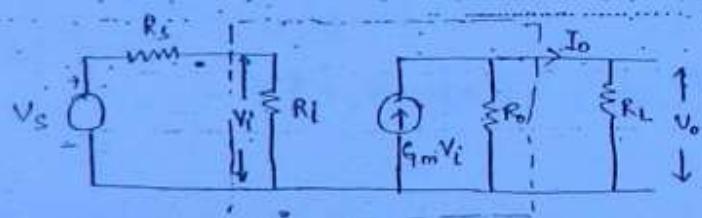
→  $R_i \gg R_s$ , ideally  $R_i = \infty$

→  $R_o \gg R_L$ , ..  $R_o = \infty$ .

$$\rightarrow I_o = \frac{G_m V_i \cdot R_o}{R_o + R_L}$$

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

$$G_m^* = \lim_{R_L \rightarrow \infty} G_m$$



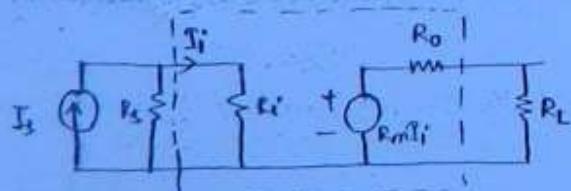
## Transresistance :-

→  $R_s \ll R_i$ , ideally  $R_s = 0$

→  $R_o \ll R_L$ , ..  $R_o = 0$

$$\rightarrow V_o = \frac{R_m I_i \cdot R_L}{R_o + R_L}$$

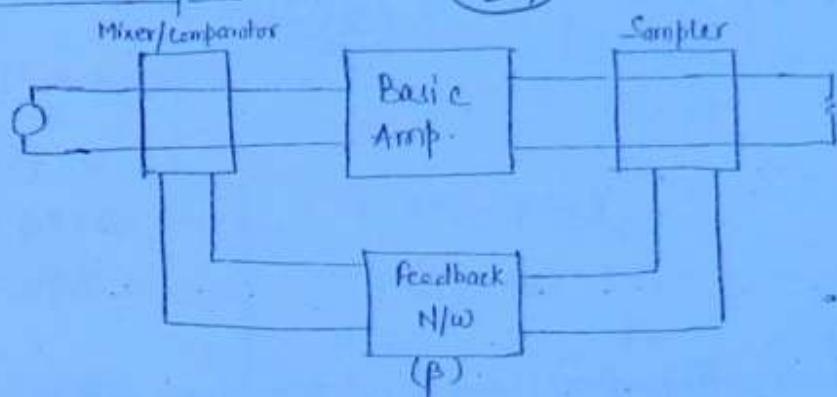
$$R_m = \frac{R_m R_L}{R_o + R_L}$$



$$R_m^* = \frac{R_m R_L}{R_L + \infty}$$

## Feedback Concept :-

(20)

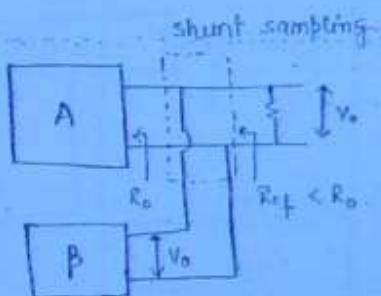


## Sampler :-

### a) Voltage Sampler :-

→ sampled voltage in feedback is same as o/p voltage.

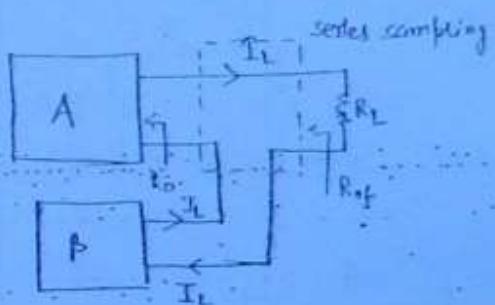
→  $R_{of} < R_o$



### b) Current Sampler :-

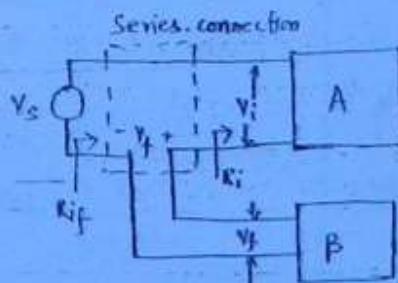
→  $R_{of} > R_o$ .

→ sampled current in feedback is same as o/p current.



## Mixer :-

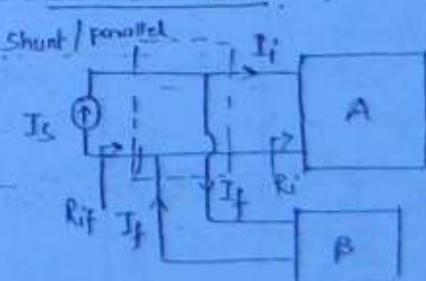
### (a) Voltage Mixer



→  $R_{if} > R_i$ ;

$$\text{Before mixing } V_i = V_s \\ \text{After } V_i = V_s - V_f$$

### (b) Current Mixer



$R_{if} < R_i'$ , Before mixing  $\Rightarrow I_i = I_s$

After  $\Rightarrow I_i' = I_s - I_o$

voltage  
current  
v<sub>f</sub> = feedback voltage i/p  
y<sub>f</sub> = shunt feedback current i/p

## Feedback Topology :-

(202)

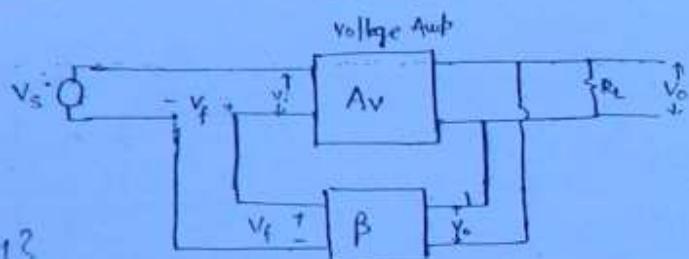
- 1) Voltage Series feedback.
- 2) Current Series "
- 3) Voltage Shunt "
- 4) Current Shunt "

Derivation of  $R_{if}$  (input resistance with feedback) and  $R_{of}$  (o/p resistance with FB)  
for Voltage Series feedback :-

$$\rightarrow V_f = \beta V_o$$

$$\Rightarrow \beta = \frac{V_f}{V_o} = \text{unitless.}$$

$$A_{if} = \frac{A_v}{1 + \beta A_v} = \frac{1}{\beta} \quad \{ \beta A_v \gg 1 \}$$



$$\rightarrow R_{if} < R_i \quad \& \quad R_{if} > R_i$$

Calculation of  $R_{if}$  -

Before feedback :-

$$V_i = V_s$$

$$i/p \text{ resistance} = \frac{V_s}{I_i} = \frac{V_i}{I_i} = R_i$$

After feedback :-

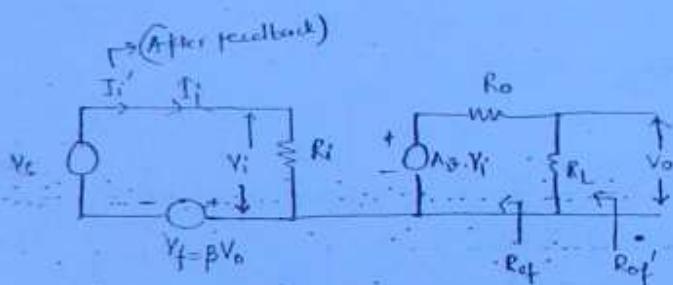
$$V_s = V_i + V_f \Rightarrow V_i = V_s - V_f$$

(There is -ve feedback).

$$i/p \text{ resistance} \approx R_{if} = \frac{V_s}{I'_i}$$

Applying KVL -

$$V_s = I'_i R_i + V_f$$



$$V_s = I'_i R_i + \beta V_o$$

$$\text{but } V_o = \frac{A_v \cdot V_i \cdot R_L}{R_o + R_L} = A_v \cdot R_o \cdot V_i$$

$$\therefore V_s = I'_i R_i + \beta \cdot A_v \cdot R_o \cdot V_i$$

$$\Rightarrow V_s = I'_i R_i + \beta A_v R_i I'_i$$

$$\Rightarrow \frac{V_s}{I'_i} = R_{if} = R_i + \beta A_v R_i$$

$$\Rightarrow R_{if} = R_i (1 + \beta A_v) \Rightarrow R_{if} > R_i$$

Calculation for  $R_{of}$

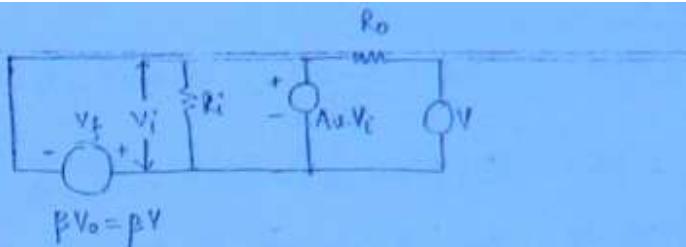
(203)

$$\rightarrow V = IR_o + A_{v2} \cdot V_i$$

$$\rightarrow V_i + V_f = 0$$

$$\Rightarrow V_i + \beta V = 0$$

$$\Rightarrow V_i = -\beta V$$



$$\beta V_o = \beta V$$

$$\therefore V = IR_o + \beta A_{v2} V$$

$$\Rightarrow V(1 + \beta A_{v2}) = IR_o$$

$$\Rightarrow R_{of} = \frac{V}{I} = \frac{R_o}{1 + \beta A_{v2}} \quad \{ R_{of} < R_o \}$$

$$R_{of}' = R_{of} \| R_L$$

	Voltage Series	Current Series	Voltage Shunt	Current Shunt
Output	V	I	V	I
Input	V	V	I	I
Basic Amplifier	Voltage Amp. $A_v = V_o/V_i$	Transconductance $G_m = I/V$	Transresistance $R_m = V/I$	Current Amp. $A_I = I_o/I_i$
Stabilised Gain	$A_{vf} = \frac{V_o}{V_s} = 1/\beta$	$G_{mf} = \frac{I_o}{V_s} = 1/\beta$	$R_{if} = \frac{V_o}{I_s} = \frac{1}{\beta}$	$A_{if} = \frac{I_o}{I_s} = \frac{1}{\beta}$
Unit of $\beta$	Unit less	ohm	mho	unit less
Another name ( $\beta_p, \beta_f$ )	Series - shunt	series - series	shunt - shunt	shunt - series
Effect on $R_i$	Yes	Yes	No	Yes
" " $R_o$	Yes	Yes	Yes	Yes

	Voltage Series	Current Series	Voltage Shunt	Current Shunt
$\rightarrow R_i \uparrow$	$V_i \xrightarrow{[A_v]} V_o$	$V \xrightarrow{[G_m]} I \xrightarrow{R_o \parallel R_f}$	$I \xrightarrow{R_i \downarrow} R_m \xrightarrow{V_o} R_{ob}$	$I \xrightarrow{R_i \downarrow} A_i \xrightarrow{I_o} R_o \uparrow$
$\rightarrow R_{if} = (1 + \beta A_v) R_i$	$\rightarrow R_{if} = (1 + \beta G_m) R_i$	$\rightarrow R_{if} = \frac{R_i}{1 + \beta G_m}$	$\rightarrow R_{if} = \frac{R_i}{1 + \beta A_v}$	
$\rightarrow R_{of} = \frac{R_o}{1 + \beta A_v}$	$\rightarrow R_{of} = (1 + \beta G_m) R_o$	$\rightarrow R_{of} = \frac{R_o}{1 + \beta G_m}$	$\rightarrow R_{of} = R_o (1 + \beta A_v)$	
$\rightarrow$ Normally, $A_v = A_{if}$ (i.e., $R_L \approx \infty$ )	$\rightarrow G_m \approx G_m$	$\rightarrow R_{if} \approx R_m$	$\rightarrow A_{if} \approx A_i$	
$\rightarrow R_{of}' = R_{of} \  R_L$				

### Workout

Q.33 : Ans.  $I_B = \frac{5 - 0.7}{10^3 K_E} = 4.3 \times 10^{-3} \text{ mA}$

(204)

$$I_C = \beta I_B = 100 \times 4.3 \times 10^{-3} = 0.43 \text{ mA}$$

$$\frac{V_0 - 12}{2K} + \frac{V_0}{4} + I_C = 0 \Rightarrow \frac{V_0}{2} + \frac{V_0}{4} - 6 + 0.43 = 0$$

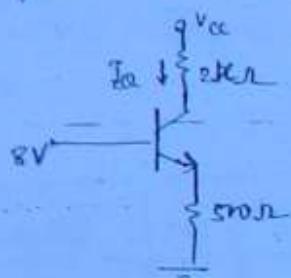
$$\Rightarrow V_0 = 7.43 \text{ V}$$

Q.17 Since, the circuit is amplifier, then  $V_{CE} > 0.2 \text{ V}$

$$\Rightarrow V_C - V_E > 0.2 \text{ V}$$

$$V_E = 8 - 0.7 = 7.3 \text{ V}$$

$$V_t = V_{CE} - I_C \cdot 2K$$



$$\Rightarrow V_{CE} = 6 - 7.3 > 0.2 \text{ V}$$

$$\Rightarrow V_{CE} > 13.5 \text{ V}$$

For pnp,  $V_{CE} < -13.5 \text{ V}$

10<sup>th</sup> Sep, 2012

### Voltage Series Feedback

Best practical examples are-

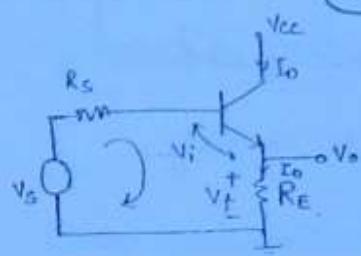
- emitter follower (cc configuration)
- source follower (CDrain ..)
- voltage follower (Non-inverting op-amp)

→ Basic Assumptions :-

- The basic amplifier is unilateral from o/p to i/p, i.e. it does not allow signal from o/p to i/p.
- Feedback n/w is unilateral, from o/p to i/p, i.e., it does not allow signal from i/p to o/p.
- $\beta$  is independent of source resistance  $R_S$  & load resistance  $R_L$ .

### Emitter Follower:

(203)



Let  $R_E$  is very small, hence drop across  $R_E$  can be neglected.

Without fb :- (or w/o RE)  $\Rightarrow V_i = V_s$ .

With fb :-  $V_i = V_s - V_f$

Since  $V_i < V_s$  with feedback, hence -ve feedback.

& series inverting (since voltage is changing).

$\rightarrow$  Also,  $V_o = V_f$

$$\Rightarrow \beta = 1 \quad \left\{ \because V_f = \beta V_o \right\}.$$

& there is voltage sampling.

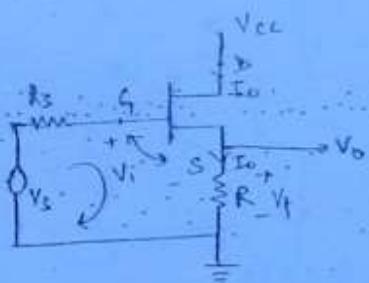
$$\rightarrow A_{f1} = \frac{V_o}{V_s} = \frac{V_f}{V_s} = 1 ; \phi = 0$$

$\rightarrow$  If we assume current sampling, then

$$V_f = \beta I_o \Rightarrow \beta = \frac{V_f}{I_o} = \frac{I_o R_E}{I_o} = R_E.$$

but  $\beta$  depends on  $R_E$ , hence not a current sampling.

### Source Follower:



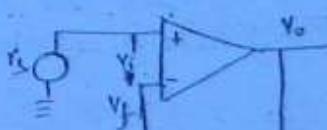
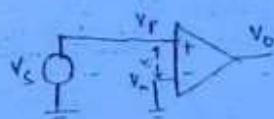
Results will be similar to Emitter follower.

For both circuits :-

- There is max. -ve feedback

- Gain is highly stable ( $\because \beta$  is independent)

### Voltage follower:



w/o feedback

$$V_{i0} = V_s - V_o = V_s$$

$$V_o = A_v \cdot V_i$$

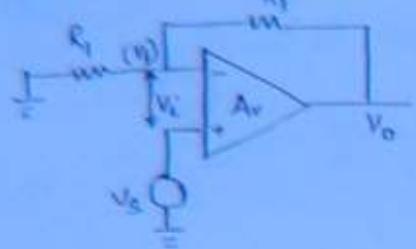
with fb  $V_i = V_s - V_f \Rightarrow$  Vfb, -ve feedback & series inver.

$$V_f = V_o \Rightarrow \beta = 1 \text{ & there is voltage sampling.}$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{V_f}{V_s} = 1 ; R_{if} = (1 + A_f) \cdot R_i = (1 + 10^4) \cdot 10^6 = 10^4 M\Omega$$

$$BWF (1 + A_f) \cdot BW = 10^4 \cdot (BW) \quad R_{of} = \frac{R_o}{1 + A_f} = \frac{100}{10^4} \approx 0.01 \Omega = 0.9 \Omega$$

## Non-Inverting op-Amps



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(contd)

$V_i = V_s - V_f \rightarrow$  -ve feedback, voltage raising comparison

$$V_f = \frac{R_f}{R_f + R_L} V_o \rightarrow V_f = \beta V_o, \Rightarrow \text{voltage sampling.}$$

$\beta$  is constant ( $R_f, R_L$  are neither source resistance, nor load resistance).

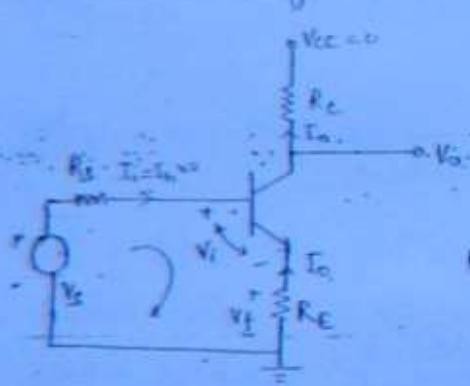
$$\rightarrow A_{Vf} = \frac{V_o}{V_f} = \left(1 + \frac{R_f}{R_L}\right) \rightarrow \text{(approximate).}$$

$$\rightarrow A_{Vf} = \frac{A_v}{1 + \beta \cdot A_v} \rightarrow \text{(exact).}$$

$$\rightarrow R_{if} = (1 + \beta A_v) \cdot R_L ; \quad R_{of} = \frac{R_o}{1 + \beta A_v} ; \quad B_{Wf} = (1 + \beta A_v) B_W.$$

## Current Series Feedback

### (i) CE with unbypassed $R_E$ :-



let drop across  $R_E \approx 0$ .

$$\text{w/o } R_E \approx V_i = V_s$$

$$\text{with } R_E \approx 0 \quad V_i = V_s - V_f$$

$\therefore$  There is series comparison

Now, let there is voltage sampling,

$$V_f = \beta V_o \quad \therefore \beta = \frac{V_f}{V_o} \Rightarrow \beta = -\frac{I_o R_E}{I_o R_C} \Rightarrow \beta = -\frac{R_E}{R_C}.$$

$$\rightarrow A_{Vf} = \frac{V_o}{V_s} = \frac{V_o}{V_f} = -\frac{R_E}{R_C}.$$

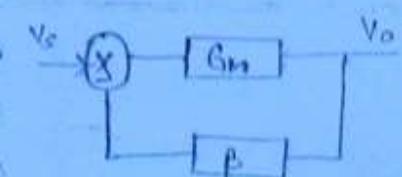
But,  $\because \beta$  is dependent on load  $R_L = R_C$ , hence our assumptions are wrong.

Hence, there is current sampling.

$$\therefore V_f = \beta I_o \Rightarrow \beta = \frac{V_f}{I_o} = \frac{E_o R_E}{I_o R_C} \Rightarrow \beta \propto \frac{1}{R_C}.$$

$$\therefore \beta = \frac{V_f}{I_o} \Rightarrow \boxed{\beta = -R_E}$$

Hence gain is independent of  $R_C$

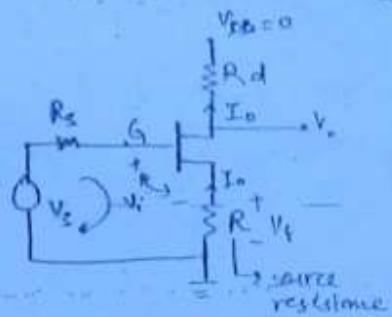


$$\beta m_f = \frac{I_o}{V_s} = \frac{1}{R_e} = -\frac{1}{R_E}$$

(207)

$$A_{vf} = \frac{V_o}{V_s} = \frac{I_o R_C}{V_s} \Rightarrow A_{vf} = -\frac{R_C}{R_E}$$

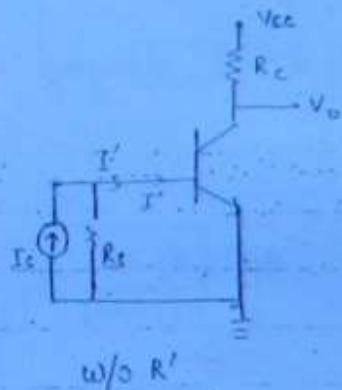
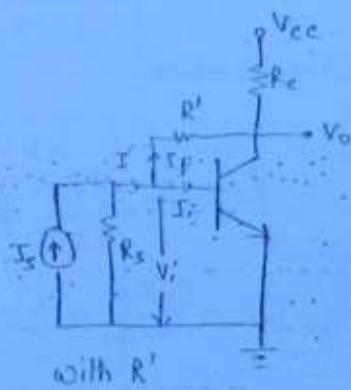
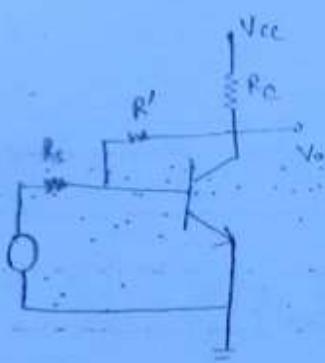
Common Source with unbypassed source resistance:-



$$\begin{aligned} \beta &= -R \\ Gm_f &= -\frac{1}{R} \\ A_{vf} &= -\frac{R_d}{R} \end{aligned}$$

Voltage Shunt Feedback :-

(a) Collector cB Bias circuit :-

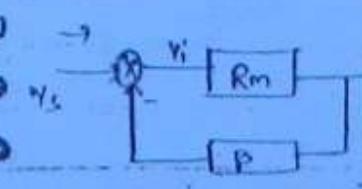


→ w/o feedback,  $I_i = I'$

→ with fb  $-I_i = I' - I_f \rightarrow I'_i \downarrow$ , hence shunt voltage comparison

→ for A CE configuration,  $|A_{vf}| \gg 1 \Rightarrow V_o \gg V_i$

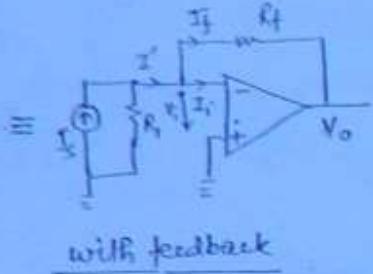
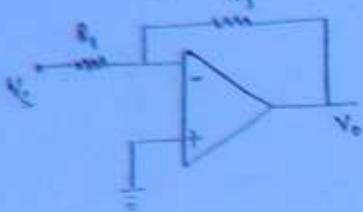
$$\rightarrow I_f = \frac{V_i - V_o}{R'} \Rightarrow I_f = -\frac{V_o}{R'} \Rightarrow \beta = -1/R'$$



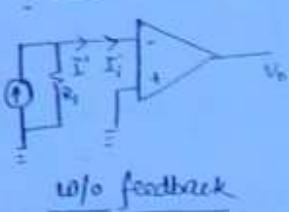
$$R_{mf} = \frac{V_o}{I_s} \Rightarrow R_{mf} = -R'$$

$$A_{vf} = \frac{V_o}{V_s} \Rightarrow \frac{V_o}{I_s R_s} \Rightarrow A_{vf} = -\frac{R'}{R_s} \quad \text{with } V_i = V_o + I_f R_f$$

Inverting Op-amp :-



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w/o feedback:  $I_i = I'$

with feedback:  $I_i = I' - I_f \Rightarrow$  shunt mixing comparison

Now,

$$V_i = V_o - V_P ; V_o = A_v \cdot V_i ; A_v < 0$$

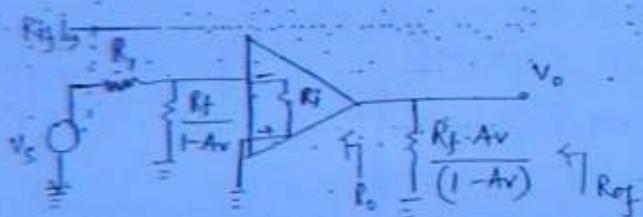
$\therefore |A_v| > 1 \Rightarrow V_o >> V_i$

$$\rightarrow I_f = \frac{V_i - V_o}{R_f} \Rightarrow I_f = -\frac{V_o}{R_f} \Rightarrow \boxed{\beta = -\frac{1}{A_v} R_f} \quad (\text{Ans})$$

$$\rightarrow R_{wf} = \frac{1}{\beta} = -\frac{A_v}{R_f} \rightarrow A_{vf} = \frac{V_o}{V_s} = \frac{V_o}{I_s R_s} \Rightarrow \boxed{A_{vf} = -\frac{R_f}{R_s}}$$

$$\Rightarrow R_{wf} = -R_f = \frac{V_o}{I_s}$$

To calculate  $R_{if}$  &  $R_{of}$ , applying Miller's theorem -



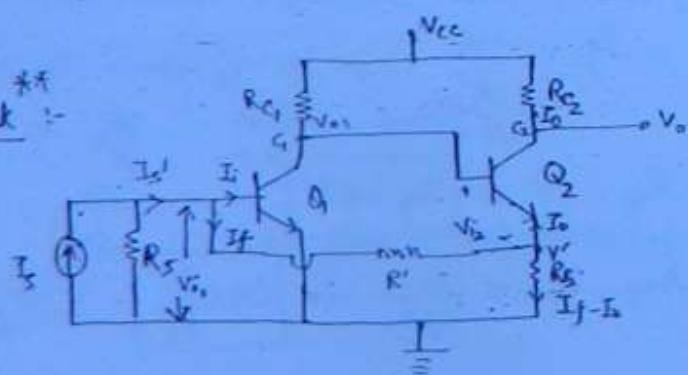
$$R_{if} = R_i + \left( \frac{R_f}{1-A_v} \parallel R_i \right)$$

$\because \frac{R_f}{1-A_v} \approx 0 \quad \{ \because |A_v| > 1 \}$

$\Rightarrow R_{if} = R_i$  hence Reduced i/p resistance

$$\Rightarrow R_{of} = R_o \parallel \left( \frac{R_f A_v}{1-A_v} \right) \approx R_o \parallel R_f \quad \text{Hence, o/p resistance also les}$$

Current Shunt Feedback :-



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w/o feedback  $\rightarrow I_i = I_s'$ with feedback  $\rightarrow I_i = I_s' - I_f \Rightarrow$  shunt mixing comparison

Now,

$$A_{Vi} = \frac{V_{o1}}{V_i} \gg 1 \Rightarrow V_{o1} \gg V_i \quad \left\{ \because CE \text{ configuration} \right\}$$

$$\text{Now, } V' = V_{o1} - V_{i2} \approx V_{o1} \quad \left\{ \because V_{i2} \ll V_{o1}, V_{i2} = \text{small signal} \right\}$$

$$\Rightarrow V = V_{o1} \gg V_i$$

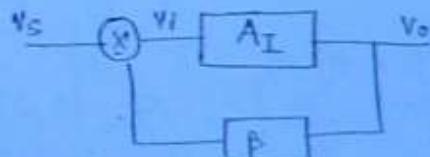
$$\rightarrow I_f = \frac{V_i - V'}{R'} \Rightarrow I_f = -\frac{V'}{R'}, \text{ but } V' = (I_f - I_o)R_E$$

$$\Rightarrow I_f = -\frac{(I_f - I_o)R_E}{R'}$$

$$\Rightarrow I_f = \frac{R_E}{R' + R_E} \cdot I_o \rightarrow \because I_f \text{ depends on } I_o, \text{ hence current sampling.}$$

$$\Rightarrow I_f = \beta I_o \Rightarrow$$

$$B = \frac{R_E}{R' + R_E}$$



$$\rightarrow A_{If} = \frac{V_o}{V_S} = \frac{1}{\beta} = \frac{1 + R'}{R_E}$$

$$\rightarrow A_{Vf} = \frac{V_o}{V_f} = \frac{I_o \cdot R_{G2}}{I_s R_S} \Rightarrow$$

$$A_{Vf} = \frac{1}{\beta} \cdot \frac{R_{G2}}{R_S}$$

## FET

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$$\rightarrow i_d = f(V_{gs}, V_{ds})$$

$$\rightarrow i_d = g_m V_{gs} + \frac{V_{ds}}{r_d} \quad \text{--- ①}$$

Change in  $i_d$  due to  $V_{gs}$  &  $V_{ds}$  -

$$di_d = g_m dV_{gs} + \frac{dV_{ds}}{r_d}$$

When  $V_{ds} = \text{constant}$ ;  $\boxed{g_m = \left. \frac{di_d}{dV_{gs}} \right|_{V_{ds}=\text{constant}}} = \text{Transconductance}$

When  $V_{gs} = \text{constant}$ ;  $\boxed{r_d = \left. \frac{dV_{ds}}{di_d} \right|_{V_{gs}=\text{constant}}} = \text{drain resistance}$

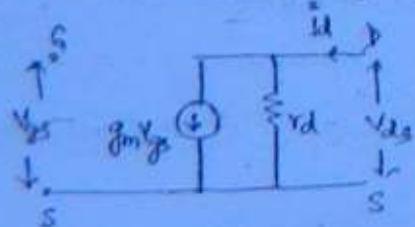
$$\rightarrow \mu = \text{amplification factor} \Rightarrow \boxed{\mu = g_m \times r_d}$$

$$\rightarrow g_m = g_{mo} \left( 1 - \frac{V_{gs}}{V_p} \right); \quad g_{mo} = \frac{2I_{DSS}}{|V_p|} = g_m \Big|_{V_{gs}=0}$$

$$\rightarrow I_{DSS} = I_{DSS} \left( 1 - \frac{V_{gs}}{V_p} \right)^2; \quad I_{DSS} = I_{DSS} \Big|_{V_{gs}=0}; \quad I_{DSS} = \text{saturation drain current}$$

$$\rightarrow g_{mo} = \frac{2}{|V_p|} \sqrt{I_D \cdot I_{DSS}}$$

Small signal Model :- (at low frequency)



Workbook -

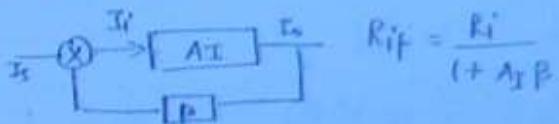
chap 7 :-

(21)

Q.1 b (All wrong answers)

Q.4 (c)

Q.7 (5)  $A_v = 50, \beta = 0.2$



$$R_{if} = \frac{R_d}{1 + A_V \beta}$$

Q.2 (a)

Q.5 (b)

Q.3 (d)

Q.6 (a)

Q.8 (b)

Q.7 (b)

$$A_v = \frac{V_o}{V_i} = \frac{I_o R_o}{I_i R_i} = \frac{A_i R_o}{R_i}$$

$$\Rightarrow A_i = \frac{50 \times 1}{2.5} = 20$$

$$\therefore R_{if} = 1/5 \Omega$$

$$G_{m1} = \frac{I_o}{V_i} = -1 \text{ mA/V}$$

Q.10 (c)

$$A_{v_f} = -f$$

$$G_{m_f} \approx 1/\beta \approx 1/R_E$$

$$D = 1 + \beta \cdot G_m = 50$$

$$\Rightarrow -1 \approx 1/R_E$$

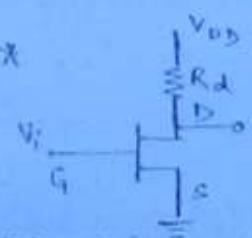
$$\beta = -R_E$$

$$\Rightarrow R_E \approx 1 \text{ k}\Omega$$

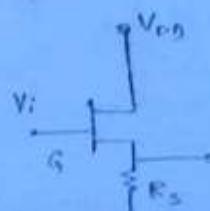
$$\text{Exact } G_{m_f} = \frac{G_m}{1 + \beta G_m} \Rightarrow -1 = \frac{G_m}{50} \Rightarrow G_m = -50$$

$$1 + \beta(-50) = 50 \Rightarrow \beta = -\frac{49}{50} = -R_E$$

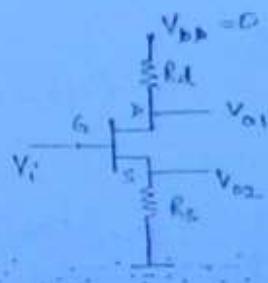
$$\Rightarrow R_E = 0.98 \text{ k}\Omega$$



Common Source



Common Drain



$V_o = V_{o1}$  = CS with drain resistance  $R_d$

$V_o = V_{o2}$  = CD with drain resistance  $R_d$ .

Fig.1

\* Small Signal Analysis :-

KVL at i/p :-

$$V_{o1} = -i_d R_d \quad \text{--- (1)}$$

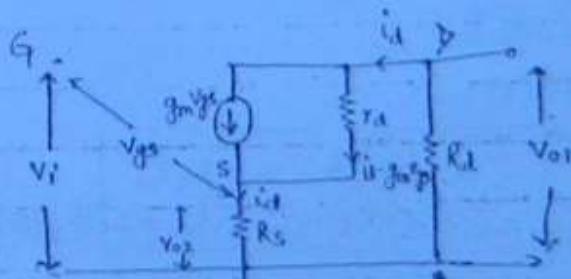
$$V_{o2} = i_d R_s \quad \text{--- (2)}$$

$$V_{gs} = V_i - i_d R_s \quad \text{--- (3)}$$

KVL at o/p

$$-i_d R_{d1} - r_d (i_d - g_m V_{gs}) - i_d R_2 = 0$$

$$\Rightarrow i_d (R_d + r_d + R_s) - g_m V_{gs} = 0$$



(from fig. 1)

$$\because \mu = g_m \times r_d \text{ & eqn (3)} -$$

$$i_d (R_d + r_d + R_s) - \mu (V_i - i_d R_s) = 0$$

$$\Rightarrow i_d = \frac{\mu \cdot V_i}{R_d + r_d + (1+\mu)R_s} \quad \text{--- (4)}$$

(212)

for CS with source resistance  $R_s$  —

$$V_{o1} = -i_d \cdot R_d$$

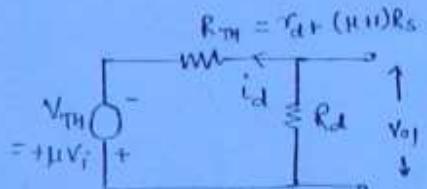
$$\Rightarrow V_{o1} = \frac{-\mu \cdot V_i \cdot R_d}{R_d + r_d + (1+\mu)R_s} \quad \text{--- (5)} \Rightarrow \boxed{AV = \frac{-\mu \cdot R_d}{R_d + r_d + (1+\mu)R_s}} ; \quad \boxed{\phi = 180^\circ} \quad \text{--- (6)}$$

→ if  $(\mu+1)R_s \gg (R_d + r_d)$  —

$$\boxed{AV = -\frac{R_d}{R_s}} \quad \text{--- (7)} \rightarrow \text{Independent of any parameter for FET, hence gain is highly stable.}$$

Thevenin's equivalent :-

$$V_{o1} = \frac{R_d}{R_d + R_{TH}} \cdot V_{TH} \quad \text{--- (8)}$$



from (5) & (8) —

$$\begin{cases} R_{TH} = R_o = r_d + (\mu+1)R_s \\ V_{TH} = -\mu V_i \end{cases} \quad \text{--- (9)}$$

⇒ op. resistance increases due to current series feedback. (Effect on op. resistance is neglected as it is already  $\gg$ )

for Common source ( $\text{w/o } R_s$ ) :-

Put  $R_s=0$  in eqn (6) —

$$\boxed{AV = -\frac{g_m r_d \cdot R_d}{R_d + r_d} = -g_m R_d'} \quad ; \quad \boxed{R_d' = R_d || r_d}$$

From eqn (9) —

$$\boxed{R_o = r_d = R_{TH}}$$

use if  $r_d$  is not given, then take it  $\infty$ .

For Common Drain with drain resistance  $R_d$  :-

$$V_{o2} = i_d \cdot R_s$$

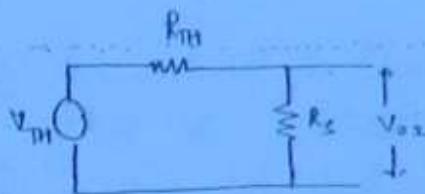
(213)

$$\therefore V_{o2} = \frac{\mu V_i R_s}{R_d + r_d + (\mu+1) R_s} \quad \text{--- (12)}$$

$$\Rightarrow A_v = \frac{V_{o2}}{V_i} \Rightarrow \boxed{A_v = -\frac{\mu \cdot R_s}{R_d + r_d + (\mu+1) R_s}} \quad \text{--- (13)} , \quad \boxed{\phi_{shift} = 0^\circ} \\ |\Delta v| < 1$$

Thevenin's Equivalent :-

$$\frac{V_{TH} \cdot R_s}{R_s + R_{TH}} = V_{o2} \quad \text{--- (14)}$$



Dividing numerator & denominator by  $(\mu+1)$  in eqn (12) -

$$V_{o2} = \frac{\left(\frac{\mu}{\mu+1}\right) V_i \cdot R_s}{\frac{R_d + r_d}{(\mu+1)} + R_s} \quad \text{--- (15)}$$

Comparing (14) & (15) :-

$$\boxed{V_{TH} = \left(\frac{\mu}{\mu+1}\right) \cdot V_i} ; \quad \boxed{R_{TH} = \frac{R_d + r_d}{\mu+1} = R_o} \quad \text{--- (16)}$$

For common drain (w/o  $R_d$ ) :-

Putting  $R_d=0$  in eqn. (13) -

$$A_v = \frac{\mu R_s}{r_d + (\mu+1) R_s} ; \quad \text{if } (\mu+1) R_s \gg r_d \quad \& \quad \mu \gg 1,$$

$\Rightarrow \boxed{A_v \approx 1}$  → Circuit is called source follower.

from eqn (16) -

$$\rightarrow R_o = \frac{r_d}{\mu+1} \approx \frac{r_d}{\mu} \approx \frac{r_d}{g_m r_d} \Rightarrow \boxed{R_o = \frac{1}{g_m}}$$

## Source Self Biasing

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### DC analysis

$$V_{GS} = I_D R_S + V_{DS} + I_D R_D$$

$$\Rightarrow V_{GS} = V_{DS} + I_D R_D$$

$$\Rightarrow V_{GS} = V_{DS} - I_D R_S$$

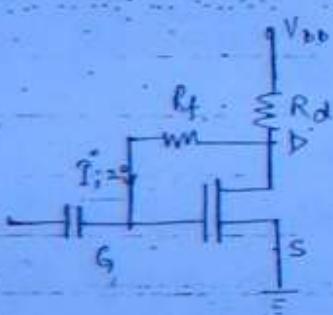
if  $V_{GS}=0$ ;  $V_{DS} = -I_D R_S$ .

if  $R_S=0$ ;  $V_{GS} = V_{DS}$ . — (fixed Biased ckt)

→ Self bias technique cannot be used to establish an operating point for enhancement-type MOSFET as voltage drop across  $R_S$  is in a direction to reverse bias the gate and forward gate bias is required for E-MOSFET.

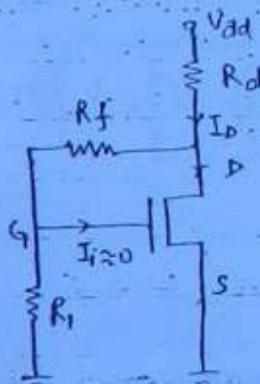
→ This is used for JFET or depletion type MOSFET.

## Drain-Gate Biasing for Enhancement Type MOSFET



$$V_{DS} = I_D R_F + V_{GS}$$

$$\Rightarrow V_{DS} = V_{GS}$$



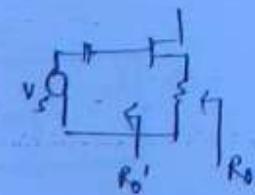
$$V_{GS} = \frac{R_G}{R_G + R_F} \cdot V_{DS}$$

## Workbook (Chap-4)

$$(1)(b) \mu = \frac{dV_{GS}}{dV_{GE}}$$

$$(2)(c) R_o = R_o' \parallel R_S$$

$$R_o' = \frac{1}{g_m} = \frac{100}{3}$$



$$R_o = \frac{100}{3} \parallel 3000 = 300$$

Ques 5:  $\frac{V_{D1}}{V_i} = -g_m R_d' \quad , \quad R_d' = r_d \parallel R_d = \infty \parallel R_d = R_d$

$$\Rightarrow R_d' = 3\text{ k}\Omega.$$

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$$\therefore V_{GS} = 0, \quad V_{GS} = -I_D \cdot R_S \\ = -2.5\text{ V}$$

$$g_m = \frac{2I_{DSS}}{|V_P|} \left[ 1 - \frac{V_{GS}}{V_P} \right] = \frac{2 \times 10}{5} \left( 1 - \frac{2.5}{5} \right) = 2$$

$$\therefore A_V = -2 \times 3 = -6$$

Ques 4:  $V_{GS} = V_{GS} = -2\text{ V}$

Ques 6 (a)

$$g_m = \frac{2 \times 10}{8} \left( 1 - \frac{2}{8} \right) = 1.875 \text{ mS}$$

$$A_V = -g_m R_d' \quad ; \quad R_d' = 20\text{ k} \parallel 2\text{ k} \\ \Rightarrow A_V = -3.41$$

Ques 7 (b)  $A_V = -g_m R_d' = -g_m (R_d)$ .

When ~~for~~ for ac analysis,  $V_{DD} = 0$ , C  $\rightarrow$  short,  $R_D = 5\text{ k} \parallel 10\text{ k}$ .

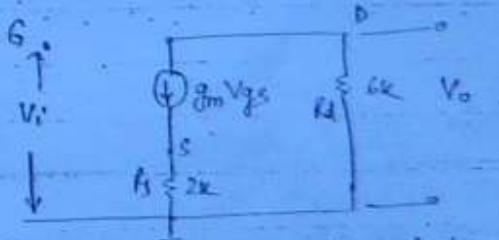
$$|A| = 2 \times (5115) = 5$$

Ques 8 (c)  $R'_i = 20 \parallel 100\text{ k} \parallel \infty$

$$R'_i = 16.67 \text{ k}\Omega$$

Ques 9 (d)  $A_V \approx -\frac{R_D}{R_C} = -3 = -2.66$  (slight cut off)

By model



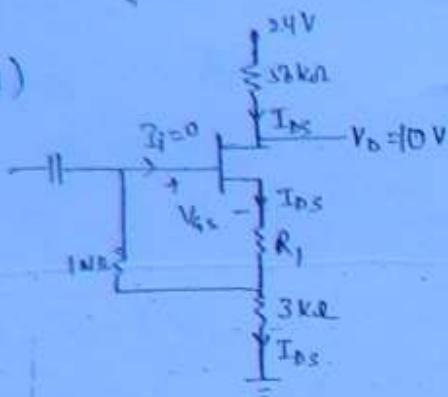
$$V_d = -g_m V_{GS} \cdot R_d. \quad (\because r_d \text{ not given})$$

$$V_i = V_{GS} + g_m V_{GS} \cdot R_S$$

$$A_V = \frac{V_o}{V_i} = \frac{-g_m R_d}{1 + g_m R_S} = \frac{-4 \times 6}{1 + 4 \times 2} = -2.6$$

Conventional:

Soln (1)



Assuming FET is in saturation

$$V_{GS} + V_{DS} = I_{DS} \cdot R_1 = 0$$

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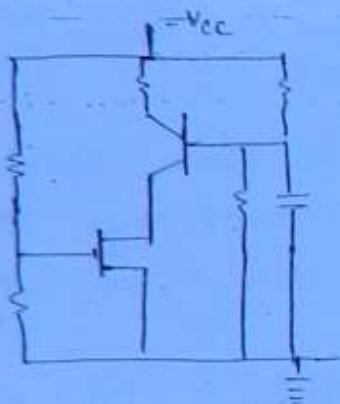
$$I_{DS} = \frac{24-10}{5k} = 1 \text{ mA}$$

$$I_{DS} = \frac{20 \times 2}{1} \left( 1 - \frac{V_{GS}}{(-1)} \right)^2$$

$$\Rightarrow 0.25 = 2(1 + V_{GS})^2 \Rightarrow V_{GS} = \frac{1}{2\sqrt{2}} - 1$$

Chapter 3:

Ques 2:



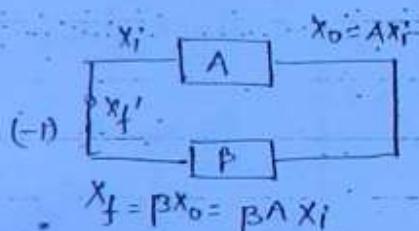
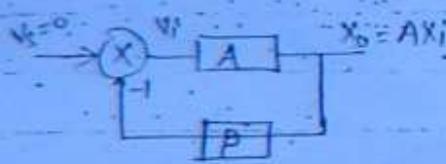
$$g_m = \frac{P}{1 + \beta} \cdot g_{mF}$$

$$= \frac{9.9}{100} \times 2$$

$$= 1.98 \text{ mA/V}$$

1st Sep, 2012

Oscillators (Sinusoidal)



$$\rightarrow \text{loop gain} = \frac{X_f}{X_i} = -AP$$

$X_f = X_i$  then there is

$\rightarrow$  If finite off w/o any ifp., ckt acts as oscillator

$$\therefore \text{loop gain} = Y = (-AP) = 1 \rightarrow \text{Barkhausen Criterion}$$

Phase shift  $\phi = 0, 360 \text{ or } 2\pi$ .

$$\rightarrow |\text{loop gain}| = AP = 1$$

Now,  $A_f = \frac{A}{1+A_f\beta}$ . for system satisfying Barkhausen criteria-

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$$A_f = \frac{A}{1-1} = \infty.$$

Barkhausen Criterion :- It states that -

- 1) Total phase shift around a loop as signal proceeds from  $i_{lp}$  through amplifier, feedback now and back to  $i_{lp}$  again, completing a Loop is multiple integral of  $2\pi$ , ie,

$$\boxed{\phi = 2n\pi} ; n=0, 1, 2, \dots$$

- 2) The magnitude of product of open loop gain of amplifier,  $A$  and feedback factor  $\beta$  is unity.

$$\boxed{|A\beta|=1}.$$

Practical Consideration :-

Practically magnitude of loop gain, ie,  $|A\beta|$  should be kept slightly greater than unity. Then amplitude of oscillation is controlled by onset of non-linearity present in system, in other words, in a practical oscillator, loop gain is kept slightly greater than one to overcome the circuits internal losses.

Oscillators :-

- Oscillator is basically a waveform generator, used in designing of signal generator and function generator.
- It is also defined as an amplifier with  $\infty$  gain.

Amplifier

- 1) Gain is finite
- 2) Negative feedback
- 3) Excellent stability

Oscillator

- 1) Gain is  $\infty$ .
- 2) Positive feedback.
- 3) less stable.

→ External i/p signal is compulsory

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→ External i/p signal is not reqd.  
i/p signal will be noise.

Note :

→ An amplifier can be converted into an oscillator by applying +ve feedback & increasing the gain to  $\infty$ .



### Oscillators

#### AF oscillator

$f_o \rightarrow 20\text{Hz}$  to  $20\text{kHz}$

##### RC phase shift osc.

- By using Op-Amp
- By using FET
- By using BJT

##### Wein Bridge oscillator

#### RF oscillator

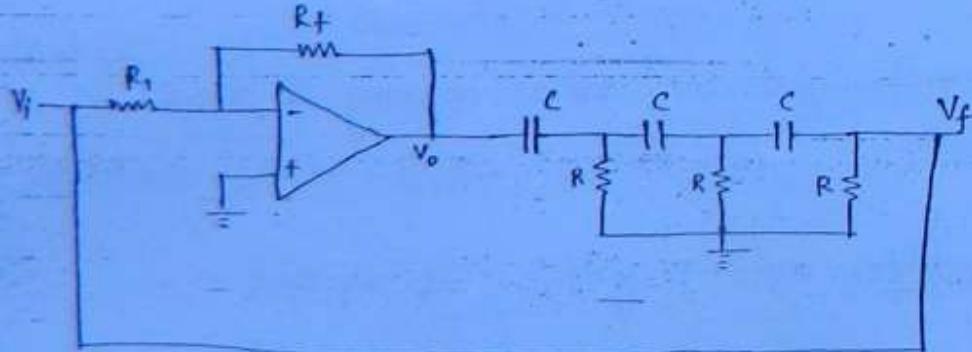
$f_o > 20\text{kHz}$

- Hartley
- Colpitt
- Clapp

→ Crystal oscillator

### RC Phase Shift oscillator

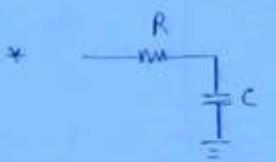
→ By using Op-Amp —



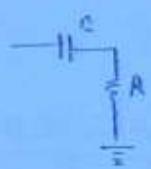
→ If  $V_f = V_i$ ; circuit acts as an oscillator.

→ feedback mechanism:- Voltage series

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$$\phi = -\tan^{-1}(\omega RC)$$

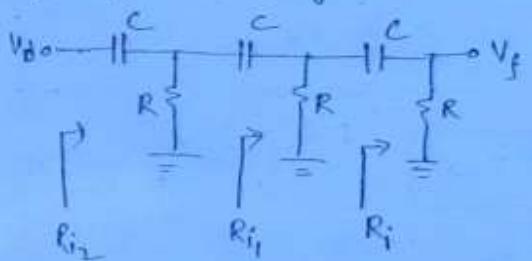


$$\phi = \tan^{-1}(\frac{1}{\omega RC})$$

→ Preferable as lower values of  $R, C$  are required to maintain higher phase shift.

→ To get the overall gain equal to  $180^\circ$ , the phase shift is distributed among all the stages.

→ In this  $RC$  phase shift, three stages are added but the individual phase shift of each stage is not  $60^\circ$ . This is due to loading effect.



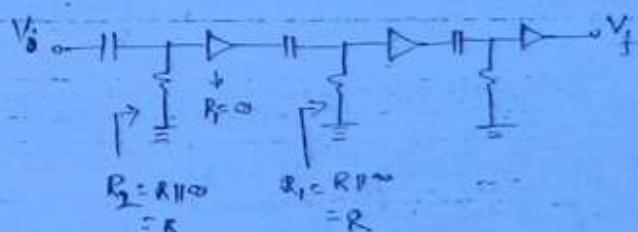
$$R_{i1} = R_i || R < R \quad \left. \begin{array}{l} \text{Hence, phase shift of} \\ \text{individual stage will} \end{array} \right\}$$

$$R_{i2} = R || R_{i1} < R. \quad \left. \begin{array}{l} \text{be } > 60^\circ \text{ in this case} \\ \text{and hence, overall } \phi > 180^\circ. \end{array} \right\}$$

$$\rightarrow \beta = V_f / V_o.$$

→ To calculate set the overall  $\phi=180^\circ$ , calculate  $V_f$  and set imaginary part equal to 0 and set value of  $R, C$  for given  $\omega$  such that real part  $\alpha$  is  $-ve$ . In this way, total  $\phi=180^\circ$  but individual  $\phi$  of stages is not known.

→ We can use buffer in b/w the stages to prevent loading effect but not used due to its complexity.



Voltage follower = buffer.

→ freq. of oscillation :-

$$\beta = \frac{V_f}{V_o} = X + jY \quad \text{on putting } Y=0 -$$

$$\rightarrow \text{freq of oscillation} ; \quad f_0 = \frac{\omega_0}{2\pi} = \frac{1}{2\pi RC \sqrt{G}}$$

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Substituting  $f_0$  in  $\beta$  -

$$\boxed{\beta = X = -\frac{1}{29}} \quad \Rightarrow \text{-ve real part} \Rightarrow 180^\circ \text{ phase shift}$$

$\rightarrow$  Condition for oscillation -  $|A\beta| = 1$ .

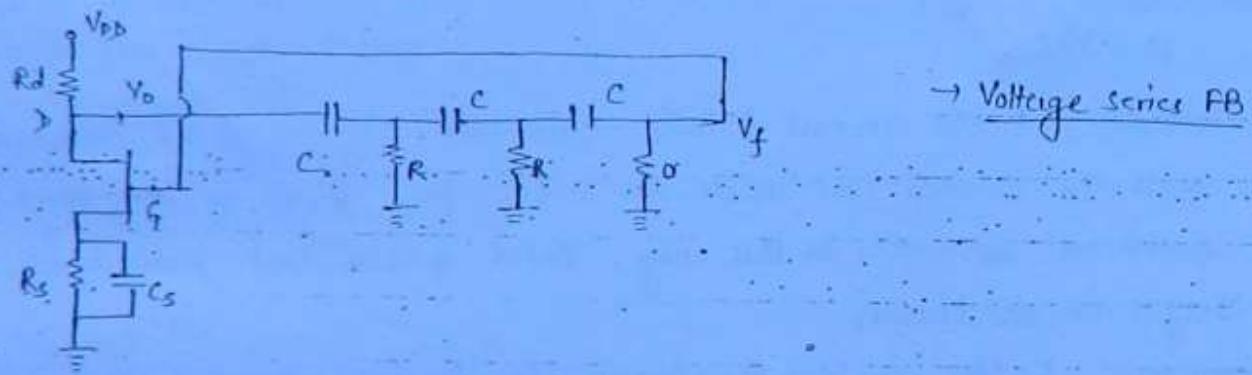
$$\Rightarrow |A| = 29$$

for inverting op-amp,  $A = -\frac{R_f}{R_i}$

$$\Rightarrow \frac{R_f}{R_i} = 29 \Rightarrow \boxed{R_f = 29 R_i} \quad \xrightarrow{\text{Imp.}}$$

Practically,  $|A\beta| \geq 1 \Rightarrow \boxed{R_f \geq 29 R_i} \quad \left\{ \text{only slightly} \geq 29 R_i \right\}$ .

By Using FET :-

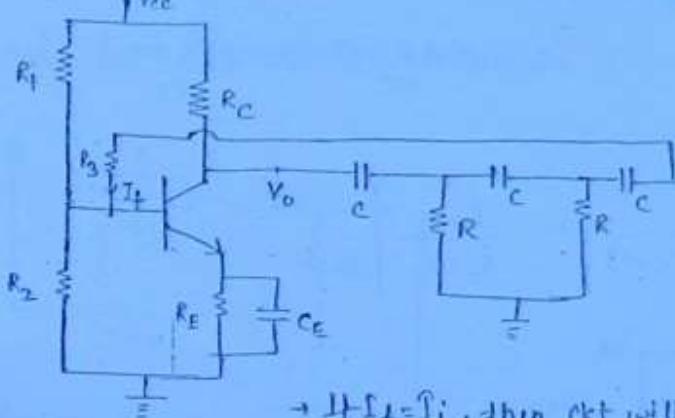


$\rightarrow$  Condition for oscillation :-  $|A\beta| = 1$

$$\rightarrow A_y = -\frac{V_{ds}}{V_{gs}} = \mu$$

$\Rightarrow \boxed{\mu = 29} \rightarrow \text{amplification factor} ; \quad \text{Practically} , \boxed{\mu \geq 29}$

By using BJT :-



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$$= h_{fe} \left[ \frac{V_0}{R_{C1}} \parallel \frac{C}{R_1} \parallel \frac{C}{R_2} \parallel \frac{C}{R_3} \parallel \frac{T_f}{R_{i'}} \right]$$

$$R_i' = R_1 \parallel R_2 \parallel h_{ie} \approx h_{ie}; R_3 + R_i' = R_u$$

$\rightarrow 4I_f = I_i$ , then ckt will act as oscillator.

→ Type of feedback - **Voltage Shunt**

$$\rightarrow f_o = \frac{\omega_0}{2\pi} = \frac{1}{2\pi R C \sqrt{4K+6}}; K = \frac{R_C}{R}$$

$$\rightarrow \text{Putting } |A_{pl}|=1; h_{fe} = 4K + 23 + \frac{29}{K}$$

$$\rightarrow \text{Diff. w.r.t. } K, \frac{dh_{fe}}{dK} = 0 \Rightarrow K \approx 2.7; h_{femin} = 44.54$$

Note

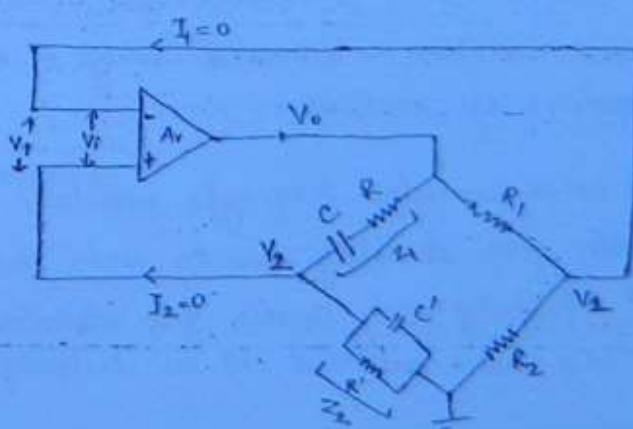
→ An FET with  $\mu < 23$  cannot be used in RC phase shift oscillator.

→ A Tr with small signal CE short ckt current gain, i.e.,  $h_{fe}$  less than 44.54 cannot be used in this oscillator.

→ RC phase shift oscillator is considered as a fixed freq oscillator since to change  $f_o$ , we have to change value of  $R$  &  $C$  of all three sections simultaneously, but this is practically very difficult.

Wein Bridge Oscillator :-

If  $V_i = V_f$ , then ckt acts as an oscillator.



$$\rightarrow V_f = V_2 - V_1 ; \Rightarrow V_f = \frac{Z_2}{Z_1 + Z_2} V_o - \frac{R_2}{R_1 + R_2} V_o$$

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$$\therefore \beta = \frac{V_f}{V_o} = \left[ \frac{Z_2}{Z_1 + Z_2} - \frac{R_2}{R_1 + R_2} \right] ; Z_2 = R' \parallel \frac{1}{sC'} ; Z = R + \frac{1}{Cs}$$

$$\Rightarrow \beta = \left[ \frac{\omega R' C}{\omega (RC + R'C + R'C') - j(1 - \omega^2 RR'C')} - \frac{R_2}{R_1 + R_2} \right]$$

Imp. part = 0

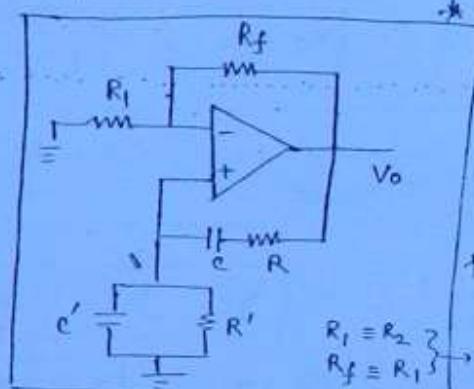
$$\Rightarrow f_o = \frac{\omega_0}{2\pi} = \frac{1}{2\pi \sqrt{RR'C'C'}} \quad \text{--- Imp}$$

Putting  $\omega_0$  in  $\beta$  —

$$\beta = \left[ \frac{R'C}{RC + R'C + R'C'} - \frac{R_2}{R_1 + R_2} \right]$$

Condition for oscillation —  $|Av\beta| = 1$

$$\Rightarrow \beta = \frac{1}{|Av|} = \frac{1}{\infty} = 0$$



\* Check workbook  
Ques 3 (conversion)

→ Another circuit for Wein bridge osc.  
 $R_1 = R_2$ ?  
 $R_f = R_1$ ?  
analogous from original ckt.

$$\Rightarrow \left[ \frac{R'C}{RC + R'C + R'C'} - \frac{R_2}{R_1 + R_2} = 0 \right] \quad \text{--- Imp.}$$

4  $R = R'$ ,  $C = C'$  —

$$f_o = \frac{1}{2\pi RC} \quad \text{obj.}$$

$$\text{Condition: } \frac{1}{3} - \frac{R_2}{R_1 + R_2} = 0 \Rightarrow \boxed{R_1 = 2R_2} \quad \text{obj.}$$

→ An oscillator circuit in which a balanced bridge is used as a feedback n/w. is called wein bridge oscillator.

→ Advantages: It is a variable freq. type oscillator

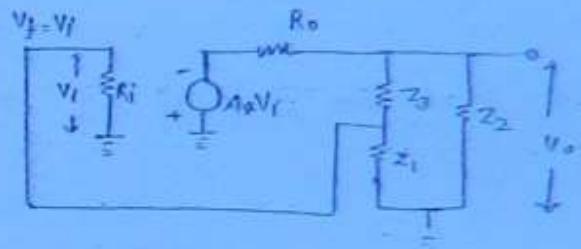
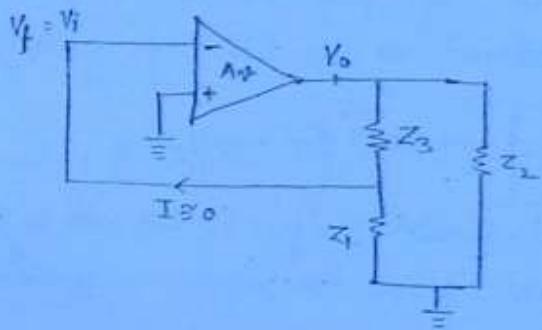
- Better freq. stability due to wein bridge.

→ Application — 1) Popularly used audio freq. oscillator  
2) As a master oscillator ckt in designing of signal generator

## RF oscillator :-

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General form of oscillator circuit :-



$$\rightarrow Z_L = (Z_1 + Z_3) \parallel Z_2 \quad \left\{ \because T \approx 0 \right\} \quad -①$$

$$\rightarrow Z_1, Z_2 \text{ & } Z_3 \text{ are all reactive} ; \quad Z_1 = jX_1 ; \quad Z_2 = jX_2 ; \quad Z_3 = jX_3 \quad -②$$

$$\rightarrow V_f = \frac{Z_1}{Z_1 + Z_3} \cdot V_o \quad \Rightarrow \quad \beta = \frac{V_f}{V_o} = \frac{Z_1}{Z_1 + Z_3} \quad -③$$

$$\rightarrow \text{Overall gain, } A_V = \frac{V_o}{V_i}$$

$$\begin{aligned} \text{From equivalent ckt, } \quad V_o &= -\frac{A_V \cdot V_i \cdot Z_L}{R_o + Z_L} \\ \Rightarrow A_V &= \frac{-A_V \cdot Z_L}{R_o + Z_L} \end{aligned}$$

$$\begin{aligned} \text{Now, } A_V \beta &= -\frac{A_V \cdot Z_L}{R_o + Z_L} \times \frac{Z_1}{Z_1 + Z_3} \\ &= \frac{-A_V \cdot Z_1 Z_2}{R_o (Z_1 + Z_2 + Z_3) + Z_2 (Z_1 + Z_3)} \end{aligned}$$

$$\left\{ \text{on putting } Z_L = \frac{Z_2 (Z_1 + Z_3)}{Z_1 + Z_2 + Z_3} \right\}$$

On substituting from eqn (2) -

$$A_V \beta = \frac{A_V \cdot X_1 X_2}{j R_o (X_1 + X_2 + X_3) - X_2 (X_1 + X_3)}$$

for freq. oscillation,  $\Im m = 0$

$$\Rightarrow \boxed{X_1 + X_2 + X_3 = 0}$$

Now, on substituting in  $\beta$  -

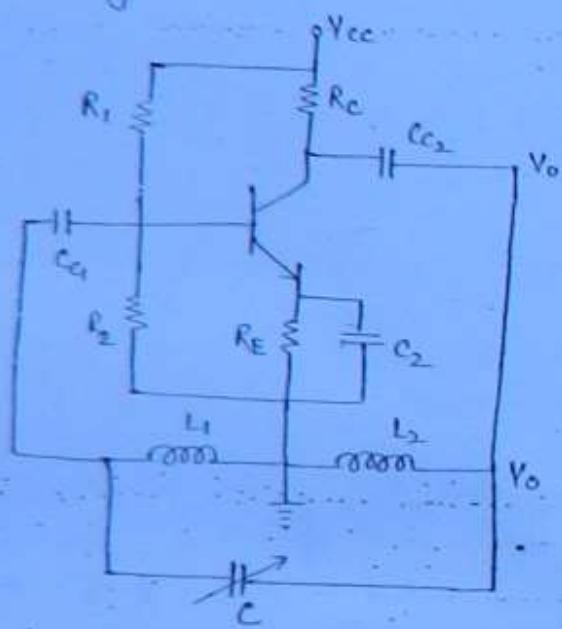
$$A_V \beta = \frac{A_V x_1 x_2}{-x_2(x_1+x_3)} = \frac{-A_V x_1}{(x_1+x_3)}$$

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$$\Rightarrow A_V \beta = \frac{A_V x_1}{x_2}$$

Now, for oscillation,  $|A_V \beta| > 1 \Rightarrow \boxed{A_V \geq \frac{x_2}{x_1}}$  → cond'n for oscillation

### Hartley Oscillator:



$$z_1 = j\omega L_1 ; z_2 = j\omega L_2 , z_3 = -j \frac{1}{\omega C}$$

freq. of oscillation :-

$$x_1 + x_2 + x_3 = 0$$

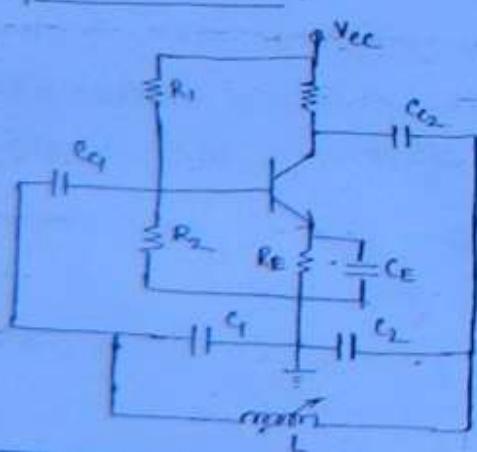
$$\Rightarrow \omega L_1 + \omega L_2 - \frac{1}{\omega C} = 0$$

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

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

Condition for oscillation :-  $A_V > \frac{x_2}{x_1} \Rightarrow \boxed{A_V \geq \frac{L_2}{L_1}}$

### Colpitt Oscillator:



$$z_1 = -j/\omega C_1 , z_2 = -j/\omega C_2 , z_3 = j\omega L$$

$$x_1 + x_2 + x_3 = 0$$

$$\Rightarrow \omega = \frac{1}{\sqrt{L \cdot \frac{C_1 C_2}{C_1 + C_2}}} ; \boxed{f = \frac{1}{2\pi \sqrt{L \cdot \frac{C_1 C_2}{C_1 + C_2}}}}$$

Condition:  $A_V \geq \frac{x_1}{x_2} \Rightarrow A_{V2} \geq \frac{C_1}{C_2}$  — Imp 225

Common Points :-

- They are variable freq. type RF oscillator
- Working principle is parallel resonance.

### Hartley Oscillator

→ It is also called Tapped inductor type oscillator

Advantage :- Capacitive tuning, ie., no wear & tear problem.

Disadvantage :- Bulky & expensive because of two inductors.

Application :- i) In designing of local oscillator ckt in receiver.

### Colpitt Oscillator

→ It has better freq. stability and it is obtained by reducing net capacitance of modified tank ckt.

$$C_{eq} = \frac{C_1 C_2}{C_1 + C_2}; Q = \frac{1}{\omega R C}; \text{ as } C \downarrow, Q \uparrow \text{ & stability} \uparrow$$

Advantage :-

- It is smaller in size and economical.

Disadvantage :- Inductive Tuning, ie., wear & tear problem.

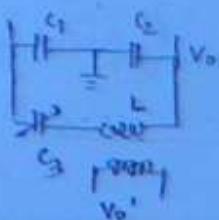
Application :- i) As a local oscillator in receiver.

### Clapp Oscillator

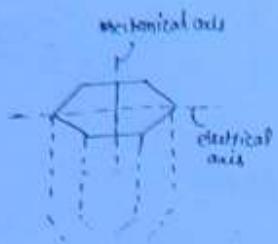
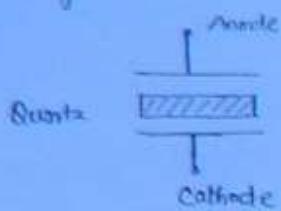
It is a modification of colpitt osc. where variable inductor is replaced by a variable capacitor  $C_3$  in series with an inductor  $L$  and if  $\phi$  is inductively obtained.

? Working principle is series resonance, therefore

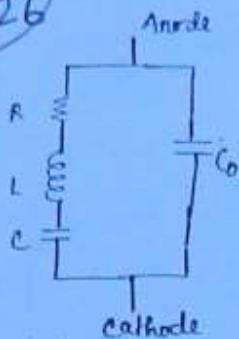
$$f_0 = \frac{1}{2\pi\sqrt{LC_3}}$$



### Crystal oscillator :-



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AC equivalent circuit

$\rightarrow$  internal losses or viscous damping.

$L \rightarrow$  Mass of crystal

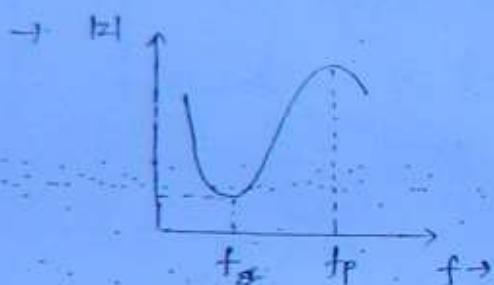
$C \rightarrow$  Stiffness =  $\frac{1}{\text{Spring constant}}$ ;  $C_0 = \text{capacitance b/w anode \& cathode plate}$ .

Series resonance :- Due to RLC in series ,

$\rightarrow$  Impedance  $\Rightarrow$  minimum.

$$\rightarrow f_s = \frac{1}{2\pi\sqrt{LC}}$$

$\rightarrow f_p > f_s$ , and freq. of oscillator varies b/w  $f_s$  and  $f_p$ .



Parallel Resonance :-

$\rightarrow$  Impedance maximum

$$\rightarrow f_p = \frac{1}{2\pi\sqrt{LC_{eq}}}; C_{eq} = \frac{C C_0}{C + C_0}$$

frequency of oscillation

$\rightarrow f_o$  depends of  $l, b, t \rightarrow$  Physical dimensions

$$\rightarrow f_o \propto \frac{1}{\text{thickness}} \Rightarrow \text{on higher frequencies, crystal becomes weak.}$$

$\rightarrow$  It is a fixed frequency type RF oscillator.

$\rightarrow$  It works on principle of piezoelectric effect.

$\rightarrow$  It has two resonating freq., i.e.  $f_s$  &  $f_p$ . Oscillating frequency lies b/w  $f_s$  &  $f_p$ .

$\rightarrow$  Due to high quality factor Q of a resonance ckt, it provides very good freq. stability.

$\rightarrow$  freq. of oscillation, generated by crystal depends on its physical dimensions but mainly on thickness.

$\rightarrow$  On high freq.,  $t$  should be small but it makes crystal mechanically weak.

Advantage:-

- Excellent freq. stability
- Simplest RF oscillator ckt.

Disadvantage:-

- Fixed freq type oscillator

- Application :-
- 1) To generate carrier in AM & FM transmission.
  - 2) In designing of timer circuit.

Frequency Stability -

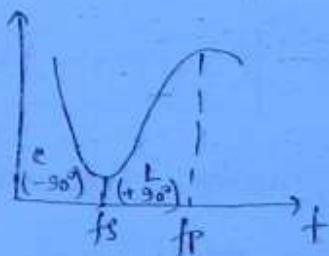
- freq. stability of an oscillator is measure of its ability to maintain as nearly a fixed freq as possible over as long a time interval as possible
- If  $d\theta$  is small change in phase angle and corresponding freq. change is  $df$ , then

$$\frac{d\theta}{df} = \text{figure of merit} \quad \text{and its value should be high.}$$

Ideally,  $\frac{d\theta}{df} = \infty$ .

(resistive  $\beta_{HW}$ )

- Inverting op-amp is preferred as compared to non-inverting as it has  $\beta_{HW}$  which is adaptable due to freq. change, ie, when  $f$  changes due to temp variations,  $\beta_{HW}$  will adjust itself so that overall change in freq. is very small.



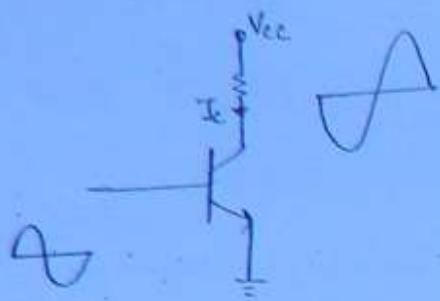
$$df = f_p^+ - (-f_s^-)$$

$$d\theta = 90^\circ - (-90^\circ) = 180^\circ$$

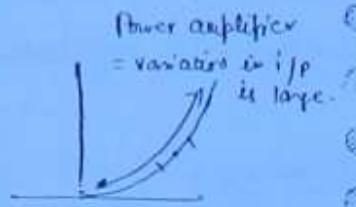
$$\frac{d\theta}{df} = \frac{180^\circ}{f_p^+ - f_s^-} = \frac{180^\circ}{0} = \infty \quad (\text{ideal}) \quad \text{for crystal oscillator}$$

## Power Amplifiers :-

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$$P_{dc} = V_{dc} \cdot I_{dc} = V_{cc} \cdot I_{CA}$$



- 1) It is last stage in multistage amplifier.
- 2) Power amplification is defined as ability of amplifier to convert available o/p dc power into ac signal power with the application of i/p signal.

### Small signal Amp.

→ i/p signal amplitudes are very small ( $\mu V$  or  $mV$ )

→ operated only in linear region

→ Important specifications are -

-  $A_I, A_V, R_f, R_o, \phi$

→ Analysis of amp. will be done by using graphical as well as mathematical analysis

→ Transistors used in power amp. are called power tri.

→ Power amplifiers are designed mostly by BJT & they are generally in class C mode

### Harmonic Distortion :-

→ In a power amp, signal amplitudes are very large, hence signal is

operated in linear & non-linear portion of i/p charac. curve., so we get harmonics in o/p & harmonic distortion is present at o/p.

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- Harmonic distortion is a non-linear distortion.

fourier series expn of collector current of power transistor :-

$$\rightarrow I_C = \underbrace{I_{C0}}_{DC} + \underbrace{B_1 \cos \omega t}_{\text{fundamental}} + \underbrace{B_2 \cos 2\omega t + \dots}_{\text{Harmonics}} \rightarrow \omega t, \text{ Amplitude} \downarrow.$$

$$\rightarrow 2^{\text{nd}} \text{ Harmonic distn} \rightarrow D_2 = \left| \frac{B_2}{B_1} \right|$$

$$\rightarrow 3^{\text{rd}} \dots \dots \dots \rightarrow D_3 = \left| \frac{B_3}{B_1} \right|$$

$\rightarrow$  AC power o/p due to fundamental component

$$P_{AC} = I_{rms}^2 \cdot R_o = \left( \frac{B_1}{2} \right)^2 \cdot R_o \quad \left\{ = P_1 \right\}$$

$\rightarrow$  Total Harmonic Power (THP) -

$$P_T = \frac{B_1^2}{2} \cdot R_o + \frac{B_2^2}{2} \cdot R_o + \dots$$

$$\Rightarrow P_T = \frac{B_1^2}{2} \cdot R_o \left[ 1 + \left( \frac{B_2}{B_1} \right)^2 + \left( \frac{B_3}{B_1} \right)^2 + \dots \right]$$

$$\Rightarrow P_T = P_1 \left[ 1 + D_2^2 + D_3^2 + \dots \right]$$

- Total Harmonic Distortion (THD) -

$$D = \sqrt{D_2^2 + D_3^2 + \dots}$$

$$\therefore P_T = P_1 (1 + D^2) \quad (\text{imp})$$

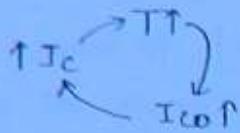
$$\text{for THD} = 10\% , D = 0.1$$

$$\therefore P_T = P_1 (1 + 0.01) = 1.01 P_1$$

$\Rightarrow P_T \approx P_1$ , ie, if THD is kept  $\leq 10\%$ , then THP is almost equal to fundamental power.

$\frac{dT_j}{dt} \rightarrow$  rate of heat dissipation

### Thermal Runaway :-



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→ The process where a transistor is subjected to self destruction due to excess heat produced in CB junction.

→ It is due to  $T_{eo}$ .

→ BJT suffers from thermal runaway. In FET there is no thermal runaway.

### Condition to eliminate thermal runaway:-

- A point of  $T_j$  is so selected that  $V_{CE} \leq \frac{V_{CC}}{2}$ .

$$\left| \frac{dP_c}{dT_j} \leq \frac{1}{\theta} \right|$$

$P_c$  = max. collector power dissipation in W.

$T_j$  = junc<sup>n</sup> temp. at collector juc<sup>n</sup>

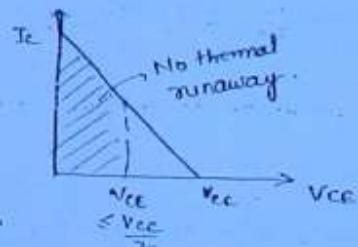
$\theta$  = Thermal resistivity in  $^{\circ}\text{C}/\text{watts}$ .

→  $T_j - T_A \propto P_D$

$$\Rightarrow T_j - T_A = \theta P_D$$

( $\theta$  should be small)

$T_A$  = ambient temp.



→ Area of collector  $\propto$ ,  $\theta \downarrow$ .

→ Diff. (1) w.r.t.  $T_j$  —

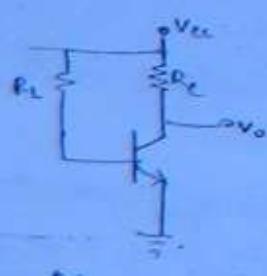
$$1 = \theta = \theta \frac{dP_D}{dT_j} \Rightarrow$$

$$\left| \frac{dP_D}{dT_j} = \frac{1}{\theta} \right|$$

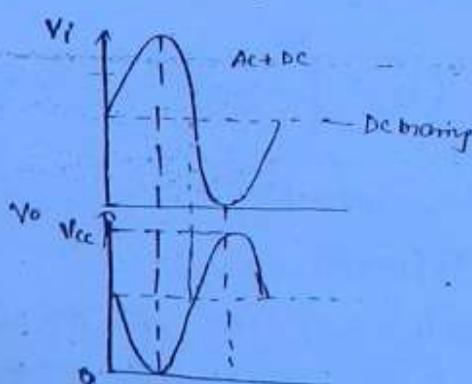
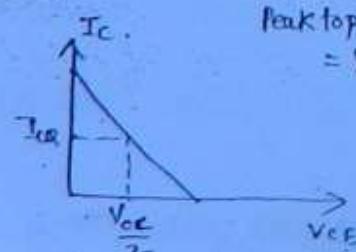
Rate of heat dissipation in atmosphere.

### Classifications of Amplifiers :-

#### Class A operation —



Cond<sup>n</sup> angle =  $2\pi$   
Peak-to-peak  
 $= V_{CC}$



Transformer is 100% efficient

$$\eta = \frac{\text{AC Power}}{\text{DC Power}}$$

- Collector current flows for entire  $360^\circ$  of i/p signal; conduct angle =  $2\pi$   
→ Q point is located at centre of ac load line

(23)

Advantage :- → Minimum distortion

→ Excellent thermal stability, ie, no thermal runaway problem

Disadvantage :- → Small power conversion efficiency

→ Reduced power gain, - Introduces Power drain -

When signal is not applied, transistor is consuming max. power & it is called power drain. When signal is applied, tr. is using less power.

Application :- designing of audio preq. amp.

Note - Class A amplifier is always designed with a single amplifier, ie, single ended, ie, one Tr. per stage.

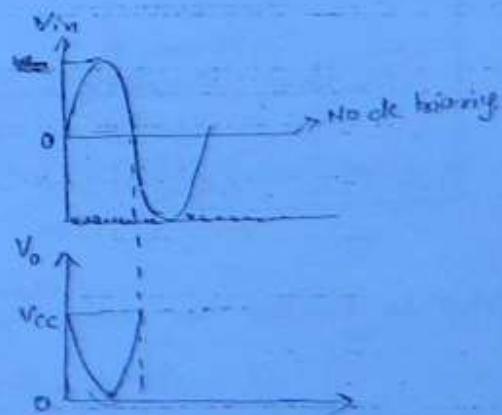
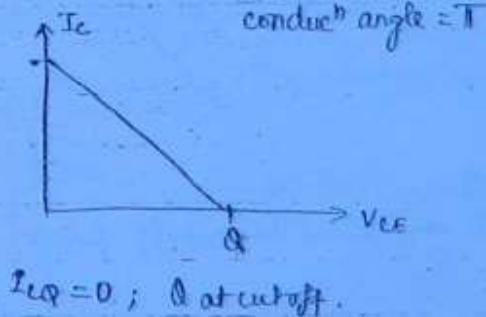
→ Power rating of transformer  $P_{D(\max)}$  → maximum allowable heat dissipation, is defined at room temp, ie.,  $25^\circ\text{C}$ .

→ In class A operation, power dissipated by Tr is equal to max. signal power o/p.

→ For class A,  $P_D = P_{D(\max)}$ ; ie, max. power o/p.

∴ To design a class A amp. with  $20\text{W}$  app signal power, Tr must dissipate  $20\text{W}$  of power.

Class B operation :-



amplifier  
for class B  
amplifier

$\Rightarrow 2V_{cc}$

- Collector current flows exactly for  $180^\circ$  of i/p signal
- Q point is located at cutoff
- It is a double ended amplifier, i.e., two transistors in one stage.

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Advantage :- Higher efficiency. ( $78.5\%$ )

- Power drain is eliminated.

Disadvantage :- Higher distortion

- Thermal stability is less.

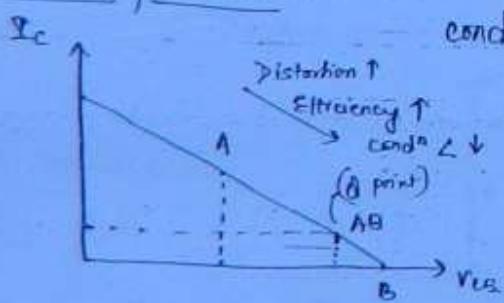
- Introduces crossover distortion (CD)  $\rightarrow$  major disadvantage.

Application :- Used in designing of Power amp., for ex., push-pull power amp., complementary symmetry push pull power amp.

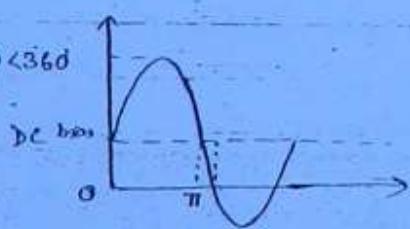
- When signal is applied, Tr is consuming power & when signal is absent, Tr. will not consume any power, therefore no power drain.
- Power dissipated by single Tr. in ckt,
- $$P_D = 0.2 P_{max} \quad P_{max} = \text{max. o/p signal power.}$$
- Power dissipated by circuit, i.e., by two tr.
- $$P_D = 0.4 P_{max}$$

For eg., to design a class B amplifier, with  $30W$  o/p signal power, power dissipated by single transistor should be  $4W$ .

Class AB operation :-



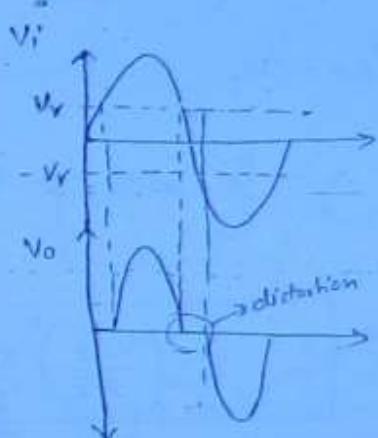
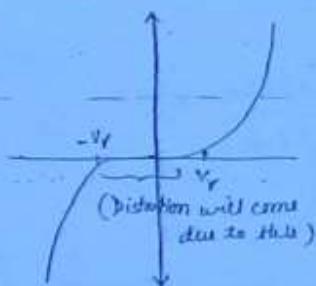
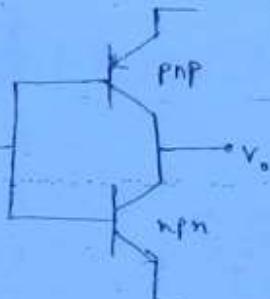
condn. angle -  $180^\circ < \phi < 360^\circ$



$\rightarrow$  Q point is located in active region but very close to cutoff point.

- Distortion & noise interference is more as compared to class A & less when compared to class B.
- It is used in power amp for ex. push pull power amp. (23)
- The main advantage of class AB operation is it eliminates COB.
- Max. efficiency is approx. 60%.

### Crossover Distortion :-

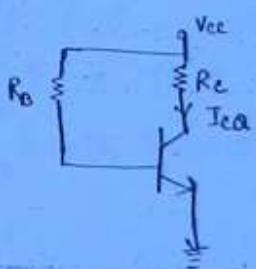


- It is a distortion arising when conduction transfer from one tr. to other.
- It is a non-linear distortion.
- It is due to operating the signal over non-linear characteristic curve. portion of
- Class B introduce COB. and class AB eliminates COB.
- The most suitable remedy to minimise COB is to use Ge Tr. in place of Si Tr. But this will reduce power handling capability of circuit.

12/07/2012

### Class A amplifier:-

#### Direct Coupled Amplifier :-



$$I_{cA} = \frac{I_{emax}}{2} = \frac{V_{cc}}{2R_c} \quad (\because Q\text{-point is in centre})$$

$$\underline{\text{DC Power}} \therefore P_{dc} = V_{dc} \cdot I_{dc}$$

$$\Rightarrow P_{dc} = V_{cc} \cdot I_{cA} \Rightarrow P_{dc} = \frac{V_{cc}^2}{2R_c}$$

AC power  
RMS

$$P_{AC} = V_{rms} \cdot I_{rms} = \left[ \frac{V_{rms}^2}{R_L} \right]$$

(234)

Peak  $P_{AC} = \frac{V_p}{\sqrt{2}} \cdot \frac{I_p}{\sqrt{2}} = \frac{V_p I_p}{2} = \left[ \frac{V_p^2}{2R_L} \right]$

Peak-to-RMS  $P_{AC} = \frac{V_{p-p}}{2\sqrt{2}} \times \frac{I_{p-p}}{2\sqrt{2}} = \frac{V_{p-p} I_{p-p}}{8} = P_p \left[ \frac{V_{p-p}^2}{8R_L} \right]$

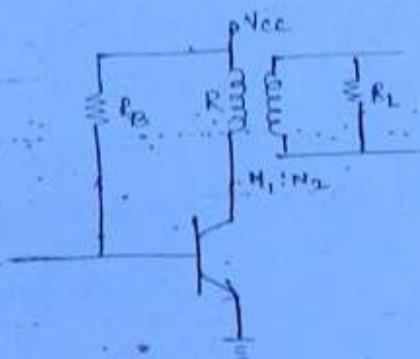
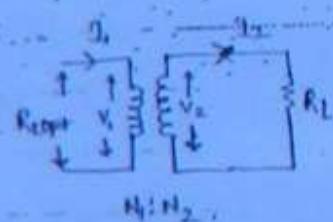
\* Ideally,  $V_{p-p} = V_{cc}$

Efficiency:  $\eta = \frac{P_{AC}}{P_{DC}} \times 100 \Rightarrow \eta = \frac{V_{p-p}^2 / 8R_L \times 100}{V_{cc}^2 / 2R_L}$

$$\Rightarrow \eta = \frac{1}{4} \left( \frac{V_{p-p}}{V_{cc}} \right)^2 \times 100 \%$$

$$\rightarrow \eta_{max} = \frac{1}{4} \times 100 \times \left( \frac{V_{cc}}{V_{cc}} \right)^2 \Rightarrow \boxed{\eta_{max} = 25\% ; \text{ Practically } \Rightarrow 10-15\%}$$

Transformer Coupled Amplifier:



$$\frac{V_1}{V_2} = \frac{N_1}{N_2}, \quad \frac{I_2}{I_1} = \frac{N_2}{N_1}; \quad R_{load} = \text{optimum resistance or reflected resistance}$$

$$R_{load} = \frac{V_1}{I_1} \Rightarrow R_L = \frac{V_2}{I_2} \Rightarrow \frac{R_{load}}{R_L} = \left( \frac{N_1}{N_2} \right)^2$$

$$\Rightarrow \boxed{R_{load} = \left( \frac{N_1}{N_2} \right)^2 \times R_L} \quad (\text{Imp})$$

Transformer provides -  
- DC isolation  
-  $R_L$  adjustment

From Graphical analysis :

$$\eta = 50 \times \left( \frac{V_{CEmax} - V_{CEmin}}{V_{CEmax} + V_{CEmin}} \right)^2 \%$$

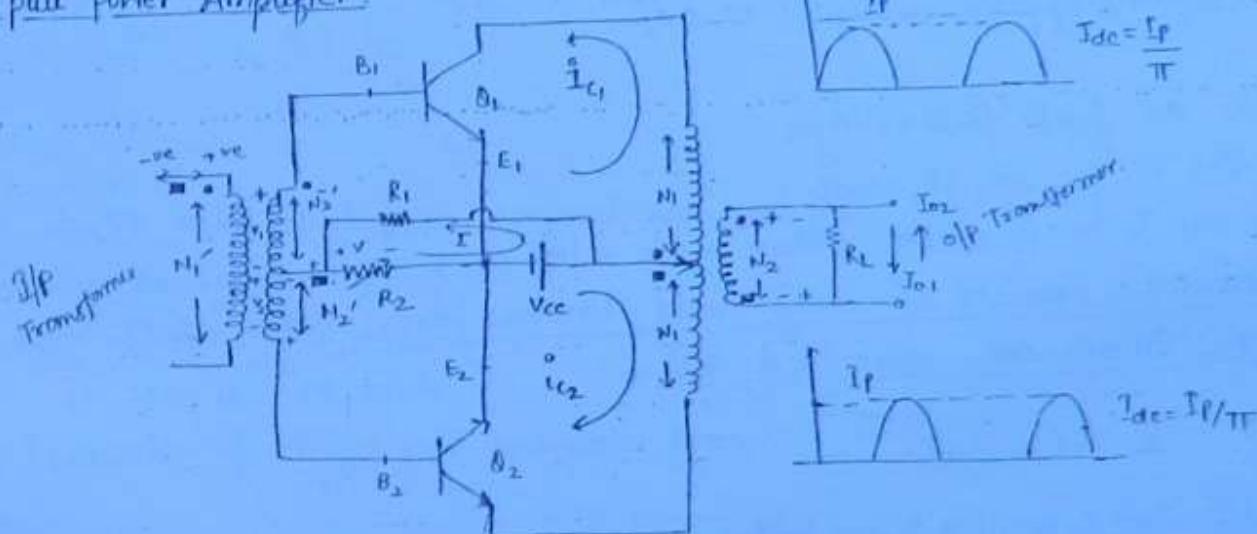
→ ideally,  $V_{CEmin} = 0$

$$\Rightarrow \eta_{max} = 50\%$$

Practically,  $\eta \approx 30-35\%$

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### Push-pull Power Amplifier:



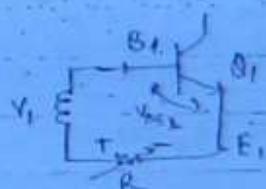
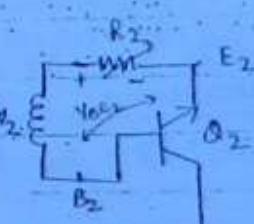
During +ve half,  $Q_1$  (ON),  $Q_2$  (OFF).

$$I_{o1} \propto i_c$$

During -ve half,  $Q_1$  (OFF),  $Q_2$  (ON)

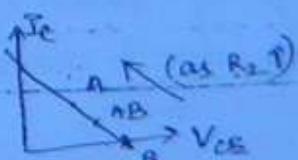
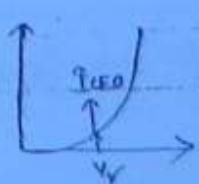
$$I_{o2} \propto i_c$$

$$\therefore I_o \propto (i_{c1} - i_{c2}) \Rightarrow I_o = K(i_c - i_c)$$



When signal is absent —  $V_1 = V_2 = 0$ .

- let  $R_2 = 0$ ,  $\Rightarrow V_{BE1} = V_{BE2} = 0 \rightarrow$  Both  $Q_1$  &  $Q_2$  in cutoff  $\rightarrow$  class B operation.
- $R_2 \uparrow \rightarrow IR_2 \uparrow$  and when  $IR_2 = V_F \therefore V_{BE1} = V_{BE2} = V_F \rightarrow$  class AB operation.
- $R_2 \uparrow, IR_2 \uparrow$  and  $V_{BE} \uparrow$  and  $\&$  will move towards saturation  $\rightarrow$  class A operation.



→  $I_{CEO}$  is the standby current during class AB operation.

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→ It is double ended amp.

→ It can be class B or class AB operated.

→ Designed with identical transformers Tr.

→ The circuit operates in class B when  $R_2 = 0$ .

→ For class AB operation, voltage drop across  $R_2$  is adjusted to be opposite equal to  $V_T$ , where a small standby current flows at zero excitation.

→ The function of centre tapped secondary coil of i/p transformer is to provide two equal & opposite voltages  $V_1$  &  $V_2$ .

→  $V_1$  &  $V_2$  are push pull voltages

→ Both the Tr. are in CE mode.

→ When one Tr. is in active, other is in cutoff.

→ o/p current consists of only odd harmonic terms since, in o/p I, even harmonic terms are cancelled out.

Proof:  $\Rightarrow \because I_o = K(i_{C1} - i_{C2})$  (from notes)

$$i_{C1} = B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t + \dots$$

$$i_{C2} = B_0 + B_1 \cos(\omega t + \pi) + B_2 \cos(2\omega t + \pi) + \dots$$

$$= B_0 - B_1 \cos \omega t + B_2 \cos 2\omega t + \dots$$

$$\therefore I_o = 2K(B_1 \cos \omega t + B_3 \cos 3\omega t + B_5 \cos 5\omega t + \dots)$$

→ First available harmonic distortion  $D_3 = \left| \frac{B_3}{B_1} \right|$  (very small)

Note: If  $B_1, B_3, B_5$  are not identical, then even harmonics will be present in o/p & distortion will be large.

Advantages: 1) Higher power due to double ended.

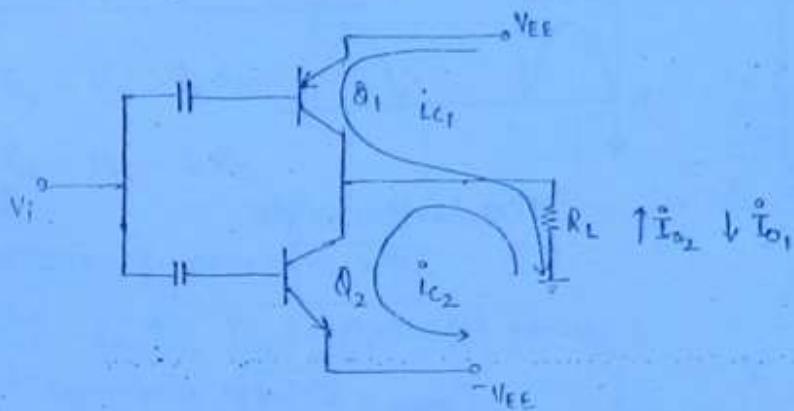
2) " Efficiency if class B operated

3) less distortion due to cancellation of even harmonics.

Disadvantage :- Very bulky & highly expensive due to requirement of bulky transformer.

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### Complementary - Symmetry Push Pull Power Amplifier :-



For  $V_i > 0$  -

$$Q_1 = \text{ON}, Q_2 = \text{OFF}, i_{o_2} = i_{e_2}$$

for  $V_i < 0$  -

$$Q_1 = \text{OFF}, Q_2 = \text{ON}, i_{o_1} = i_{e_1}$$

- It is double ended amplifier designed with matched pairs of Tr.
- Popularly used Power amp ckt.
- Always class B operated.
- Both Tr. are in CE mode.
- o/p T consists of only odd harmonic terms.

Advantage :- - same as push-pull B amplifier.

- circuit is smaller in size & economical due to elimination of bulky transformer.

Disadvantage :- - Requires two power supply  
- introduces COB

$$\text{Efficiency} : \eta = \frac{P_{ac}}{P_{dc}} \times 100\%$$

→ ideally,  $V_p = V_{cc}$ .

$$\therefore P_{ac} = \frac{V_p^2}{2R_L}$$

$$\therefore \boxed{\eta = \frac{\pi}{4} \times \left( \frac{V_p}{V_{cc}} \right) \times 100\%}$$

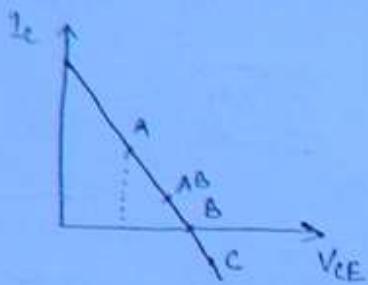
when  $Q_1$  ON      when  $Q_2$  ON

$$P_{dc} = V_{cc} \times \frac{I_p}{\pi} + V_{cc} \times \frac{I_p}{\pi}$$

$$\therefore \boxed{P_{dc} = \frac{2V_{cc} \cdot I_p}{\pi} = \frac{2V_{cc} \cdot V_p}{\pi R_L}} \quad \text{---(imp)}$$

$$\eta_{max} = \frac{\pi}{4} \times 100\% \Rightarrow \boxed{\eta_{max} = 78.5\%}$$

### Class C amplifier :-



→ Duty cycle :

$$\frac{\tau_p}{T} \times 100\% = D$$

→ Power dissipation across Tr. during  $\tau_p$  -

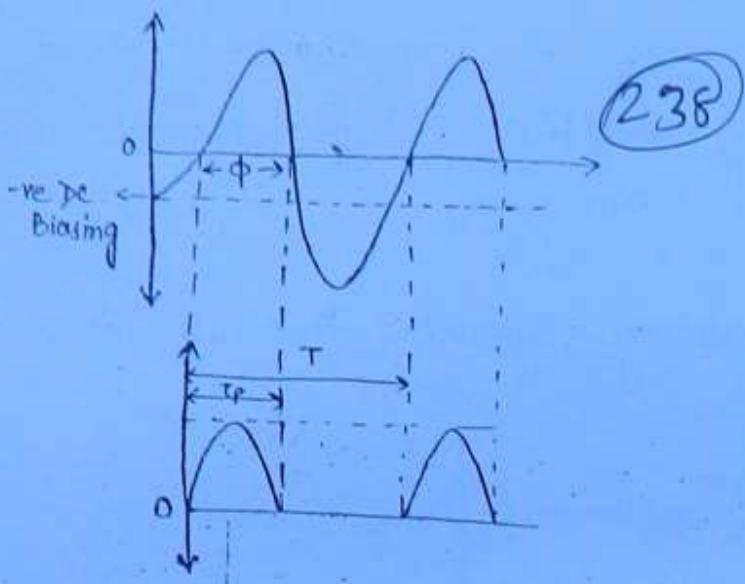
$$P_D = V_{CE} \cdot I_c$$

→ Energy dissipation across Tr. during  $\tau_p$  -

$$E_D = P_D \cdot \tau_p$$

→ Avg. power dissipation during one cycle :-

$$P_{D\text{avg}} = \frac{E_D}{T} = \frac{P_D \cdot \tau_p}{T} \Rightarrow P_{D\text{avg}} = P_D \cdot D$$



→ conduction angle

$$\phi < 180^\circ$$

→ Efficiency -

$$\eta_{\text{max}} = 87.5\%$$

→ Distortion is very high.

### Class D Amplifier :-

- They are special amplifier designed to operate with digital pulse signal.
- Efficiency of class D is above 90%.
- It is not a power amplifier.
- Widely used in commercial application.

## Multivibrator by using Transistors:

2.39

### Bistable Multivibrator -

#### a) Fixed Bias Binary -

$$V_{C_2} = V_{CC} - I_2 R_C$$

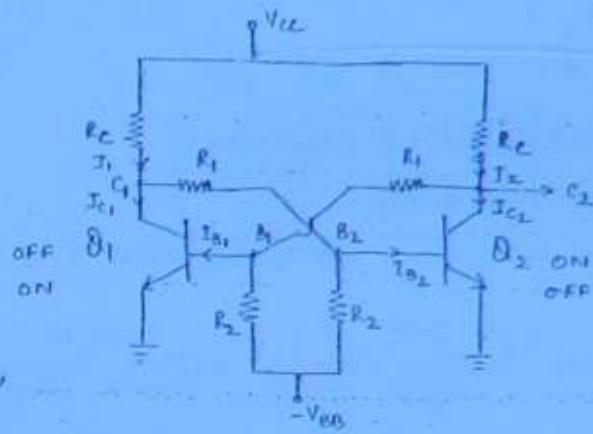
$$V_{C_1} = V_{CC} - I_1 R_C$$

→ Because of noise -

$$\rightarrow V_{B_2} \uparrow, I_{B_2} \uparrow, I_{C_2} \uparrow, I_2 \uparrow, V_{C_2} \downarrow, V_{B_1} \downarrow$$

"Regenerative action"

$$V_{C_1} \uparrow, I_1 \downarrow, I_{C_1} \downarrow, I_{B_1} \downarrow$$



→ Finally  $Q_2$  in saturation and  $Q_1$  in cutoff.

$$Q = V_{C_2} = V_{CESAT} = 0 ; \bar{Q} = V_{C_1} \approx V_{CC} = 1$$

When a -ve pulse is applied -

[at  $B_2$ ]  $\rightarrow V_{B_2} \downarrow, I_{B_2} \downarrow, I_{C_2}, I_2 \downarrow, V_{C_2} \uparrow, V_{B_1} \uparrow, I_{B_1} \uparrow, I_{C_1}, I_1 \uparrow, V_{C_1} \uparrow$

"Regenerative action"

Finally  $Q_2$  in cutoff &  $Q_1$  in saturation -

$$Q = V_{C_2} \approx V_{CC} = 1 ; \bar{Q} = V_{C_1} = V_{CESAT} = 0$$

When  $Q_2 = \text{on}$ ;  $Q_1 = \text{off}$   $\rightarrow Q_2$  should be well in saturation &  $Q_1$  should be well in cutoff.

$$V_{C_2} = V_{CESAT}, V_{C_1} = \frac{V_{CC}R_1}{R_1+R_C} + \frac{V_{BESAT}R_C}{R_1+R_C} \approx V_{CC}$$

$$\rightarrow I_0 = I_{B_1} = 0$$

$$\rightarrow V_{C_1} = \frac{V_{CESAT}R_2}{R_1+R_2} - \frac{V_{BB} \cdot R_1}{R_1+R_2} ; V_{B_1} \approx 0 \text{ when } V_{BB} = 0.0 \rightarrow \text{Noise Margin}$$

$V_{B_1}$  = well in cutoff when  
V<sub>BB</sub> is present  
(Regd)

⇒ Noise margin  
more  
0.5  
 $\uparrow = +2.5$   
 $-2$  (more)

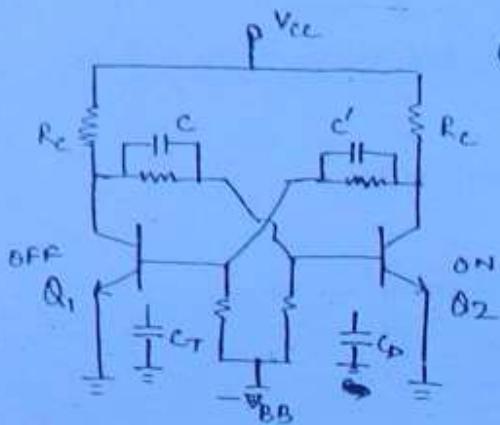
## Commutating capacitors - (c & c')

→  $C_s$  &  $C_t$  are transition & diffusion capacitances of ON & OFF Tr. respectively.

→  $c$  &  $c'$  are speed up capacitors.

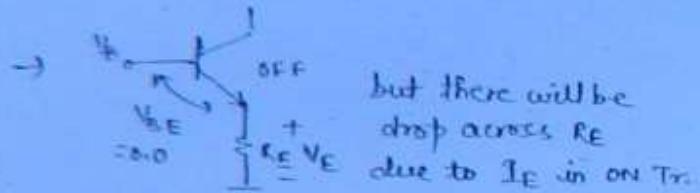
→  $c$  &  $c'$  → very small

→ Due to  $c$  &  $c'$  → ↓ in transition time or ↓ propagation delay.



(240)

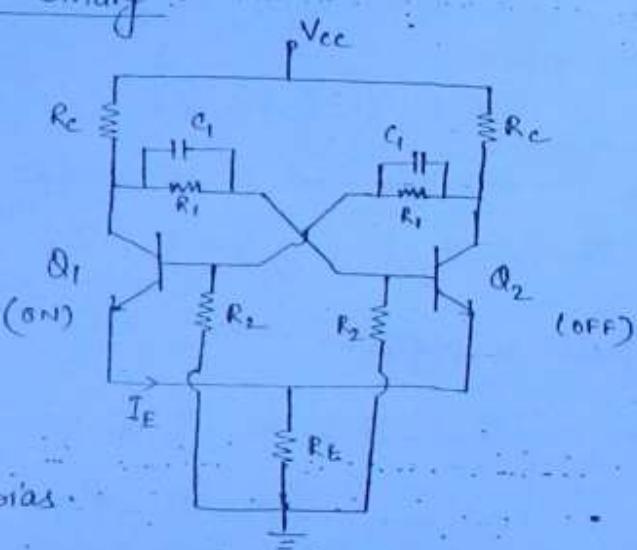
## Self Biased Binary / Emitter Coupled Binary



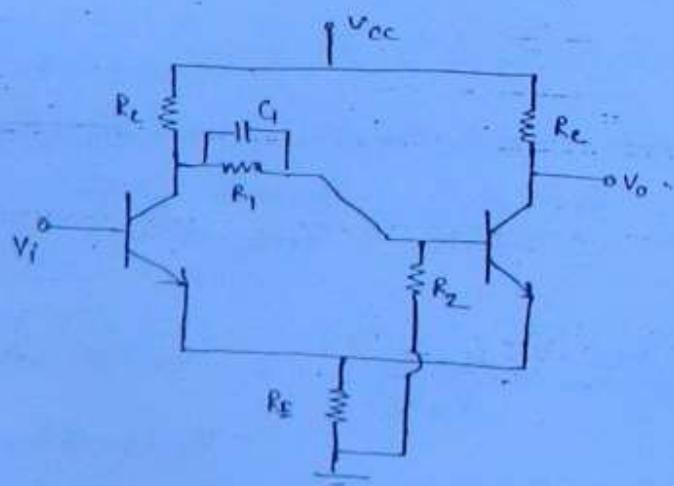
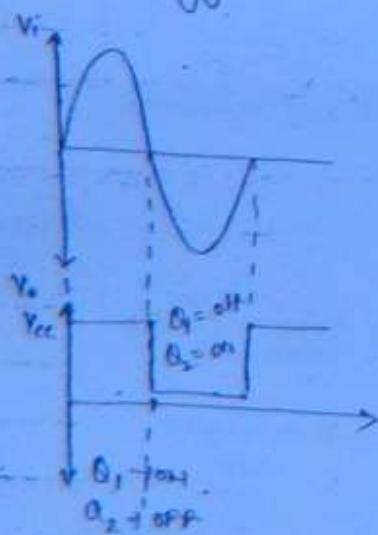
and it min. voltage reqd. to ON  
Q<sub>2</sub> is atleast  $(V_E + 0.5)$  and hence

-  $V_{BE}$  is not required in this ckt.

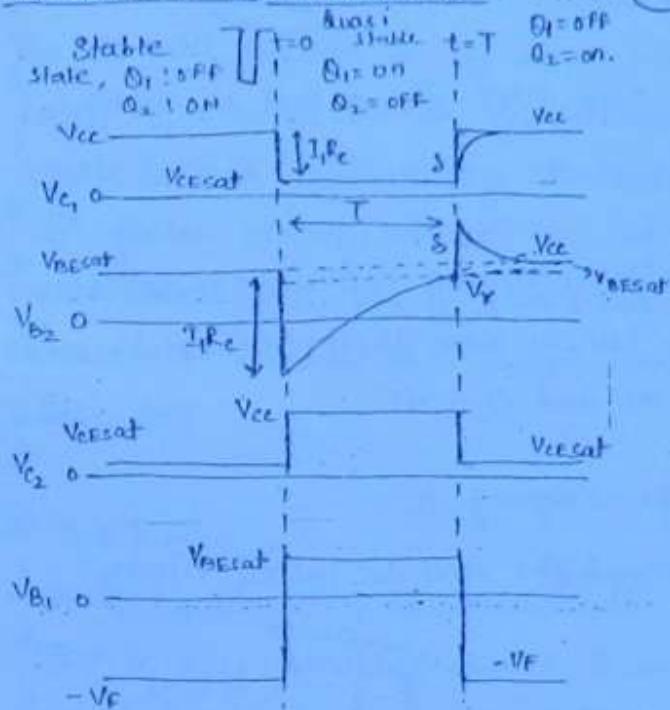
→ Other operation is same as fixed bias.



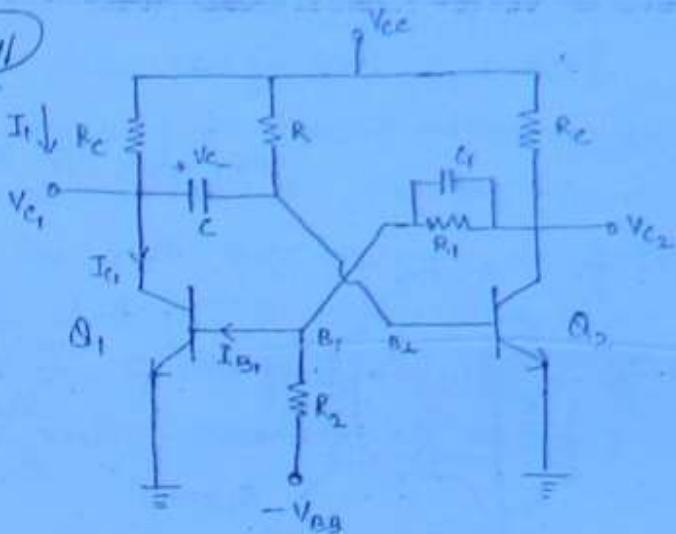
## Schmitt Trigger:



## Monostable Multivibrator :-



(24)



For  $t < 0$  ckt is in stable state.

$Q_1: OFF, Q_2: ON$

$$V_{C_2} = V_{CESAT}; V_{B_2} = V_{BESAT}$$

$\therefore$  capacitor will act as open circuit.

$$I_{C_1} = I_{B_1} = 0, \therefore \text{current through } R_C = 0$$

$$\text{Hence } V_{C_1} = V_{C_2} = V_{CC}$$

$$\text{and } V_{B_1} = \frac{V_{CESAT} \cdot R_2}{R_1 + R_2} - \frac{V_{BB} \cdot R_1}{R_1 + R_2} = -V_F$$

$\rightarrow$  At  $t \leq 0$ , voltage across C -

$$V_C = V_{CC} - V_{CESAT}$$

$$V_{C_1} = V_{C_2} = V_{CESAT}$$

$\rightarrow$  For  $t > 0$ . A trigger is applied and  $Q_2: OFF$  &  $Q_2: ON$ .

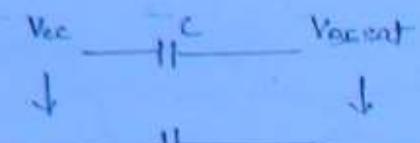
When  $Q_2 = OFF$ ,  $V_{C_2} = V_{CC}$  and it will be transferred to  $B_1$  by commutating capacitor.

& Now,  $Q_1$  will be ON due to this. &  $V_{C_1} = V_{CESAT}$ .

$\rightarrow$  Capacitor C is called timing element (capacitor) and its value is very large and it does not allow sudden change.

$\rightarrow$  A current  $I_1$  will flow in  $Q_1$ .

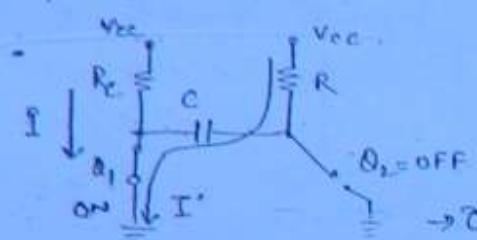
$$I_1 = \frac{V_{CC} - V_{CESAT}}{R_C}$$



(242)

$V_{c1} = V_{cc} - I_1 R_1 \Rightarrow V_{c1} < 0$  and  $Q_2$  is well in cutoff.

Now, for  $0 \leq t \leq T$

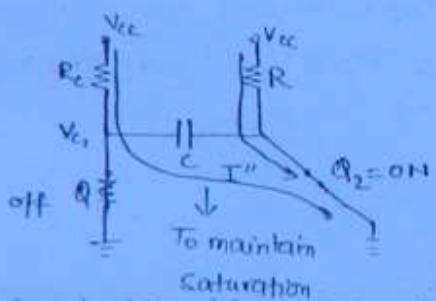


I will start charging capacitor towards  $V_{cc}$ , but as soon as it reaches  $V_r$ ,  $Q_1 = ON$  and  $Q_2 = OFF$ .  
 $\rightarrow C = RC$  - for charging C.

$R$  &  $C$  → very high. and are called timing elements.

for  $t > T$

At  $t = T'$ ,



$\Delta t = T'$   $\because C$  is large, it does not allow sudden change ( $S \gg V_r$ )

Now,  $I''$  will start charging C and  $V_{c1}$  will start ↑ towards  $V_{cc}$  and  $V_r + S$  will start discharging and will settle at  $V_{cesat}$ . (see graph)

### Important Points:

→ Waveform:

for  $t < 0$  :- The circuit is in stable state with  $Q_2(ON)$  &  $Q_1(OFF)$ . Capacitor C will act as open ckt.

for  $0 < t < T$  + Quasi Stable state.

- On application of -ve trigger, at  $t=0$  to base  $B_2$ , a regenerative action takes place driving  $Q_2$  below cutoff. Now voltage at  $C_2$  rises to  $V_{cc}(OFF)$  and because of cross coupling b/w  $G$  &  $B_2$ ,  $Q_1$  comes into saturation.

- Now current  $I_1$  exist in  $R_C$  of  $Q_1$  and  $V_{c1}$  drops abruptly by an amount  $I_1 R_C$  upto  $V_{cesat}$ . The voltage at  $B_2$  drops by same amount  $I_1 R_C$  since  $C_1$  &  $B_2$  are capacitively coupled.

- Now the multivibrator is in Quasi stable state with  $Q_1$  (on),  $Q_2$  (off).
- The circuit will remain in Q<sub>2</sub> state for only a finite time T because (24) base  $B_2$  is connected to  $V_{cc}$  through a resistance R, therefore  $V_{B_2}$  starts to rise exponentially towards  $V_{cc}$  with time constant  $RC$  & when it passes cutin voltage  $V_f$  of  $Q_2$  at  $t=T$ , a regenerative action will take place as a result of which  $Q_1$  will go into cutoff &  $Q_2$  comes into conduction and multivibrator returns to its initial stable state.

For  $t \geq T$  -

- At  $t=T^+$ ,  $Q_1$  = off,  $Q_2$  = conducted.  $V_{C_1}$  drops to  $V_{ccsat}$ .  $V_{C_1}$  returns to  $-V_f$ . Now  $V_{C_1}$  rises abruptly since  $Q_1$  is off. This  $T$  in  $V_{C_1}$  transmitted to base of  $Q_2$  and  $Q_2$  goes into oversaturation. Hence an overshoot S develops in  $V_{B_2}$  at  $t=T^+$  which decays as  $\exp(-t/T)$  after capacitor recharge.

Derivation of T :

C null charge

$$V_{B_2} = V_f - (V_f - V_i) e^{-t/T}$$

$$\rightarrow V_f = V_{cc}, \quad V_i = V_{ccsat} - I_1 R_C \quad \text{where } I_1 R_C = V_{cc} - V_{ccsat}$$

$$\rightarrow T = R_C$$

$$\therefore V_{cc} - (V_{cc} - V_{ccsat} + V_{cc} - V_{ccsat}) e^{-t/R_C} = V_{B_2}$$

$$\Rightarrow V_{B_2} = V_{cc} - [2V_{cc} - (V_{ccsat} + V_{ccsat})] e^{-t/R_C}$$

$$\text{at } t=T, V_{B_2} = V_f$$

$$V_f = V_{cc} - [2V_{cc} - (V_{ccsat} + V_{ccsat})] e^{-T/R_C}$$

$$\Rightarrow T = R_C \ln \left[ \frac{2V_{cc} - (V_{ccsat} + V_{ccsat})}{V_{cc} - V_f} \right]$$

$$\Rightarrow T = RC \ln 2 + RC \ln \left( \frac{V_{cc} - \frac{V_{BEsat} + V_{cesat}}{2}}{V_{cc} - V_f} \right)$$

(244)

For Si -

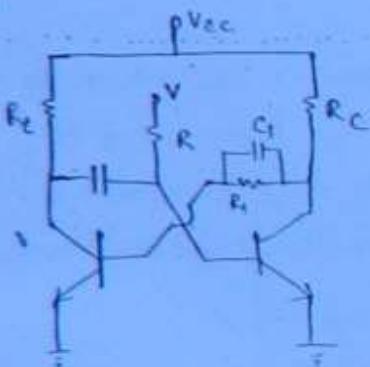
$$\frac{V_{BEsat} + V_{cesat}}{2} = \frac{0.8 + 0.2}{2} = 0.5^{\circ} = V_f.$$

$$T = RC \ln 2 + R_C \ln \left( \frac{V_{cc} - V_f}{V_{cc} - V_r} \right)$$

$$\Rightarrow T = RC \ln 2 = 0.69 RC$$

(Workbook)

\*  
Chap 11  
Conv - 1



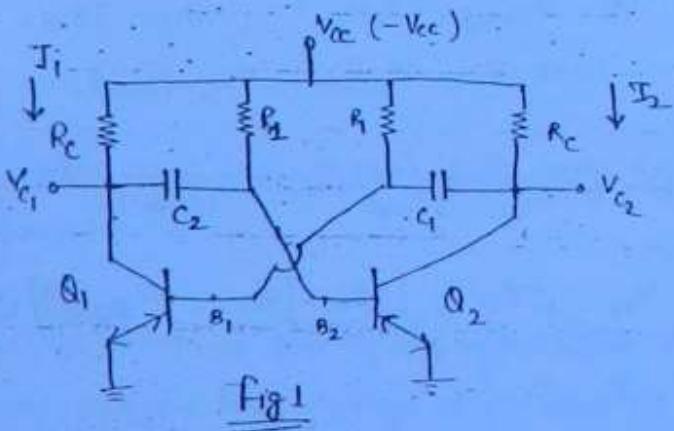
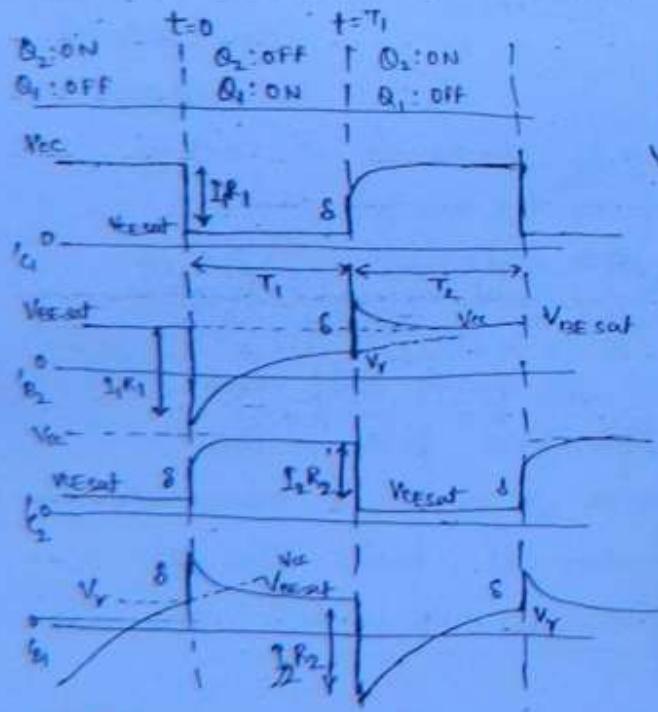
In the above derivation change

$$V_f = V \quad \& \text{ assume } V_{BEsat} + V_{cesat} \ll V_{cc} \quad \& V_f \ll V.$$

$$\Rightarrow T = RC \ln \left( 1 + \frac{V_f}{V} \right)$$

Diagram for voltage controlled f (or T)

### Astable Multivibrator

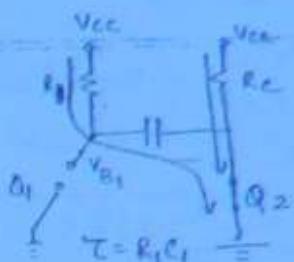


for  $t < 0$   $Q_1: OFF ; Q_2: ON$

$$V_{C_2} = V_{CEsat} ; \quad V_{B2} = V_{CEsat}$$

$$I_{C_1} = I_{B_1} = 0 ; \quad V_{C_1} = V_{CC} \quad \& \quad V_{B_1} < 0.$$

(245)



$\rightarrow C_1$  will charge &  $V_{B_1} \uparrow$  till  $V_r$  & then

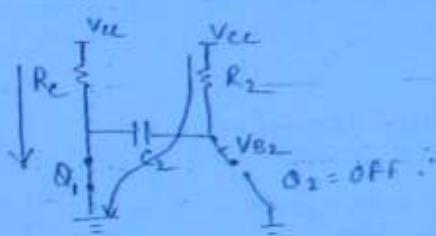
states will change with sudden change in  $V_{C_2}$  &  $V_{B_2}$ .

$\rightarrow Q_1: ON \quad Q_2: OFF \quad$  For  $t \leq 0$

for  $0 < t \leq T_1$  —  $Q_2: OFF ; Q_1: ON$

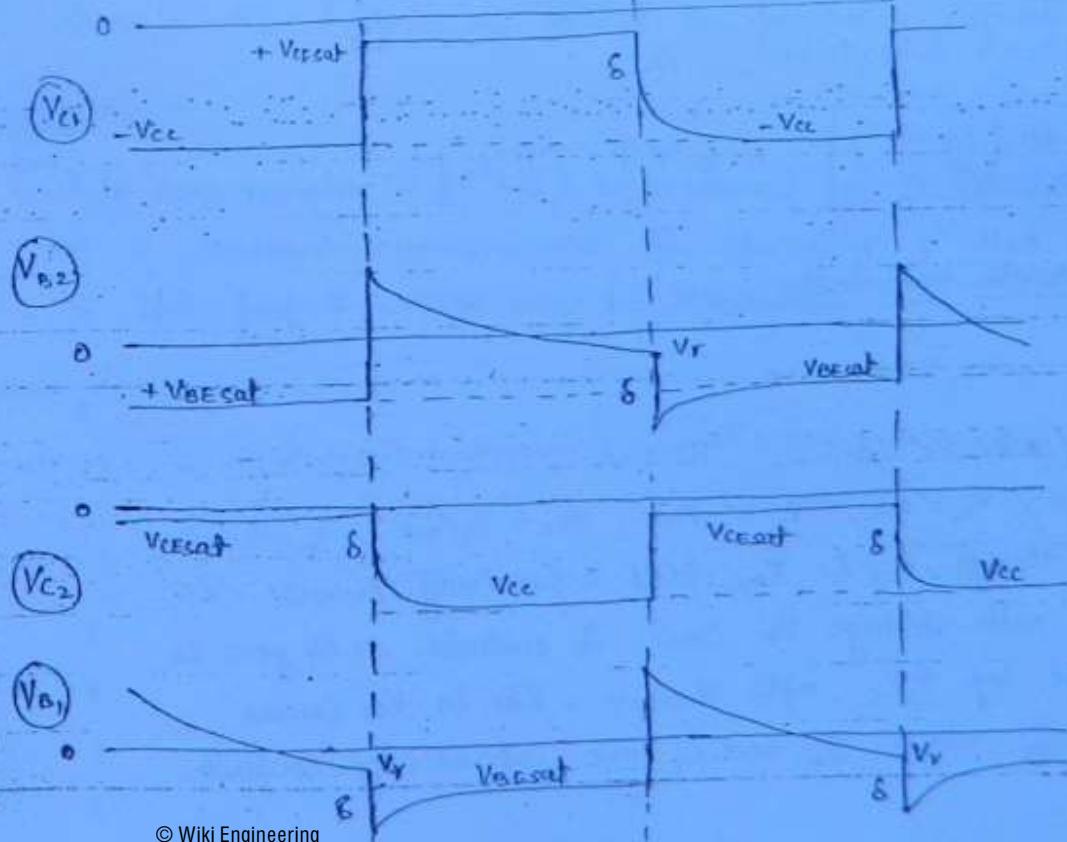
Now  $C_2$  will start charging

& similar process as above will be repeated



$\rightarrow$  for  $n-p-n$  Tr

$Q_2: ON$	$t=0$	$Q_2: OFF$	$t=T_1$	$Q_2: ON$	$t=T_1+T_2$
$Q_1: OFF$		$Q_1: ON$		$Q_1: OFF$	



$$\rightarrow T_1 = 0.69 R_2 C_2 \quad ; \quad T_2 = 0.69 R_1 C_1 \quad \left\{ \because T_1 \neq T_2, D \neq 50\% \Rightarrow \text{Asymmetrical square wave} \right.$$

(246)

$$\rightarrow T = T_1 + T_2 = 0.69 (R_1 C_1 + R_2 C_2)$$

$$\rightarrow f = \frac{1}{T} = \frac{1.44}{R_1 C_1 + R_2 C_2} ; \text{ if } R_1 = R_2 = R \text{ & } C_1 = C_2 = C$$

↓  
for asymmetrical sq. wave

$$\text{then } T = 1.38 R C$$

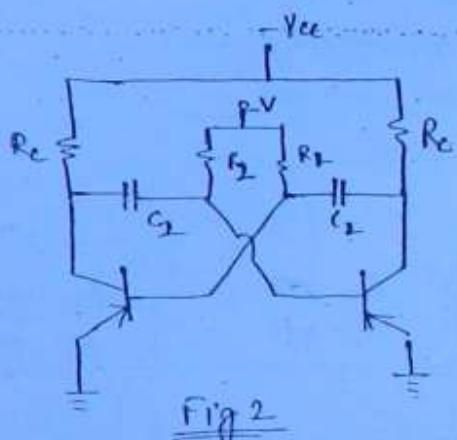
↓  
for symmetrical sq. wave

→ Voltage to freq. converter :-

$$T_1 = R_2 C_2 \ln \left( 1 + \frac{V_{cc}}{V} \right)$$

$$T_2 = R_1 C_1 \ln \left( 1 + \frac{V_{cc}}{V} \right).$$

$$T = (R_1 C_1 + R_2 C_2) \ln \left( 1 + \frac{V_{cc}}{V} \right)$$



→  $f = \frac{1}{T}$  & if  $R_1 = R_2 = R$  &  $C_1 = C_2 = C$ , then

$$T = 2RC \ln \left( 1 + \frac{V_{cc}}{V} \right) \quad \therefore \text{as } V \uparrow, T \downarrow \text{ & } f \uparrow;$$

Important Points for Astable Multivibrator :-

Waveform (PnP)

for  $t < 0$   $Q_1$ : OFF,  $Q_2$ : ON

Hence, for  $t < 0$ ,  $V_{Q_1} = +V_c$ ,  $V_{C_1} = -V_{cc}$ ,  $V_{B_2} = V_{BEsat}$ ,  $V_{B_1} = V_{CEsat}$

→ Capacitor  $C_1$  charges through  $R_2$  &  $V_{B_1}$  falls exponentially towards  $-V_{cc}$ .

→ At  $t=0$ ,  $V_{B_1}$  reaches cutin voltage  $V_r$  and  $Q_1$  conducts. As  $Q_1$  goes to saturation,  $V_{C_1}$  rises by  $I_1 R_C$  upto  $V_{CEsat}$ . Rise in  $V_{C_1}$  causes equal rise  $I_1 R_C$  in  $V_{B_2}$  since  $B_2$  and  $C_1$  are capacitively coupled.

- Rise in  $V_{B_2}$  cuts off  $Q_2$  and its collector falls towards  $-V_{cc}$ . This fall in  $V_{C_2}$  is coupled through capacitor  $C_1$  to base  $B_1$ , causing undershoot  $\delta$ . in  $V_{B_1}$  and abrupt anomalous drop by same amount  $\delta$  in  $V_{C_2}$ .

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- The voltage  $V_{B_2}$  is  $V_{cc}at + \frac{1}{R_2}t$  at  $t=0^+$  and decays exponentially with time constant  $R_2 C_2$  towards  $-V_{cc}$ . At  $t=T_p$ ; base  $B_2$  reaches cutin level  $V_r$  and reverse transition takes place.

In Fig(1), the frequency of oscillation may be varied over the range from Hz to MHz by adjusting  $R$  or  $C$ . It is also possible to change  $T$  electrically by connecting  $R_1, R_2, C_2$  to an auxiliary voltage  $-V$  (fig.2) (The collector supply remains  $-V_{cc}$ ). Then,

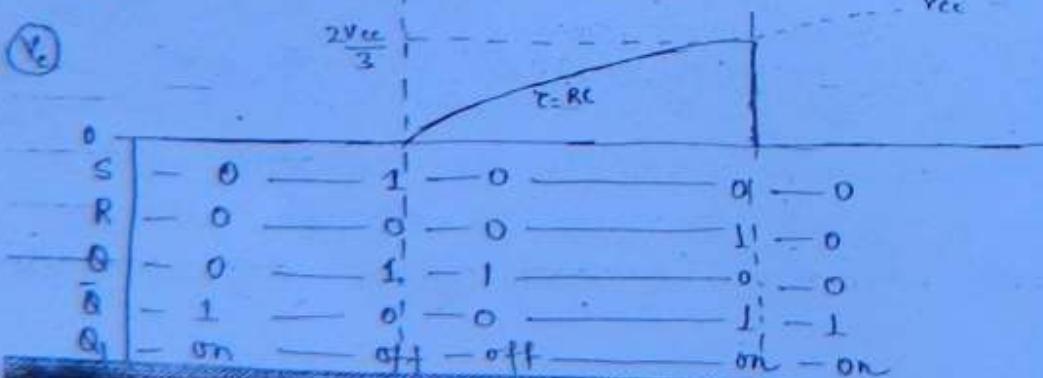
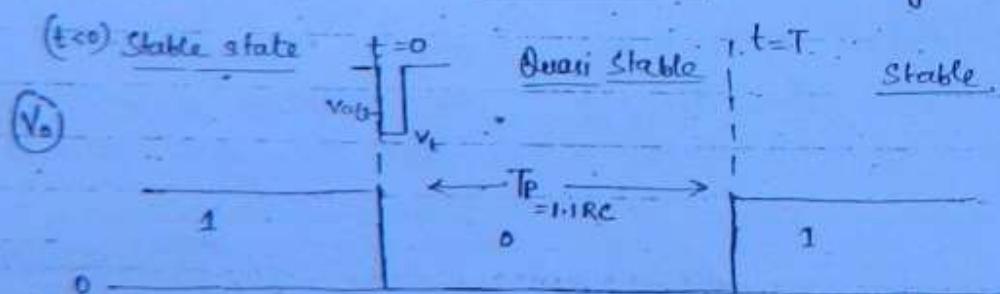
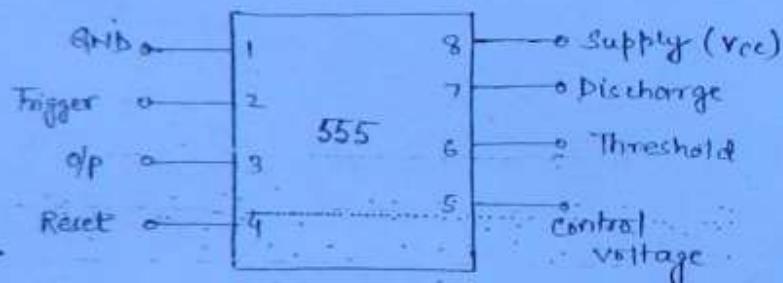
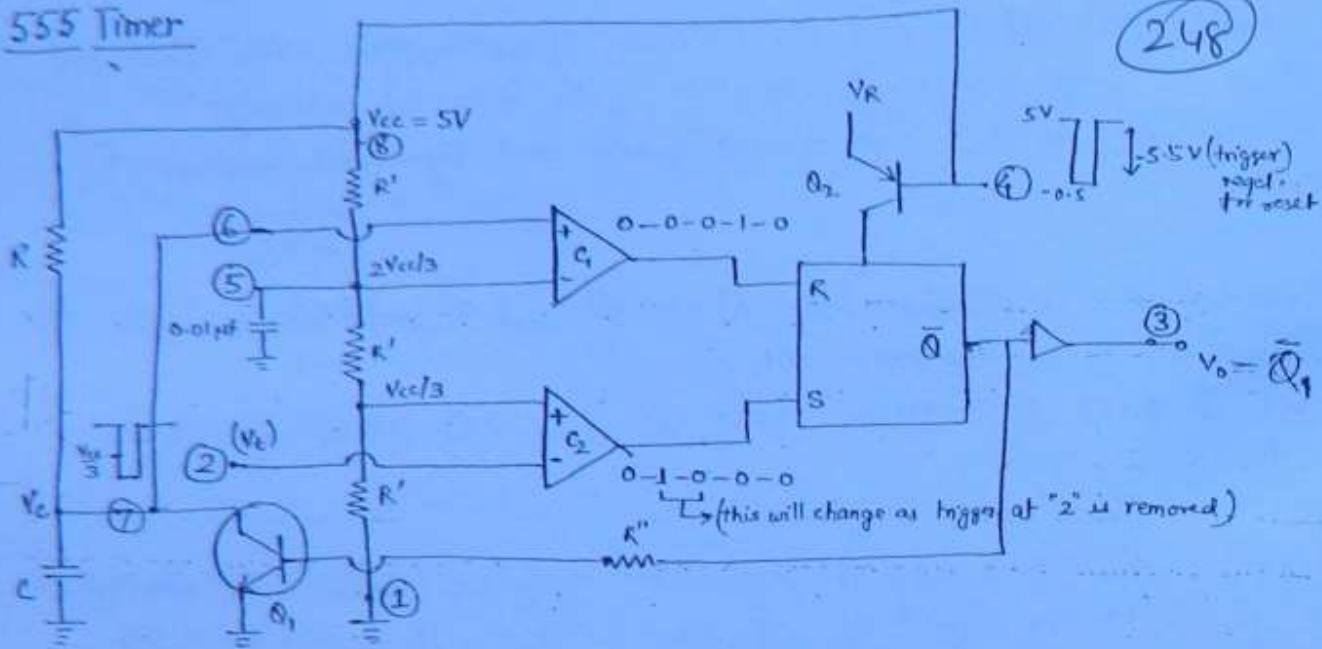
$$T = 2RC \ln \left( 1 + \frac{V_{cc}}{V} \right).$$

Such a ckt (fig.2) is voltage to frequency converter.

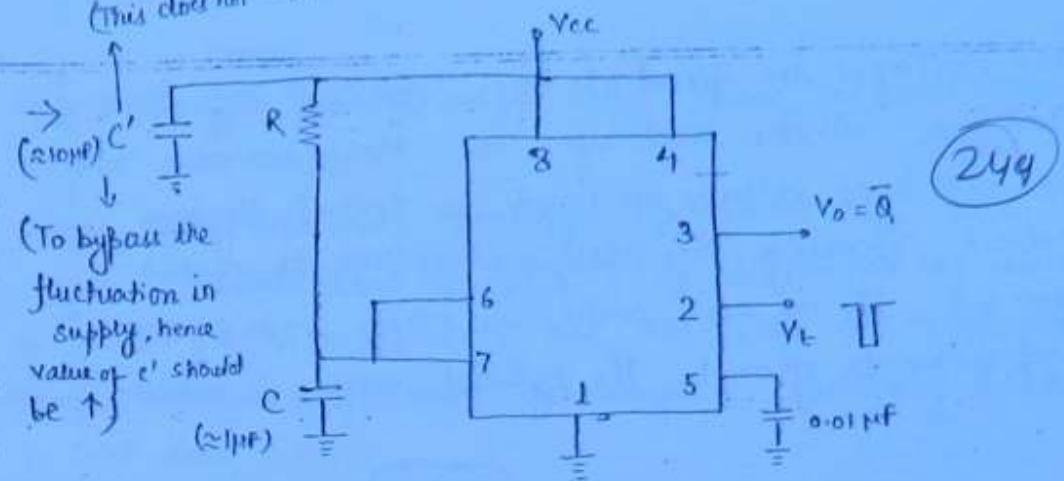
- Note
- If each resistor  $R$  ( $R_1, R_2$ ) is replaced by a transistor which acts as a constant current source for charging  $C$  then excellent linearity b/w freq & voltage may be obtained.

## 555 Timer

(248)



(This does not effect the timing)



### 555 in Monostable Mode

Derivation of  $T_p$  -

$\rightarrow V_C \rightarrow$  change from 0 to  $\frac{2V_{cc}}{3}$

$$\rightarrow V_C = V_f - (V_f - V_i)e^{-t/Rc} ; \quad V_f = V_{cc}, \quad V_i = 0, \quad R = RC$$

$$\therefore V_C = V_{cc} (1 - e^{-t/RC})$$

$$\text{At } t = T_p - \frac{2V_{cc}}{3} = V_{cc} (1 - e^{-T_p/RC})$$

$$\Rightarrow T_p = RC \ln 3$$

$$\Rightarrow T_p = 1.1 RC$$

$\rightarrow$  The device 555 is a monolithic timing ckt that can produce accurate & highly stable time delays or oscillations.

Constructional Details :-

The device consists of two comparators ( $C_1$  &  $C_2$ ) that drive set & reset terminals of a flip flop which in turn controls on & off cycles of discharge tr  $Q_1$ .

- Comparator reference voltages are fixed at  $\frac{2V_{cc}}{3}$  for  $C_1$  &  $\frac{V_{cc}}{3}$  for  $C_2$ , by means of a voltage divider made up of three series resistors  $R^*$ . These reference voltages are reqd. to control timing.
- Timing can be controlled externally by applying voltage to control voltage terminal (pin 5). If no such control is reqd., pin 5 can be bypassed by a capacitor to ground. The typical value is about 0.01μF.

(250)

### function:-

- When the voltage is applied at trigger terminal goes grows +ve & passes through reference level  $\frac{V_{cc}}{3}$ , the o/p of  $C_2$  changes its state. This change of state ( $S=1, R=0$ ) will set the flip flop with  $\bar{Q}=0$  & Tr.  $Q_1 = \text{off}$ .
- When voltage applied at threshold terminal (pin 6) grows +ve & passes through  $\frac{2V_{cc}}{3}$ , o/p of  $C_1$  changes its state ( $S=0, R=1$ ). This change of state will reset the flip flop with  $\bar{Q}=1$  and Tr.  $Q_1 = \text{on}$ .

### PIN-4 (Reset Pin)

- A separate reset terminal is provided which is used to reset the ff externally. Normally when Pin 4 is not used, it should be connected to the supply  $V_{cc}$  to avoid any false triggering. Transistor  $Q_2$  acts as a buffer, isolating the ckt from false reset.

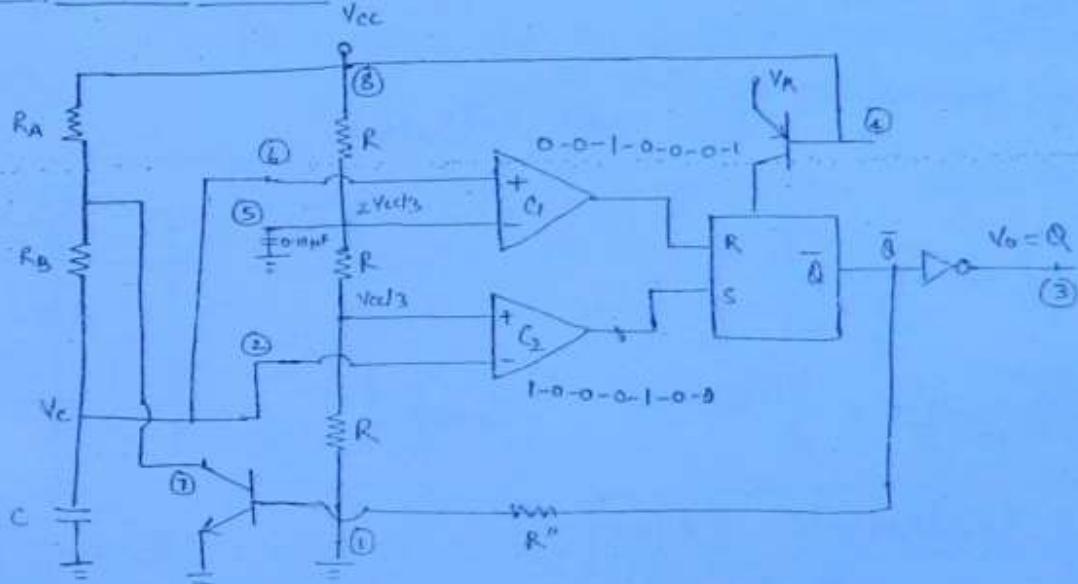
### 555 timer in Monostable state :-

- for  $t < 0$ , ckt is in stable state.  $V_t$  (trigger voltage) =  $V_{cc}$ ,  $V_o = \bar{Q} = 1$  &  $V_c = 0$ , and  $S=R=0$ .
- At  $t=0$ , on application of -ve trigger less than  $\frac{V_{cc}}{3}$  causes o/p of  $C_2$  to be high. This will set ff with  $\bar{Q}=V_o=0$  and  $Q_1 = \text{off}$ .

- Note that after termination of trigger pulse, FF will remain in  $\bar{Q}=0$  state.  
 (since  $S=0, R=0$ ). 257

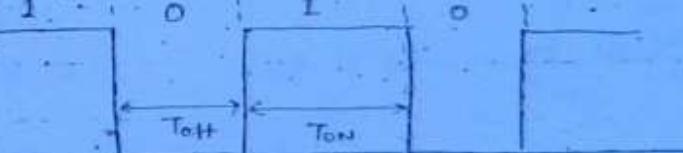
- Now, timing capacitor  $C$  charges up towards  $V_{cc}$  with  $\tau = RC$ . When  $V_c$  reaches threshold level of  $\frac{2V_{cc}}{3}$ ,  $C_1$  will switch its state. This change of state ( $R=1, S=0$ ) resets the FF. with  $\bar{Q}=V_o=1$  and  $Q_1=0$ . Then ~~the~~ saturation resistance of  $Q_1$  discharges  $C$  suddenly & ckt reach to its initial state.

### 555 Timer in Astable Mode :-



$t=0$

(1)

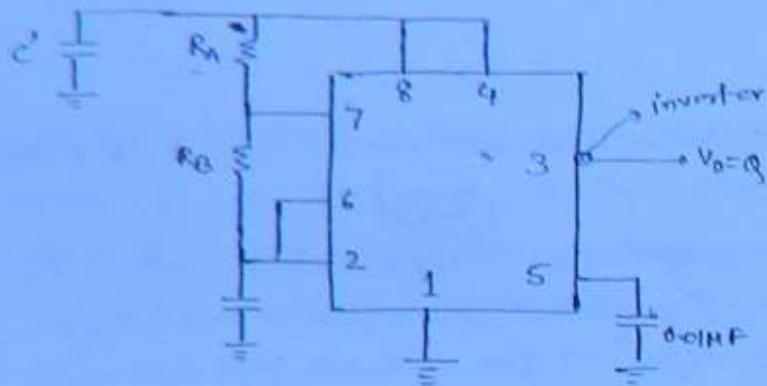


(2)

$V_c$



S	1	0	—	Q	—	0	—	1	—	0	—
R	0	—	1	—	0	—	0	—	1	—	0
Q	1	—	0	—	1	—	0	—	0	—	1
$\bar{Q}$	0	—	1	—	0	—	1	—	1	—	0
$\bar{Q}_1$	off	—	on	—	off	—	on	—	on	—	off



(252)

### Astable Mode

#### Derivation of Ton:

Capacitor charges from  $\frac{V_{cc}}{3}$  to  $\frac{2V_{cc}}{3}$  with  $\tau = (R_A + R_B)C$ .

$$V_f = V_{cc}, \quad V_i = \frac{V_{cc}}{3}$$

$$\therefore V_C = V_{cc} - \left[ V_{cc} - \frac{V_{cc}}{3} \right] e^{-t/\tau}$$

At  $t = t_{on}$

$$\Rightarrow \frac{2V_{cc}}{3} = V_{cc} - \frac{2V_{cc}}{3} \cdot e^{-t_{on}/(R_A + R_B)C}$$

$$\Rightarrow \boxed{t_{on} = 0.69 (R_A + R_B)C = \tau \ln 2}$$

#### Derivation of Toff:

Capacitor discharges from  $\frac{2V_{cc}}{3}$  to  $\frac{V_{cc}}{3}$  with  $\tau' = R_B C$ .

$$V_i = \frac{2V_{cc}}{3}, \quad V_f = 0.$$

At  $t = T_{off}$ -

$$\frac{V_{cc}}{3} = 0 - \left( 0 - \frac{2V_{cc}}{3} \right) e^{-T_{off}/\tau'}$$

$$\Rightarrow \boxed{T_{off} = 0.69 R_B C}$$

$\rightarrow \because [T_{on} > T_{off}] \Rightarrow > 50\% \Rightarrow$  Asymmetrical sq wave.

$$\therefore \tau = T_{on} + T_{off} = 0.69 (R_A + 2R_B) C$$

$$f = \frac{1}{T} = \frac{1.44}{(R_A + 2R_B)C}$$

Duty Cycle

$$D = \frac{R_A + R_B}{R_A + 2R_B} \times 100\%$$

When  $V_O = Q$ ,  $D > 50\%$ .

When  $V_O = \bar{Q}$ ,  $D < 50\%$ .

→ For  $R_A = 0\Omega$ ,

$$D = 50\%$$

- but  $R_A$  cannot be 0 since pin 7 will be directly connected with  $V_{CC}$  and Tr. will burn.

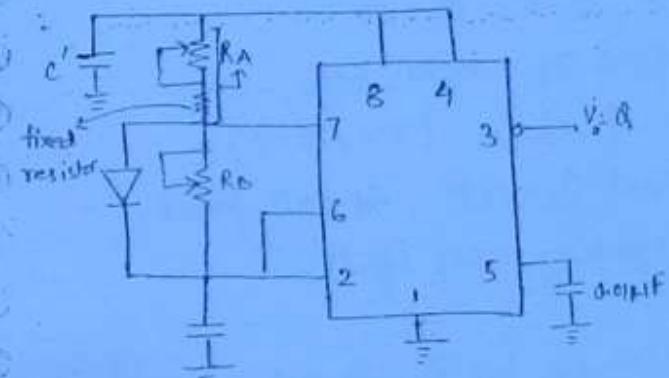


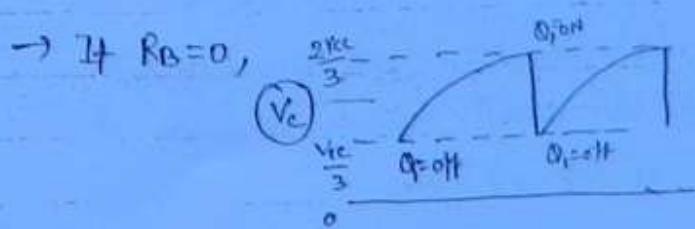
Fig 2.

$$\rightarrow T = 0.69(R_A + R_B)C$$

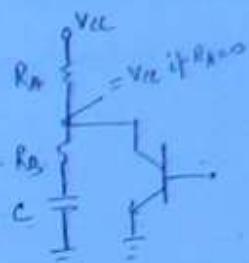
$$f = \frac{1}{T} = \frac{1.44}{(R_A + R_B)C}$$

$$D = \frac{T_{on}}{T} \Rightarrow D = \frac{R_A}{R_A + R_B} \times 100\%$$

Now if  $R_A = R_B$ ,  $D = 50\%$ .



(253)

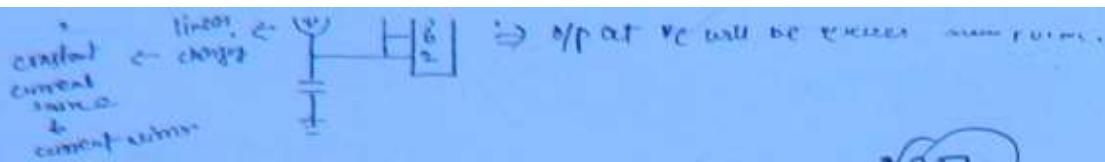


Case I

- ①  $Q_1 = \text{off}$
- C charges from  $\frac{V_{CC}}{3}$  to  $\frac{2V_{CC}}{3}$
- $T = R_A C$
- $T_{on} = 0.69 R_A C$

Case II

- ②  $Q_1 = \text{on}$
- C discharges from  $\frac{2V_{CC}}{3}$  to  $\frac{V_{CC}}{3}$
- $T = R_B C$
- $T_{off} = 0.69 R_B C$



(254)

- Potentiometer is provided in o/p as duty cycle can be adjusted by changing  $R_A \& R_B$ . (also all diodes & resistor are not ideal and  $R_A \& R_B$  can be set accordingly to get <sup>symm.</sup> o/p).
- A fixed resistor is added to prevent pin 7 to directly connected to  $V_{cc}$  even if potentiometer at  $R_A$  is set at 0.

### Important Points:-

For Fig ①

- In this mode, timing capacitor C charges up towards  $V_{cc}$  through  $R_A + R_B$  upto  $\frac{2V_{cc}}{3}$  then  $C_1$  switches its state. This change of state ( $s=0, R=1$ ) reset the FF with  $V_O = Q=0, \bar{Q}=1, Q_1=on$ . Then capacitor C discharges through  $R_B$  &  $Q_1$  upto  $\frac{V_{cc}}{3}$ . then  $C_2$  switches its state. This change of state ( $s=1, R=0$ ) set the FF with  $V_O = Q_0=1, \bar{Q}=0$  and  $Q_1=off$ . At this point capacitor starts to charge again, thus completing the cycle.

- Duty cycle will always be  $<$  or  $>$  50% for fig ①. To achieve 50% duty cycle we should make  $R_A=0$ , however with  $R_A=0$  pin 7 is directly connected to  $+V_{cc}$  and this may damage tr.  $Q_1$  when  $Q_1$  is ON.

→ In fig ② -

- Capacitor charges through  $R_A$  and diode D upto  $\frac{2V_{cc}}{3}$  and discharge through  $R_B$  and  $Q_1$  upto  $\frac{V_{cc}}{3}$ . Then cycle repeats.

- To obtain a square wave o/p  $R_A$  must be a combination of a fixed resistor & potentiometer so that potentiometer can be adjusted for exact sq. wave. Fixed resistor will avoid direct connection of pin 7 to  $V_{cc}$  when potentiometer is set at 0.5.

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(255)

### Application

- 1) It is used as Freq. modulator, i.e. voltage to freq. converter.  
2) It is used as missing pulse detector.

directly connecting voltage with charge  
or variation units from  
 $2V_{CE}$  or  $V_{CE}$ .