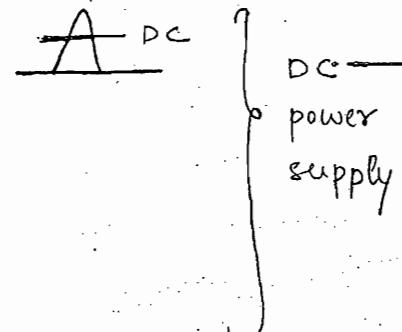
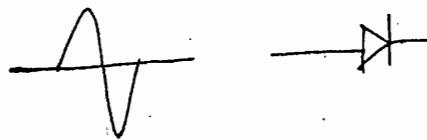


Diodes: Applications:

(1). Rectifiers:



(2). Filter:

(3). Voltage Regulator:

(4). Clippers

(5). Clampers

(6). peak detector

(7). Voltage Multiplier

(8). Diode as a digital logic gate. (AND gate & OR gate)

(9). Diode is a analog gate. (sampling gate)

(10). Diode as a varactor diode.

(11). Zener diode. (Voltage Limiters).

(12). Diode Resistance

(i). Static Resistance

(ii). Dynamic Resistance: small s/g analysis of a diode

(13). Diode capacitance

(i). Transition capacitance (C_T)(ii). Diffusion capacitance (C_D)

BJT

(1). BJT device analysis

(2). BJT Biasing (DC)

(3). Small s/g Amplifiers (voltage amplifiers)

(4). Large s/g Amplifiers (power amplifiers)

(5). feedback theory

Negative feedback theory (Amplifiers)

Low freq. Analysis

High freq. Analysis

Frequency Response

Positive feedback theory (oscillators)

Integrated theory (Op Amps) :

www.gatenotes.in

(1). Multistage Amplifiers

(i) Effect of cascading on Bandwidth.

(ii) Important cascading designs.

(a). Cascode Amplifier (CE-CC)

(OR)

Wide Band amplifier

(b). Darlington pair / high I/p impedance (CC-CC)

(2). Coupling Techniques :

(i) RC coupling

(ii) Direct coupling

(3). Differential Amplifiers

(4). Applications of OP-Amp

FET / MOSFET :

(1) FET device

(2) FET Biasing

(3) FET Amplifiers

} FET

MOSFET

(1) MOSFET device

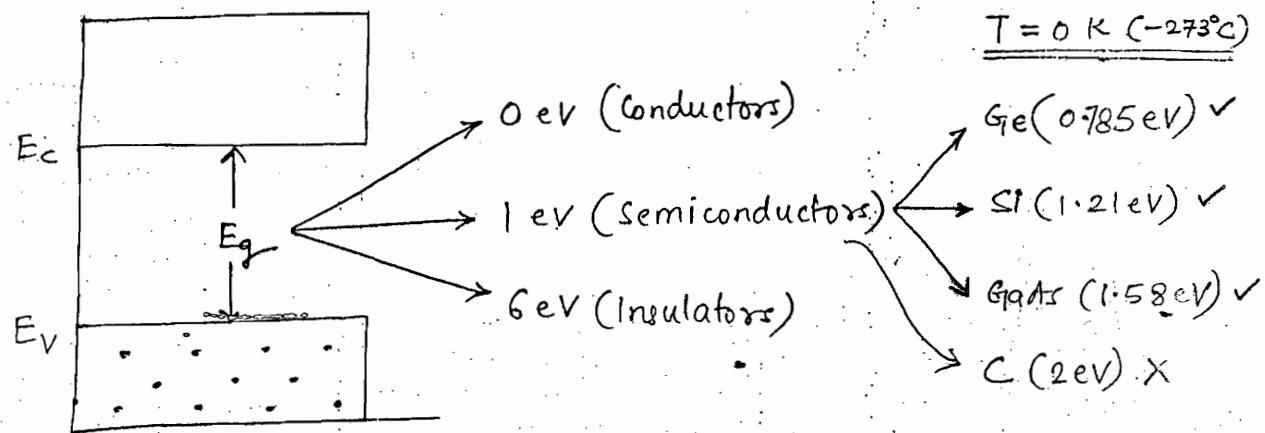
(2) MOSFET Biasing

(3) MOSFET Amplifiers

Introduction to Electronics:

Q. Why 'Si' and 'Ge' are generally preferred compare to GaAs ?

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At Room Temperature 27°C (300K) :

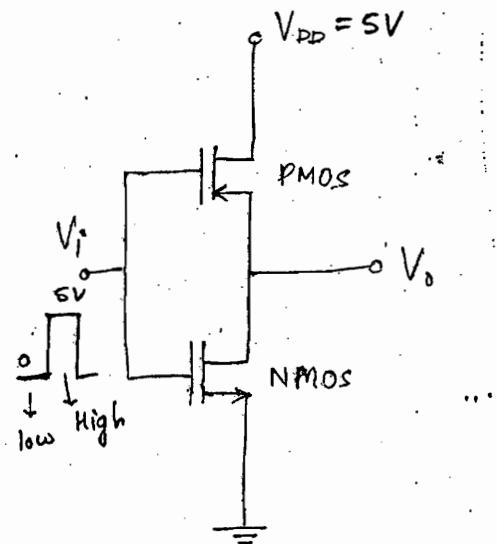
C is a bad semi-conductor.

A good semi-conductor must conduct at room temp..

- She energy gap value of Si and Ge are less compared to GaAs , we expect more conduction in case of Si & Ge.

Q. Why GaAs is used in present CMOS Technology?

CMOS



$$V_i = 0 \text{ (Low)}$$

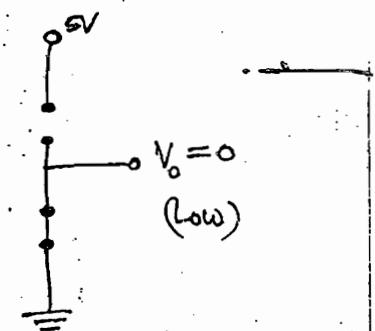
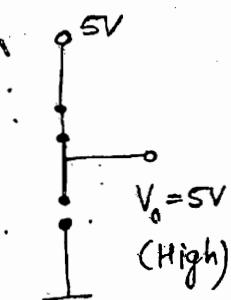
PMOS \rightarrow ON

NMOS \rightarrow OFF

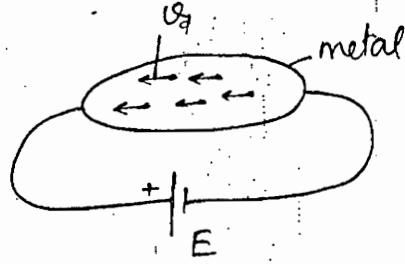
$$V_i = 5V \text{ (High)}$$

PMOS \rightarrow OFF

NMOS \rightarrow ON



Mobility:



$$V_d \propto E$$

$$V_d = \mu E$$

$$\text{Mobility, } \mu = \frac{\text{drift velocity}}{\text{electric field}} \quad (\text{m}^2/\text{v sec})$$

Mobility of electron (μ_e): (Room temp = 27°C)

$$\mu_e (\text{Si}) = 1300 \text{ cm}^2/\text{V.sec}$$

$$\mu_e (\text{Ge}) = 3,800 \text{ cm}^2/\text{V.sec}$$

$$\mu_e (\text{GaAs}) = 8,500 \text{ cm}^2/\text{V.sec} \quad (\text{High switching speed})$$

Temperature with standing capability:

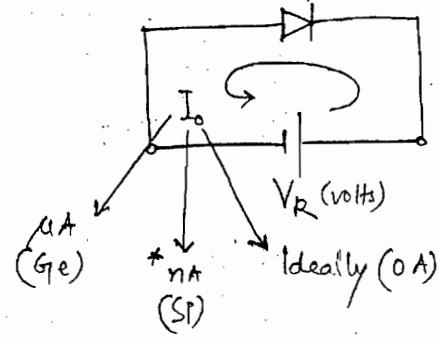
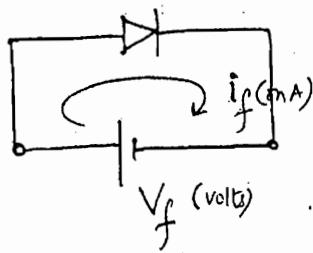
$$T(\text{Ge}) \approx 100^\circ\text{C}$$

$$T(\text{Si}) \approx 200^\circ\text{C}$$

$$* T(\text{GaAs}) \approx 200^\circ\text{C}$$

Q. Why 'Si' is more important than 'Ge'?

(D). $I_o \rightarrow$ Reverse Saturation (or) Leakage current:



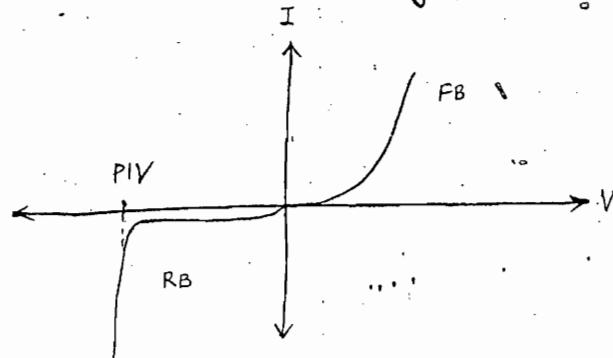
(2). Temperature with standing capability:

Power dissipation (Heating effect)

$$T_{\text{Ge}} \approx 100^\circ\text{C}$$

$$* T_{\text{Si}} \approx 200^\circ\text{C}$$

(3). Peak Inverse Voltage (PIV):



$$\text{PIV (Si)} \approx 1000 \text{ V}$$

$$\text{PIV (Ge)} \approx 400 \text{ V}$$

Q. Give the important properties of GaAs?

(1). At high frequency applications GaAs is used
(mobility of GaAs is more).

$$M_e (\text{GaAs}) \rightarrow 8,500 \text{ cm}^2/\text{V-sec.}$$

$$M_h (\text{GaAs}) \rightarrow 400 \text{ cm}^2/\text{V-sec.}$$

for Si:

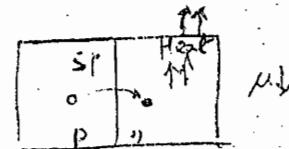
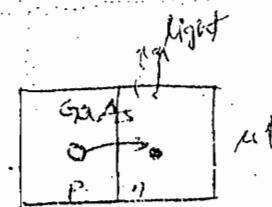
$$M_e (\text{Si}) \rightarrow 1300 \text{ cm}^2/\text{V-sec}$$

$$M_h (\text{Si}) \rightarrow 500 \text{ cm}^2/\text{V-sec}$$

for Ge:

$$M_e (\text{Ge}) \rightarrow 3,800 \text{ cm}^2/\text{V-sec}$$

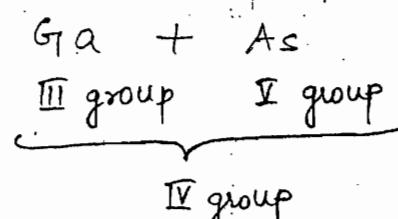
$$M_h (\text{Ge}) \rightarrow 1,800 \text{ cm}^2/\text{V-sec}$$



(3). GaAs is a best example for direct band gap.

Si & Ge are best examples for indirect band gap.

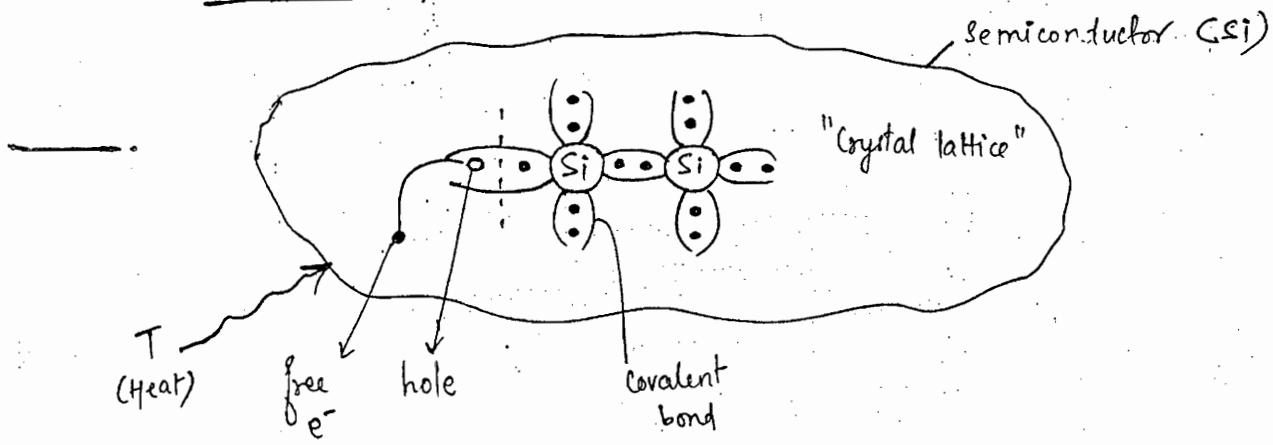
(4). GaAs is a compound semiconductor.



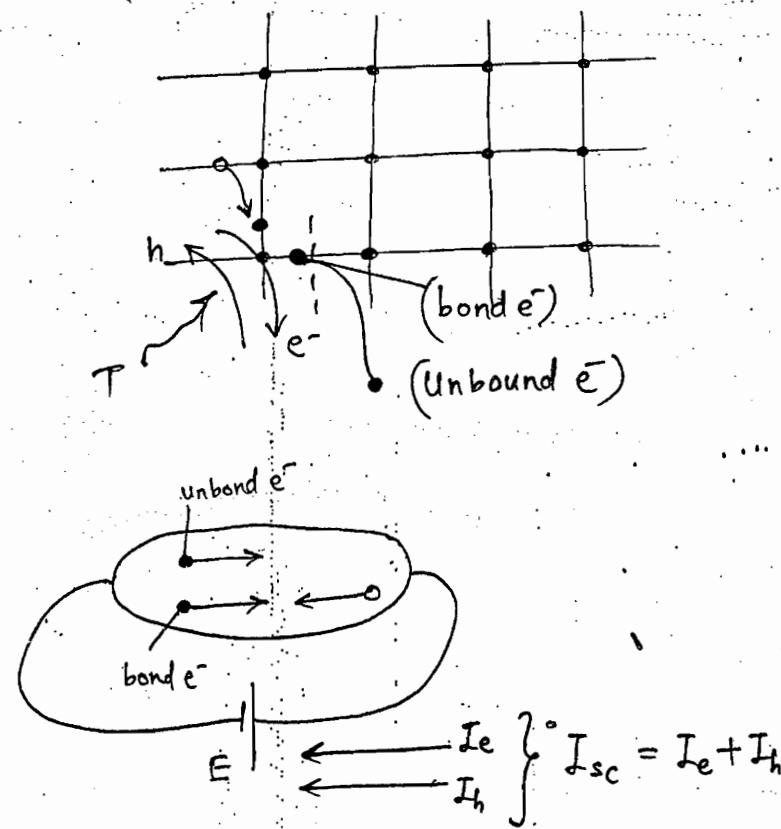
(5). GaAs is used in optics.

Q. Why mobility of e^- > mobility of holes ?

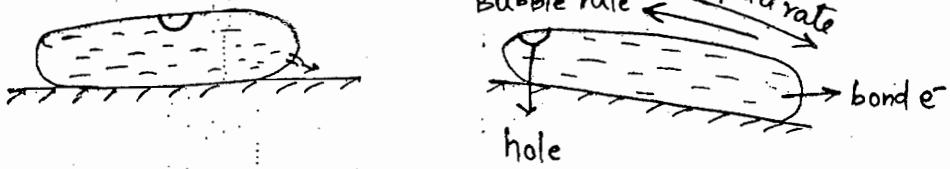
Hole concept:



Crystal Lattice:

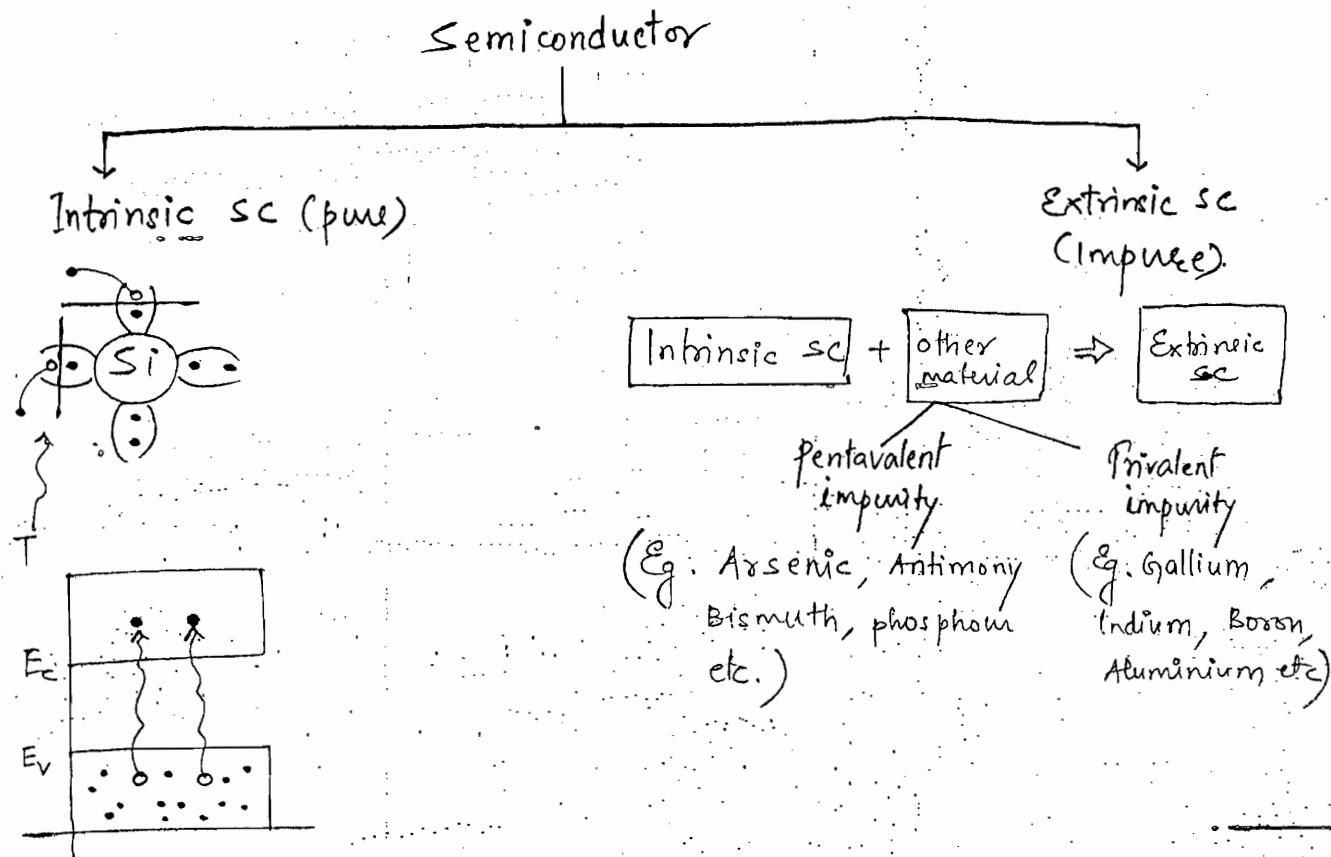


Eg: Glass-Bubble \rightarrow Liquid rate is not easy to analyse, so bubble rate (hole) is easy to analyse. the liquid rate (bond e^-).



The mobility of an unbond e^- (free e^-) is always greater than bond e^- (valence electrons). 7

Q. Give the classification of semiconductor :



$n \rightarrow$ concentration of e^- / cm^3

$p \rightarrow$ conc. of holes $/ \text{cm}^3$

$n_i \rightarrow$ Intrinsic conc. ($e-h$) $/ \text{cm}^3$.

$$n_i = p = n$$

$$n_i \propto \text{temp}$$

Si:

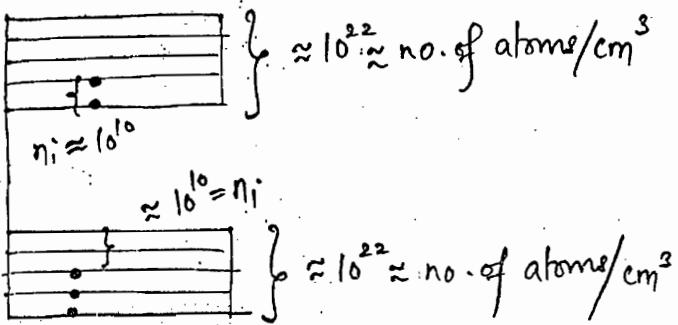
$$n_i(300\text{ K}) \rightarrow 1.5 \times 10^{10} / \text{cm}^3$$

$$\text{No. of atoms} / \text{cm}^3 \rightarrow 5.0 \times 10^{22} / \text{cm}^3$$

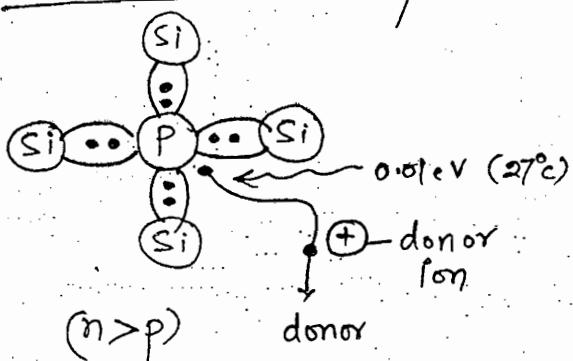
Ge:

$$n_i(300\text{ K}) \rightarrow 2.5 \times 10^{13} / \text{cm}^3. (\text{Eg} \downarrow)$$

$$\text{No. of atoms} / \text{cm}^3 \rightarrow 4.4 \times 10^{22} / \text{cm}^3 (\text{size of Ge} \uparrow)$$



Pentavalent Impurity:

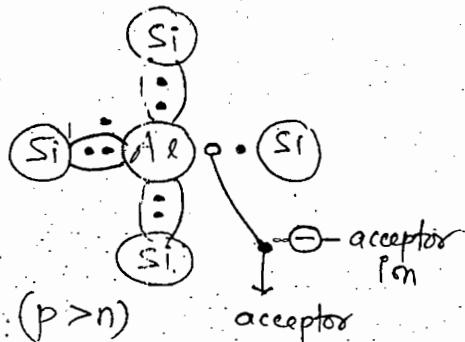


Majority carriers \rightarrow Electrons

Minority carriers \rightarrow holes

N-type Semiconductor

Triivalent Impurity:



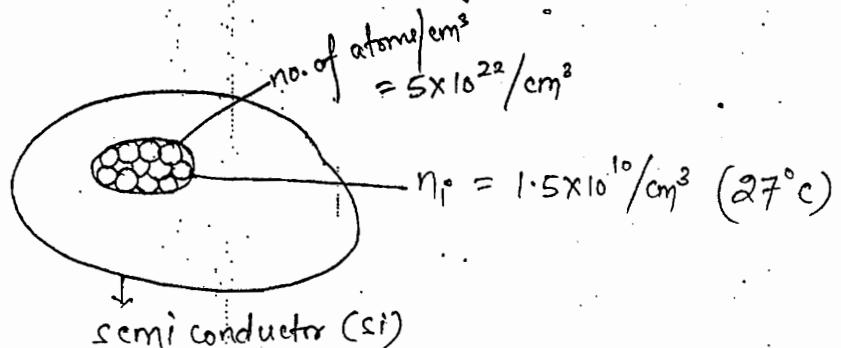
Majority carriers \rightarrow Holes

Minority carriers \rightarrow Electrons

P-type Semiconductor

- * That's why N-type device is always preferred to P-type because electron mobility is higher than hole mobility.

Q. Explain about doping concept in Semiconductors?



Doping : $N_D \rightarrow \text{Donor atoms}/\text{cm}^3$

(1) Ordinary p-n diodes:

10^8 sc atoms \rightarrow 1 impurity [1 'p' (pentavalent)]

5×10^{22} sc atoms $\rightarrow N_D$

$$N_D = \frac{5 \times 10^{22}}{10^8} = 5 \times 10^{14}/\text{cm}^3$$

$$N_D > N_i$$

(2) Zener Diode:

10^6 s.c. atoms \rightarrow 1 impurity {1 'p'}

5×10^{22} s.c. atoms $\rightarrow N_D$

$$N_D = \frac{5 \times 10^{22}}{10^6} = 5 \times 10^{16}/\text{cm}^3$$

$$N_D \gg N_i$$

(3) Tunnel diode:

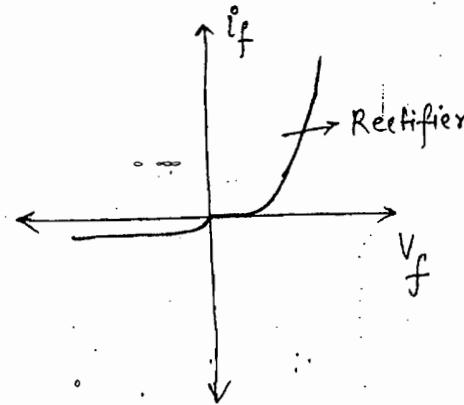
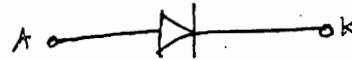
10^3 s.c. atoms \rightarrow 1 impurity

5×10^{22} s.c. atoms $\rightarrow N_D$

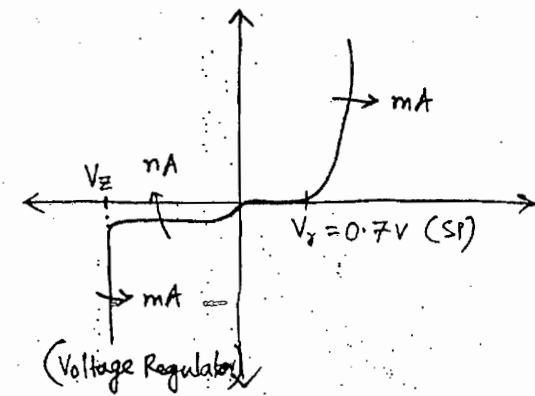
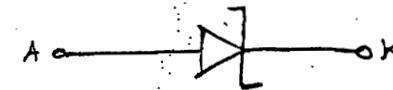
$$N_D = \frac{5 \times 10^{22}}{10^3} = 5 \times 10^{19}/\text{cm}^3$$

$$N_D \ggg N_i$$

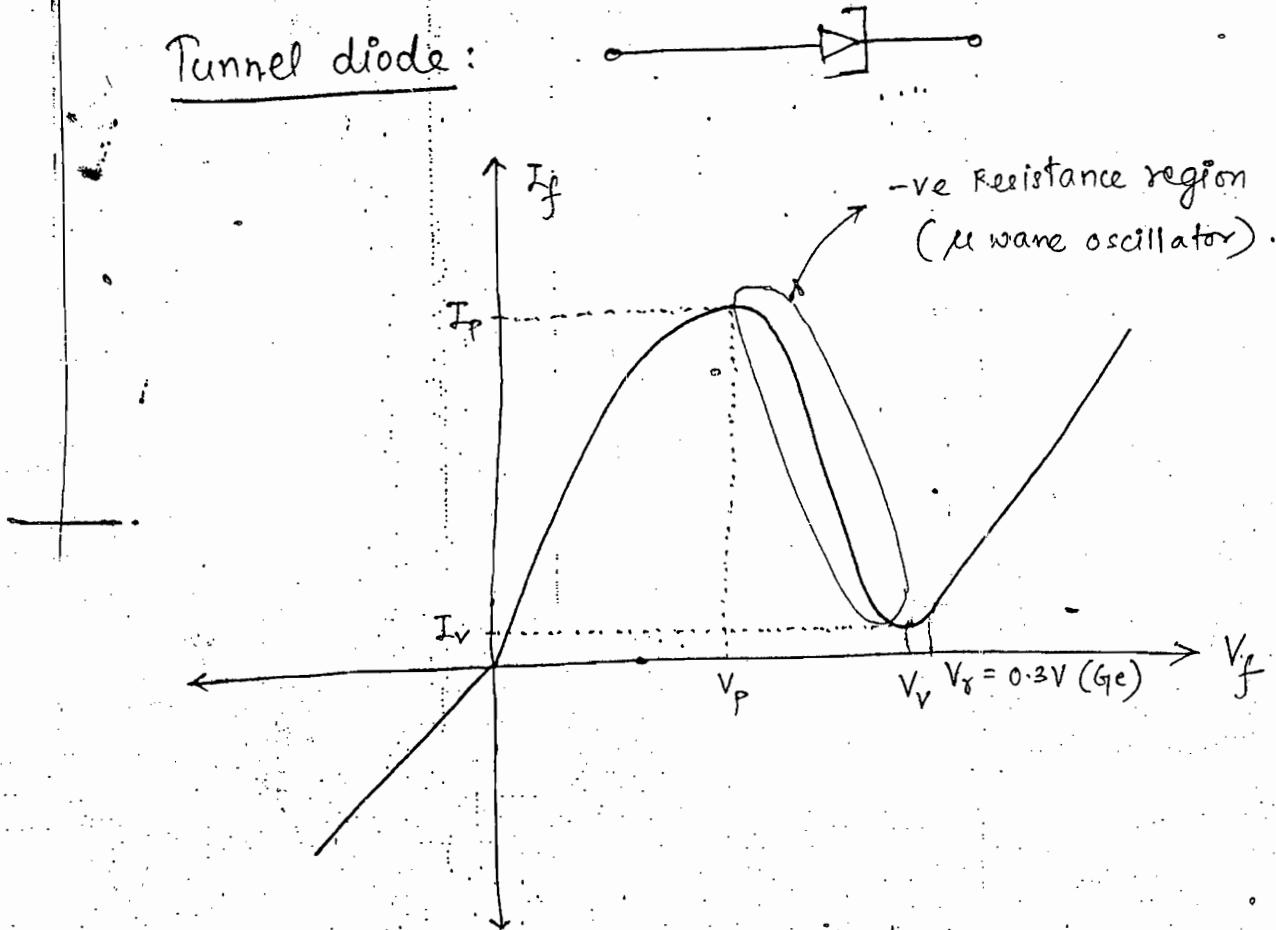
Ordinary p-n diode:



Zener Diode:



Tunnel diode:

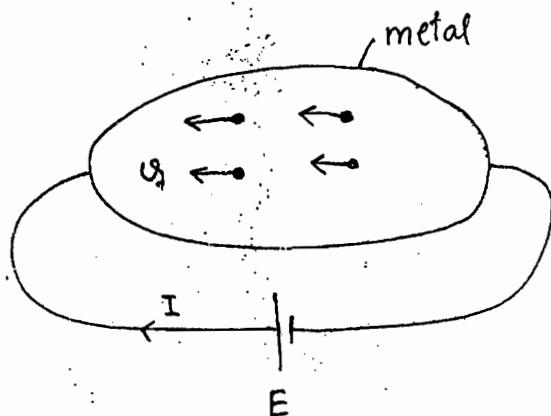


$$R = \frac{dv}{di} = \frac{V_r - V_p}{I_r - I_p} = -\text{ve}$$

Q. Explain about drift current in Semiconductors?

Drift current

Drift \rightarrow movement
(Greek word)



$$v_d \propto E$$

$$v_d = \mu E$$

$$\mu = \frac{v_d}{E} \cdot m^2/v\text{-sec}$$

- (i) "The current is produced due to the drifting of free electrons is called as Drift current".

- (2). The current can occur in metals & semi-conductors.
- (3). Drift current mechanism can also be called as "potential gradient".

Current Density: (J)

$$J = \frac{I}{A} \text{ A/m}^2$$

$$I = \frac{Ne}{t}$$

$$I = \frac{NeV_d}{L}$$

$$J = \frac{NeV_d}{AL} \quad (\because n = N/AL)$$

$J = nev_s$

($\because v_s \rightarrow$ charge density)
 $p = ne$

$$J = p v$$

Metals:

$$J = ne\mu E$$

($\because V_d = ME$)

$$J = \sigma E$$

($\because \sigma = ne\mu$)

Semi-conductors:

$$J_{sc} = J_n + J_p$$

$$J_n = nq\mu_n E$$

$$J_p = pq\mu_p E$$

$$J_{sc} = \underbrace{(n\mu_n + p\mu_p)}_{\sigma_{s.c.}} \cdot q E$$

$$\therefore \sigma_{s.c.} = (n\mu_n + p\mu_p)q$$



(a) Intrinsic s.c. - ($n = p = n_i$)

$$\sigma_{\text{intrinsic}} = n_i (\mu_n + \mu_p) q$$

(a) Extrinsic s.c. - ($n \neq p$)

(i) N-type : ($n > p$)

$$\sigma_{\text{N-type}} = n q / \mu_n$$

$$\approx N_D q / \mu_n$$

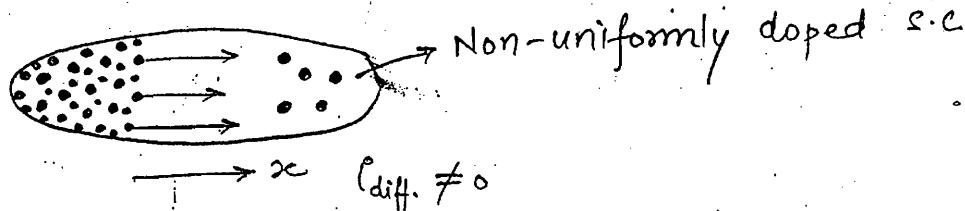
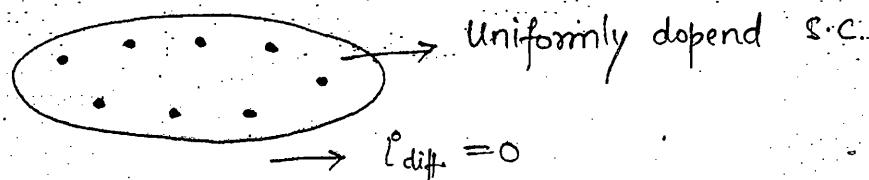
(ii) P-type : ($p > n$)

$$\sigma_{\text{P-type}} = p q / \mu_p$$

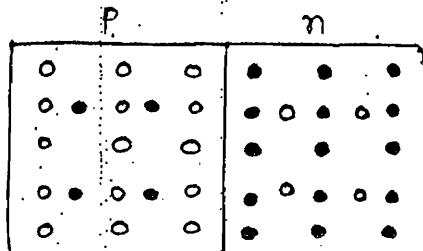
$$\approx N_A q / \mu_p$$

Q. Explain about Diffusion current in semi-conductors?

Diffusion current:



e.g.



: pn jxn is a best example for non-uniformly doped s.c.

$$J_n \propto q \frac{dn}{dx}$$

$$J_n = D_n q \frac{dn}{dx} \quad (D_n \rightarrow \text{diffusion constant for } e^-)$$

$$I_n = Aq D_n \frac{dn}{dx}$$

$$J_p \propto q \frac{dp}{dx}$$

$$J_p = -D_p q \frac{dp}{dx} \quad (D_p \rightarrow \text{diffusion constant for holes})$$

$$I_p = -Aq D_p \frac{dp}{dx}$$

Diffusion current \rightarrow The rate of change of concentration w.r.t. distance x is called as diffusion current.

Diffusion current mechanism can also be called as "concentration gradient".

Analysis:

According to kinetic gas theory -

$$D \propto \mu \Rightarrow \frac{D}{\mu} = V_T = \text{constant}$$

$$D = \mu V_T$$

V_T \rightarrow Volt equivalent temp. (or) thermal voltage

$$V_T = \frac{kT}{qV} \quad \text{where, } k \rightarrow \text{Boltzmann constant}$$

$T \rightarrow \text{Temperature}$

In semi-conductors:

$$\frac{D_n}{\mu_n} = \frac{D_p}{\mu_p} = V_T \quad \text{— Einstein Relation}$$

$$V_T = \frac{KT}{qV}$$

$$= \frac{T}{q/k}$$

As $qV = e^- = 1.6 \times 10^{-19} C$

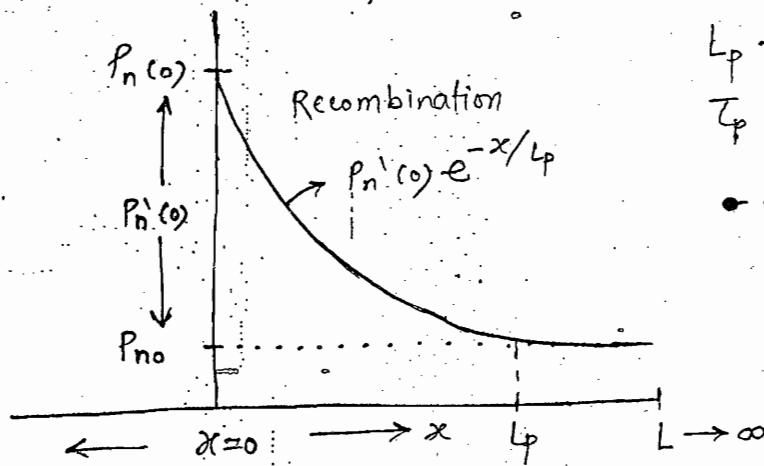
$$K = 1.38 \times 10^{-23} J/K$$

$$V_T = \frac{T(K)}{11,600} \text{ volts}$$

d. Calculate volt equivalent Temp. at room temp.?

A. $V_T = \frac{300}{11,600} = 25.86 \text{ mV} \approx 26 \text{ mV} / (25 \text{ mV})$

Graphical Analysis:



$$L_p \rightarrow \text{Diffusion length}$$

$$\tau_p \rightarrow \text{Carrier life time}$$

• --- $\begin{matrix} \uparrow \\ L \\ \downarrow \\ \tau_p \end{matrix} \rightarrow \circ$

$$L_p^2 = D_p \tau_p$$

Diffusion rate $\propto \frac{1}{\text{Recombination rate}}$

Case - (1)

$$x = 0$$

$$P = P_n'(0) \cdot e^{-0/L_p}$$

$$= P_n'(0) \cdot 1$$

$$= P_n'(0)$$

Case - (2)

$$x = L_p$$

$$P = P_n'(0) \cdot e^{-L_p/L_p}$$

$$= P_n'(0) \cdot \frac{1}{e}$$

$$= \frac{1}{e} \cdot P_n'(0)$$

Case - (3)

$$x = \infty$$

$$P = P_n'(0) \cdot e^{-\infty/L_p}$$

$$= P_n'(0) \cdot \frac{1}{e^\infty}$$

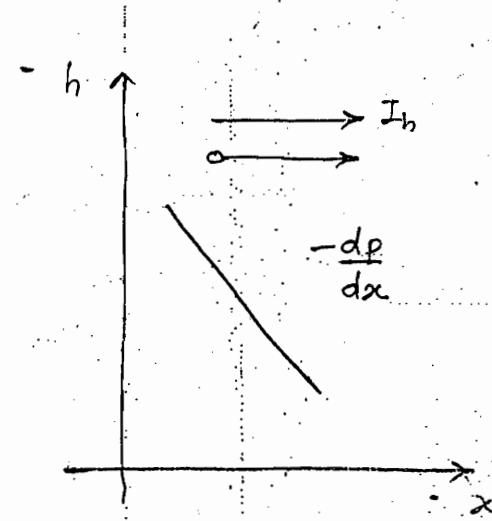
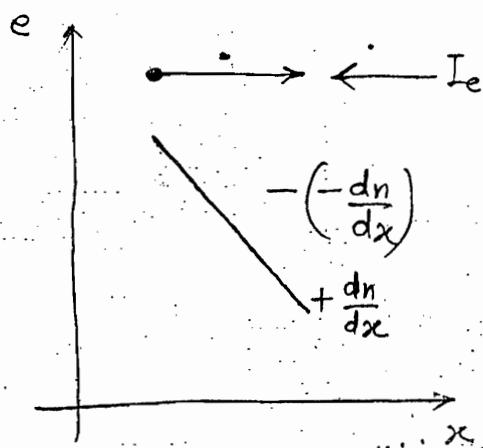
$$= 0$$

(1). When the thickness of material is more , recombination rate will be more , diffusion rate will be less.

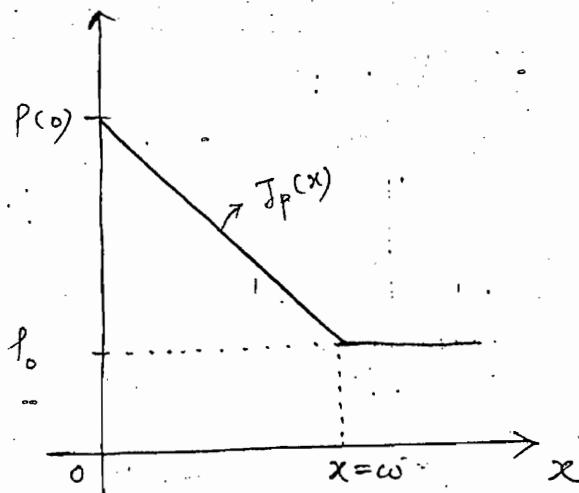
(Non-linear graph)

(2). When the thickness of material is very less , diffusion rate will increase but recombination rate will decrease.

(Graph is linear).

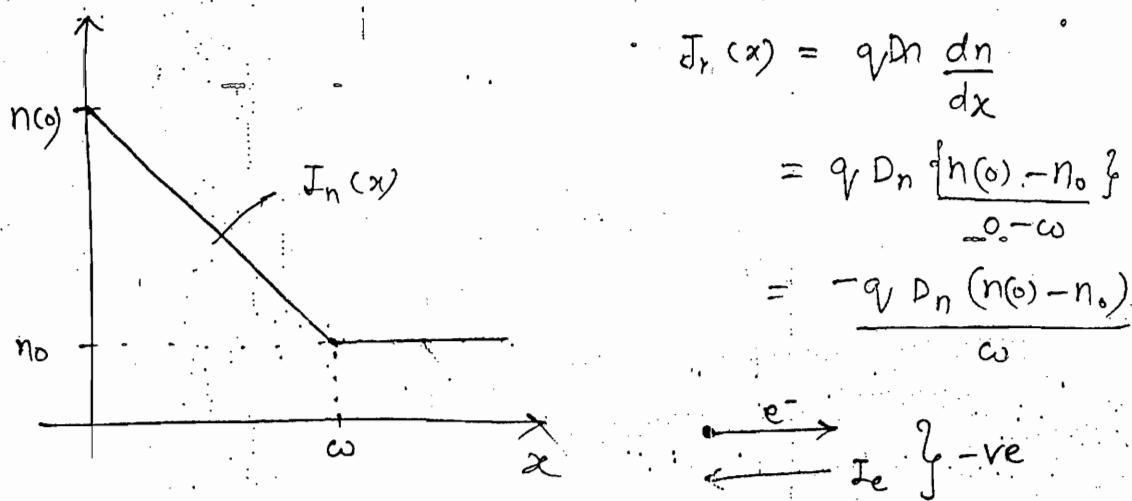


Problems in diffusion current :



$$\begin{aligned} J_p(x) &= -qV D_p \frac{dP(x)}{dx} \\ &= -qV D_p \frac{P(0) - P_0}{\omega} \\ &= \frac{qV D_p [P(0) - P_0]}{\omega} \end{aligned}$$

$\rightarrow I_h$ } +ve



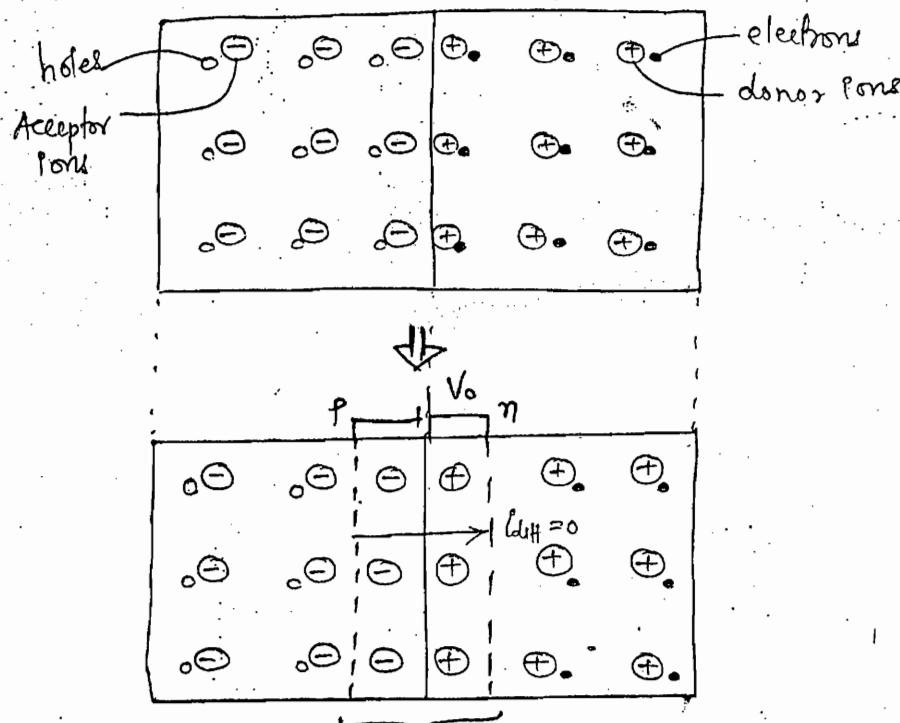
Q. Explain about pn junction theory in 3 conditions :

(1) Open circuit pn junction

(2) Forward Biased

(3) Reversed Biased.

(1) Open circuit pn junction —



Depletion Region

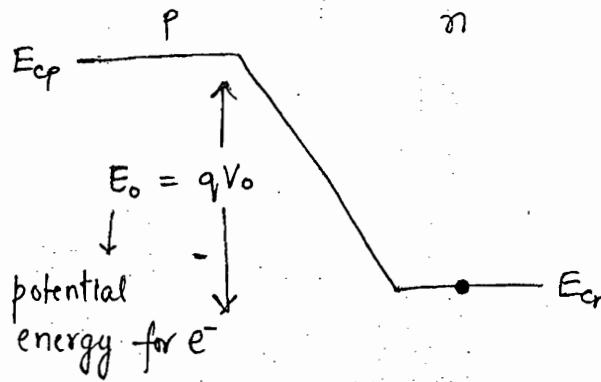
(or) Space charge region (or) Transition region

V_0 (or) V_j (or) V_s \rightarrow potential barrier (or) junction voltage (or)
contact potential (or) cut-in voltage

$$V_o = V_T \ln \left(\frac{N_A N_D}{n_i^2} \right)$$

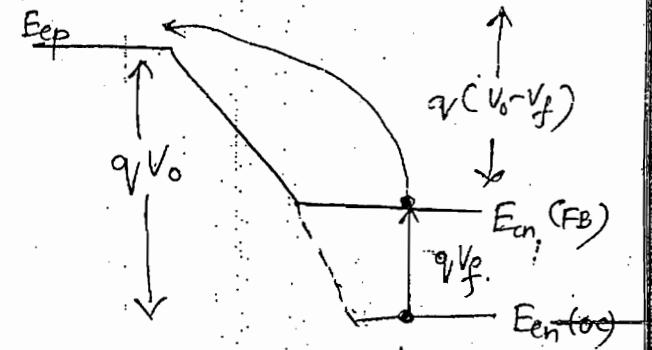
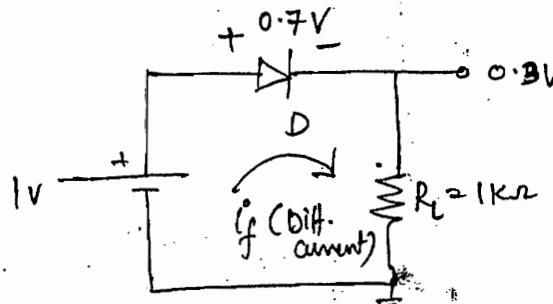
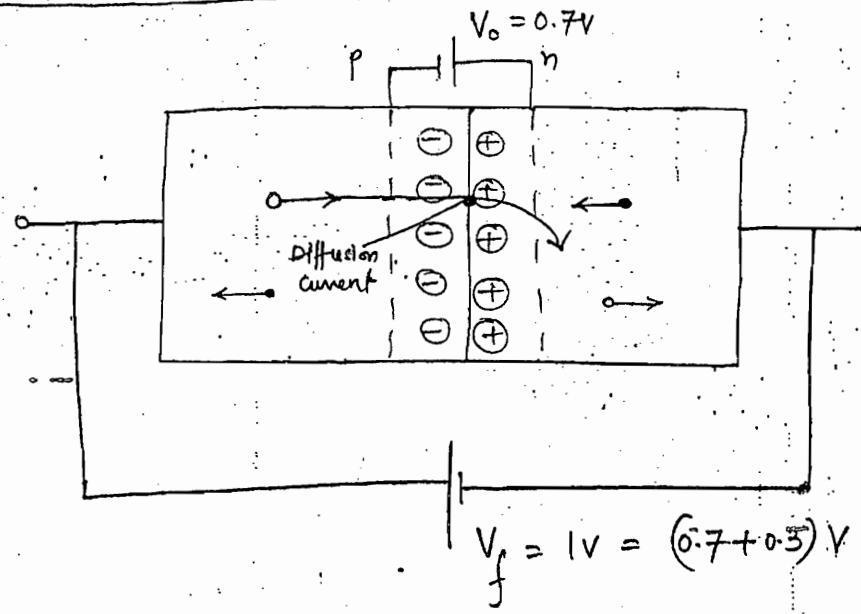
V_o (Si) $\rightarrow 0.5V$ to $0.7V$

V_o (Ge) $\rightarrow 0.1V$ to $0.3V$

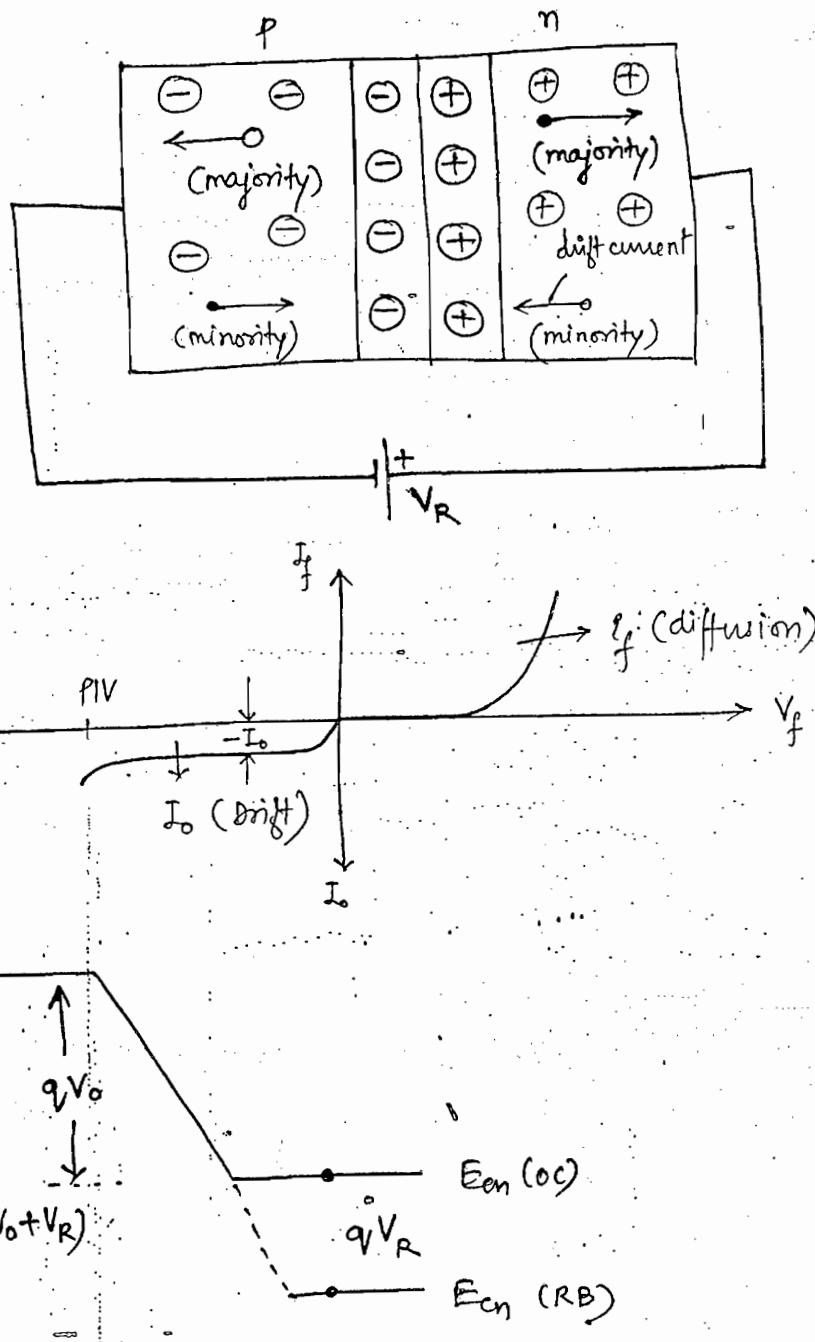


A

(2) forward Bias:



(3). Reverse Bias —

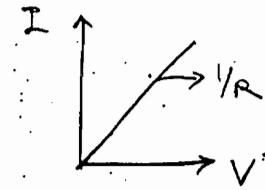
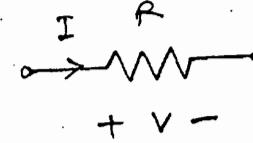


- * Due to appearance of immobile charge ions, resistance offered is more in R.B whereas in F.B resistance offered is less
- * Depletion layer in R.B has more width due to immobile charge ions (\uparrow)

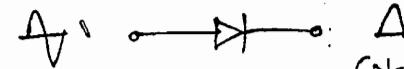
Diode Applications :

Linear Resistor

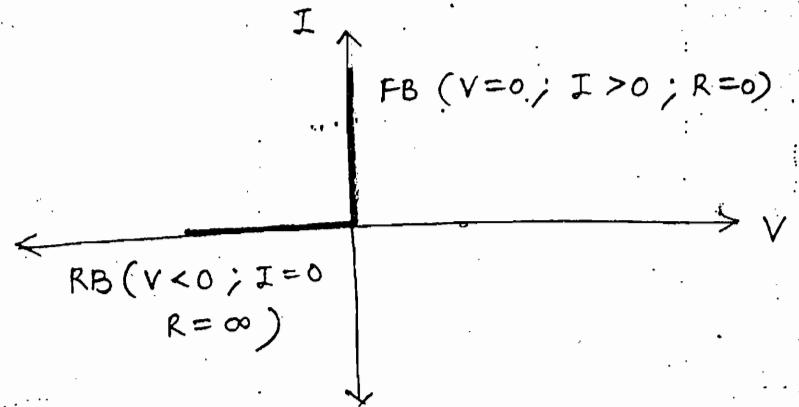
$$V = IR$$



Ideal Diode :

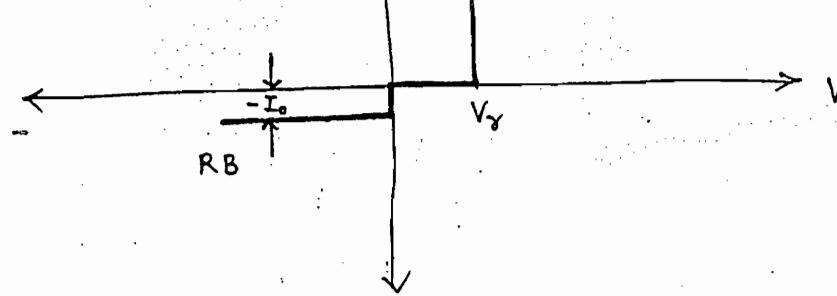


(Non-linear)

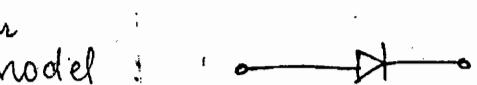


Practical diode :

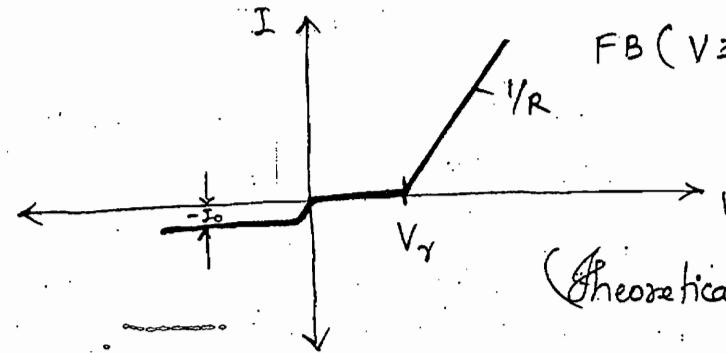
FB ($V \geq V_g ; I > 0 ; R = 0$)



Piece-wise model

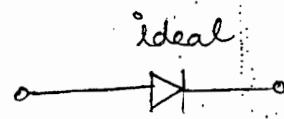
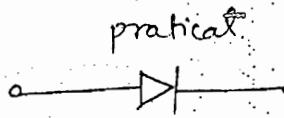
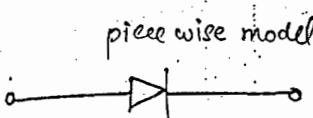


FB ($V \geq V_g ; I > 0 ; R \neq 0$)



(Theoretical concept.)

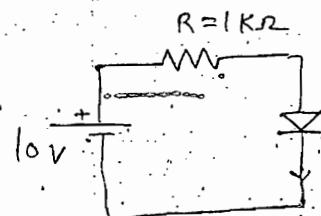
Equivalent models for a diode:

 \Rightarrow s_c  \Rightarrow s_c V_f  \Rightarrow i_f R_p s_c

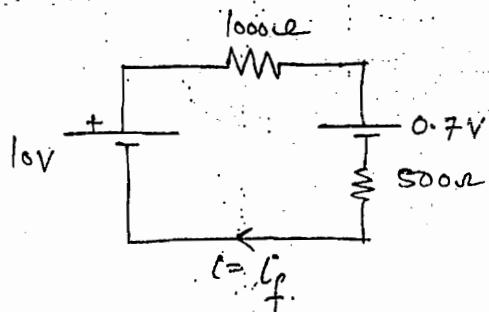
Problems:

Q1. V-I characteristics of pn diode is shown below.

$$i = \begin{cases} \frac{V - 0.7}{500} & V > 0.7 \\ 0 & V < 0.7 \end{cases}$$

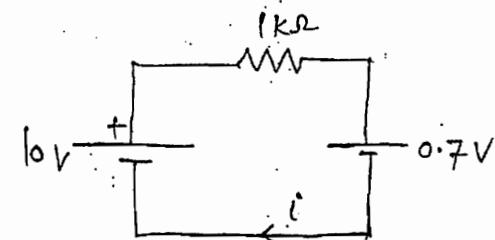


Sol. piece-wise linear model:



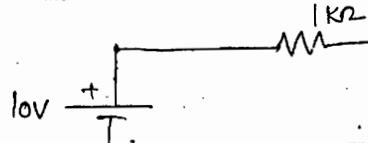
$$i = \frac{10 - 0.7}{1000 + 500} = 6.2 \text{ mA}$$

Practical model:



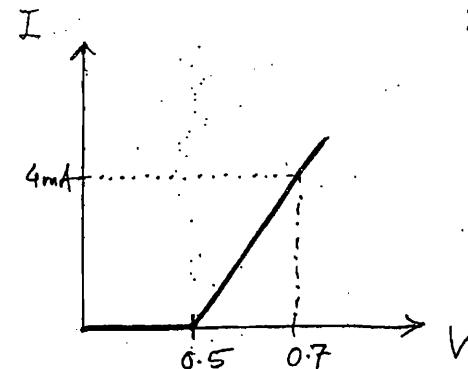
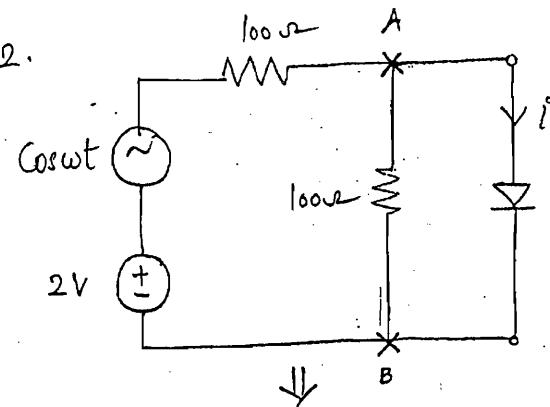
$$i = \frac{10 - 0.7}{1000} = 9.3 \text{ mA}$$

Ideal model:

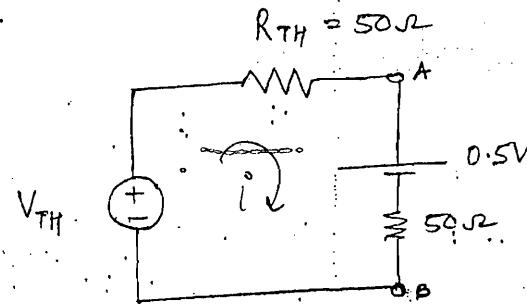


$$i = \frac{10}{1000} = 10 \text{ mA}$$

Q2.



Sol.



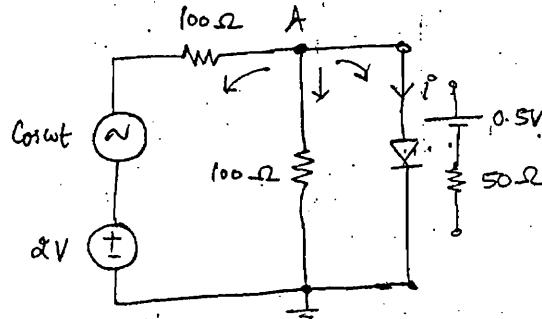
$$V_{TH} = \frac{(\cos \omega t + 2)}{2}$$

$$= 1 + \frac{\cos \omega t}{2}$$

$$i = \frac{\left(1 + \frac{\cos \omega t}{2}\right) - 0.5}{(50 + 50)} = \frac{0.5 + 0.5 \cos \omega t}{100}$$

$$i = \frac{0.5}{100} (1 + \cos \omega t) \text{ A}$$

$$\therefore i = 5(1 + \cos \omega t) \text{ mA}$$

Alter:

$$\frac{V_A - 2 - \cos \omega t}{100} + \frac{V_A}{100} + \frac{V_A - 0.5}{50} = 0$$

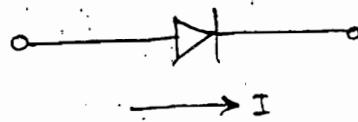
$$4V_A - 2 - \cos \omega t - 1 = 0$$

$$V_A = \frac{3 + \cos \omega t}{4}$$

$$i = \frac{V_A - 0.5}{50} = \frac{\frac{3 + \cos \omega t}{4} - 0.5}{50} = \frac{(1 + \cos \omega t)}{200} \text{ A}$$

$$= 5(1 + \cos \omega t) \text{ mA}$$

Diode current Eqn:



$$I = I_0 (e^{\frac{V}{\eta V_T}} - 1)$$

η = Idealized factor

$$\begin{aligned} \eta &= 1 \text{ (integrated theory)} \\ &= 2 \text{ (discrete theory)} \end{aligned}$$

where, $I \rightarrow$ total current through a diode

$I_0 \rightarrow$ Reverse saturation current
(or) minority current

$V \rightarrow$ applied voltage drop across a diode.

$V \rightarrow +ve$ (FB) $\rightarrow 0.7V_{dd}$

$V \rightarrow -ve$ (RB)

$V_T \rightarrow$ thermal voltage

$$V_T = \frac{kT}{11600}$$

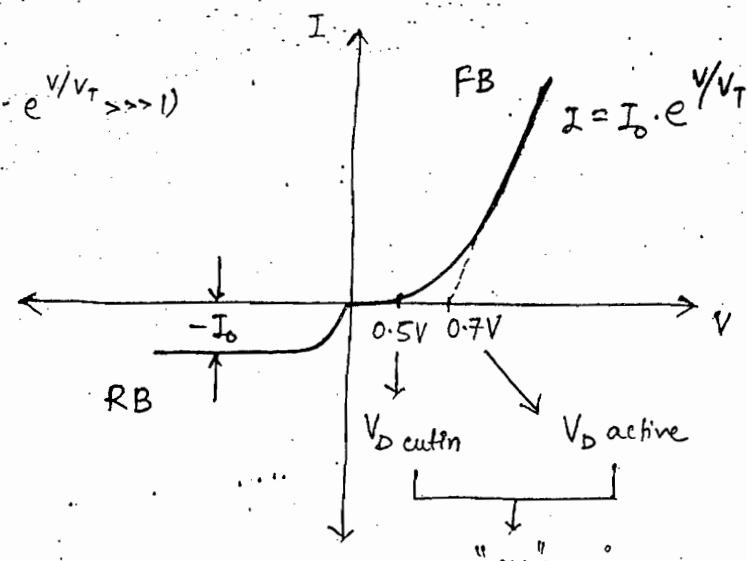
At room temperature, $V_T = 26mV$

Case (1): FB

$$V = +ve (0.7V)$$

$$I = I_0 (e^{\frac{V}{V_T}} - 1) \quad (\because e^{V/V_T} \gg 1)$$

$$I = I_0 \cdot e^{\frac{V}{V_T}}$$



Case (2): RB

$$V = -ve$$

$$I = I_0 (e^{-\frac{V}{\eta V_T}} - 1) \quad (\because e^{-V/\eta V_T} \ll 1)$$

$$I = -I_0$$

Reverse saturation current

Temperature dependence parameters in diode eqn:

$$I = I_0 \cdot (e^{\frac{V}{V_T}} - 1)$$

↓

depends on T ($V_T = \frac{kT}{11600}$)
 depends on T
 (thermal agitation,
 Breakage of covalent
 Bonds)

I_0 vs Temp.:

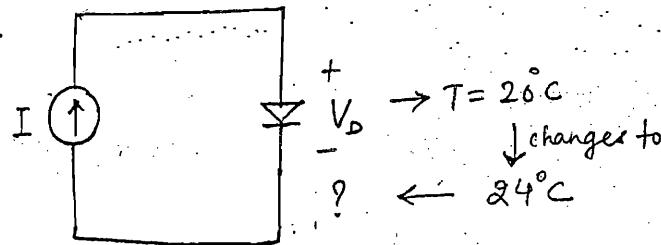
" I_0 increases by 7% per $^{\circ}\text{C}$ rise in temp."

(or)

" I_0 doubles for every 10°C rise in temp."

$$\therefore (1.07)^{10} = 1.967 \approx 2$$

V_D vs Temp.:



$$\frac{dV_D}{dT} = -2.5 \text{ mV}/^{\circ}\text{C} \quad (\text{OR}) \quad -2 \text{ mV}/^{\circ}\text{C}$$

$$\frac{dV_D}{dT} = -2.5 \text{ mV}/^{\circ}\text{C}$$

$$(24 - 20) \times -2.5 \text{ mV} = dV_D$$

$$dV_D = -10 \text{ mV}$$

$$V_D' - V_D = -10 \text{ mV}$$

$$V_D' = (0.7 - 10 \times 10^{-3}) \text{ Volts} = 0.69 \text{ V}$$

$$= (700 - 10) \text{ mV} = 690 \text{ mV}$$

V_T vs temp:

$$\Delta V_T = \frac{I \uparrow}{11600}$$

Q6. $I_0 = 10 \text{ pA} \quad (20^\circ\text{C})$

$I_0 = ? \quad (40^\circ\text{C})$

Exact: $\frac{dI_0}{dT} = 0.07 \text{ pA}/^\circ\text{C}$

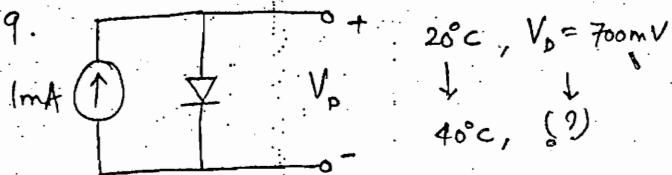
$$\therefore I_0' = I_0 \cdot 1.07 (40 - 20) = (1.07)^{20} I_0 = 3.89 I_0 = 38.9 \text{ pA}$$

Approx: $I_0 \rightarrow$ double for every 10°C rise in temp.

$I_0 \rightarrow 20 \text{ pA} \text{ at } 30^\circ\text{C}$

$I_0' \rightarrow 40 \text{ pA} \text{ at } 40^\circ\text{C}$

Q9.



$$20^\circ\text{C}, V_B = 700 \text{ mV}$$

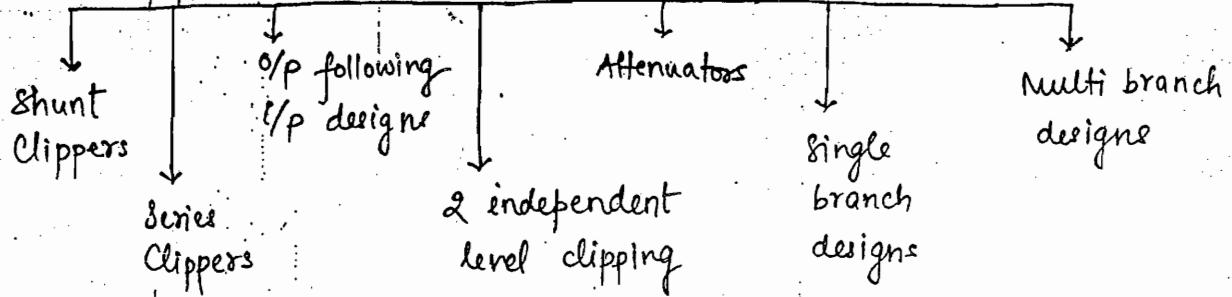
$$40^\circ\text{C}, (?)$$

$$\frac{dV_B}{dT} = -2.5 \text{ mV}/^\circ\text{C}$$

$$dV_B = -2.5 \times (20) = -50 \text{ mV}$$

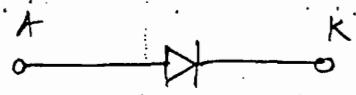
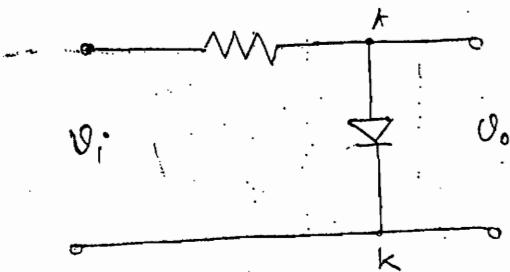
$$\therefore V_B = 650 \text{ mV at } 40^\circ\text{C}$$

Clippers:



Shunt Clippers:

Model 1:



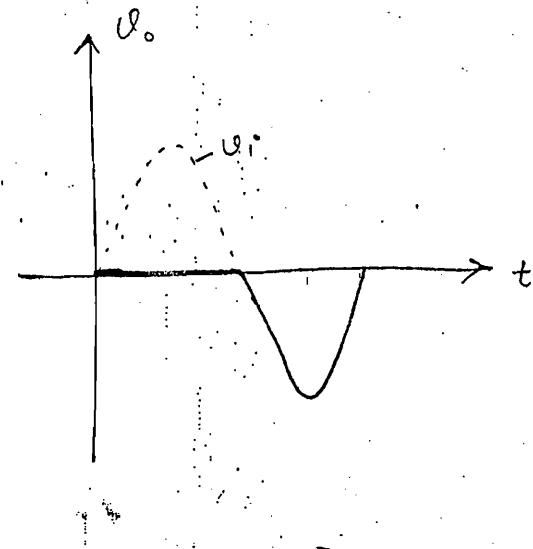
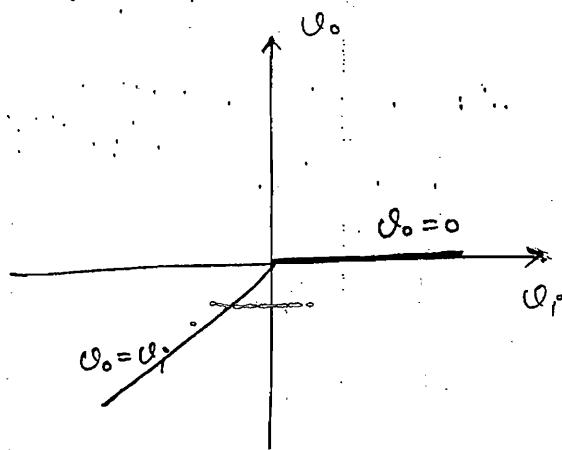
$$V_A > V_K \rightarrow \text{sc}$$

$$V_A < V_K \rightarrow \text{oc}$$

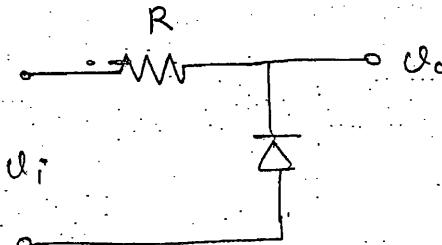
$$V_i < 0 \rightarrow D(\text{off}) \Rightarrow V_o = V_i$$

$$V_i > 0 \rightarrow D(\text{on}) \Rightarrow V_o = 0$$

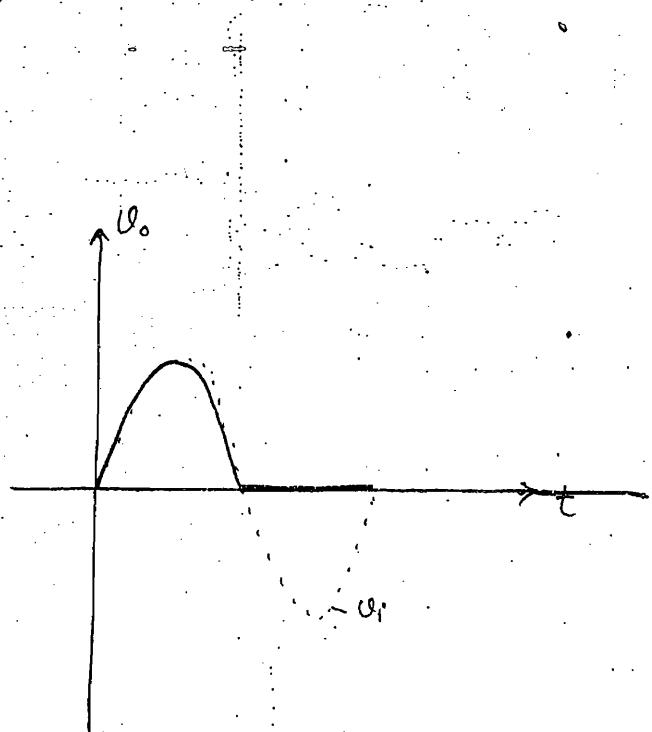
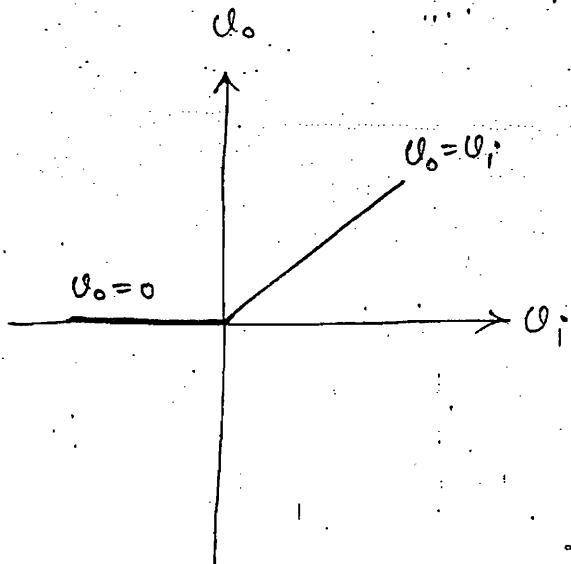
Transfer characteristic :

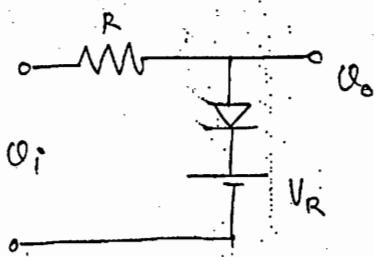


Model - 2 :

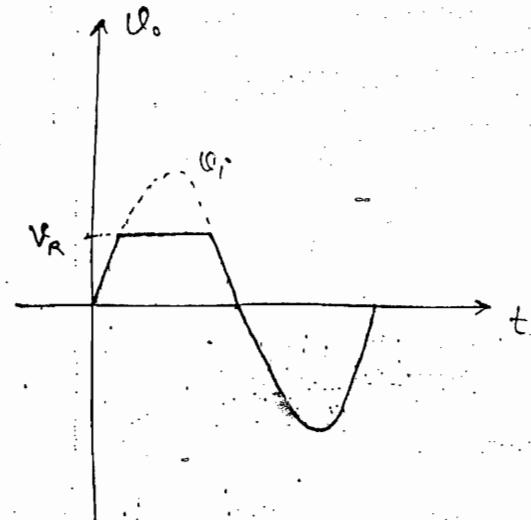
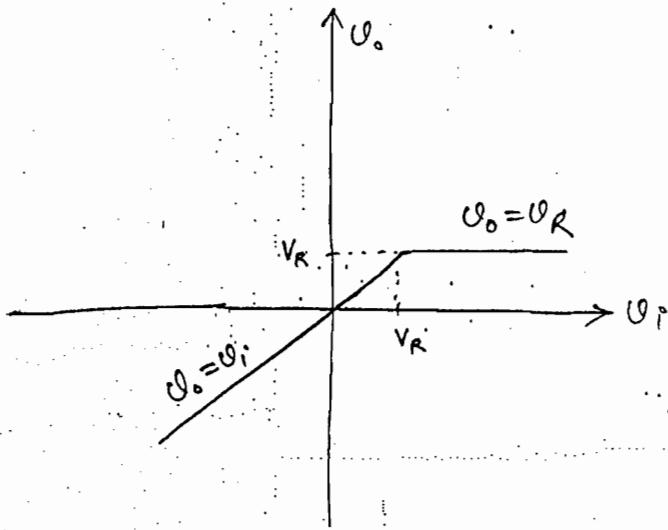
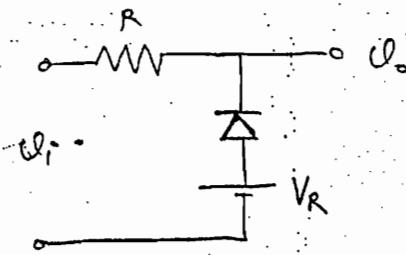


$\delta_i < 0 \quad D(\text{on}) \rightarrow V_o = 0$,
 $\delta_i > 0 \quad D(\text{off}) \rightarrow V_o = V_i$

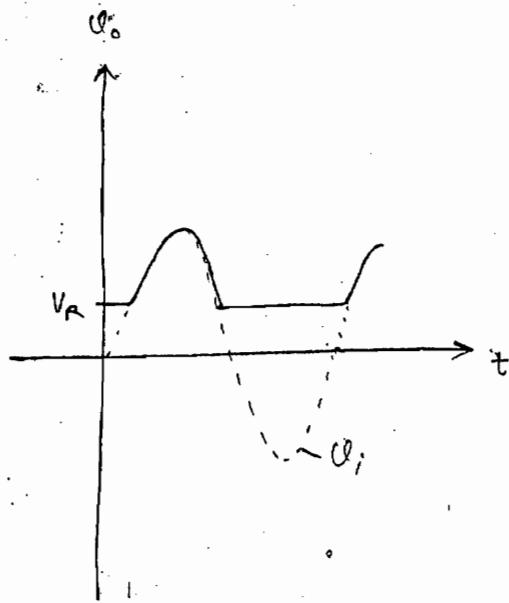
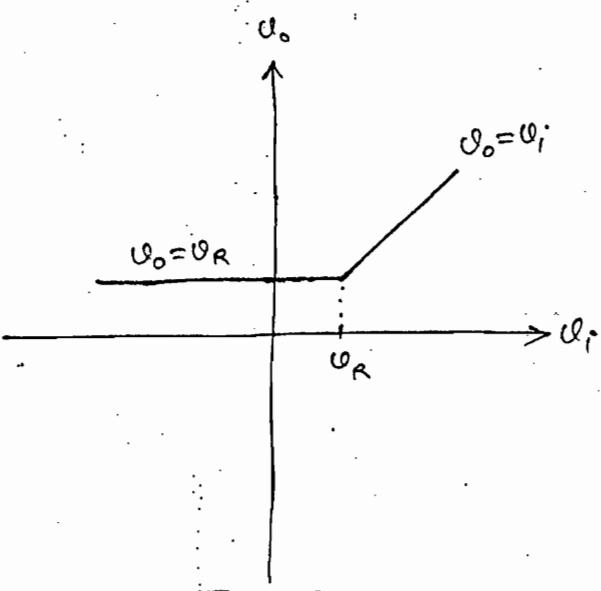


Model - 3 :

$U_i < V_R \rightarrow D(\text{off}) \rightarrow U_o = U_i$
 $U_i > V_R \rightarrow D(\text{on}) \rightarrow U_o = V_R$

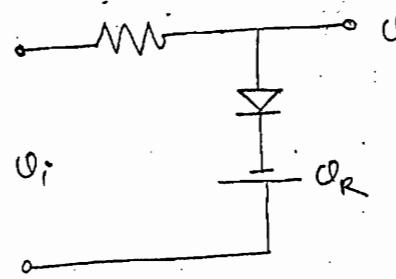
Model - 4 :

$U_i < V_R \rightarrow D(\text{on}) \rightarrow U_o = U_i$
 $U_i > V_R \rightarrow D(\text{off}) \rightarrow U_o = V_R$



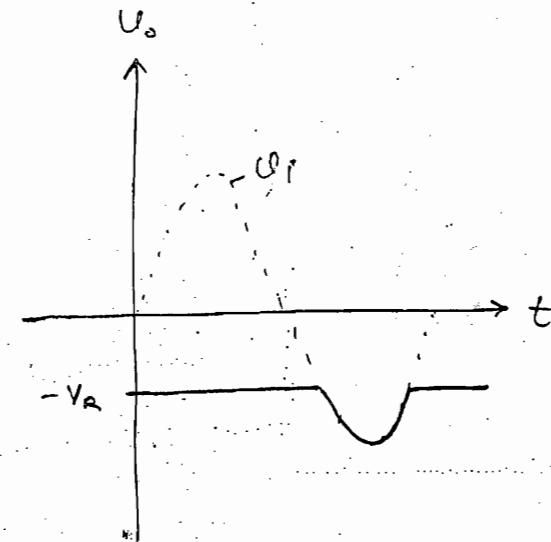
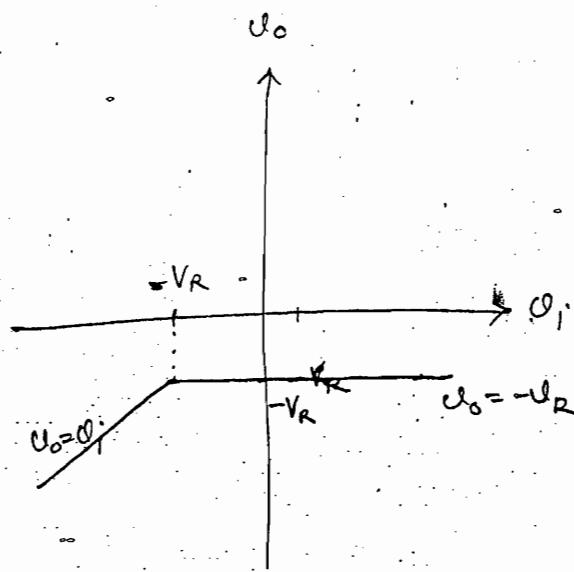
Model - 5 :

27.

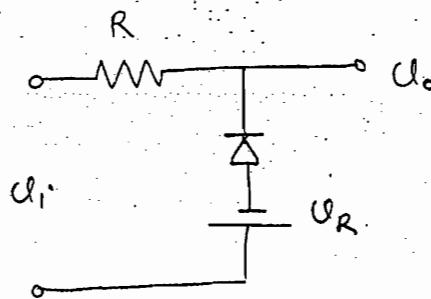


$$v_i < -V_R \rightarrow D(\text{off}) \rightarrow v_o = v_i$$

$$v_i > -V_R \rightarrow D(\text{on}) \rightarrow v_o = -V_R$$

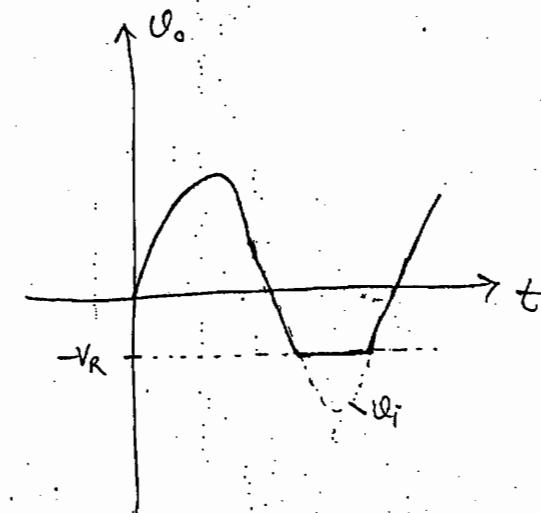
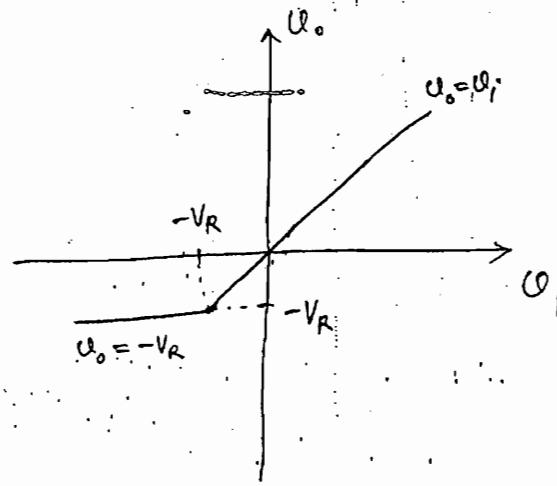


Model - 6 :



$$v_i < -V_R \rightarrow D(\text{on}) \rightarrow v_o = -V_R$$

$$v_i > -V_R \rightarrow D(\text{off}) \rightarrow v_o = v_i$$

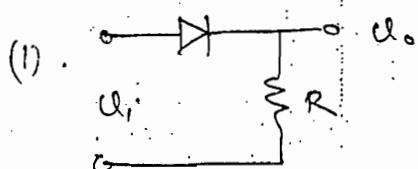


Conclusion:

- (1) When the diode is in downward diode, the sig will be transmitted below the ref. voltage
- (2) When the diode is in upward diode, the sig will be transmitted above the ref. voltage

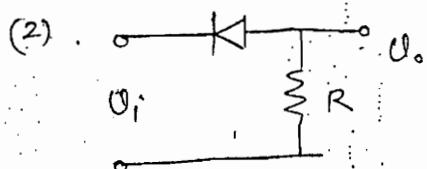
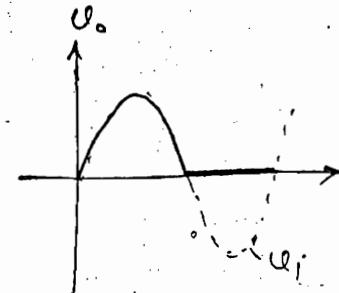
Series Clippers:

Models :



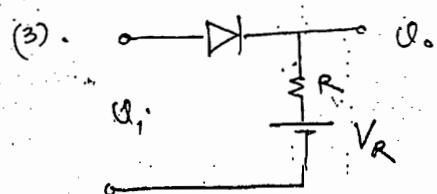
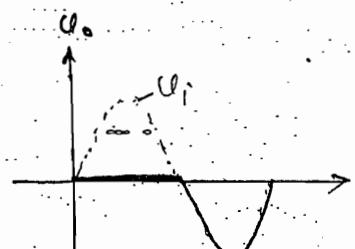
$$U_i < 0 \rightarrow D(\text{off}) \Rightarrow U_o = 0$$

$$U_i > 0 \rightarrow D(\text{on}) \Rightarrow U_o = U_i$$



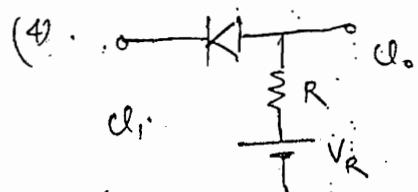
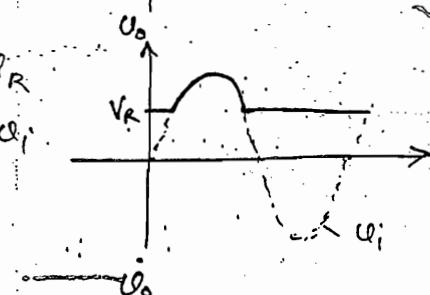
$$U_i < 0 \rightarrow D(\text{on}) \Rightarrow U_o = U_i$$

$$U_i > 0 \rightarrow D(\text{off}) \Rightarrow U_o = 0$$



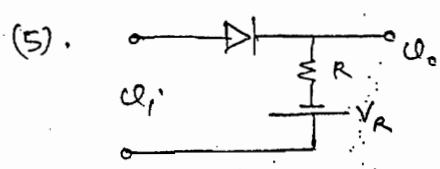
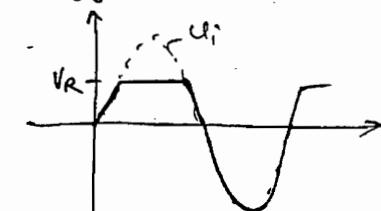
$$U_i < U_R \rightarrow D(\text{off}) \Rightarrow U_o = U_R$$

$$U_i > U_R \rightarrow D(\text{on}) \Rightarrow U_o = U_i$$



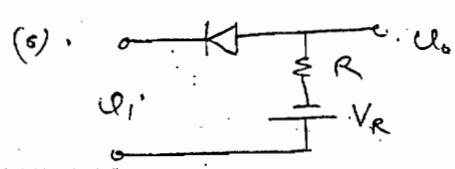
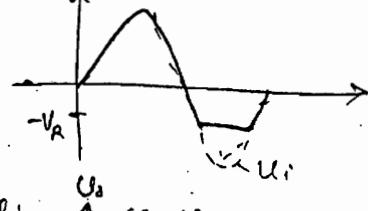
$$U_i < U_R \rightarrow D(\text{on}) \Rightarrow U_o = U_i$$

$$U_i > U_R \rightarrow D(\text{off}) \Rightarrow U_o = U_R$$



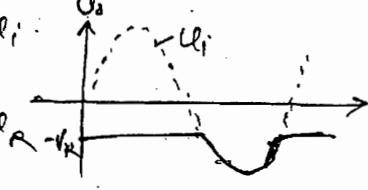
$$U_i < -U_R \rightarrow D(\text{off}) \Rightarrow U_o = -U_R$$

$$U_i > -U_R \rightarrow D(\text{on}) \Rightarrow U_o = U_i$$



$$U_i < -U_R \rightarrow D(\text{on}) \Rightarrow U_o = U_i$$

$$U_i > -U_R \rightarrow D(\text{off}) \Rightarrow U_o = -U_R - V_R$$

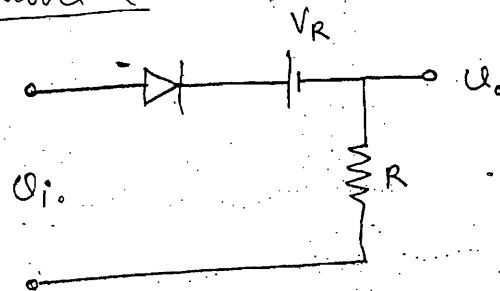


Conclusions:

- (1). When the diode is in forward bias to the P/P sig, the sig will be transmitted above the ref. voltage.
- (2). When the diode is in reverse bias to the P/P sig, the sig will be transmitted below the ref. voltage.

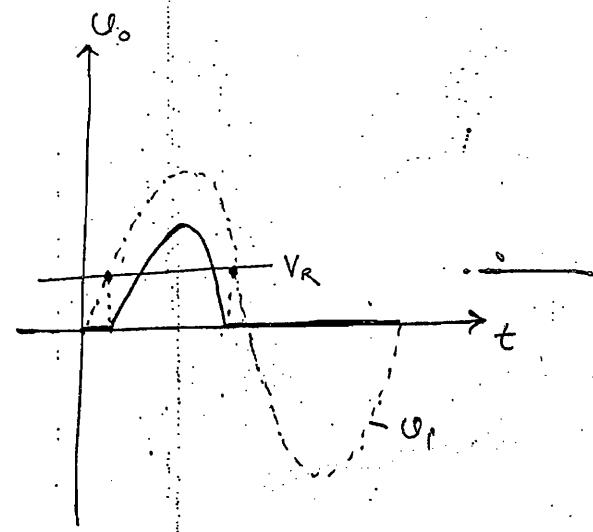
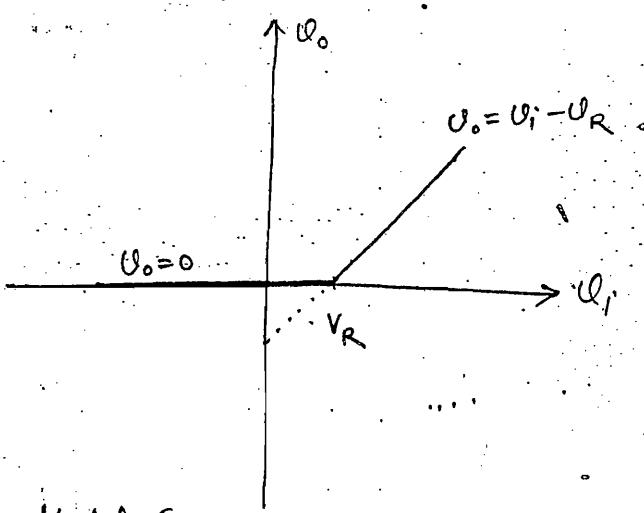
O/P following C/P design:

Model - (1):

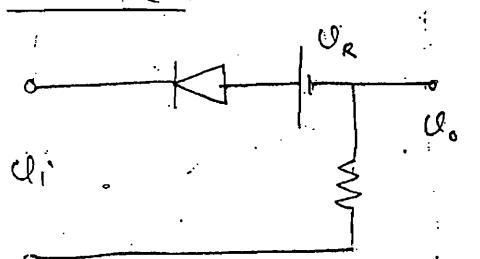


$$v_i < V_R \rightarrow D(\text{off}) \Rightarrow v_o = 0$$

$$v_i > V_R \rightarrow D(\text{on}) \Rightarrow v_o = v_i - V_R$$

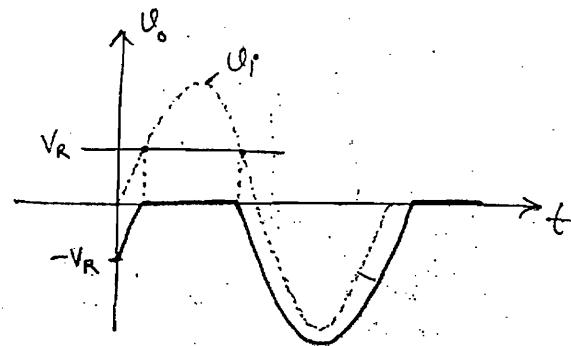
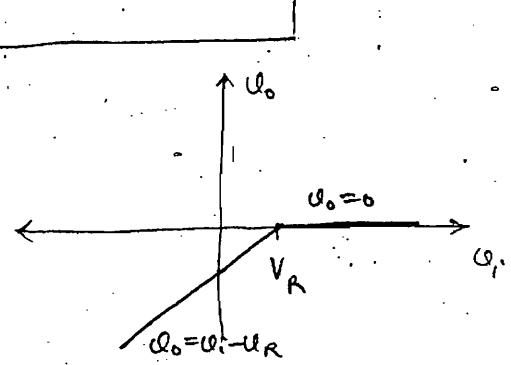


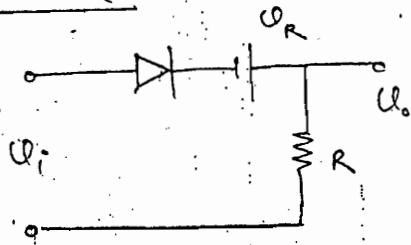
Model - (2):



$$v_i < V_R \rightarrow D(\text{on}) \Rightarrow v_o = v_i - V_R$$

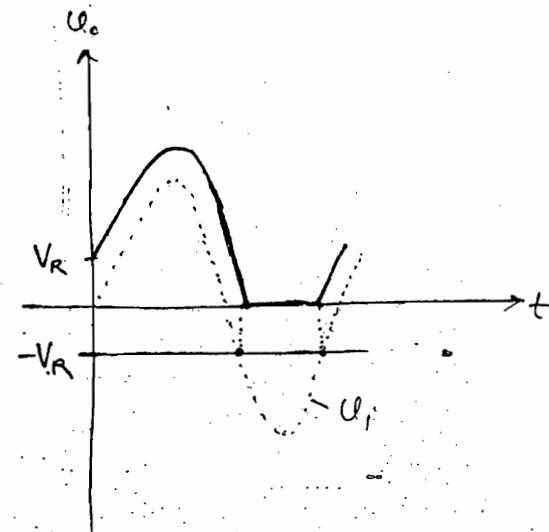
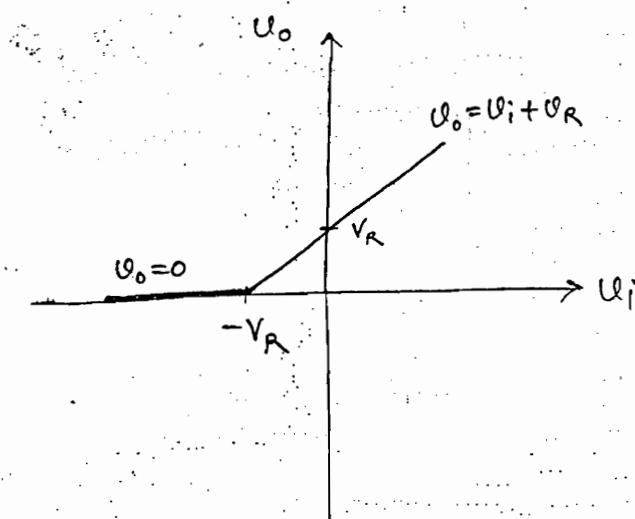
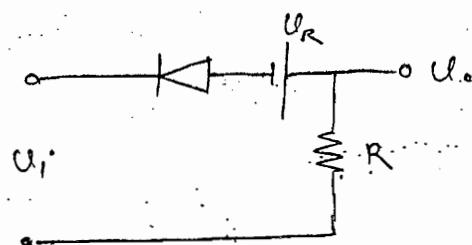
$$v_i > V_R \rightarrow D(\text{off}) \Rightarrow v_o = 0$$



Model - (3) :

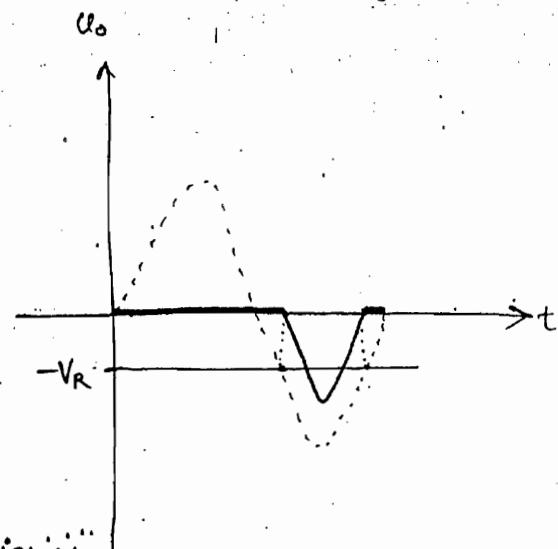
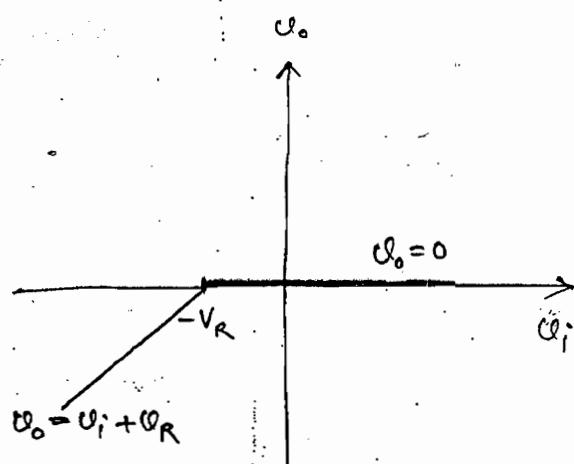
$U_i < -V_R \rightarrow D(\text{off}) \Rightarrow U_o = 0$

$U_i > -V_R \rightarrow D(\text{on}) \Rightarrow U_o = U_i + V_R$

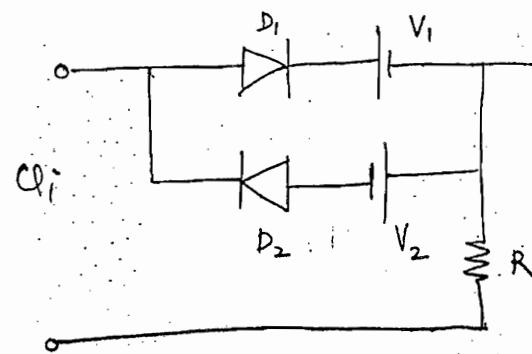
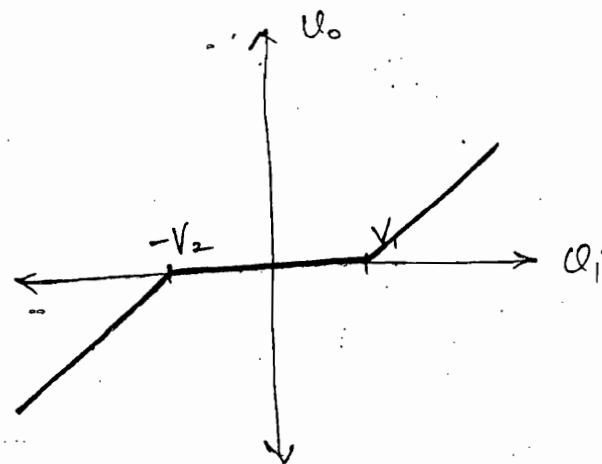
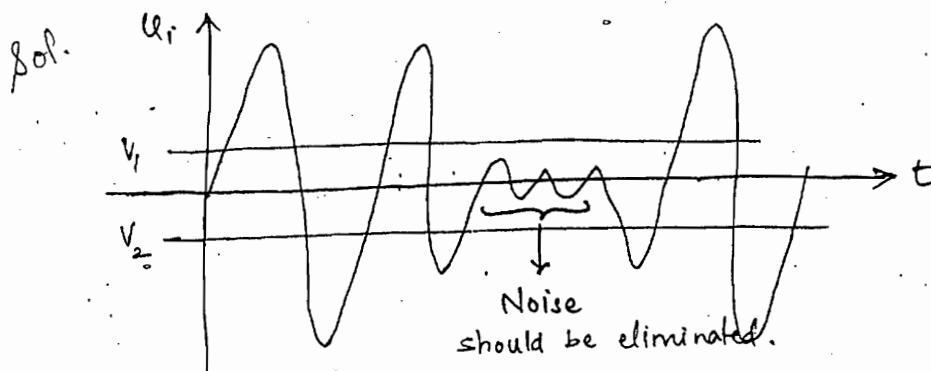
Model - (4) :

$U_i < -V_R \rightarrow D(\text{on}) \Rightarrow U_o = U_i + V_R$

$U_i > -V_R \rightarrow D(\text{off}) \Rightarrow U_o = 0$

Applications of o/p following i/p designii

- Q. Design a noise clipper ckt.. which has to eliminate the noise in b/w 2 threshold levels.



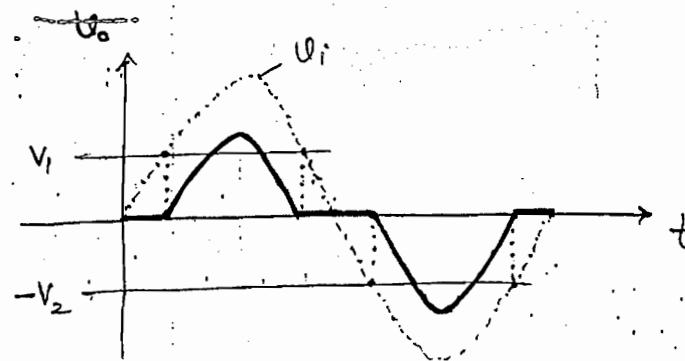
$$U_i < -V_2, D_1(\text{off}), D_2(\text{on})$$

$$U_o = U_1 + U_2$$

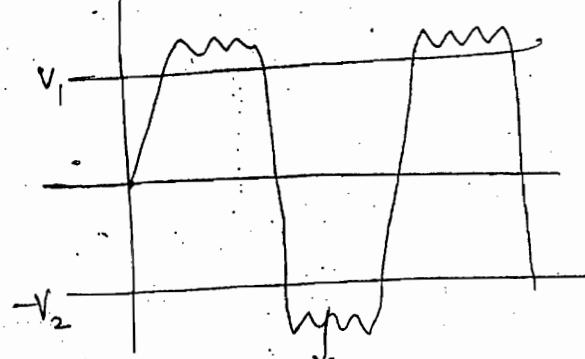
$$-V_2 < U_i < V_1, D_1(\text{off}), D_2(\text{on})$$

$$U_o = 0$$

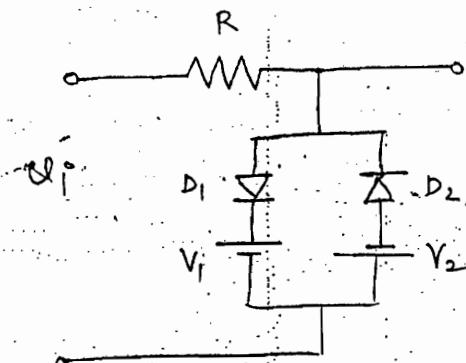
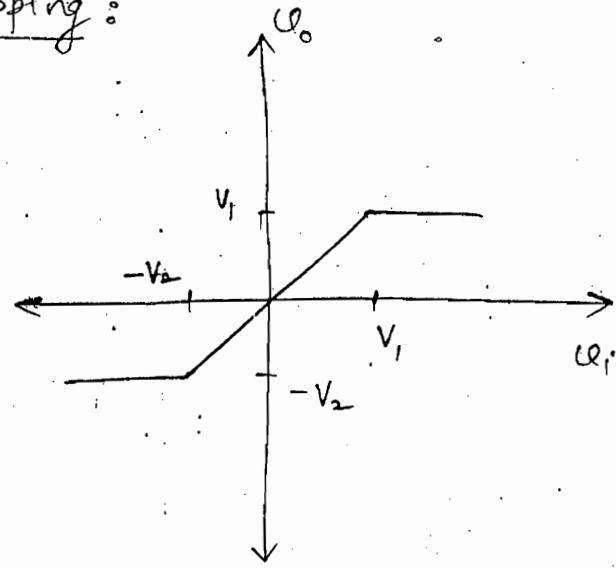
$$U_i > V_1, D_1(\text{on}), D_2(\text{off}), U_o = U_i - V_1$$



(A) Two Independent level Clipping:



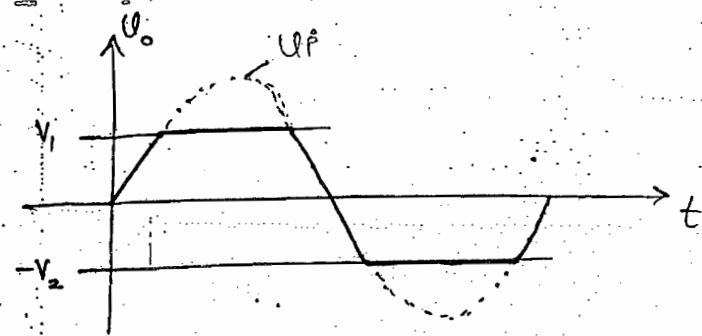
Noise should be eliminated



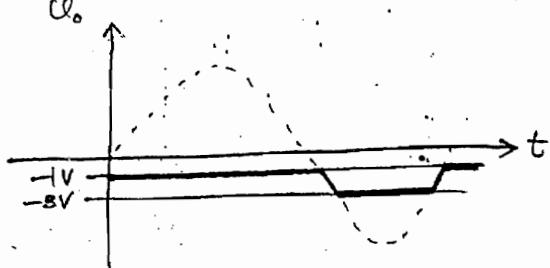
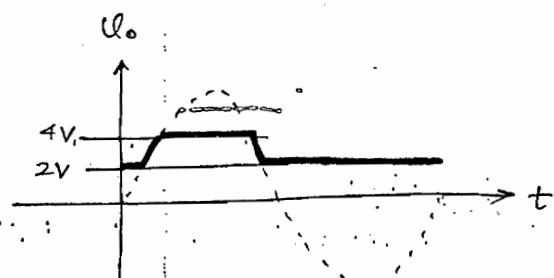
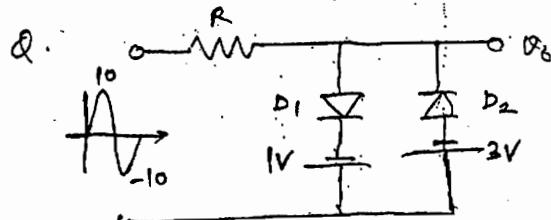
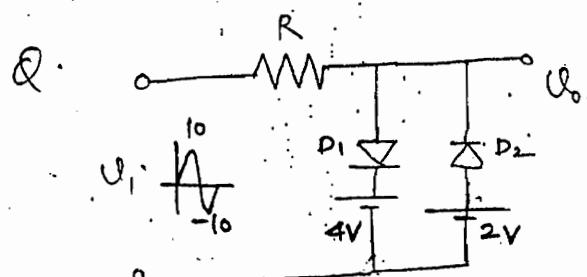
$$U_i < -V_2, D_1(\text{off}), D_2(\text{on}), U_o = -V_2$$

$$-V_2 < U_i \leq V_1, D_1(\text{off}), D_2(\text{on}), U_o = U_i$$

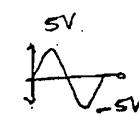
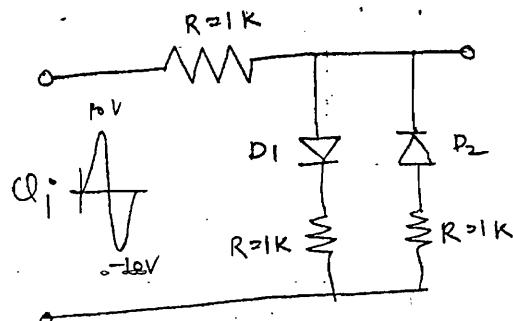
$$U_i > V_1, D_1(\text{on}), D_2(\text{off}), U_o = V_1$$



* $V_{D_1} > V_{D_2}$

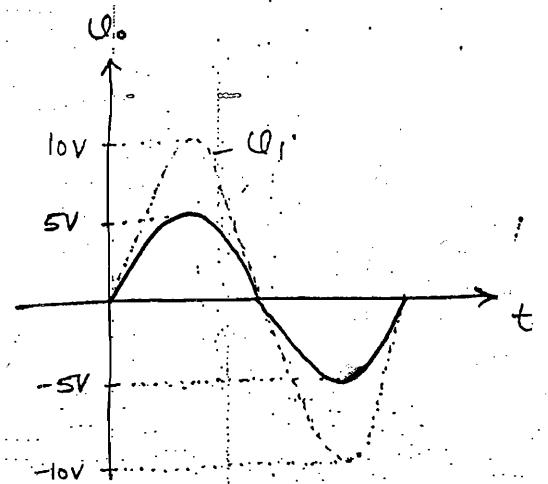
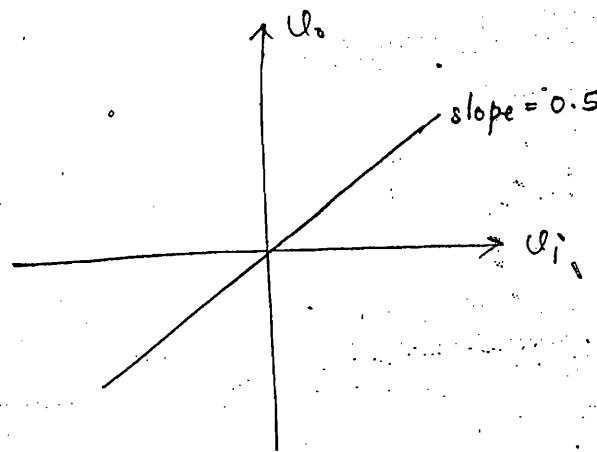


(5). Attenuator:

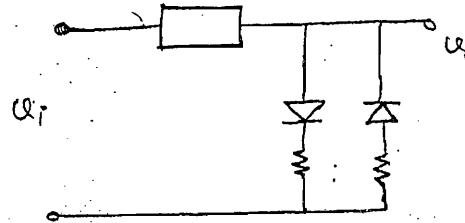


$$U_i < 0, D_1(\text{off}), D_2(\text{on}), U_o = \frac{U_i}{2}$$

$$U_i > 0, D_1(\text{on}), D_2(\text{off}), U_o = \frac{U_i}{2}$$

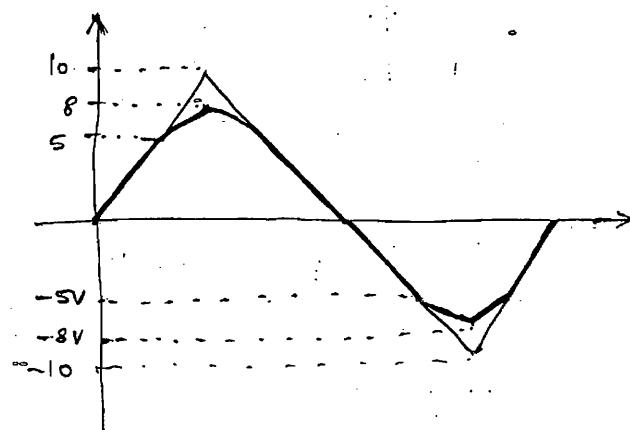


attenuation depending on this
↓ Resistance



Q. Design an IC diode function generator to generate triangular wave into sine wave.

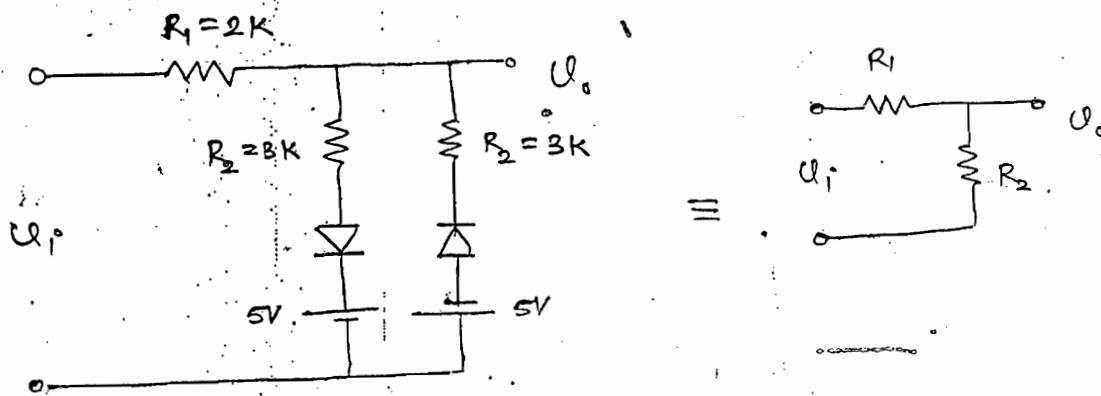
Sol.



$$\frac{\Delta U_o}{\Delta U_i} = \frac{8 - 5}{10 - 5}$$

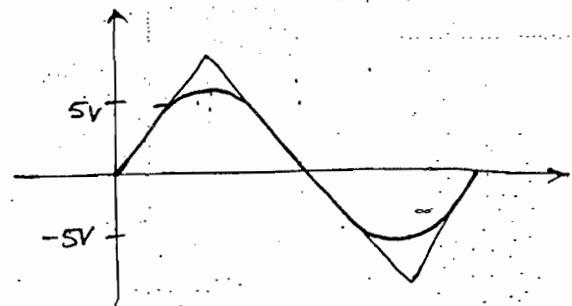
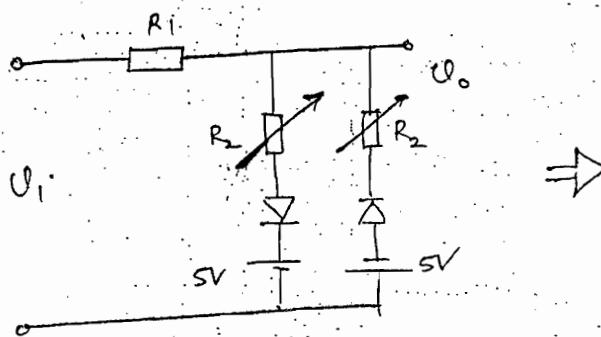
$$= \frac{3}{5}$$

$$= \frac{3K}{2K+3K}$$

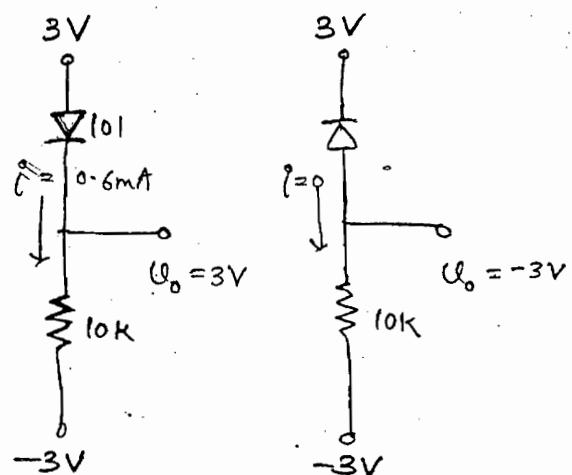
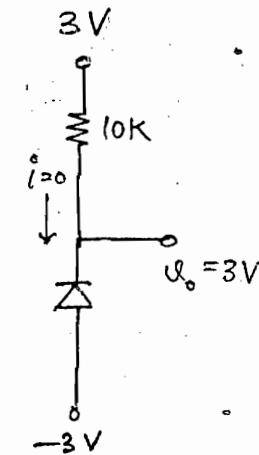
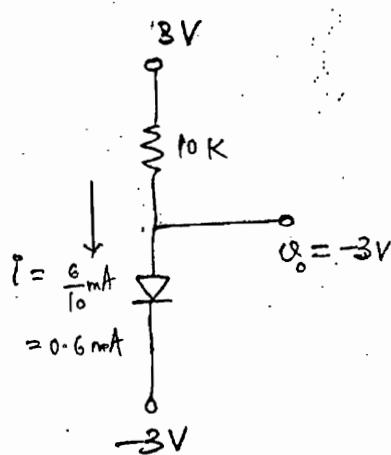


$$U_o = U_i \times \frac{R_2}{R_1 + R_2}$$

$$\frac{U_o}{U_i} = \frac{R_2}{R_1 + R_2} = \frac{\Delta U_o}{\Delta U_i} = \frac{3}{2+3} \Rightarrow R_1 = 2K, R_2 = 3K$$



Single Branch designs :

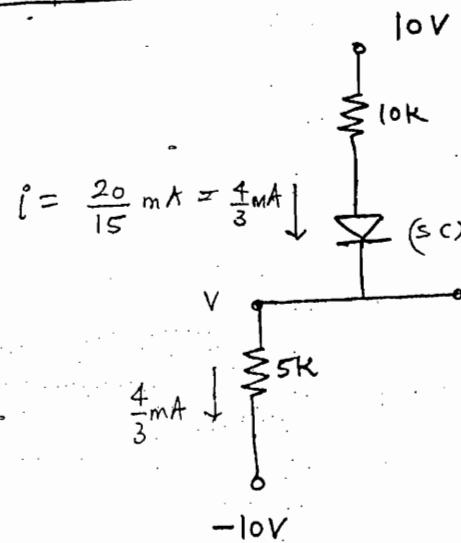


* In single branches, current cannot be changed, only voltage can be changed by different configurations.

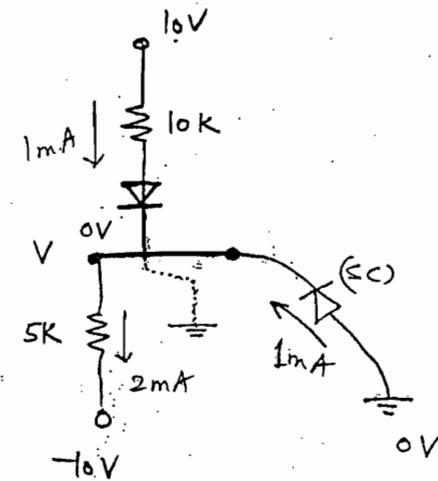
Multiple Branch:

Preference - Single Branch.

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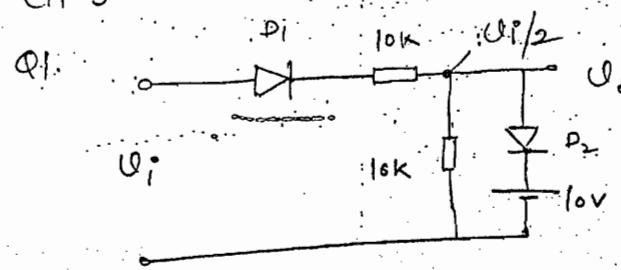
$$V = 10 - \frac{10 \times 20}{15} = -3.3 \text{ V}$$



$$V = 0 \text{ V} \text{ (due to ground)}$$

* In multiple branches, currents are different and can be changed by changing configurations.

CH-3

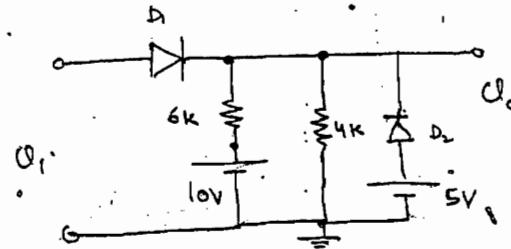


$$V_i < 0 \rightarrow D_1 (\text{off}), D_2 (\text{off}), V_o = 0 \quad \checkmark$$

$$0 < V_i < 10 \rightarrow D_1 (\text{on}), D_2 (\text{off}), V_o = V_i/2$$

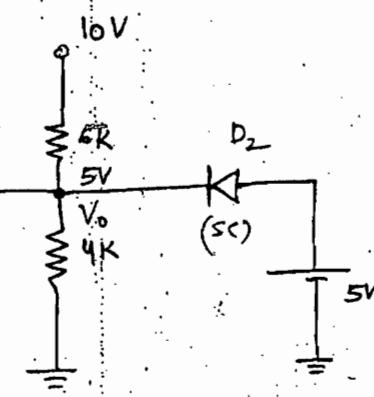
$$V_i > 10 \rightarrow D_1 (\text{off}), D_2 (\text{on}), V_o = 10 \text{ V}$$

Q7.

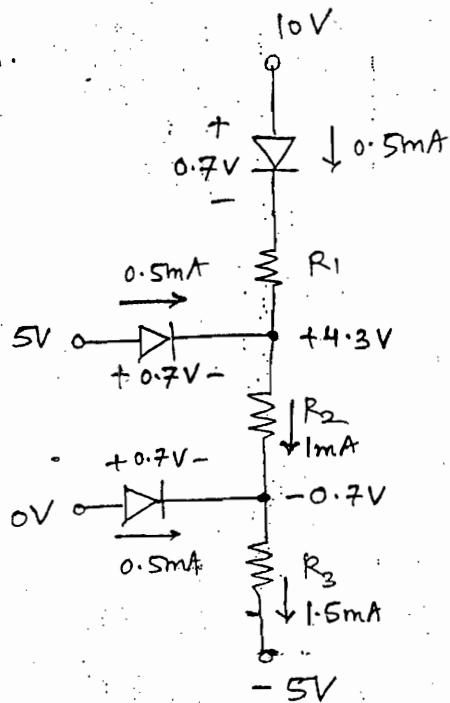


$$\text{If } V_i > 5 \Rightarrow D_1 (\text{on}), V_o = V_i$$

$$\text{If } V_i < 5 \Rightarrow D_1 (\text{off}), V_o = 5 \text{ V}$$



Q8.



$$\frac{10 - 0.7 - 4.3}{0.5 \text{ mA}} = R_1$$

$$\therefore R_1 = 10 \text{ k}\Omega$$

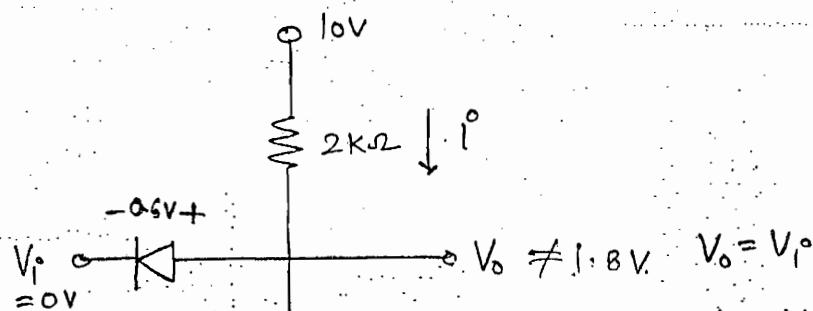
$$R_2 = \frac{4.3 - (-0.7)}{0.5 \text{ mA} + 0.5 \text{ mA}}$$

$$= \frac{5}{1 \text{ mA}}$$

$$= 5 \text{ k}\Omega$$

$$R_3 = \frac{-0.7 - (-5)}{1.5 \text{ mA}} = 2.86 \text{ k}\Omega$$

Q11.

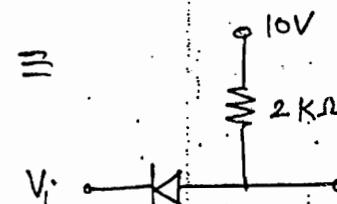


Do not
get that
much
energy to
drop that
at each
diode

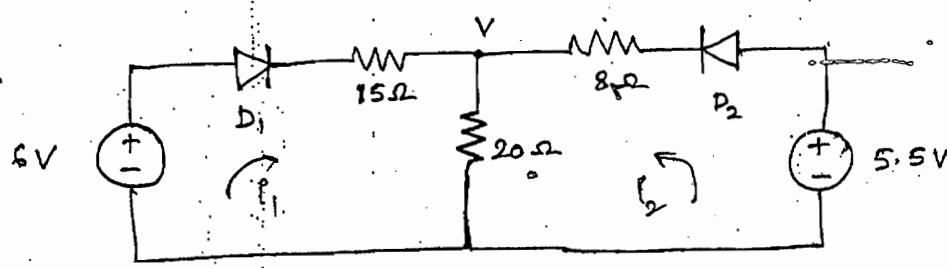
$$\therefore V_o = 0.6V$$

$$i^{\circ} = \frac{10 - 0.6}{2k}$$

$$= 4.7 \text{ mA}$$



Q12.



$$\frac{V - 6 + 0.6}{15} + \frac{V}{20} + \frac{V - 5.5 + 0.6}{8} = 0$$

$$V\left(\frac{1}{15} + \frac{1}{20} + \frac{1}{8}\right) = \frac{6}{15} + \frac{5.5}{8} - \frac{0.6}{15} + \frac{0.6}{8}$$

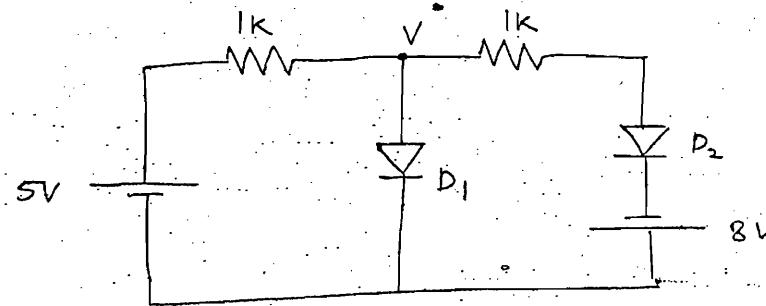
$$V = \cancel{\frac{6+5.5}{15+20+8}} = 4.024 V$$

~~Q12~~

$$P_1 = \frac{6 - 0.6 - 4.024}{15} = +ve \quad D_1 (\text{ON})$$

$$P_2 = \frac{5.5 - 0.6 - 4.024}{8} = +ve \quad D_2 (\text{ON})$$

Q13.



$$\frac{V - 5}{1} + \frac{V + 8}{1} = 0$$

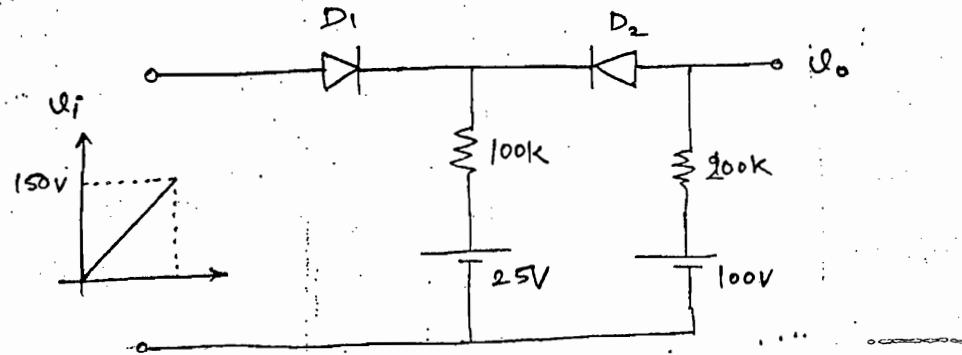
$$2V = -3 \Rightarrow V = -1.5 \text{ volts} \Rightarrow D_1 (\text{off})$$

Q14.

Special problems in clippers:

Ramp I/P problems:

Q.



$u_i = 0$: , D_1 (off), D_2 (on)

$$\text{Circuit diagram: } \begin{array}{c} u_o \\ | \\ 100k \parallel 25 \quad | \\ | \quad | \\ 25 \quad 100 \end{array}$$

$$R = \frac{100 - 25}{300} = \frac{75}{300} = 0.25 \text{ mA}$$

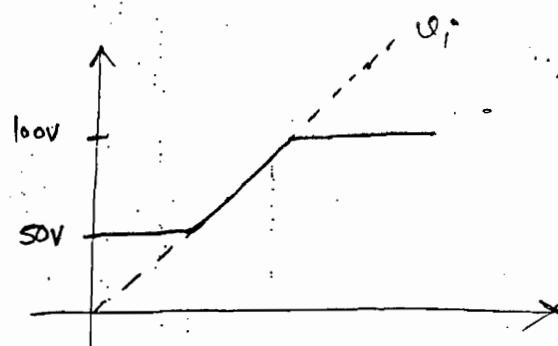
$$u_o = 100 - 200 \times 0.25 \\ = 50 \text{ V.}$$

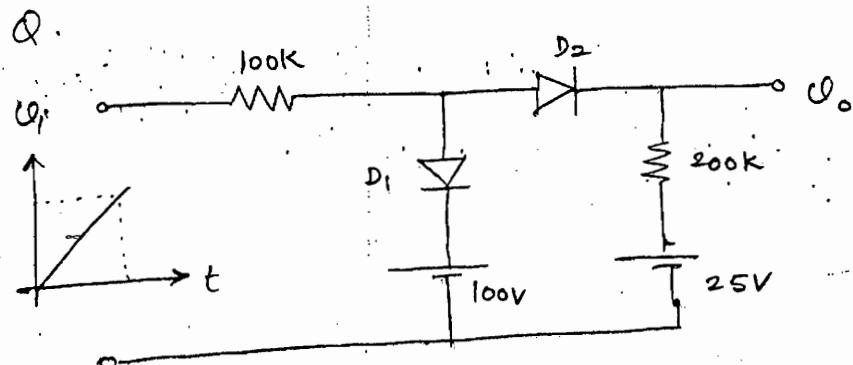
$\therefore 0 < u_i < 50 \text{ V} \Rightarrow D_1 (\text{off}) \Rightarrow D_2 (\text{on}) \Rightarrow u_o = 50 \text{ V}$

$u_i > 50 \text{ V}$: D_1 (on) , D_2 (on) $\Rightarrow u_o = u_i$

$u_i < 100 \text{ V}$: D_1 (on) , D_2 (on) $\Rightarrow u_o = u_i$

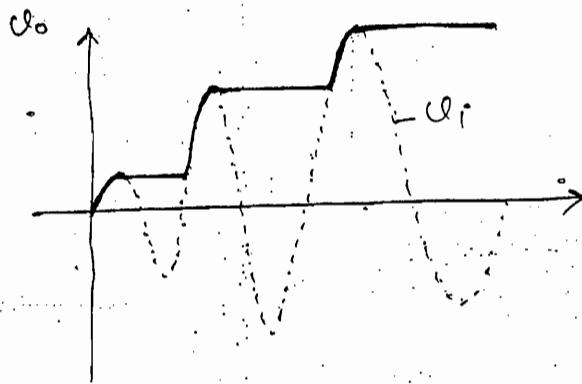
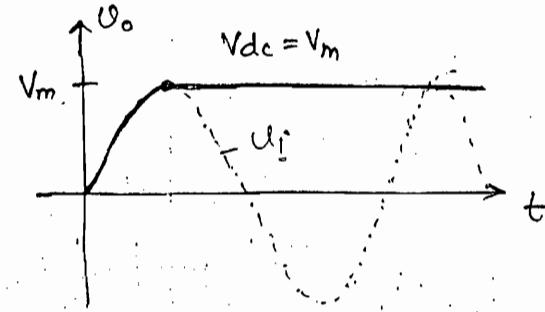
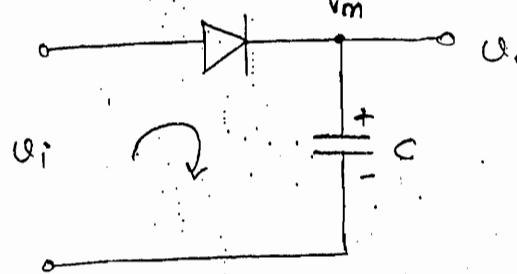
$u_i > 100 \text{ V}$: D_1 (on) , D_2 (off) $\Rightarrow u_o = 100 \text{ V}$



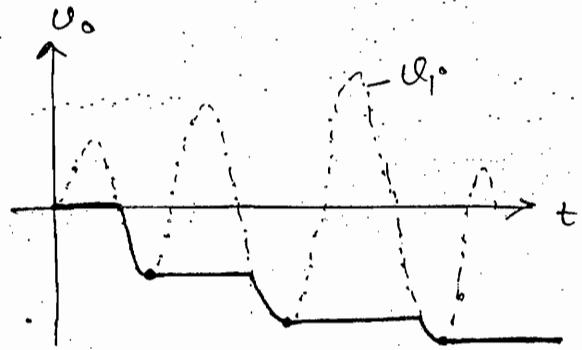
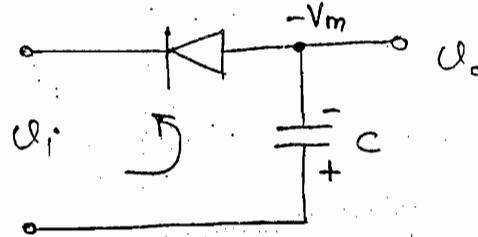


leak detector, clamer , voltage Multiplier :

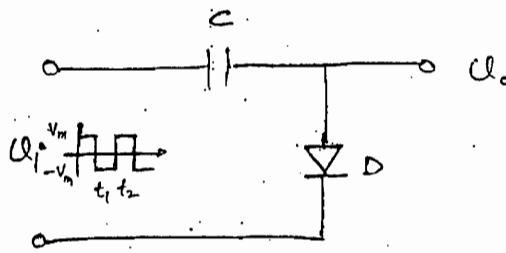
Positive peak detector:



Negative peak detector:



Negative Clammer:



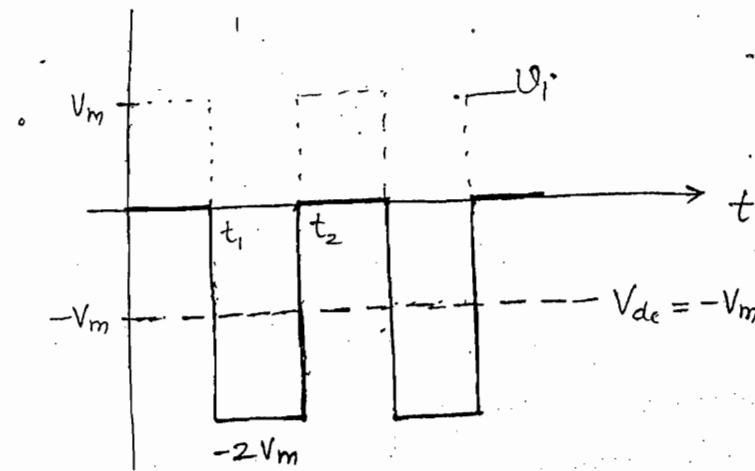
$0 \rightarrow t_1$:

$C \rightarrow$ get charged to V_m ; D (on)

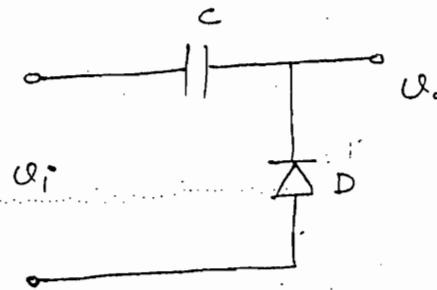
$$u_o = 0$$

$t_1 \rightarrow t_2$: D (off)

$$u_o = -2V_m$$



Positive Clamper:



$0 \rightarrow t_1 : D \text{ (off)}$

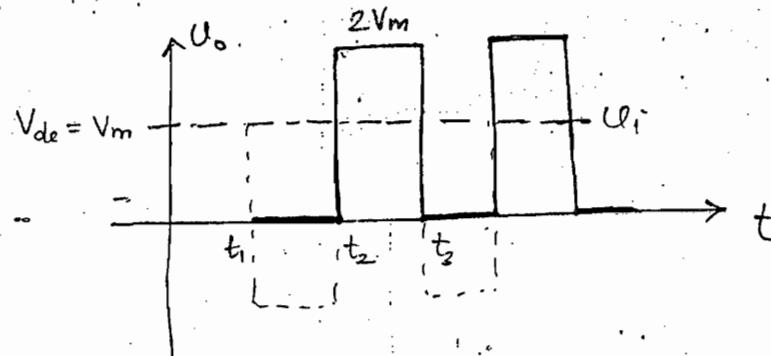
Start with t_1

C \rightarrow unchanged.

$t_1 \rightarrow t_2 : D \text{ (on)}$

C \rightarrow charges to V_m , $U_o = 0$

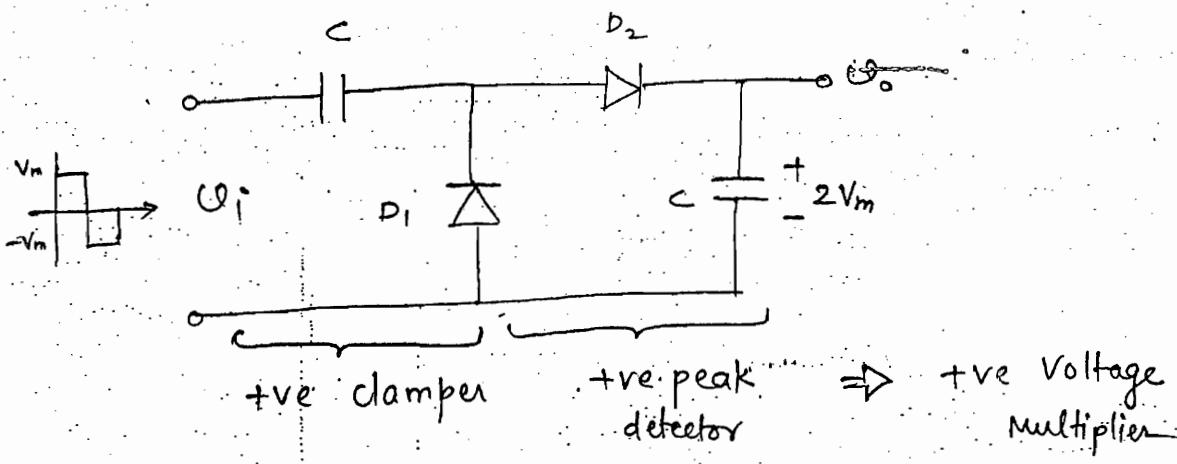
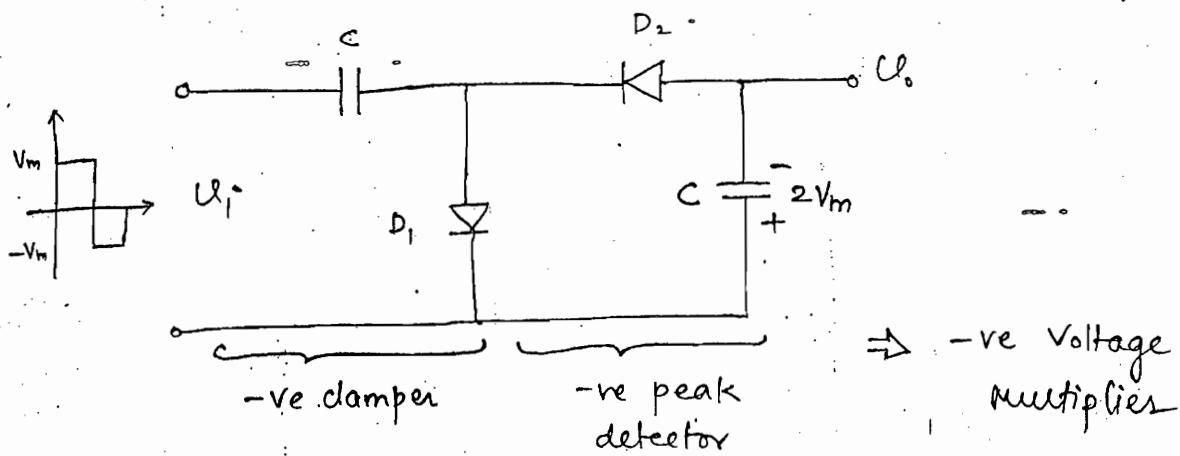
$t_2 \rightarrow t_3 : D \text{ (off)}$, $U_o = 2V_m$



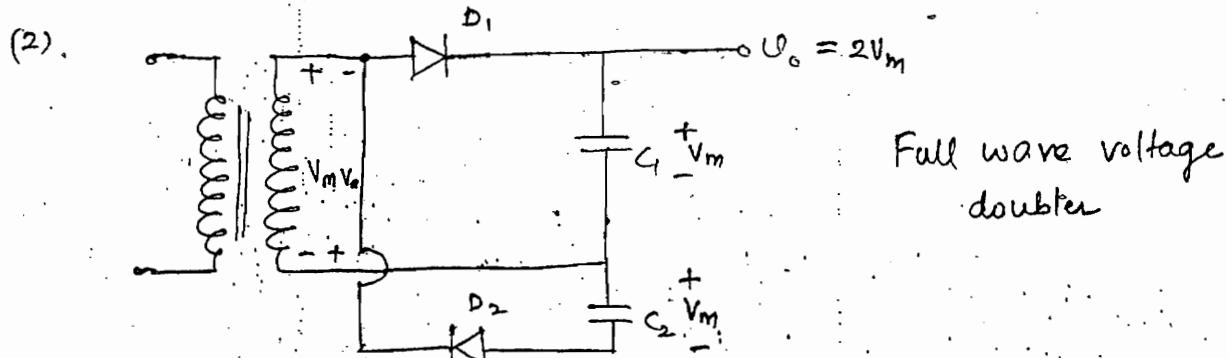
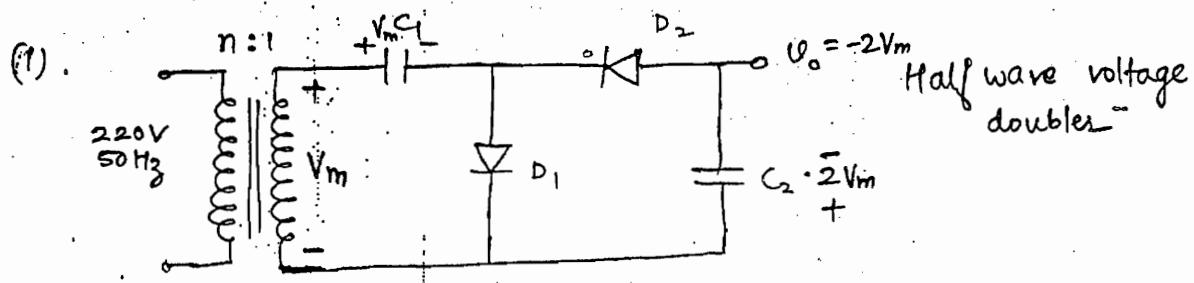
Techniques in clamps:

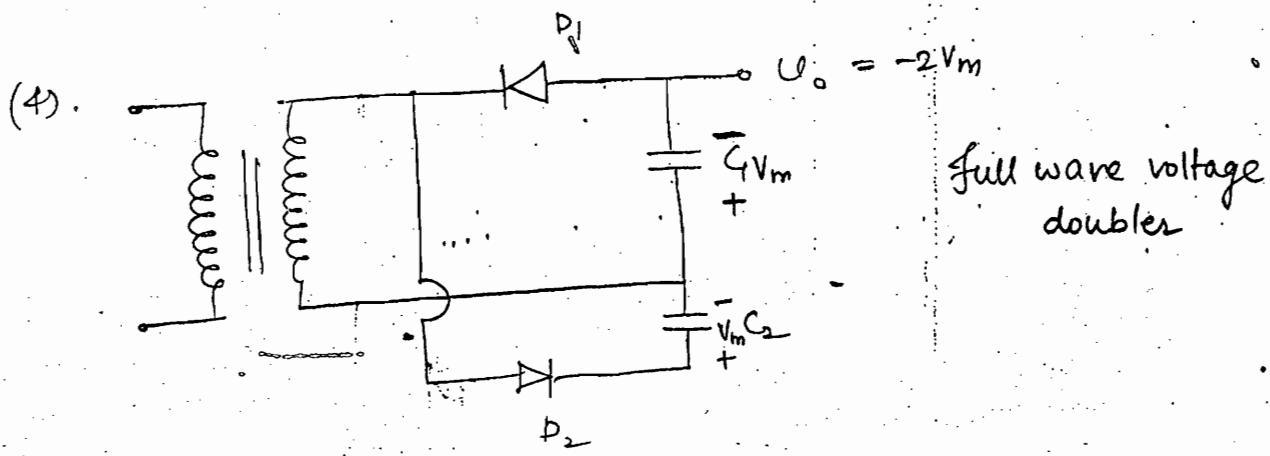
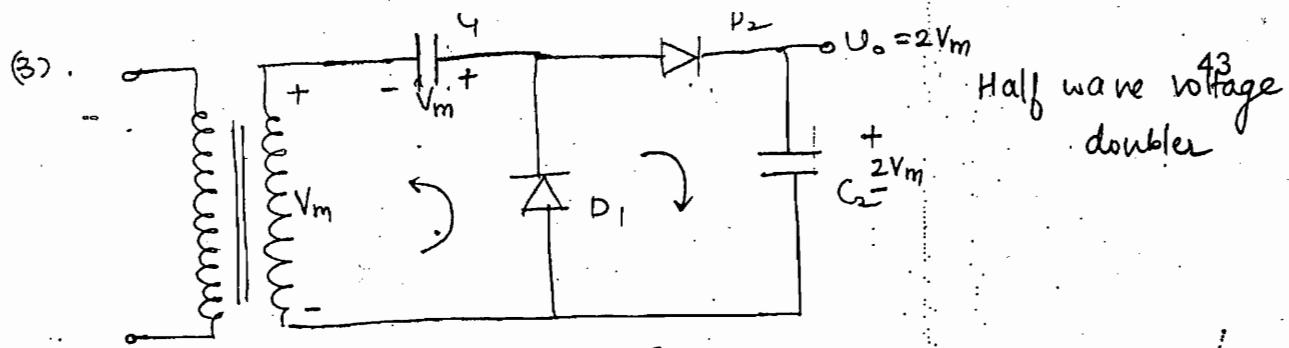
- (1). When the diode is in downward dirn, the total sig will be clamped (pulling) below the ref. voltage.
- (2). When the diode is in \uparrow dirn, the total sig will be clamped above the ref. voltage

Voltage multiplier:



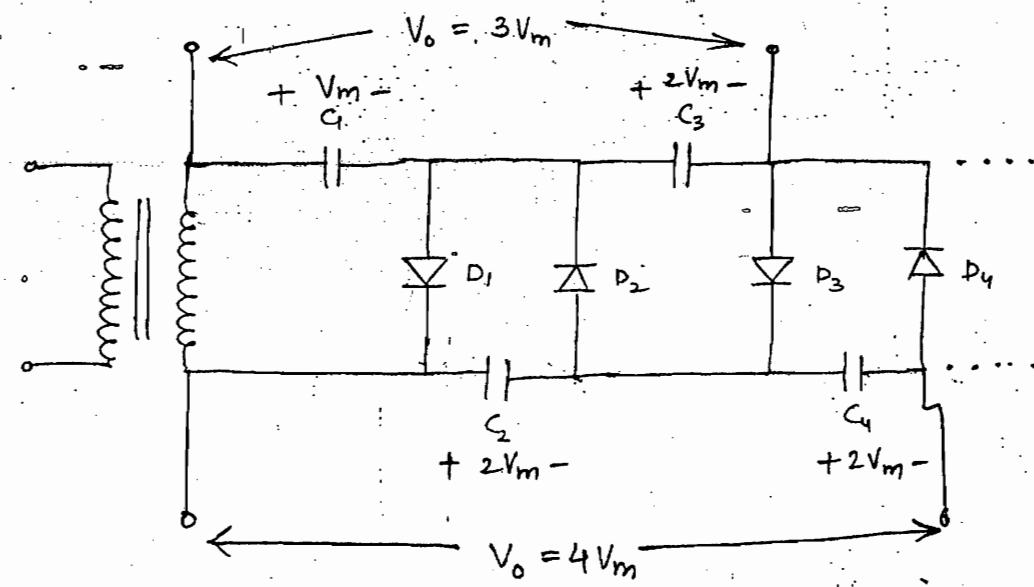
Voltage multiplier problems:



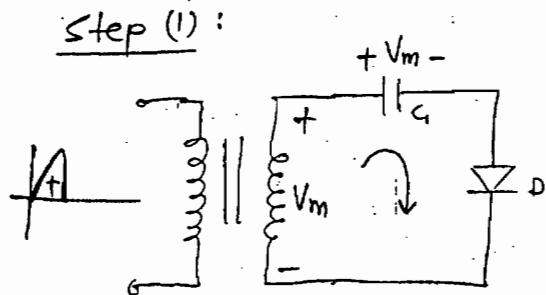


Q. Design the voltage multiplier ckt which generates n multiples of DC.

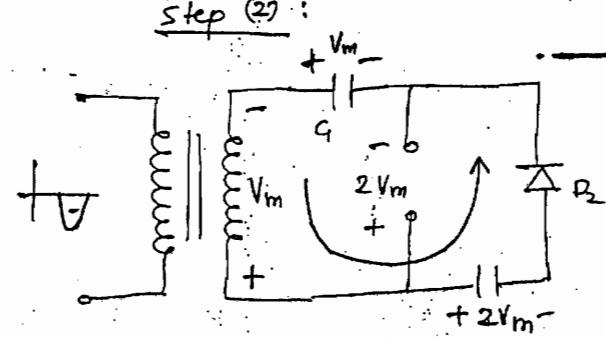
Sol.



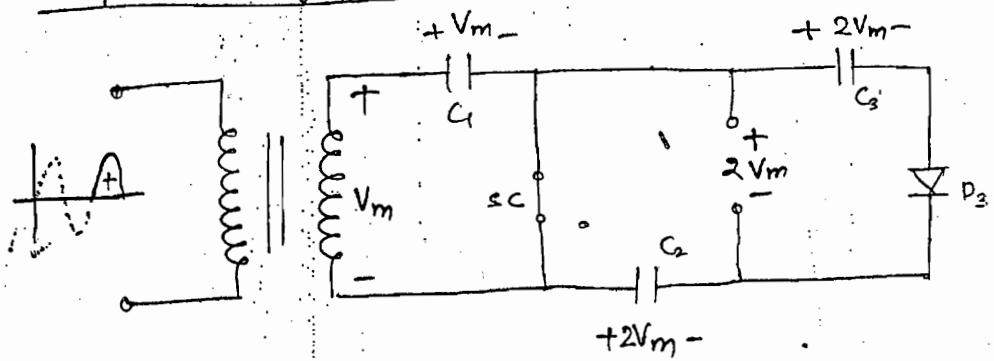
Step (1) :



Step (2) :

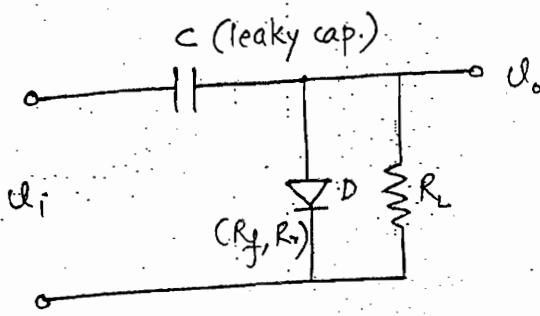
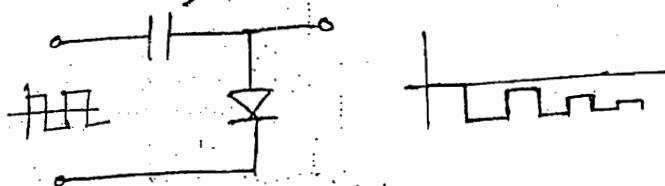


Step (3): 2nd cycle (+ve)



Clampers:

practical cap. (leaky cap.)



R_f = forward resistance
= $10\ \Omega$

R_r = reverse resistance
= $10\ M\Omega$

$$R_f \xrightarrow{(10\ \Omega)} R \xrightarrow{(10\ K\Omega)} R_r$$

Geometric mean gives the value of (R_f, R_r) .

$$C = \sqrt{ab}$$

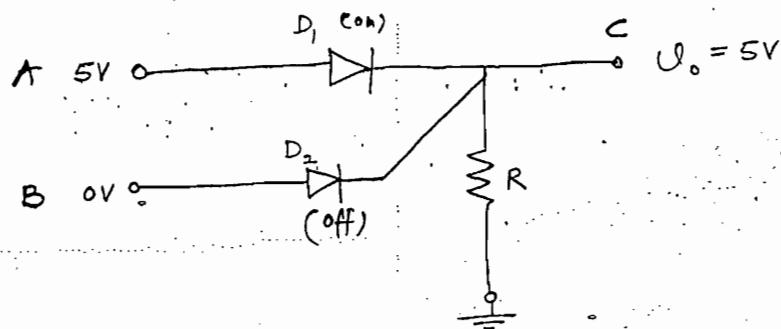
$$R = \sqrt{R_f \cdot R_r}$$

$$R = \sqrt{10 \times 10 \times 10^6} = 10^4 = 10\ K\Omega$$

Diode as a digital logic gate :

"OR" logic : $(A + B = C)$

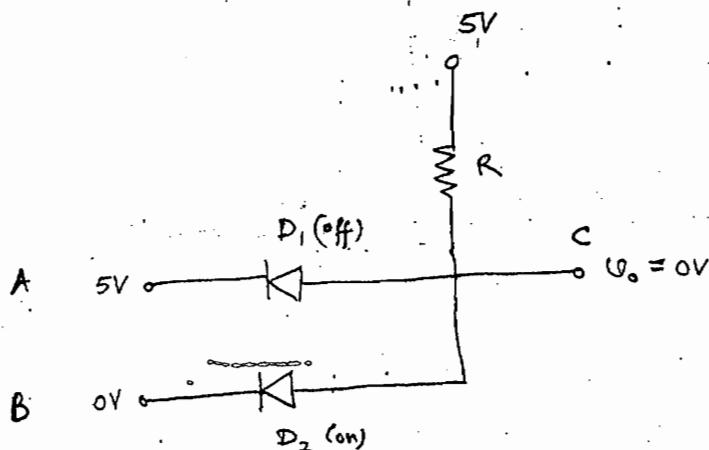
A	B	C
0	0	0
0	1	1
1	0	1
1	1	1



U/Copier Lab

"AND" logic : $(A \cdot B = C)$

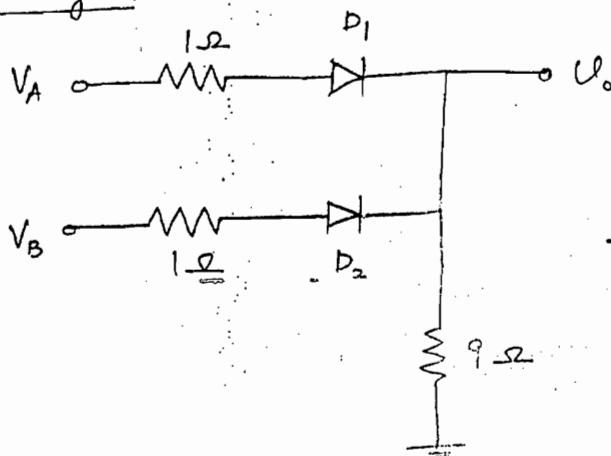
A	B	C
0	0	0
0	1	0
1	0	0
1	1	1



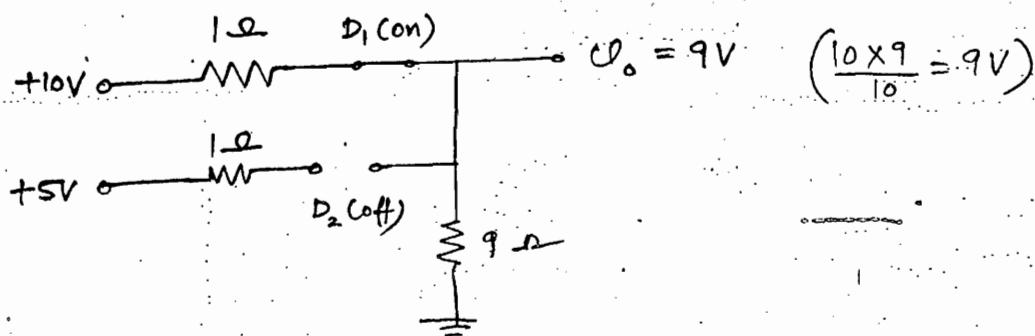
Laser/Inkjet/Copier Label A

Problems:

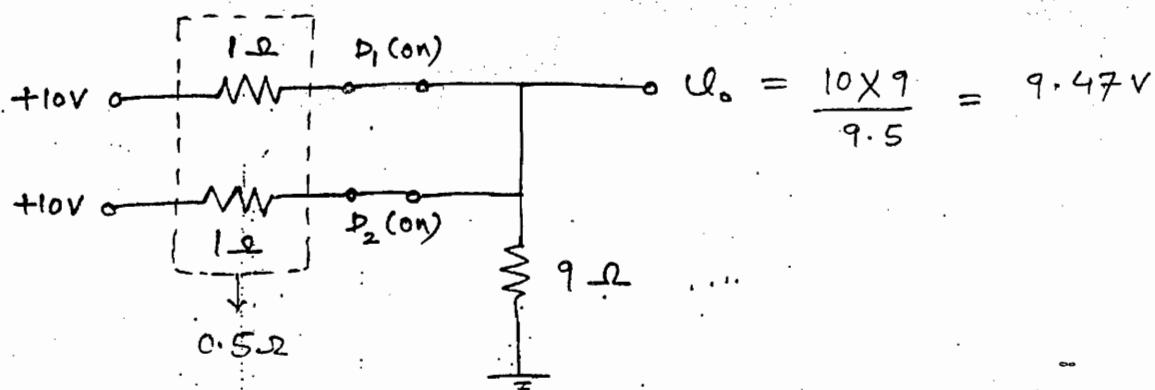
"OR" logic:



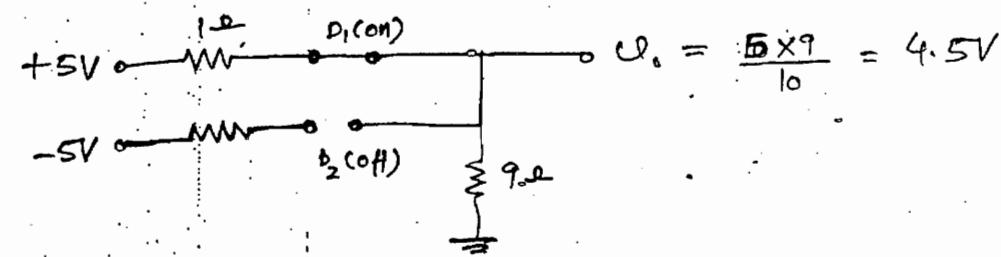
Case : (1) : $V_A = 10V, V_B = 5V$



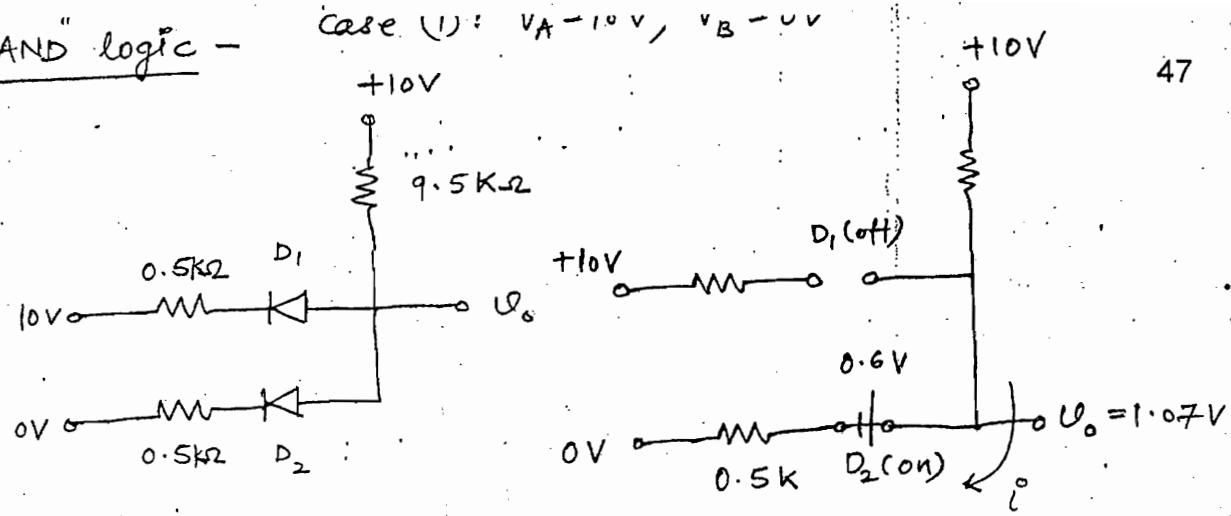
Case (2) : $V_A = 10V, V_B = 10V$



Case (3) : $V_A = 5V, V_B = -5V$

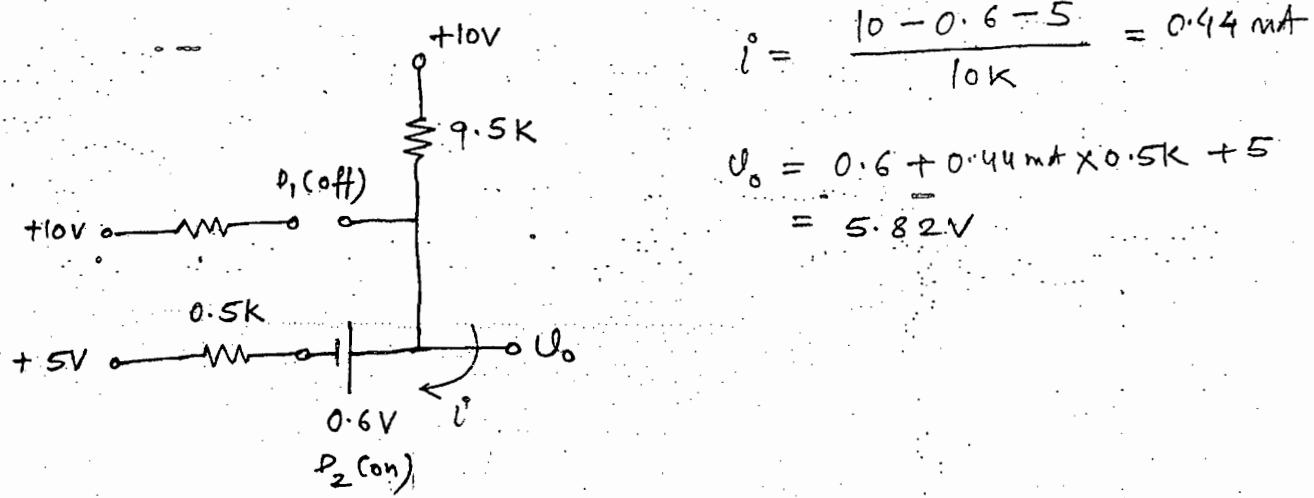


"AND" logic -

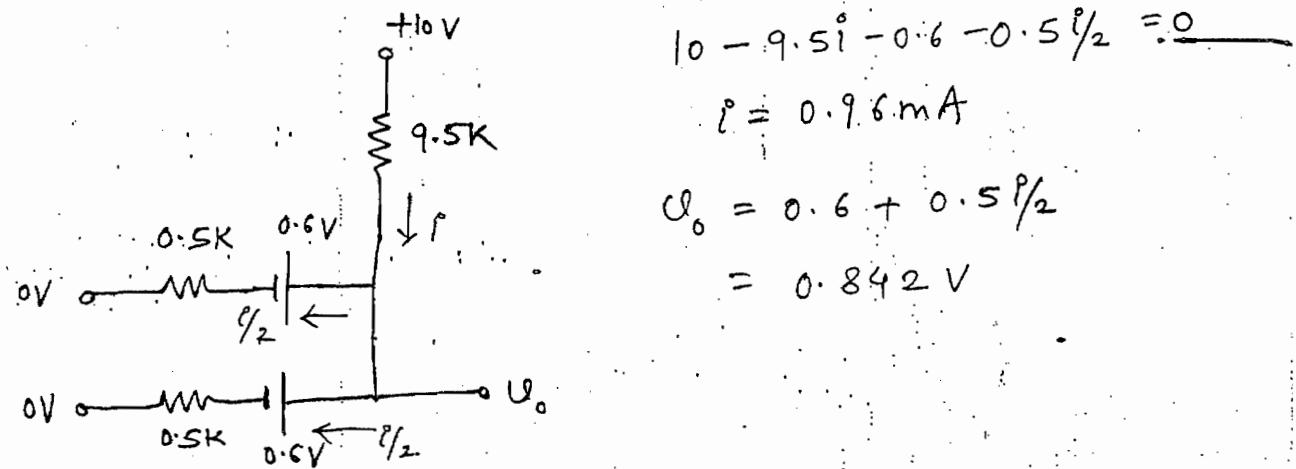


47

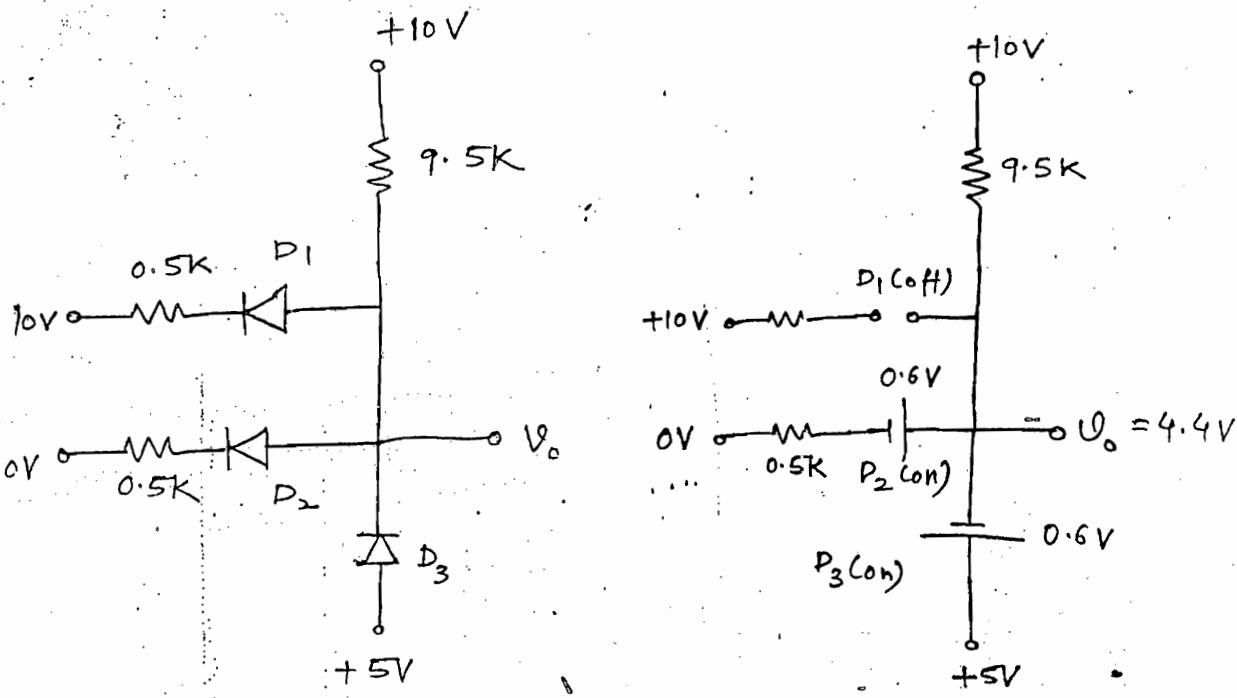
Case (2) : $V_A = 10V$ and $V_B = 5V$



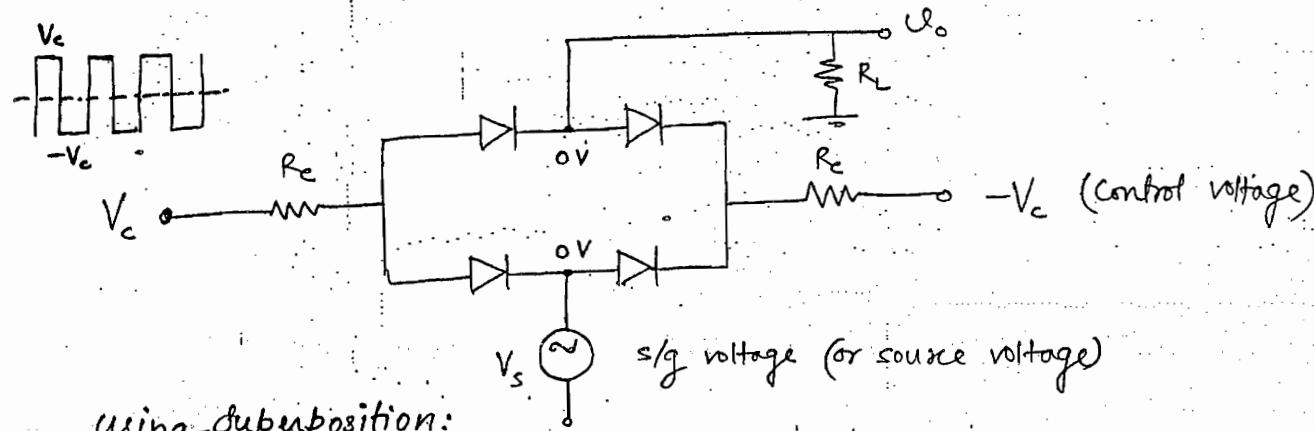
Case (3) : $V_A = 0V$ and $V_B = 0V$



Q.

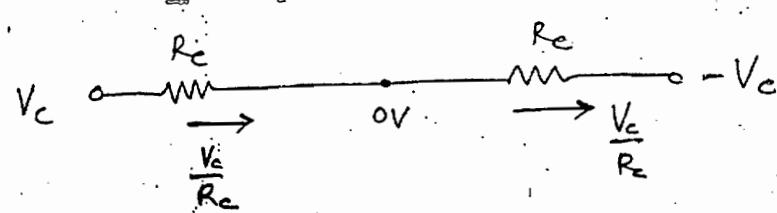
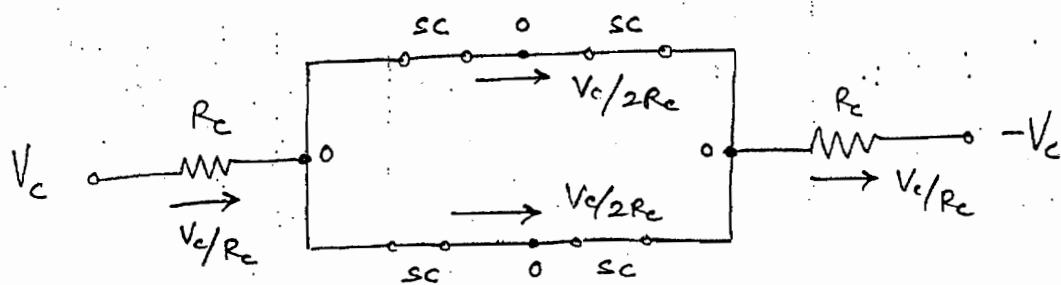


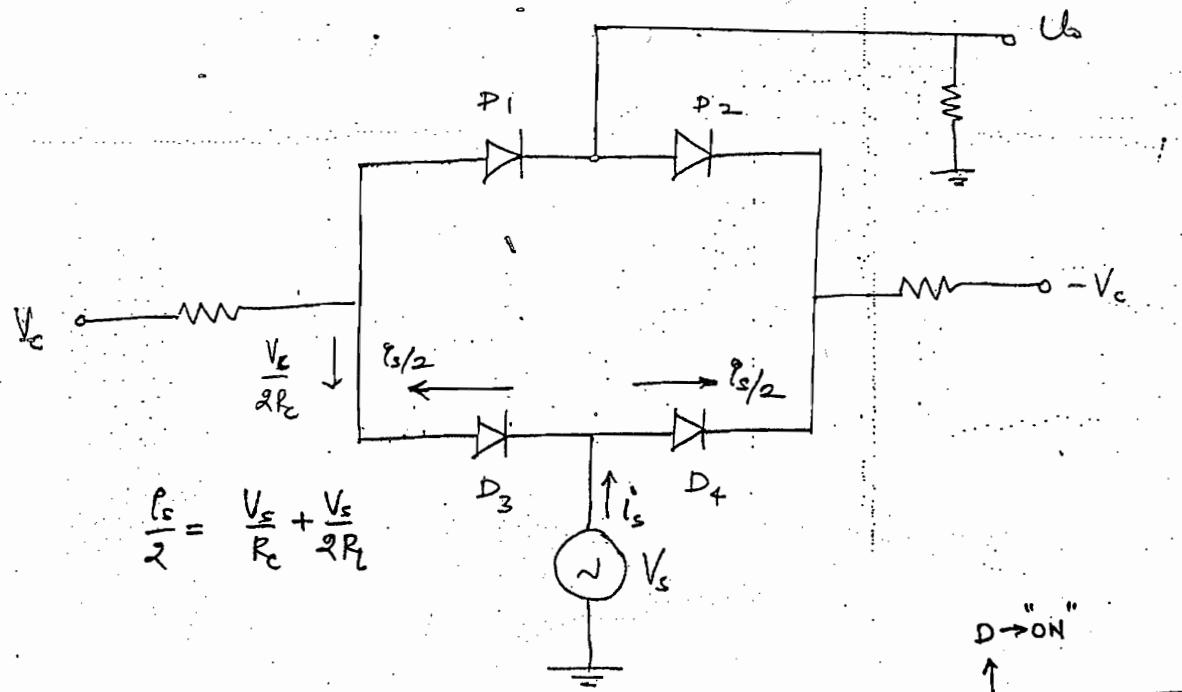
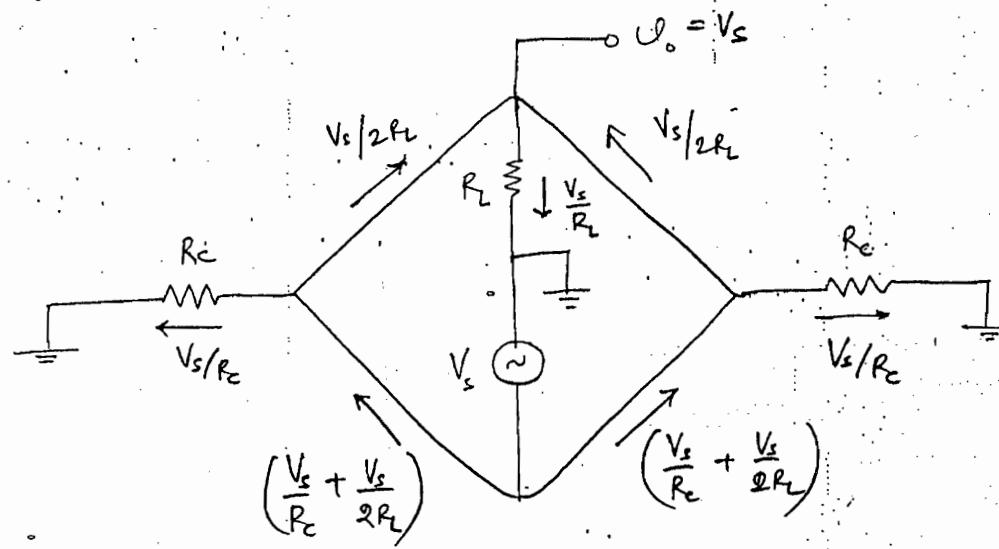
Diode as an analog gate:



using superposition:

DC Analysis:

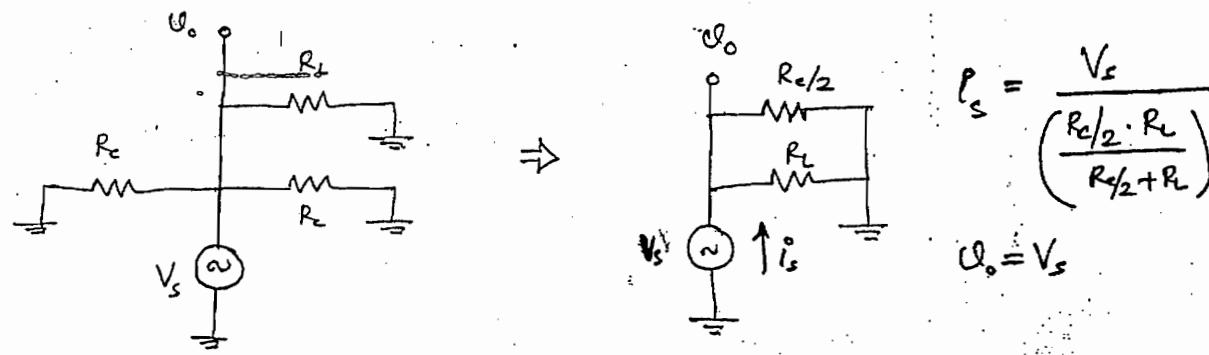


AC Analysis:

$$\frac{V_C}{2R_C} > \frac{i_s}{2} \quad (\text{Diode will be ON})$$

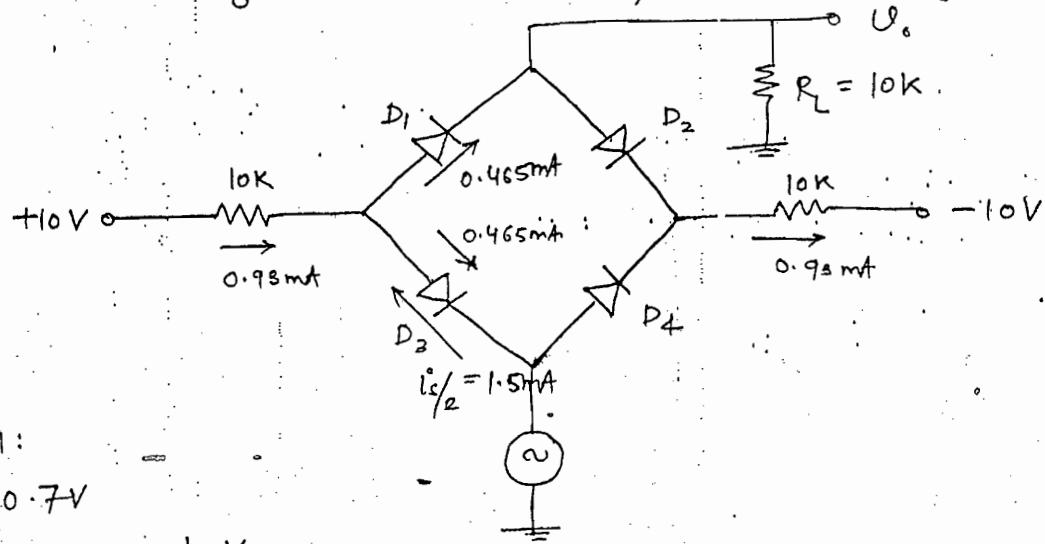
$\therefore \frac{V_C}{2R_C} > \frac{V_s}{R_C} + \frac{V_s}{2R_L}$

D → "ON"



problem:

Draw the Transfer characteristic for the ckt given below:

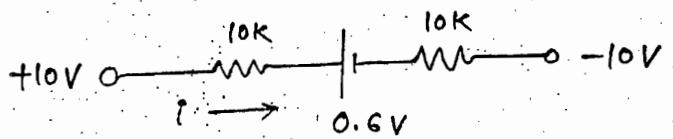


Given:

$$V_T = 0.7V$$

$$-10V \leq V_s \leq 10V$$

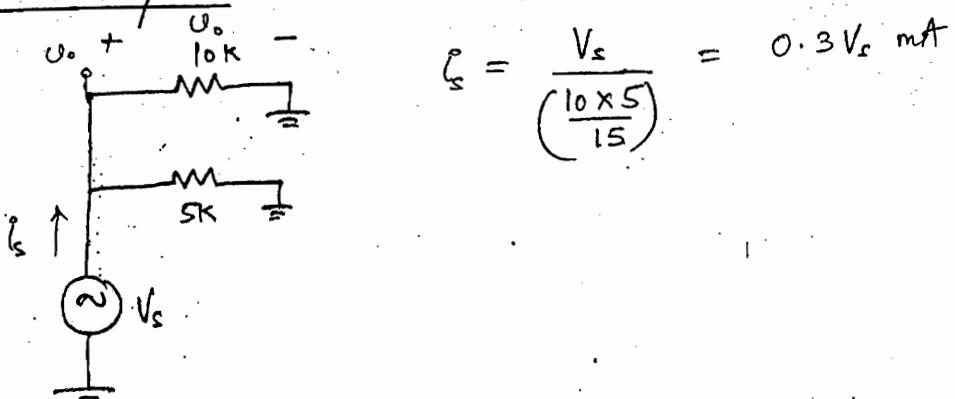
Sol. D.C Analysis:



$$I = \frac{10 - 1.4 - (-10)}{20k} = 0.93 \text{ mA}$$

$$I_{\frac{1}{2}} = 0.465 \text{ mA}$$

A.C Analysis:



Assume : peak level that is. $V_s = 10V$

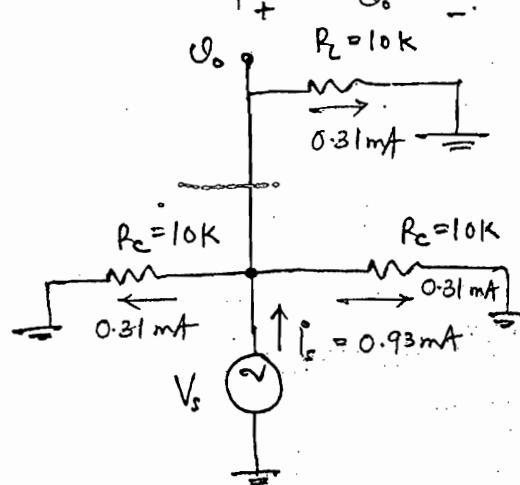
$$\therefore I_s = 0.3(10) = 3 \text{ mA}$$

$$I_{\frac{1}{2}} = 1.5 \text{ mA}$$

At this source current, D_3 (off). s/g can be passed only if $I_{\frac{1}{2}}$ is less than η , i.e. 0.465 mA .

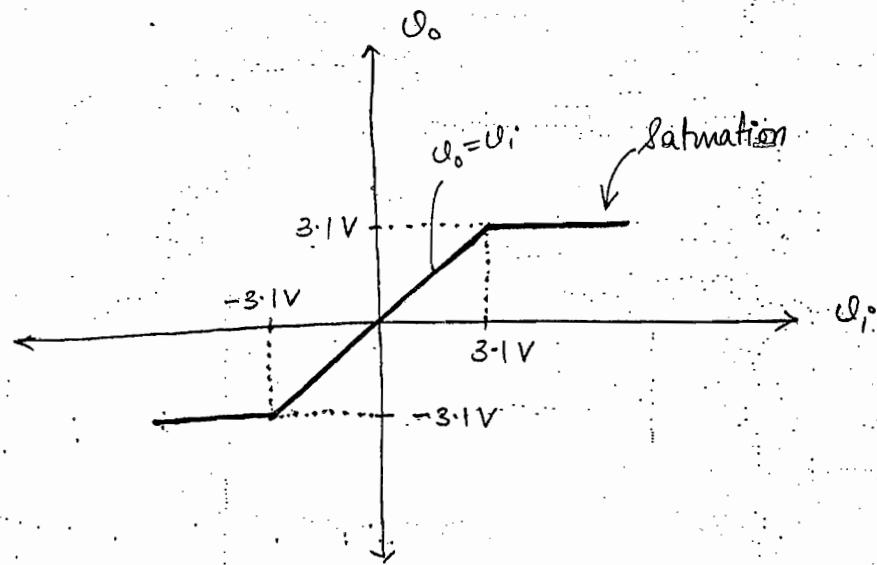
$$I_s/2 = 0.465 \text{ mA}$$

$$\Rightarrow I_s = 0.93 \text{ mA}$$



$$U_o = 0.31 \text{ mA} \times 10 \text{ k} \\ = 3.1 \text{ Volts.}$$

Transfer characteristic:

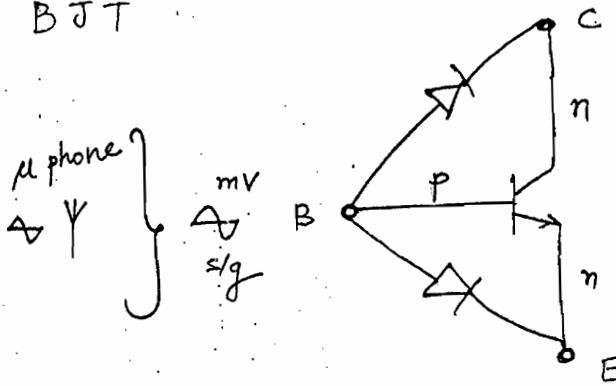


Conclusion:

In Analog gate design, The control voltage should be always greater than 3 times of s/g voltage to make all the diodes ON.

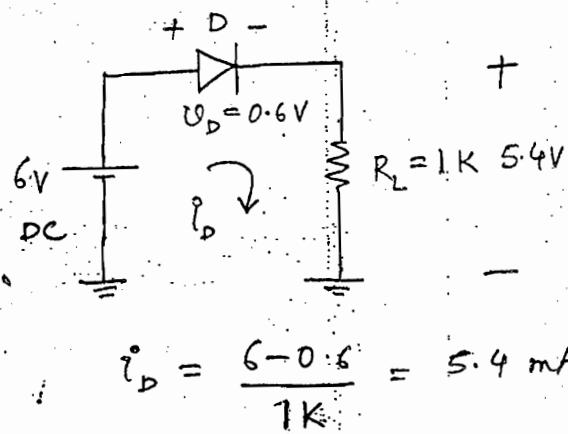
Small s/g Analysis of a Diode:

BJT



Diode Resistance

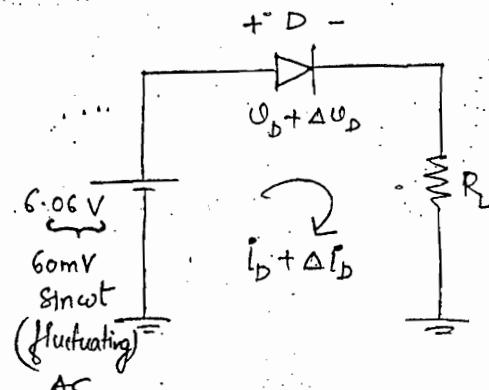
Static Resistance



$$R_D = \frac{V_D}{I_D} = \frac{0.6 \text{ V}}{5.4 \text{ mA}}$$

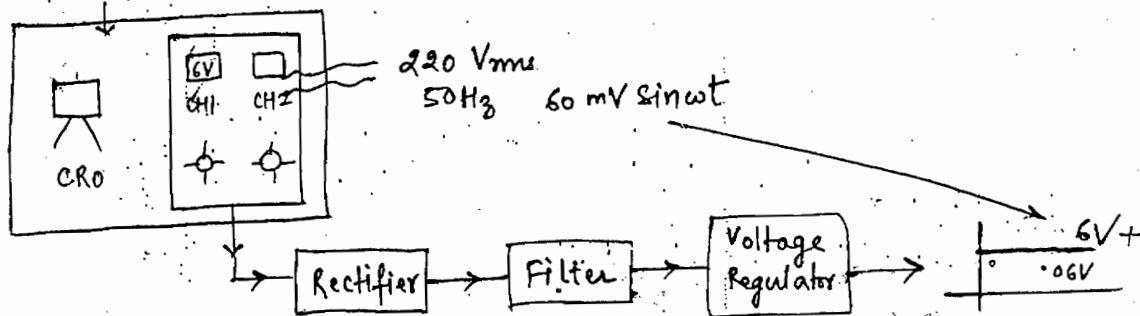
static
DC
Resistance = 111 Ω

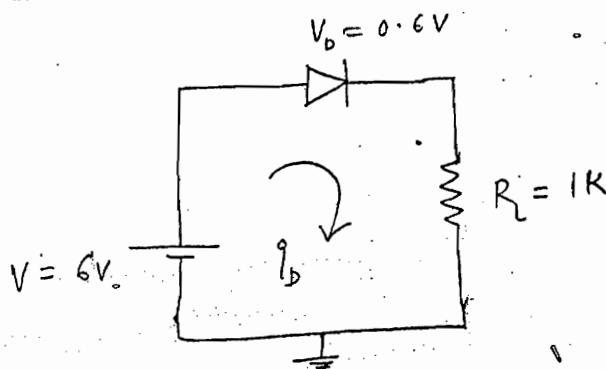
Dynamic Resistance



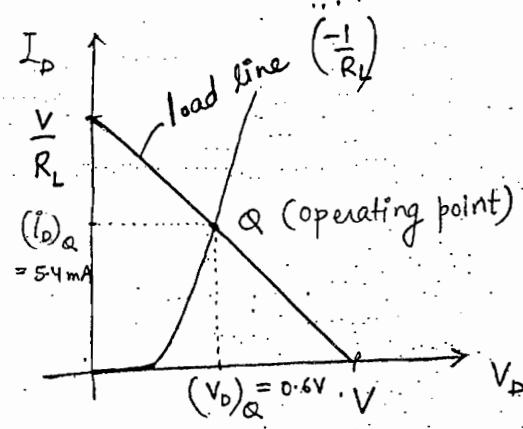
dynamic/Ac Resistance

Regulated Power supply (DC):



Analysis:Static ResistanceMathematical Analysis:

$$\begin{aligned} I_D &= \frac{V - V_D}{R_L} \\ &= \frac{6 - 0.6}{1\text{K}} \\ &= 5.4\text{mA} \end{aligned}$$

Graphical Analysis:

$$\begin{aligned} I_D &= \frac{V - V_D}{R_L} \\ I_D &= -\frac{V_D}{R_L} + \frac{V}{R_L} \\ y &= mx + c \end{aligned}$$

$$m = \text{slope} = -\frac{1}{R_L}$$

$$R_D = \frac{V_D}{I_D} = \frac{0.6}{5.4\text{mA}} = 111\Omega$$

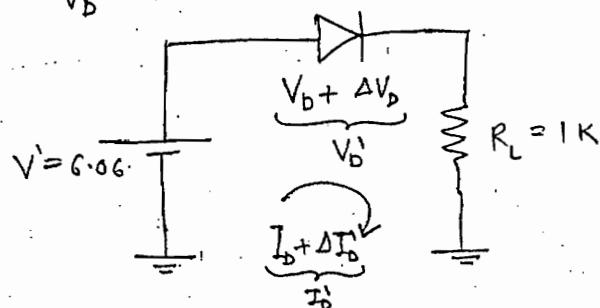
* As I_D increases, R_D decreases.

Design:

Suppose due to a small s/g variation ($V_i = 60\text{mV sin}\omega t$)

the DC voltage $V = 6\text{V}$ is changed to $V' = 6.06\text{V}$.

then find the new diode current I'_D & new diode voltage V'_D



$$\text{Sol. } V' = V + 4V$$

$$6.06V = 6V + 0.06V$$

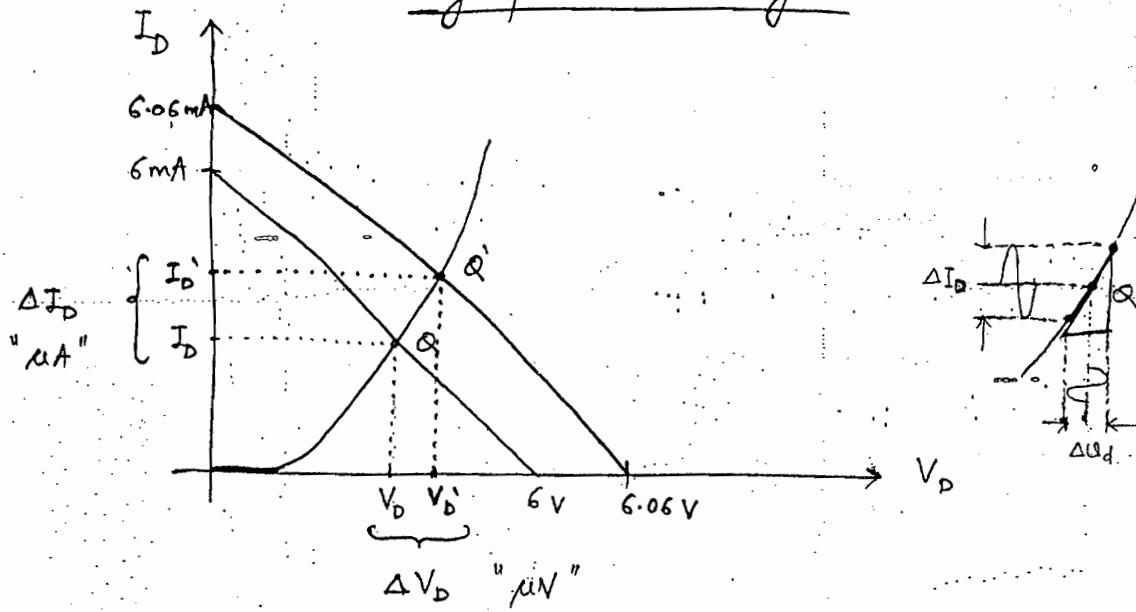
$$I'_D = I_D + \Delta I_D$$

$$? \quad \downarrow \quad ? \\ 5.4mA$$

$$V'_D = V_D + \Delta V_D$$

$$? \quad \downarrow \quad ? \\ 0.6V$$

graphical analysis:



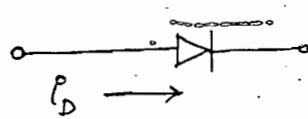
As the change in V_D and I_D are at a level i.e. very small. That's why we do not go for graphical methods.

Note:

For small s/g analysis of a device (Diode, BJT, FET, MOSFET) graphical method is not suitable because the variations are in the order of "μV" and "μA". Mathematical analysis is best suitable for small s/g study.

* Lower the Q-point; (smaller current or lower voltage), the higher is the ac resistance.

* As with the dc and ac res. levels, the lower the level of currents, the higher is the resistance level.



$$I_D = I_0 e^{V_D/V_T}$$

$$I_D + \Delta I_D = I_0 e^{(V_D + \Delta V_D)/V_T}$$

$$= I_0 \cdot e^{V_D/V_T} \cdot e^{\Delta V_D/V_T}$$

$$= I_D \cdot e^{\Delta V_D/V_T}$$

$$\Delta V_D \ll V_T \quad (\text{small s/g analysis})$$

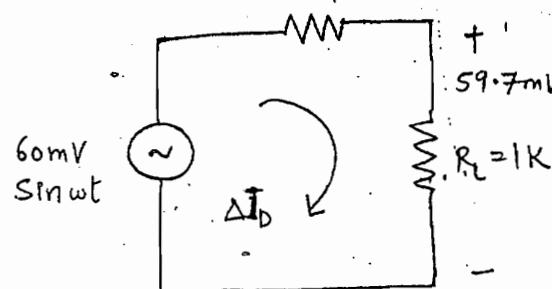
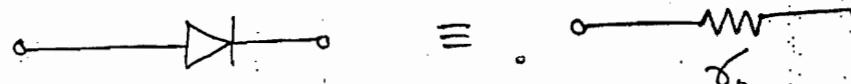
$$I_D + \Delta I_D = I_D \left(1 + \frac{\Delta V_D}{V_T}\right)$$

$$I_D + \Delta I_D = I_D + I_D \cdot \frac{\Delta V_D}{V_T}$$

$$\Delta I_D = I_D \cdot \frac{\Delta V_D}{V_T}$$

$$\gamma_D = \frac{\Delta V_D}{\Delta I_D} = \frac{V_T}{I_D(\text{DC})}$$

Small s/g equivalent model :



$$\gamma_D = \frac{V_I}{I_D}$$

$$= \frac{26mV}{5.4mA}$$

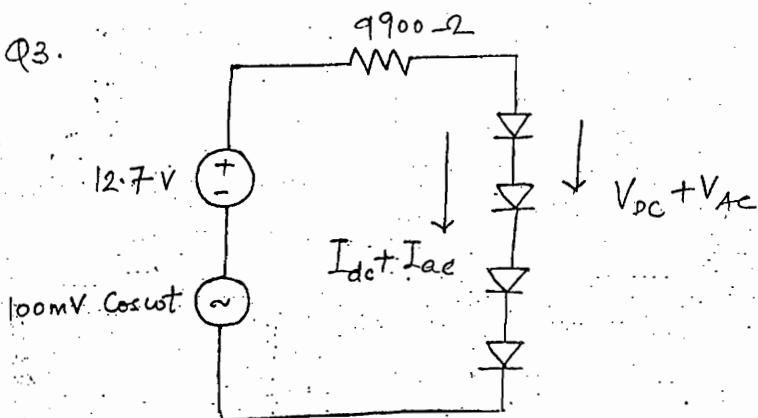
$$= 5\Omega$$

$$\Delta I_D = \frac{\Delta V}{\gamma_D + R_L} = \frac{60mV \sin \omega t}{1k\Omega} = 60 \mu A \sin \omega t$$

$$\begin{aligned}
 \Delta V_D &= 4I_D - \gamma_D \\
 &= 60 \mu\text{A} \sin\omega t \cdot (5 \Omega) \\
 &= 300 \mu\text{V} \cdot \sin\omega t \\
 &= 0.3 \text{ mV} \sin\omega t
 \end{aligned}$$

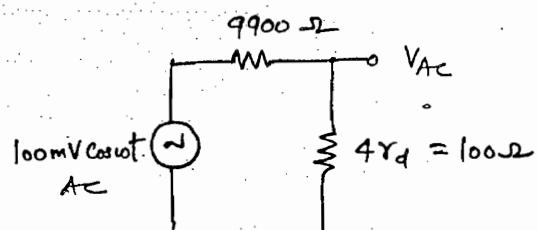
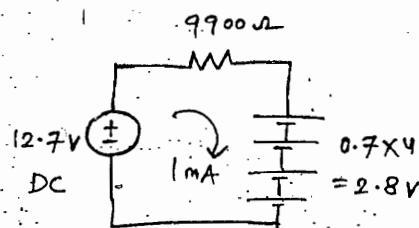
Hence, diode is a battery DC wise whereas AC wise it is a resistor.

Q3.



$$\text{Sol. } I_{dc} = \frac{12.7 - 4 \times 0.7}{9900} = 1 \text{ mA}$$

$$\begin{aligned}
 \gamma_D &= \frac{V_T}{I_{dc}} \\
 &= \frac{25 \text{ mV}}{1 \text{ mA}} \\
 &= 25 \Omega
 \end{aligned}$$



$$\begin{aligned}
 V_{AC} &= 100 \text{ mV} \cos\omega t \cdot \frac{100 \Omega}{(9900 + 100) \Omega} \\
 &= 100 \text{ mV} \cos\omega t \cdot \frac{100}{10,000} \\
 &= 1 \text{ mV} \cos\omega t
 \end{aligned}$$

Function Capacitance:

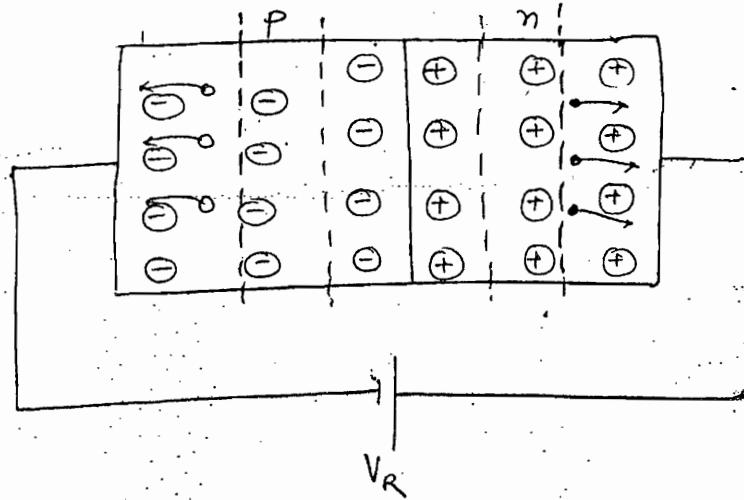
↓
Transition capacitance (C_T)
(RB)

(Transition means depletion
which always occurs in
RB).

Diffusion capacitance (C_D)
(FB)

(Diffusion occurs in FB
because there is no
depletion region).

Transition capacitance (C_T): (or) Depletion Region capacitance.



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"The rate of change of charge w.r.t. to applied voltage" - Capacitance

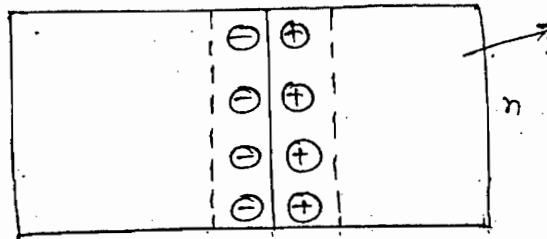
"The rate of change of immobile charge in the depletion region w.r.t. to reverse bias voltage"
- Transition capacitance

$$C_{II} = \frac{A\epsilon}{d}$$

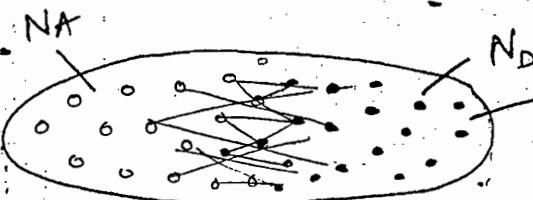
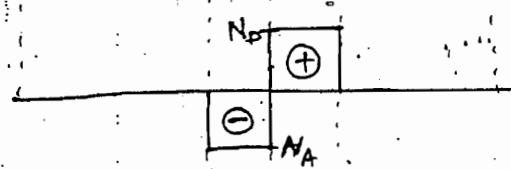
$$C_T = \frac{A\epsilon}{\omega} \propto \frac{1}{\omega}$$

$$\omega \propto \sqrt{V_j} \rightarrow \text{Alloy type} \quad \left. \begin{array}{l} \\ \end{array} \right\} \text{OC jn}$$

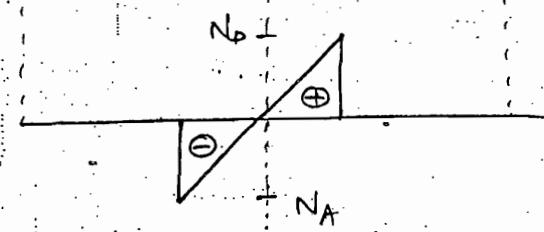
$$\omega \propto \sqrt[3]{V_j} \rightarrow \text{grown jn type}$$



Alloy type (or)
Step graded jxn (or)
Sudden change (or)
Abrupt change



Grown jxn (or)
Linearly graded (or)
Diffused type.



From OC to RB -

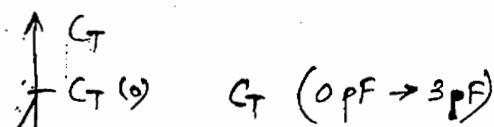
$$\omega \propto \sqrt{V_j + V_R} \quad - \text{alloy type}$$

$$\omega \propto \sqrt[3]{V_j + V_R} \quad - \text{grown jxn type}$$

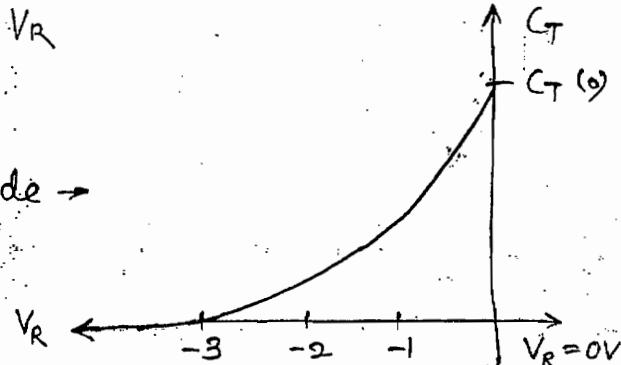
$$C_T \propto \frac{1}{\sqrt{V_R}}$$

(if V_j is neglected, generally V_j is very small)

$$C_T \propto \frac{1}{\sqrt[3]{V_R}}$$



ordinary diode -



Variactor diode

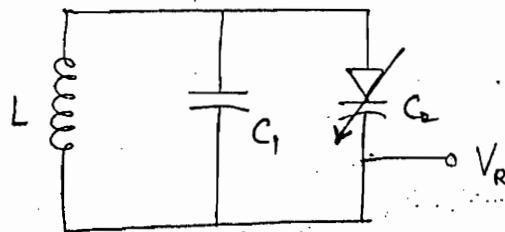
(or)
Varicap



Power diode : $C_T : (5 \text{ pF} \rightarrow 300 \text{ pF})$

Tuning ckt:

VCO → Voltage Controlled Oscillator

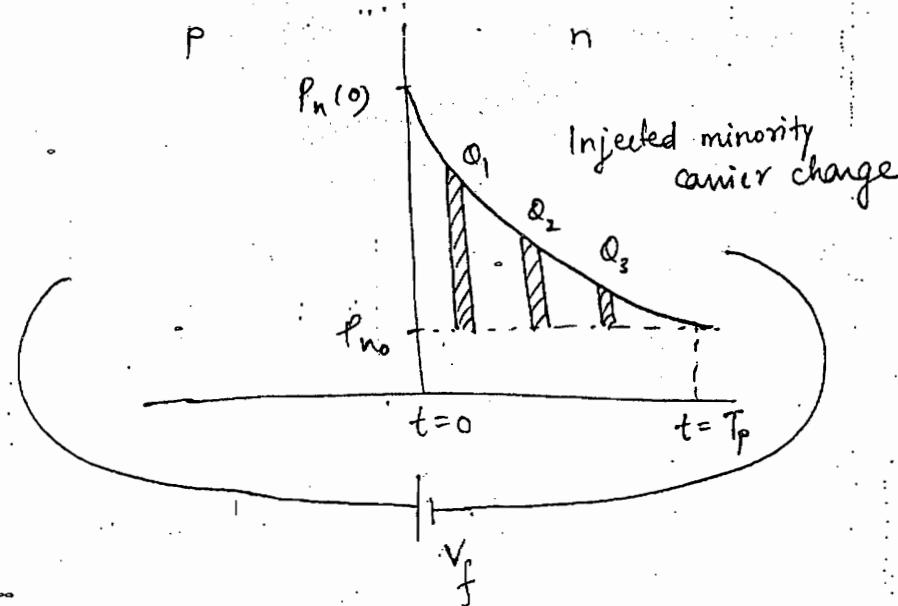


$$f = \frac{1}{2\pi\sqrt{LC_T}}$$

$$\& C_T = C_1 + C_2$$

depends on V_R

Diffusion capacitance (C_D) : (or) storage capacitance



"The rate of change of injected minority carrier charge w.r.t. forward Bias voltage" - C_D

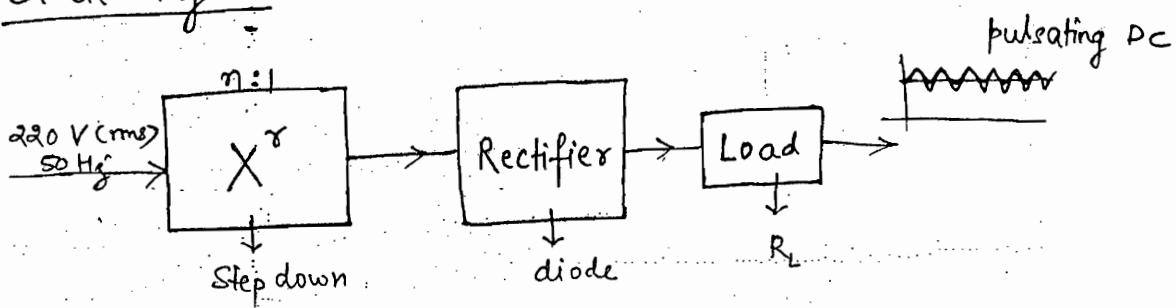
$$Q = I_p \times T_p$$

$$C_D = \frac{dQ}{dV} = T_p \frac{dI_p}{dV} = \frac{T_p I_p}{V_T} = \frac{T_p I_{de}}{V_T}$$

$$C_D \approx 100 \text{ pF}$$

Rectifiers:

Block Diagram:



If is a ckt which converts ac to pulsating dc.

Practical Rectifier designs are:

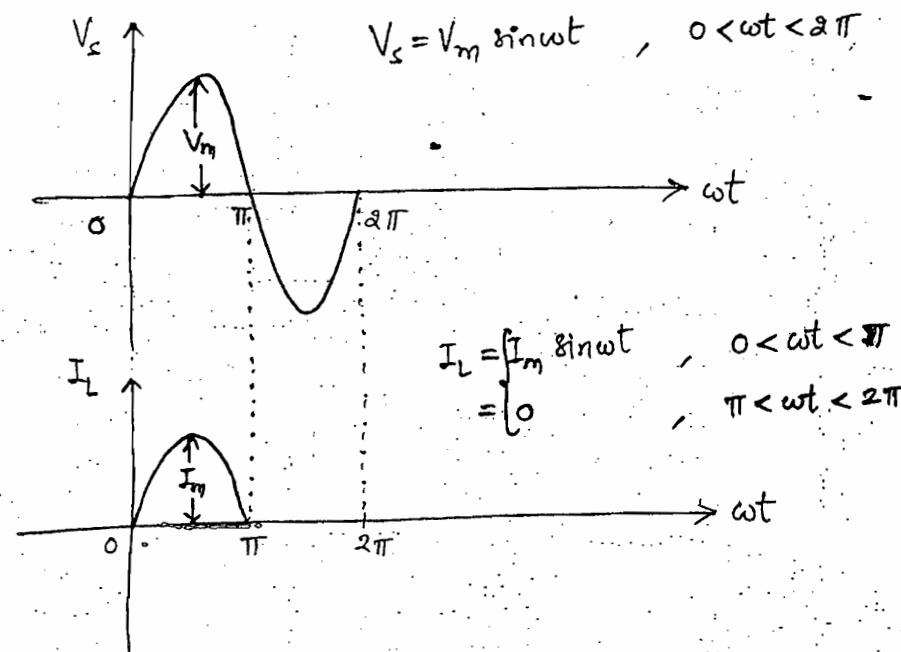
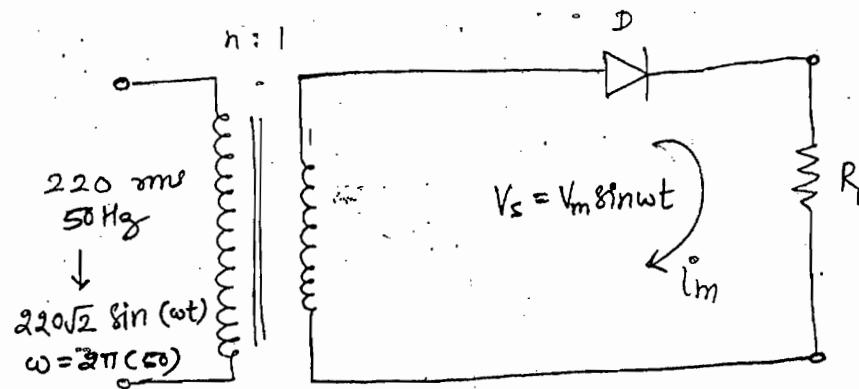
- (1). Half Wave Rectifier
- (2). Full Wave Rectifier with centre tap
- (3). Bridge Rectifier

Rectifier parameters:

- (1). Average current (I_{de})
- (2). DC o/p voltage (V_{de})
- (3). RMS load current (I_{rms})
- (4). Ripple factor
- (5). Voltage regulation
- (6). Rectification Efficiency (η)
- (7). Transformer utilization factor (TOF).
- (8). Peak Inverse Voltage (PIV)
- (9). Form factor

(1). Half Wave Rectifier : (HWR)

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(1). I_{dc} :

$$I_{dc} = \frac{\text{area under the curve}}{2\pi}$$

$$\therefore I_{dc} = \frac{1}{2\pi} \int_0^{2\pi} I_L \cdot d\omega \quad (\omega = \omega t)$$

$$= \frac{1}{2\pi} \int_0^{\pi} I_L \cdot d\omega + 0$$

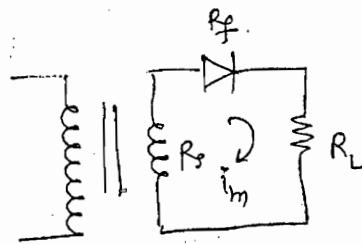
$$I_{dc} = \frac{I_m}{\pi}$$

(2). V_{dc} :

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

$$= \frac{I_m}{\pi} \cdot R_L$$

$$V_{dc} = \frac{V_m \cdot R_L}{\pi (R_f + R_s + R_L)}$$



$$V_{dc} = \frac{V_m}{\pi} \left(1 + \frac{R_s + R_f}{R_L} \right)$$

$$V_{dc} \Big|_{R_L \rightarrow \infty} = \frac{V_m}{\pi} \quad (\text{no load condition})$$

(3). I_{rms} :

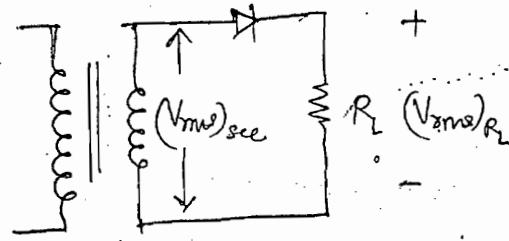
$$I_{rms} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} I^2 d\alpha}$$

$$= \sqrt{\frac{I_m^2}{2\pi} \int_0^{2\pi} \sin^2 \alpha \cdot d\alpha}$$

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

$$(V_{rms})_R = V_m / 2$$

$$(V_{rms})_{sec} = V_m / \sqrt{2}$$



(4). Ripple factor:

$$\%f = \frac{\text{rms value of alternating component}}{\text{average value}}$$

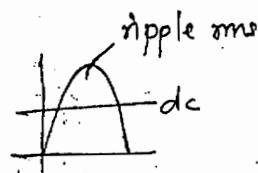
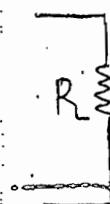
$I' \rightarrow$ ac current

$I_{dc} \rightarrow$ dc current

$I \rightarrow$ total current

$I'_{rms} \rightarrow$ rms value of ac current

$$\%f = \frac{I'_{rms}}{I_{dc}}$$



$$I_{rms} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} I^2 dx}$$

$$I = I_{dc} + I'$$

$$I' = I - I_{dc}$$

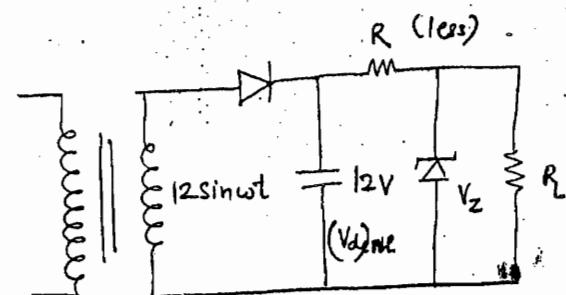
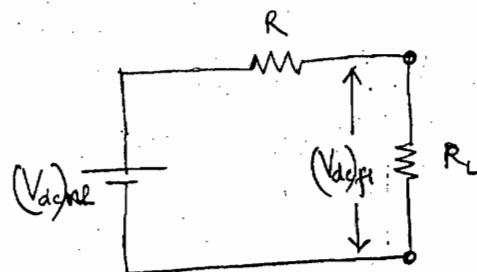
$$I'^2 = \frac{1}{2\pi} \int_0^{2\pi} (I - I_{dc})^2 dx$$

$$= \underbrace{\frac{1}{2\pi} \int_0^{2\pi} I^2 dx}_{I_{rms}^2} - \underbrace{2I_{dc} \int_0^{2\pi} I dx}_{I_{dc}} + \underbrace{\frac{I_{dc}^2}{2\pi} \int_0^{2\pi} dx}_{1}$$

$$\alpha_f = \frac{I_{rms}}{I_{dc}} = \sqrt{\frac{I_{rms}^2 - I_{dc}^2}{I_{dc}}} = \sqrt{\left(\frac{I_{rms}}{I_{dc}}\right)^2 - 1}$$

$$\begin{aligned}\alpha_f &= \sqrt{\left(\frac{Im/2}{Im/\pi}\right)^2 - 1} \\ &= \sqrt{\left(\frac{II}{2}\right)^2 - 1} \\ &= 1.21 \quad (\text{or}) \quad 121\%\end{aligned}$$

(5). Voltage Regulation:



$$\% VR^n = \frac{(V_{de})_{ne} - (V_{de})_{fe}}{(V_{de})_{fe}}$$

* Ideally $\% VR^n$ should be zero.

$$I_{dc} = \frac{I_m}{\pi}$$

$$I_{dc} = \frac{V_m}{\pi (R_f + R_s + R_L)}$$

$$I_{dc} (R_f + R_s) + I_{dc} R_L = \frac{V_m}{\pi}$$

↓ ↓
 $(V_{dc})_{fl}$ $(V_{dc})_{nl}$

$$\% VR^n = \frac{I_{dc} (R_f + R_s)}{\frac{V_m}{\pi} - I_{dc} (R_f + R_s)}$$

(6). Rectification Efficiency (η) :

$\eta = \frac{\text{DC power delivered to load}}{\text{AC I/p power}}$

$$= \frac{P_{dc}}{P_{AC}} = \frac{I_{dc}^2 \cdot R_L}{I_{m\text{av}}^2 \cdot (R_f + R_s + R_L)}$$

$$= \frac{I_{dc}^2}{I_{m\text{av}}^2} \cdot \frac{1}{\left(1 + \frac{R_f + R_s}{R_L}\right)}$$

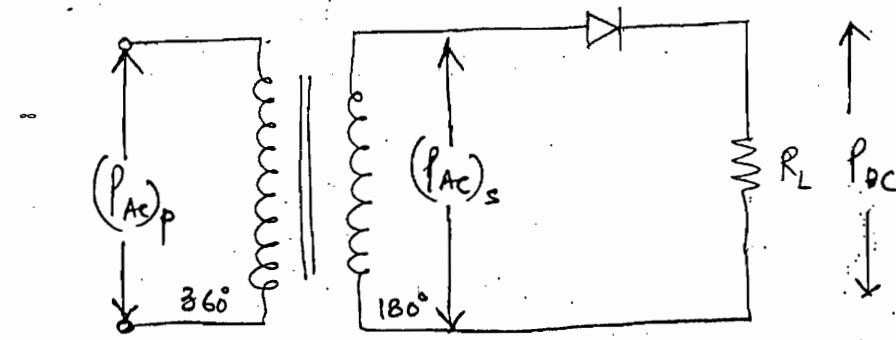
If $R_L \gg (R_f + R_s)$

$$\eta \approx \frac{I_{dc}^2}{I_{m\text{av}}^2}$$

$$= \frac{\left(\frac{I_m}{\pi}\right)^2}{\left(\frac{I_m}{2}\right)^2}$$

$$= \frac{4}{\pi^2}$$

$$= 40.6\% \quad (\text{max})$$

(7). TUF:

$$TUF = \frac{(TUF)_p + (TUF)_s}{2} \quad (X)$$

$$(TUF)_p = \frac{P_{DC}}{(P_{AC})_p} = \eta = 40.6\%$$

$$(TUF)_s = \frac{P_{DC}}{(P_{AC})_s} = \frac{I_{dc}^2 \cdot R_L}{(V_{rms} \cdot I_{rms})_{see}}$$

$$= \frac{I_{dc}^2 \cdot R_L}{\frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{2}}$$

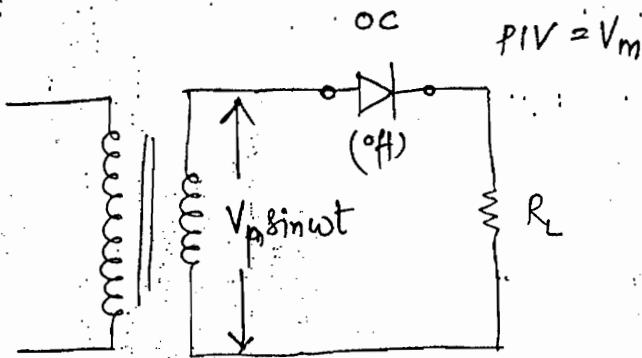
$$= \frac{I_{dc}^2 \cdot R_L}{\frac{I_m^2}{2\sqrt{2}} (R_f + R_s + R_L)}$$

$$= \frac{I_{dc}^2}{\frac{I_m^2}{2\sqrt{2}} \left(1 + \frac{R_f + R_s}{R_L} \right)}$$

If $R_L \gg (R_f + R_s)$

$$(TUF)_s = \frac{\left(\frac{I_m}{\pi}\right)^2}{\frac{I_m^2}{2\sqrt{2}}} = \frac{2\sqrt{2}}{\pi^2} = 28\%$$

$$\therefore TUF = (TUF)_s = 28\%$$

(8). PIV:

$$PIV = V_m$$

(9). form factor (FF):

$$FF = \frac{\text{rms value}}{\text{avg. value}} = \frac{I_{\text{rms}}}{I_{\text{dc}}}$$

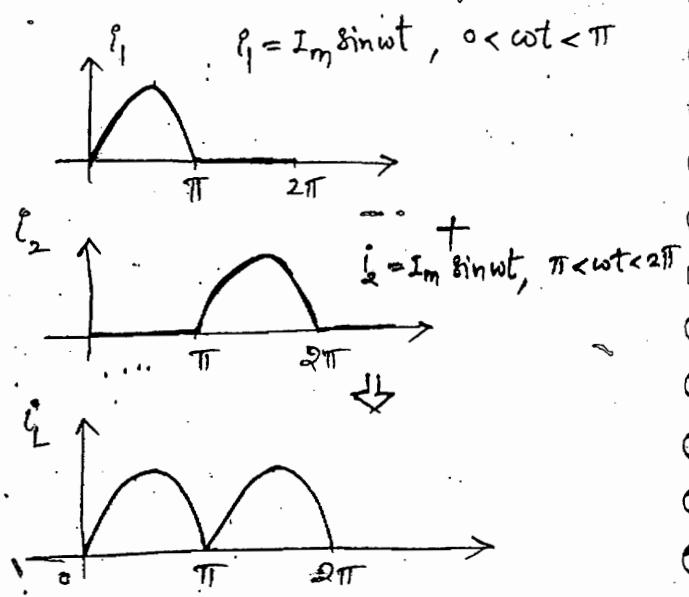
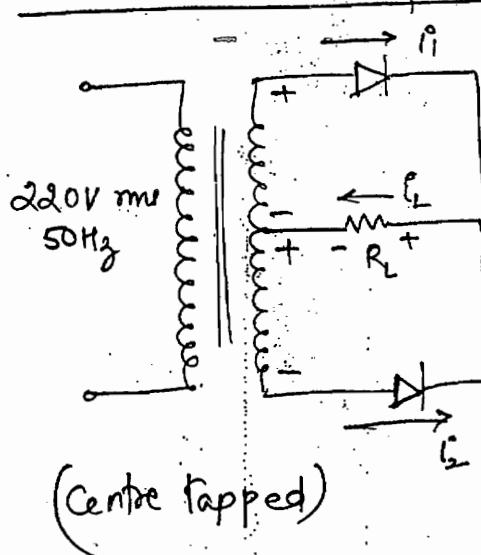
$$= \frac{I_m/2}{I_m/\pi}$$

$$= \pi/2 = 1.57$$

(10). peak factor (PF):

$$\text{peak factor} = \frac{\text{peak value}}{\text{rms value}}$$

$$= \frac{I_m}{I_m/2} = 2$$

(2). Full Wave Rectifier:

Rectifier parameters

$$(1): I_{dc} = \frac{2 I_m}{\pi}$$

$$(2): (V_{dc})_{re} = \frac{2 V_m}{\pi}$$

$$(3): I_{rms} = \frac{I_m}{\sqrt{2}}$$

$$\begin{aligned}(4): \gamma_f &= \sqrt{\left(\frac{I_{rms}}{I_{dc}}\right)^2 - 1} \\ &= \sqrt{\frac{\left(\frac{I_m}{\sqrt{2}}\right)^2}{\left(\frac{2 I_m}{\pi}\right)^2} - 1} \\ &= \sqrt{\frac{\pi^2}{8} - 1} = 0.48 \\ &= 48\%\end{aligned}$$

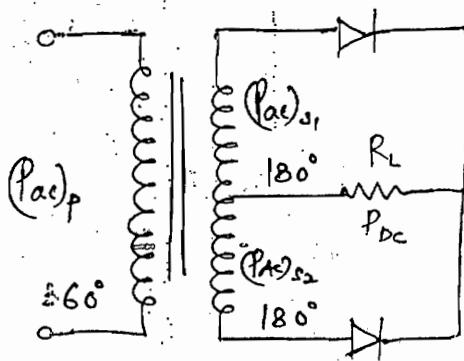
$$(5): I_{dc} (r_f + R_s) + I_{dc} \cdot R_L = V_m / \pi$$

$$\downarrow \quad \downarrow \quad \downarrow \quad \downarrow$$

$$\frac{2 I_m}{\pi} \quad R_s / \pi \quad \frac{2 I_m}{\pi} \quad \frac{2 V_m}{\pi}$$

$$VR^n =$$

$$\begin{aligned}(6): \eta &= \frac{\rho_{DC}}{\rho_{AC}} \\ &= \frac{I_{dc}^2}{I_{rms}^2} \\ &= \frac{\left(\frac{2 I_m}{\pi}\right)^2}{\left(\frac{I_m}{\sqrt{2}}\right)^2} = \frac{8}{\pi^2} \\ &= 81.2\%\end{aligned}$$

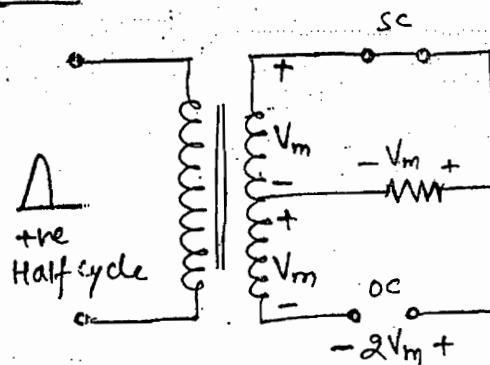
(7). TOF:

$$(TOF)_p = \frac{P_{DC}}{(P_{AC})_p} = \eta = 81.2\%$$

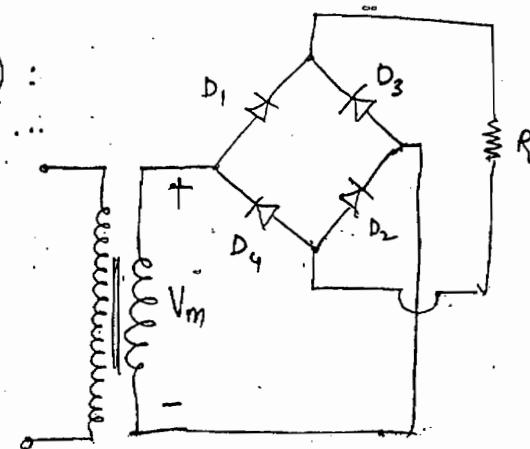
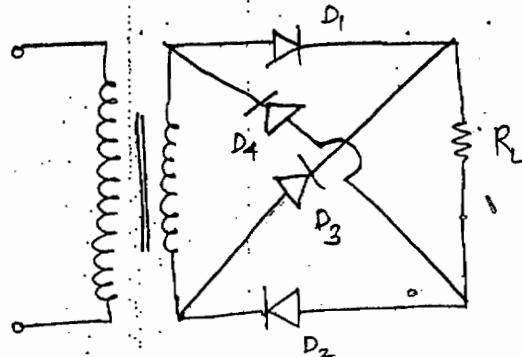
$$(TOF)_s = \frac{P_{DC}}{(P_{AC})_{\text{tot}}} = 57.3\%$$

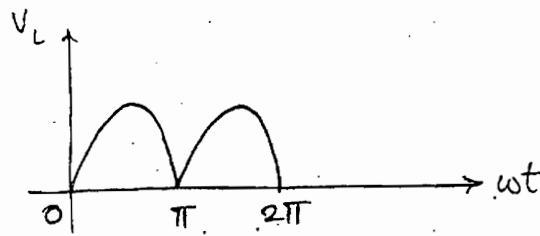
$$TOF = \frac{81.2 + 57.3}{2} = 69.3\% \quad (\text{disadvantage})$$

* Avg can be taken here bcoz 180° each get added and mean tak be taken.

(8) PIV:

$$PIV = 2V_m \quad (\text{disadvantage})$$

(3). full Wave Rectifier (Bridge Type):



Common parameters as full wave Rectifier (centre tapped)

$$(1) I_{dc} = \frac{2I_m}{\pi}$$

$$(2) (V_{dc})_{ne} = \frac{2V_m}{\pi}$$

$$(3) I_{rms} = \frac{I_m}{\sqrt{2}} \quad \& \quad V_{rms} = \frac{V_m}{\sqrt{2}}$$

$$(4) \eta_f = 48\%$$

$$(5) \eta = 81.2\%$$

Other parameters :

$$(6) I_{dc} (R_f + \frac{R_s}{2}) + I_{dc} R_L = \frac{2V_m}{\pi}$$

\downarrow

$2R_f \quad R_s$

$$I_{dc} (2R_f + R_s) + I_{dc} R_L = \frac{2V_m}{\pi}$$

$$VR^n =$$

(7) TUF :

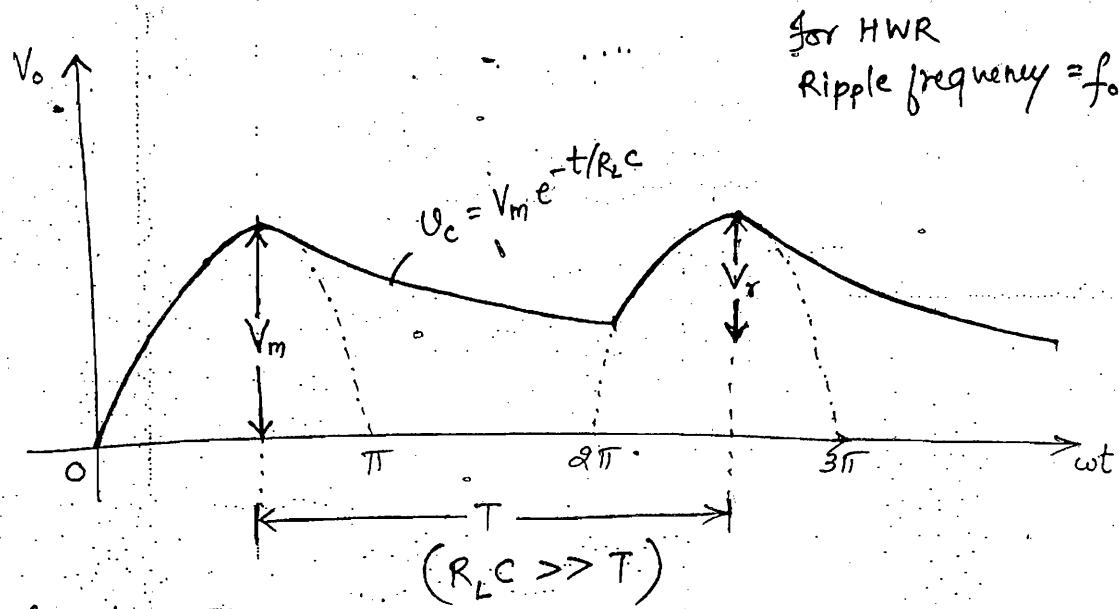
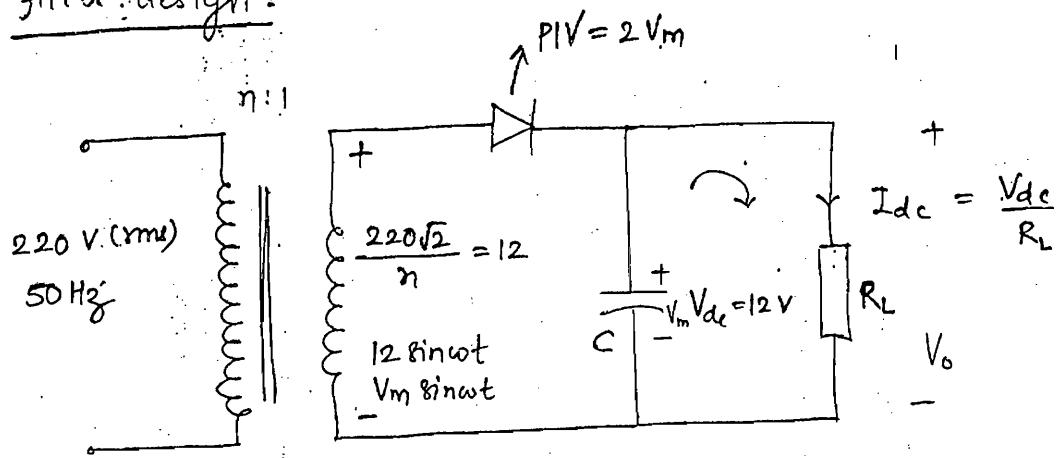
$$(TUF)_p = \frac{P_{DC}}{(P_{AC})_p} = \eta = 81.2\% \quad (360^\circ - 360^\circ)$$

$$(TUF)_s = \frac{P_{DC}}{(P_{AC})_s} = \frac{P_{DC}}{(P_{AC})_p} = \eta = 81.2\%$$

$$\therefore TUF = 81.2\%$$

$$(8) PIV = V_m \quad (\text{advantage})$$

filter design:



$$f = \frac{1}{T} = 50$$

$$T = \frac{1}{f} = \frac{L}{50} = 20 \text{ ms}$$

$$V_c = V_m e^{-t/R_L C}$$

if $R_L C \gg T$

$$V_c = V_m \left(1 - \frac{t}{R_L C}\right)$$

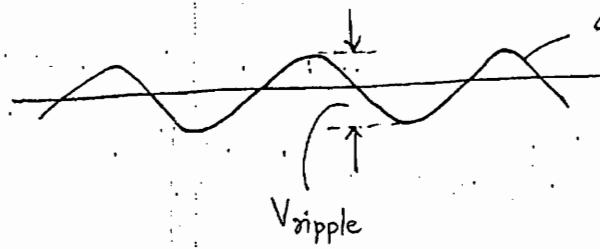
$$V_c = V_m - \frac{V_m T}{R_L C}$$

\downarrow

V_{dc} V_{ripple}

$$\downarrow V_{ripple} = \frac{V_m T}{(R_L C)}$$

$$\text{angular wave } (V_{\text{rms}} = \frac{V_m}{\sqrt{3}})$$



$$\gamma_f = \frac{V_{\text{ripple (rms)}}}{V_{\text{dc}}}$$

$$= \left(\frac{V_m T}{R_L C} \right) \cdot \frac{1}{\sqrt{3}} \frac{1}{V_{\text{dc}}}$$

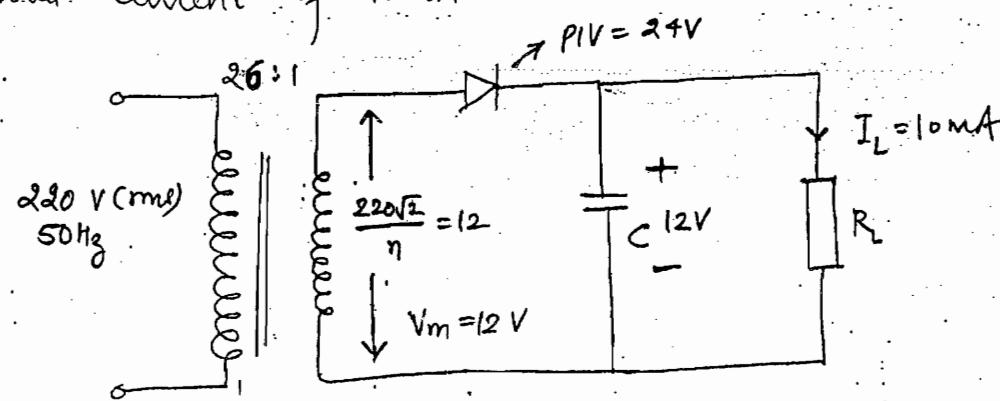
$$\gamma_f = \frac{T}{\sqrt{3} R_L C} = \frac{T}{\sqrt{3} C}$$

$$\text{if } R_L = \infty, \gamma_f = 0 \text{ & } V_c = V_m$$

Problem:

Design a DC power supply of 12V where the γ_f should not exceed 5% for a 220V (rms), 50Hz supply if the load current of 10mA.

Sol.



(1). Turns ratio :

$$\frac{220\sqrt{2}}{\eta} = 12 \Rightarrow \eta = 26$$

(2). R_L :

$$R_L = \frac{V_{\text{dc}}}{I_L} = \frac{12V}{10\text{mA}} = 1.2\text{ k}\Omega$$

(3). δf :

$$\delta f = \frac{V_{\text{ripple}} (\text{rms})}{V_{\text{dc}}}$$

$$\frac{5}{100} = \frac{V_{\text{ripple}}}{12\sqrt{3}}$$

$$V_{\text{ripple}} = \frac{60\sqrt{3}}{100} \approx 1 \text{ volt}$$

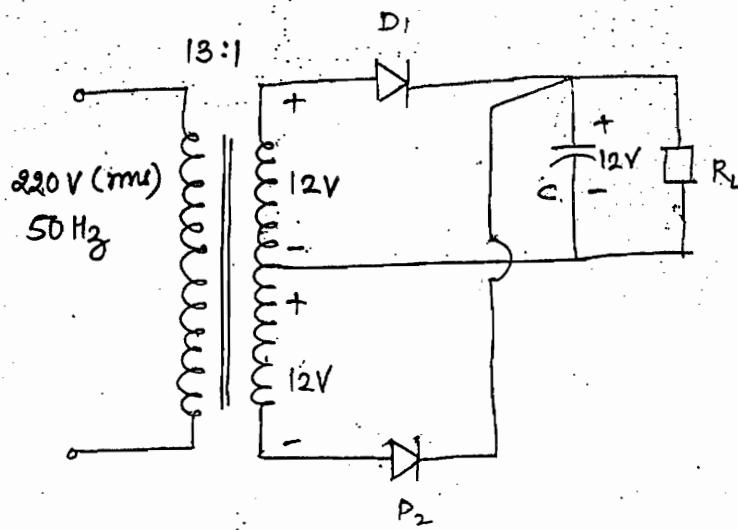
(4). V_{ripple} :

$$V_{\text{ripple}} = \frac{V_m T}{R_L C}$$

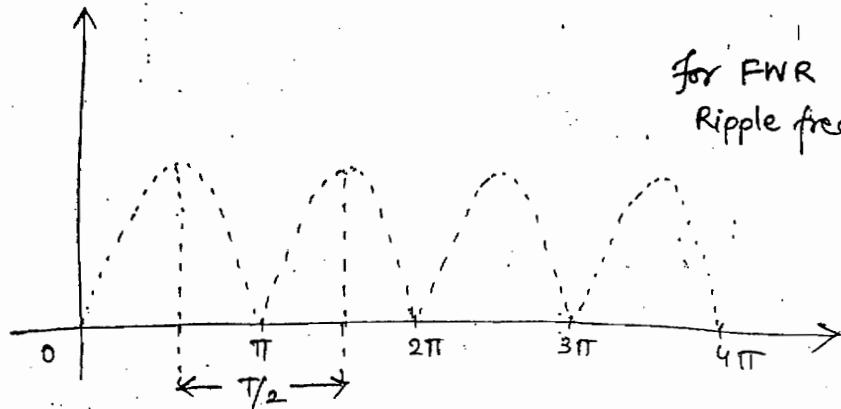
$$I = \frac{I_L T}{C}$$

$$C = 10 \text{ mA} \times 20 \text{ ms}$$

$$\therefore C = 200 \mu\text{F}$$

FWR with filter (capacitor):

for FWR
Ripple frequency = $2f$.



$$(1) \frac{\alpha 20\sqrt{2}}{n} = 24 \Rightarrow n = 13$$

$$(2) R_L = \frac{V_{de}}{I_L} = \frac{12V}{10mA} = 1.2K\Omega$$

$$(3) r_f = \frac{V_{ripple} (\text{rms})}{V_{de}}$$

$$\frac{5}{100} = \frac{V_{ripple}}{12\sqrt{3}} \Rightarrow V_{ripple} \approx 1 \text{ volt}$$

$$(4) V_{ripple} = \frac{V_m}{R_L C} \cdot \left(\frac{I}{2}\right)$$

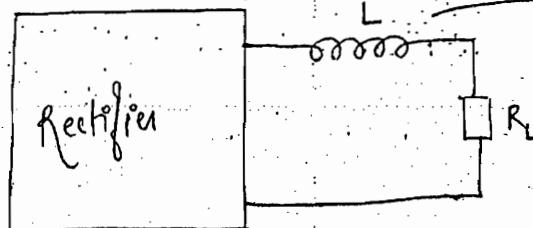
$$I = \frac{I}{C} \cdot \left(\frac{I}{2}\right)$$

$$C = 10mA \times 10ms$$

$$\therefore C = 100 \mu F$$

Conclusion:

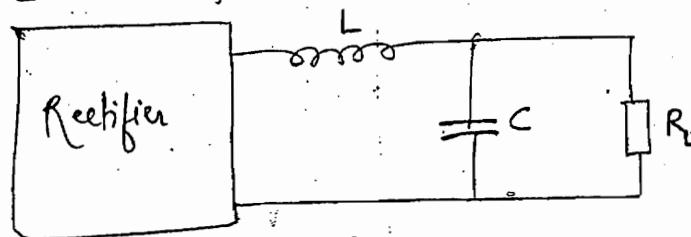
(Inductor filter):



\rightarrow size↑, $r_f \propto R_L$

- * Heavy load techniques are not possible with inductor.
- * Good for No load.

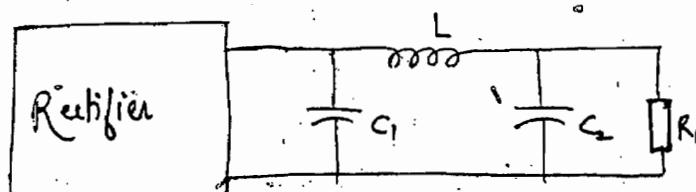
L section filter:



$$r_f = \frac{\sqrt{2}}{3} \cdot \frac{X_C}{X_L}$$

* for variable load

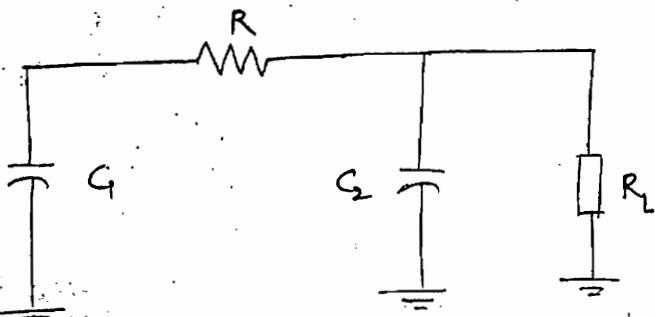
II Section filter:



$$r_f = \frac{\sqrt{2} \cdot X_C \cdot X_C}{R_L \cdot X_L}$$

* for low loads.

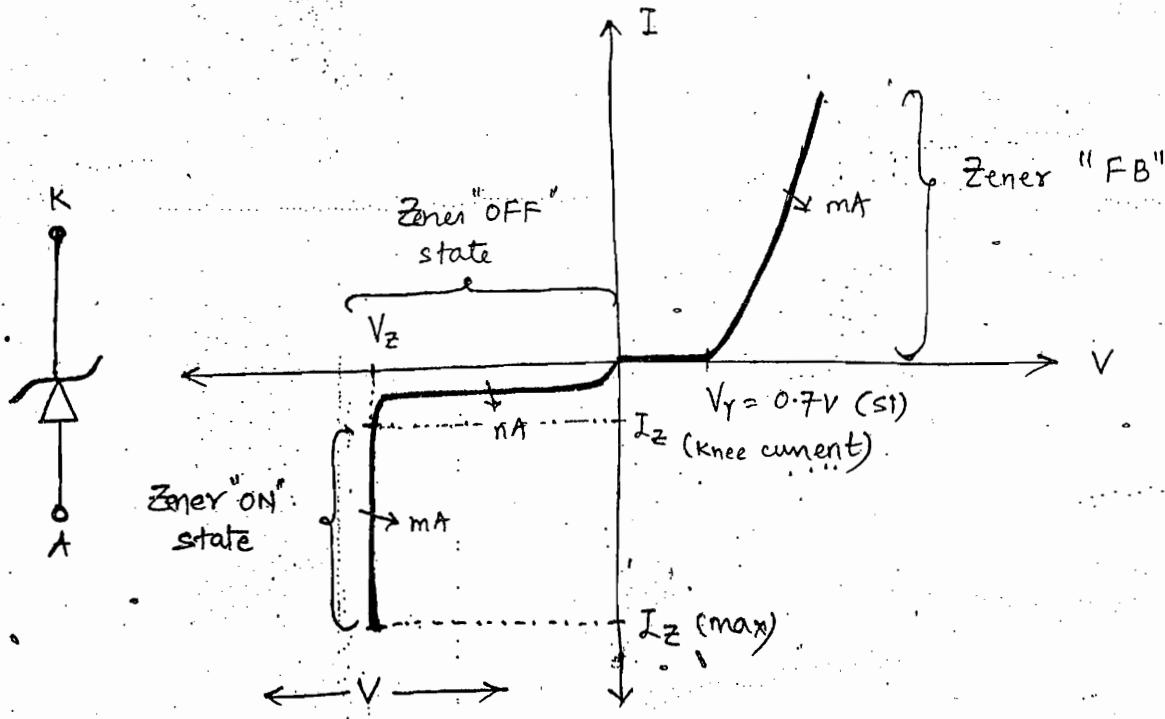
RC filter:



$$r_f = \frac{\sqrt{2} X_{C_1} X_{C_2}}{R_L \cdot R}$$

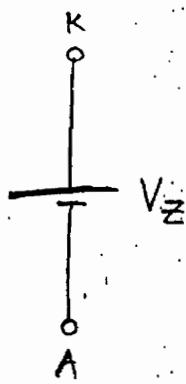
* for low loads

Zener Voltage Regulator -



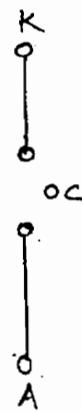
$V > V_z$

"ON" state

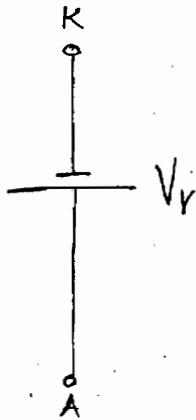


$0 < V < V_z$

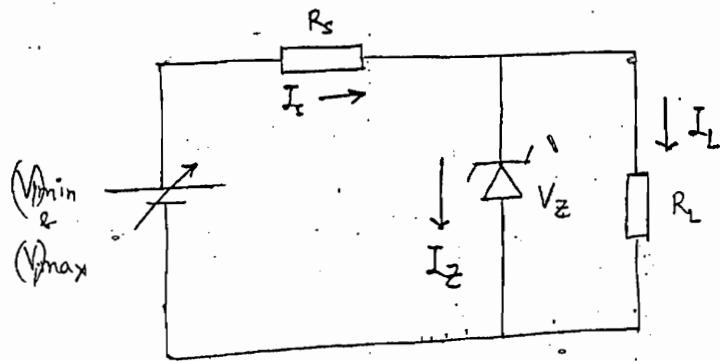
"OFF" state



$V \rightarrow FB$



(i). Zener diodes will be manufactured by Si material whereas tunnel diode will be manufactured by Ge or GaAs.



$$(I_s)_{\min} = \frac{(V_i)_{\min} - V_z}{R_s}$$

$$(I_s)_{\max} = \frac{(V_i)_{\max} - V_z}{R_s}$$

$$(I_z)_{\min} = (I_s)_{\min} - I_L \quad \left. \begin{array}{l} \\ \end{array} \right\} \quad V_i \rightarrow \text{variable}$$

$$(I_z)_{\max} = (I_s)_{\max} - I_L \quad \left. \begin{array}{l} \\ \end{array} \right\} \quad R_L \rightarrow \text{fixed}$$

$$(I_z)_{\min} = (I_s)_{\min} - (I_L)_{\max} \quad \left. \begin{array}{l} \\ \end{array} \right\} \quad V_i \rightarrow \text{variable}$$

$$(I_z)_{\max} = (I_s)_{\max} - (I_L)_{\min} \quad \left. \begin{array}{l} \\ \end{array} \right\} \quad R_L \rightarrow \text{variable}$$

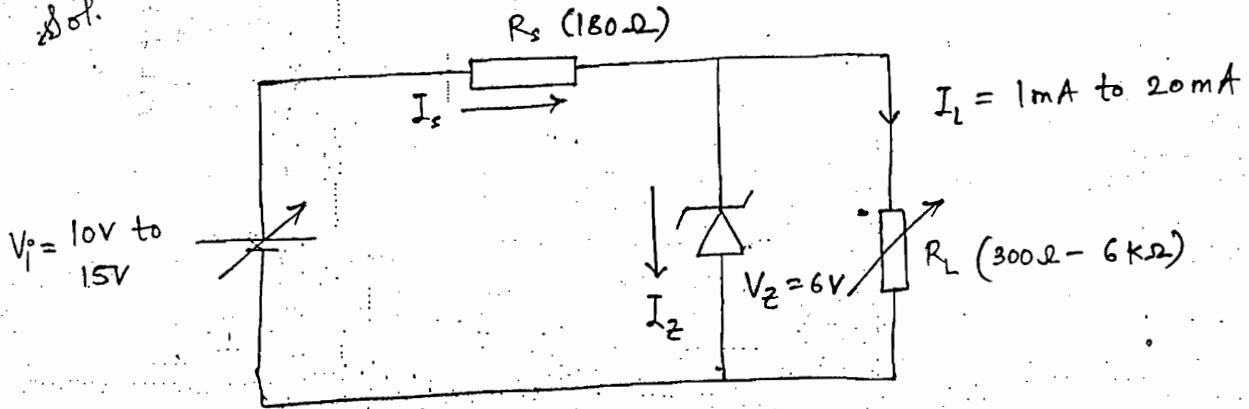
$$(I_z)_{\min} = I_s - (I_L)_{\max} \quad \left. \begin{array}{l} \\ \end{array} \right\} \quad V_i \rightarrow \text{fixed}$$

$$(I_z)_{\max} = I_s - (I_L)_{\min} \quad \left. \begin{array}{l} \\ \end{array} \right\} \quad R_L \rightarrow \text{variable}$$

Problem :

Design a DC regulated power supply of 6V across the load with variable load current of 1mA to 20mA. Zener diode specifications are $(I_z)_{\min} = 0.1 \text{ mA}$ & $(I_z)_{\max} = 50 \text{ mA}$ with o/p unregulated DC voltage of 10V to 15V.

Sol.



$$I_s = \frac{V_i - V_z}{R_s}$$

$$(I_s)_{\min} = \frac{(V_i)_{\min} - V_z}{(R_s)_{\max}}$$

$$(I_s)_{\min} - (I_L)_{\max} > (I_z)_{\min}$$

$$\frac{(V_i)_{\min} - V_z}{(R_s)_{\max}} - (I_L)_{\max} > (I_z)_{\min}$$

$$\frac{10 - 6}{(R_s)_{\max}} - 20 \text{ mA} > 0.1 \text{ mA}$$

$$\frac{4}{(R_s)_{\max}} > 20.1 \text{ mA}$$

$$\therefore (R_s)_{\max} < 199 \Omega$$

Now, similarly, $(I_s)_{\max} - (I_L)_{\min} < (I_z)_{\max}$

$$\frac{(V_i)^{\max} - V_z}{(R_s)_{\min}} - (I_L)_{\min} < (I_Z)_{\max}$$

$$\frac{15-6}{(R_s)_{\min}} - 1 \text{ mA} < 50 \text{ mA}$$

GS

$$(R_s)_{\min} > \frac{9}{51 \text{ mA}}$$

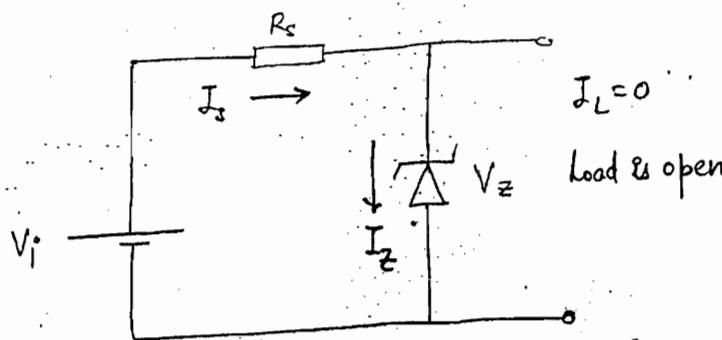
$$\therefore (R_s)_{\min} > 176 \Omega$$

$$\therefore R_s \in (176, 199) \Omega$$

$$176 < R_s < 199$$

Hence, we choose $R_s = 180 \Omega$. (Better)

Small s/g Analysis of zener diode :



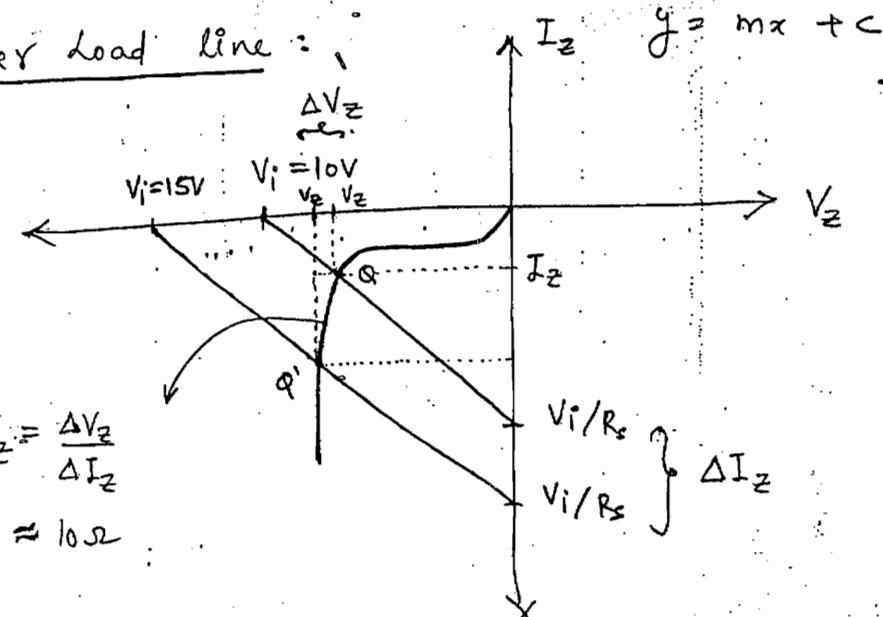
$$V_i^o = I_s R_s + V_z$$

$$I_s = I_z$$

$$\therefore V_i^o = I_z R_s + V_z$$

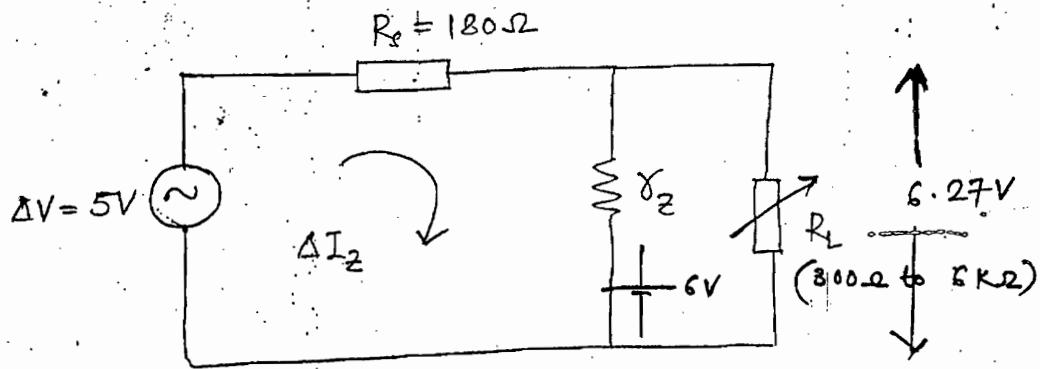
$$I_z = \left(-\frac{1}{R_s}\right) V_z + \frac{V_i^o}{R_s}$$

Zener Load line :



$$\gamma_z = \frac{\Delta V_z}{\Delta I_z}$$

$$\approx 10 \Omega$$



$$\Delta I_Z \approx \frac{\Delta V}{R_s + Z_2} \quad (Z_2 \ll R_L)$$

$$\Delta I_Z \approx \frac{\Delta V}{R_s}$$

$$\text{And, } \Delta V_Z = \Delta I_Z \times Z_2$$

$$= \frac{\Delta V}{R_s} \cdot Z_2$$

$$= \frac{5}{180} \times 10$$

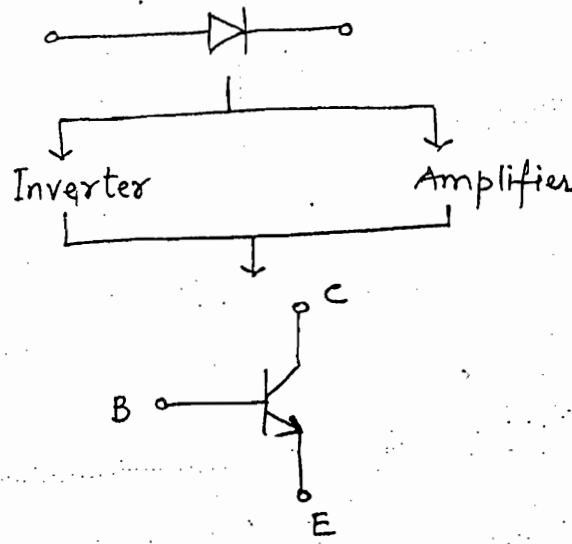
$$\Rightarrow \frac{5}{18} = 0.27V$$

BJT

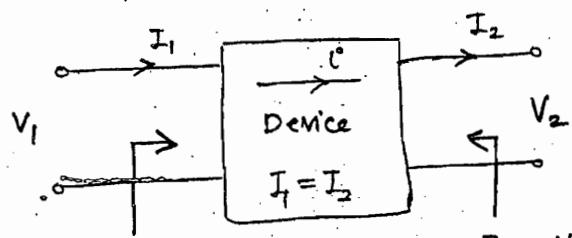
(1) Problems with diode:

Diode cannot be used as an inverter

Diode cannot be used as an amplifier.



Two port N/W:



$$Z_i = \frac{V_1}{I_{FB}} \quad (1 \text{ K}\Omega)$$

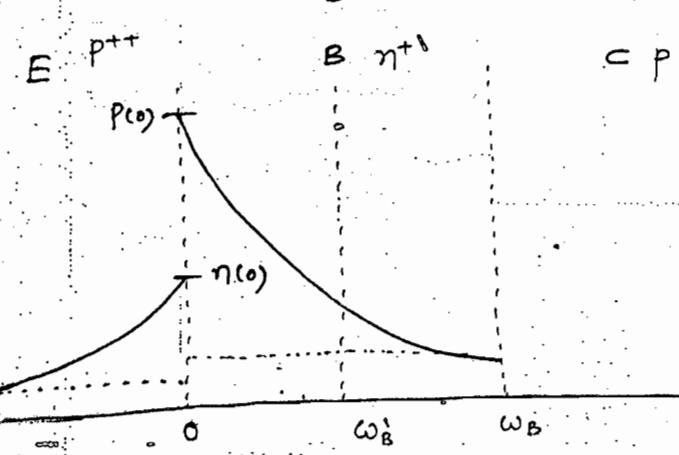
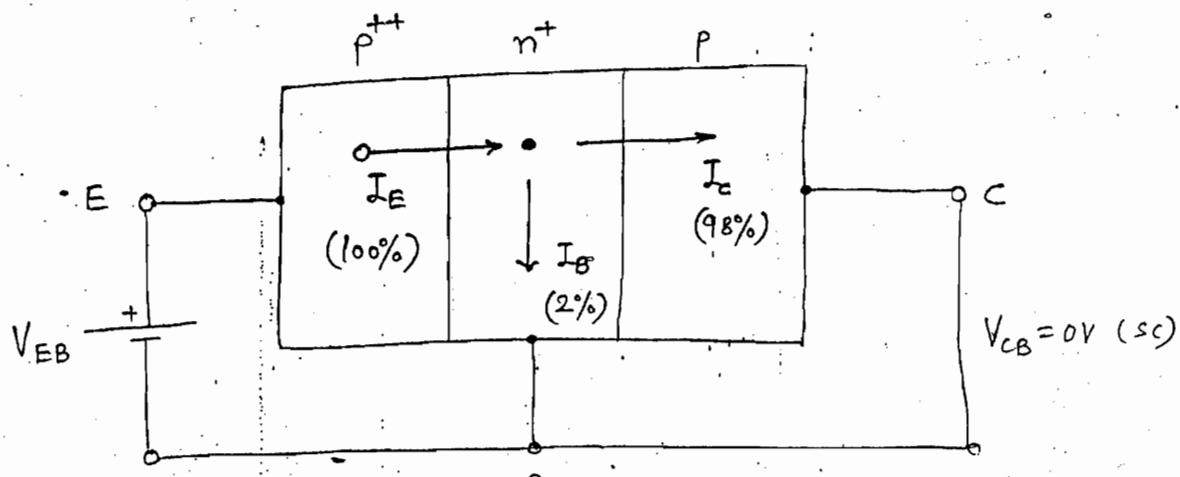
$$Z_o = \frac{V_2}{I_2} \quad (1 \text{ M}\Omega)$$

$$\begin{aligned} P_v &= \frac{P_o}{P_i} = \frac{I_2^2 Z_o}{I_{FB}^2 Z_i} = \frac{Z_o}{Z_i} \\ &= \frac{1 \text{ M}\Omega}{1 \text{ K}\Omega} = 10^3 \end{aligned}$$

$$P_o = 1000 P_i \rightarrow \text{Amplifier}$$

* Transistor is a transfer of current.

Basic principle of BJT :



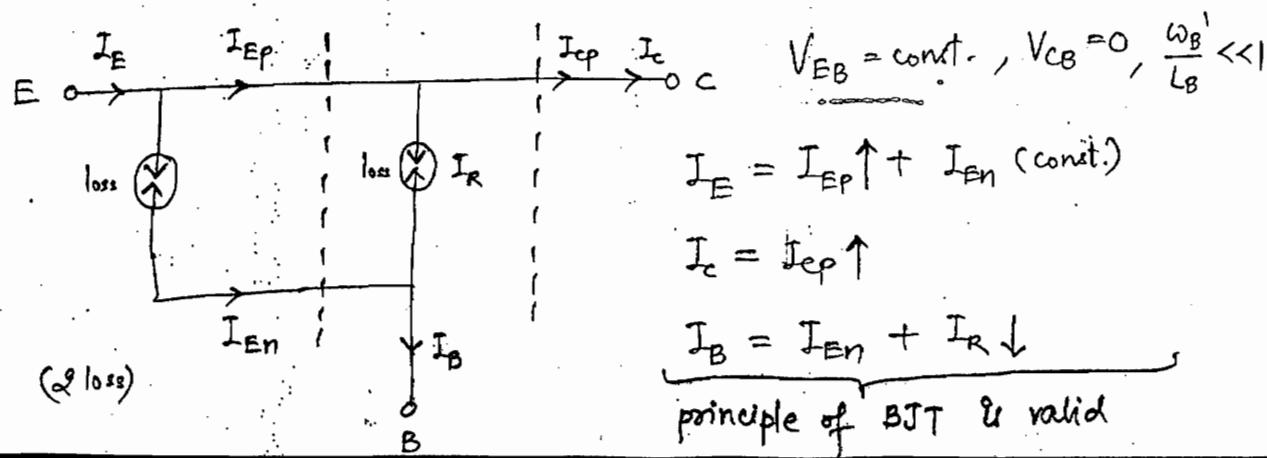
$$V_{CB} = 0V \quad (1) \quad \frac{\omega_B'}{L_B} \gg 1 \quad (2) \quad \frac{\omega_B'}{L_B} \ll 1 \quad (\text{width of Base is narrow})$$

Diffusion :

$$I_p = -AqD_p \frac{dp}{dx}$$

Diffusion $\propto \frac{1}{\text{Recombination rate}}$

$$I_n = AqD_n \frac{dn}{dx}$$



(1). Emitter injection efficiency-

$$\begin{aligned}\gamma &= \frac{I_{EP}}{I_E} \\ &= \frac{I_{EP}}{I_{EP} + I_{En}} \\ &= \frac{1}{1 + \frac{I_{En}}{I_{EP}}} \approx 1\end{aligned}$$

(2). Base transport factor (β^*)

$$\beta^* = \frac{I_{cp}}{I_{EP}} = \frac{I_{EP} - I_R}{I_{EP}} = 1 - \frac{I_R}{I_{EP}} \approx 1$$

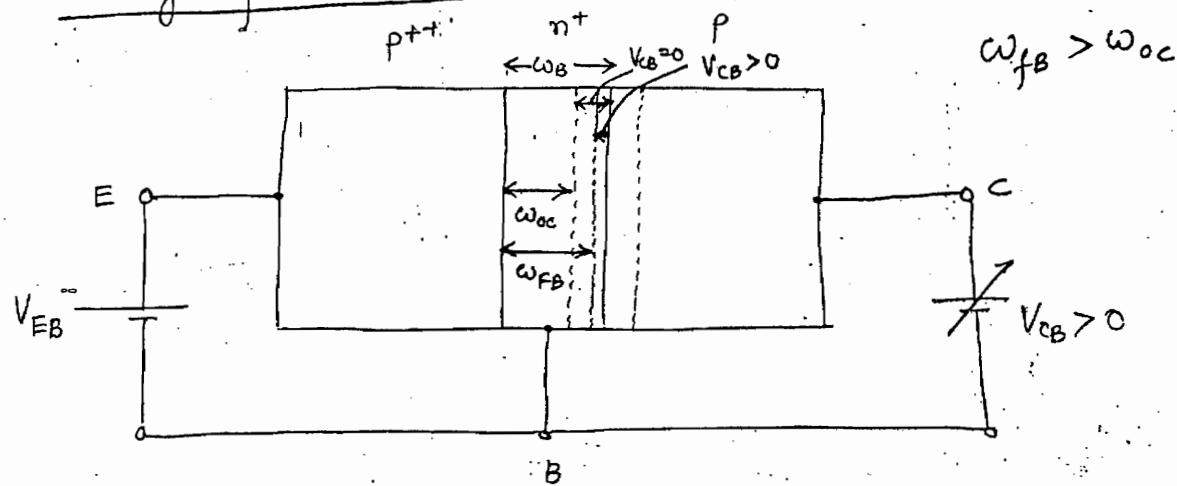
(3). Large s/g current gain (α)

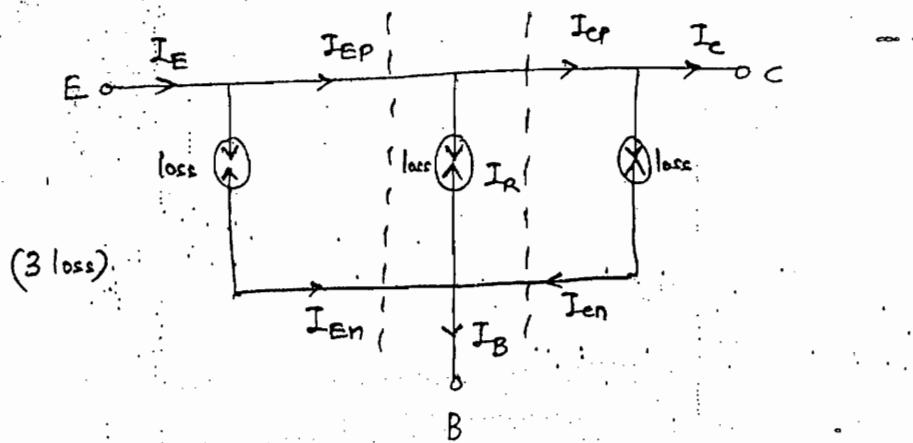
$$\alpha = \frac{I_C}{I_E} = \frac{I_{cp}}{I_{EP}} : \frac{I_{EP}}{I_E}$$

$$\therefore \alpha = \beta^* \gamma$$

In Best transistors, $\alpha = 0.98$ i.e. $I_C = 0.98 I_E$

Testing of BJT in Saturation:





$$V_{EB} = \text{const.}; \quad V_{CB} > 0; \quad \frac{w_B}{L_B} \ll 1$$

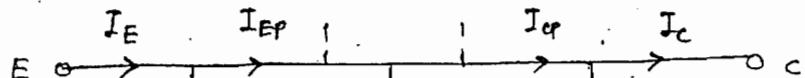
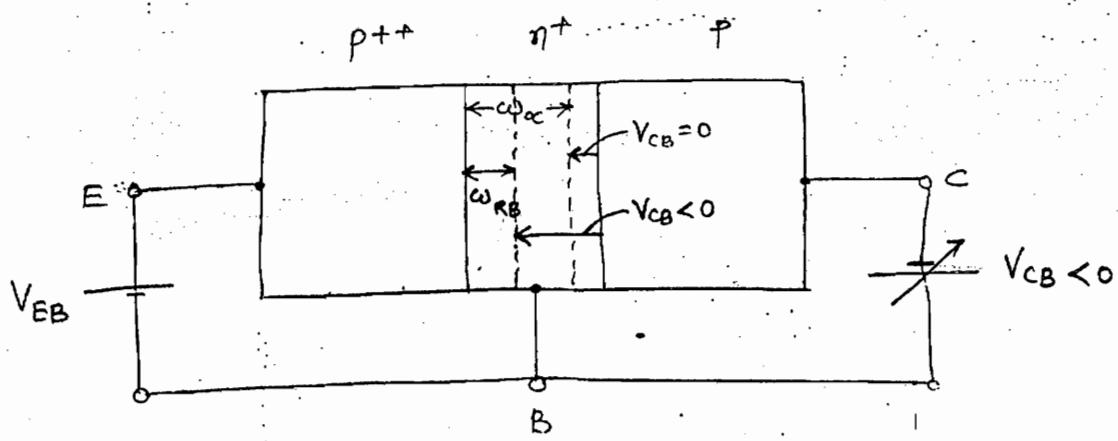
$$\downarrow I_E = I_{EP} \downarrow + I_{En} \text{ (const.)}$$

$$\downarrow I_C = I_{CP} \downarrow - I_{en} \uparrow$$

$$\uparrow I_B = I_{En} + I_{Rb} \uparrow + I_{en} \uparrow \text{ (const.)}$$

} BJT is a failure in saturation

Testing of BJT in Active Region:



$$V_{EB} = \text{const.}; \quad V_{CB} < 0; \quad \frac{w_B}{L_B} \ll 1$$

$$\uparrow I_E = I_{EP} \uparrow + I_{En}$$

$$\downarrow I_B = I_{En} \text{ (const.)} + I_{Rb} \downarrow - I_{ea} \text{ (const.)}$$

$$\uparrow I_C = I_{EP} \uparrow + I_{Co} \text{ (const.)}$$

BJT conditions :

(1). Doping levels

$$N_E \gg N_B$$

$$N_B > N_C$$

(2) ω_B should be narrower (Recombination losses reduced)

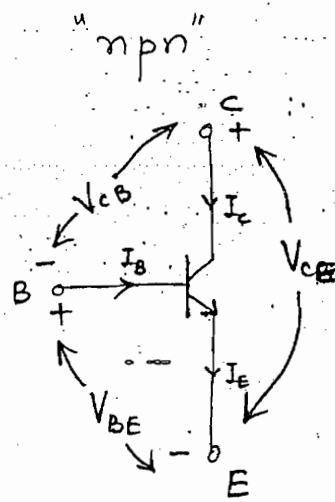
$$I_E = I_B + I_C$$

$$I_E \approx I_C$$

(3). $(\text{Area})_{\text{Collector}} > (\text{Area})_{\text{emitter}} > (\text{Area})_{\text{Base}}$
 • (70%) (25%) (5%)

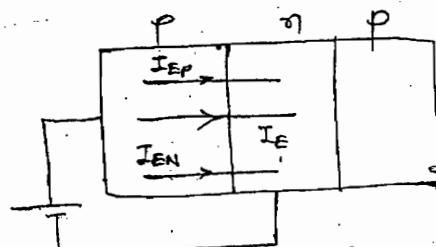
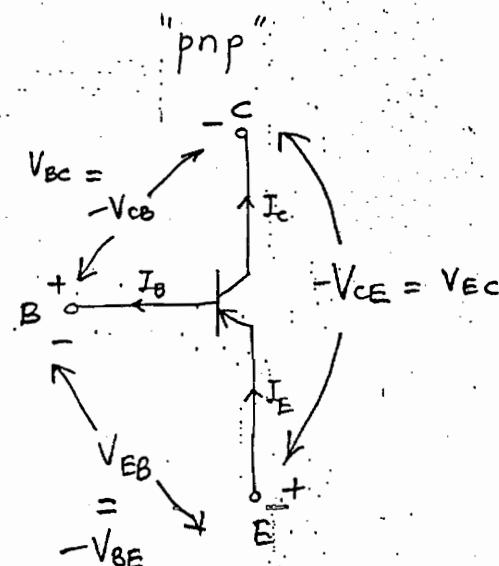
Becoz' heat generation is more in collector region
due to R_B , Hence, area has to be more for collector.

Symbols & Notations of BJT :



$$V_{CB} + V_{BE} = V_{CE}$$

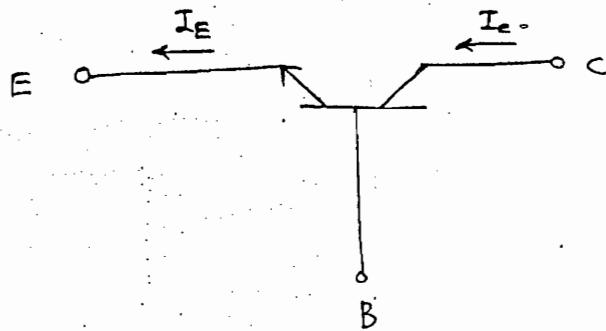
always



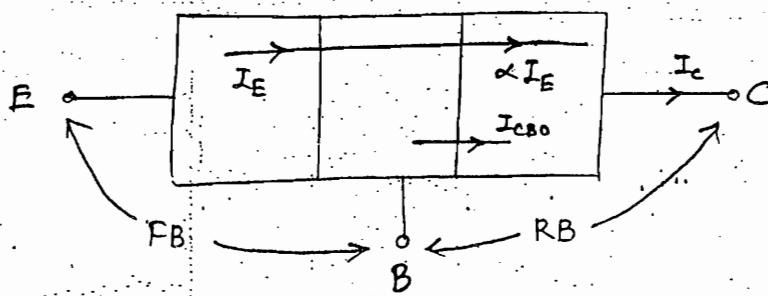
(Testing of symbol pnp)

B J T Configuration:

Common Base (CB):



$$\alpha = \frac{I_C}{I_E} ; \alpha \rightarrow (0.90 \text{ to } 0.99)$$



I_{CBO} → Collector to Base leakage current when emitter is open.

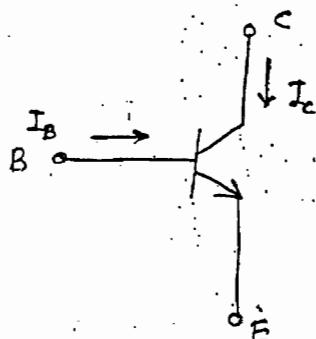
$$I_C = \alpha I_E + I_{CBO}$$

$$I_C = \alpha (I_B + I_D) + I_{CBO}$$

$$I_C (1 - \alpha) = \alpha I_B + I_{CBO}$$

$$I_C = \left(\frac{\alpha}{1 - \alpha} \right) I_B + \left(\frac{1}{1 - \alpha} \right) I_{CBO}$$

Common Emitter (CE):



$$\beta = \frac{I_C}{I_B} ; \beta \rightarrow (20 \text{ to } 500)$$

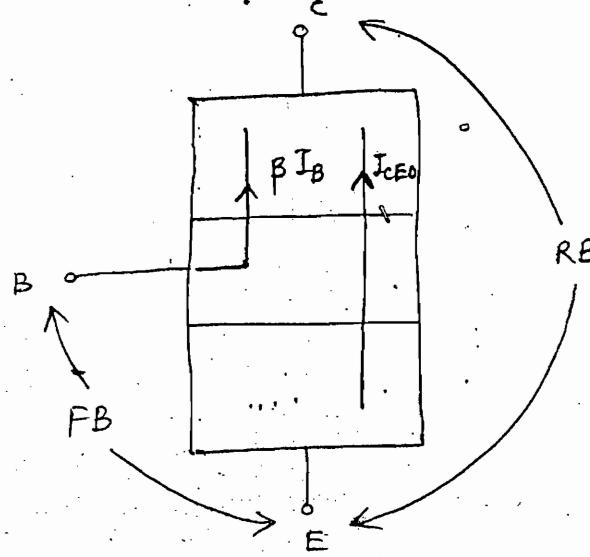
$$\beta = \frac{I_C}{I_E - I_C}$$

$$\beta = \frac{I_C / I_E}{1 - I_C / I_E} = \frac{\alpha}{(1 - \alpha)}$$

$$\therefore \beta = \frac{\alpha}{1-\alpha}$$

$$\& \quad \alpha = \frac{\beta}{1+\beta}$$

and $\frac{1}{(1-\alpha)} = 1 + \beta$



$I_{CEO} \rightarrow$ Collector to emitter leakage current when base is open.

$$I_C = \beta I_B + I_{CEO}$$

Relation b/w I_{cbo} and I_{ebo} :

$$CB: \quad I_C = \left(\frac{\alpha}{1-\alpha}\right) I_B + \left(\frac{1}{1-\alpha}\right) I_{cbo}$$

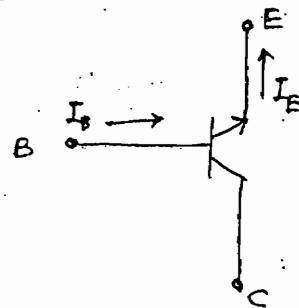
$$I_C = \beta I_B + (1+\beta) I_{cbo} \quad (1)$$

$$CE: \quad I_C = \beta I_B + I_{ebo} \quad (2)$$

On comparing, $I_{ebo} = (1+\beta) I_{cbo}$

$$I_{ebo} = \left(\frac{1}{1-\alpha}\right) I_{cbo}$$

Common Collector (CC):



$$r = \frac{I_E}{I_B}; \quad r \rightarrow (20 \text{ to } 500)$$

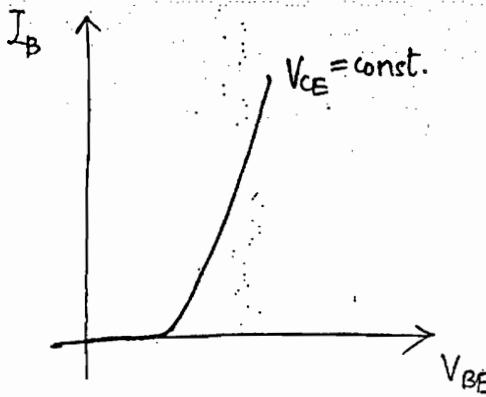
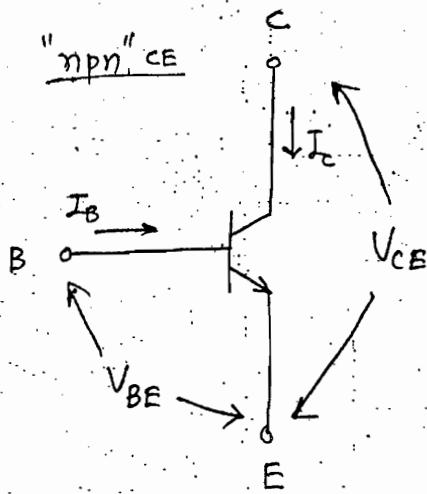
$$r = \frac{I_E}{I_E - I_C}$$

$$Y = \frac{1}{1-\alpha} = 1 + \beta$$

$$Y \approx \beta$$

* BJT operating Regions :

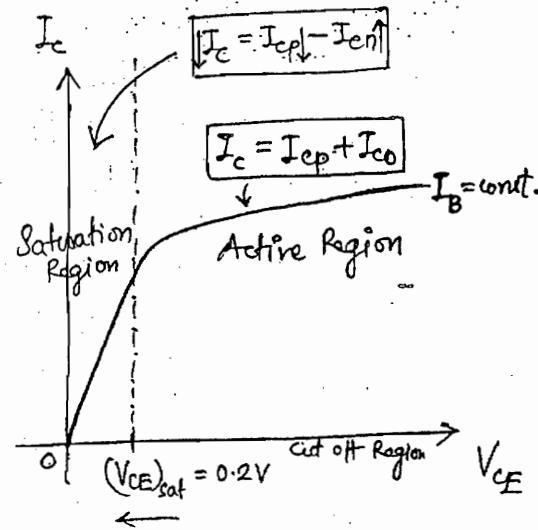
Regions	J_E	J_C	Applications
(1) Active Region	FB	RB	Amplifiers
(2) Saturation Region	FB	FB	"on"
(3) Cut-off Region	RB	RB	"OFF"
(4) Reverse active Region	RB	FB	X



(I/p characteristics shows the diode characteristics and it is of no use, it does not represent BJT)

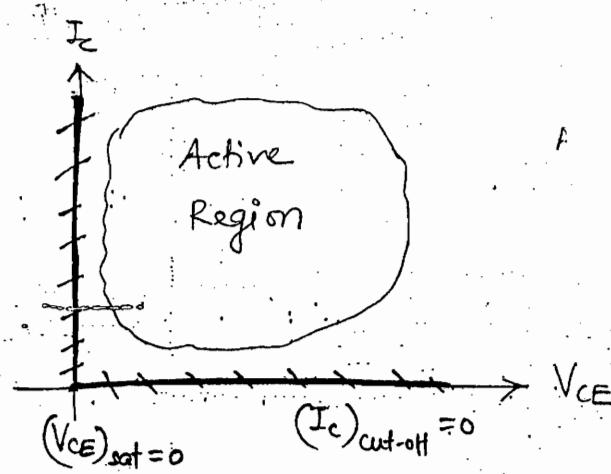
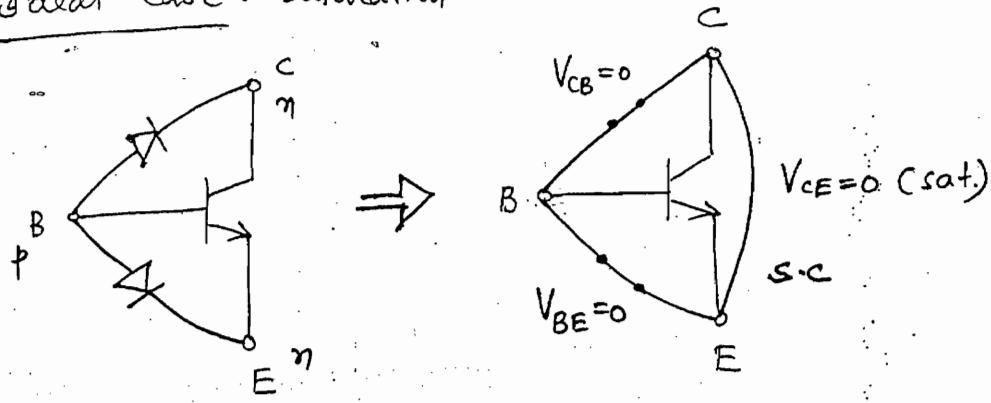
I/p parameters $\rightarrow V_{BE}, I_B$

O/p parameters $\rightarrow V_{CE}, I_C$

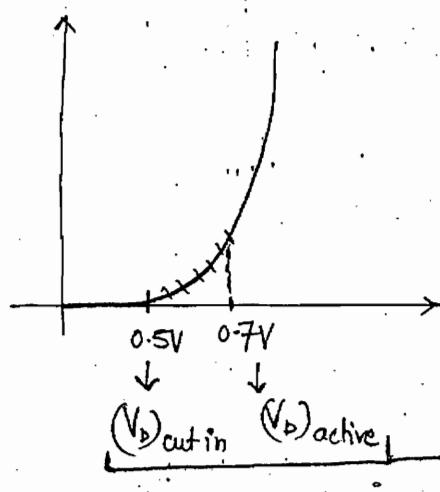
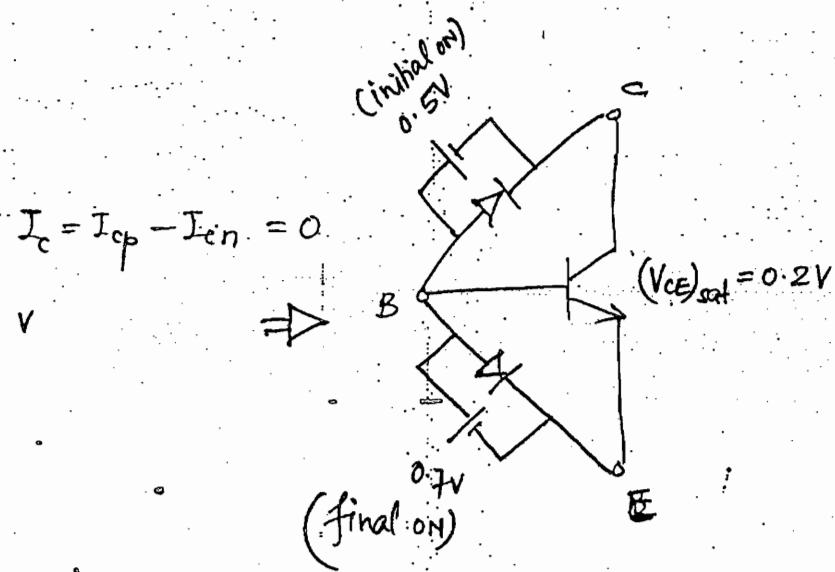
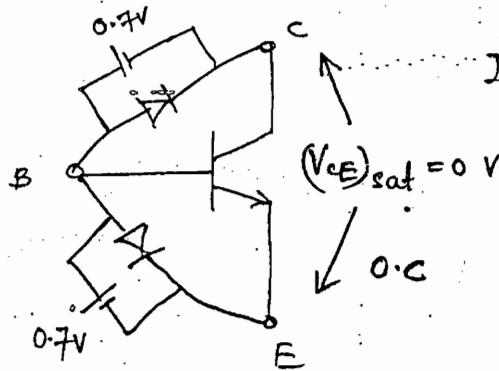


(O/p characteristics are used to study BJT and BJT is represented by these characteristics).

Ideal case: saturation

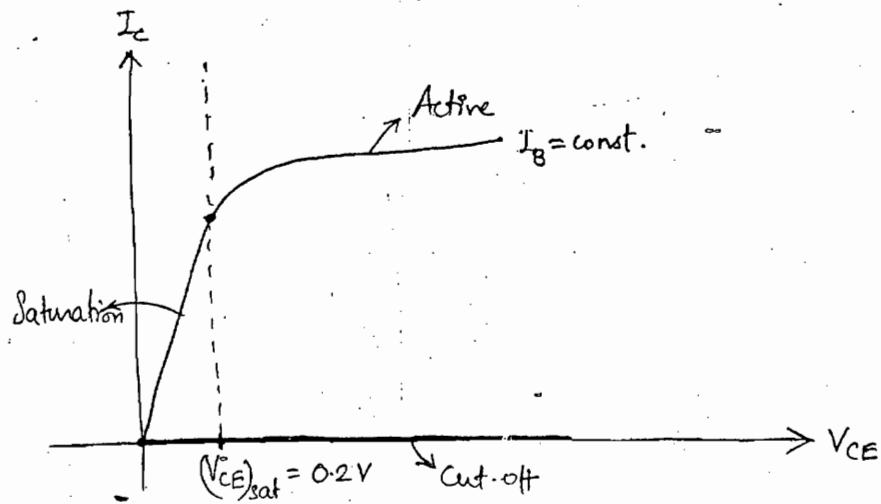
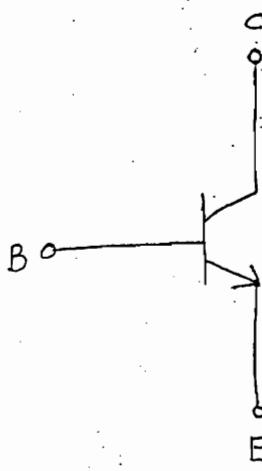


Practical case:

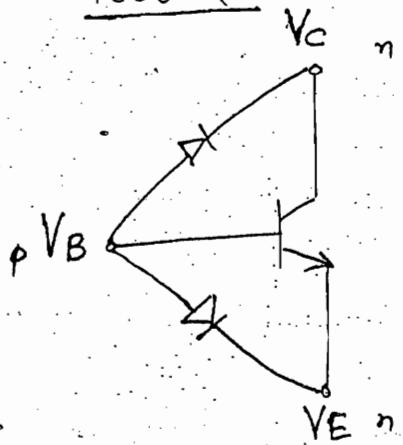


Diode \rightarrow ON $\leftarrow 0.5\text{ V}$ (Initial on) $\leftarrow 0.7\text{ V}$ (Final on)

Testings of BJT in different operating Regions:



Test - (1):



Saturation \rightarrow

$$V_B > V_E$$

$$V_B > V_C$$

Cut-off \rightarrow

$$V_B < V_E$$

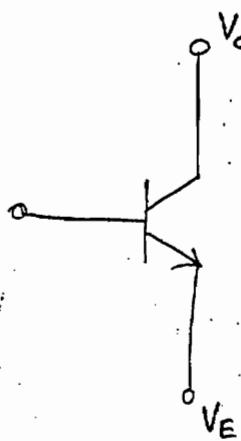
$$V_B < V_C$$

Active Region \rightarrow

$$V_B > V_E$$

$$V_B < V_C$$

Test - (2):



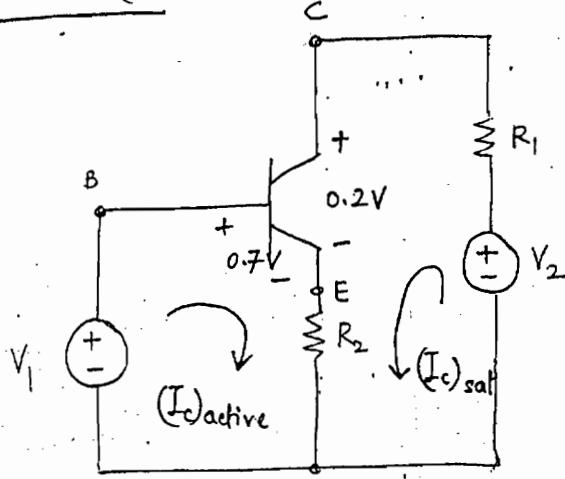
$$V_{CE} = V_C - V_E$$

Saturation \rightarrow

$$V_{CE} < (V_{CE})_{sat}$$

Active Region \rightarrow

$$V_{CE} > (V_{CE})_{sat}$$

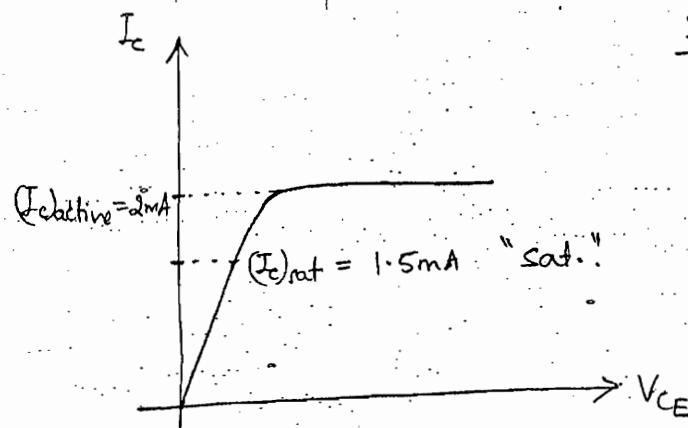
Test - (3) :

$$(V_{BE})_{active} = 0.7V \quad (\text{I/p loop})$$

$$(V_{CE})_{sat.} = 0.2V \quad (\text{o/p loop})$$

$$(I_c)_{active} = \frac{V_1 - 0.7}{R_2} = 2\text{mA (say)}$$

$$(I_c)_{sat} = \frac{V_2 - 0.2}{R_1 + R_2} = 1.5\text{ mA (say)}$$

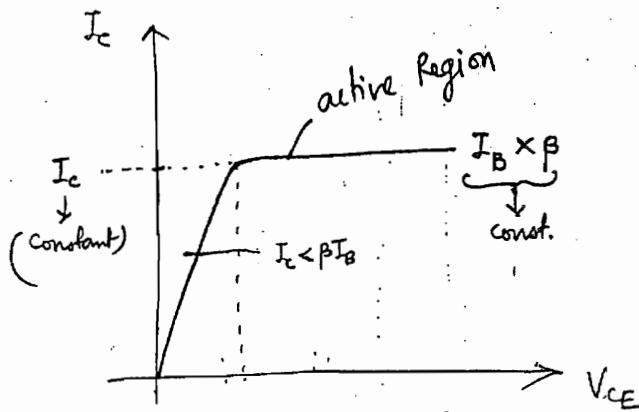


Saturation \rightarrow

$$(I_c)_{sat} < (I_c)_{active}$$

Active Region \rightarrow

$$(I_c)_{sat} > (I_c)_{active}$$

Test - (4) :

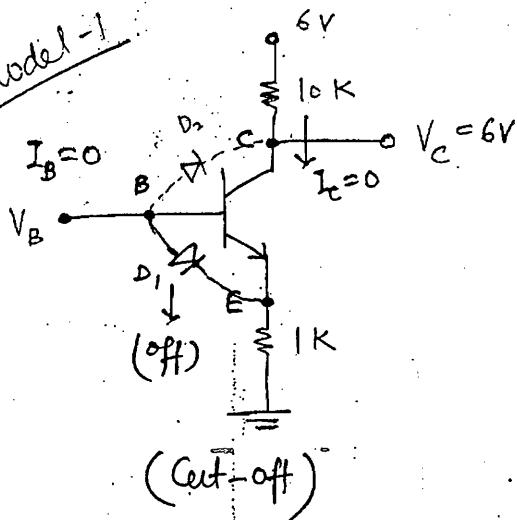
Saturation \rightarrow

$$I_c < \beta I_B$$

Active Region \rightarrow

$$I_c = \beta I_B \quad (\text{const.})$$

Q2. Model -1



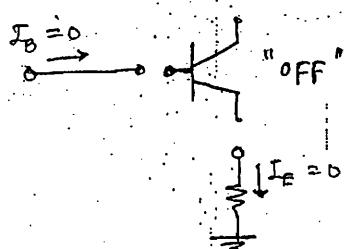
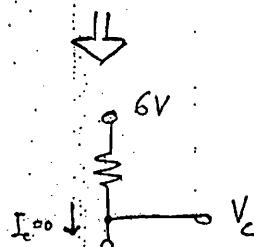
for $V_B = 0V$

$D_1 \rightarrow (\text{off})$

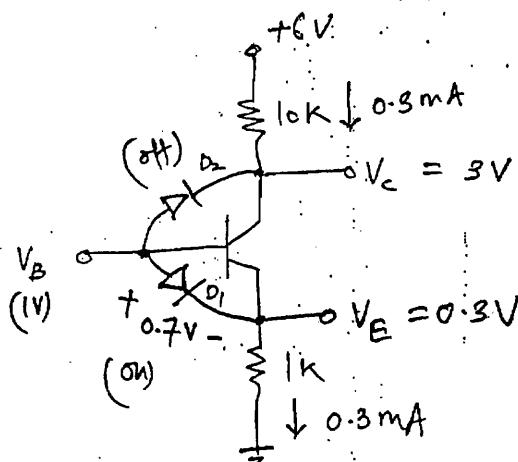
$D_2 \rightarrow (\text{off})$ automatically.

There is no chance of reverse active.
Hence, there is a cut-off.

$$I_B = I_C = I_E = 0.$$



for $V_B = 1V$



Active Region:

$D_1: (\text{on})$

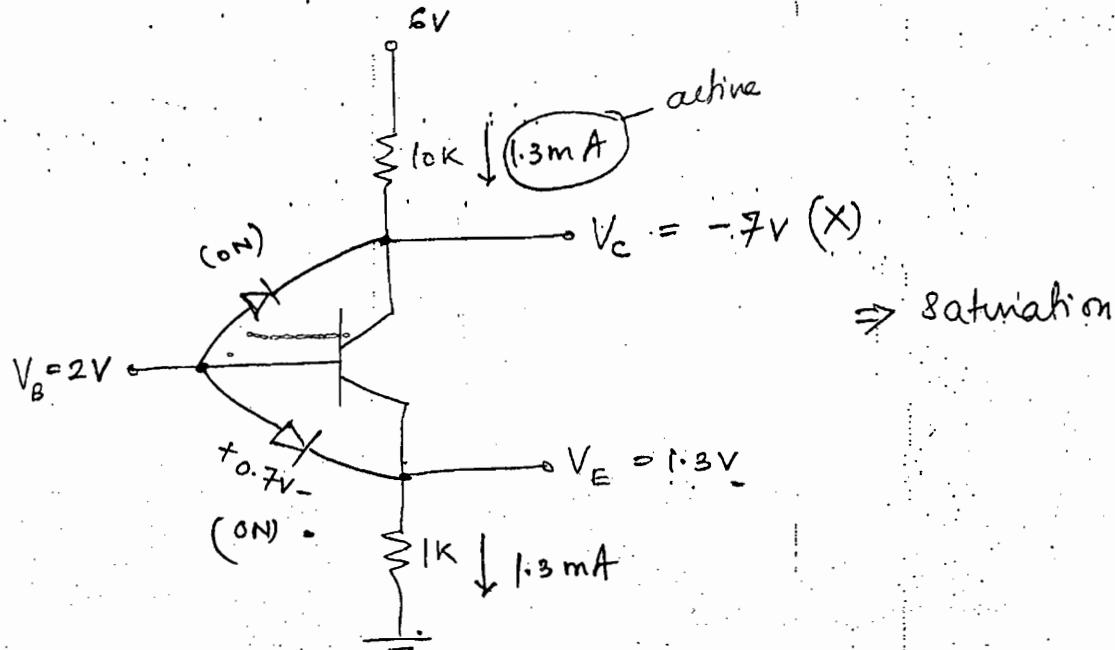
$D_2: (\text{off})$

$$\begin{aligned} V_C &= 6 - 0.3mA \times 10k \\ &= 6 - 3 \\ &= 3 \text{ volt}. \end{aligned}$$

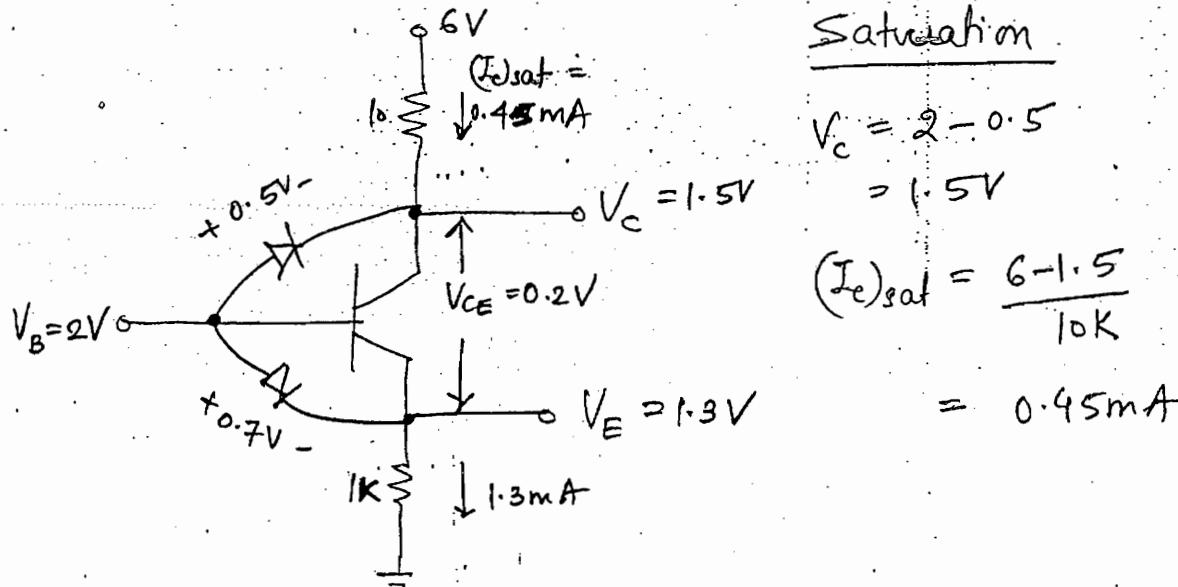
$$I_E = \frac{V_E}{1k} = \frac{0.3}{1k} = 0.3mA$$

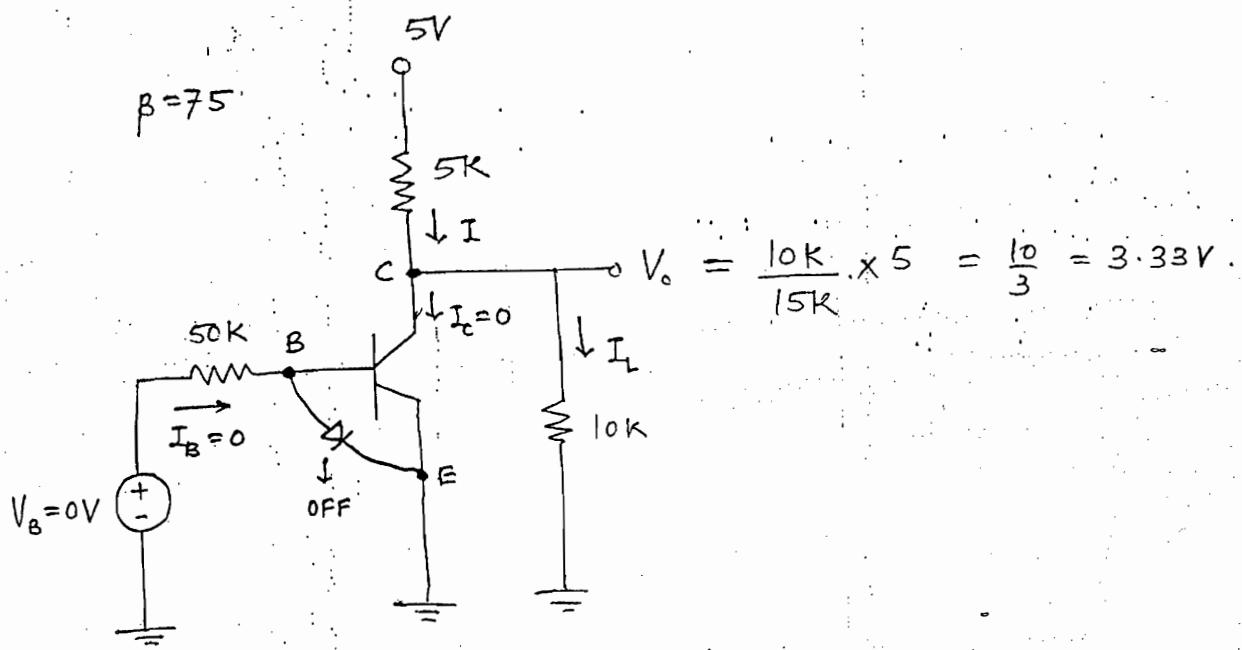
$$I_C \approx I_E = 0.3mA$$

For $V_B = 2V$

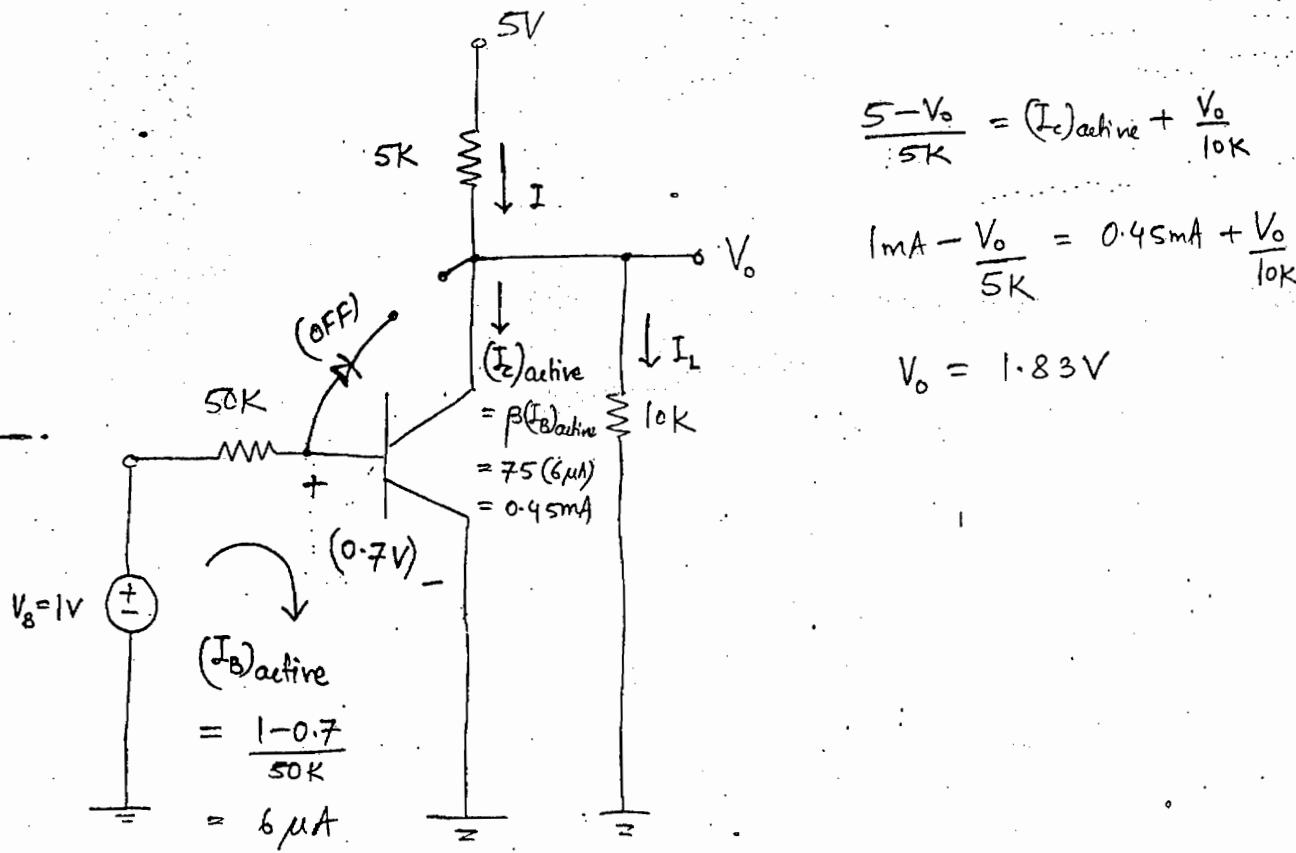


Saturation

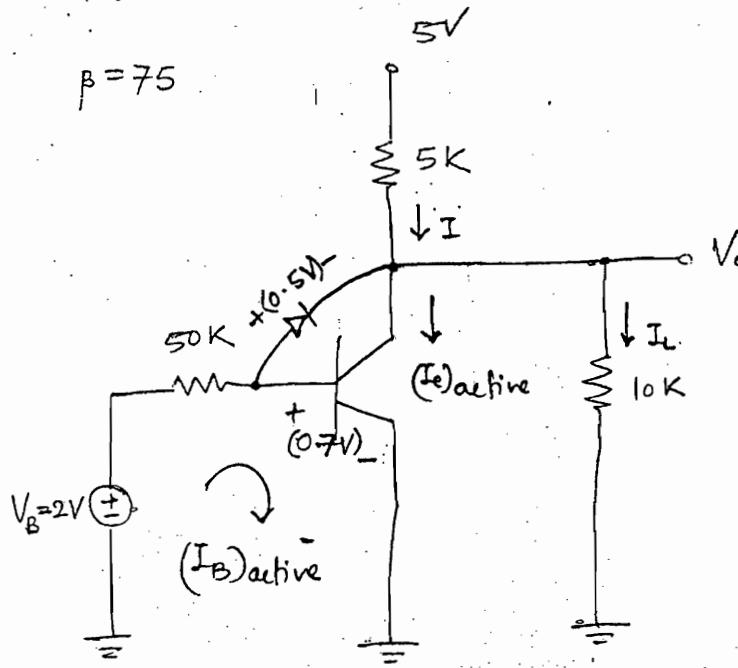


Model - (2)for $V_B = 0V$: (Cut off)

$$V_o = \frac{10k}{15k} \times 5 = \frac{10}{3} = 3.33V$$

for $V_B = 1V$: (Active)

For $V_B = 2V$



V_o is ?

- (a) 3 V (b) 5 V
 (c) 0.2 V (d) 1.3 V

(practical dkt)

Ideally diode \rightarrow sc.
 then $(V_{CE})_{sat} = 0V$.

$$(I_B)_{active} = \frac{2 - 0.7}{50K} = 26 \mu A$$

$$(I_C)_{active} = \beta \cdot (I_B)_{active} = 75 \times 26 \mu A = 1.95 mA$$

$$\text{Now, } \frac{5 - V_o}{5K} = 1.95 + \frac{V_o}{10K}$$

$$1mA - \frac{V_o}{5} = 1.95mA + \frac{V_o}{10K} \Rightarrow V_o = -Ve$$

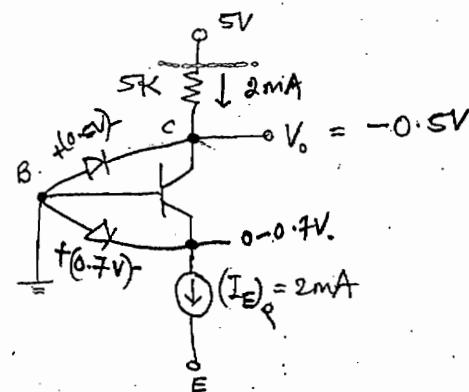
saturation

$$\Rightarrow V_o = -3.16 V$$

$$\text{Saturation} \rightarrow V_o = (V_{CE})_{sat} = 0V \text{ (ideally)}$$

$$V_o = (V_{CE})_{sat} = 0.2 V \text{ (practically)}$$

Q1.



$$V_o = 5 - 5K \times 2mA$$

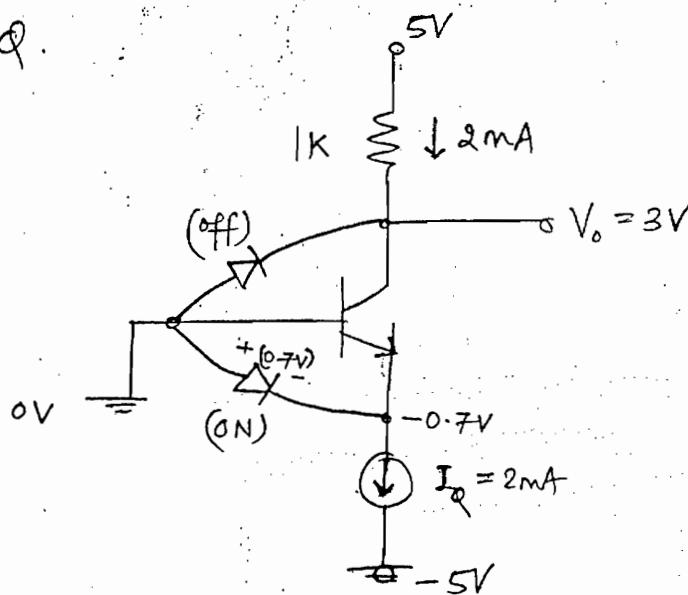
$$V_o = 5 - 10$$

$$V_o = -5V (X)$$

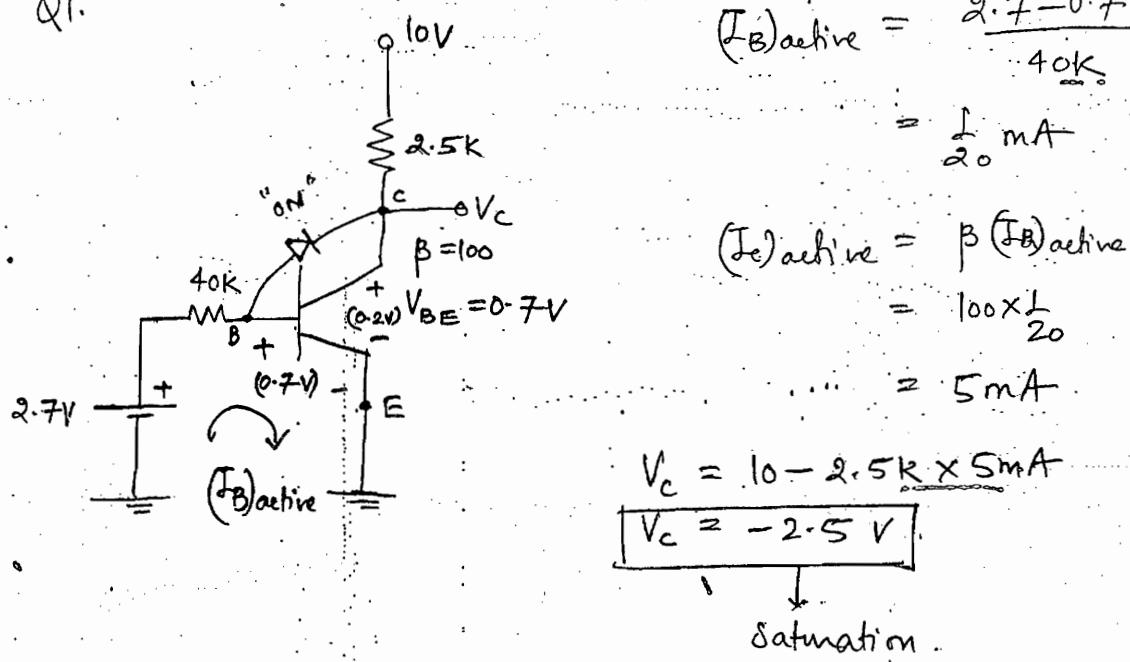
$$V_o = -Ve \rightarrow \text{saturation}$$

$$\therefore V_o = -0.5V$$

Q.



Q1.

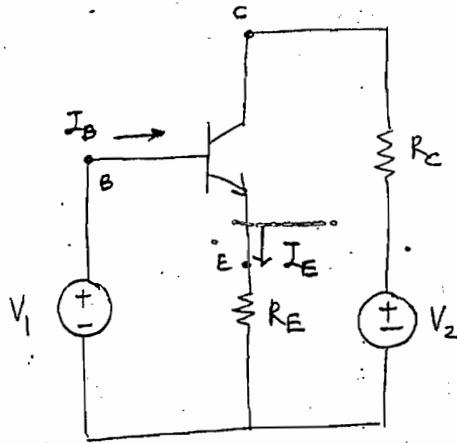


$$\therefore V_{CE} = 0.2\text{V}$$

$$\frac{10 - 0.2}{2.5\text{K}} = (I_c)_{\text{sat.}}$$

$$(I_c)_{\text{sat.}} = \frac{9.8}{2.5\text{K}} = 3.9 \text{ mA}$$

Model (1) :



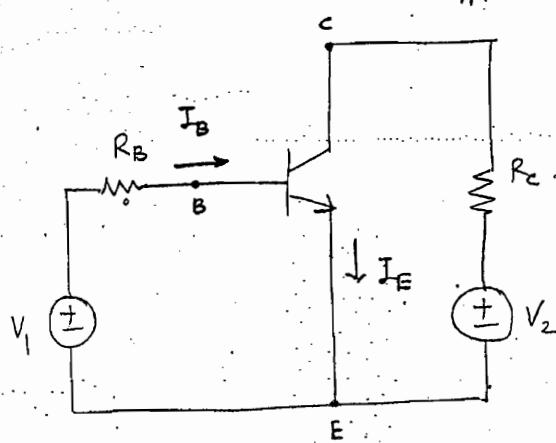
$$\begin{aligned}I_E &= I_B + I_C \\&\Rightarrow I_B + \beta I_B \\&= (1+\beta) I_B\end{aligned}$$

$$V_1 = V_{BE} + I_E R_E$$

(or)

$$V_1 = V_{BE} + I_B R_E (1+\beta)$$

Model (2) :

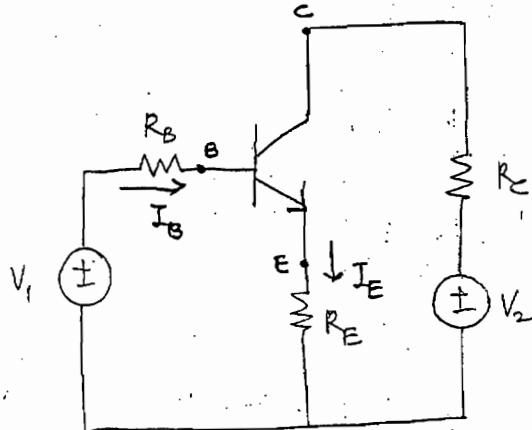


$$V_1 = I_B R_B + V_{BE}$$

(or)

$$\begin{aligned}V_1 &= \frac{I_E R_B}{(1+\beta)} + V_{BE} \\&= I_E \left(\frac{R_B}{1+\beta} \right) + V_{BE}\end{aligned}$$

Model (3) :

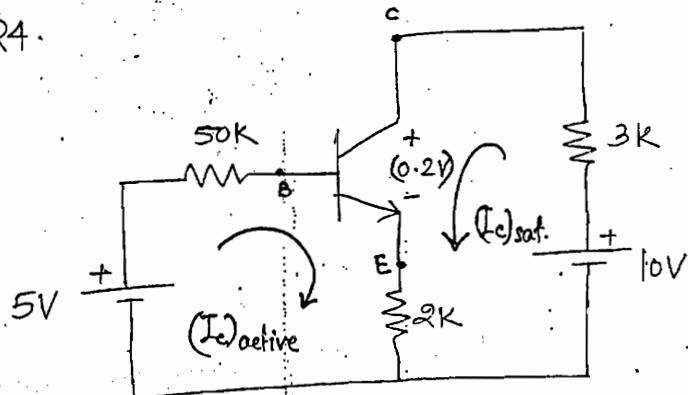


$$V_1 = I_B R_B + V_{BE} + I_B R_E (1+\beta)$$

(or)

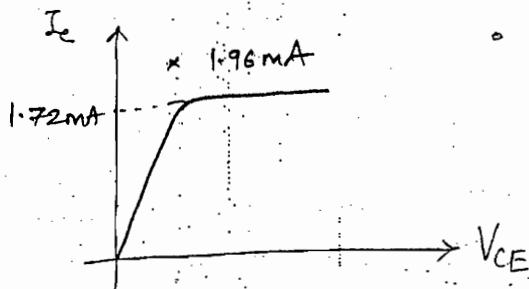
$$V_1 = I_E \left(\frac{R_B}{1+\beta} \right) + V_{BE} + I_E R_E$$

Q4.



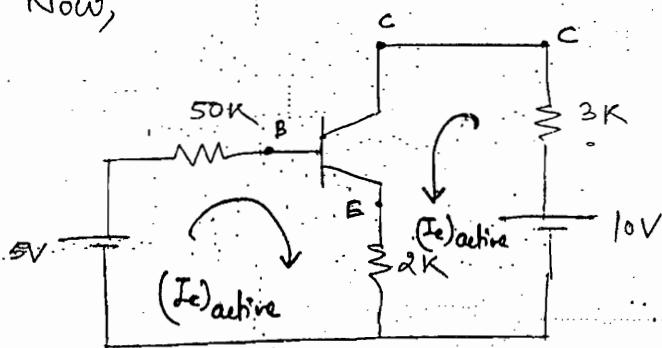
$$(I_c)_{\text{active}} = \frac{5 - 0.7}{\left(\frac{50k}{1+100}\right) + 2k} \\ = 1.72 \text{ mA}$$

$$(I_c)_{\text{sat}} = \frac{10 - 0.2}{5k} \\ = 1.96 \text{ mA}$$



$(I_c)_{\text{sat}} < 1.72 \text{ mA}$
 But $(I_c)_{\text{sat}} > 1.72 \text{ mA}$
 \therefore Active Region.

Now,



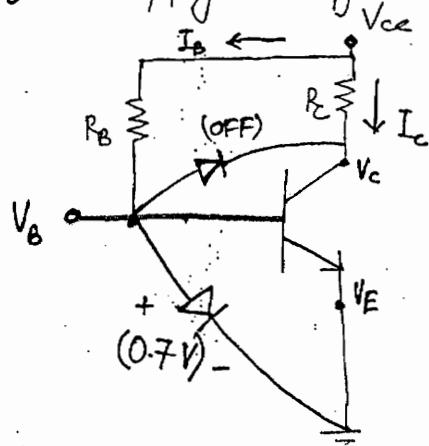
$$V_E = 1.72 \text{ mA} \times 2k \\ = 3.44 \text{ V}$$

$$V_C = 10 - 3k \times 1.72 \text{ mA} \\ = 4.84 \text{ V}$$

$$V_{CE} = V_C - V_E \\ = 4.84 - 3.44 \\ = 1.4 \text{ V}$$

Single supply design in BJT to operate in Active Region:

(1).

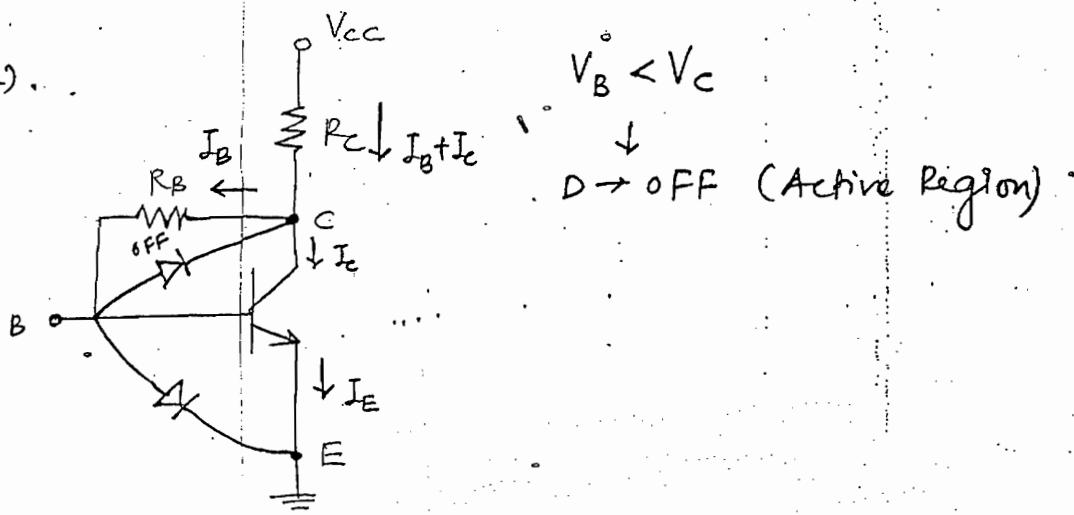


$$R_B > R_C$$

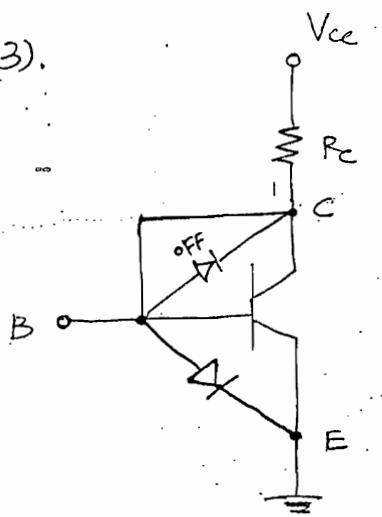
$$V_B < V_C \rightarrow \text{D (off)}$$

Active Region.

(2).

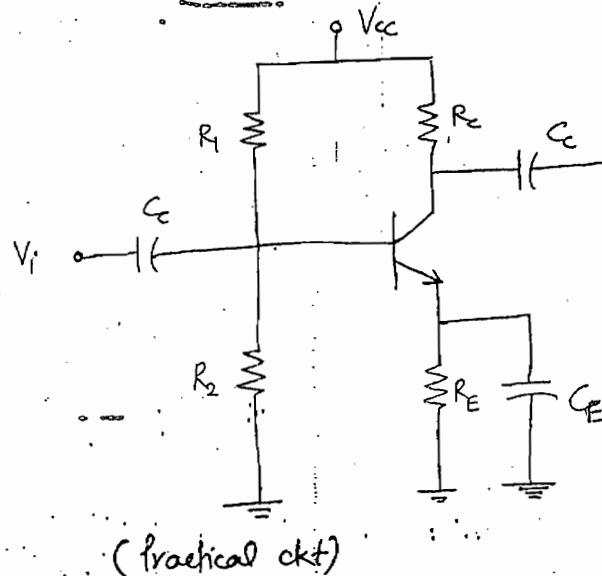


(3).



DC load line and operating point (Q) :

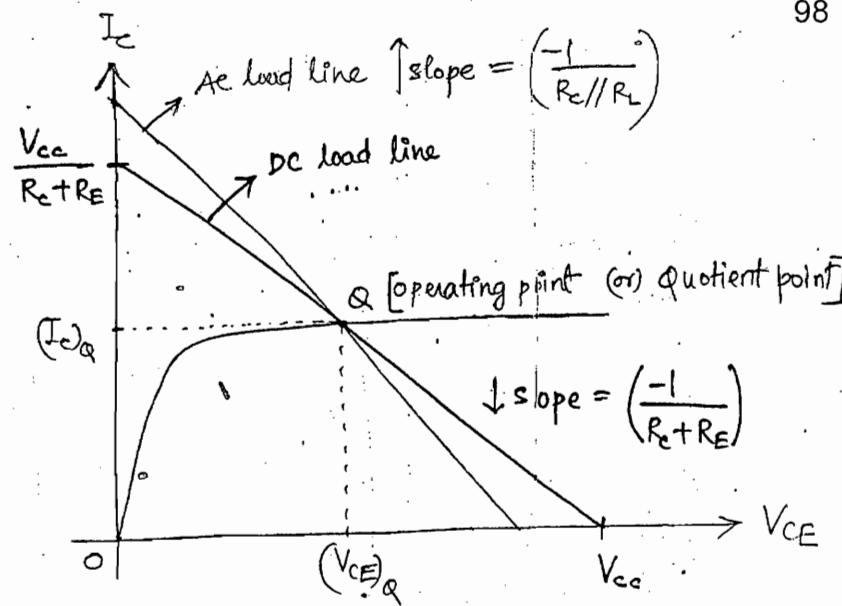
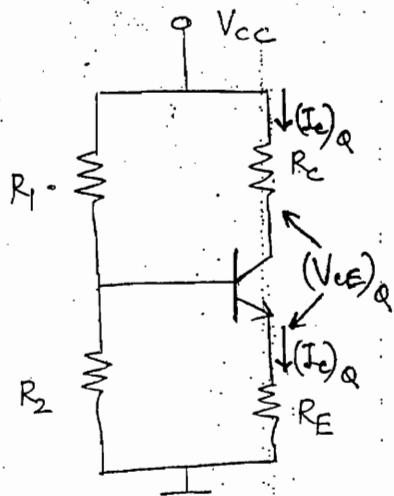
CE Amplifier:



DC Analysis:

(1). AC should be grounded.

(2). $X_C \propto \frac{1}{f} \rightarrow 0 \text{ (DC)}$ $\rightarrow \infty$ ($C \rightarrow \infty$)



O/P loop:

$$V_{cc} = I_c (R_c + R_E) + V_{CE}$$

$$I_c = -\frac{1}{(R_c + R_E)} \cdot V_{CE} + \frac{V_{cc}}{R_c + R_E}$$

(Quotient \rightarrow greek word
meaning \rightarrow Inactive)

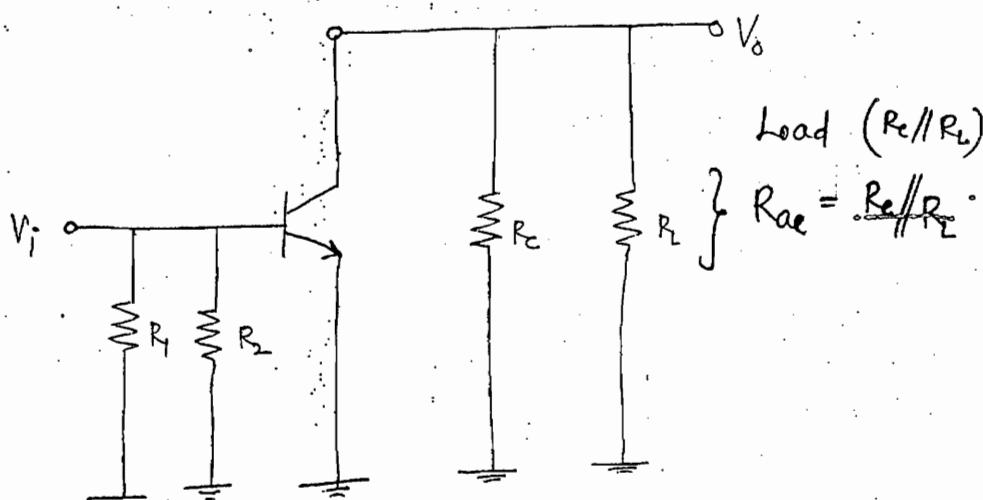
Load $(R_c + R_E)$

AC Analysis:

AC load line —

(1). ~~Ec~~ should be grounded.

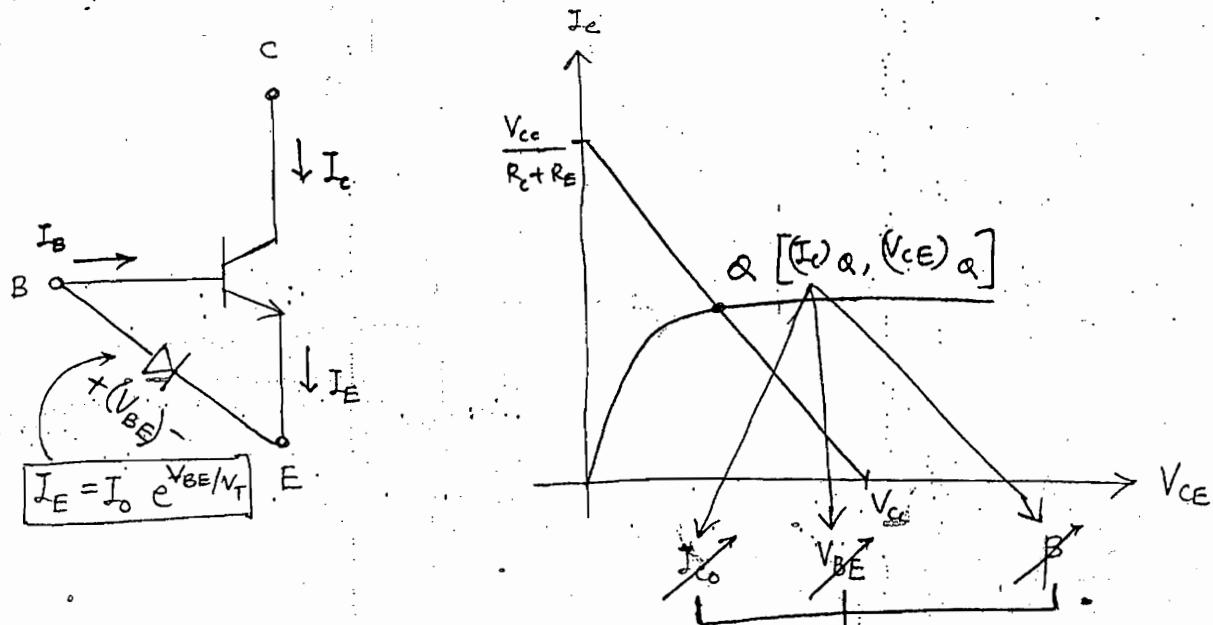
(2). $X_C \propto \frac{1}{f} \rightarrow 0 \quad (C \rightarrow \infty)$



Range: $R_c, R_E, R_L \rightarrow (0 \text{ to } 10k\Omega)$

Temperature dependence on BJT parameters:

99



$$(1) \cdot I_C = \beta I_B + (1 + \beta) I_{CBO}$$

$$(2) \cdot I_C = I_0 e^{V_{BE}/V_T}$$

$$(3) \cdot I_C = \beta I_B$$

"Biasing" is a ckt or technique which makes the the α point stable w.r.t. temperature variations.

I_{C0} Vs Temperature:

' I_{C0} ' increases by 7% for every $^{\circ}\text{C}$ rise in temp.

' I_{C0} ' doubles for every 10°C rise in temp.

V_{BE} Vs Temp. :

$$\frac{dV_{BE}}{dT} = -2.5 \text{ mV/}^{\circ}\text{C}$$

$T \uparrow, V_{BE} \downarrow, I_C \downarrow$

β Vs temp. :

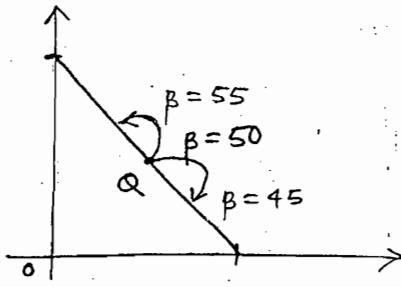
$$\beta = \frac{I_C}{I_B}, T \uparrow, I_{C0} \uparrow, I_C \uparrow, \beta \uparrow$$

neglected

B replacement problem:

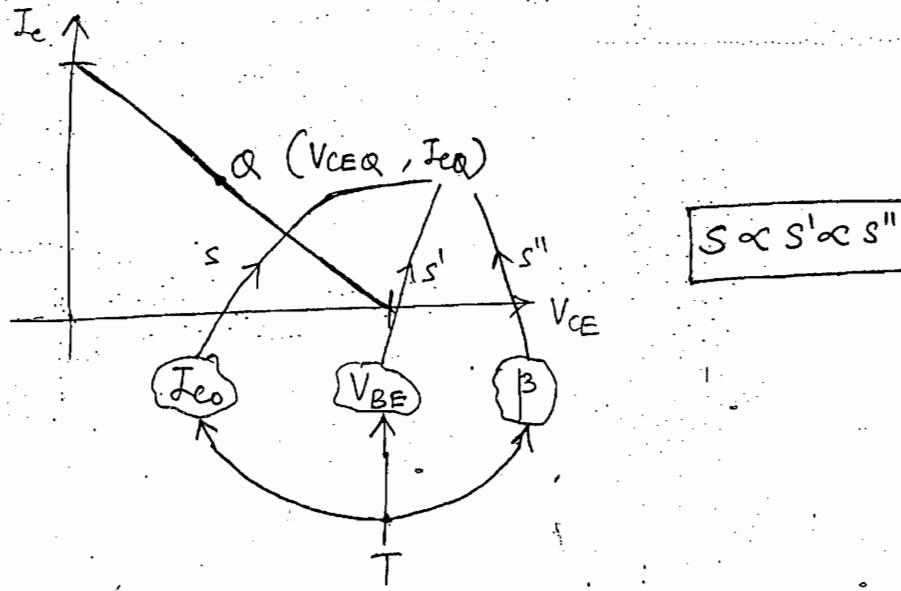
β value changes and hence point Q changes.

Replacement of transistor shift the Q point because of mismatching.



Stability factor (s):

"It is a measure of variation of Q point w.r.t temp."



$$s = \left. \frac{\partial I_c}{\partial I_{c0}} \right|_{V_{BE} \text{ & } \beta = \text{const.}}$$

$$s' = \left. \frac{\partial I_c}{\partial V_{BE}} \right|_{I_{c0} \text{ & } \beta = \text{const.}}$$

$$s'' = \left. \frac{\partial I_c}{\partial \beta} \right|_{I_{c0} \text{ & } V_{BE} = \text{const.}}$$

Conclusion:

- (1). The most dominating parameter w.r.t temp. is I_{CO} .
- (2). The stability factor ideally it should be zero, practically it should be min^m as possible.

$$I_C = \beta I_B + (1+\beta) I_{CO}$$

Differentiate w.r.t. I_C . We get,

$$\frac{dI}{dI_C} = \beta \frac{\partial I_B}{\partial I_C} + (1+\beta) \cdot \frac{\partial I_{CO}}{\partial I_C}$$

$$\frac{dI}{dI_C} = \beta \frac{\partial I_B}{\partial I_C} + \frac{(1+\beta)}{S}$$

$$S = \frac{1+\beta}{1 - \beta \frac{\partial I_B}{\partial I_C}}$$

$$\begin{array}{ccc} S=0 & \xrightarrow{\hspace{1cm}} & S=1 \\ (\text{ideal}) & & (\text{practical}) \end{array}$$

Compensation

Biasing ckt:Stabilization Techniques

fixed bias
($S \approx 10^3$)

Collector to base
Bias
($S \approx 45$)

Voltage divider
Bias
($S \approx 1$)
Self Bias
(or)
Emitter Bias
(or)
Universal Bias

Compensation techniques

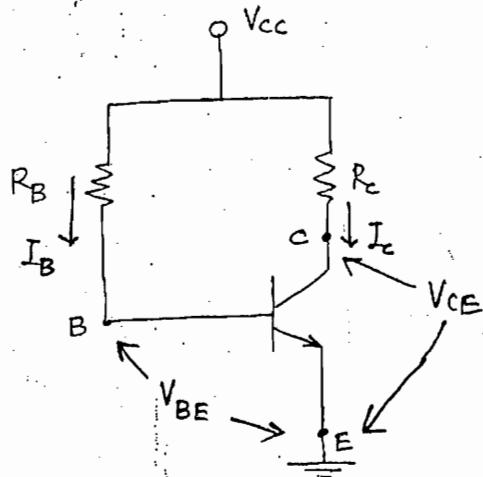
Diode

Sensistor

Thermistor

temp. sensitive devices

Fixed Bias :



$$I_B = \frac{V_{CC} - V_{BE}}{R_B}$$

If $V_{CE} \gg V_{BE}$

$$I_B \approx \frac{V_{CC}}{R_B} \rightarrow \text{fixed}$$

$$I_C = \beta I_B$$

↓ ↓
fixed fixed

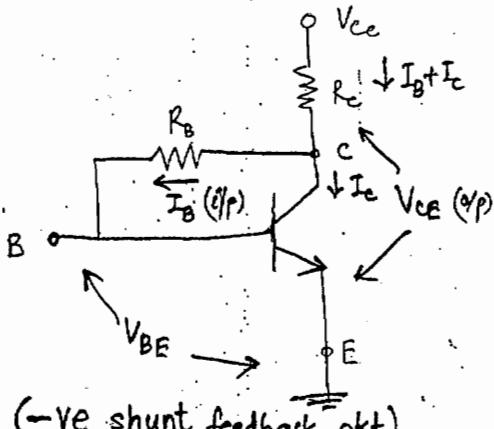
i/p loop :

$$V_{CE} = I_B R_B + V_{BE}$$

o/p loop :

$$V_{CE} = I_C R_C + V_{CE}$$

Collector to Base Bias : (-ve feedback control s)



(-ve shunt feedback ckt)

i/p loop :

$$V_{CC} = (I_B + I_C) R_C + I_B R_B + V_{BE}$$

$$V_{CC} = I_B (R_B + R_C) + I_C R_C + V_{BE}$$

$$I_B = \frac{V_{CC} - I_C R_C - V_{BE}}{R_B + R_C}$$

O/p loop:

$$V_{CE} = (I_B + I_C) R_E + V_{CE}$$

$$V_{CC} = I_C R_E = V_{CE} + I_B R_E$$

$$\therefore I_B = \frac{V_{CE} + I_B R_E - V_{BE}}{R_B + R_E}$$

↓
I/P

Stabilization:

$$T \uparrow, I_B \uparrow, I_C \uparrow, (I_B + I_C) R_E \uparrow, V_{CE} \downarrow$$

$$I_B \downarrow \quad I_C \downarrow$$

Now, $V_{CE} = I_B (R_B + R_E) + I_C R_E + V_{BE}$

Differentiate w.r.t I_C : We get,

$$0 = (R_B + R_E) \frac{\partial I_B}{\partial I_C} + R_E + 0$$

$$\frac{\partial I_B}{\partial I_C} = \frac{-R_E}{(R_B + R_E)}$$

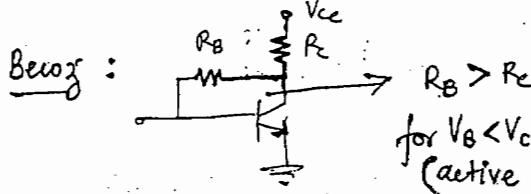
$$S = \frac{1 + \beta}{1 - \beta \frac{\partial I_B}{\partial I_C}}$$

$$S = \frac{1 + \beta}{1 + \beta \frac{R_E}{R_B + R_E}} = \frac{(1 + \beta)(R_B + R_E)}{R_B + (1 + \beta)R_E} = \frac{(1 + \beta) \left(1 + \frac{R_B}{R_E}\right)}{\left(1 + \beta + \frac{R_B}{R_E}\right)}$$

$$S = \frac{1 + \beta}{1 + \beta \frac{1}{\left(\frac{R_B}{R_E} + 1\right)}}$$

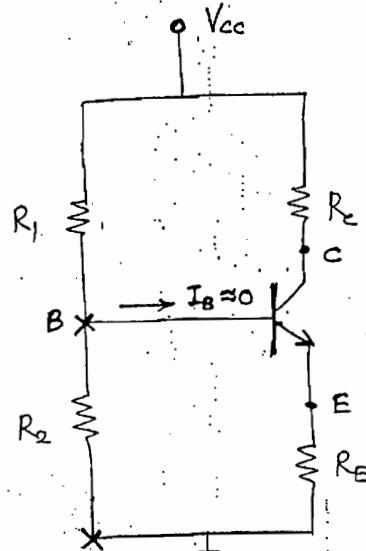
≈ 1 if $\frac{R_B}{R_E} = 0$ $R_B = 0$
 $R_E = \infty$

$\therefore R_B / R_E \ll 1 \rightarrow$ failure condition



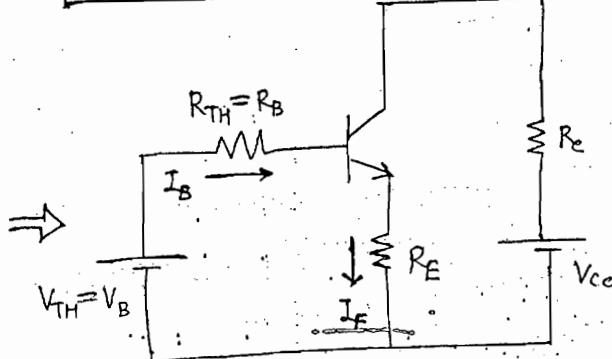
$$\frac{R_B}{R_E} \gg 1 \quad \text{ckt condition}$$

Voltage Divider Bias -



$$V_B = \frac{V_{CC} \times R_2}{R_1 + R_2}$$

$$R_B = R_1 // R_2$$



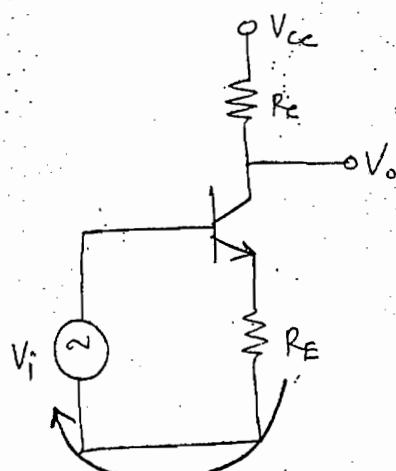
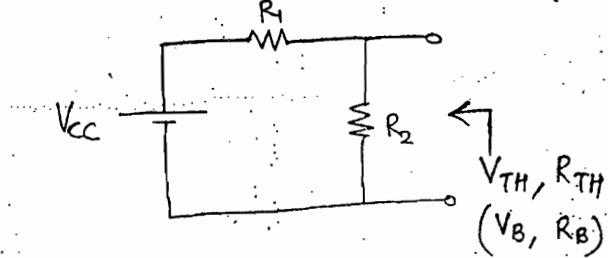
C/p loop :

$$V_B = I_B \cdot R_B + V_{BE} + (I_B + I_c) R_E$$

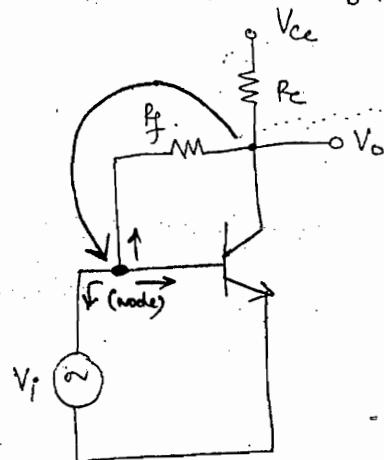
$$V_B = I_B \cdot (R_B + R_E) + I_c R_E + V_{BE}$$

$$0 = (R_B + R_E) \frac{\partial I_B}{\partial I_c} + R_E = 0$$

$$\frac{\partial I_B}{\partial I_c} = -\frac{R_E}{R_B + R_E}$$



Series feedback



Shunt feedback

$$S = \frac{1 + \beta}{1 - \beta \frac{\partial I_B}{\partial I_c}} = \frac{1 + \beta}{1 + \beta \left(\frac{R_E}{R_B + R_E} \right)} = \frac{1 + \beta}{1 + \beta \frac{1}{1 + \frac{R_B}{R_E}}}$$

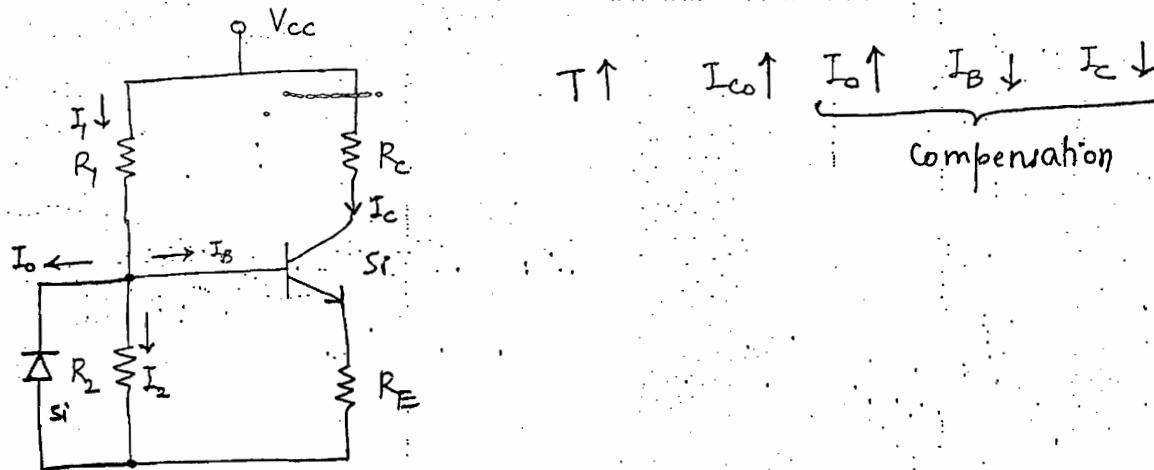
$$\text{if } R_B / R_E \ll 1 \Rightarrow S \rightarrow 1$$

Design conditions:

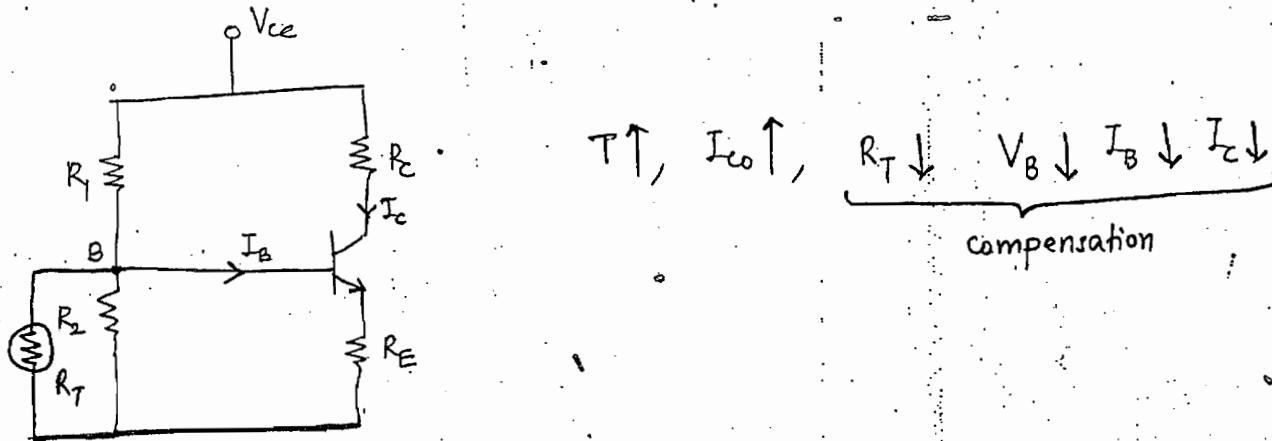
- (1). If $\frac{R_B}{R_E} \ll 1$, then $S \rightarrow \text{unity}$.
- (2). To make R_B less, take a condition $R_Y > R_Z$.
- (3). If R_E increases, -ve fb in the ckt increases which will affect the s/g gain in A_e analysis. So solve this problem always keep a capacitor (By pass capacitor C_E) in shunt to R_E .

Compensation Techniques -

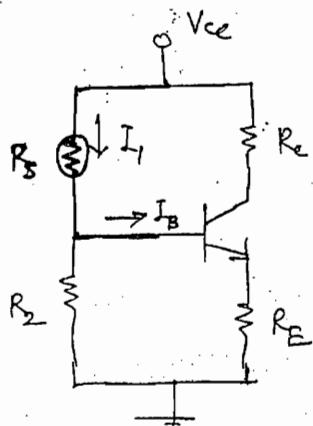
Compensation through Diode -



Compensation through thermistor:



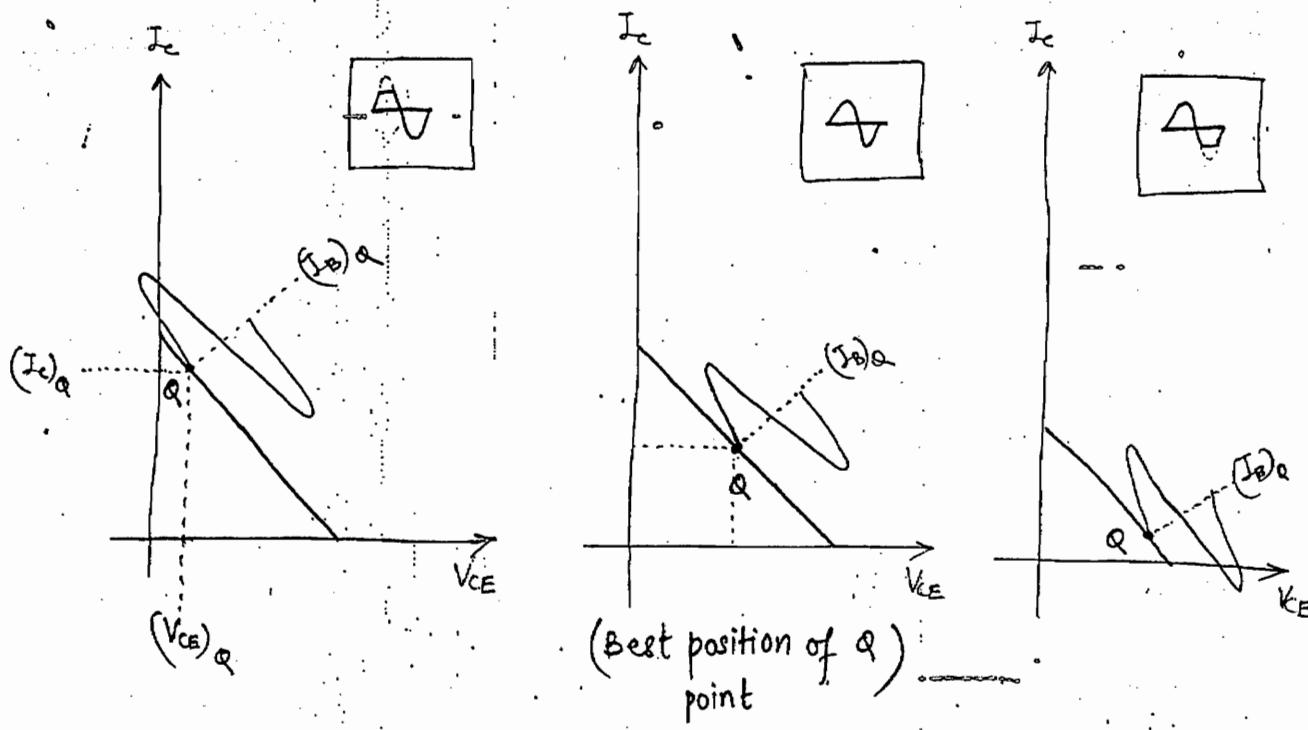
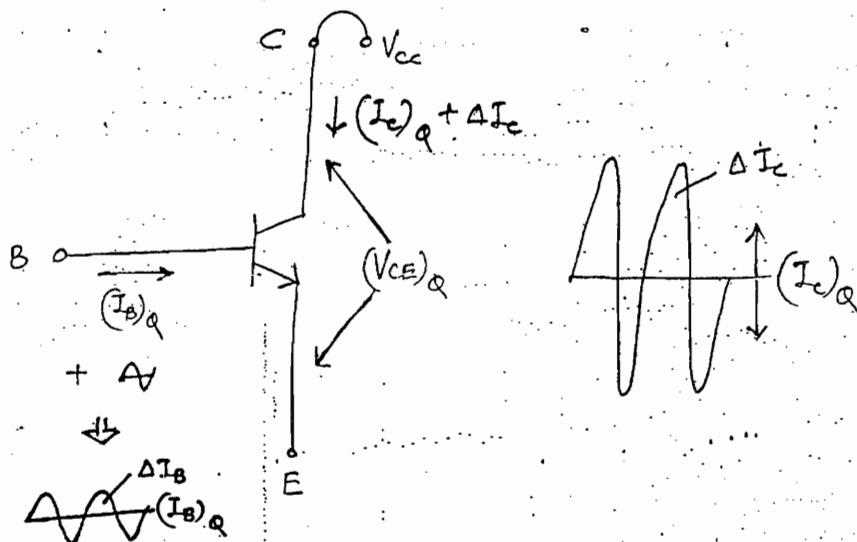
Compensation through Resistor:



$T \uparrow I_{Co} \uparrow R_s \uparrow I_c \downarrow, I_B \downarrow, I_E \downarrow$

compensation

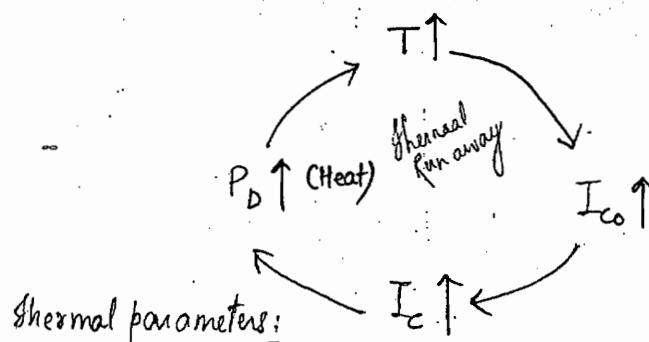
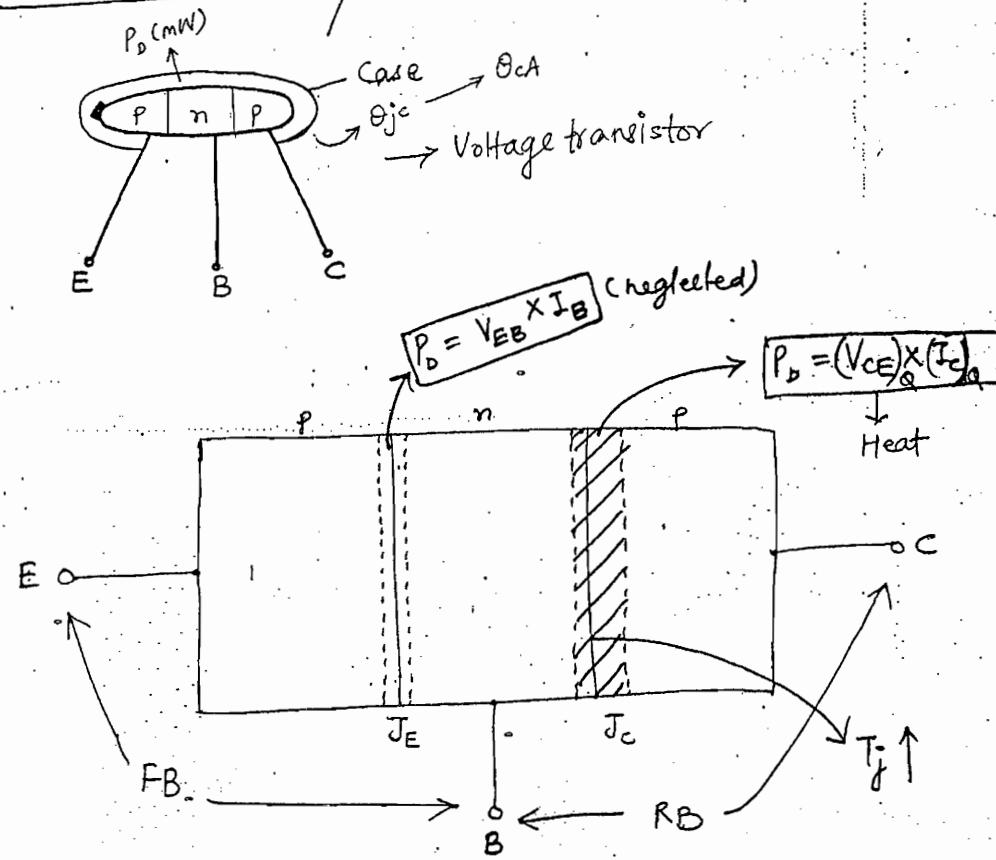
Position of Q point on DC load line:



Conclusion:

The amplifier analysis of the Q point should be design at the centre of DC load line.

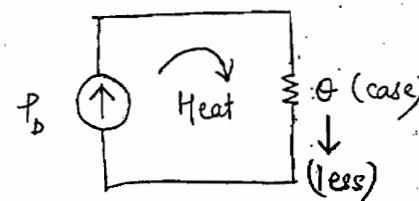
$$(V_{CE})_Q = \frac{V_{CC}}{2}$$

Thermal Runaway:Thermal parameters:

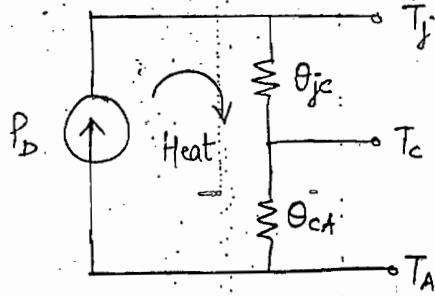
$$T_j - T_A \propto P_D$$

$$\frac{T_j - T_A}{P_D} = \theta_j$$

$$V = RI \quad (\text{analogue})$$



$$\theta = \text{thermal resistance} = \frac{T_j - T_A}{P_D} \text{ } ^\circ\text{C/mW}$$

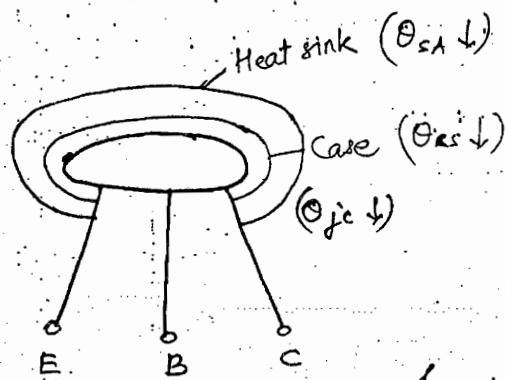


Problem with cases:

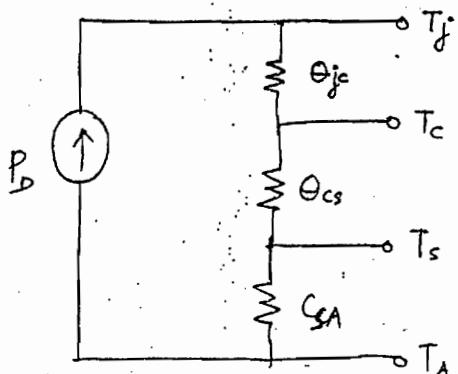
θ_{JC} can be less, but θ_{CA} is not much less, it is somewhat more. ($\theta_{JC} \downarrow$ & $\theta_{CA} \uparrow$)

Heat must be delivered to atmosphere simultaneously by absorbing from device.

Power transistor:



for power transistor,



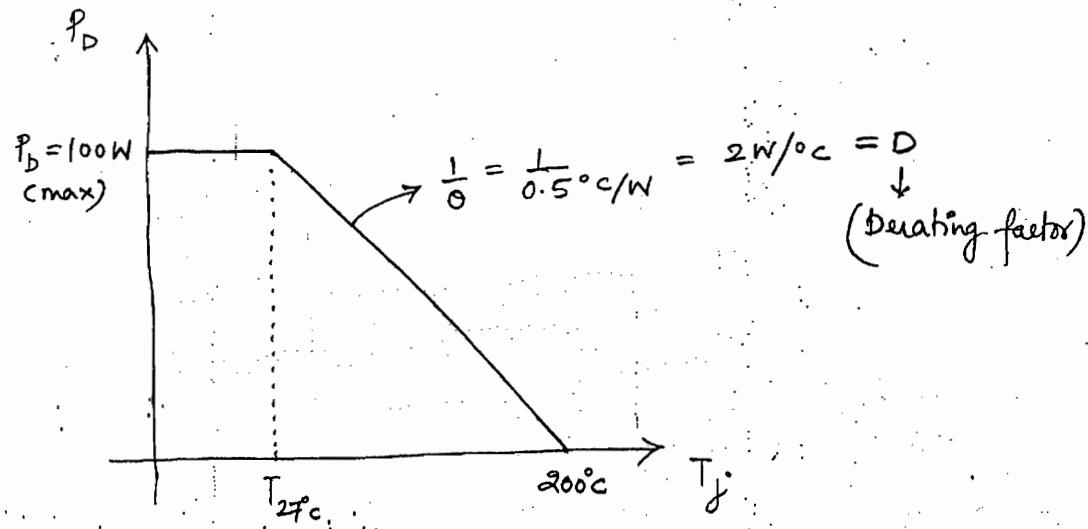
$$P_D = \frac{T_J - T_A}{\theta_{JC} + \theta_{CS} + \theta_{SA}}$$

for Voltage transistor,

$$P_D = \frac{T_J - T_A}{\theta_{JC} + \theta_{CA}}$$

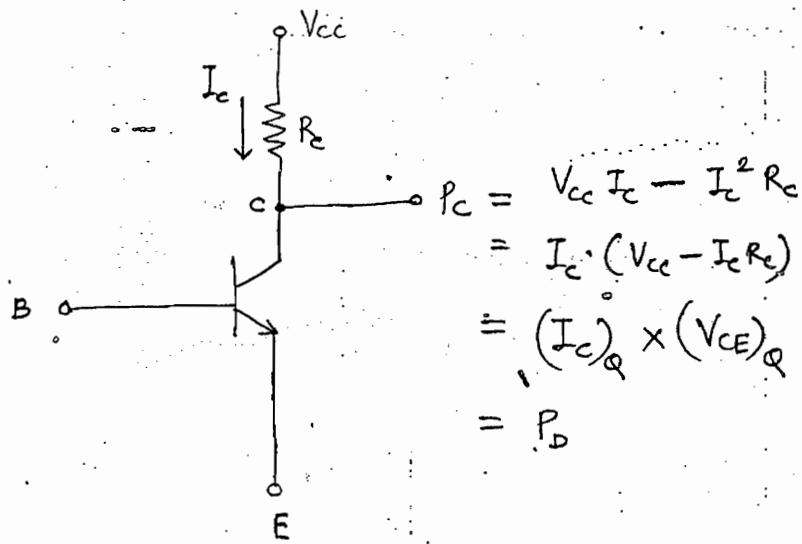
- * Heat sink can deliver heat as fast as case absorbs from the device. Hence, there is no problem if total $\theta_{PA} = \theta_{JC} + \theta_{CS} + \theta_{SA}$ to be less.

Specification graph for BJT -



Condition for thermal stability:

$$\frac{\partial P_c}{\partial T_j} < \frac{\partial P_D(\text{steady state})}{\partial T_j}$$



"The heat generated at the collector jxn should not exceed the heat dissipated at the collector jxn.
 (cmax)

$$T_j - T_A = \theta \cdot P_D$$

Differentiate w.r.t T_j . We get,

$$\frac{I_c}{\theta} = \Theta \cdot \frac{\partial P_D}{\partial T_j}$$

$$\frac{\partial P_D}{\partial T_j} = \frac{1}{\theta} = \text{slope}$$

$$\therefore \frac{\partial P_C}{\partial T_j} < \frac{1}{\theta}$$

$$(1) \quad \frac{\partial P_C}{\partial I_c} \cdot \frac{\partial I_c}{\partial I_{c0}} \cdot \frac{\partial I_{c0}}{\partial T_j} = \frac{\partial P_C}{\partial T_j}$$

Now, $P_C = V_{CC} I_c - I_c^2 R_C$

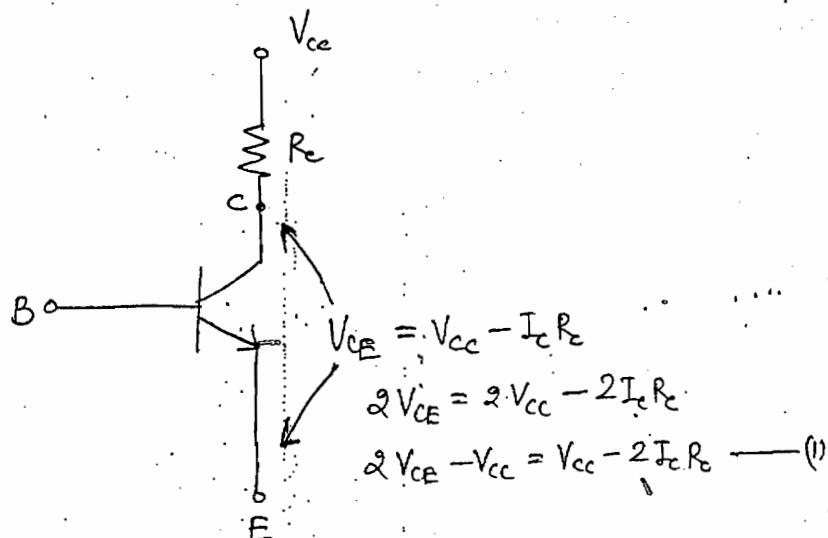
Differentiate w.r.t. I_c

$$(1). \quad \frac{\partial P_C}{\partial I_c} = V_{CE} - 2 I_c R_C$$

$$(2). \quad \frac{\partial I_c}{\partial I_{c0}} = S$$

$$(3). \quad \frac{\partial I_{c0}}{\partial T_j} = 0.07 I_{c0}$$

$$\therefore (V_{CC} - 2 I_c R_C) \times S \times (0.07 I_{c0}) < \frac{1}{\theta}$$



$$\therefore (2 V_{CE} - V_{CC}) (S) (0.07 I_{c0}) < \frac{1}{\theta}$$

Case - (1) : $(V_{CE})_Q = 0$

$$\underbrace{(-V_{CE}) \cdot (s) \cdot (0.07 I_C)}_{(-ve)} < \frac{1}{\theta} \downarrow (+ve)$$

\therefore thermally stable if $(V_{CE})_Q = 0$

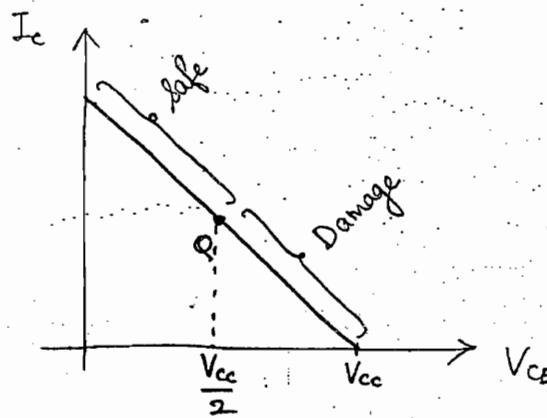
Case - (2) : $(V_{CE})_Q = V_{CC}/2$

$$0 < \frac{1}{\theta} \Rightarrow \text{thermally stable if } (V_{CE})_Q \leq V_{CC}/2$$

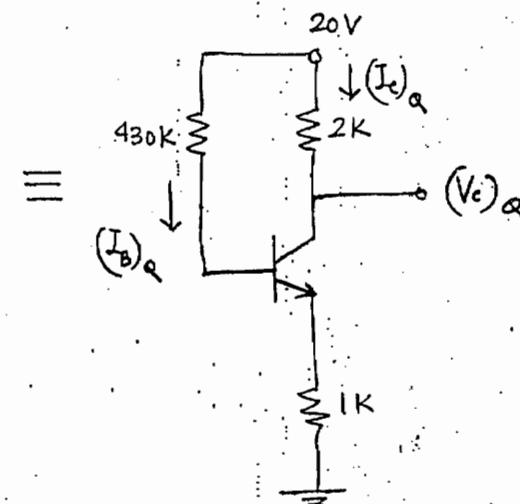
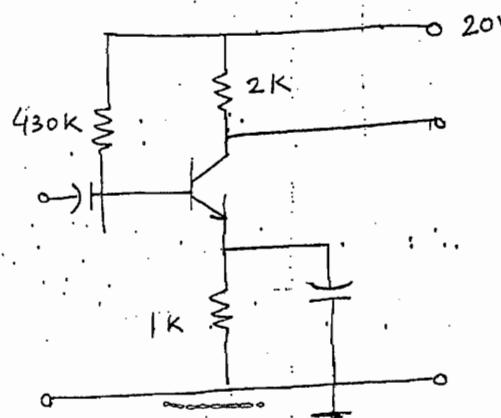
Case - (3) : $V_{CE} > V_{CC}/2$

There are chances of thermal runaway.

Thermal Runaway if $(V_{CE})_Q > V_{CC}/2$



Q9.

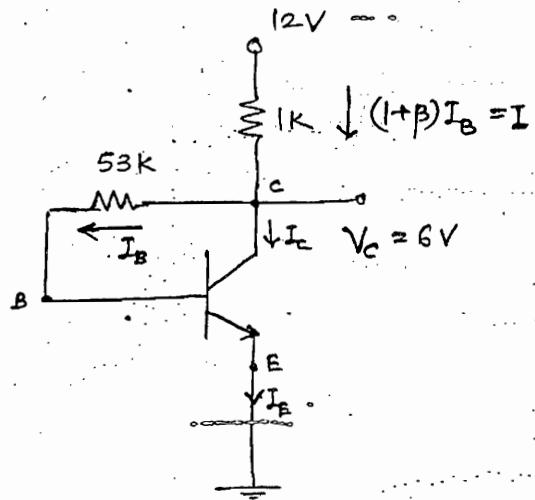
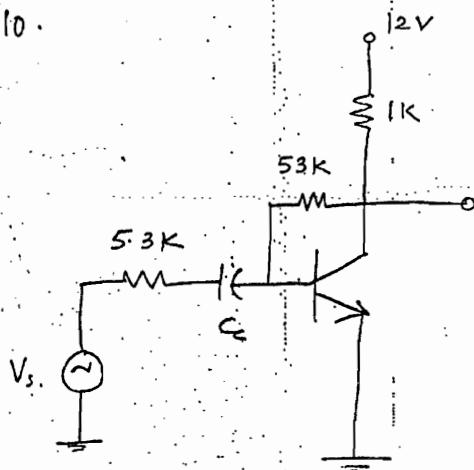


$$(I_B)_Q = \frac{20 - 0.7}{430K + (1+50)1K} = 40\mu A$$

$$\begin{aligned}(I_C)_Q &= \beta(I_B)_Q \\ &= 50 \times 40 \mu A \\ &= 2mA\end{aligned}$$

$$\begin{aligned}(V_C)_Q &= V_{CC} - (I_C)_Q \cdot R_E \\ &= 20 - 2mA \times 9K \\ &= 16V\end{aligned}$$

Q10.



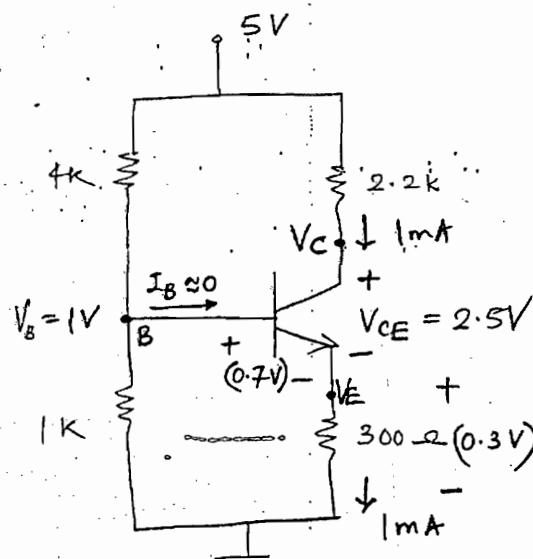
$$(I_B)_Q = \frac{12 - 0.7}{53K + (1+60)1K} = 99\mu A$$

$$\begin{aligned}(I_C)_Q &= \beta \cdot (I_B)_Q \\ &= 60 \times 99 \mu A \\ &= 5.94mA\end{aligned}$$

$$I_E = (1+60) \times 99 \mu A = 6mA$$

$$\begin{aligned}V_C &= 12 - 6mA \times 1K \\ &= 6V\end{aligned}$$

Q11.



$$\begin{aligned} V_{CE} &= 5 - 2.2 \times 1mA - 0.3 \times 1mA \\ &= 5 - 2.2 - 0.3 \\ &= 2.5V \end{aligned}$$

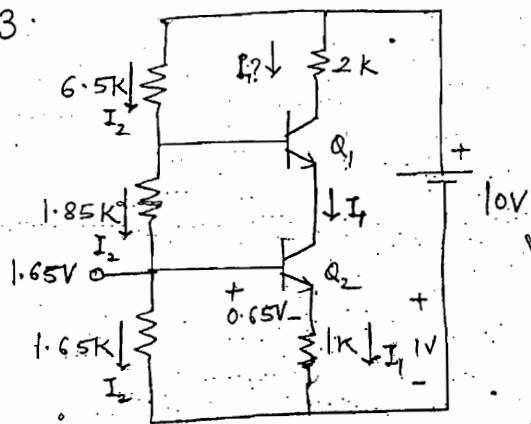
$$V_B = \frac{1}{5} \times 5 = 1V$$

$$V_E = 1 - 0.7 = 0.3V$$

$$I_E = \frac{0.3}{300} = 1mA$$

$$\therefore I_c = 1mA$$

Q3.



$B \rightarrow$ High value (given)

$$\uparrow B = \frac{I_c}{I_B}$$

Assume $I_B \approx 0$

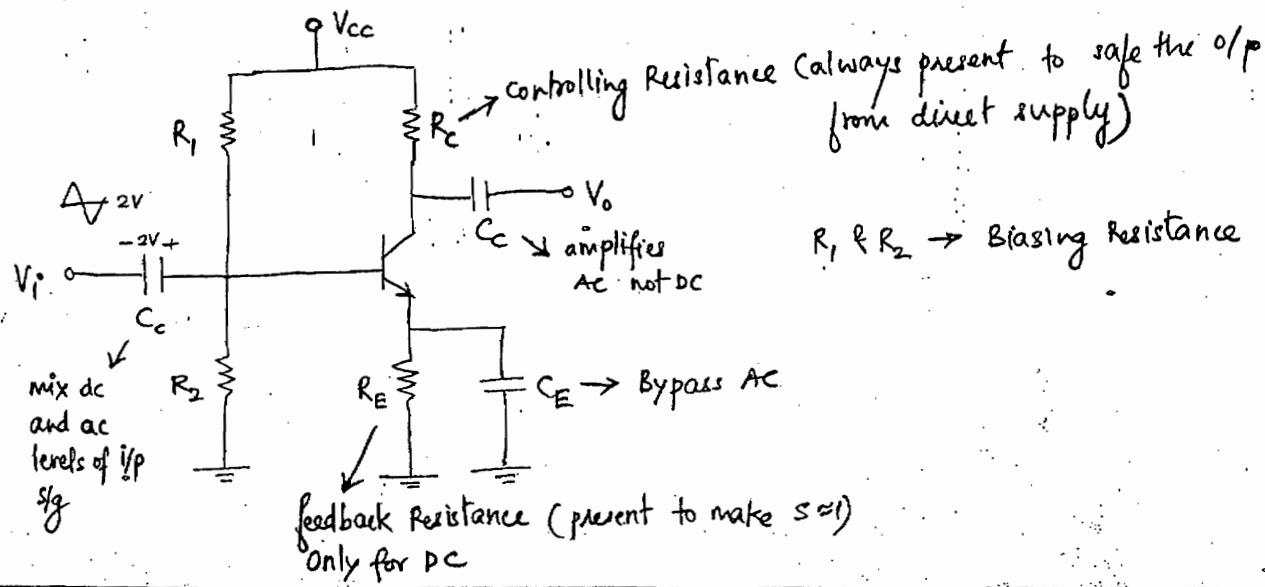
$$1.65 - 0.65 = 1V$$

$$I_q = \frac{1V}{1K} = 1mA$$

BJT Applications:

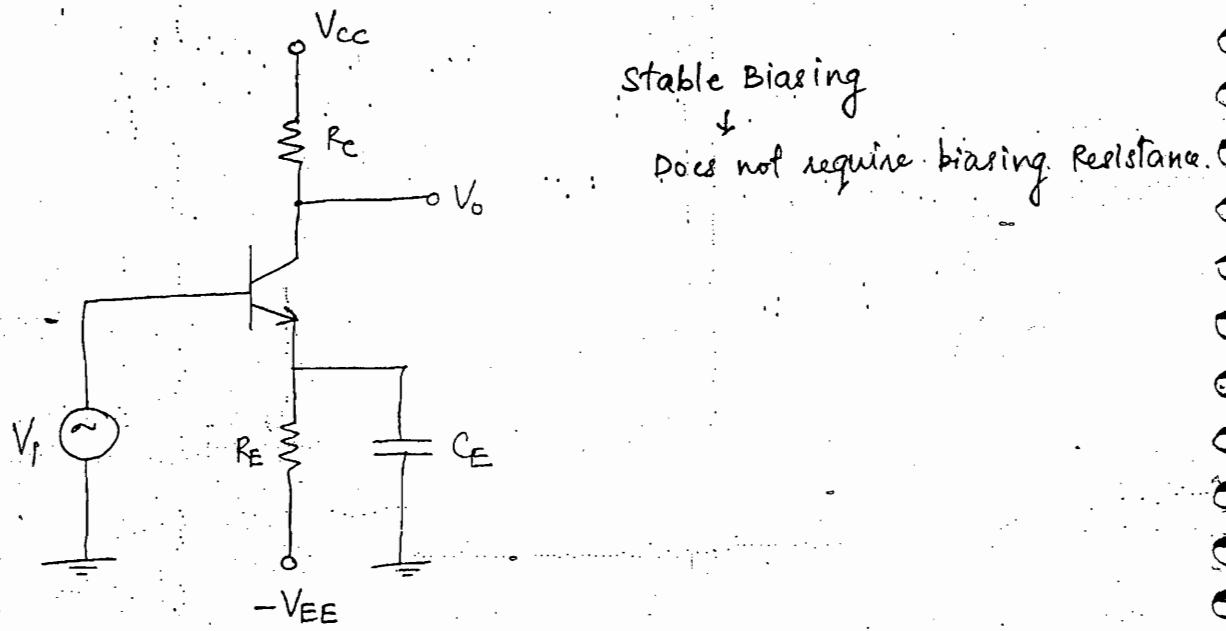
CE Amplifier:

RC coupled BJT Amplifier / AC amplifiers:



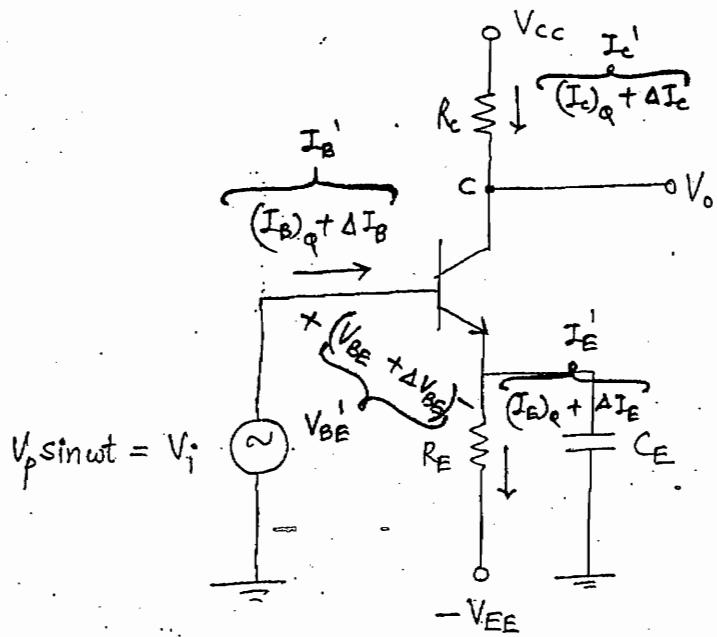
- (1). If it is amplifier DC wise.
- (2). App - Voltage Amplifier

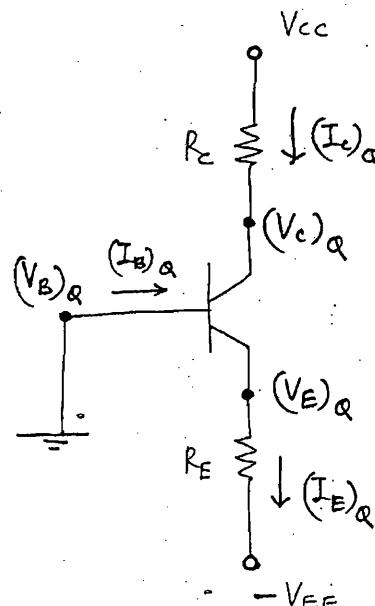
Direct Coupled BJT Amplifier (DC Amplifiers):



- (1). App - Integrated ckt (OP-Amp)
- (2). They have high gain (very high)
- (3). It is amplifier DC wise

Analysis of CE amplifier:



DC Model

$$(V_E)_Q = -V_{BE}$$

$$(V_B)_Q = 0$$

$$(V_C)_Q = V_{CC} - (I_C)_Q R_C$$

$$(I_E)_Q = \frac{V_{EE} - V_{BE}}{R_E}$$

$$(I_C)_Q = \alpha (I_E)_Q$$

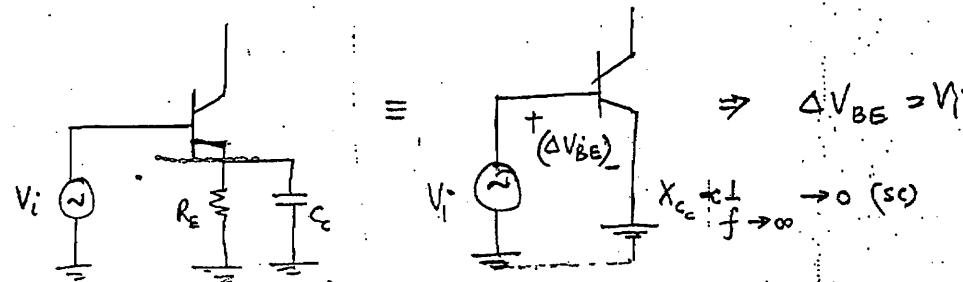
$$(I_B)_Q = \frac{(I_E)_Q}{(1 + \beta)}$$

$$(I_E)_Q = I_{E0} \cdot e^{V_{BE}/V_T}$$

$$(I_E)_Q + \Delta I_E = I_{E0} \cdot e^{(V_{BE} + \Delta V_{BE})/V_T}$$

$$(I_E)_Q + \Delta I_E = I_{E0} \cdot e^{V_{BE}/V_T} \cdot e^{\Delta V_{BE}/V_T}$$

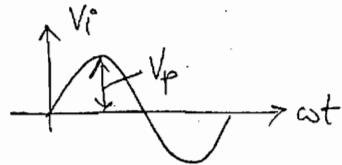
$$I'_E = (I_E)_Q \cdot e^{\Delta V_{BE}/V_T}$$

AC analysis -

$$I'_E = (I_E)_Q \cdot e^{V_i/V_T}$$

→ Non-linear device

Assume : $V_i = V_p \sin \omega t$



$$I_E' = (I_E)_Q \cdot e^{\frac{V_p \sin \omega t}{V_T}}$$

$$I_E' = (I_E)_Q \left\{ 1 + \frac{V_p \sin \omega t}{V_T} + \frac{1}{2!} \frac{V_p^2 \sin^2 \omega t}{V_T^2} + \dots \right\}$$

$$I_E' = (I_E)_Q \left\{ 1 + \frac{V_p \sin \omega t}{V_T} + \frac{V_p^2}{2V_T^2} \frac{(1 - \cos 2\omega t)}{2} + \dots \right\}$$

$$I_E' = (I_E)_Q \underbrace{\left(1 + \frac{V_p^2}{4V_T^2} \right)}_{\text{DC drift (unwanted)}} + \underbrace{\frac{(I_E)_Q \cdot V_p \sin \omega t}{V_T}}_{\text{fundamental comp. (useful)}} - \underbrace{\frac{(I_E)_Q \cdot V_p^2 \cos 2\omega t}{4V_T^2}}_{\text{Distortions (2nd Harmonic comp.)}} + \dots \text{etc.}$$

$$\% \text{ 2nd Harmonic distortion} = \frac{|B_2|}{|B_1|}$$

$$= \frac{(I_E)_Q \cdot V_p^2 / 4V_T^2}{(I_E)_Q \cdot V_p / V_T}$$

Harmonics are not
bcoz of DC

$$= \frac{V_p}{4V_T}$$

$$= \frac{V_p}{100 \text{ mV}} \quad (\text{at Room temp.})$$

s/g

$$\begin{cases} X_1 \rightarrow 1 \text{ mV} \sin \omega t \Rightarrow \% D = 1\% \\ X_2 \rightarrow 10 \text{ mV} \sin \omega t \Rightarrow \% D = 10\% \end{cases}$$

Harmonics are bcoz of s/g

$\therefore 1 \text{ mV}$ is a good choice for AC analysis.

Now, if

$$V_p \lll 4V_T$$

$$\frac{V_p}{4V_T} \lll 1 \quad (\text{neglected})$$

$$\therefore I_E' = (I_E)_Q + \left[\frac{(I_E)_Q}{V_T} \right] \cdot V_p \sin \omega t$$

And now,

$$I_C' = \alpha I_E'$$

$$= \alpha \cdot (I_E)_Q \cdot \left(1 + \frac{V_p \sin \omega t}{V_T} \right)$$

$$V_o = V_{CC} - I_C' R_C$$

$$= V_{CC} - \alpha (I_E)_Q R_C - \alpha \frac{(I_E)_Q}{V_T} R_C V_p \sin \omega t$$

$$= V_{CC} - \underbrace{(I_C)_Q R_C}_{(V_{CC})_Q} - \underbrace{\frac{(I_C)_Q}{V_T} R_C V_p \sin \omega t}_{V_i}$$

$$A_V = g_m R_C$$

$$V_o = (V_C)_Q - g_m R_C \cdot V_p \sin \omega t$$

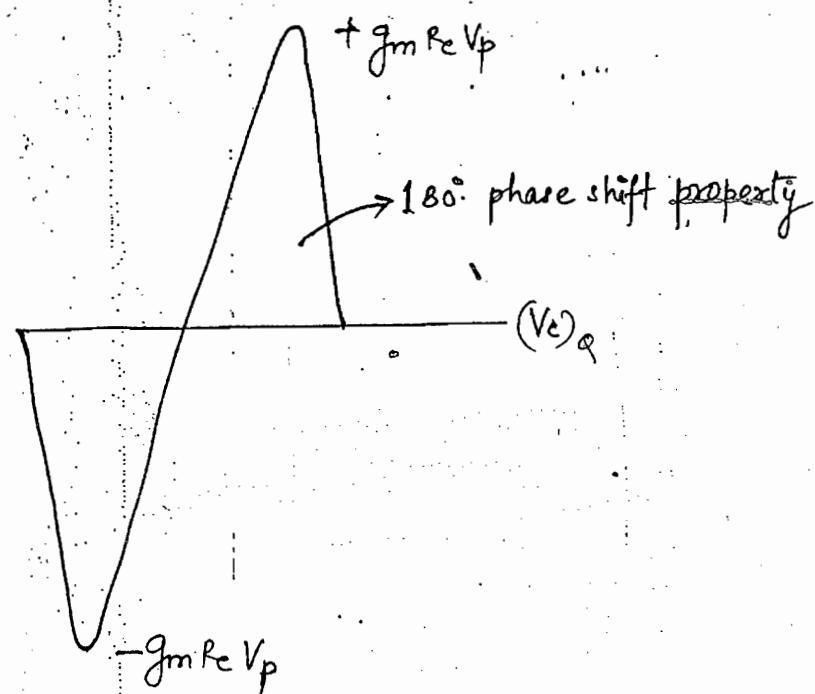
$$\text{amplification} \rightarrow A_V = g_m R_C \quad \text{and} \quad g_m = \frac{(I_C)_Q}{V_T} \propto (I_C)_Q$$

Hence, amplification in AC necessitates the requirement of DC i.e. $(I_C)_Q$ is because of DC and the factor which is responsible for gain.

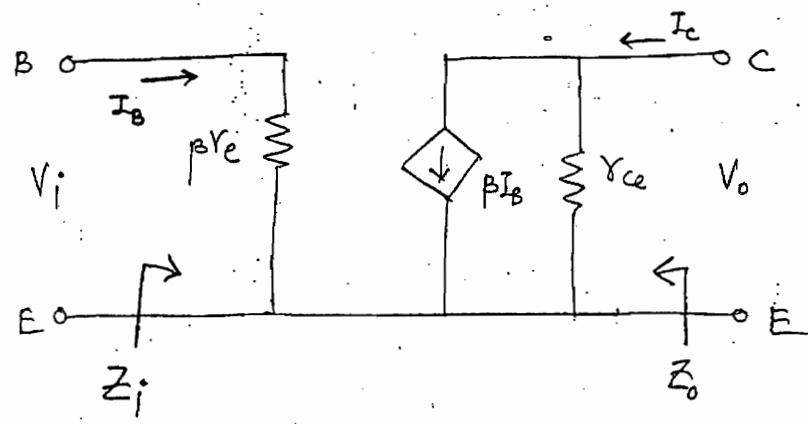
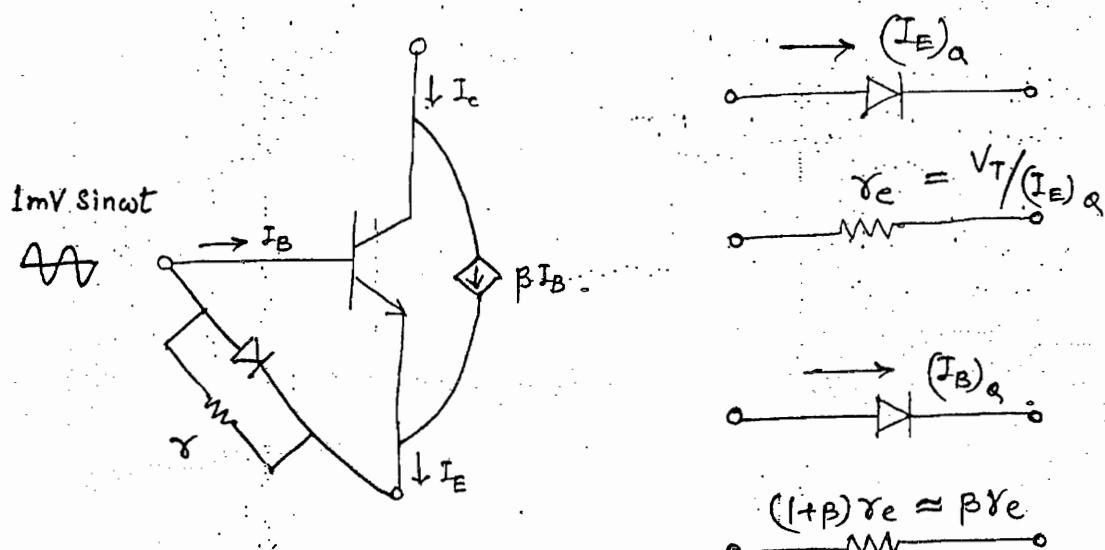
$$V_o = \underbrace{(V_C)_Q}_{\text{Direct coupled}} - \underbrace{g_m R_C \cdot V_p \sin \omega t}_{\text{RC coupled}}$$

$$\text{for RC coupled, } V_o = -g_m R_C \cdot V_p \sin \omega t$$

$$\text{for Direct coupled, } V_o =$$



Properties of CE amplifier -



$$(1) Z_i = \beta Y_e$$

Assume, $\beta = 100$
 $Y_e = 10 \Omega$

$$\therefore Z_i = 1 \text{ k}\Omega$$

$$(2) Z_o = Y_{ce} = 80 \text{ k}\Omega$$

$$(3) V_o = (-\beta I_B) \cdot Y_{ce}$$

$$= -\beta \cdot \frac{V_i}{\beta Y_e} \cdot Y_{ce}$$

$$\therefore V_o = -\frac{Y_{ce}}{Y_e} \cdot V_i$$

$$\therefore A_V = \frac{V_o}{V_i} = -\frac{Y_{ce}}{Y_e} > 1$$

180° phase shift

$$(4) A_I = \frac{I_E}{I_B} = \beta > 1$$

Conclusions:

- (1) The i/p impedance is moderate.
- (2) The o/p impedance is also moderate.
- (3) The voltage gain is greater than one.
- (4) The current gain is greater than one.
Hence, power gain is also greater than one.

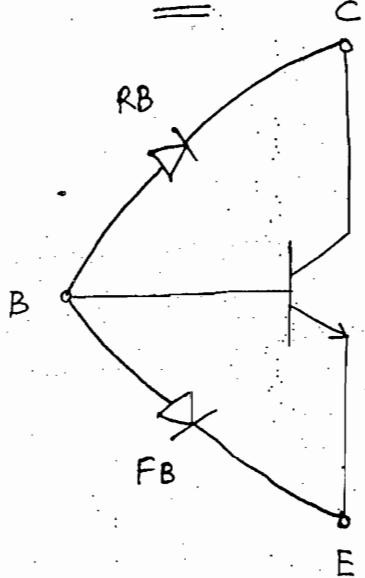
Note: In most of practical applications, CE is used because the power gain is high.

* 95% applications use CE configuration.

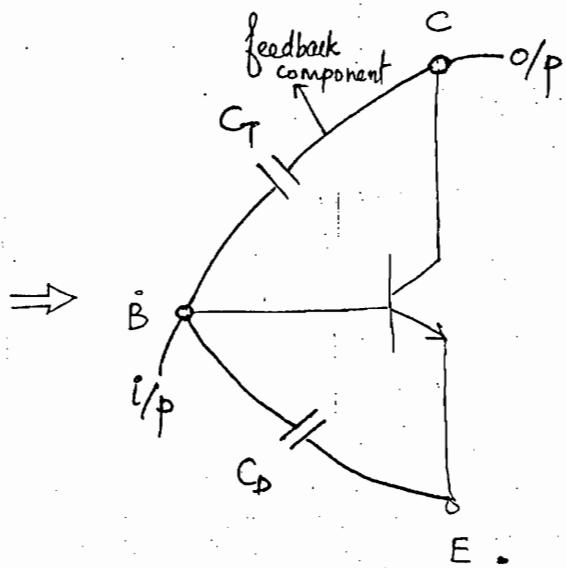
* CE is not used for high frequency s/g - CB is preferred in such cases.

CB amplifier in High frequency:

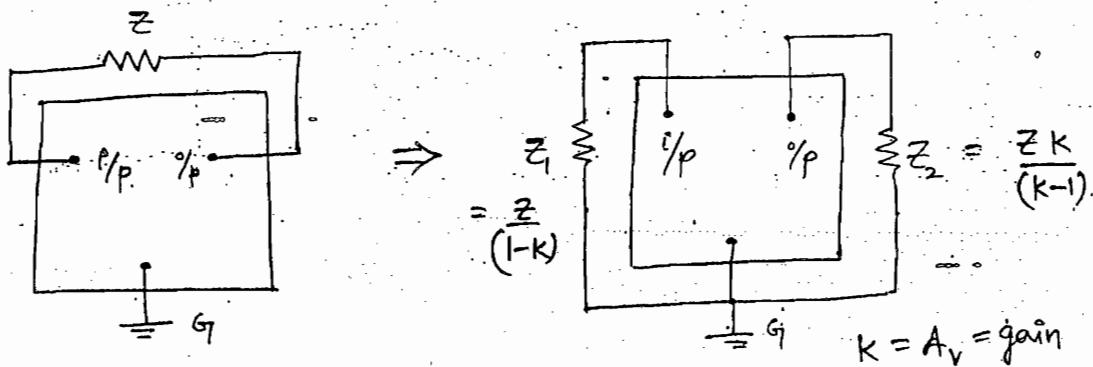
CE:



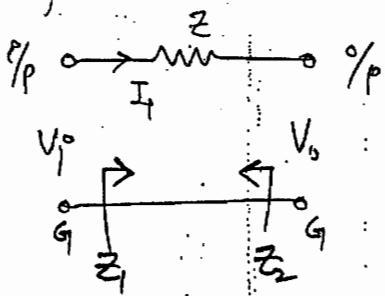
(Active)



Miller's theorem:



Proof:



$$I = \frac{V_i - V_o}{z} \quad \text{OR} \quad I = -\frac{V_o}{z} \left(1 - \frac{V_i}{V_o}\right)$$

$$z = \frac{V_o}{I} \left(1 - \frac{V_o}{V_i}\right) \quad z = z_2 \left(1 - \frac{1}{k}\right)$$

$$z = z_1 \left(1 - k\right)$$

$$z_2 = \frac{z k}{(k-1)}$$

$$Z_1 = \frac{Z}{(1-K)}$$

$$\frac{1}{C_{T_1}} = \frac{1}{G_T (1-A_V)}$$

$$C_{T_1} = G_T (1-A_V)$$

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

for high value of K (generally)
 $K-1 \approx K$

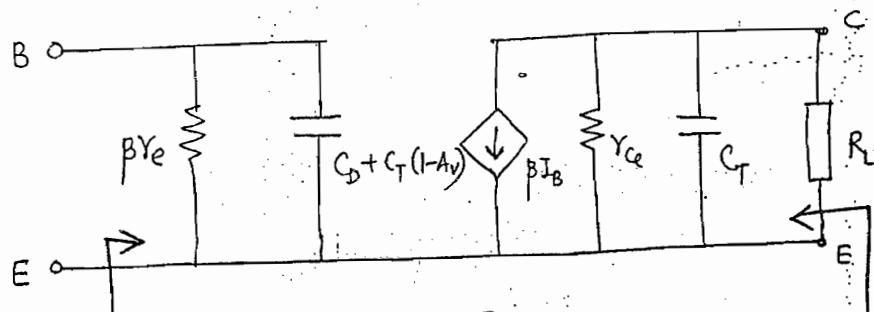
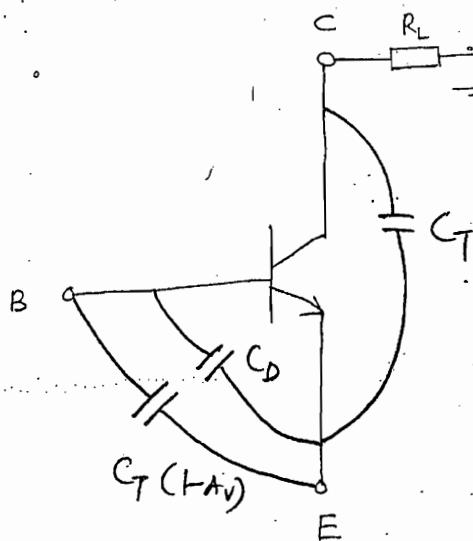
$$\therefore Z_2 = Z$$

$$\frac{1}{C_{T_2}} = \frac{1}{G_T}$$

$$C_{T_2} = G_T$$

$$C_{in} = C_D + C_T (1-A_V)$$

(high value)



$(R_C // R_L \approx R_E)$
 That's why
 R_E is not shown)

$$\uparrow T_o = (\beta Y_e) [C_D + C_T (1 - A_V)]$$

$$T_o = G_T R_L$$

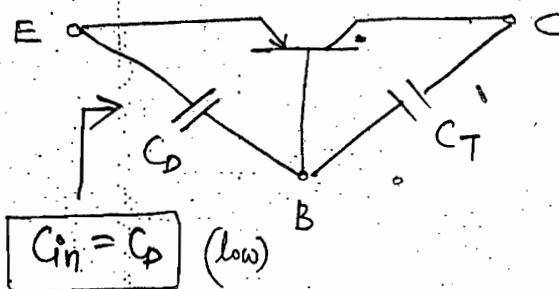
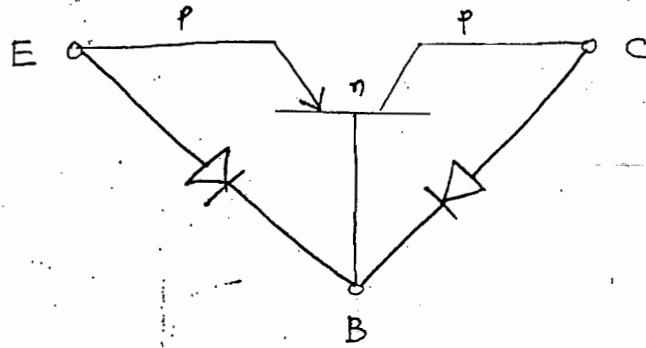
$$\omega_{BW} = \frac{1}{T_{total}}$$

$$T_{total} = T_i + T_o \approx T_i$$

$$\text{and } T_i \gg T_o$$

$$\therefore \omega_{BW} = \frac{1}{T_i} \uparrow \quad (\text{Hence, not preferred for high frequency})$$

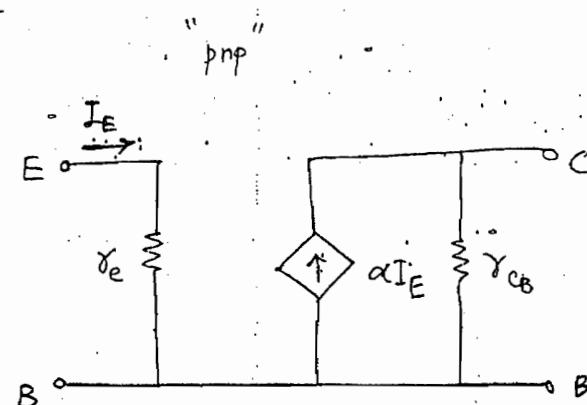
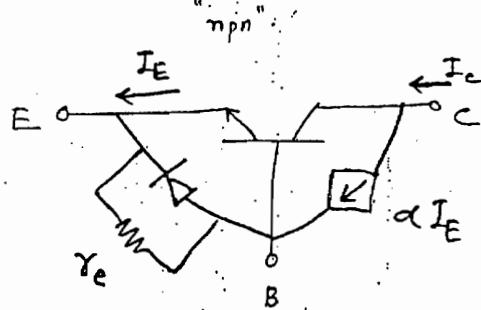
$$\omega_{BW} = \frac{1}{(\beta Y_e) [C_D + C_T (1 - A_V)]}$$

CB:

$$\uparrow \omega_{BW} = \frac{1}{T_C} \quad (\text{Hence, preferred for high frequency})$$

$$\omega_{BW} = \frac{1}{(B\gamma_e)(C_D)}$$

Properties of CB amplifier:



$$(1). Z_i = \gamma_e$$

$$\gamma_e \rightarrow (1 \Omega \text{ to } 25 \Omega)$$

$$Z_i \approx 10 \Omega$$

$$(2). Z_o = \gamma_{CB}$$

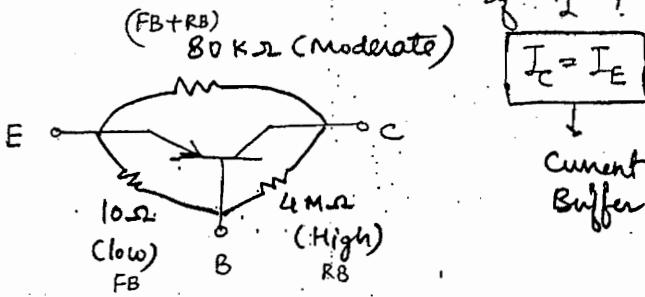
$$= 4 M\Omega$$

$$(3) A_V = \frac{\gamma_{CB}}{\gamma_e} > 1 \text{ (No phase shift)}$$

$$= \frac{4 M\Omega}{10\Omega}$$

$$\approx 400 \text{ K}\Omega$$

$$(4) A_I = \frac{I_C}{I_E} = d \leq 1 \quad 123$$



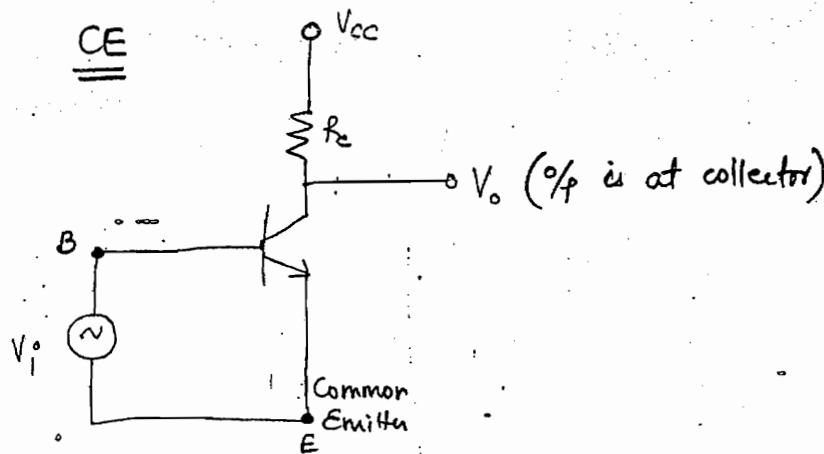
Conclusions:

- (1) The o/p impedance is low.
- (2) The i/p impedance is high.
- (3) The voltage gain is greater than one
- (4) The current gain is less than or equal to one (buffer).

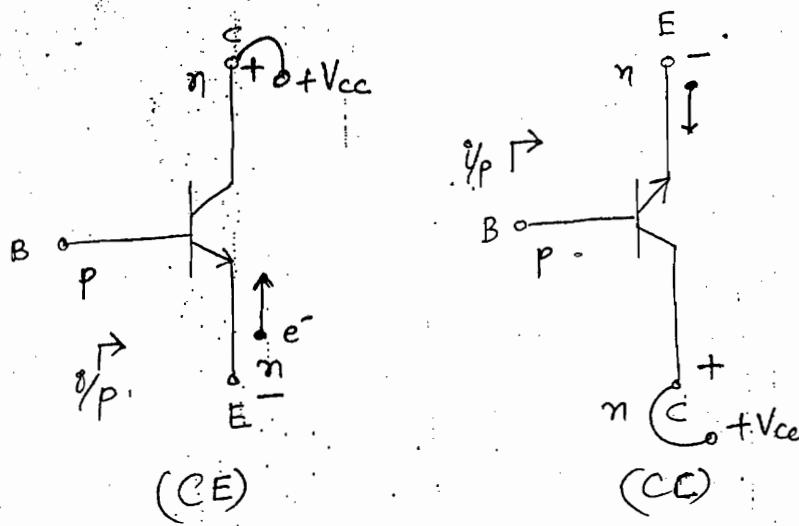
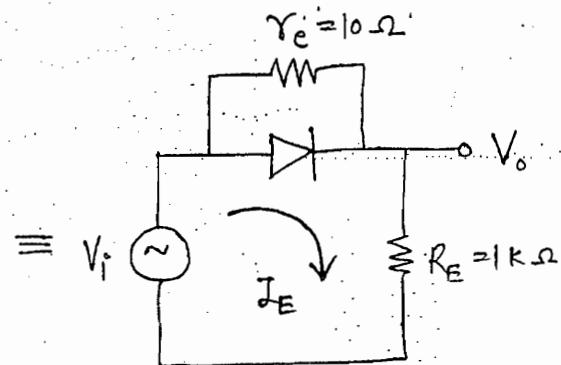
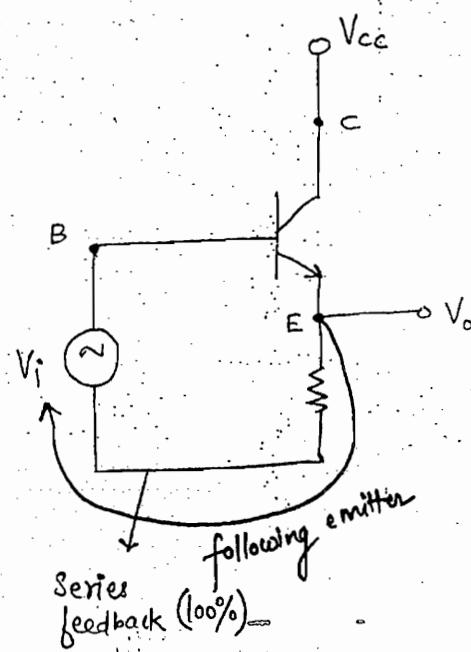
Note: In practical applications, c_B amplifier is used whenever high BW is required.

* Power gain is not much high bcoz of $A_I \leq 1$.

CC Amplifier: (or) Emitter follower



for CC, o/p must be at emitter (simultaneously load will also be at emitter)

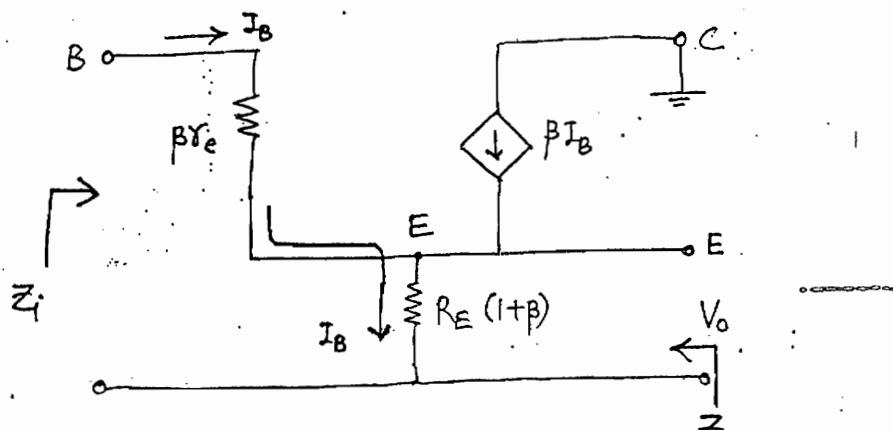
**CC:**

$$V_o = \frac{R_E}{R_E + \gamma_e} \times V_i$$

But $R_E \gg \gamma_e$

$$\therefore V_o = V_i \rightarrow \text{Voltage Buffer}$$

Properties of CC amplifier:



$$(1) \cdot Z_i = \beta r_e + (1+\beta) R_E$$

Assume, $\beta = 100$

$$r_e = 10 \Omega$$

$$R_E = 10 k\Omega$$

$$\begin{aligned} Z_i &= (100)(10) + (100)(10k) \\ &= 10k + 10^3 k\Omega \\ &\approx 10^3 k\Omega \approx 1 M\Omega \end{aligned}$$

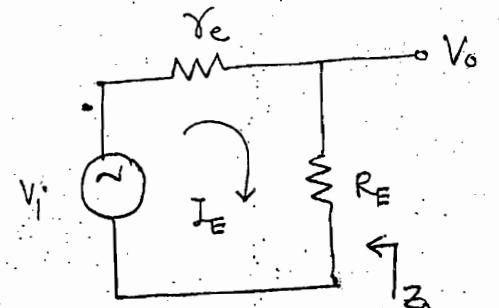
i.e. $Z_i = \beta R_E$

$$(2) \cdot Z_o = \frac{V_o}{I_E}$$

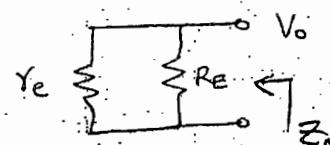
$$Z_o = r_e // R_E$$

$$\approx r_e$$

$$\approx 10 \Omega$$



$$\Downarrow V_i = 0$$



$$(3) \cdot A_V = \frac{V_o}{V_i} \leq 1$$

$$(4) \cdot A_I = \frac{I_E}{I_B}$$

$$= 1 + \beta$$

$$\approx \beta > 1$$

Conclusions:

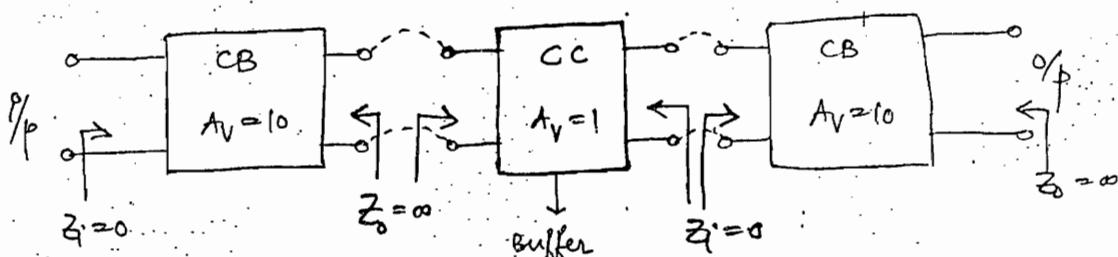
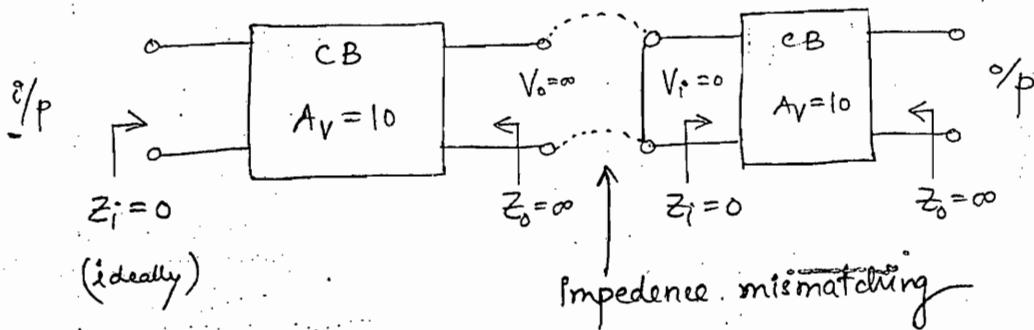
- (1) The i/p impedance is high.
- (2) The o/p impedance is low (Impedance matching)
- (3) The voltage gain is unity (Voltage Buffer)
- (4) The current gain is greater than unity.

Note: In practical applications, cc is used for "impedance matching".

* Here also, power gain is not much high.

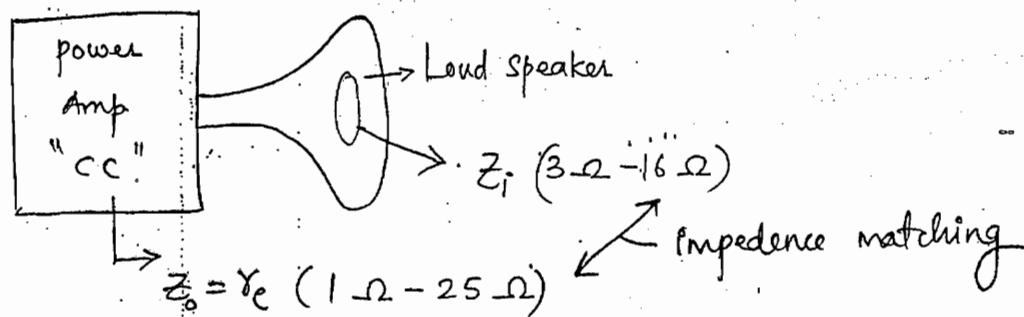
Common collector in Impedance Matching:

App-(1)



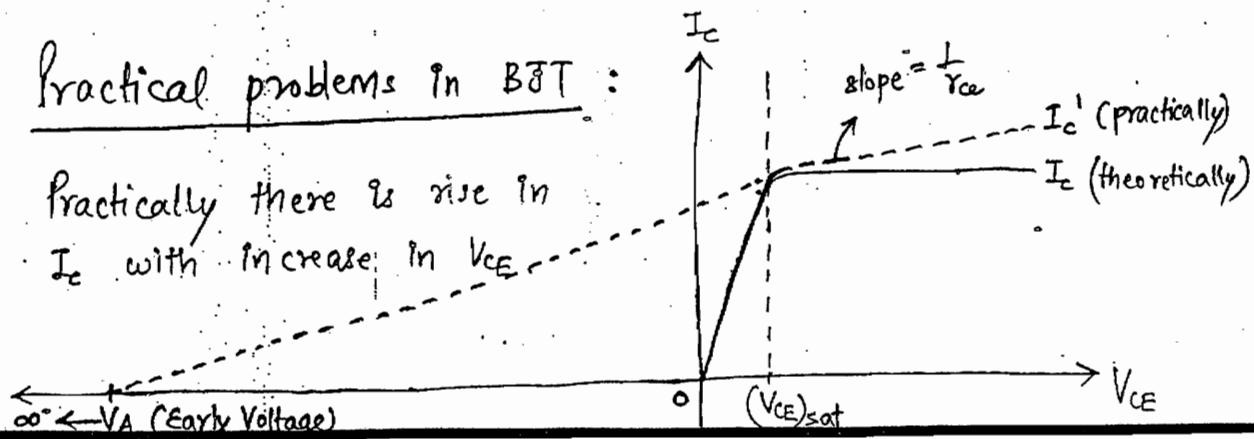
App-(2)

(Impedance Matching)

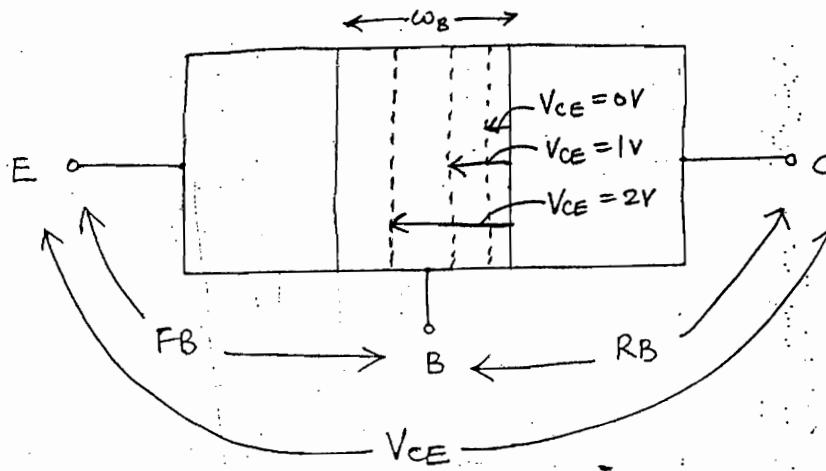


Practical problems in BJT:

Practically there is rise in I_c with increase in V_{CE} .



Early Effect (on Base width modulation) : (early \rightarrow Scientist name)



As $V_{CE} \uparrow$, $\omega_B \downarrow$, $I_R \downarrow$, $I_B \downarrow$, $I_{EP} \uparrow$, $I_E \uparrow$, $I_{cp} \uparrow$, $I_c \uparrow$

$$I_c = I_0 \cdot e^{\frac{V_{BE}}{V_T}} \rightarrow \text{Theoretical eqn}$$

$$I_c' = I_0 \cdot e^{\frac{V_{BE}}{V_T} \left(1 + \frac{V_{CE}}{V_A} \right)} \rightarrow \text{Practical eqn}$$

%p Resistance:

$$\gamma_{\bar{Q}} = \gamma_0 = \frac{\partial V_{CE}}{\partial I_c}$$

$$\left[\gamma_e = \frac{V_T}{(I_E)_Q} \right]$$

$$\frac{1}{\gamma_0} = \frac{\partial I_c'}{\partial V_{CE}} = \frac{\partial}{\partial V_{CE}} \left[I_0 \cdot e^{\frac{V_{BE}}{V_T} \left(1 + \frac{V_{CE}}{V_A} \right)} \right]$$

$$= \frac{I_0 \cdot e^{\frac{V_{BE}}{V_T}}}{V_A}$$

$$= \frac{(I_c)_Q}{V_A}$$

$$\therefore \gamma_0 = \frac{V_A}{(I_c)_Q}$$

Now, Assume: $V_A = 1000$, $(I_c)_Q = 1mA$

$$\therefore \gamma_0 = \frac{1000 \text{ V}}{1 \text{ mA}} = 1 \text{ M}\Omega$$

* Early voltage (V_A) tells the quality of BJT (good or bad).

Analysis - (2) : Mathematical Analysis

$$I_c = I_s \cdot e^{V_{BE}/V_T}$$

(I_s or $I_0 \rightarrow$ saturation current)

$$I_s = \frac{A q D_B n_p(0)}{w_B}$$

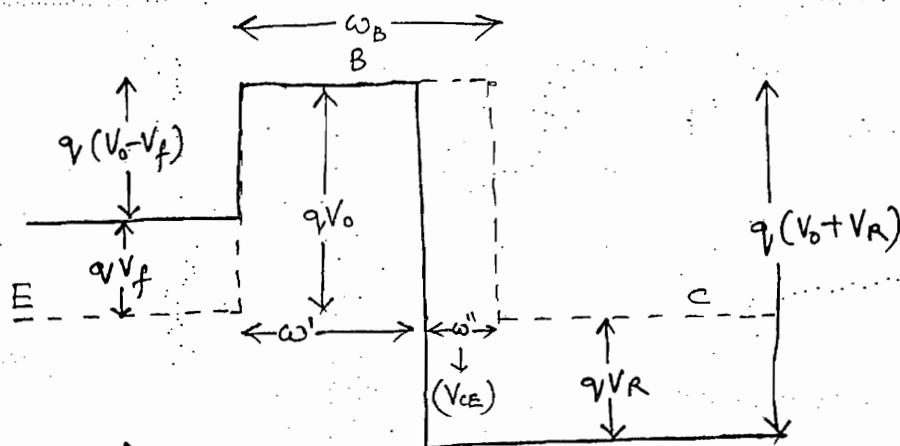
Injected minority current (or) diffusion current

$$I_s \propto \text{Area}$$

$$I_s \propto \frac{1}{w_B}$$

As $V_{CE} \uparrow$, $w_B \downarrow$, $I_s \uparrow$, $I_c \uparrow$

Analysis - (3) : Energy Band theory



As $V_{CE} \uparrow$,

$$w_B = w' \downarrow + w'' \uparrow$$

↓
const.

Early Effect

Effective base width

$w' \rightarrow$ (mobile charge width)

$w'' \rightarrow$ (immobile charge width)
penetration width

$w_B \rightarrow$ physical width of base

$$\boxed{w_B = w''}$$

$$\boxed{w' = 0}$$

→ punch through
(or)

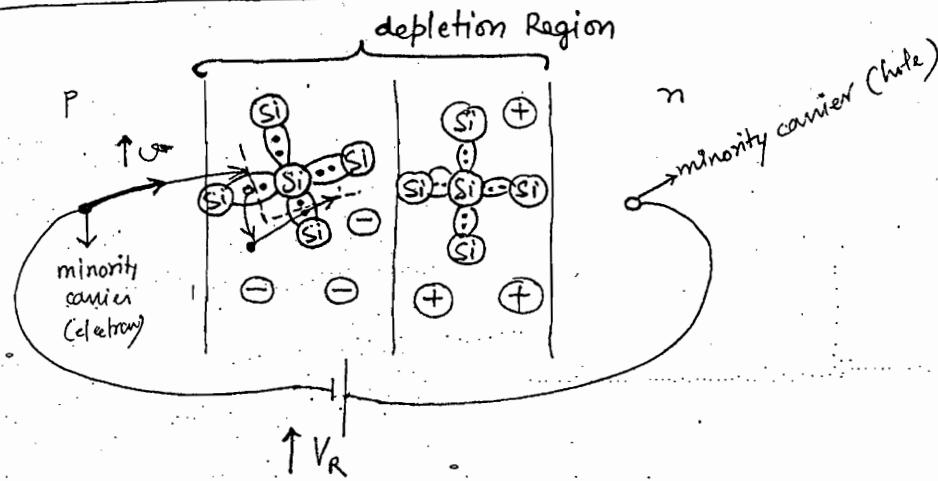
Reach through

Special Breakdown

* Punch through takes place due RB application.

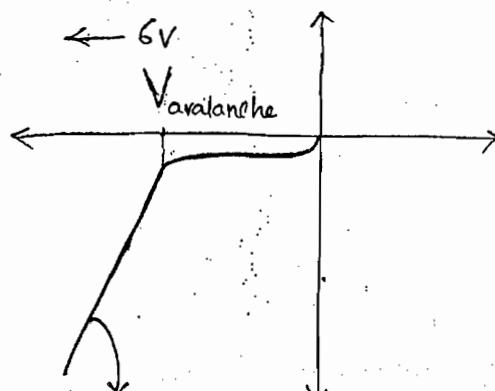
- (1). Thermal runaway.
- (2). Punch through (or) Reach through.
- (3). Avalanche breakdown.

Avalanche breakdown in a diode:



Explanation:

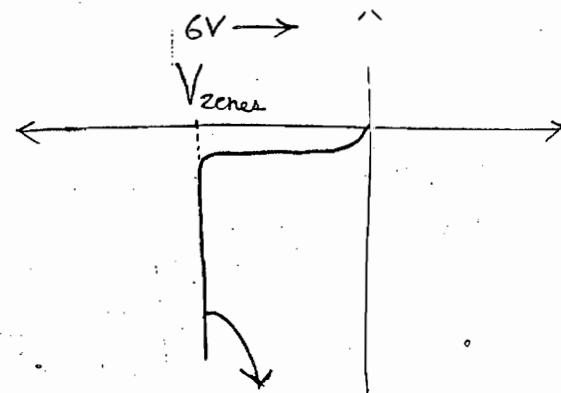
- (1). In the depletion region, there will be some stable atoms along with immobile ions.
 - (2). When the RB voltage increases, the velocity of minority charge carrier increases which will break the covalent bond in the depletion region.
 - (3). Because of this breakdown new electron hole pairs will be generated which will increase the minority population, this type of mechanism is called as avalanche breakdown. (Avalanche multiplication).
- * Becoz of high level of doping in zener, breakdown takes place by electric field applied (RB) directly. There is no such multiplication of minority charge carriers.



Linear Breakdown

(or)

Avalanche Breakdown



Instant Breakdown

(or)

Zener Breakdown

(Very fast phenomena)

Rapid current

$$V_Z < 6V$$

Above 6V, Avalanche \rightarrow

CB

I_C

I_{CBO}

BV_{CBO}

V_{CB}

Damage later (slow current)

$$I_{CEO} > I_{CBO} \quad [I_{CEO} = (\eta\beta) I_{CBO}]$$

$$BV_{CEO} < BV_{CBO}$$

$$BV_{CEO} = \frac{BV_{CBO}}{\sqrt{\beta}}$$

CE

I_C

I_{CEO}

BV_{CEO}

(Rapid current)

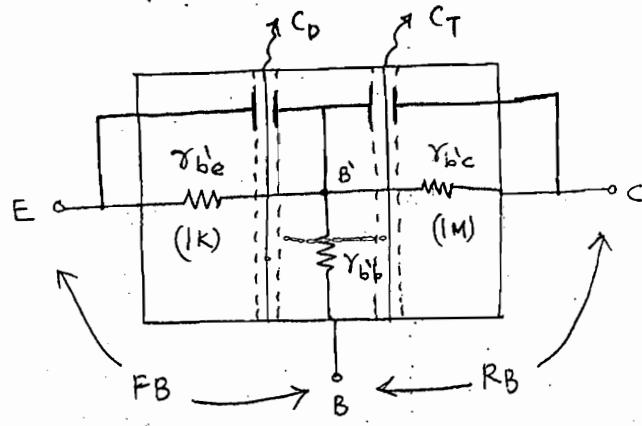
Damage soon

$\beta \rightarrow$ current gain in CE

$\eta \rightarrow$ avalanche multiplication factor

$\eta \rightarrow (3 \text{ to } 6)$

Low frequency Analysis : $\infty f \rightarrow 0$

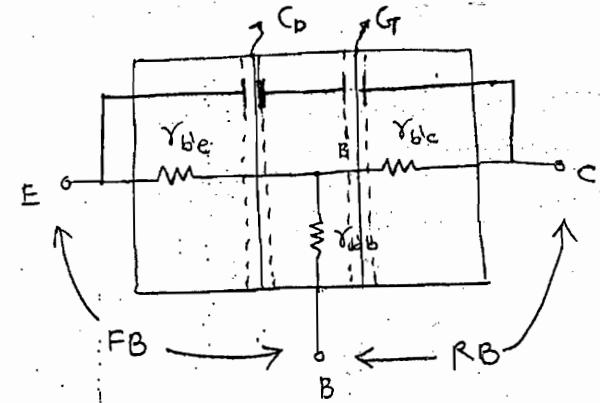


$$X_C \propto \frac{1}{f} \rightarrow \infty \text{ (OC)}$$

At low frequency analysis,
C_D & G_T are assumed to
be open (neglect).

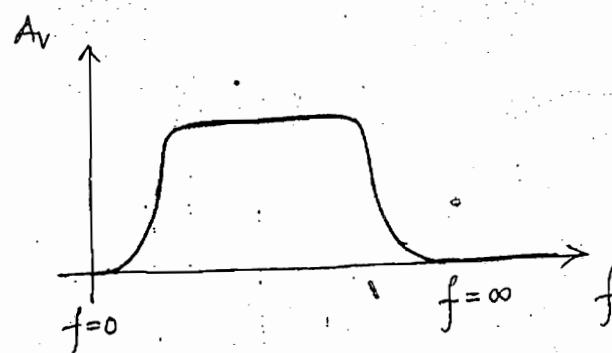
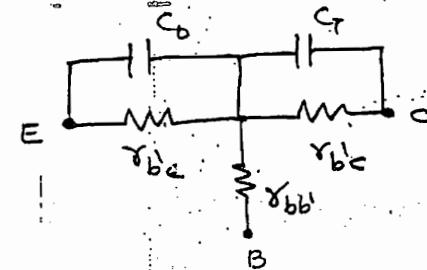
High frequency Analysis

$$\text{AC } f \rightarrow \infty$$



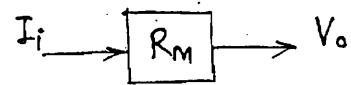
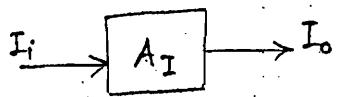
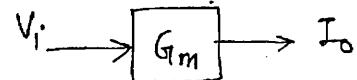
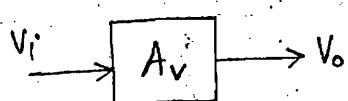
$$X_C \propto \frac{1}{f} \rightarrow 0 \text{ (SC)}$$

At high frequency analysis,
C_D & G_T are taken in
to account (practically).

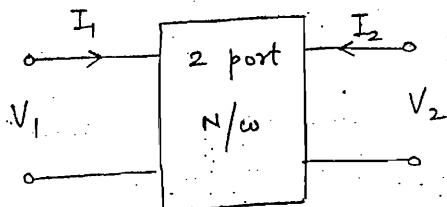


* Types of Basic amplifiers :

- (1). Voltage amplifier
- (2). Current amplifier
- (3). Transconductance amplifier
- (4). Tranresistance amplifier



Linear 2 port N/W -



Z parameters:

$$V_1 = Z_i I_1 + Z_{xy} I_2$$

$$V_2 = Z_f I_1 + Z_o I_2$$

$$V_1 = f(I_1, I_2)$$

$$V_2 = f(I_1, I_2)$$

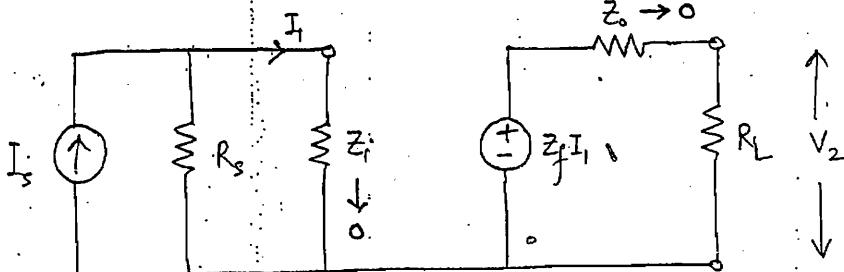
$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_i & Z_{xy} \\ Z_f & Z_o \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

An ideal amplifier is a unilateral device

$$\therefore Z_y = 0 ; Z_i = 0 ; Z_o = 0$$

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ Z_f & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

$$V_2 = Z_f \cdot I_1 \rightarrow \text{"CCVS"}$$



Transresistance Amplifier "CCVS"

y parameters:

$$I_1 = f(V_1, V_2)$$

$$I_2 = f(V_1, V_2)$$

$$I_1 = Y_i V_1 + Y_o V_2$$

$$I_2 = Y_f V_1 + Y_o V_2$$

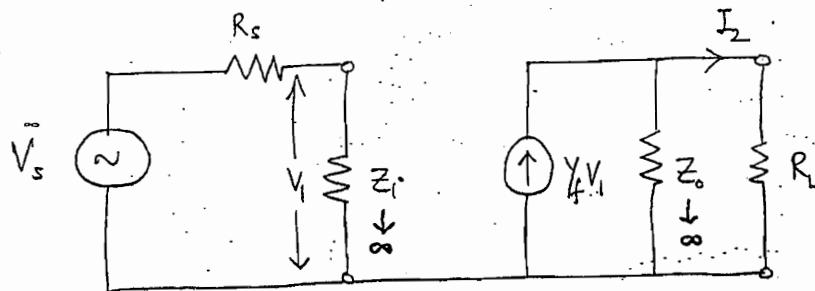
$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_i & Y_o \\ Y_f & Y_o \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$

An ideal amplifier is a unilateral device

$$Y_i = 0 ; Y_o = 0 ; Y_f = 0$$

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ Y_f & 0 \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$

$$I_2 = Y_f V_1 \rightarrow "VCCS"$$



Trans conductance amplifier "VCCS" (FET/MOSFET).

H parameters:

$$V_1 = f(I_1, V_2)$$

$$I_2 = f(I_1, V_2)$$

$$V_1 = h_{i1} I_1 + h_{r1} V_2$$

$$I_2 = h_{f1} I_1 + h_{o1} V_2$$

$$\text{Case (1)} : V_2 = 0$$

$$\text{Case (1)} : V_2 = 0$$

$$h_i = \frac{V_1}{I_1} \rightarrow \text{i/p impedance}$$

$$h_f = \frac{I_2}{I_1} \rightarrow \text{forward current gain}$$

$$\text{Case (2)} : I_1 = 0$$

$$h_r = \frac{V_1}{V_2} \rightarrow \text{Reverse voltage gain}$$

$$h_o = \frac{I_2}{V_2} \rightarrow \text{o/p admittance}$$

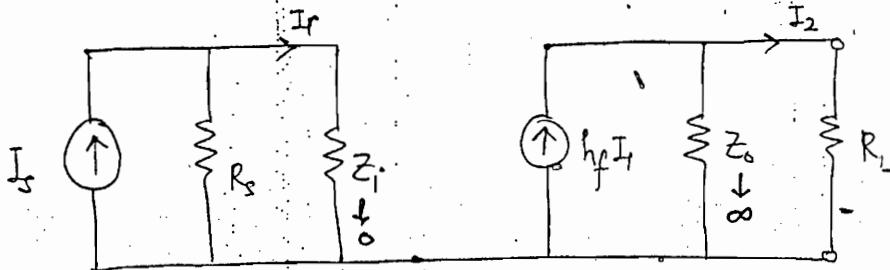
An ideal amplifier is a unilateral device

$$h_i = 0; \quad h_v = 0; \quad h_o = 0$$

$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ h_f & 0 \end{bmatrix} \begin{bmatrix} I \\ V_2 \end{bmatrix}$$

$$I_2 = h_f I_1 \rightarrow \text{"CCCS"}$$

$$I_c = \beta I_b$$



Current Amplifier "CCCS" (BJT)

g parameters:

$$I_1 = f(V_1, I_2)$$

$$V_2 = f(V_1, I_2)$$

$$I_1 = g_f V_1 + g_o I_2$$

$$V_2 = g_f V_1 + g_o I_2$$

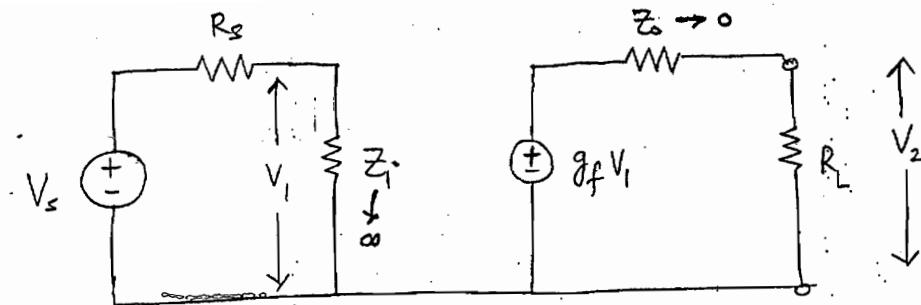
$$\begin{bmatrix} I_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} g_f & g_o \\ g_f & g_o \end{bmatrix} \begin{bmatrix} V_1 \\ I_2 \end{bmatrix}$$

An ideal amplifier is a unilateral device

$$g_i = 0; \quad g_o = 0; \quad g_o = 0$$

$$\begin{bmatrix} I_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ g_f & 0 \end{bmatrix} \begin{bmatrix} V_1 \\ I_2 \end{bmatrix}$$

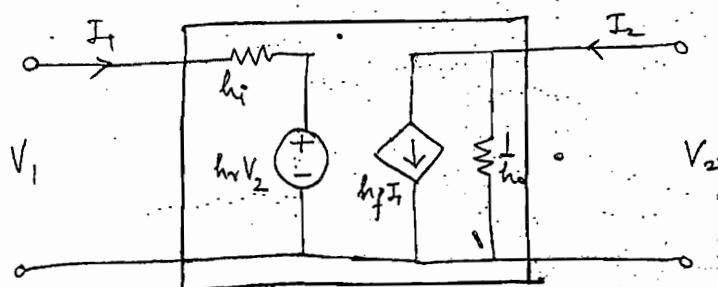
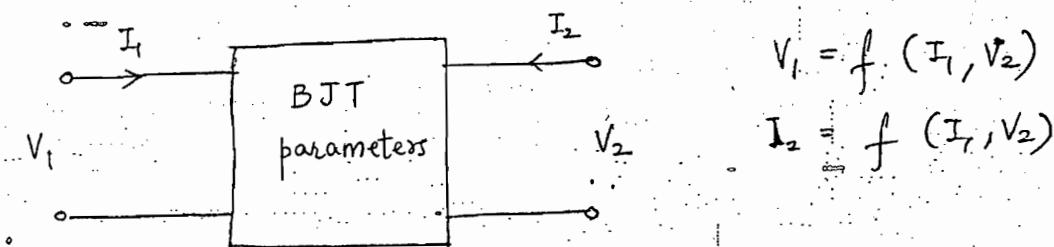
$$V_2 = g_f V_1 \rightarrow \text{"VCVS"}$$



Voltage Amplifier "VCVS" (Op-Amp)

Low frequency analysis of BJT :

BJT as a h-parameter model :



Low frequency model of BJT :

Typical values of "h-parameters :

h-parameters	CE.	CC	CB	
h_i	h_{ic} last letter of CE	1100Ω	1100Ω	22Ω
h_f	h_{fe} last letter of CE	50	-51	-0.98
h_r	h_{re} last letter of CE	2.4×10^{-4}	≈ 1	2.9×10^{-4}
h_o	h_{ob} last letter of CB	$24 \times 10^{-6} \text{ mho}$	$24 \times 10^{-6} \text{ mho}$	$0.49 \times 10^{-6} \text{ mho}$

Conversion formulae:

CE to CB

$$h_{eb} = \frac{h_{ie}}{1+h_{fe}}$$

$$h_{fb} = \frac{-h_{fe}}{1+h_{fe}}$$

$$h_{rb} = \frac{h_{ie} \cdot h_{oe}}{1+h_{fe}} - h_{re}$$

$$h_{ob} = \frac{h_{oe}}{1+h_{fe}}$$

CB to CE:

$$h_{ie} = \frac{h_{ib}}{1+h_{fb}}$$

$$h_{fe} = \frac{-h_{fb}}{1+h_{fb}}$$

$$h_{re} = \frac{h_{ib} \cdot h_{ob}}{1+h_{fb}} - h_{rb}$$

$$h_{oe} = \frac{h_{ob}}{1+h_{fb}}$$

CE to CC

$$h_{ic} = h_{ie}$$

$$h_{fc} = -(1+h_{fe})$$

$$h_{rc} = 1$$

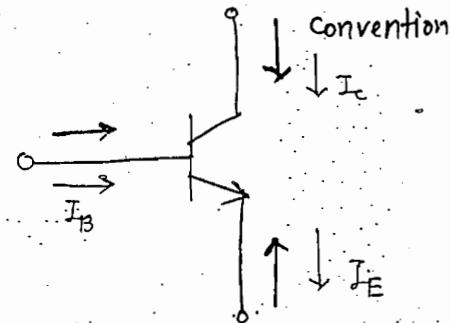
$$h_{oc} = h_{oe}$$

Concept of -ve gain:

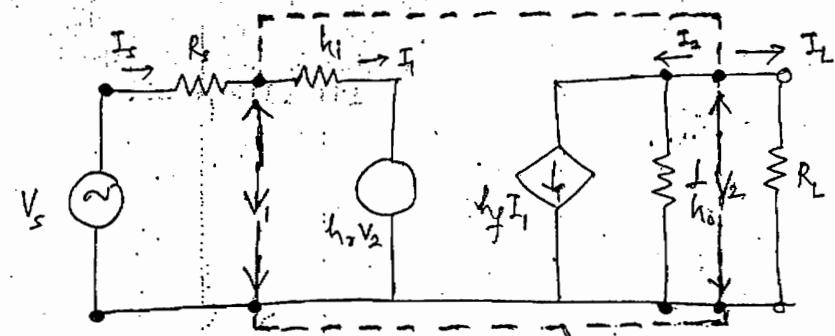
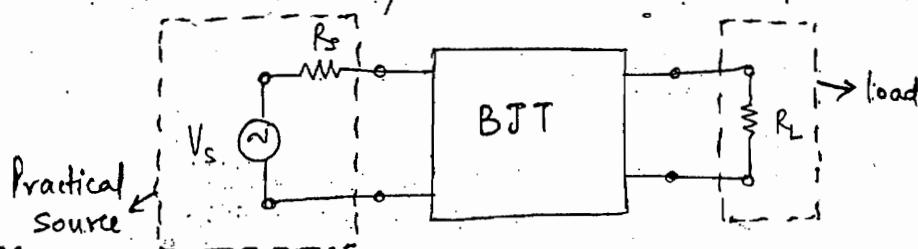
$$\text{gain } (I_B, I_C) = +\text{ve}$$

$$\text{gain } (I_B, I_E) = -\text{ve}$$

$$\text{gain } (I_C, I_E) = -\text{ve}$$



Amplifier analysis at low frequencies : (Exact model):



Characteristics of Amplifier (BJT)

- (1) I/p impedance, $Z_i = \frac{V_1}{I_1}$
- (2) O/p impedance, $Z_o = \frac{V_2}{I_2}$
- (3) Volt. gain, $A_v = \frac{V_2}{V_1}$
- (4) Current gain, $A_I = \frac{I_2}{I_1}$

Ckt parameters (ckt)

- (1). Effective I/p impedance, Z_i'
- (2). Effective O/p impedance, Z_o'
- (3). Voltage amplification, $A_{vS} = \frac{V_2}{V_S}$
- (4). Current amplification, $A_{IS} = \frac{I_2}{I_S}$

BJT characteristics Analysis:

(1) Current gain, A_I :

$$I_2 = h_f I_1 + h_o V_2$$

$$V_2 = -I_2 R_L$$

$$I_2 = h_f I_1 - h_o R_L I_2$$

$$I_2 (1 + h_o R_L) = h_f I_1$$

$$\frac{I_2}{I_1} = \frac{h_f}{1 + h_o R_L}$$

$$A_I = \frac{I_2}{I_1} = -\frac{I_2}{I_1} = \frac{-h_f}{1 + h_o R_L}$$

$$A_I = \frac{-h_f}{1 + h_o R_L}$$

(2) I/p impedance, Z_i :

$$V_1 = h_i I_1 + h_r V_2$$

$$\frac{V_1}{I_1} = h_i + h_r \left(-\frac{I_2 R_L}{I_1} \right)$$

$$Z_i = h_i + h_r A_I R_L$$

$$\therefore Z_i = h_i + h_r R_L A_I$$

A1

(3) Voltage gain, A_V :

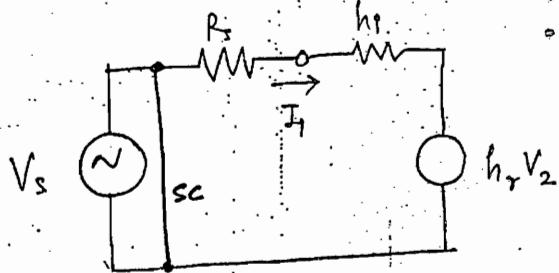
$$\begin{aligned} A_V &= \frac{V_2}{V_1} \\ &= \frac{-I_2 R_L}{V_1} \\ &= \frac{-I_2 / I_1 \cdot R_L}{V_1 / I_1} \end{aligned}$$

$$A_V = \frac{A_I \cdot R_L}{Z_i}$$

(4) %p impedance, Z_o :

$$Z_o = V_2 / I_2$$

$$Y_o = I_2 / V_2 = h_f \left(\frac{I_1}{V_2} \right) + h_o$$



$$V_s = (h_i + R_s) I_1 + h_o V_2$$

$$\frac{I_1}{V_2} = -\frac{h_o}{h_i + R_s}$$

$$Y_o = h_f \left(\frac{-h_o}{h_i + R_s} \right) + h_o$$

$$Z_o = \frac{1}{h_o - \frac{h_f h_o}{h_i + R_s}}$$

Conclusion:

Exact model are valid for all types of configurations ie. CE, CB, CC

Conditions :

$$(1) \quad h_{oe} R_L \leq 0.1$$

$$R_L < 5K\Omega$$

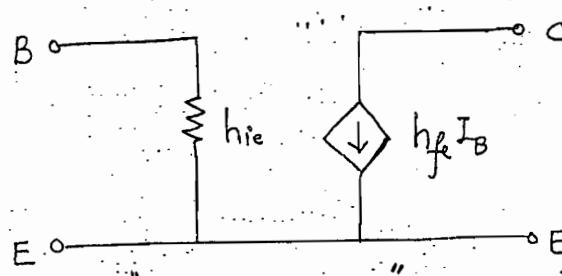
$$\text{Eq. } A_I = \frac{-h_{fe}}{1+h_{oe}R_L} = \frac{-h_{fe}}{1+0.1} = -0.91 h_{fe}$$

* In electronics, less 10% error can be neglected.

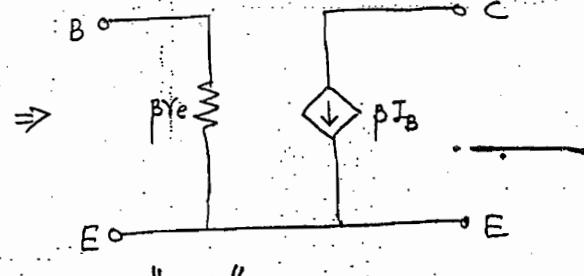
* If $R_L > 5K$, error will come.

(2). h_{oe} & h_{re} should be neglected.

(3). It is valid only for CE mode

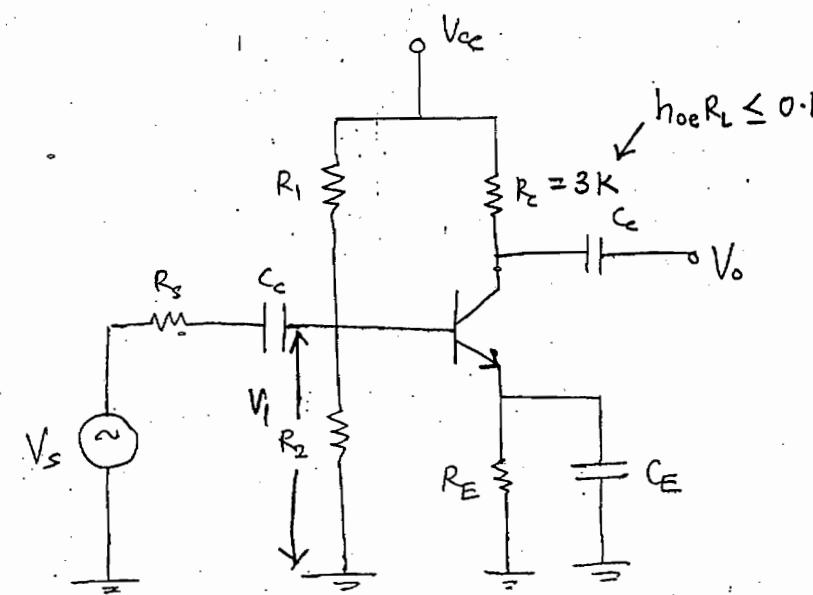


"h parameter" model

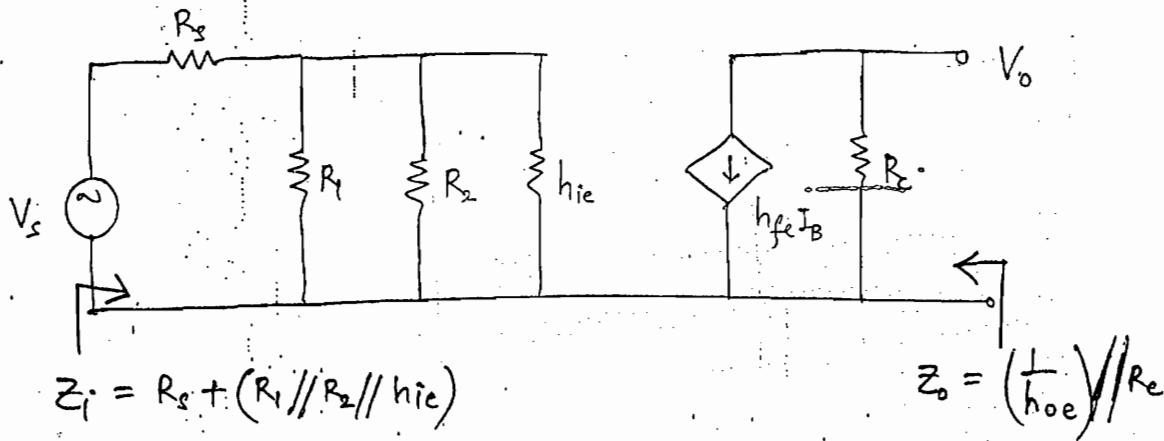


" γ_e " model

CE amplifier :



Ac model, DC \rightarrow grounded, capacitor \rightarrow SC



$$A_{V_s} = \frac{V_o}{V_i} \cdot \frac{V_i}{V_s}$$

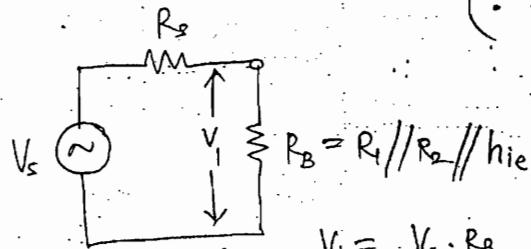
$$= A_V \cdot \frac{V_i}{V_s}$$

$$A_{V_s} = \frac{A_V \cdot R_B}{(R_s + R_B)}$$

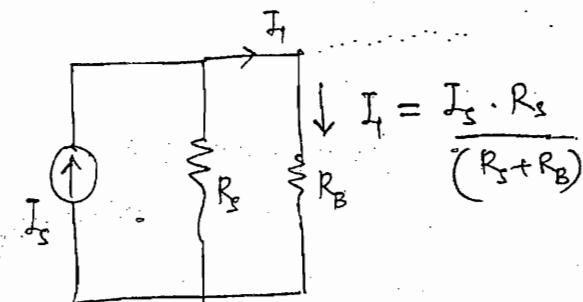
$$A_{I_s} = \frac{I_L}{I_I} \cdot \frac{I_I}{I_s}$$

$$= A_I \cdot \frac{I_I}{I_s}$$

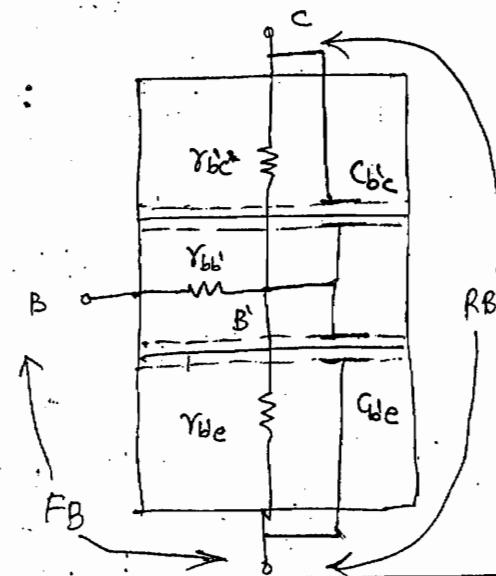
$$A_{I_s} = \frac{A_I \cdot R_I}{(R_s + R_B)}$$

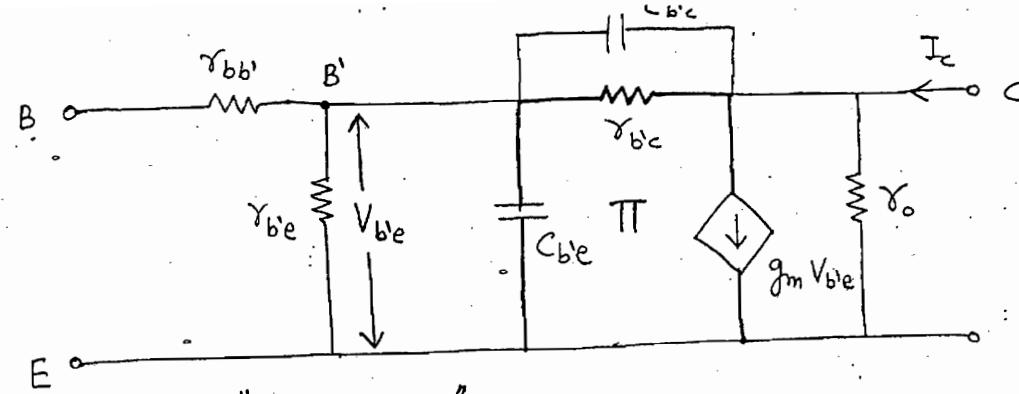


$$V_i = \frac{V_s \cdot R_B}{(R_s + R_B)}$$



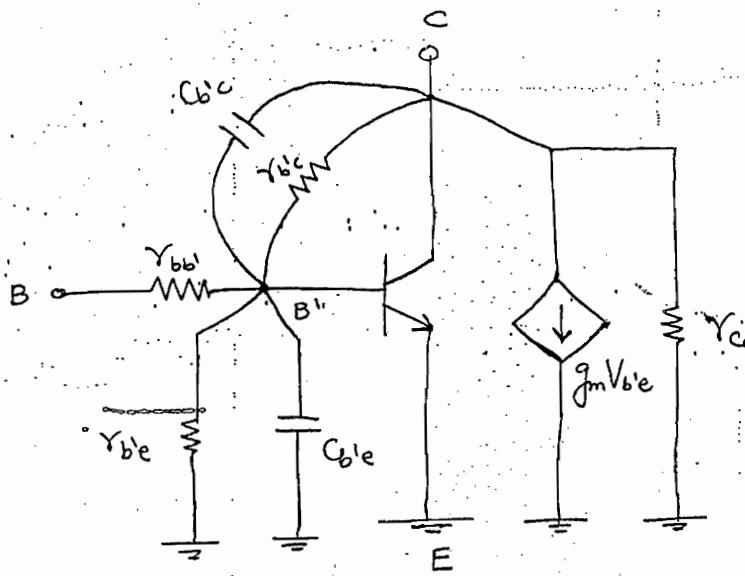
High frequency analysis of BJT :





"Hybrid- π " model

- * This model is valid for both high and low frequency.
- But for low frequency, capacitor \rightarrow open
- and for high frequency, ckt is taken as it is.



Hybrid π parameters:

- $\gamma_{bb'}$ \rightarrow Base spreading resistance (less)
- $\gamma_{be'}$ \rightarrow I/p resistance (γ_π)
- $\gamma_{bc'}$ \rightarrow feedback resistance (high)
- γ_{ce} \rightarrow O/p resistance (γ_o)
- $C_b \rightarrow C_{be'}$ \rightarrow diffusion capacitance (stronger)
- $C_t \rightarrow C_{bc'}$ \rightarrow transition capacitance (weaker)
- $g_m \rightarrow$ trans conductance $\left[(I_c)_{Q, \max} = 50 \text{ mA} \right]$

Typical values

100 (\approx sc)
1 K Ω
4 M Ω (\approx o/i)
80 K Ω
100 pF
3 pF
50 mA/V

Expression for Hybrid π parameters:

Note: Hybrid π parameters depends on 3 important parameters

(1). Collector current (I_c)_Q

(2). Temperature

(3). Collector to Emitter Voltage V_{CE}

$$(1) \cdot \underline{g_m} : g_m = \frac{(I_c)_Q}{V_T} = \frac{(I_c)_Q}{T/11,600}$$

As $T \uparrow$, $g_m \downarrow$ $A_V = -g_m R_L$

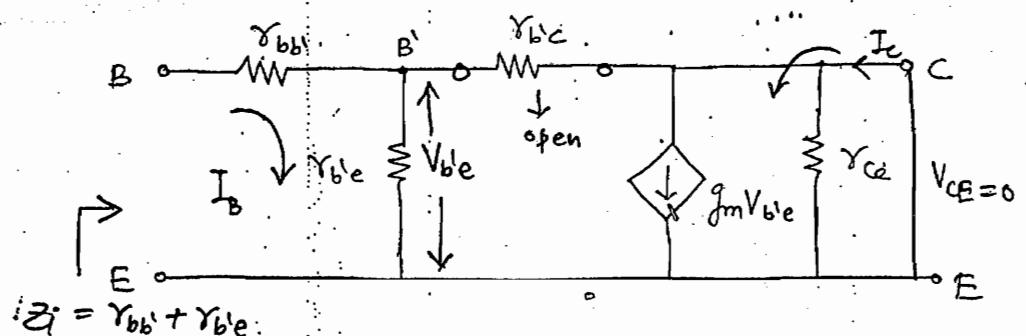
As $(I_c)_Q \uparrow$, $g_m \uparrow$ if load is const. $A_V \uparrow$ if $g_m \uparrow$

(2). $\gamma_{b'e}$:

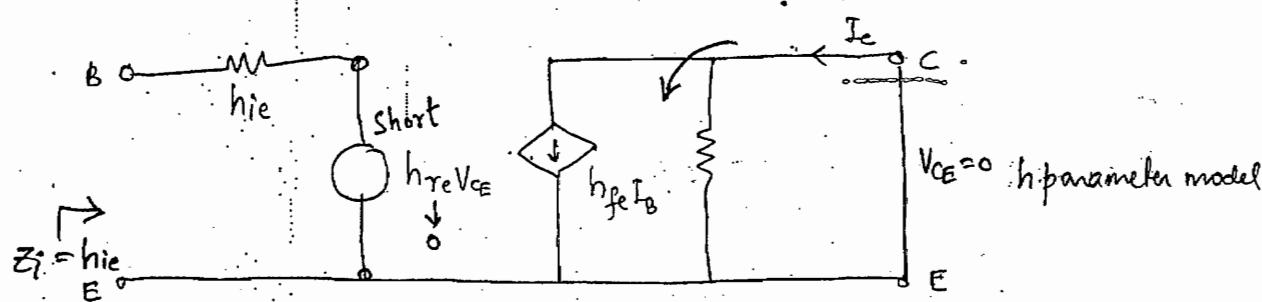
$$\gamma_{b'e} = \frac{h_{fe}}{g_m} \rightarrow h \text{ parameter} \quad (\text{at moderate frequency})$$

↓
π model

Converting high freq. model into low frequency.



$\gamma_{bb'} \rightarrow E/p$ Put $V_{CE} = 0$ (sc)



$$I_C = g_m V_{B'E}$$

$$I_C = g_m \cdot I_B V_{B'E}$$

$$\gamma_{B'E} = \left(\frac{I_C}{I_B} \right) \frac{1}{g_m} \quad \text{--- (1)}$$

$$\therefore \boxed{\gamma_{B'E} = \frac{h_{fe}}{g_m}}$$

And,

$$g_m = \frac{(I_e)_\alpha}{V_T}$$

$$I_C = h_{fe} I_B$$

$$\frac{I_C}{I_B} = h_{fe} \quad \text{--- (2)}$$

$$\text{and } h_{fe} = \beta$$

$$\frac{1}{g_m} = \frac{V_T}{(I_e)_\alpha} \approx \frac{V_T}{(I_E)_\alpha} = \gamma_e$$

$$\therefore \boxed{\gamma_{B'E} = \frac{h_{fe}}{g_m} = \beta \gamma_e}$$

(3). $\gamma_{bb'}$: Put $V_{CE} = 0$ (again) ($\because \gamma_{bb'} \rightarrow \text{i/p parameter}$)

$$Z_i = \gamma_{bb'} + \gamma_{B'E}$$

$$Z_i = h_{ie}$$

$$\gamma_{bb'} + \gamma_{B'E} = h_{ie}$$

$$\boxed{\gamma_{bb'} = h_{ie} - \gamma_{B'E}}$$

↓ ↓ ↓
 100Ω 1.1kΩ 1kΩ

(4). $\gamma_{B'C}$:

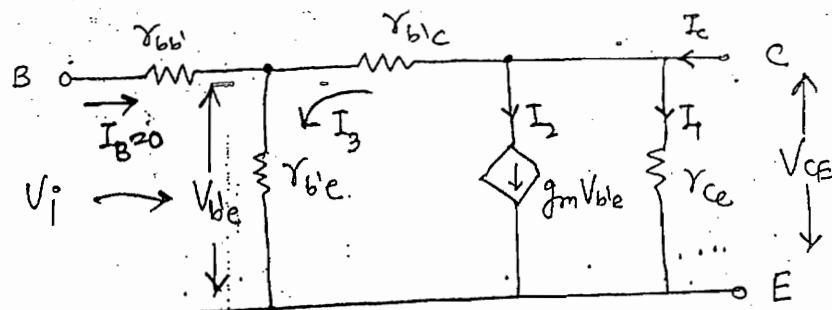
$$\frac{\gamma_{B'C}}{(1-A_V)} \approx \frac{\gamma_{B'C} A_V}{(A_V - 1)} \approx \gamma_{B'C}$$

↓
 (o/p parameter)

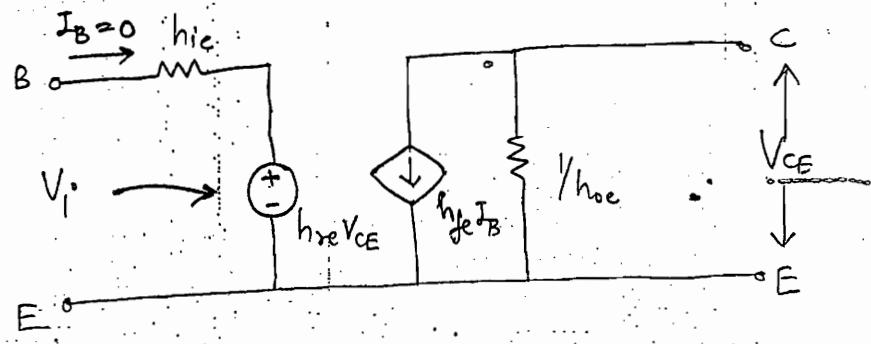
$\gamma_{B'C}$ ↓
 ↓
 o/p

↓
 $\gamma_{B'C}$
 ↓
 negligible

Put $I_B = 0$



$$V_p = V_{be} = \frac{V_{CE} \times r_{be}}{r_{be} + r_{bc}} = V_{CE} \cdot \frac{r_{be}}{r_{be} + r_{bc}} \quad (\because r_{bc} \gg r_{be})$$



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

$$\therefore V_{CE} \cdot \frac{r_{be}}{r_{be} + r_{bc}} = h_{re} \cdot V_{CE}$$

$$\boxed{r_{bc} = \frac{r_{be}}{h_{re}}}$$

(5) r_{ce} :

$$\frac{I_c}{V_{CE}} = h_{re}$$

$$I_c = \frac{V_{CE}}{r_{ce}} + g_m V_{be} + \frac{V_{CE}}{r_{be} + r_{bc}}$$

$$I_c = \frac{V_{CE}}{r_{ce}} + g_m \left(\frac{V_{CE} r_{be}}{r_{be} + r_{bc}} \right) + \frac{V_{CE}}{r_{be} + r_{bc}}$$

$$\frac{I_c}{V_{CE}} = \frac{1}{r_{ce}} + \frac{g_m r_{be} + 1}{r_{be} + r_{bc}}$$

$$\frac{I_c}{V_{CE}} = \frac{1}{r_{ce}} + \frac{h_{fe} + 1}{r_{be} + r_{bc}}$$

$$(\because h_{fe} = g_m r_{be})$$

$$h_{re} = \frac{1}{r_{ce}} + \frac{h_{fe}}{r_{bc}} = g_{ce} + h_{fe} g_{bc}$$

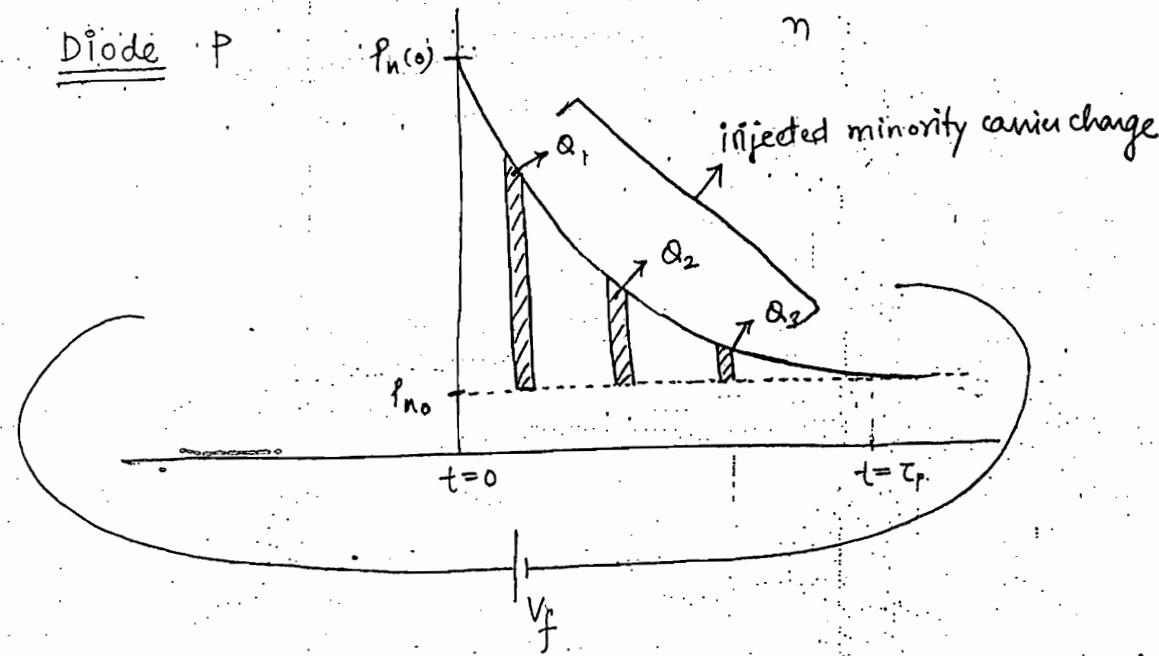
$$(\because r_{bc} \rightarrow \text{neglected})$$

$$(r_{ce} = \frac{1}{g_{ce}} \text{ & } r_{bc} = g_{bc})$$

$$g_{ce} = h_{oe} - h_{fe} g_{b'e}$$

$$\gamma_{ce} = \frac{1}{g_{ce}} = \frac{1}{h_{oe} - h_{fe} g_{b'e}}$$

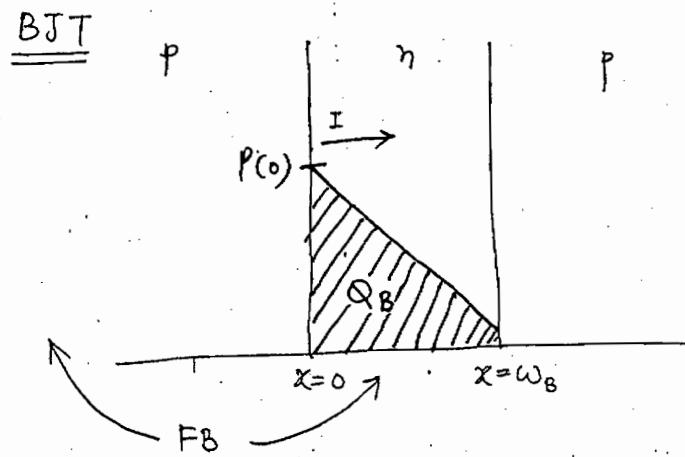
(6). Diffusion capacitance, C_D : ($C_{b'e}$)



$$Q = I_p T_p$$

$$C_D = \frac{dQ}{dV} = T_p \cdot \frac{dI_p}{dV}$$

$$\therefore C_D = \frac{T_p I_{dc}}{V_T}$$



$$Q_B = \frac{1}{2} P(o) \cdot (Aq w_B) \cdot \frac{cm^2}{m^2}$$

$$I = - Aq D_B \frac{dp}{dx}$$

$$= - Aq D_B \left\{ \frac{P(o) - 0}{0 - w_B} \right\}$$

$$= \frac{Aq D_B \{ P(o) \}}{w_B}$$

$$Q_B = \frac{1}{2} P(0) \cdot A q D_B$$

$$\therefore Q_B = \frac{I \omega_B^2}{2 D_B}$$

And, $C_D = \frac{d Q_B}{d V} = \frac{d \frac{\omega_B^2 I}{2 D_B}}{d V} = \frac{\omega_B^2}{2 D_B} \cdot \frac{d I}{d V} = \frac{\omega_B^2}{2 D_B} g_m$

$$C_D = \frac{\omega_B^2}{2 D_B} \cdot \frac{1}{\gamma_e}$$

$$C_D = \frac{\omega_B^2 \cdot (I_c)_0}{2 D_B} \cdot \frac{V_T}{V_T}$$

$$T_B = \frac{\omega_B^2}{2 D_B} \rightarrow \text{Base transit time}$$

As $\frac{\omega_B^2}{2 D_B} = \frac{\omega_B^2}{2 \mu V_T} \rightarrow \frac{m^2}{m^2/V \cdot \text{sec}} \rightarrow \text{sec i.e. time}$

Q. As V_{CE} ↑ what happens to C_D in hybrid II model?

Sol. As $V_{CE} \uparrow$, $\omega_B \downarrow$, $C_D \downarrow$. (Early Effect)

$$\omega_B = \omega_B' + \omega_B''$$

↓
this width we are talking about.

(7). $C_{b'c}$: Transition capacitance (C_T) :

$$C_T = \frac{\epsilon A}{\omega}$$

$$G \propto \frac{1}{\sqrt{V_R}} \rightarrow \text{Alloy type}$$

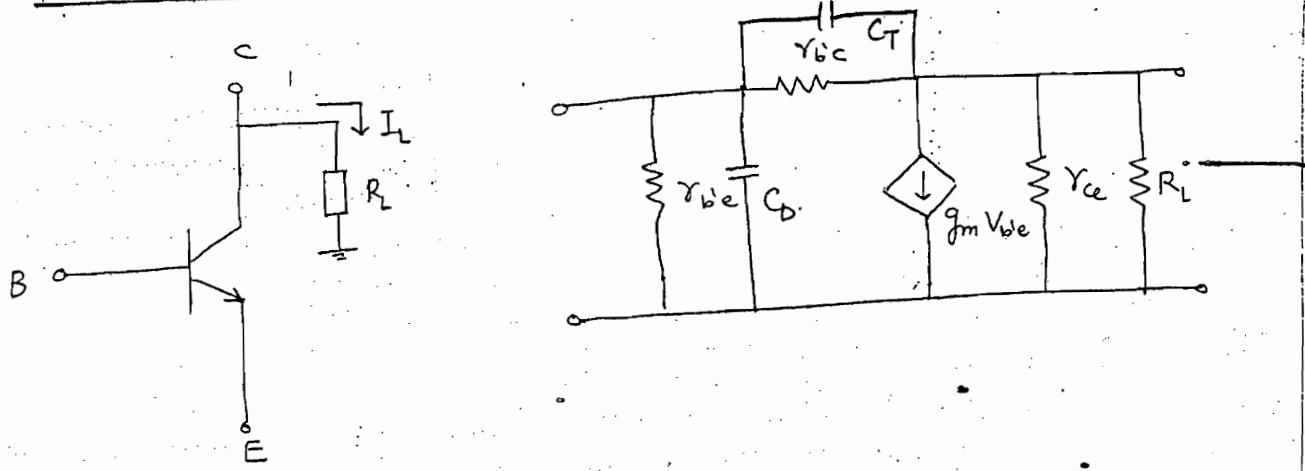
$$G_T \propto \frac{1}{\sqrt[3]{V_R}} \rightarrow \text{Grown jxn type}$$

$$C_{b'c} \propto \frac{1}{(V_{CE})^n}$$

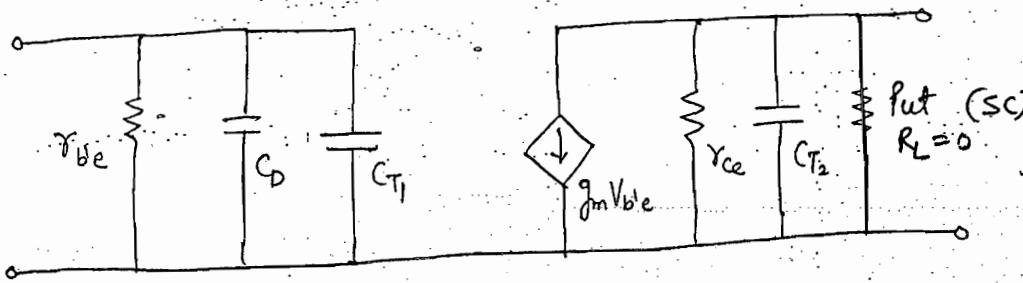
" , $n = 1/2 \rightarrow \text{alloy type}$
 $n = 1/3 \rightarrow \text{grown jxn type}$

* $\gamma_{b'e}$ is independent of temp. & $(I_e)_Q$, it only depends upon V_{ce} supply. (the only parameter).

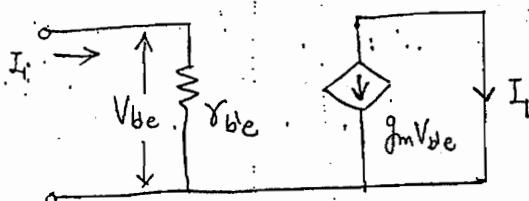
frequency response of an amplifier:



CE SC current gain (max^m in sc condition):



At $f = 0$: (Cap \rightarrow open)



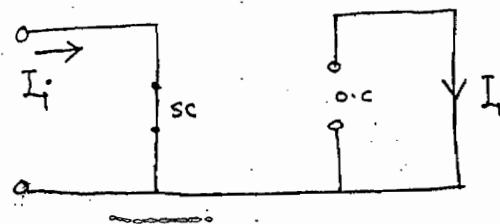
$$A_I = \frac{I_L}{I_i} = -\frac{g_m V_{be}}{V_{be}/\gamma_{b'e}}$$

$$= -g_m \gamma_{b'e}$$

$$= -h_{fe}$$

$$A_I = -\beta$$

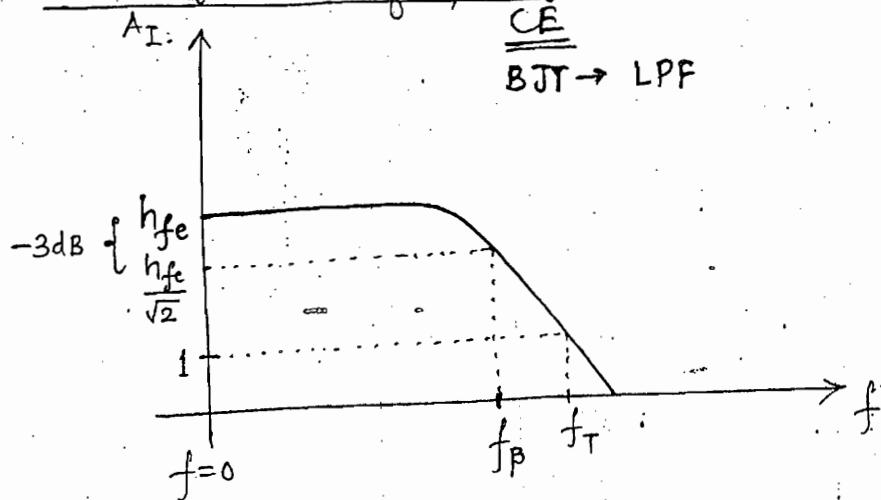
At $f = \infty$: (Cap \rightarrow short)



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

$$A_I = 0$$

Current gain V. frequency:

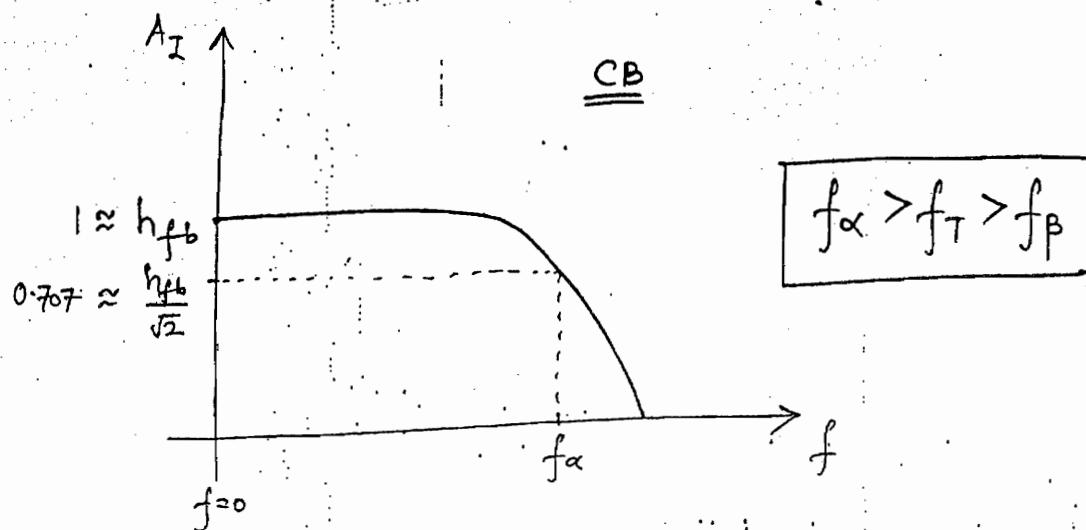


* dB values are used to compress the values of scale markings. ($f \rightarrow \text{GHz, MHz}$)

Scaling is required for simplicity:

* $20\log(1) = 0$ i.e. above f_T , the gain is zero.

Hence, f_T is given by manufacturer, upto which a sig can be applied to the given BJT.



$$f_T = h_{fe} f_B$$

$$f_\alpha \approx 1.01 f_T \quad (\text{for } h_{fe} = \infty)$$

$$f_\alpha = (1 + h_{fe}) f_B = \left(\frac{1}{h_{fe}} + 1\right) f_T$$

f_B (β cut-off frequency) :

It is the frequency at which common emitter SC current gain reduces by 3 dB of its value.

f_T (unity gain BW frequency) :

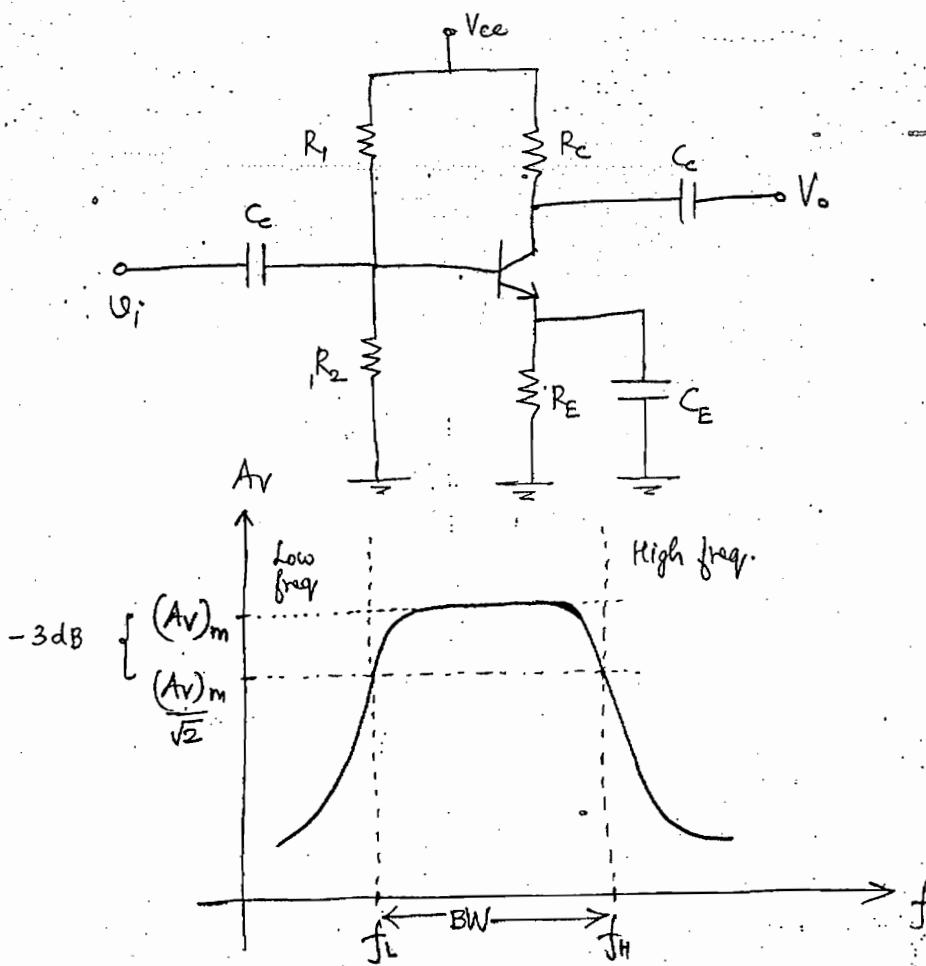
It is the frequency at which CE SC A_I falls to unity that is called as f_T .

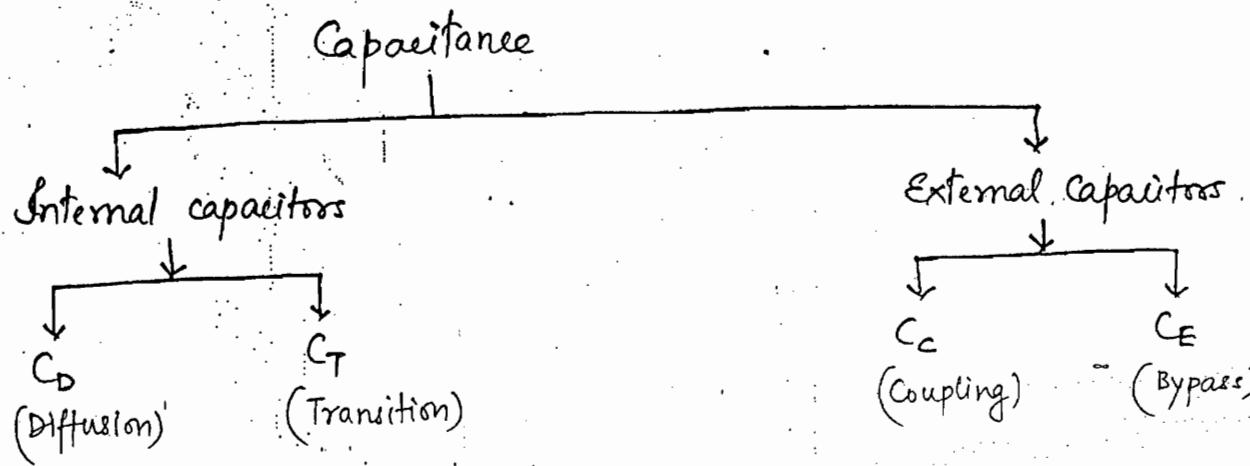
f_A (α cut-off frequency) :

It is the frequency at which CB SC A_I reduces by 3 dB of its value.

Voltage gain V_o frequency :

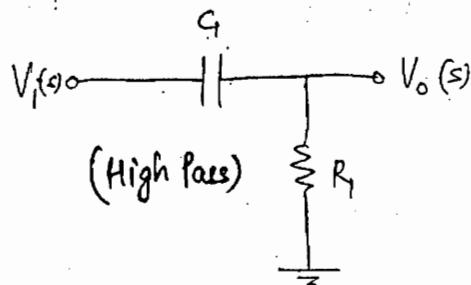
CE amplifier :





Low frequency Range:

- (1). At low freq. range , internal capacitors C_D & C_T which are assumed to be OC are neglected in the analysis.
- (2). But when the external capacitors C_C & C_E are assumed to be OC , they will create a problem for s/s response .
- (3). The s/s response at low freq. range is a High pass response .
- (4). The equivalent model of HP ckt is given as :



$$\begin{aligned} \frac{V_o(s)}{V_i(s)} &= \frac{R}{R - j X_C} \\ &= \frac{s}{(s + \frac{1}{R_C})} \end{aligned}$$

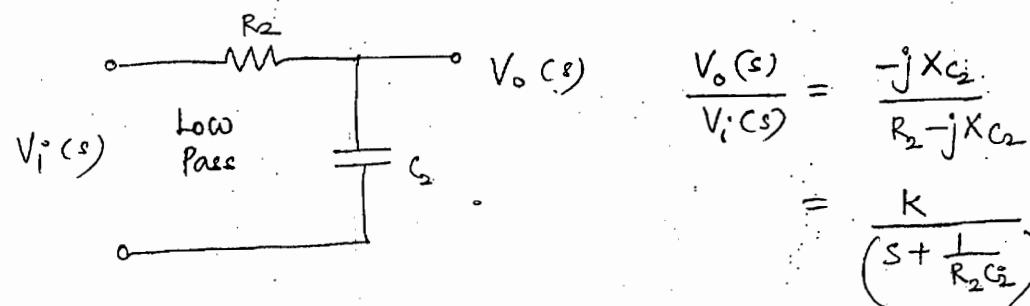
zero $\rightarrow 1$
pole $\rightarrow 1$

High frequency Range:

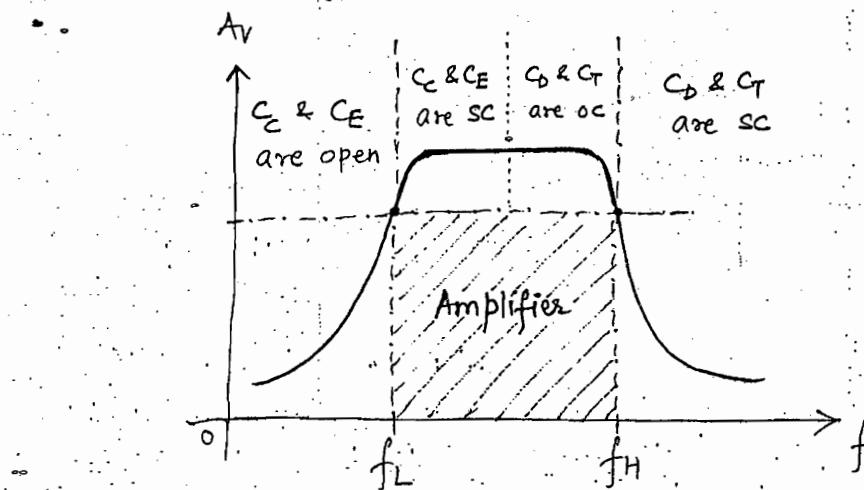
- (1). At high freq. range , external capacitors C_C & C_E which are assumed to be SC are neglected in analysis
- (2). But when the internal capacitors C_D & C_T are assumed to be SC , they will create a problem for s/s response .

(3). The BJT response at high frequency range is a low pass response.

(4). The equivalent model of LP is given as:



Mid band range:



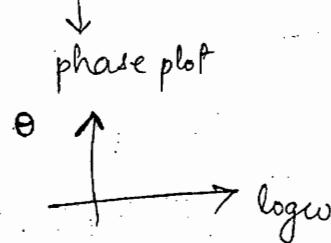
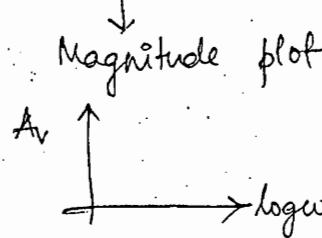
(1). At low freq. range, mid band range frequencies are considered as high frequencies. Therefore, external capacitors C_E and C_E are assumed to be SC

(2). At high freq. range, mid band range frequencies are considered as low frequencies. Therefore, internal capacitors C_D & C_T are assumed to be OC.

(3). That means. In the mid band range all the internal and external capacitors are neglected. Therefore the gain is independent of frequency.

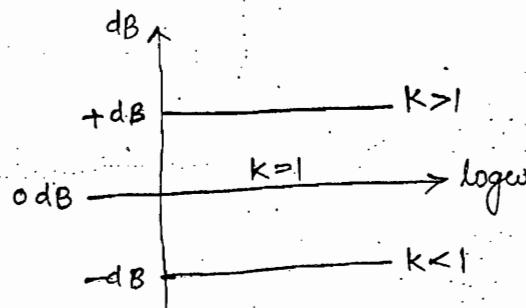
-3 dB concept:

Bode plots



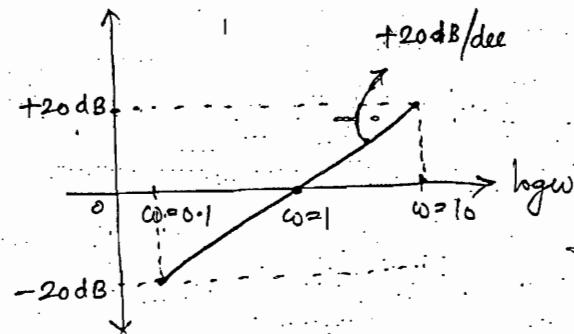
$$(1). \quad G(s) = K \text{ (constant)}$$

$$|G(j\omega)|_{dB} = 20 \log K$$



$$(2). \quad G(s) = s \text{ (zero)}$$

$$|G(j\omega)|_{dB} = 20 \log \omega$$



$$(3). \quad G(s) = \frac{1}{(1+sT)} \text{ (pole)}$$

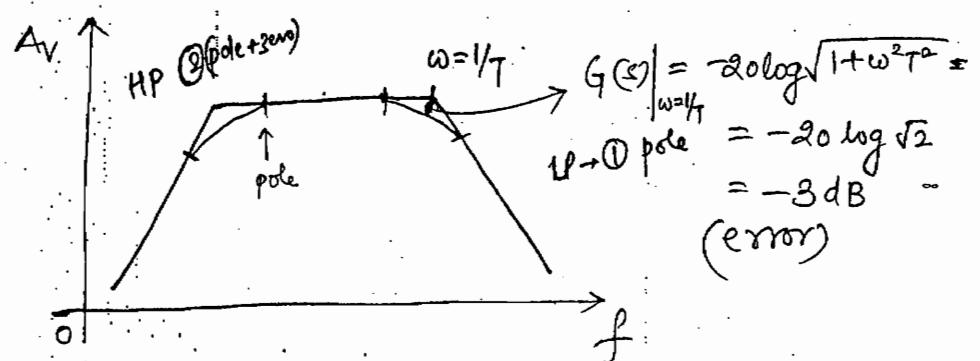
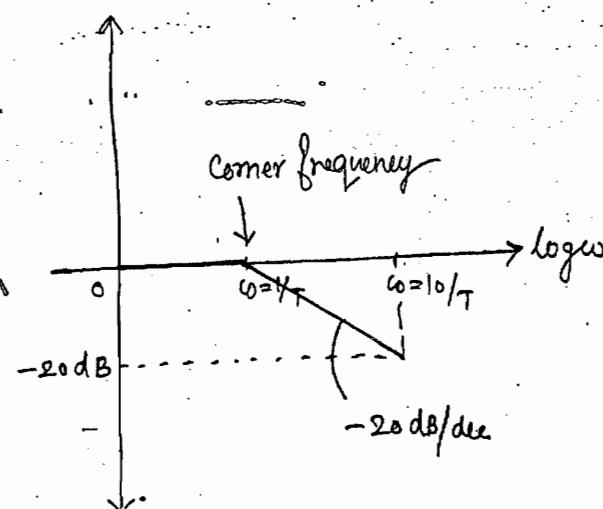
$$|G(j\omega)|_{dB} = -20 \log \sqrt{1+\omega^2 T^2}$$

$\omega T \ll 1$ (low freq.)

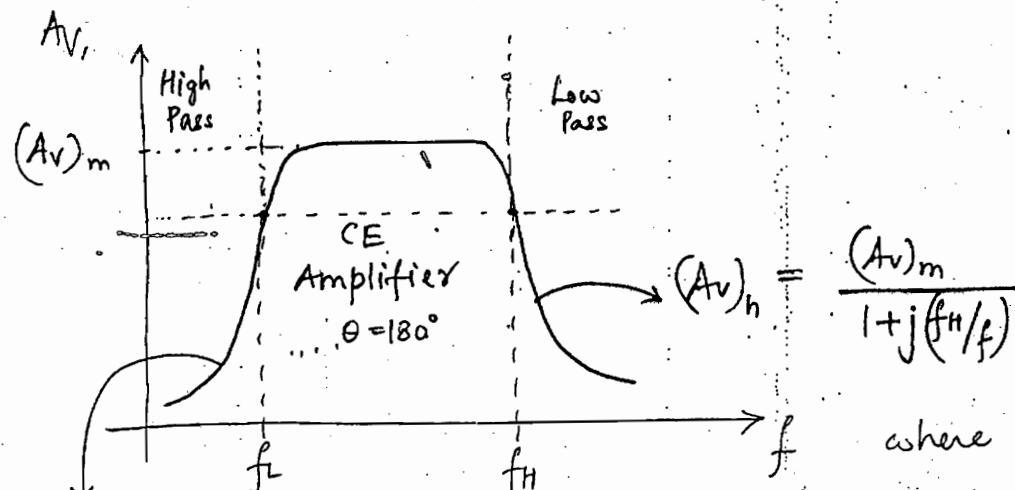
$$|G(j\omega)|_{dB} = 0 \text{ dB}$$

$\omega T \gg 1$ (high freq.)

$$|G(j\omega)|_{dB} = -20 \log \omega T$$



Phase Analysis :



$$(A_V)_L = \frac{(A_V)_m}{1 - j(f_L/f)}$$

$$\text{where } f_L = \frac{1}{2\pi R_c \text{internal}}$$

$$\theta_L = +\tan^{-1}(f_L/f)$$

when $f \rightarrow f_L$

$$\theta_L = 45^\circ$$

$$\theta_{LCE} = 180^\circ + \theta_L \\ = 225^\circ$$

$$\theta_{LCB} = 45^\circ$$

$$\theta_H = -\tan^{-1}(f_H/f)$$

when $f \rightarrow f_H$

$$\theta_H = -45^\circ$$

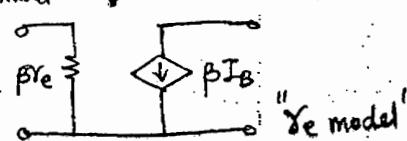
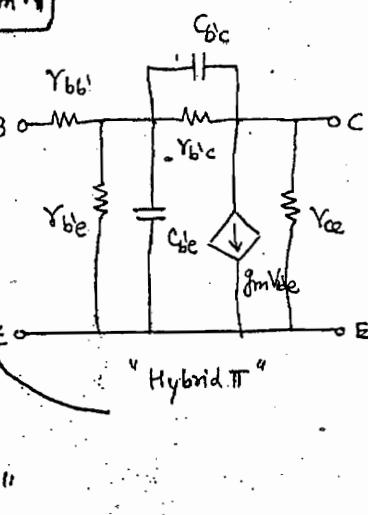
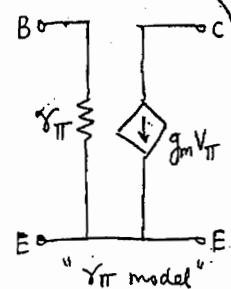
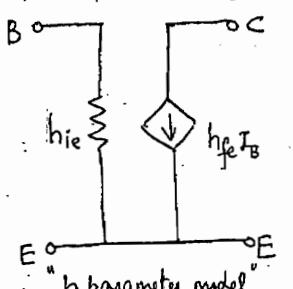
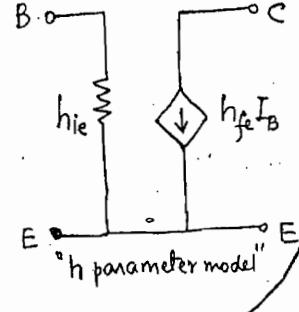
$$\theta_{HCE} = 180^\circ + \theta_H \\ = 135^\circ$$

$$\theta_{HCB} = -45^\circ$$

C_D & C_T
are OC

$$h_{ie} = \beta Y_e = Y_{\pi}$$

$$h_{fe} I_B = \beta I_B = g_m V_T$$



$$A_v = -g_m R_e$$

$$A_v = -\frac{R_e}{r_e} \rightarrow "r_e"$$

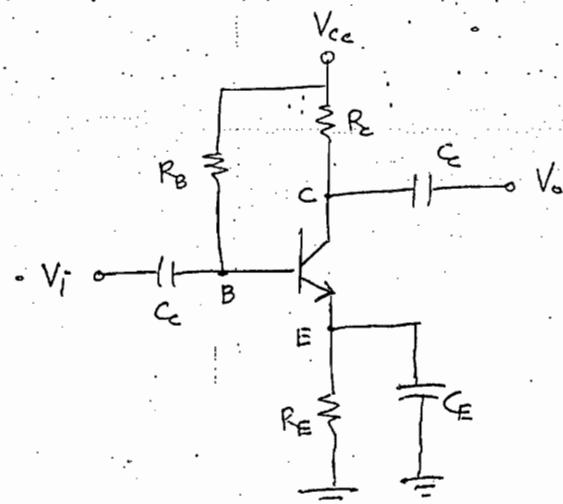
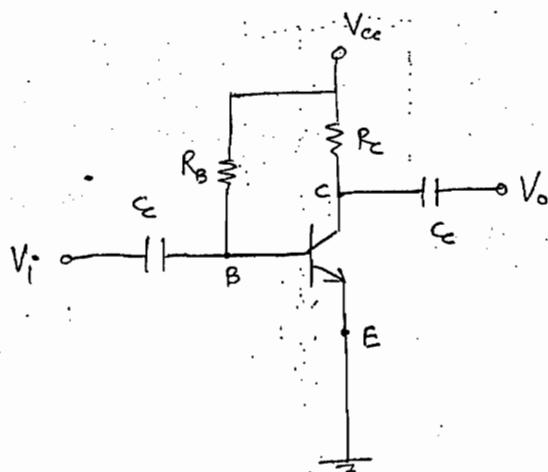
$$A_v = -\frac{h_{fe} R_e}{h_{ie}} \rightarrow "h \text{ parameter}"$$

$$A_v = -\frac{B R_e}{r_\pi} \rightarrow "r_\pi"$$

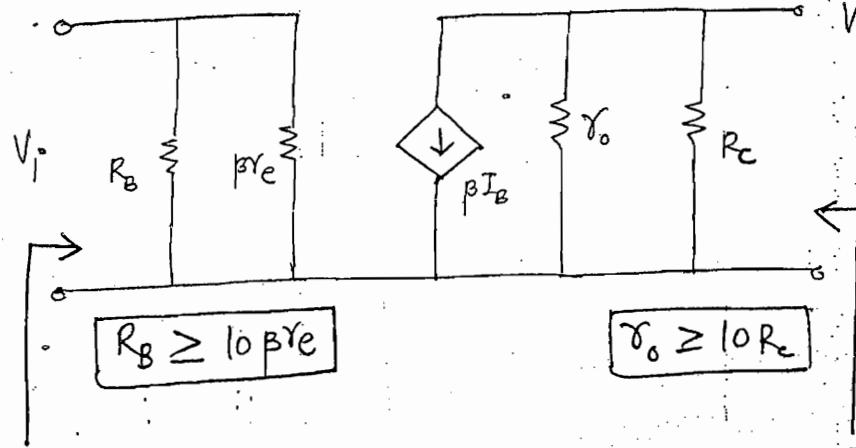
Practical Amplifier designs:

- (1). CE bypass amplifier
- (2). CE, unbypass amplifier
- (3). CC amplifier
- (4). CB amplifier

(1). CE bypass amplifier:



If R_E is not equal to zero, then there must be a bypass capacitor C_E in parallel to it.



$$Z_i = R_B // \beta r_e \approx \beta r_e \text{ (low)}$$

$$Z_o = r_o // R_C \approx R_C$$

$$V_o = -\beta I_B R_C$$

$$I_B = \frac{V_i}{\beta r_e}$$

$$A_v = \frac{V_o}{V_i} = -\frac{R_C}{r_e} = -g_m R_C \quad (\propto \frac{1}{r_e})$$

Heavy values of gain are unstable

i.e. if $A_v = 1000 \rightarrow$ varies (999.8 or 1000.5)

$\therefore A_v \uparrow$, stability \downarrow

Conclusions:

(i). Drawbacks of CE bypass amplifier :

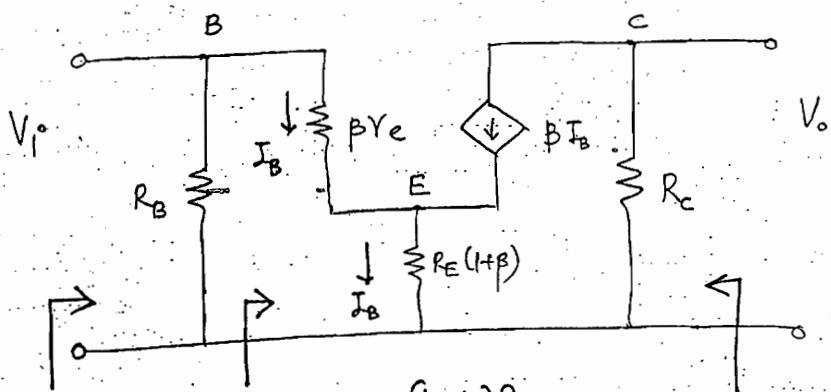
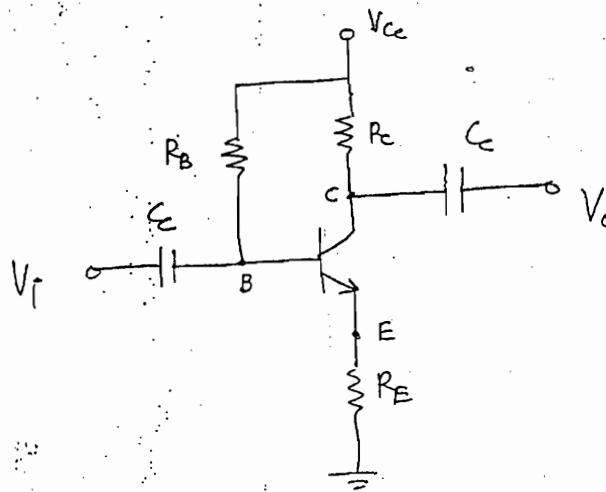
(i) The i/p impedance is low. (For an ideal amplifier i/p impedance should be ∞) .

(ii). The gain is not stable becoz of dynamic resistor r_e .

* $A_v = \infty$, stability = 0

* $A_v = 0$, stability = ∞

(2) CE unbypassed amplifier:



$$Z'_i = R_B \parallel z_i \quad z_i = \beta Y_e + (1+\beta) R_E \\ \approx \beta R_E \quad (\text{High})$$

Assume: $\beta = 100$, $r_e = 1 \Omega$, $R_E = 10 k\Omega$

$$z_i = 100 \Omega + (101) 10 k\Omega \\ = 100 \Omega + 1 M\Omega \\ \approx 1 M\Omega$$

$$V_o = (-\beta I_B) R_C$$

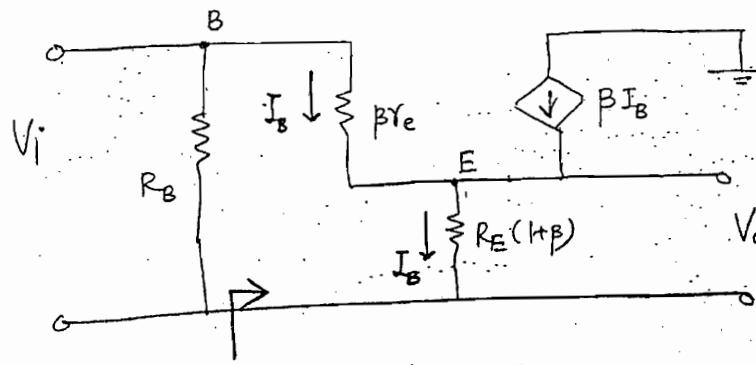
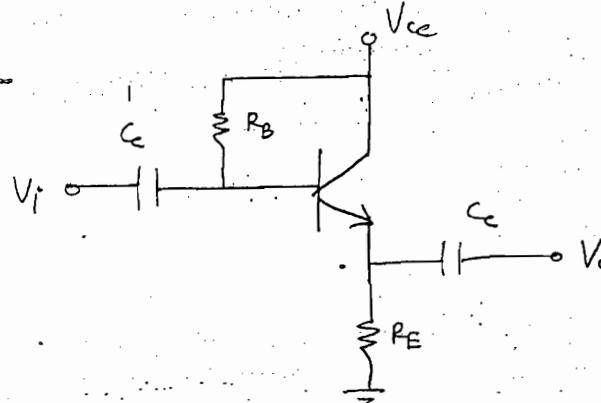
$$I_B = \frac{V_i}{\beta R_E}$$

$$A_v = \frac{V_o}{V_i} = -\frac{R_C}{R_E} \quad (\propto r_e)$$

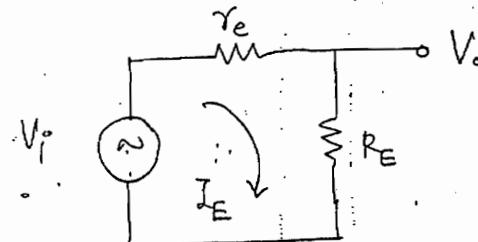
As $A_v \downarrow$, stability \uparrow .

Conclusions:

- (1). The I/p impedance is high.
- (2). Gain is stable because of absence of dynamic resistance r_e .
- (3). Emitter follower amplifier :



$$Z_i = \beta r_e + (1+\beta) R_E \\ \approx \beta R_E$$



$$V_o = V_i \times \frac{R_E}{(r_e + R_E)}$$

$$V_o = V_i$$

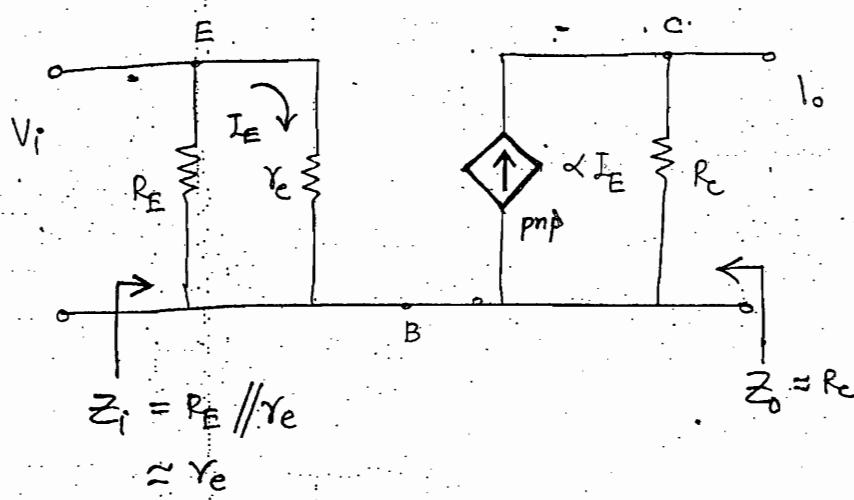
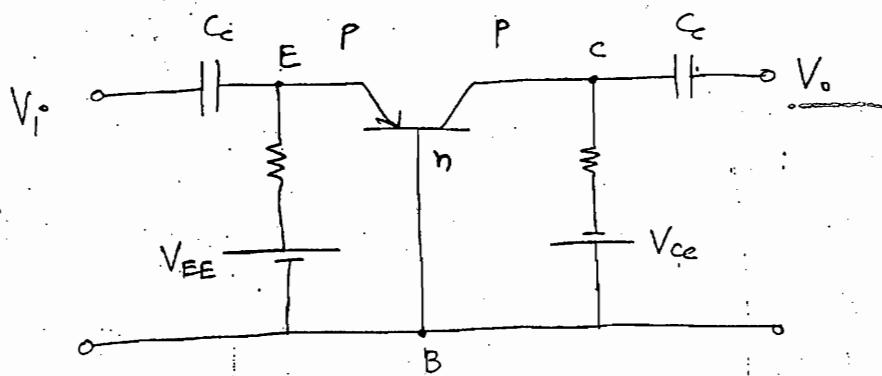
\rightarrow Voltage Buffer

$(\because R_E \gg r_e)$

$$Z_o = r_e // R_E \approx r_e \text{ (impedance matching)}$$

short: $A_V = \frac{V_o}{V_i} = \frac{(I_o)R_L}{I_i Z_i} = \frac{A_I R_L}{Z_i}$

$$\text{Here, } A_V = 1 \Rightarrow Z_i = A_I R_L = \beta \cdot R_E$$

(4). CB amplifier -

$$V_o = (\alpha I_E) R_C$$

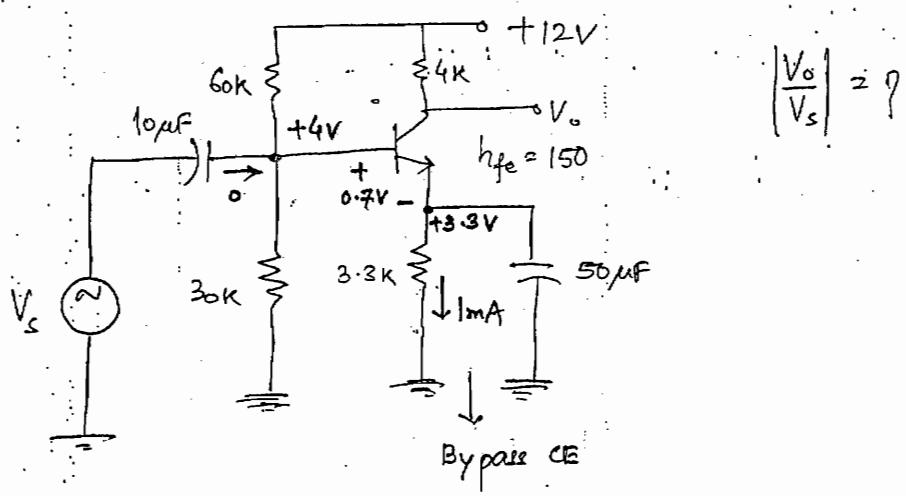
$$I_E = \frac{V_i}{r_e}$$

$$A_v = \frac{V_o}{V_i} = \frac{R_C \cdot \alpha}{r_e} \approx \frac{R_C}{r_e} \quad (\because \alpha \approx 1)$$

Problems :

CH-6 conv.

Q3.



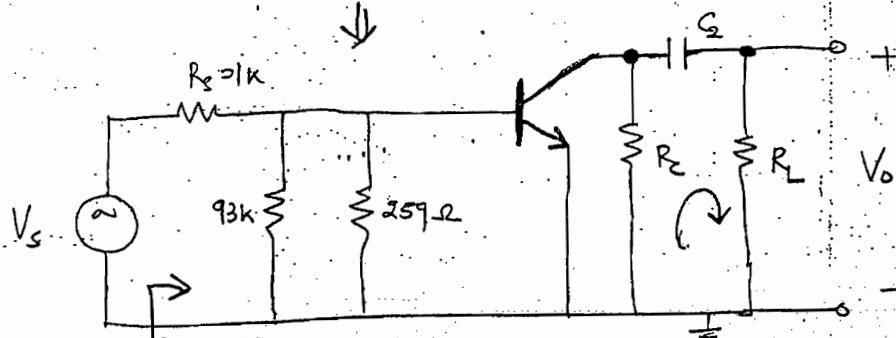
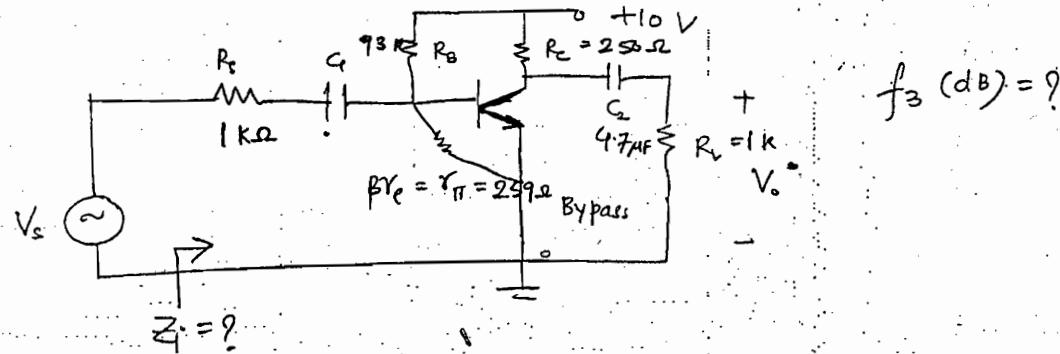
$$\text{Sol. } A_V = -\frac{R_e}{r_e} = -\frac{4K}{r_e}$$

$$r_e = \frac{V_T}{(I_e)_Q} = \frac{2.6 \text{ mV}}{1 \text{ mA}} = 2.6 \Omega$$

$$\therefore |A_V| = \left| \frac{-4 \text{ K} \cdot \Omega}{2.6 \Omega} \right| = 153.8$$

$$\text{If } V_T = 2.5 \text{ mV}, \quad |A_V| = \left| \frac{-4 \text{ K} \cdot \Omega}{2.5 \Omega} \right| = 160$$

Q4.

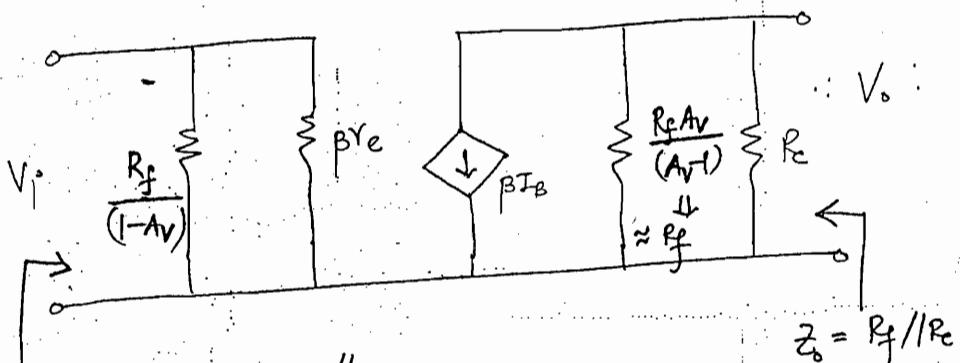
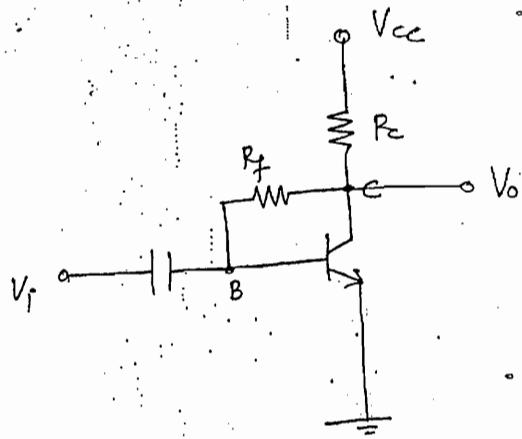


$$\begin{aligned} Z_i &= 1k + (93k // 259\Omega) \\ &= 1k + 259\Omega \\ &= 1259\Omega \end{aligned}$$

$$f_3(\text{dB}) = \frac{1}{2\pi(R_C + R_L)C}$$

$$= \frac{1}{2\pi (250 + 1000) \times 4.7 \times 10^{-6}}$$

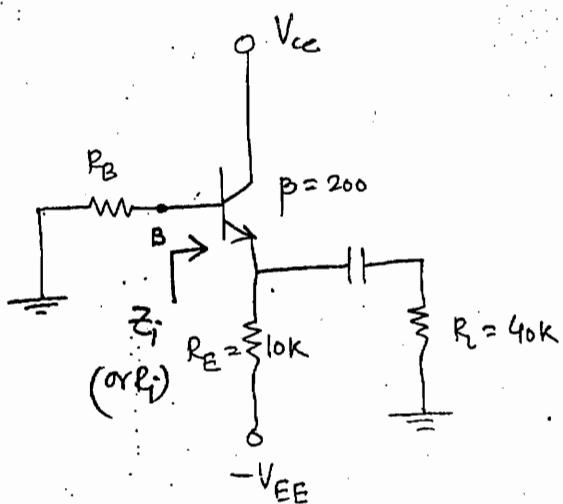
$$= 27.09 \text{ Hz}$$



$$A_V = -\frac{R_L \parallel R_f}{r_e}$$

$$Z_i = \left(\frac{R_f}{1 - A_V} \right) \parallel R_e$$

Q1.

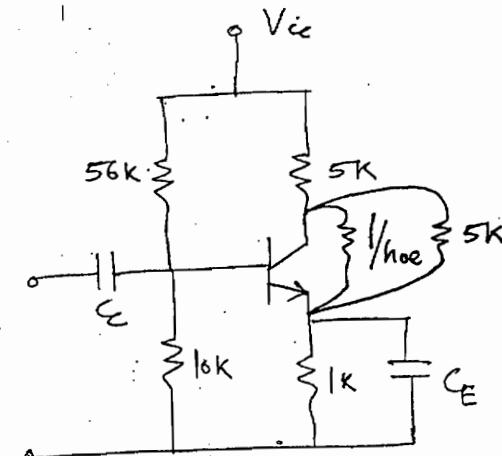


$$\begin{aligned} r_i &= \beta \cdot \left(R_L \parallel R_E \right) \\ &= 200 \cdot \left(\frac{10k \cdot 40k}{50k} \right) \end{aligned}$$

$$= 1600 \text{ k}$$

$$= 1.6 \text{ M}\Omega$$

Q2.



$$h_{fe} = 100$$

$$h_{ie} = 2 \text{ k}\Omega$$

$$h_{re} = 0$$

$$h_{oe} = 0.05 \text{ m mho} \\ = 0.05 \times 10^{-3} \text{ } \Omega$$

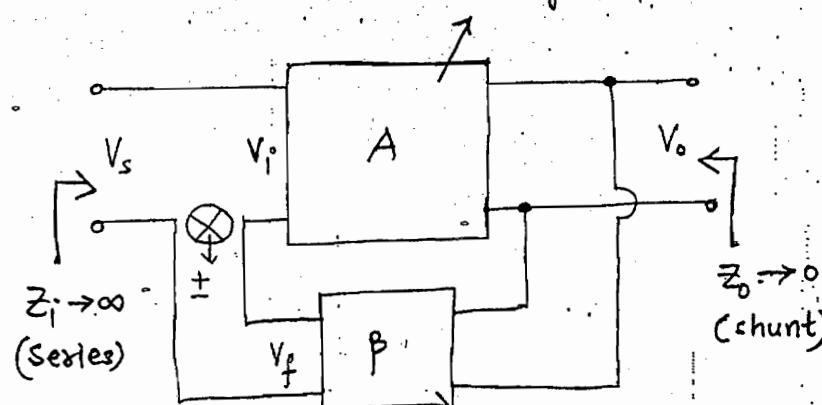
$$\begin{aligned} Z_0 &= 1/h_{oe} \parallel 5\text{k} = \frac{1}{0.05 \times 10^{-3}} \parallel 5000 \\ &= 20\text{k} \parallel 5\text{k} \\ &= 4\text{k} \Omega \end{aligned}$$

$$* f_T = \frac{1}{2\pi R C} = \frac{g_m}{2\pi(C_b + G)}$$

feedback amplifiers:

feedback theory:

Voltage amplifier (ideal amplifier)



$$V_f = \beta V_o \quad \text{feedback fraction (or) feedback ratio}$$

(or) reverse transmission factor

$$V_i = V_s \pm V_f$$

$$V_i = V_s + V_f \rightarrow +ve \text{ fb}$$

$$V_i = V_s - V_f \rightarrow -ve \text{ fb}$$

Positive fb

$$V_o = AV_i$$

$$V_i = V_s + V_f$$

$$V_o = A(V_s + V_f)$$

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

$$V_o(1 - A\beta) = AV_s$$

$$\boxed{\frac{V_o}{V_s} = \frac{A}{1 - \beta A}}$$

Negative fb

$$V_o = AV_i$$

$$V_i = V_s - V_f$$

$$V_o = A(V_s - V_f)$$

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

$$V_o(1 + \beta A) = AV_s$$

$$\boxed{\frac{V_o}{V_s} = \frac{A}{1 + \beta A}}$$

Conclusions:

$$(1) A_{pf} > A > A_{nf}$$

$$(2) A_{nf} = \frac{A}{1 + \beta A}$$

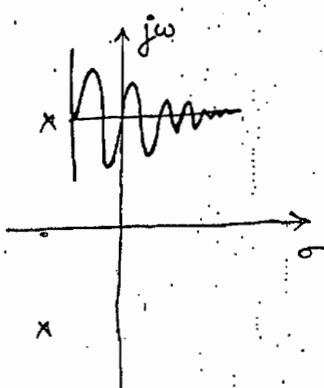
$$\beta A \ggg 1$$

$$A_{nf} \approx \frac{1}{\beta} \quad (\text{stability})$$

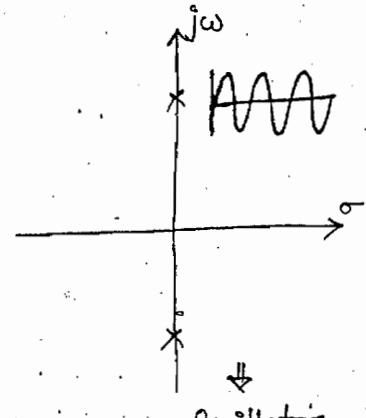
* -ve fb theory is applied for stable s/s like amplifiers.

$$(3) A_{pf} = \frac{A}{1 - \beta A}$$

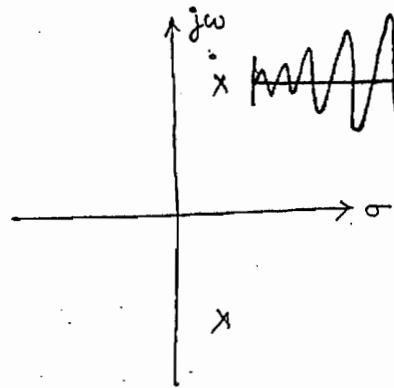
$$\underline{\beta A < 1}$$



$$\underline{\beta A = 1}$$



$$\underline{\beta A > 1}$$



$$A_{pf} = \frac{A}{\beta}, (\text{No } Y_P)$$

* The fb theory is applied for unstable s/s like oscillators. ¹⁶³

Advantages of -ve fb amplifiers :

(1). Stability of AC gain :

Suppose there is a small change in the internal resistance 'r_e' of an amplifier, then the fractional change of gain with feedback is

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

Differentiate w.r.t. A .

$$\frac{\partial A_f}{\partial A} = \frac{(1 + \beta A)(1) - A \cdot \beta}{(1 + \beta A)^2}$$

$$\frac{dA_f}{A_f} = \frac{1}{(1 + \beta A)^2} \cdot \frac{dA}{A_f}$$

$$\frac{dA_f}{A_f} = \frac{1}{(1 + \beta A)} \cdot \frac{dA}{A}$$

↓
fractional change
of gain with fb. = $\boxed{\frac{1}{(1 + \beta A)}}$. fractional change of
gain w/o fb

↓
Sensitivity

$(1 + \beta A) \rightarrow$ Desensitivity

* for a good s/s, sensitivity $\downarrow \Rightarrow$ Desensitivity \uparrow
(i.e. less sensitive to noise, temp. etc)

(2). Increase in I/p impedance :

$$V_i = V_s - V_f$$

$$V_i = V_s - \beta V_o$$

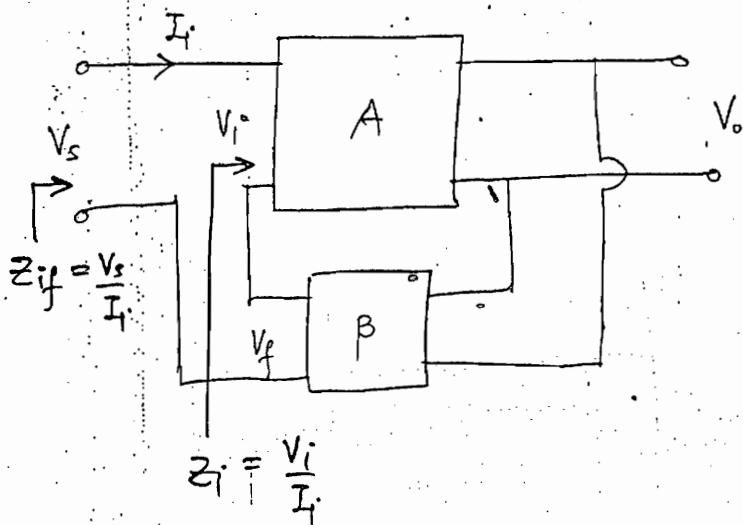
$$V_i = V_s - \beta A \cdot V_o$$

$$V_i(1 + \beta A) = V_s$$

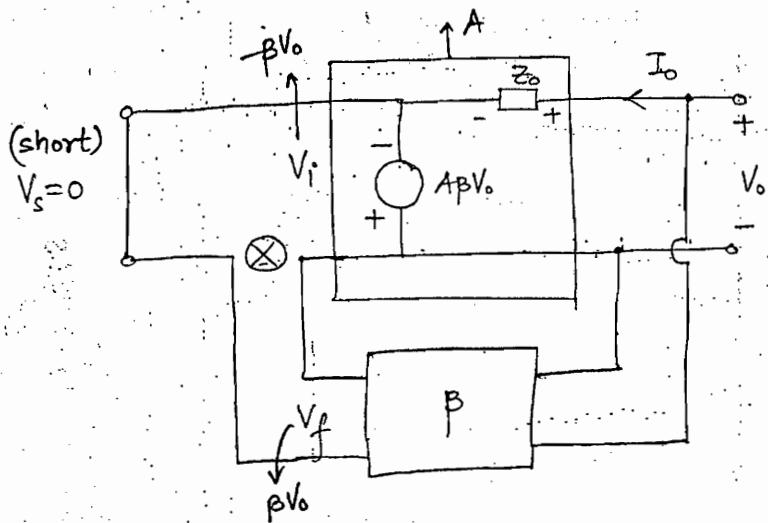
$$\frac{V_i}{I_i} (1 + \beta A) = \frac{V_s}{I_i}$$

$$Z_i(1 + \beta A) = Z_{if}$$

$$\therefore Z_{if} = (1 + \beta A) \cdot Z_i$$



(3). Decrease in o/p impedance:

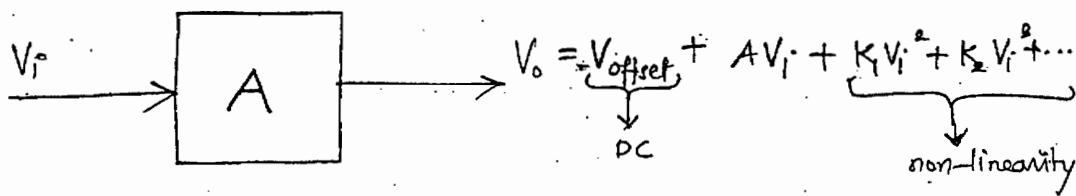


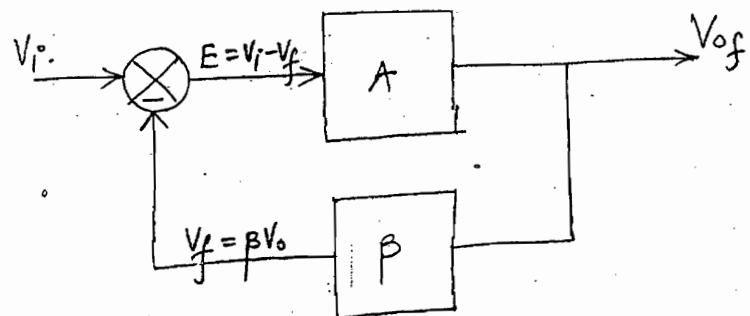
$$V_o + \beta A V_o = I_o Z_o$$

$$-V_o (1 + \beta A) = I_o Z_o$$

$$Z_{if} = \frac{V_o}{I_o} = \frac{Z_o}{(1 + \beta A)}$$

(4). Linearity:





$$V_{of} = V_{offset} + AE + K_1 E^2 + K_2 E^3 + \dots$$

$$V_{of} = V_{offset} + A(V_i - V_f) + K_1 E^2 + K_2 E^3 + \dots$$

$$V_o(1 + \beta A) = V_{offset} + AV_i + K_1 E^2 + K_2 E^3 + \dots$$

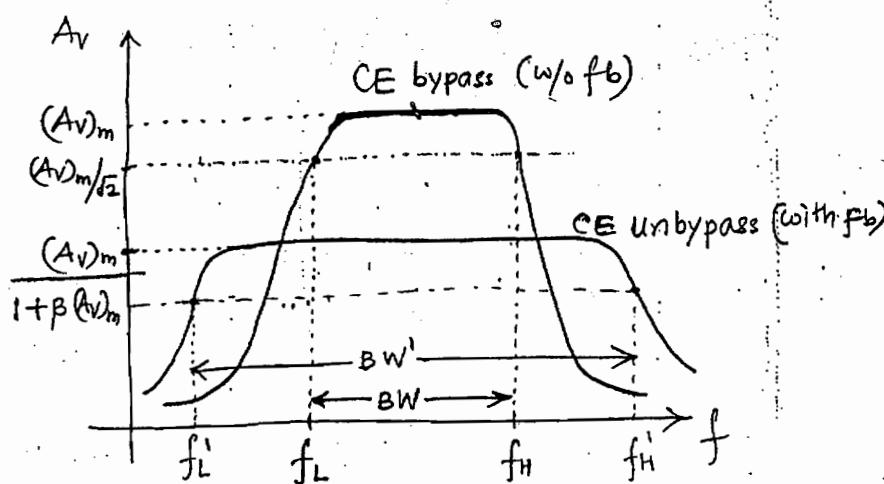
$$V_o = \frac{V_{offset}}{(1 + \beta A)} + \frac{A}{(1 + \beta A)} \cdot V_i + \frac{K_1 E^2 + K_2 E^3 + \dots}{(1 + \beta A)}$$

↓ tends to zero

$$V_o \approx \frac{A}{(1 + \beta A)} \cdot V_i \quad \text{and } \beta A \gg 1$$

$$\therefore V_o \approx \frac{1}{\beta} V_i \rightarrow \text{linear relation}$$

(5) Increase in BW



Lower cut off frequency

$$(A_v)_e = \frac{(A_v)_m}{1 - j(f_L/f)} \quad ; \quad (A_v)_{ef} = \frac{(A_v)_e}{1 + \beta(A_v)}$$

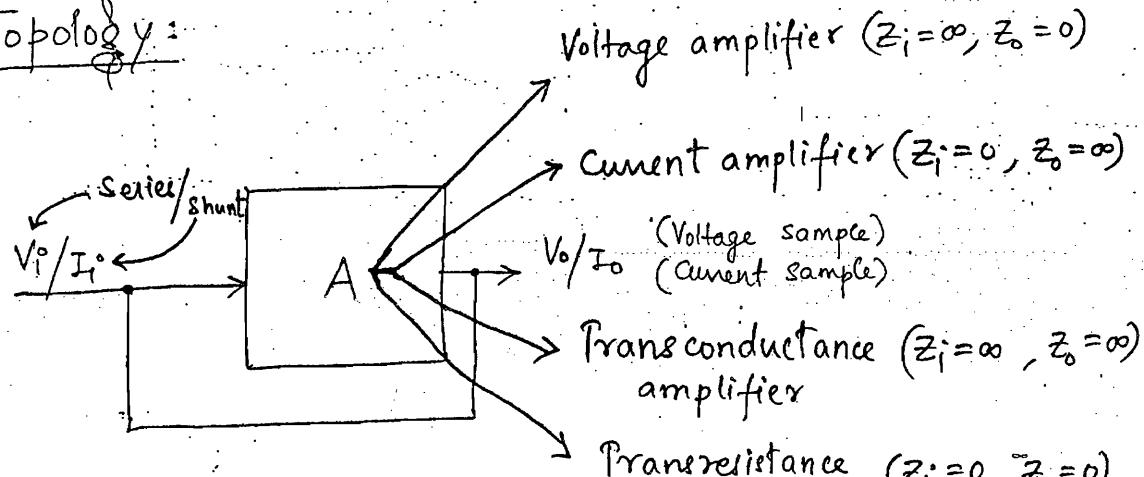
$$(Av)_f = \frac{(Av)_m}{1 - j(f_L/f)} = \frac{(Av)_m}{1 + \beta \cdot \frac{(Av)_m}{1 - j(f_L/f)}} = \frac{(Av)_m}{1 + \beta(Av)_m - j(f_L/f)}$$

$$(Av)_{if} = \frac{(Av)_m}{1 + \beta(Av)_m} \cdot \frac{1 - j \left[\frac{f_L}{1 + \beta(Av)_m} \right] \cdot f}{f}$$

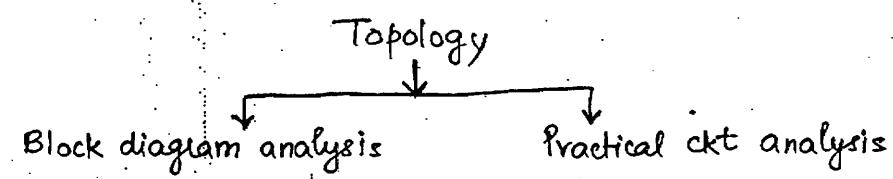
$$\therefore f_L' = \frac{f_L}{1 + \beta(Av)_m}$$

Similarly, $f_H' = f_H (1 + \beta(Av)_m)$

Topology:

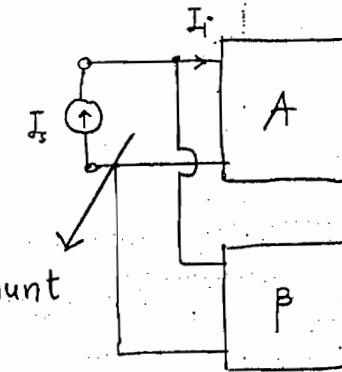
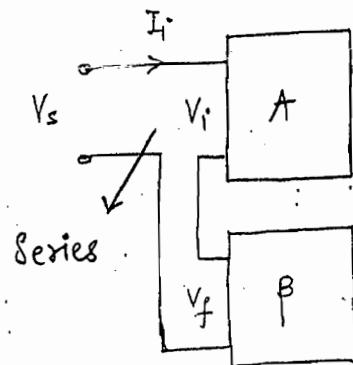


	O/P	I/P
(1).	Voltage	Series
(2).	Voltage	Shunt
(3).	Current	series
(4).	Current	shunt

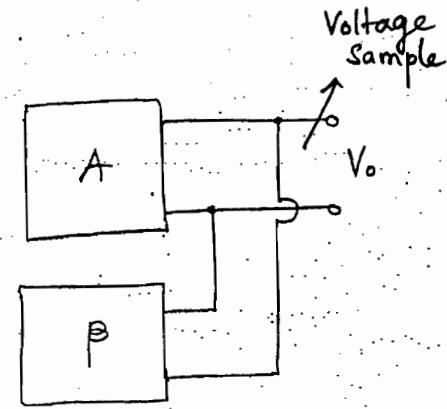
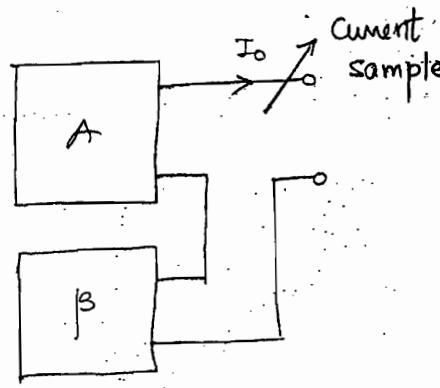


Block Diagram Analysis :

i/p side

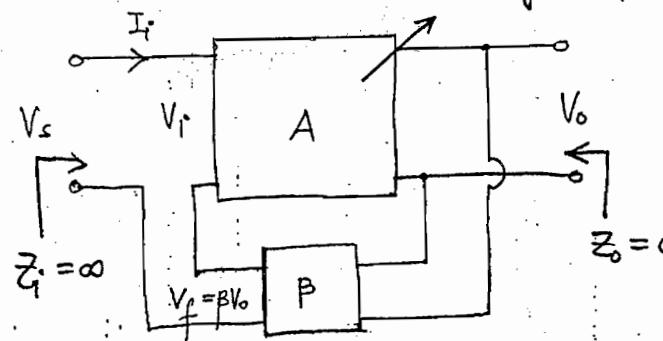


%/p side



(I). Voltage Series :

Voltage amplifier



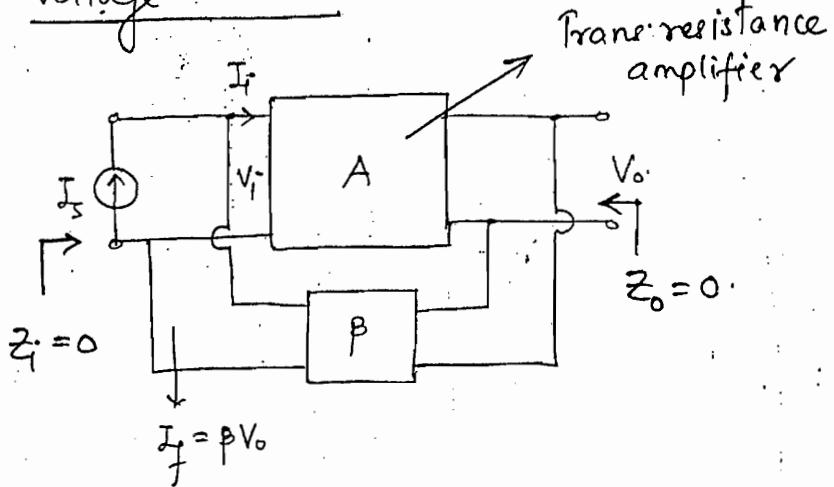
Voltage Series
(%/p)
(or)

Series Voltage
(%/p)
(or)

Series Shunt
(%/p)
(or)

Voltage Voltage
(%/p)

(2). Voltage shunt:



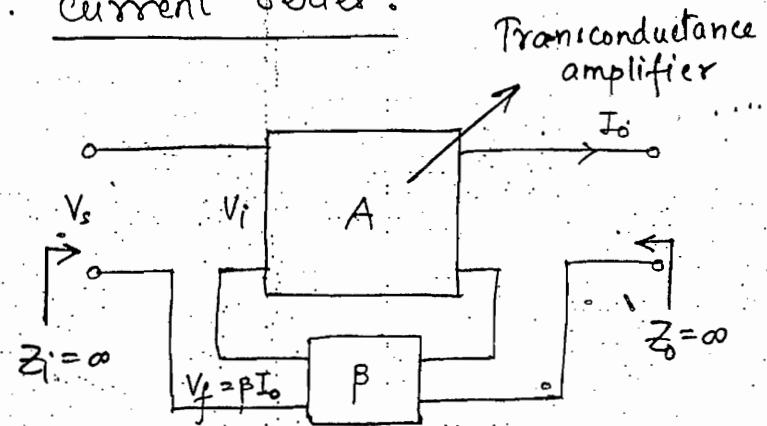
Voltage shunt
(or)

Shunt Voltage
(or)

Shunt shunt
(or)

Voltage current

(3). Current series:



Current series

(or)

Series current

(or)

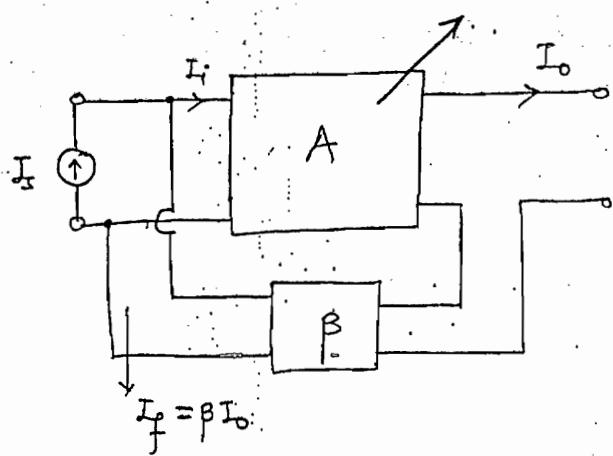
series series

(or)

Current voltage

(4). Current shunt:

current amplifier



Current shunt

(or)

shunt current

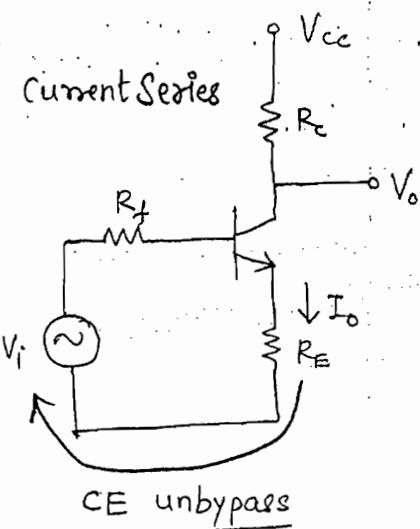
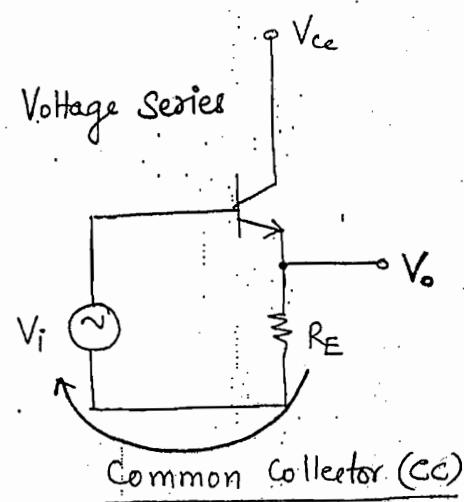
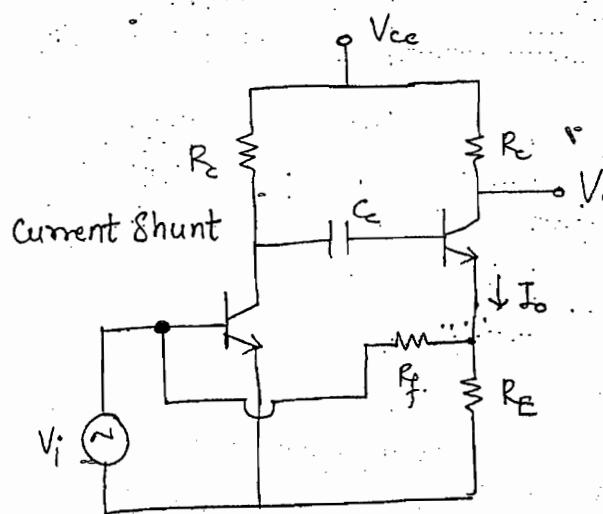
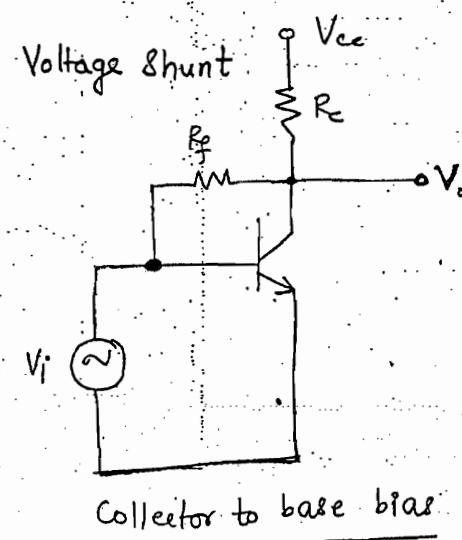
(or)

shunt series

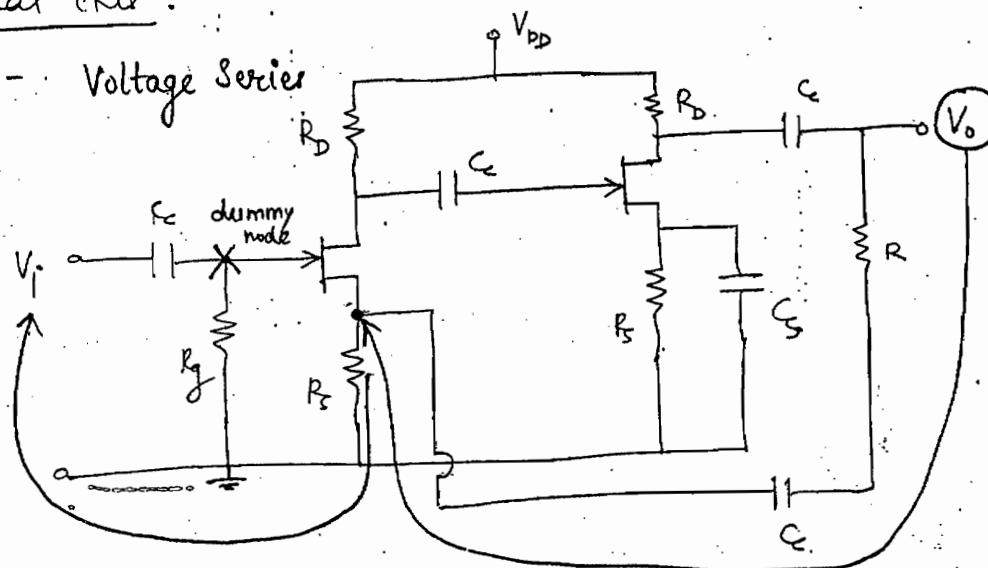
(or)

current current

Practical ckt analysis :

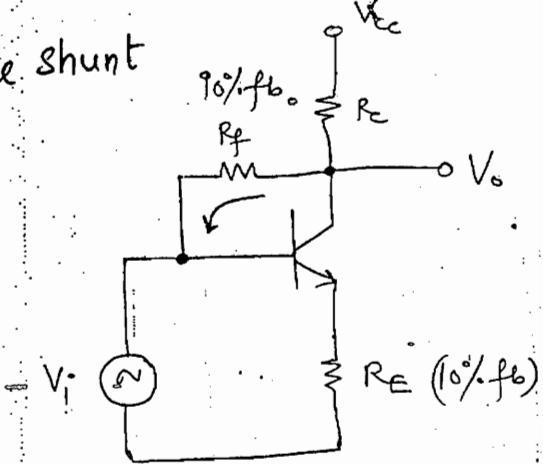
ckt - 1ckt - 2ckt - 3ckt - 4

Typical ckt's :

FET - Voltage Series

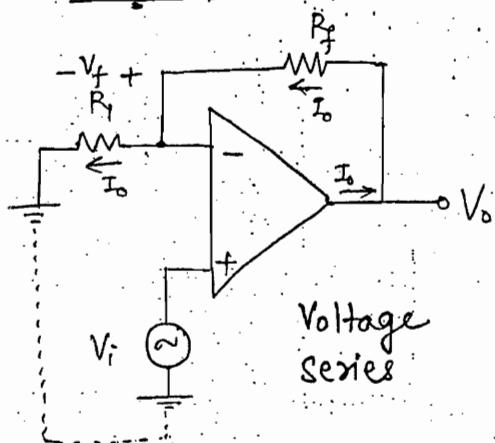
BJT -

Voltage Shunt

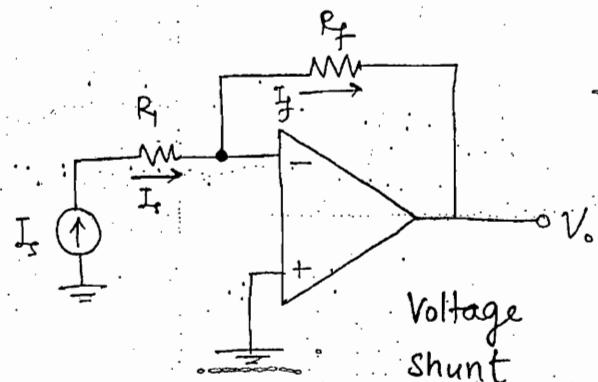


* If shunt not works completely, then series fb are required (2 things at a time).

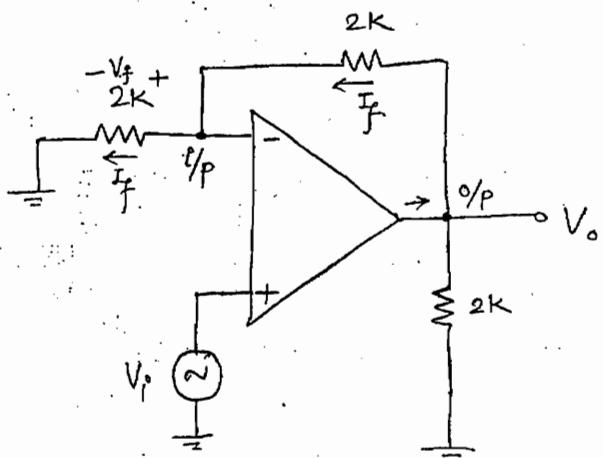
Note: If 2 fb i.e. series & shunt are connected at a time, shunt dominates series fb.

OP-Amps -

Non-inverting Amplifier

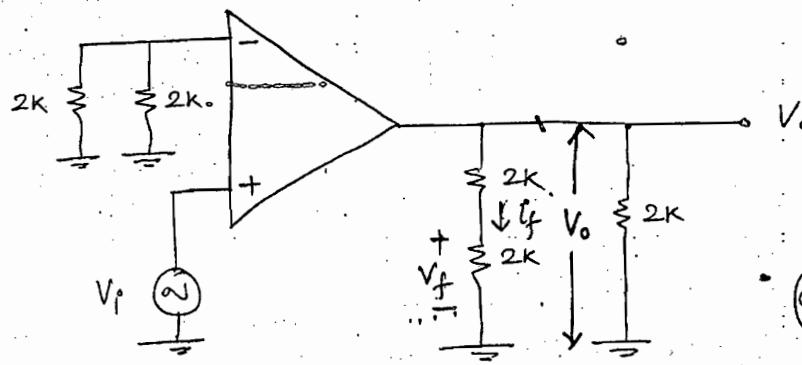
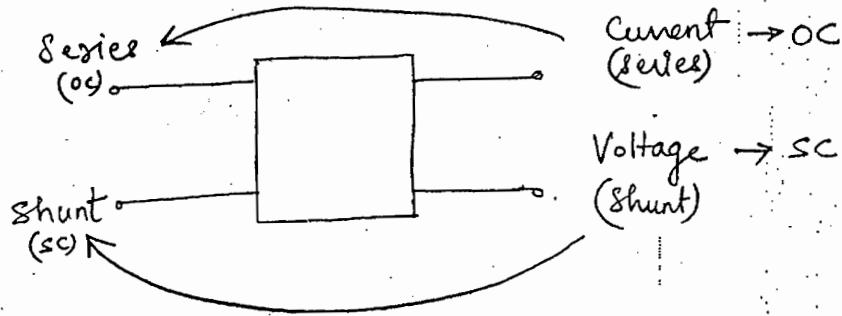


Inverting Amplifier

Problems in (re) fb theory :Model - (1) :

Voltage Series

Calculate A_f and β ?



$$\beta = \frac{V_f}{V_o} = \frac{2K}{2K+2K} = \frac{1}{2}$$

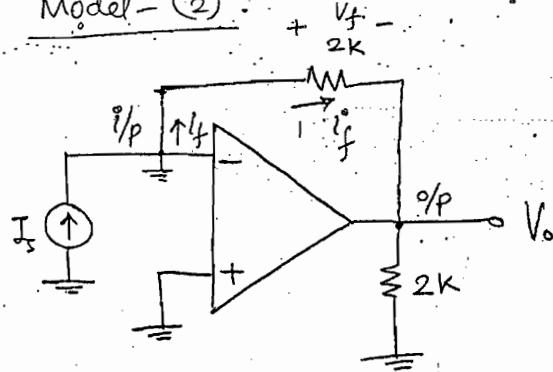
$$A_f = \frac{1}{\beta} = 2$$

$$(or) \quad \beta = \frac{I_f}{V_o} = \frac{1}{4K\Omega}$$

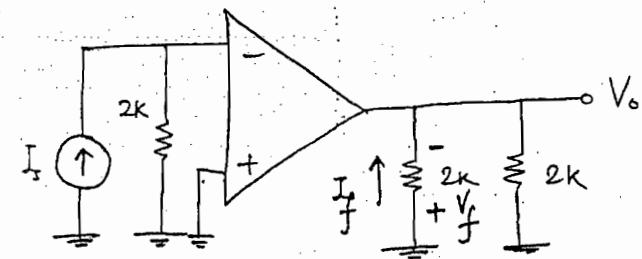
or By virtual ground theory,

$$A_f = \frac{R_f}{R_i} + 1 = \frac{2K}{2K} + 1 = 2.$$

Model - (2) :

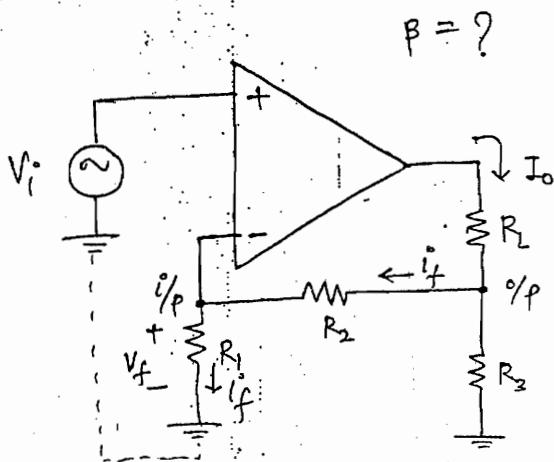


Voltage Shunt



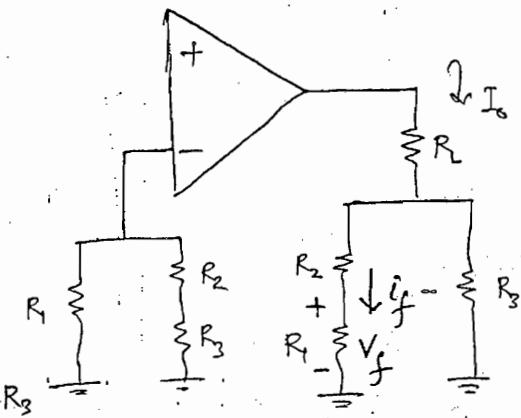
$$\beta = \frac{V_f}{V_o} = -1$$

$$(or) \quad \beta = \frac{I_f}{V_o} = -\frac{1}{2K}$$

Model - (3) :

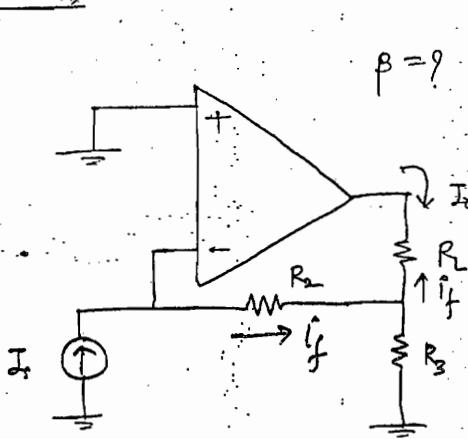
$\beta = ?$

Current series

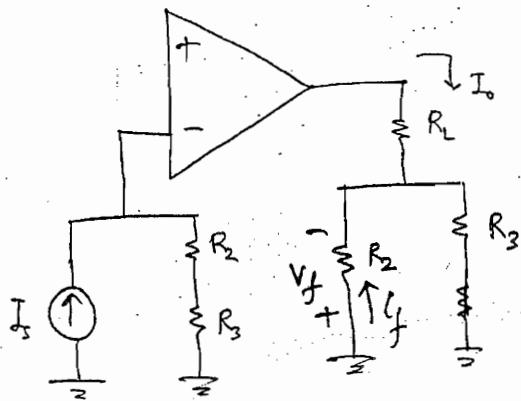


$$i_f = \frac{I_o \times R_3}{(R_1 + R_2 + R_3)} \Rightarrow \beta = \frac{i_f}{I_o} = \frac{R_3}{R_1 + R_2 + R_3}$$

$$(or) \beta = \frac{V_f}{I_o} = \frac{i_f \cdot R_1}{I_o} = \frac{R_1 R_3}{(R_1 + R_2 + R_3)}$$

Model - (4) :

Current shunt

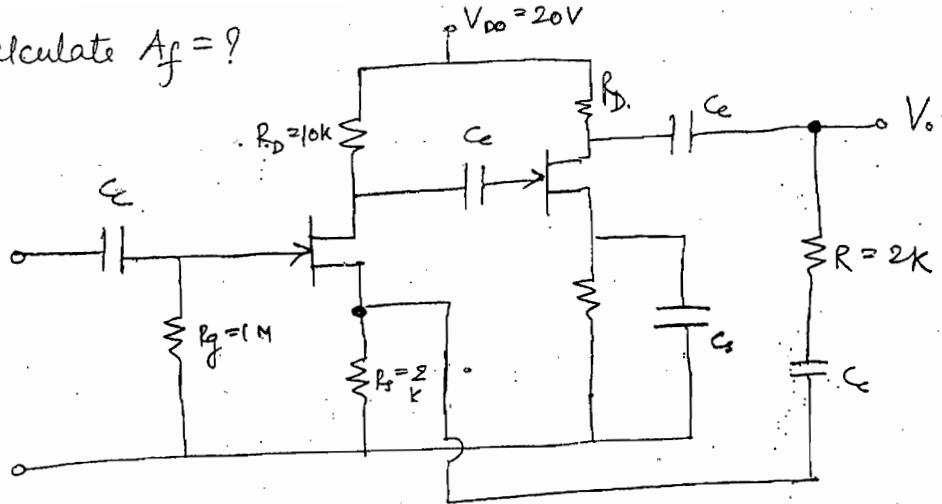


$$i_f = \frac{-I_o \times R_3}{(R_2 + R_3)} \Rightarrow \beta = \frac{i_f}{I_o} = \frac{-R_3}{R_2 + R_3}$$

$$(or) \beta = \frac{V_f}{I_o} = \frac{i_f \cdot R_2}{I_o} = \frac{-R_2 R_3}{R_2 + R_3}$$

Q. Calculate $A_f = ?$

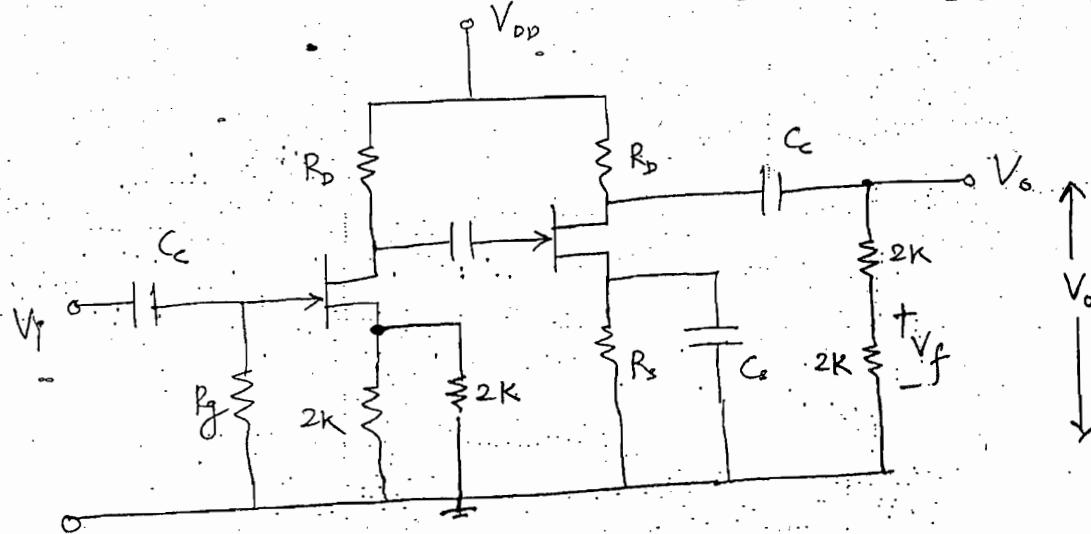
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\Downarrow

O/P \downarrow
Voltage
 \downarrow
SC

O/P
 \downarrow
series
 \downarrow
OC



$$\beta = \frac{V_f}{V_o} = \frac{2k}{2k+2k} = \frac{1}{2}$$

$$A_f = \frac{1}{\beta} = 2$$

Procedure for feedback :

- So get the O/P side, O/P should be neglected.
- for voltage series or voltage shunt, O/P node should be grounded.
- for current series or current shunt, O/P loop should be open.

To get the o/p side, i/p should be neglected.

→ for voltage series or current series, i/p loop should be open.

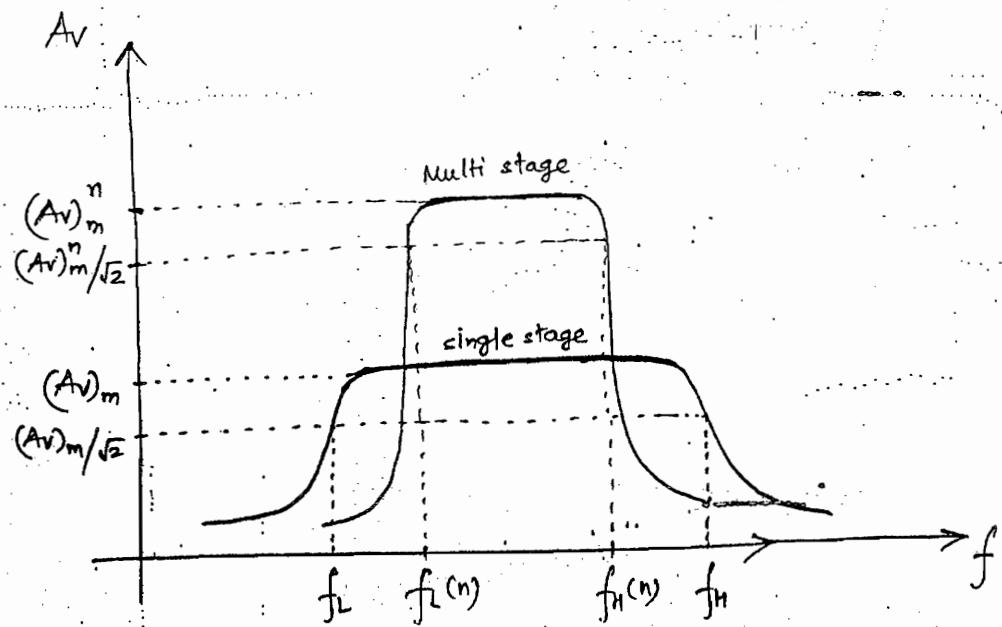
→ for voltage shunt or current shunt, i/p node should be grounded.

Integrated Theory (OP-Amps) :

Multistage Amplifiers:

(Integration → multistage)

Effect of cascading on BW: (identical stages)



High cut-off frequency:

$$\text{Single stage} \rightarrow (Av)_h = \frac{(Av)_m}{1 + j(f/f_H)}$$

$$\left| \frac{(Av)_h}{(Av)_m} \right| = \sqrt{\frac{1}{1 + (f/f_H)^2}}$$

$$n \text{ identical stages} \rightarrow \left| \frac{(Av)_h}{(Av)_m} \right|^n = \left| \frac{1}{\sqrt{1 + (f/f_H)^2}} \right|^n$$

when $f \rightarrow f_H^{(n)}$

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

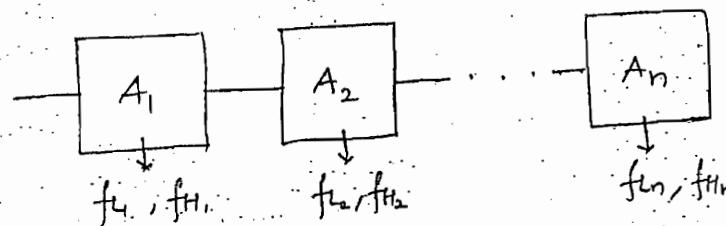
$$1 + \left\{ \frac{f_H(n)}{f} \right\}^2 = 2^{V_n}$$

$$f_H(n) = \sqrt{2^{V_n} - 1} \cdot f_H$$

Similarly,

$$f_L(n) = \frac{f_L}{\sqrt{2^{V_n} - 1}}$$

Effect of cascading on BW : (Non identical stages)



$$A^n = A_1 \cdot A_2 \cdot A_3 \cdots \cdots \cdot A_n$$

$$A_L^n = A_{L1} \cdot A_{L2} \cdot A_{L3} \cdots \cdots \cdot A_{Ln}$$

$$A_H^n = A_{H1} \cdot A_{H2} \cdot A_{H3} \cdots \cdots \cdot A_{Hn}$$

$$\frac{A^n}{\sqrt{1 + \left(\frac{f_L^n}{f} \right)^2}} = \frac{A_1}{\sqrt{1 + \left(\frac{f_{L1}}{f} \right)^2}} \cdot \frac{A_2}{\sqrt{1 + \left(\frac{f_{L2}}{f} \right)^2}} \cdots \frac{A_n}{\sqrt{1 + \left(\frac{f_{Ln}}{f} \right)^2}}$$

$$1 + \left(\frac{f_L^n}{f} \right)^2 = 1 + \left(\frac{f_{L1}}{f} \right)^2 + \left(\frac{f_{L2}}{f} \right)^2 + \cdots + \left(\frac{f_{Ln}}{f} \right)^2 + \underbrace{\left(\frac{f_{L1} f_{L2}}{f^2} \right)^2 + \cdots}_{\text{neglected}}$$

$$f_L^n = \sqrt{f_{L1}^2 + f_{L2}^2 + \cdots + f_{Ln}^2}$$

$$\text{Similarly, } f_H^n = \frac{1}{\sqrt{\left(\frac{1}{f_{H1}} \right)^2 + \left(\frac{1}{f_{H2}} \right)^2 + \cdots + \left(\frac{1}{f_{Hn}} \right)^2}}$$

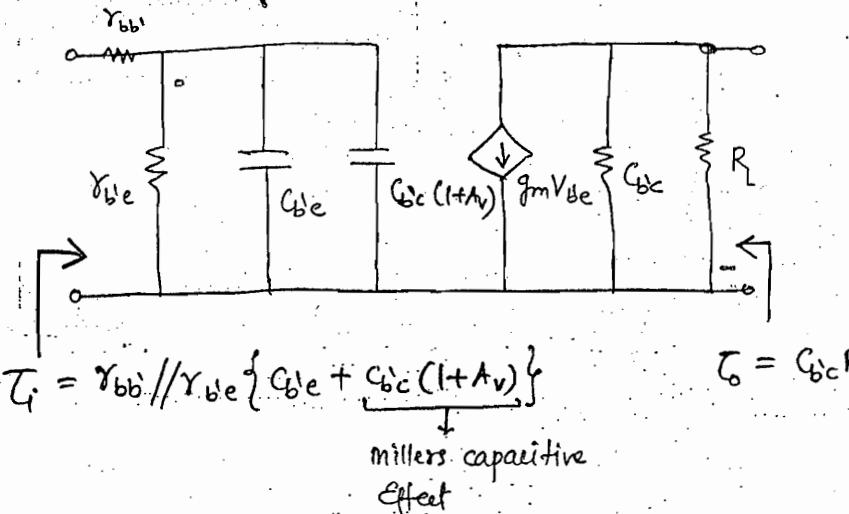
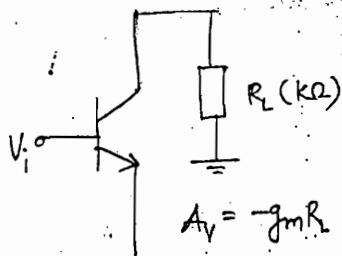
Important cascading designs:

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- (1). Cascode amplifiers : $(CE - CB)$ (Wide Band Amplifiers)
- (2). Darlington pair : $(CC - CC)$ (High I/p impedance ckt)

- (1). Cascode amplifiers ($CE - CB$)

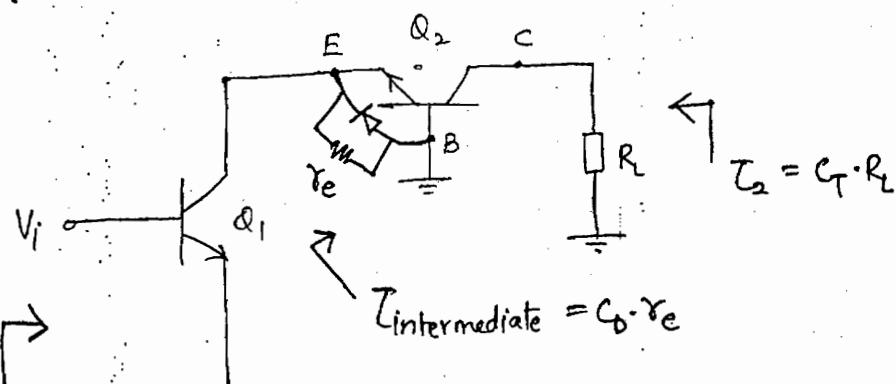
CE:



$$\text{As } A_V = 1000 \quad \therefore T_1 \uparrow = \frac{1}{\omega_{BW} \downarrow} \quad (T_1 \gg T_0)$$

To reduce, T_1 , there must be reduction in A_V to improve ω_{BW} .

CE - CB:



$$\downarrow T_1 = Y_{BB} // Y_{B'e} \{ C_{B'e} + C_{Bc}(a) \} = \frac{1}{\omega_{BW} \uparrow}$$

$$A_{V1} = -g_m R_L$$

$$= -g_m Y_e$$

$$= -\frac{Y_e}{Y_0} = 1$$

$$A_{V2} = g_m R_L$$

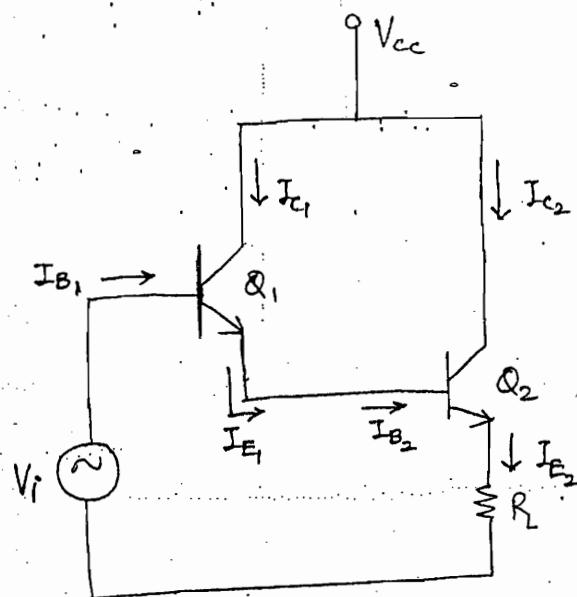
$$A_{\text{cascade}} = A_{V1} \cdot A_{V2}$$

$$= -1 \cdot g_m R_L$$

$$= -g_m R_L \quad (= A_V)_{CE}$$

Hence, with CE-CE, ω_{BW} increases with keeping 177
the gain ($A_{\text{cascade}} = A_V = -g_m R_L$) same as with
single CE.

(2). Darlington pair: (CC-CC)



A

$$(4). Z_o = R_E // Y_e$$

$$\approx Y_e$$

$$\therefore Z_o = Y_e \rightarrow \text{Impedance matching}$$

$$(1). A_I = A_{I_1} \cdot A_{I_2}$$

$$\begin{aligned} A_{I_1} &= \frac{I_{E_1}}{I_{B_1}} \\ &= \frac{I_{B_1} + I_{C_1}}{I_{B_1}} \\ &= 1 + \frac{I_{C_1}}{I_{B_1}} \\ &= 1 + \beta \end{aligned}$$

$$\begin{aligned} A_{I_2} &= \frac{I_{E_2}}{I_{B_2}} \\ &= \frac{I_{B_2} + I_{C_2}}{I_{B_2}} \\ &= 1 + \frac{I_{C_2}}{I_{B_2}} \\ &= 1 + \beta \end{aligned}$$

$$\therefore A_I = (1 + \beta)^2 \approx \beta^2 \Rightarrow A_I = \beta^2$$

$$(2). A_V = A_{V_1} \cdot A_{V_2}$$

$$\therefore A_V = 1 \rightarrow \text{Buffer action}$$

$$(3). Z_i = A_I \cdot R_L$$

$$\therefore Z_i = \beta^2 \cdot R_L$$

$$\text{Assume: } \beta = 100, R_L = 1 \text{ k}\Omega$$

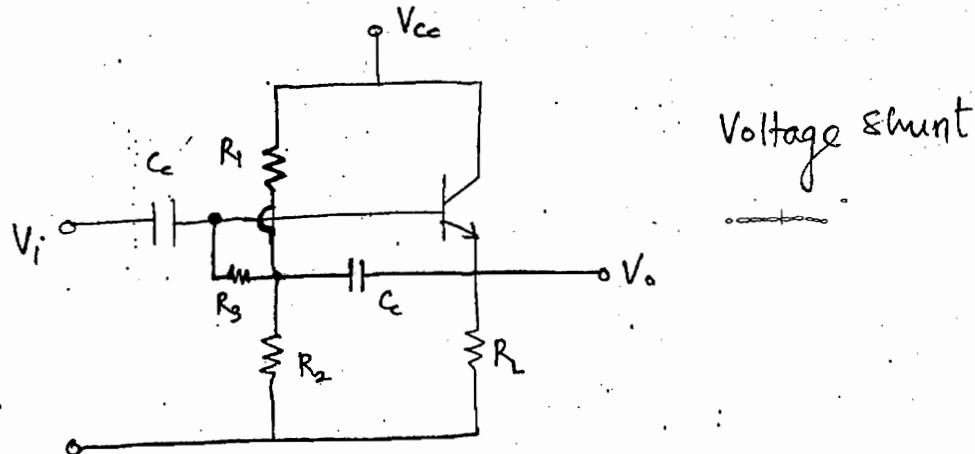
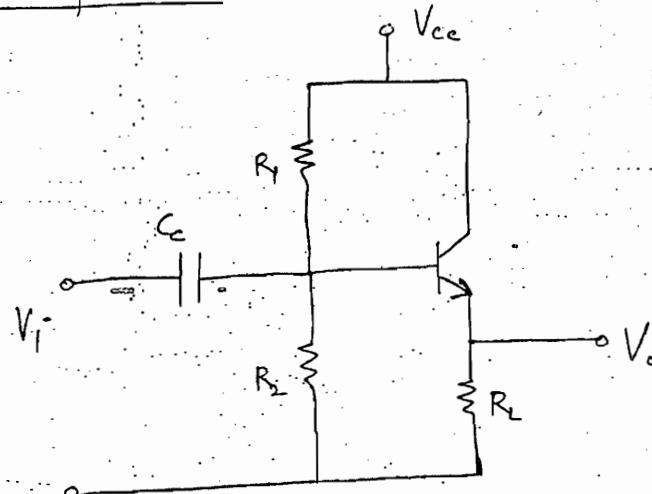
$$Z_i = (100)^2 1 \text{ k}\Omega = 10 \text{ M}\Omega \text{ (High)}$$

Disadvantages:

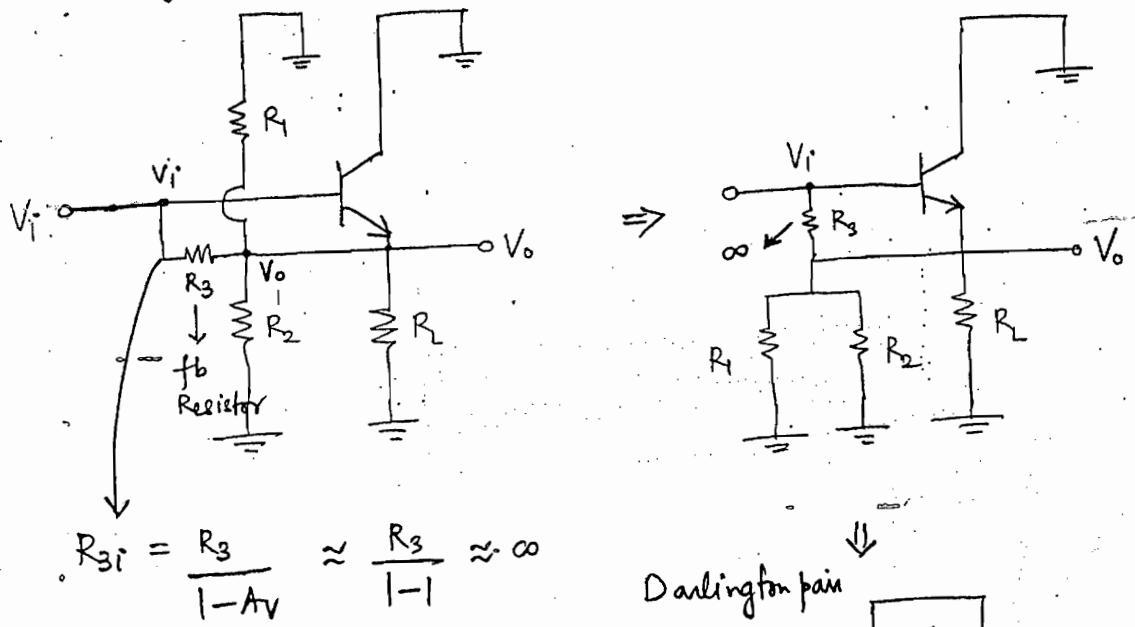
- (1). In practical design, the I/p impedance will be low for a darlington pair becoz of biasing resistors.

e.g. $Z_i \parallel R_1 \parallel R_2$ if $R_1 = 100k\Omega$
 $R_2 = 10k\Omega$
 $Z_i = Z_i \parallel 100k \parallel 10k$
 $\approx 10k$

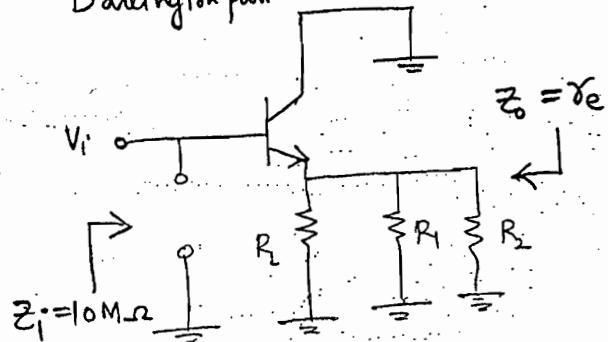
Emitter follower : Boot strap



AC analysis:



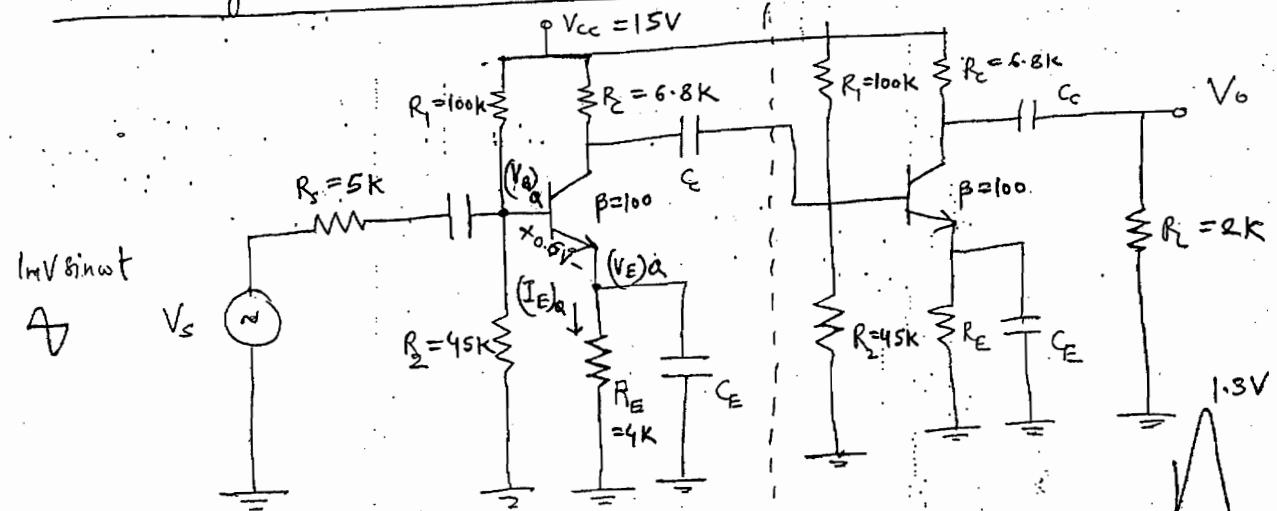
Darlington pair



Coupling designs :

- (1). RC coupling
- (2). Direct coupling

Two stage RC coupled CE amplifier :



$$\text{Calculate : } A_{Vs} = \frac{V_o}{V_s}$$

DC analysis: α_1 & α_2

$$(V_B)_Q = \frac{15 \times 45K}{(100K + 45K)} = 4.6V$$

$$(V_E)_Q = 4.6 - 0.6 = 4V$$

$$(I_E)_Q = \frac{(V_E)_Q}{R_E} = \frac{4}{4K} = 1mA$$

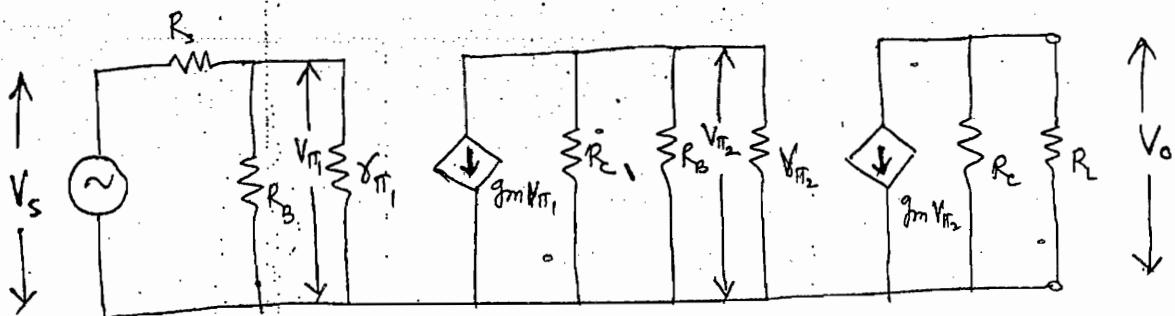
$$g_m = \frac{(I_c)_Q}{V_T} \approx \frac{1mA}{25mV} = 40mV^{-1}$$

$$\gamma_e = \frac{1}{g_m} = 25\Omega$$

$$\beta \gamma_e = \gamma_{\pi} = h_{ie} = 100 \times 25 = 2.5k\Omega$$

AC analysis:

$$R_B = R_1 // R_2 = 100K // 45K = 32K$$



$$A_{VS} = \frac{V_o}{V_s} = \frac{V_o}{V_{\pi 2}} \cdot \frac{V_{\pi 2}}{V_{\pi 1}} \cdot \frac{V_{\pi 1}}{V_s}$$

$$\frac{V_{\pi 1}}{V_s} = \frac{R_B // \gamma_{\pi 1}}{R_B // \gamma_{\pi 1} + R_E} = \frac{32K // 2.5K}{32K // 2.5K + 5K} = 0.316K$$

$$V_{\pi 2} = -g_m V_{\pi 1} \cdot (R_E // R_B // \gamma_{\pi 2})$$

$$\frac{V_{\pi 2}}{V_{\pi 1}} = -g_m (R_E // R_B // \gamma_{\pi 2}) = -40mV \cdot (6.8K // 32K // 2.5K) = -69$$

$$V_o = -g_m V_{\pi_2} (R_L // R_L)$$

$$\frac{V_o}{V_{\pi_2}} = -g_m (R_L // R_L) = -40 \text{ mV} (6.8 \text{ K} // 2 \text{ K}) = -61$$

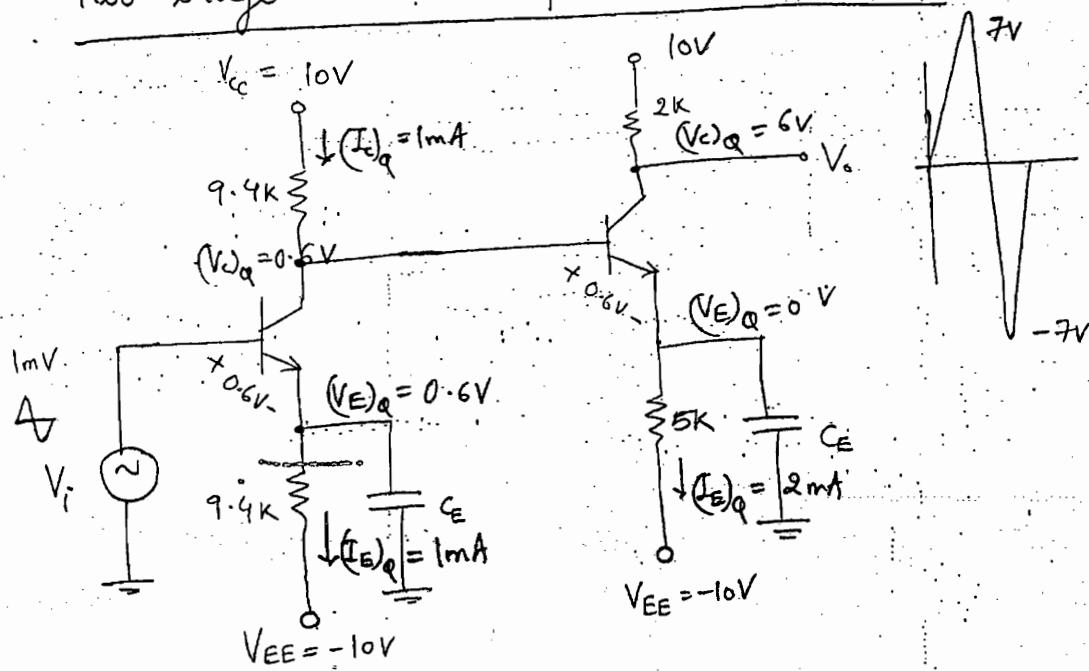
$$A_{Vs} = 0.316 \times -69 \times -61 \\ = 1330$$

$$\therefore \frac{V_o}{V_s} = 1330$$

$$\text{Eq. } V_o = 1330 \times 1 \text{ mV} = 1.3 \text{ V}$$

ME, 20
SU
E

Two stage direct coupled BJT amplifier -



Q₁:

$$g_{m1} = \frac{(I_c)_q}{V_T} = \frac{1 \text{ mA}}{25 \text{ mV}} = 40 \text{ mV}$$

$$\gamma_{e_1} = \frac{1}{g_{m1}} = 25 \Omega$$

$$\beta \gamma_{e_1} = \gamma_{\pi_1} = 2.5 \text{ k}\Omega$$

Q₂:

$$g_{m2} = \frac{(I_c)_q}{V_T} = \frac{2 \text{ mA}}{25 \text{ mV}}$$

$$= 80 \text{ mV}$$

$$\gamma_{e_2} = \frac{1}{g_{m2}} = 12.5 \Omega$$

$$\beta \gamma_{e_2} = \gamma_{\pi_2} = 1.25 \text{ k}\Omega$$

$$\begin{aligned} AV_1 &= -g_m (R_e // r_{T_2}) \\ &= -40 \cdot (9.4 \text{ k} // 1.25 \text{ k}) \\ &= -44 \end{aligned}$$

$$\begin{aligned} AV_2 &= -g_m R_e \\ &= -80 \cdot 2 \text{ k} \\ &= -160 \end{aligned}$$

$$\begin{aligned} \therefore AV &= AV_1 \cdot AV_2 \\ &= -44 \times -160 \\ &= 7040 \end{aligned}$$

* This is the choice of op-amp. This type of amplifier
one used in instrumentation and biomedical measurements.
Direct coupled technique is better than RC coupled
technique.

Conclusion:

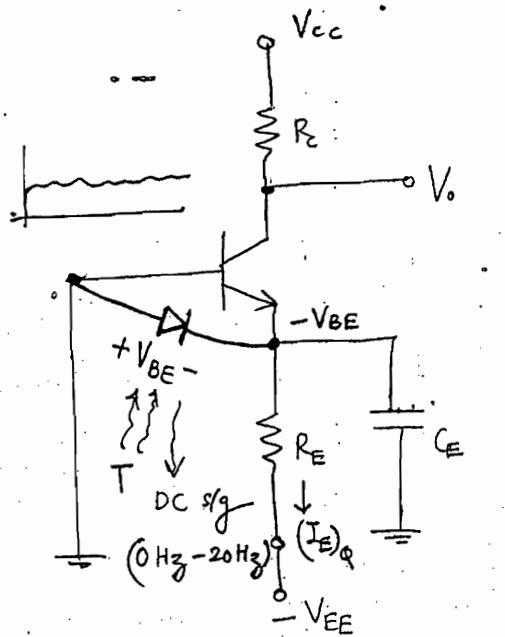
In integrated ckt designs, 2 important conditions
to select an amplifier are:

- (1). It should give high gain.
- (2). The size should be less.

Note: So in integrated ckt's, direct coupled amplifier
became superior than RC coupled amplifiers becoz of
high gain and small in size.

Direct coupled amplifier:

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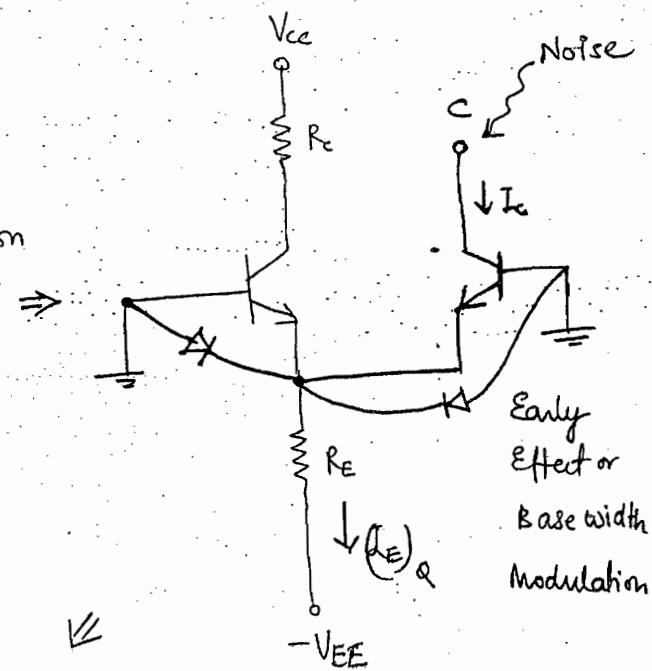
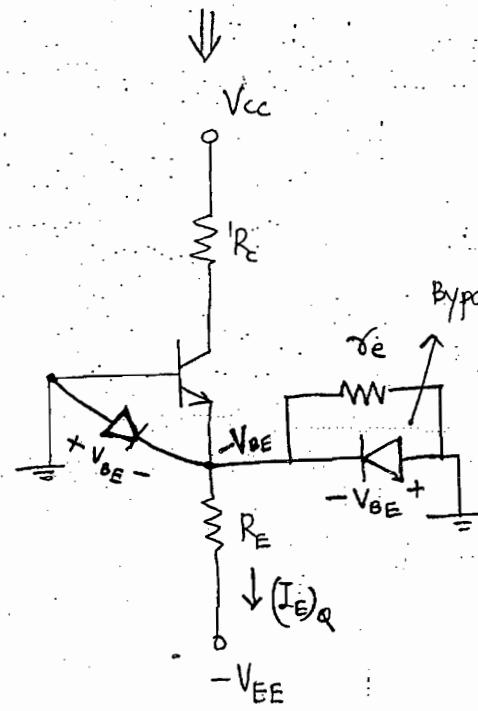
$$(1). \frac{dV_{BE}}{dT} = -2.5 \text{ mV}/^\circ\text{C}$$

↓
dc drift problem.

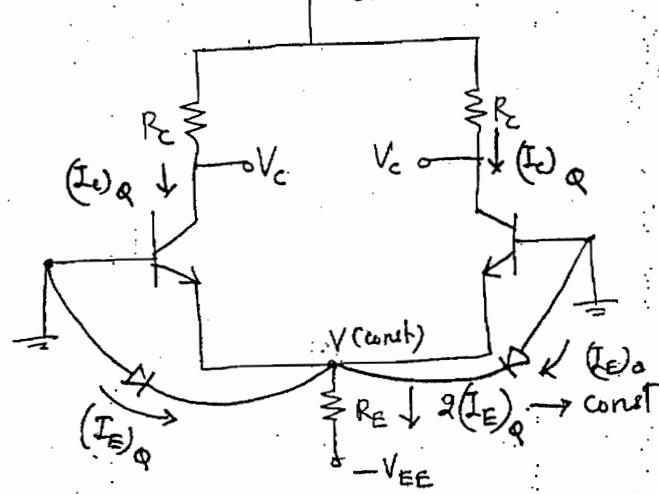
$$(2). f_{min} = \frac{1}{2\pi R_E C_E}$$

for $f_{min} \rightarrow 0 \Rightarrow C_E \rightarrow \infty$

↓
(Impossible to fabricate
on IC practically).

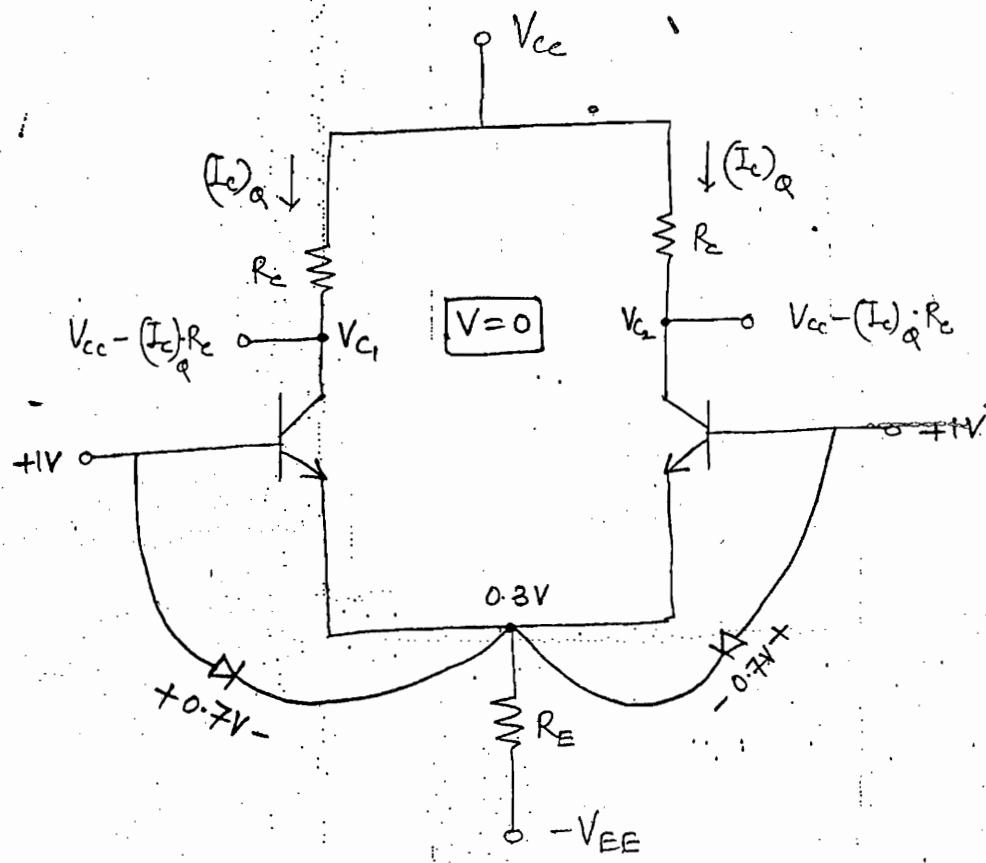


Differential amplifier



Testing of differential amplifier for various I/p:

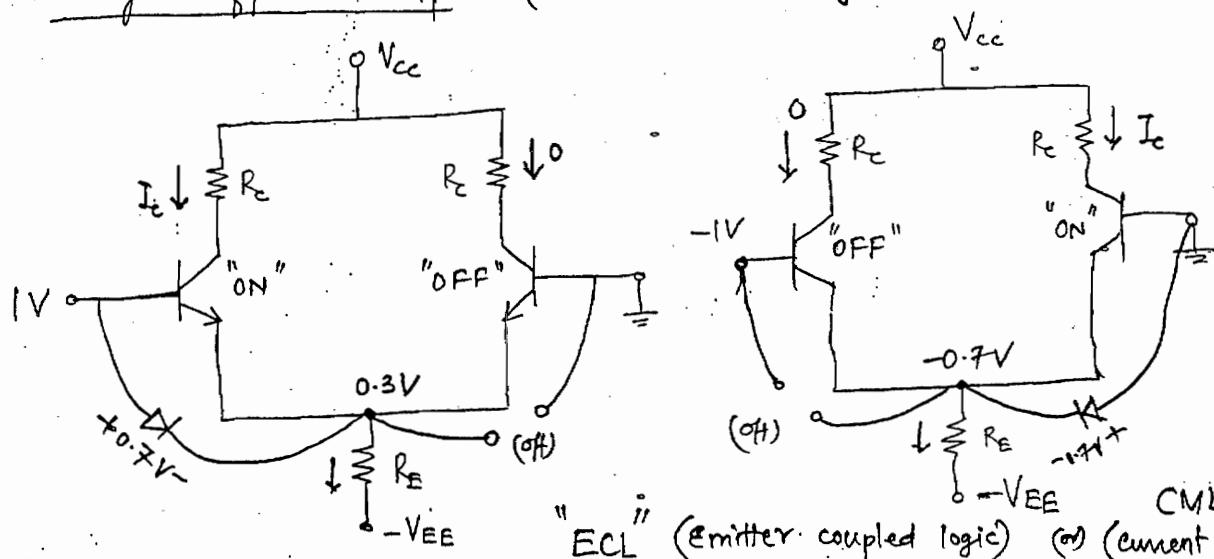
Common I/p:

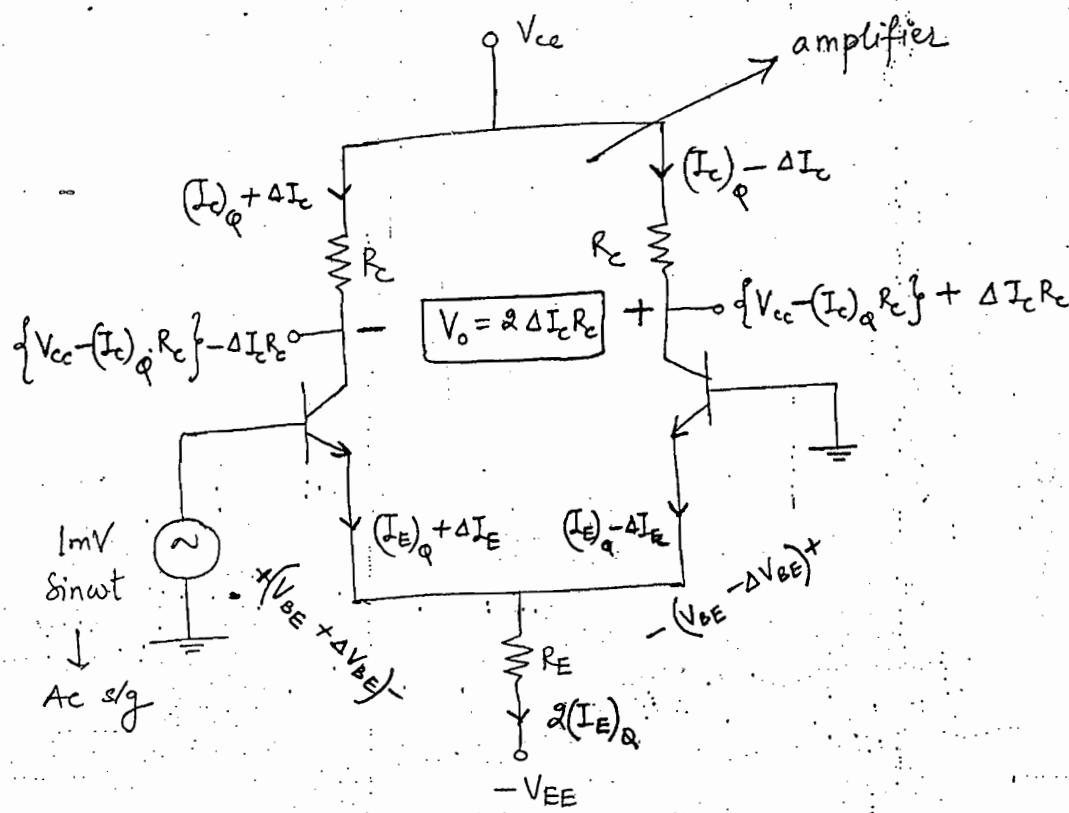


Common s/g are nothing but noise (or) interference s/g which is eliminated by differential amplifier.

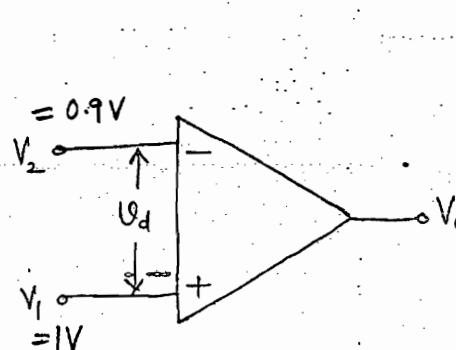
$$\therefore V = V_{C1} - V_{C2} = 0 \text{ for common s/g.}$$

Large difference I/p: ($\geq IV$) \rightarrow switching



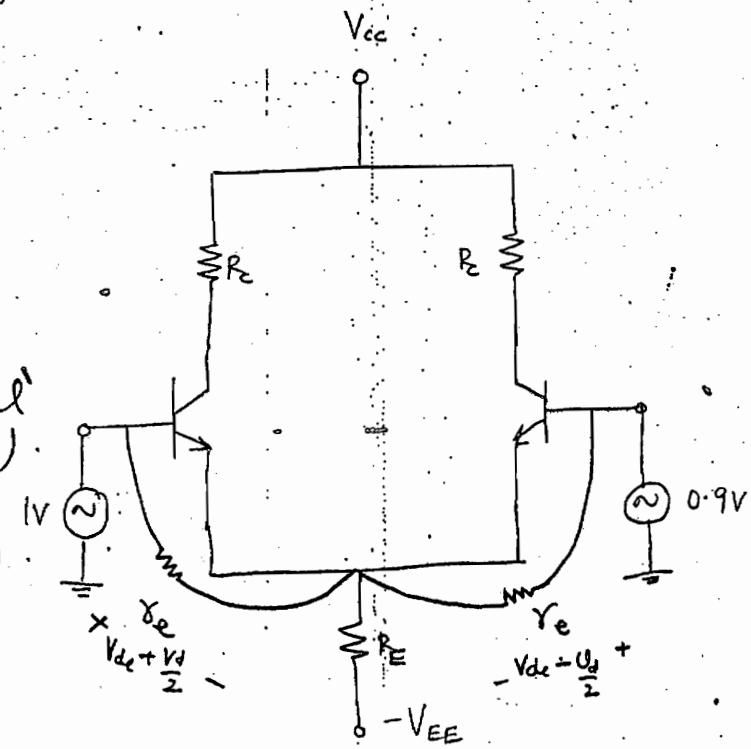


mathematical analysis of differential amplifier :



$$V_d = 1 - 0.9 = V_1 - V_2 \text{ (differential voltage)} \\ = 0.1V$$

$$V_c = V_{dc} = \frac{V_1 + V_2}{2} \text{ (common voltage)} \\ = \frac{1.9}{2} \\ = 0.95V$$



$$V_1 = \frac{V_1 - V_2}{2} + \frac{V_1 + V_2}{2}$$

$$= \frac{U_d}{2} + V_{de}$$

$$V_2 = \frac{V_2 - V_1}{2} + \frac{V_1 + V_2}{2}$$

$$= -\frac{U_d}{2} + V_{de}$$

Mathematics involved inside:

$$V_1 = \frac{1-0.9}{2} + \frac{1+0.9}{2} = \frac{0.1}{2} + 0.95$$

$$-V_2 = \frac{0.9-1}{2} + \frac{1+0.9}{2} = -\frac{0.1}{2} + 0.95$$

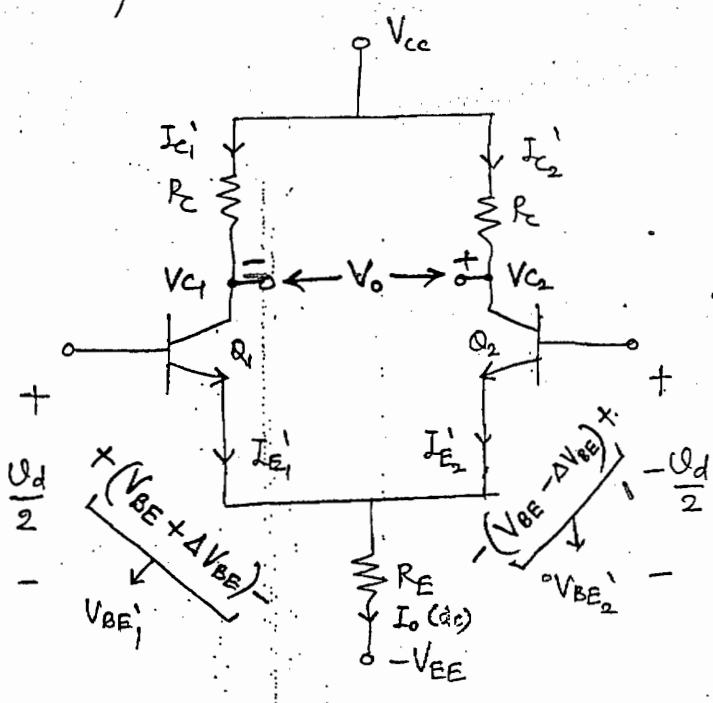
$$U_d = V_1 - V_2$$

$$= \frac{0.1}{2} + 0.95 - (-\frac{0.1}{2}) - 0.95$$

$$= 0.1 \text{ V}$$

Transfer characteristic of OP-Amp :

Analysis:



$$\begin{aligned} V_0 &= V_Q - V_{C2} \\ &= (V_{CC} - I_{e1'} R_E) - (V_{ce} - I_{e2'} R_E) \\ &= (I_{e2'} - I_{e1'}) R_E \end{aligned}$$

$$I_{e1}' = (I_e)_Q + 4I_e$$

$$I_{e2}' = (I_e)_Q - 4I_e$$

$$(I_E)_Q = I_{E0} \cdot e^{\frac{V_{BE}}{V_T}}$$

$$I'_E = I_{E0} \cdot e^{\frac{V_{BE} + \Delta V_{BE}}{V_T}}$$

$$I'_E = I_{E0} \cdot e^{\frac{V_{BE_1}}{V_T}}$$

$$I'_E = I_{E0} \cdot e^{\frac{V_{BE_2}}{V_T}}$$

$$\frac{I'_E}{I'_E} = e^{\underbrace{(V_{BE_1} - V_{BE_2})}_{\text{Qd}} / V_T}$$

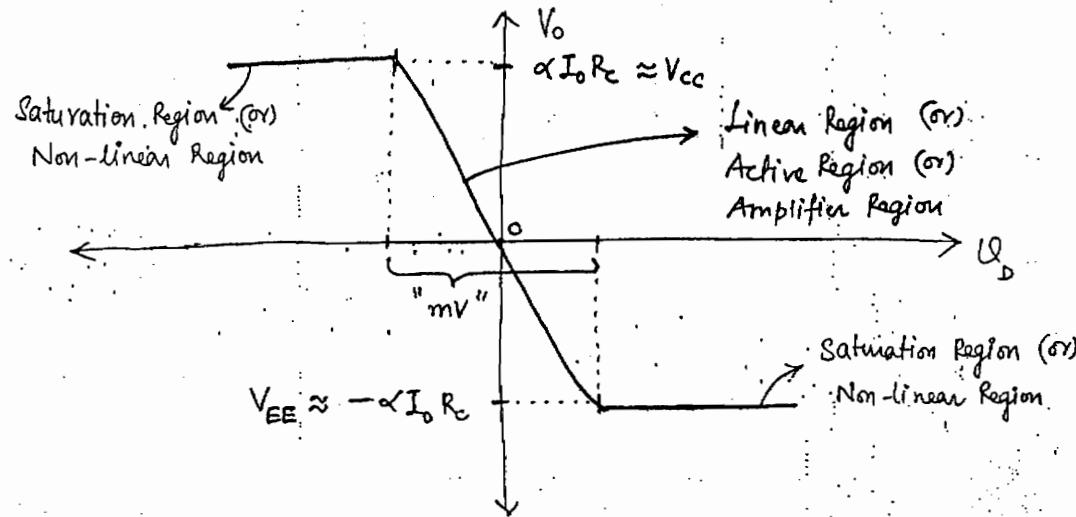
$$\frac{I'_E}{I'_E} = e^{\frac{Qd}{V_T}}$$

$$\frac{I'_E - I'_E}{I'_E + I'_E} = \frac{e^{Qd/V_T} - 1}{e^{Qd/V_T} + 1}$$

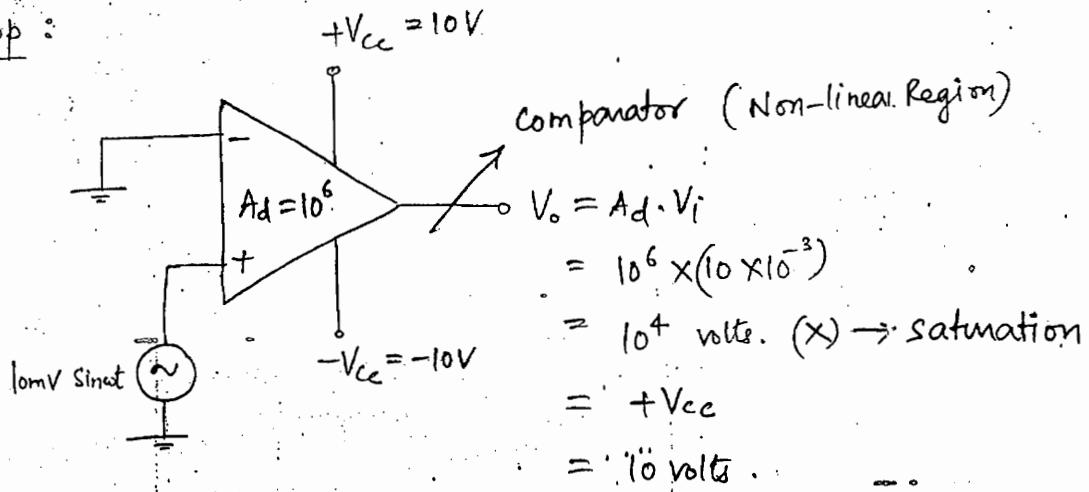
$$(1). \quad I'_E - I'_E = \frac{I'_c - I'_c}{\alpha} \quad (2). \quad I'_E + I'_E = I_o \text{ (dc)}$$

$$\frac{I'_c - I'_c}{\alpha I_o} = \frac{e^{Qd/V_T} - 1}{e^{Qd/V_T} + 1}$$

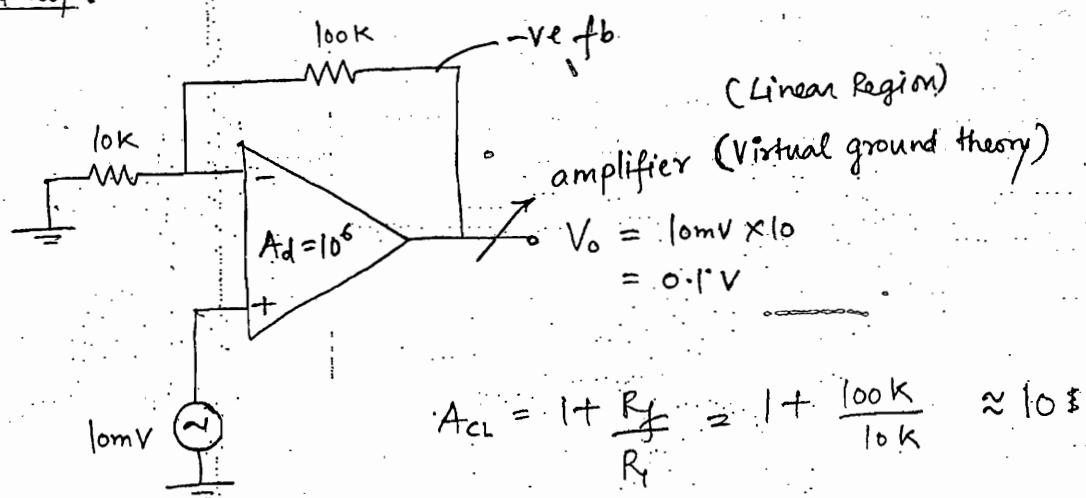
$$V_o = -\alpha I_o R_c \cdot \frac{(e^{Qd/V_T} - 1)}{(e^{Qd/V_T} + 1)}$$



Open loop :



Closed loop :



* Under Linear region, $V_d \rightarrow 0$ (Virtual ground theory)

Advantages of differential amplifier compared to direct coupled amplifier :

- (1). It gives best noise rejection capability.
- (2). There is no need of coupling or bypass capacitors.
- (3). The O/p impedance is high. Typically $1M\Omega$ we can design.

Okt configurations :

At the O/p side we have -

(1) single O/p

(2) Dual O/p

At the o/p side, we have

(1). Balanced o/p (o/p is taken across C_1 & C_2)

(2). Unbalanced o/p (o/p is taken across C_1 to ground or C_2 to ground)

The configurations are:

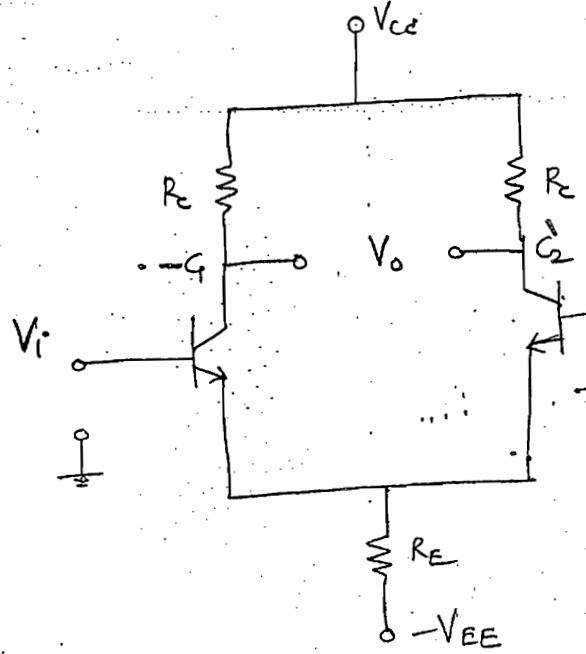
(1). Single I/p balanced o/p

(2). Single I/p unbalanced o/p

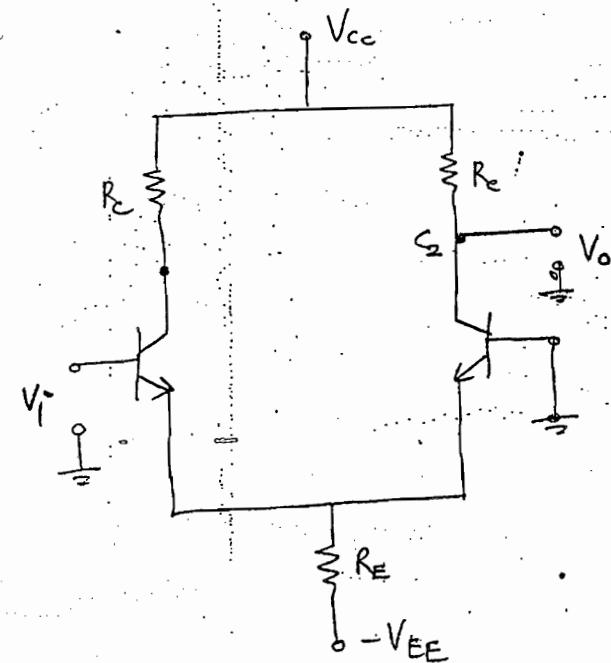
(3). Dual I/p balanced o/p

(4). Dual I/p unbalanced o/p

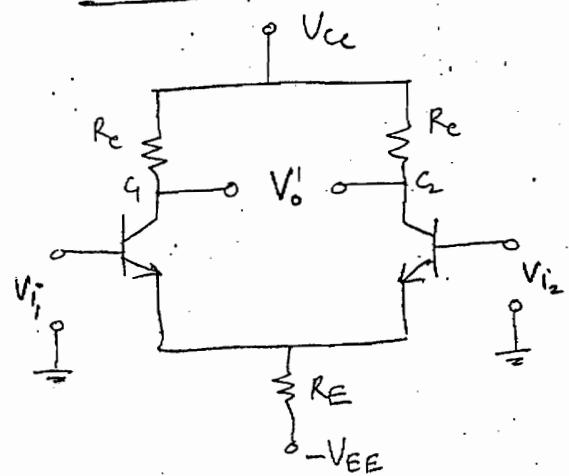
Single I/p balanced o/p:



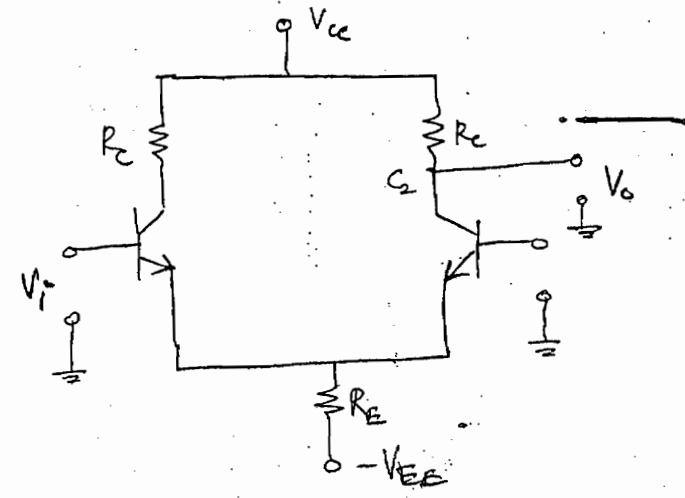
Single I/p unbalanced o/p:



Dual I/p balanced o/p:



Dual I/p unbalanced o/p:



Design problems in differential amplifier:

Q1. Calculate the voltage gain of a single I/p balanced O/p or dual I/p balanced O/p differential amplifier with $R_E = 1\text{ k}\Omega$ and $(I_E)_Q = 26\text{ mA}$.

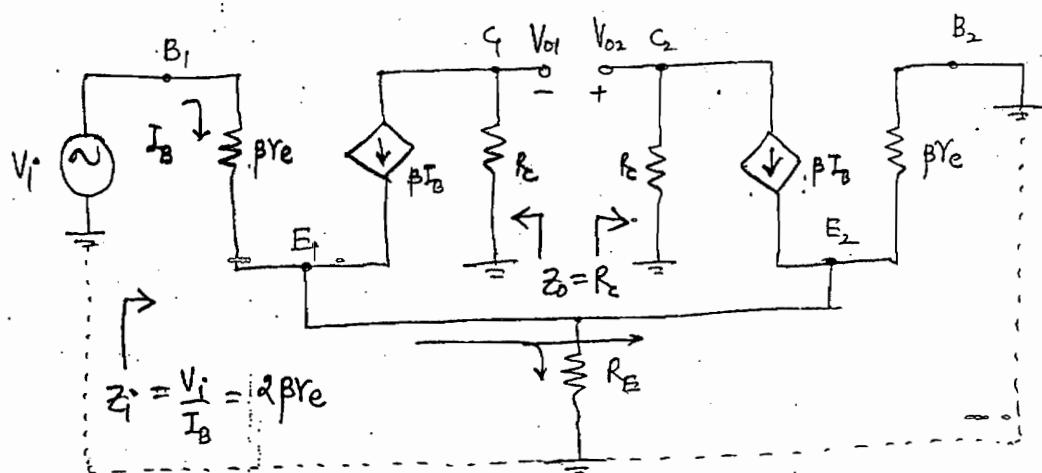
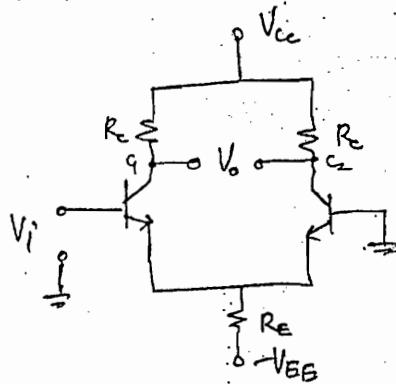
Q2. Calculate the voltage gain of single I/p unbalanced O/p or dual I/p unbalanced O/p differential amplifier with $R_E = 1\text{ k}\Omega$ and $(I_E)_Q = 26\text{ mA}$.

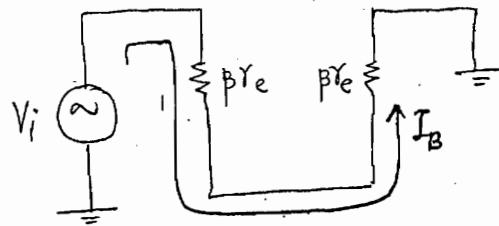
Q3. Calculate the O/p impedance of a differential amplifier ckt with $\beta = 50$ and $(I_E)_Q = 26\text{ mA}$.

Q4. Calculate the O/p impedance of a diff. amplifier ckt with $R_E = 1\text{ k}\Omega$.

Q5. Design the operating point of a diff. amplifier ckt

AC analysis:





$$Z_i = 2\beta Y_E \\ \because R_E \gg \beta Y_E$$

$$V_{O1} = -\beta I_B R_C$$

$$V_{G1} = -\beta \frac{V_i}{2\beta Y_E} R_E$$

$$\frac{V_{O1}}{V_i} = -\frac{R_C}{2Y_E} \rightarrow \text{unbalanced gain}$$

$$\text{Similarly, } \frac{V_{O2}}{V_i} = +\frac{R_C}{2Y_E} \rightarrow \text{unbalanced gain}$$

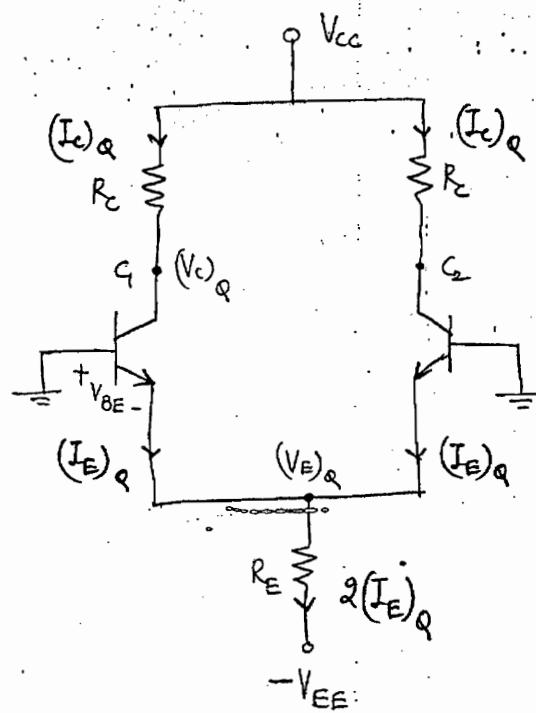
$$\text{And, } A_V = A_{V1} - A_{V2}$$

$$A_V = -\frac{R_C}{2Y_E} - \frac{R_C}{2Y_E}$$

$$A_V = -\frac{R_C}{Y_E} \rightarrow \text{Balanced gain}$$

Sol(5).

DC Analysis: operating point



$$V_{EE} - V_{BE} - 2(I_E)_Q \cdot R_E = 0$$

$$(I_E)_Q = \frac{V_{EE} - V_{BE}}{2R_E}$$

$$(V_{CE})_Q = (V_C)_Q - (V_E)_Q \\ = V_{CC} - (I_C)_Q \cdot R_C + V_{BE}$$

$$(V_{CE})_Q = V_{CC} + V_{BE} - (I_C)_Q \cdot R_C$$

Sol(1). Balanced o/p:

$$|A_d| = \left| \frac{R_e}{r_e} \right| = \frac{1K}{1\Omega} = 1000$$

$$r_e = \frac{V_T}{(I_E)_Q} = \frac{26}{26} = 1\Omega$$

Sol(2). Unbalanced o/p:

$$|A_d| = \left| \frac{R_e}{r_e} \right| = \frac{1K}{2} = 500$$

$$\text{Sol(3). } Z_i = \alpha_B r_e \\ = 2 \times 50 \times 1 = 100 \Omega$$

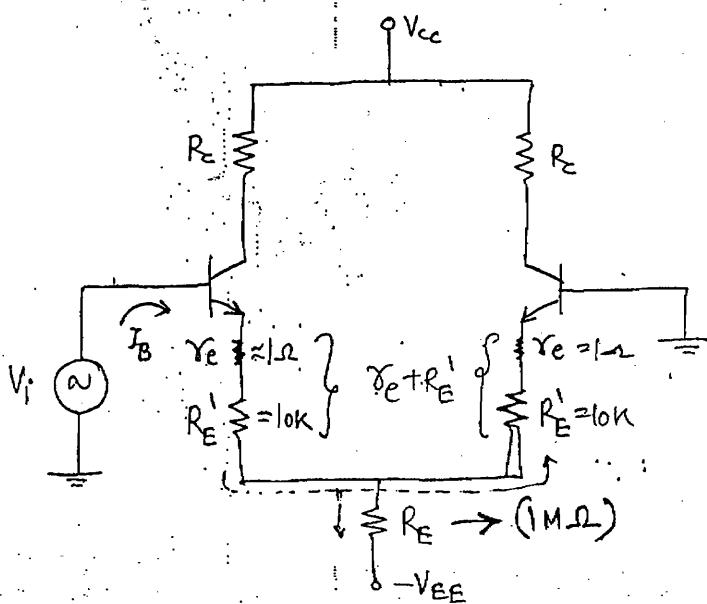
$$\text{Sol(4). } Z_o = R_c = 1K\Omega$$

Conclusion:

- (1). The gain of amp. is unstable.
 (2). The i/p impedance is low.

To solve this problem, we use feedback.

Swamping resistor technique:



$$A_d = \frac{R_c}{r_e}$$

↓
 Swamping Resistor technique

$$A_d = \frac{R_c}{r_e + R_E}$$

$$A_d \approx \frac{R_c}{R_E'} \rightarrow \text{stable}$$

$$Z_i = 2\beta r_e \downarrow$$

Swamping

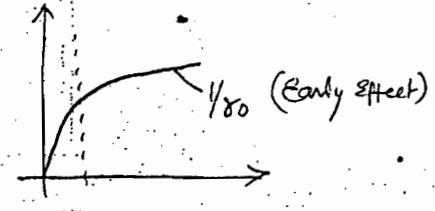
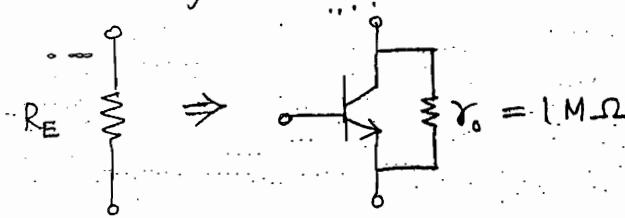
$$Z_i = 2\beta(r_e + R_E')$$

$$Z_i \approx 2\beta R_E' \quad (\text{High})$$

$$\text{If } \beta = 50, R_E' = 10\text{ k}\Omega \Rightarrow Z_i = 2 \times 50 \times 10\text{ k}\Omega = 1\text{ M}\Omega$$

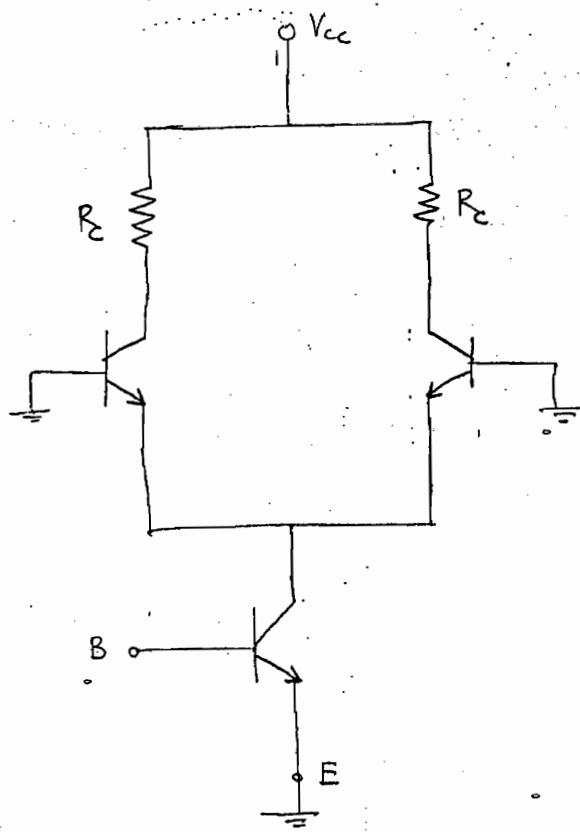
* The only problem is with this, we have to keep $R_E \rightarrow 1\text{ M}\Omega$
i.e. to block I_B , $R_E >> (r_e + R_E')$.

In place of R_E , put transistor to solve the above problem.

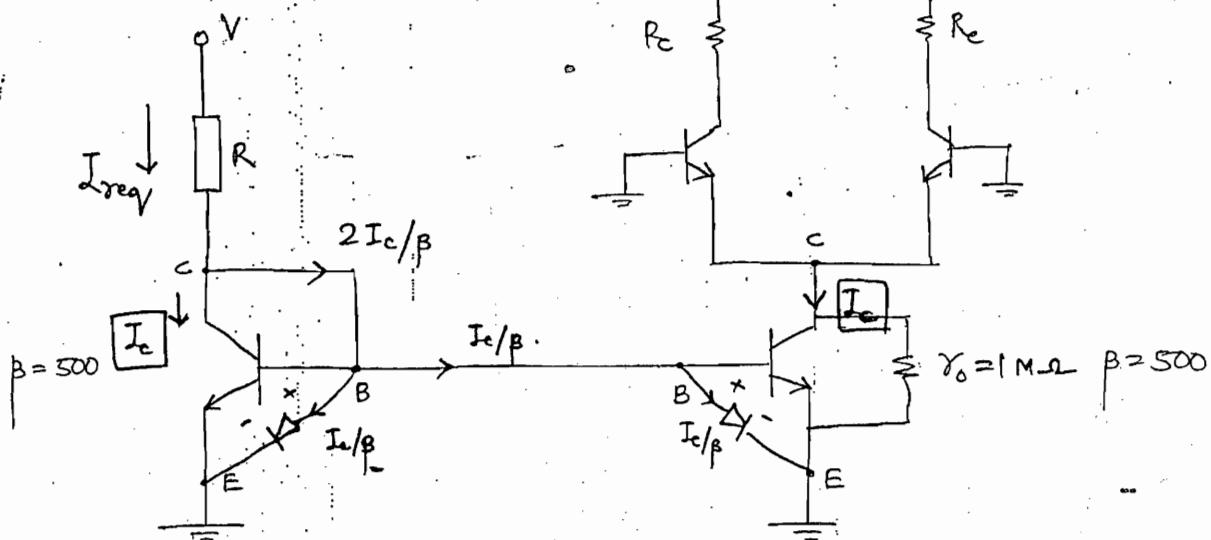


Integrated ckt Biasing

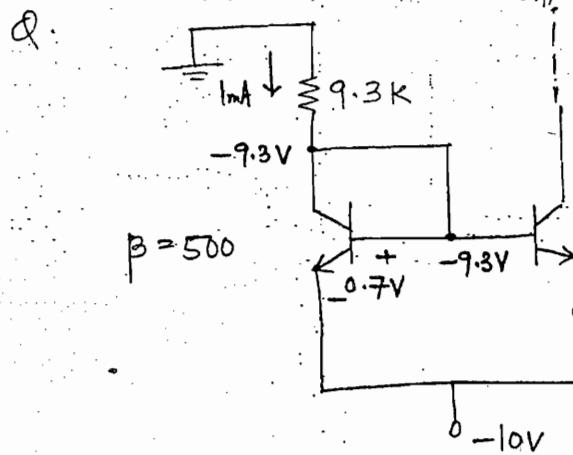
current mirror biasing:



(1). Simple current mirror:

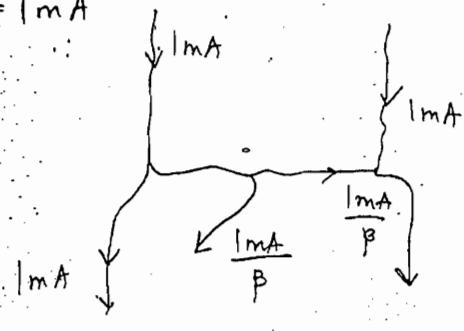


$I_B = I_c / \beta$ for reduce I_B . $\Rightarrow \beta \uparrow$ (high value)

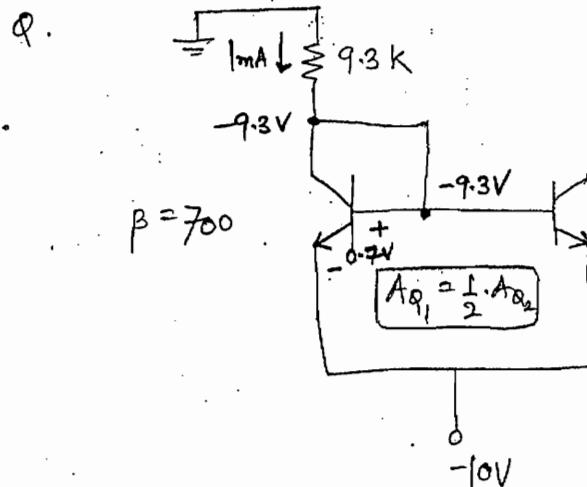


Dif. Omp.

$$\downarrow I_o = ? \quad I_o = 1 \text{ mA} \quad \beta = 500$$

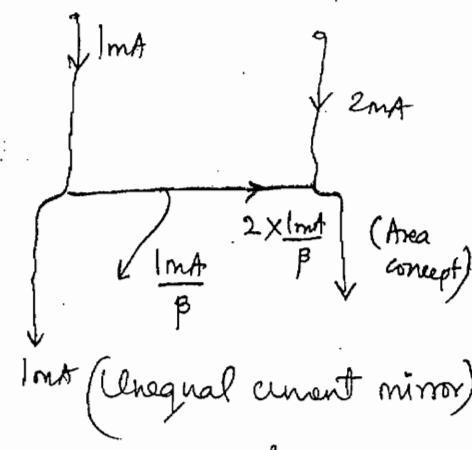


(Equal current mirror)



$$\downarrow I_o = ? \quad I_o \approx 2 \text{ mA}$$

$$\beta = 715$$



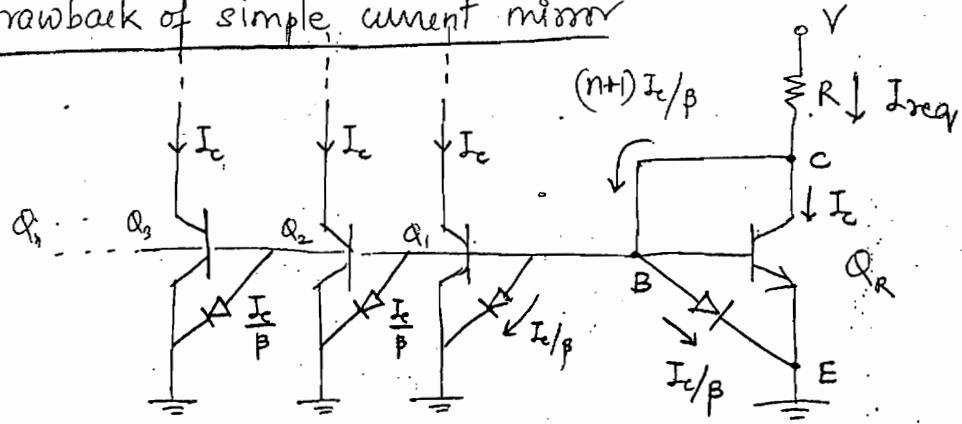
1mA (Unequal current mirror)

* β are equal (theoretically)

v β are slightly different (practically) \rightarrow don't bother (approx ans) ✓

Drawback of simple current mirror

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$$I_{req} = I_c + (n+1) \frac{I_c}{\beta}$$

$$I_{req} = I_c \left[1 + \frac{(n+1)}{\beta} \right]$$

e.g. if $n=99$, then $\ell \beta = 500$.

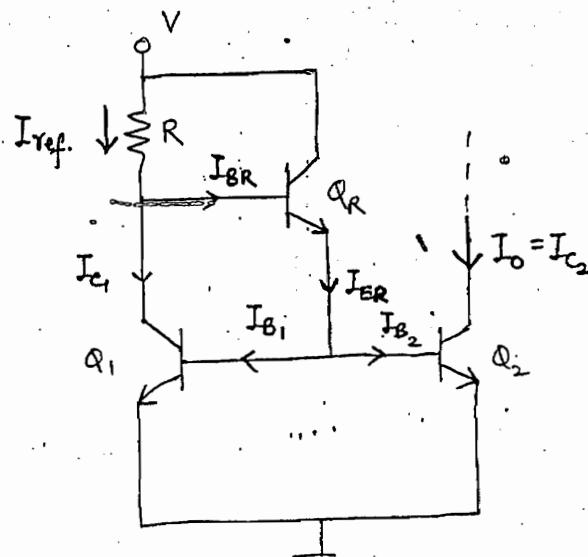
$$I_{req} = I_c \left(1 + \frac{100}{500} \right) = \frac{6}{5} I_c$$

$I_{req} \neq I_c$ (accuracy is lost)

Hence, $n \leq 10$ transistors in the biasing for better accuracy.

To improve accuracy, there is modification required.

(2). Modified current mirror :



$$\begin{aligned} I_{req} &= I_{BR} + I_C \\ &= I_C + I_{ER}/(1+\beta) \\ &= I_C + \frac{I_{B1} + I_{B2}}{(1+\beta)} \end{aligned}$$

Acc. to current mirror,

$$I_{C1} = I_{C2} \quad \& \quad I_{B1} = I_{B2}$$

$$\begin{aligned} \text{Now, } I_{req} &= I_{C2} + \frac{2 I_{B2}}{(1+\beta)} \\ &= I_{C2} + \frac{2 I_{C2}}{\beta(1+\beta)} \end{aligned}$$

$$I_{ref} = I_S + \frac{2 I_C}{\beta + \beta^2}$$

$$I_{ref} \approx I_S \left(1 + \frac{2}{\beta^2} \right)$$

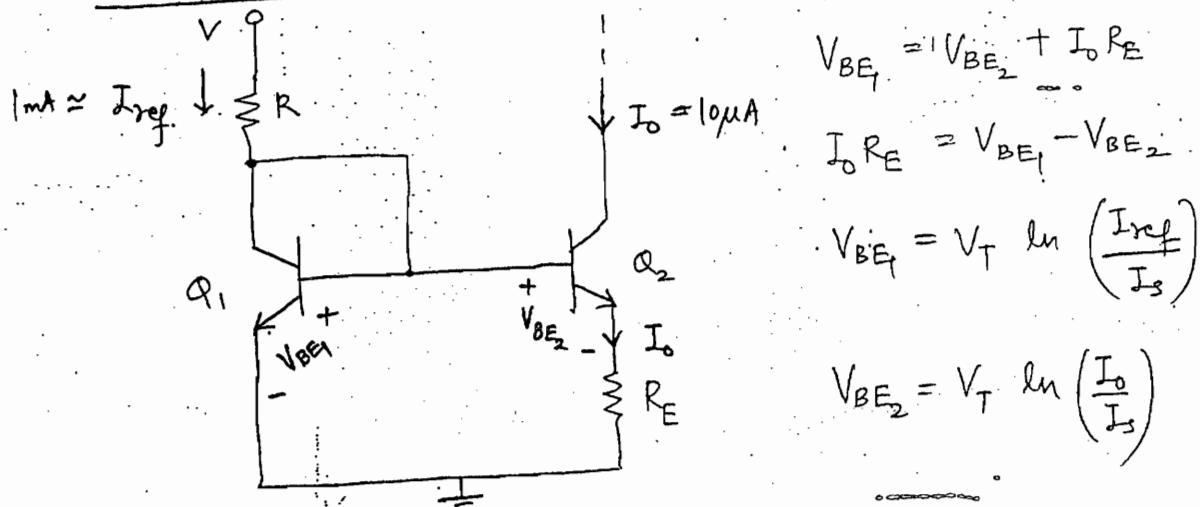
$$\therefore I_o = I_S = \frac{I_{ref}}{\left(1 + \frac{2}{\beta^2} \right)}$$

for n transistors,

$$I_G = \frac{I_{ref}}{1 + (n+1) \frac{2}{\beta^2}}$$

Here, accuracy is much improved.

(3). micro level current mirror (mA to μA mirror)



$$\therefore I_o R_E = V_T \ln \left(\frac{I_{ref}}{I_o} \right)$$

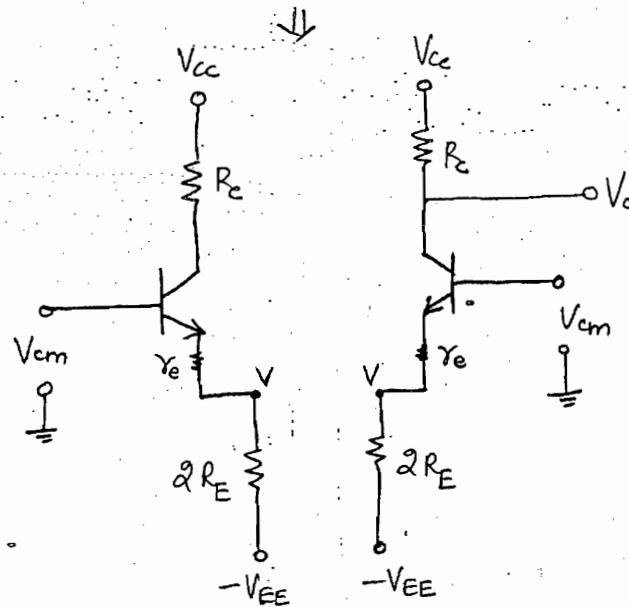
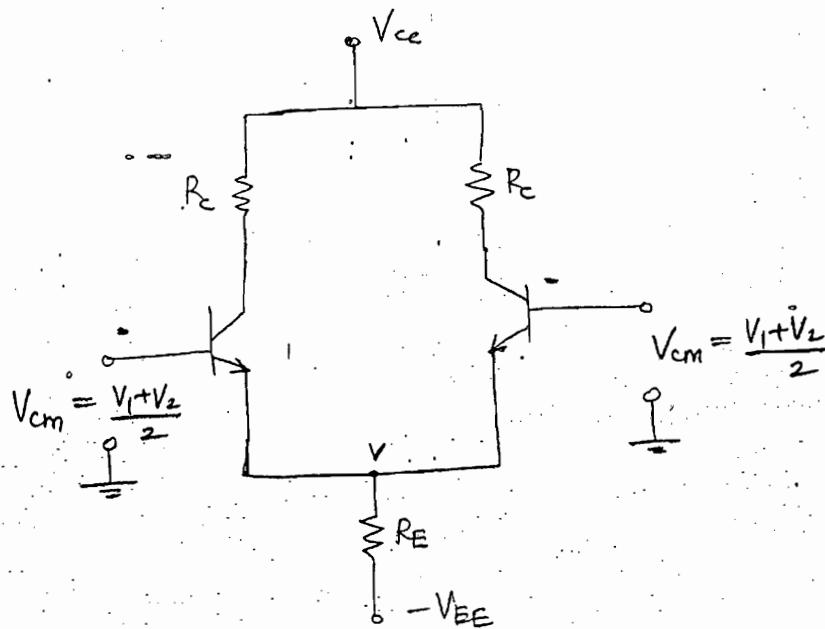
$$\text{e.g. } 10 \times 10^{-6} \cdot R_E = 26 \text{ mV} \cdot \ln \left(\frac{1 \text{ mA}}{10 \times 10^{-6} \text{ A}} \right)$$

$$\text{: (designing } R_E \text{)} \rightarrow R_E = \frac{26 \times 10^{-3}}{10^{-7}} \ln \left(\frac{10^{-3}}{10^{-6}} \right) = 11 \text{ k}\Omega$$

\Downarrow
Widlar Current source
 \downarrow
Scientist

CMRR :

$$\text{CMRR} = \frac{A_d}{A_c} \Rightarrow (\text{CMRR})_{\text{dB}} = 20 \log \left(\frac{A_d}{A_c} \right)$$

Common mode gain (A_c) :

$$A_{cm} = \frac{V_o}{V_{cm}} = \frac{-R_e}{r_e + 2R_E}$$

$\Rightarrow 2R_E \gg r_e$

$$\therefore A_{cm} = \frac{-R_e}{2R_E} \rightarrow \text{Unbalanced}$$

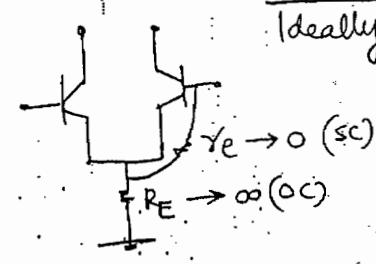
$$A_d = \frac{-R_c}{2r_e} \rightarrow \text{Unbalanced}$$

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

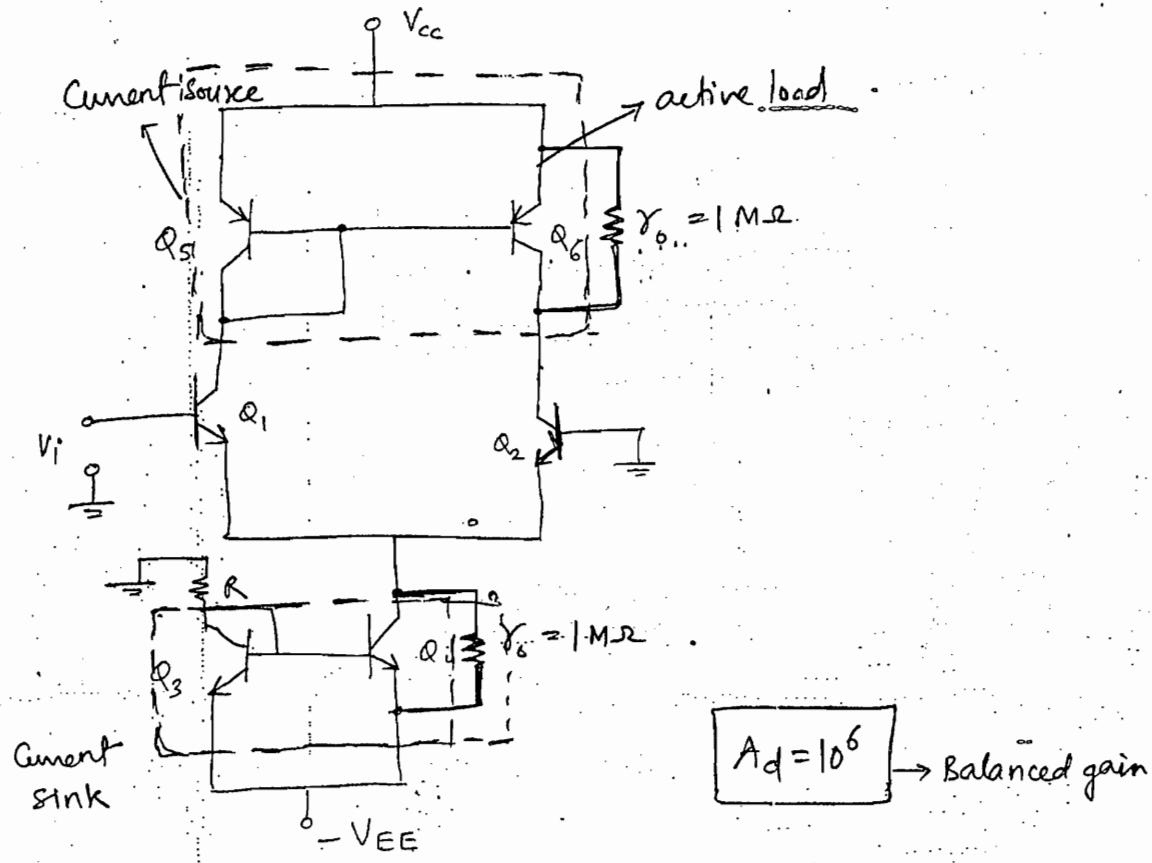
$$\text{CMRR} = \frac{-R_c / 2r_e}{-R_c / 2R_E}$$

$$\boxed{\text{CMRR} = \frac{R_E / r_e}{1}} = g_m R_E$$

For A_c -
Ideally $(\text{CMRR} = \infty)$



Integrated IC differential amplifier:



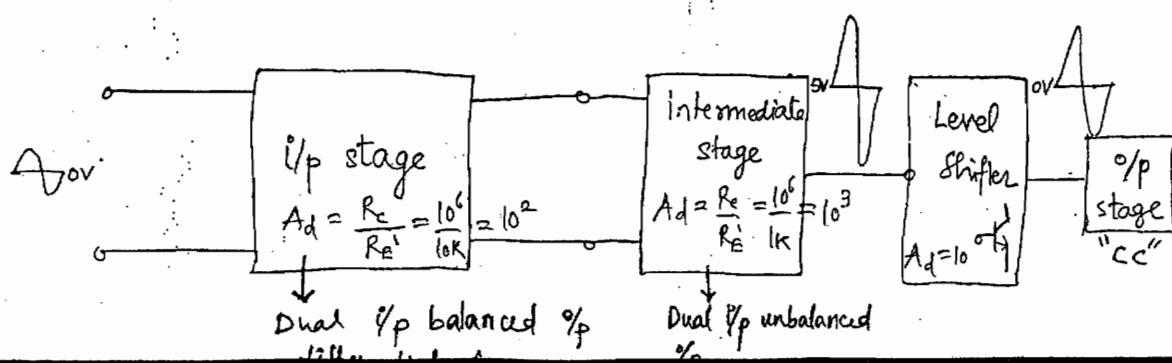
$$A_d = \frac{R_c}{2Y_e} = \frac{Y_0}{2Y_e} = \frac{1 \text{ M}\Omega}{2 \times 1 \Omega} = 0.5 \times 10^6$$

$$A_C = \frac{R_c}{2R_E} = \frac{Y_0}{2Y_0} = 0.5$$

$$\text{CMRR} = \frac{A_d}{A_C} = \frac{0.5 \times 10^6}{0.5} = 10^6$$

Operational amplifier:

Block diagram of Op-Amp:



Applications of Op-Amp :

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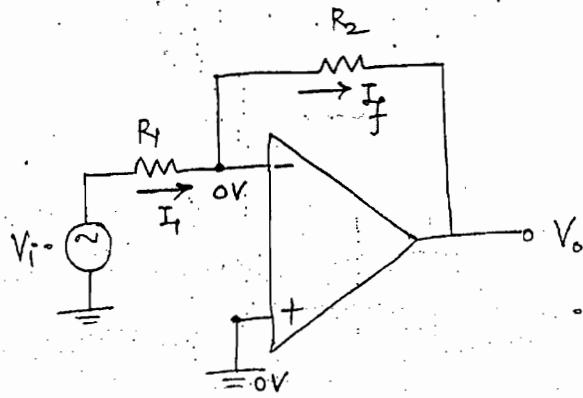
- (1). Inverting amplifier
- (2). Non-inverting amplifier
- (3). Phase shifter
- (4). Voltage followers
- (5). Differential amplifiers
- (6). Subtractor
- (7). Adder (inverting)
- (8). Non-inverting adder
- (9). Current to voltage converter
- (10). Voltage to current converter

- * (11). Voltage limiters
- (12). Logarithmic amplifiers
- (13). Antilog amplifiers
- * (14). precision rectifiers
- (15). Instrumentational amplifiers
- (16). Modulators (Analog multipliers)
- (17). Demodulators (Analog dividers)
- * (18). filter designs (active filters)
 - LPF
 - HPF
 - BPF
 - BRF
 - All pass filter
- (19). Integrators
- (20). Differentiators.

Non-linear applications :

- (1). Generating a square wave
 - Comparators (ZCD)
 - Schmitt trigger.
 - Astable multivibrator.
- (2). Mono stable multivibrator
- (3). 555 timer
- (4). oscillators
 - RC oscillators
 - RC phase shift oscillator
 - Wein bridge oscillator
 - LC oscillators
 - Hartley
 - Colpitts

(1) Inverting amplifier:

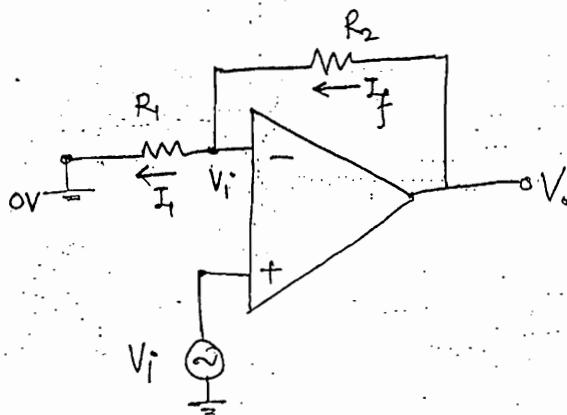


$$I_f = I_f$$

$$\therefore \frac{V_i - 0}{R_1} = \frac{0 - V_o}{R_2}$$

$$\frac{V_o}{V_i} = -\frac{R_2}{R_1}$$

(2) Non-inverting amplifier:

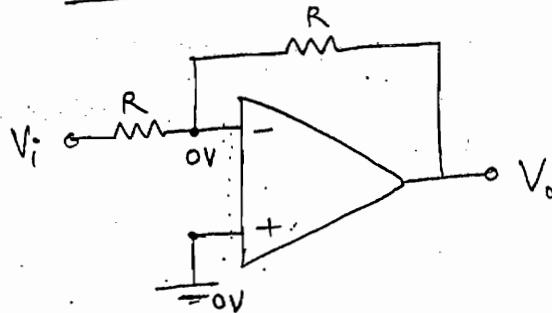


$$I_f = I_f$$

$$\frac{V_i - 0}{R_1} = \frac{V_o - V_i}{R_2}$$

$$\frac{V_o}{V_i} = 1 + \frac{R_2}{R_1}$$

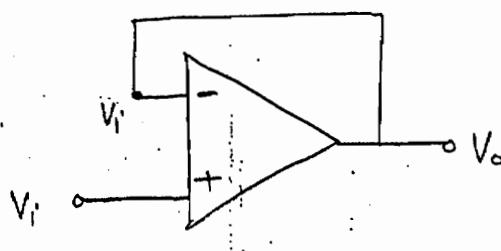
(3) Phase shifter:



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

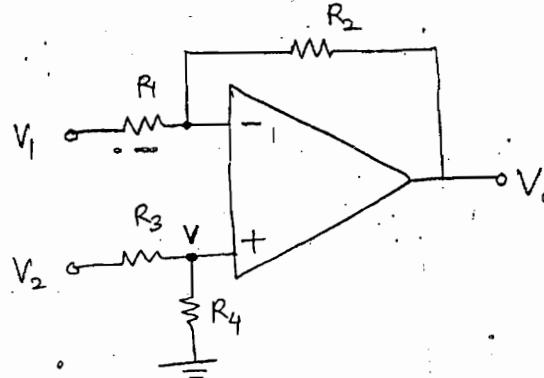
$$V_o = -V_i$$

(4) Voltage follower:



$$V_o = V_i$$

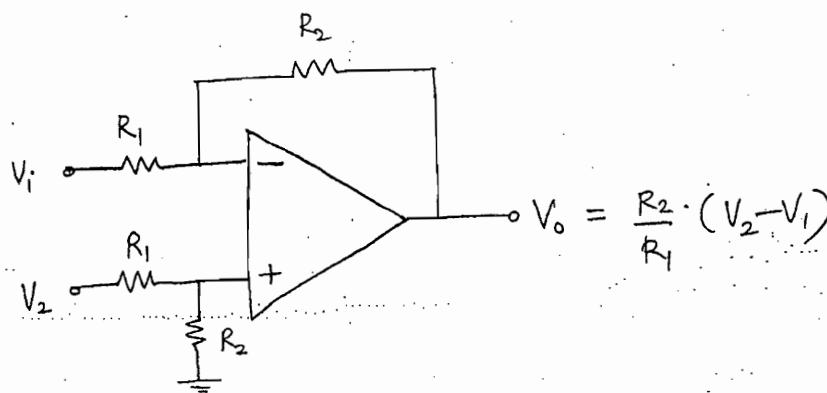
(5). Differential amplifier:



By superposition,

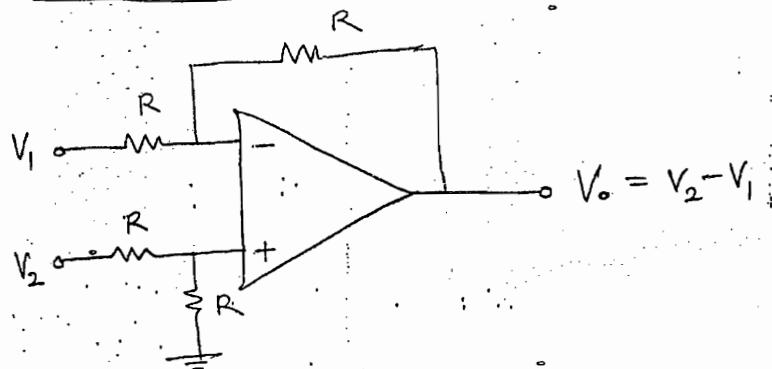
$$V_0 = -\frac{R_2}{R_1} \cdot V_1 + \left(1 + \frac{R_2}{R_1}\right) V$$

$$\text{where, } V = \frac{V_2 \cdot R_4}{(R_3 + R_4)}$$



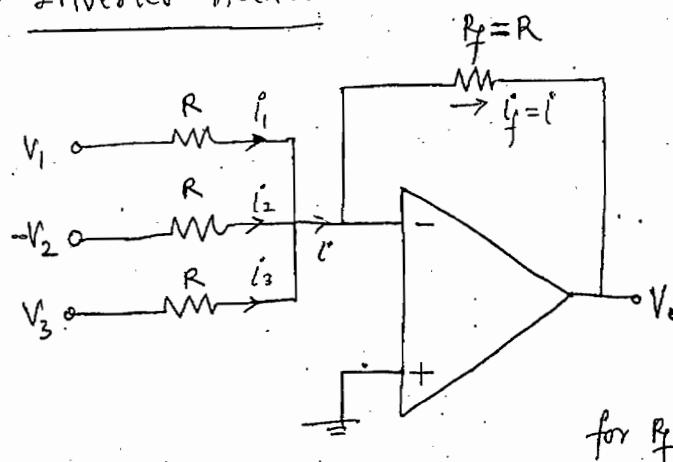
$$V_0 = \frac{R_2}{R_1} \cdot (V_2 - V_1)$$

(6). Subtractor :



$$V_0 = V_2 - V_1$$

(7). Inverter Adder :



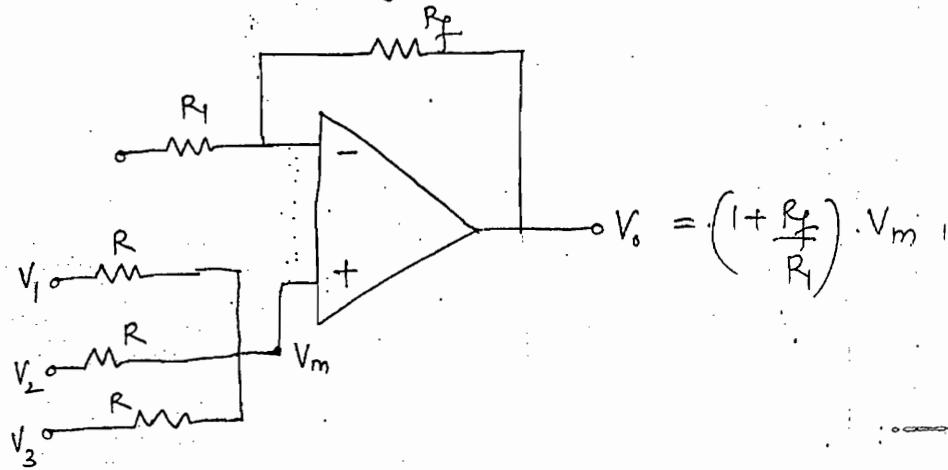
$$i_f = i$$

$$i_1 + i_2 + i_3 = \frac{0 - V_0}{R_f}$$

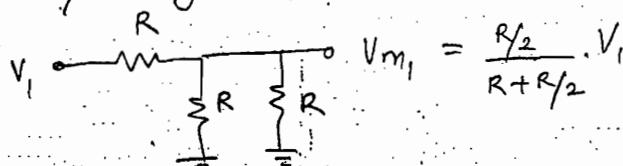
$$\frac{V_1 + V_2 + V_3}{R} = -\frac{V_0}{R_f}$$

$$V_0 = -\frac{R_f}{R} (V_1 + V_2 + V_3)$$

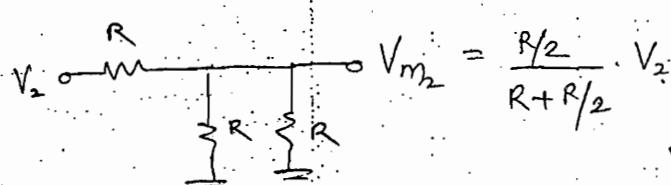
$$\text{for } R_f = R; \quad V_0 = -(V_1 + V_2 + V_3)$$

(B) Non-inverting Adder:

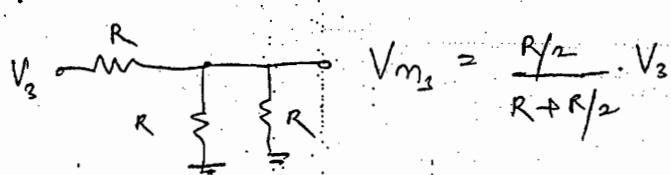
By using superposition,



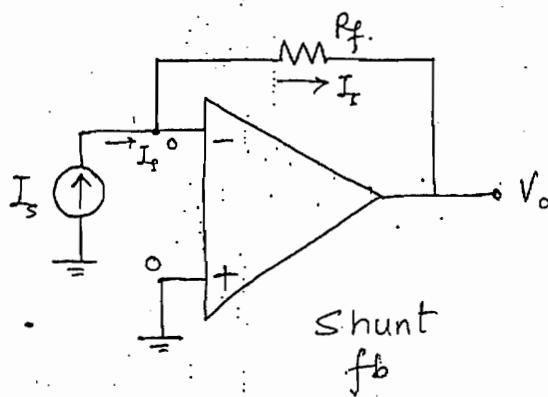
$$\begin{aligned} \therefore V_m &= V_{m1} + V_{m2} + V_{m3} \\ &= \frac{R/2}{R+R/2} \cdot (V_1 + V_2 + V_3) \end{aligned}$$



$$\begin{aligned} &= \frac{R}{3R} \cdot (V_1 + V_2 + V_3) \\ &= \frac{(V_1 + V_2 + V_3)}{3} \end{aligned}$$

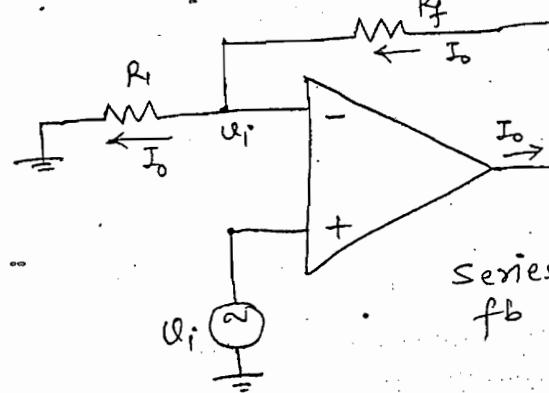


$$V_o = \left(1 + \frac{R_f}{R}\right) \left(\frac{V_1 + V_2 + V_3}{3}\right)$$

(9) Current to voltage converter: "CCVS"

$$0 - V_o = I_s \cdot R_f$$

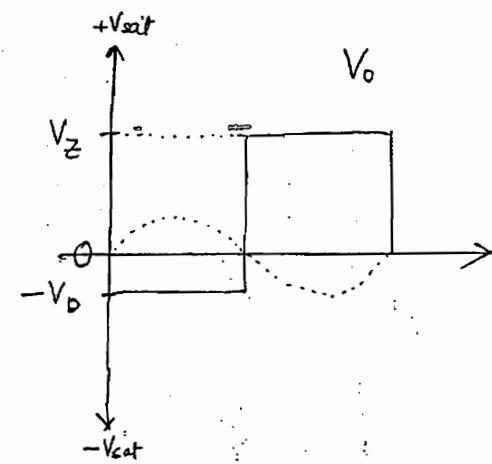
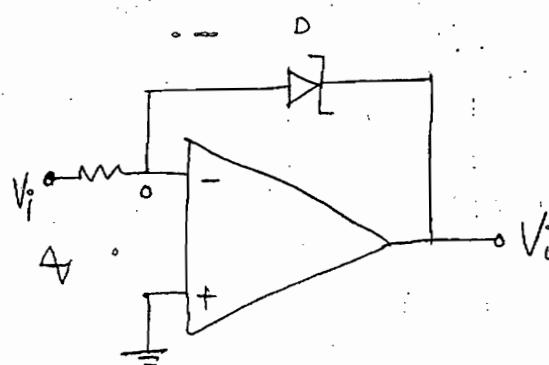
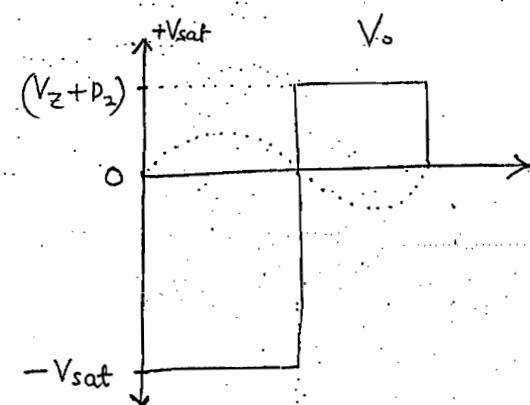
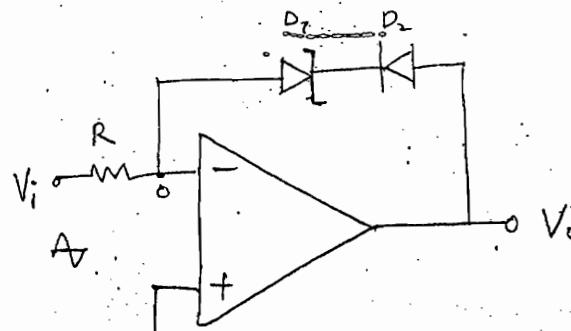
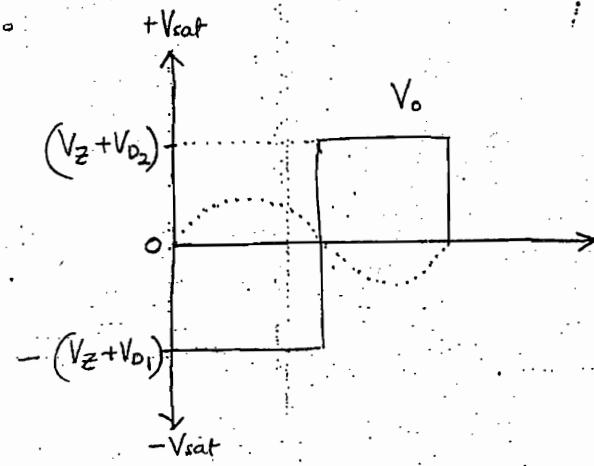
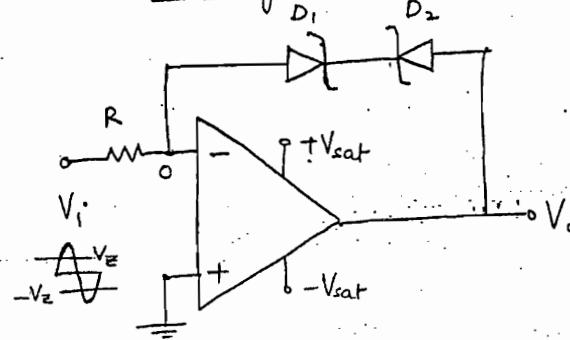
$$V_o = -I_s R_f$$

(10). Voltage to current converter: "VCCS"

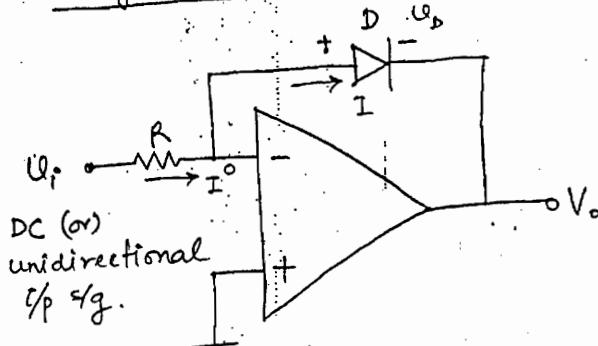
$$V_i = I_o R_f$$

$$I_o = \frac{V_i}{R_f}$$

Series
fb

(11). Voltage limiter

(12). Logarithmic amplifier: (Data compression)



$$0 - V_o = U_D$$

$$V_o = -U_D$$

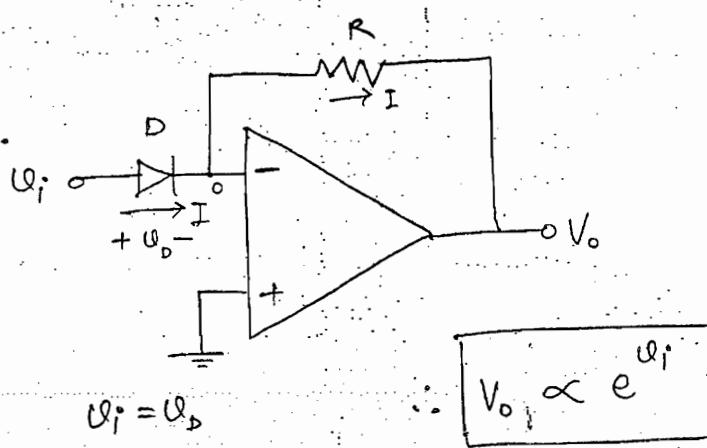
$$I = I_0 e^{U_D/V_T}$$

$$U_D = V_T \cdot \ln \left(\frac{I}{I_0} \right)$$

$$V_o = -V_T \ln \left(\frac{U_i}{I_0 R} \right)$$

$$\therefore V_o \propto \ln U_i$$

(13). Antilog amplifier: (Data expansion)



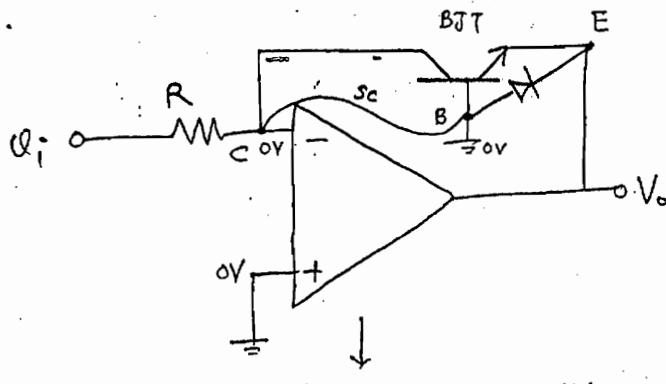
$$I = I_0 e^{U_D/V_T}$$

$$\frac{0 - V_o}{R} = I_0 e^{U_D/V_T}$$

$$V_o = -I_0 R \cdot e^{U_D/V_T}$$

$$\therefore V_o \propto e^{U_i}$$

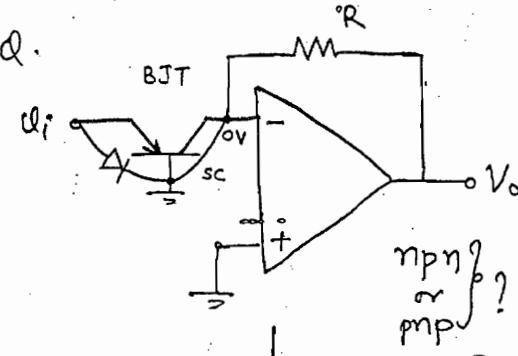
Q.



Logarithmic amplifier

(Replacement of diode with BJT)

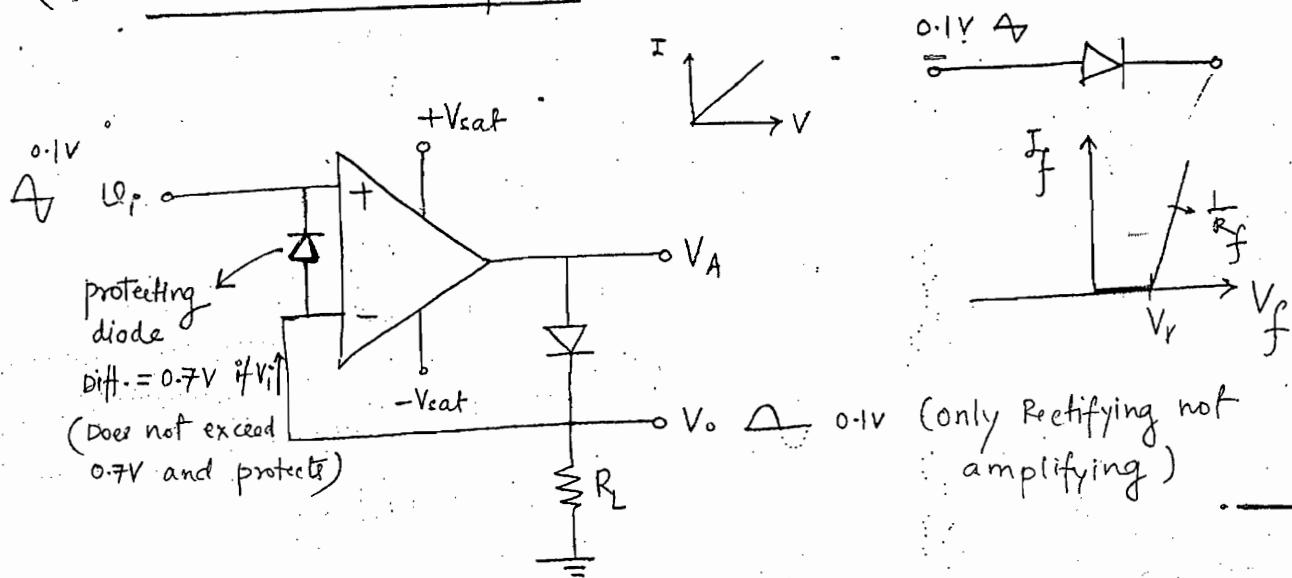
Q.



Antilog amplifier

(Replacement of diode
with BJT)

BJT must be "pnp".

(14) Precision rectifiers:+ve Half cycle (CL)

$$V_A = +V_{sat}$$

 $D \rightarrow \text{ON}$

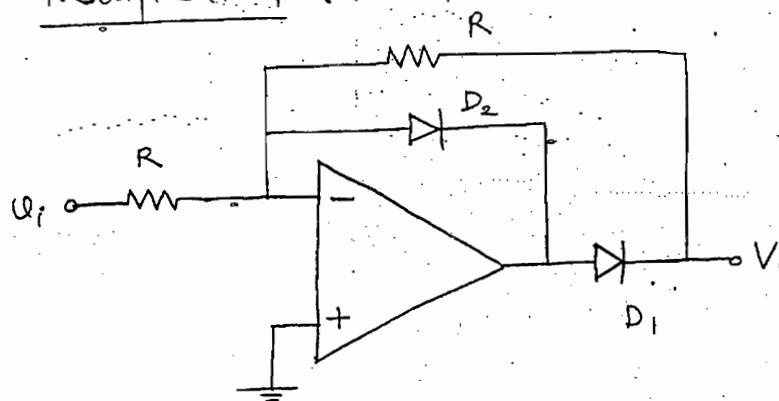
$$V_o = V_i \text{ (Linear)}$$

-ve Half cycle (OL)

$$V_A = -V_{sat}$$

 $D \rightarrow \text{OFF}$

$$V_o = 0 \text{ (Non-linear)}$$

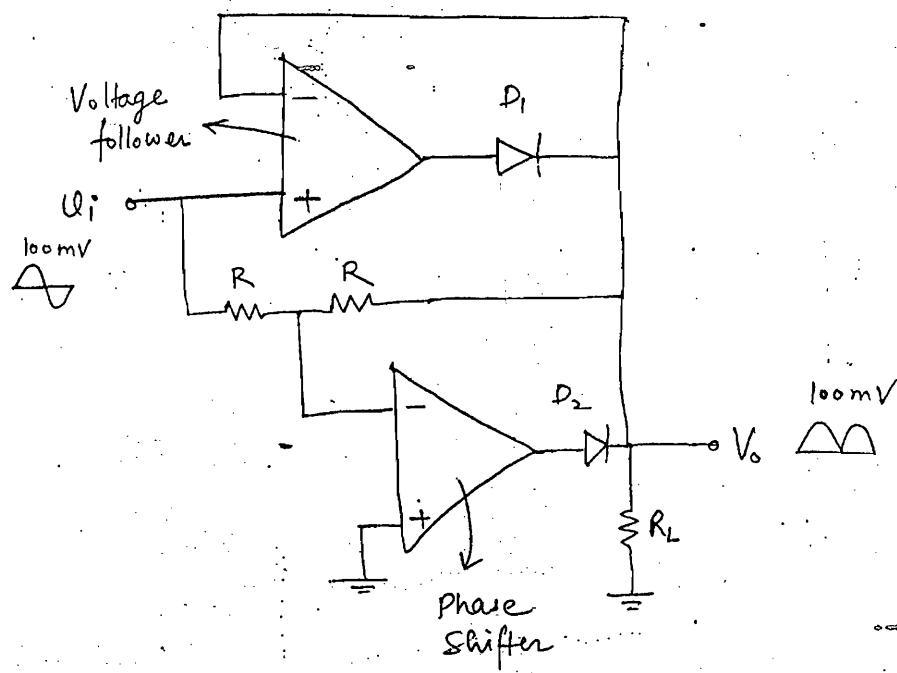
Modification+ve Half cycle $D_2 \rightarrow \text{ON}, D_1 \rightarrow \text{OFF}$

$$V_o = 0$$

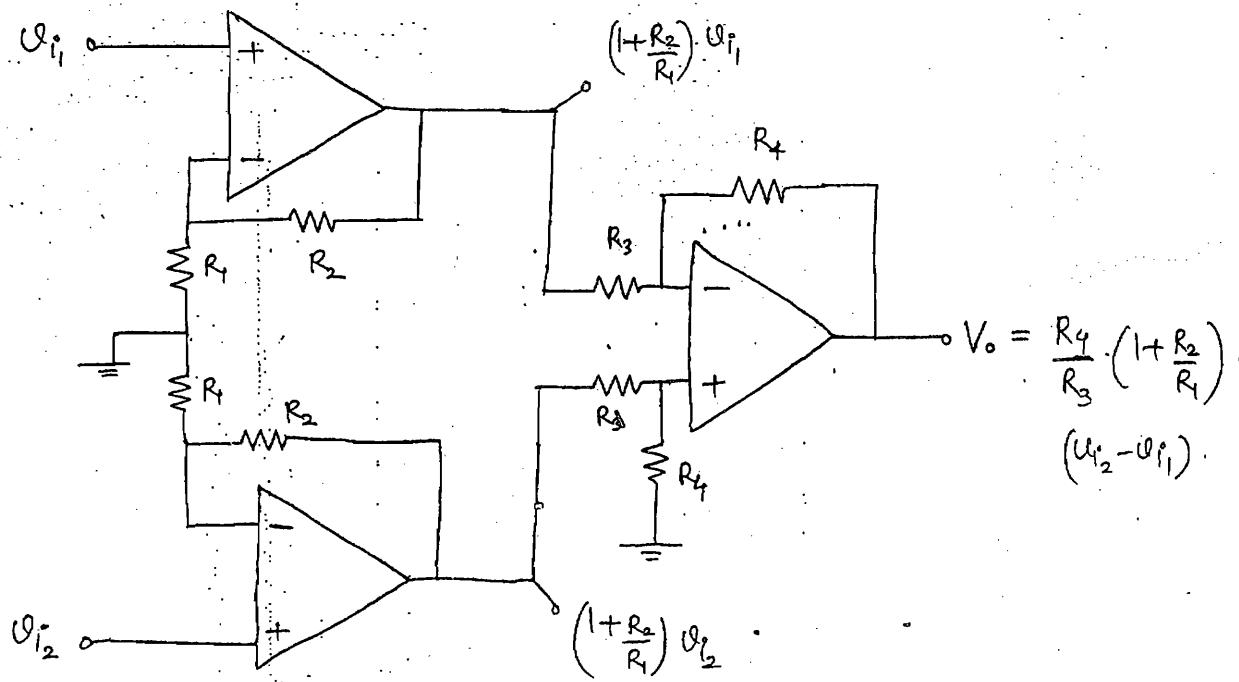
-ve Half cycle $D_2 \rightarrow \text{OFF}, D_1 \rightarrow \text{ON}$

$$V_o = -V_i \text{ (Phase shifter)}$$

Full wave rectifier:

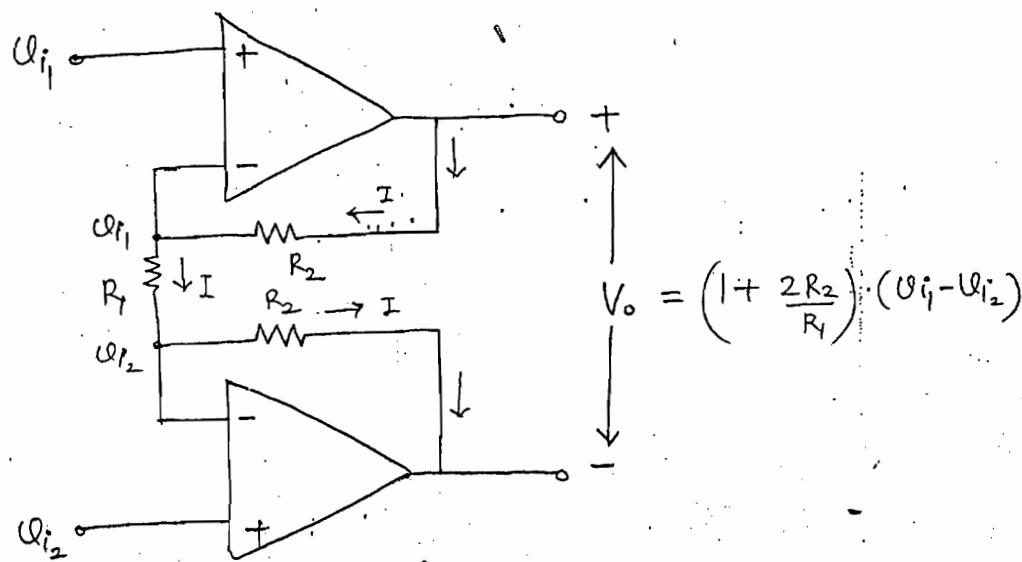


(15) Instrumentational amplifier:



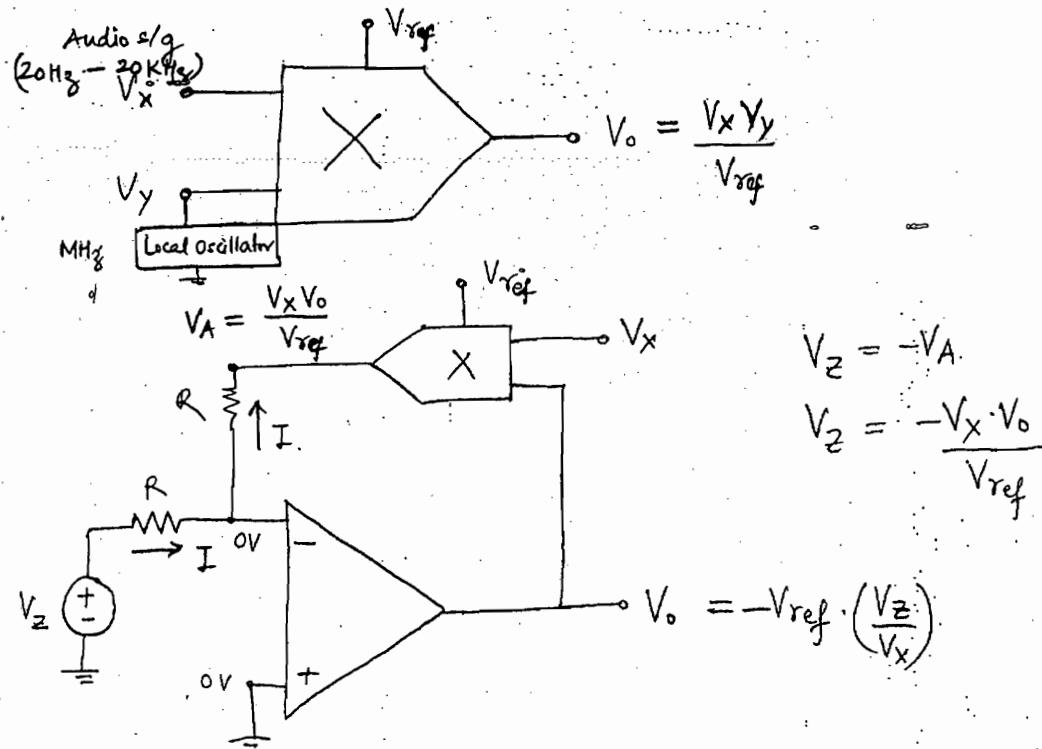
$$A_d = \frac{V_o}{U_d} = \frac{R_4}{R_3} \cdot \left(1 + \frac{R_2}{R_1}\right)$$

Modified ckt : (Better)

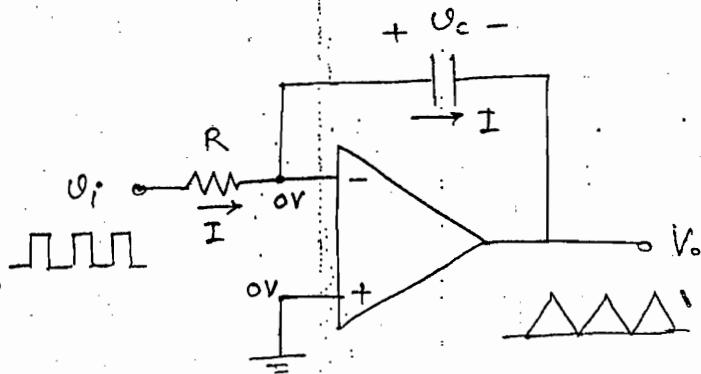


$$A_d = \frac{V_o}{V_d} = \frac{V_o}{(V_{i_1} - V_{i_2})} = \frac{I(R_1 + 2R_2)}{I(R_1)} = 1 + \frac{2R_2}{R_1} \quad (\text{High})$$

(16). Modulator : (Analog Multiplier)



(17) Demodulator : (Analog Divider)

(8). Integrator : (LPF)

$$+V_c -$$

\int

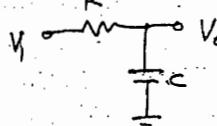
$$V_c = \frac{1}{C} \int i dt$$

$$V_o = -V_c$$

$$= -\frac{1}{C} \int i dt$$

$$V_o = -\frac{1}{RC} \int i dt$$

Transfer function:



$$TF = \frac{V_o(s)}{V_i(s)} = \frac{-1}{SCR}$$

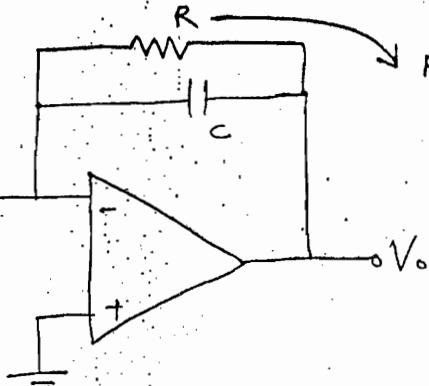
phase shift

$$\frac{1}{SCR} = \frac{1}{j\omega CR} = \frac{-j}{\omega CR}$$

↓
phase = -90°

(Phase lag design)

→ R is meant for DC conditions.

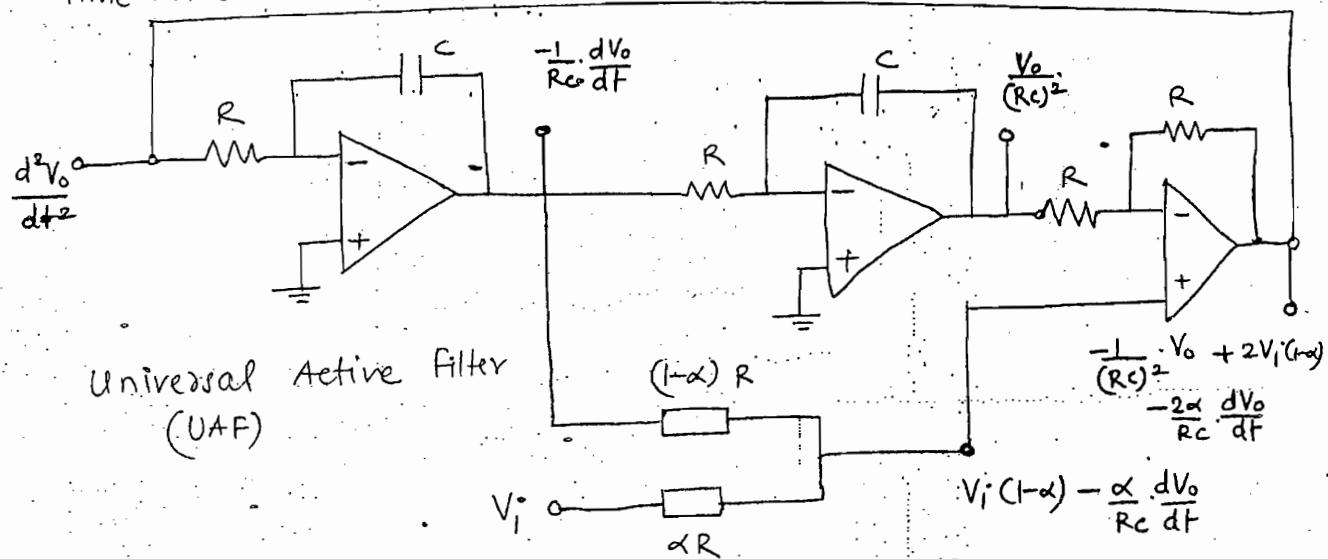
(9). Differentiator:

Simulate a 2nd order differential eqn using op-amps.

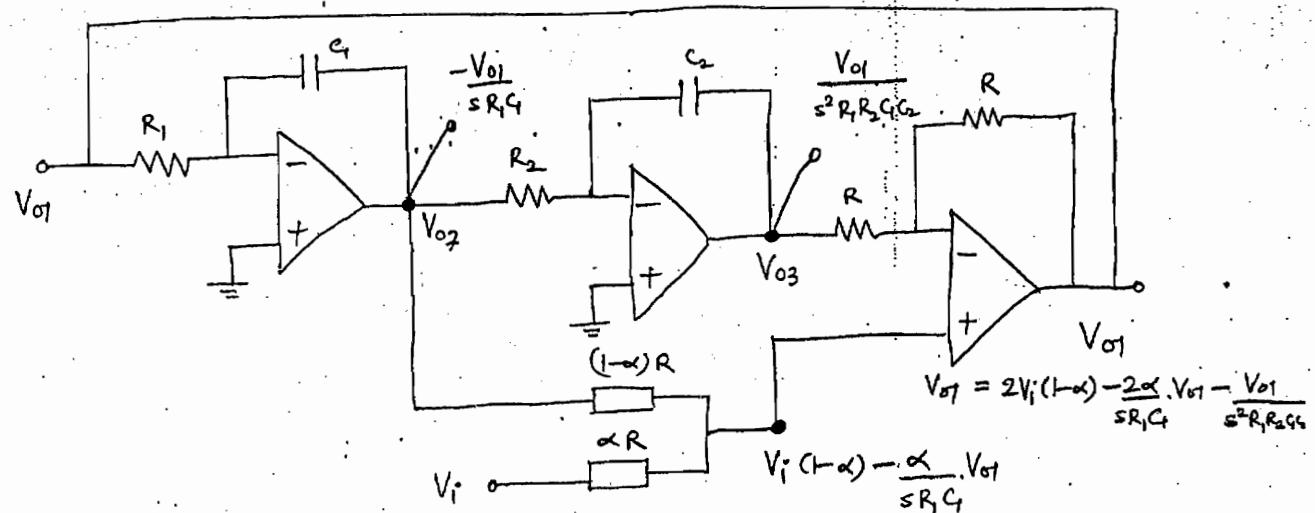
$$\frac{d^2 V_o}{dt^2} + \alpha \cdot \frac{dV_o}{dt} + \gamma V_o = V_i$$

$$\frac{d^2 V_o}{dt^2} = V_i - \alpha \cdot \frac{dV_o}{dt} - \gamma V_o$$

Time domain analysis



frequency domain analysis:



$$V_{o1} \left(1 + \frac{2\alpha}{sR_1C_1} + \frac{1}{s^2 R_1 R_2 C_1 C_2} \right) = 2V_i(1-\alpha)$$

$$\frac{V_{o1}}{V_i} = \frac{2(1-\alpha) \cdot s^2 (R_1 R_2 C_1 C_2)}{s^2 R_1 R_2 C_1 C_2 + 2\alpha s R_2 C_2 + 1} \quad (\text{HPF})$$

$$\frac{V_{02}}{V_i} = \frac{-2(1-\alpha) \cdot s R_2 C_2}{s^2 R_1 R_2 G C_2 + 2\alpha s R_2 C_2 + 1} \quad (\text{BPF})$$

$$\frac{V_{03}}{V_i} = \frac{2(1-\alpha)}{s^2 R_1 R_2 G C_2 + 2\alpha s R_2 C_2 + 1} \quad (\text{LPF})$$

LP + HP \Rightarrow BRF (Notch filter)

$$\frac{V_{04}}{V_i} = \frac{2(1-\alpha) (s^2 R_1 R_2 G C_2 + 1)}{s^2 R_1 R_2 G C_2 + 2\alpha s R_2 C_2 + 1}$$

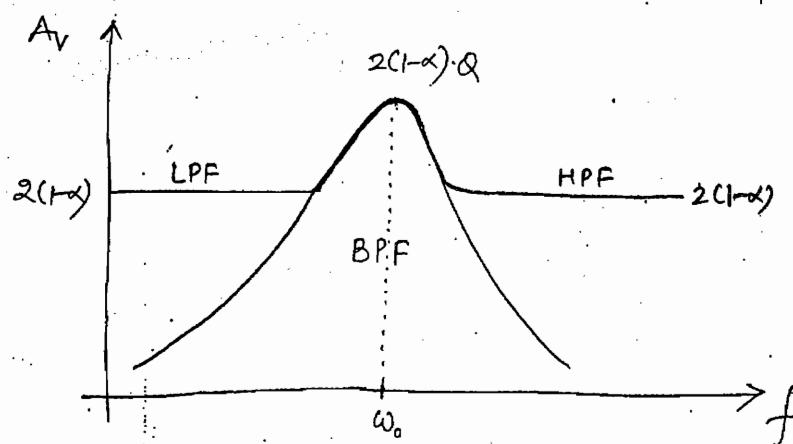
Band pass:

$$\left| \frac{V_{02}}{V_i} \right| = \frac{2(1-\alpha) \cdot s R_2 C_2}{s^2 R_1 R_2 G C_2 + 2\alpha s R_2 C_2 + 1}$$

Resonant frequency, $\omega_0 = \frac{1}{\sqrt{R_1 R_2 G C_2}}$

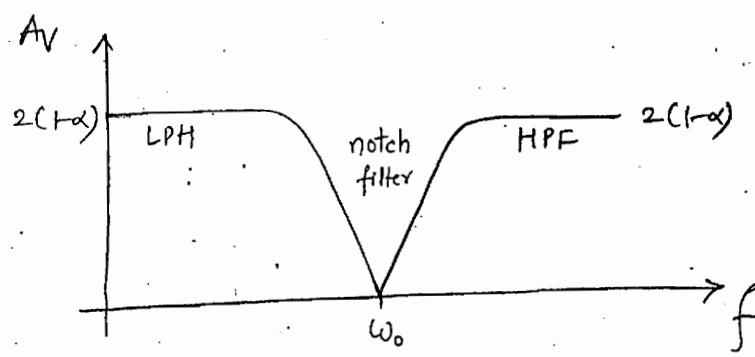
At ω_0 , gain is highest,

$$\left| \frac{V_{02}}{V_i} \right| = \frac{2(1-\alpha)}{2\alpha} = 2(1-\alpha) \cdot Q$$



Band Reject filter:

$$\frac{V_{04}}{V_i} = \frac{2(1-\alpha) (s^2 R_1 R_2 G C_2 + 1)}{s^2 R_1 R_2 G C_2 + 2\alpha s R_2 C_2 + 1}$$



$$\left| \frac{V_{04}}{V_i} \right| = \frac{2(1-\alpha)}{\omega \alpha} \leq R_1 G \rightarrow \text{zero}$$

$$\therefore 2(1-\alpha) \leq R_1 G \cdot Q$$

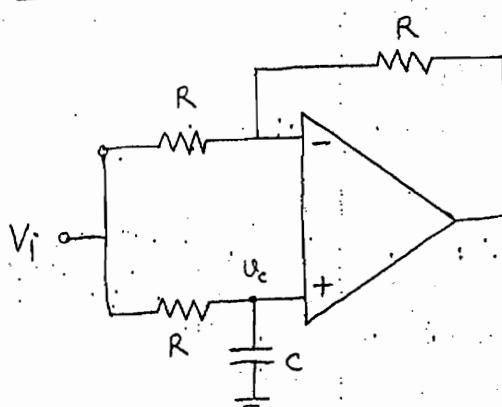
* LC filters are best due to high value Q (1000's multiple)
 RC filters have Q value in the range of 10's multiple.

for loads:

(1) R_L load $\rightarrow A_V$ may be 100's and 1000's.

(2) $\circ - \square \infty \square - \circ \rightarrow A_V = \frac{A_I \cdot Z_L}{Z_I} \text{ at } \omega_0, Z_L = \infty$
 $\therefore A_V = \infty$

All pass filter : (Phase corrector)



By superposition,

$$V_o = -\frac{R}{R} V_i + \left(1 + \frac{R}{R}\right) V_c$$

$$= -V_i + 2V_c$$

$$= -V_i + 2 \cdot V_i \left(\frac{-jX_c}{R-jX_c} \right)$$

$$= V_i \left[-1 + \frac{2}{1+j2\pi f RC} \right]$$

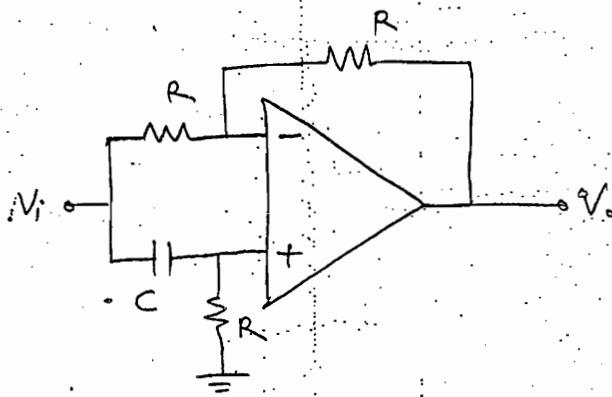
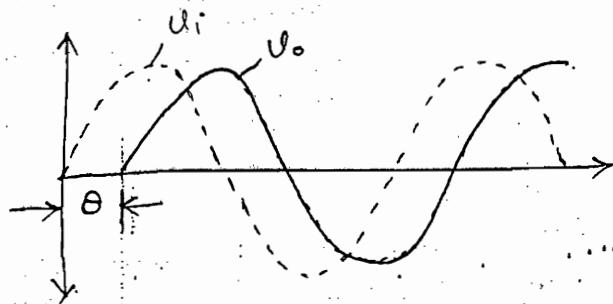
$$= V_i \cdot \frac{(1-j2\pi f RC)}{(1+j2\pi f RC)}$$

$$\frac{V_o}{V_i} = \frac{1-j2\pi fRC}{1+j2\pi fRC}$$

$$\frac{V_o}{V_i} = \left| \frac{V_o}{V_i} \right| \angle \theta = 1 \cdot \angle -2\tan^{-1}(2\pi fRC)$$

↓
phase lag.

$$\therefore \theta = -2\tan^{-1}(2\pi fRC)$$



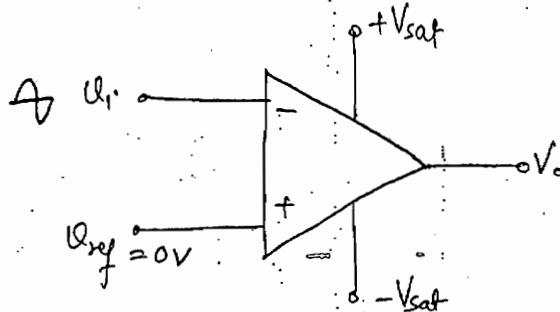
$$\theta = +2\tan^{-1}(2\pi fRC)$$

1st order APP \rightarrow can correct $\pm 180^\circ$.

2nd order APP \rightarrow can correct $\pm 360^\circ$.

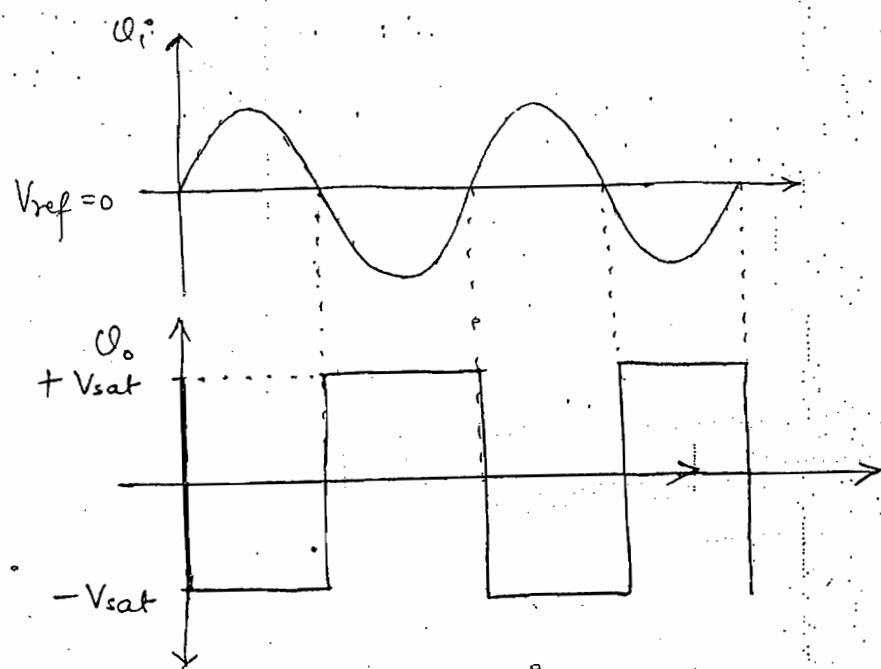
Non-linear applications:

Comparator: Zero crossing detector (ZCD)

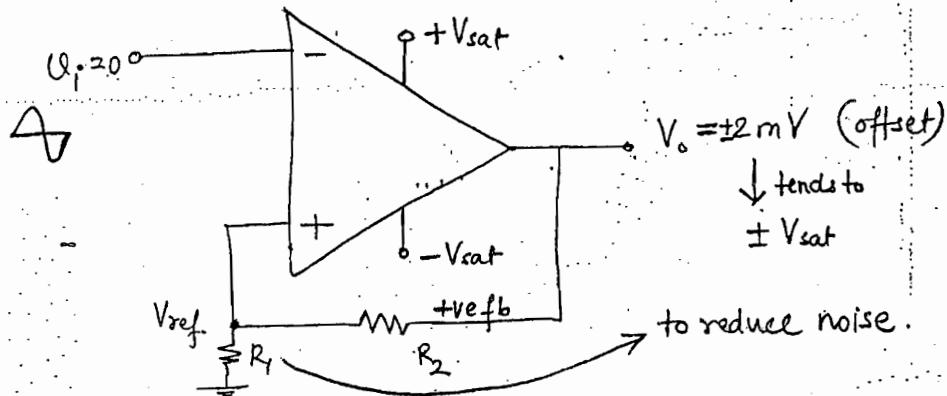


$$u_i > 0 \Rightarrow u_d = -ve \Rightarrow V_o = -V_{sat}$$

$$u_i < 0 \Rightarrow u_d = +ve \Rightarrow V_o = +V_{sat}$$



when \$V_{ref}\$ is not given : Schmitt trigger



\$V_{io} \rightarrow\$ I/p offset voltage "nV"

\$I_{io} \rightarrow\$ I/p offset current "nA"

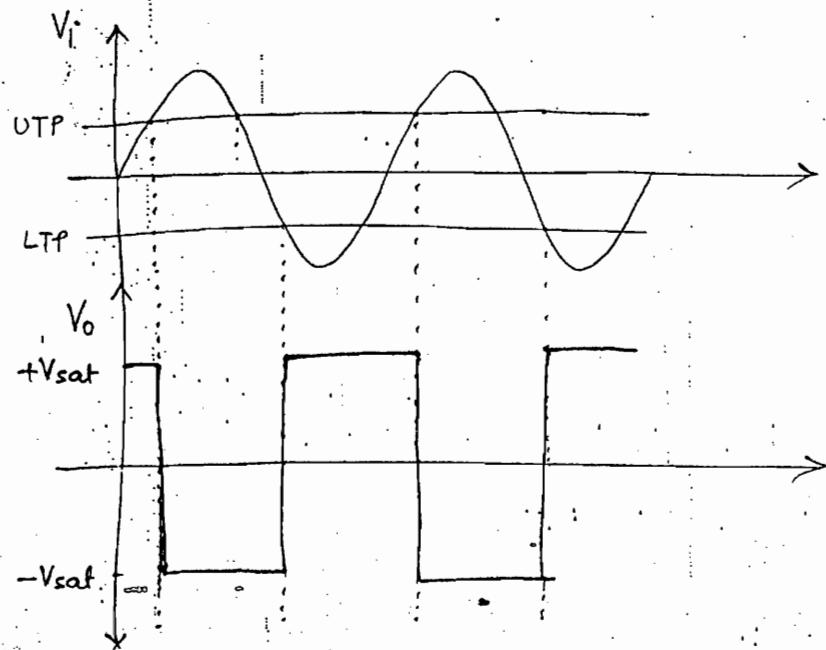
$$\text{At } V_p = 0, \quad V_{io} \times 10^6 = V_o = nV \times 10^6 = mV.$$

$$\text{when, } V_o = +V_{sat} \Rightarrow V_{ref_1} = \frac{R_1}{R_1 + R_2} V_{sat}$$

$$\text{when, } V_o = -V_{sat} \Rightarrow V_{ref_2} = \frac{R_1}{R_1 + R_2} -V_{sat}$$

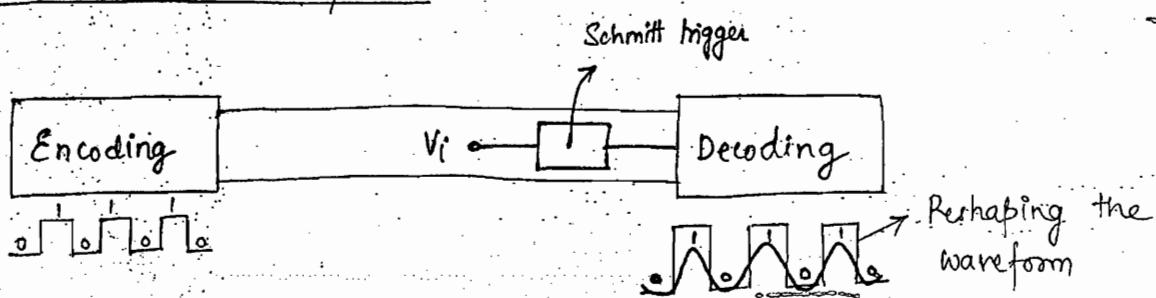
$$\text{Upper trigger point, UTP} = +V_{sat} \cdot \frac{R_1}{R_1 + R_2}$$

$$\text{Lower trigger point, LTP} = -V_{sat} \cdot \frac{R_1}{R_1 + R_2}$$

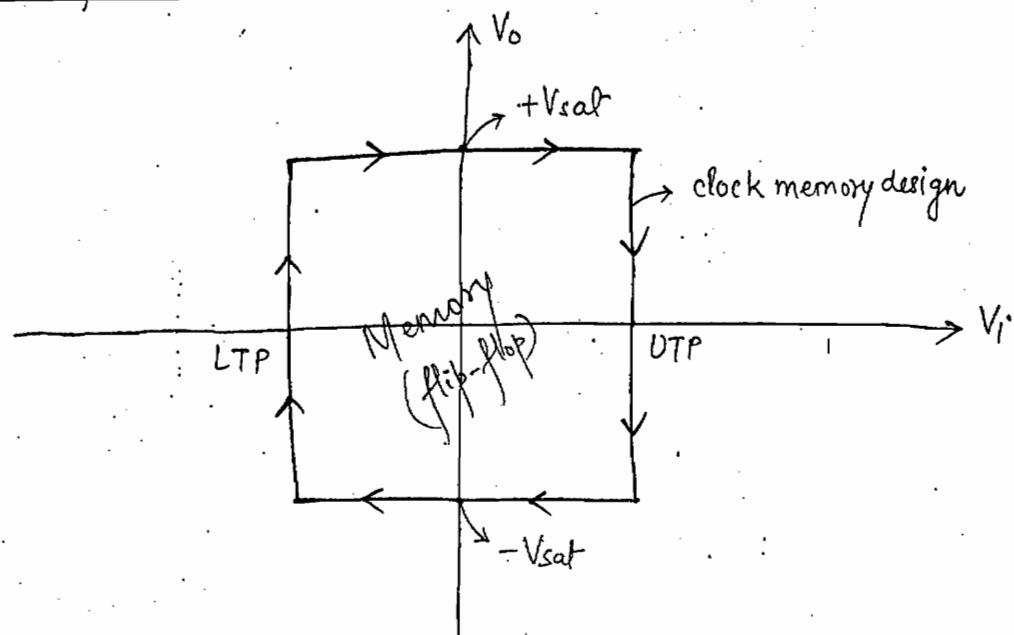


Applications of Schmitt trigger :

Communication system :

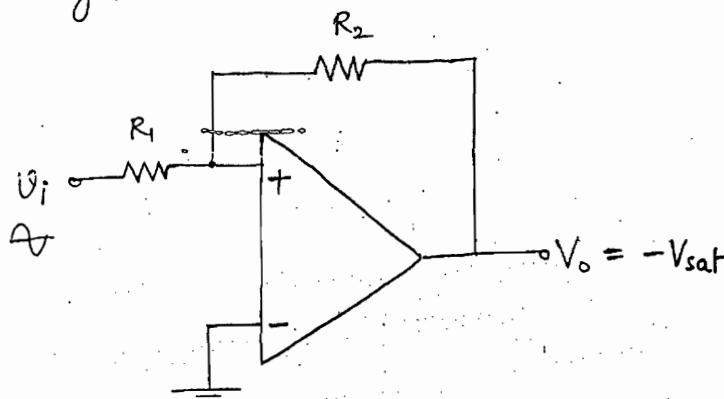


Transfer characteristics :



a. Design a schmitt trigger which gives anti clock memory design.

Sol.



When $V_o = -V_{sat}$

$$U_i \cdot \frac{R_2}{R_1 + R_2} + V_o \cdot \frac{R_1}{R_1 + R_2} = 0$$

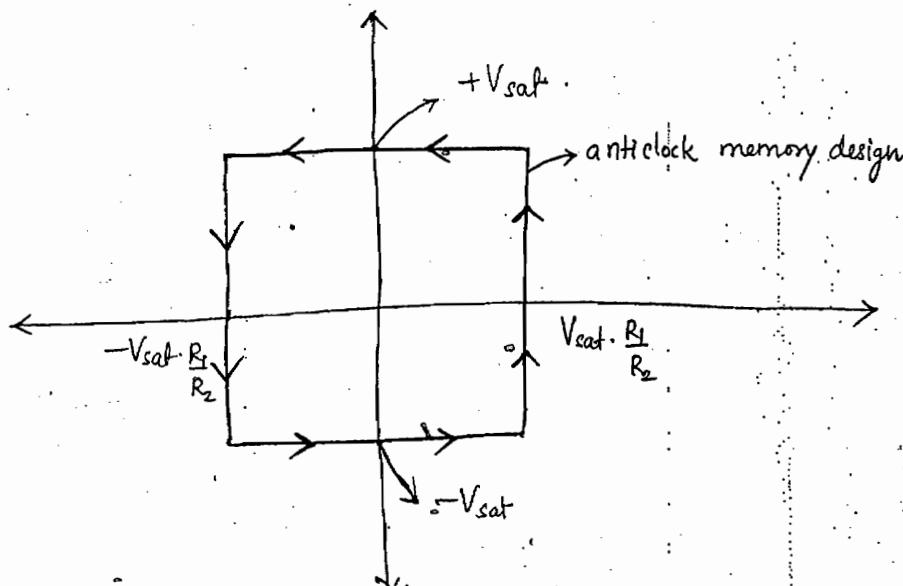
$$U_i \cdot \frac{R_2}{R_1 + R_2} - V_{sat} \cdot \frac{R_1}{R_1 + R_2} = 0$$

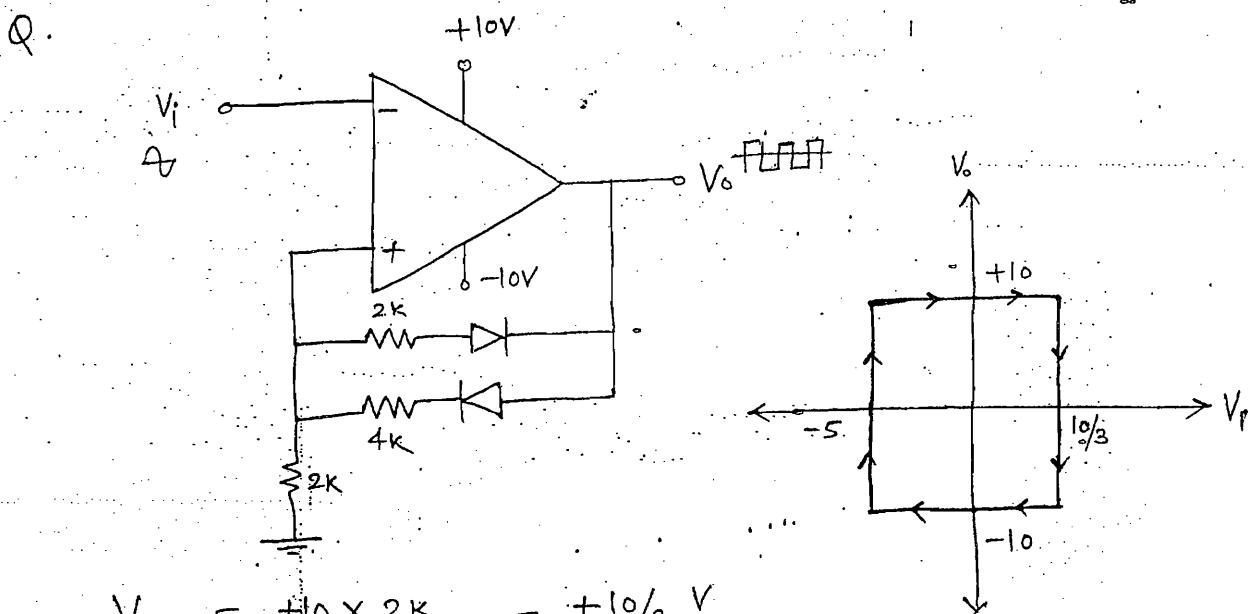
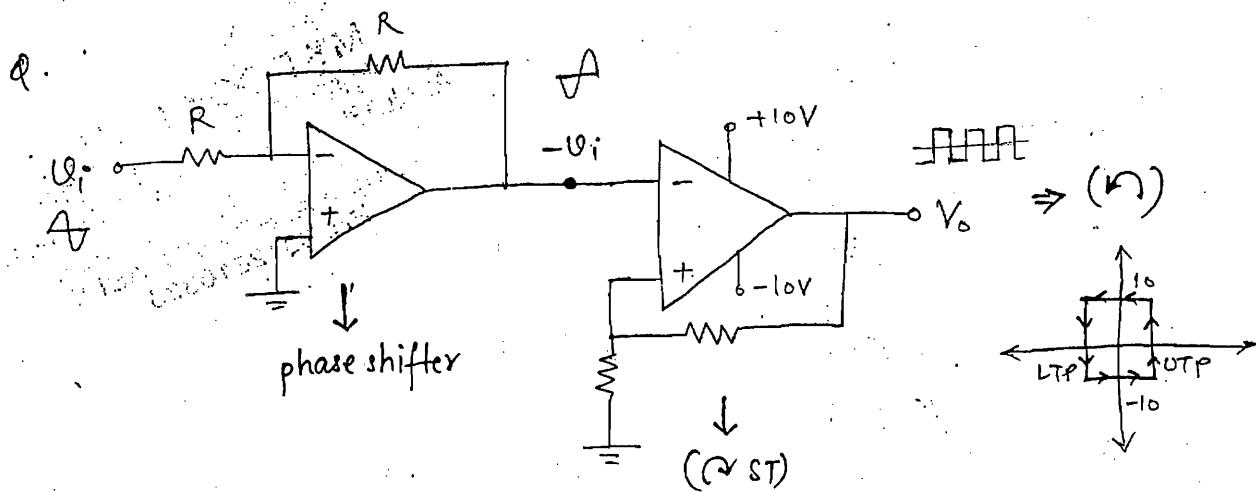
$$U_i = +V_{sat} \cdot \frac{R_1}{R_2}$$

When $V_o = +V_{sat}$

$$U_i \cdot \frac{R_2}{R_1 + R_2} + V_o \cdot \frac{R_1}{R_1 + R_2} = 0$$

$$U_i = -V_{sat} \cdot \frac{R_1}{R_2}$$

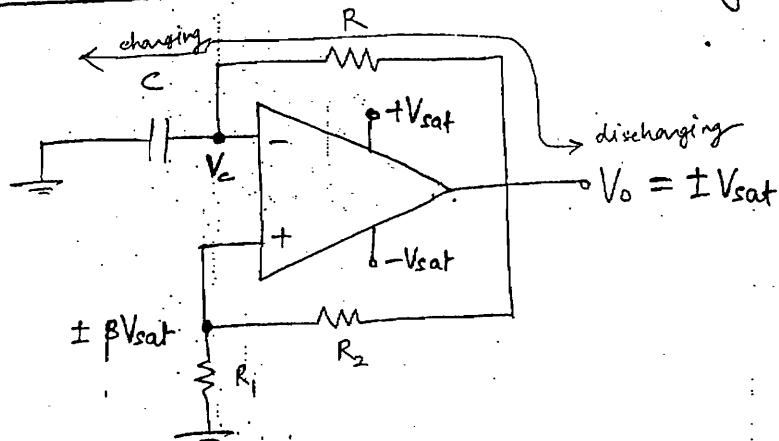


Problems:

$$V_{UTP} = +10 \times \frac{2k}{2k+4k} = +10/3 \text{ V}$$

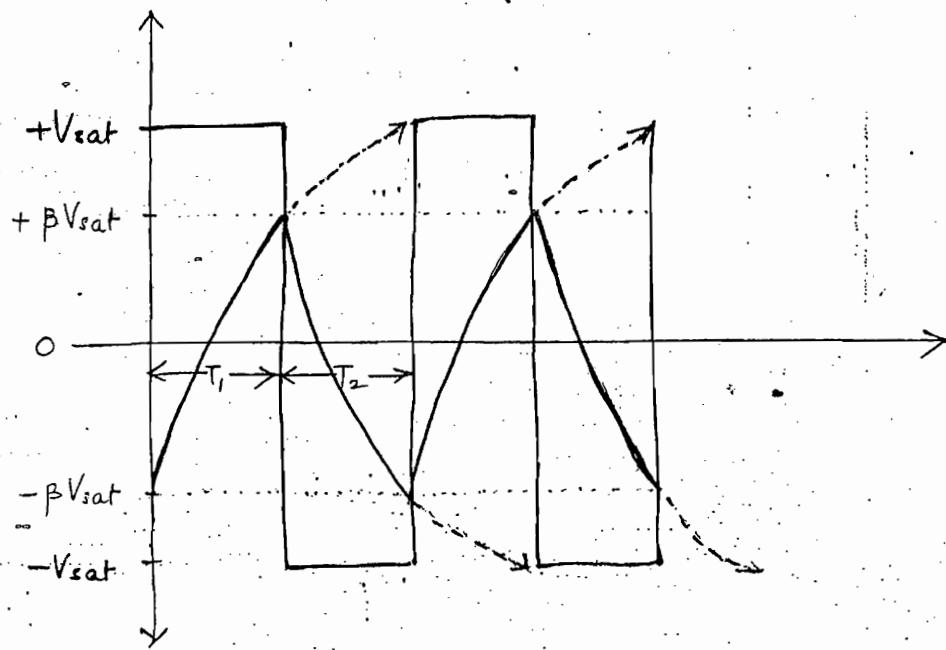
$$V_{LTP} = -10 \times \frac{2k}{2k+4k} = -5 \text{ V}$$

Astable Multivibrator (or) free running oscillator:



$$V_{U_{TP}} = +V_{sat} : \frac{R_1}{R_1+R_2} = \beta V_{sat}$$

$$V_{L_{TP}} = -V_{sat} \cdot \frac{R_1}{R_1+R_2} = -\beta V_{sat}$$



$$V_c = V_{final} (V_{final} - V_{initial}) e^{-t/RC}$$

At $t = T_2$,

$$V_c = -\beta V_{sat} \quad ; \quad V_{final} = -V_{sat} \quad ; \quad V_{initial} = +\beta V_{sat}$$

$$-\beta V_{sat} = -V_{sat} + (1+\beta) V_{sat} \cdot e^{-T_2/RC}$$

$$(1-\beta) V_{sat} = (1+\beta) \cdot V_{sat} e^{-T_2/RC}$$

$$T_2 = RC \ln \left(\frac{1+\beta}{1-\beta} \right)$$

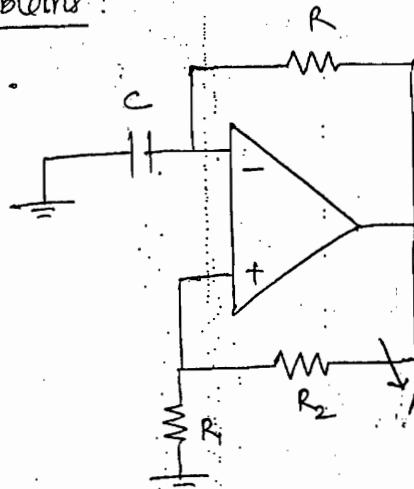
$$T_1 = RC \ln \left(\frac{1+\beta}{1-\beta} \right)$$

$$T = T_1 + T_2 = 2RC \ln \left(\frac{1+\beta}{1-\beta} \right)$$

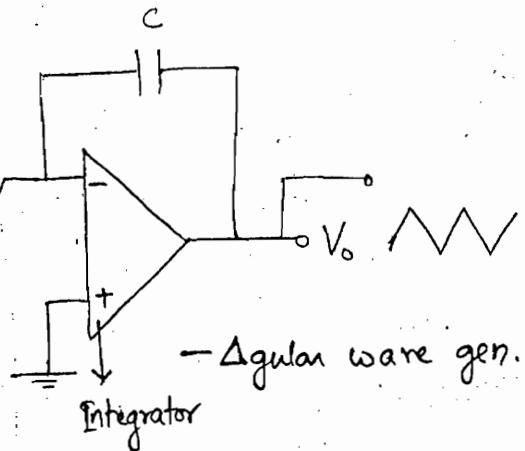
$$f = \frac{1}{2RC \ln \left(\frac{1+\beta}{1-\beta} \right)}$$

Problems:

Q.

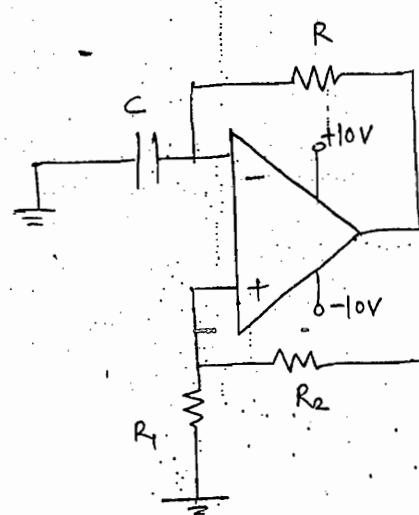


Astable multivibrator



Integrator

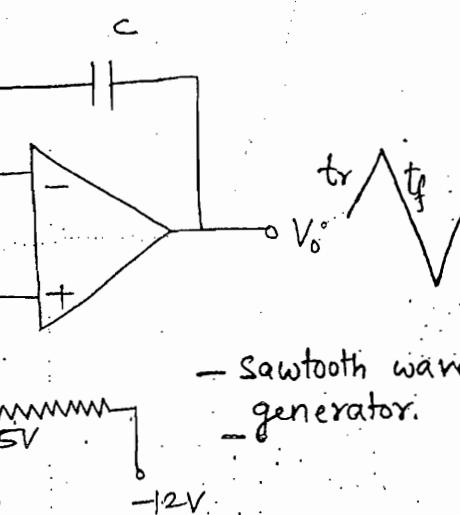
Q.

5V
-5VSV
SV

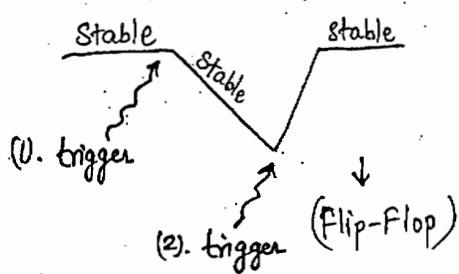
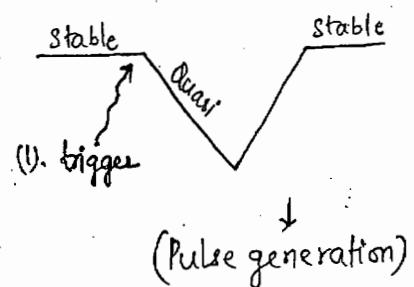
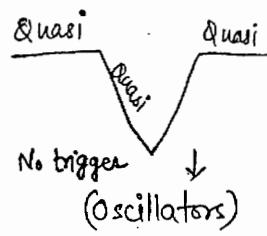
2.5V

+12V

-12V

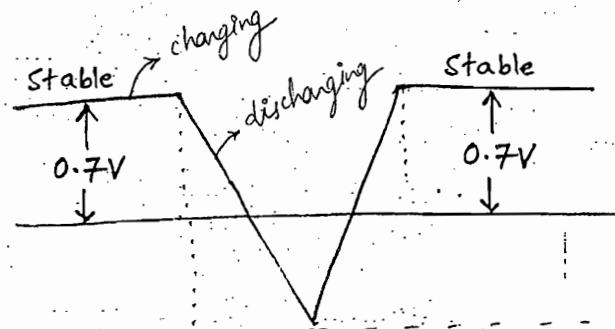
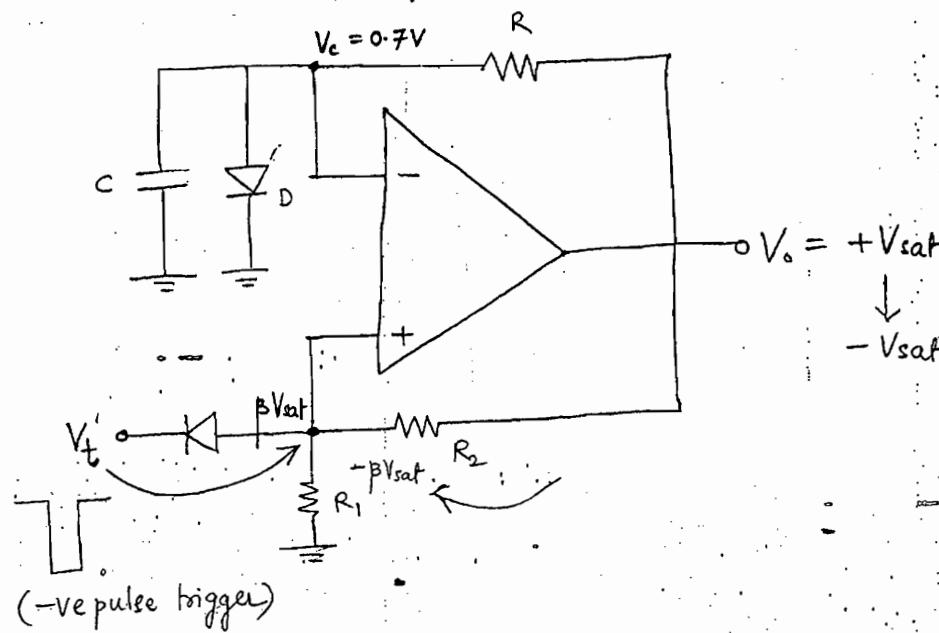
- Sawtooth wave
generator.when the wiper moves towards +ve side. $t_f > t_r$ when the wiper moves towards -ve side. $t_r > t_f$ Mono stable multivibrator:

without capacitor, there is no multivibrator.

Note: Bistable multivibratorMonostable multivibratorAstable multivibrator

Monostable Multivibrator:

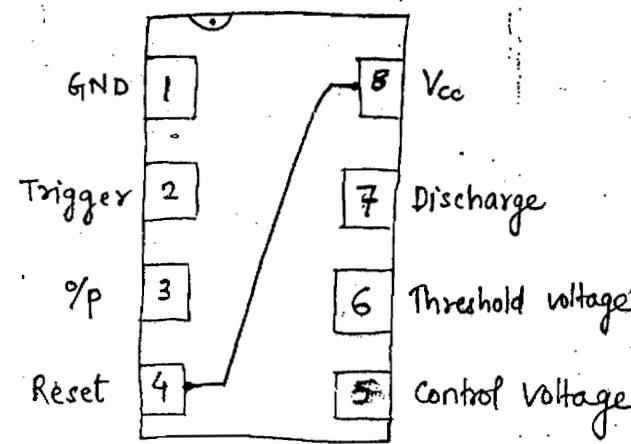
219

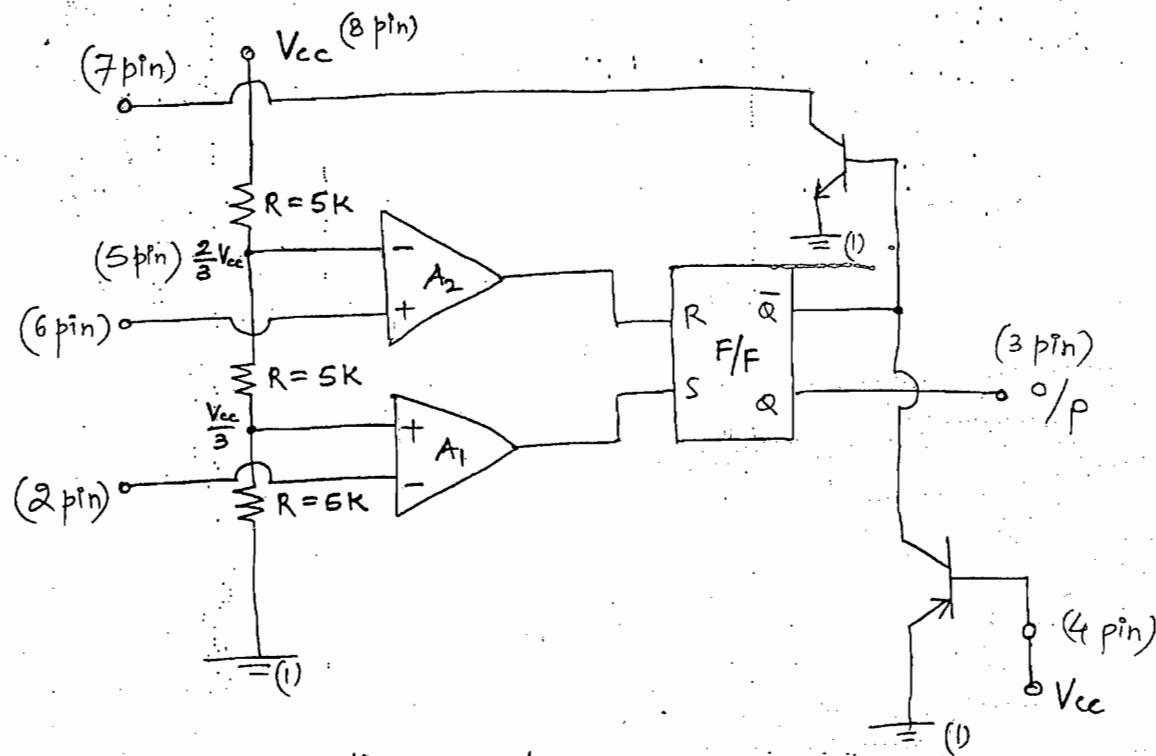


$$T_p = 0.693 RC$$

555 timer

Pin diagram →



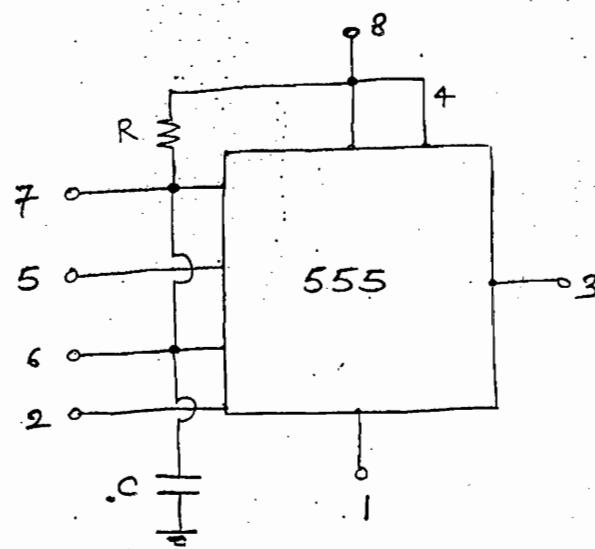


555 timer operating modes :

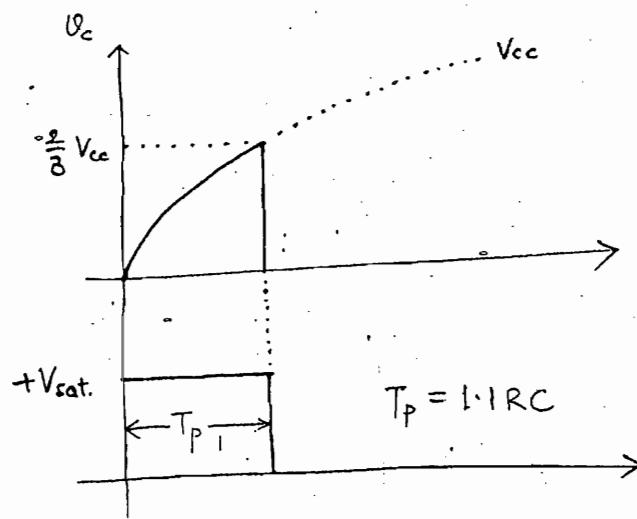
→ Monostable mode

→ Astable mode

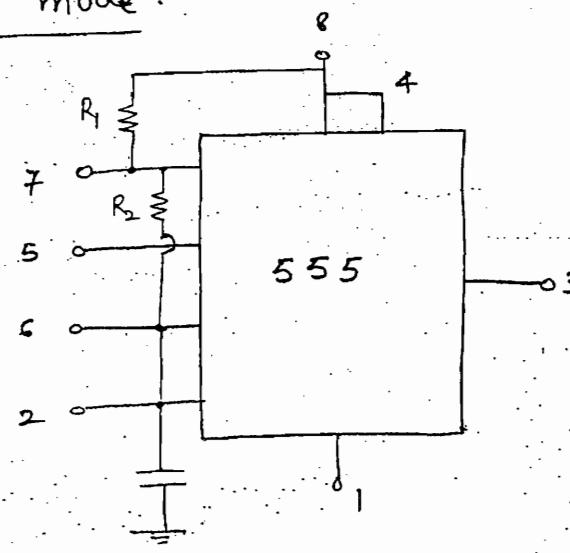
(i). Monostable 555 timer :



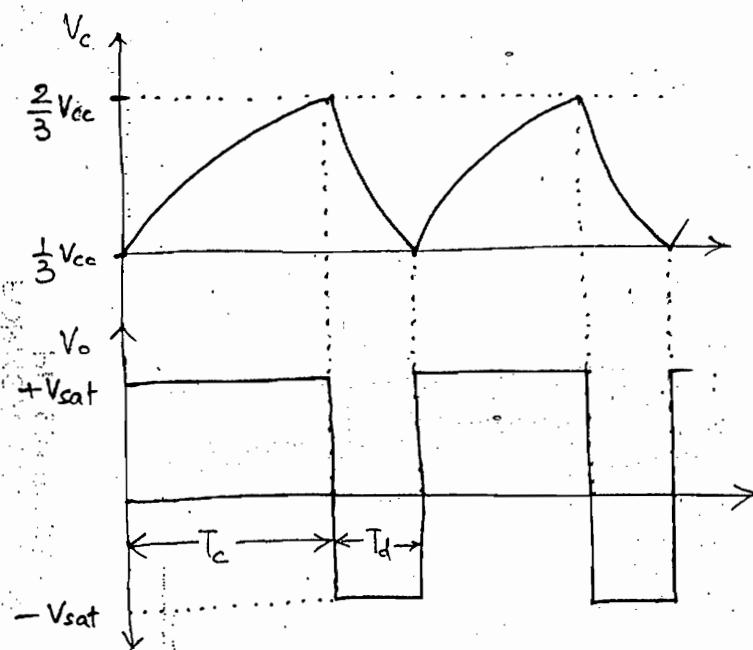
Condition	Q	\bar{Q}	%/P	Capacitor position
Stand by.	0	1	0	discharging - stable ($V_c = 0$)
trigger $< \frac{1}{3} V_{cc}$	1	0	1	charging - quasi ($V_c = \frac{2}{3} V_{cc}$)
$V_c > \frac{2}{3} V_{cc}$	0	1	0	discharging - stable ($V_c = 0$)



(2). Astable mode :



Condition	Q	\bar{Q}	%P	Capacitor position
Stand by	0	1	0	discharging, $V_c = \frac{1}{3} V_{cc}$
$V_c < \frac{1}{3} V_{cc}$	1	0	1	charging, $V_c = \frac{2}{3} V_{cc}$
$V_c > \frac{2}{3} V_{cc}$	0	1	0	discharging, $V_c = \frac{1}{3} V_{cc}$



$$T = T_c + T_d$$

$$T_c = 0.693 (R_1 + R_2) C$$

$$T_d = 0.693 \cdot R_2 C$$

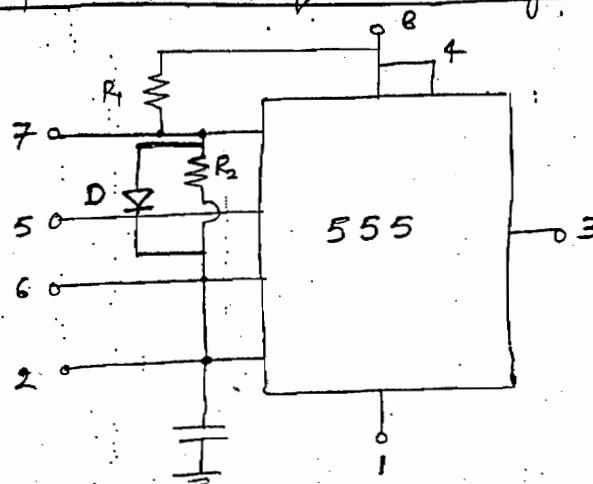
$$T = 0.693 (R_1 + 2R_2) C$$

$$\text{Duty cycle} = \frac{T_c}{T_c + T_d}$$

$$DC = \frac{0.693 (R_1 + R_2) C}{0.693 (R_1 + 2R_2) C}$$

$$DC = \frac{R_1 + R_2}{R_1 + 2R_2}$$

Modification \rightarrow Square wave generator



$$T_c = 0.693 R_1 C$$

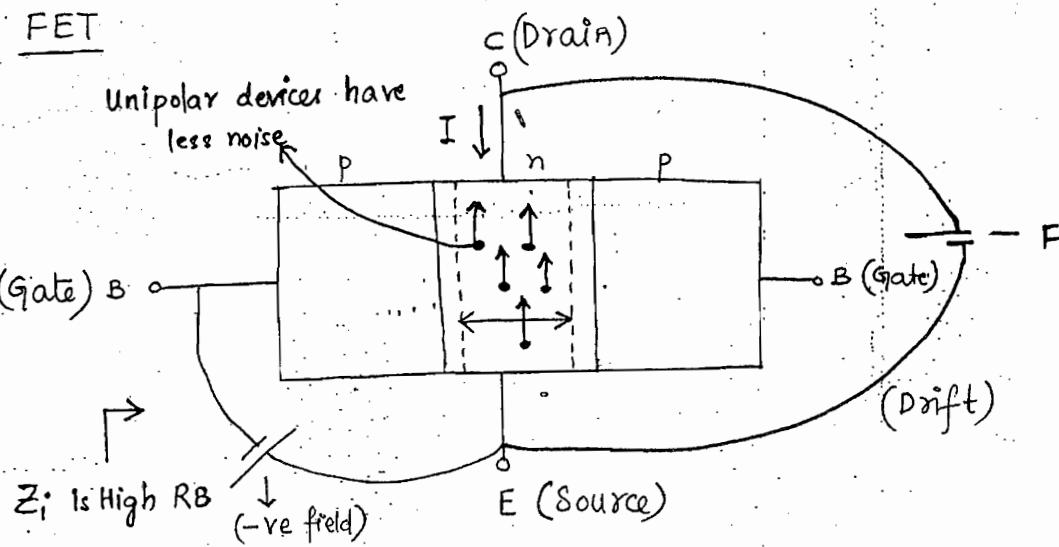
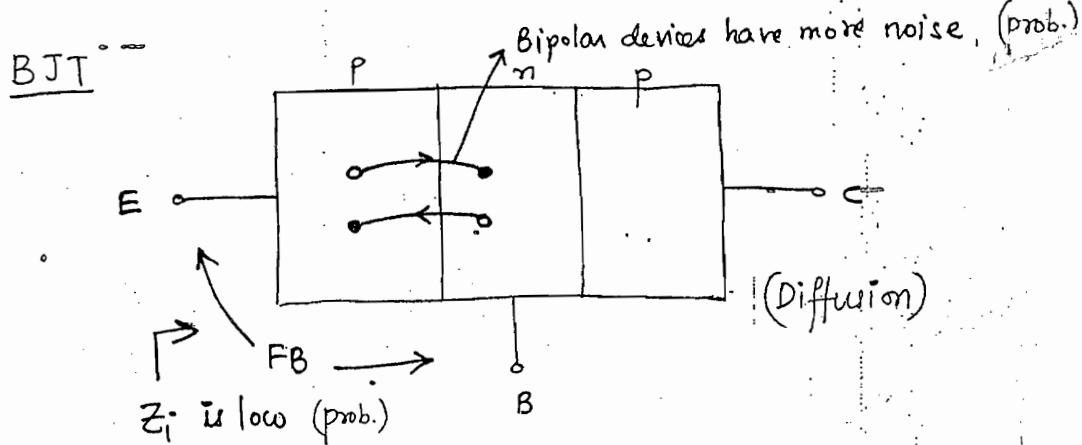
$$DC = \frac{R_1}{R_1 + R_2}$$

$$\text{if } R_1 = R_2 = R$$

$$DC = \frac{R}{R+R} = 0.5$$

Square wave

Introduction :



FET is a replacement of BJT. Both problems in BJT are solved in FET.

Bipolar jxn transistor is having following disadvantages :

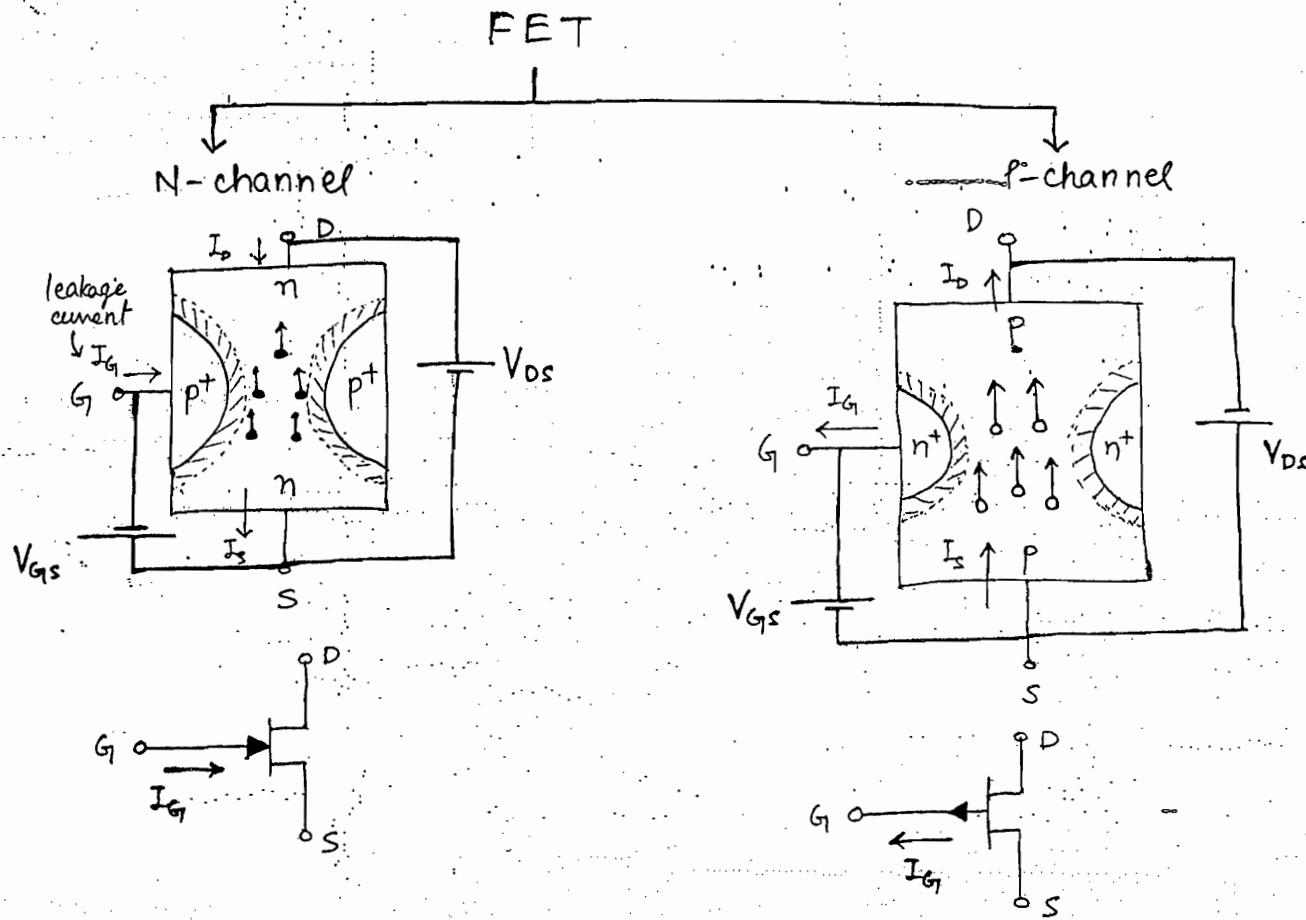
- (1). The I/p impedance is low.
- (2). The noise level is high.

Field effect transistor is having following advantages compared to BJT :

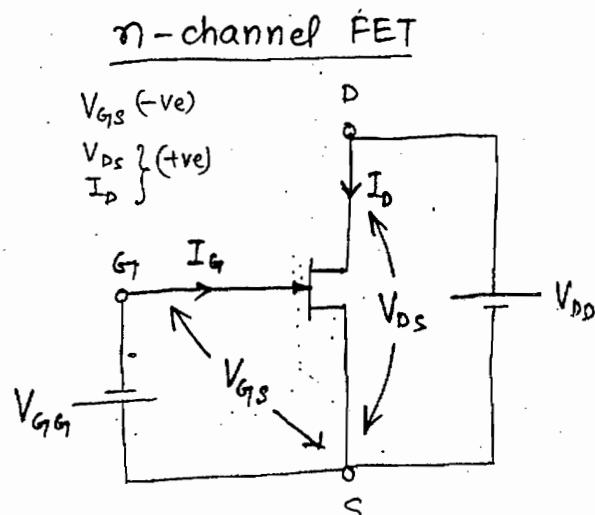
- (1). The I/p impedance is high becz the I/p is always R_B .
- (2). The noise level is comparatively low becz it is a unipolar device.

Note: BJT is a current controlled device whereas FET is a voltage controlled device.

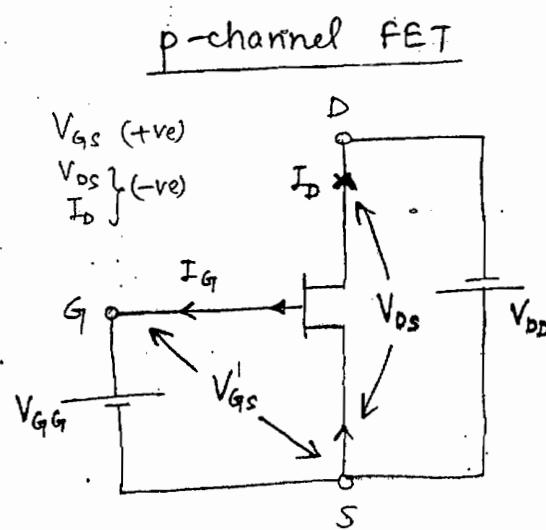
Types of FET are :



Important points in FET :



ℓ/p parameters $\rightarrow V_{GS}, I_G$
 ℓ/o parameters $\rightarrow V_{DS}, I_D$



(1). Gate current I_G is called as leakage current or reverse saturation current. In FET analysis, we are assuming $I_G \approx 0$.

(2). There will be no I/p characteristic study of FET.

(3). - There are 2 types of graphs:

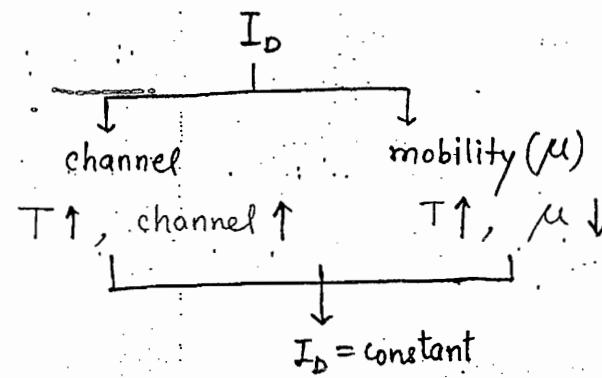
(i). Drain characteristics (V_{DS} vs I_D at constant V_{GS})

(ii). Transfer characteristics (V_{GS} vs I_D at constant V_{DS})

(4). For FET analysis, transfer characteristics study became dominant than drain characteristics.

(5). The Q point in FET is defined as $Q [I_D]_Q, (V_{GS})_Q$.

(6). FET is always thermally stable.



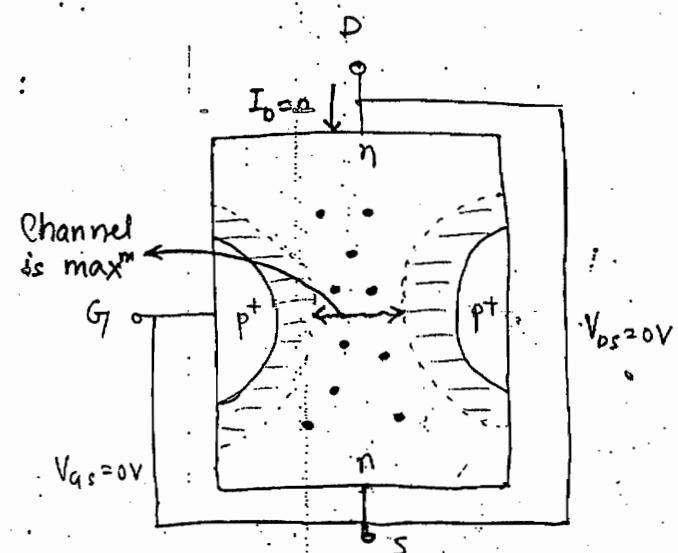
(7). For FET biasing, we have to fix V_{GS} voltage but not drain current I_D .

Working principle of FET :

Case - (1) :

$$V_{GS} = 0V$$

$$V_{DS} = 0V$$



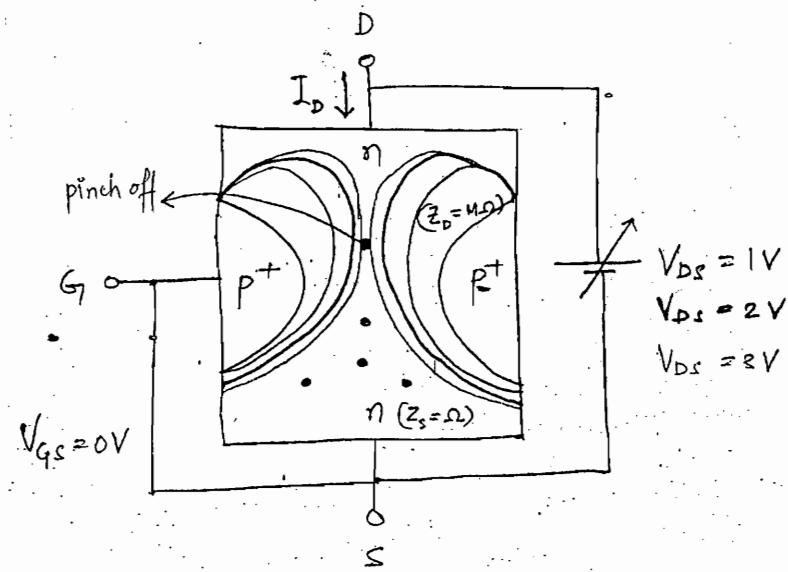
Conclusions:

- (1). When $V_{GS} = 0V$, max^m channel can be achieved
- (2). When $V_{DS} = 0V$, the drain current I_D becomes zero.

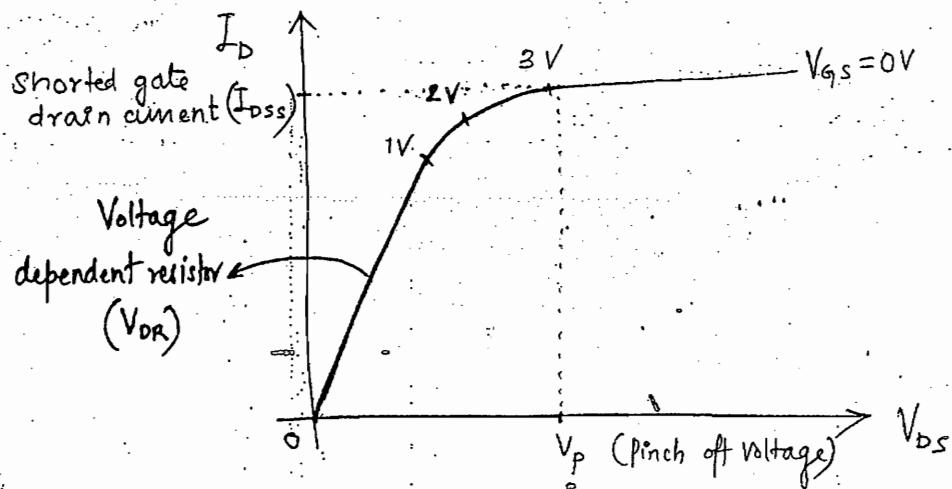
Case - (2):

$$V_{GS} = 0V$$

$$V_{DS} \neq 0V$$



Characteristics (V_{DS} vs I_D):

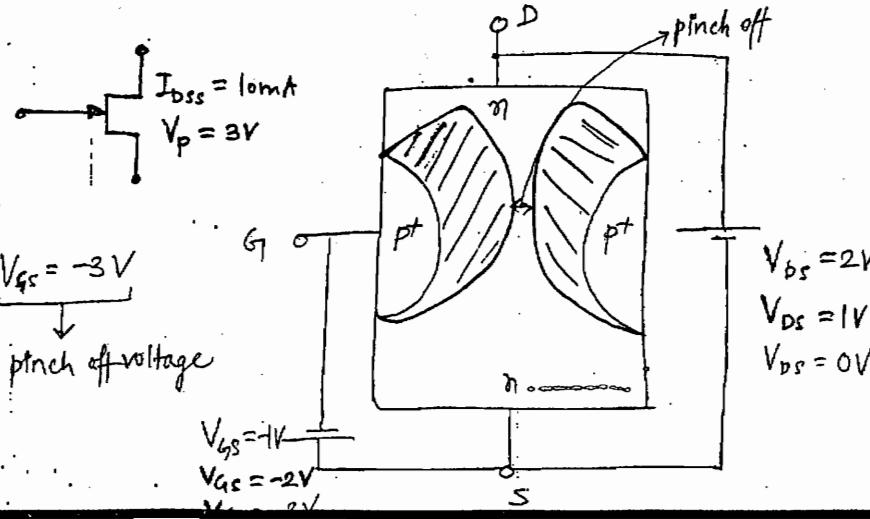


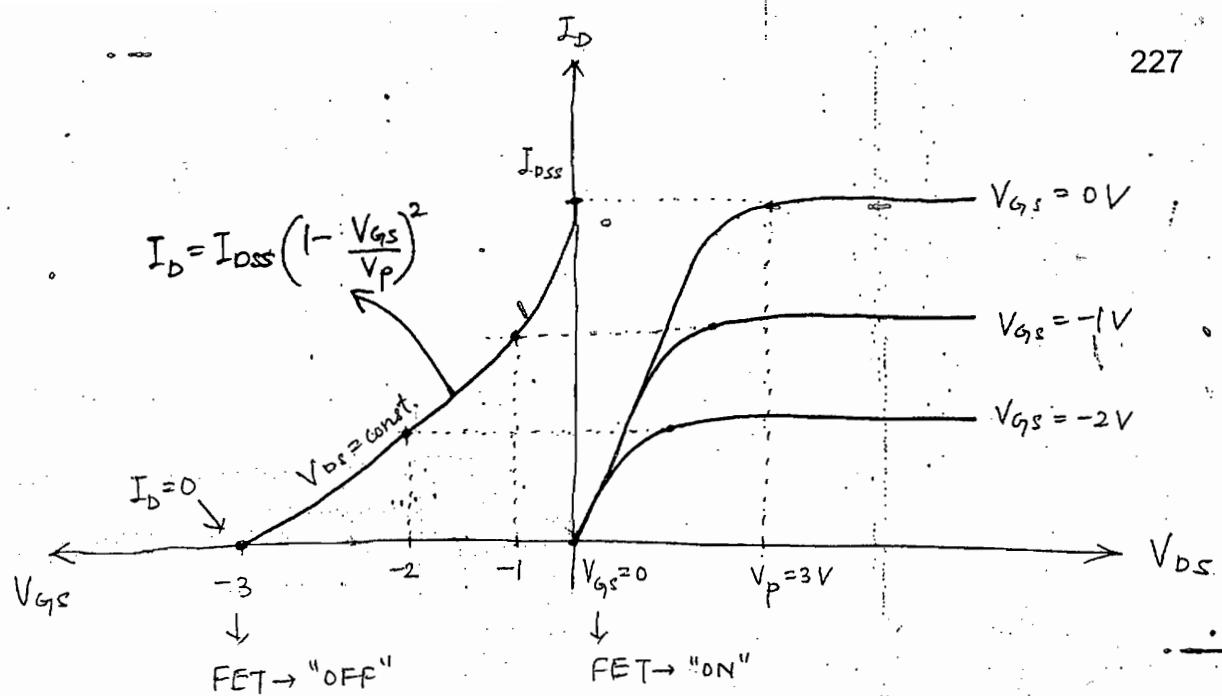
Case - (3):

$$V_{GS} \neq 0V$$

$$V_{DS} \neq 0V$$

When $V_{DS} = 0V$; $V_{GS} = -3V$



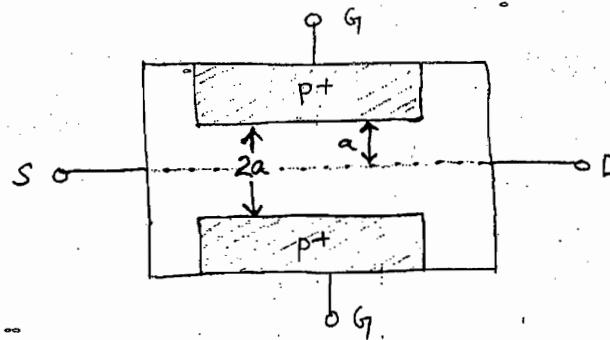


(1). Transfer characteristic eqn is

$$I_D = I_{DSS} \left(1 - \frac{V_{GS}}{V_p}\right)^2$$

$$(2) |V_{GS}|_{off} = |V_p|$$

$$(3) V_p = \frac{q^2 \gamma N_A}{2\epsilon} \quad (\text{for p-channel}) \quad V_p = \frac{q^2 \gamma N_D}{2\epsilon} \quad (\text{for n-channel})$$



$2a \rightarrow$ full channel Height
 $a \rightarrow$ Half channel Height.

- * On increasing doping, penetration decreases and Hence pinch off voltages increases.

FET parameters:

$$(1). \text{ AC drain resistance, } r_d = \left. \frac{\Delta V_{DS}}{\Delta I_D} \right|_{V_{GS}=\text{const.}}$$

$$(2). \text{ Transconductance, } g_m = \left. \frac{\Delta I_D}{\Delta V_{GS}} \right|_{V_{DS}=\text{const.}}$$

$$(3). \text{ Amplification factor, } \mu = \left. \frac{\Delta V_{DS}}{\Delta V_{GS}} \right|_{I_D=\text{const.}}$$

$$\boxed{\mu = g_m \cdot r_d}$$

Expression for g_m :

$$I_D = I_{DSS} \left(1 - \frac{V_{GS}}{V_p} \right)^2$$

$$\frac{dI_D}{dV_{GS}} = -2 \frac{I_{DSS}}{V_p} \left(1 - \frac{V_{GS}}{V_p} \right)$$

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = -2 \frac{I_{DSS}}{V_p} \left[1 - \frac{(V_{GS})_Q}{V_p} \right]$$

$$\boxed{g_m = g_{m0} \cdot \left[1 - \frac{(V_{GS})_Q}{V_p} \right]}$$

where, $\boxed{g_{m0} = -\frac{2I_{DSS}}{V_p}}$

(max^m transconductance)

And,

$$I_D = I_{DSS} \left(1 - \frac{V_{GS}}{V_p} \right)^2 \Rightarrow \left(1 - \frac{V_{GS}}{V_p} \right) = \sqrt{\frac{I_D}{I_{DSS}}}$$

$$g_m = -2 \frac{I_{DSS}}{V_p} \left(1 - \frac{V_{GS}}{V_p} \right)$$

$$= -2 \frac{I_{DSS}}{V_p} \cdot \sqrt{\frac{I_D}{I_{DSS}}}$$

$$\therefore g_m = -2 \frac{\sqrt{I_{DSS} \cdot (I_D)_Q}}{V_p} = \frac{2}{|V_p|} \cdot \sqrt{I_{DSS} \cdot (I_D)_Q}$$

FET Biasing :

conditions :

$$(1) \cdot I_G \approx 0$$

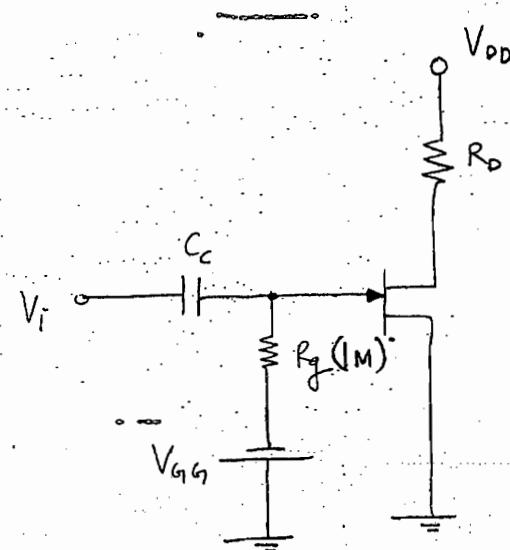
$$(2) \cdot I_D \approx I_s$$

$$(3) \cdot I_D = I_{DSS} \left(1 - \frac{V_{GS}}{V_p}\right)^2$$

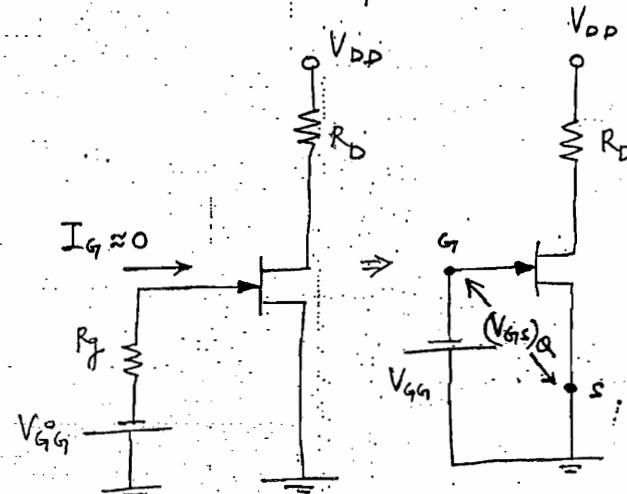
$$(4) \cdot Q \text{ point} \rightarrow Q \left[(I_D)_Q, (V_{GS})_Q \right]$$

(1). Fixed Biasing -

Model - (1) : Common source bypass



DC analysis



• R_g are meant for AC sig only

(1). $(V_{GS})_Q = -V_{GG}$ (fixed) - Biasing by keeping a battery.

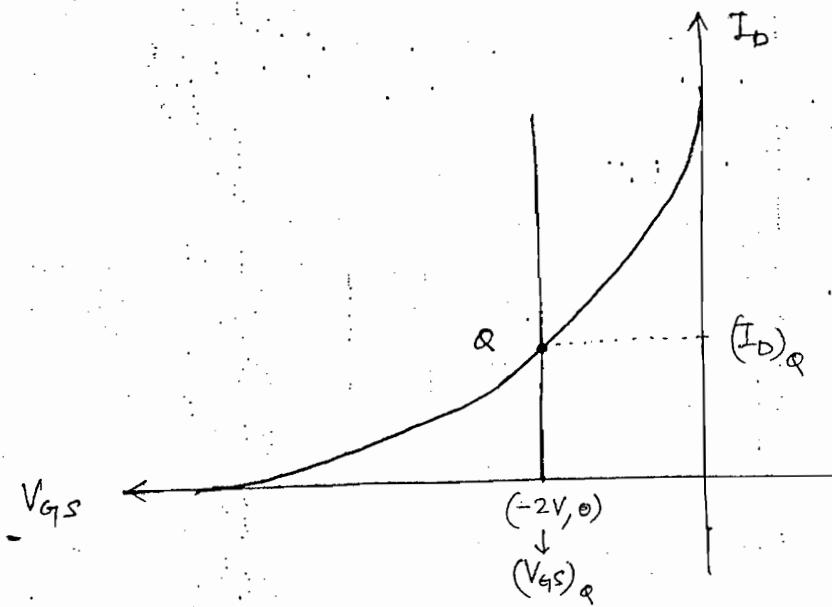
$$(2). (I_D)_Q = I_{DSS} \left[1 - \left(\frac{(V_{GS})_Q}{V_p}\right)^2\right]$$

$$(3). (V_{DS})_Q = V_{DD} - (I_D)_Q \cdot R_D$$

$$(4). (V_D)_Q = (V_{DS})_Q$$

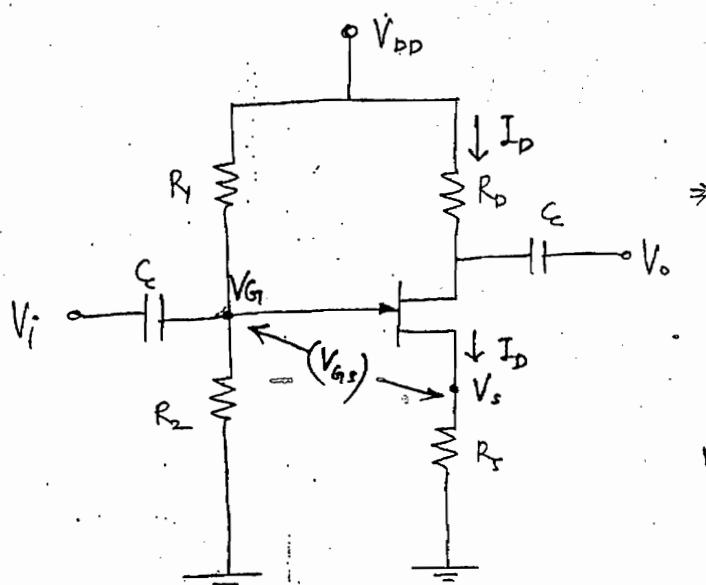
$$(5). (V_S)_Q = \cancel{(V_S)_Q} 0$$

$$(6). (V_G)_Q = (V_{GS})_Q$$

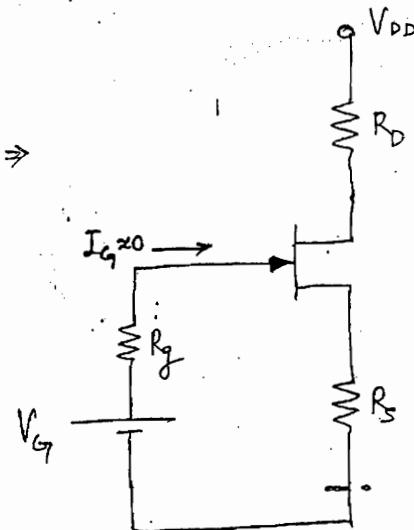


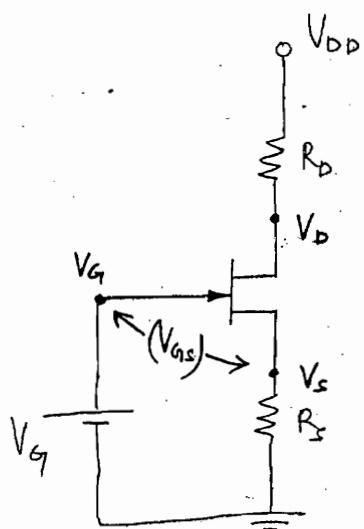
This biasing is not good becoz battery can vary the Q point as there are fluctuations in battery voltage. This is not a stable biasing. Hence, some modifications are required. Bypass amplifiers are always unstable.

(2) Voltage Divider Biasing:



DC analysis



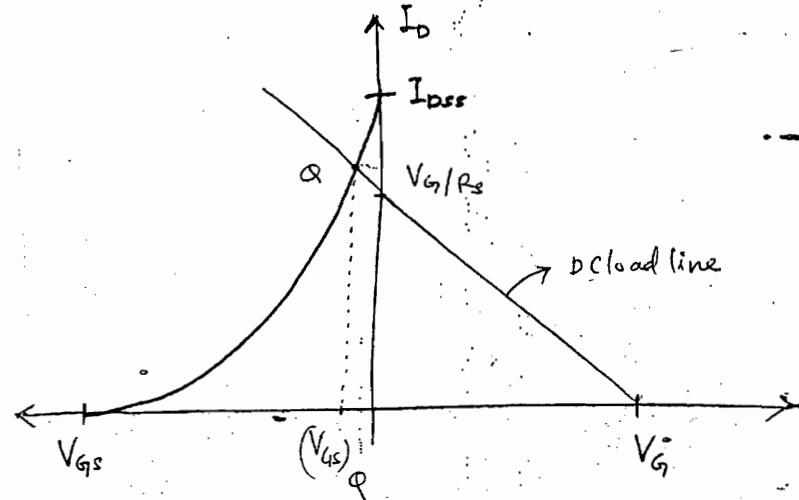


$$V_G = \frac{V_{DD} \times R_2}{R_1 + R_2}$$

$$R_g = R_1 // R_2$$

$$V_G = V_{GS} + I_D \cdot R_s$$

$$V_{GS} = V_G - I_D R_s$$



$$(1) \quad I_D = I_{DSs} \left(1 - \frac{V_{GS}}{V_p}\right)^2 \quad \text{and} \quad V_{GS} = V_G - I_D R_s$$

Solving, we get,

$$(I_D)_Q = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

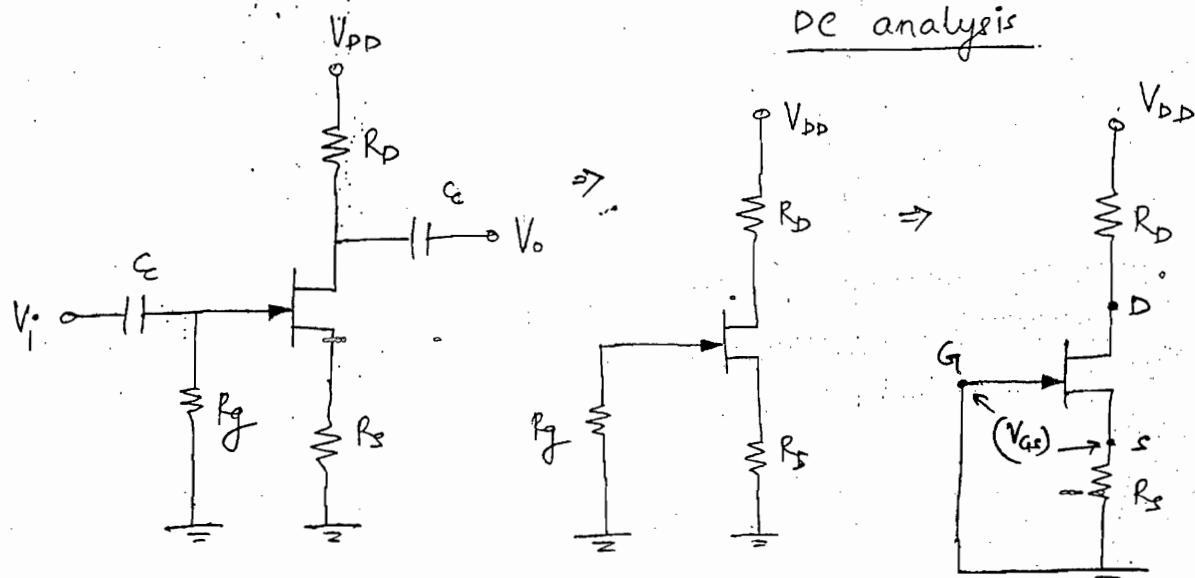
+ve root (n-channel)

-ve root (p-channel)

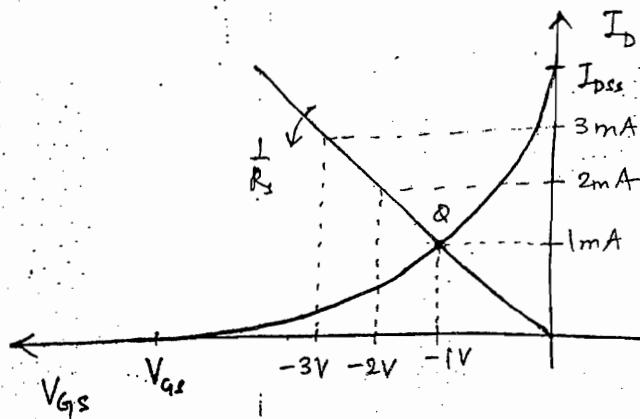
$$(2) \quad (V_{GS})_Q = V_G - (I_D)_Q \cdot R_s$$

- fixed biasing uses 2 batteries whereas this biasing is using single battery so it is an improved biasing but due to the presence of battery, it is also not much stable. Hence, more modification is required.

(3). Self Biasing (stable Biasing): Common source unbypass



$$V_{GS} = -I_D R_s \rightarrow \text{Ohm's law}$$



$$(1) \quad I_D = I_{DSS} \left(1 - \frac{V_{GS}}{V_p}\right)^2 \quad \text{and} \quad V_{GS} = -I_D R_s$$

$$(2) \quad (I_D)_Q = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a} \quad \begin{array}{l} \text{+ve root (n-channel)} \\ \text{-ve root (p-channel)} \end{array}$$

$$(3) \quad (V_{GS})_Q = - (I_D)_Q \cdot R_s$$

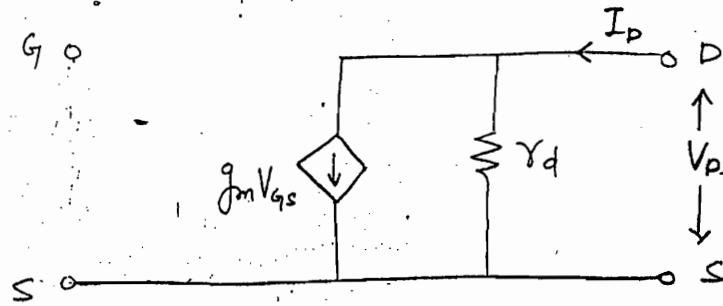
- Self biasing is a stable biasing and improved biasing as it does not depend upon battery voltage.
- UnBypass amplifiers are always stable.

FET model :

$$I_D = f(V_{GS}, V_{DS})$$

$$I_D = \frac{\partial I_D}{\partial V_{GS}} \cdot V_{GS} + \frac{\partial I_D}{\partial V_{DS}} \cdot V_{DS}$$

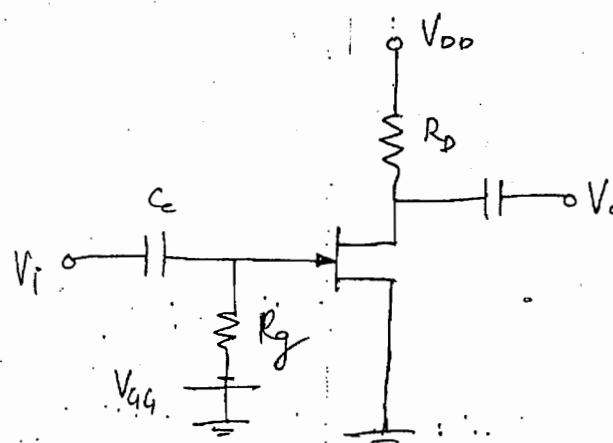
$$\underline{I_D = g_m \cdot V_{GS} + \frac{V_{DS}}{r_d}}$$

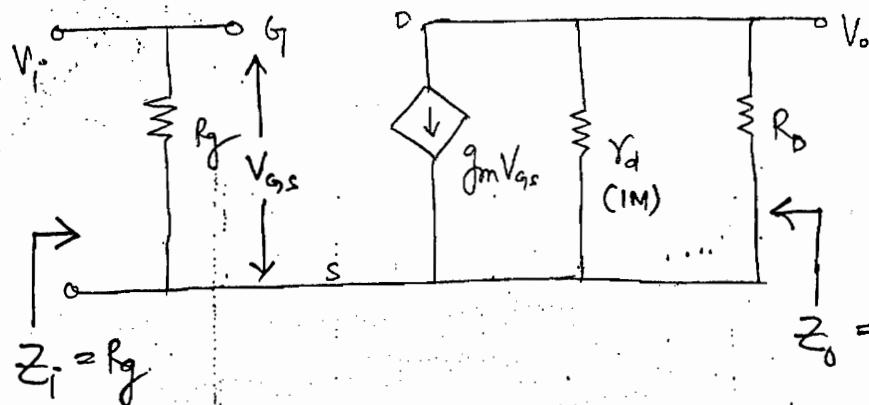


FET amplifiers :

- (1). CS bypass amplifier.
- (2). CS unbypass amplifier.
- (3). CD amplifier.
- (4). CG amplifier.

(1). CS bypass amplifier :





$$r_d \geq 10 R_D$$

$$Z_o = r_d // R_D \approx R_D$$

Voltage gain:

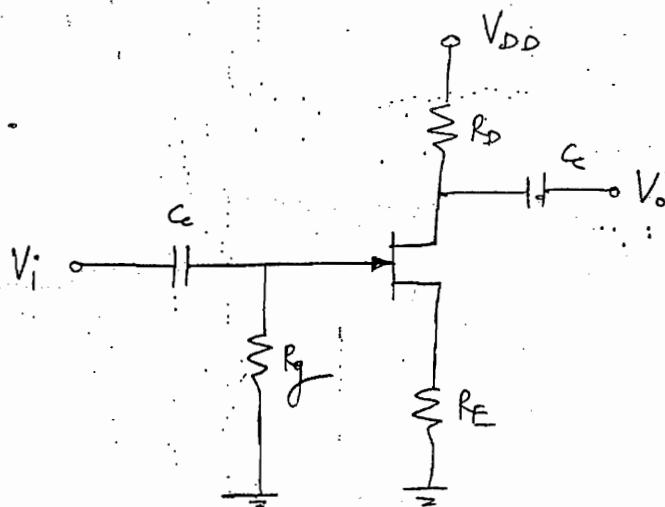
$$V_o = -g_m \cdot V_{GS} \cdot (r_d // R_D)$$

$$V_{GS} = V_i$$

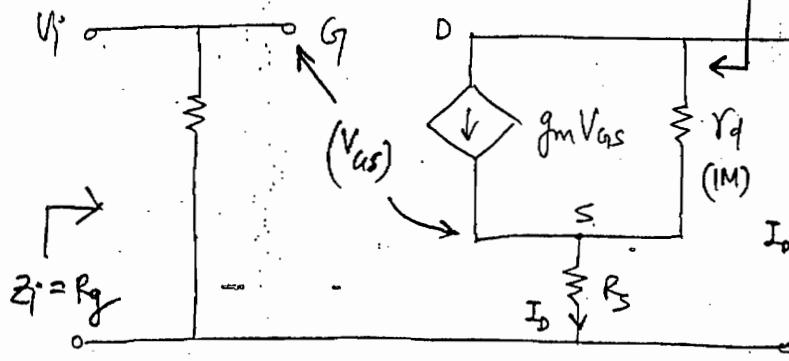
$$V_o = -g_m V_i R_D$$

$$A_V = \frac{V_o}{V_i} = -g_m R_D$$

CS unbypassed amplifier:



$$Z_o = r_d + (1 + \mu) R_E$$



$$\begin{aligned} Z'_o &= R_D // Z_o \\ &= R_D // [r_d + (1 + \mu) R_E] \\ &\approx R_D \end{aligned}$$

Voltage gain:

$$V_o = -I_D R_D \quad \text{--- (1)}$$

$$I_D R_D + (I_D - g_m V_{GS}) \gamma_d + I_D R_S = 0 \quad \text{--- (2)}$$

$$V_i^* = V_{GS} + I_D R_S \quad \text{--- (3)}$$

$$V_{GS} = V_i^* - I_D R_S$$

$$I_D R_D + [I_D - g_m (V_i^* - I_D R_S)] \gamma_d + I_D R_S = 0$$

$$I_D R_D + I_D \gamma_d - g_m \gamma_d (V_i^* - I_D R_S) + I_D R_S = 0$$

$$I_D (R_D + \gamma_d + R_S + g_m \gamma_d R_S) = g_m \gamma_d V_i^*$$

$$I_D = \frac{g_m \gamma_d V_i^*}{(R_D + R_S + \gamma_d + g_m \gamma_d R_S)}$$

$$V_o = -I_D R_D$$

$$V_o = \frac{-g_m \gamma_d R_D \cdot V_i^*}{(R_D + R_S + \gamma_d + g_m \gamma_d R_S)}$$

$$\frac{V_o}{V_i^*} = \frac{-g_m \gamma_d R_D}{(R_D + R_S + \gamma_d + g_m \gamma_d R_S)}$$

$$= \frac{-g_m R_D}{\left(\frac{R_D}{\gamma_d} + \frac{R_S}{\gamma_d} + 1 + g_m R_S\right)}$$

↓ ↓
neglected neglected.

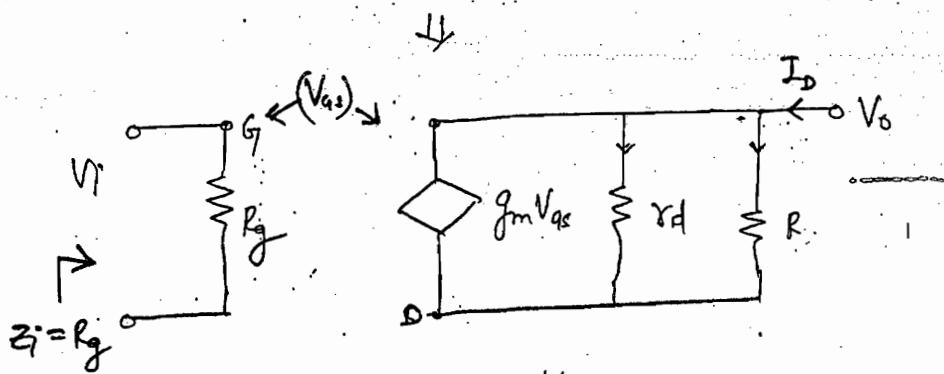
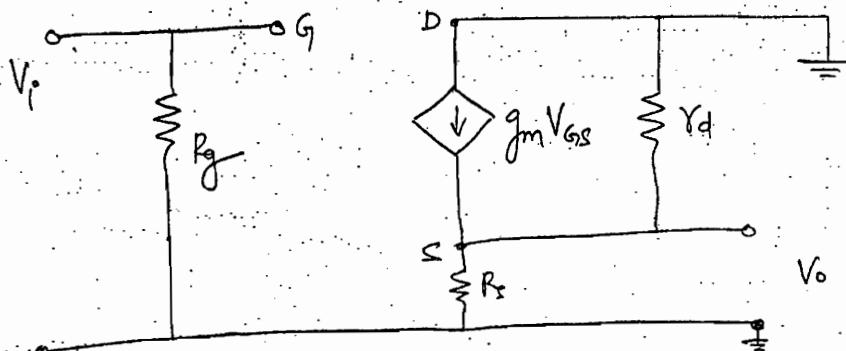
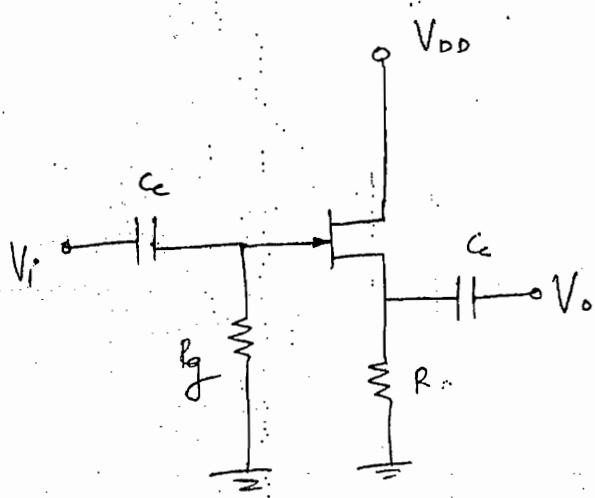
$$A_V = \frac{V_o}{V_i^*} = \frac{-g_m R_D}{(1 + g_m R_S)}$$

$$A_V = \frac{-g_m R_D}{1 + g_m R_S}$$

if $g_m R_S \ggg 1$, then

$$A_V = -\frac{R_D}{R_S} \rightarrow \text{Stable}$$

(3) Common Drain amplifier:



$$I_D + g_m V_{GS} = \frac{V_o}{R_d} + \frac{V_o}{R_s}$$

$$V_i = V_{GS} + V_o \Rightarrow V_o = -V_{GS} \Rightarrow V_{GS} = -V_o$$

↓

$$I_D - g_m V_o = \frac{V_o}{R_d} + \frac{V_o}{R_s}$$

$$I_D = \left(g_m + \frac{1}{R_d} + \frac{1}{R_s} \right) V_o$$

%p Impedance :

$$Z_o = \frac{V_o}{I_D} = \frac{1}{\frac{1}{r_d} + \frac{1}{R} + g_m} \approx \frac{1}{\frac{1}{R} + g_m} \approx \frac{1}{g_m}$$

neglect

(Impedance matching).

Voltage gain :

$$V_o = g_m V_{GS} \cdot (r_d // R_s)$$

$$V_i^o = V_{GS} + V_o$$

$$V_i^o = V_{GS} + g_m V_{GS} - (r_d // R)$$

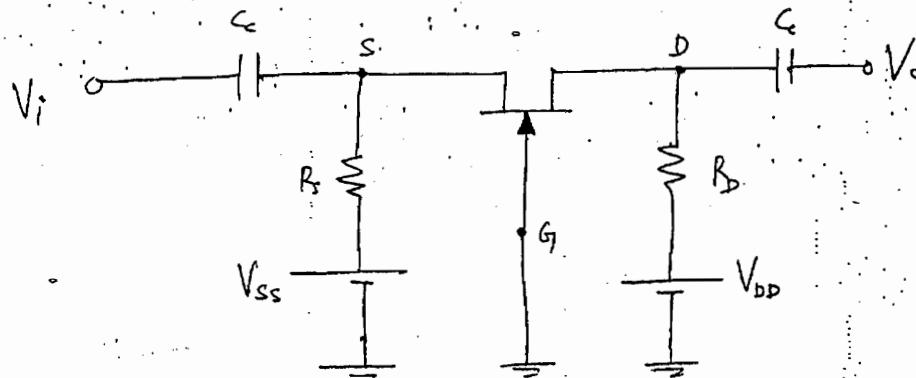
$$V_{GS} \equiv \frac{V_i^o}{1 + g_m (r_d // R)}$$

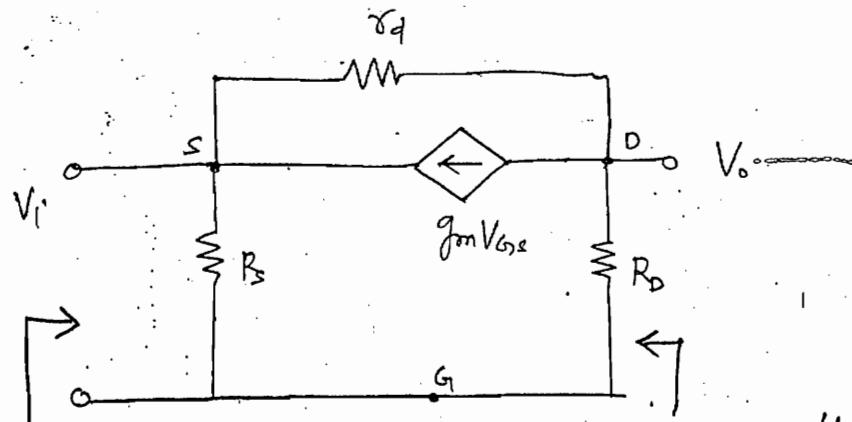
$$V_o = g_m \cdot V_i^o \cdot (r_d // R) \\ \frac{1}{1 + g_m (r_d // R)}$$

$$A_V = \frac{V_o}{V_i^o} = \frac{g_m \cdot R_s}{(1 + g_m R_s)}$$

If $g_m R_s \gg 1$, then

$$A_V = \frac{V_o}{V_i^o} \approx 1 \quad \begin{matrix} \text{Voltage} \\ \text{Buffer} \end{matrix}$$

(4). Common gate amplifier:



$$Z_i = R_s \parallel \frac{1}{g_m}$$

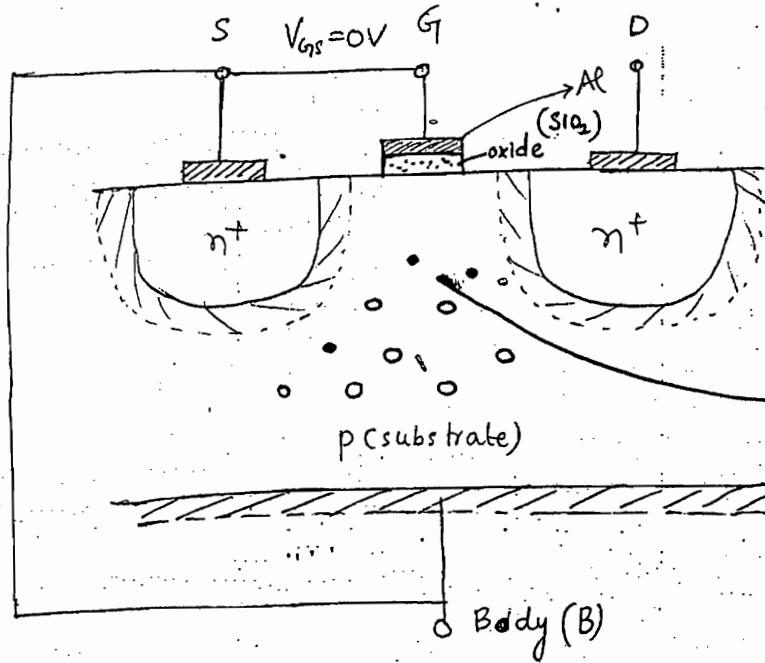
$$Z_o = r_d \parallel R_D \approx R_D$$

$$A_v = g_m R_D$$

Conclusion:

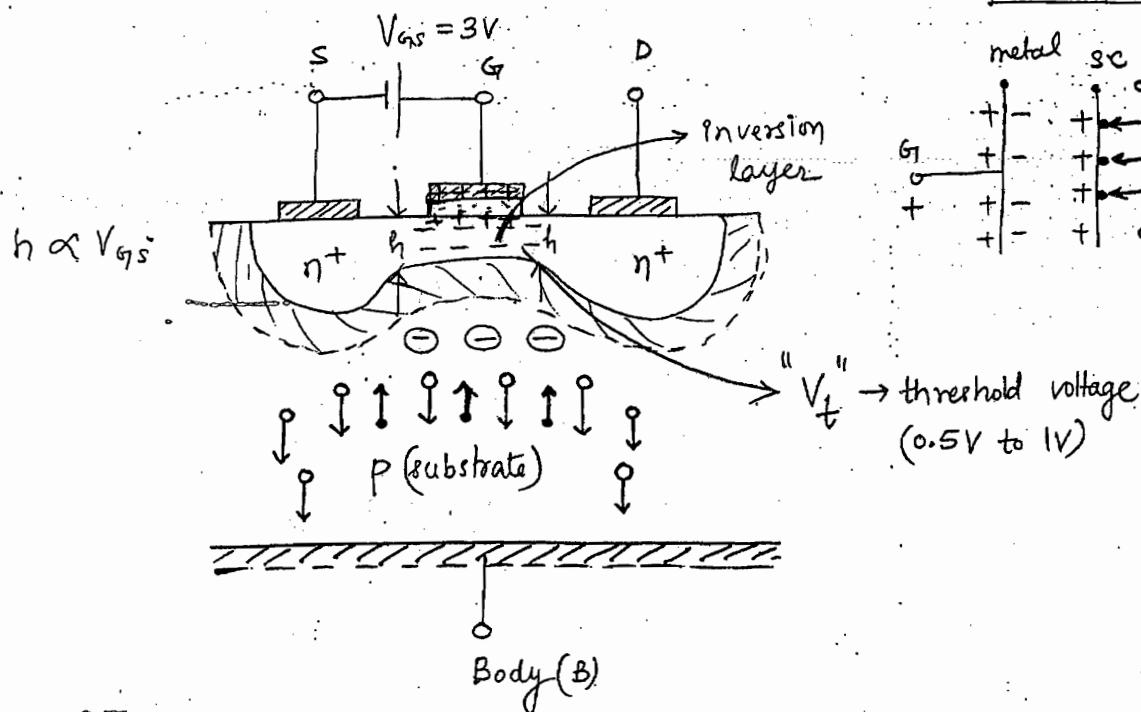
	Z_i	Z_o	A_v
(1). CS bypass	R_g	$r_d \parallel R_D$	$-g_m R_D$
(2). CS unbypass	R_g	$r_d + (1+\mu) R_s \parallel R_D$	$\frac{-g_m R_D}{(1+g_m R_s)}$
(3). CD	R_g	$R_s \parallel \frac{1}{g_m}$	$\frac{g_m R_s}{(1+g_m R_s)}$
(4). CG	$R_s \parallel \frac{1}{g_m}$	$r_d \parallel R_D$	$+g_m R_D$

n-channel Enhancement MOSFET (NMOS)

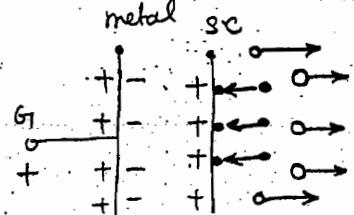


Holes \rightarrow majority charge carriers
Electrons \rightarrow minority charge carriers

channel is not existing at $V_{GS} = 0V$
 $\therefore I_D = 0$



Mos capacitor



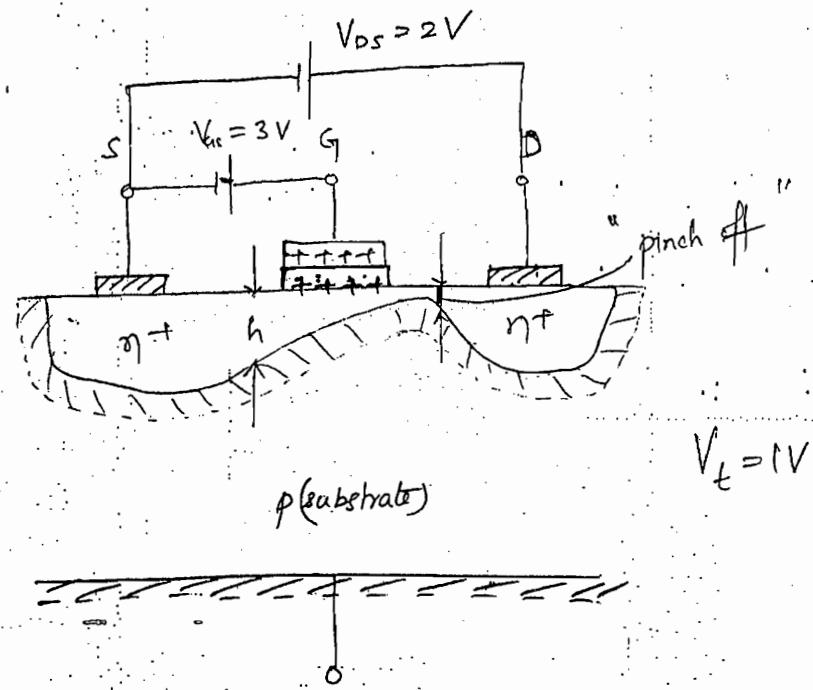
$(V_{GS} - V_t) \rightarrow$ effective voltage (or) over drive voltage

- V_t is the min^m V_{GS} voltage at which conducting channel will be formed.

$$\begin{aligned}V_{DG} &= V_{DS} + V_{SG} \\&= V_{DS} - V_{GS}\end{aligned}$$

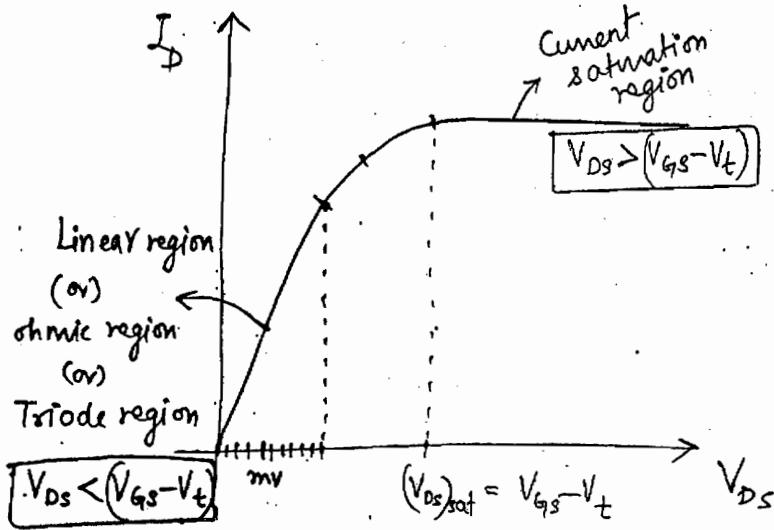
$$V_{GD} = V_{GS} - V_{DS}$$

$$\text{Here, } V_{DS} \geq 0 \Rightarrow V_{GD} = V_{GS}$$



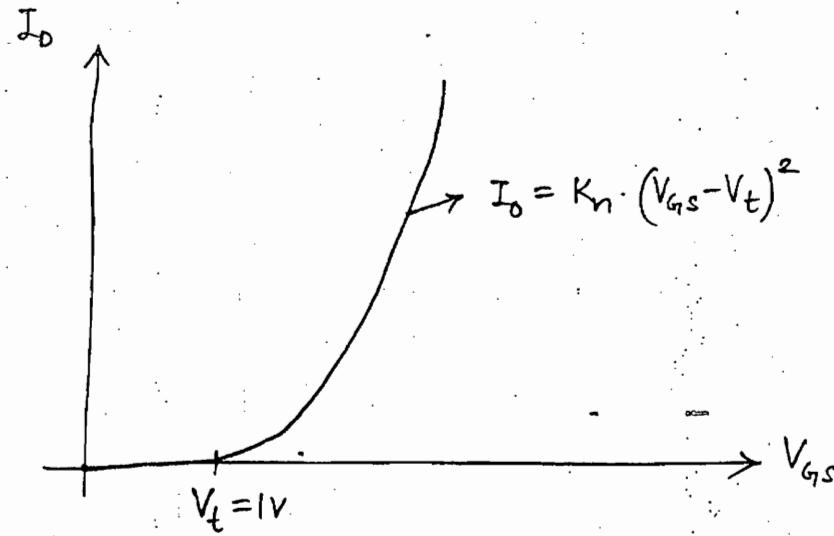
$$\begin{aligned}V_{GD} &= V_{GS} - V_{DS} \\&= 3V - 2V \\&= 1V = V_t\end{aligned}$$

$$(V_{DS})_{\text{sat.}} = V_{GS} - V_t$$



Similarity

Triode \rightarrow vacuum
Voltage control
FET, MOSFET



V-I characteristic egn of MOSFET :

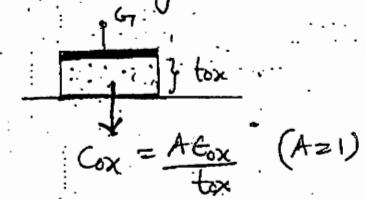
"NMOS"

$$I_D = \mu_n C_{ox} \left(\frac{C_0}{L} \right) \left\{ (V_{GS} - V_t) V_{DS} - \frac{1}{2} V_{DS}^2 \right\}$$

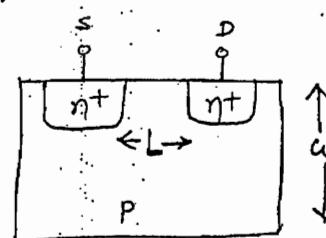
where, $\mu_n \rightarrow$ mobility of electrons

$C_{ox} \rightarrow$ capacitance of oxide layer

$$C_{ox} = \epsilon_{ox} / t_{ox}$$



$$\begin{aligned} & \& \epsilon_{ox} = 3.9 \epsilon_0 \\ & & = 3.9 \times 8.85 \times 10^{-12} \text{ F/m} \\ & & = 3.45 \times 10^{-11} \text{ F/m.} \end{aligned}$$



$(\frac{w}{L})$ \rightarrow aspect ratio

$\therefore (\frac{w}{L})$ is very less for modern mosfets.

$(V_{GS} - V_t) \rightarrow$ effective voltage (or) over drive voltage

- $V_{DS} \rightarrow$ drain to source voltage

Triode region:-

$$V_{DS} < V_{GS} - V_t \text{ , if } V_{DS} \text{ (mV)}$$

$$I_D = \mu_n C_{ox} \left(\frac{w}{L}\right) \cdot (V_{GS} - V_t) \cdot V_{DS}$$

$$R_{DS} = \frac{V_{DS}}{I_D} = \frac{1}{\mu_n \cdot \left(\frac{w}{L}\right) \cdot (V_{GS} - V_t)}$$

Current Saturation:

$$V_{DS} \geq V_{GS} - V_t$$

$$(V_{DS})_{sat} = V_{GS} - V_t$$

$$I_D = \frac{1}{2} \mu_n C_{ox} \left(\frac{w}{L}\right) \cdot (V_{GS} - V_t)^2$$

\downarrow
 K_n (Transconductance parameter) mA/V²

* g_m should be more, so K_n should be more,

$\mu_n \rightarrow$ faster, gain \rightarrow higher

* A good MOSFET should have high value of K_n .

$$I_D = K_n (V_{GS} - V_t)^2$$

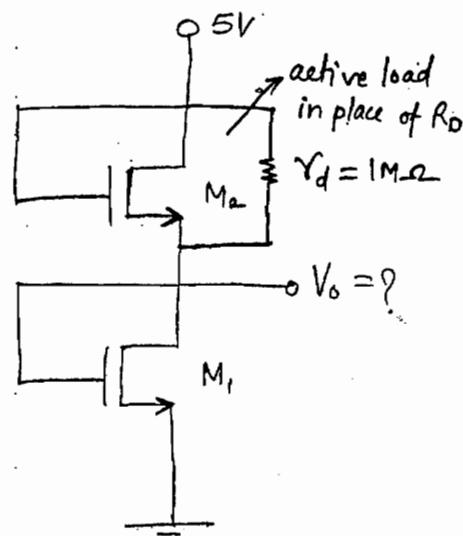
$$\therefore g_m = \frac{\partial I_D}{\partial V_{GS}} = 2K_n (V_{GS} - V_t)$$

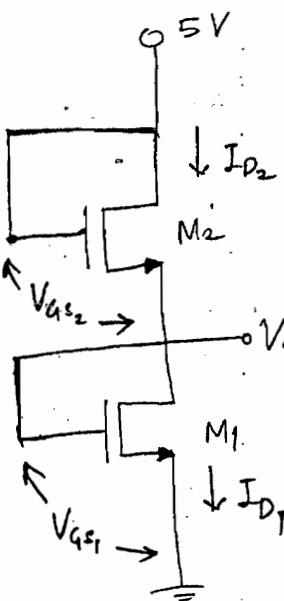
Active load MOSFET problems:

Q. given : $K_1 = 9 \text{ mA/V}^2$

$$K_2 = 36 \text{ mA/V}^2$$

$$V_t = 1 \text{ V}$$





$$I_{D_1} = I_{D_2}$$

$$K_1 (V_{GS_1} - V_t)^2 = K_2 (V_{GS_2} - V_t)^2$$

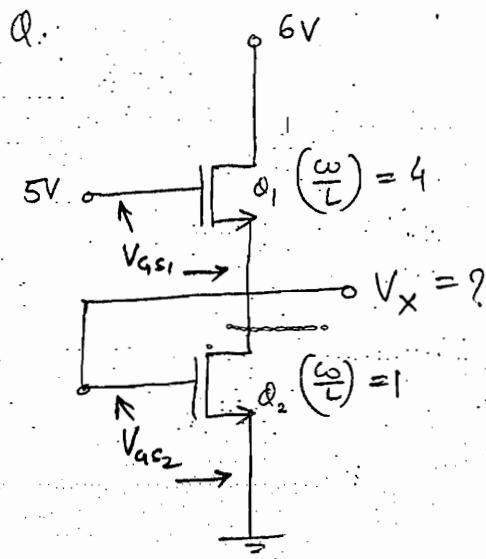
$$9 (V_o - 1)^2 = 36 (5 - V_o - 1)^2$$

$$V_o^2 + 1 - 2V_o = 4 (V_o^2 + 16 - 8V_o)$$

$$V_o^2 - 2V_o + 1 = 4V_o^2 - 32V_o + 64$$

$$3V_o^2 - 32V_o + 63 = 0$$

$$V_o = 3 \text{ Volts} \quad 7V(x)$$



$$Q_1 \& Q_2 \Rightarrow \mu_n C_{ox} = 100 \mu A/V^2$$

$$V_t = 1V$$

$$I_{D_1} = I_{D_2}$$

~~$$\frac{1}{2} \mu_n C_{ox} \left(\frac{\omega}{L}\right)_1 (V_{GS_1} - V_t)^2 =$$~~

~~$$\frac{1}{2} \mu_n C_{ox} \left(\frac{\omega}{L}\right)_2 (V_{GS_2} - V_t)^2$$~~

$$4 (5 - V_x - 1)^2 = (V_x - 1)^2$$

$$4 (V_x - 4)^2 = V_x^2 + 1 - 2V_x$$

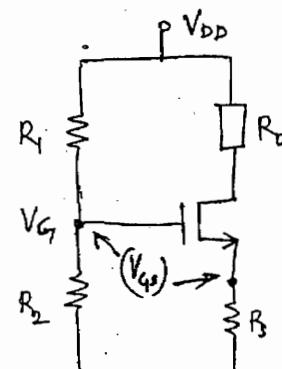
$$4V_x^2 + 64 - 32V_x = V_x^2 - 2V_x + 1$$

$$3V_x^2 - 30V_x + 63 = 0$$

$$\therefore V_x = 3 \text{ volts}$$

Biasing:

MOSFET Biasing:



$$I_D = k_n (V_{GS} - V_t)^2$$

↓
should be independent of
 k_n and V_t

$$V_G = V_{GS} + I_D R_S \Rightarrow V_G \approx I_D R_S$$

small \rightarrow neglected. \Rightarrow

$$(I_D)_Q = \frac{V_G}{R_S}$$

$$\& (V_{GS})_Q = \sqrt{\frac{(I_D)_Q}{k_n}} + V_t$$

- * This is the idea behind good biasing for MOSFET i.e. $(V_{GS} \approx 0)$.
- * A good MOSFET requirement is high value of k_n and low value of V_t .

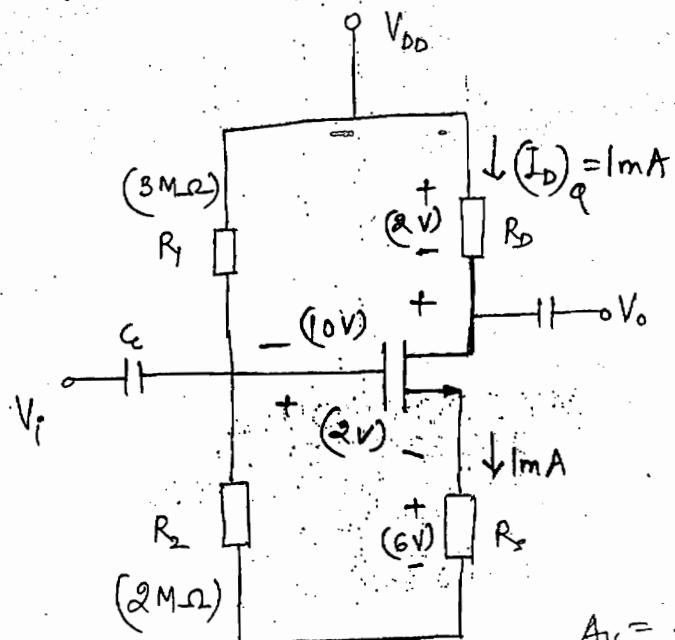
Q. Design an NMOS amplifier with given specifications:

$$V_{DD} = 20V, (I_D)_Q = 1mA, k_n = 1mA/V^2, V_t = 1V$$

$$(V_{DQ})_Q = 10V$$

Calculate the voltage gain of NMOS amplifier

Sol.



$$(V_{GS})_Q = \sqrt{\frac{I}{k_n}} + V_t = 2V$$

$$R_S = \frac{6V}{1mA} = 6K$$

$$R_3 = \frac{2V}{1mA} = 2K$$

$$V_G = V_{DD} \cdot \frac{R_2}{R_1 + R_2}$$

$$\frac{R_2}{R_1 + R_2} = \frac{8}{20} = \frac{2}{5}$$

$$A_V = -g_m R_D$$

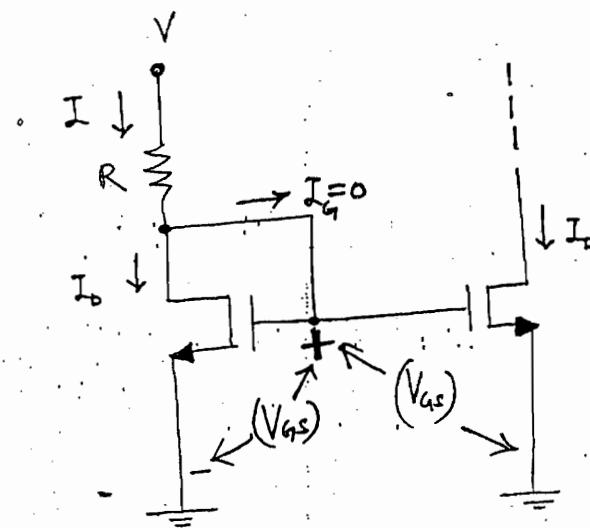
$$g_m = 2k_n (V_{GS} - V_t)^2$$

$$= 2(1)(2-1)^2$$

$$= 2 mS$$

$$A_V = -2 \times 2 = -4$$

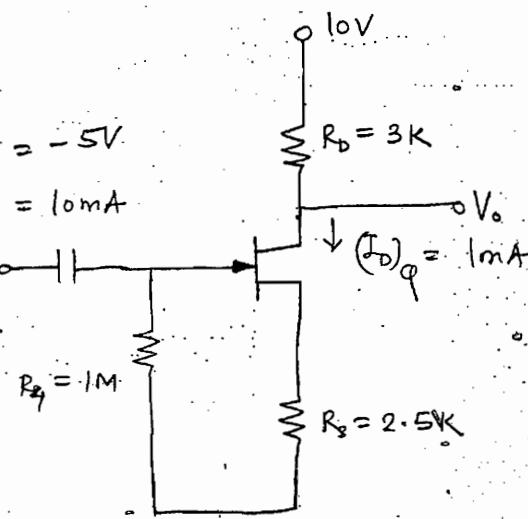
Current mirror biasing :



Q11.

$$V_p = -5V$$

$$I_{DSS} = 10mA$$



$$(V_{GS})_Q = -(I_D)_Q \cdot R_f$$

$$= -1mA \cdot 2.5k\Omega$$

$$= -2.5$$

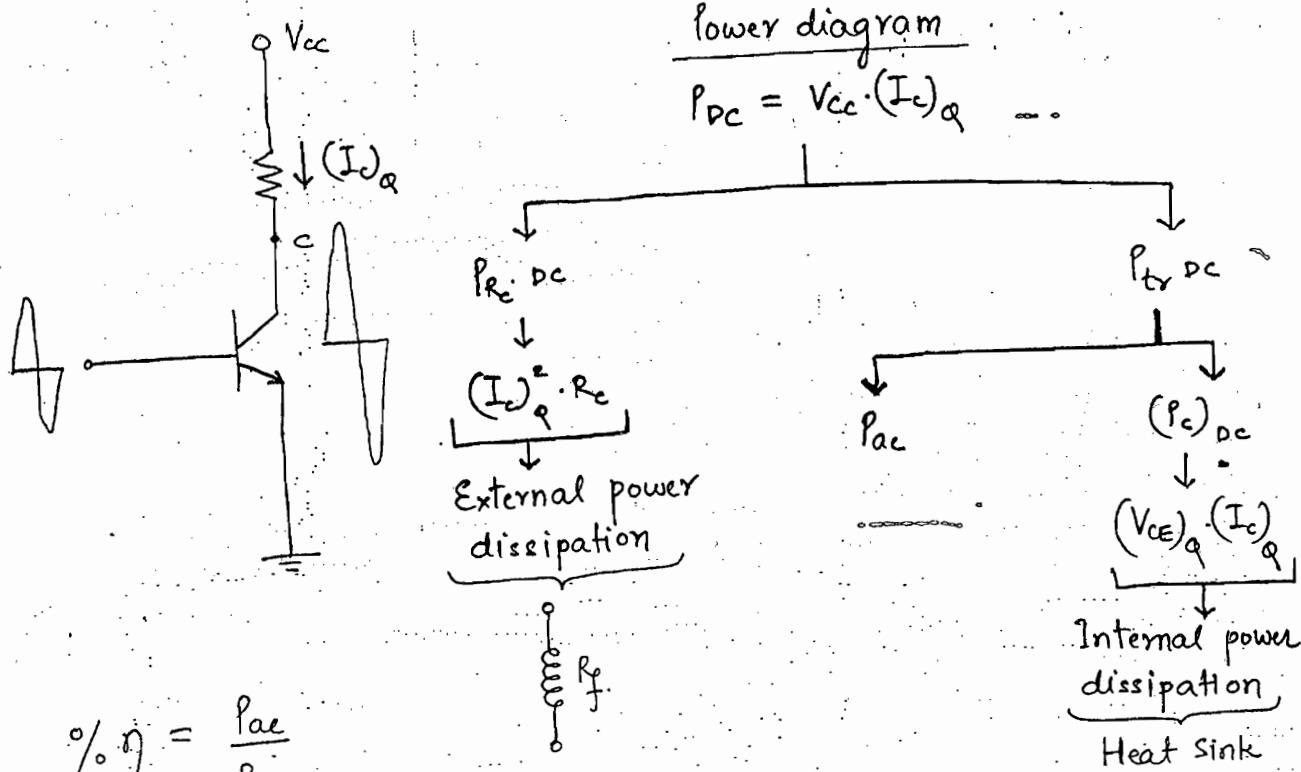
$$g_m = -\frac{2I_{DSS}}{V_p} \left(1 - \frac{(V_{GS})_Q}{V_p} \right)$$

$$= -\frac{2 \times 10}{-5} \left[1 + \frac{2.5}{5} \right]$$

$$= 2$$

$$A_v = g_m R_L = -2 \times 3k\Omega = -6$$

Power Amplifiers



$$P_{DC} = P_{ac} + \text{losses}$$

$$P_{DC} = P_{ac} + P_D$$

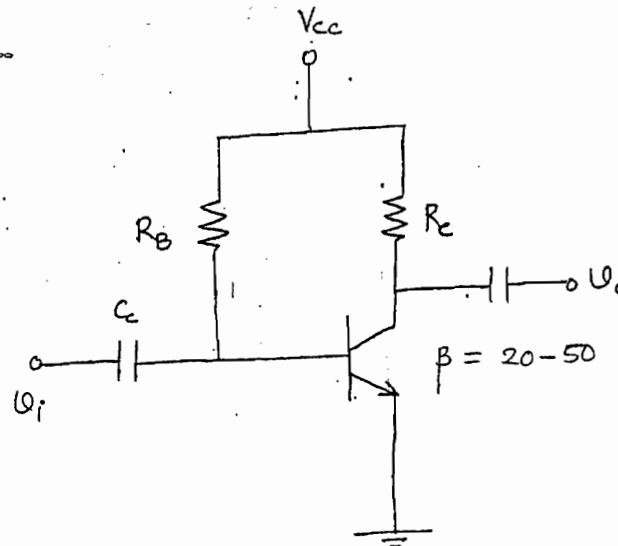
$$\text{If } P_D \rightarrow 0 \Rightarrow \eta \rightarrow 100\%$$

Power amplifier is a device which converts DC power to ac power and whose action is controlled by ac e/p sig.

Class A amplifier:

- (1). Class A series fed power amplifier
- (2). Class A \$X^*\$ coupled power amplifier

(1) Class A series fed power amplifier:



Base width is high, so recombination high, I_B is high
so B is less.

$$\text{DC I/p} \rightarrow P_{dc} = V_{cc} \cdot (I_c)_Q$$

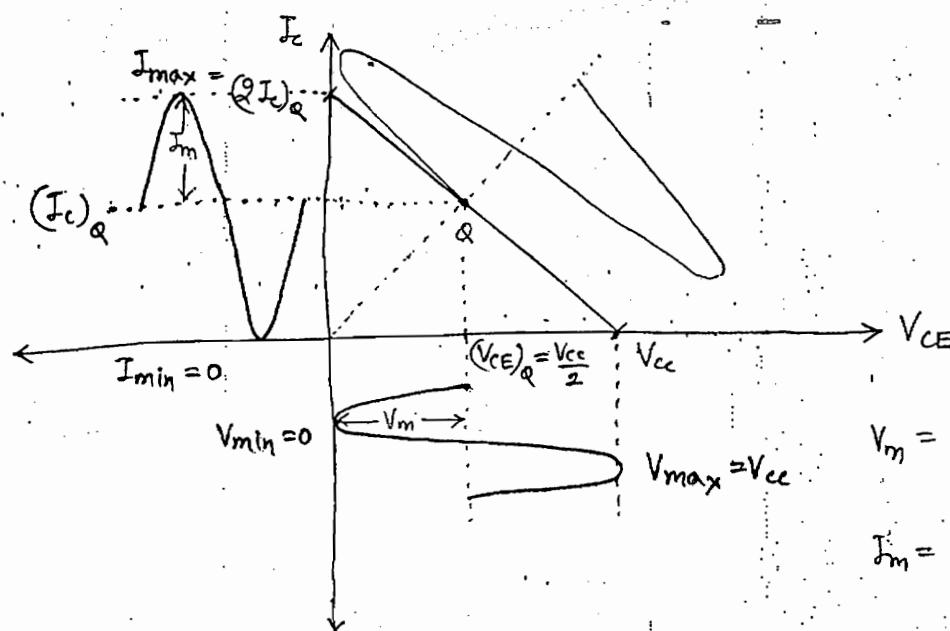
$$\begin{aligned}\text{AC o/p} \rightarrow P_{ac} &= V_{m\text{re}} \cdot I_{m\text{re}} \\ &= \frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}} \\ &= \frac{V_m I_m}{2}\end{aligned}$$

DC conditions:

$$(I_B)_Q = \frac{V_{cc} - V_{BE}}{R_b}$$

$$(I_c)_Q = \beta (I_B)_Q$$

$$(V_{CE})_Q = V_{cc} - (I_c)_Q \cdot R_e$$

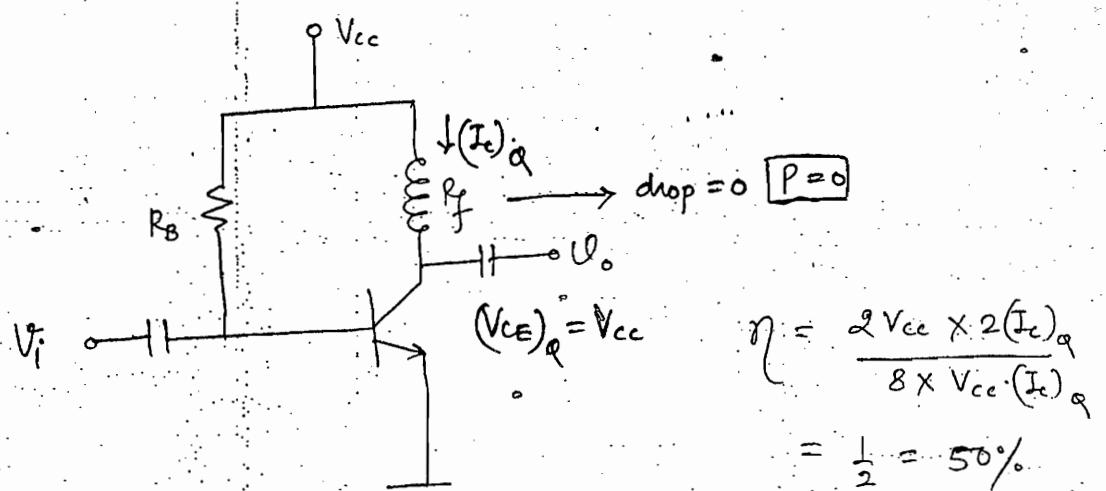


$$\eta = \frac{P_{ac}}{P_{dc}} = \frac{1}{8} \cdot \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{V_{ce} \cdot (I_c)_Q}$$

$$= \frac{1}{8} \cdot \frac{(V_{ce} - 0)[(2 I_c)_Q - 0]}{V_{ce} \cdot (I_c)_Q}$$

$$= \frac{1}{8} \cdot 2 = \frac{1}{4} = 25\%$$

(2). X^r coupled:



Conclusion:

In class A series fed power amplifier the efficiency maxm is 25%. So improve this efficiency, replace the resistor R_B with an inductor L_f where the efficiency goes to 50%.

Power dissipation -

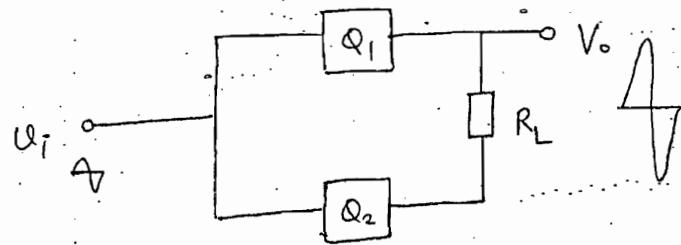
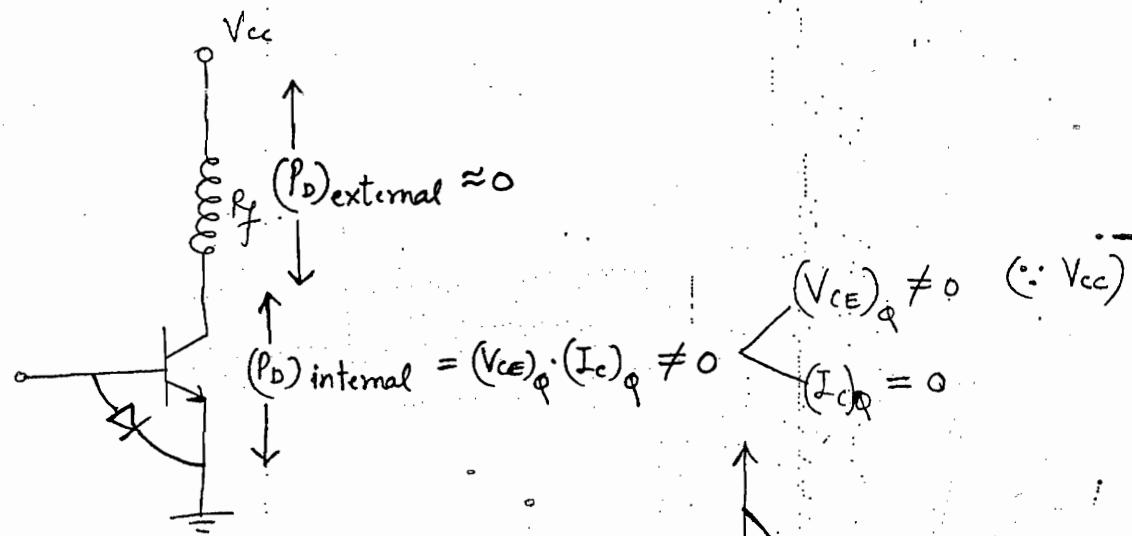
$$P_{dc} = P_{ac} + P_D$$

$$P_D = P_{dc} - P_{ac}$$

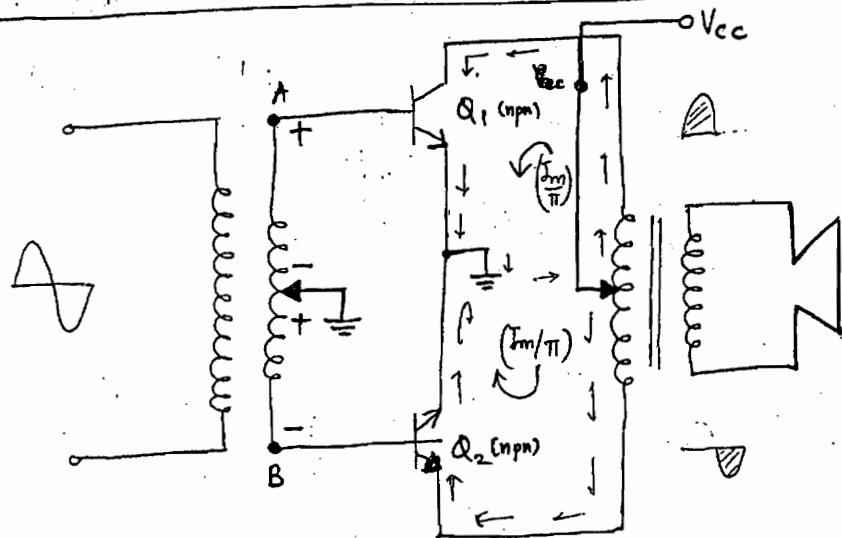
$$= V_{ce} \cdot (I_c)_Q - \frac{V_m I_m}{2}$$

At zero sig conditions, $P_D = P_{dc}$ ($\because P_{ac} = 0$)

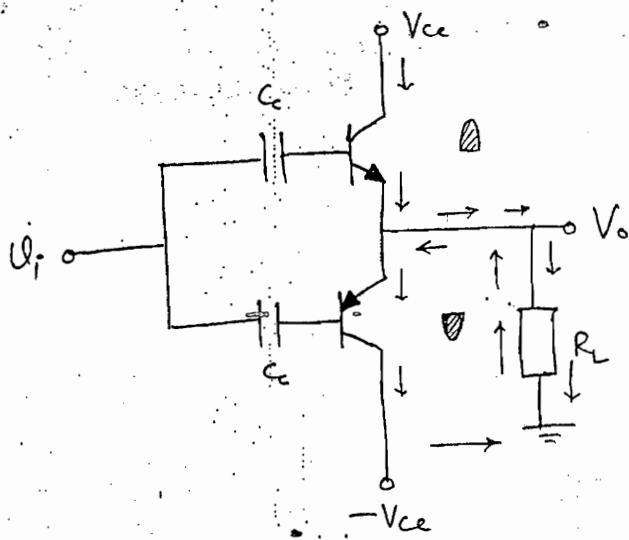
Class B power amplifiers :



Push pull class B power amplifier : (single Battery)



Complementary symmetry class B : (2 Batteries)



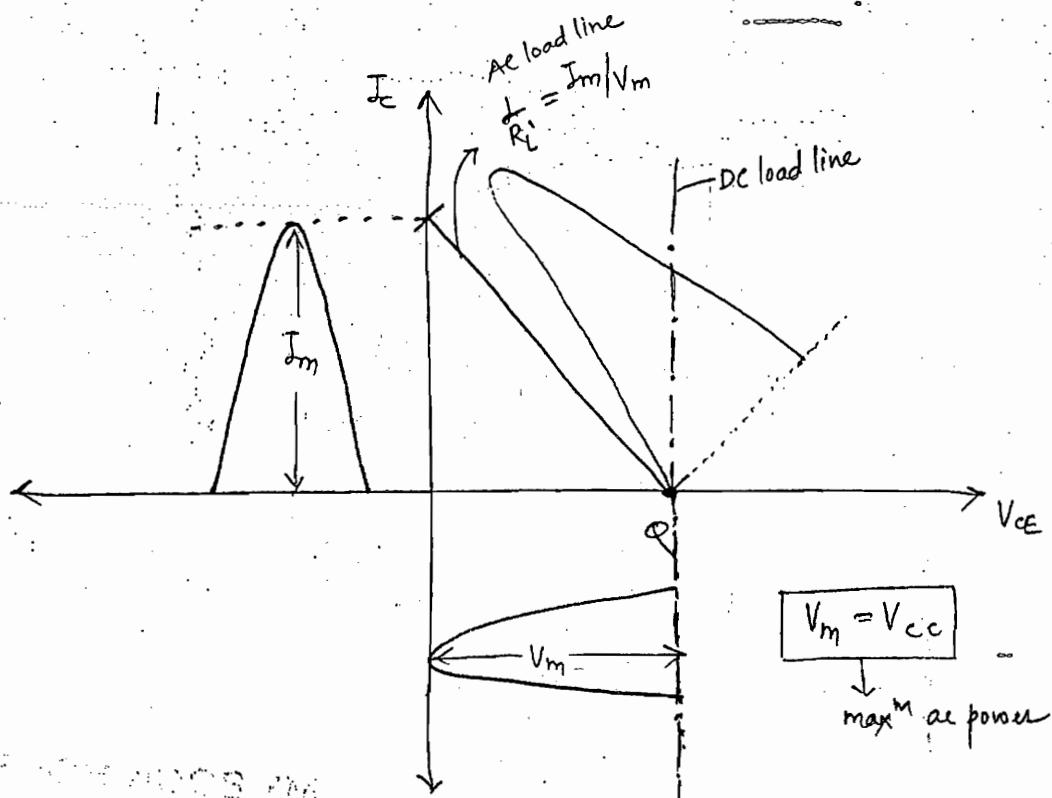
$$(I_c)_Q = 0$$

$$P_{DC} = V_{ce} \cdot I_{dc}$$

$$= V_{ce} \left(\frac{2I_m}{\pi} \right)$$

Efficiency : $P_{ac} = \frac{V_m I_m}{2}$

$$\% \eta = \frac{P_{ac}}{P_{DC}} = \frac{V_m I_m / 2}{V_{ce} \cdot 2I_m / \pi} = \frac{\pi}{4 \cdot V_{ce}}$$



$$\% \eta = \frac{V_{ce} \cdot \pi}{4 \cdot V_{ce}} = \frac{\pi}{4} = 78.5\%$$

Power dissipation in class B :

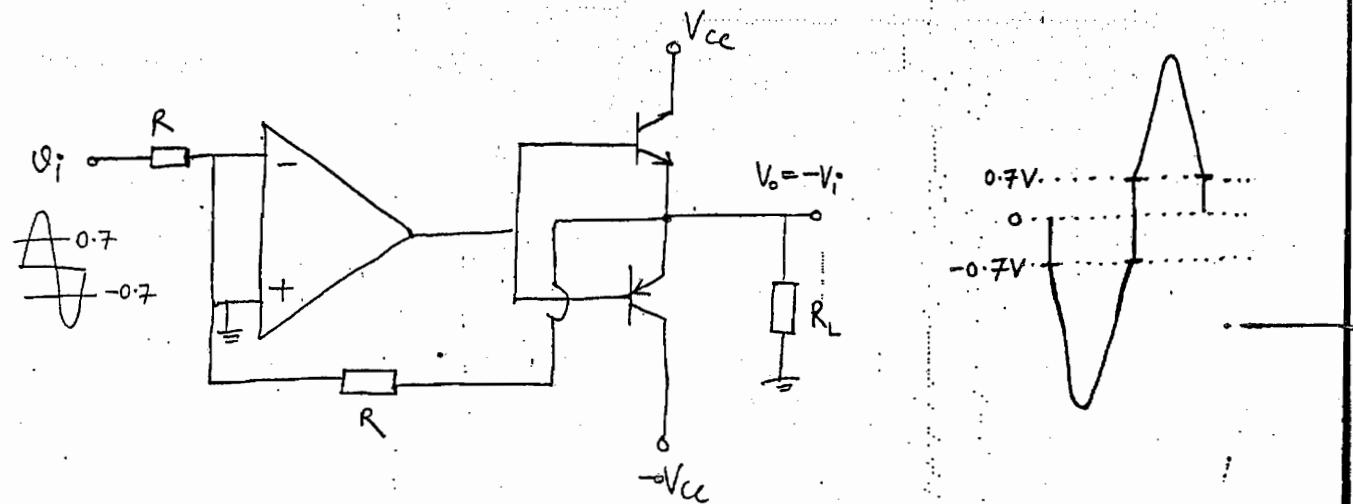
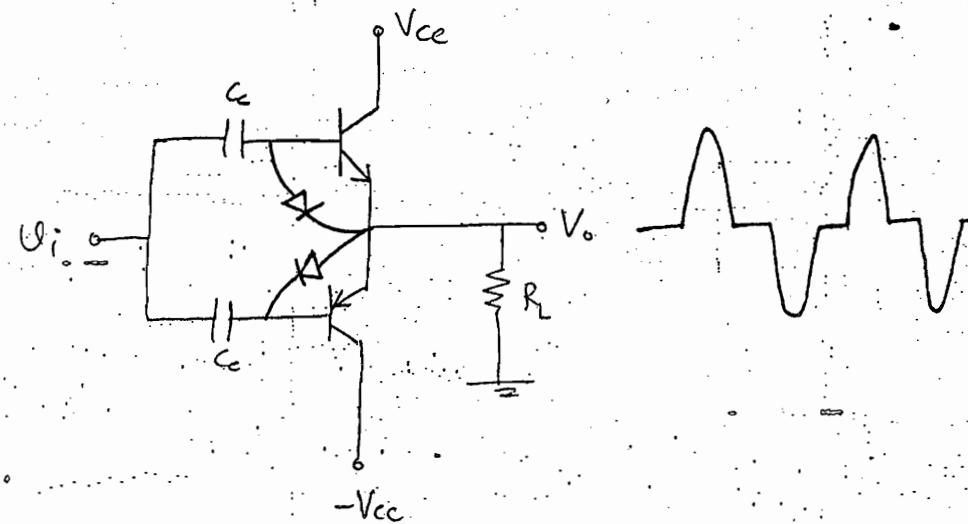
$$P_D = P_{DC} - P_{AC}$$

$$= V_{CE} \cdot \frac{2 I_m}{\pi} - \frac{V_m I_m}{2}$$

At s/g conditions, $V_m = 0, I_m = 0$.

$P_D = 0$ (Sleeping mode)

Cross over distortion:



Oscillators

$$V_{O1} = \frac{2(1-\alpha) S^2 R^2 C^2}{S^2 R^2 C^2 + 2\alpha SRC + 1} \quad (\text{H.P.})$$

$$V_i = S^2 R^2 C^2 + 2\alpha SRC + 1$$

$$V_{O2} = -\frac{2(1-\alpha) SRC}{S^2 R^2 C^2 + 2\alpha SRC + 1} \quad (\text{B.P.})$$

$$V_i = S^2 R^2 C^2 + 2\alpha SRC + 1$$

$$V_{O3} = \frac{2(1-\alpha)}{S^2 R^2 C^2 + 2\alpha SRC + 1} \quad (\text{L.P.})$$

$$V_i = S^2 R^2 C^2 + 2\alpha SRC + 1$$

Band-pass \Rightarrow

$$V_{O2} = -\frac{2(1-\alpha) SRC}{S^2 R^2 C^2 + 2\alpha SRC + 1}$$

$$V_i = \frac{-2(1-\alpha) SRC}{S^2 R^2 C^2 + 2\alpha SRC + 1}$$

$$= -\frac{2(1-\alpha) SRC}{S^2 R^2 C^2 + 2\alpha SRC + 1}$$

$$= -\frac{2(1-\alpha) SRC}{-\omega^2 R^2 C^2 + 2\alpha SRC + 1}$$

$$\left| \begin{array}{l} \omega_0 = 1 \\ RC \end{array} \right.$$

$$= -\frac{2(1-\alpha) SRC}{2\alpha SRC}$$

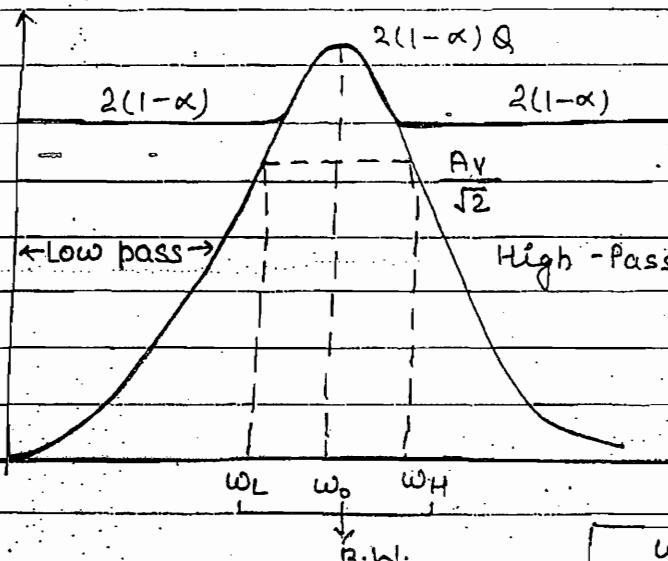
$$= \frac{1-\alpha}{\alpha} = -\frac{2(1-\alpha) Q}{2\alpha}$$

$$Q \rightarrow \text{Quality factor} ; \quad \left\{ Q = \frac{1}{2\alpha} \right\} \text{ pole } Q$$

$$\omega_0 = \frac{1}{RC} \quad \dots \text{Pole Frequency}$$

$$\begin{aligned} V_{O1} &= \frac{2(1-\alpha) s^2}{\omega_0^2} & V_{O3} &= \frac{2(1-\alpha)}{\omega_0^2 + \frac{s}{\omega_0 Q} + 1} \\ V_i & & V_i & \end{aligned}$$

$$\frac{s^2}{\omega_0^2} + \frac{s}{\omega_0 Q} + 1$$



$$\frac{\omega_0}{B.W.} = Q$$

Pole-Equation \Rightarrow

$$\frac{s^2}{\omega_0^2} + \frac{s}{\omega_0 Q} + 1 = 0 \quad \text{Assume } \frac{s}{\omega_0} = x$$

$$\rightarrow x^2 + \frac{x}{Q} + 1 = 0$$

$$\frac{s_{1,2}}{\omega_0} = -\frac{1}{Q} \pm \sqrt{\left(\frac{1}{Q}\right)^2 - 4(1)x(1)}$$

$$2x1$$

$$\frac{s_{1,2}}{\omega_0} = -\frac{1}{Q} \pm \sqrt{\frac{1}{Q^2} - 4}$$

$$\text{real-part} \Rightarrow 1 > 4, \quad Q < 1 ; \quad Q < \frac{1}{2}$$

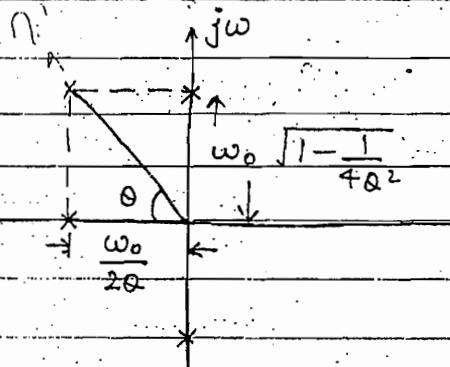
Imaginary part \Rightarrow

$$Q > 1/2$$

$$\Rightarrow S_{1,2} = -\frac{i}{\omega_0} \pm j \sqrt{\frac{4-1}{Q^2}}$$

$$\Rightarrow S_{1,2} = -\frac{i}{\omega_0} \pm j \sqrt{\frac{1-1}{4Q^2}}$$

$$\Rightarrow S_{1,2} = -\frac{\omega_0}{2Q} \pm \frac{\omega_0 j \sqrt{1-1}}{4Q^2}$$



$$\theta = \tan^{-1} \frac{\omega_0 \sqrt{1-1/4Q^2}}{\omega_0/2Q}$$

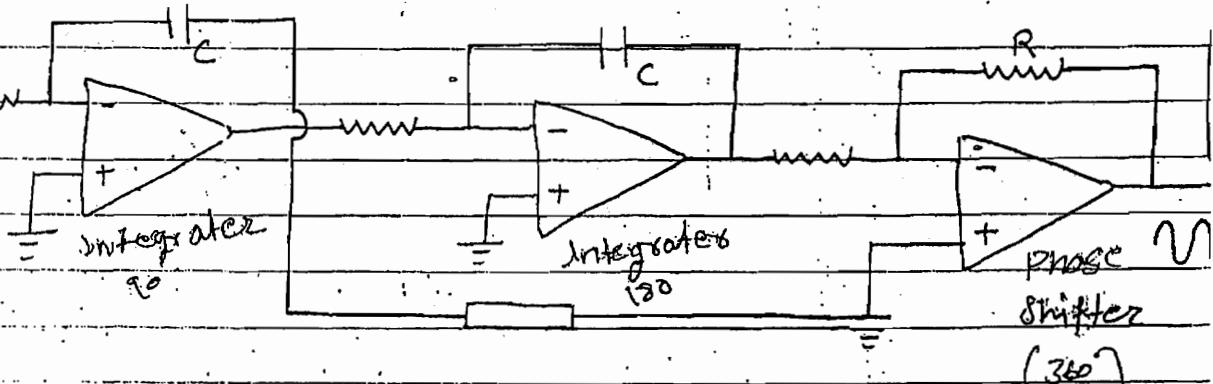
$$\theta = \tan^{-1} 2Q \sqrt{1-1/4Q^2}$$

$$Q \rightarrow \infty$$

$$\theta = \tan^{-1} \infty$$

$$[\theta = 90^\circ]$$

$$Q = \frac{1}{2\alpha} = \infty, \quad (\alpha = 0)$$



Harmonic Oscillator

(OR)

Quadrature Oscillator [* Should be zero then Pole will come on imaginary axis]

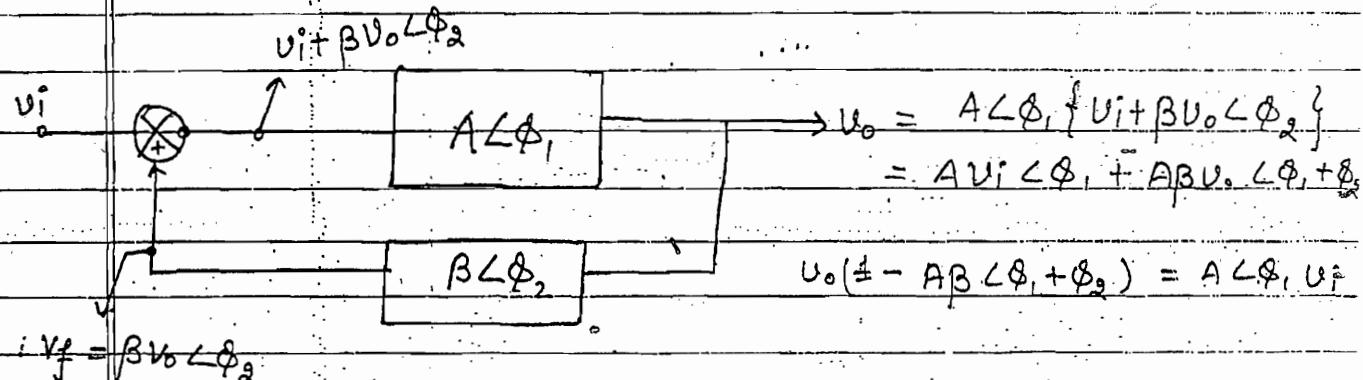
$$\frac{\partial^2 v_o}{\partial t^2} + \alpha \frac{\partial v_o}{\partial t} + \gamma v_o = v_i$$

\downarrow

$$\frac{\partial^2 v_o}{\partial t^2} + \gamma v_o = 0$$

harmonic oscillators :

* Positive feedback concept \Rightarrow



$$v_o = A \angle \phi_1 \quad \therefore 1 - AB \angle \phi_1 + \phi_2 = 0$$

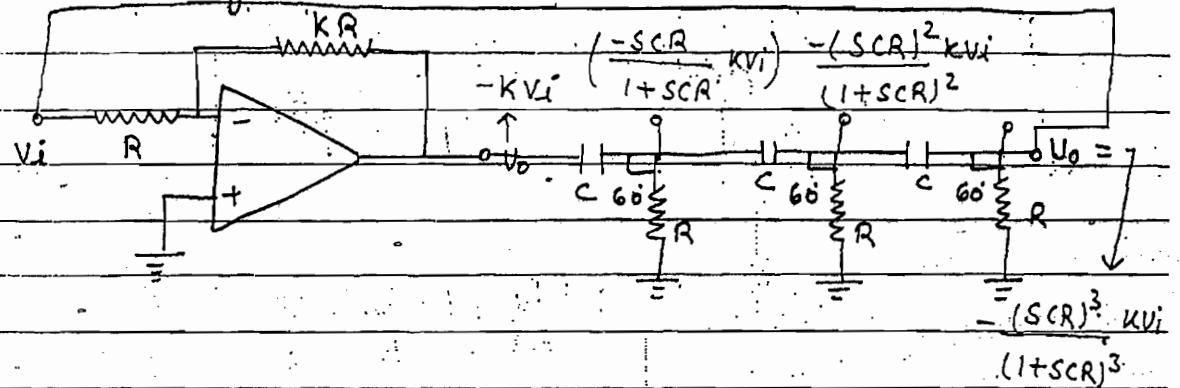
$$v_i \quad 1 - AB \angle \phi_1 + \phi_2$$

conditions $\Rightarrow |AB| = 1$

$$A = \frac{v_o}{v_i}, \quad B = \frac{v_f}{v_o}$$

$$AB = \frac{v_o}{v_i} \cdot \frac{v_f}{v_o} = \frac{v_f}{v_i} = 1 \quad [v_f = v_i]$$

$\angle \phi_1 + \phi_2 = 0^\circ$ (or) multiples of 2π .
Barkhausen criteria.

* RC-Phase Shift Oscillators \Rightarrow


$$-\frac{(SCR)^3}{(1+SCR)^3} KVi = V_o$$

$$\rightarrow (1+SCR)^3 + (SCR)^3 K = 0$$

$$\rightarrow 1 + 3s^2C^2R^2 + 3SCR + (SCR)^3(K+1) = 0$$

$$\rightarrow 1 - 3\omega^2C^2R^2 + 3j\omega CR - j\omega^3R^3C^3(K+1) = 0$$

$$1 - 3\omega^2C^2R^2 = 0 \quad (\text{Real part})$$

$$\left\{ \begin{array}{l} \omega_0 = 1 \\ \sqrt{3}RC \end{array} \right.$$

$$3\omega_0CR - \omega_0^3R^3C^3(K+1) = 0$$

$$\rightarrow 1 - \omega_0^2R^2C^2(K+1) = 0 \quad \text{at gain} = 8 \text{ will get}$$

$$\rightarrow \omega_0 = \frac{1}{\sqrt{3}RC} \quad \text{on img. axis/pulse}$$

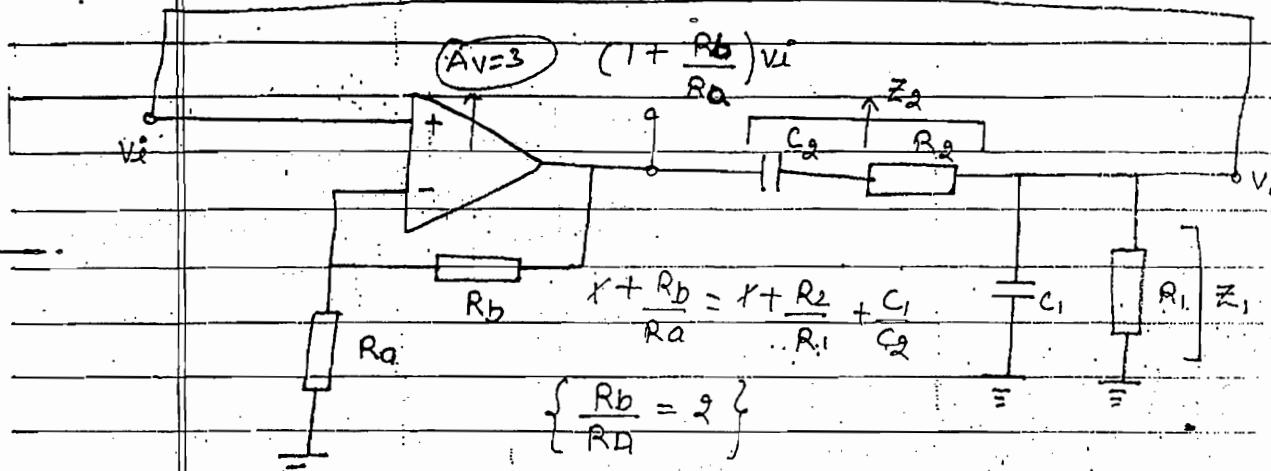
$$1 - 1/(K+1) = 0 \quad , \quad K = 8$$

Conclusions \Rightarrow

- RC-phase shift osci. are used for audio freq. ranges (20 Hz - 20 kHz).
- Quality factor for ω_0/ω will be in multiples of 10.
- RC-phase shift Oscill. can't be design for dynamic range of frequency.

* Wein-Bridge Oscillator \Rightarrow

$$\left(1 + \frac{R_b}{R_a} \right) \left(1 + \frac{R_2}{R_1} + \frac{C_1}{C_2} \right) = 1$$



$$V_o = \left(1 + \frac{R_b}{R_a} \right) V_i - Z_1 = \left(1 + \frac{R_b}{R_a} \right) V_i - \frac{Z_1}{1 + Z_2 Y_1}$$

$$V_o = \left(1 + \frac{R_b}{R_a} \right) V_i \Rightarrow \left(1 + \frac{R_b}{R_a} \right) \cdot \frac{1 + (R_2 + \frac{1}{C_2}) (\frac{1}{R_1} + sC_1)}{1 + R_2 + C_1 + \frac{1}{C_2 R_1} + sC_1 R_2}$$

$$V_o = \left(1 + \frac{R_b}{R_a} \right) V_i$$

$$\frac{1 + R_2}{R_1} + \frac{C_1}{C_2} + j \left[\frac{-1}{w C_2 R_1} + w C_1 R_2 \right]$$

$$j = \text{Imaginary} \Rightarrow$$

$$-1 + w C_1 R_2 = 0$$

$$w C_2 R_1$$

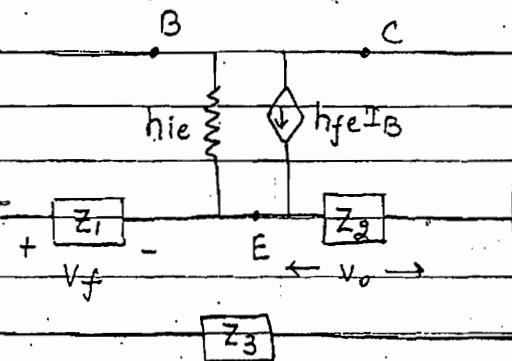
$$\omega_0 = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$R_1 = R_2 = R, \quad C_1 = C_2 = C$$

$$\omega_0 = \frac{1}{RC}$$

$$f_0 = \frac{1}{2\pi R C}$$

3). LC - Oscillators \Rightarrow



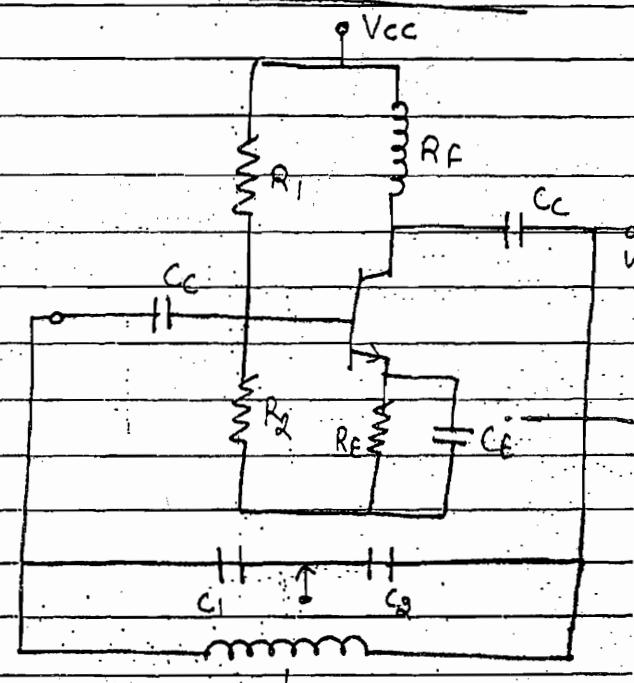
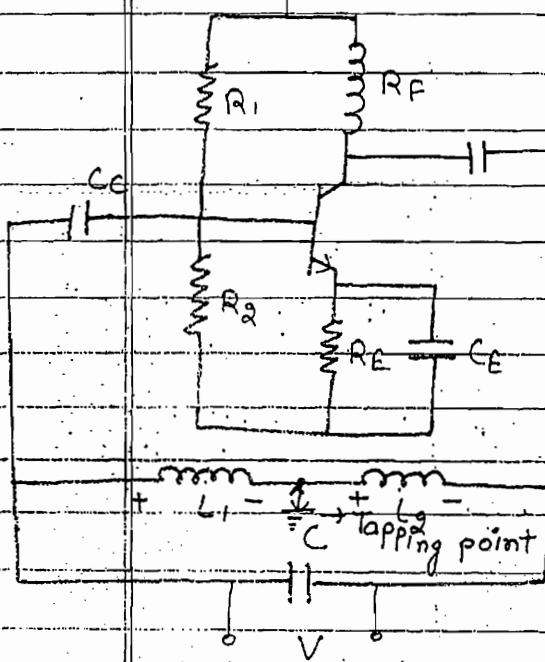
Condition:

Z_1 & Z_2 are same reactive components. Z_3 is opposite compo. (reactive only).

Colpitts Oscill.

Harley

Oscill.



Analysis of LC - oscillators \Rightarrow

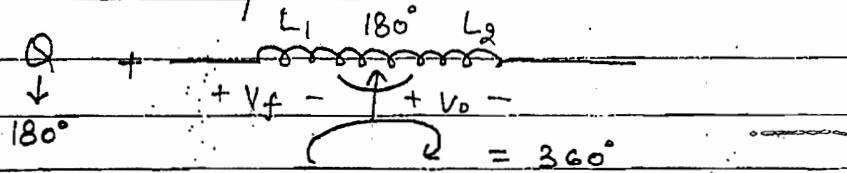
Q:1 what is the func. of inductor (RF) in LC oscili. ckt?

Ans: inductor RF is used as an isolation b/w dc-supply & ac-signal.

Q:2 Explain phase relation in LC oscili.?

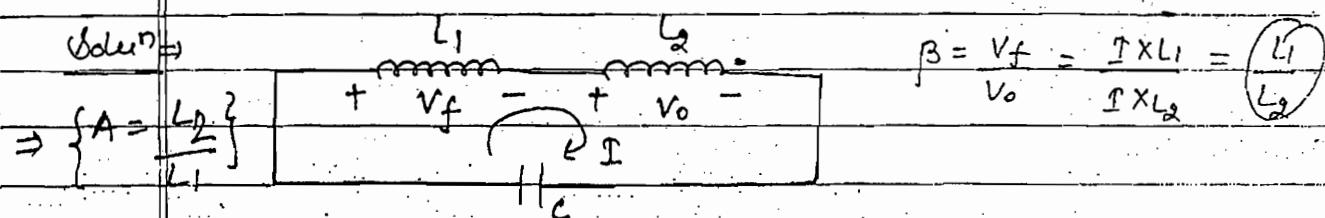
Ans:

Soluⁿ \Rightarrow Hartley \Rightarrow



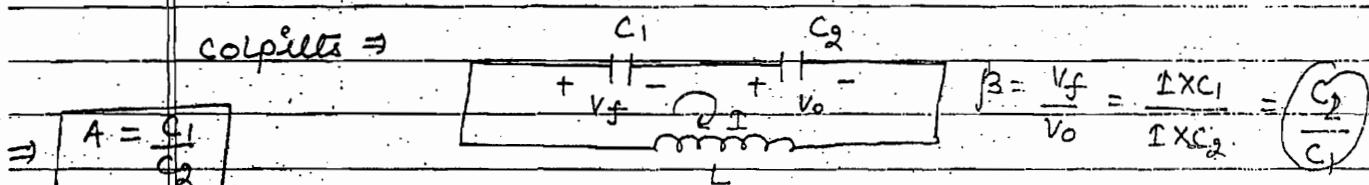
Ques: 3 Explain conditions for gain A and β for Hartley osci.

Soluⁿ \Rightarrow



$$\beta = \frac{V_f}{V_0} = \frac{I \times L_1}{I \times L_2} = \frac{L_1}{L_2}$$

Colpitts \Rightarrow



$$\beta = \frac{V_f}{V_0} = \frac{I \times C_1}{I \times C_2} = \frac{C_1}{C_2}$$

Ques: 3 Explain expressions for freq. of oscil. in LC circ?

Soluⁿ \Rightarrow $f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{L_T C}}$

$$L_T = L_1 + L_2 + 2m, \quad m \rightarrow \text{negligible}$$

$$L_T = L_1 + L_2$$

Colpitts \Rightarrow $f = \frac{1}{2\pi\sqrt{L_C C_T}}, \quad \frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2}$

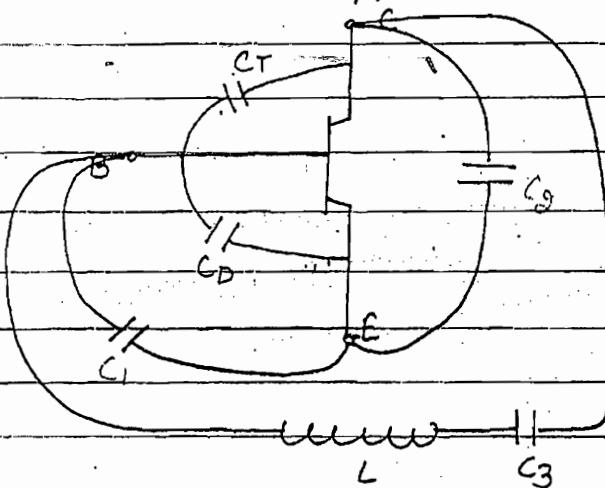
Ques: why LC oscilla. can't be use for low freq. of oscilla.

Soluⁿ \Rightarrow $f = \frac{1}{2\pi\sqrt{LC}}$; $f \rightarrow$ less, LC compo. become high.

* RC never use for high freq.

Ques Explain about Clapp oscillator design?

Soln =



$$f = \frac{1}{2\pi\sqrt{LC_T}}$$

$$C_T = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

$$C_1 \gg C_3, C_2 \gg C_3$$

$$\frac{1}{C_1} = \frac{1}{C_3}$$

C_1, C_2 are getting effect by high

freq. [C_D and C_T Problem]

but C_3 is free from them

$$\left\{ \begin{array}{l} f = \frac{1}{2\pi\sqrt{LC_3}} \end{array} \right.$$

- A quality factor of an LC-rlw is in multiples of 100. This LC oscill. can stabilized the freq. upto 5MHz.

- modify version of colpitts is called clapp oscillator which can stabilized freq. upto 10 MHz.

- To stabilized freq. above 10 MHz crystal design will play a role which is known as piezo electric effect principle. When a mechanical force is applied one face of crystal, a electrical signal will be generated on other face of crystal and vice versa. Typically Q (quality factor) of crystal will be around 20,000 (multiples of 1000).

FET / MOSFET

- FET Biasing
- FET amplifiers

O.P. Rapp

- differential Amplifiers
- op amp application