 ANUPAM SHUKLA Rs.60

NAME:-

SUBJECT:- Analog Electronics – [E.E]

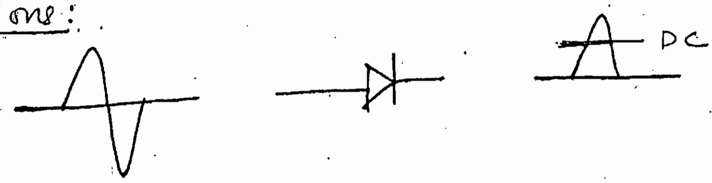
INSTITUTE:-

ANALOG ELECTRONICS

ANUPAM SHUKLA

Diodes: Applications:

(1). Rectifiers:



(2). Filter :

(3). Voltage Regulator:

} DC power supply

(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) — Low freq. Analysis; High freq. Analysis; Frequency Response
(4). Large s/g Amplifiers (Power amplifiers)
(5). Feedback theory
Negative feedback theory (Amplifiers)

Positive feedback theory (oscillators)

Integrated theory (op Amps):

(1). Multistage Amplifiers

(i) · Effect of cascading on Bandwidth.

(ii) Important cascading designs.

(a). Cascode Amplifier (CE-CB)

(OR)

Wide Band Amplifier

(b). Darlington pair / high i/p impedence (CC-CC)

(2). Coupling Techniques:

(i). RC coupling

(ii). Direct coupling

(3). Differential Amplifier

(4). applications of op-Amp

FET / MOSFET:

(1) FET device
(2). FET Biasing
(3). FET Amplifiers.

$\Big\}$ FET

MOSFET.

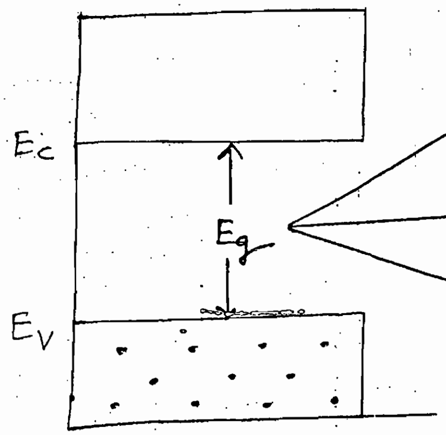
(1). MOSFET device

(2). MOSFET Biasing

(3). MOSFET Amplifiers

Introduction to Electronics :

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



0 eV (Conductors)

1 eV (Semiconductors)

6 eV (Insulators)

At $T = 0\,K\ (-273°C)$

Ge (0.785 eV) ✓

Si (1.21 eV) ✓

GaAs (1.58 eV) ✓

C (2 eV) ✗

At Room temperature 27°C (300 K) :

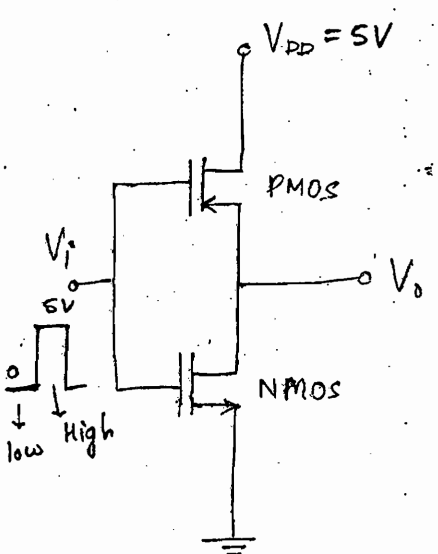
C is a bad semi-conductor.

A good semi-conductor must conduct at room temp.

A. The 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 ?

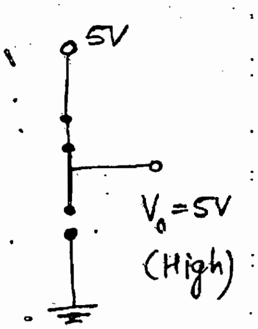
CMOS



$V_i = 0$ (Low)

PMOS → ON

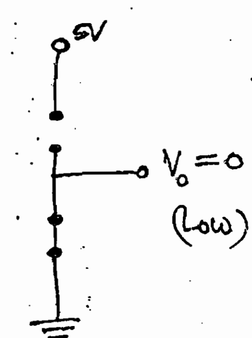
NMOS → OFF



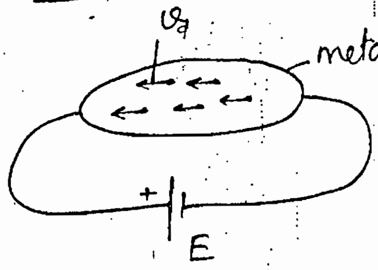
$V_i = 5V$ (High)

PMOS → OFF

NMOS → ON



Mobility :



$V_d \propto E$

$v_d = \mu E$

Mobility, $\mu = \dfrac{\text{Drift velocity}}{\text{Electric field}}$ $(m^2/V\,sec)$

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

$\mu_e\,(Si) = 1300\ cm^2/V\cdot sec$

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

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

Temperature with standing capability

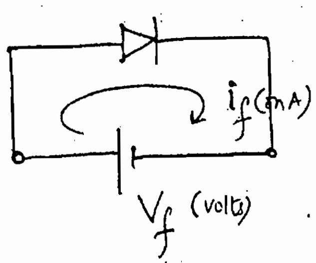
$T\,(Ge) \approx 100°c$

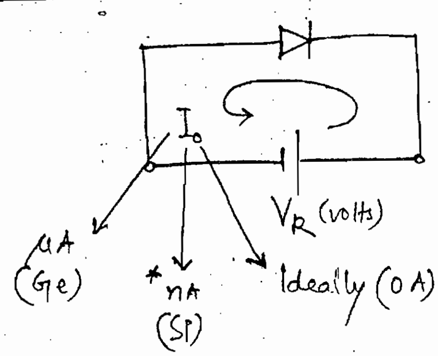
$T\,(Si) \approx 200°c$

* $T\,(GaAs) \geq 200°c$

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

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





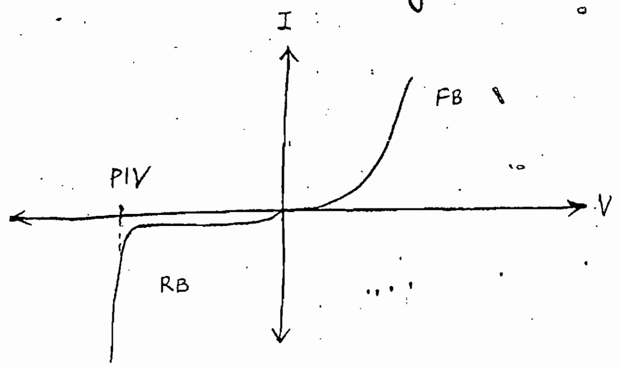
(2). Temperature with standing capability :

Power dissipation (Heating effect)

$T_{Ge} \approx 100°c$

* $T_{Si} \approx 200°c$

(3). Peak Inverse Voltage (PIV):



* PIV (Si) ≈ 1000 V

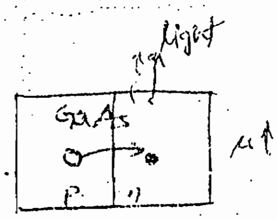
PIV (Ge) ≈ 400 V

Q. Give the important properties of GaAs ?

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

(2). μ_e (GaAs) → 8,500 cm^2/V-sec.

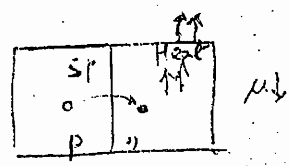
μ_h (GaAs) → 400 cm^2/V-sec



for Si :

μ_e (Si) → 1300 cm^2/V-sec

μ_h (Si) → 500 cm^2/V-sec



for Ge :

μ_e (Ge) → 3,800 cm^2/V-sec

μ_h (Ge) → 1,800 cm^2/V-sec

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

Si & Ge are best examples for indirect band gap.

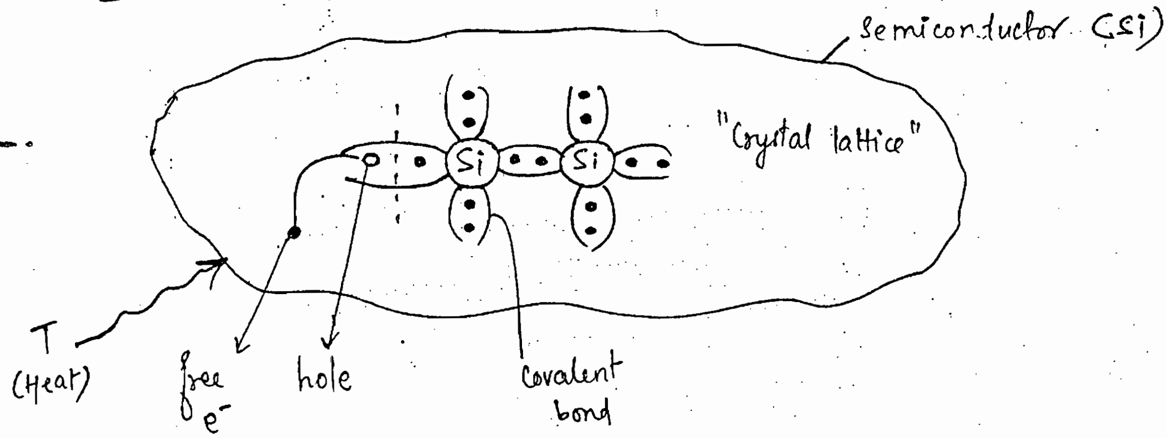
(4). GaAs is a compound semiconductor.

$$\underbrace{\underset{\text{III group}}{Ga} + \underset{\text{V group}}{As}}_{\text{IV group}}$$

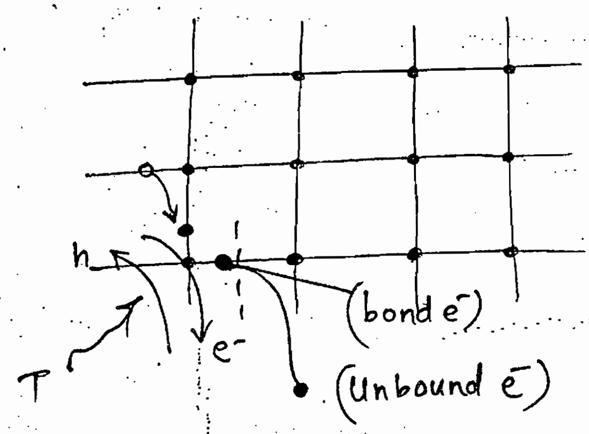
(5). GaAs is used in optics.

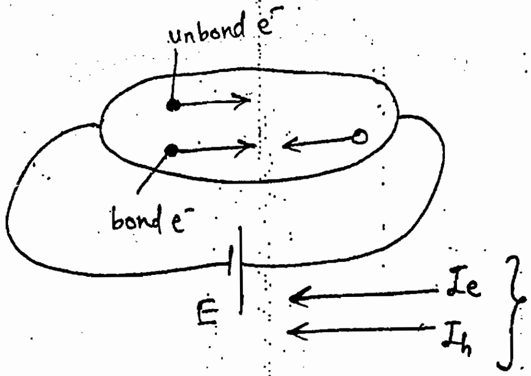
Q. Why mobility of e^- > mobility of holes . ?

Hole concept :



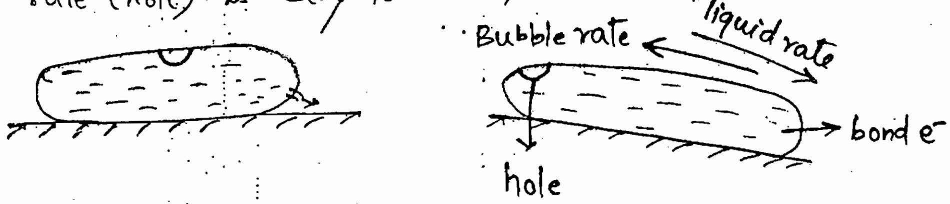
Crystal Lattice :





$$\left.\begin{matrix} I_e \\ I_h \end{matrix}\right\} I_{sc} = I_e + I_h$$

Eg: Glass-Bubble → 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).

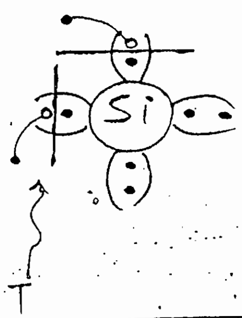
Q. Give the classification of semiconductor :

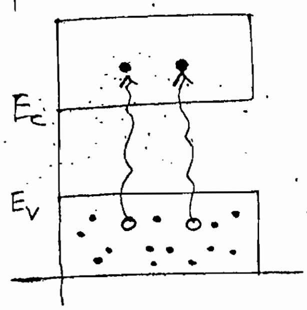
Semiconductor

- Intrinsic sc (pure)
- Extrinsic sc (Impure)

$\boxed{\text{Intrinsic sc}} + \boxed{\text{other material}} \Rightarrow \boxed{\text{Extrinsic sc}}$

other material:
- Pentavalent impurity (Eg. Arsenic, Antimony, Bismuth, phosphorus etc.)
- Trivalent impurity (Eg. Gallium, Indium, Boron, Aluminium etc)





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

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

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

$n_i = p = n$

$n_i \propto$ temp

Si :

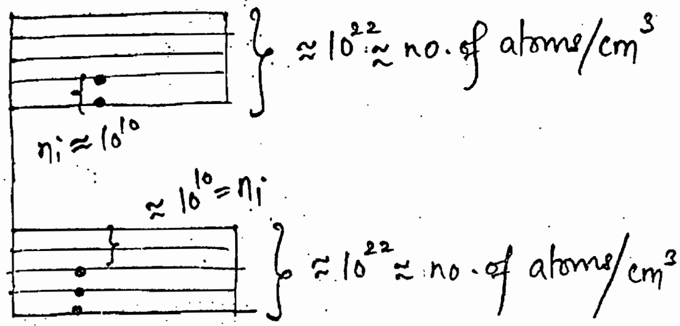
$n_i\,(300\,K) \rightarrow 1.5 \times 10^{10}/cm^3$

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

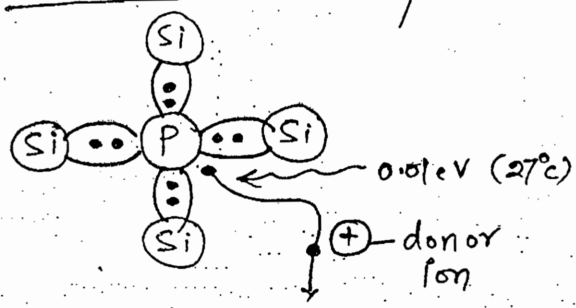
Ge :

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

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



Pentavalent Impurity:



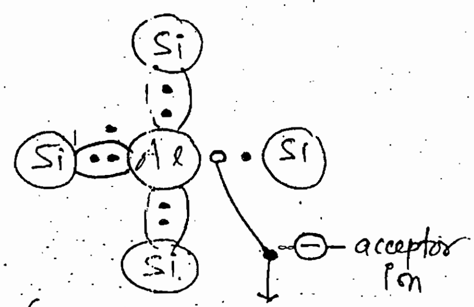
$(n > p)$

Majority carriers → Electrons

Minority carriers → holes

N-type semi-conductor

Trivalent Impurity:



$(p > n)$

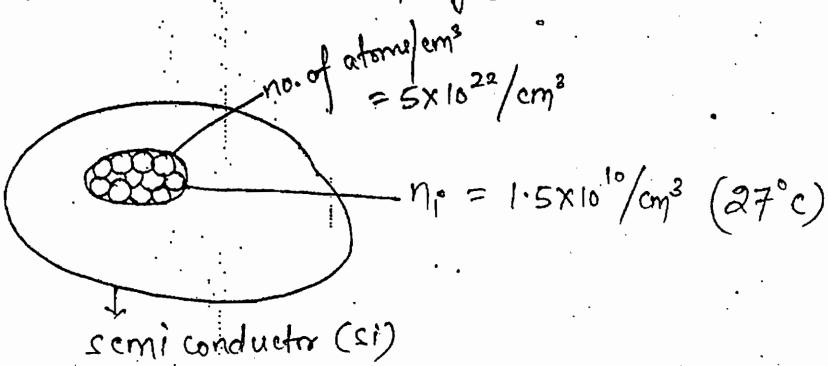
Majority carriers → Holes

Minority carriers → Electrons

P-type semi-conductor

* 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 semi conductors?



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

(1). <u>Ordinary p-n diodes:</u>

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

5×10^{22} sc atoms → N_D

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

$N_D > N_i$ ....

(2). <u>Zener Diode:</u>

10^6 s.c atoms → 1 impurity {1 'p'}

5×10^{22} s.c atoms → N_D

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

$N_D >> N_i$

(3). <u>Tunnel diode:</u>

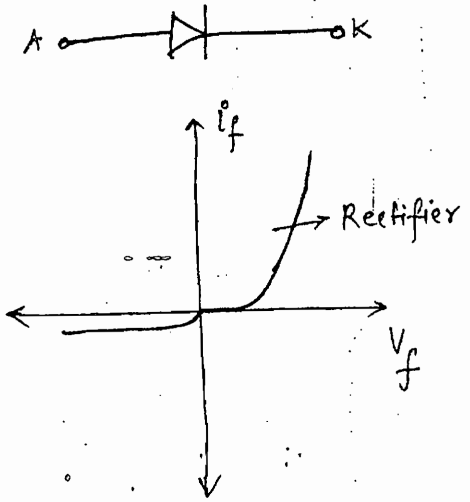
10^3 s.c atoms → 1 impurity

5×10^{22} s.c. atoms → N_D

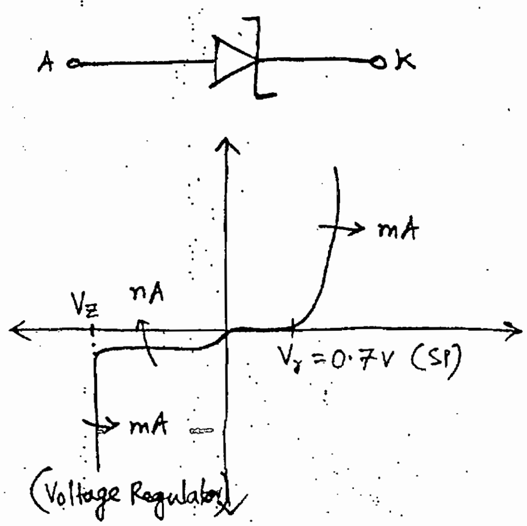
$$N_D = \frac{5 \times 10^{22}}{10^3} = 5 \times 10^{19}/cm^3$$

$N_D >>> N_i$

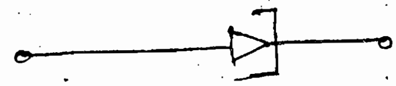
<u>Ordinary p-n diode:</u>

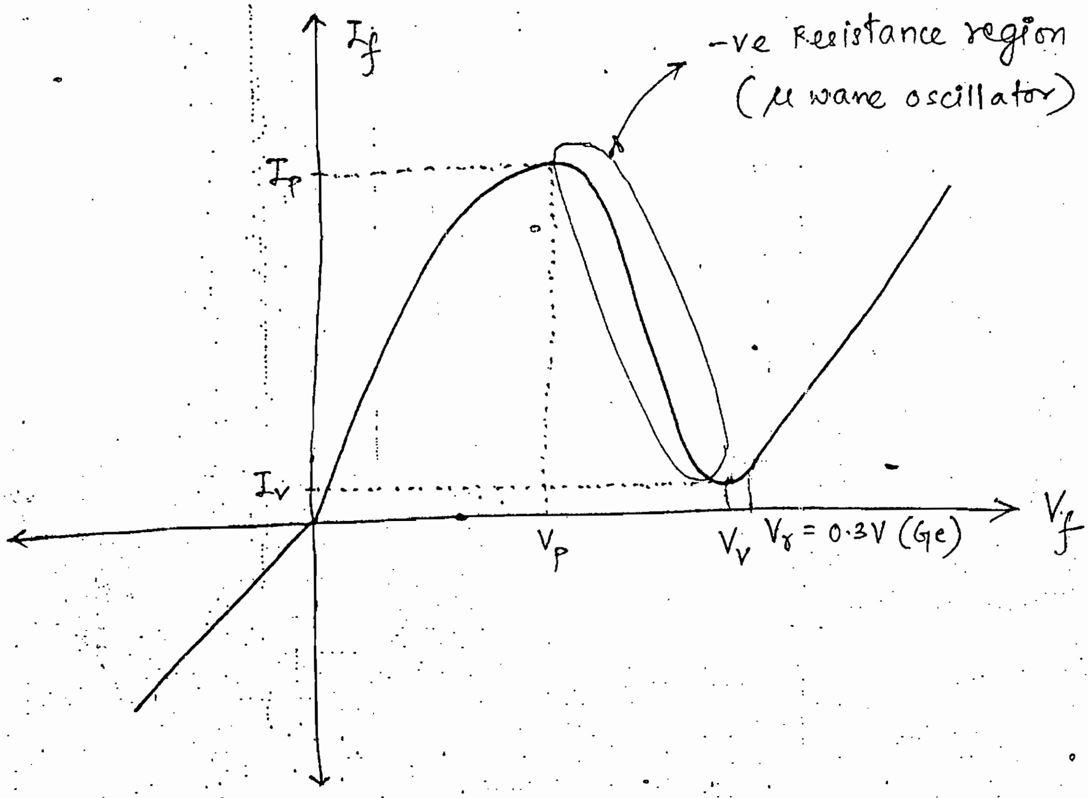


<u>Zener Diode:</u>



Tunnel diode :



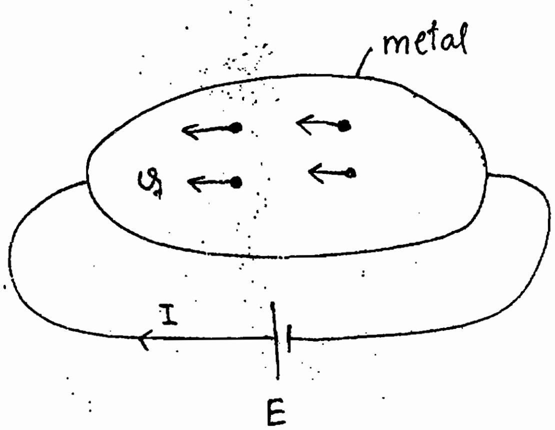


$$R = \frac{dV}{dI} = \frac{V_v - V_p}{I_v - I_p} = -ve.$$

Q. Explain about drift current in semiconductors ?

Drift current

Drift → movement
(greek word)



$v_d \propto E$

$v_d = \mu E$

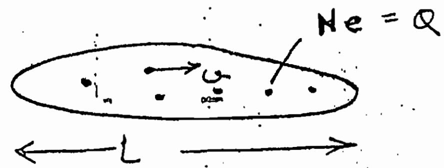
$\mu = \frac{v_d}{E}$ $m^2/V\text{-sec}$

(1). "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)



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

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

$$I = \frac{Ne\,v_d}{L}$$

$$J = \frac{Ne\,v_d}{AL} \qquad (\because n = N/AL)$$

$$\boxed{J = ne\,v_d} \qquad (\because \rho \rightarrow \text{charge density}) \quad \rho = ne$$

$$J = \rho v$$

Metals:

$$J = ne\mu E \qquad (\because v_d = \mu E)$$

$$J = \sigma E \qquad (\because \sigma = ne\mu)$$

Semi-conductors:

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

$$J_n = nq\mu_n E$$

$$J_p = pq\mu_p E$$

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

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

(a) Intrinsic s.c – $(n = p = n_i)$

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

(a) Extrinsic s.c – $(n \neq p)$

(i) · N-type : $(n > p)$

$$\sigma_{N\text{-type}} = nq\mu_n$$
$$\approx N_D q \mu_n$$

(ii) · P-type : $(p > n)$

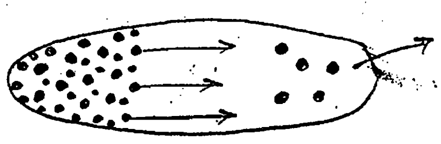
$$\sigma_{p\text{-type}} = pq\mu_p$$
$$\approx N_A q \mu_p$$

Q. Explain about Diffusion current in semi-conductors?

Diffusion current :

→ Uniformly dopend s.c.

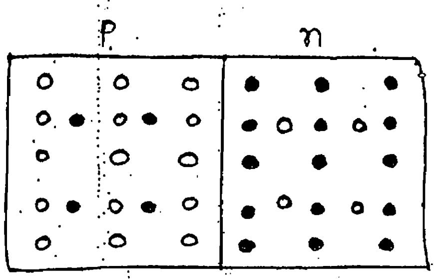
→ $i_{diff} = 0$



→ Non-uniformly doped s.c

→ x $\quad i_{diff.} \neq 0$

Eg.



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

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

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

$I_n = A q D_n \frac{dn}{dx}$

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

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

$I_p = -A q D_p \frac{dp}{dx}$

<u>Diffusion current</u> $\rightarrow$ The rate of change of concentration w.r.to distance x is called as diffusion current. Diffusion current mechanism can also be called as "concentration gradient".

<u>Analysis:</u>

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

$V_T = \frac{KT}{q}$ where, $K \rightarrow$ Boltzman constant
$T \rightarrow$ Temperature

<u>In semi-conductors:</u>

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

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

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

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

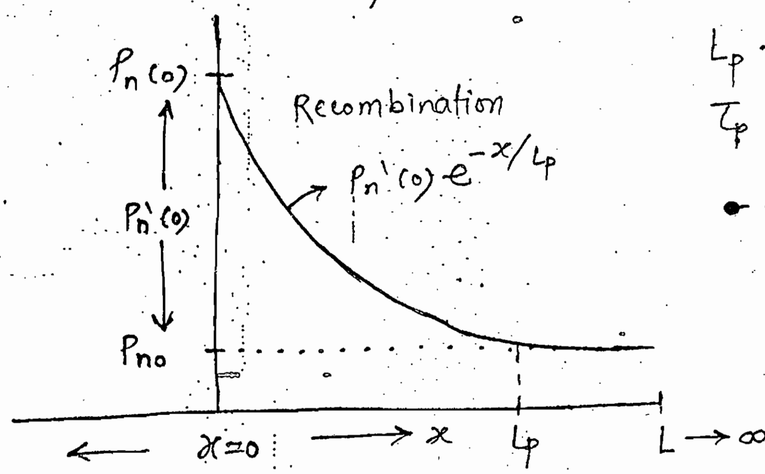
$K = 1.38 \times 10^{-23}\ J/K$

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

Q.- Calculate volt equivalent Temp. at room temp. ?

A. $V_T = \frac{300}{11600} = 25.86\ mV \approx 26\ mV / (25\ mV)$

Graphical Analysis:



$L_p \to$ Diffusion length

$\tau_p \to$ Carrier life time

$L_p^2 = D_p \tau_p$

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

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)

$x = \infty$

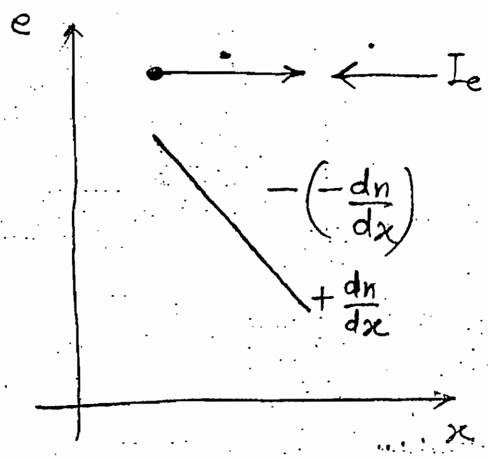
$P = P_n'(0) \cdot e^{-\infty/L_p}$

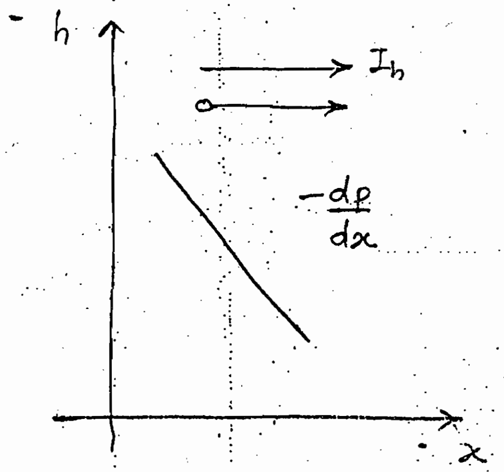
$= P_n'(0) \cdot \frac{1}{e^\infty}$

$= 0$

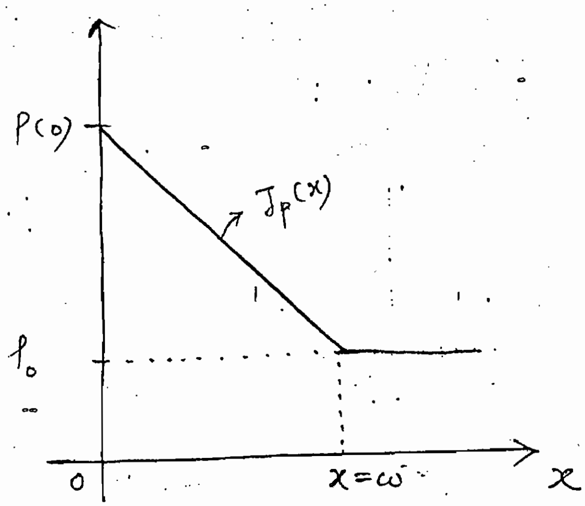
(1). When the thickness of material is more, recombination rate will be more, diffusion rate will be less.
(Non-linear graph)

(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:

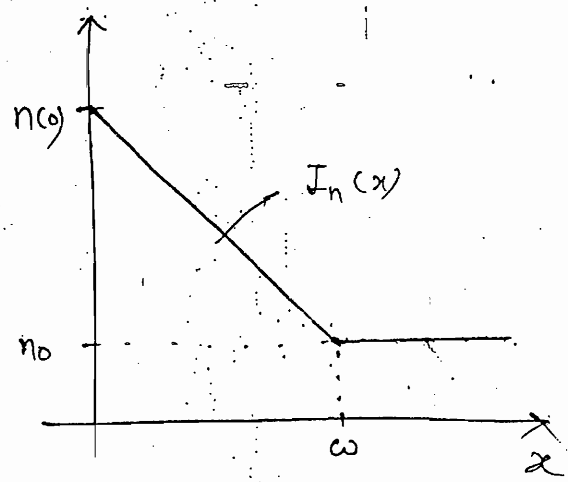


$$J_p(x) = -q D_p \frac{dp(x)}{dx}$$

$$= -q D_p \frac{P(0) - P_0}{0 - \omega}$$

$$= \frac{q D_p [P(0) - P_0]}{\omega}$$

$$\left.\begin{array}{l} \circ\longrightarrow \\ \longrightarrow I_h \end{array}\right\} +ve$$



$$J_n(x) = qD_n \frac{dn}{dx}$$

$$= qD_n \frac{\{n(0) - n_0\}}{0 - \omega}$$

$$= \frac{-qD_n (n(0) - n_0)}{\omega}$$

$\xrightarrow{e^-}$, $\xleftarrow{\quad} I_e$ $\Big\}$ $-ve$

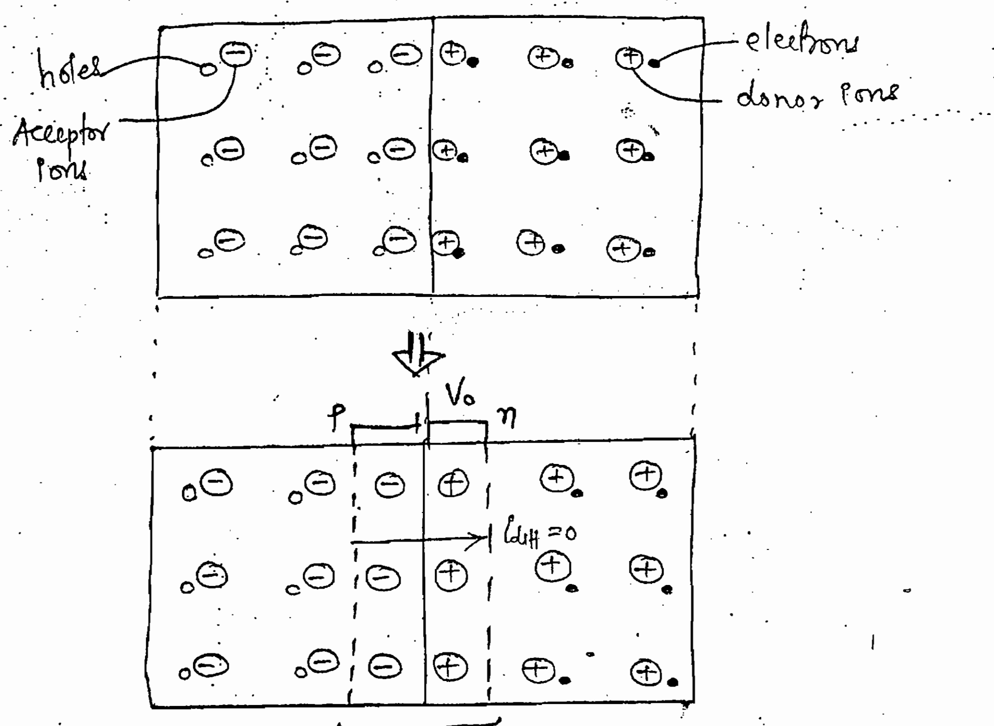
Q. Explain about pn jxn theory in 3 conditions :

(1) open ckt pn jxn

(2). forward Biased

(3). Reversed Biased.

(1) Open ckt pn jxn —



Depletion Region

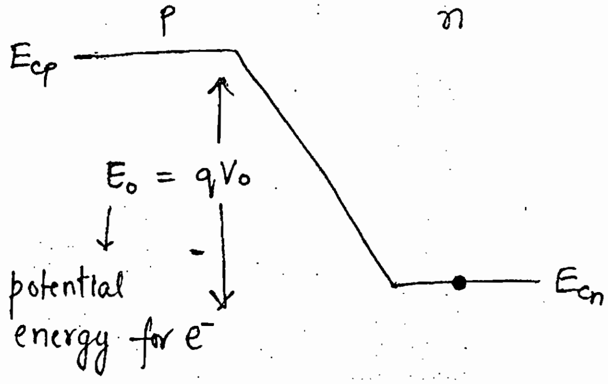
(or) space charge Region (or) Transition region

V_0 (or) V_j (or) V_γ → potential Barrier (or) jxn voltage (or) Contact potential (or) Cut in Voltage

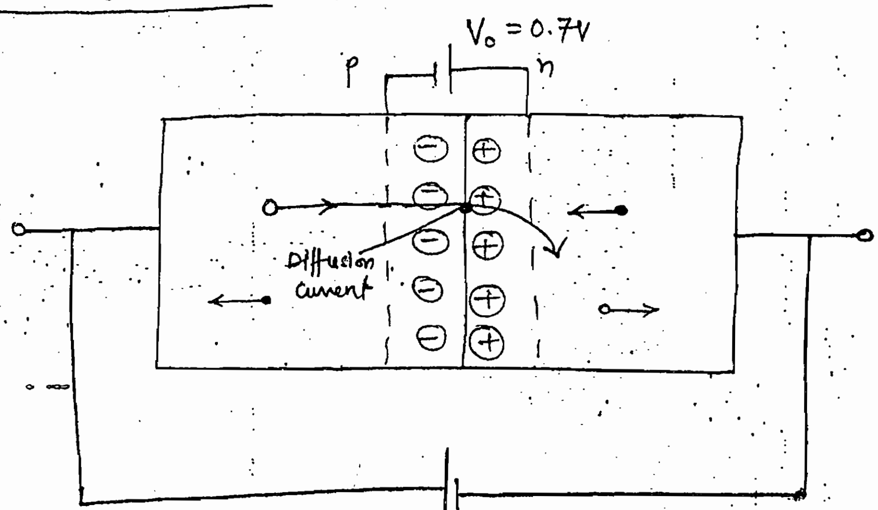
$$V_0 = V_T \ln\left(\frac{N_A N_D}{n_i^2}\right)$$

$V_0 \text{ (Si)} \rightarrow 0.5V \text{ to } 0.7V$

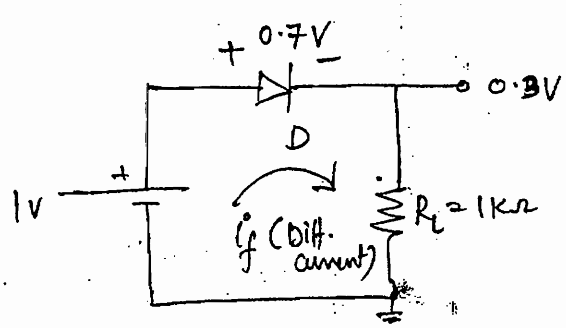
$V_0 \text{ (Ge)} \rightarrow 0.1V \text{ to } 0.3V$

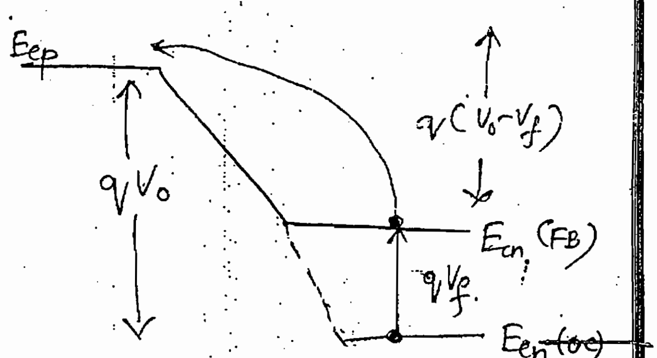


(2). Forward Bias:

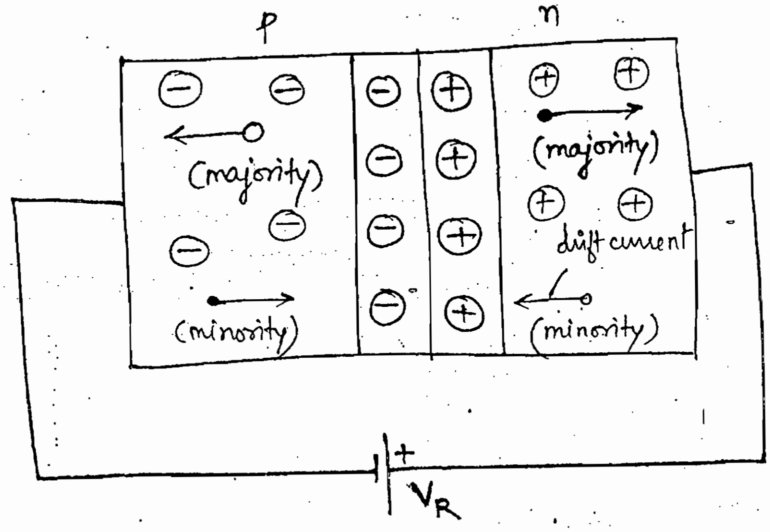


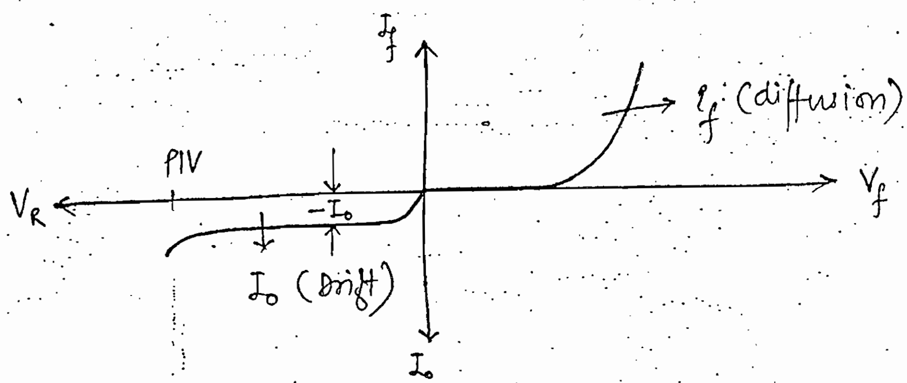
$V_f = 1V = (0.7 + 0.3)V$

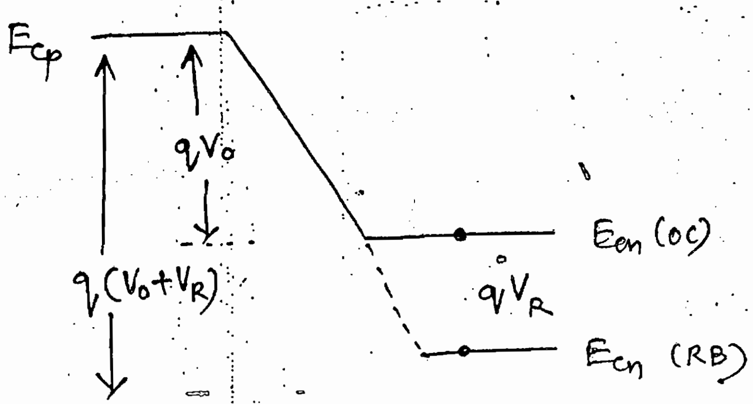




(3). Reverse Bias —







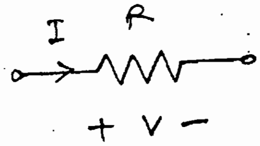
* Due to appearance of immobile charge ions, resistance offered is more in R.B whereas in F.B resistance offered is less

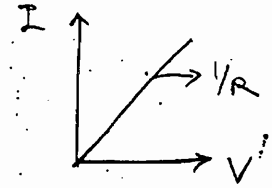
* Depletion layer in R.B has more width due to immobile charge ions (↑).

Diode Applications :

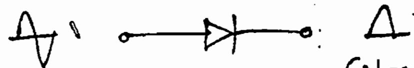
Linear Resistor

$V = IR$

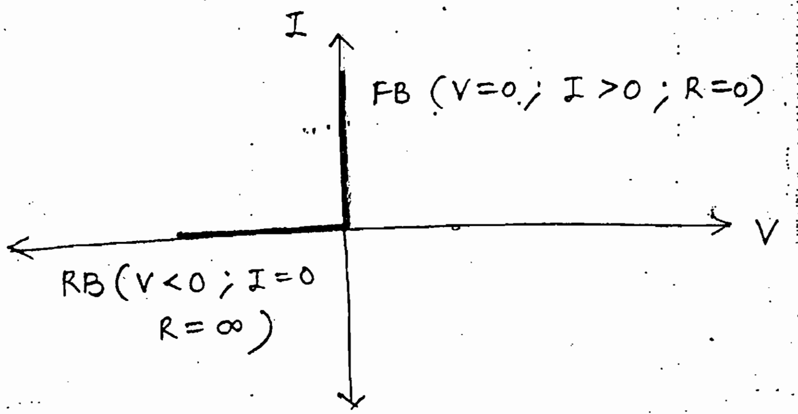




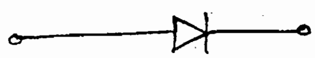
Ideal Diode :

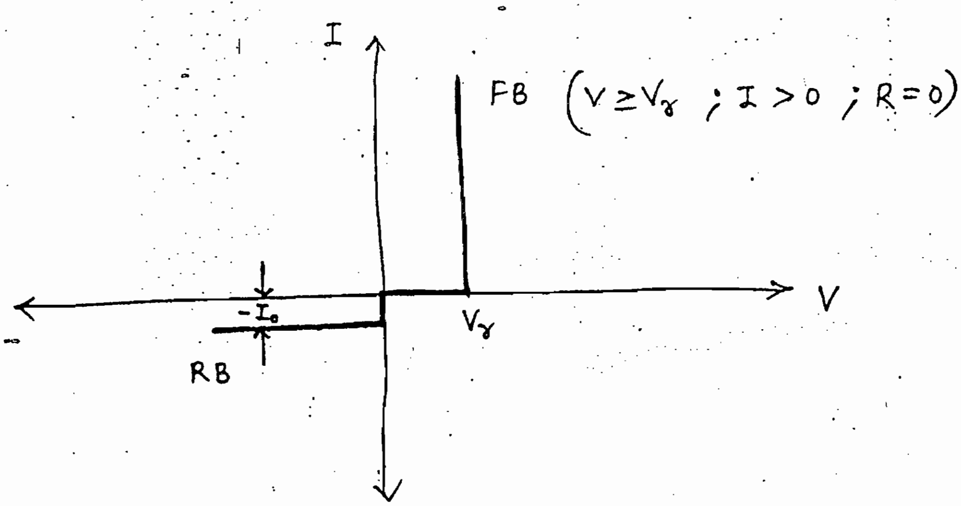


(Non-linear)

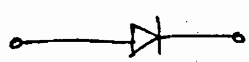


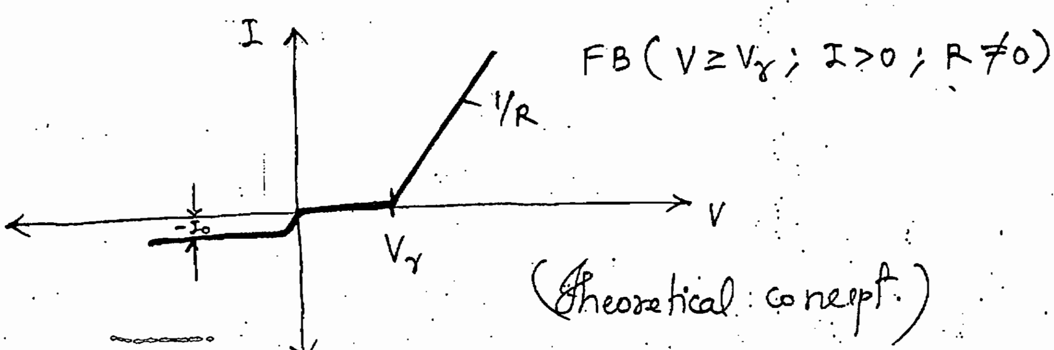
Practical diode :





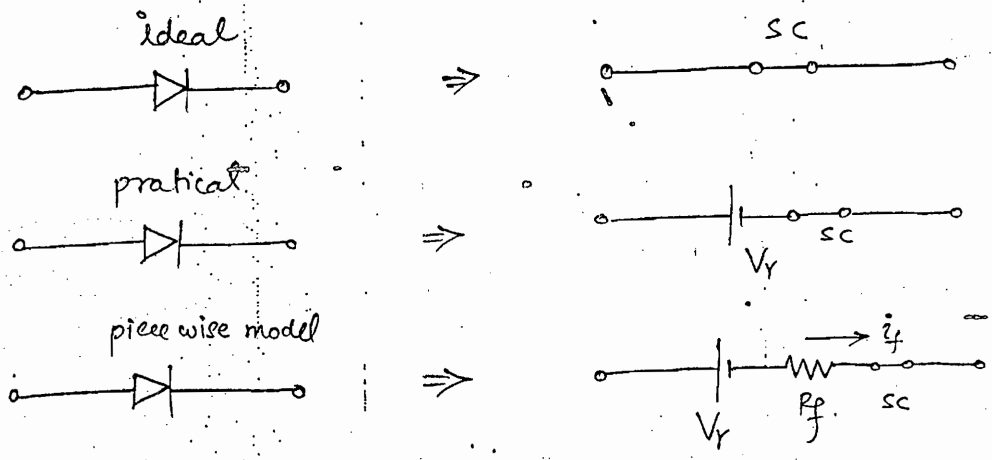
Piece-wise linear model :





(Theoretical concept)

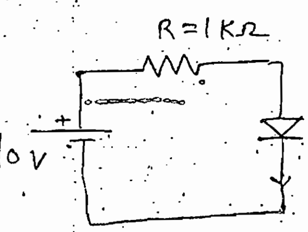
Equivalent models for a diode :



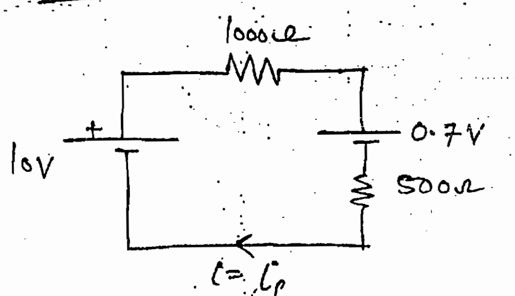
Problems:

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

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

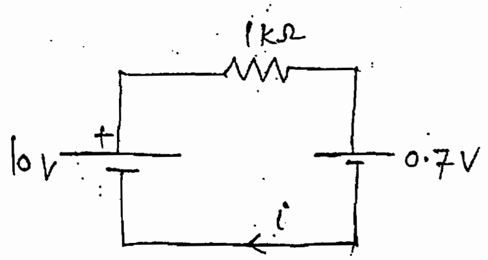


Sol. Piece-wise linear model:



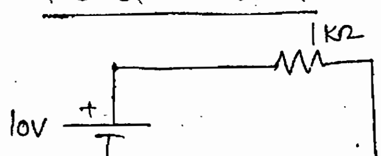
$$i = \frac{10-0.7}{1000+500} = 6.2\ mA.$$

Practical model:



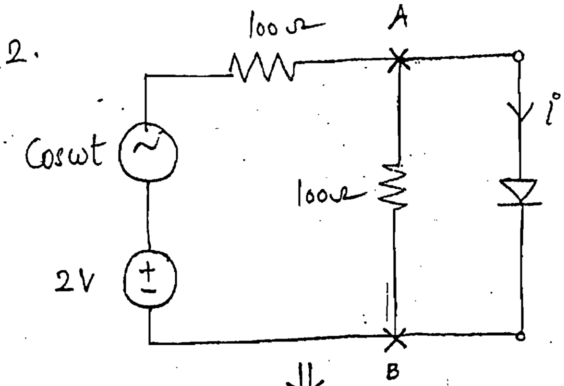
$$i = \frac{10-0.7}{1000} = 9.3\ mA$$

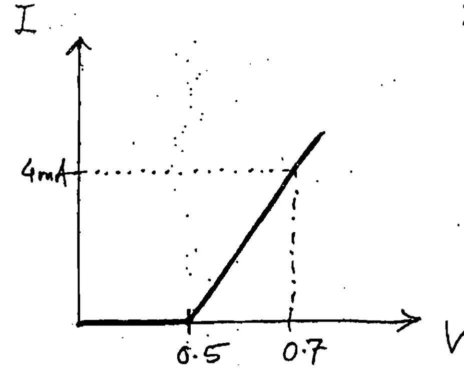
Ideal model:



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

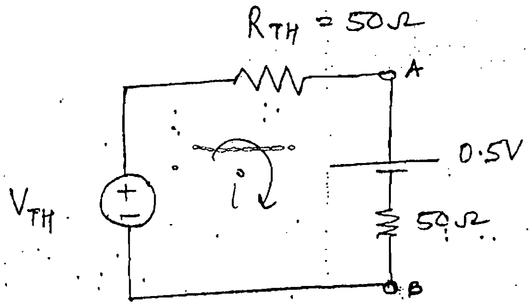
Q2.





$\Downarrow$

Sol.



$$\therefore r_f = \frac{\Delta V}{\Delta i} = \frac{0.7-0.5}{4mA - 0}$$

$$\therefore r_f = 50\,\Omega \quad \left\{ \text{from graph} \right\}$$

$$\therefore V_\gamma = 0.5V$$

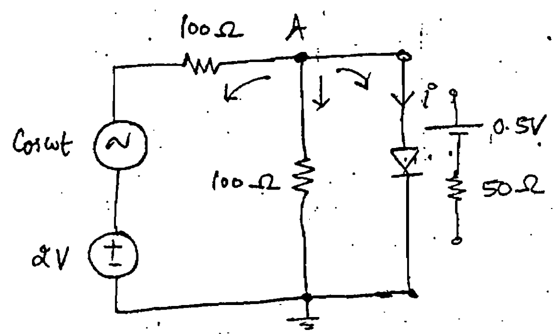
$$V_{TH} = \frac{(\cos\omega t + 2)}{2} = 1 + \frac{\cos\omega t}{2}$$

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

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

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

Alter:



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

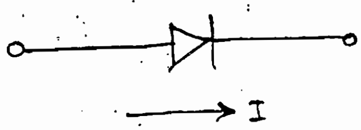
$$4V_A - 2 - \cos\omega t - 1 = 0$$

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

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

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

Diode current Eqn:



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

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) $\rightarrow$ 0.7 volts

$V \rightarrow$ −ve (RB)

$V_T \rightarrow$ thermal voltage

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

At room temperature, $V_T = 26\,mV$

η = Idealized factor

$\eta = 1$ (integrated theory) ✓

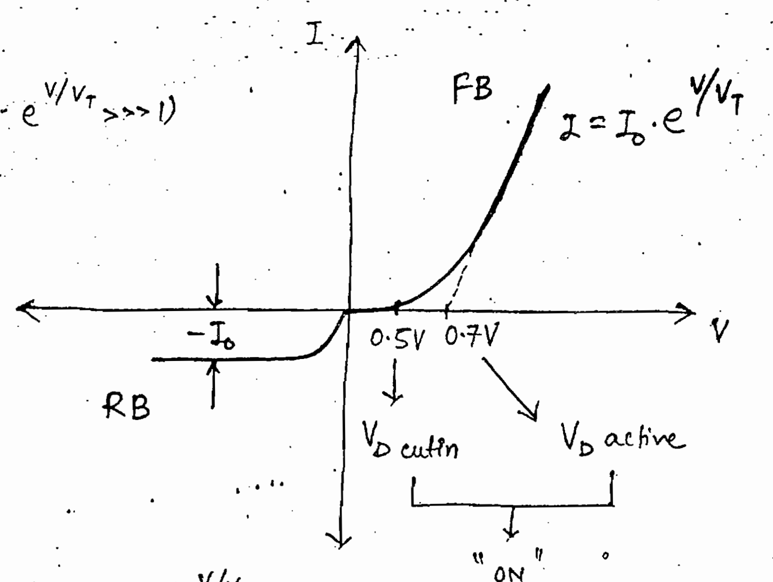
$= 2$ (discrete theory)

Case (1): FB

$V = +ve$ (0.7V)

$I = I_0 \left(e^{V/V_T} - 1\right)$ $\quad (\because e^{V/V_T} >>> 1)$

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



Case (2): RB

$V = -ve$

$I = I_0 \left(e^{-V/\eta V_T} - 1\right)$ $\quad (\because e^{-V/V_T} <<< 1)$

$$\boxed{I = -I_0}$$

Reverse saturation current.

Temperature dependence parameters in diode Eqn:

$$I = I_0 \left(e^{V/V_T} - 1 \right)$$

V_T — depends on T

— depends on T $\left(V_T = \frac{T}{11600} \right)$

I_0 — depends on T (thermal agitation, Breakage of covalent Bonds)

I_0 Vs temp:

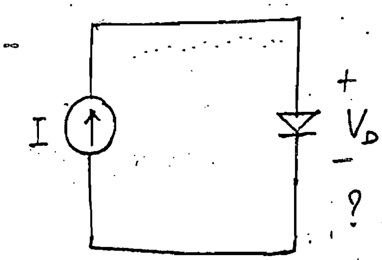
" I_0 increases by 7% per °C rise in temp."

(or)

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

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

V_D Vs Temp:



$V_D \rightarrow T = 20°C$

↓ changes to

$? \leftarrow 24°C$

$$\frac{dV_D}{dT} = -2.5 \text{ mV/°c} \quad (\text{OR}) \quad -2 \text{ mV/°c}$$

$$\frac{dV_D}{dT} = -2.5 \text{ mV/°c}$$

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

$dV_D = -10 \text{ mV}$

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

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

$= (700 - 10)$ mV $= 690$ mV

V_T Vs temp:

$$\uparrow V_T = \frac{T\uparrow}{11600}$$

Q6. $I_o = 10$ pA (20°c)

$I_o = ?$ (40°c)

Exact: $\frac{dI_o}{dT} = 0.07$ pA/°c

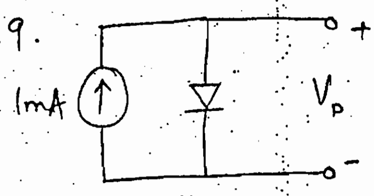
$I_o' = I_o \cdot 1.07^{(40-20)} = (1.07)^{20} I_o = 3.89\, I_o = 38.9$ pA

Approx: $I_o \rightarrow$ double for every 10°c rise in temp.

$I_o \rightarrow 20$ pA at 30°C ....

$I_o' \rightarrow 40$ pA at 40°c.

Q9.



20°c, $V_D = 700$ mV
↓ ↓
40°c, (?)

$\frac{dV_D}{dT} = -2.5$ mV/°c

$dV_D = -2.5 \times (20) = -50$ mV

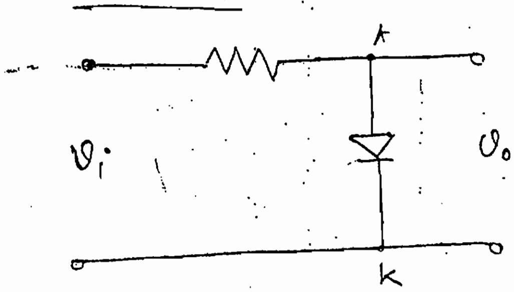
$\therefore V_D' = 650$ mV at 40°c

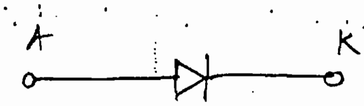
Clippers:

- Shunt Clippers
- Series Clippers
- O/p following I/p designs
- 2 independent level clipping
- Attenuators
- Single branch designs
- Multi branch designs

Shunt Clippers:

Model 1:





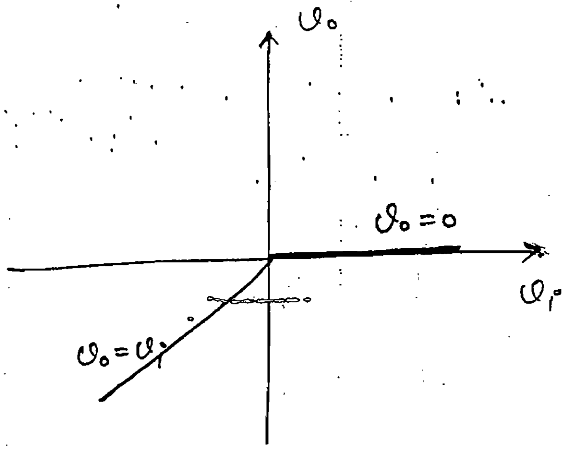
$V_A > V_K \rightarrow$ SC

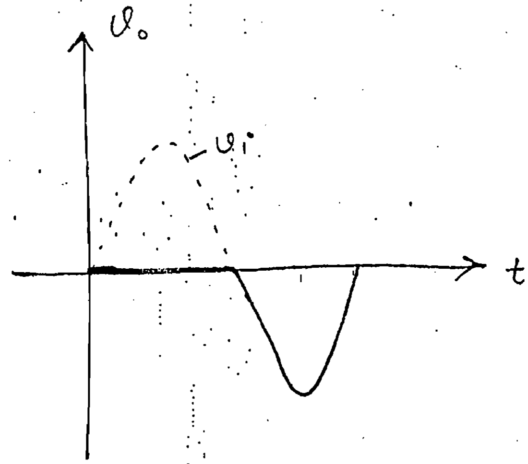
$V_A < V_K \rightarrow$ OC

$v_i < 0 \rightarrow$ D (off) $\Rightarrow V_o = v_i$

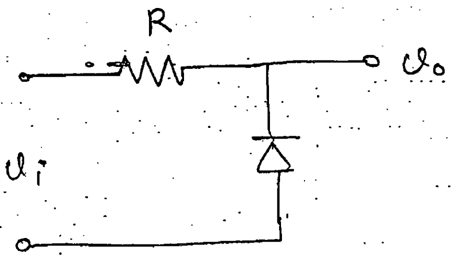
$v_i > 0 \rightarrow$ D (on) $\Rightarrow v_o = 0$

Transfer characteristics :



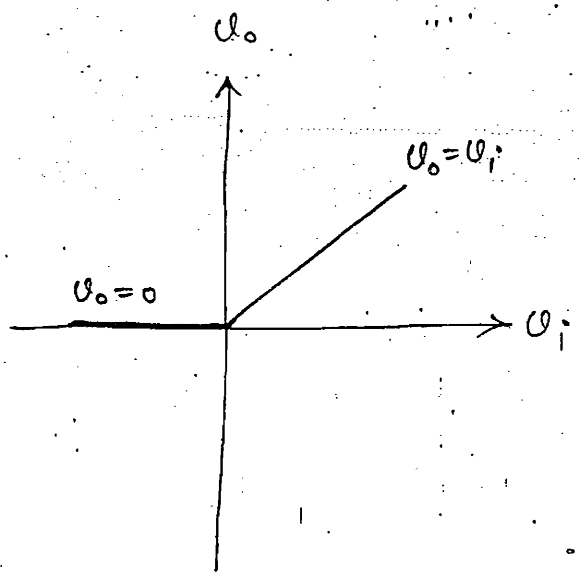


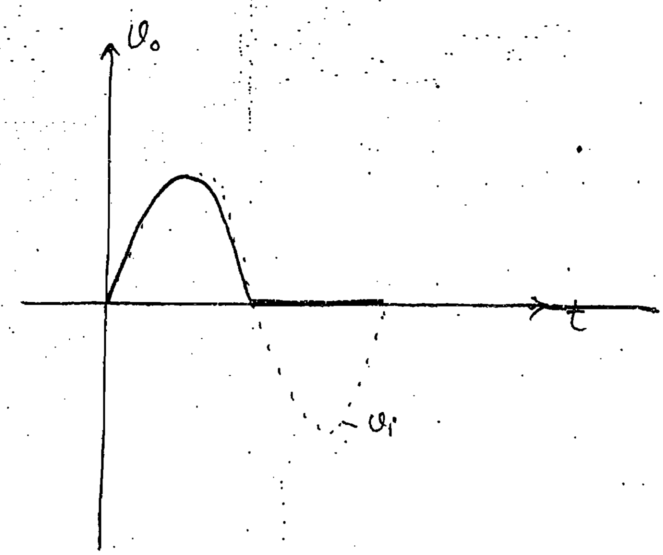
Model - 2 :



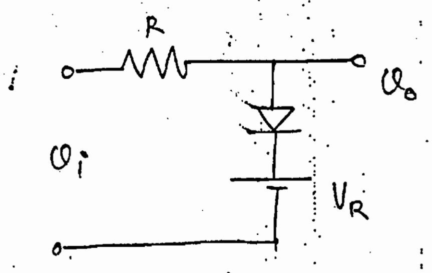
$v_i < 0 \quad D\ (\text{on}) \rightarrow v_o = 0$,

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



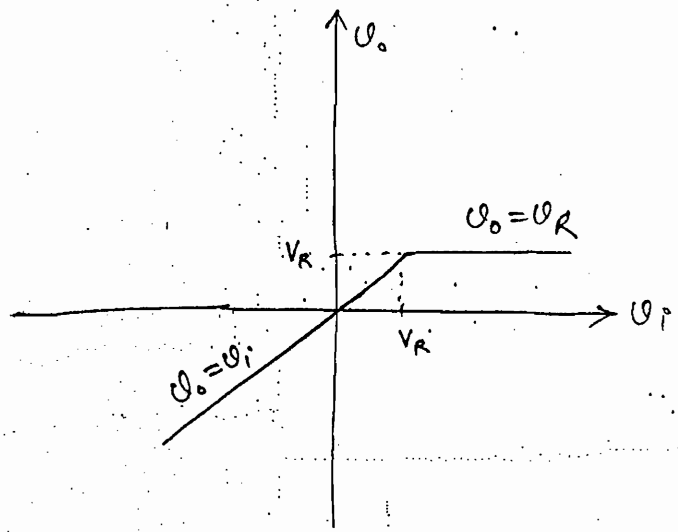


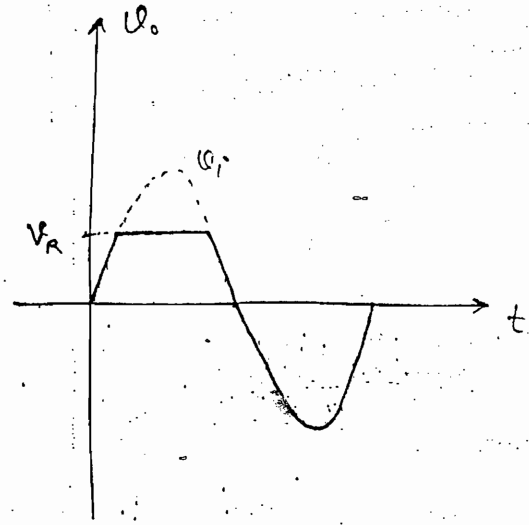
Model - 3 :



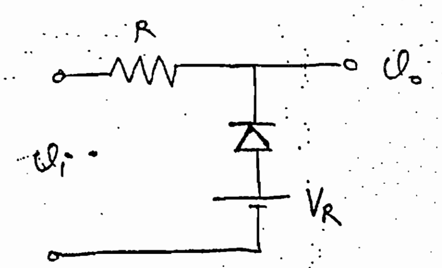
$v_i < V_R \rightarrow D(\text{off}) \rightarrow v_o = v_i$

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



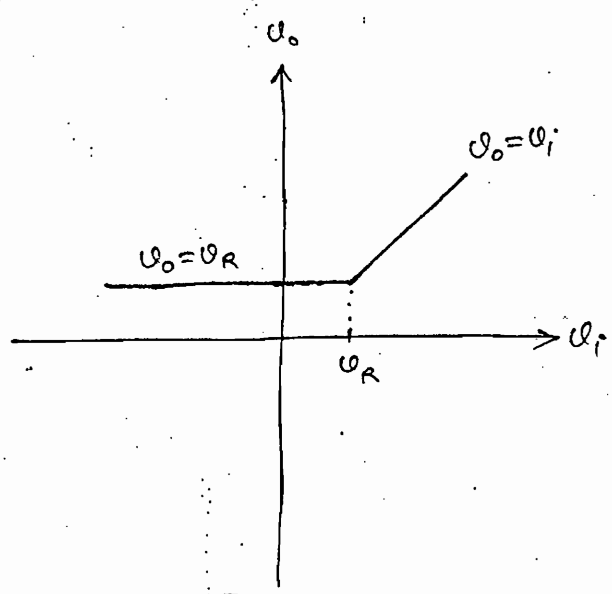


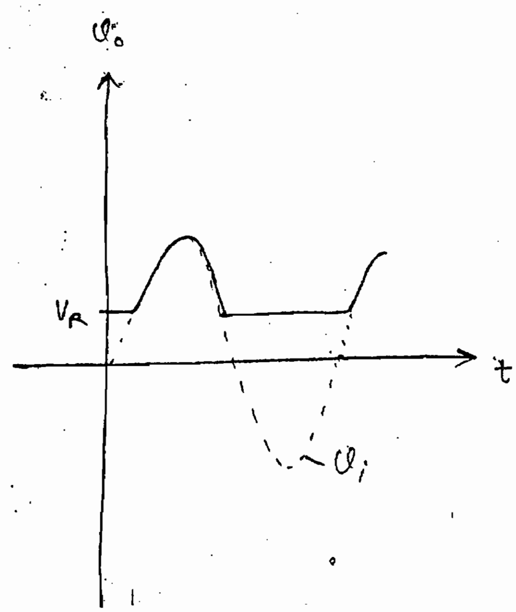
Model - 4 :



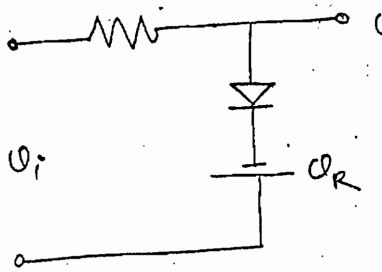
$v_i < v_R \rightarrow D(\text{on}) \rightarrow v_o = v_R$

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



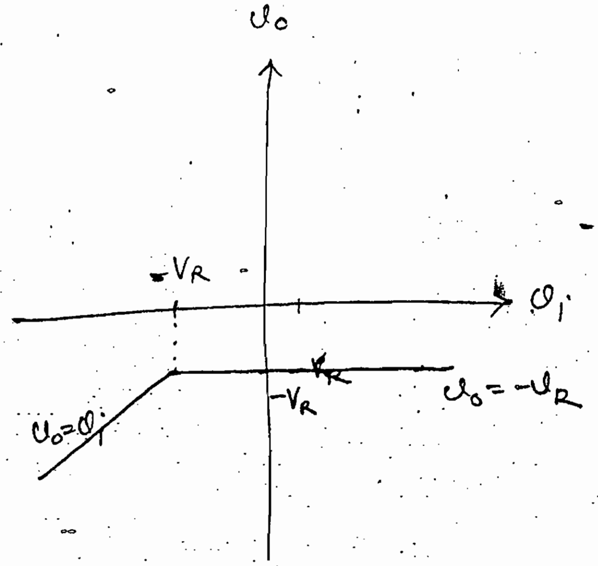


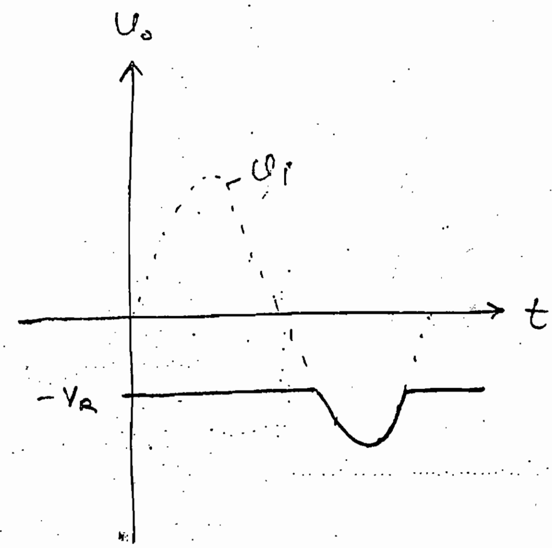
Model - 5 :



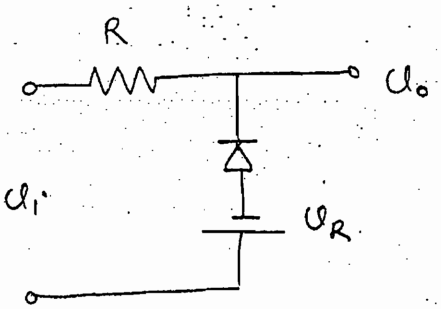
$v_i < -v_R \rightarrow D\,(\text{off}) \rightarrow v_o = v_i$

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



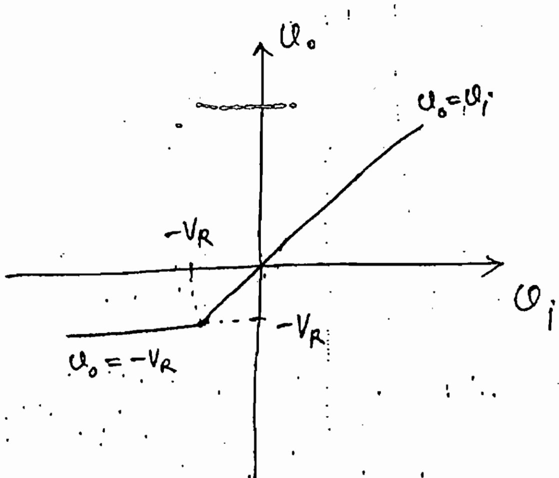


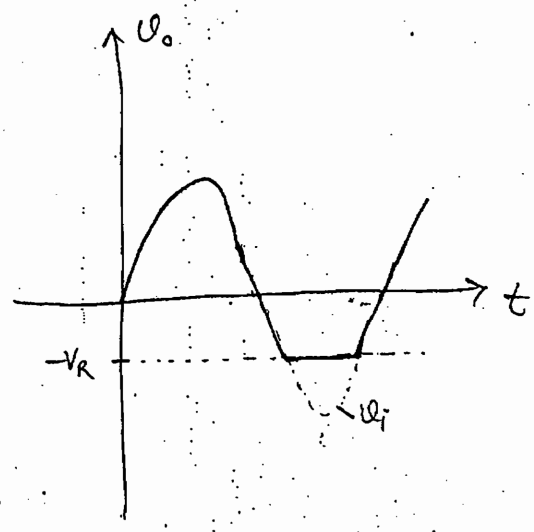
Model - 6 :



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

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





Conclusion:

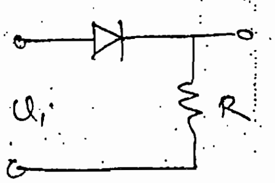
(1): When the diode is in downward dirn, the s/g will be transmitted below the ref. voltage.

(2) When the diode is in upward dirn, the s/g will be transmitted above the ref. voltage.

Series Clippers:

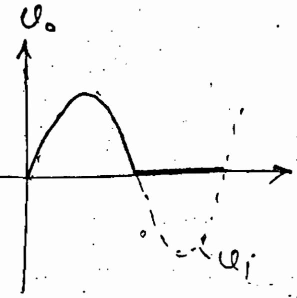
Models:

(1).

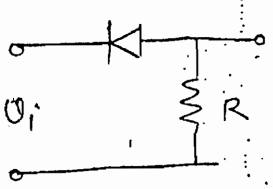


$v_i < 0 \rightarrow D(\text{off}) \Rightarrow v_o = 0$

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

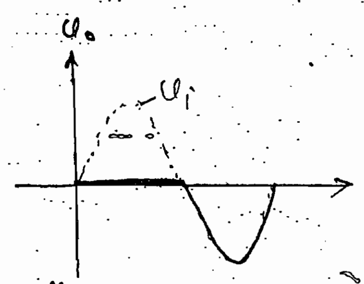


(2).

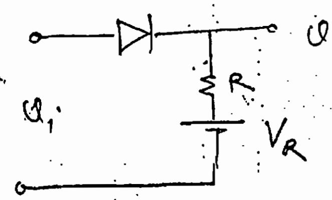


$v_i < 0 \rightarrow D(\text{ON}) \Rightarrow v_o = v_i$

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

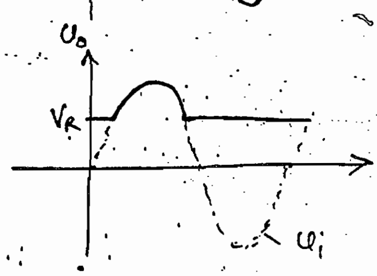


(3).

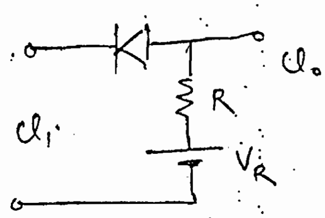


$v_i < V_R \rightarrow D(\text{off}) \Rightarrow v_o = V_R$

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

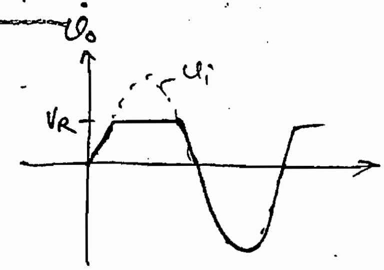


(4).

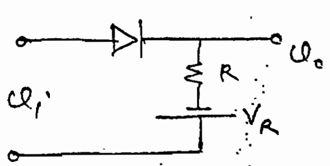


$v_i < V_R \rightarrow D(\text{ON}) \Rightarrow v_o = v_i$

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

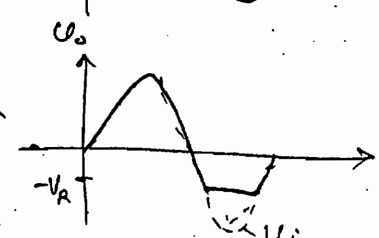


(5).

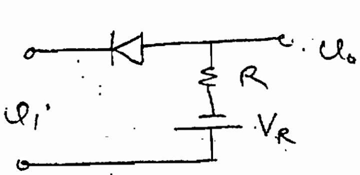


$v_i < -V_R \rightarrow D(\text{off}) \Rightarrow v_o = -V_R$

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

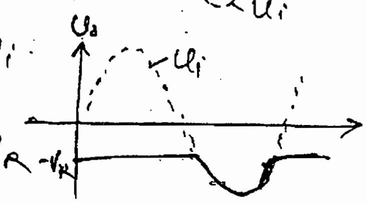


(6).



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

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



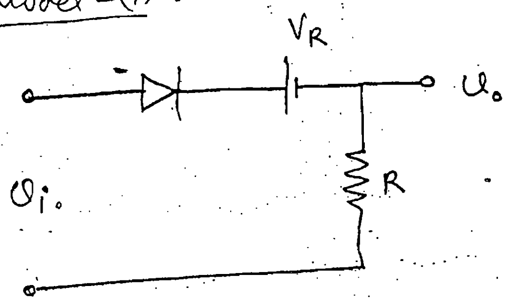
Conclusions:

(1). When the diode is in forward dirⁿ to the i/p s/g, the s/g will be transmitted above the ref. voltage.

(2). When the diode is in reverse dirⁿ to the i/p s/g, the s/g will be transmitted below the ref. voltage.

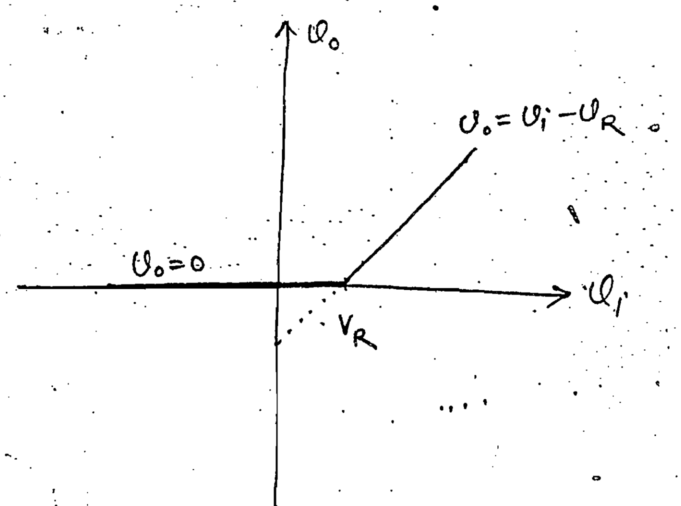
O/p following i/p designs:

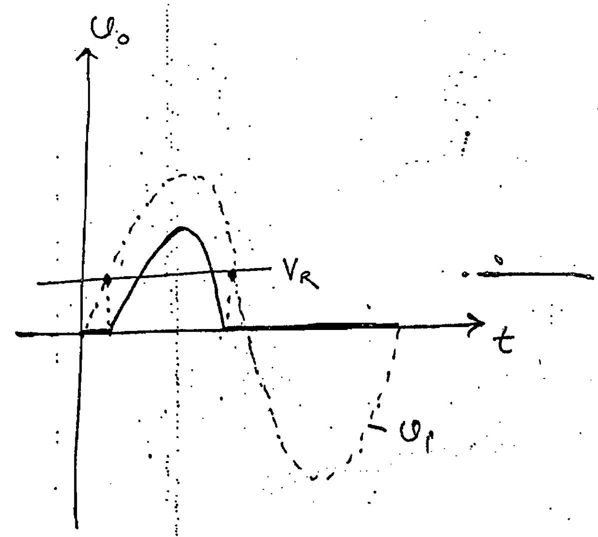
Model-(1):



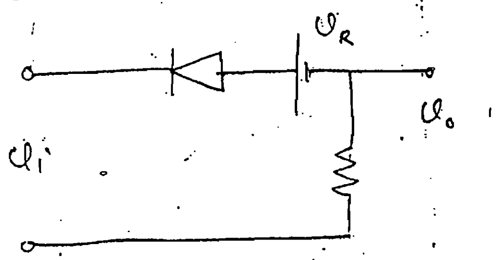
$v_i < V_R \rightarrow D\ (\text{off}) \Rightarrow v_o = 0$

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



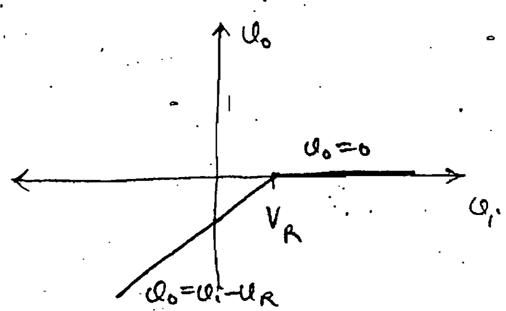


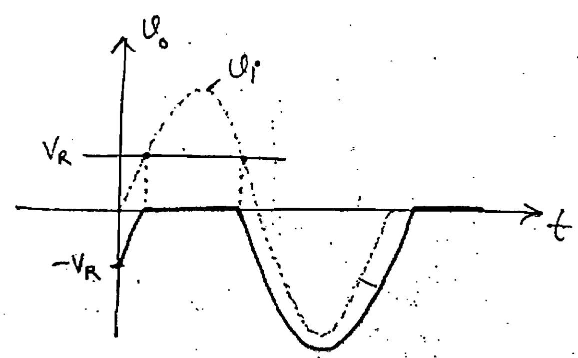
Model-(2):



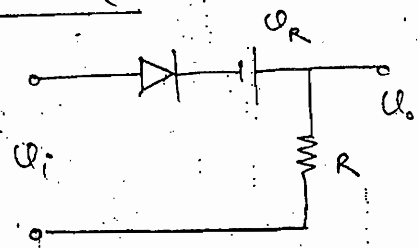
$v_i < V_R \rightarrow D\ (\text{on}) \Rightarrow v_o = v_i - v_R$

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



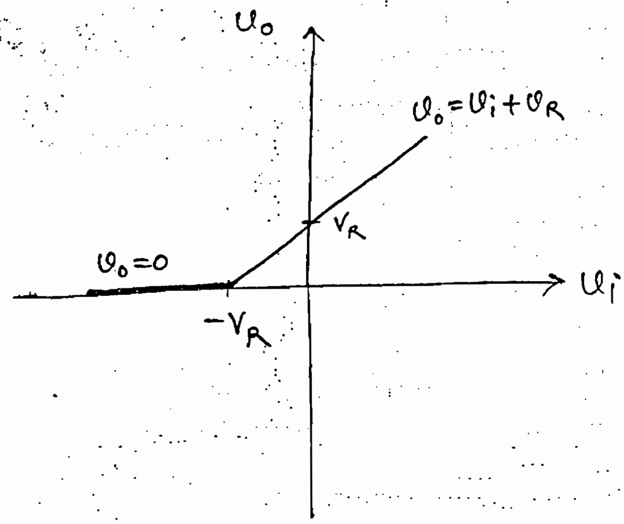


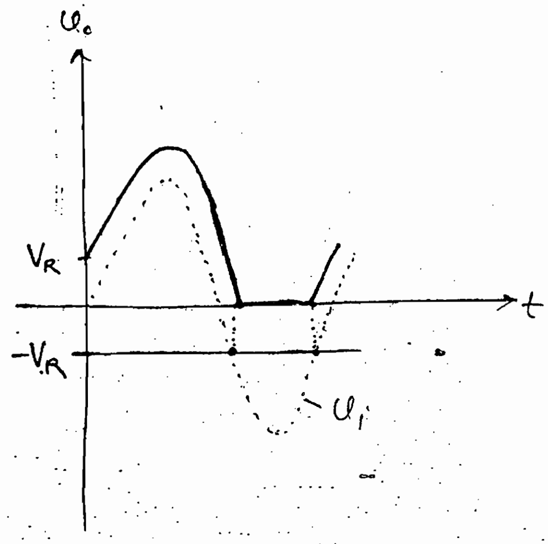
Model -(3):



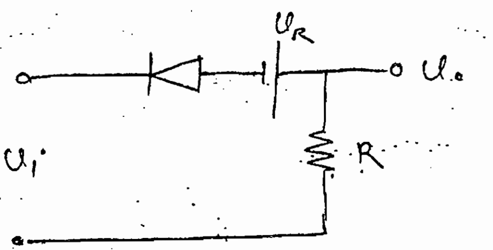
$v_i < -v_R \rightarrow D(\text{off}) \Rightarrow v_o = 0$

$v_i > -v_R \rightarrow D(\text{ON}) \Rightarrow v_o = v_i + v_R$



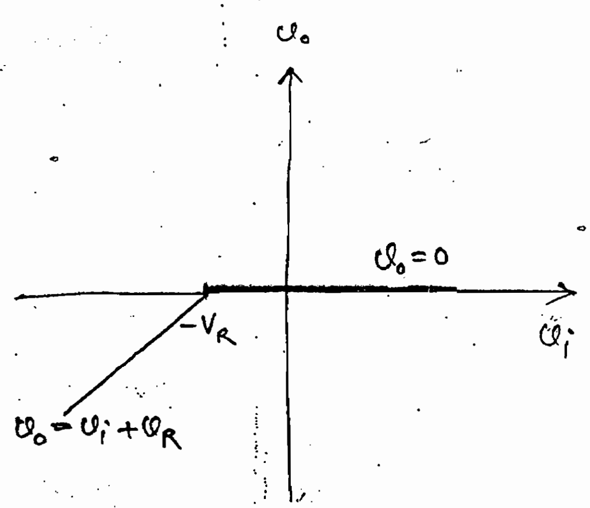


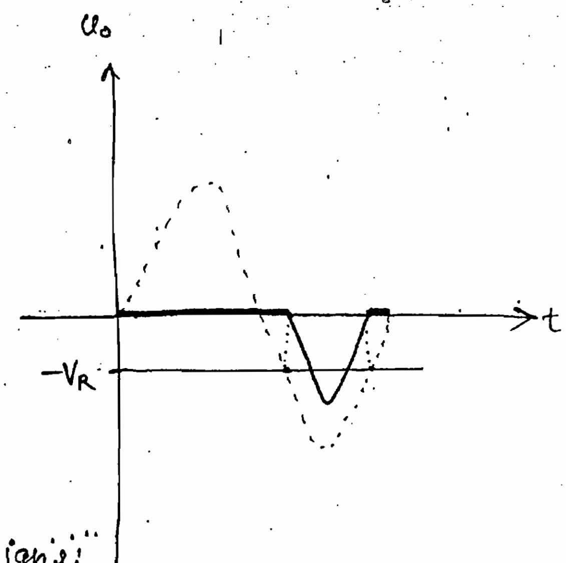
Model -(4):



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

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

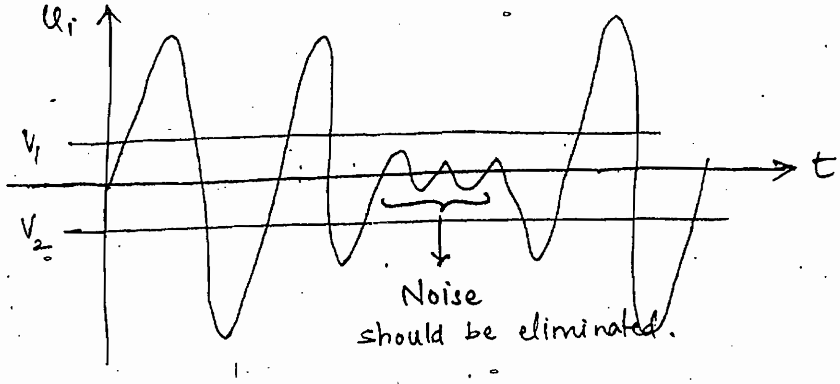


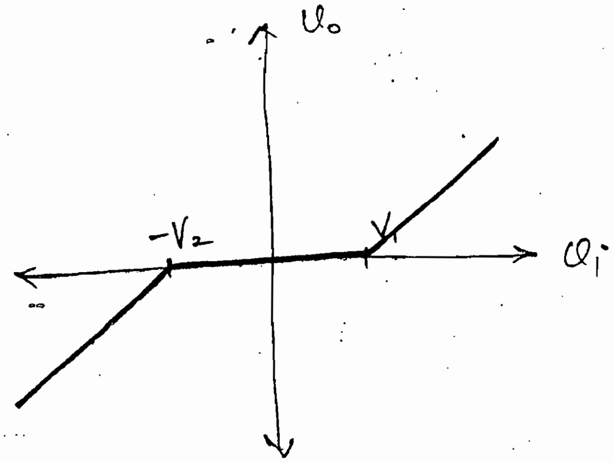


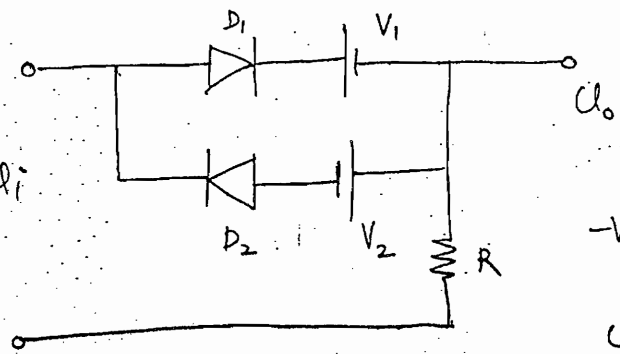
Applications of o/p following i/p designs:

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

Sol.







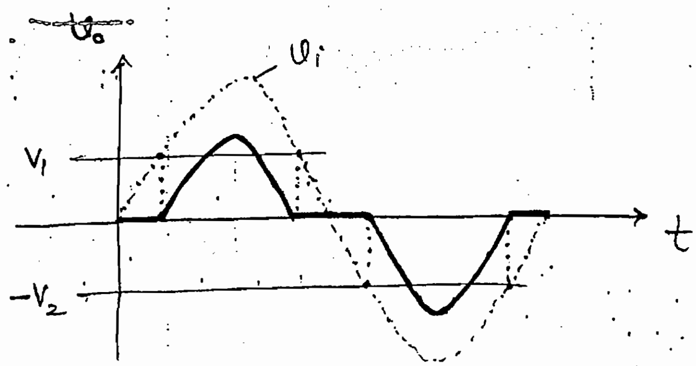
$v_i < -V_2$, D_1 (off), D_2 (on)

$v_o = v_i + V_2$

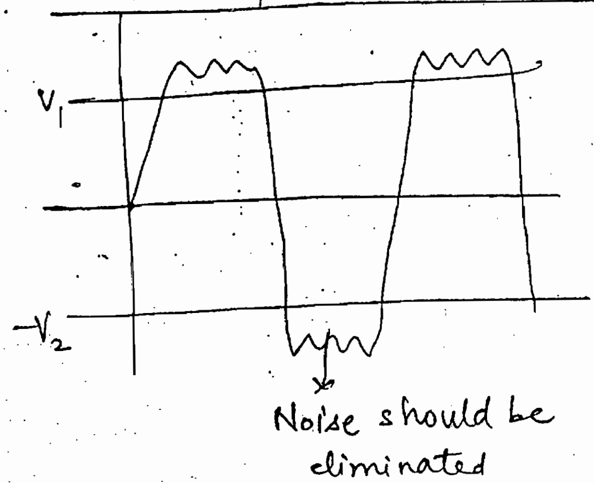
$-V_2 < v_i < V_1$; D_1 (off), D_2 (off)

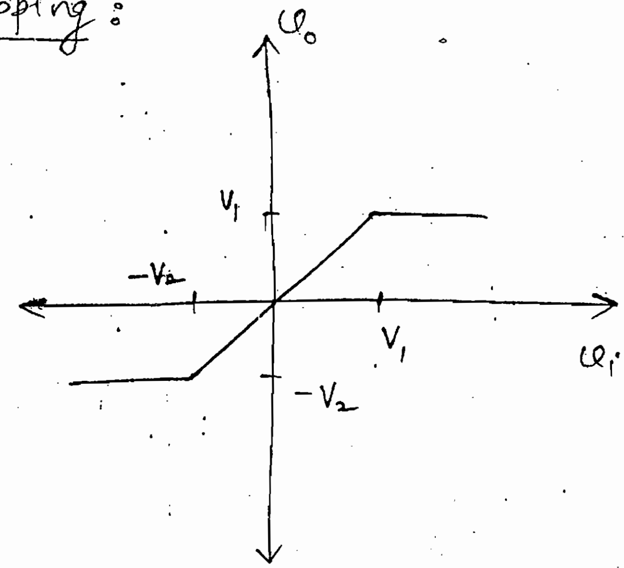
$v_o = 0$

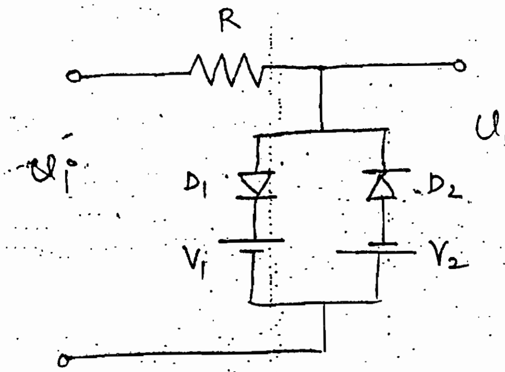
$v_i > V_1$, D_1 (on), D_2 (off), $v_o = v_i - V_1$



(4). Two Independent level Clipping :



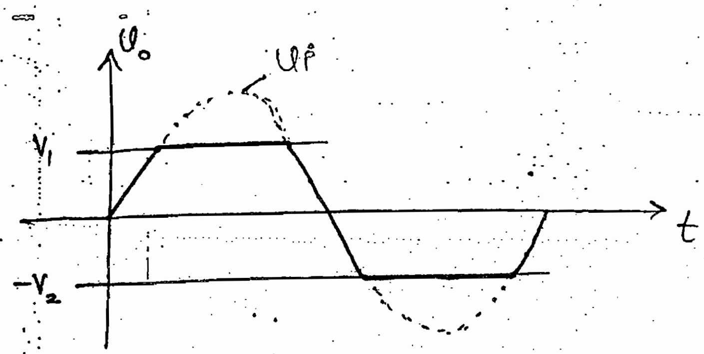




$v_i < -V_2$, D_1 (off) , D_2 (on) , $v_o = -V_2$

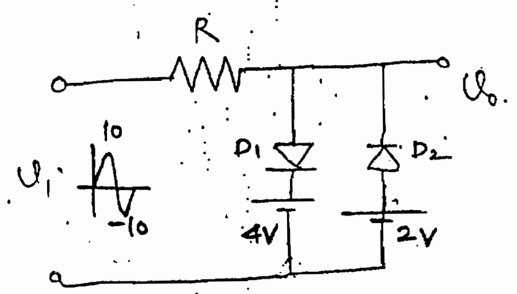
$-V_2 < v_i < V_1$, D_1(off) , D_2 (off) , $v_o = v_i$

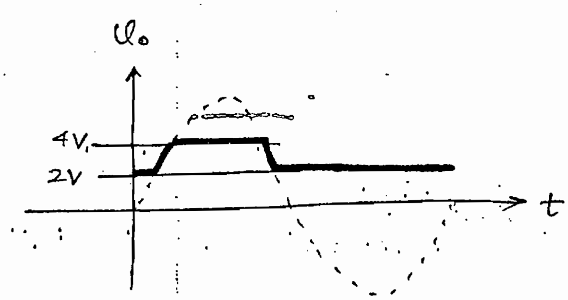
$V_i > V_1$, D_1 (on) , D_2 (off) , $v_o = V_1$



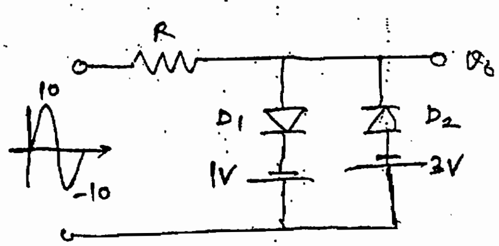
* $\boxed{V_{D_1} > V_{D_2}}$

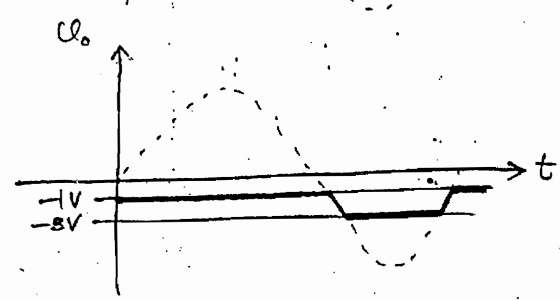
Q.



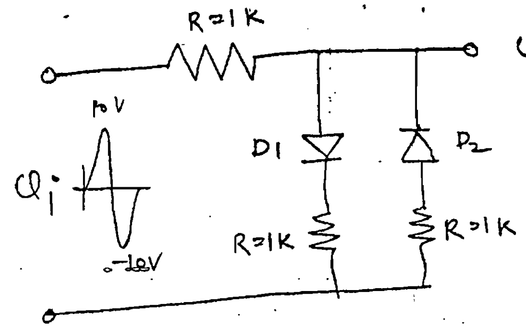


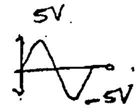
Q.





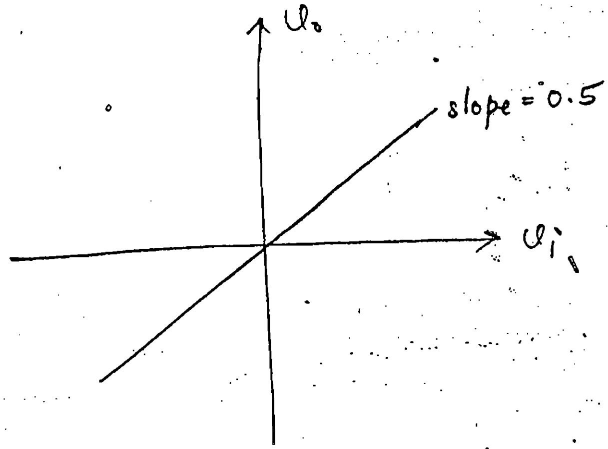
(5). Attenuator:

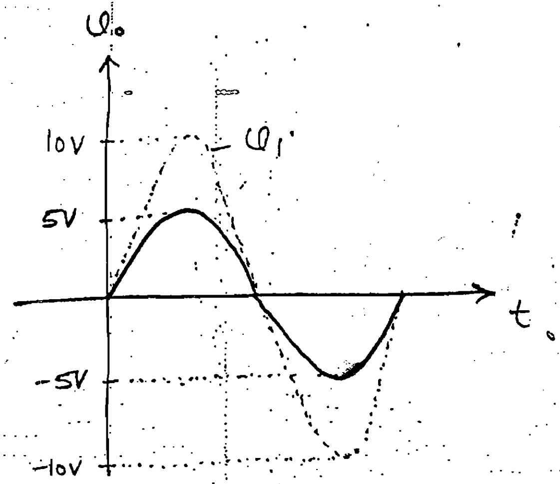


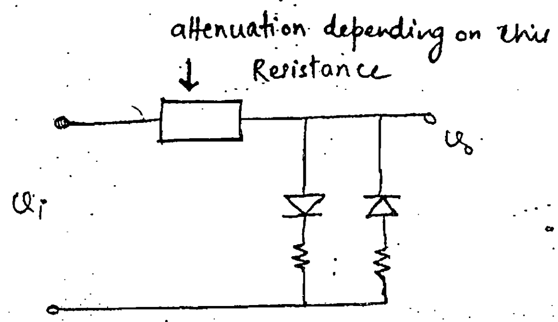


$v_i < 0$, D_1(off), D_2(on), $v_o = \frac{v_i}{2}$

$v_i > 0$, D_1(off), D_2(off), $v_o = \frac{v_i}{2}$

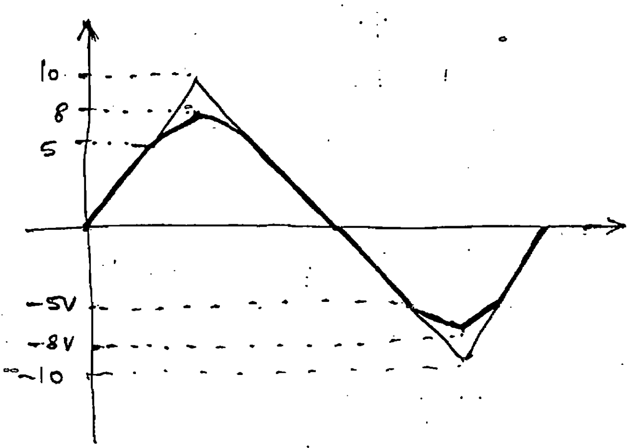




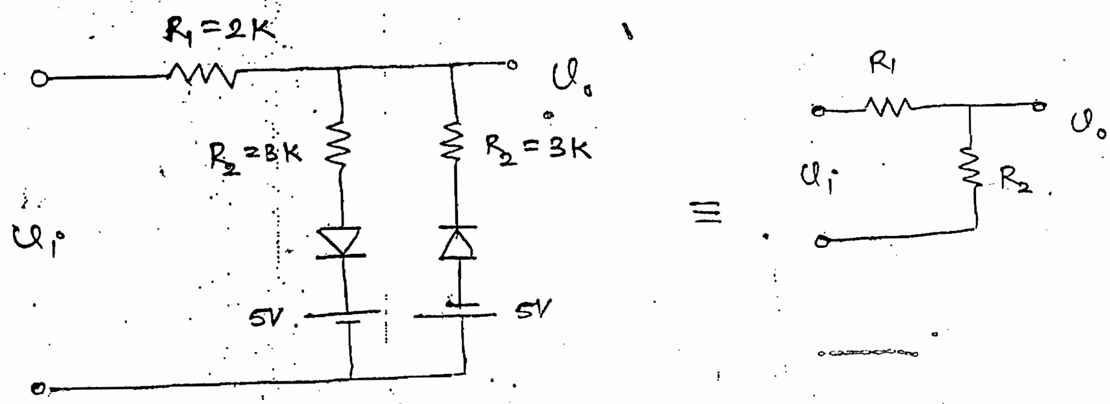


Q. Design an IC diode function generator to generate Δgular wave into sine wave.

Sol.

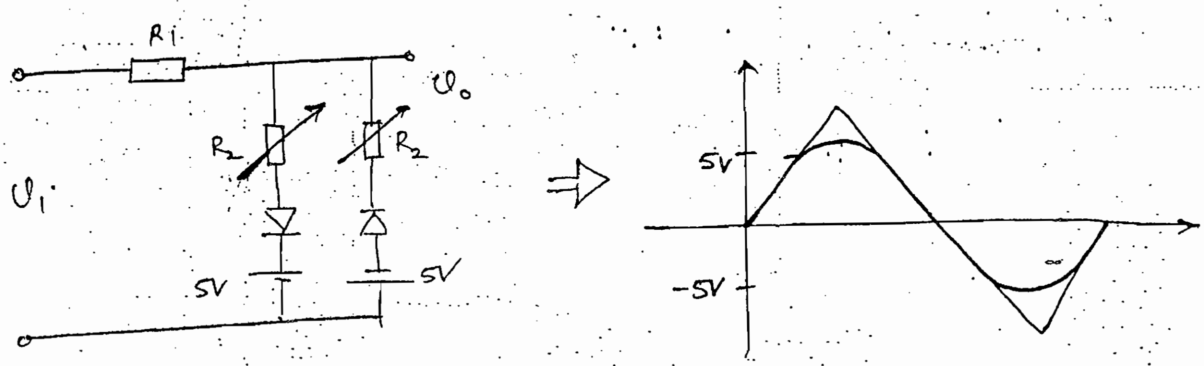


$$\frac{\Delta v_o}{\Delta v_i} = \frac{8-5}{10-5} = \frac{3}{5} = \frac{3K}{2K+3K}$$

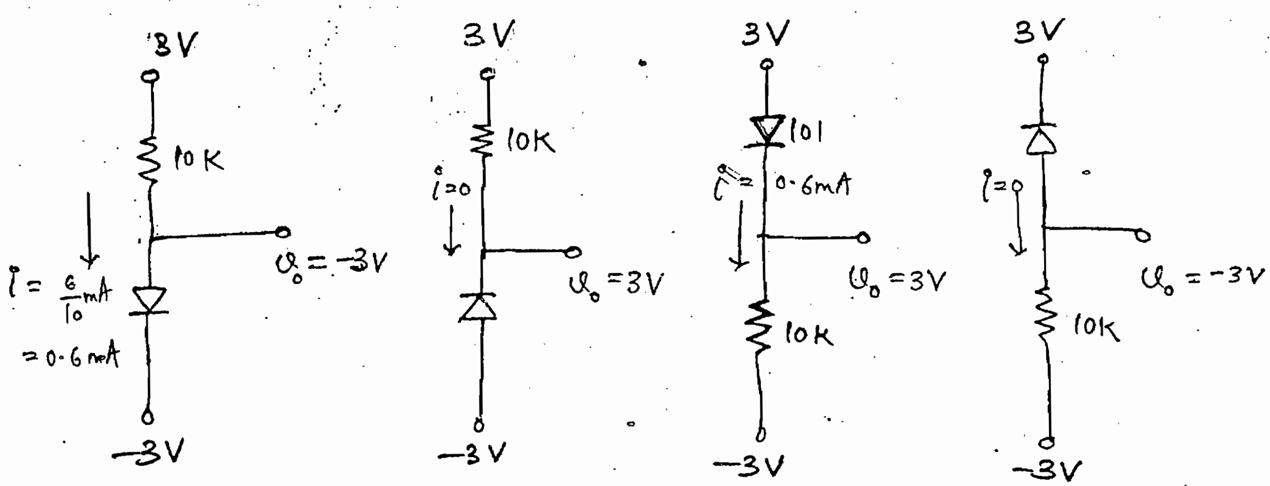


$$v_o = v_i \times \frac{R_2}{R_1+R_2}$$

$$\frac{v_o}{v_i} = \frac{R_2}{R_1+R_2} = \frac{\Delta v_o}{\Delta v_i} = \frac{3}{2+3} \Rightarrow \begin{array}{l} R_1 = 2K \\ R_2 = 3K \end{array}$$

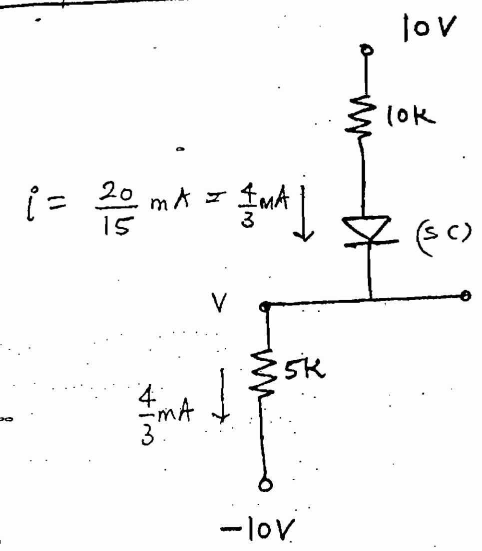


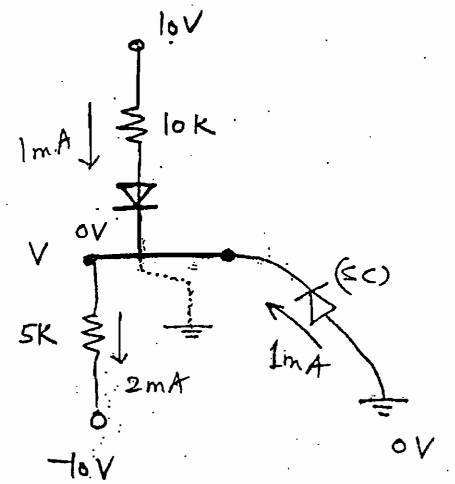
Single Branch designs :



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

Multiple Branch: Reference – Single Branch.





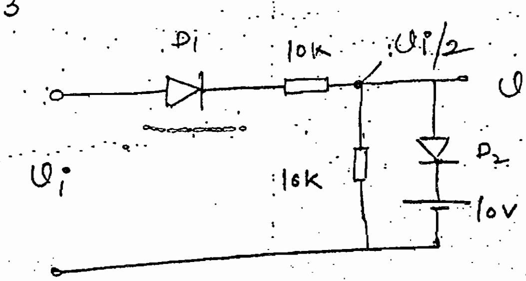
$$V = 10 - \frac{10 \times 20}{15} = -3.3\,V$$

$V = 0\,V$ (due to ground)

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

CH-3

Q1.

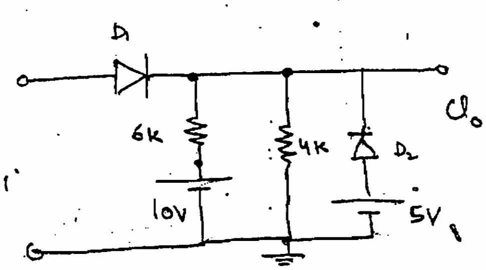


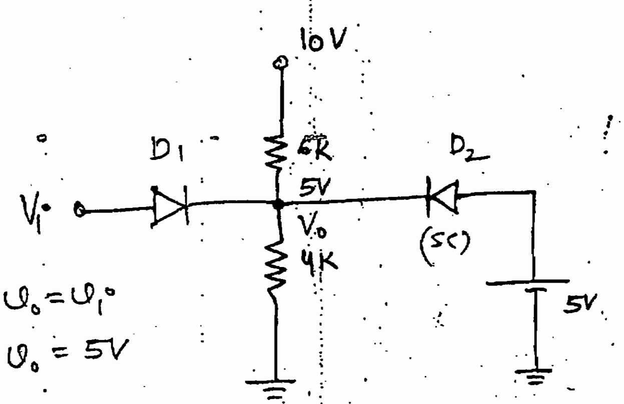
$v_i < 0 \rightarrow D_1$ (off), D_2 (off), $v_o = 0$ ✓

$0 < v_i < 20 \rightarrow D_1$ (on), D_2 (off), $v_o = v_i/2$

$v_i > 20 \rightarrow D_1$ (off), D_2 (on), $v_o = 10\,V$

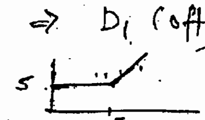
Q7.



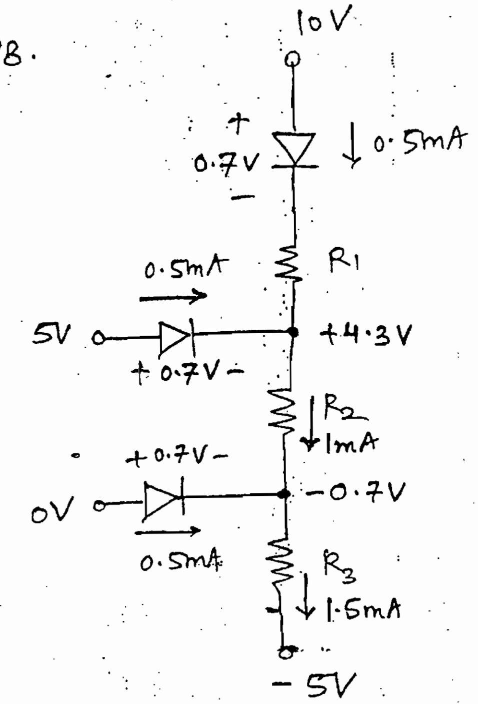


If $V_i > 5 \Rightarrow D_1$ (ON); $v_o = v_i$

If $V_i < 5 \Rightarrow D_1$ (off), $v_o = 5V$



Q8.



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

$$\therefore R_1 = 10\,K\Omega$$

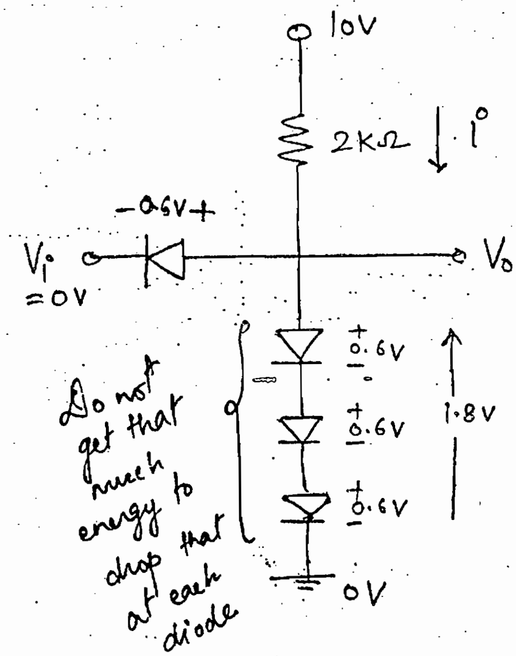
$$R_2 = \frac{4.3 - (-0.7)}{0.5\,mA + 0.5\,mA)}$$

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

$$= 5\,K\Omega$$

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

Q11.



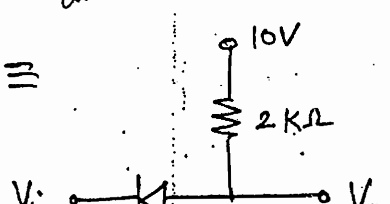
$V_o \neq 1.8V$. $V_o = V_i$

i.e. $V_o = 0.6V$

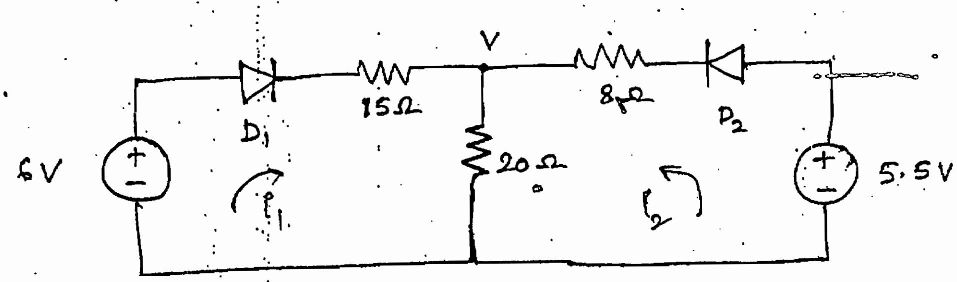
$$i = \frac{10 - 0.6}{2K}$$

$$= 4.7\,mA$$

$\equiv$



Q12.



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

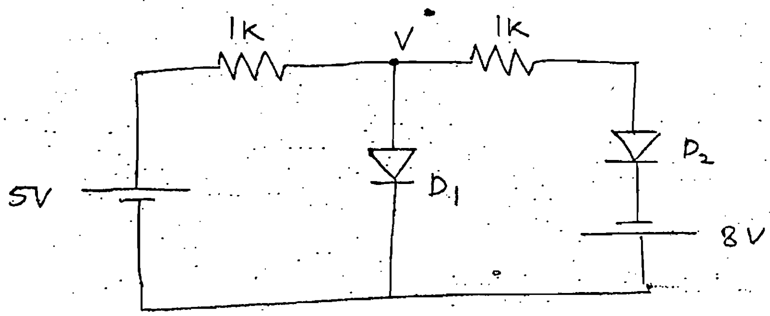
$$V\left(\frac{1}{15}+\frac{1}{20}+\frac{1}{8}\right)=\frac{6}{15}+\frac{5.5}{8}-\frac{0.6}{15}+\frac{0.6}{8}$$

$$V = \quad = 4.024\ V$$

$$i_1 = \frac{6-0.6-4.024}{15} = +ve \quad D_1\ (ON)$$

$$i_2 = \frac{5.5-0.6-4.024}{8} = +ve \quad D_2\ (ON)$$

Q13.



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

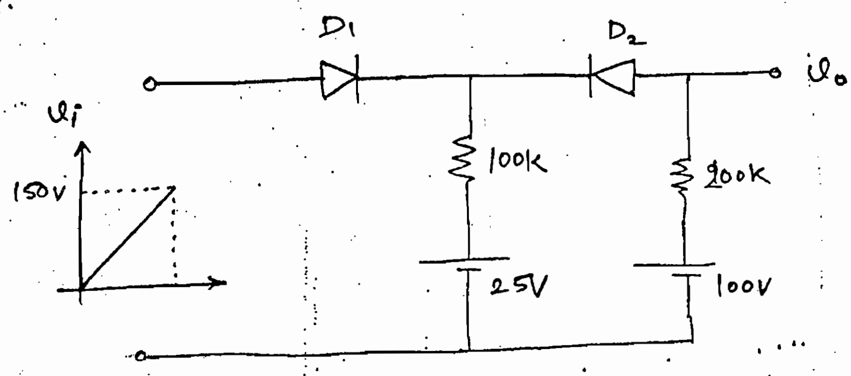
$$2V=-3 \Rightarrow V=-1.5 \text{ volts} \Rightarrow D_1\ (off)$$

Q14.

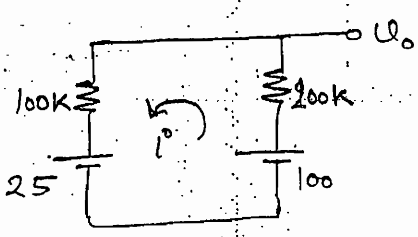
Special problems in clippers:

Ramp i/p problems:

Q.



$v_i = 0$:, D_1 (off), D_2 (ON)



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

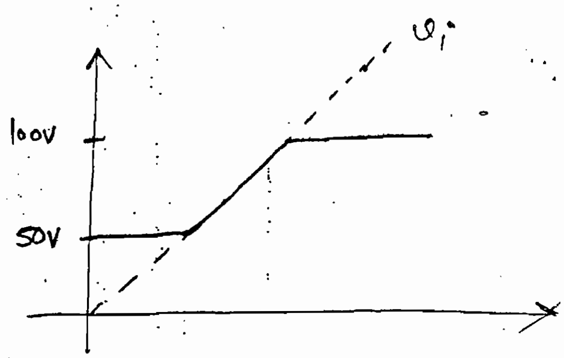
$$v_o = 100 - 200 \times 0.25$$
$$= 50V.$$

$\therefore$ $0 < v_i < 50$ V $\Rightarrow$ D_1 (off) $\Rightarrow$ D_2 (ON) $\Rightarrow$ $v_o = 50V$

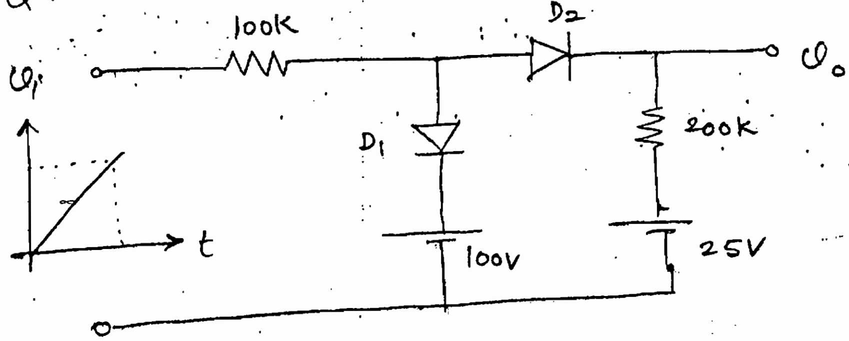
$v_i > 50V$: D_1 (ON) , D_2 (on) $\Rightarrow$ $v_o = v_i$

$v_i < 100V$: D_1 (ON) , D_2 (ON) $\Rightarrow$ $v_o = v_i$

$v_i > 100V$: D_1 (ON), D_2 (off) $\Rightarrow$ $v_o = 100V$

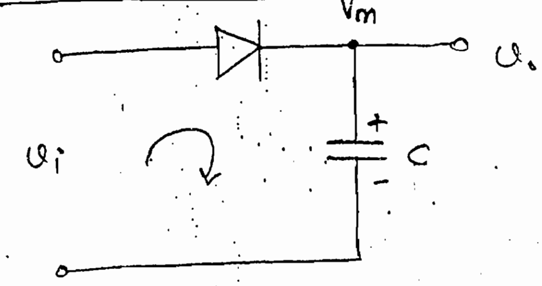


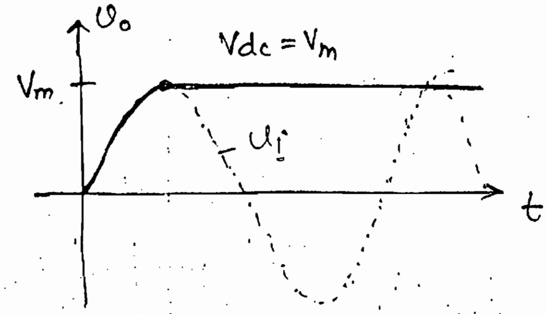
Q.

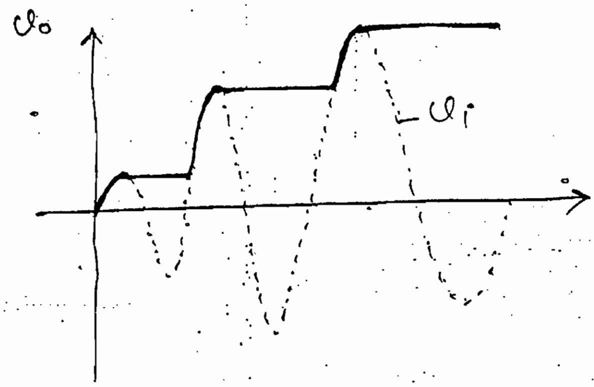


Peak detector, clamper, voltage Multiplier :

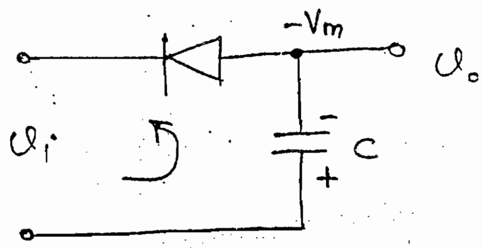
Positive peak detector :

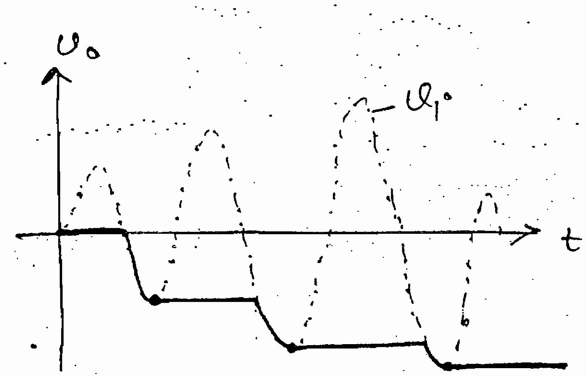




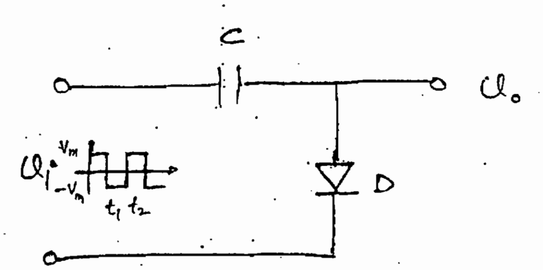


Negative peak detector :





Negative Clamper :



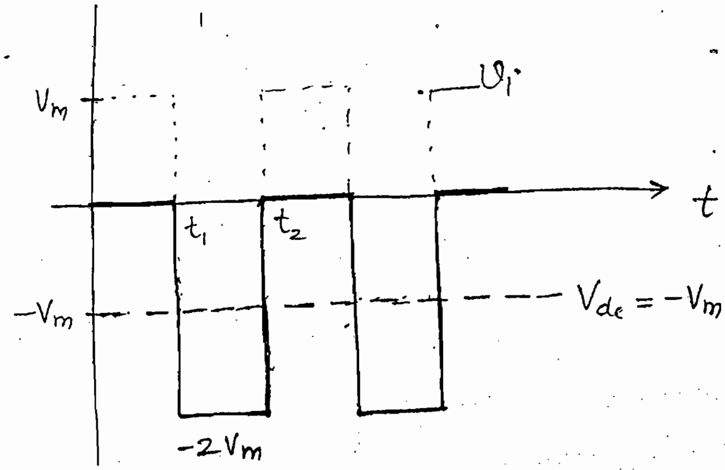
$0 \rightarrow t_1$:

C → get charged to V_m ; D (ON)

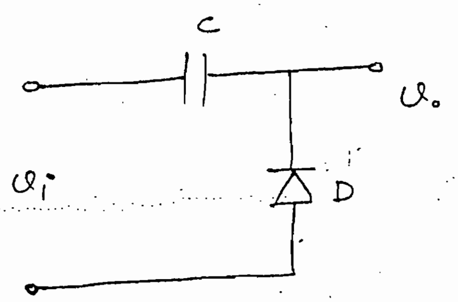
$v_o = 0$

$t_1 \rightarrow t_2$: D (off)

$v_o = -2V_m$



Positive Clamper:



$0 \rightarrow t_1$: D (off)

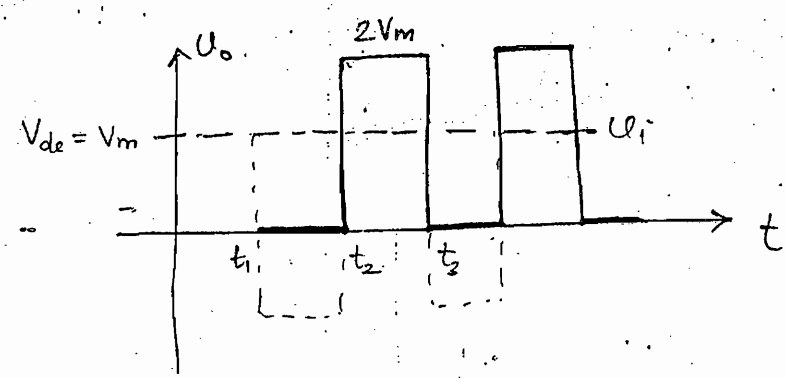
Start with t_1

$C \rightarrow$ unchanged.

$t_1 \rightarrow t_2$: D (on)

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

$t_2 \rightarrow t_3$: D (off) , $v_o = 2V_m$

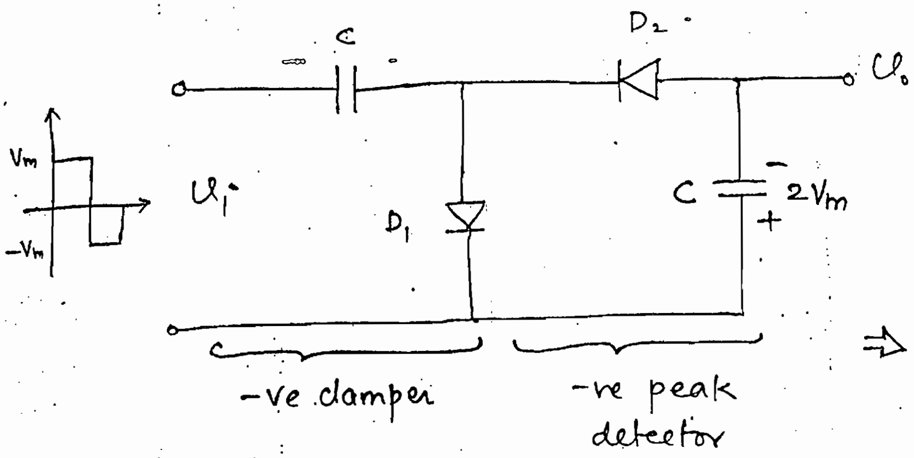


Techniques in clampers:

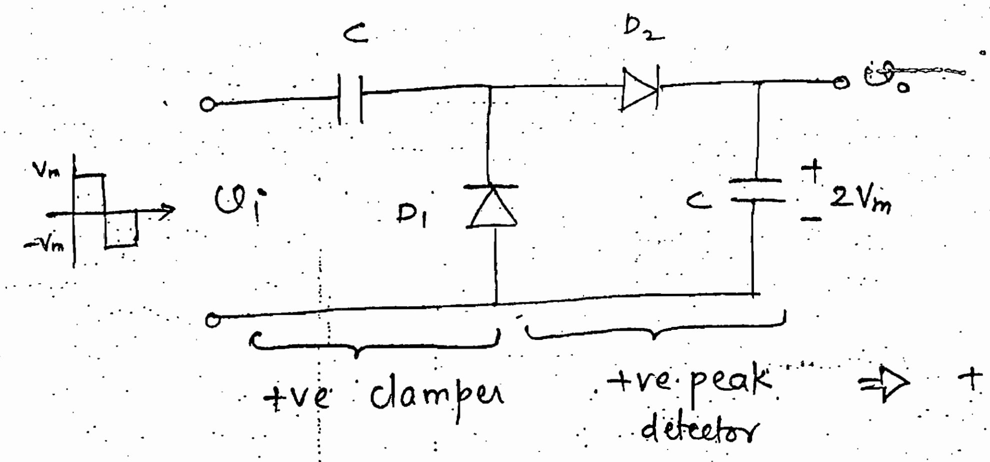
(1). When the diode is in downward dir^n, the total s/g will be clamped (pulling) below the ref. voltage.

(2). When the diode is in ↑ dir^n, the total s/g will be clamped above the ref. voltage.

Voltage multiplier:



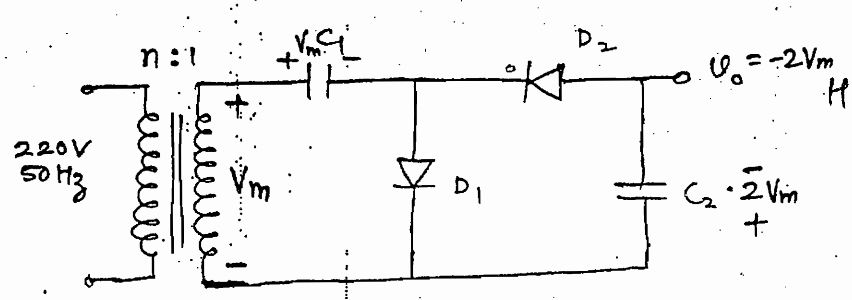
⇒ -ve Voltage Multiplier



⇒ +ve Voltage Multiplier

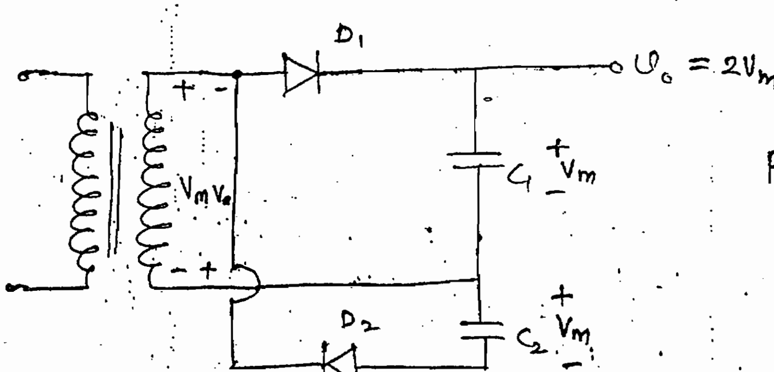
Voltage Multiplier probleme:

(1).



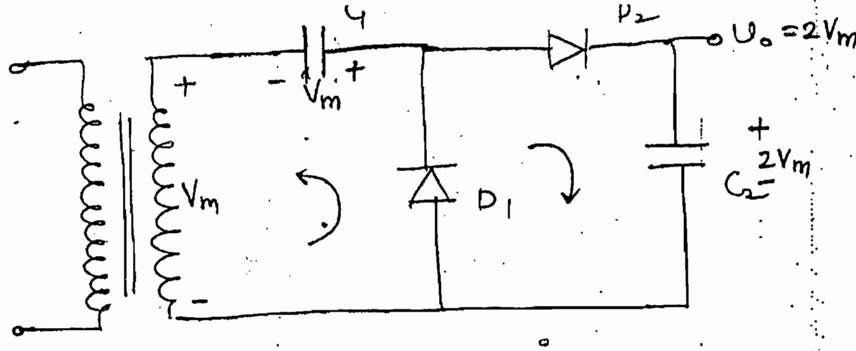
Half wave voltage doubler

(2).



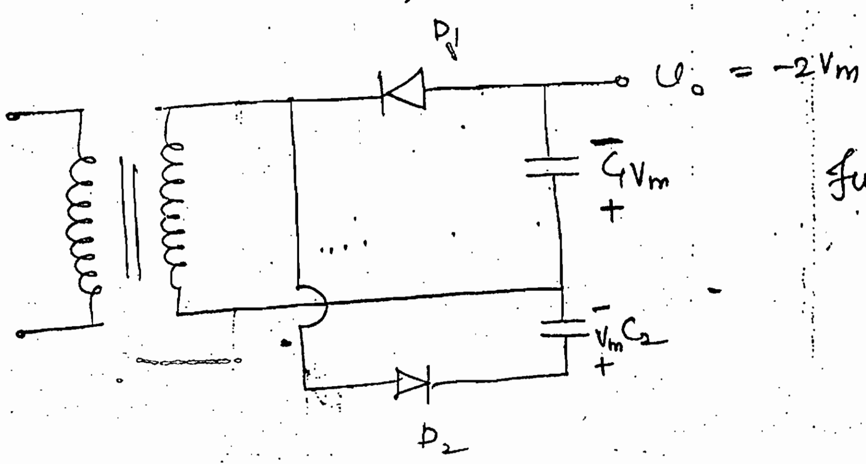
Full wave voltage doubler

(3).



Half wave voltage doubler

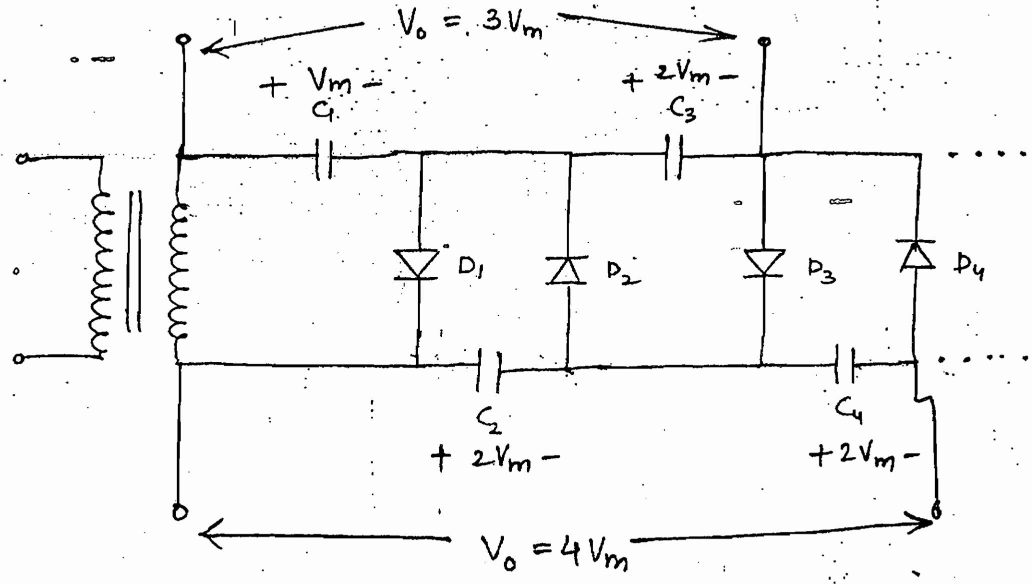
(4).



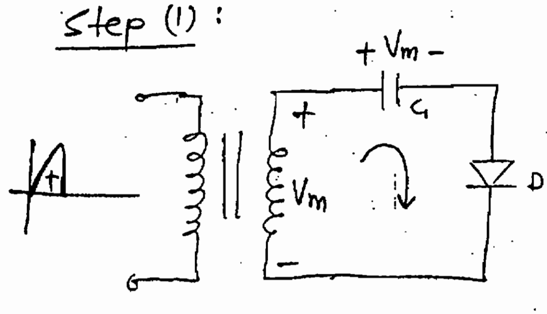
Full wave voltage doubler

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

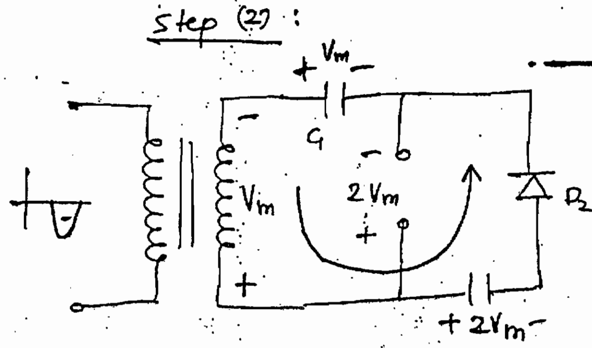
Sol.



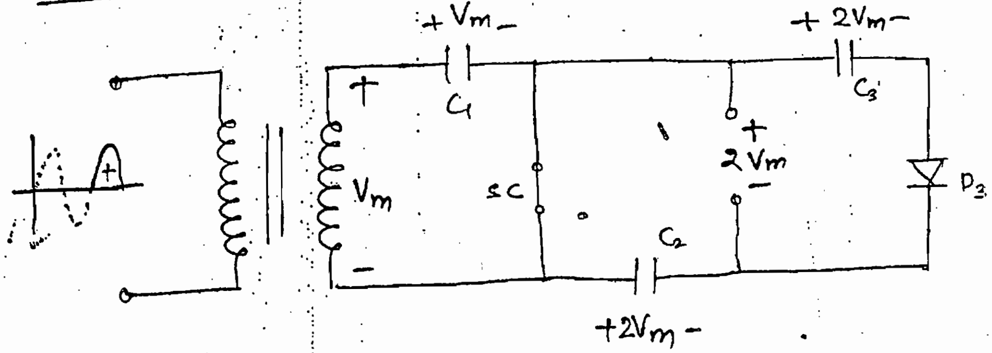
Step (1):



Step (2):

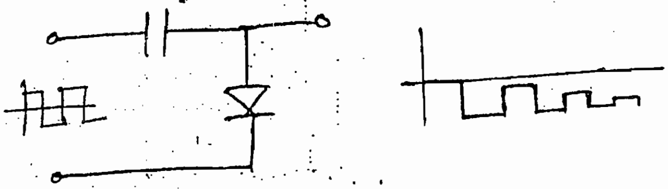


Step (3): 2nd cycle (+ve)

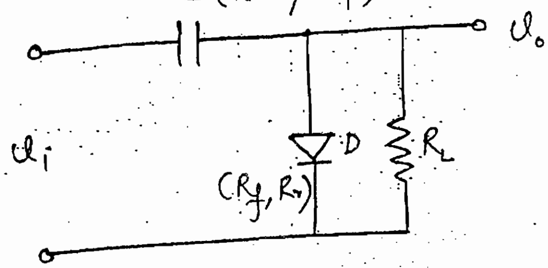


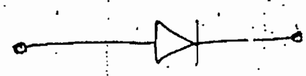
Clampers:

practical cap. (leaky cap.)



C (leaky cap.)





R_f = forward resistance
$= 10\,\Omega$

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

$$\underset{(10\,\Omega)}{R_f} \longrightarrow \underset{(10\,K\Omega)}{R} \longrightarrow \underset{(10\,M\Omega)}{R_r}$$

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

$C = \sqrt{ab}$

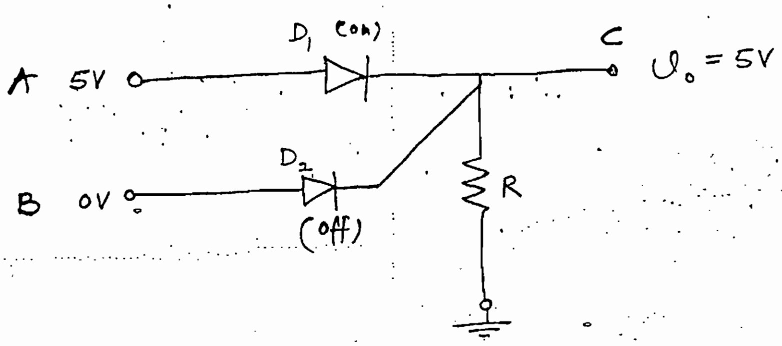
$R = \sqrt{R_f \cdot R_r}$

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

Diode as a digital logic gate :

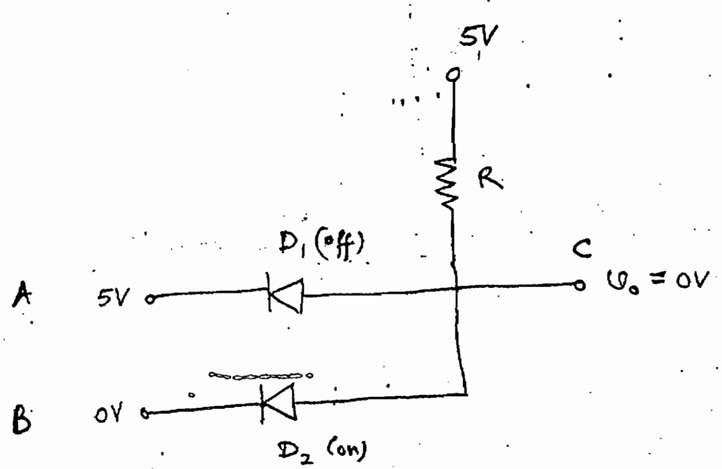
"OR" logic : (A + B = C)

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



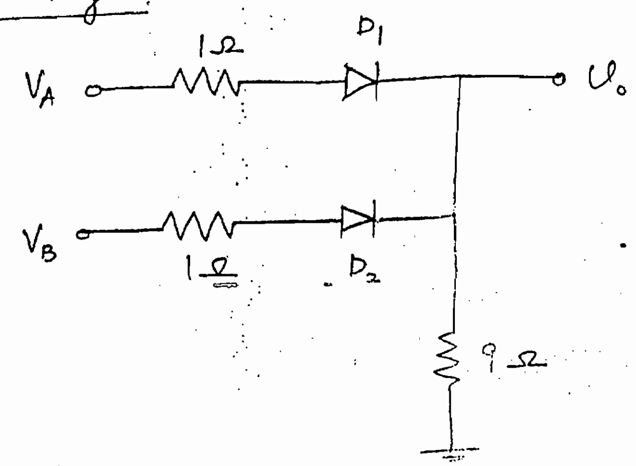
"AND" logic : (A · B = C)

A	B	C
0	0	0
0	5	0
5	0	0
5	5	5

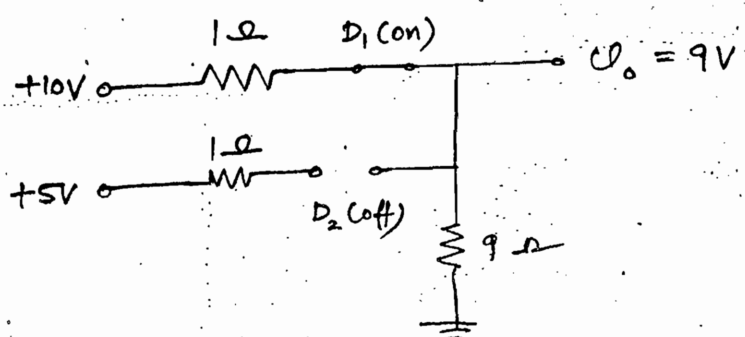


Problems:

"OR" logic:

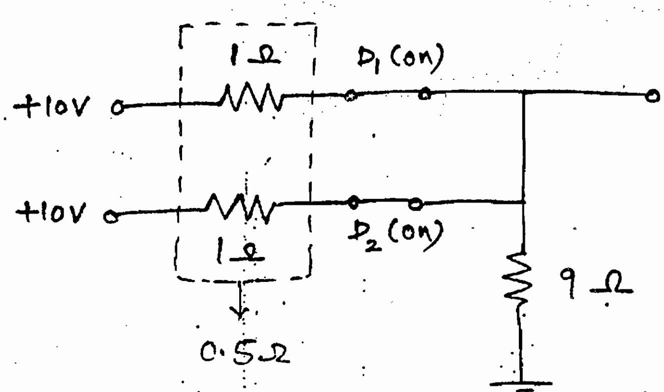


Case :(1): $V_A = 10\,V$, $V_B = 5\,V$



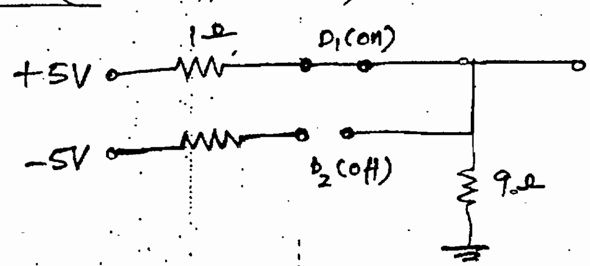
$\left(\frac{10 \times 9}{10} = 9V\right)$

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



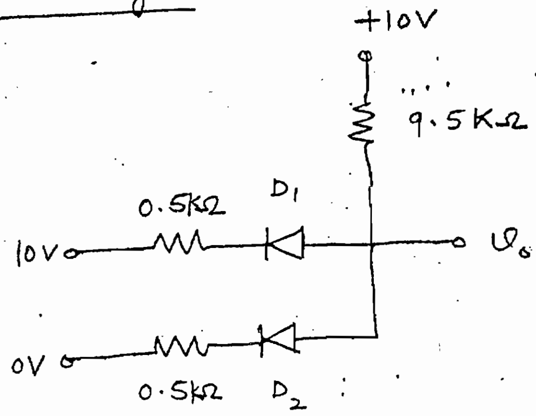
$v_o = \frac{10 \times 9}{9.5} = 9.47V$

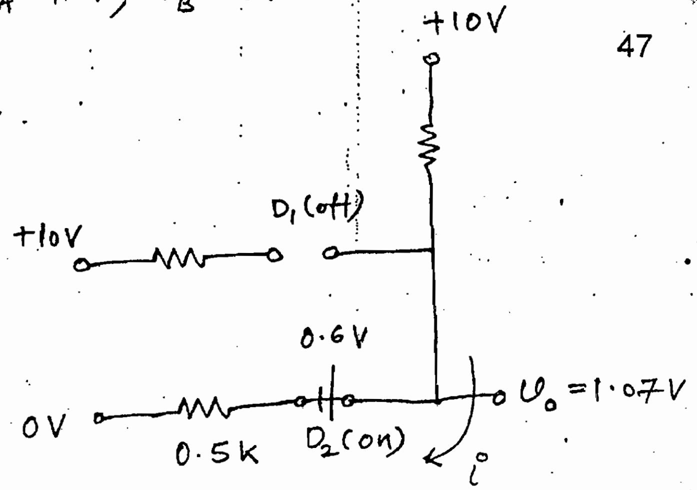
Case (3): $V_A = 5V$, $V_B = -5V$



$v_o = \frac{5 \times 9}{10} = 4.5V$

"AND" logic - Case (1): $V_A = 10V$, $V_B = 0V$



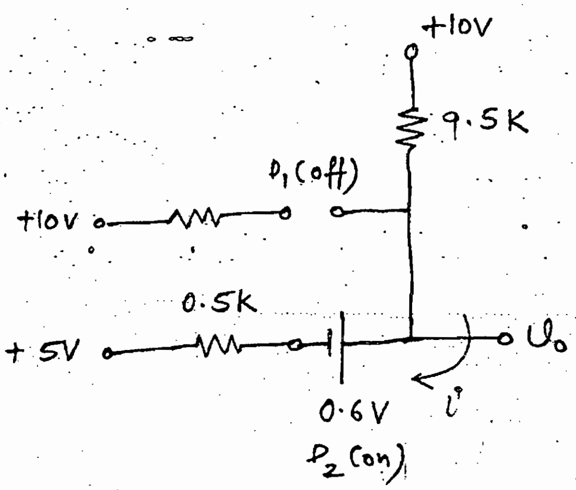


Assume $V_\gamma = 0.6V$.

$$i = \frac{10 - 0.6}{10K} = 0.94\,mA$$

$$V_o = 0.6 + 0.94\,mA \times 0.5K = 1.07V$$

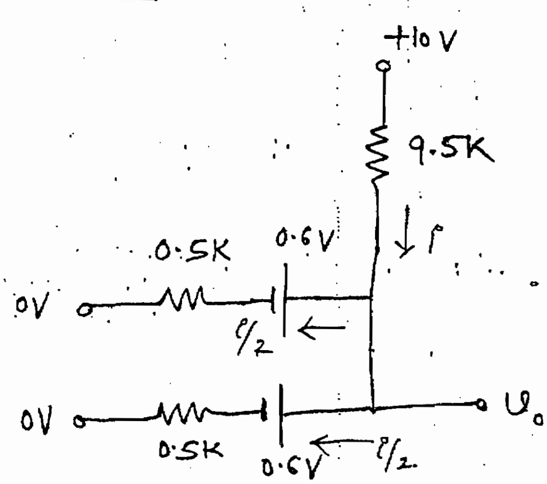
Case (2): $V_A = 10V$ and $V_B = 5V$



$$i = \frac{10 - 0.6 - 5}{10K} = 0.44\,mA$$

$$V_o = 0.6 + 0.44\,mA \times 0.5K + 5 = 5.82V$$

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

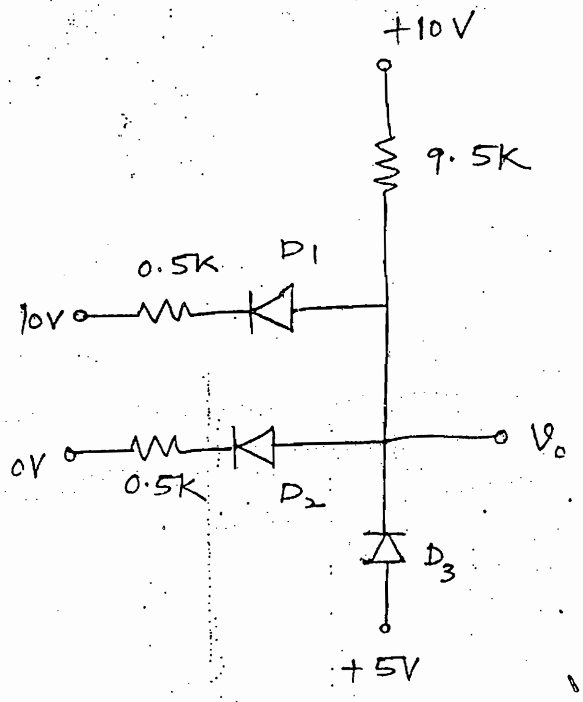


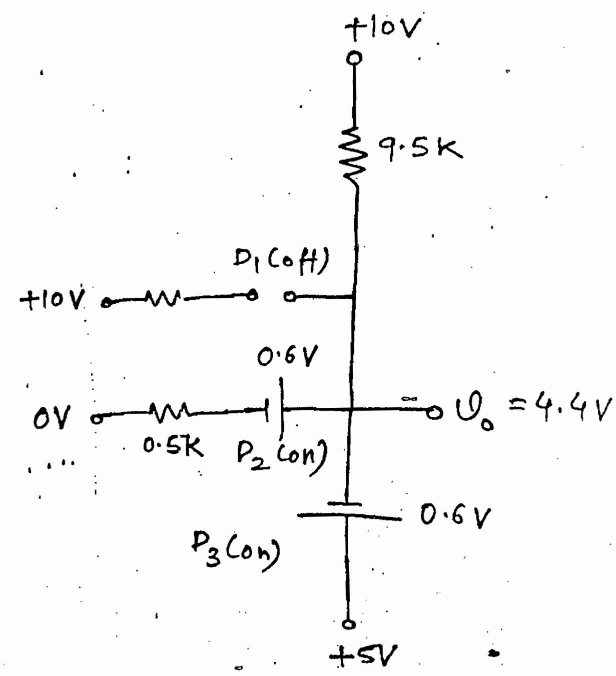
$$10 - 9.5i - 0.6 - 0.5\,i/2 = 0$$

$$i = 0.96\,mA$$

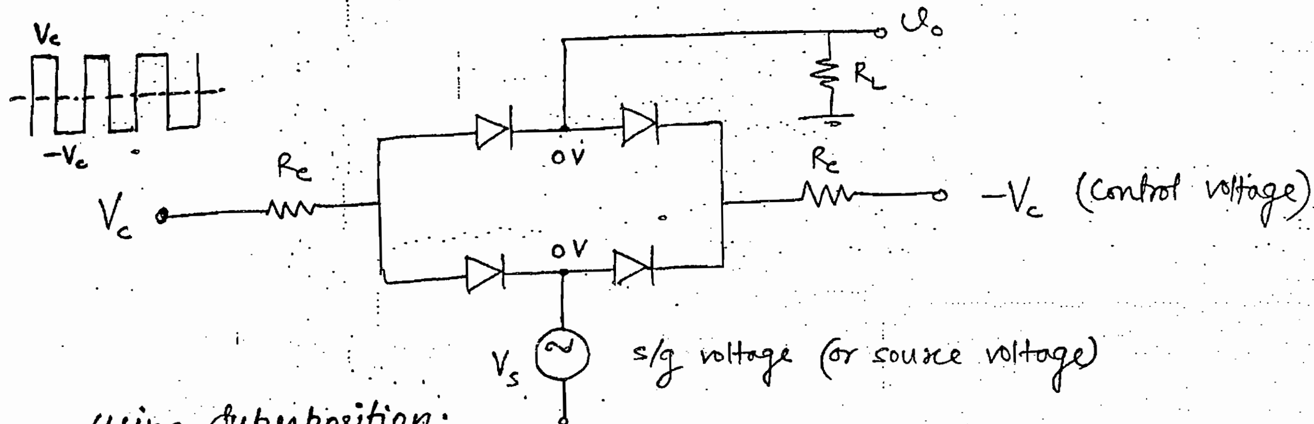
$$V_o = 0.6 + 0.5\,i/2 = 0.842\,V$$

Q.



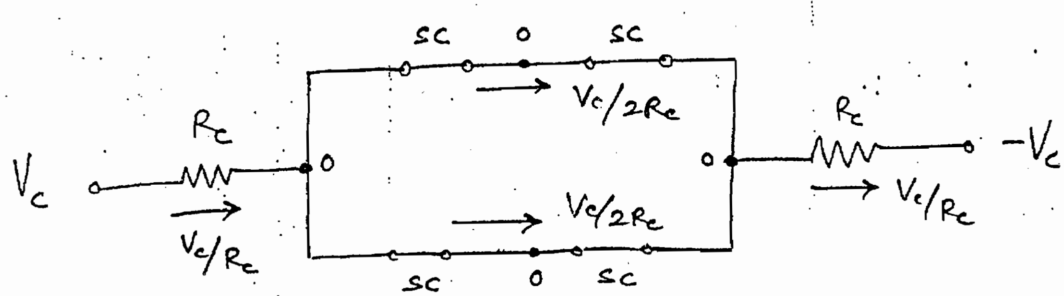


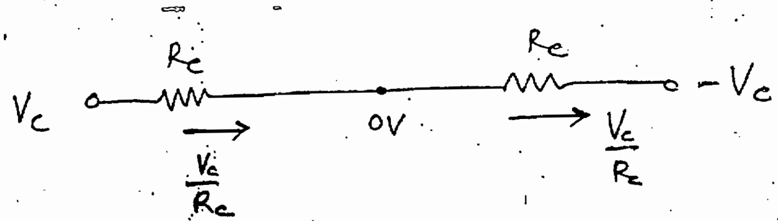
Diode as an analog gate:



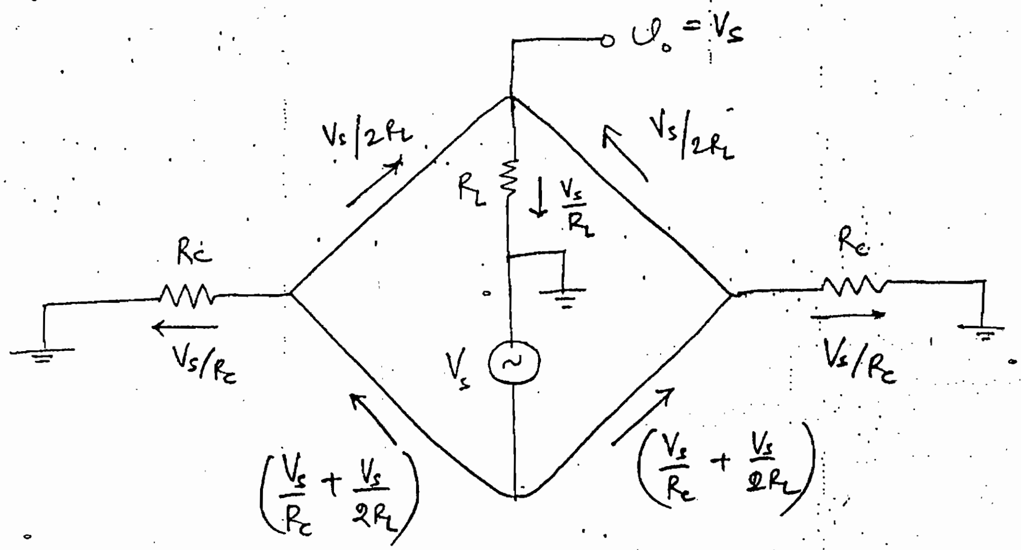
using superposition:

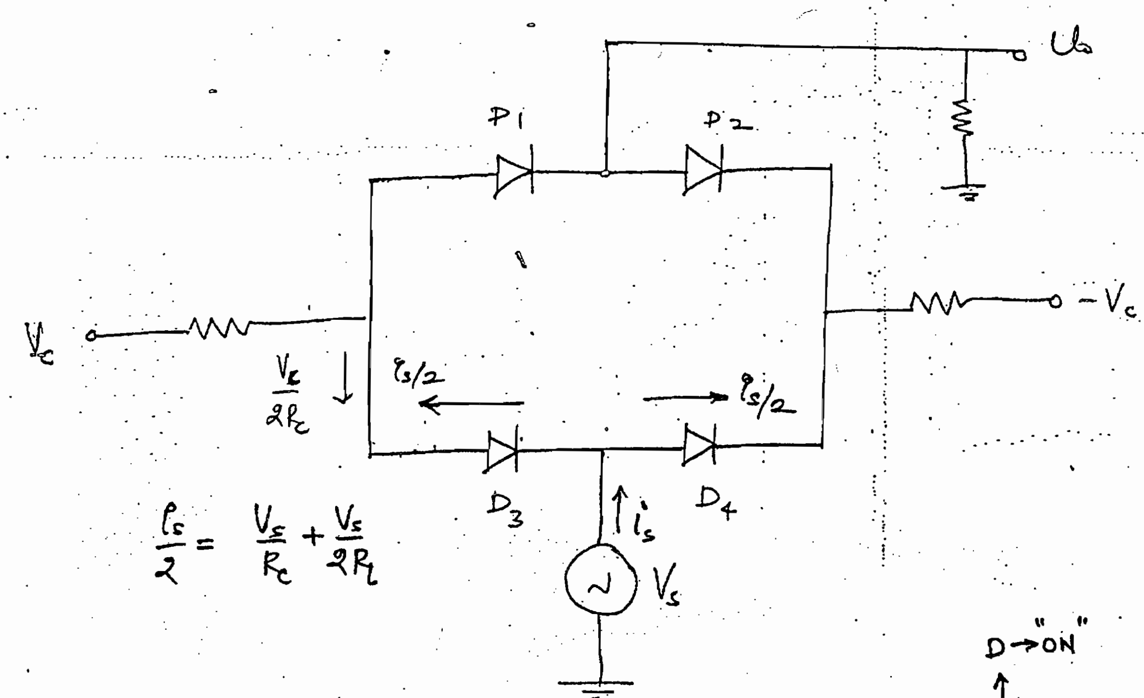
DC Analysis:





AC Analysis:



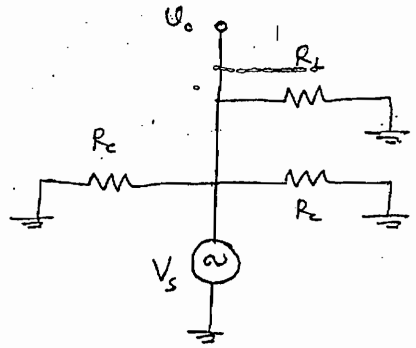


$$\frac{i_s}{2} = \frac{V_s}{R_c} + \frac{V_s}{2R_L}$$

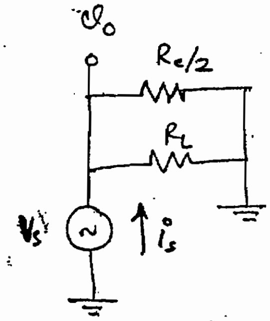
$$\frac{V_c}{2R_c} > \frac{i_s}{2}$$ (Diode will be ON)

D → "ON"

$$\therefore \boxed{\frac{V_c}{2R_c} > \frac{V_s}{R_c} + \frac{V_s}{2R_L}}$$



⇒

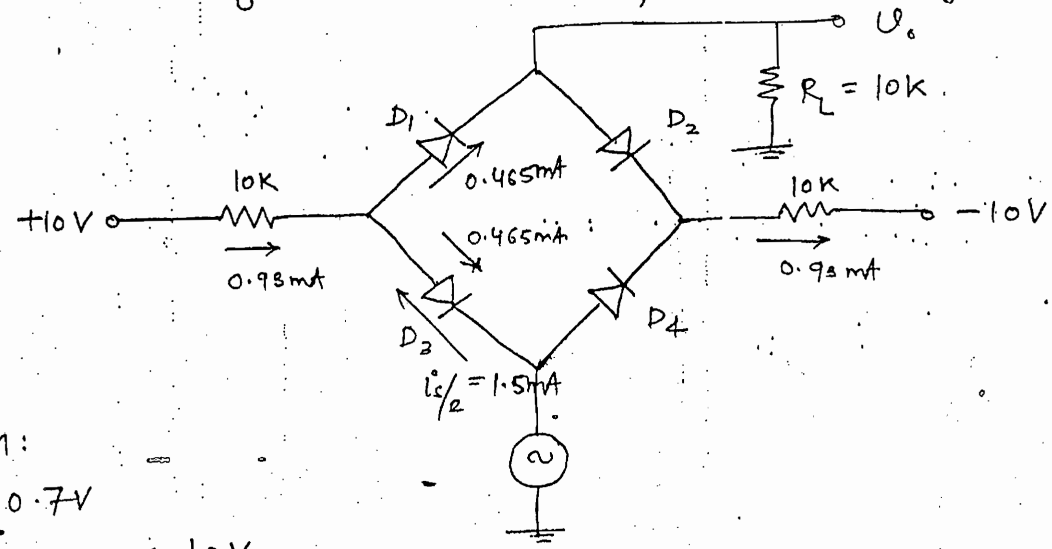


$$i_s = \frac{V_s}{\left(\frac{R_c/2 \cdot R_L}{R_c/2 + R_L}\right)}$$

$$v_o = V_s$$

Problem:

Draw the Transfer characteristics for the ckt given below:

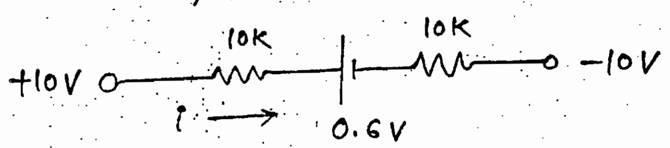


Given:

$V_\gamma = 0.7V$

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

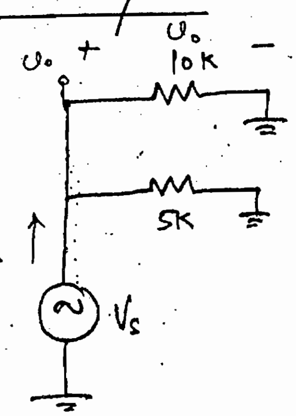
Sol. **D.C Analysis:**



$$i = \frac{10 - 1.4 - (-10)}{20K} = 0.93\ mA$$

$$i/2 = 0.465 mA$$

AC Analysis:



$$i_s = \frac{V_s}{\left(\frac{10 \times 5}{15}\right)} = 0.3 V_s\ mA$$

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

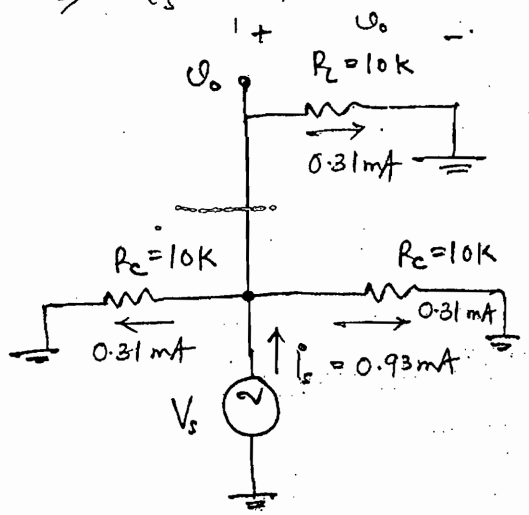
$\therefore i_s = 0.3\ (10) = 3\ mA$

$i_s/2 = 1.5mA$

At this source current, D_3 (off). s/g can be passed only if $i_s/2$ is less than $i/2$ i.e. 0.465mA.

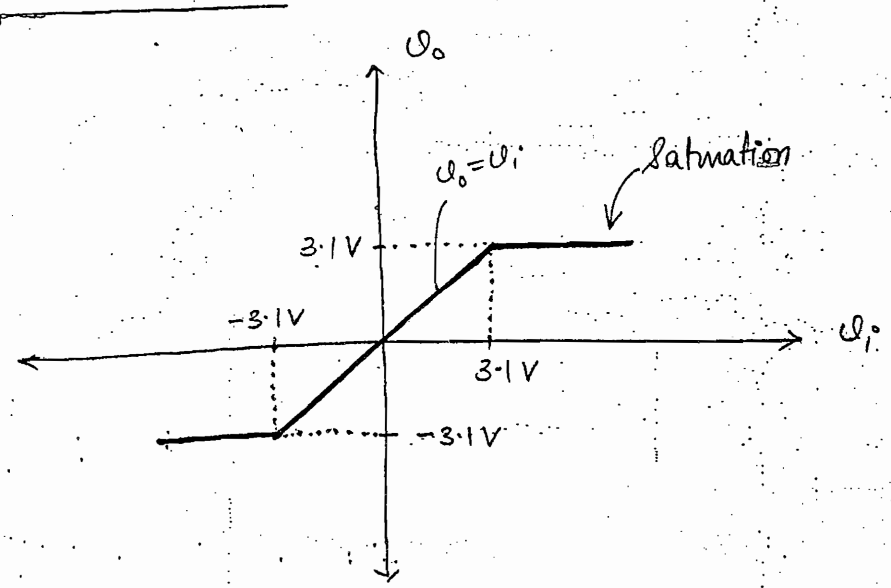
$i_s/2 = 0.465\,mA$

$\Rightarrow i_s = 0.93\,mA$



$v_o = 0.31\,mA \times 10\,K$
$= 3.1$ Volts.

Transfer characteristics :

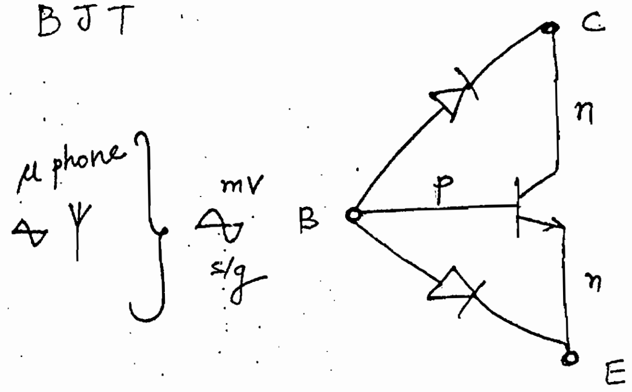


Conclusion:

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

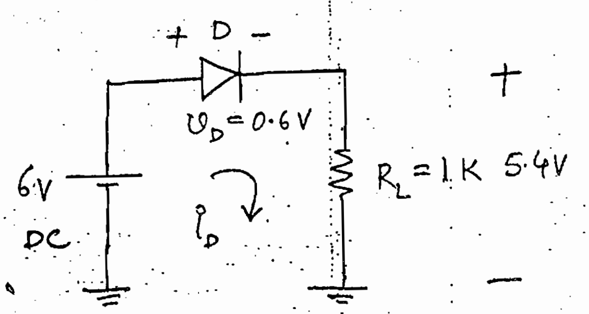
Small s/g Analysis of a Diode:

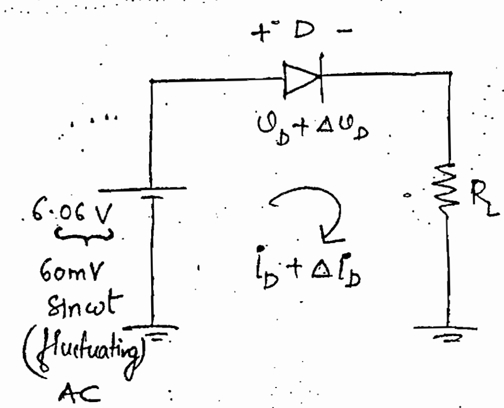
BJT



Diode Resistance

Static Resistance | Dynamic Resistance





$$i_D = \frac{6-0.6}{1K} = 5.4 \text{ mA}.$$

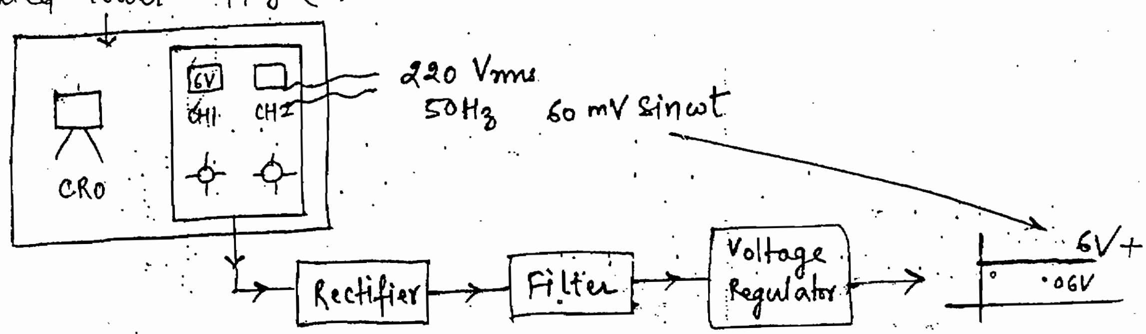
$$R_D = \frac{V_D}{I_D} = \frac{0.6 \text{ V}}{5.4 \text{ mA}}$$

static/DC Resistance = 111 Ω

$$r_D = \frac{\Delta V_D}{\Delta I_D}$$

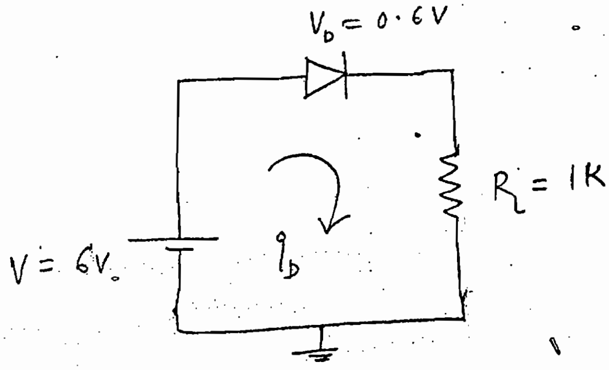
dynamic/AC Resistance

Regulated Power supply (DC):



Analysis:

Static Resistance



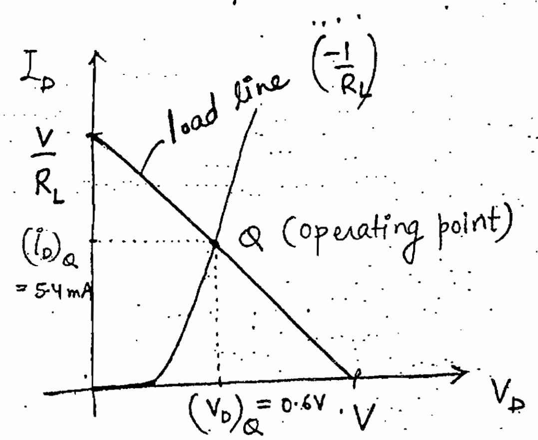
Mathematical Analysis:

$$I_D = \frac{V - V_D}{R_L}$$

$$= \frac{6 - 0.6}{1K}$$

$$= 5.4 mA$$

Graphical Analysis:



$$I_D = \frac{V - V_D}{R_L}$$

$$I_D = -\frac{V_D}{R_L} + \frac{V}{R_L}$$

$$y = mx + c$$

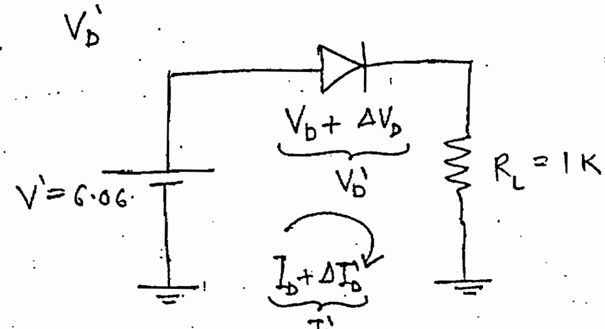
$$m = \text{slope} = -1/R_L$$

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

* As I_D increases, R_D decreases.

Design:

Suppose due to a small s/g variation ($V_i = 60mV \sin\omega t$) The DC voltage $V = 6$ volts is changed to $V' = 6.06$ volts. Then find the new diode current I_D' & new diode voltage V_D'



Sol. $V' = V + \Delta V$

$6.06V = 6V + 0.06V$

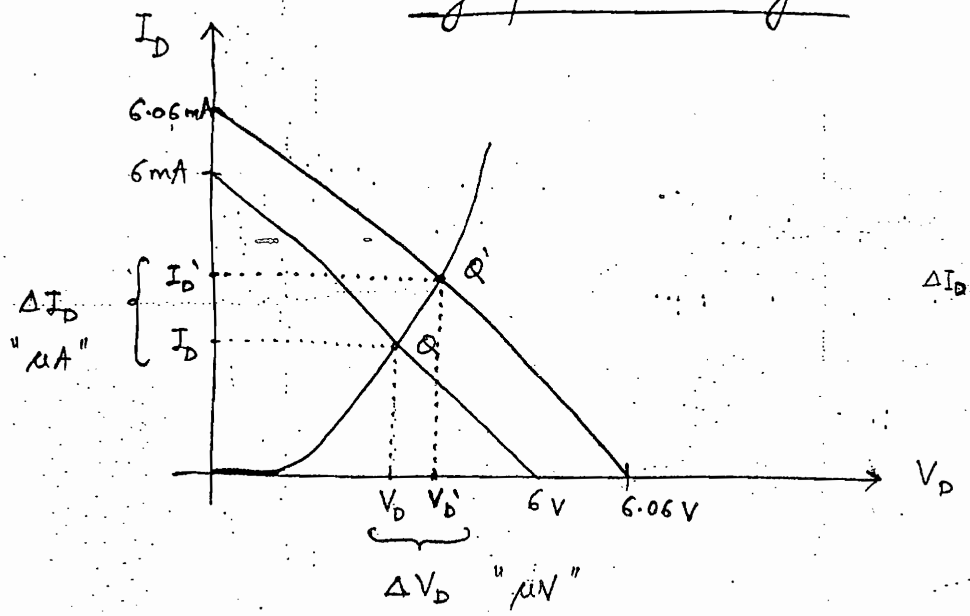
$I_D' = I_D + \Delta I_D$

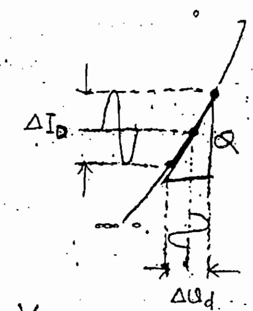
? $\quad$ 5.4 mA $\quad$?

$V_D' = V_D + \Delta V_D$

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

<u>graphical Analysis:</u>





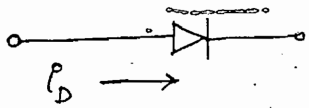
As the change in V_D and I_D are at μ level ie. very small. Thats why we do not go for graphical methods.

<u>Note</u>:

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.

Mathematical Analysis:



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

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

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

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

$\Delta V_D <<< V_T$ (small s/g analysis)

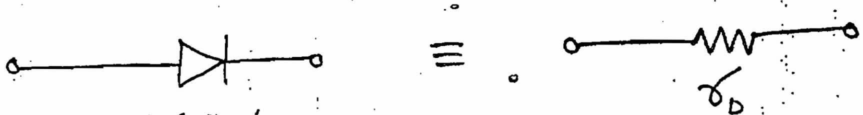
$$I_D + \Delta I_D = I_D \left(1 + \frac{\Delta V_D}{V_T}\right)$$

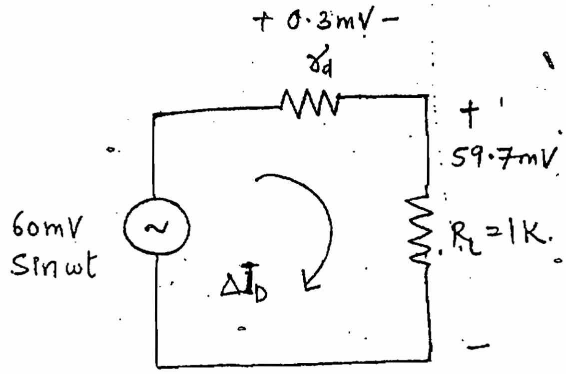
$$\cancel{I_D} + \Delta I_D = \cancel{I_D} + I_D \cdot \frac{\Delta V_D}{V_T}$$

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

$$\boxed{r_D = \frac{\Delta V_D}{\Delta I_D} = \frac{V_T}{I_D(DC)}}$$

Small s/g Equivalent model:





$$r_D = \frac{V_T}{I_D}$$

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

$$= 5\,\Omega$$

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

$$\Delta V_D = \Delta I_D \cdot r_D$$

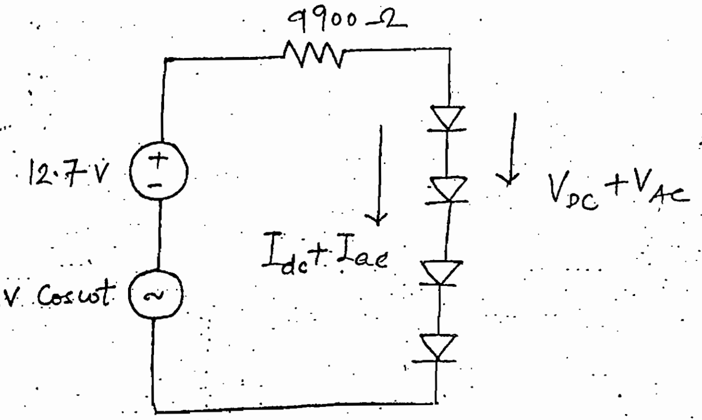
$$= 60\,\mu A \sin\omega t \cdot (5\,\Omega)$$

$$= 300\,\mu V \cdot \sin\omega t$$

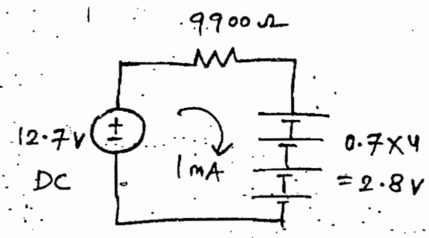
$$= 0.3\,mV \sin\omega t$$

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

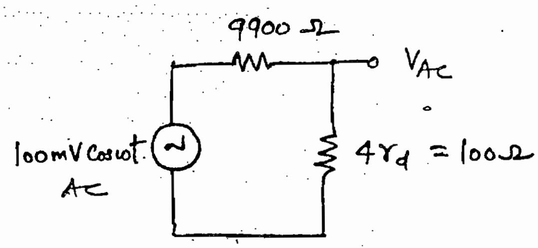
Q3.



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



$$r_D = \frac{V_T}{I_{dc}}$$



$$= \frac{25\,mV}{1\,mA}$$

$$= 25\,\Omega$$

$$V_{AC} = 100\,mV \cos\omega t \times \frac{100\,\Omega}{(9900+100)\,\Omega}$$

$$= 100\,mV \cos\omega t \cdot \frac{\times 100}{10{,}000}$$

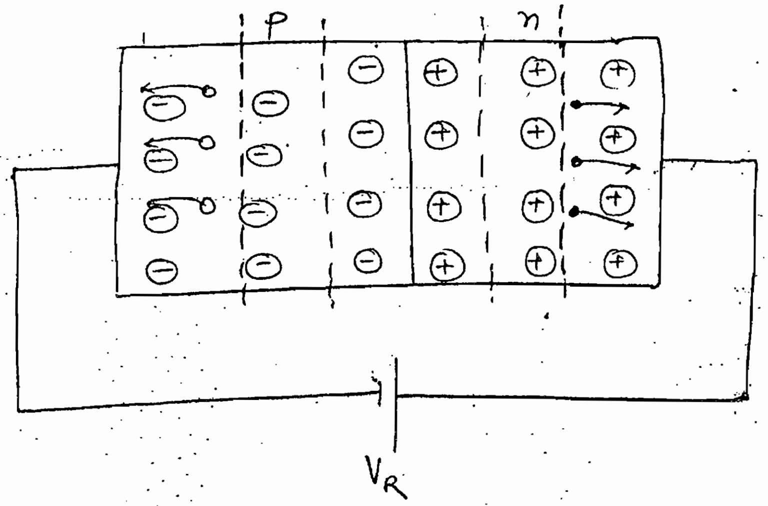
$$= 1\,mV \cos\omega t$$

Junction Capacitance:

Transition Capacitance (C_T)
(RB)
(Transition means depletion which always occur 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



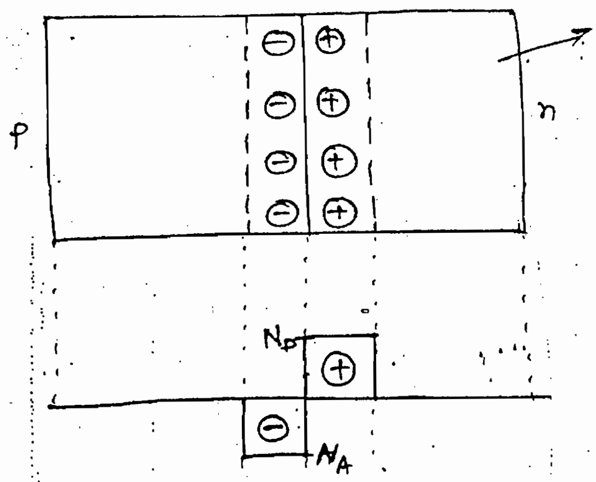
"The rate of change of charge w.r.to applied voltage" – Capacitance

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

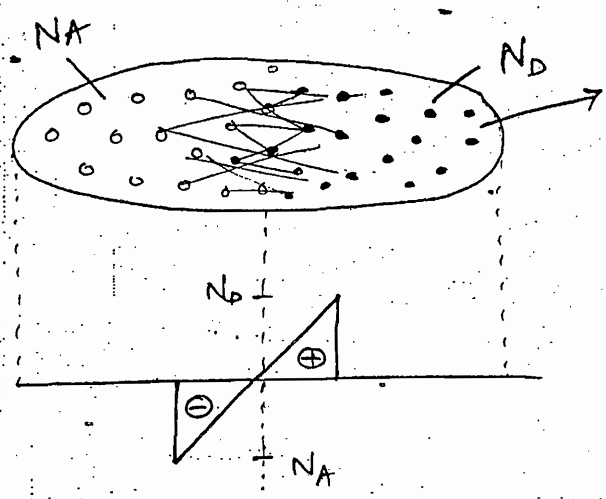
$$C_{||} = \frac{A\epsilon}{d}$$

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

$\omega \propto \sqrt{V_j} \rightarrow$ Alloy type
$\omega \propto \sqrt[3]{V_j} \rightarrow$ Grown jxn type
} OC jxn



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



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

from OC to RB –

$\omega \propto \sqrt{V_j + V_R}$ – alloy type

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

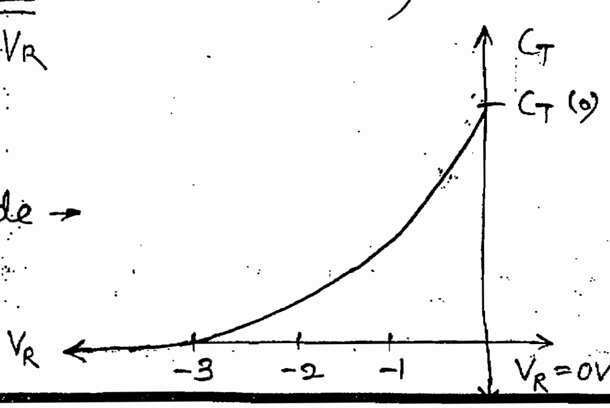
$C_T \propto \frac{1}{\sqrt{V_R}}$

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

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

C_T (0 pF → 3 pF)

ordinary diode →

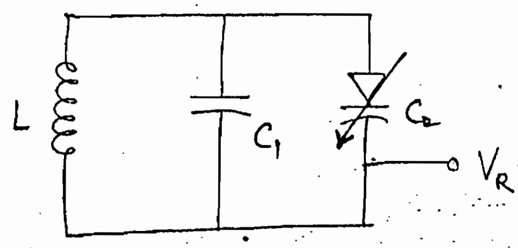


Varactor diode (or) Varicap

Power diode : C_T (5 pF → 300 pF)

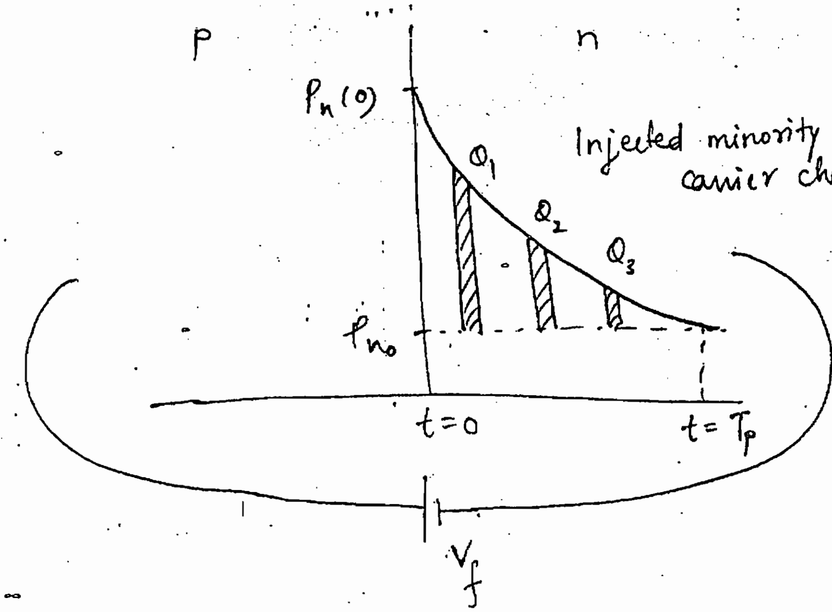
Tuning ckt :

VCO → Voltage Controlled Oscillator



$$f = \frac{1}{2\pi\sqrt{LC_T}} \quad \& \quad C_T = C_1 + \underset{\text{depends on } V_R}{\underbrace{C_2}}$$

Diffusion capacitance (C_D) : (or) storage capacitance



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

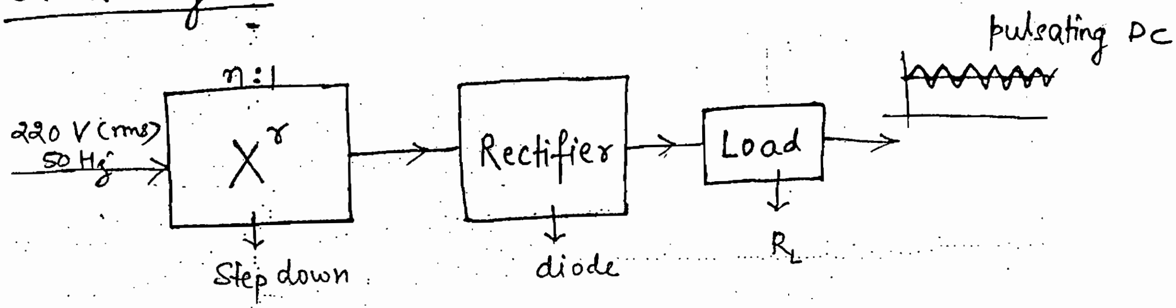
$$Q = I_p \times T_p$$

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

$$C_D \cong 100\ pF$$

Rectifiers:

Block Diagram:



It is a ckt which converts AC to pulsating DC.

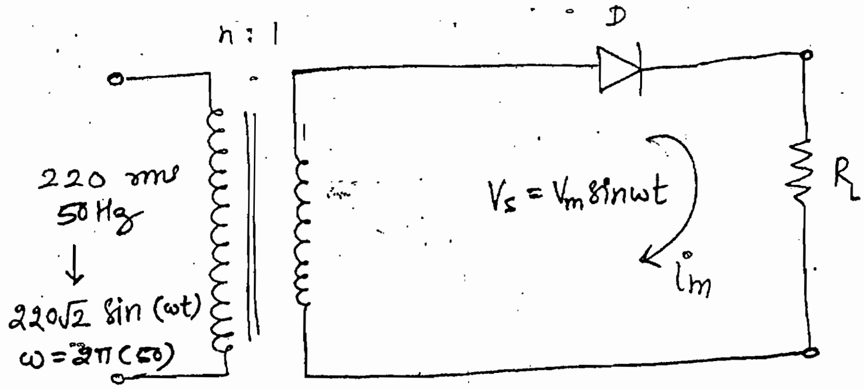
Practical Rectifier designs are:

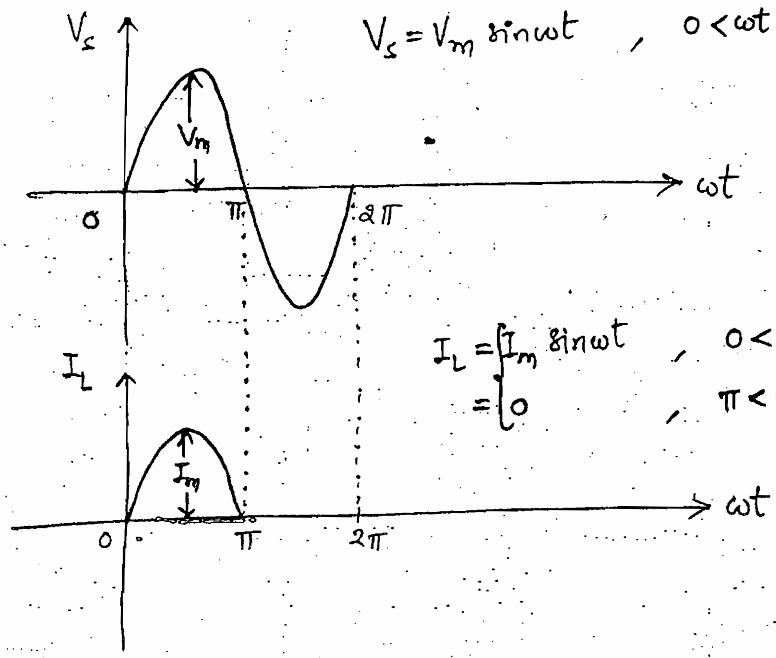
(1). Half wave Rectifier
(2). Full wave Rectifier with centre tap
(3). Bridge Rectifier

Rectifier parameters:

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

(1). Half Wave Rectifier : (HWR)





$V_s = V_m \sin\omega t \quad , \quad 0 < \omega t < 2\pi$

$$I_L = \begin{cases} I_m \sin\omega t & , \quad 0 < \omega t < \pi \\ 0 & , \quad \pi < \omega t < 2\pi \end{cases}$$

(1). $\underline{I_{dc}}$:

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

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

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

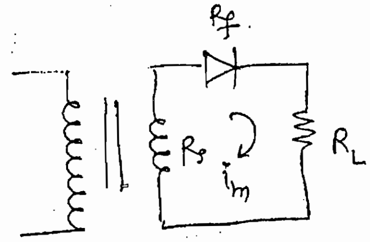
$$I_{dc} = \frac{I_m}{\pi}$$

(2). $\underline{V_{dc}}$:

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

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

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



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

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

(3). $\underline{I_{rms}}$:

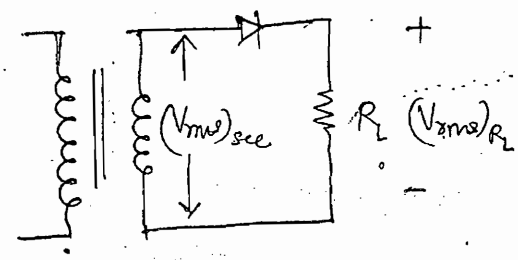
$$I_{rms} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} I^2 \, d\alpha}$$

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

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

$$(V_{rms})_{R_L} = V_m/2$$

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



(4). Ripple factor :

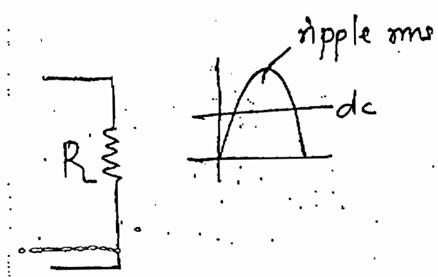
$$rf = \frac{\text{rms value of alternating component}}{\text{average value}}$$

$I' \to$ ac current

$I_{dc} \to$ dc current

$I \to$ total current

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



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

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

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

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

$$I'^2_{rms} = \frac{1}{2\pi} \int_0^{2\pi} (I - I_{dc})^2 \, d\alpha$$

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

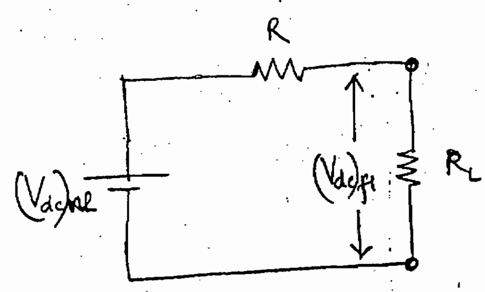
$$\gamma f = \frac{I'_{rms}}{I_{dc}} = \frac{\sqrt{I_{rms}^2 - I_{dc}^2}}{I_{dc}} = \sqrt{\left(\frac{I_{rms}}{I_{dc}}\right)^2 - 1}$$

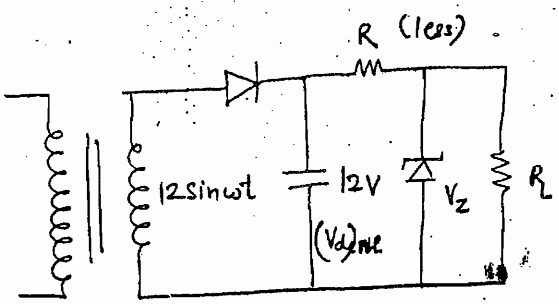
$$\gamma f = \sqrt{\left(\frac{I_m/2}{I_m/\pi}\right)^2 - 1}$$

$$= \sqrt{\left(\frac{\pi}{2}\right)^2 - 1}$$

$$= 1.21 \quad (\text{or}) \quad 121\%$$

(5) <u>Voltage Regulation:</u>





$$\% VR^n = \frac{(V_{dc})_{nl} - (V_{dc})_{fl}}{(V_{dc})_{fl}}$$

* Ideally % VR^n should be zero.

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

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

$$\underbrace{I_{dc}(R_f + R_s) + I_{dc} \cdot R_L}_{(V_{dc})_{fl}} = \underbrace{\frac{V_m}{\pi}}_{(V_{dc})_{nl}}$$

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

(6). <u>Rectification Efficiency (η)</u>:

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

$$= \frac{P_{DC}}{P_{AC}} = \frac{I_{dc}^2 \cdot R_L}{I_{rms}^2 \cdot (R_f + R_s + R_L)}$$

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

If $R_L >>> (R_f + R_s)$

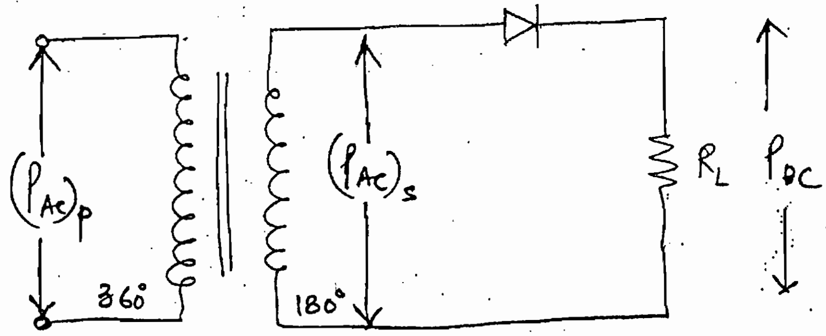
$$\eta \cong \frac{I_{dc}^2}{I_{rms}^2}$$

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

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

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

(7). TUF :



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

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

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

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

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

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

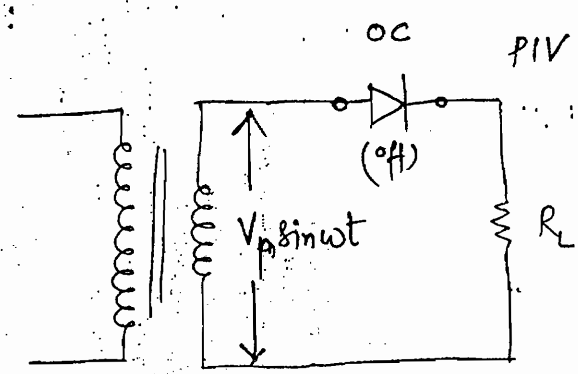
If $R_L >>> (R_f + R_s)$,

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

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

(8). PIV :

$PIV = V_m$



(9). Form factor (FF) :

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

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

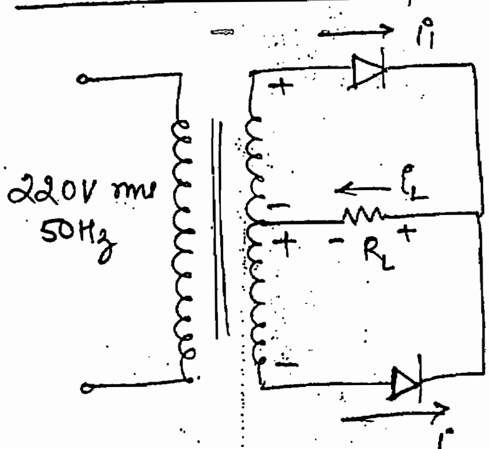
$$= \pi/2 = 1.57$$

(10). Peak factor (PF) :

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

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

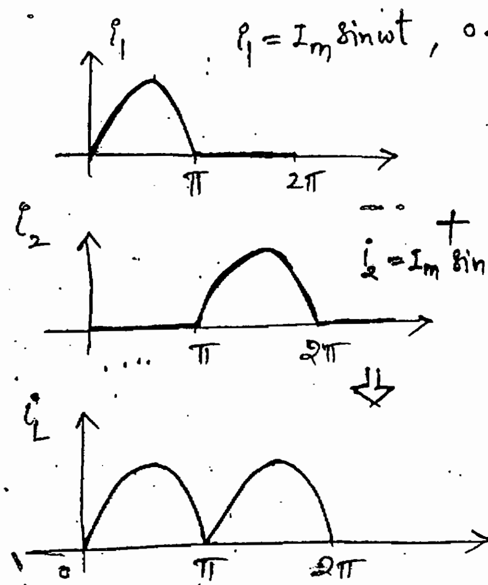
(2). Full Wave Rectifier :



(Centre Tapped)

$i_1 = I_m \sin \omega t \;, \; 0 < \omega t < \pi$

$i_2 = I_m \sin \omega t, \; \pi < \omega t < 2\pi$



Rectifier parameters

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

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

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

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

$= \sqrt{\frac{(I_m/\sqrt{2})^2}{(2 I_m/\pi)^2} - 1}$

$= \sqrt{\frac{\pi^2}{8} - 1} = 0.48$

$= 48\%$

(5). $\underbrace{I_{dc}}_{2 I_m/\pi} \, (\underbrace{R_f + R_s}_{R_s/2}) + \underbrace{I_{dc} R_L}_{2 V_m/\pi} = \underbrace{V_m/\pi}_{2 V_m/\pi}$

$VR^n =$

(6). $\eta = \frac{P_{DC}}{P_{AC}}$

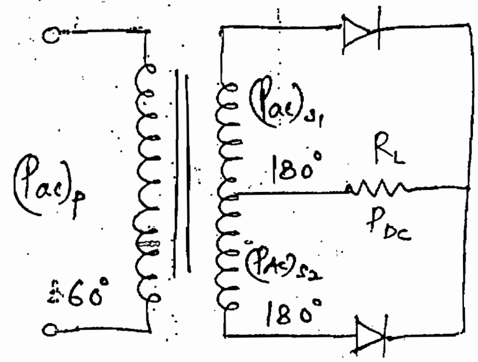
$= \frac{I_{dc}^2}{I_{rms}^2}$

$= \frac{(2 I_m/\pi)^2}{(I_m/\sqrt{2})^2} = \frac{8}{\pi^2}$

$= 81.2\%$

AU

(7). TUF :



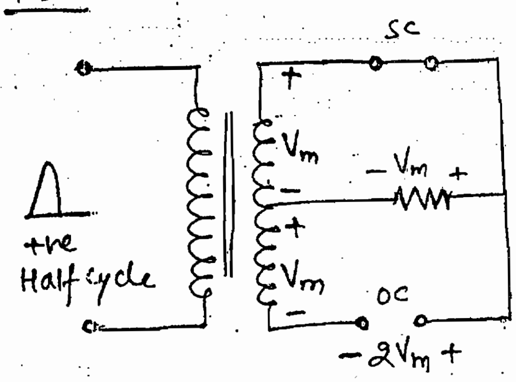
$$(TUF)_p = \frac{P_{DC}}{(P_{AC})_p} = \eta = 81.2\%$$

$$(TUF)_s = \frac{P_{DC}}{(P_{ac})_{sec_1} + (P_{AC})_{sec_2}} = 57.3\%$$

$$TUF = \frac{81.2 + 57.3}{2} = 69.3\%$$ (disadvantage)

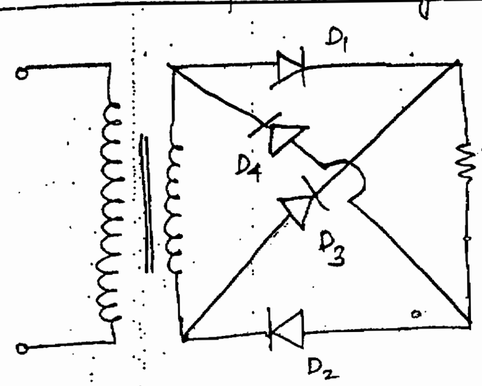
* Avg can be taken here becoz 180° each get added and mean can be taken.

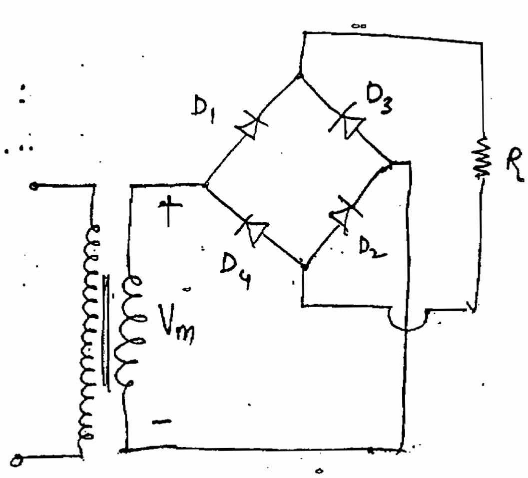
(8) PIV :

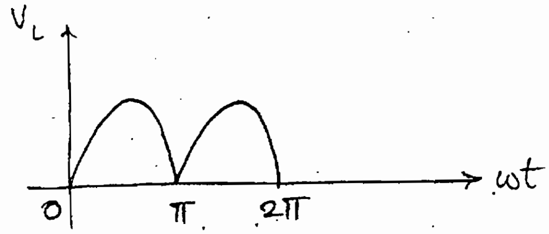


$PIV = 2V_m$ (disadvantage)

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







Common parameters as full wave Rectifier (centre tapped)

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

(2). $(V_{dc})_{NL} = \dfrac{2V_m}{\pi}$

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

(4). $rf = 48\%$

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

Other parameters :

(6). $I_{dc}\,(\underbrace{R_f}_{2R_f} + \underbrace{\tfrac{R_s}{2}}_{R_s}) + I_{dc} R_L = \dfrac{2V_m}{\pi}$

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

$VR^n =$

(7). TUF :

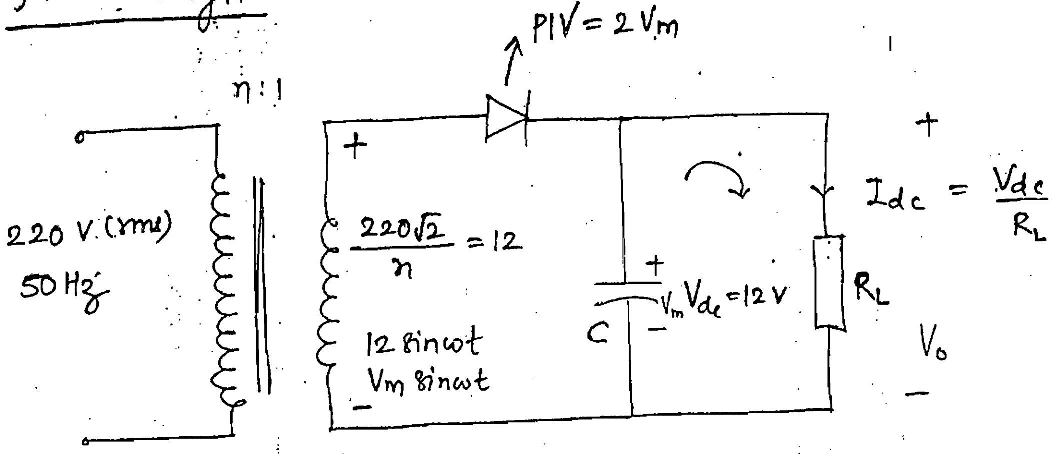
$$(TUF)_p = \frac{P_{DC}}{(P_{AC})_p} = \eta = 81.2\% \qquad (360^\circ - 360^\circ)$$

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

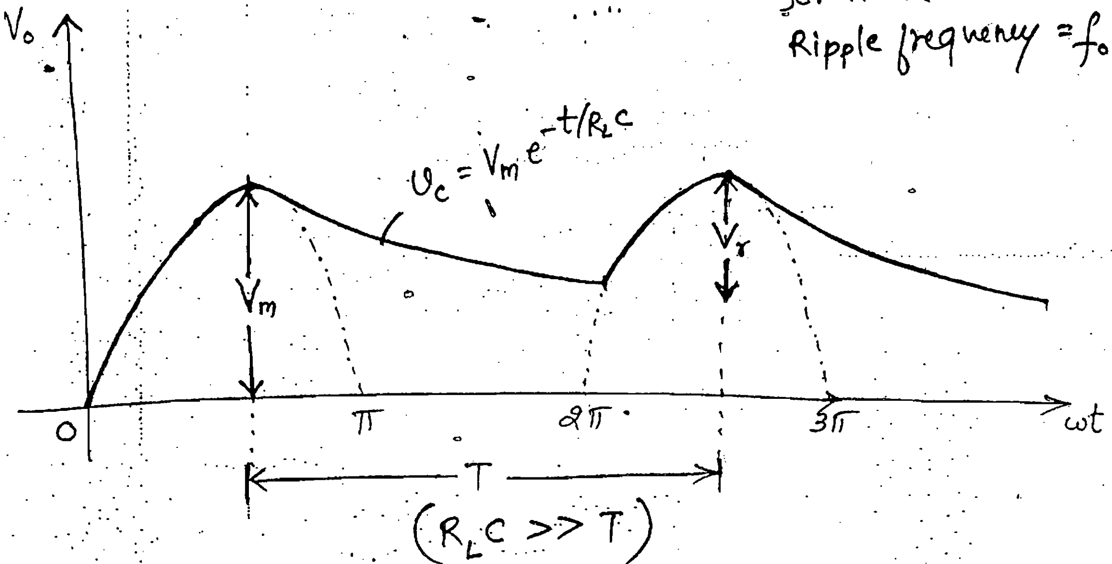
$\therefore$ TUF $= 81.2\%$

(8). $PIV = V_m$ (advantage)

Filter design:



For HWR
Ripple frequency $= f_0$



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

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

$$V_c = V_m \cdot e^{-t/R_LC}$$

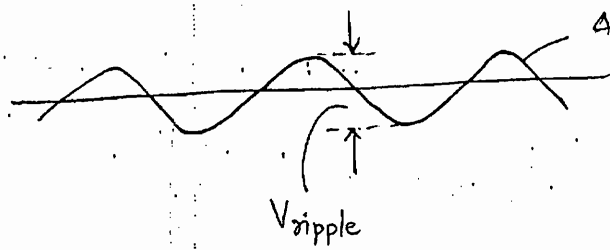
if $R_LC >> T$

$$V_c = V_m\left(1 - \frac{T}{R_LC}\right)$$

$$V_c = \underbrace{V_m}_{V_{dc}} - \underbrace{\frac{V_m T}{R_LC}}_{V_{ripple}}$$

$$\downarrow V_{ripple} = \frac{V_m T}{(R_LC)\uparrow}$$

triangular wave $\left(V_{rms} = \frac{V_m}{\sqrt{3}}\right)$



$$rf = \frac{V_{ripple\,(rms)}}{V_{dc}}$$

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

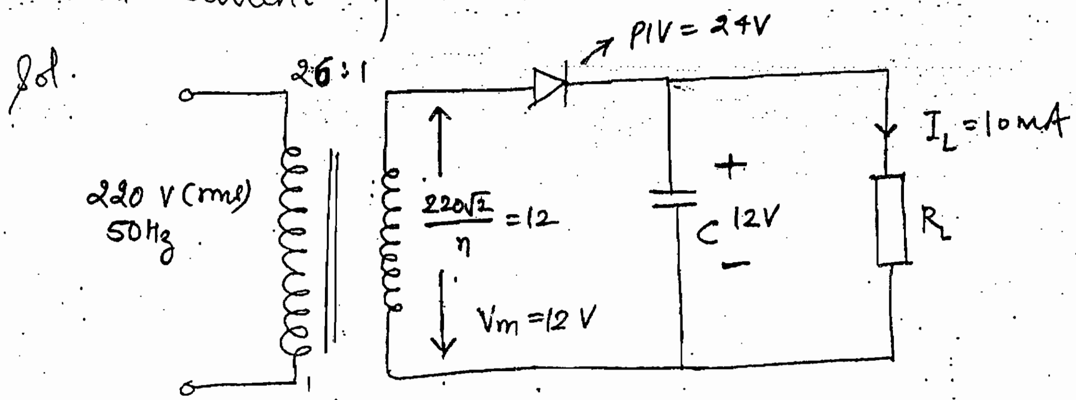
$$rf = \frac{T}{\sqrt{3} R_L C} = \frac{T}{\sqrt{3}\,\tau}$$

If $R_L = \infty$, $rf = 0$ & $v_c = V_m$.

Problem:

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

Sol.



(1). Turns ratio :

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

(2). R_L :

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

(3). **rf:**

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

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

$$V_{ripple} = \frac{60\sqrt{3}}{100} \cong 1\ \text{Volt}$$

(4). **Vripple:**

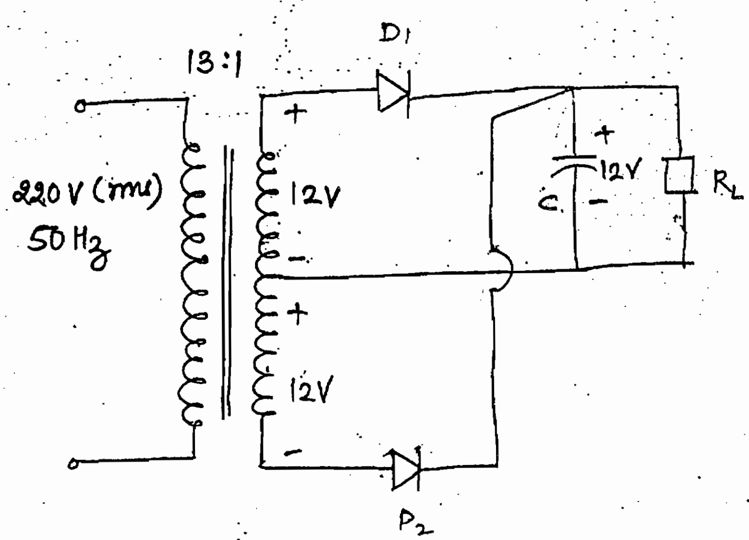
$$V_{ripple} = \frac{V_m T}{R_L C}$$

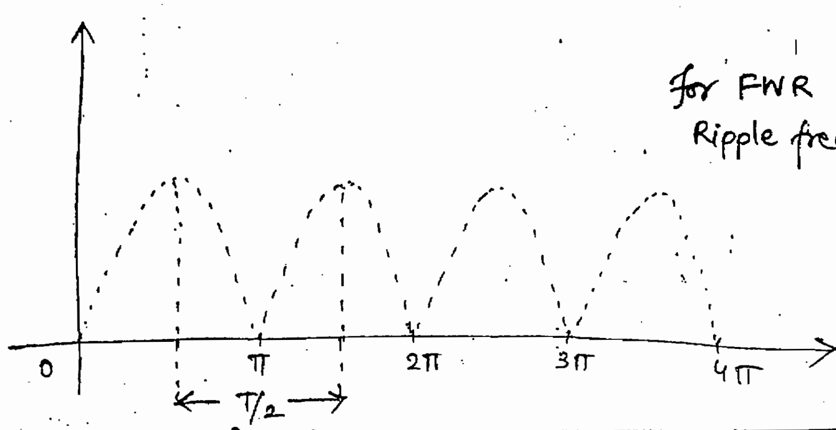
$$1 = \frac{I_L T}{C}$$

$$C = 10\,mA \times 20\,ms$$

$$\therefore C = 200\,\mu F$$

FWR with filter (capacitor):





for FWR
Ripple frequency = $2f_0$

(1). $\dfrac{220\sqrt{2}}{n} = 24 \Rightarrow n = 13$

(2). $R_L = \dfrac{V_{dc}}{I_L} = \dfrac{12\,V}{10\,mA} = 1.2\,K\Omega$

(3). $rf = \dfrac{V_{ripple}\ (rms)}{V_{dc}}$

$\dfrac{5}{100} = \dfrac{V_{ripple}}{12\sqrt{3}} \Rightarrow V_{ripple} \cong 1\ volt$

(4). $V_{ripple} = \dfrac{V_m}{R_L C} \cdot \left(\dfrac{T}{2}\right)$

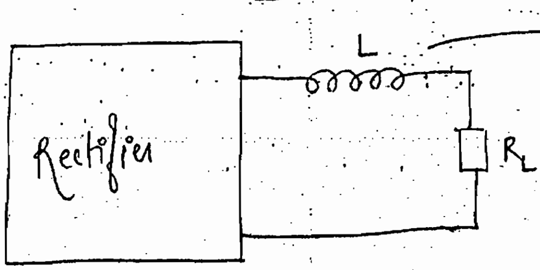
$1 = \dfrac{I_L}{C} \cdot \left(\dfrac{T}{2}\right)$

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

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

Conclusion:

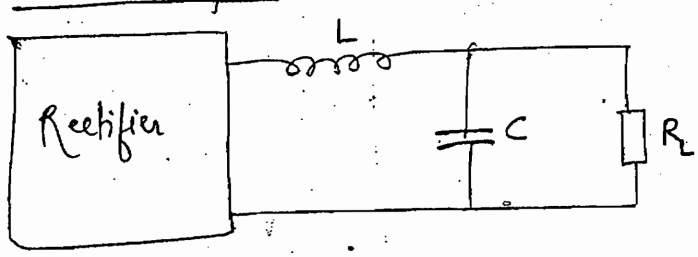
(Inductor filter):



$\rightarrow$ size↑, $rf \propto R_L$

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

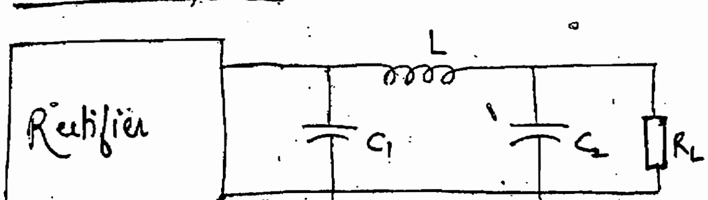
L section filter:



$rf = \dfrac{\sqrt{2}}{3} \cdot \dfrac{X_C}{X_L}$

* for variable load

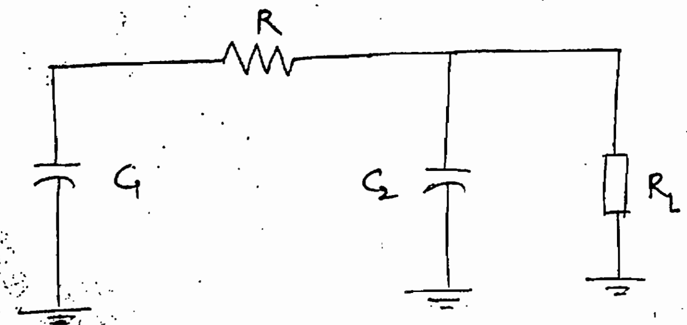
π Section filter:



$rf = \dfrac{\sqrt{2} \cdot X_{C_1} X_{C_2}}{R_L \cdot X_L}$

* for low loads.

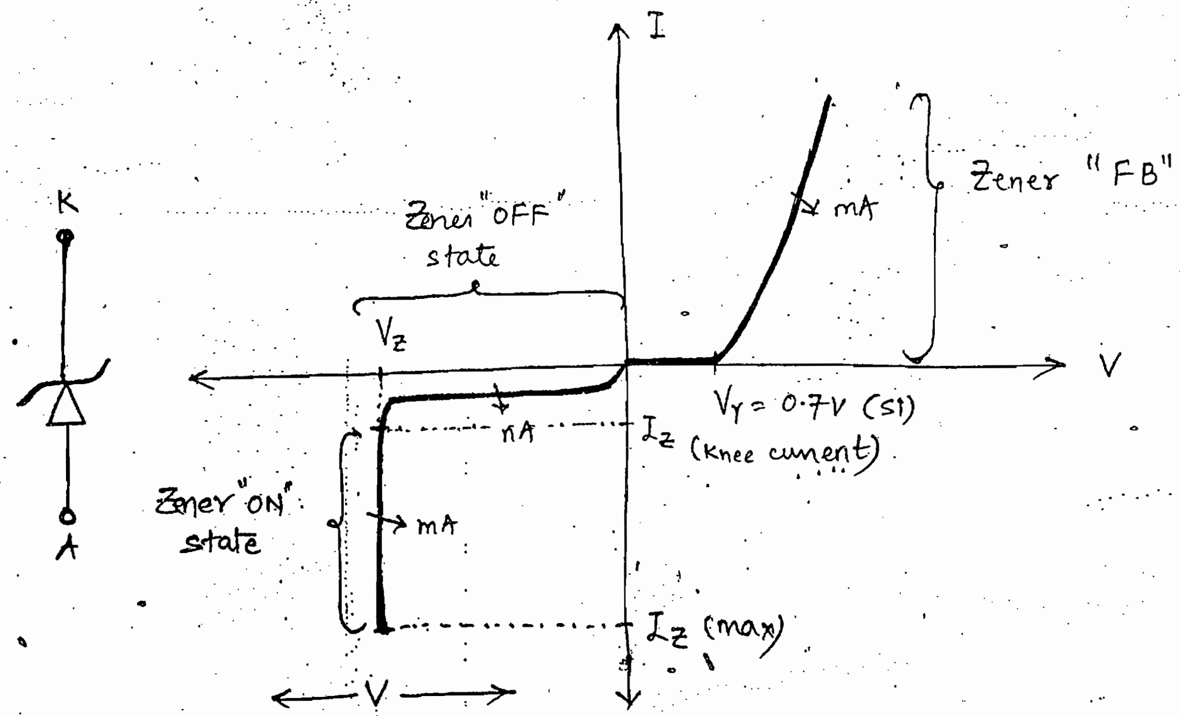
RC filter:



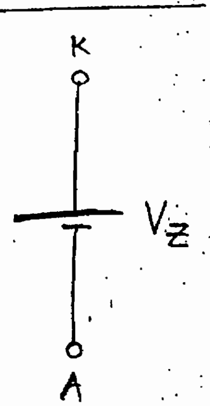
$$r_f = \frac{\sqrt{2}\, X_{C_1} \cdot X_{C_2}}{R_L \cdot R}$$

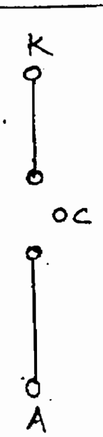
* for low loads

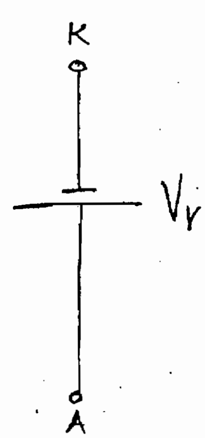
Zener Voltage Regulator –



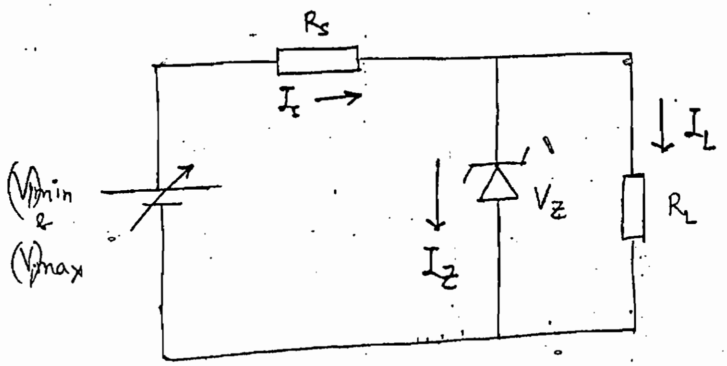
$V > V_Z$	$0 < V < V_Z$	$V \to FB$
"ON" state	"OFF" state	







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



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

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

$$\left.\begin{aligned}(I_Z)_{min} &= (I_s)_{min} - I_L \\ (I_Z)_{max} &= (I_s)_{max} - I_L\end{aligned}\right\} \quad V_i \rightarrow \text{variable} \ \& \ R_L \rightarrow \text{fixed}$$

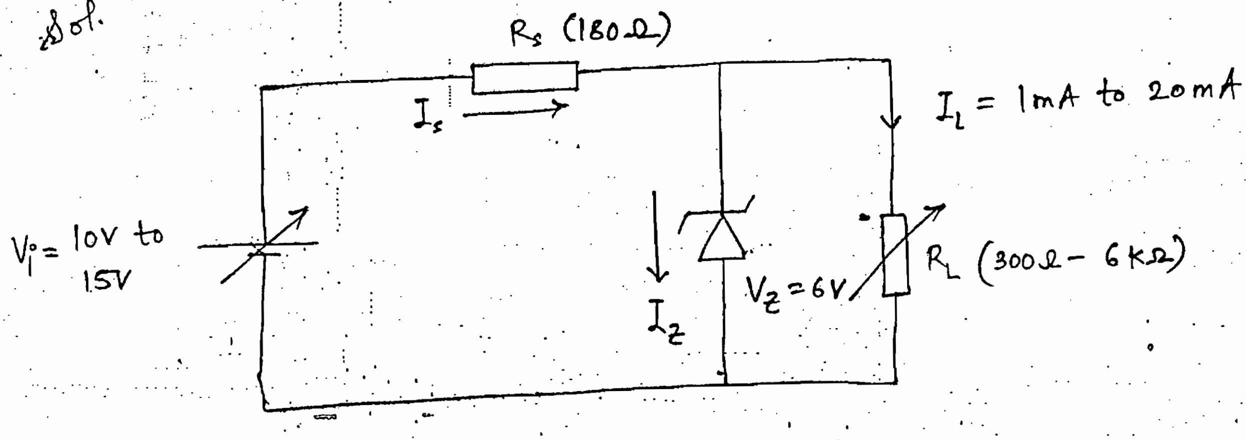
$$\left.\begin{aligned}(I_Z)_{min} &= (I_s)_{min} - (I_L)_{max} \\ (I_Z)_{max} &= (I_s)_{max} - (I_L)_{min}\end{aligned}\right\} \quad V_i \rightarrow \text{variable} \ \& \ R_L \rightarrow \text{variable}$$

$$\left.\begin{aligned}(I_Z)_{min} &= I_s - (I_L)_{max} \\ (I_Z)_{max} &= I_s - (I_L)_{min}\end{aligned}\right\} \quad V_i \rightarrow \text{fixed} \ \& \ R_L \rightarrow \text{variable}$$

Problem :

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

Sol.



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

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

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

$$\frac{(V_i)_{min} - V_Z}{(R_s)_{max}} - (I_L)_{max} > (I_Z)_{min}$$

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

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

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

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

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

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

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

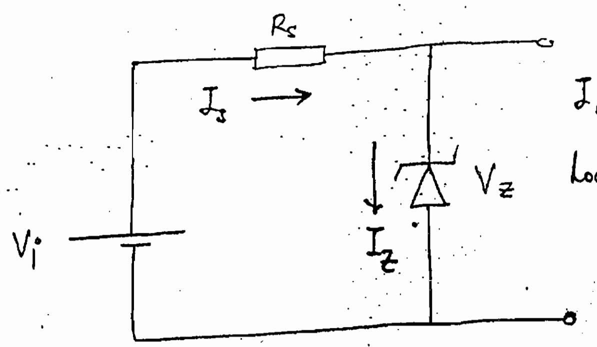
$$\therefore (R_s)_{min} > 176\ \Omega$$

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

$$176 < R_s < 199$$

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

Small s/g Analysis of zener diode :



$I_L = 0$

Load is open

$$V_i = I_s R_s + V_z$$

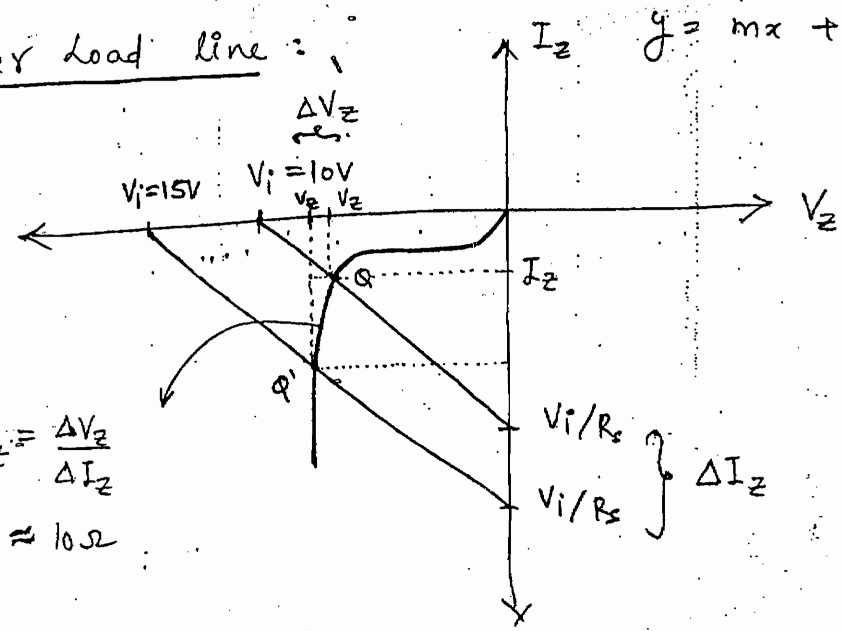
$$I_s = I_z$$

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

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

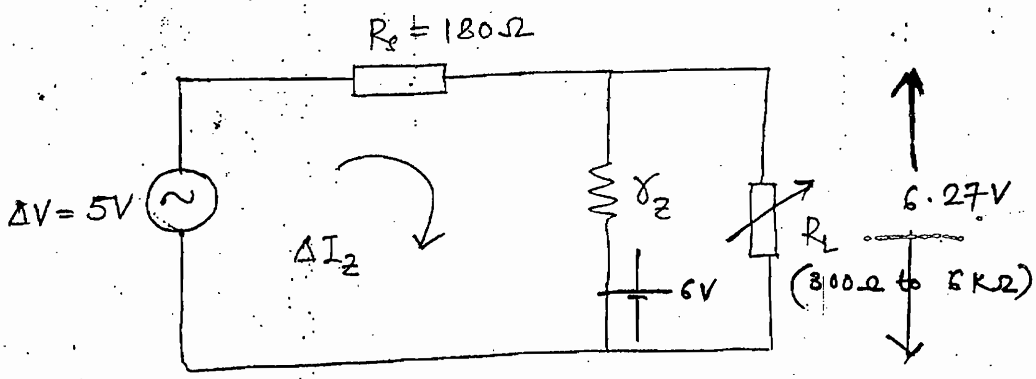
$$y = mx + c$$

Zener Load line :



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

$$\approx 10\,\Omega$$



$$\Delta I_z \simeq \frac{\Delta V}{R_s + r_z} \qquad (r_z << R_L)$$

$$\Delta I_z \simeq \frac{\Delta V}{R_s}.$$

And, $\Delta V_z = \Delta I_z \times r_z$

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

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

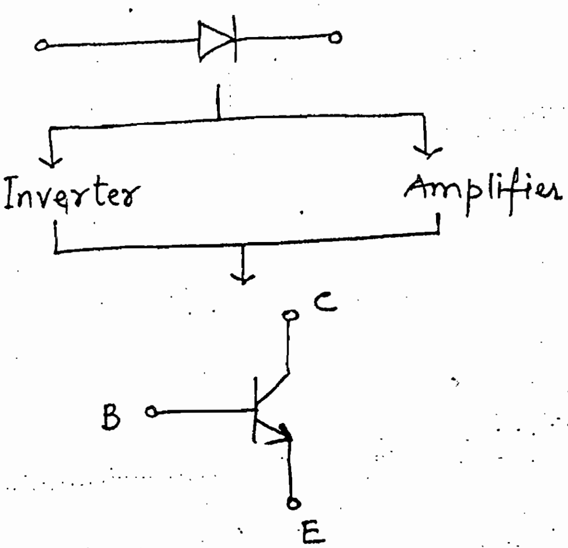
$$= \frac{5}{18} = 0.27\ V$$

BJT

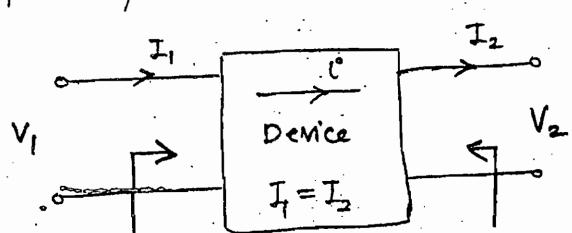
(1). Problems with diode:

Diode cannot be used as an inverter

Diode cannot be used as an Amplifier.



Two port N/w:



$Z_i = \frac{V_1}{I_1}$ $(1 K\Omega)_{FB}$ $\qquad$ $Z_0 = \frac{V_2}{I_2}$ $(1 M\Omega)_{RB}$

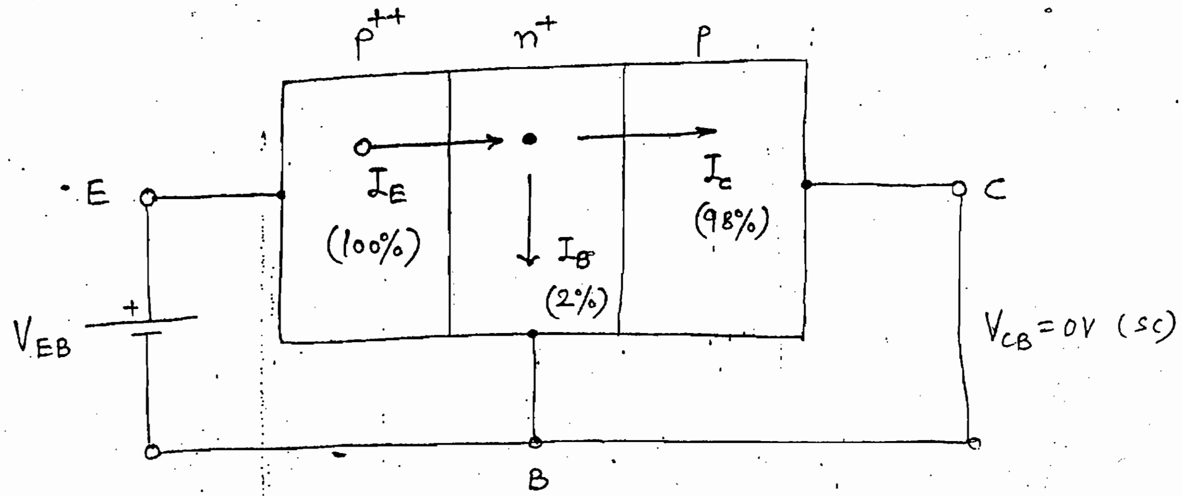
$$P_V = \frac{P_0}{P_i} = \frac{I_2^2 Z_0}{I_1^2 Z_i} = \frac{Z_0}{Z_i}$$

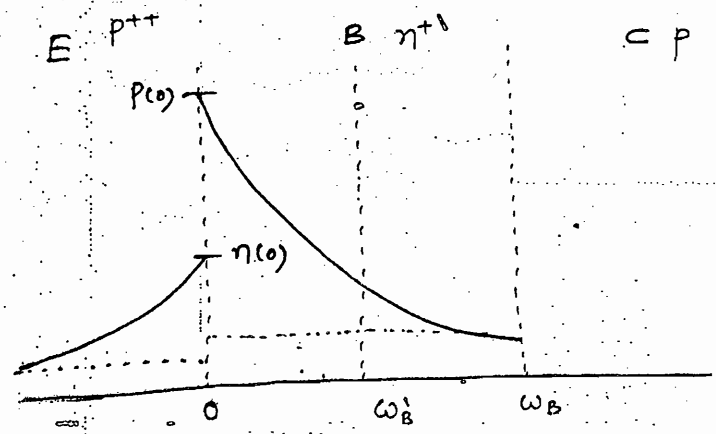
$$= \frac{1 M\Omega}{1 K\Omega} = 10^3$$

$\therefore \boxed{P_0 = 1000 P_i} \rightarrow$ Amplifier

$\ast$ Transistor is a transfer of current.

Basic principle of BJT :





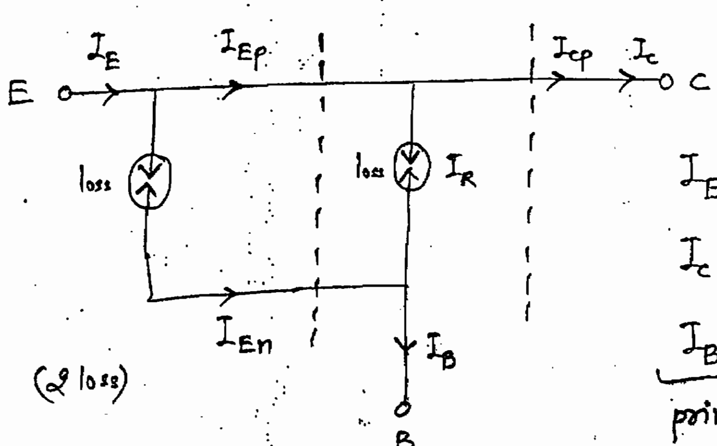
$V_{CB} = 0V$ (1). $\frac{\omega_B}{L_B} >> 1$ (2). $\frac{\omega_B'}{L_B} << 1$ (width of Base is narrow)

Diffusion :

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

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

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



V_{EB} = const. , $V_{CB} = 0$, $\frac{\omega_B'}{L_B} << 1$

$$I_E = I_{Ep}\uparrow + I_{En} \text{ (const.)}$$

$$I_c = I_{ep}\uparrow$$

$$\underbrace{I_B = I_{En} + I_R\downarrow}_{\text{principle of BJT is valid}}$$

(1). Emmitter injection efficiency-

$$\gamma = \frac{I_{Ep}}{I_E}$$

$$= \frac{I_{Ep}}{I_{Ep} + I_{En}}$$

$$= \frac{1}{1 + \frac{I_{En}}{I_{Ep}\uparrow}} \approx 1$$

(2). Base transport factor (β^*):

$$\beta^* = \frac{I_{cp}}{I_{Ep}} = \frac{I_{Ep} - I_R}{I_{Ep}} = 1 - \frac{I_R\downarrow}{I_{Ep}\uparrow} \approx 1$$

(3). Large s/g current gain (α):

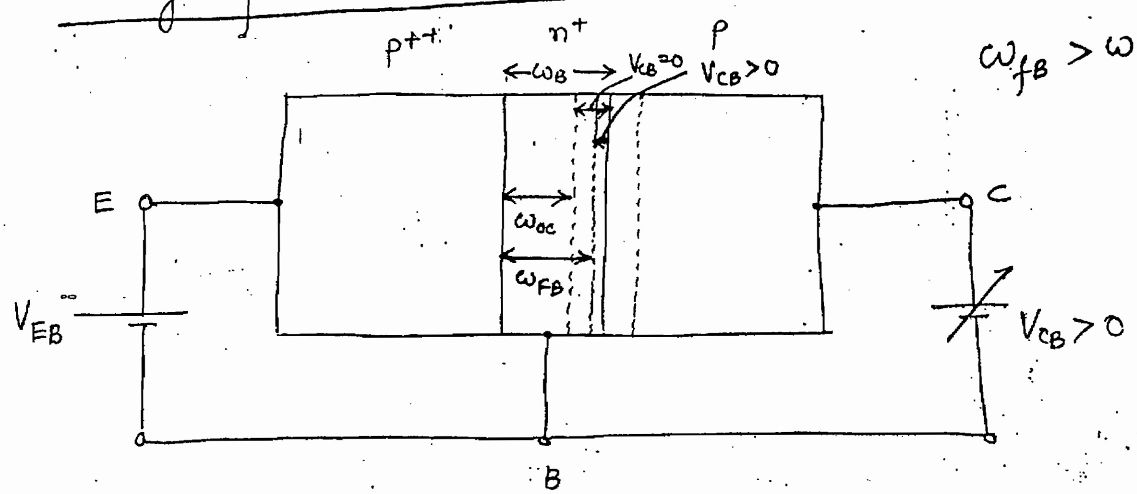
$$\alpha = \frac{I_c}{I_E} = \frac{I_{cp}}{I_{Ep}} \cdot \frac{I_{Ep}}{I_E}$$

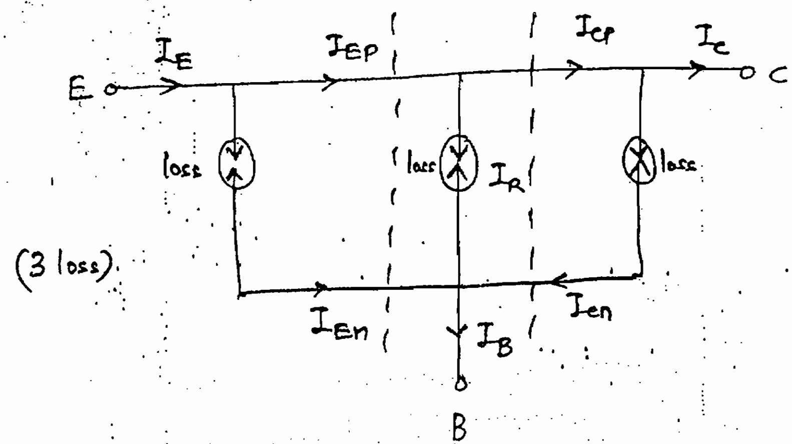
$$\therefore \alpha = \beta^* \gamma$$

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

Testing of BJT in Saturation:

$\omega_{fB} > \omega_{oc}$





$V_{EB} = \text{const.}$; $V_{CB} > 0$; $\frac{\omega_B}{L_B} << 1$

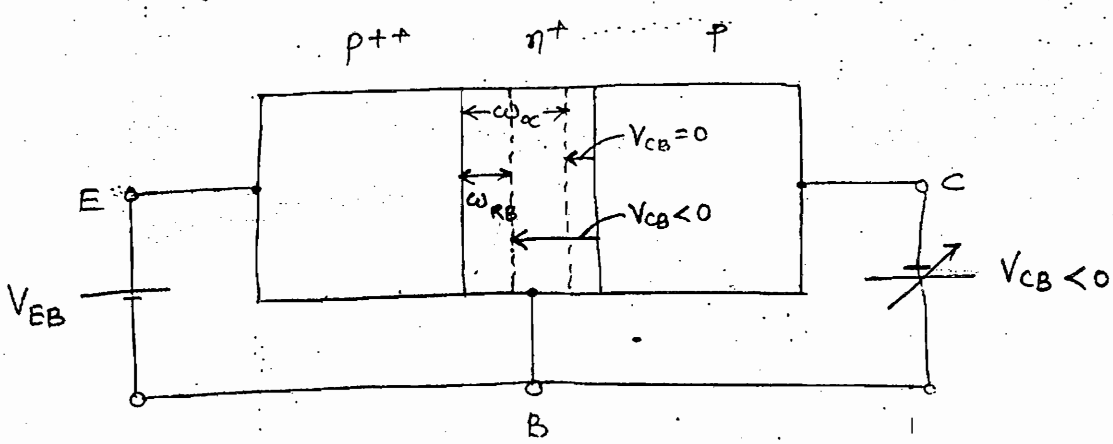
$\downarrow I_E = I_{EP}\downarrow + I_{En}$ (const.)

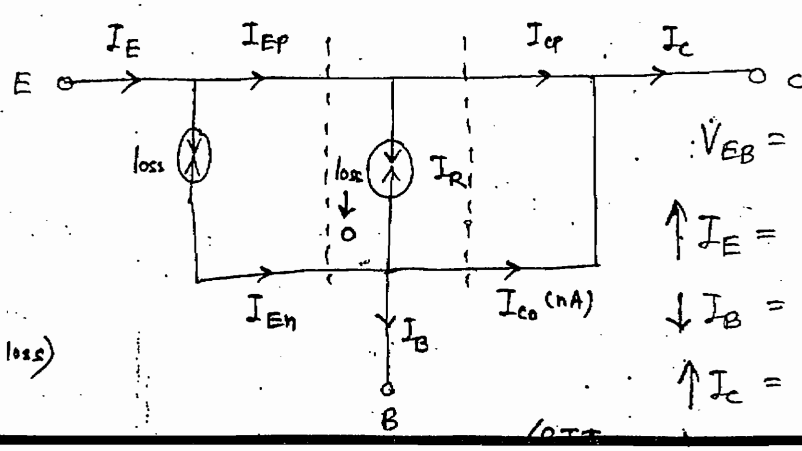
$\downarrow I_c = I_{cP}\downarrow - I_{cn}\uparrow$

$\uparrow I_B = \underset{\text{(const.)}}{I_{En}} + I_R\uparrow + I_{cn}\uparrow$

BJT is a failure in saturation

Testing of BJT in Active Region:





(only 1 loss)

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

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

$\downarrow I_B = \underset{\text{(const.)}}{I_{En}} + I_R\downarrow - \underset{\text{(const.)}}{I_{co}}$

$\uparrow I_c = I_{EP}\uparrow + \underset{\text{(const.)}}{I_{co}}$

BJT conditions :

(1). Doping levels

$$N_E >> N_B$$
$$N_B > N_C$$

(2) :- W_B should be narrower (Recombination losses reduces)

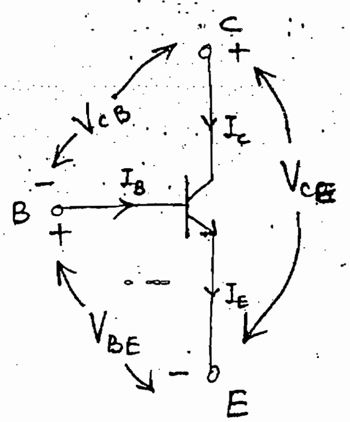
$$I_E = I_B + I_C$$
$$I_E \approx I_C$$

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

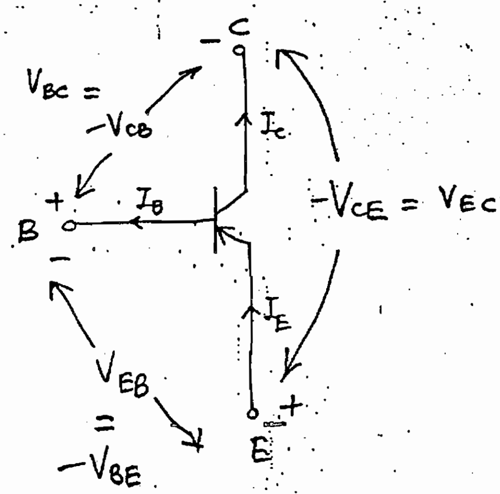
Becoz heat generation is more in collector region due to RB, Hence, area has to be more for collector.

Symbols & Notations of BJT :

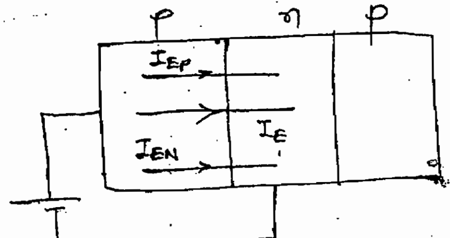
"npn"



"pnp"



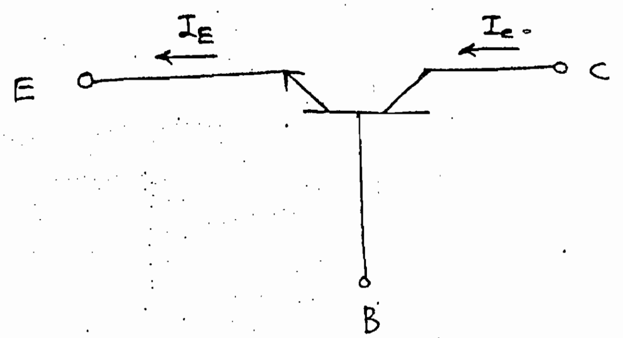
$$\boxed{V_{CB} + V_{BE} = V_{CE}}$$ always



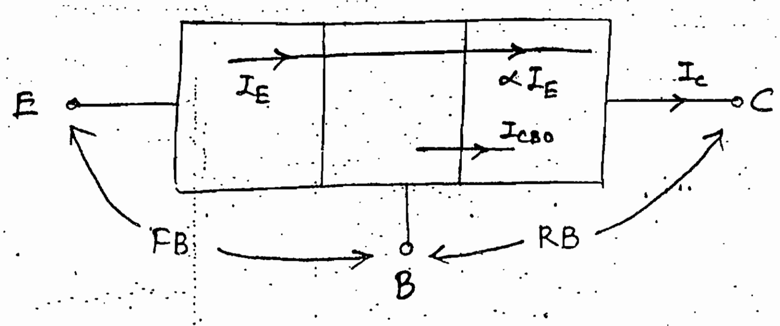
(Testing of symbol pnp)

BJT configuration :

Common Base (CB) :



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



$I_{CBO} \rightarrow$ Collector to Base leakage current when emitter is open.

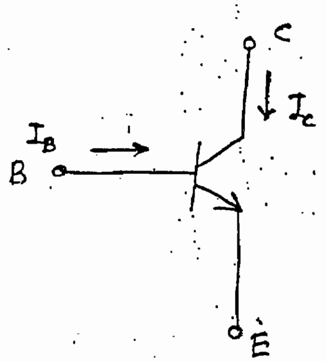
$$I_c = \alpha I_E + I_{CBO}$$

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

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

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

Common Emitter (CE) :



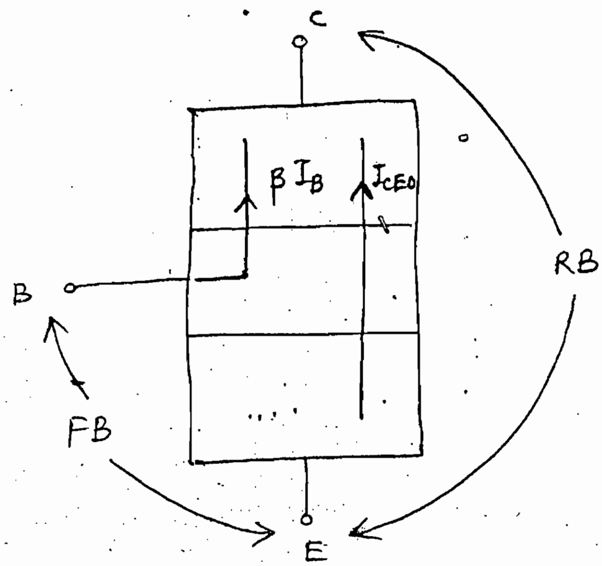
$$\beta = \frac{I_c}{I_B} \quad ; \quad \beta \rightarrow (20 \text{ to } 500)$$

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

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

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

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_{CEO} :

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

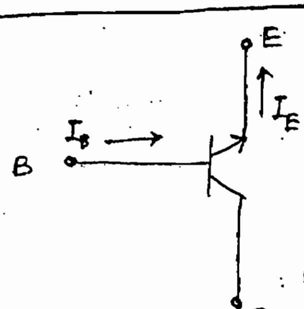
$$I_C = \beta I_B + (1+\beta) I_{CBO} \quad \text{——(I)}$$

CE : $$I_C = \beta I_B + I_{CEO} \quad \text{——(II)}$$

On comparing, $$I_{CEO} = (1+\beta) I_{CBO}$$

$$I_{CEO} = \frac{1}{(1-\alpha)} I_{CBO}$$

Common Collector (CC) :



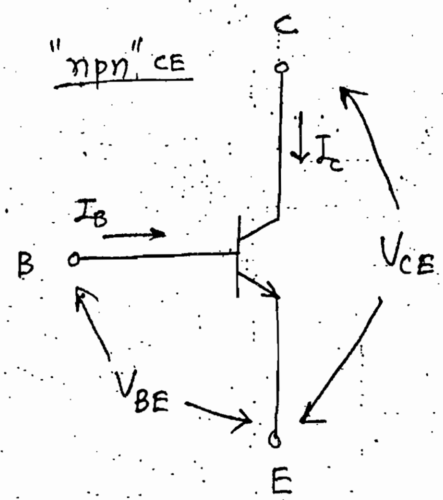
$$\gamma = \frac{I_E}{I_B} \quad ; \quad \gamma \rightarrow (20 \text{ to } 500)$$

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

$$\boxed{\gamma = \frac{1}{1-\alpha} = 1+\beta} \qquad \boxed{\gamma \approx \beta}$$

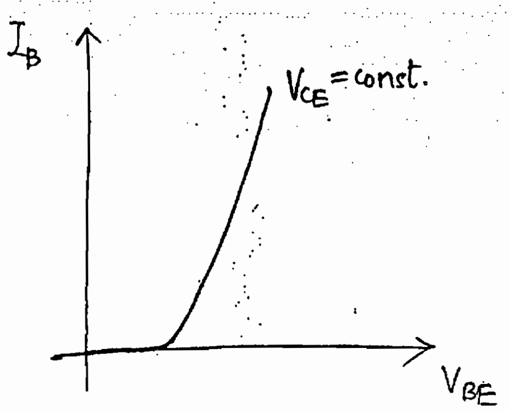
*BJT operating Regions:

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

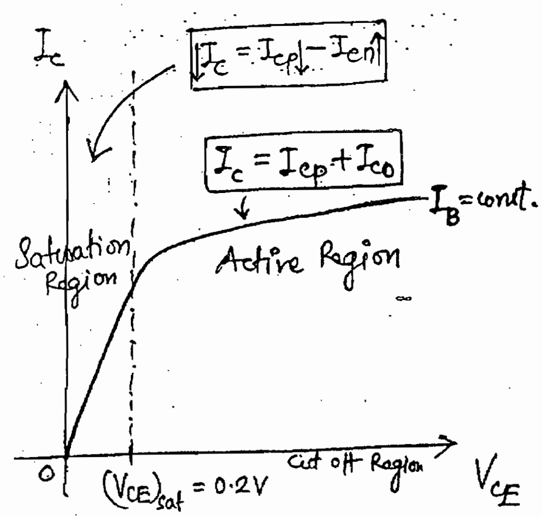


I/p parameters → V_{BE}, I_B

O/p parameters → V_{CE}, I_C

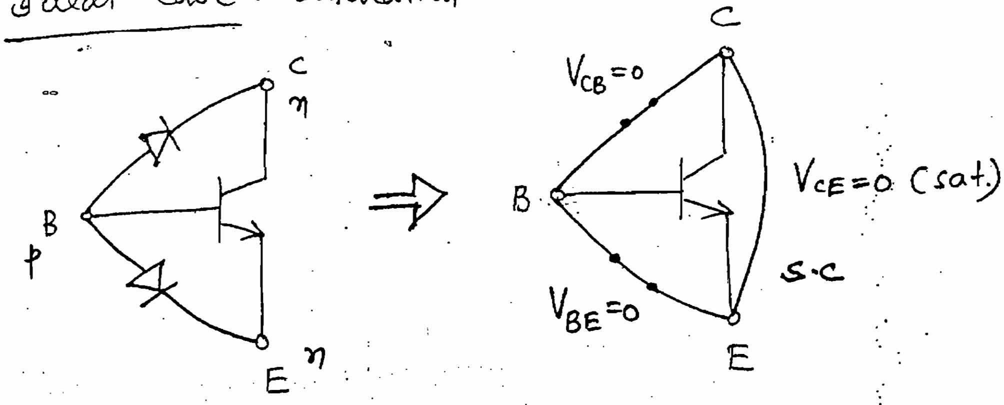


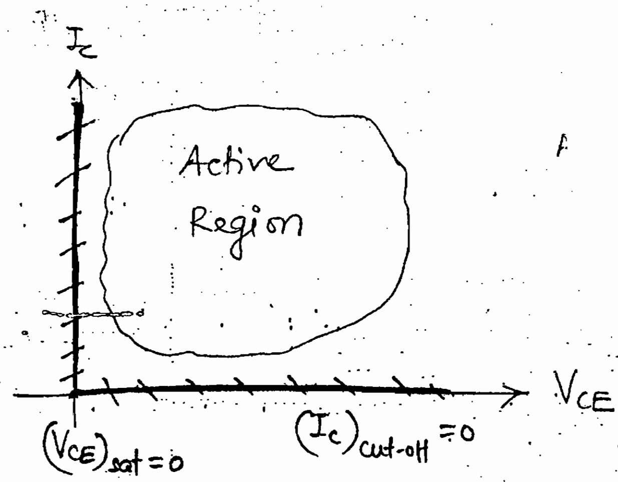
(I/p characteristics shows the diode characteristics and it is of no use, it does not represent BJT)



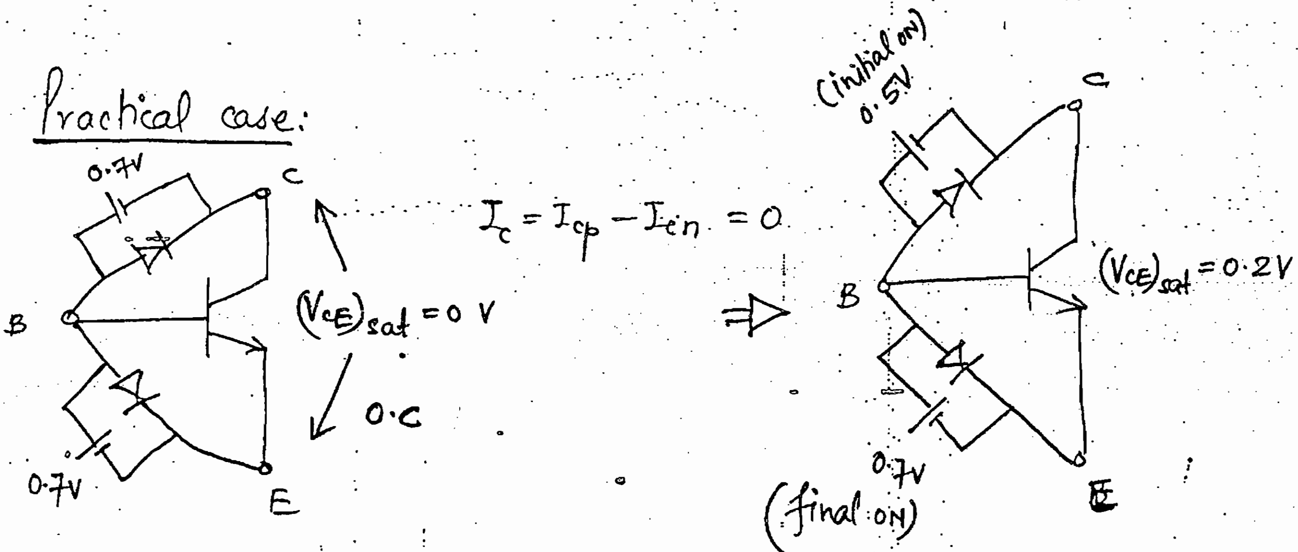
(O/p characteristics are used to study BJT and BJT is represented by these characteristics).

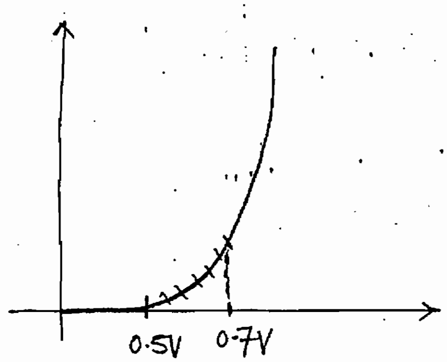
Ideal case : saturation





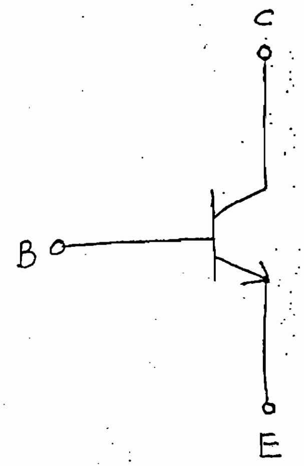
Practical case:

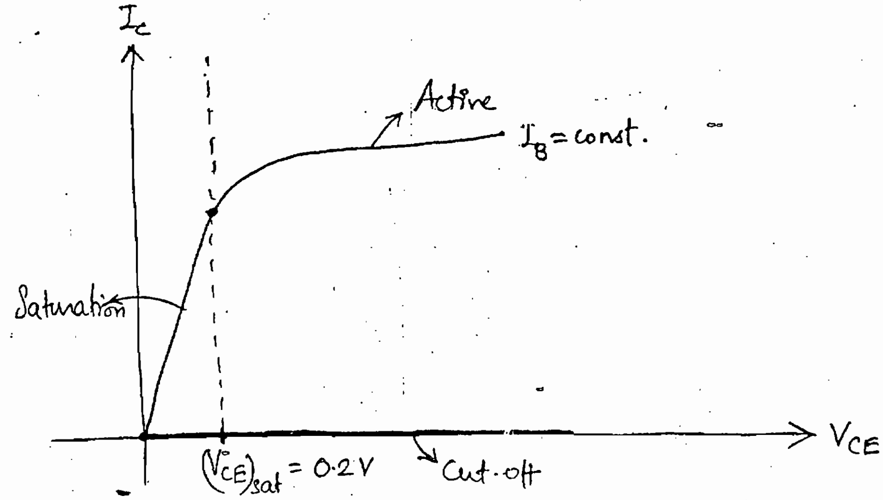




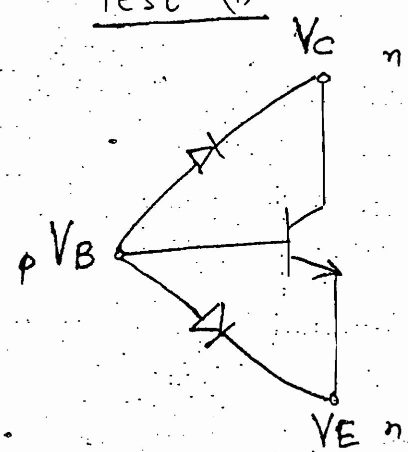
$(V_D)_{cut\,in}$ $(V_D)_{active}$ — Diode → "ON" < 0.5V (Initial ON) / 0.7V (Final ON)

Testings of BJT in different operating Regions:





Test-(1):



Saturation →

$V_B > V_E$

$V_B > V_C$

Cut-off →

$V_B < V_E$

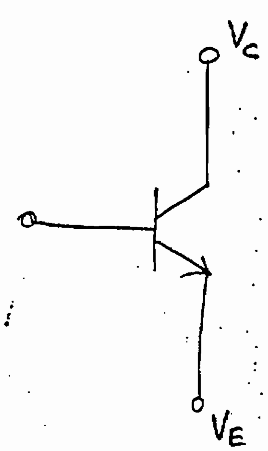
$V_B < V_C$

Active Region →

$V_B > V_E$

$V_B < V_C$

Test-(2):



$V_{CE} = V_C - V_E$

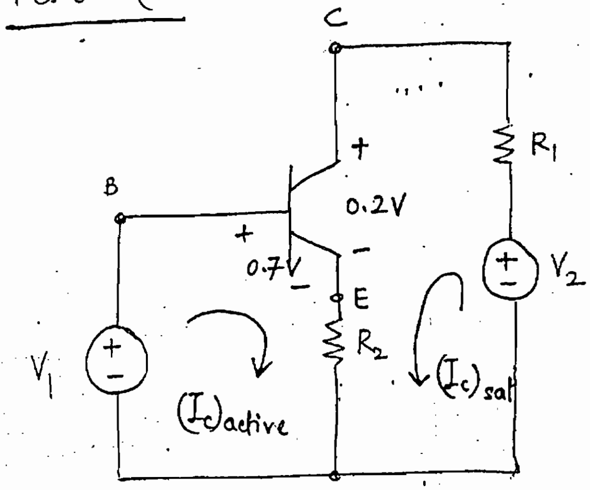
Saturation →

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

Active Region →

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

Test – (3):

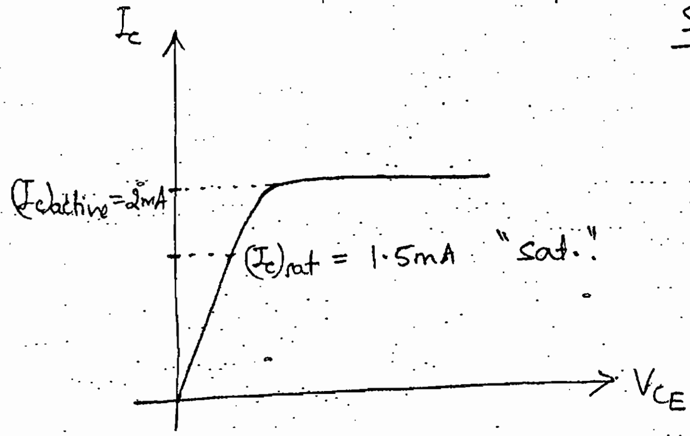


$(V_{BE})_{active} = 0.7V$ (i/p loop)

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

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

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



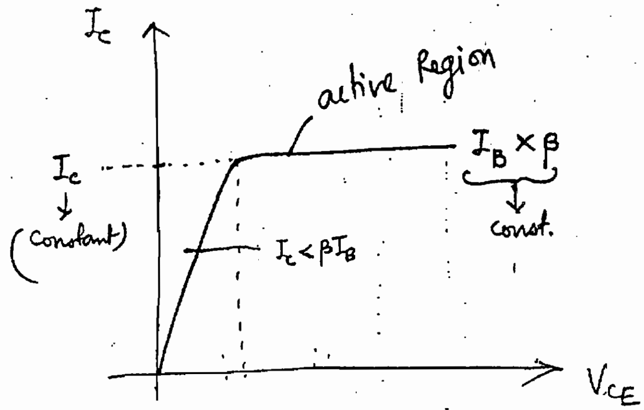
Saturation →

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

Active Region →

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

Test –(4):



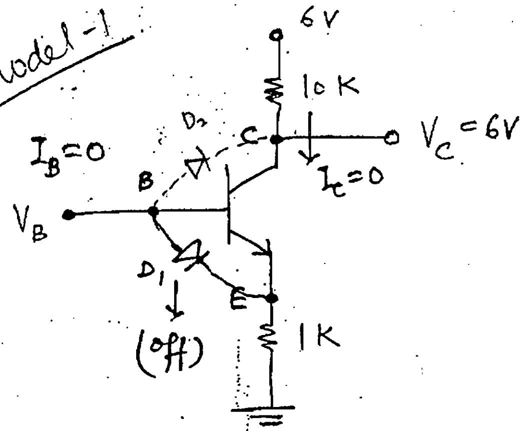
Saturation →

$I_c < \beta I_B$

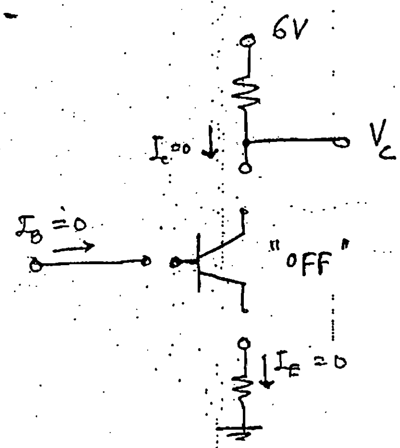
Active Region →

$I_c = \beta I_B$ (const.)

Q2. Model-1



(Cut-off)



For $V_B = 0V$

$D_1 \rightarrow$ (off)

$D_2 \rightarrow$ (off) automatically.

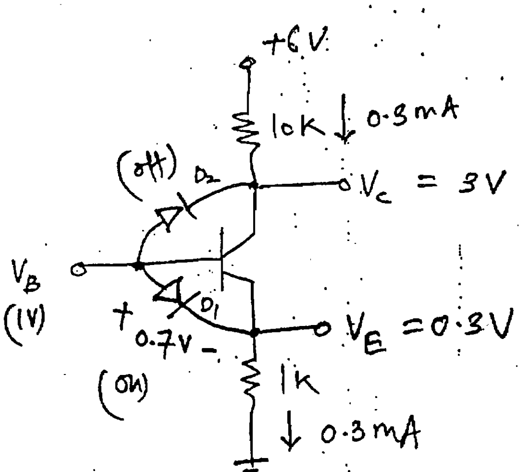
There is no chance of reverse active.

Hence, there is a cut-off

$I_B = I_C = I_E = 0.$

$V_C \doteq 6V$ (directly)

For $V_B = 1V$



Active Region:

D_1: (on)

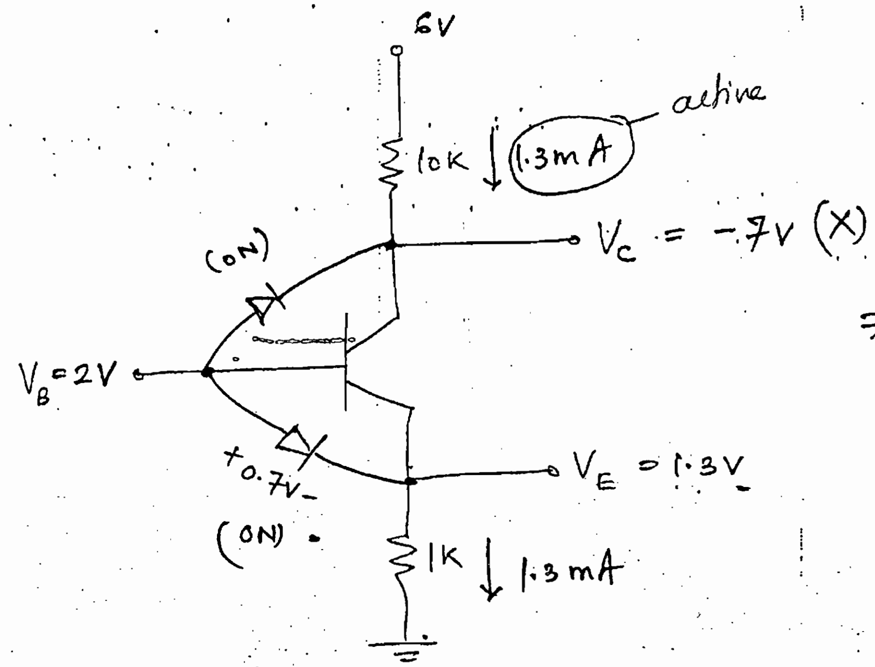
D_2 (off)

$$V_C = 6 - 0.3mA \times 10K$$
$$= 6 - 3$$
$$= 3 \text{ volts}.$$

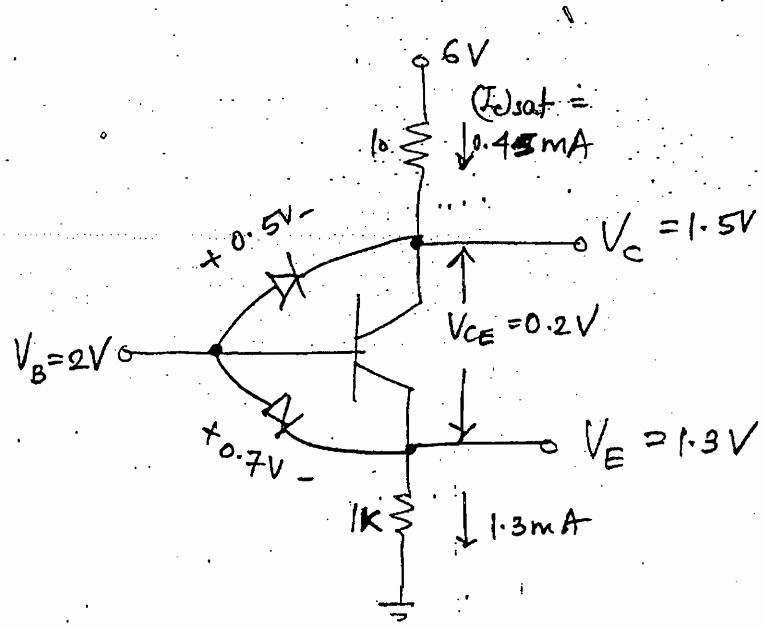
$$I_E = \frac{V_E}{1K} = \frac{0.3}{1K} = 0.3mA$$

$$I_C \approx I_E = 0.3mA$$

For $V_B = 2V$



$\Rightarrow$ Saturation



Saturation

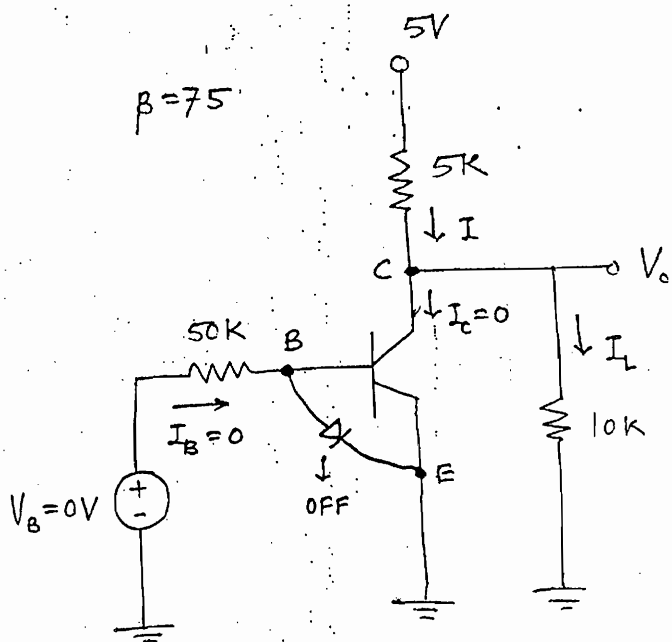
$$V_C = 2 - 0.5$$
$$= 1.5V$$

$$(I_C)_{sat} = \frac{6 - 1.5}{10K}$$
$$= 0.45mA$$

Model-(2):

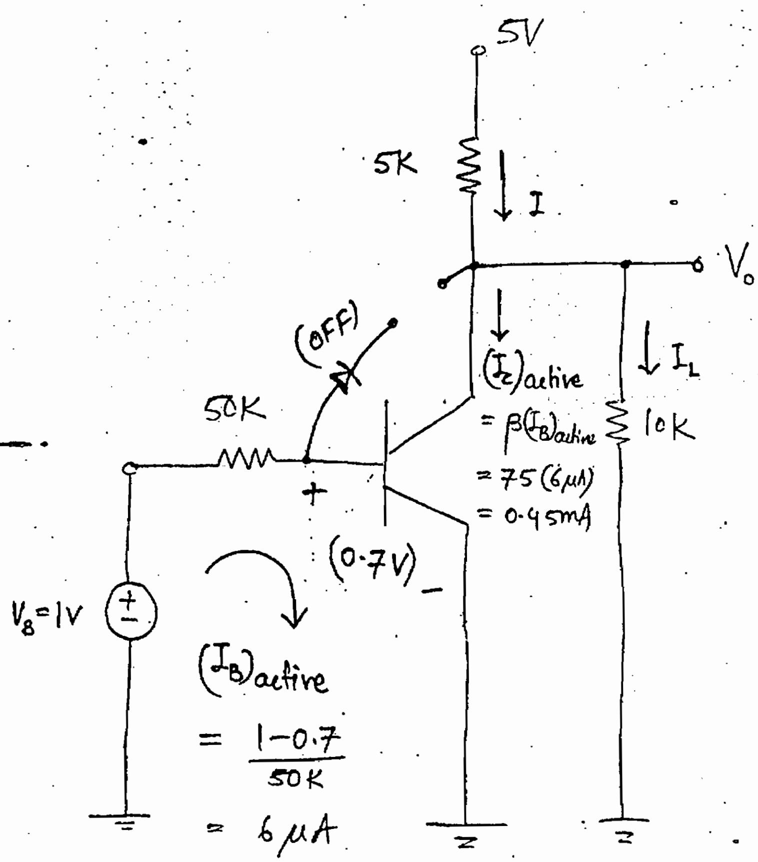
for $V_B = 0V$: (Cut off)

$\beta = 75$



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

for $V_B = 1V$: (Active)



$(I_c)_{active} = \beta (I_B)_{active} = 75(6\mu A) = 0.45mA$

$(I_B)_{active} = \frac{1-0.7}{50K} = 6\mu A$

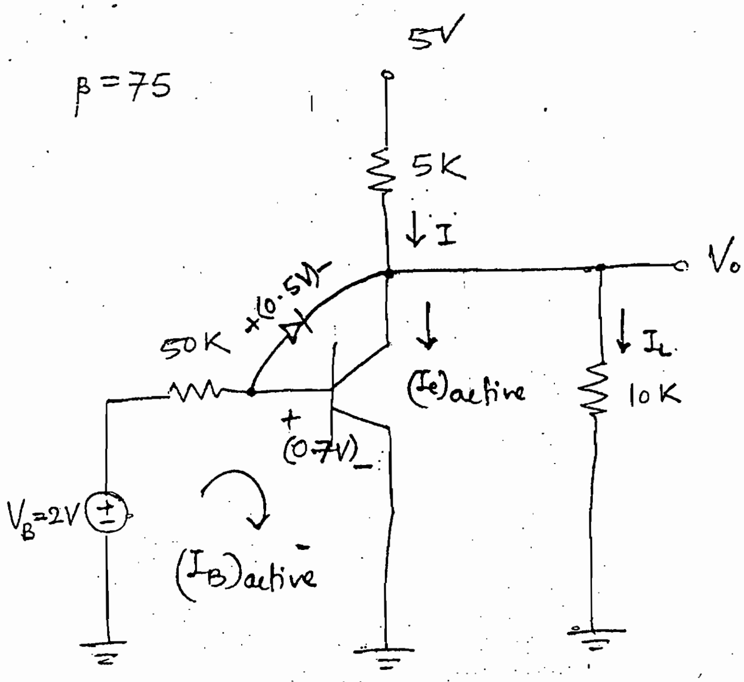
$$\frac{5 - V_o}{5K} = (I_c)_{active} + \frac{V_o}{10K}$$

$$1mA - \frac{V_o}{5K} = 0.45mA + \frac{V_o}{10K}$$

$$V_o = 1.83V$$

For $V_B = 2V$

$\beta = 75$



V_o is ?

(a) 3 V (b) 5V

(c) 0.2 V (d) 1.3V

(practical ckt)

Ideally diode → sc.

then $(V_{CE})_{sat} = 0V$.

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

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

Now,

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

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

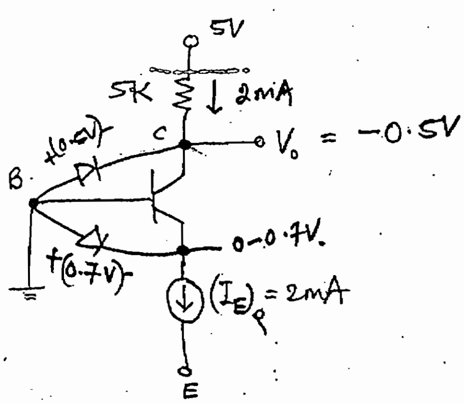
saturation

$$\Rightarrow V_o = -3.16\,V$$

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

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

Q1.



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

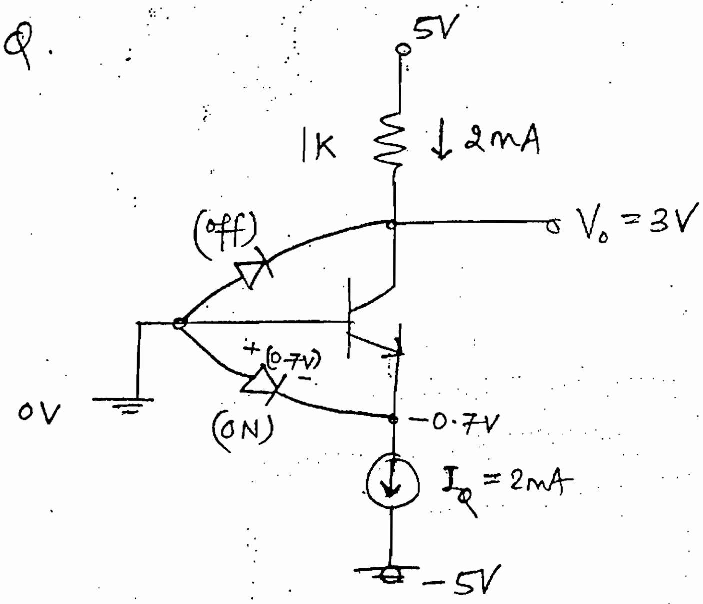
$V_o = 5 - 10$

$V_o = -5V$ (X)

$\boxed{V_o = -ve}$ → saturation

$\therefore V_o = -0.5V$

Q.

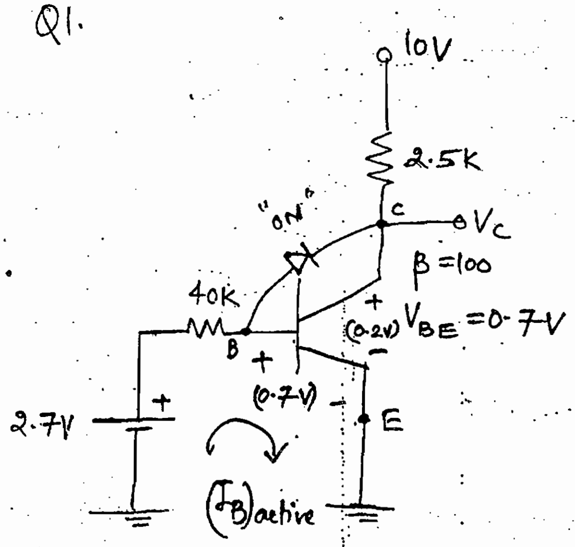


$V_o = 5 - 2mA \times 1K$
$= 5 - 2$
$= 3V$

$\boxed{V_o = 3V}$

active Region.

Q1.



$(I_B)\text{active} = \dfrac{2.7 - 0.7}{40K}$

$= \dfrac{1}{20}\,mA$

$(I_c)\text{active} = \beta\,(I_B)\text{active}$

$= 100 \times \dfrac{1}{20}$

$= 5mA$

$V_c = 10 - 2.5K \times 5mA$

$\boxed{V_c = -2.5\ V}$

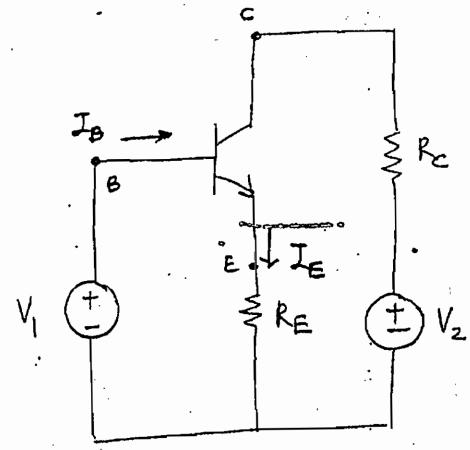
Saturation.

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

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

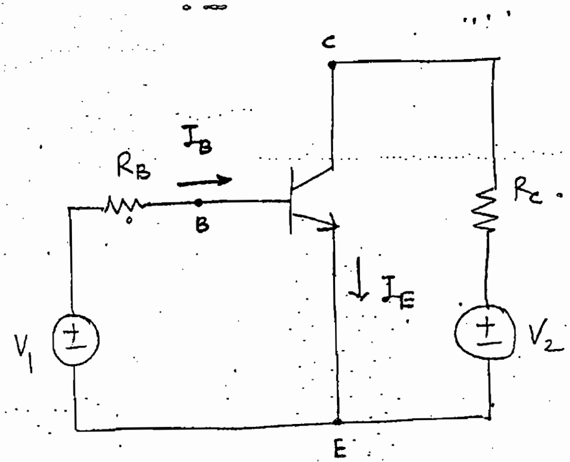
$(I_c)\ \text{sat.} = \dfrac{9.8}{2.5K} = 3.9\ mA$

Model (1) :



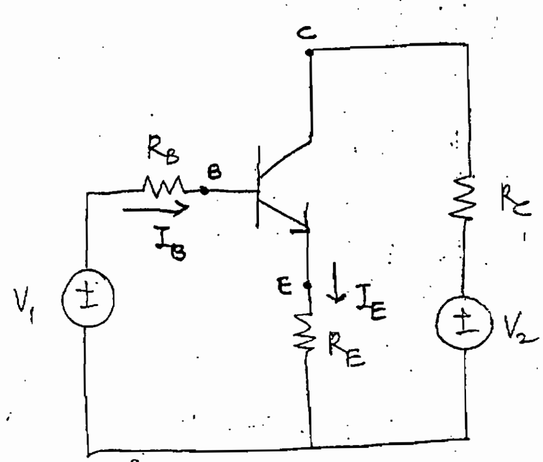
$$I_E = I_B + I_C$$
$$= I_B + \beta I_B$$
$$= (1+\beta) I_B$$
$$V_1 = V_{BE} + I_E R_E$$
(or)
$$V_1 = V_{BE} + I_B R_E (1+\beta)$$

Model (2) :



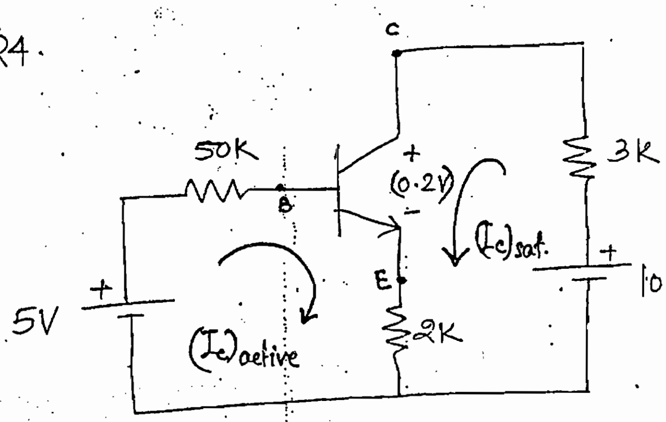
$$V_1 = I_B R_B + V_{BE}$$
(or)
$$V_1 = \frac{I_E R_B}{(1+\beta)} + V_{BE}$$
$$= I_E \left(\frac{R_B}{1+\beta}\right) + V_{BE}$$

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.

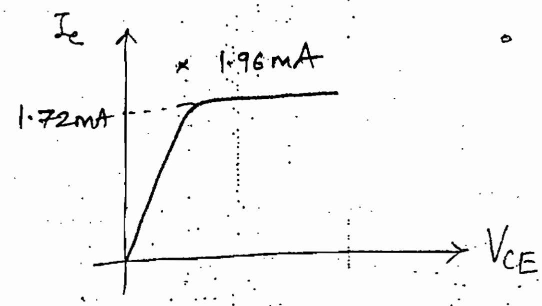


$$(I_c)_{active} = \frac{5-0.7}{\left(\frac{50K}{1+100}\right) + 2k}$$

$$= 1.72mA$$

$$(I_c)_{sat} = \frac{10-0.2}{5K}$$

$$= 1.96mA$$

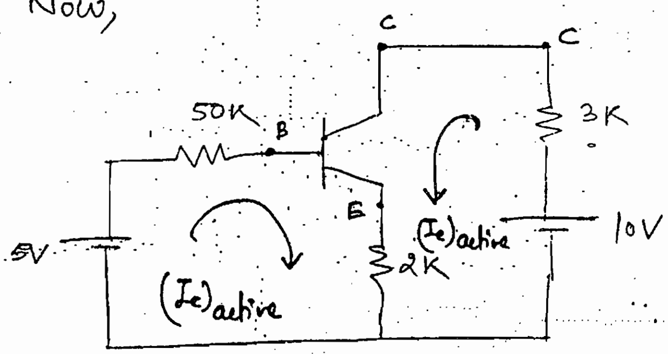


$(I_c)_{sat.} < 1.72\ mA$

But $(I_c)_{sat} > 1.72mA$

$\therefore$ Active Region.

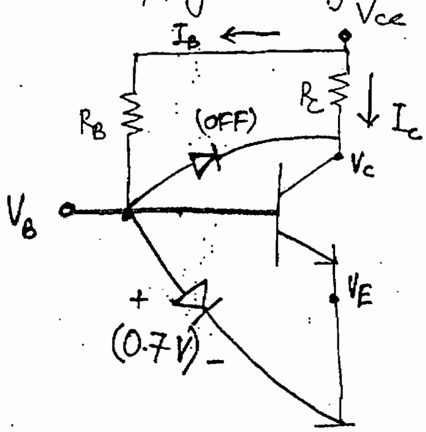
Now,



$$V_E = 1.72mA \times 2K$$
$$= 3.44\ V$$
$$V_C = 10 - 3K \times 1.72mA$$
$$= 4.84\ V$$
$$V_{CE} = V_C - V_E$$
$$= 4.84 - 3.44$$
$$= 1.4\ V$$

Single supply design in BJT to operate in Active Region:

(1).

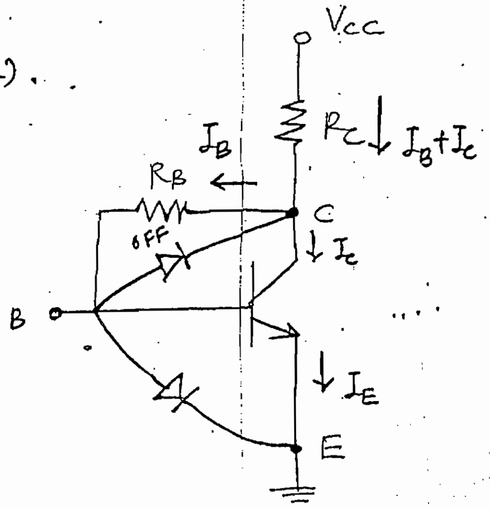


$R_B > R_C$

$V_B < V_C \rightarrow D\ (off)$

Active Region.

(2).

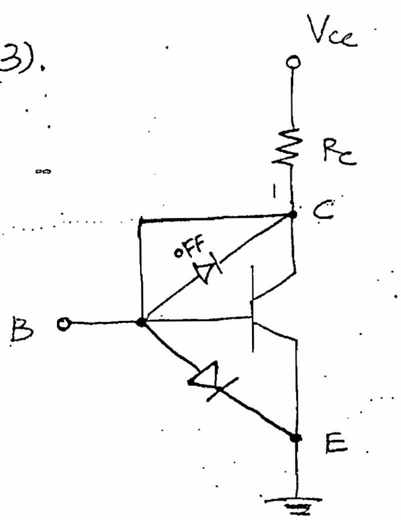


$V_B < V_C$

↓

D → OFF (Active Region)

(3).



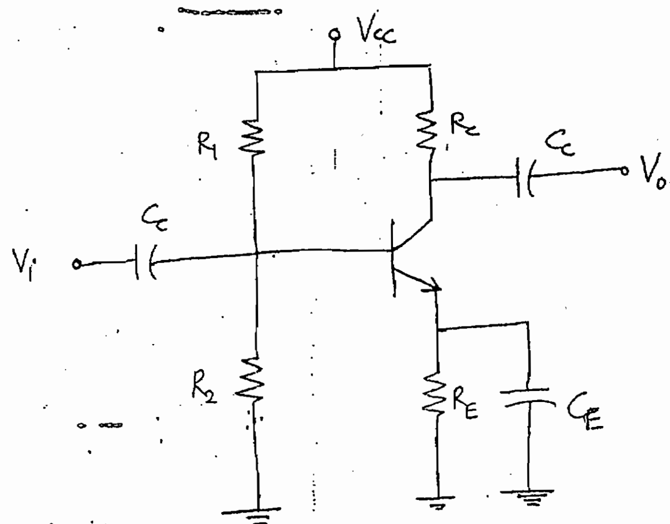
$V_B = V_C$; D → OFF (Active Region)

(Active Region)

BJT as a Diode.

DC load line and operating point (Q):

CE Amplifier:



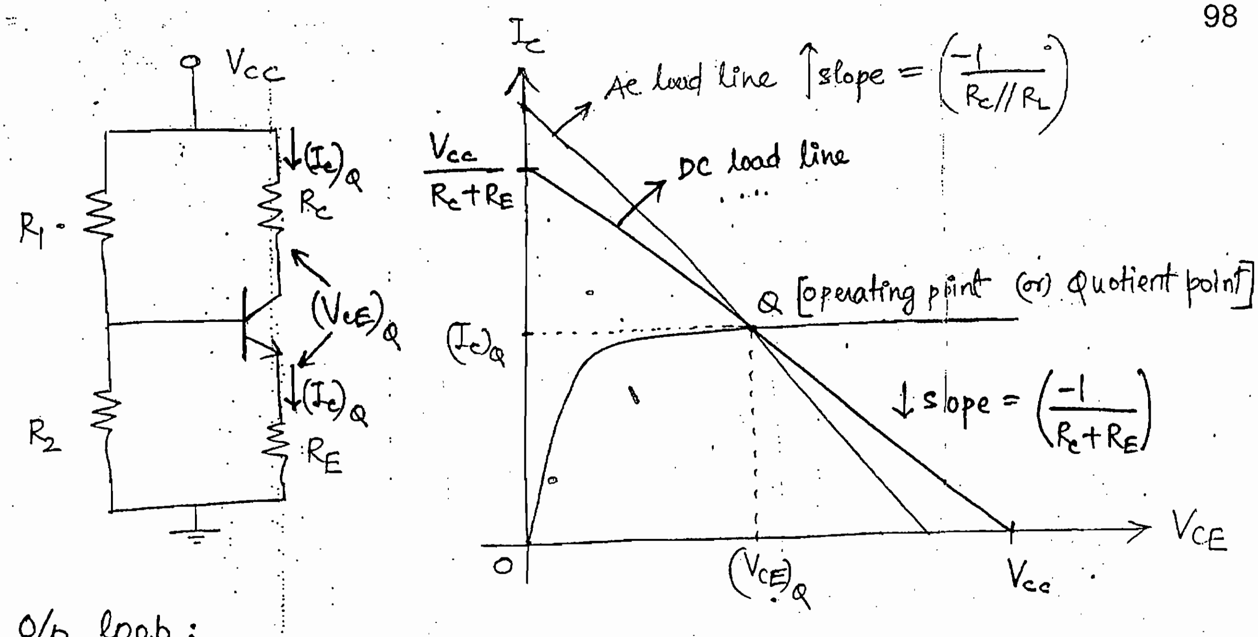
(Practical ckt)

DC Analysis:

(1). AC should be grounded.

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

($C \rightarrow OC$)



<u>O/p loop:</u>

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

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

(Quotient → greek word
↓
meaning → Inactive)

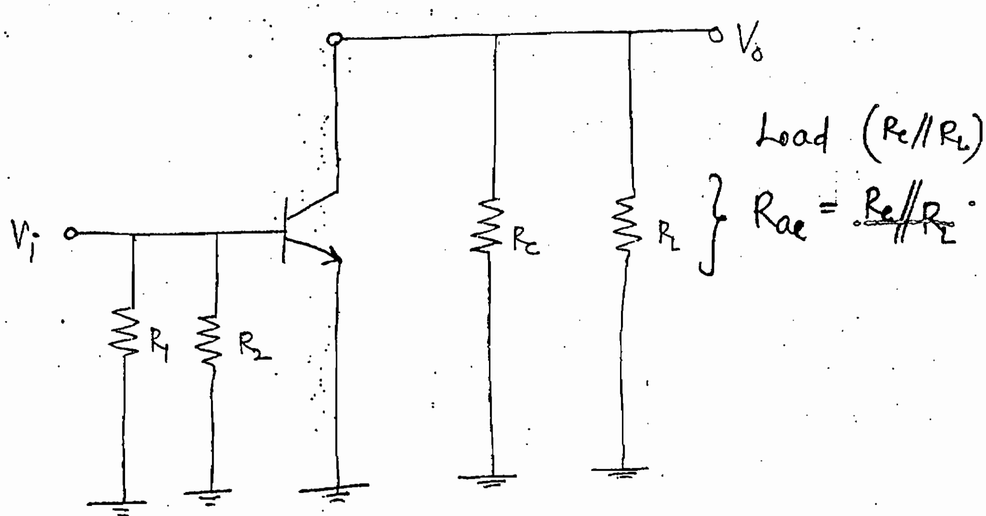
Load (R_c+R_E)

<u>AC Analysis:</u>

<u>AC load line</u> –

(1). DC should be grounded.

(2). $X_c \propto \frac{1}{f \to \infty} \to 0 \quad (C \to SC)$

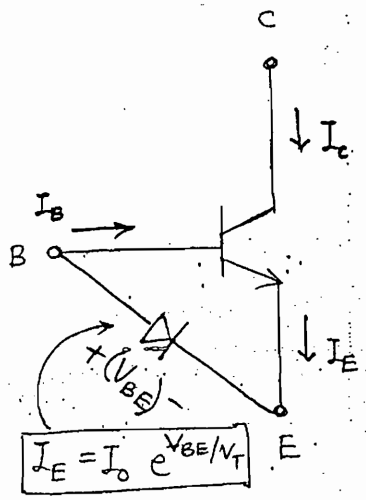


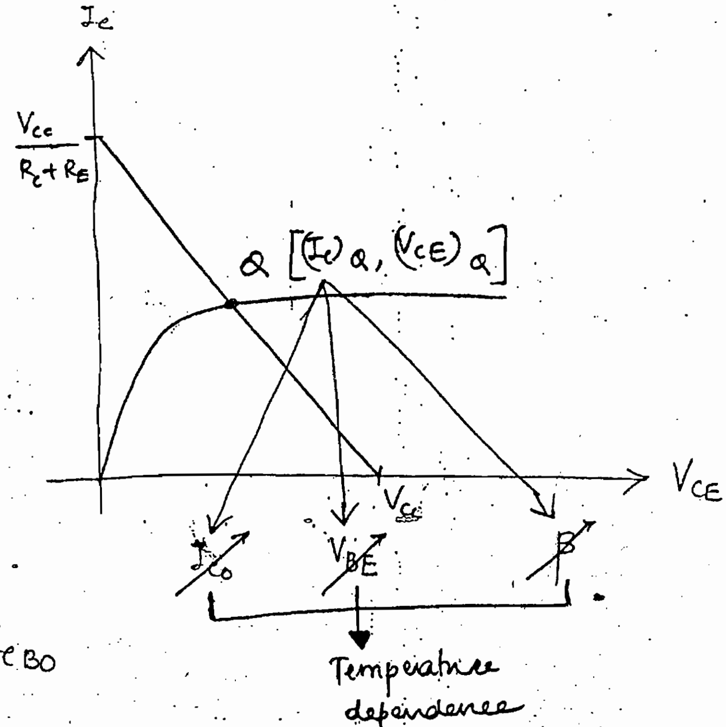
Load $(R_c // R_L)$

$R_{ac} = R_c // R_L$

Range: R_c, R_E, R_L → (0 to 10kΩ)

Temperature dependence on BJT parameters:





(1). $I_c = \beta I_B + (1+\beta) I_{CBO}$

(2). $I_c = I_o \cdot e^{V_{BE}/V_T}$

(3). $I_c = \beta I_B$

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

I_{co} Vs Temperature:

'I_{co}' increases by 7% for every °c rise in temp.

'I_{co}' doubles for every 10°c rise in temp.

V_{BE} Vs Temp.:

$$\frac{dV_{BE}}{dT} = -2.5\,mV/°c$$

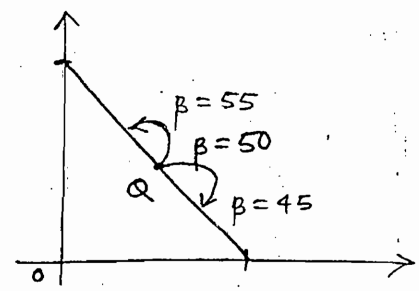
$T\uparrow$, $V_{BE}\downarrow$, $I_c\downarrow$

β Vs temp.:

$\beta = \frac{I_c}{I_B}$; $T\uparrow$, $\underbrace{I_{co}\uparrow, I_c\uparrow}_{\text{neglected}}$, $\beta\uparrow$,

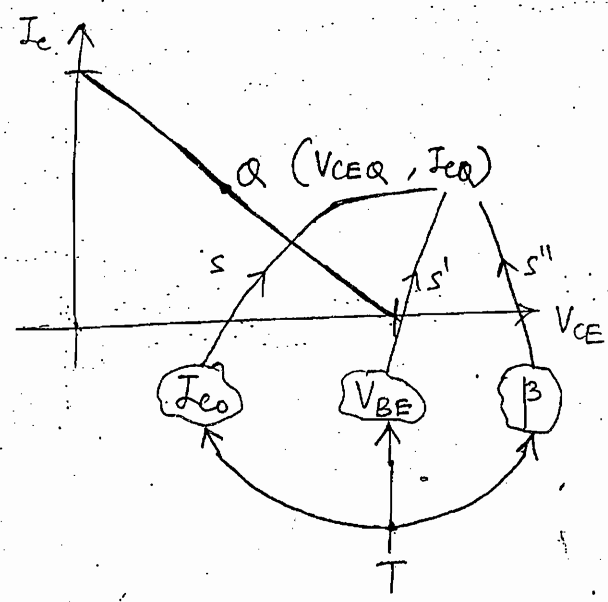
β replacement problem:

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".



$\boxed{S \propto S' \propto S''}$

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

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

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

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,

$$1 = \beta \frac{\partial I_B}{\partial I_c} + (1+\beta) \cdot \frac{\partial I_{co}}{\partial I_c}$$

$$1 = \beta \frac{\partial I_B}{\partial I_c} + \frac{(1+\beta)}{S}$$

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

$S=0$ (ideal) $\longrightarrow$ $S=1$ (practical)

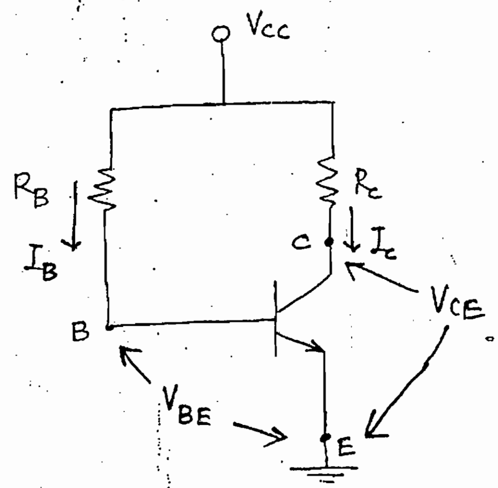
Compensation

Biasing ckts:

- Stabilization Techniques
  - Fixed Bias ($S \approx 101$)
  - Collector to base Bias ($S \approx 45$)
  - Voltage divider Bias ($S \simeq 1$) (or) Self Bias (or) Emitter Bias (or) Universal Bias
- Compensation techniques
  - Diode
  - Thermistor
  - Sensistor

$\underbrace{\text{Diode, Thermistor, Sensistor}}_{\text{temp. sensitive devices}}$

Fixed Bias :



i/p loop :

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

o/p loop :

$V_{cc} = I_c R_c + V_{CE}$

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

If $V_{cc} >> V_{BE}$

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

$$\underset{\text{fixed}}{\underset{\downarrow}{I_c}} = \beta \underset{\text{fixed}}{\underset{\downarrow}{I_B}}$$

$$I_B \approx \frac{V_{cc}}{R_B}$$

$$\frac{\partial I_B}{\partial I_c} = 0$$

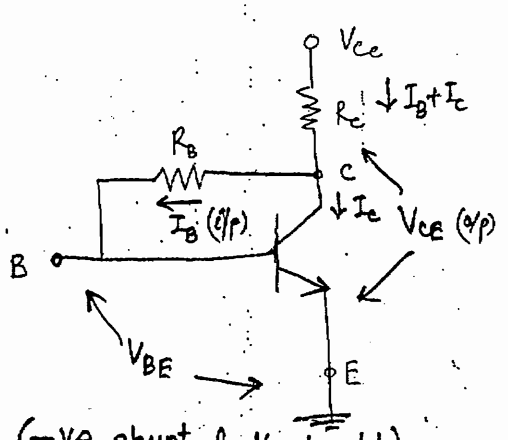
$$S = \frac{1+\beta}{1-\beta \frac{\partial I_B}{\partial I_c}}$$

$$\boxed{S = 1+\beta}$$

↓

Stability is not there for the ckt.

Collector to Base Bias : (–ve feedback Control s/s)



(–ve shunt feedback ckt)

i/p loop :

$V_{cc} = (I_B + \cancel{I_c}) \cancel{R_c} + I_B R_B + V_{BE}$

$V_{cc} = I_B (R_B + R_c) + I_c R_c + V_{BE}$

$$I_B = \frac{V_{cc} - I_c R_c - V_{BE}}{R_B + R_c}$$

O/p loop:

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

$$V_{CC} - I_C R_C = V_{CE} + I_B R_C$$

$$\therefore \underset{\downarrow \; i/p}{I_B} = \frac{\overset{\nearrow o/p}{V_{CE}} + I_B R_C - V_{BE}}{R_B + R_C}$$

Stabilization:

$T\uparrow$, $I_{CO}\uparrow$, $I_C\uparrow$, $(I_B + I_C) R_C \uparrow$, $V_{CE}\downarrow$

$I_B\downarrow$ $I_C\downarrow$

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

differentiate w.r.t I_C: We get,

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

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

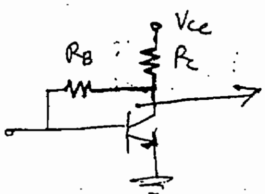
$$S = \frac{1+\beta}{1 - \beta \frac{\partial I_B}{\partial I_C}}$$

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

$$\boxed{S = \frac{1+\beta}{1 + \beta \frac{1}{\left(\frac{R_B}{R_C} + 1\right)}}} \rightarrow \approx 1 \text{ if } \frac{R_B}{R_C} = 0 \left\langle \begin{matrix} R_B = 0 \\ R_C = \infty \end{matrix} \right.$$

$\therefore$ $\boxed{R_B / R_C \ll 1}$ → failure condition

Becoz:



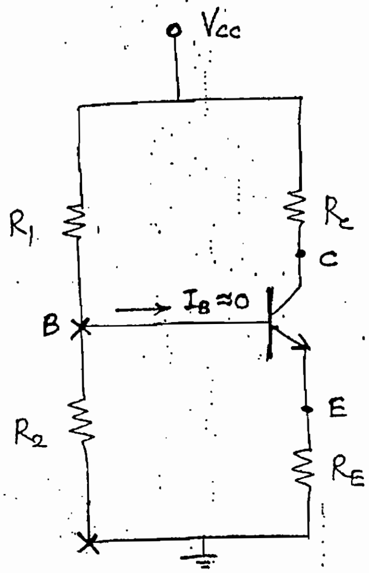
→ $R_B > R_C$ $\boxed{\frac{R_B}{R_C} \gg 1}$ — ckt condition

for $V_B < V_C$

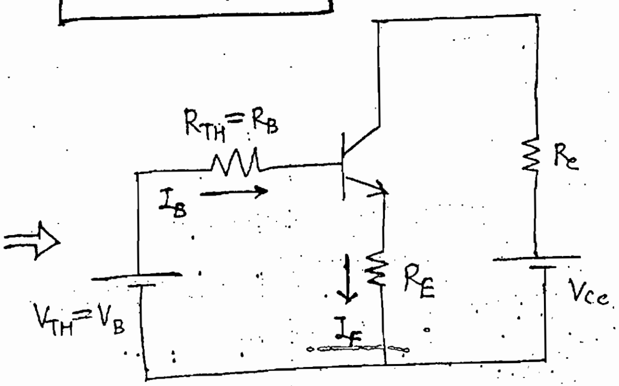
(active Region requirement condition)

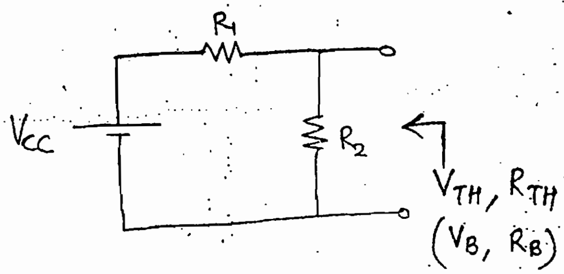
Voltage Divider Bias–

$$\boxed{V_B = \frac{V_{cc} \times R_2}{R_1 + R_2}} \qquad \boxed{R_B = R_1 // R_2}$$



⇒





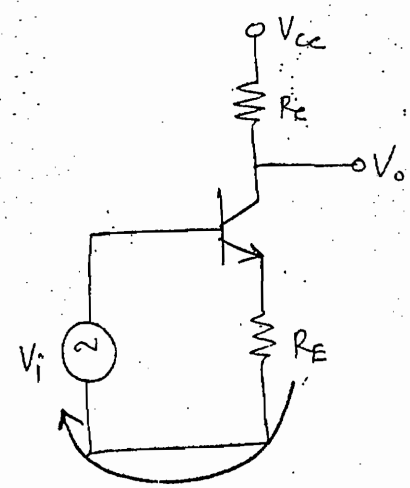
I/p loop :

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

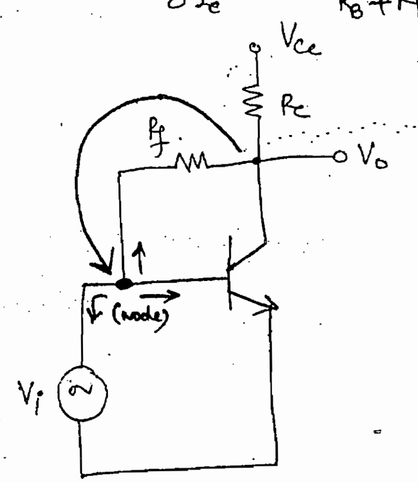
$$V_B = I_B \cdot (R_B + R_E) + I_c R_E + V_{BE}$$

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

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



Series feedback.



Shunt feedback

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

If $R_B / R_E << 1 \Rightarrow S \to 1$

Design conditions:

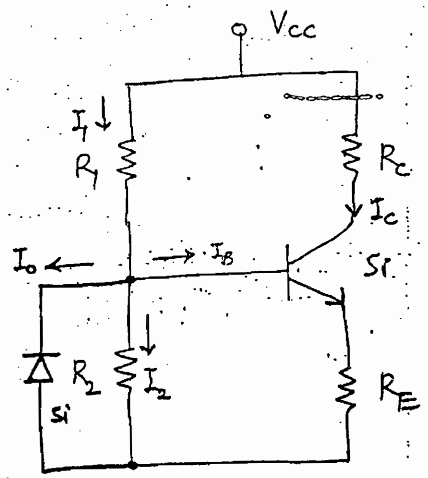
(1). If $\frac{R_B}{R_E} << 1$, then $S \rightarrow$ unity.

(2). To make R_B less, take a condition $R_1 > R_2$.

(3). If R_E increases, -ve fb in the ckt increases which will affect the s/g gain in AC analysis. To solve this problem always keep a capacitor (By pass capacitor C_E) in shunt to R_E.

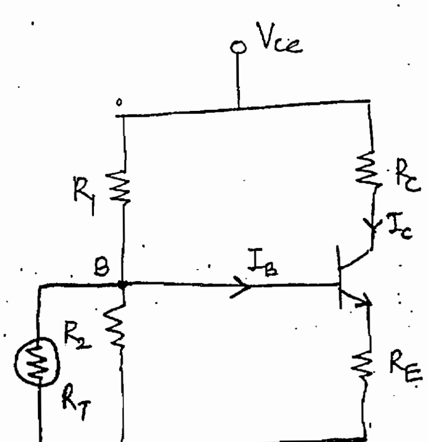
Compensation Techniques –

Compensation through Diode –



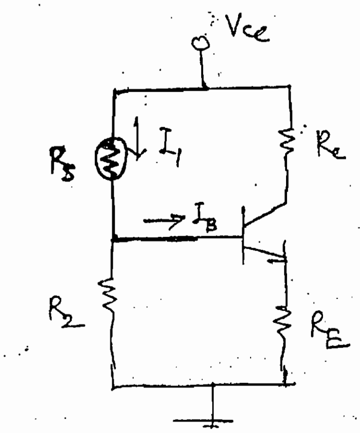
$T\uparrow \quad I_{co}\uparrow \quad \underbrace{I_D\uparrow \quad I_B\downarrow \quad I_C\downarrow}_{\text{Compensation}}$

Compensation through thermistor:



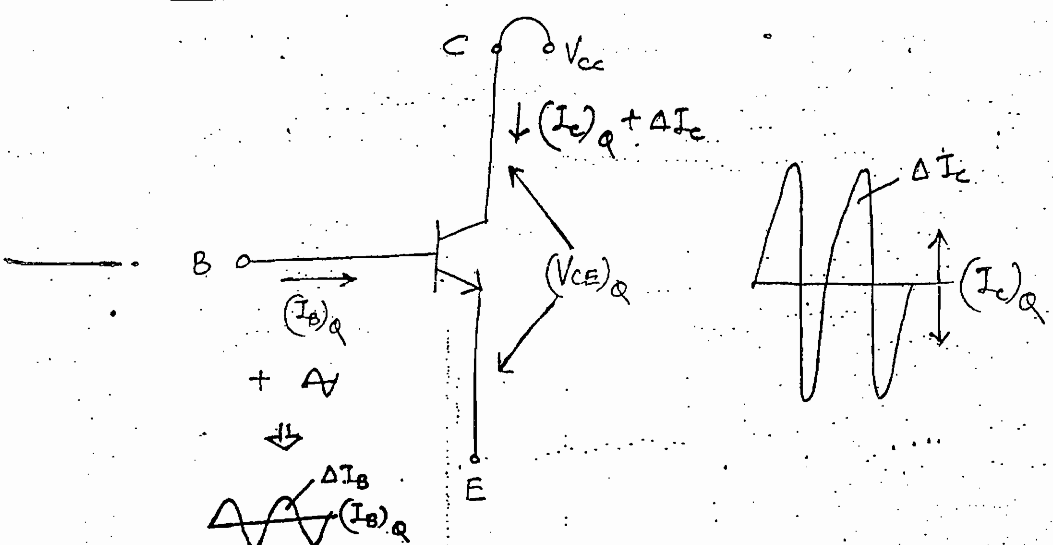
$T\uparrow, \ I_{co}\uparrow, \ \underbrace{R_T\downarrow \quad V_B\downarrow \quad I_B\downarrow \quad I_C\downarrow}_{\text{compensation}}$

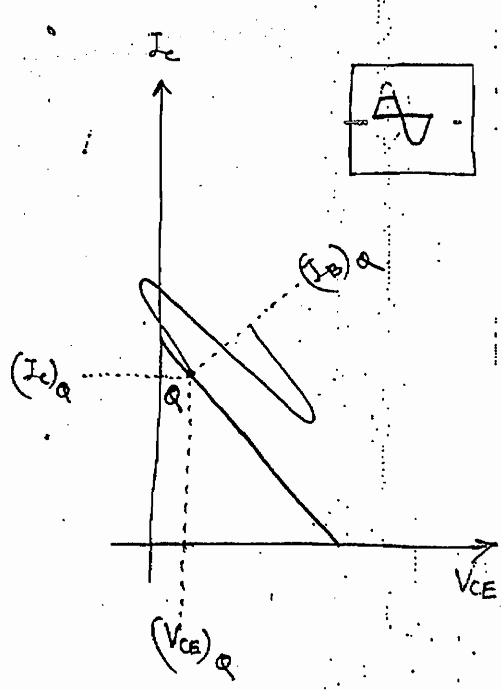
Compensation through Sensistor :

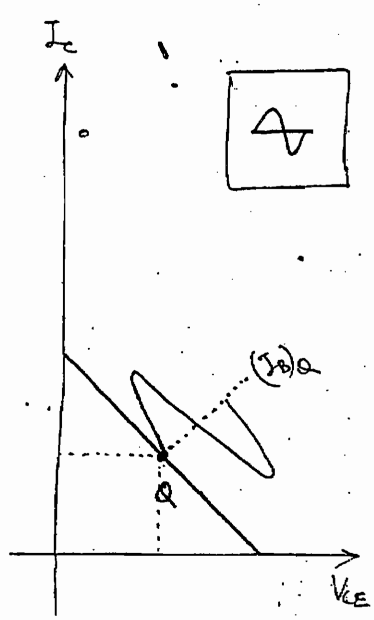


$$T\uparrow\ I_{co}\uparrow\ \underbrace{R_s\uparrow\ I_1\downarrow\ I_B\downarrow\ I_C\downarrow}_{\text{compensation}}$$

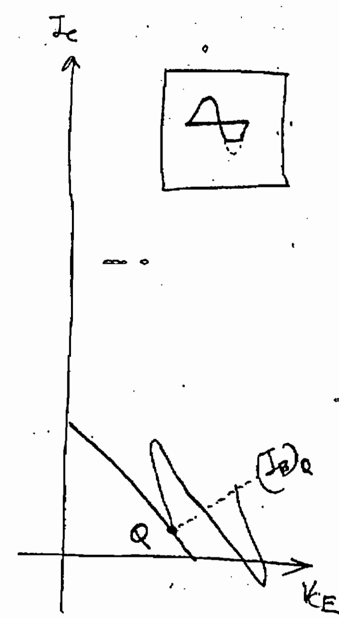
Position of Q point on DC load line :







(Best position of Q point)

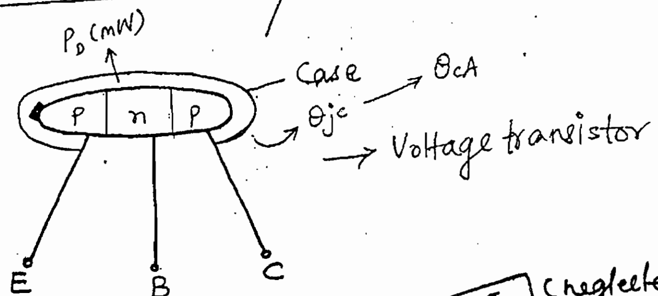


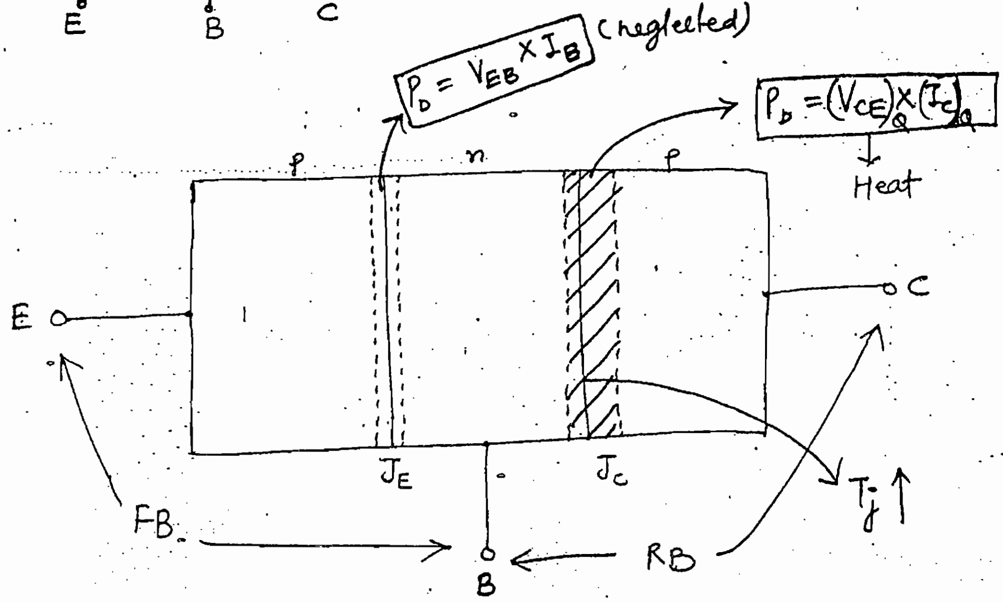
Conclusion:

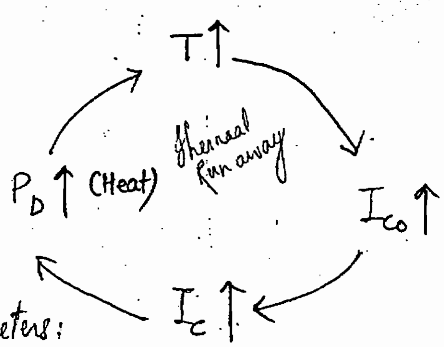
The amplifier analysis of the Q point should be design at the centre of DC load line.

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

Thermal Runaway:





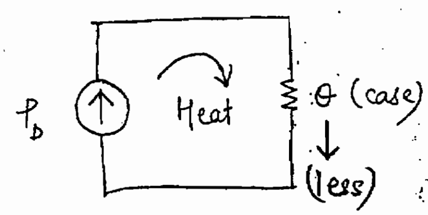


Thermal parameters:

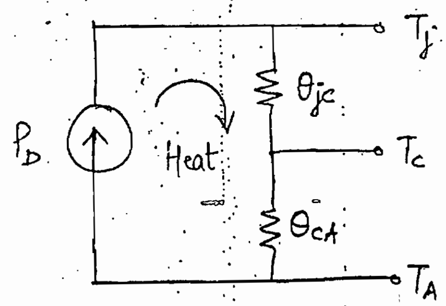
$$T_j - T_A \propto P_D$$

$$\underbrace{T_j - T_A}_{V} = \underbrace{\theta_j}_{R} \, \underbrace{P_D}_{I}$$

$V = RI$ (analogue)



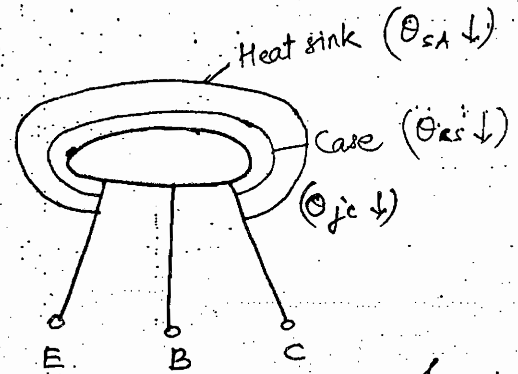
θ = thermal Resistance $= \frac{T_j - T_A}{P_D}$ °c/mW

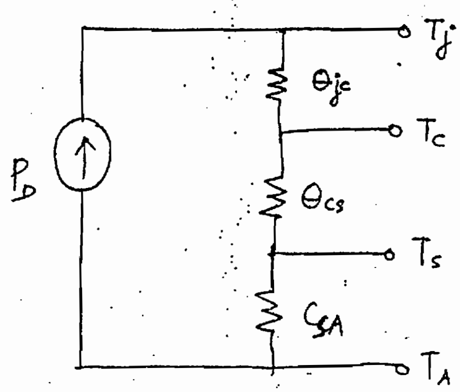


Problem with cases :

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,

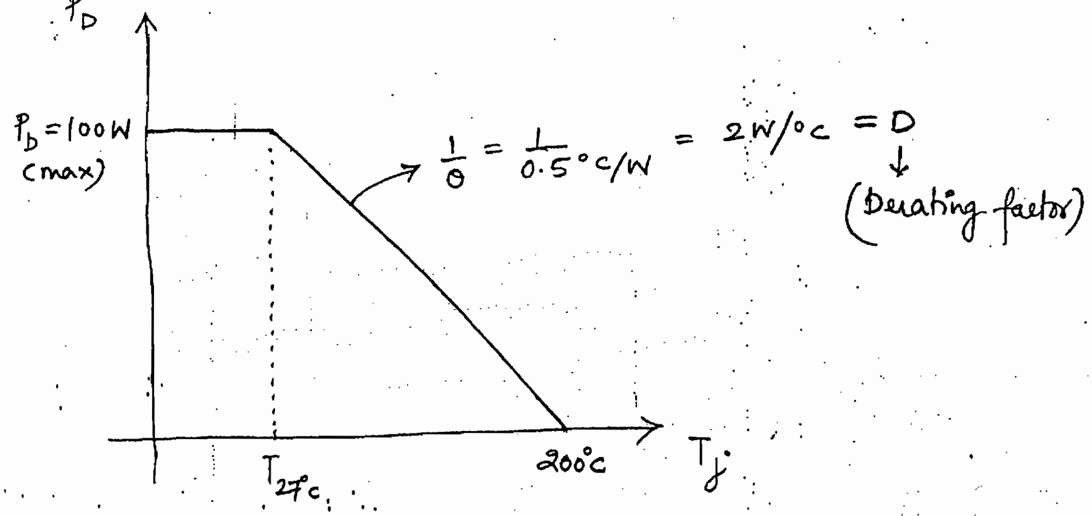
$$P_D = \frac{T_j - T_A}{\theta_{jc} + \theta_{cs} + \theta_{SA}}$$

For voltage transistor,

$$P_D = \frac{T_j - T_A}{\theta_{jc} + \theta_{CA}}$$

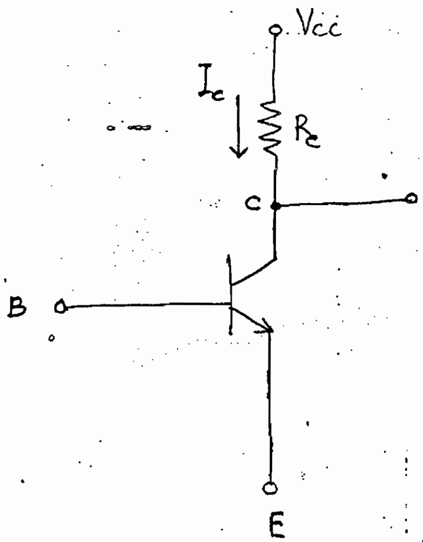
* Heat sink can deliver heat as fast as case absorbs from the device. Hence, there is no problem to total $\theta_{jA} = \theta_{jc} + \theta_{cs} + \theta_{SA}$ to be less.

Specification graph for BJT –



Condition for thermal stability:

$$\frac{\partial P_C}{\partial T_j} < \frac{\partial P_{D(\text{steady state})}}{\partial T_j}$$



$$\begin{aligned} P_c &= V_{cc} I_c - I_c^2 R_c \\ &= I_c \cdot (V_{cc} - I_c R_c) \\ &= (I_c)_Q \times (V_{CE})_Q \\ &= P_D \end{aligned}$$

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

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

Differentiate w.r.t T_j. We get,

$$1 - 0 = \theta \cdot \frac{\partial P_D}{\partial T_j}$$

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

$$\therefore \quad \frac{\partial P_c}{\partial T_j} < \frac{1}{\theta}$$

$$\frac{\partial P_c}{\partial I_c} \cdot \frac{\partial I_c}{\partial I_{co}} \cdot \frac{\partial I_{co}}{\partial T_j} = \frac{\partial P_c}{\partial T_j}$$

Now, $\quad P_c = V_{cc} I_c - I_c^2 R_c$

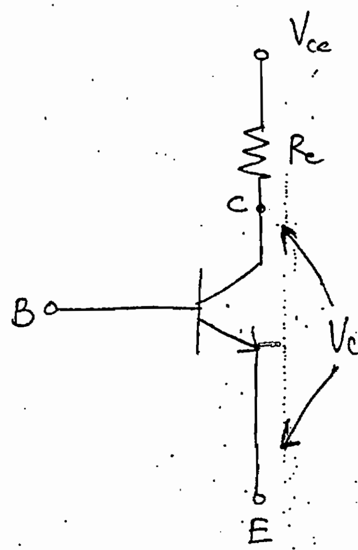
Differentiate w. r. t. I_c

(1). $\quad \frac{\partial P_c}{\partial I_c} = V_{ce} - 2 I_c R_c$

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

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

$$\therefore \quad \boxed{(V_{cc} - 2 I_c R_c) \times (S) \times (0.07 I_{co}) < \frac{1}{\theta}}$$



$V_{CE} = V_{cc} - I_c R_c$

$2 V_{CE} = 2 V_{cc} - 2 I_c R_c$

$2 V_{CE} - V_{cc} = V_{cc} - 2 I_c R_c$ ——(1)

$$\therefore \quad \boxed{(2 V_{CE} - V_{cc})\,(S) \cdot (0.07 I_{co}) < \frac{1}{\theta}}$$

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

$$\underbrace{(-V_{ce})\cdot(S)\cdot(0.07\, I_{co})}_{(-ve)} < \underset{(+ve)}{\frac{1}{\theta}}$$

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

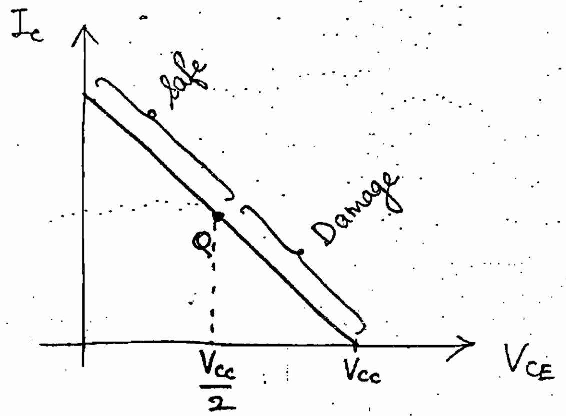
Case -(2) : $(V_{CE})_Q = V_{cc}/2$

$0 < \frac{1}{\theta}$ $\Rightarrow$ Thermally stable if $(V_{CE})_Q \le V_{cc}/2$

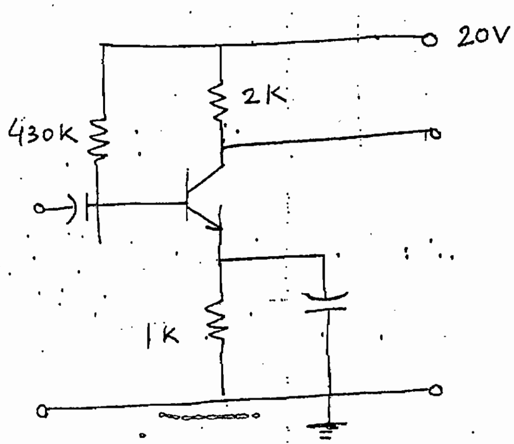
Case -(3) : $V_{CE} > V_{cc}/2$

There are chances of thermal run away.

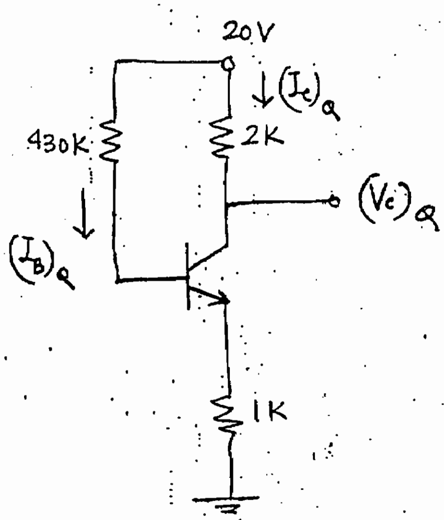
Thermal Runaway if $(V_{CE})_Q > V_{cc}/2$



Q9.



$\equiv$

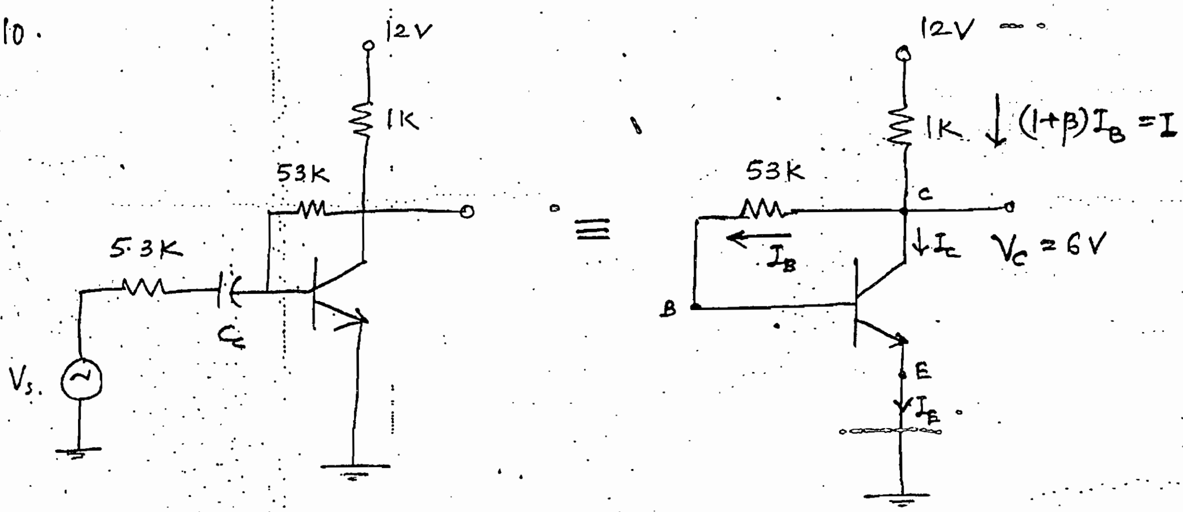


$$(I_B)_Q = \frac{20 - 0.7}{430K + (1+50)\,1K} = 40\,\mu A$$

$$(I_C)_Q = \beta (I_B)_Q$$
$$= 50 \times 40\,\mu A$$
$$= 2\,mA$$

$$(V_C)_Q = V_{cc} - (I_C)_Q \cdot R_C$$
$$= 20 - 2\,mA \times 2K$$
$$= 16\,V$$

Q10.



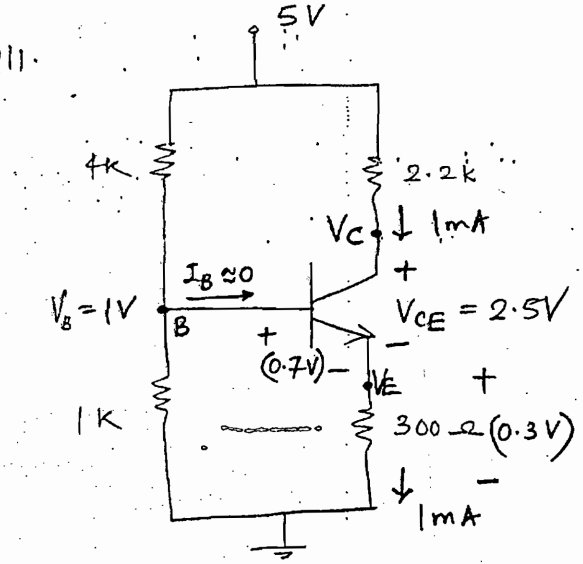
$$(I_B)_Q = \frac{12 - 0.7}{53K + (1+60)\,1K} = 99\,\mu A$$

$$(I_C)_Q = \beta \cdot (I_B)_Q$$
$$= 60 \times 99\,\mu A$$
$$= 5.94\,mA$$

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

$$V_C = 12 - 6\,mA \times 1K$$
$$= 6\,V$$

Q11.



$V_{CE} = 5 - 2.2K \times 1mA - 0.3k \times 1mA$

$= 5 - 2.2 - 0.3$

$= 2.5V$

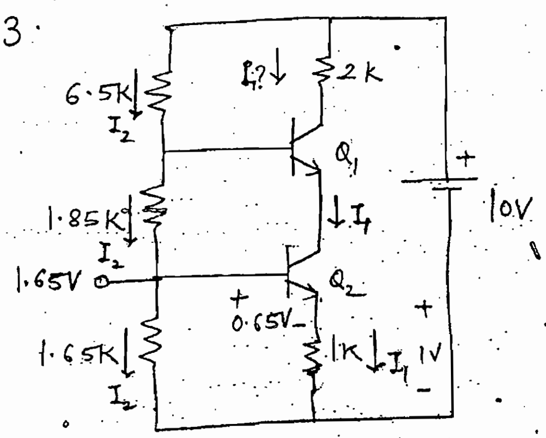
$V_B = \frac{1}{5} \times 5 = 1V$

$V_E = 1 - 0.7 = 0.3V$

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

$\therefore I_C = 1mA$

Q3.



$\beta \rightarrow$ High value (given)

$\uparrow \beta = \frac{I_C}{I_B \downarrow}$

Assume $I_B \approx 0$

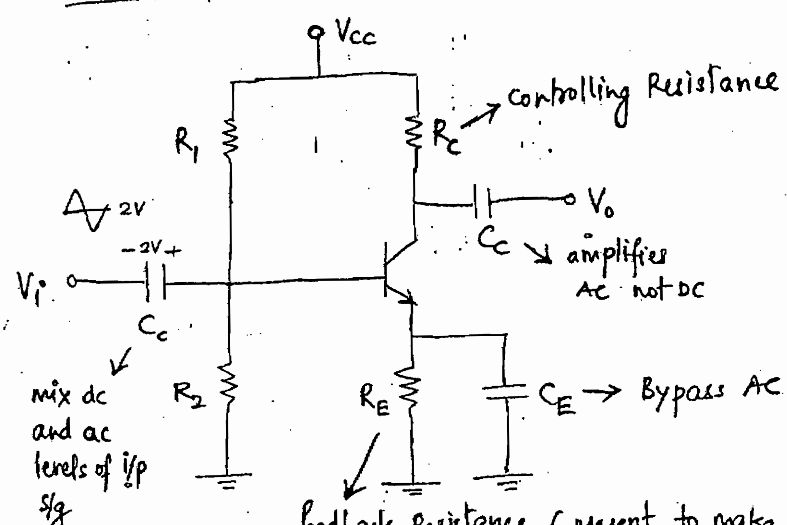
$1.65 - 0.65 = 1V$

$I_1 = \frac{1V}{1K} = 1mA$

BJT Applications:

CE Amplifier:

RC coupled BJT Amplifier / AC amplifiers:



$R_C \rightarrow$ controlling Resistance (always present to safe the o/p from direct supply)

$C_c \rightarrow$ amplifies AC not DC

R_1 & $R_2 \rightarrow$ Biasing Resistance

$C_c \rightarrow$ mix dc and ac levels of i/p s/g

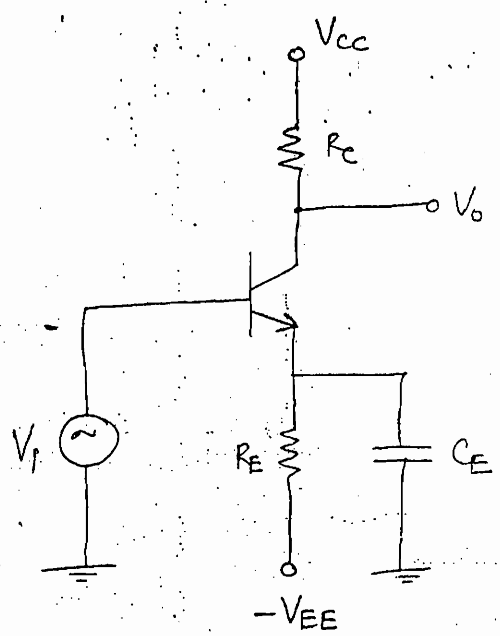
$C_E \rightarrow$ Bypass AC

$R_E \rightarrow$ feedback Resistance (present to make $S \approx 1$) Only for DC

(1). It is amplifier AC wise.

(2). App – Voltage Amplifier

Direct Coupled BJT Amplifier (DC Amplifiers):



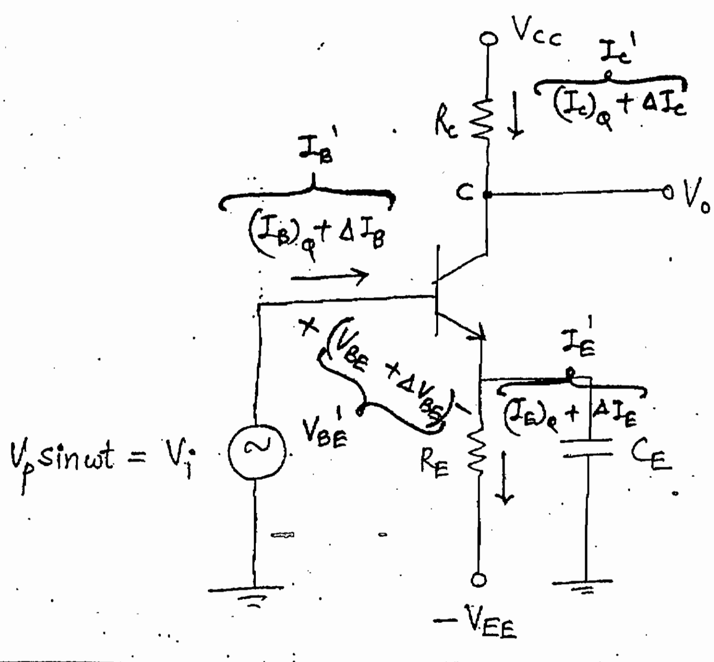
Stable Biasing
↓
Does not require biasing Resistance.

(1). App – Integrated ckts (OP-Amp)

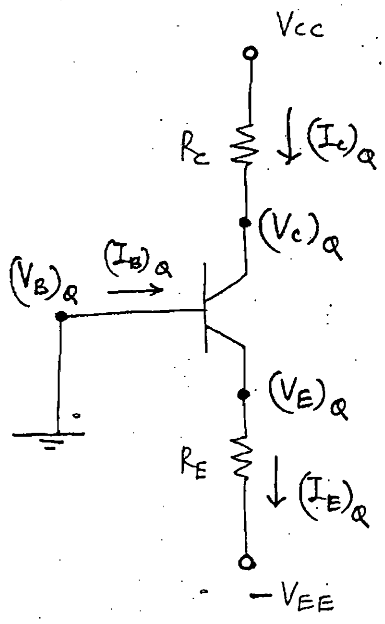
(2). They have high gain (very high)

(3). It is amplifier DC wise

Analysis of CE amplifier:



DC Model



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

$$(V_B)_Q = 0$$

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

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

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

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

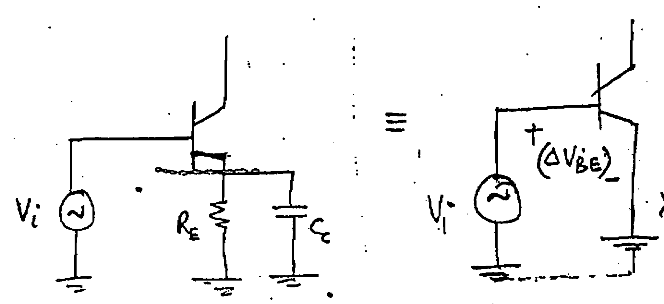
$$(I_E)_Q = I_{E0} \cdot e^{V_{BE}/V_T}$$

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

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

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

AC analysis –

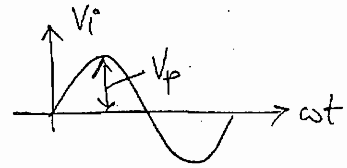


$X_{C_c} \propto \frac{1}{f} \xrightarrow{f \to \infty} 0$ (sc)

$\Rightarrow \Delta V_{BE} = V_i$

$$\therefore \boxed{I_E' = (I_E)_Q \cdot e^{V_i/V_T}} \rightarrow \text{Non-linear device}$$

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



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

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

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

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

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

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

Harmonics are not becoz of DC $\longrightarrow$

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

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

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

$\downarrow$

Harmonics are becoz of s/g

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

Now, If $\boxed{V_p <<< 4V_T}$

$V_p/4V_T <<< 1$ (neglected)

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

And now,

$$I_c' = \alpha I_E'$$

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

$$V_o = V_{cc} - I_c' R_c$$

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

$$= \underbrace{V_{cc} - (I_c)_Q R_c}_{(V_{CE})_Q} - \underbrace{\frac{(I_e)_Q}{V_T} \cdot R_c}_{\boxed{A_v = g_m R_c}} \cdot \underbrace{V_p \sin\omega t}_{V_i}$$

$$\boxed{V_o = (V_c)_Q - g_m R_c \cdot V_p \sin\omega t}$$

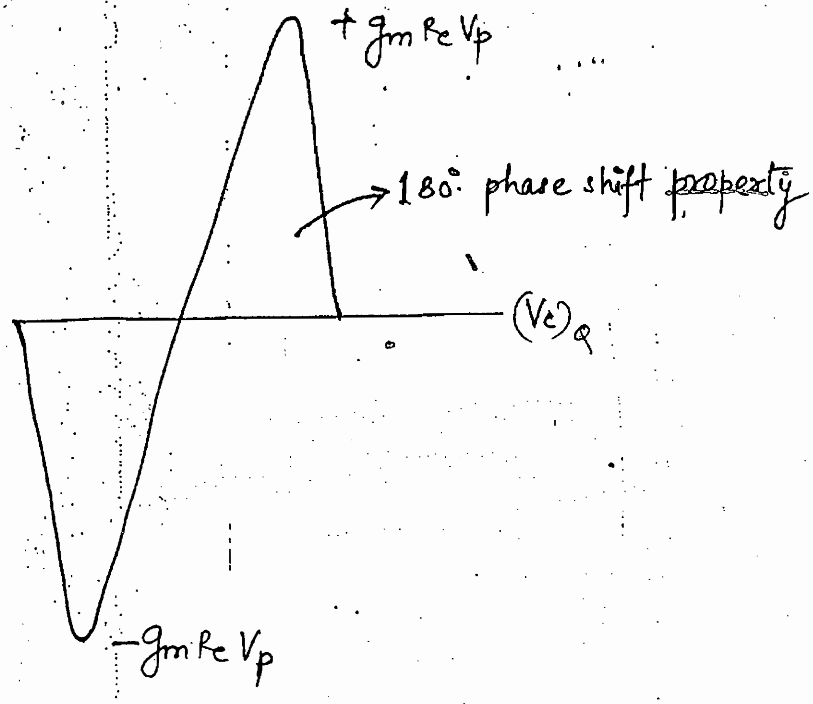
Amplification $\rightarrow A_V = g_m R_c$ and $g_m = \frac{(I_c)_Q}{V_T} \propto (I_c)_Q$

Hence, amplification in AC necessicates the requirement of DC i.e. $(I_c)_Q$ is becoz of DC and the factor which is responsible for gain.

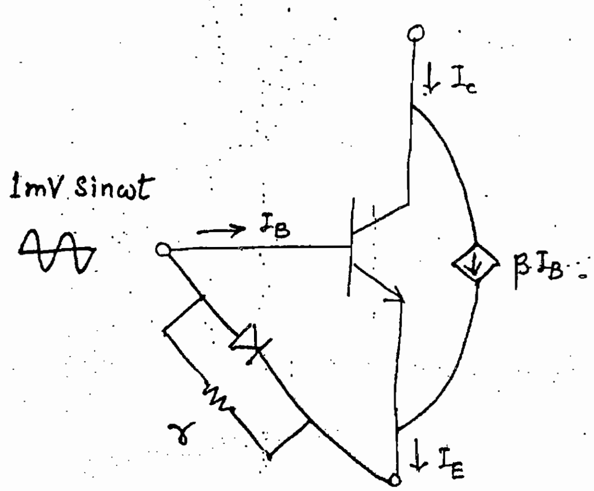
$$V_o = \underbrace{(V_c)_Q}_{\downarrow \text{ Direct coupled}} - \underbrace{g_m R_c \cdot V_p \sin\omega t}_{\downarrow \text{ RC coupled}}$$

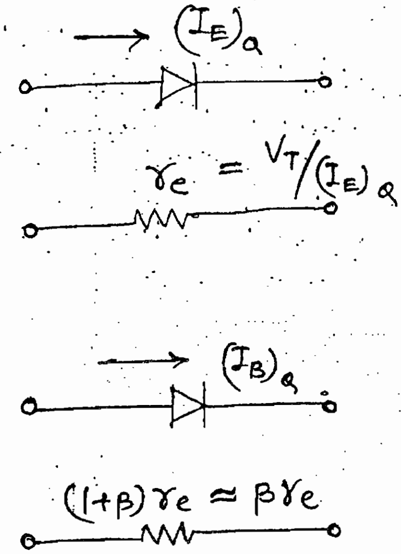
for RC coupled, $V_o = -g_m R_c \cdot V_p \sin\omega t$

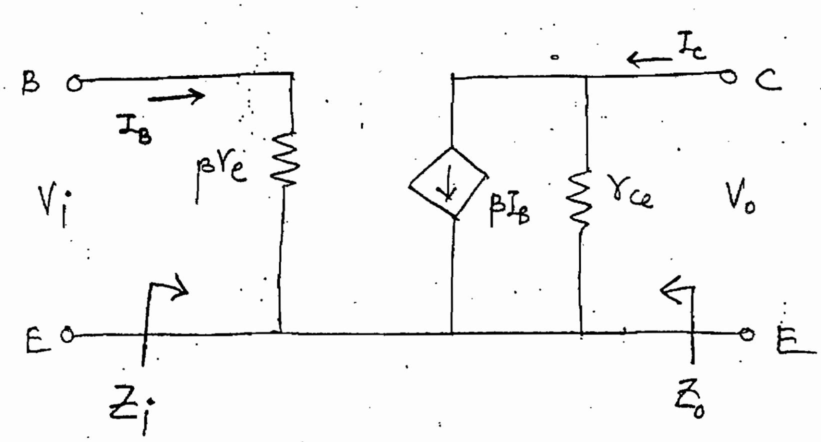
for Direct coupled, $V_o =$



Properties of CE amplifier –







(1) $Z_i = \beta r_e$

Assume, $\beta = 100$

$r_e = 10\,\Omega$

$\therefore Z_i = 1\,K\Omega$

(2). $Z_o = r_{ce} = 80\ K\Omega$

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

$= -\beta \cdot \dfrac{V_i}{\beta r_e} \cdot r_{ce}$

$\therefore V_o = -\dfrac{r_{ce}}{r_e} \cdot V_i$

$\therefore A_V = \dfrac{V_o}{V_i} = -\dfrac{r_{ce}}{r_e} > 1$

↓
180° phase shift

(4). $A_I = \dfrac{I_B}{I_B} = \beta > 1$

Conclusions:

(1). The i/p impedence is moderate.

(2). The o/p impedence 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.

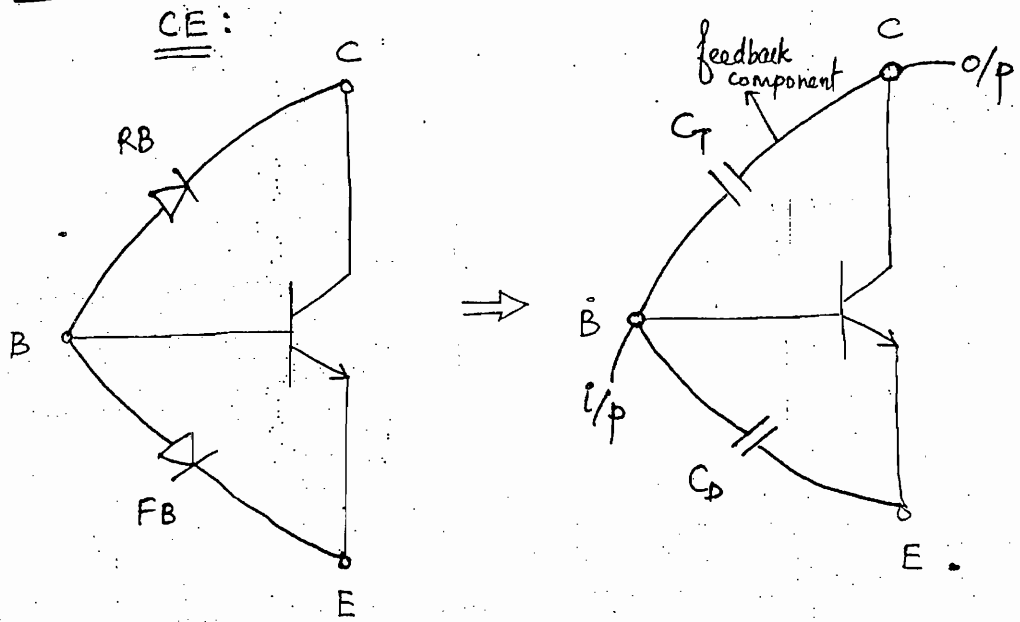
Note: In most of practical applications, CE is used becoz the power gain is high.

* 95% applications we CE configuration.

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

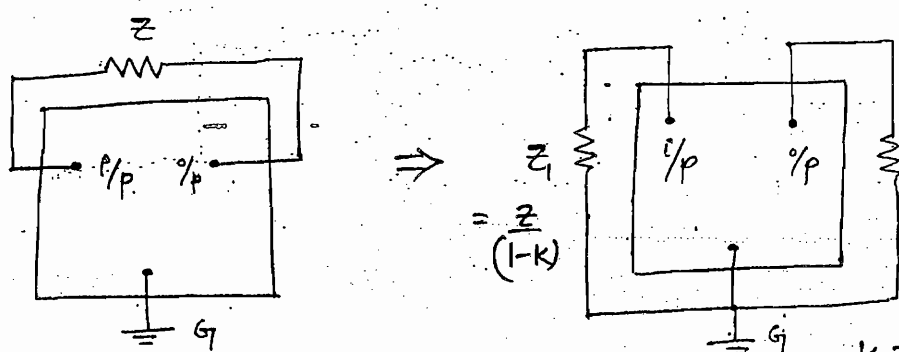
CB amplifier in High frequency :

CE :



(Active)

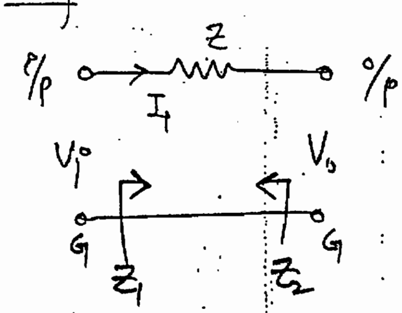
Millers theorem :



$$Z_1 = \frac{Z}{(1-K)} \qquad Z_2 = \frac{ZK}{(K-1)}$$

$K = A_v = \text{gain}$

Proof :



$$I_1 = \frac{V_i - V_o}{Z} \quad \text{OR} \quad I_1 = \frac{-V_o}{Z}\left(1 - \frac{V_i}{V_o}\right)$$

$$Z = \frac{V_i}{I_1}\left(1 - \frac{V_o}{V_i}\right) \qquad Z = Z_2\left(1 - \frac{1}{K}\right)$$

$$\boxed{Z = Z_1(1-K)} \qquad \boxed{Z_2 = \frac{ZK}{(K-1)}}$$

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

$$\frac{1}{C_{T_1}} = \frac{1}{C_T\,(1-A_v)}$$

$$C_{T_1} = C_T\,(1-A_v)$$

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

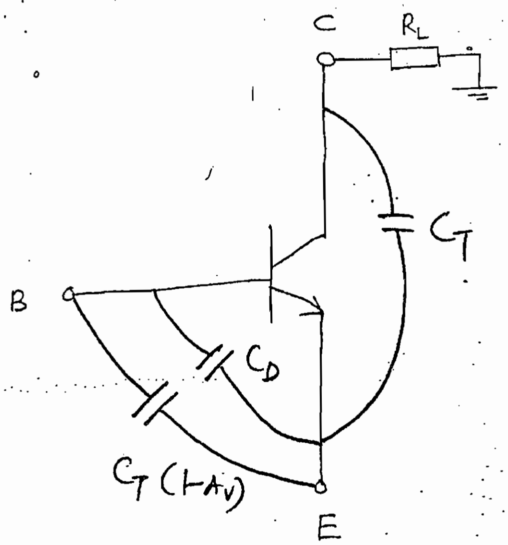
for high value of K (generally)

$$K-1 \approx K$$

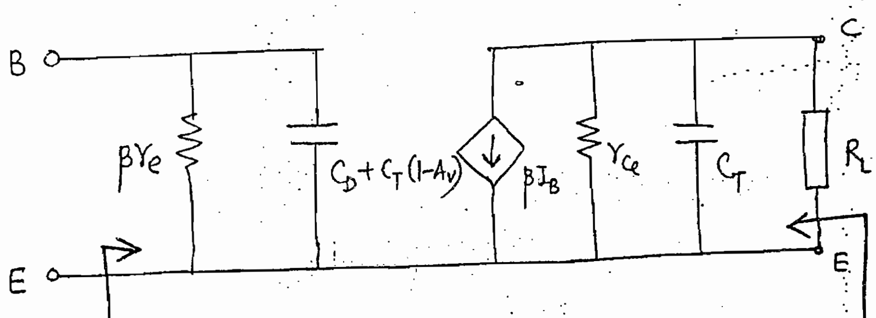
$$\therefore\ Z_2 = Z$$

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

$$C_{T_2} \approx C_T$$



$$\boxed{C_{in} = C_D + C_T\,(1-A_v)}$$ (high value)



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

$$\uparrow T_i = (\beta r_e)\,[C_D + C_T\,(1-\underset{(-ve)}{A_v})]$$

$$T_o \approx C_T R_L$$

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

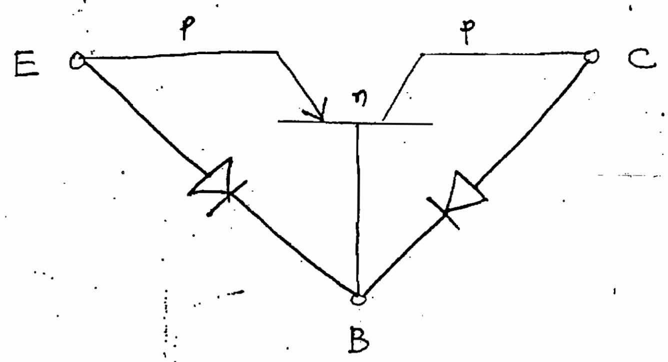
$$T_{total} = T_i + T_o \approx T_i$$

and $T_i >>> T_o$

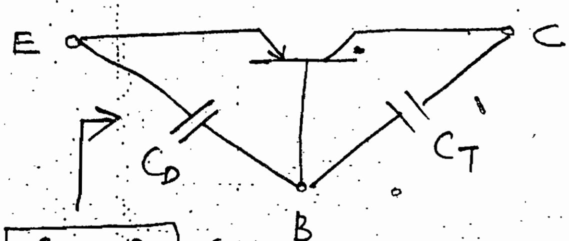
$$\therefore\ \omega_{BW}\downarrow = \frac{1}{T_i\uparrow}$$ (Hence, not preferred for high frequency)

$$\omega_{BW} = \frac{1}{(\beta r_e)\,[C_D + C_T\,(1-A_v)]}$$

CB ::



$\Downarrow$

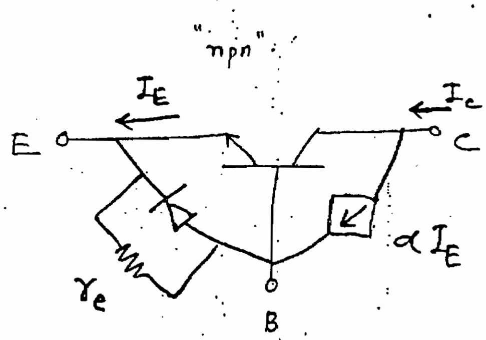


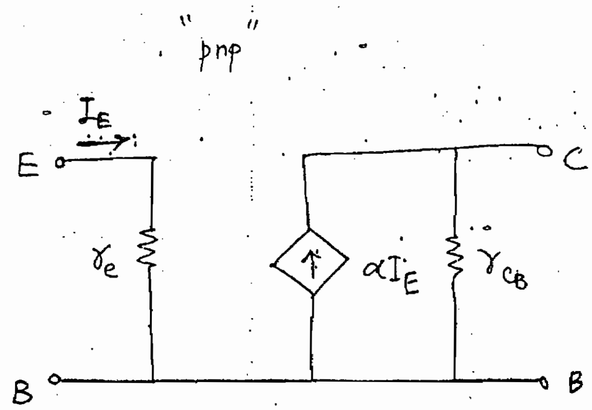
$\boxed{C_{in} = C_D}$ (low)

$$\uparrow \omega_{BW} = \frac{1}{T_i \downarrow}$$ (Hence, preferred for high frequency)

$$\omega_{BW} = \frac{1}{(\beta r_e)(C_D)}$$

Properties of CB amplifier :





(1). $Z_i = r_e$

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

$Z_i \approx 10\,\Omega$

(2). $Z_o = r_{CB}$

$= 4\,M\Omega$

(3). $A_V = \frac{r_{CB}}{r_e} > 1$ (No phase shift)

$\simeq \frac{4\,M\Omega}{10\,\Omega}$

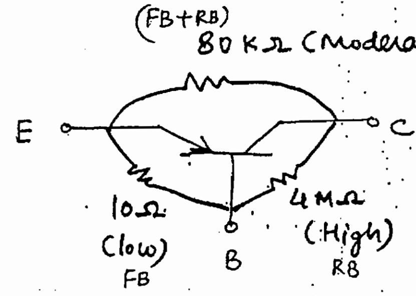
$\simeq 400\,K\Omega$

(4). $A_I = \frac{I_C}{I_E} = \alpha \leq 1$

If $A_I = 1$

$\boxed{I_C = I_E}$

Current Buffer



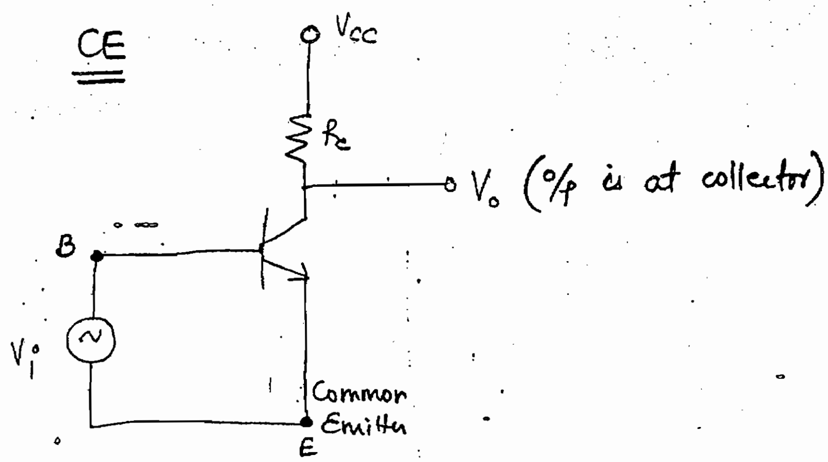
Conclusions:

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

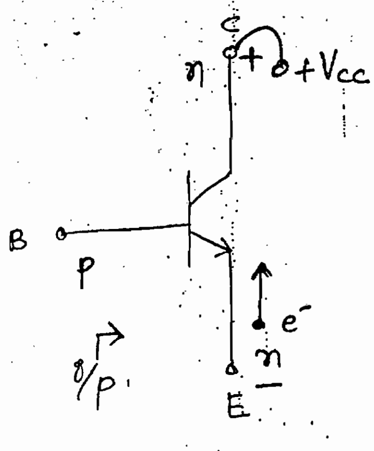
Note: In practical applications, CB amplifier is used whenever high BW is required.

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

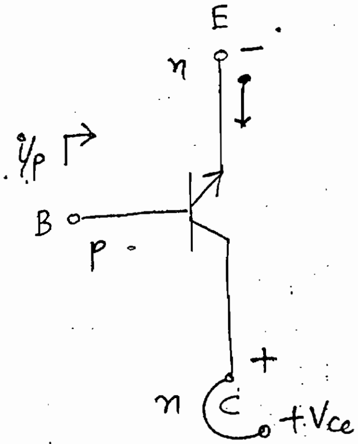
CC Amplifier: (or) Emitter follower.



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

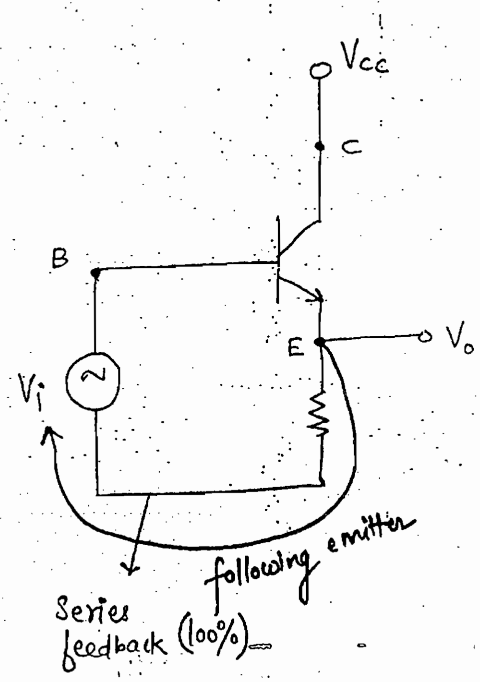


(CE)

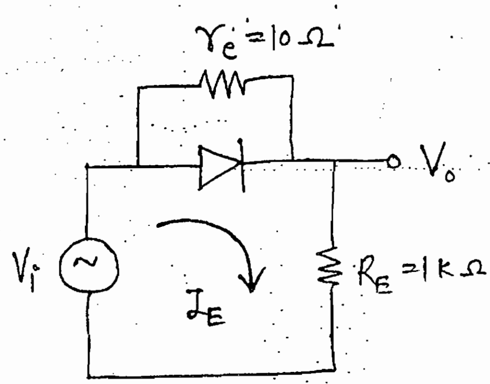


(CC)

CC :



$\equiv$

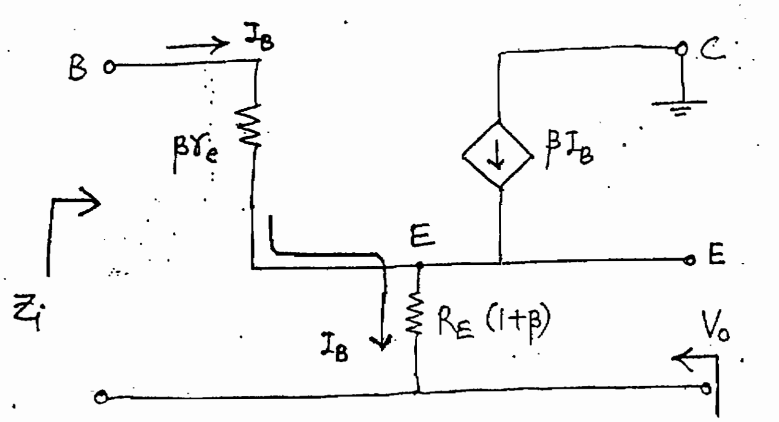


$$V_o = \frac{R_E}{R_E + r_e} \times V_i$$

But $R_E >>> r_e$

$\therefore$ $\boxed{V_o = V_i}$ → Voltage Buffer

Properties of CC amplifier:



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

Assume, $\beta = 100$

$r_e = 10\ \Omega$

$R_E = 10\ K\Omega$

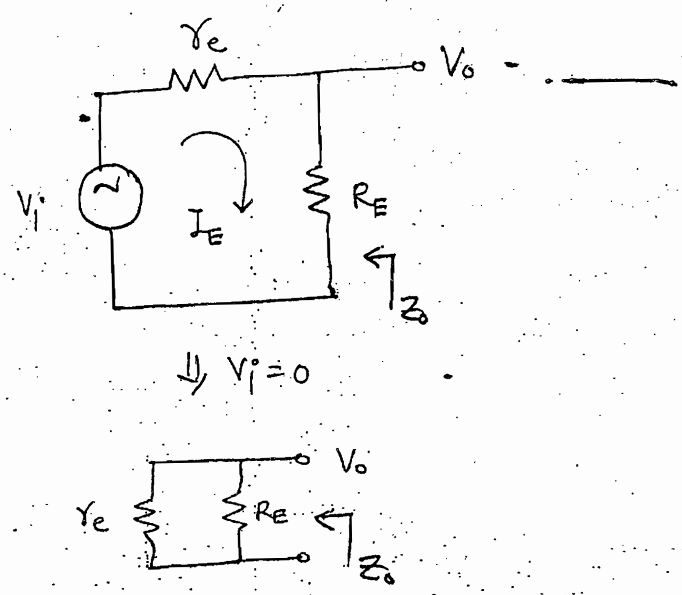
$$Z_i = (100)(10) + (100)(10K)$$

$$= 10K + 10^3\ K\Omega$$

$$\approx 10^3\ K\Omega \approx 1\ M\Omega$$

i.e. $Z_i = \beta R_E$

(2). $Z_o = \dfrac{V_o}{I_E}$



$Z_o = r_e // R_E$

$\approx r_e$

$\approx 10\ \Omega$

(3). $A_V = \dfrac{V_o}{V_i} \leq 1$

(4). $A_I = \dfrac{I_E}{I_B}$

$= 1+\beta$

$\approx \beta > 1$

Conclusions:

(1). The I/p impedence is high.

(2). The O/p impedence is low (impedence matching)

(3). The voltage gain is unity (voltage buffer)

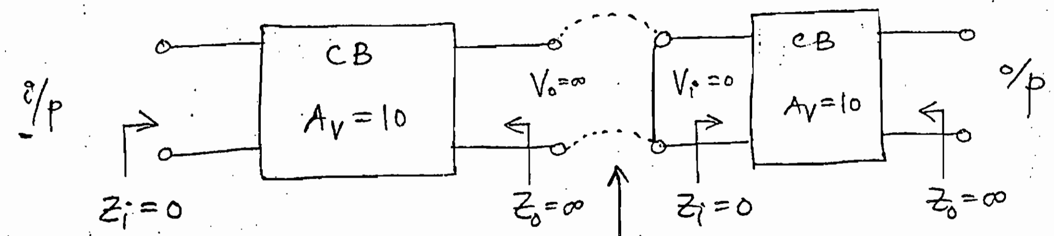
(4). The current gain is greater than unity.

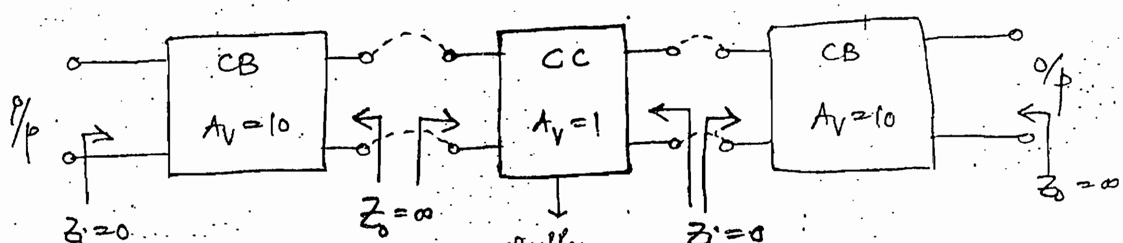
<u>Note:</u> In practical applications, CC is used for "impedence matching".

* Here also, power gain is not much high.

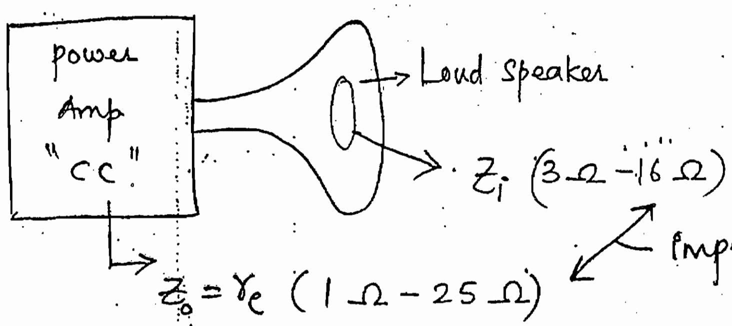
Common Collector in Impedence Matching :

<u>App-(1)</u>



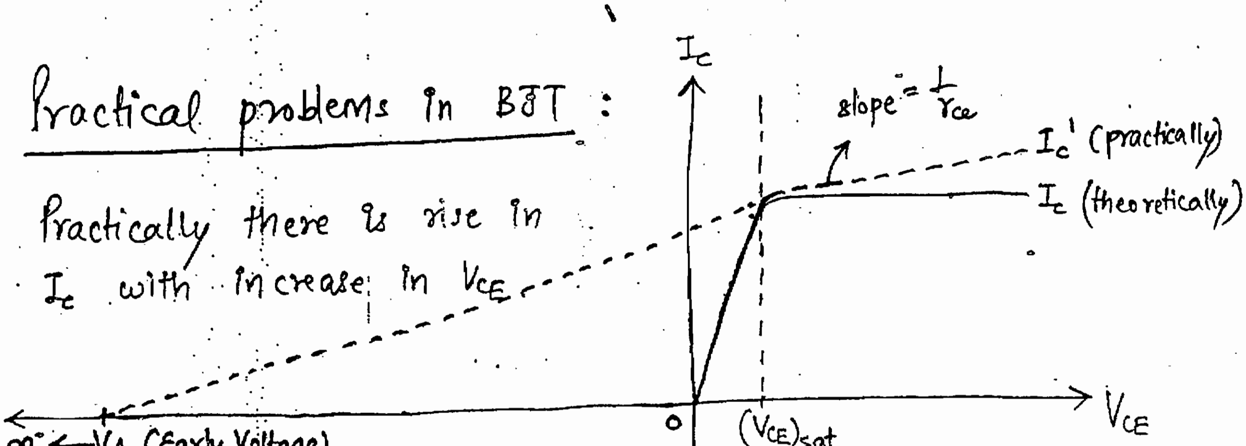


<u>App-(2)</u>



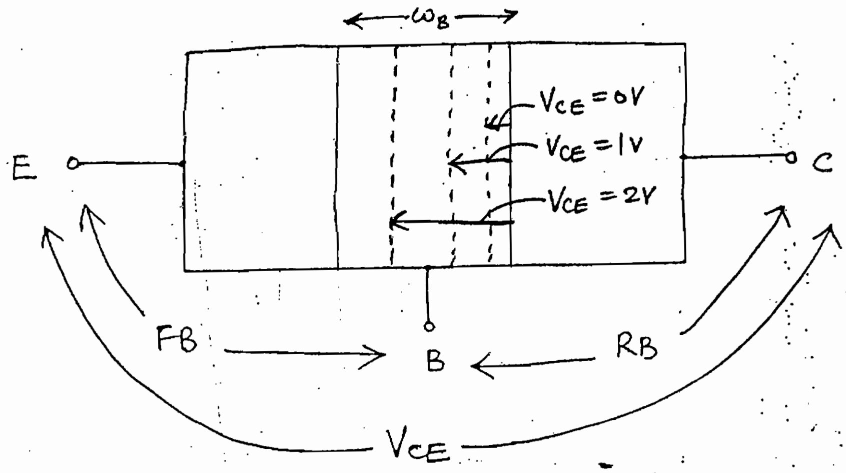
Practical problems in BJT :

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



Analysis-(i) : Block diagram Analysis

Early Effect (or) Base width Modulation : (Early → Scientist name)



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

$$\boxed{I_c = I_0 \cdot e^{V_{BE}/V_T}} \rightarrow \text{Theoritical eq}^n$$

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

o/p Resistance :

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

$$r_{ce} = r_0 = \frac{\partial V_{CE}}{\partial I_c'}$$

$$\frac{1}{r_0} = \frac{\partial I_c'}{\partial V_{CE}} = \frac{\partial}{\partial V_{CE}}\left[I_0 \cdot e^{V_{BE}/V_T}\left(1+\frac{V_{CE}}{V_A}\right)\right]$$

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

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

$$\therefore \boxed{r_0 = \frac{V_A}{(I_c)_Q}}$$

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

$$\therefore r_0 = \frac{1000\ V}{1\ mA} = 1M\Omega$$

* 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 → saturation current)

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

↓

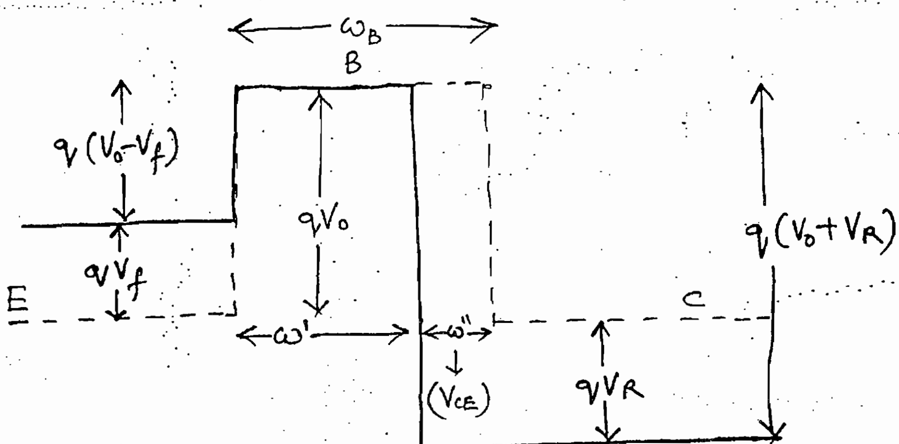
Injected minority current (or) diffusion current

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

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

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

Analysis – (3) : Energy Band theory



As $V_{CE}\uparrow$,

$$\boxed{\underbrace{\omega_B}_{\text{const.}} = \omega'\downarrow + \omega''\uparrow}$$

↓

Early Effect .

Effective base width

ω' → (mobile charge width)

ω'' → (Immobile charge width) penetration width

ω_B → physical width of base

$$\boxed{\begin{aligned}\omega_B &= \omega'' \\ \omega' &= 0\end{aligned}}$$ → punch through (or) Reach through / Special Breakdown

* Punch through takes place due RB application.

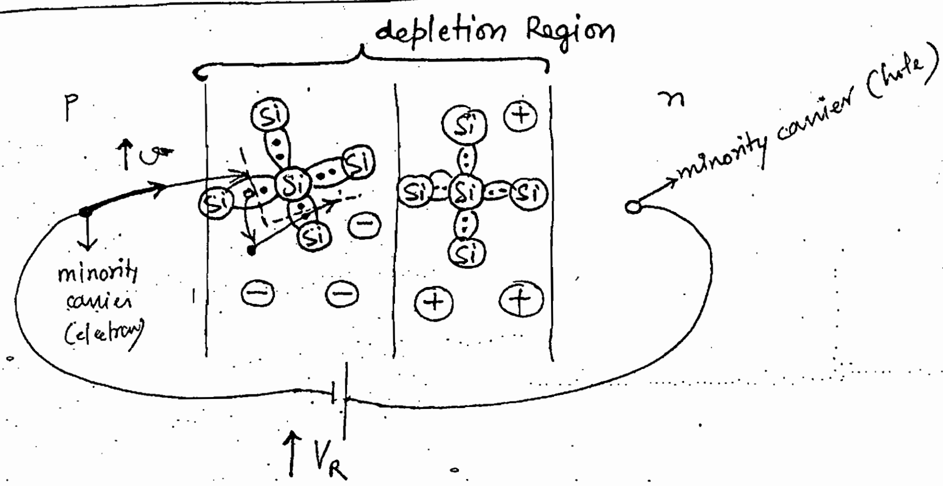
Breakdown mechanisms in BJT :

(1). Thermal runaway.

(2). Punch through (or) Reach through.

(3). Avalanche breakdown.

Avalanche breakdown in a Diode :



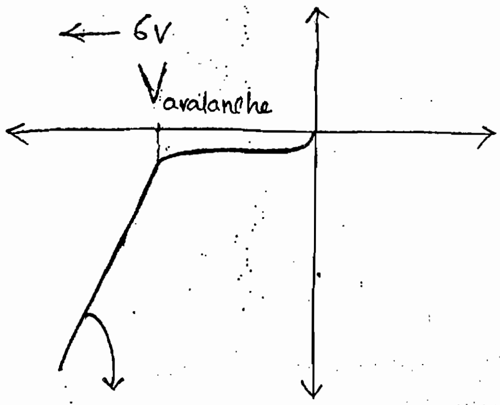
Explanation :

(1). In the depletion region, there will be some stable atoms along with immobile ions.

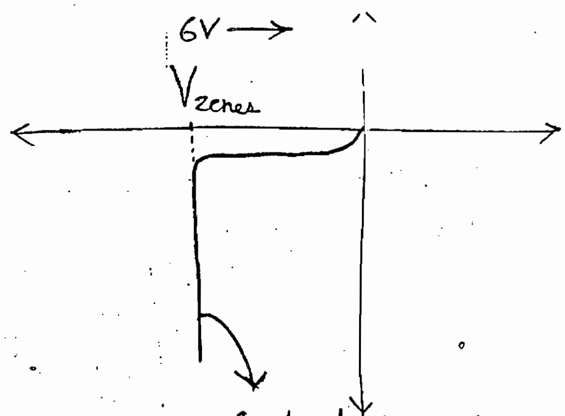
(2). When the RB voltage ↑ses, the velocity of minority charge carrier ↑ses which will break the covalent bond in the depletion region.

(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

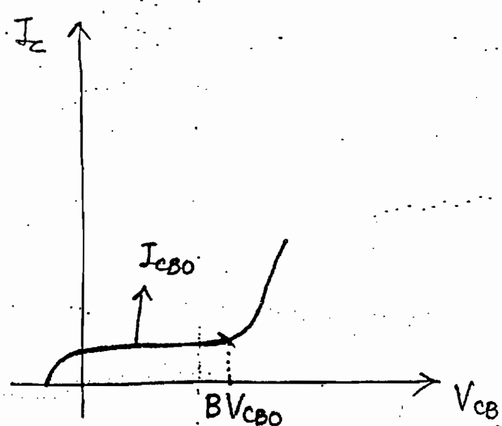


Instant Breakdown
(or)
Zener Breakdown
(Very fast phenomena)
Rapid current

$V_z < 6V$

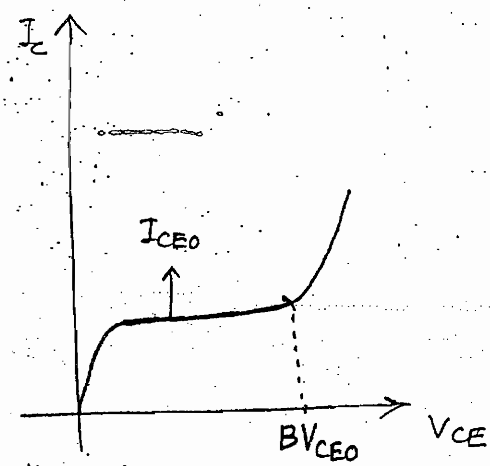
Above 6V, Varalanche →

CB



Damage later (slow current)

CE



(Rapid current)
Damage soon

$I_{CEO} > I_{CBO}$ $\quad [I_{CEO} = (1+\beta) I_{CBO}]$

$BV_{CEO} < BV_{CBO}$

$$BV_{CEO} = \frac{BV_{CBO}}{\sqrt[\eta]{\beta}}$$

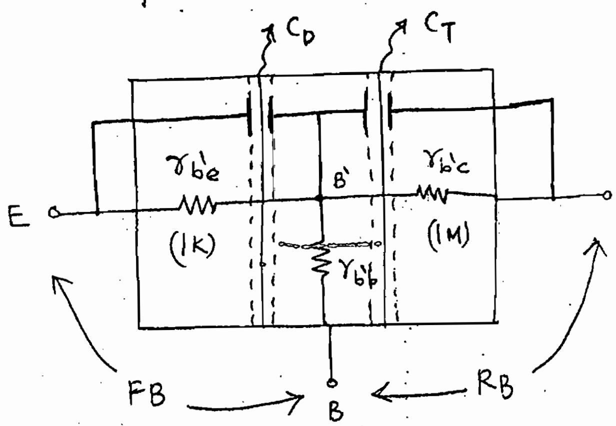
$\beta \rightarrow$ current gain in CE

$\eta \rightarrow$ avalanche multiplication factor

$\eta \rightarrow (3 \text{ to } 6)$

Small s/g Analysis of BJT :

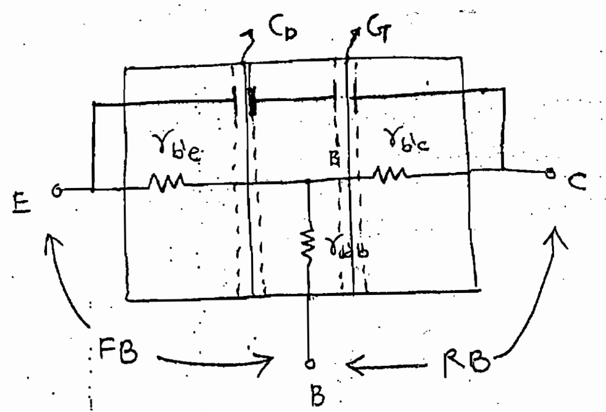
Low frequency Analysis : $f \to 0$



$$X_C \propto \frac{1}{f \to 0} \longrightarrow \infty \ (OC)$$

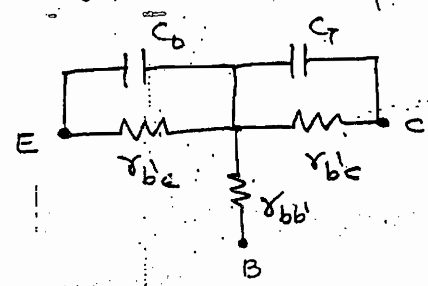
At low frequency analysis, C_D & C_T are assumed to be open (neglect).

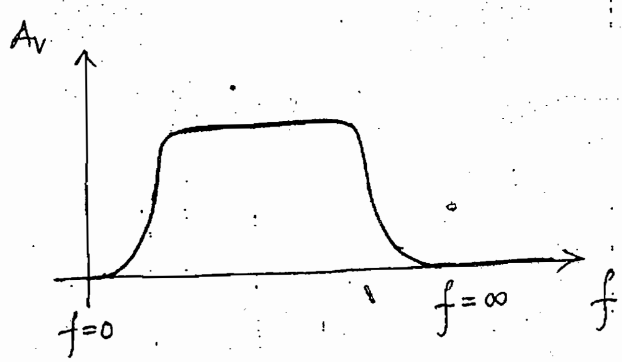
High frequency Analysis $f \to \infty$



$$X_C \propto \frac{1}{f \to \infty} \longrightarrow 0 \ (SC)$$

At high frequency analysis, C_D & C_T are taken in to account. (practically)





* Types of Basic amplifiers :

(1). Voltage amplifier

(2). current amplifier

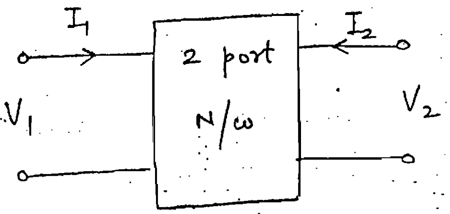
(3). Transconductance amplifier

(4). Transresistance amplifier

$V_i \rightarrow \boxed{A_V} \rightarrow V_o$ $\qquad$ $V_i \rightarrow \boxed{G_m} \rightarrow I_o$

$I_i \rightarrow \boxed{A_I} \rightarrow I_o$ $\qquad$ $I_i \rightarrow \boxed{R_M} \rightarrow V_o$

Linear 2 port N/ω –



$V_1 = f(I_1, I_2)$

$V_2 = f(I_1, I_2)$

Z parameters:

$V_1 = Z_i I_1 + Z_r I_2$

$V_2 = Z_f I_1 + Z_o I_2$

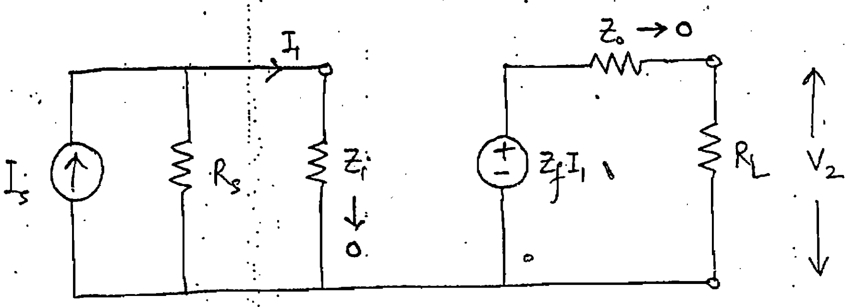
$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_i & Z_r \\ Z_f & Z_o \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

An ideal amplifier is a unilateral device

$\therefore Z_r = 0 \; ; \; Z_i = 0 \; ; \; Z_o = 0$

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ Z_f & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

$\boxed{V_2 = Z_f \cdot I_1} \rightarrow$ "ccvs"



Transresistance Amplifier "ccvs"

Y parameters :

$I_1 = f(V_1, V_2)$
$I_2 = f(V_1, V_2)$

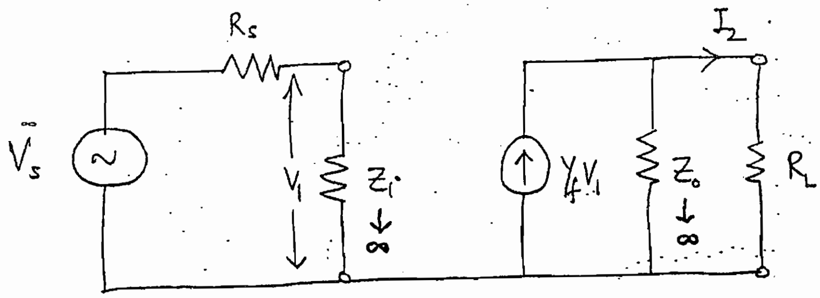
$$I_1 = Y_i V_1 + Y_r V_2$$
$$I_2 = Y_f V_1 + Y_o V_2$$

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_i & Y_r \\ Y_f & Y_o \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$

An ideal amplifier is a unilateral device

$Y_i = 0$; $Y_r = 0$; $Y_o = 0$

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ Y_f & 0 \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$ $\boxed{I_2 = Y_f V_1} \rightarrow$ "VCCS"



Trans conductance amplifier "VCCS" (FET/MOSFET).

H parameters :

$V_1 = f(I_1, V_2)$
$I_2 = f(I_1, V_2)$

$V_1 = \cancel{h_i I_1} + h_r V_2$
$I_2 = h_f I_1 + h_o V_2$

~~Case (1) : $V_2 = 0$~~

Case (1) : $V_2 = 0$

$h_i = \frac{V_1}{I_1} \rightarrow$ i/p impedence

$h_f = \frac{I_2}{I_1} \rightarrow$ forward current gain

Case (2) : $I_1 = 0$

$h_r = \frac{V_1}{V_2} \rightarrow$ Reverse voltage gain

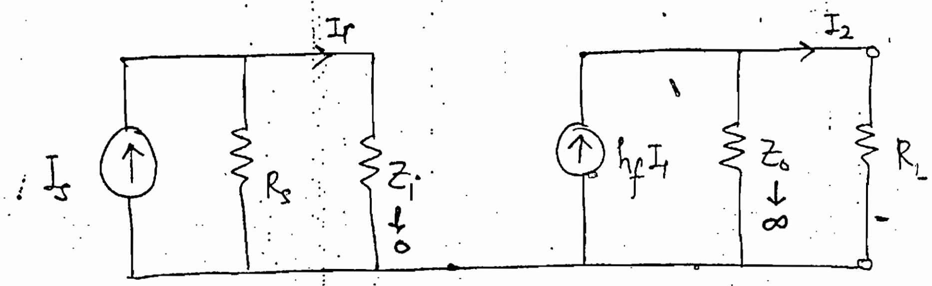
$h_o = \frac{I_2}{V_2} \rightarrow$ o/p admittance

An ideal amplifier is a unilateral device

$h_i = 0$; $h_r = 0$; $h_o = 0$

$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ h_f & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} \qquad \boxed{I_2 = h_f I_1} \rightarrow \text{"CCCS"}$$

$$I_C = \beta I_B$$



Current Amplifier "CCCS" (BJT)

g parameters:

$I_1 = f(V_1, I_2)$

$V_2 = f(V_1, I_2)$

$I_1 = g_i V_1 + g_r I_2$

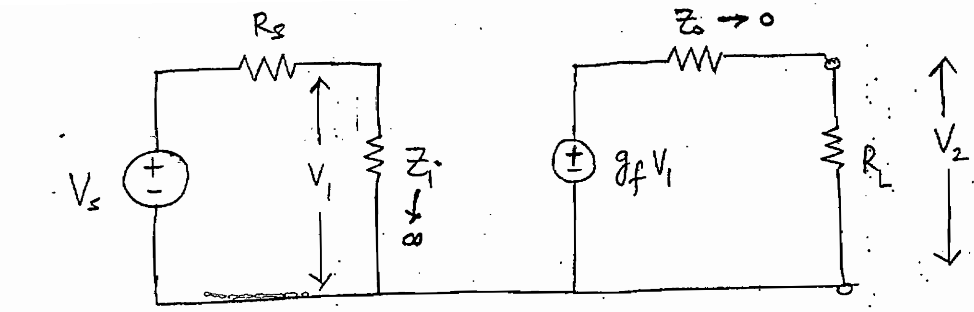
$V_2 = g_f V_1 + g_o I_2$

$$\begin{bmatrix} I_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} g_i & g_r \\ g_f & g_o \end{bmatrix} \begin{bmatrix} V_1 \\ I_2 \end{bmatrix}$$

An ideal amplifier is a unilateral device

$g_i = 0$; $g_r = 0$; $g_o = 0$

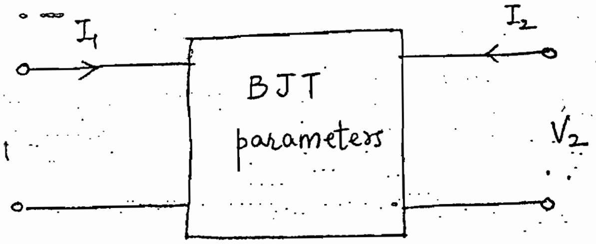
$$\begin{bmatrix} I_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ g_f & 0 \end{bmatrix} \begin{bmatrix} V_1 \\ I_2 \end{bmatrix} \qquad \boxed{V_2 = g_f V_1} \rightarrow \text{"VCVS"}$$



Voltage Amplifier "VCVS" (OP-Amp)

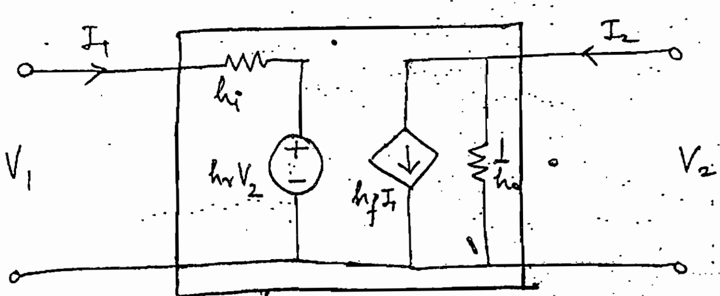
Low frequency analysis of BJT :

BJT as a h-parameter model :



$V_1 = f(I_1, V_2)$

$I_2 = f(I_1, V_2)$



Low frequency model of BJT

Typical values of h-parameters :

h-parameters		CE	CC	CB
h_i	h_{ic} ↓ last letter of CE	$1100\ \Omega$	$1100\ \Omega$	$22\ \Omega$
h_f	h_{ie} ↓ last letter of CE	50	-51	-0.98
h_r	h_{ib} ↓ last letter of CB	2.4×10^{-4}	≈ 1	2.9×10^{-4}
h_o		24×10^{-6} mho	24×10^{-6} mho	0.49×10^{-6} mho

Conversion formulaes:

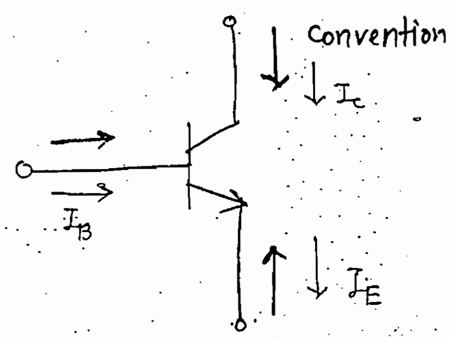
CE to CB	CB to CE	CE to CC
$h_{ib} = \dfrac{h_{ie}}{1+h_{fe}}$	$h_{ie} = \dfrac{h_{ib}}{1+h_{fb}}$	$h_{ic} = h_{ie}$
$h_{fb} = \dfrac{-h_{fe}}{1+h_{fe}}$	$h_{fe} = \dfrac{-h_{fb}}{1+h_{fb}}$	$h_{fc} = -(1+h_{fe})$
$h_{rb} = \dfrac{h_{ie}\cdot h_{oe}}{1+h_{fe}} - h_{re}$	$h_{re} = \dfrac{h_{ib}\cdot h_{ob}}{1+h_{fb}} - h_{rb}$	$h_{rc} \equiv 1$
$h_{ob} = \dfrac{h_{oe}}{1+h_{fe}}$	$h_{oe} = \dfrac{h_{ob}}{1+h_{fb}}$	$h_{oc} = h_{oe}$

Concept of −ve gain:

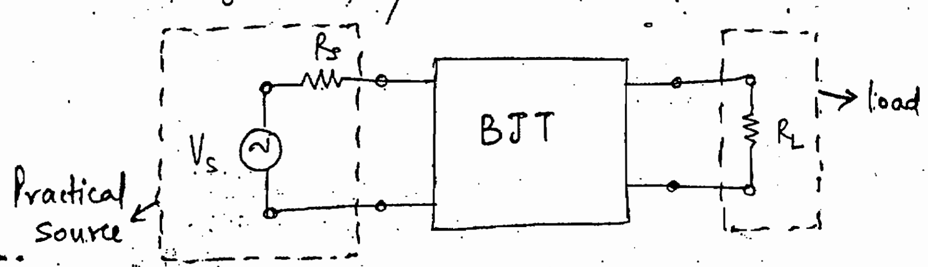
gain (I_B, I_C) = +ve

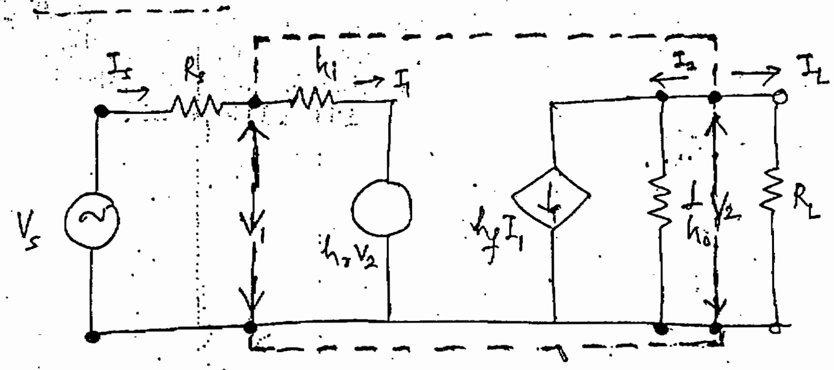
gain (I_b, I_E) = −ve

gain (I_C, I_E) = −ve



Amplifier analysis at low frequencies : (Exact model):





Characteristics of Amplifier (BJT)

(1) I/p impedence, $Z_i = \frac{V_1}{I_1}$

(2) O/p impedence, $Z_o = \frac{V_2}{I_2}$

(3). Volt. gain, $A_V = \frac{V_2}{V_1}$

(4). Current gain, $A_I = \frac{I_L}{I_1}$

ckt parameters (s/ckt)

(1). Effective i/p impedence, Z_i'

(2). Effective o/p impedence, Z_o'

(3). Voltage amplification, $A_{Vs} = \frac{V_2}{V_s}$

(4). Current amplification, $A_{Is} = \frac{I_L}{I_s}$

BJT characteristics Analysis:

(1) Current gain, A_I :

$$V_1 = h_i I_1 + h_r V_2$$
$$I_2 = h_f I_1 + h_o V_2$$

$$I_2 = h_f I_1 + h_o V_2$$
$$V_2 = -I_2 R_L$$
$$I_2 = h_f I_1 - h_o R_L I_2$$
$$I_2 (1 + h_o R_L) = h_f I_1$$
$$\frac{I_2}{I_1} = \frac{h_f}{1 + h_o R_L}$$
$$A_I = \frac{I_L}{I_1} = \frac{-I_2}{I_1} = \frac{-h_f}{1 + h_o R_L}$$

$$\therefore \boxed{A_I = \frac{-h_f}{1 + h_o R_L}}$$

(2) I/p impedence, Z_i :

$$V_i = h_i I_1 + h_r V_2$$
$$\frac{V_1}{I_1} = h_i + h_r \left(\frac{-I_2 R_L}{I_1}\right)$$

AI

$$Z_i = h_i + h_r A_I \cdot R_L$$
$$\therefore \boxed{Z_i = h_i + h_r R_L A_I}$$

(3) Voltage gain, A_V :

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

$$= \frac{-I_2 R_L}{V_1}$$

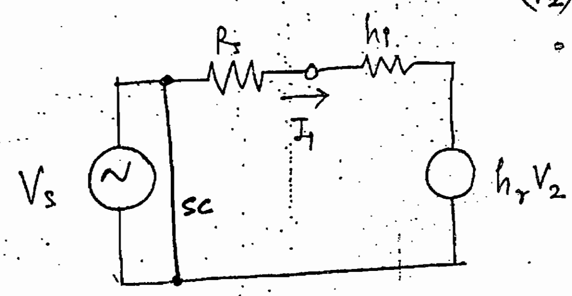
$$= \frac{-I_2/I_1 \cdot R_L}{V_1/I_1}$$

$$\boxed{A_V = \frac{A_I \cdot R_L}{Z_i}}$$

(4). o/p impedence, Z_o :

$$Z_o = V_2/I_2$$

$$Y_o = I_2/V_2 = h_f\left(\frac{I_1}{V_2}\right) + h_o$$



$$V_s = (h_i + R_s) I_1 + h_r V_2$$
$$\downarrow$$
$$0$$

$$\frac{I_1}{V_2} = \frac{-h_r}{h_i + R_s}$$

$$Y_o = h_f\left(\frac{-h_r}{h_i + R_s}\right) + h_o$$

$$\boxed{Z_o = \frac{1}{h_o - \frac{h_f h_r}{h_i + R_s}}}$$

Conclusions:

Exact model are valid for all types of configurations i.e. CE, CB, CC

Approximate model for CE

Conditions :

(1). $h_{oe} R_L \leq 0.1$ $\boxed{R_L < 5K\Omega}$

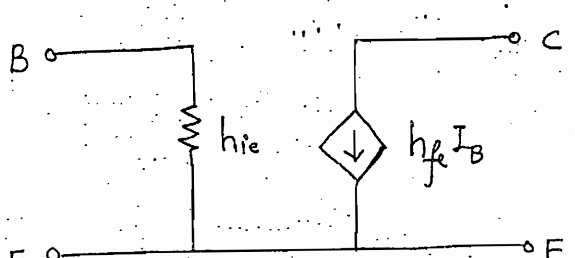
$$\text{Eg. } A_I = \frac{-h_{fe}}{1+h_{oe}R_L} = \frac{-h_{fe}}{1+0.1} = -0.91\, h_{fe}$$

* In electronics, less 10% error can be neglected.

* If $R_L > 5K$, error will come.

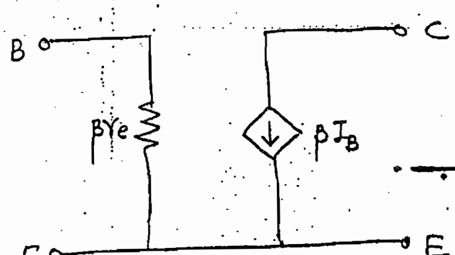
(2). h_{oe} & h_{re} should be neglected.

(3). It is valid only for CE mode



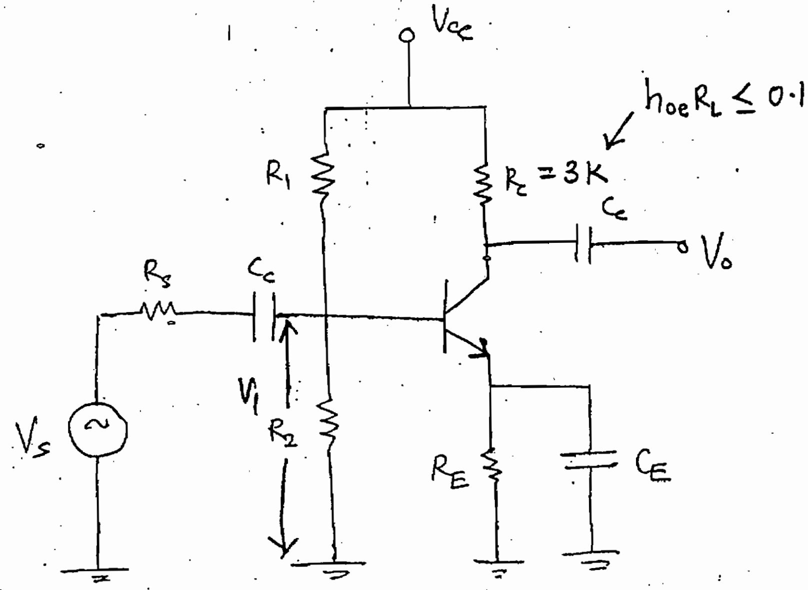
"h parameter" model

$\Rightarrow$

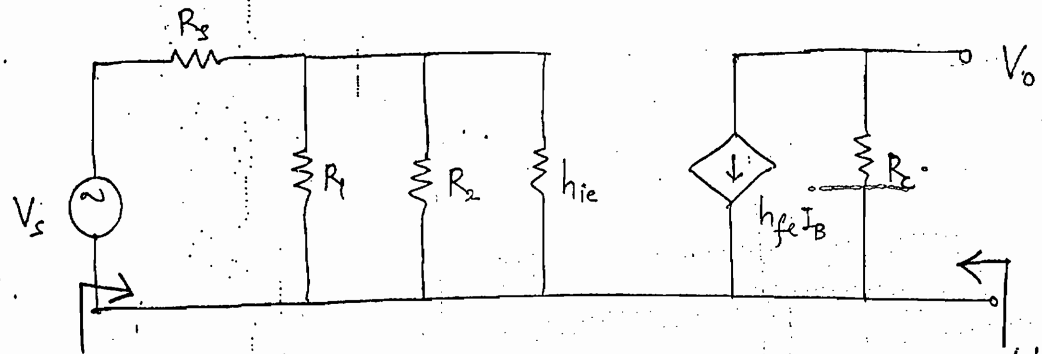


"r_e" model

CE amplifier :



AC model; DC → grounded, capacitor → SC



$$Z_i = R_s + (R_1 // R_2 // h_{ie})$$

$$Z_o = \left(\frac{1}{h_{oe}}\right) // R_c$$

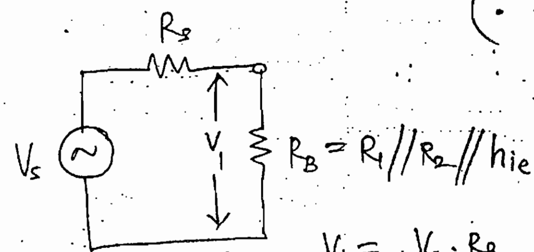
$$\approx R_c$$

$(\because 80K // 3K \approx 3K)$

$$A_{Vs} = \frac{V_o}{V_1} \cdot \frac{V_1}{V_s}$$

$$= A_V \cdot \frac{V_1}{V_s}$$

$$A_{Vs} = \frac{A_V \, R_B}{(R_s + R_B)}$$

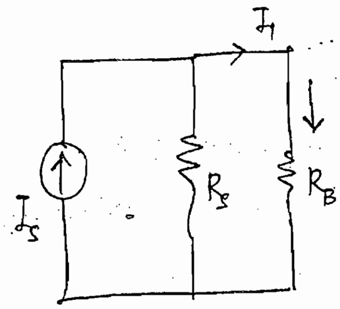


$$V_1 = \frac{V_s \cdot R_B}{(R_s + R_B)}$$

$$A_{Is} = \frac{I_L}{I_1} \cdot \frac{I_1}{I_s}$$

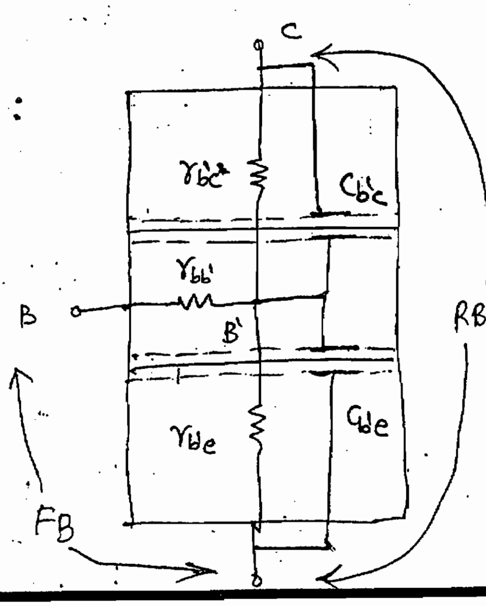
$$= A_I \cdot \frac{I_1}{I_s}$$

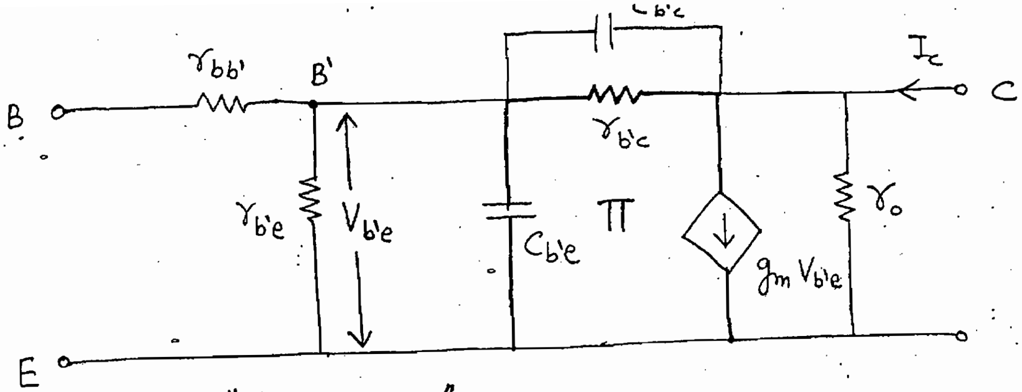
$$A_{Is} = \frac{A_I \cdot R_s}{(R_s + R_B)}$$



$$I_1 = \frac{I_s \cdot R_s}{(R_s + R_B)}$$

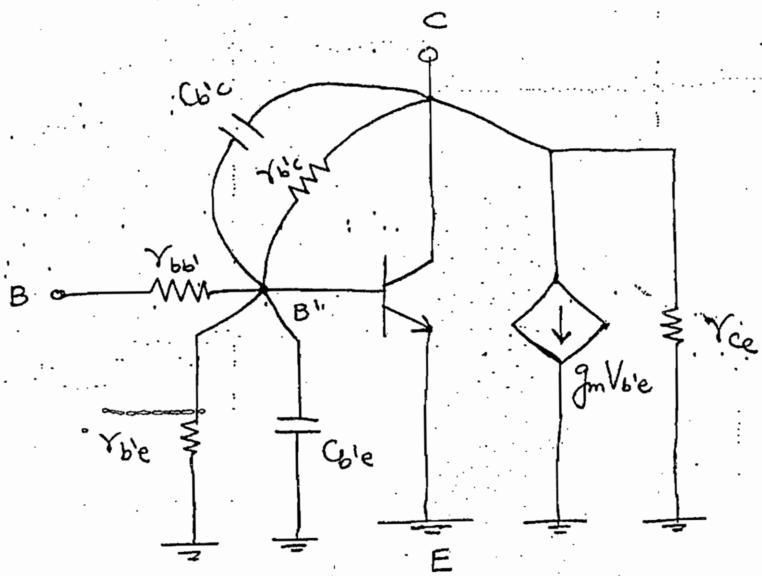
High frequency analysis of BJT :





"Hybrid – π" model

* This model is valid for both high and low frequency.
- But for low frequency, capacitor → open
and for high frequency, ckt is taken as it is.



Hybrid π parameters:	Typical values
$r_{bb'}$ → Base spreading resistance (less)	100 (≈ s.c)
$r_{b'e}$ → i/p resistance (r_π)	1 KΩ
$r_{b'c}$ → feedback resistance (high)	4 MΩ (≈ o.c)
r_{ce} → o/p resistance (r_o)	80 KΩ
C_D → $C_{b'e}$ → diffusion capacitance (stronger)	100 pF
C_T → $C_{b'c}$ → transition capacitance (weaker)	3 pF
g_m → trans conductance $\left[(I_c)_Q\big\|_{max} = 50\,mA\right]$	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). g_m :

$$g_m = \frac{(I_c)_Q}{V_T} = \frac{(I_c)_Q}{T/11{,}600}$$

As $T\uparrow$, $g_m\downarrow$

As $(I_c)_Q\uparrow$, $g_m\uparrow$

$A_V = -g_m R_c$

$\downarrow$ if load is const.

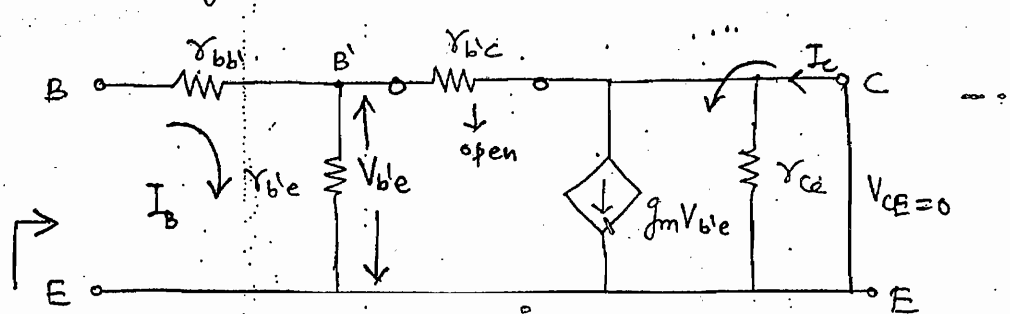
$A_V\uparrow$ if $g_m\uparrow$

(2). $r_{b'e}$:

$$\underset{\substack{\downarrow\\ \pi \text{ model}}}{r_{b'e}} = \frac{h_{fe}}{g_m}$$

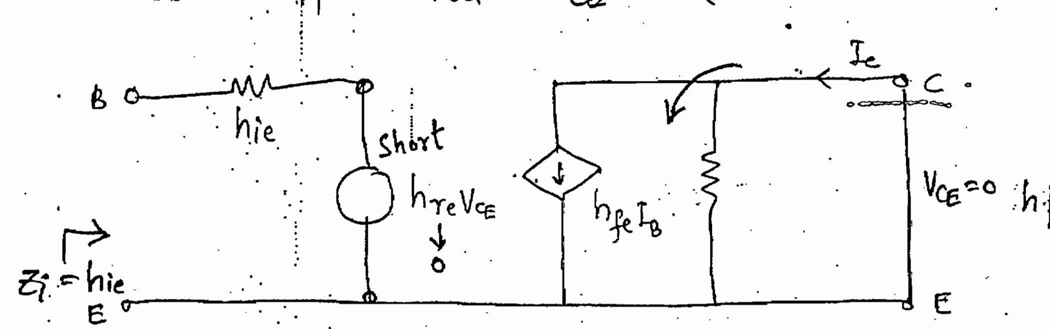
$h_{fe} \rightarrow$ h parameter (at moderate frequency)

Converting high freq. model into low frequency.



$Z_i = r_{bb'} + r_{b'e}$

$r_{bb'} \rightarrow$ i/p   Put $V_{CE} = 0$ (SC)



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

$$I_c = h_{fe} I_B$$

$$I_c = g_m \cdot I_B r_{b'e}$$

$$\frac{I_c}{I_B} = h_{fe} \quad \text{— (2)}$$

$$r_{b'e} = \left(\frac{I_c}{I_B}\right)\frac{1}{g_m} \quad \text{— (1)}$$

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

And,

$$g_m = \frac{(I_c)_Q}{V_T} \qquad \text{and} \quad h_{fe} = \beta$$

$$\frac{1}{g_m} = \frac{V_T}{(I_c)_Q} \approx \frac{V_T}{(I_E)_Q} = r_e$$

$$\therefore \boxed{r_{b'e} = \frac{h_{fe}}{g_m} = \beta r_e}$$

(3). $r_{bb'}$: Put $V_{CE} = 0$ (again) ($\because r_{bb'} \rightarrow$ i/p parameter)

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

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

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

$$\underset{100\,\Omega}{r_{bb'}} = \underset{1.1\,K\Omega}{h_{ie}} - \underset{1\,K\Omega}{r_{b'e}}$$

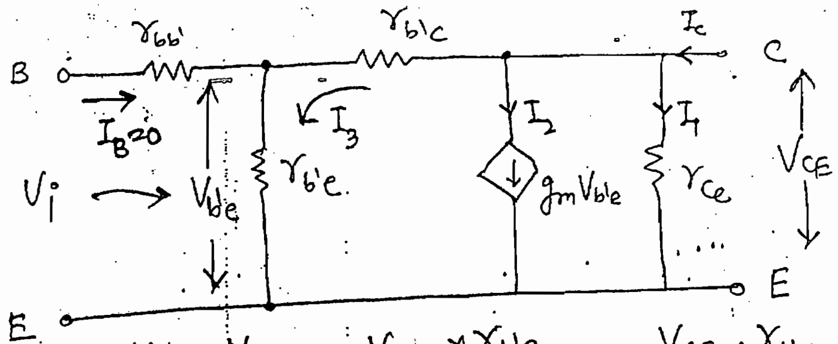
(4). $r_{b'c}$:

$r_{b'c}$ → i/p, o/p

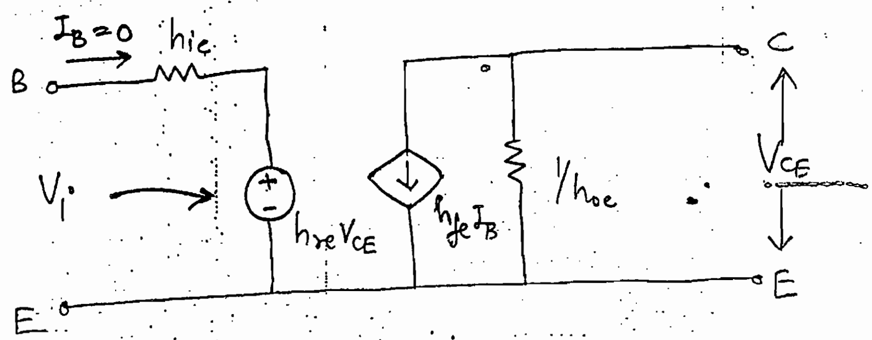
i/p: $\dfrac{r_{b'c}}{(1-A_V)}$ → negligible

o/p: $\dfrac{r_{b'c}\, A_V}{(A_V - 1)} \approx r_{b'c}$ → (o/p parameter)

Put $I_B = 0$:



$$V_i = V_{b'e} = \frac{V_{CE} \times r_{b'e}}{r_{b'e} + r_{b'c}} = V_{CE} \cdot \frac{r_{b'e}}{r_{b'c}} \qquad (\because r_{b'c} >> r_{b'e})$$



$$V_i = h_{re} V_{CE}$$

$$\therefore \quad V_{CE} \cdot \frac{r_{b'e}}{r_{b'c}} = h_{re} \cdot V_{CE}$$

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

(5). $\underline{\underline{r_{ce}}}$:

$$\therefore \frac{I_c}{V_{CE}} = h_{oe}$$

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

$$I_c = \frac{V_{CE}}{r_{ce}} + g_m \left(\frac{V_{CE}\, r_{b'e}}{r_{b'e} + r_{b'c}} \right) + \frac{V_{CE}}{r_{b'e} + r_{b'c}}$$

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

$$\frac{I_c}{V_{CE}} = \frac{1}{r_{ce}} + \frac{h_{fe} + 1}{r_{b'e} + r_{b'c}} \qquad (\because h_{fe} = g_m r_{b'e})$$

$(\because r_{b'e} \rightarrow \text{neglected})$

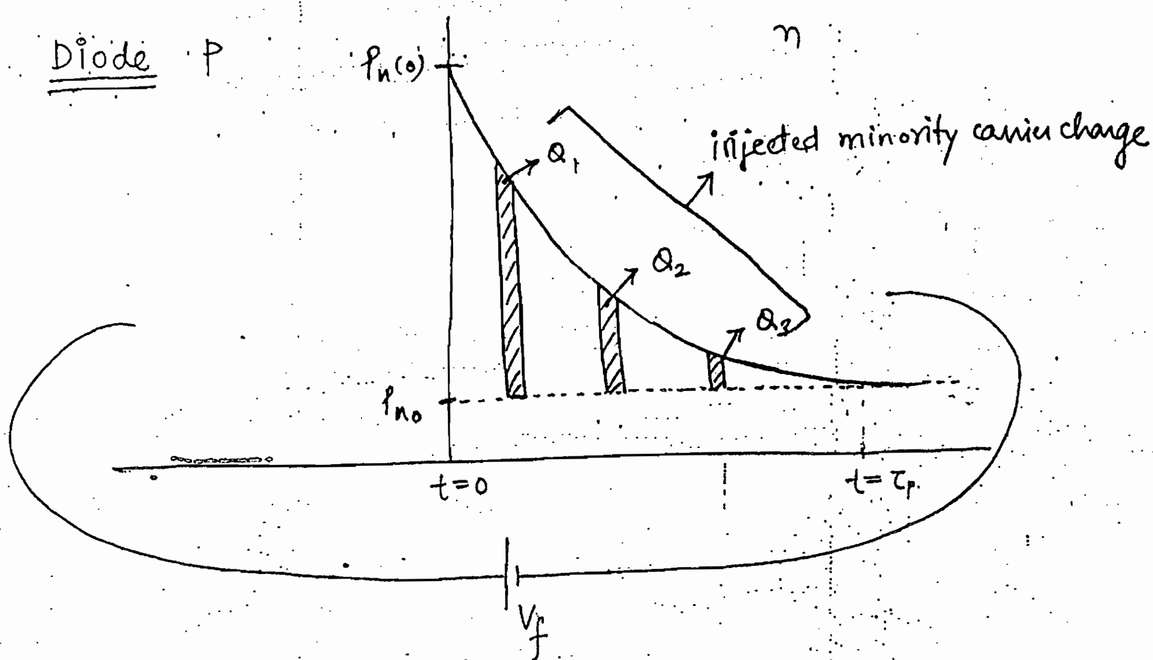
$$h_{oe} = \frac{1}{r_{ce}} + \frac{h_{fe}}{r_{b'c}} = g_{ce} + h_{fe}\, g_{b'c} \qquad \left(r_{ce} = \frac{1}{g_{ce}} \;\&\; r_{b'c} = g_{b'c} \right)$$

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

$$\therefore \boxed{r_{ce} = \frac{1}{g_{ce}} = \frac{1}{h_{oe} - h_{fe}\, g_{b'c}}}$$

(6). Diffusion capacitance, C_D : ($C_{b'e}$)

Diode

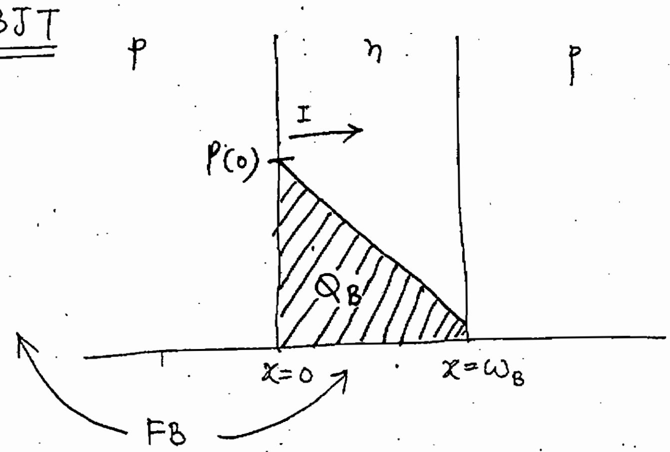


$$Q = I_p \tau_p$$

$$C_D = \frac{dQ}{dV} = \tau_p \cdot \frac{dI_p}{dV}$$

$$\therefore \boxed{C_D \doteq \frac{\tau_p I_{dc}}{V_T}}$$

BJT



$$Q_B = \frac{1}{2} P(0) \cdot (A q \omega_B) \quad \frac{C\,m^3}{m^3}$$

$$I = -A q D_B \frac{dp}{dx}$$

$$= -A q D_B \left\{ \frac{P(0) - 0}{0 - \omega_B} \right\}$$

$$= \frac{A q D_B\, P(0)}{\omega_B}$$

$$Q_B = \frac{1}{2} P(0) \cdot Aq\omega_B$$

$$\therefore Q_B = \frac{I \omega_B^2}{2 D_B}$$

And, $C_D = \frac{dQ_B}{dV} = \frac{d}{dV}\frac{\omega_B^2 I}{2 D_B} = \frac{\omega_B^2}{2 D_B} \cdot \frac{dI}{dV} = \frac{\omega_B^2}{2 D_B} \cdot g_m$

$$\boxed{C_D = \frac{\omega_B^2}{2 D_B} \cdot \frac{1}{r_e}}$$

$$C_D = \underbrace{\frac{\omega_B^2}{2 D_B}} \cdot \frac{(I_c)_Q}{V_T}$$

$$\boxed{T_B = \omega_B^2 / 2 D_B} \rightarrow \text{Base transit time}$$

As $\frac{\omega_B^2}{2 D_B} = \frac{\omega_B^2}{2 \mu V_T} \rightarrow \frac{m^2}{m^2/V\cdot sec \cdot V} \rightarrow sec$ i.e. time

Q. As V_{CE} ↑ses what happens to C_D in Hybrid π model?

Sol. As $V_{CE} \uparrow$, $\omega_B \downarrow$, $C_D \Downarrow$. (Early Effect)

$$\omega_B = \omega_B' + \omega_B''$$

ω_B' ↓ this width we are talking about.

(7). $C_{b'c}$: <u>Transition capacitance (C_T)</u> :

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

$$C_T \propto \frac{1}{\sqrt{V_R}} \rightarrow \text{Alloy type}$$

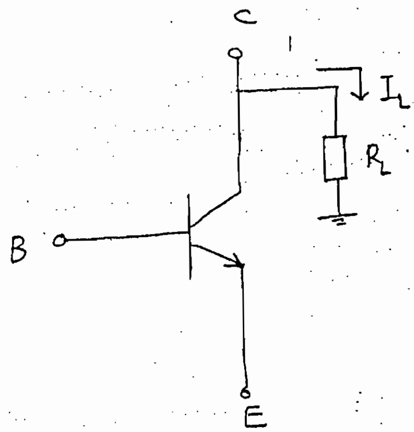
$$C_T \propto \frac{1}{\sqrt[3]{V_R}} \rightarrow \text{Grown jxn type}$$

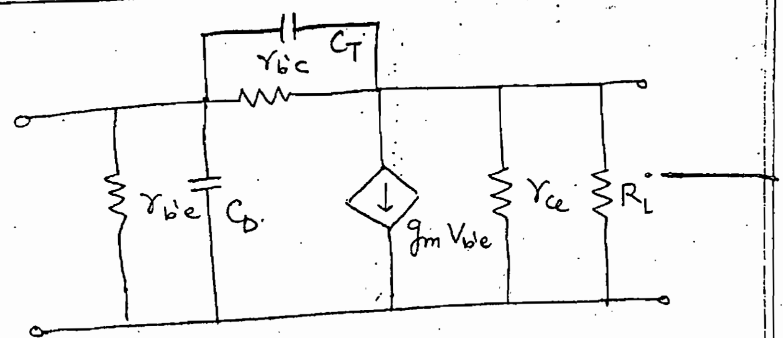
$\boxed{C_{b'c} \propto \frac{1}{(V_{CE})^n}}$ ", $n = 1/2 \rightarrow$ alloy type

$n = 1/3 \rightarrow$ grown jxn type

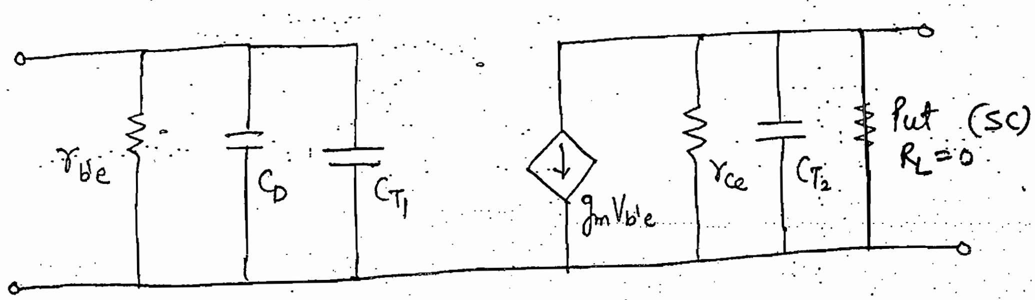
* $C_{b'c}$ is independent of temp. & $(I_E)_Q$, it only depends upon V_{CE} supply. (the only parameter).

Frequency response of an amplifier:



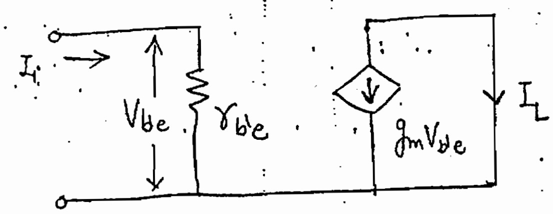


CE SC current gain (max^m in sc condition):



$\Downarrow$

At $f = 0$: (Cap → open)



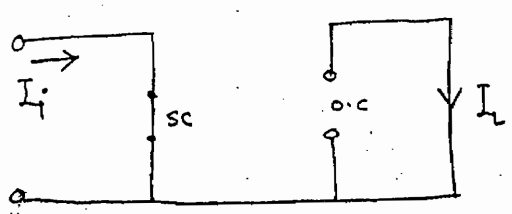
$$A_I = \frac{I_L}{I_i} = \frac{-g_m V_{b'e}}{V_{b'e}/r_{b'e}}$$

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

$$= -h_{fe}$$

$$\boxed{A_I = -\beta}$$

At $f = \infty$: (cap → short)



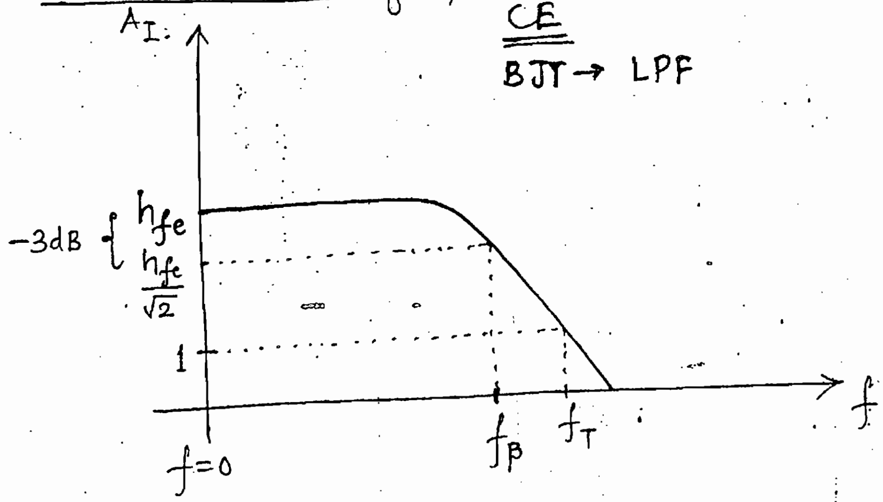
$$A_I = \frac{I_L}{I_i}$$

$$\boxed{A_I = 0}$$

Current gain Vs frequency:

CE

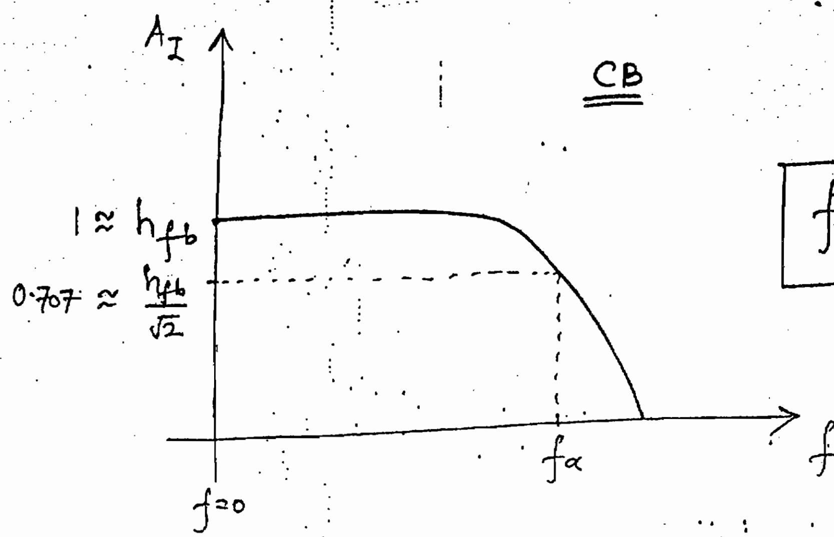
BJT → LPF



* dB values are used to compress the values of scale markings. (f → GHz, MHz.)
  Scaling is required for simplicity.

* $20\log(1) = 0$ ie. above f_T, the gain is zero.
  Hence, f_T is given by manufacturer, upto which a s/g can be applied to the given BJT.

CB



$$\boxed{f_\alpha > f_T > f_\beta}$$

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

$$f_\alpha \approx 1.01 f_T \quad (\text{for } h_{fe} = 100)$$

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

f_β (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):

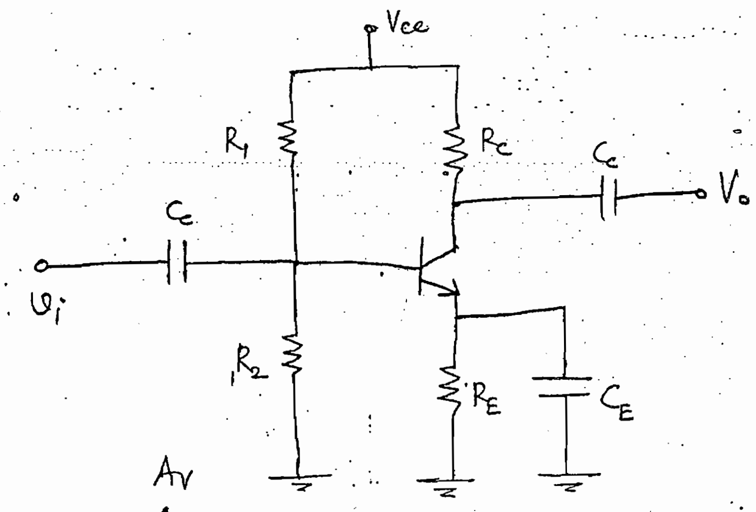
It is the frequency at which CE sc A_I falls to unity that is called as f_T.

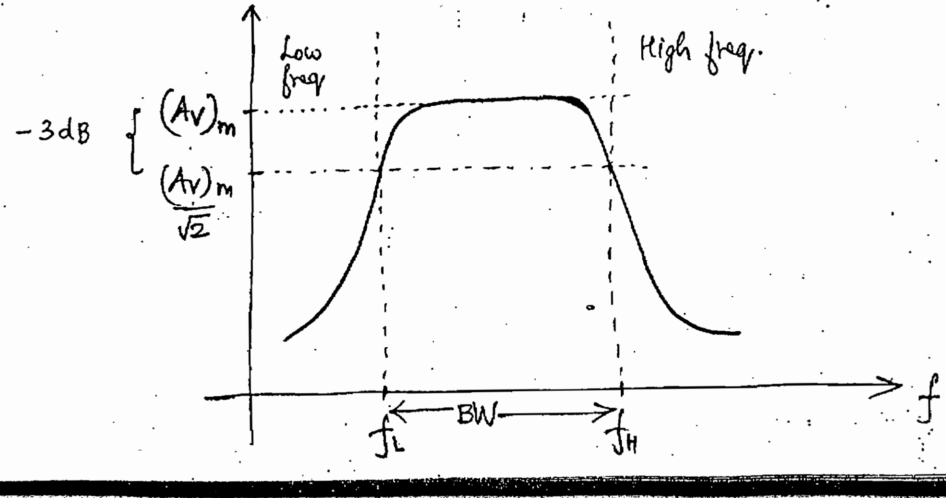
f_α (α cut-off frequency):

It is the frequency at which CB sc A_I reduces by 3 dB of its value.

Voltage gain Vs frequency:

CE amplifier:





Capacitance

- Internal capacitors
  - C_D (Diffusion)
  - C_T (Transition)
- External capacitors
  - C_C (Coupling)
  - C_E (Bypass)

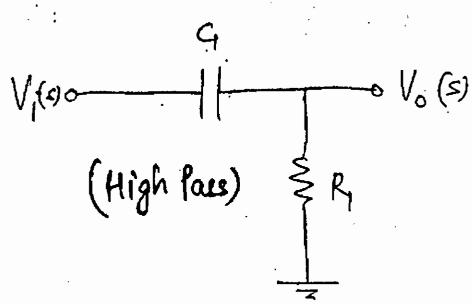
Low frequency Range:

(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:



$$\frac{V_o(s)}{V_i(s)} = \frac{R_1}{R_1 - jX_{C_1}}$$

$$= \frac{s}{\left(s + \frac{1}{R_1 C_1}\right)}$$

zero → 1
pole → 1

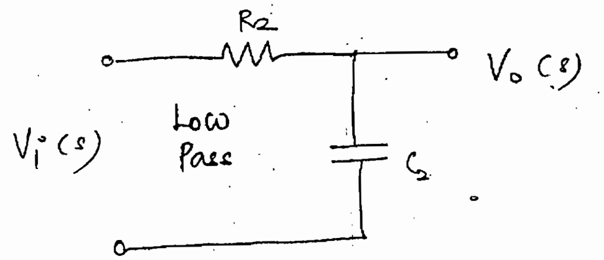
High frequency Range:

(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.

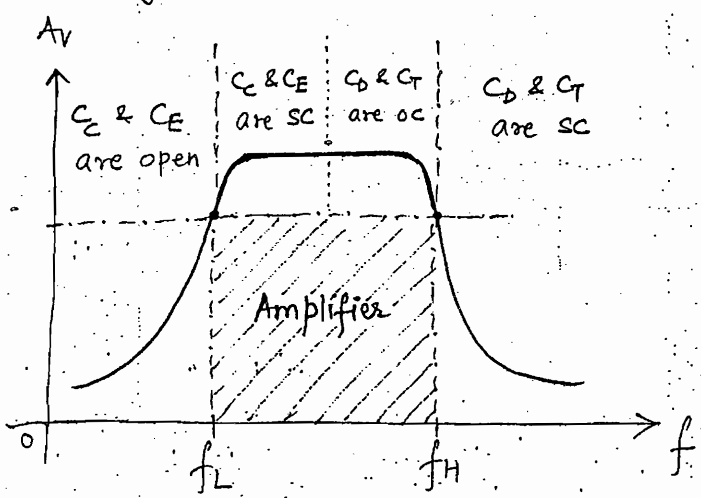
(4). The equivalent model of LP is given as:



$$\frac{V_o(s)}{V_i(s)} = \frac{-jX_{C_2}}{R_2 - jX_{C_2}}$$

$$= \frac{K}{\left(s + \frac{1}{R_2 C_2}\right)}$$ Pole → 1

Mid band range :



(1). At low freq. range, mid band range frequencies are considered as high frequencies. Therefore, external capacitors C_c 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.

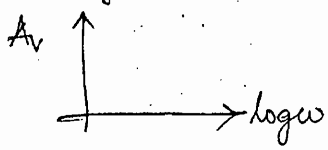
(3). That means in the mid band range all the internal and external capacitors are neglected. Therefore the gain is independent of frequency.

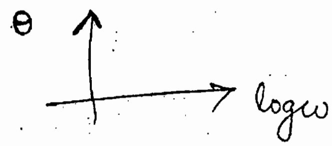
−3 dB concept:

Bode plots

→ Magnitude plot

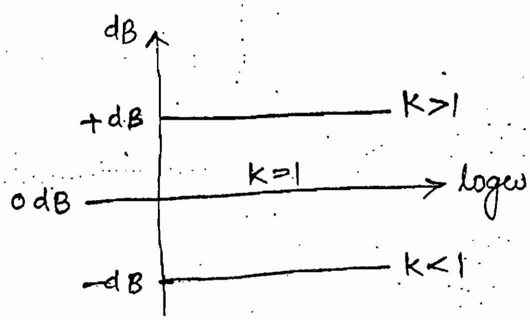
→ phase plot





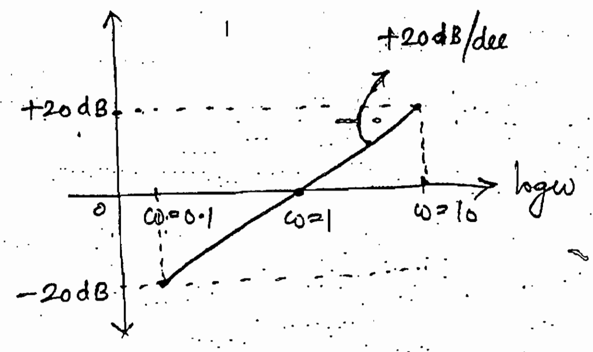
(1). $G(s) = K$ (constant)

$|G(j\omega)|_{dB} = 20 \log K$



(2). $G(s) = s$ (zero)

$|G(j\omega)|_{dB} = 20\log\omega$



(3). $G(s) = \frac{1}{(1+sT)}$ (pole)

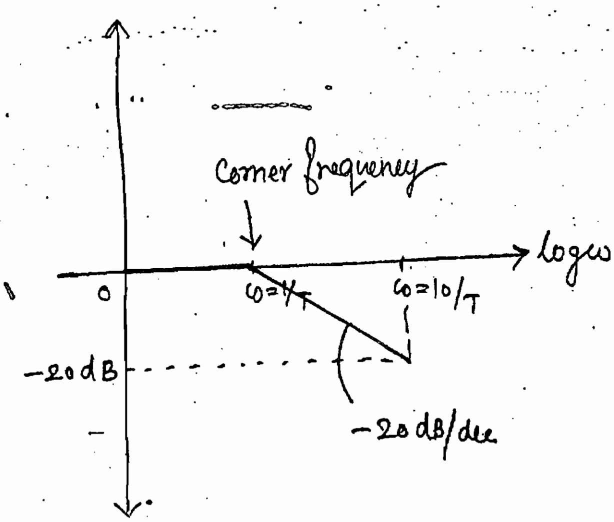
$|G(j\omega)|_{dB} = -20\log\sqrt{1+\omega^2T^2}$

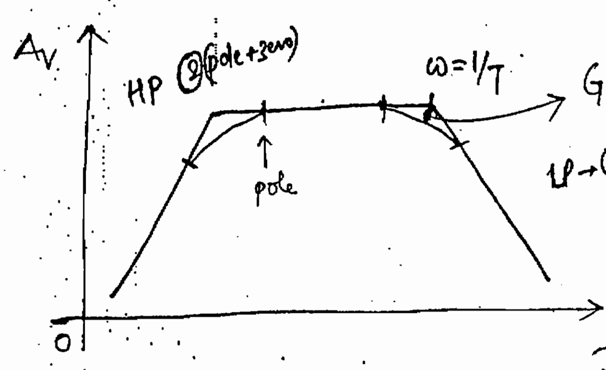
$\omega T << 1$ (low freq.)

$|G(j\omega)|_{dB} = 0$ dB

$\omega T >> 1$ (high freq.)

$|G(j\omega)|_{dB} = -20\log\omega T$





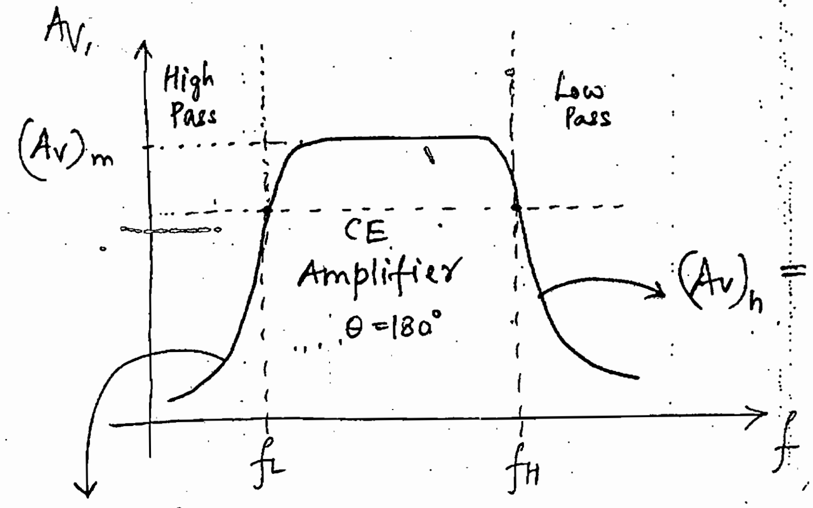
$G(s)\big|_{\omega=1/T} = -20\log\sqrt{1+\omega^2T^2}$

LP → ① pole $= -20\log\sqrt{2}$

$= -3\,dB$

(error)

Phase Analysis :



$$(A_v)_h = \frac{(A_v)_m}{1 + j(f_H/f)}$$

where

$$f_H = \frac{1}{2\pi R C_{internal}}$$

$$(A_v)_l = \frac{(A_v)_m}{1 - j(f_L/f)}$$

where $f_L = \dfrac{1}{2\pi R C_{external}}$

$\theta_l = +\tan^{-1}(f_L/f)$

when $f \to f_L$

$\theta_l = 45°$

$\theta_{lCE} = 180° + \theta_l$

$= 225°$

$\theta_{lCB} = 45°$

$\theta_h = -\tan^{-1}(f_H/f)$

when $f \to f_H$

$\theta_h = -45°$

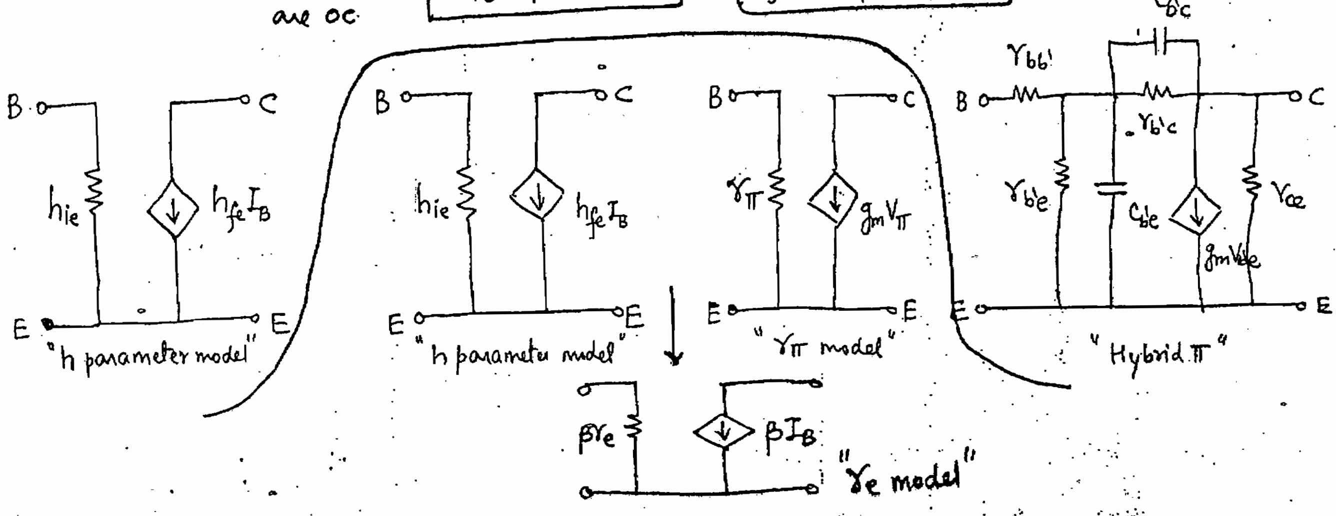
$\theta_{hCE} = 180° + \theta_h$

$= 135°$

$\theta_{hCB} = -45°$

C_D & C_T are oc

$\boxed{h_{ie} = \beta r_e = r_\pi}$ $\boxed{h_{fe} I_b = \beta I_B = g_m V_\pi}$



$A_v = -g_m R_c$

$$\boxed{A_v = -\frac{R_c}{r_e}} \rightarrow \text{" } r_e \text{ "}$$

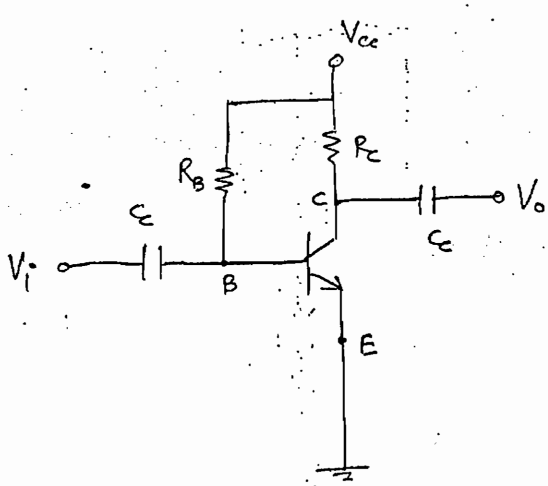
$$\boxed{A_v = -\frac{h_{fe} R_c}{h_{ie}}} \rightarrow \text{" h parameter "}$$

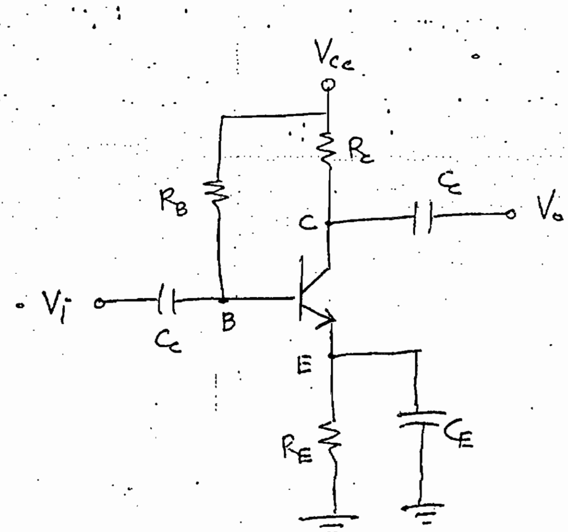
$$\boxed{A_v = \frac{-\beta R_c}{r_\pi}} \rightarrow \text{" } r_\pi \text{ "}$$

Practical Amplifier designs :

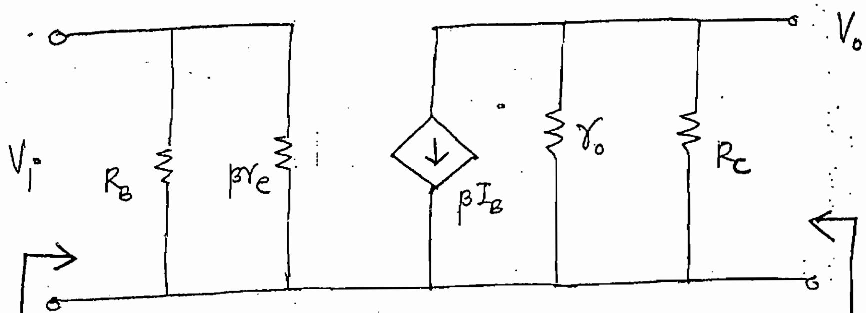
(1). CE bypass amplifier
(2). CE unbypass amplifier
(3). CC amplifier
(4). CB amplifier

(1). CE bypass amplifier :





If R_E is not equal to zero, then there must be a bypass capacitor C_E in // to it.



$\boxed{R_B \geq 10\beta r_e}$ $\boxed{r_o \geq 10 R_c}$

$Z_i = R_B // \beta r_e \approx \beta r_e$ (low) $Z_o = r_o // R_c \approx R_c$

$$V_o = -\beta I_B R_c$$

$$I_B = \frac{V_i}{\beta r_e}$$

$$A_v = \frac{V_o}{V_i} = \frac{-R_c}{r_e} = -g_m R_c \qquad \left(\propto \frac{1}{r_e}\right)$$

Heavy values of gain are unstable

i.e. if $A_v = 1000 \longrightarrow$ varies (999.8 or 1000.5)

$\therefore$ $A_v \uparrow$, stability $\downarrow$

Conclusions:

(1). Drawbacks of CE bypass amplifier :

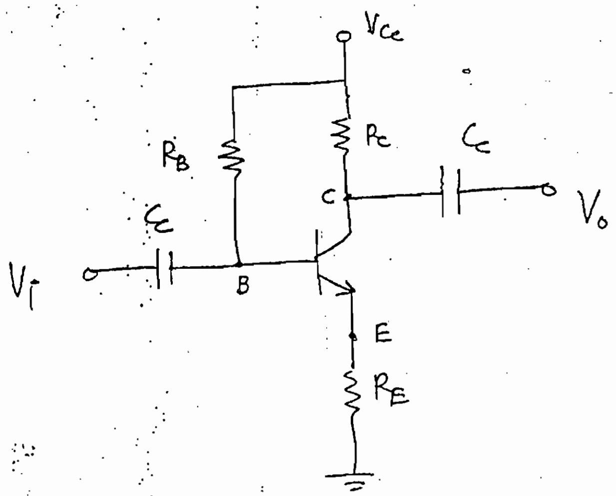
(i) The i/p impedence is low. (For an ideal amplifier i/p impedence should be ∞).

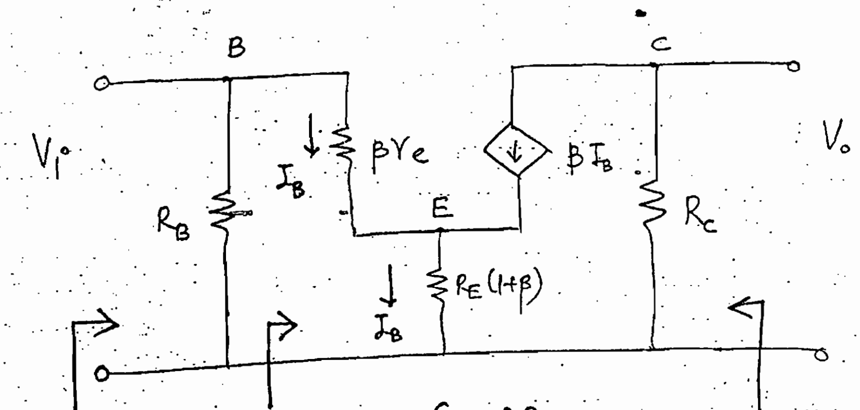
(ii). The gain is not stable becoz of dynamic resistor r_e.

* $A_v = \infty$, stability $= 0$
* $A_v = 0$, stability $= \infty$

A

(2). CE unbypass amplifier:





$Z_i' = R_B \| Z_i$ $\quad Z_i = \beta r_e + (1+\beta)R_E$ $\qquad Z_o \approx R_C$

$\approx \beta R_E$ (High)

Assume: $\beta = 100$, $r_e = 1\,\Omega$, $R_E = 10\,k\Omega$

$$Z_i = 100\,\Omega + (101)\,10\,k\Omega$$
$$= 100\,\Omega + 1\,M\Omega$$
$$\approx 1\,M\Omega.$$

$$V_o = (-\beta I_B) R_C$$

$$I_B = \frac{V_i}{\beta R_E}$$

$$A_v = \frac{V_o}{V_i} = -\frac{R_C}{R_E} \quad (\propto r_e^{\circ})$$

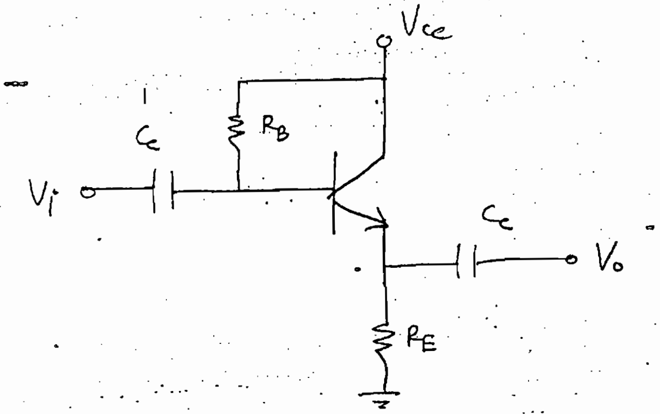
As $A_v \downarrow$, stability $\uparrow$.

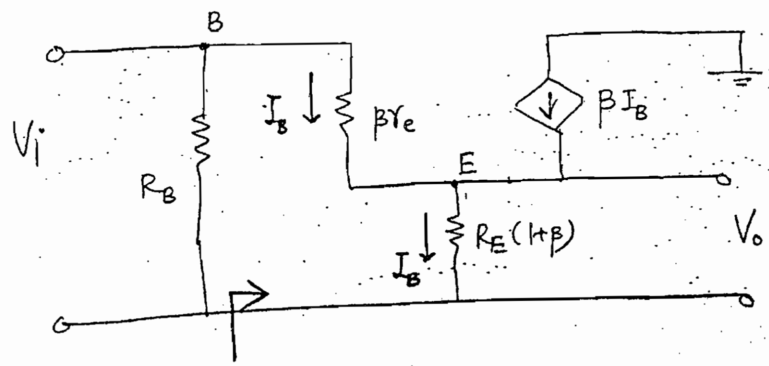
Conclusions:

(1). The i/p impedence is high.

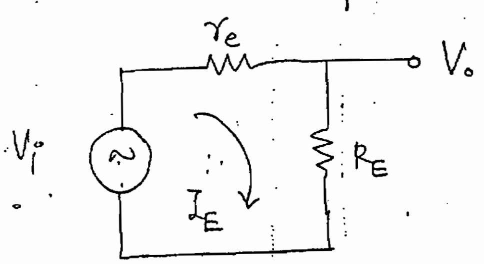
(2). Gain is stable because of absence of dynamic resistance r_e.

(3). <u>Emitter follower amplifier</u> :





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



$$V_o = \frac{V_i \times R_E}{(r_e + R_E)}$$

$$\boxed{V_o = V_i} \quad (\because R_E >> r_e)$$

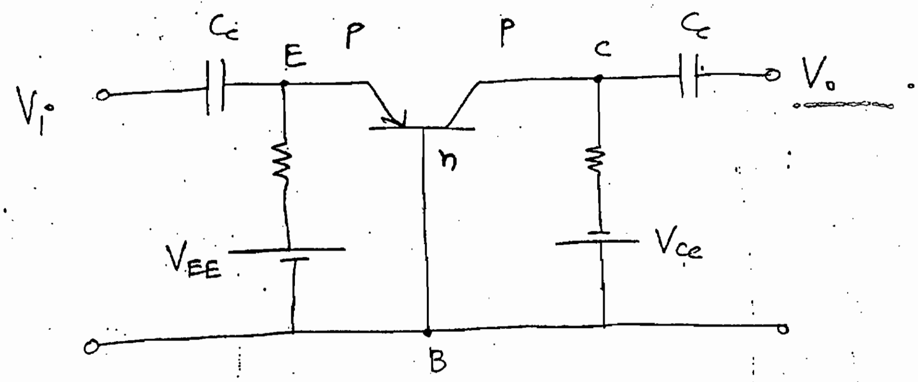
→ Voltage Buffer

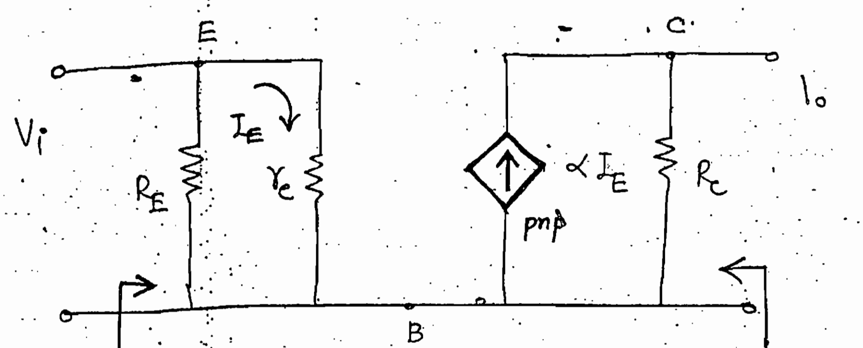
$Z_o = r_e // R_E \approx r_e$ (impedance matching).

short: $A_V = \frac{V_o}{V_i} = \left(\frac{I_o}{I_i}\right)\frac{R_L}{Z_i} = \frac{A_I \cdot R_L}{Z_i}$

Here, $A_V = 1 \Rightarrow Z_i = A_I R_L = \beta \cdot R_E$

(4). CB amplifier –





$Z_i = R_E \,//\, r_e$
$\simeq r_e$

$Z_o \approx R_c$

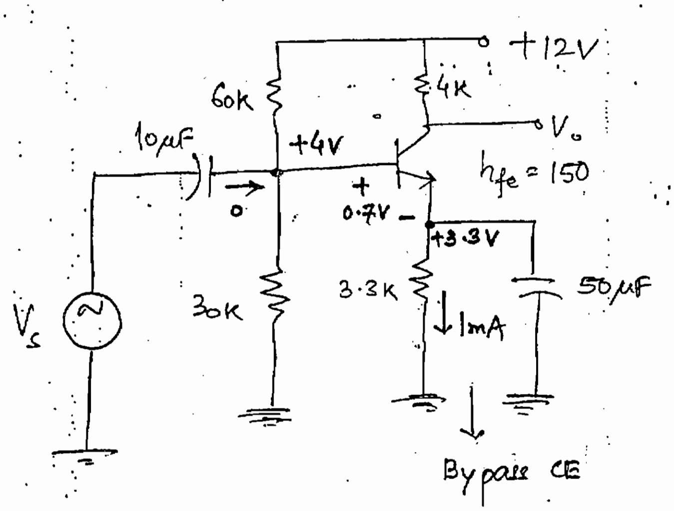
$V_o = (\alpha I_E) R_c$

$I_E = \dfrac{V_i}{r_e}$

$A_V = \dfrac{V_o}{V_i} = \dfrac{R_c}{r_e}\cdot\alpha \approx \dfrac{R_c}{r_e} \qquad (\because \alpha \approx 1)$

Problems :

CH-6 conv.
Q3.



$\left|\dfrac{V_o}{V_s}\right| = ?$

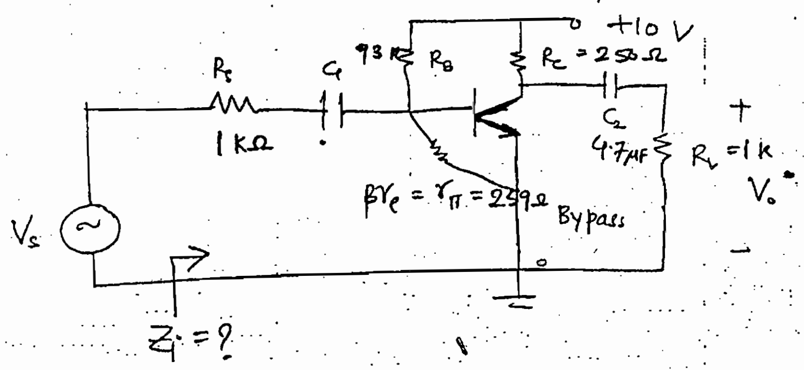
Sol. $A_V = -\frac{R_C}{r_e} = \frac{-4k}{r_e}$

$r_e = \frac{V_T}{(I_E)_Q} = \frac{26\,mV}{1\,mA} = 26\,\Omega$

$\therefore |A_V| = \left|\frac{-4\,k\Omega}{26\,\Omega}\right| = 153.8$

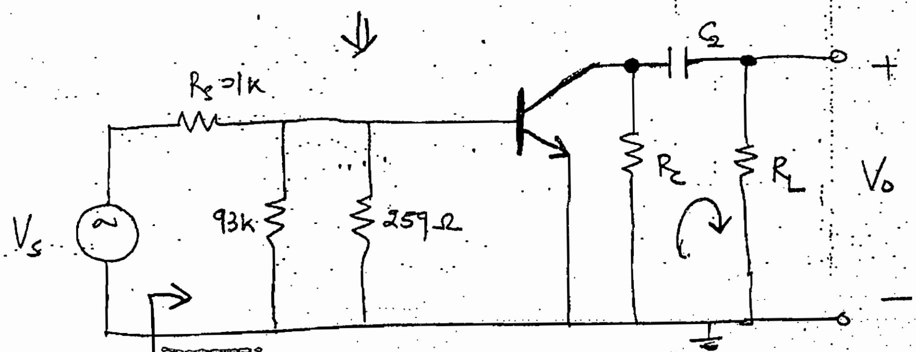
If $V_T = 25mV$, $|A_V| = \left|\frac{-4k\Omega}{25\Omega}\right| = 160$

Q4.



$f_3\,(dB) = ?$

⇓

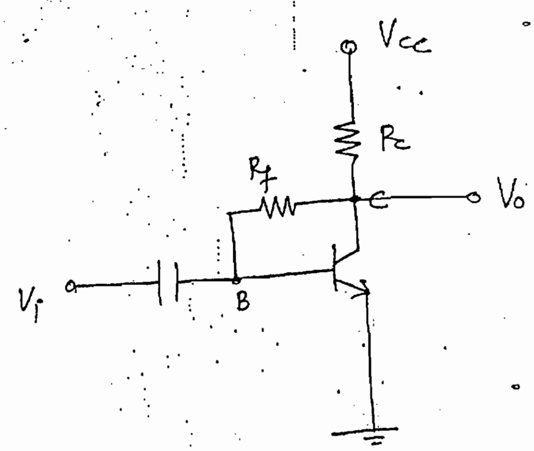


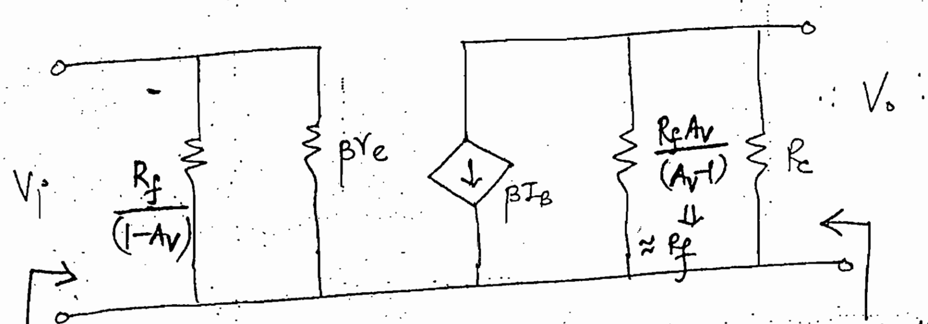
$Z_i = 1k + (93k // 259\,\Omega)$
$= 1k + 259\,\Omega$
$= 1259\,\Omega$

$f_3\,(dB) = \frac{1}{2\pi\,(R_C + R_L)\,C}$

$= \frac{1}{2\pi\,(250 + 1000) \times 4.7 \times 10^{-6}}$

$= 27.09\ Hz$



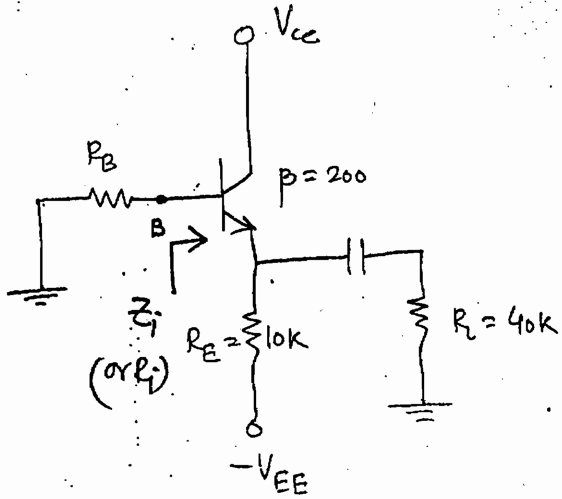


$Z_o = R_f // R_c$

$$A_V = \frac{-R_c // R_f}{r_e}$$

$$Z_i = \left(\frac{R_f}{1-A_V}\right) // \beta r_e$$

Q1.



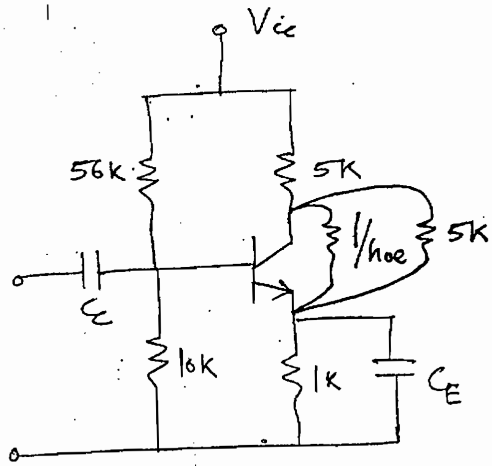
$$R_i = \beta \cdot (R_L // R_E)$$

$$= 200 \left(\frac{10K \cdot 40K}{50K}\right)$$

$$= 1600\ K$$

$$= 1.6\ M\Omega$$

Q2.



$h_{fe} = 100$

$h_{ie} = 2K\Omega$

$h_{re} = 0$

$h_{oe} = 0.05 \ m\,mho$

$= 0.05 \times 10^{-3} \ \mho$

$$Z_0 = \frac{1}{h_{oe}} \,//\, 5K = \frac{1}{0.05\times10^{-3}} \,//\, 5000$$

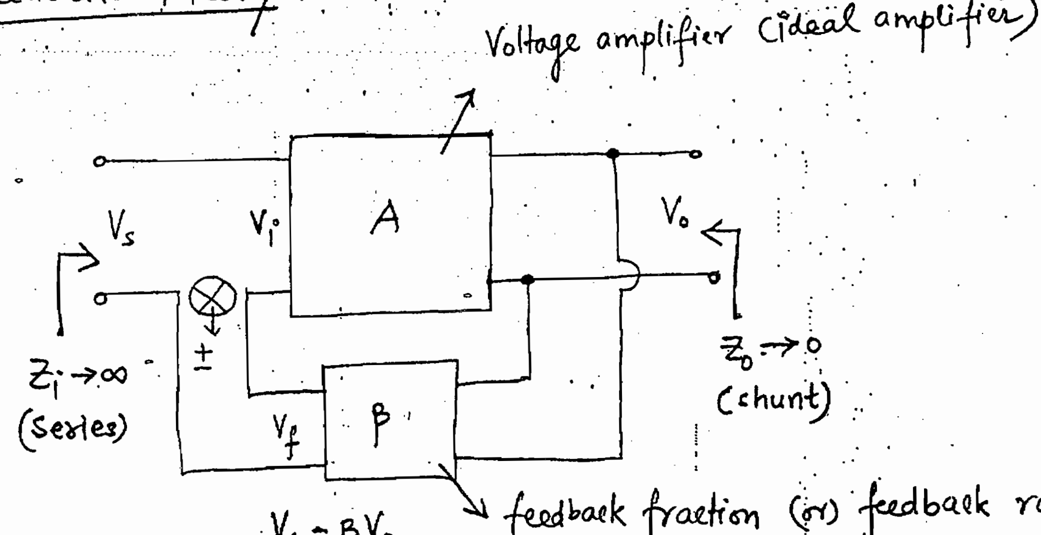
$$= 20K \,//\, 5K$$

$$= 4K\Omega$$

$$* \quad f_T = \frac{1}{2\pi RC} = \frac{g_m}{2\pi(C_D + C_T)}$$

feedback amplifiers :

feedback theory :



$V_f = \beta V_o$

$V_i = V_s \pm V_f$

$V_i = V_s + V_f \rightarrow +ve \ fb$

$V_i = V_s - V_f \rightarrow -ve \ fb$

Positive fb

$$V_o = A V_i$$
$$V_i = V_s + V_f$$
$$V_o = A(V_s + V_f)$$
$$V_o = A(V_s + \beta V_o)$$
$$V_o(1 - A\beta) = A V_s$$
$$\boxed{\frac{V_o}{V_s} = \frac{A}{1 - \beta A}}$$

Negative fb

$$V_o = A V_i$$
$$V_i = V_s - V_f$$
$$V_o = A(V_s - V_f)$$
$$V_o = A(V_s - \beta V_o)$$
$$V_o(1 + \beta A) = A V_s$$
$$\boxed{\frac{V_o}{V_s} = \frac{A}{1 + \beta A}}$$

Conclusions:

(1). $A_{pf} > A > A_{nf}$

(2). $A_{nf} = \dfrac{A}{1 + \beta A}$

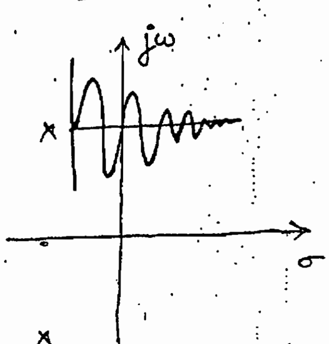
$\beta A >>> 1$

$A_{nf} \approx \dfrac{1}{\beta}$ (Stability)

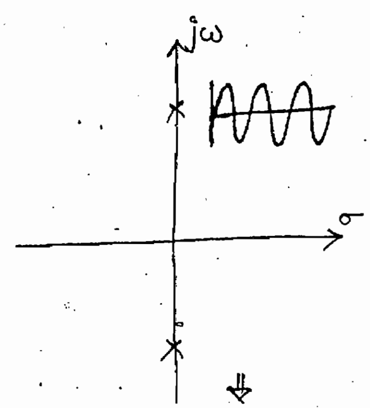
* −ve fb theory is applied for stable s/s like amplifiers.

(3). $A_{pf} = \dfrac{A}{1 - \beta A}$

$\beta A < 1$



$\beta A = 1$

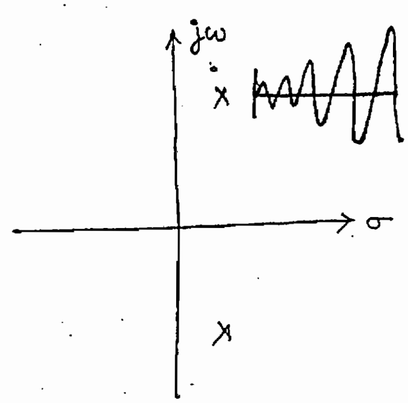


⇓

Oscillators

$A_{pf} = \dfrac{A}{0}$ (No i/p)

$\beta A > 1$



* +ve fb theory is applied for unstable s/s like oscillators.

Advantages of -ve fb amplifiers :

(1). Stability of AC gain :

Suppose there is a small change in the internal resistance 're' of an amplifier, then the fractional change of gain with feedback is

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

Differentiate w.r.to A

$$\frac{\partial A_f}{\partial A} = \frac{(1+\beta A)(1) - A.\beta}{(1+\beta A)^2}$$

$$\frac{dA_f}{\cancel{A_f}} = \frac{1}{(1+\beta A)^2} \cdot \frac{dA}{A_f}$$

$$\frac{dA_f}{A_f} = \frac{1}{(1+\beta A)} \cdot \frac{dA}{A}$$

fractional change of gain with fb $= \underbrace{\boxed{\frac{1}{(1+\beta A)}}}_{\text{Sensitivity}} \cdot$ fractional change of gain w/o fb

$(1+\beta A) \rightarrow$ Desensitivity

* for a good s/s , Sensitivity $\downarrow \Rightarrow$ Desensitivity $\uparrow$
(ie. less sensitive to noise, temp. etc)

(2). Increase in I/p impedence :

$$V_i = V_s - V_f$$

$$V_i = V_s - \beta V_o$$

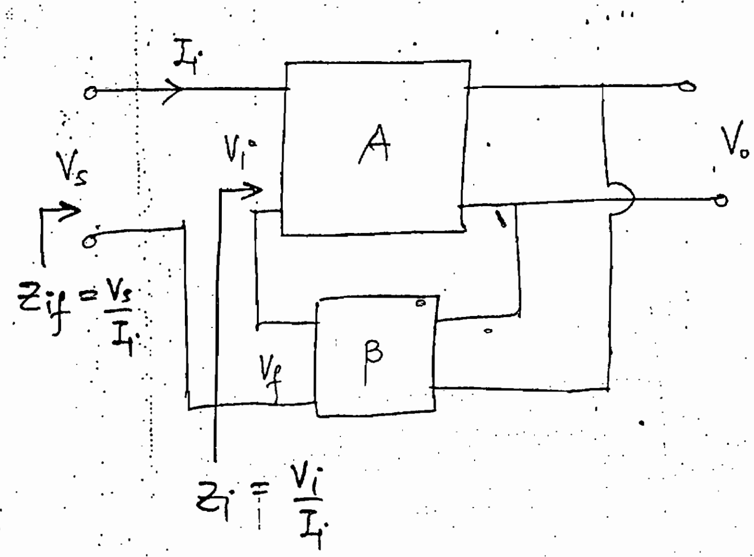
$$V_i = V_s - \beta A \cdot V_i$$

$$V_i(1+\beta A) = V_s$$

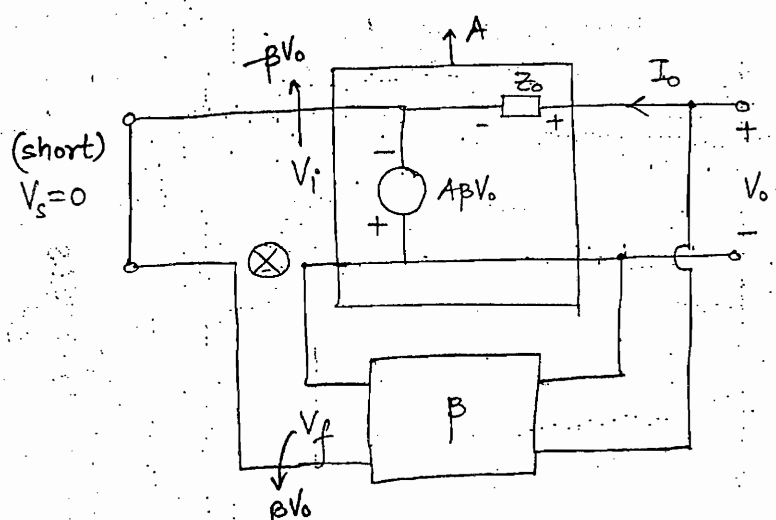
$$\frac{V_i}{I_i}(1+\beta A) = \frac{V_s}{I_i}$$

$$Z_i(1+\beta A) = Z_{if}$$

$$\therefore Z_{if} = (1+\beta A) \cdot Z_i$$



(3). Decrease in o/p impedence:

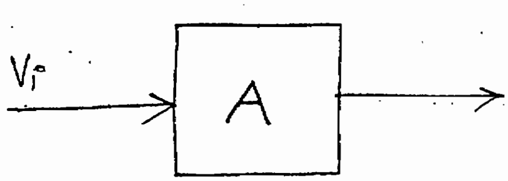


$$V_o + \beta A V_o = I_o Z_o$$

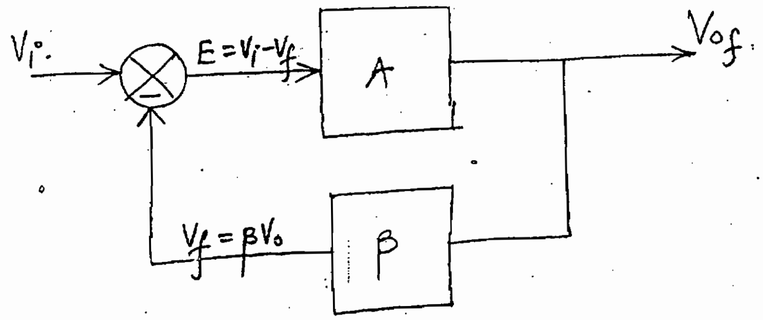
$$V_o(1+\beta A) = I_o Z_o$$

$$Z_{of} = \frac{V_o}{I_o} = \frac{Z_o}{(1+\beta A)}$$

(4). Linearity:



$$V_o = \underbrace{V_{offset}}_{DC} + AV_i + \underbrace{K_1 V_i^2 + K_2 V_i^3 + \cdots}_{\text{non-linearity}}$$



$$V_{of} = V_{offset} + AE + K_1 E^2 + K_2 E^3 + \cdots$$

$$V_{of} = V_{offset} + A(V_i - V_f) + K_1 E^2 + K_2 E^2 + \cdots$$

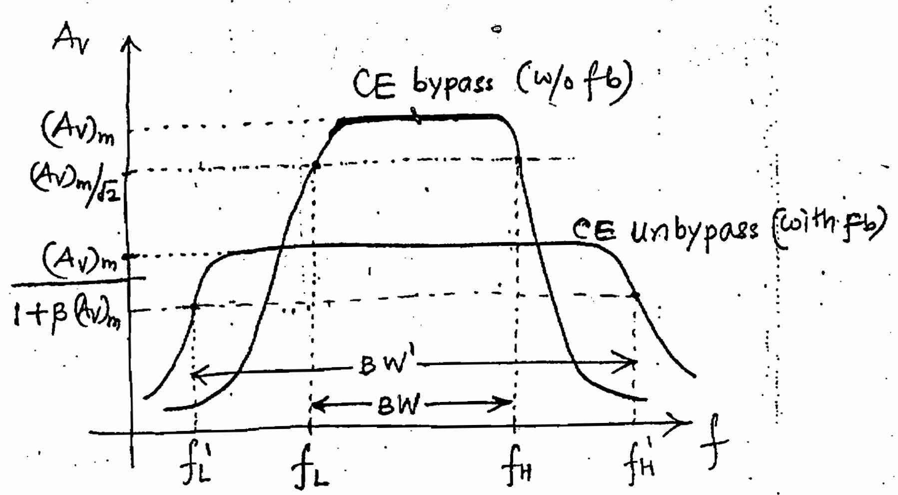
$$V_o(1 + \beta A) = V_{offset} + AV_i + K_1 E^2 + K_2 E^2 + \cdots$$

$$V_o = \underbrace{\frac{V_{offset}}{(1+\beta A)}}_{\text{tends to zero}} + \frac{A}{(1+\beta A)} \cdot V_i + \frac{K_1 E^2 + K_2 E^3 + \cdots}{(1+\beta A)}$$

$$V_o \approx \frac{A}{(1+\beta A)} \cdot V_i \qquad \text{and} \quad \beta A >>> 1$$

$$\therefore \boxed{V_o \approx \frac{1}{\beta} V_i} \rightarrow \text{linear relation}$$

(5) . Increase in BW -



Lower cut off frequency

$$(A_V)_l = \frac{(A_V)_m}{1 - j(f_L/f)} \quad ; \quad (A_V)_{lf} = \frac{(A_V)_l}{1+\beta (A_V)_l}$$

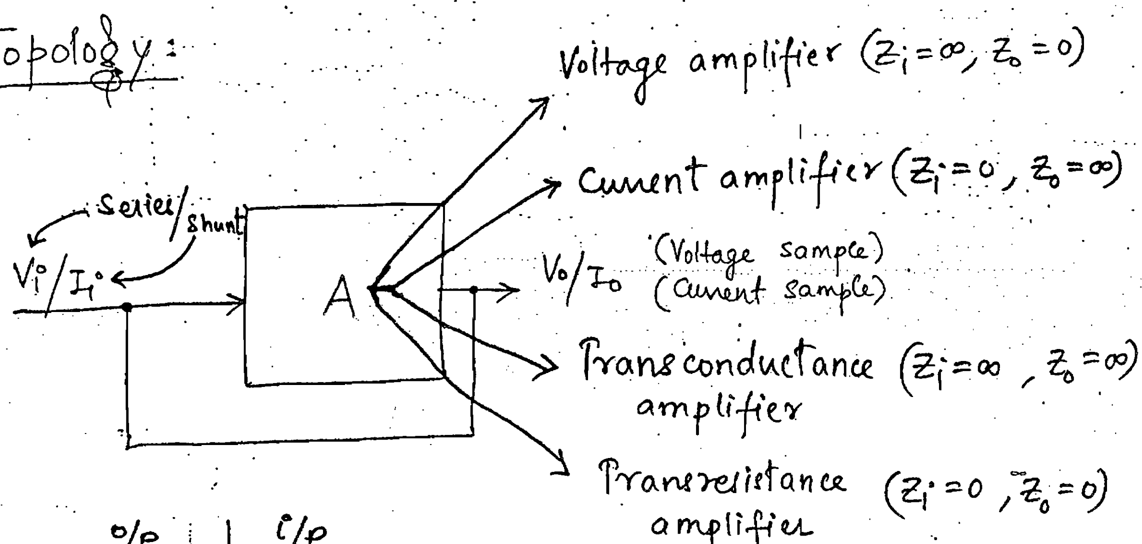
$$(A_v)_f = \frac{\frac{(A_v)_m}{1-j(f_L/f)}}{1+\beta\cdot\frac{(A_v)_m}{1-j(f_L/f)}} = \frac{(A_v)_m}{1+\beta(A_v)_m - j(f_L/f)}$$

$$(A_v)_{Lf} = \frac{\frac{(A_v)_m}{1+\beta(A_v)_m}}{1-j\left[\frac{f_L}{1+\beta(A_v)_m}\right]\cdot\frac{1}{f}}$$

$$\therefore\quad f_L' = \frac{f_L}{1+\beta(A_v)_m}$$

Similarly, $f_H' = f_H\,(1+\beta(A_v)_m)$

Topology:

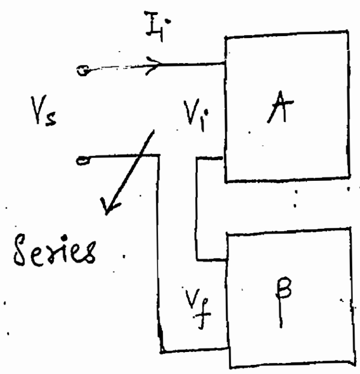


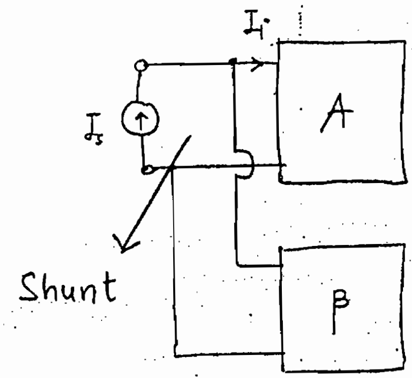
	o/p	i/p
(1).	Voltage	Series
(2).	Voltage	Shunt
(3).	Current	series
(4).	Current	shunt

Topology → Block diagram analysis, Practical ckt analysis

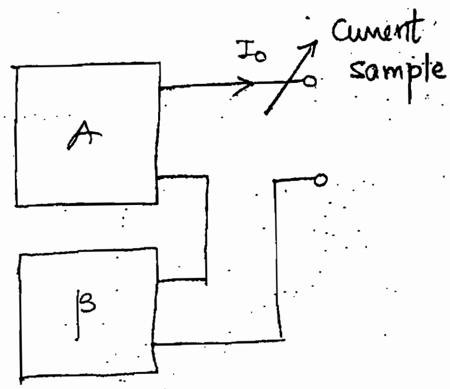
Block Diagram Analysis :

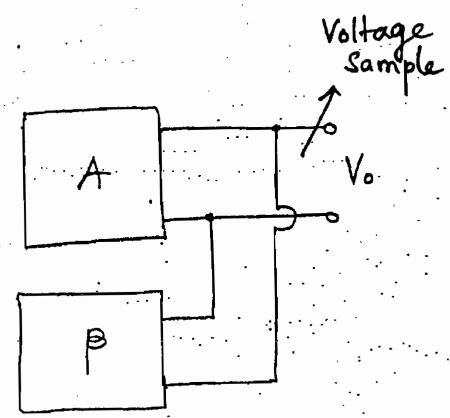
i/p side



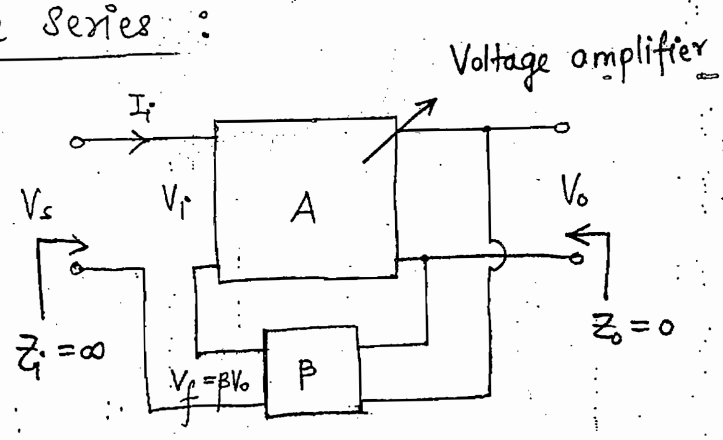


o/p side





(1). Voltage Series :



Voltage Series
(o/p)

(or)

Series Voltage
(i/p)

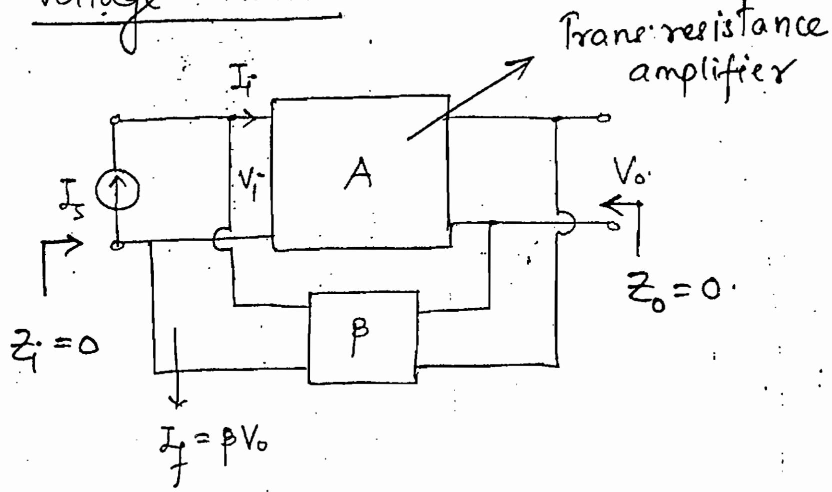
(or)

Series Shunt
(i/p)

(or)

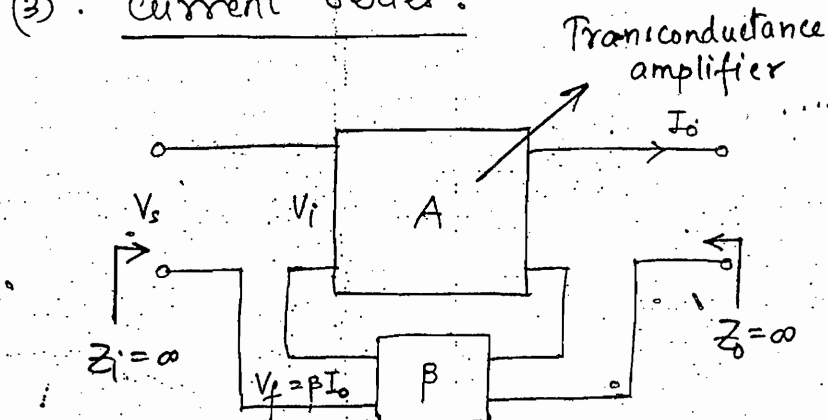
Voltage voltage
(o/p

(2). Voltage Shunt :



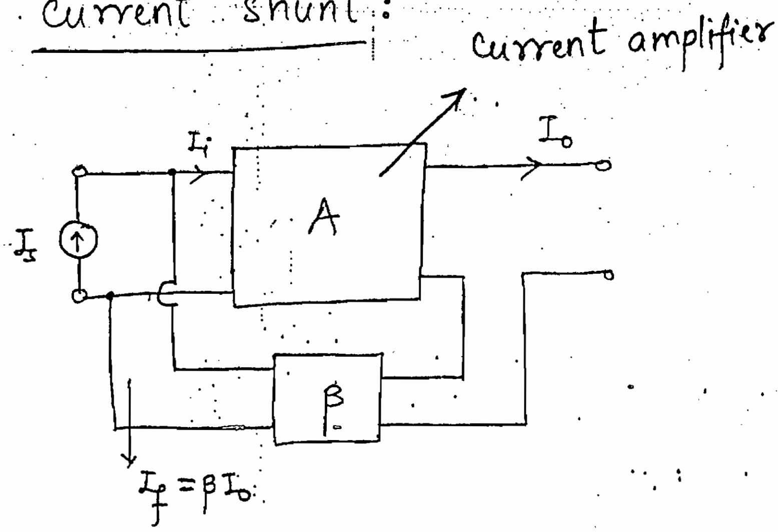
Voltage shunt
(or)
Shunt Voltage
(or)
Shunt shunt
(or)
Voltage current

(3). Current series :



Current Series
(or)
Series current
(or)
Series Series
(or)
Current Voltage

(4). Current shunt :

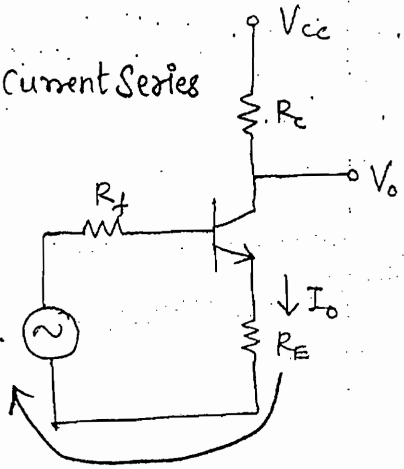


Current shunt
(or)
shunt current
(or)
shunt series
(or)
current current

Practical ckt analysis :

ckt-1

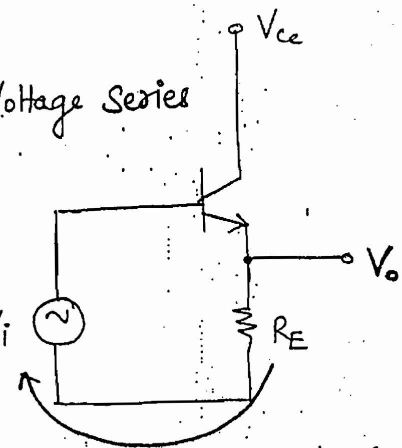
Current Series



CE unbypass

ckt-2

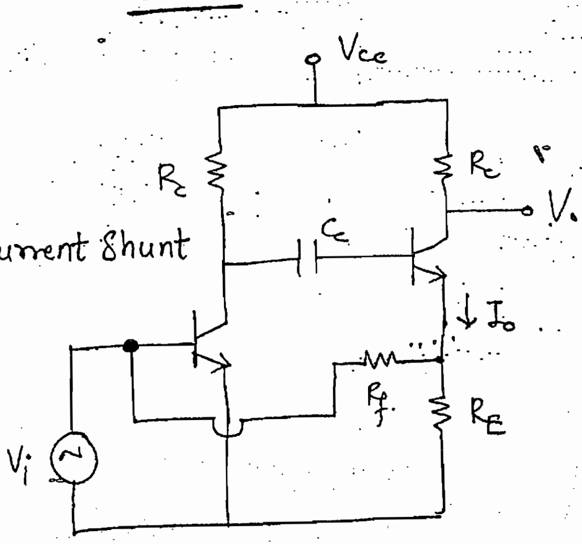
Voltage Series



Common Collector (CC)

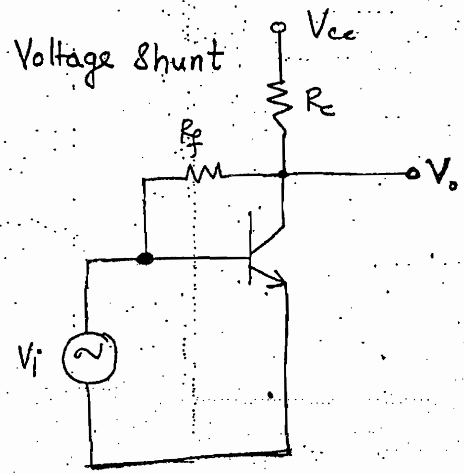
ckt-3

Current Shunt



ckt-4

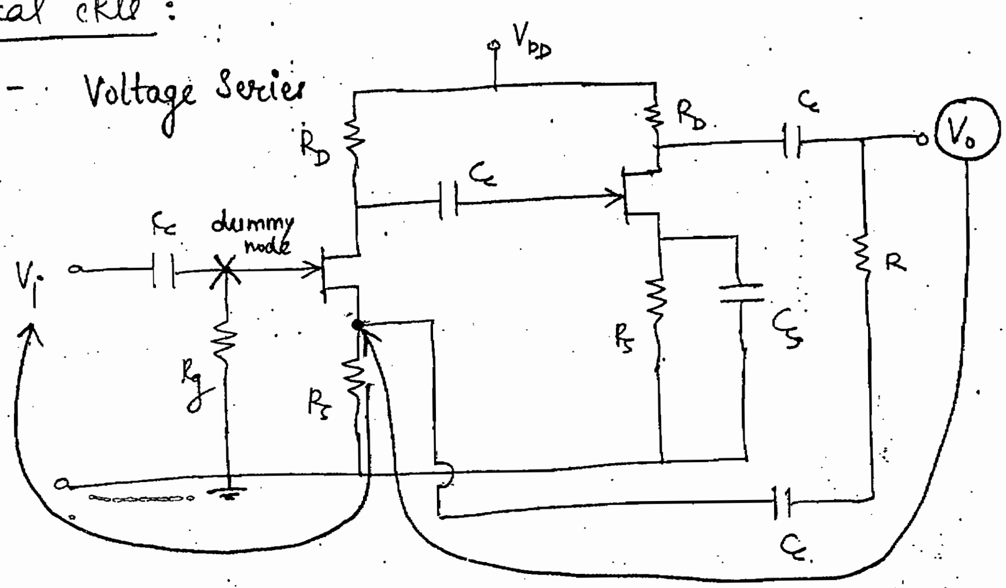
Voltage Shunt



Collector to base bias

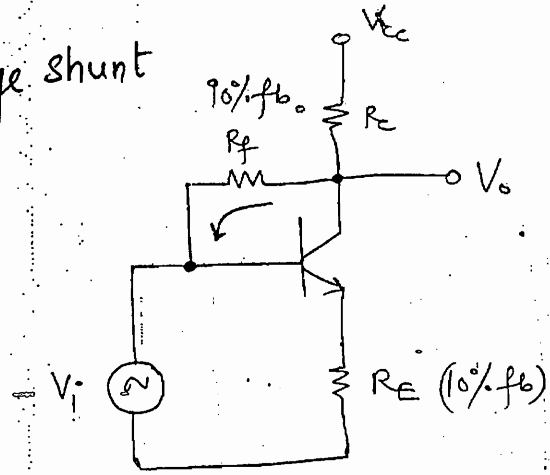
Typical ckts :

FET - Voltage Series



BJT –

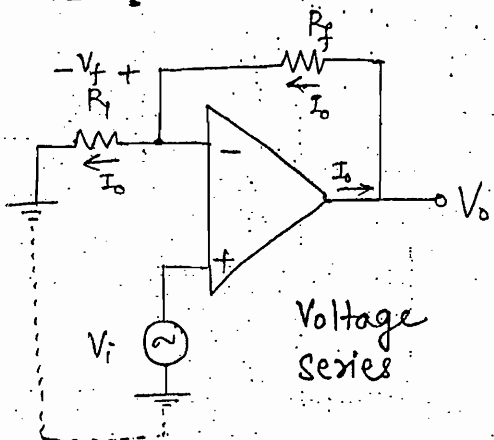
Voltage shunt



* If shunt not works completely, then series fb are required (2 things at a time).

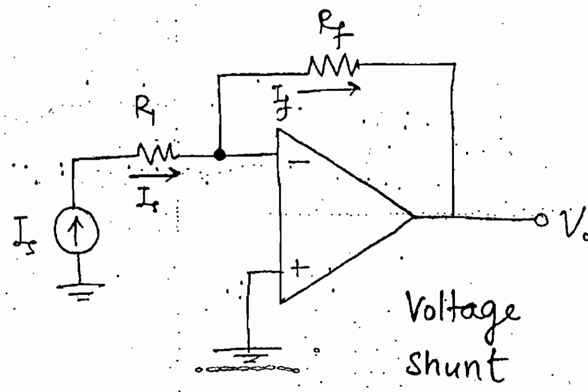
Note: If 2 fb i.e. series & shunt are connected at a time, shunt dominates series fb.

OP – Amps –



Voltage series

Non-inverting Amplifier



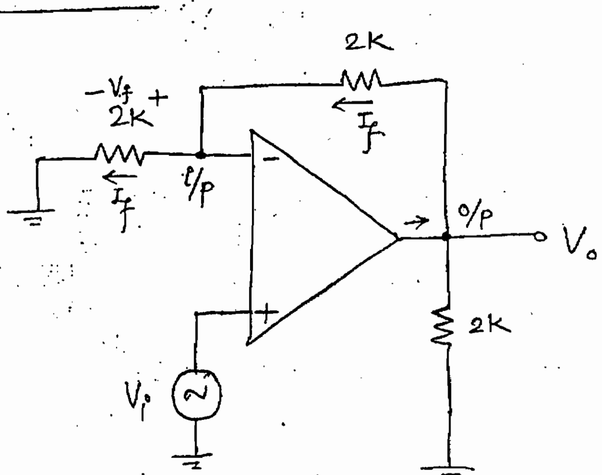
Voltage shunt

Inverting Amplifier

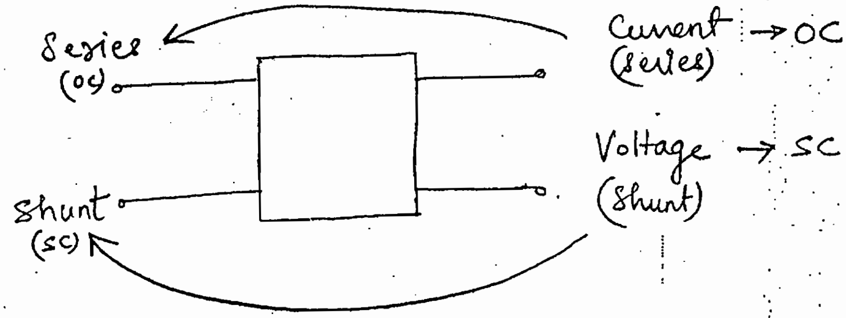
Problems in (–ve) fb theory:

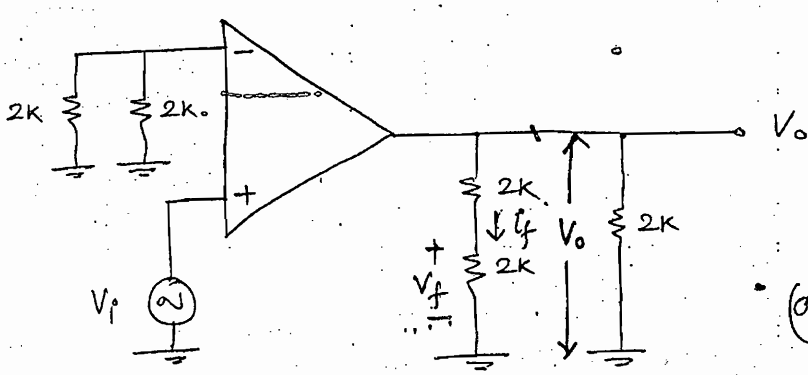
Model – (1):

Voltage Series



Calculate A_f and β ?





$$\beta = \frac{V_f}{V_o} = \frac{2K}{2K+2K} = \frac{1}{2}$$

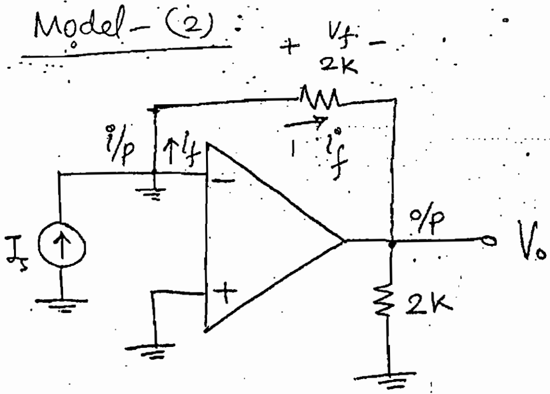
$$A_f = \frac{1}{\beta} = 2$$

(or) $\beta = \frac{I_f}{V_o} = \frac{1}{4K\Omega}$

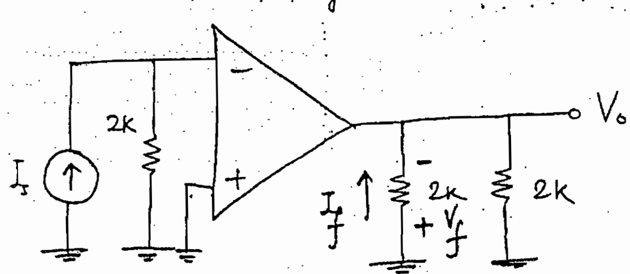
or By virtual ground theory,

$$A_f = \frac{R_f}{R_1} + 1 = \frac{2K}{2K} + 1 = 2.$$

Model-(2) :



Voltage shunt

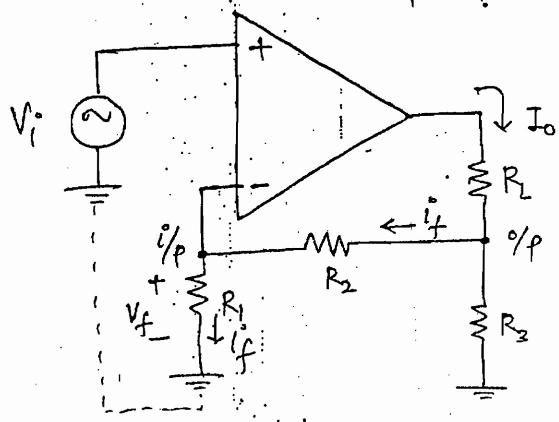


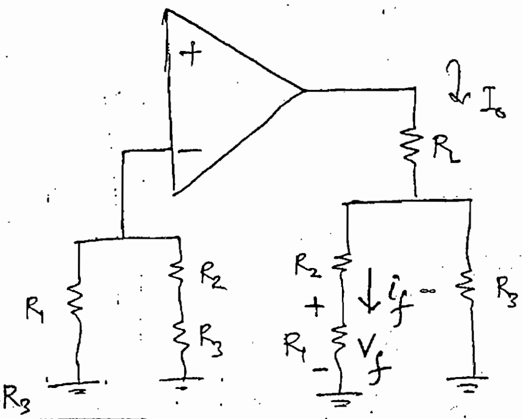
$$\beta = \frac{V_f}{V_o} = -1$$

(or) $\beta = \frac{I_f}{V_o} = \frac{-1}{2K}$

Model-(3):

$\beta = ?$ Current series



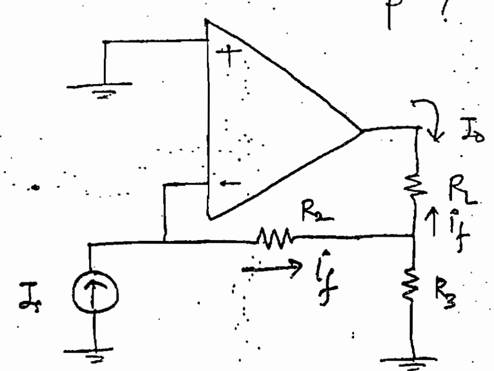


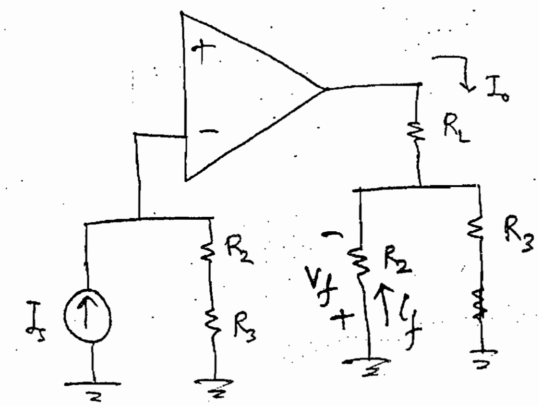
$$I_f = \frac{I_o \times R_3}{(R_1 + R_2 + R_3)} \Rightarrow \beta = \frac{i_f}{i_o} = \frac{R_3}{R_1 + R_2 + R_3}$$

$$(\text{or}) \quad \beta = \frac{V_f}{I_o} = \frac{i_f R_1}{i_o} = \frac{R_1 R_3}{(R_1 + R_2 + R_3)}$$

Model-(4):

$\beta = ?$ Current shunt

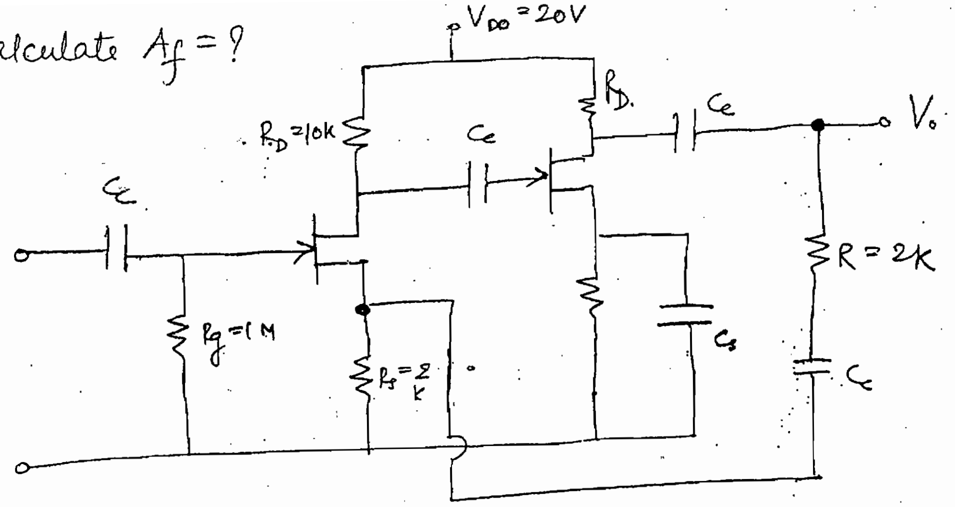




$$i_f = \frac{-I_o \times R_3}{(R_2 + R_3)} \Rightarrow \beta = \frac{i_f}{I_o} = \frac{-R_3}{R_2 + R_3}$$

$$(\text{or}) \quad \beta = \frac{V_f}{I_o} = \frac{i_f \cdot R_2}{I_o} = \frac{-R_2 R_3}{R_2 + R_3}$$

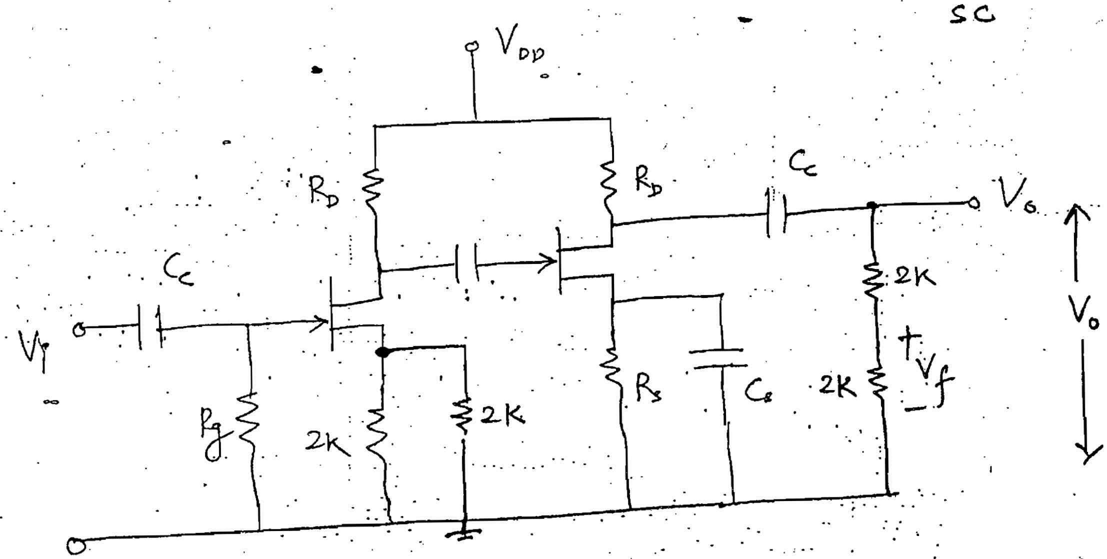
Q. Calculate $A_f = ?$



o/p → Voltage → SC

i/p → series → OC

⇓



$$\beta = \frac{V_f}{V_o} = \frac{2K}{2K + 2K} = \frac{1}{2}$$

$$A_f = \frac{1}{\beta} = 2$$

Procedure for feedback :

To get the i/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.

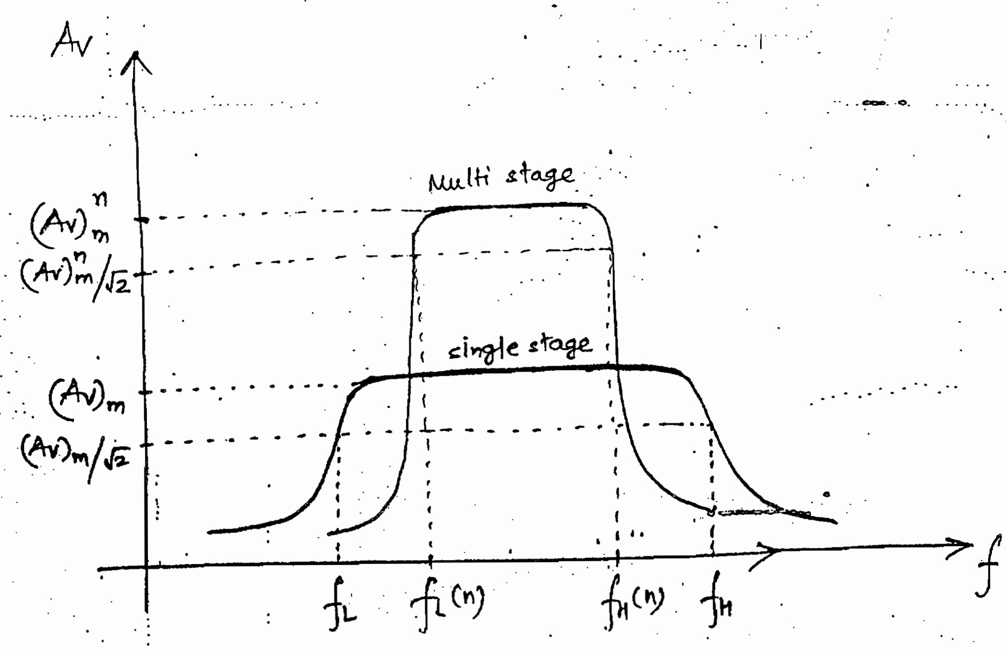
→ for voltage shunt or current shunt, i/p node should be grounded.

Integrated Theory (OP-Amps):

(Integration → multistage)

Multistage Amplifiers:

Effect of cascading on BW: (identical stages)



High cut-off frequency:

Single stage → $(A_v)_h = \dfrac{(A_v)_m}{1+j(f/f_H)}$

$$\left|\frac{(A_v)_h}{(A_v)_m}\right| = \frac{1}{\sqrt{1+(f/f_H)^2}}$$

n identical stages → $\left|\dfrac{(A_v)_h}{(A_v)_m}\right|^n = \left|\dfrac{1}{\sqrt{1+(f/f_H)^2}}\right|^n$

when $f \to f_H(n)$

$$\left|\frac{1}{\sqrt{1+\left(\frac{f_H(n)}{f_H}\right)^2}}\right|^n = \frac{1}{\sqrt{2}}$$

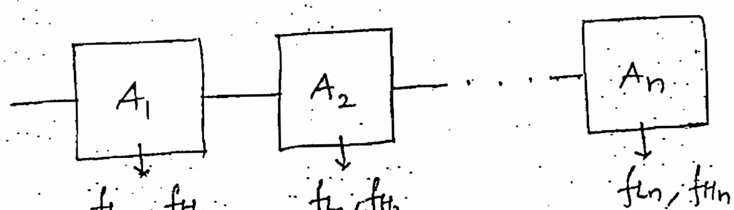
$$1+\left\{\frac{f_H(n)}{f_H}\right\}^2 = 2^{1/n}$$

$$\boxed{f_H(n) = \sqrt{2^{1/n}-1}\cdot f_H}$$

Similarly,

$$\boxed{f_L(n) = \frac{f_L}{\sqrt{2^{1/n}-1}}}$$

<u>Effect of cascading on BW</u> : (Non identical stages)



$$A^n = A_1 \cdot A_2 \cdot A_3 \cdots\cdots A_n$$

$$A_l^n = A_{l_1}\cdot A_{l_2}\cdot A_{l_3}\cdots\cdots A_{l_n}$$

$$A_h^n = A_{h_1}\cdot A_{h_2}\cdot A_{h_3}\cdots\cdots A_{h_n}$$

$$\frac{\cancel{A^n}}{\sqrt{1+\left(\frac{f_L^n}{f}\right)^2}} = \frac{\cancel{A_1}}{\sqrt{1+\left(\frac{f_{L_1}}{f}\right)^2}}\cdot\frac{\cancel{A_2}}{\sqrt{1+\left(\frac{f_{L_2}}{f}\right)^2}}\cdots\cdots\frac{\cancel{A_n}}{\sqrt{1+\left(\frac{f_{L_n}}{f}\right)^2}}$$

$$\cancel{1}+\left(\frac{f_L^n}{\cancel{f}}\right)^2 = \cancel{1}+\left(\frac{f_{L_1}}{\cancel{f}}\right)^2+\left(\frac{f_{L_2}}{\cancel{f}}\right)^2+\cdots\cdots+\left(\frac{f_{L_n}}{\cancel{f}}\right)^2+\underbrace{\left(\frac{f_{L_1}f_{L_2}}{f^2}\right)^2+\cdots}_{\text{neglected}}$$

$$f_L^n = \sqrt{f_{L_1}^2+f_{L_2}^2+\cdots+f_{L_n}^2}$$

Similarly, $$f_H^n = \frac{1}{\sqrt{(1/f_{H_1})^2+1/f_{H_2}^2+\cdots+1/f_{H_n}^2}}$$

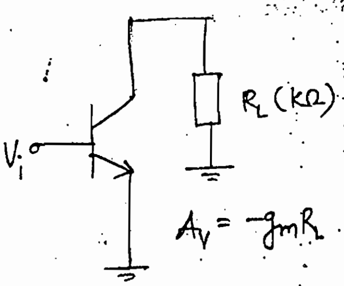
Important cascading designs:

(1). Cascode amplifiers : (CE − CB) (Wide Band Amplifiers)

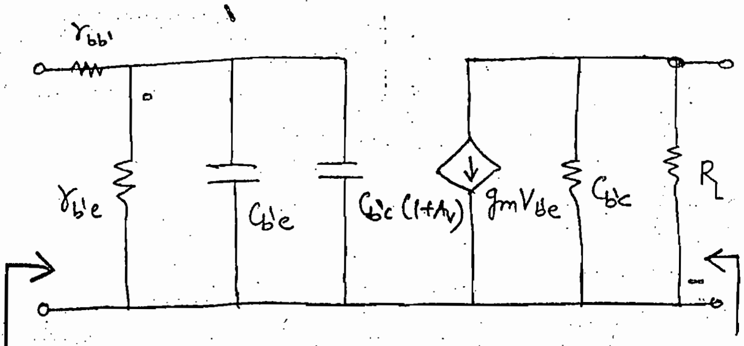
(2). Darlington pair : (CC − CE) (High i/p impedence ckt)

(1). Cascode amplifiers (CE − CB)

CE :



$A_V = -g_m R_L$

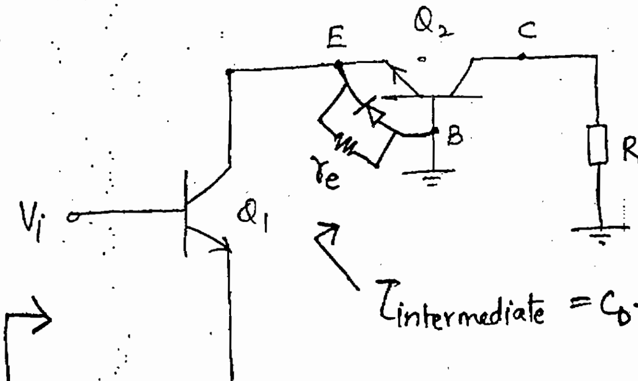


$\tau_i = r_{bb'} // r_{b'e} \{ C_{b'e} + \underbrace{C_{b'c}(1+A_V)}_{\text{millers capacitive Effect}} \}$ $\tau_o = C_{b'c} R_L$

As $A_V = 1000$ $\therefore \tau_i \uparrow = \dfrac{1}{\omega_{BW} \downarrow}$ $(\tau_i >> \tau_o)$

To reduce, τ_i there must be reduction in A_V to improve ω_{BW}.

CE − CB :



$\tau_2 = C_T \cdot R_L$

$\tau_{intermediate} = C_D \cdot r_e$

$\downarrow \tau_i = r_{bb'} // r_{b'e} \{ C_{b'e} + C_{b'c}(2) \} = \dfrac{1}{\omega_{BW} \uparrow}$

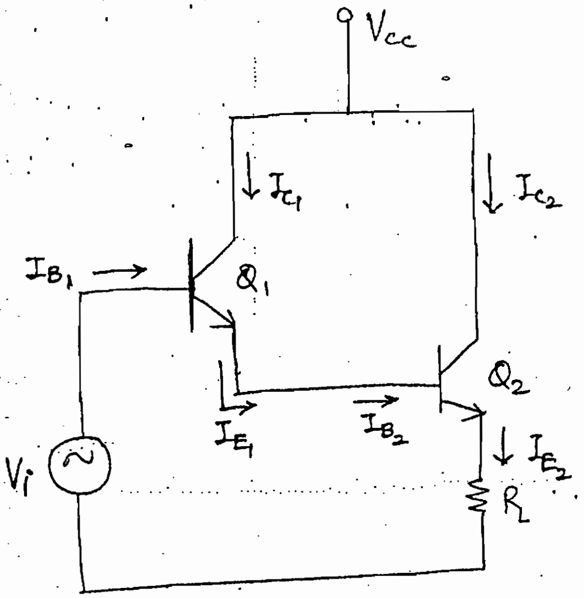
$A_{V_1} = -g_m R_L$
$= -g_m r_e$
$= -\dfrac{r_e}{r_e} = -1$

$A_{V_2} = g_m R_L$

$A_{cascade} = A_{V_1} \cdot A_{V_2}$
$= -1 \cdot g_m R_L$
$= -g_m R_L$ $(= A_V)_{CE}$

Hence, with CE−CB, ω_{BW} increases with keeping the gain ($A_{cascade} = A_V \simeq -g_m R_L$) same as with single CE.

(2). <u>Darlington pair:</u> (CC−CC)



(4). $Z_o = R_E \,//\, r_e$

$\approx r_e$

$\boxed{\therefore Z_o = r_e}$ → Impedence matching

(1). $A_I = A_{I_1} \cdot A_{I_2}$

$$A_{I_1} = \frac{I_{E_1}}{I_{B_1}} = \frac{I_{B_1} + I_{C_1}}{I_{B_1}} = 1 + \frac{I_{C_1}}{I_{B_1}} = 1 + \beta$$

$$A_{I_2} = \frac{I_{E_2}}{I_{B_2}} = \frac{I_{B_2} + I_{C_2}}{I_{B_2}} = 1 + \frac{I_{C_2}}{I_{B_2}} = 1 + \beta$$

$$\therefore A_I = (1+\beta)^2 \approx \beta^2 \Rightarrow \boxed{A_I = \beta^2}$$

(2). $A_V = A_{V_1} \cdot A_{V_2}$

$= 1 \cdot 1$

$\boxed{\therefore A_V = 1}$ → Buffer action

(3). $Z_i = A_I \cdot R_L$

$\therefore \boxed{Z_i = \beta^2 \cdot R_L}$

<u>Assume</u>: $\beta = 100$, $R_L = 1\,K\Omega$

$Z_i = (100)^2 \, 1\,K\Omega = 10\,M\Omega$ (High)

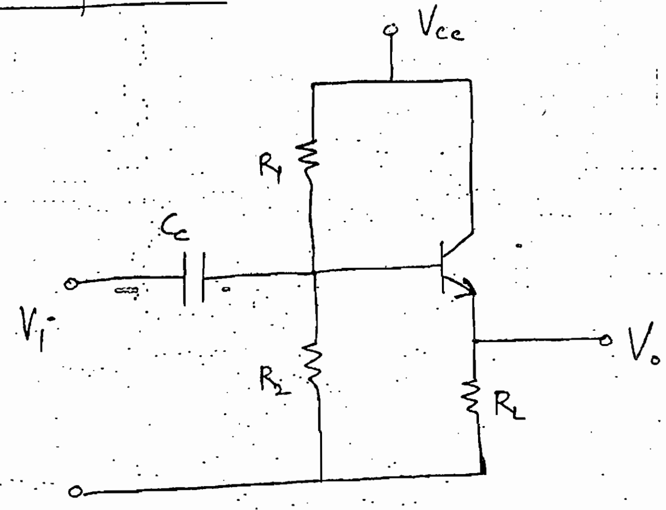
Disadvantages:

(1). In practical design, the I/p impedence will be low for a darlington pair becoz of biasing resistors.

Eg. $Z_i // R_1 // R_2$ if $R_1 = 100k\,\Omega$
$R_2 = 10k\,\Omega$

$$Z_i = Z_i // 100k // 10k$$
$$\approx 10K$$

Emitter follower : Boost strap



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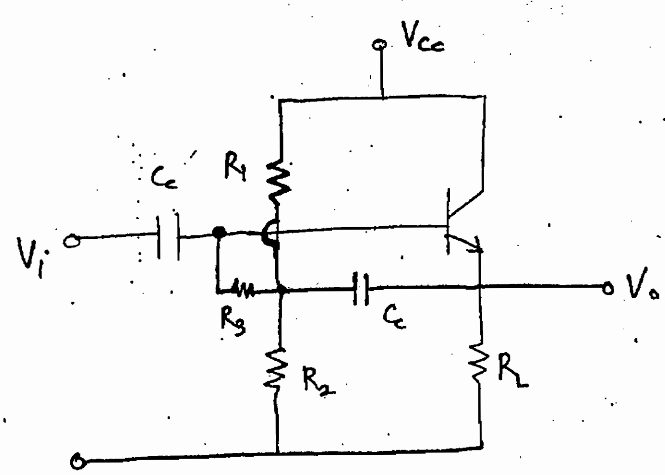
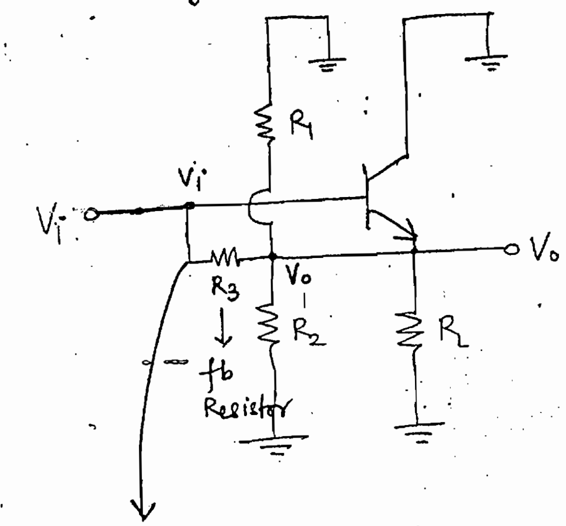
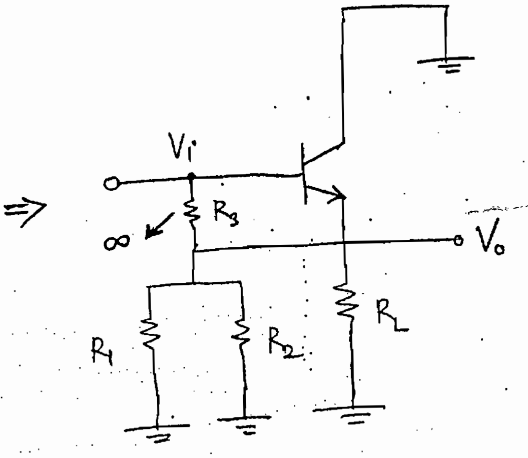


Voltage shunt

AC analysis:



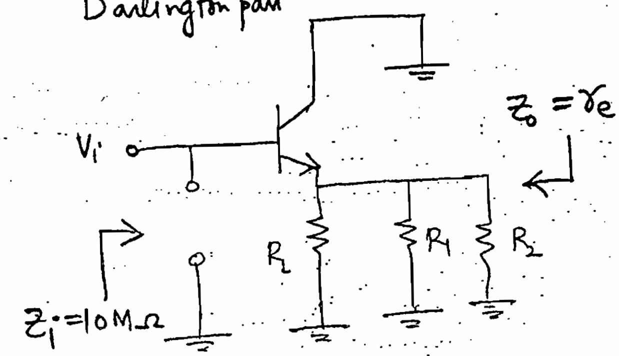
$\Rightarrow$



$$R_{3i} = \frac{R_3}{1-A_V} \approx \frac{R_3}{1-1} \approx \infty$$

$\Downarrow$

Darlington pair

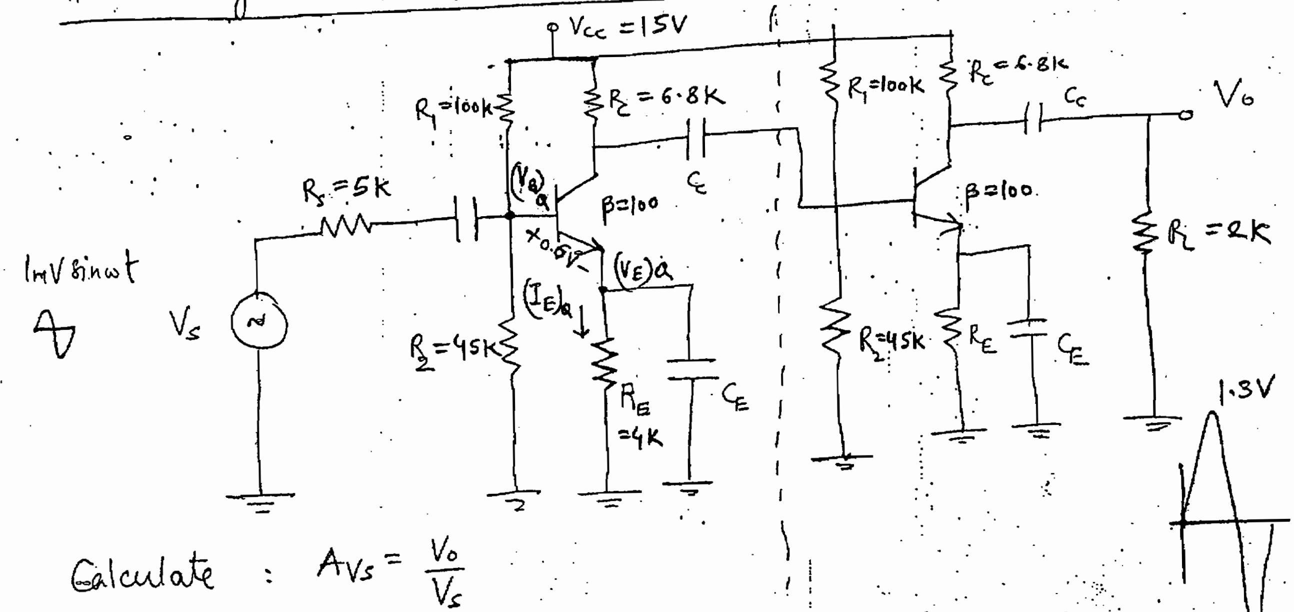


Coupling designs:

(1). RC coupling

(2). Direct coupling

Two stage RC coupled CE amplifier:



Calculate : $A_{Vs} = \frac{V_o}{V_s}$

DC analysis : Q_1 & Q_2

$$(V_B)_Q = \frac{15 \times 45K}{(100K + 45K)} = 4.6\,V$$

$$(V_E)_Q = 4.6 - 0.6 = 4V$$

$$(I_E)_Q = \frac{(V_E)_Q}{R_E} = \frac{4}{4\,K} = 1mA$$

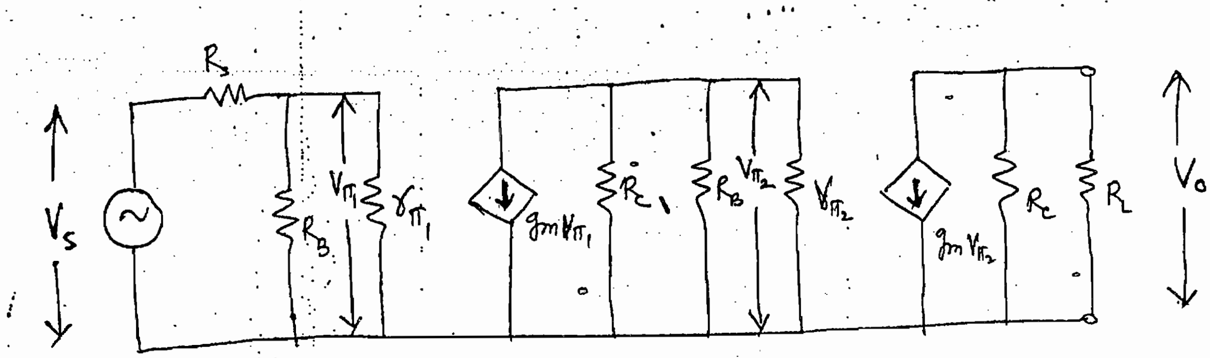
$$g_m = \frac{(I_c)_Q}{V_T} \approx \frac{1\,mA}{25\,mV} = 40\,m\mho$$

$$r_e = \frac{1}{g_m} = 25\,\Omega$$

$$\beta r_e = r_\pi = h_{ie} = 100 \times 25 = 2.5\,K\Omega$$

AC analysis :

$$R_B = R_1 // R_2 = 100K // 45K \approx 32K$$



$$A_{vs} = \frac{V_o}{V_s} = \frac{V_o}{V_{\pi_2}} \cdot \frac{V_{\pi_2}}{V_{\pi_1}} \cdot \frac{V_{\pi_1}}{V_s}$$

$$\frac{V_{\pi_1}}{V_s} = \frac{R_B // r_{\pi_1}}{R_B // r_{\pi_1} + R_s} = \frac{32K // 2.5K}{32K // 2.5K + 5K} = 0.316\,K$$

$$V_{\pi_2} = -g_m V_{\pi_1} \cdot (R_c // R_B // r_{\pi_2})$$

$$\frac{V_{\pi_2}}{V_{\pi_1}} = -g_m (R_c // R_B // r_{\pi_2}) = -40\,m\mho \cdot (6.8K // 32K // 2.5K)$$
$$= -69$$

$$V_o = -g_m V_{\pi_2} (R_c // R_L)$$

$$\frac{V_o}{V_{\pi_2}} = -g_m (R_c // R_L) = -40\,m℧ \,(6.8K // 2K) = -61$$

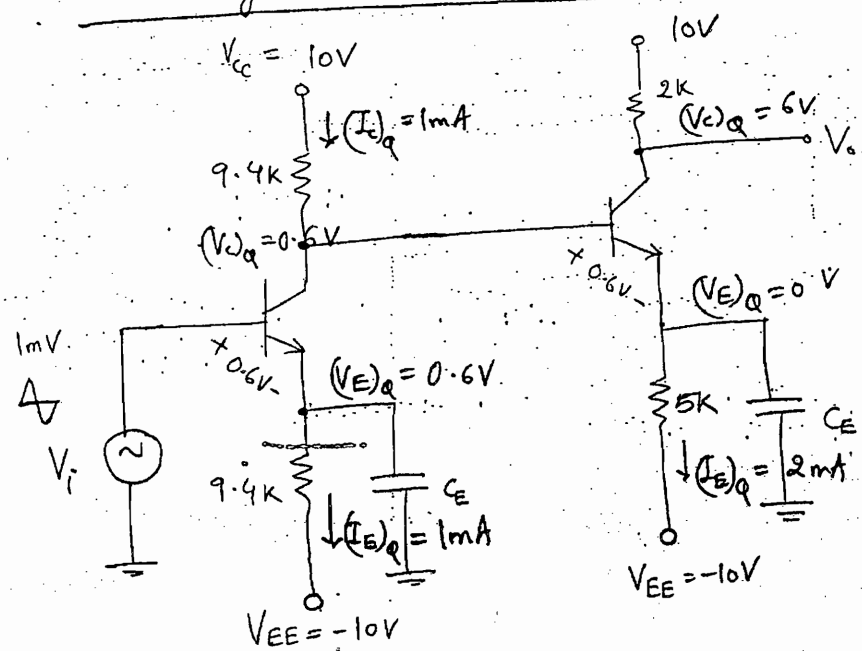
$$A_{V_s} = 0.316 \times -69 \times -61$$

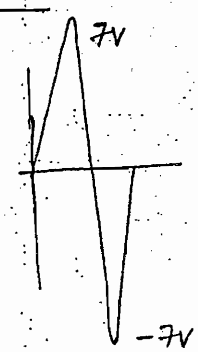
$$= 1330$$

$$\therefore \frac{V_o}{V_s} = 1330$$

Eg. $V_o = 1330 \times 1mV = 1.3V$

Two stage direct coupled BJT amplifier -





Q_1:

$$g_{m_1} = \frac{(I_c)_Q}{V_T} = \frac{1mA}{25mV} = 40\,m℧$$

$$r_{e_1} = \frac{1}{g_{m_1}} = 25\,\Omega$$

$$\beta r_{e_1} = r_{\pi_1} = 2.5K\Omega$$

Q_2:

$$g_{m_2} = \frac{(I_c)_Q}{V_T} = \frac{2mA}{25mV}$$

$$= 80\,m℧$$

$$r_{e_2} = \frac{1}{g_{m_2}} = 12.5\,\Omega$$

$$\beta r_{e_2} = r_{\pi_2} = 1.25K\Omega$$

$$A_{V_1} = -g_m (R_C // r_{\pi_2})$$
$$= -40 \cdot (9.4K // 1.25K)$$
$$= -44$$

$$A_{V_2} = -g_{m_2} \cdot R_C$$
$$= -80 \cdot 2K$$
$$= -160$$

$$\therefore A_V = A_{V_1} \cdot A_{V_2}$$
$$= -44 \times -160$$
$$= 7040$$

* This is the choice of op-amp. This type of amplifier are used in instrumentation and biomedical measurements. Direct coupled technique is better than RC coupled technique.

<u>Conclusion:</u>

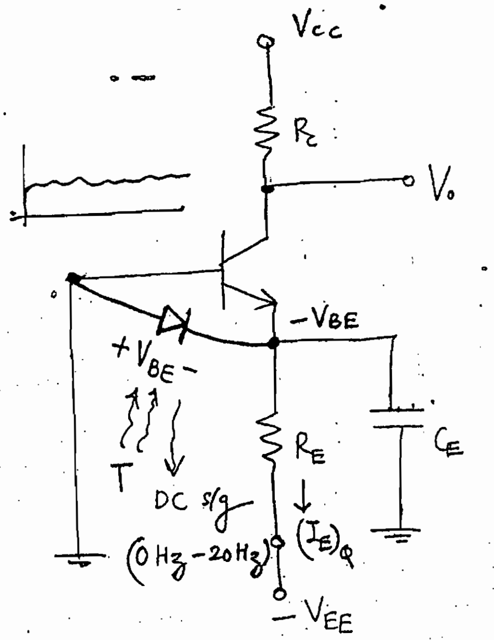
In integrated ckt designs, 2 important conditions to select an amplifier are:

(1). It should give high gain.

(2). The size should be less.

<u>Note</u>: So in integerated ckts, direct coupled amplifier became superior than RC coupled amplifies becoz of high gain and small in size.

Direct coupled amplifier:



(1). $\frac{dV_{BE}}{dT} = -2.5\, mV/^\circ C$

$\downarrow$

DC drift problem.

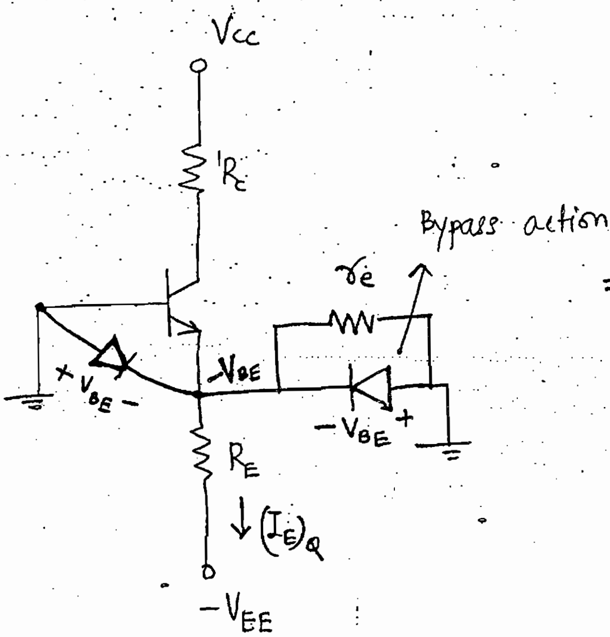
(2). $f_{min} = \frac{1}{2\pi R_E C_E}$

for $f_{min} \to 0 \Rightarrow C_E \to \infty$

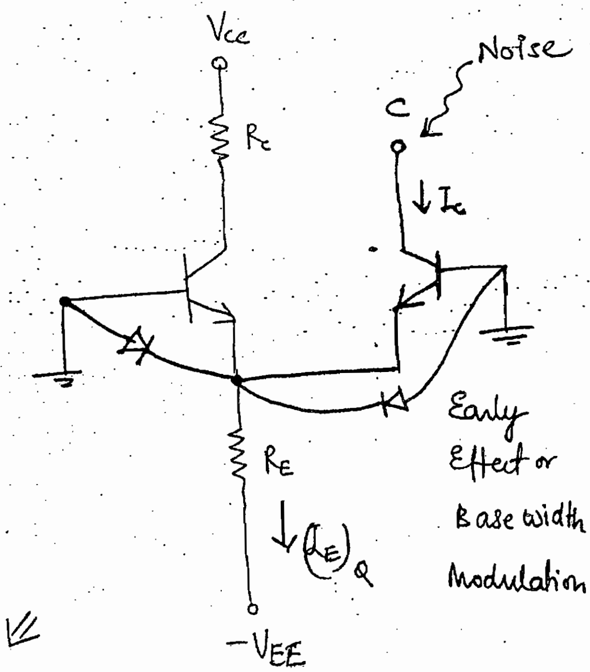
$\downarrow$

(Impossible to fabricate on IC practically).

$\Downarrow$

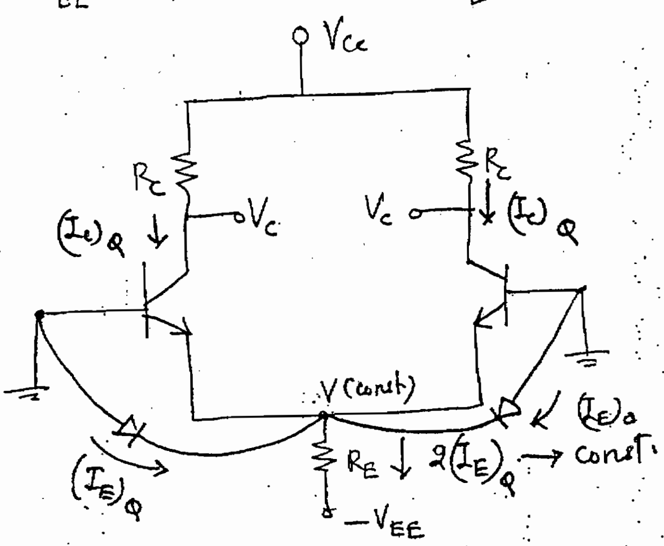


$\Rightarrow$



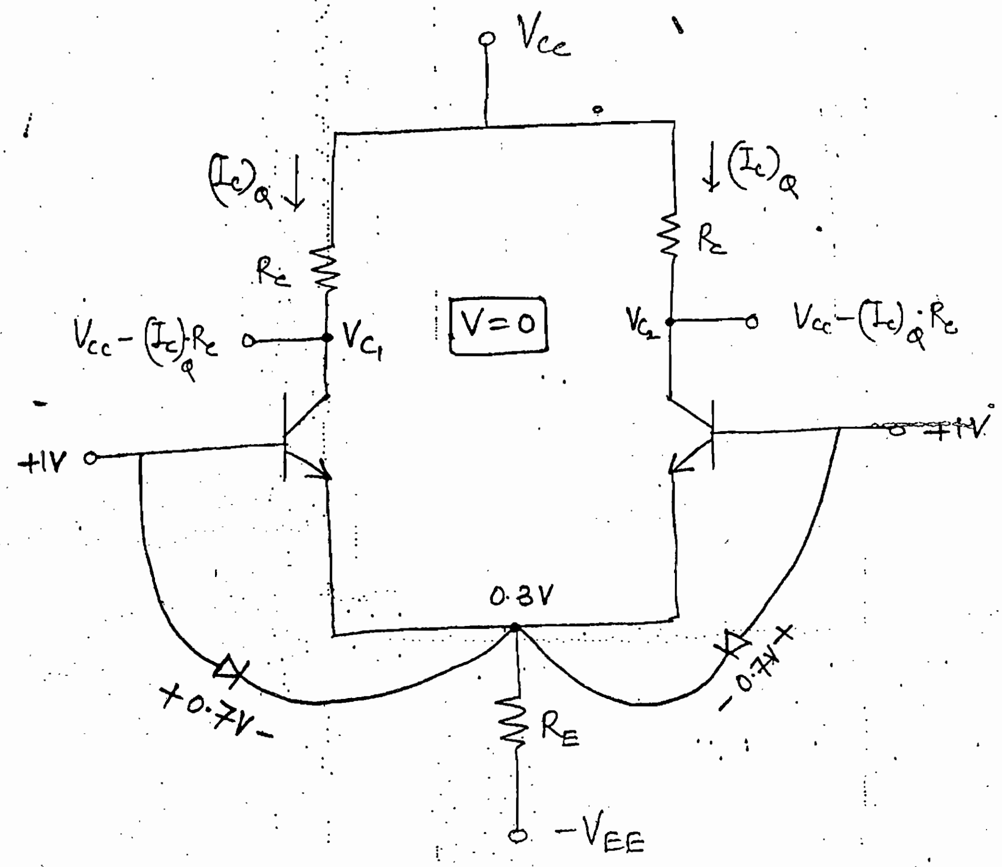
$\Swarrow$

Differential amplifier



Testing of differential amplifier for various i/p :

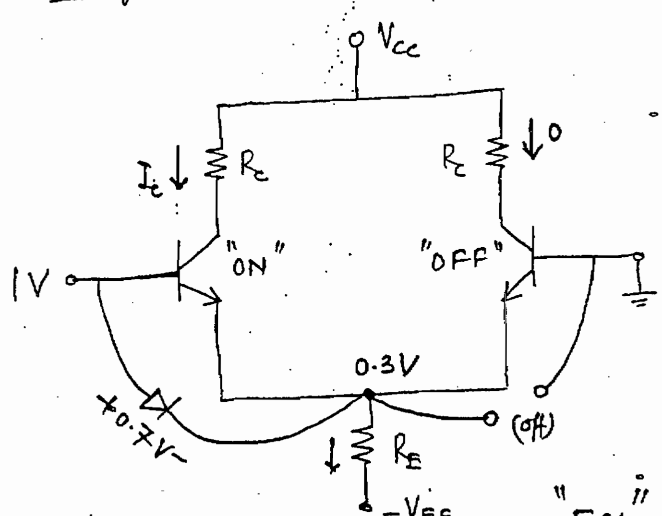
Common i/p :

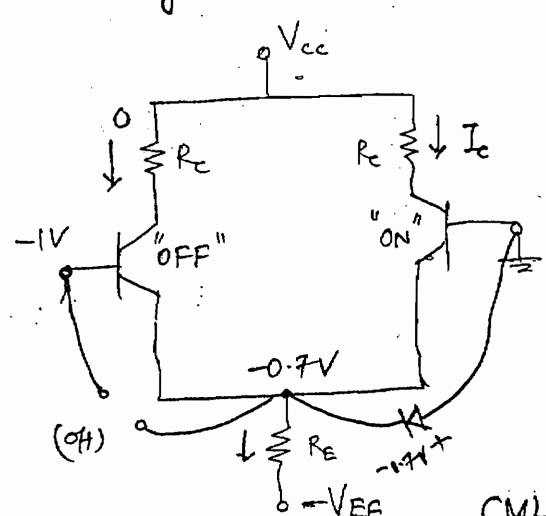


Common s/g are nothing but noise (or) interference s/g which is eliminated by differential amplifier.

$\therefore \; V = V_{c_1} - V_{c_2} = 0$ for common s/g.

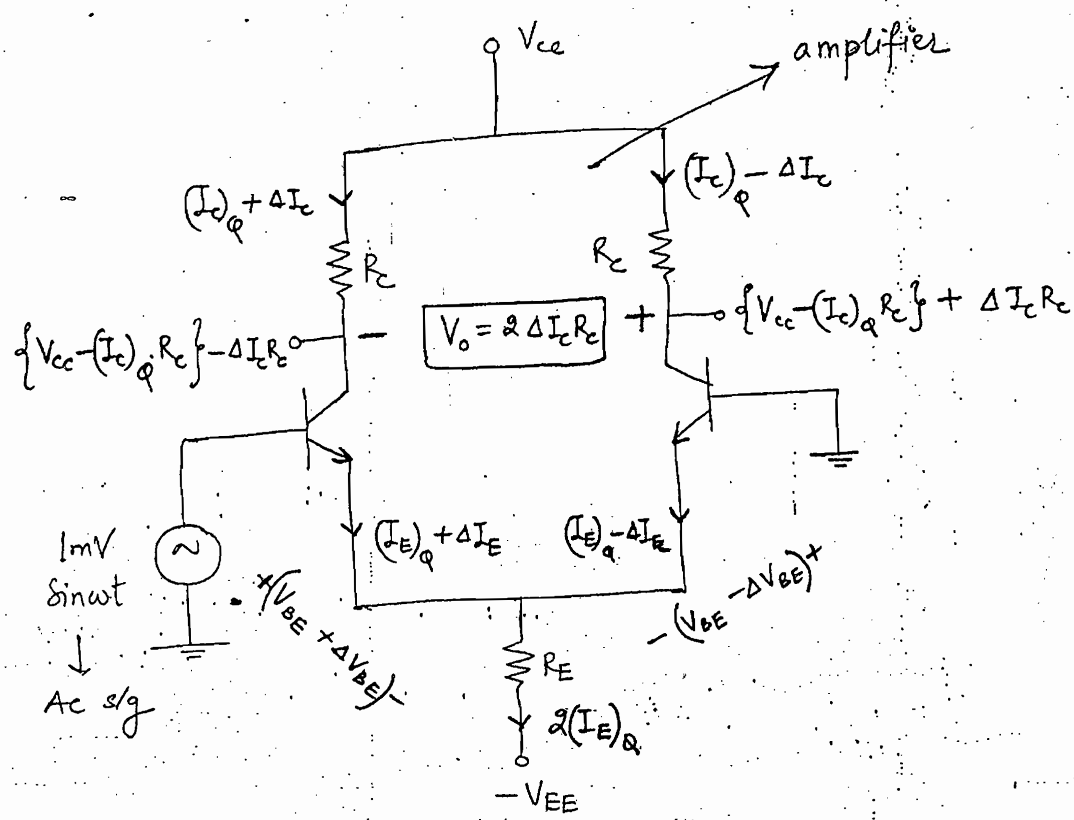
Large difference i/p : ($\geq$ 1V) → switching



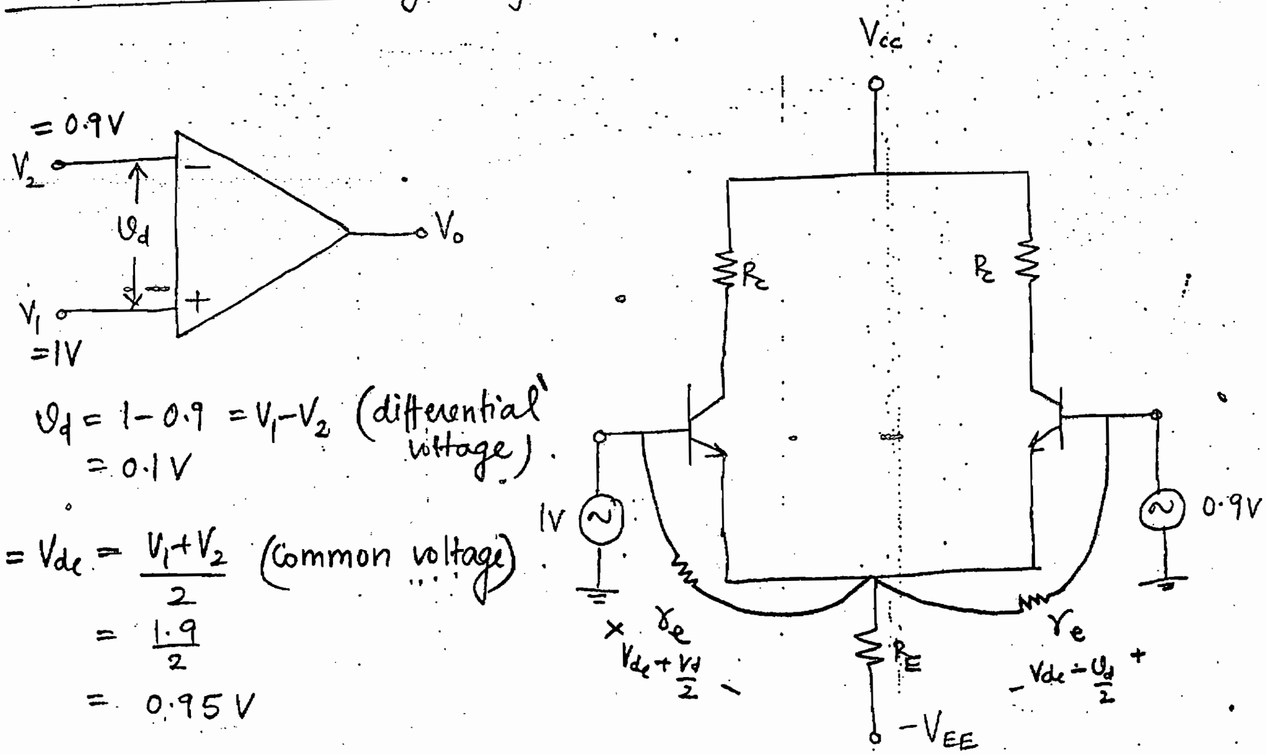


"ECL" (Emitter coupled logic) (or) CML (current mode logic)

Small difference i/p :



Mathematical analysis of differential amplifier :



$v_d = 1 - 0.9 = V_1 - V_2$ (differential voltage)
$= 0.1\,V$

$V_c = V_{dc} = \dfrac{V_1 + V_2}{2}$ (Common voltage)
$= \dfrac{1.9}{2}$
$= 0.95\,V$

$$V_1 = \frac{V_1 - V_2}{2} + \frac{V_1 + V_2}{2}$$

$$= \frac{v_d}{2} + v_{dc}$$

$$V_2 = \frac{V_2 - V_1}{2} + \frac{V_1 + V_2}{2}$$

$$= -\frac{v_d}{2} + v_{dc}$$

Mathematics involved inside:

$$V_1 = \frac{1-0.9}{2} + \frac{1+0.9}{2} = \frac{0.1}{2} + 0.95$$

$$V_2 = \frac{0.9-1}{2} + \frac{1+0.9}{2} = -\frac{0.1}{2} + 0.95$$

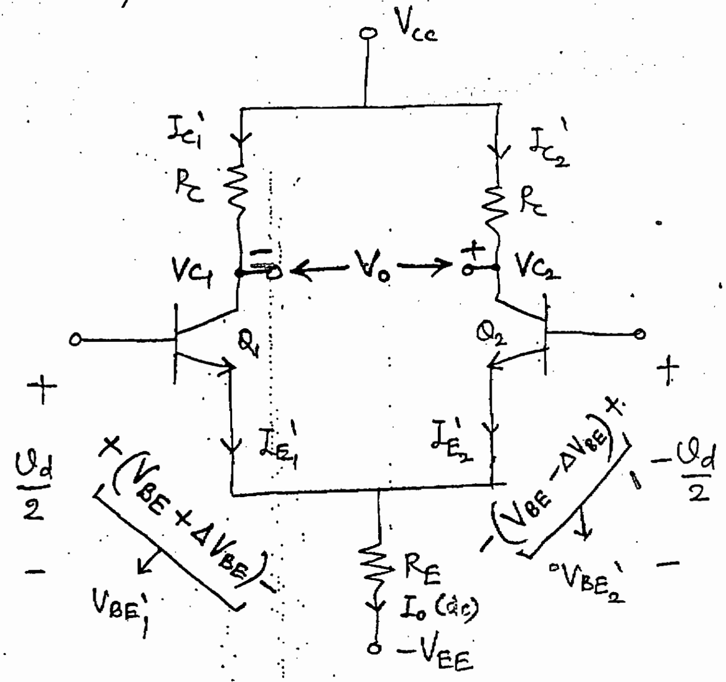
$$v_d = V_1 - V_2$$

$$= \frac{0.1}{2} + \cancel{0.95} - \left(-\frac{0.1}{2}\right) - \cancel{0.95}$$

$$= 0.1\ V$$

Transfer characteristics of op-Amp :

Analysis:



$$V_0 = V_{C_1} - V_{C_2}$$

$$= (V_{CC} - I_{C_1}' R_C) - (V_{CC} - I_{C_2}' R_C)$$

$$= (I_{C_2}' - I_{C_1}') R_C$$

$$I_{E_1}' = (I_E)_Q + \Delta I_E$$

$$I_{E_2}' = (I_E)_Q - \Delta I_E$$

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

$$I_{E_1}' = I_{E_0} \cdot e^{V_{BE} + \Delta V_{BE}/V_T}$$

$$I_{E_1}' = I_{E0} \cdot e^{V_{BE_1}'/V_T}$$

$$I_{E_2}' = I_{E0} \cdot e^{V_{BE_2}'/V_T}$$

$$\frac{I_{E_1}'}{I_{E_2}'} = e^{\underbrace{(V_{BE_1}' - V_{BE_2}')}_{v_d}/V_T}$$

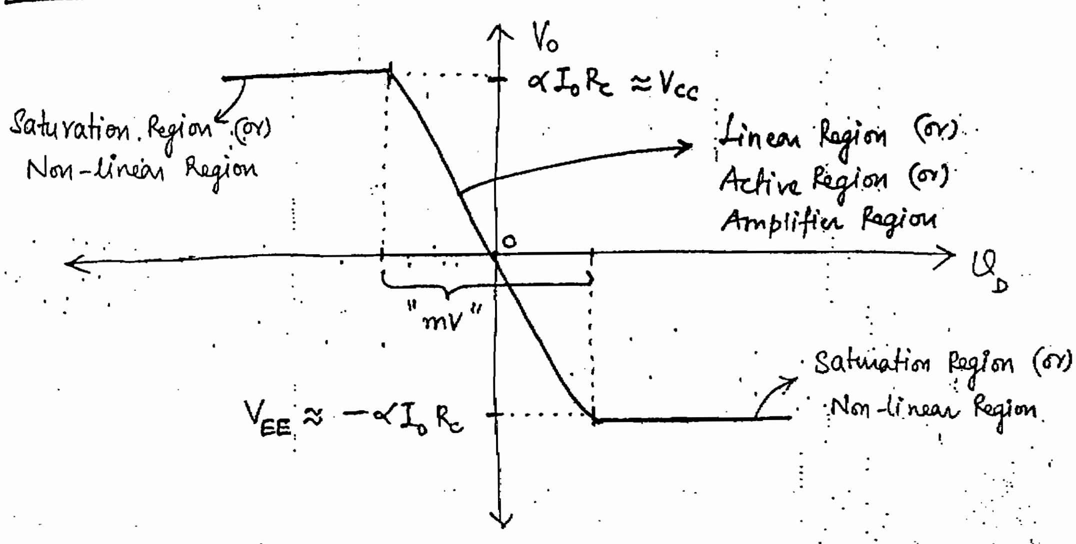
$$\frac{I_{E_1}'}{I_{E_2}'} = e^{v_d/V_T}$$

$$\frac{I_{E_1}' - I_{E_2}'}{I_{E_1}' + I_{E_2}'} = \frac{e^{v_d/V_T} - 1}{e^{v_d/V_T} + 1}$$

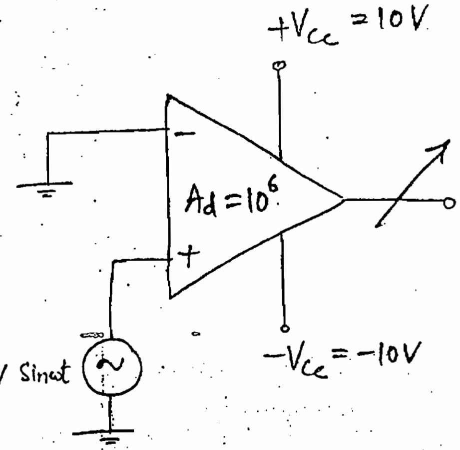
(1). $I_{E_1}' - I_{E_2}' = \dfrac{I_{c_1}' - I_{c_2}'}{\alpha}$ (2). $I_{E_1}' + I_{E_2}' = I_o$ (dc)

$$\therefore \frac{I_{c_1}' - I_{c_2}'}{\alpha I_o} = \frac{e^{v_d/V_T} - 1}{e^{v_d/V_T} + 1}$$

$$\boxed{V_o = -\alpha I_o R_c \cdot \frac{(e^{v_d/V_T} - 1)}{(e^{v_d/V_T} + 1)}}$$



Open loop :



Comparator (Non-linear Region)

$V_o = A_d \cdot V_i$

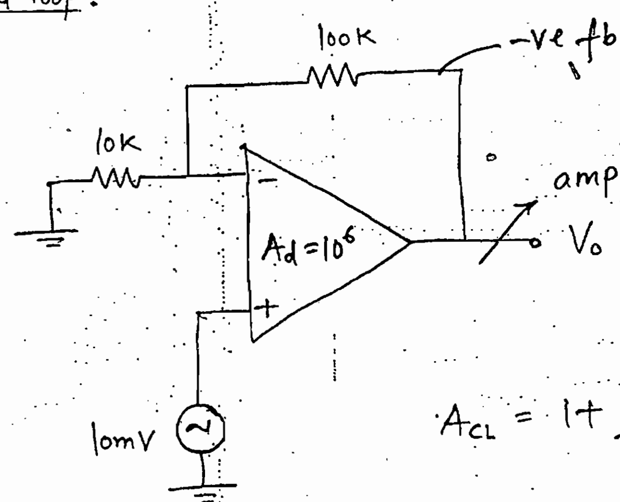
$= 10^6 \times (10 \times 10^{-3})$

$= 10^4$ volts. (×) → saturation

$= +V_{cc}$

$= 10$ volts.

Closed loop :



(Linear Region)

amplifier (Virtual ground theory)

$V_o = 10mV \times 10$

$= 0.1 V$

$A_{CL} = 1 + \frac{R_f}{R_1} = 1 + \frac{100K}{10K} \approx 10\$$

* Under Linear region, $V_d \to 0$ (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 of bypass capacitors.

(3). The i/p impedence is high. Typically 1 MΩ we can design.

Ckt configurations :

At the i/p side we have.

(1) Single i/p (2) Dual i/p

At the o/p side, we have

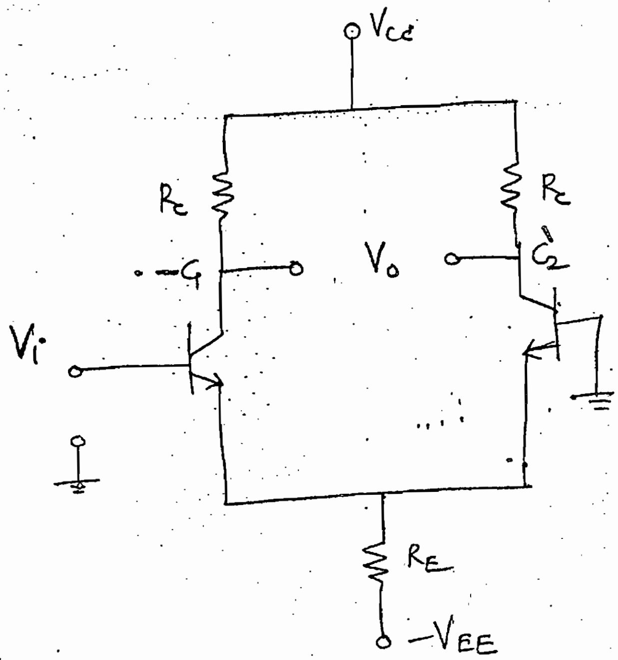
(1). Balanced o/p (o/p is taken across C_1 & C_2)

(2). Unbalanced o/p (o/p is taken across C_1 to gnd or C_2 to gnd)

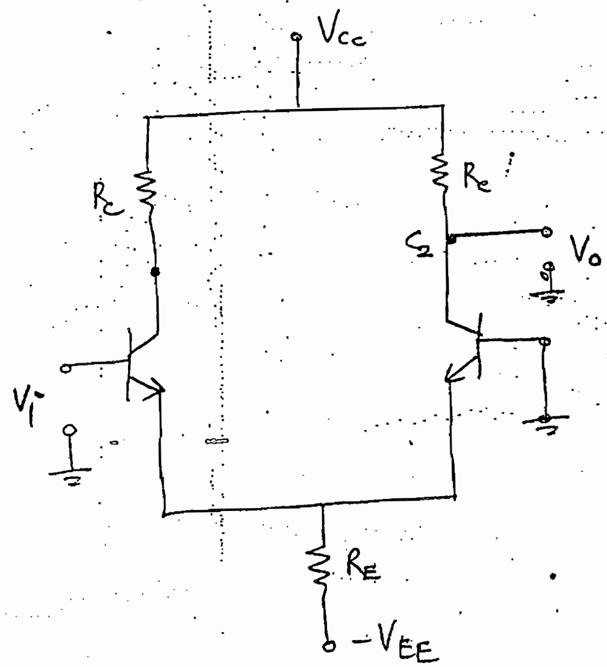
The configurations are:

(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

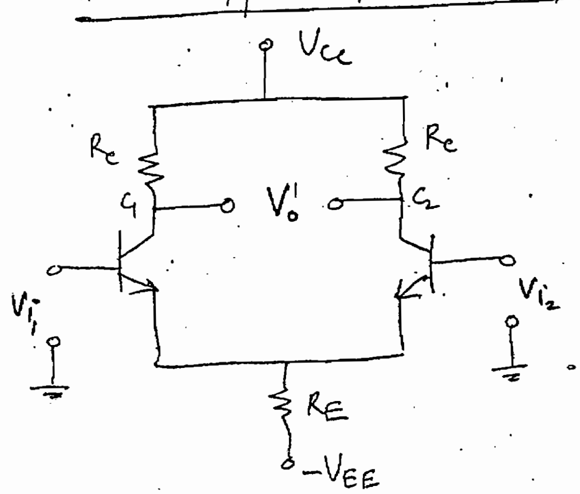
Single i/p balanced o/p:



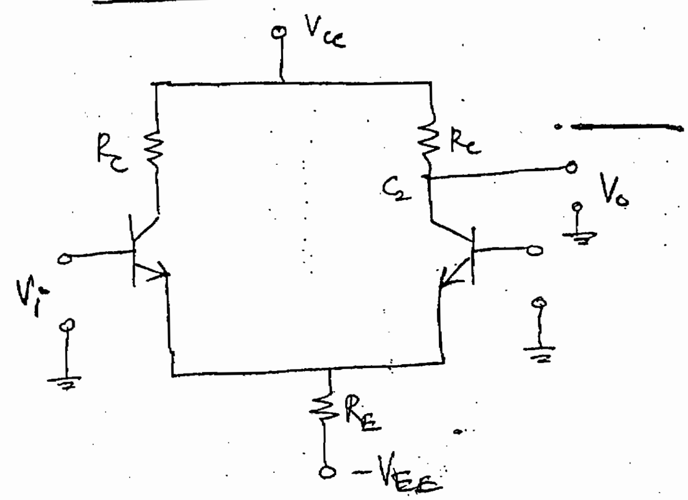
Single i/p unbalanced o/p:



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 $R_c = 1K$ and $(I_E)_Q = 26mA$.

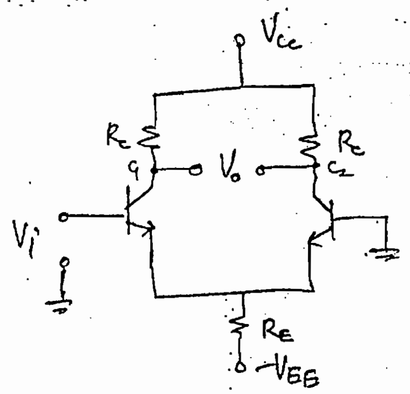
Q2. Calculate the voltage gain of single i/p unbalanced o/p or dual i/p unbalanced o/p differential amplifier with $R_c = 1K$ and $(I_E)_Q = 26mA$.

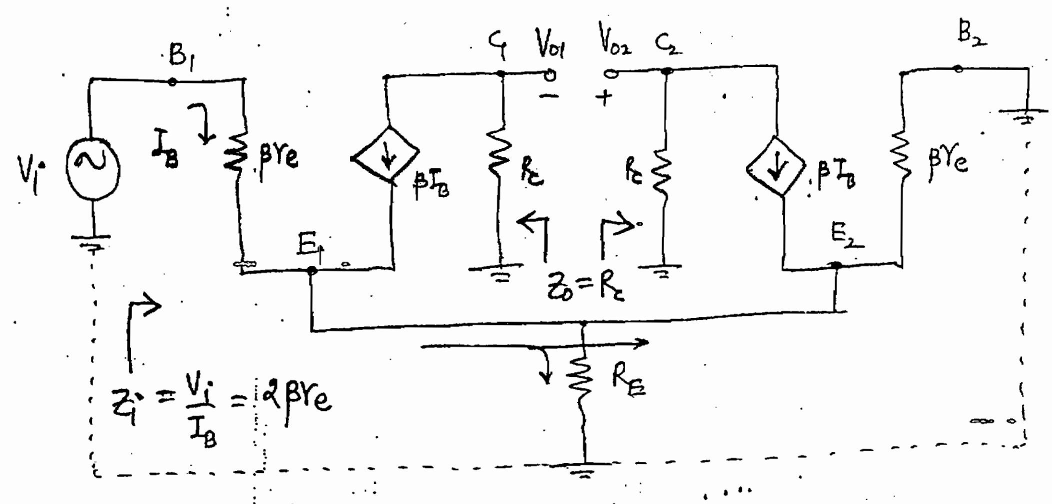
Q3. Calculate the i/p impedence of a differential amplifier ckt with $\beta = 50$ and $(I_E)_Q = 26mA$.

Q4. Calculate the o/p impedence of a diff. amplifier ckt with $R_c = 1K$.

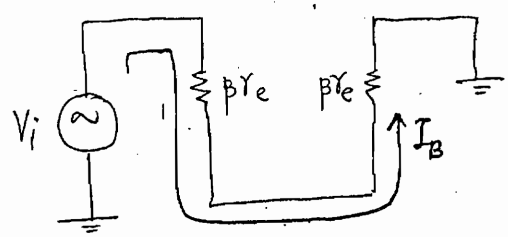
Q5. Design the operating point of a diff. amplifier ckt

AC analysis :





$$Z_i = \frac{V_i}{I_B} = 2\beta r_e$$



$Z_i = 2\beta r_e$

$\because R_E >> \beta r_e$

$$V_{o1} = -\beta I_B R_c$$

$$V_{o1} = -\beta \frac{V_i}{2\beta r_e} \cdot R_c$$

$$\frac{V_{o1}}{V_i} = \frac{-R_c}{2r_e} \rightarrow \text{unbalanced gain}$$

Similarly, $\frac{V_{o2}}{V_i} = +\frac{R_c}{2r_e}$ → unbalanced gain

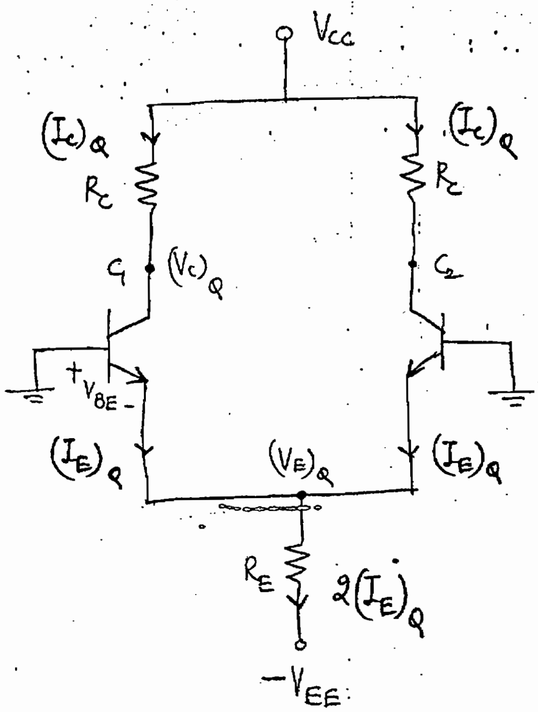
And, $A_V = A_{V_1} - A_{V_2}$

$$A_V = \frac{-R_c}{2r_e} - \frac{R_c}{2r_e}$$

$$\boxed{A_V = \frac{-R_c}{r_e}} \rightarrow \text{Balanced gain}$$

Sol (5).

DC Analysis: operating point



$$V_{EE} - V_{BE} - 2(I_E)_Q \cdot R_E = 0$$

$$\boxed{(I_E)_Q = \frac{V_{EE} - V_{BE}}{2R_E}}$$

$$(V_{CE})_Q = (V_C)_Q - (V_E)_Q$$

$$= V_{cc} - (I_c)_Q R_c + V_{BE}$$

$$\boxed{(V_{CE})_Q = V_{cc} + V_{BE} - (I_c)_Q R_c}$$

Sol.(1). **Balanced o/p:**

$$|A_d| = \left|\frac{R_c}{r_e}\right| = \frac{1K}{1\Omega} = 1000$$

$$r_e = \frac{V_T}{(I_E)_q} = \frac{26}{26} = 1\Omega$$

Sol (2). **Unbalanced o/p:**

$$|A_d| = \left|\frac{R_c}{2r_e}\right| = \frac{1K}{2} = 500.$$

Sol (3). $Z_i = 2\beta r_e$

$$= 2 \times 50 \times 1 = 100\ \Omega$$

Sol (4). $Z_o = R_C = 1\ K\Omega$

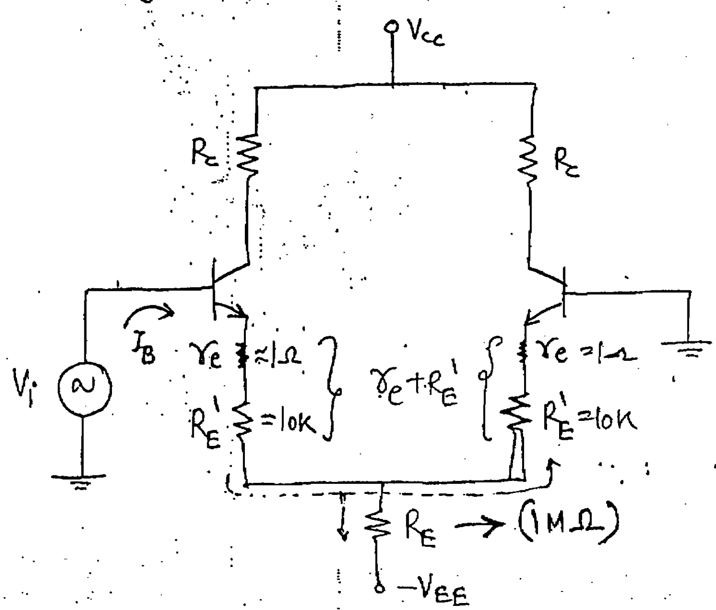
Conclusions:

(1). The gain of diff. amp is unstable.

(2). The i/p impedence is low.

To solve this problem, we use feedback.

Swamping resistor technique:



$$A_d = \frac{R_c}{r_e}$$

↓

Swamping Resistor technique

↓

$$A_d = \frac{R_c}{r_e + R_E'}$$

$$\boxed{A_d \approx \frac{R_c}{R_E'}} \rightarrow \text{stable}$$

$Z_i = 2\beta r_e$

Swamping

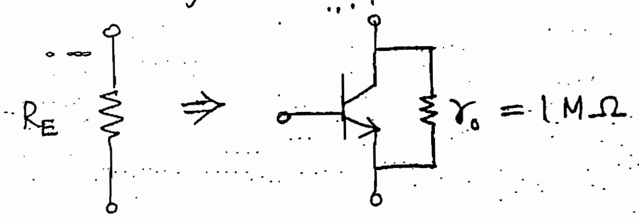
$Z_i = 2\beta(r_e + R_E')$

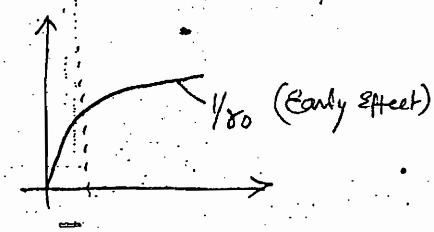
$\boxed{Z_i \approx 2\beta R_E'}$ (High)

If $\beta = 50$, $R_E' = 10\,k\Omega$ $\Rightarrow$ $Z_i = 2 \times 50 \times 10\,k\Omega = 1\,M\Omega$

* The only problem is with this, we have to keep $R_E \to 1\,M\Omega$ {i.e. to block I_B, $R_E >>> (r_e + R_E')$}.

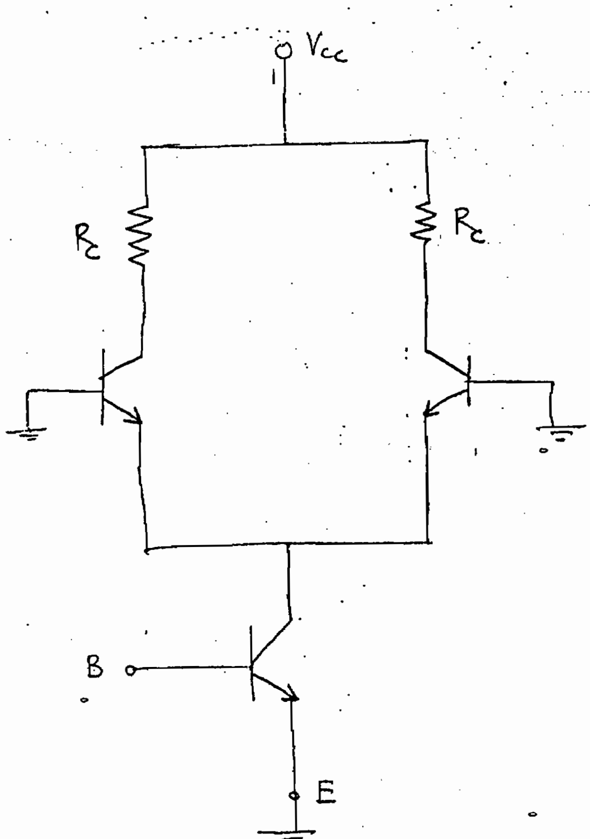
In place of R_E, put transistor to solve the above problem.



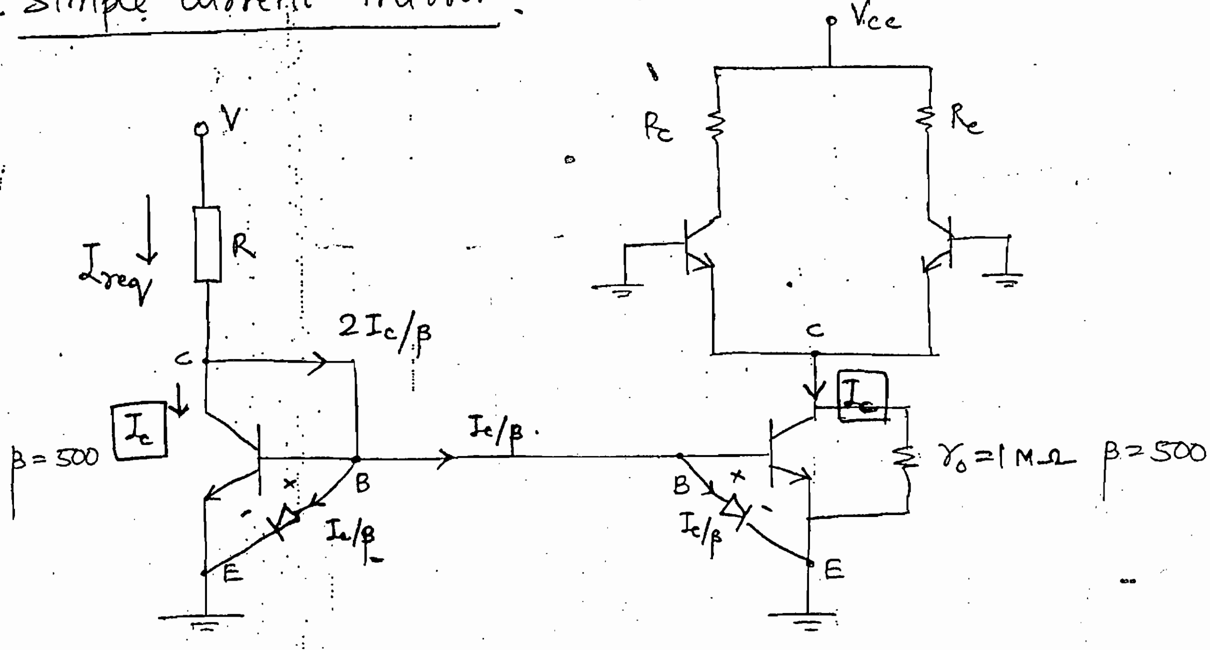


Integrated ckt Biasing :

Current mirror Biasing :

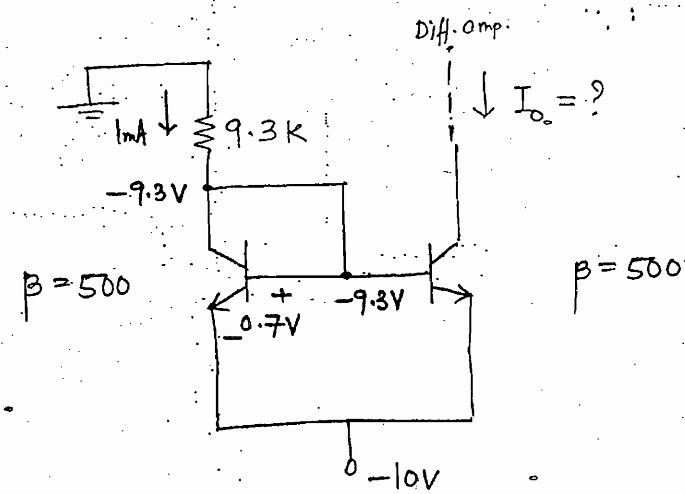


(1). Simple current mirror :

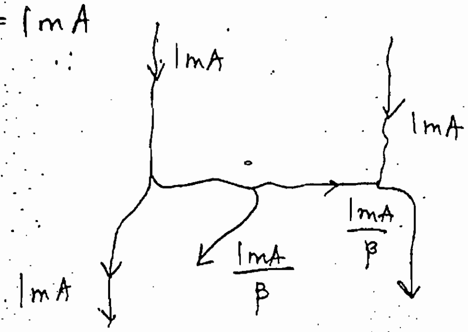


$I_B = I_c/\beta$. for reduce I_B . $\Rightarrow \beta \uparrow$. (High value)

Q.

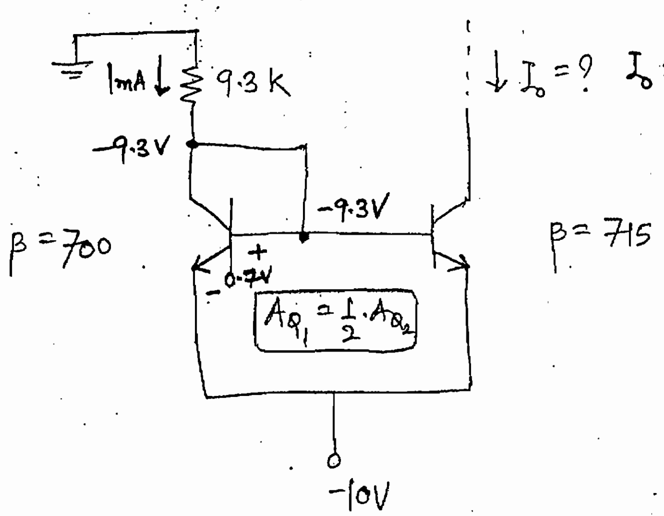


$I_o = 1\,mA$

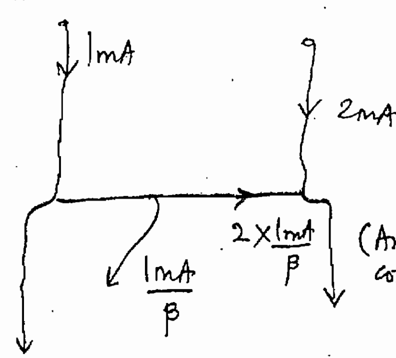


(Equal current mirror)

Q.



$I_o \approx 2mA$

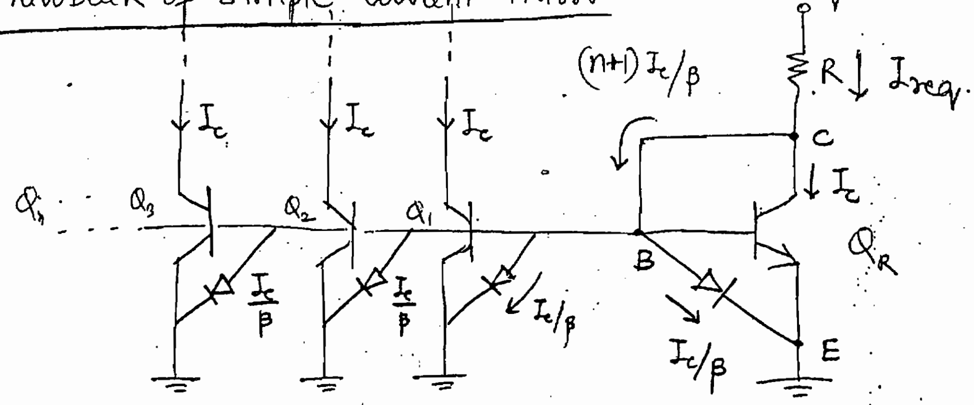


(Area concept)

(Unequal current mirror)

* β are equal (theoretically)
* β are slightly different (practically) → don't bother (approx ans) ✓

Drawback of simple current mirror



$$I_{req.} = I_c + (n+1)\frac{I_c}{\beta}$$

$$I_{req.} = I_c\left[1 + \frac{(n+1)}{\beta}\right].$$

Eg. If $n = 99$, then & $\beta = 500$.

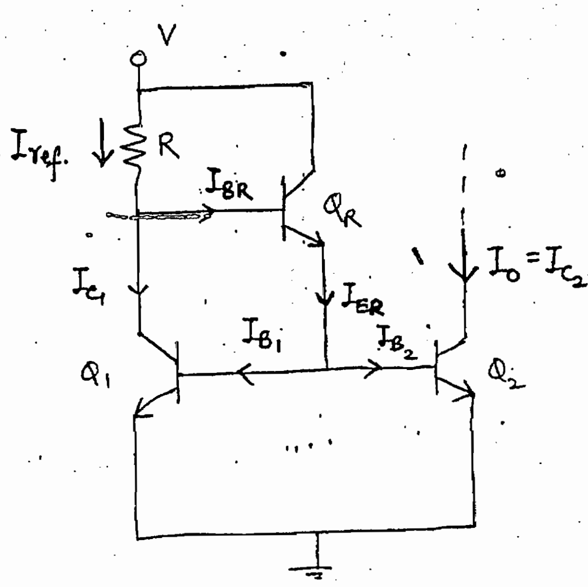
$$I_{req} = I_c\left(1 + \frac{100}{500}\right) = \frac{6}{5} I_c$$

$$I_{req} \neq I_c \quad \text{(accuracy is lost)}.$$

Hence, $n \leq 10$ transistors in the biasing for better accuracy.

To improve accuracy, there is modification required.

(2). Modified current mirror:



$$I_{ref} = I_{BR} + I_{C_1}$$

$$= I_{C_1} + I_{ER}/(1+\beta)$$

$$= I_{C_1} + \frac{I_{B_1} + I_{B_2}}{(1+\beta)}$$

Acc. to current mirror,

$$I_{C_1} = I_{C_2} \quad \& \quad I_{B_1} = I_{B_2}$$

Now, $I_{ref} = I_{C_2} + \dfrac{2 I_{B_2}}{(1+\beta)}$

$$= I_{C_2} + \frac{2 I_{C_2}}{\beta(1+\beta)}.$$

$$I_{ref} = I_{C_2} + \frac{2 I_{C_2}}{\beta + \beta^2}$$

$$I_{ref} \approx I_{C_2}\left(1 + \frac{2}{\beta^2}\right)$$

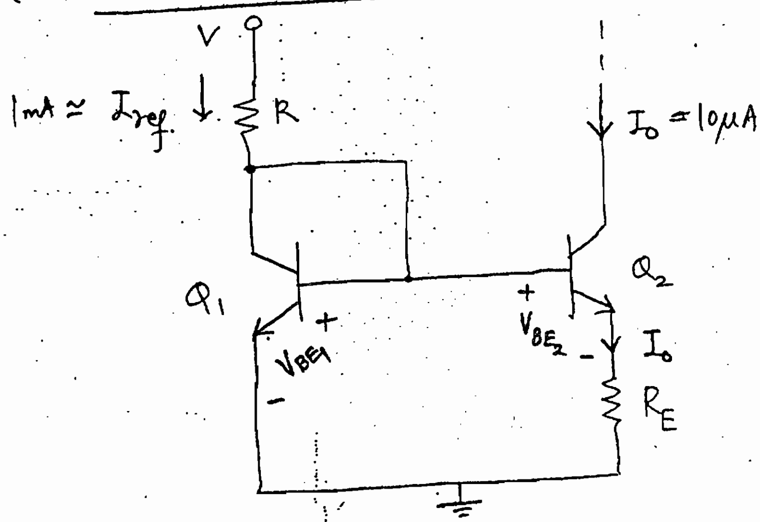
$$\therefore I_o = I_{C_2} = \frac{I_{ref}}{\left(1 + \frac{2}{\beta^2}\right)}$$

for n transistors,

$$I_{C_i} = \frac{I_{ref}}{1 + \frac{(n+1)}{\beta^2}}$$

Here, accuracy is much improved.

(3). micro level current mirror (mA to μA mirror) :



$$V_{BE_1} = V_{BE_2} + I_o R_E$$

$$I_o R_E = V_{BE_1} - V_{BE_2}$$

$$V_{BE_1} = V_T \ln\left(\frac{I_{ref}}{I_s}\right)$$

$$V_{BE_2} = V_T \ln\left(\frac{I_o}{I_s}\right)$$

$$\therefore I_o R_E = V_T \ln\left(\frac{I_{ref}}{I_o}\right)$$

Eg. $10 \times 10^{-6} R_E = 26\,mV \cdot \ln\left(\frac{1mA}{10 \times 10^{-6} A}\right)$

(designing R_E) → $R_E = \frac{26 \times 10^{-3}}{10^{-7}} \ln\left(\frac{10^{-3}}{10^{-5}}\right) = 11\,K\Omega$

⇓

Widlar current source

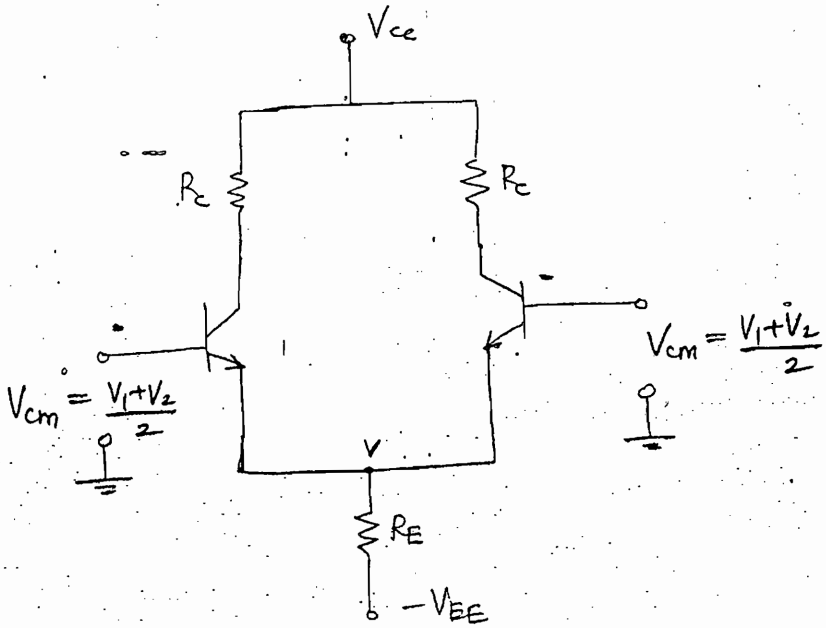
↓

scientist.

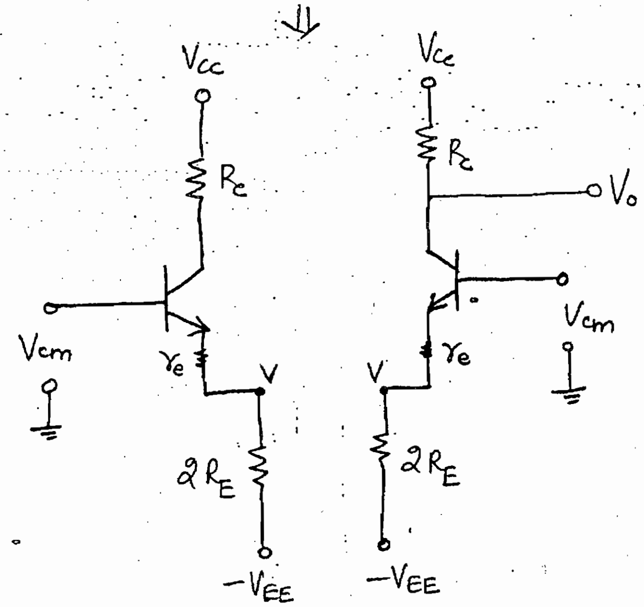
CMRR :

$$CMRR = \frac{A_d}{A_c} \Rightarrow (CMRR)_{dB} = 20\log\left(\frac{A_d}{A_c}\right)$$

Common mode gain (A_c):



⇓



$$A_{cm} = \frac{V_o}{V_{cm}} = \frac{-R_c}{r_e + 2R_E}$$

$\Rightarrow 2R_E >>> r_e$

$\therefore \boxed{A_{cm} = \frac{-R_c}{2R_E}}$ → Unbalanced

$\boxed{A_d = \frac{-R_c}{2r_e}}$ → unbalanced

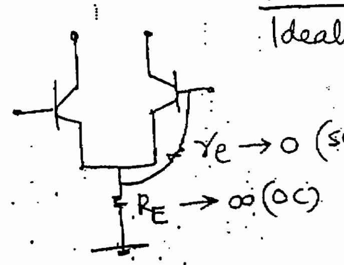
$$CMRR = A_d / A_c$$

$$CMRR = \frac{-R_c/2r_e}{-R_c/2R_E}$$

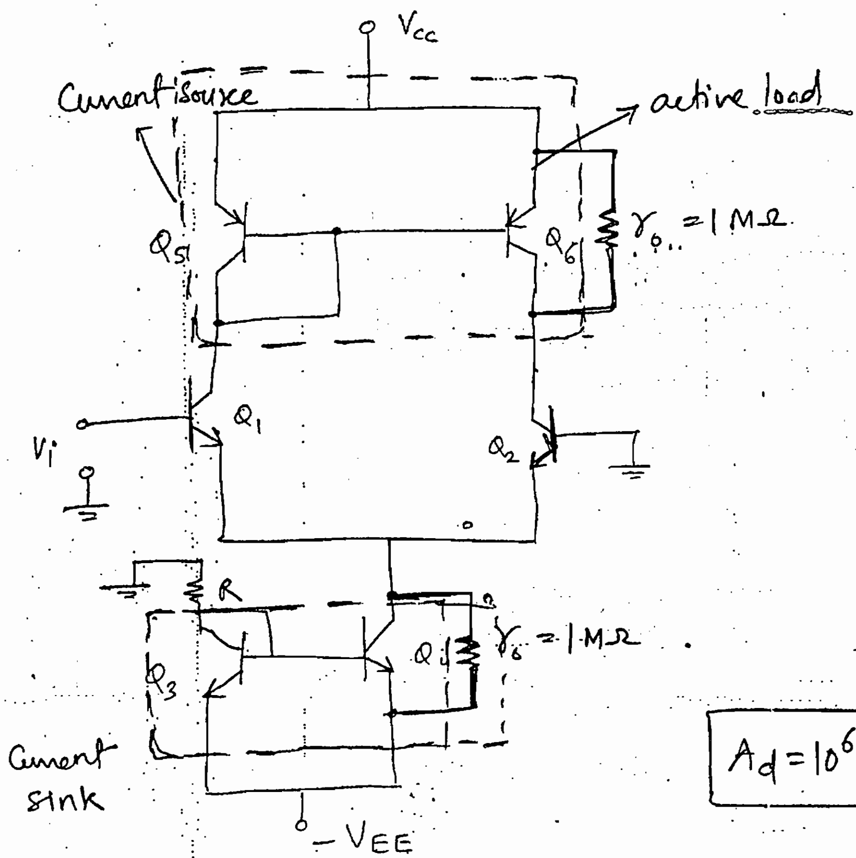
$$\boxed{CMRR = R_E/r_e} \boxed{= g_m R_E}$$

For A_c –

Ideally ($CMRR = \infty$)



Integrated IC differential amplifier:



$\boxed{A_d = 10^6}$ → Balanced gain

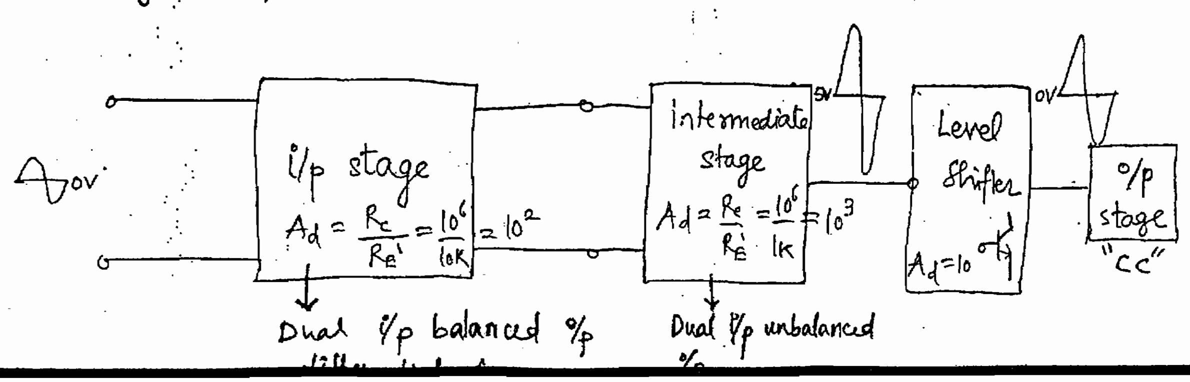
$$A_d = \frac{R_c}{2r_e} = \frac{r_o}{2r_e} = \frac{1\,M\Omega}{2 \times 1\,\Omega} = 0.5 \times 10^6$$

$$A_c = \frac{R_c}{2R_E} = \frac{r_o}{2r_o} = 0.5$$

$$CMRR = \frac{A_d}{A_c} = \frac{0.5 \times 10^6}{0.5} = 10^6$$

Operational amplifier:

Block diagram of Op-Amp:



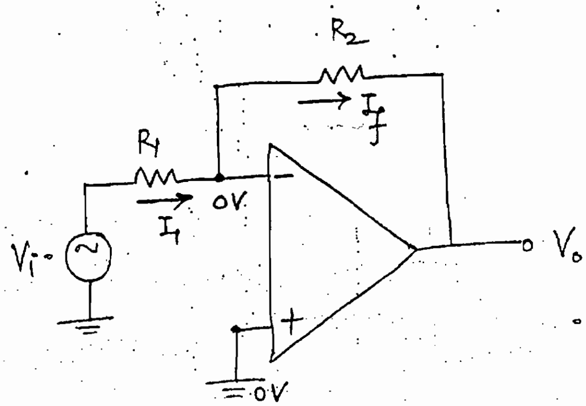
Applications of Op-Amp : Linear applications :

(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 :

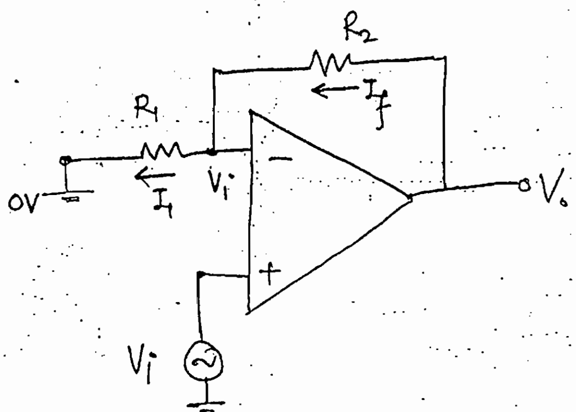


$$I_1 = I_f$$

$$\frac{V_i - 0}{R_1} = \frac{0 - V_o}{R_2}$$

$$\frac{V_o}{V_i} = -\frac{R_2}{R_1}$$

(2). Non-inverting amplifier :

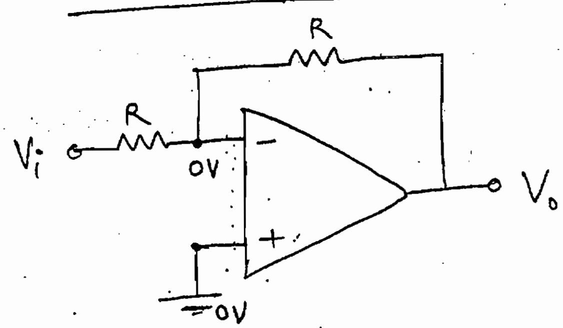


$$I_1 = I_f$$

$$\frac{V_i - 0}{R_1} = \frac{V_o - V_i}{R_2}$$

$$\frac{V_o}{V_i} = 1 + \frac{R_2}{R_1}$$

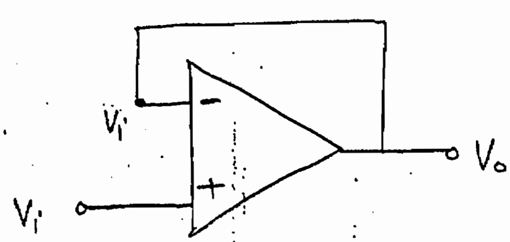
(3). Phase shifter :



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

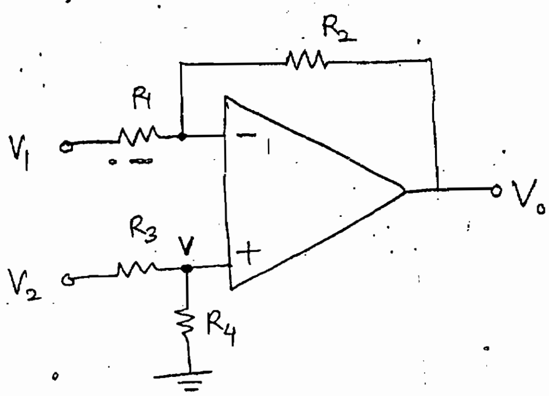
$$V_o = -V_i\,!$$

(4). Voltage follower :



$V_o = V_i$

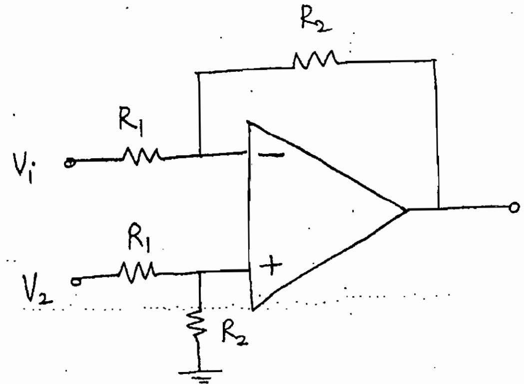
(5). Differential amplifier :



By superposition,

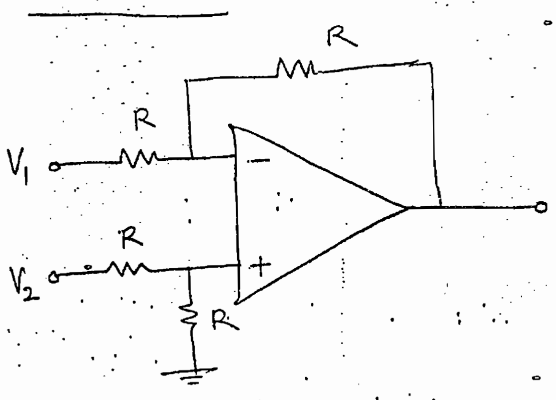
$$V_o = -\frac{R_2}{R_1} \cdot V_1 + \left(1 + \frac{R_2}{R_1}\right) V$$

where , $V = \frac{V_2 \cdot R_4}{(R_3 + R_4)}$



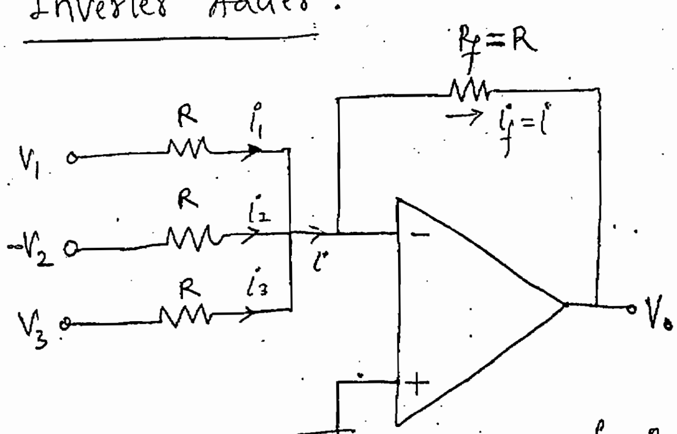
$$V_o = \frac{R_2}{R_1} \cdot (V_2 - V_1)$$

(6). Subtractor :



$$V_o = V_2 - V_1$$

(7). Inverter Adder :



$$i_f = i$$

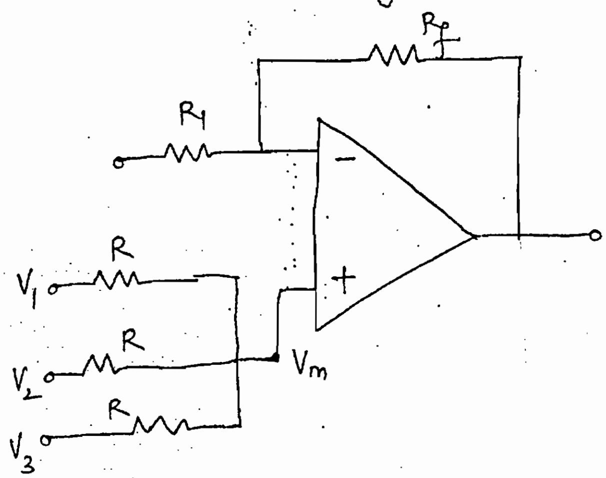
$$i_1 + i_2 + i_3 = \frac{0 - V_o}{R_f}$$

$$\frac{V_1 + V_2 + V_3}{R} = \frac{-V_o}{R_f}$$

$$V_o = \frac{-R_f}{R}(V_1 + V_2 + V_3)$$

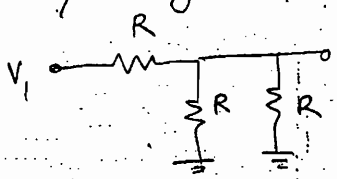
for $R_f = R$; $V_o = -(V_1 + V_2 + V_3)$

(8). Non-inverting Adder :

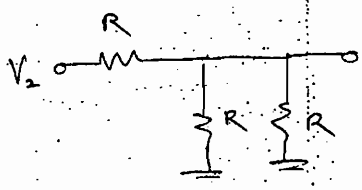


$$V_o = \left(1+\frac{R_f}{R_1}\right) \cdot V_m$$

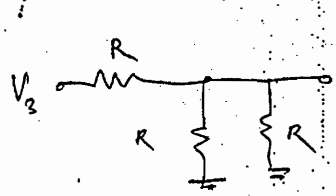
By using superposition,



$$V_{m_1} = \frac{R/2}{R+R/2} \cdot V_1$$



$$V_{m_2} = \frac{R/2}{R+R/2} \cdot V_2$$



$$V_{m_3} = \frac{R/2}{R+R/2} \cdot V_3$$

$$\therefore V_m = V_{m_1} + V_{m_2} + V_3$$

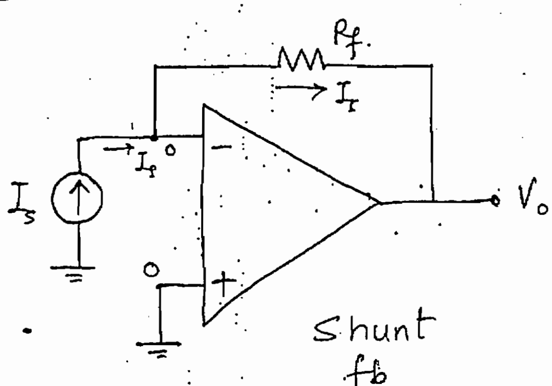
$$= \frac{R/2}{R+R/2} \cdot (V_1+V_2+V_3)$$

$$= \frac{R}{3R} (V_1+V_2+V_3)$$

$$= \frac{(V_1+V_2+V_3)}{3}$$

$$V_o = \left(1+\frac{R_f}{R_1}\right)\left(\frac{V_1+V_2+V_3}{3}\right)$$

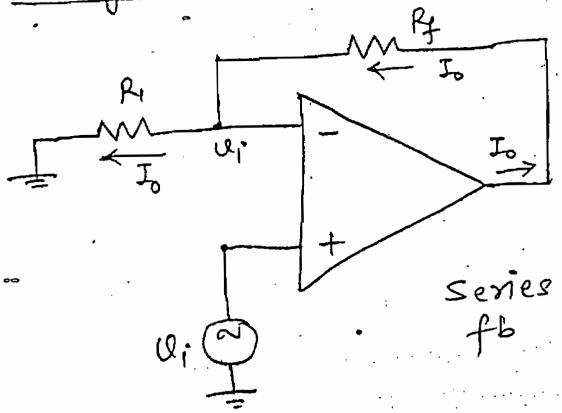
(9) Current to voltage converter : "CCVS"



$$0 - V_o = I_s \cdot R_f$$

$$V_o = -I_s R_f$$

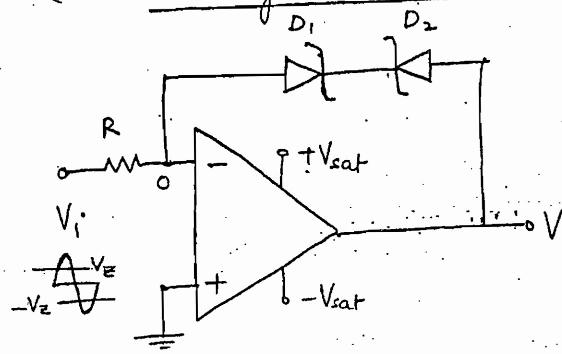
(10). Voltage to current converter : "VCCS."

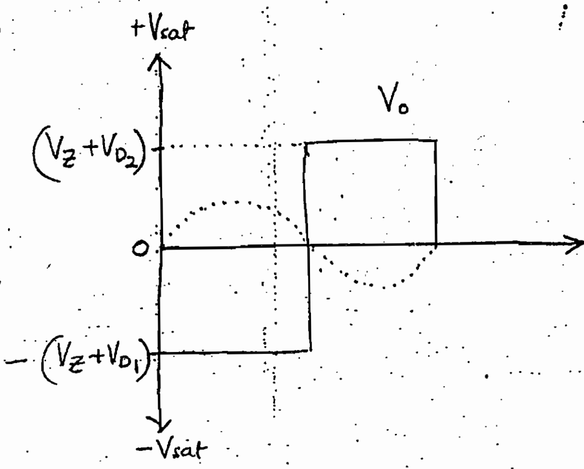


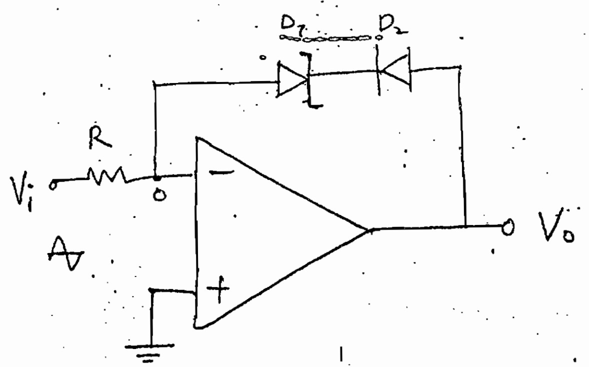
$V_i = I_o R_1$

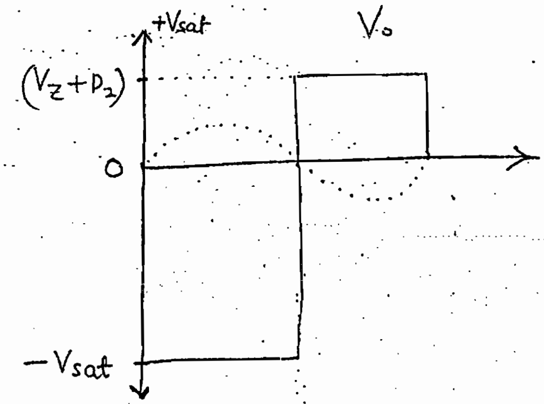
$I_o = \frac{V_i}{R_1}$

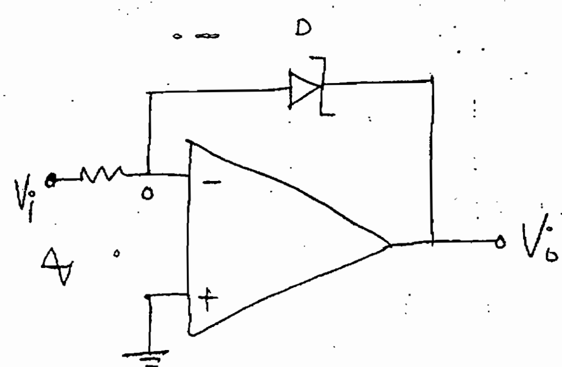
*(11). Voltage limiter :

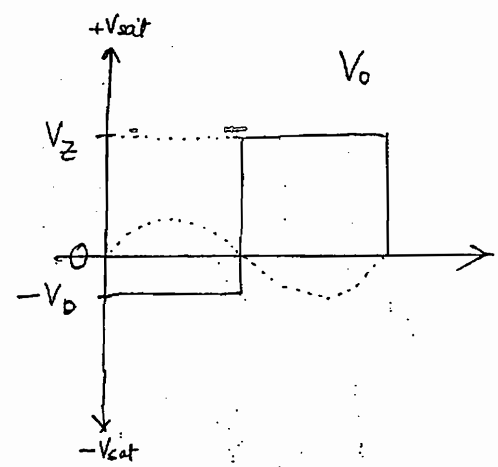




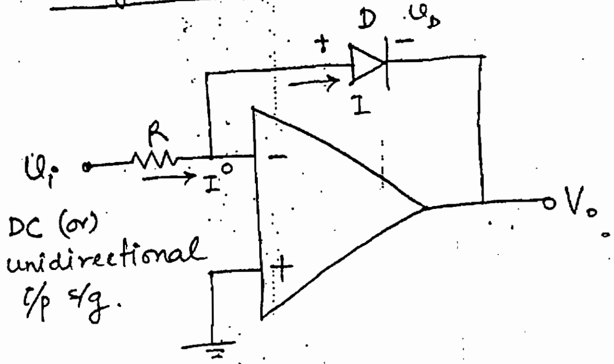








(12). Logarthmic amplifier: (Data compression)



$$0 - V_o = \vartheta_D$$

$$V_o = -\vartheta_D$$

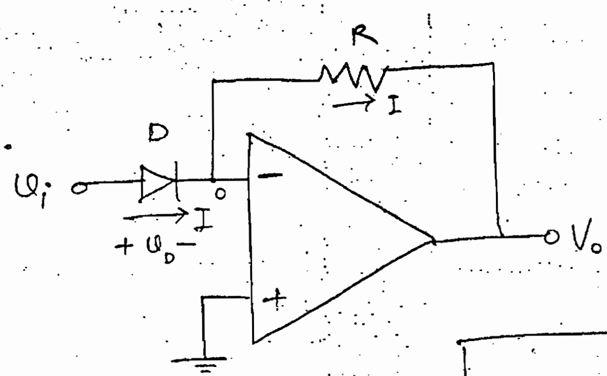
$$I = I_o e^{\vartheta_D / V_T}$$

$$\vartheta_D = V_T \cdot \ln\left(\frac{I}{I_o}\right)$$

$$V_o = -V_T \ln\left(\frac{\vartheta_i}{I_o R}\right)$$

$$\therefore \boxed{V_o \propto \ln \vartheta_i}$$

(13). Antilog amplifier: (Data expansion)



$$I = I_o \cdot e^{\vartheta_D / V_T}$$

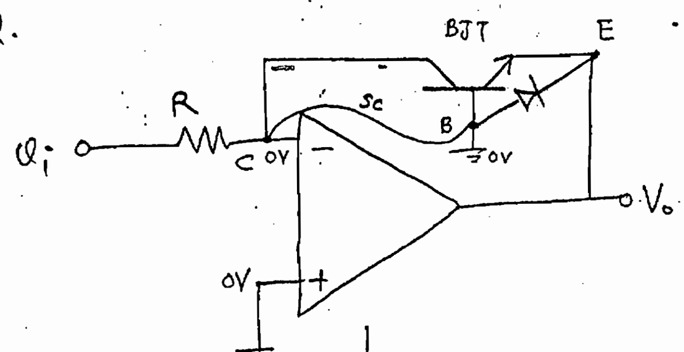
$$\frac{0 - V_o}{R} = I_o\, e^{\vartheta_i / V_T}$$

$$V_o = -I_o R \cdot e^{\vartheta_i / \vartheta_T}$$

$$\vartheta_i = \vartheta_D$$

$$\therefore \boxed{V_o \propto e^{\vartheta_i}}$$

Q.

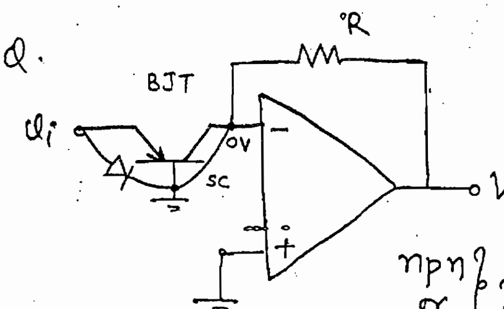


↓

Logarthmic amplifier

(Replacement of diode with BJT)

Q.



npn or pnp ?

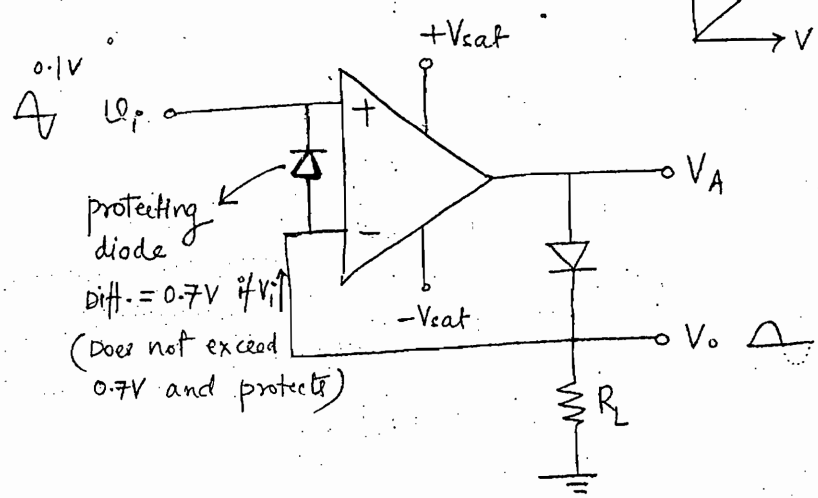
↓

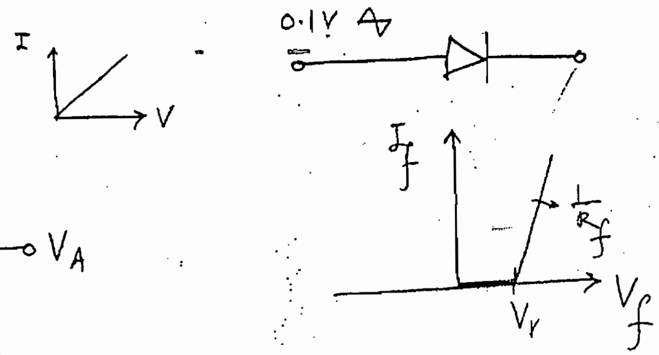
Antilog amplifier

(Replacement of diode with BJT)

BJT must be "pnp".

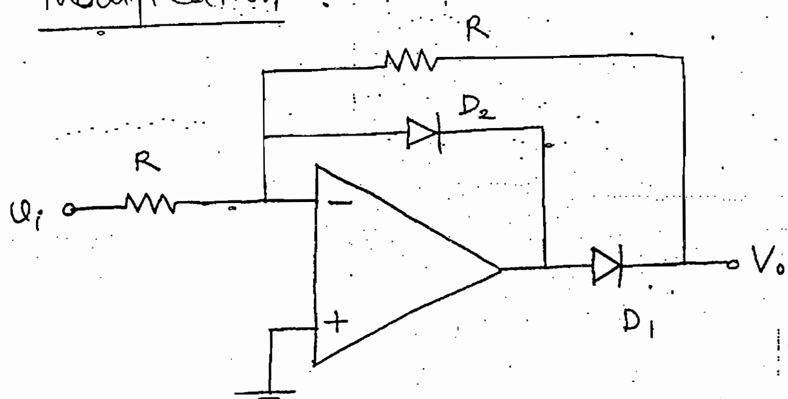
(14). Precision rectifiers:

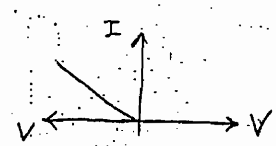




+ve Half cycle (CL)	−ve Half cycle (OL)
$V_A = +V_{sat}$	$V_A = -V_{sat}$
$D \rightarrow$ ON	$D \rightarrow$ OFF
$V_o = V_i$ (Linear)	$V_o = 0$ (Non-linear)

Modification :





+ve Half cycle

$D_2 \rightarrow$ ON , $D_1 \rightarrow$ OFF

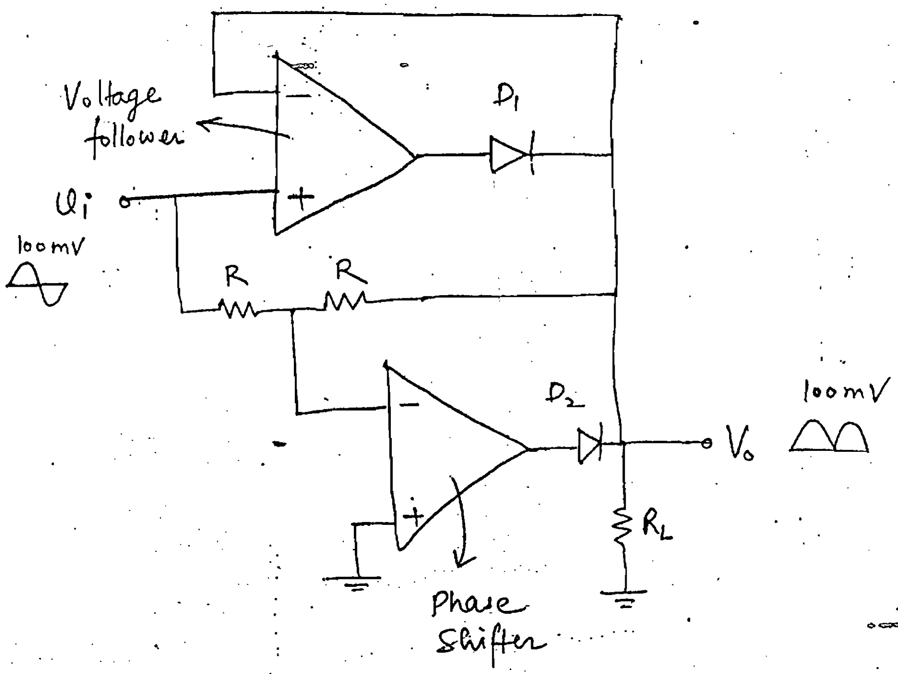
$V_o = 0$

−ve Half cycle

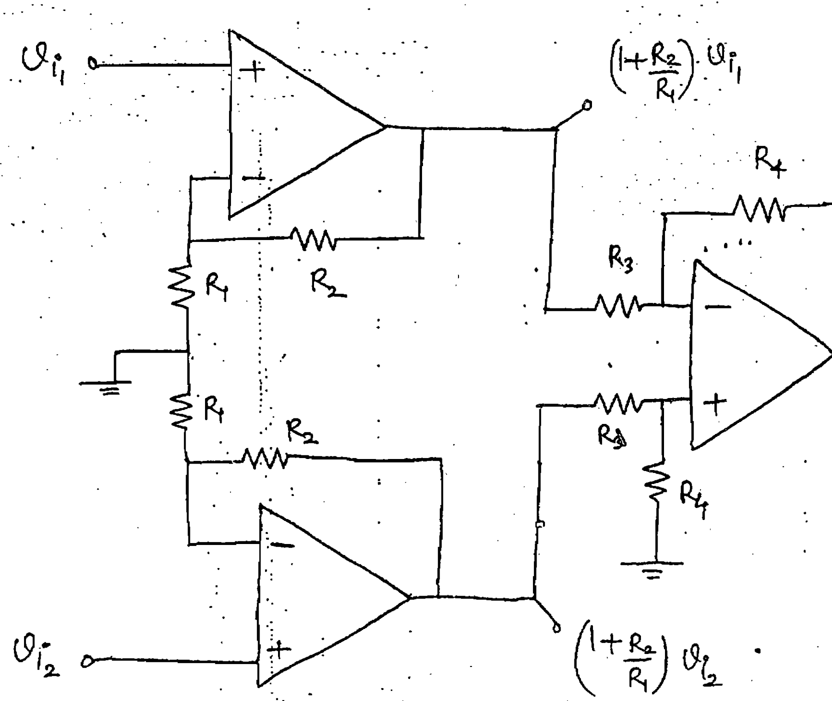
$D_2 \rightarrow$ OFF , $D_1 \rightarrow$ ON

$V_o = -V_i$ (Phase shifter)

Full wave rectifier:



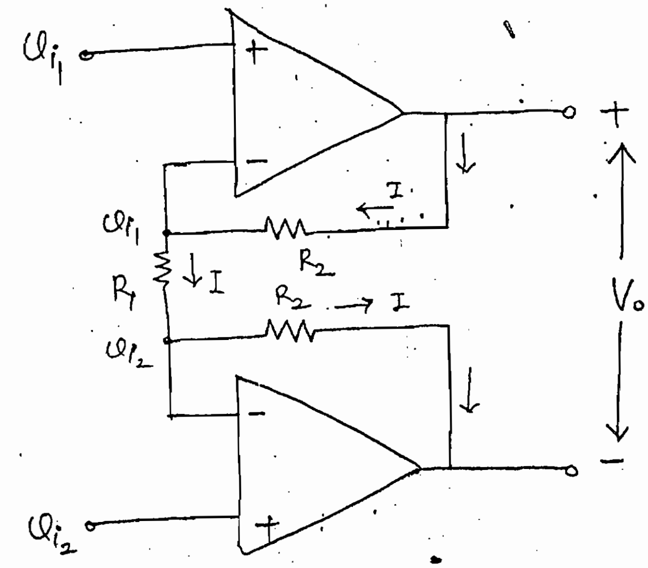
(15) Instrumentational amplifier:



$$V_o = \frac{R_4}{R_3}\left(1+\frac{R_2}{R_1}\right)\cdot(v_{i_2}-v_{i_1})$$

$$A_d = \frac{v_o}{v_d} = \frac{R_4}{R_3}\left(1+\frac{R_2}{R_1}\right)$$

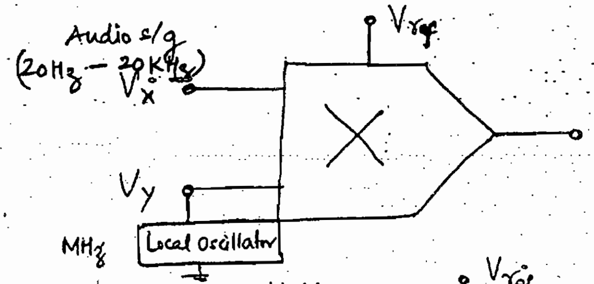
Modified ckt : (Better)



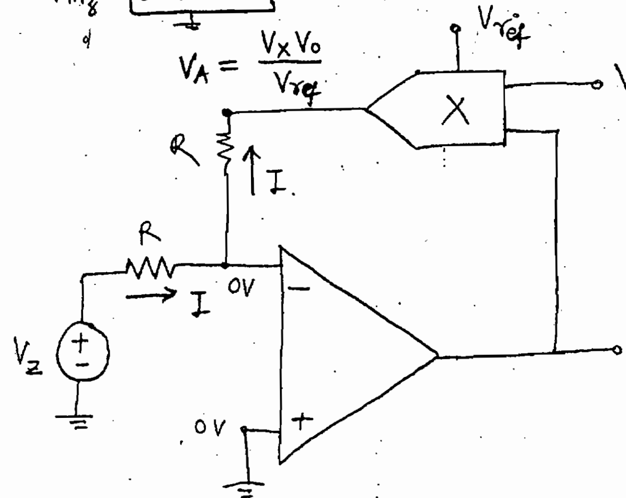
$$V_o = \left(1 + \frac{2R_2}{R_1}\right)\cdot(V_{i_1} - V_{i_2})$$

$$A_d = \frac{V_o}{V_d} = \frac{V_o}{(V_{i_1} - V_{r_2})} = \frac{I(R_1 + 2R_2)}{I(R_1)} = 1 + \frac{2R_2}{R_1} \quad \text{(High)}$$

(16). Modulator : (Analog Multiplier)



$$V_o = \frac{V_x V_y}{V_{ref}}$$



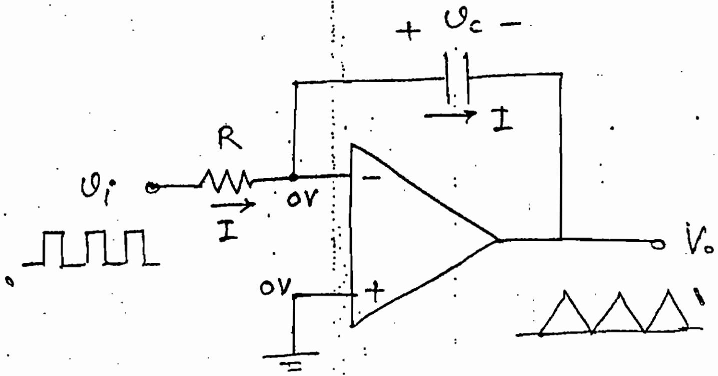
$$V_o = -V_{ref}\cdot\left(\frac{V_z}{V_x}\right)$$

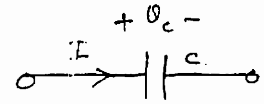
$$V_z = -V_A$$

$$V_z = \frac{-V_x \cdot V_o}{V_{ref}}$$

(17) Demodulator : (Analog Divider)

(18). Integrator : (LPF)





$$v_c = \frac{1}{c}\int_{-\infty} i\,dt$$

$$V_o = -v_c$$

$$= -\frac{1}{c}\int I\,dt$$

$$V_o = -\frac{1}{RC}\int v_i\,dt$$

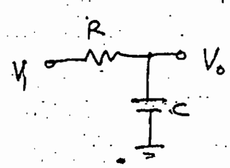
Transfer function :

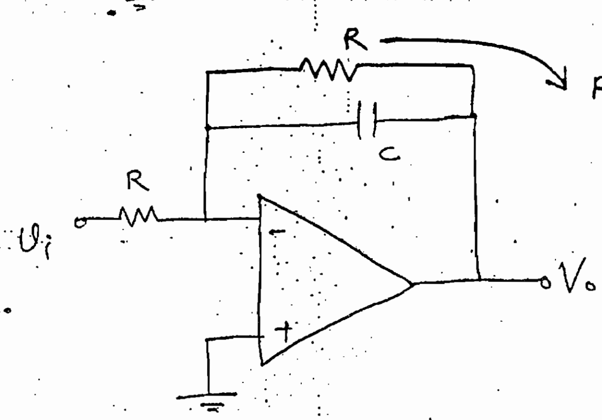
$$TF = \frac{V_o(s)}{V_i(s)} = \frac{-1}{SCR} \Rightarrow \frac{1}{SCR} = \frac{1}{j\omega CR} = \frac{-j}{\omega CR}$$

phase shift

phase = -90°

(Phase lag design)





R is meant for DC conditions.

(19). Differentiator :

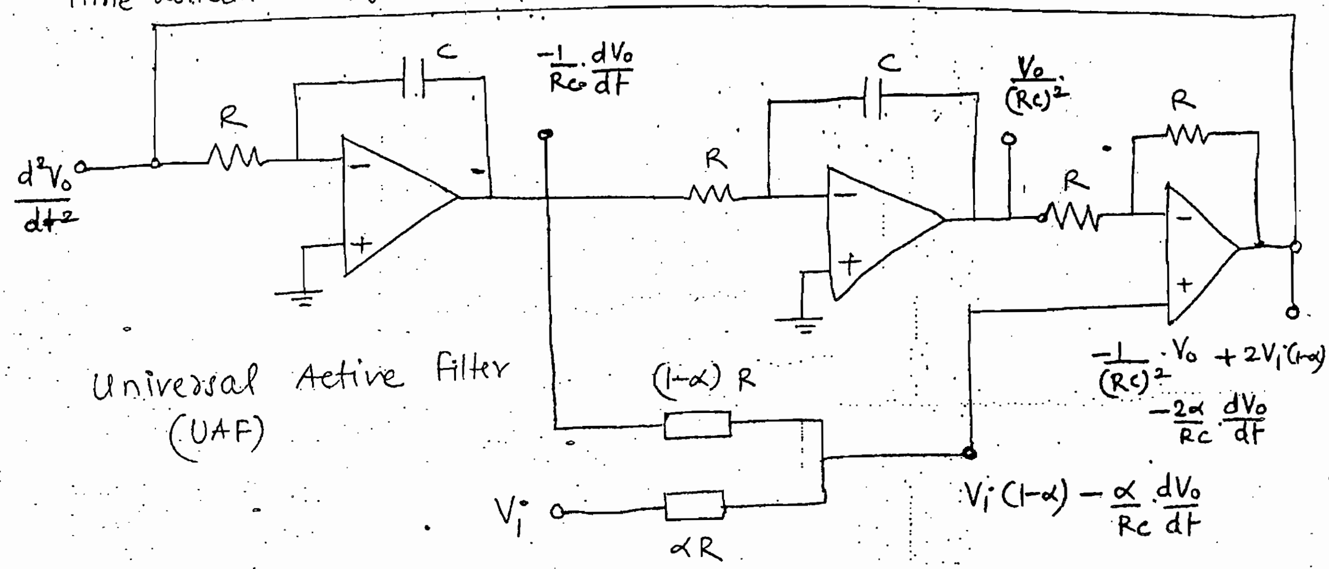
(20). Active filters :

Simulate a 2^{nd} order differential eqn using op-amps.

$$\frac{d^2V_o}{dt^2} + \alpha\frac{dV_o}{dt} + \gamma V_o = V_i$$

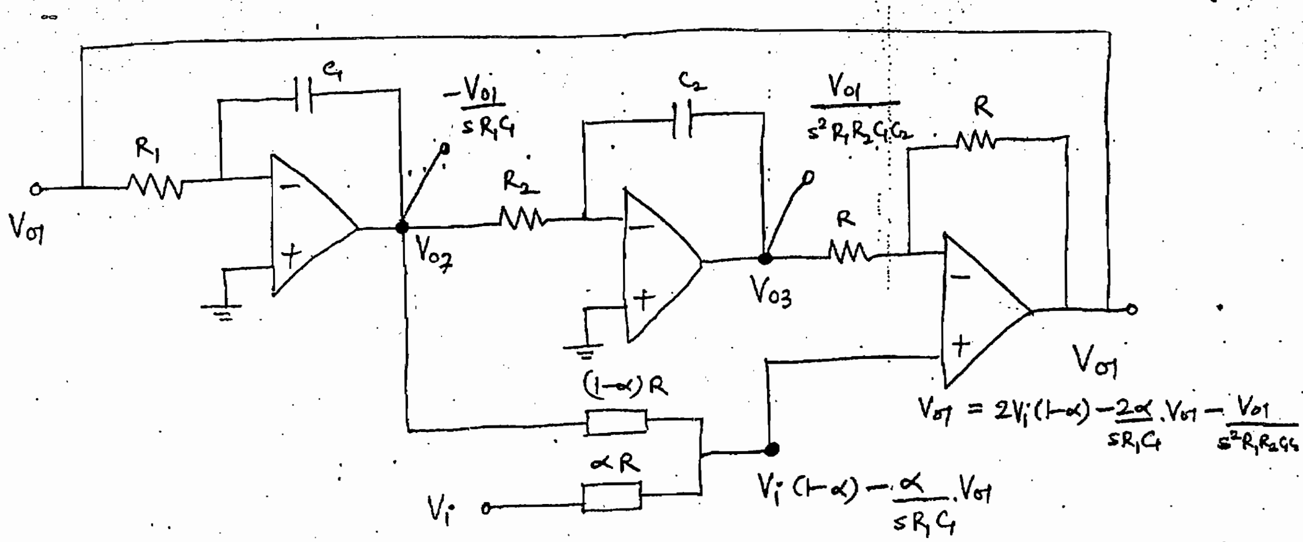
$$\frac{d^2V_o}{dt^2} = V_i - \alpha\frac{dV_o}{dt} - \gamma V_o$$

Time domain analysis



Universal Active Filter (UAF)

Frequency domain analysis :



$$V_{o1}\left(1 + \frac{2\alpha}{sR_1C_1} + \frac{1}{s^2R_1R_2C_1C_2}\right) = 2V_i(1-\alpha)$$

$$\frac{V_{o1}}{V_i} = \frac{2(1-\alpha)\cdot s^2(R_1R_2C_1C_2)}{s^2R_1R_2C_1C_2 + 2\alpha sR_2C_2 + 1} \quad \text{(HPF)}$$

$$\frac{V_{o2}}{V_i} = \frac{-2(1-\alpha)\cdot sR_2C_2}{s^2R_1R_2C_1C_2 + 2\alpha sR_2C_2 + 1} \quad \text{(BPF)}$$

$$\frac{V_{o3}}{V_i} = \frac{2(1-\alpha)}{s^2R_1R_2C_1C_2 + 2\alpha sR_2C_2 + 1} \quad \text{(LPF)}$$

LP + HP ⇒ BRF (Notch filter)

$$\frac{V_{o4}}{V_i} = \frac{2(1-\alpha)(s^2R_1R_2C_1C_2+1)}{s^2R_1R_2C_1C_2 + 2\alpha sR_2C_2 + 1}$$

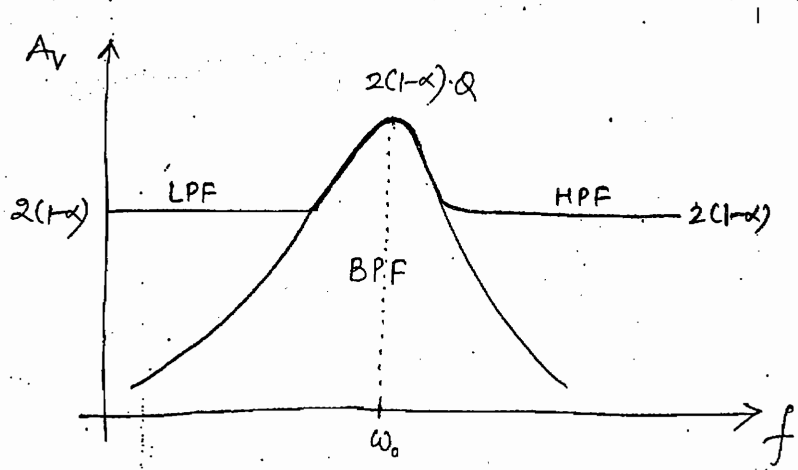
<u>Band pass</u>:

$$\left|\frac{V_{o2}}{V_i}\right| = \frac{2(1-\alpha)\cdot sR_2C_2}{s^2R_1R_2C_1C_2 + 2\alpha sR_2C_2 + 1}$$

Resonant frequency, $\omega_0 = \dfrac{1}{\sqrt{R_1 R_2 C_1 C_2}}$

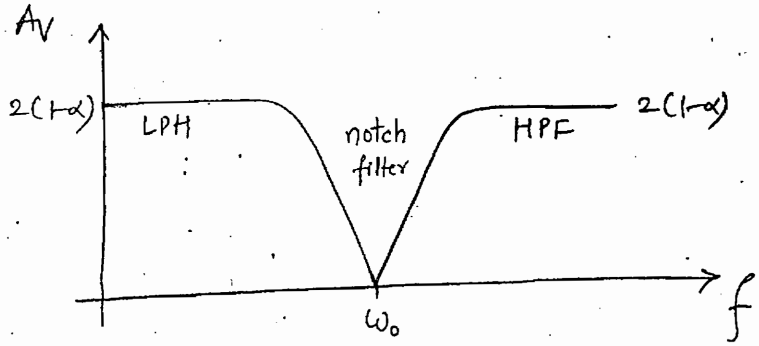
At ω_0, gain is highest,

$$\left|\frac{V_{o2}}{V_i}\right| = \frac{2(1-\alpha)}{2\alpha} = 2(1-\alpha)\cdot Q$$



<u>Band Reject filter</u>:

$$\frac{V_{o4}}{V_i} = \frac{2(1-\alpha)(s^2R_1R_2C_1C_2+1)}{s^2R_1R_2C_1C_2 + 2\alpha sR_2C_2 + 1}$$



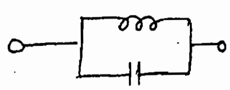
$$\left|\frac{V_{o4}}{V_i}\right| = \frac{2(1-\alpha)\, s R_1 C_1}{\underbrace{2\alpha}_{\text{zero}}}$$

$$= 2(1-\alpha)\, s R_1 C_1 \cdot Q$$

* LC filters are best due to high value Q 1000's multiple
  RC filters have Q value in the range of 10's multiple.

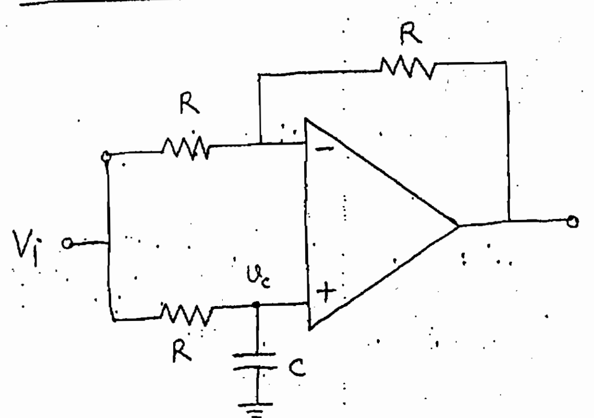
For loads :

(1) R_L load $\rightarrow$ A_V may be 100's and 1000's.

(2)  $\rightarrow$ $A_V = \dfrac{A_I \cdot Z_L}{Z_i}$ at ω_0, $Z_L = \infty$

$\therefore A_V = \infty$

All pass filter : (Phase corrector)



By superposition,

$$V_o = \frac{-R}{R} V_i + \left(1 + \frac{R}{R}\right) V_c$$

$$= -V_i + 2V_c$$

$$= -V_i + 2 \cdot V_i \left(\frac{-jX_c}{R - jX_c}\right)$$

$$= V_i \left[-1 + \frac{2}{1 + j2\pi f RC}\right]$$

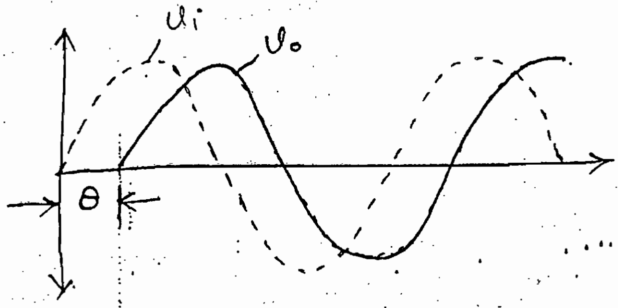
$$= V_i \cdot \left(\frac{1 - j2\pi f RC}{1 + j2\pi f RC}\right)$$

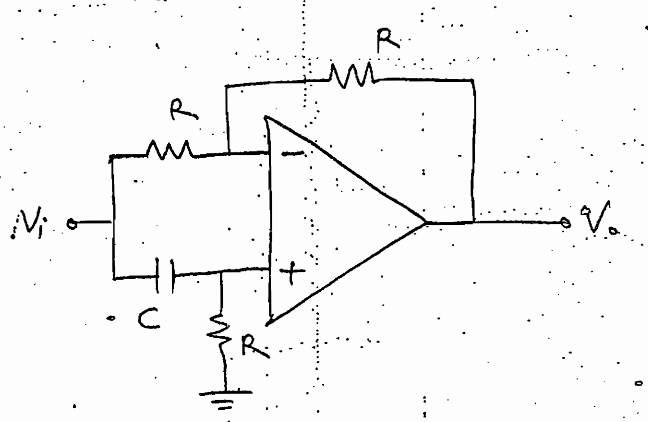
$$\frac{V_o}{V_i} = \frac{1 - j2\pi f RC}{1 + j2\pi f RC}$$

$$\frac{V_o}{V_i} = \left|\frac{V_o}{V_i}\right| \angle\theta = 1 \cdot \angle -2\tan^{-1}(2\pi f RC)$$

phase lag.

$$\therefore \theta = -2\tan^{-1}(2\pi f RC)$$





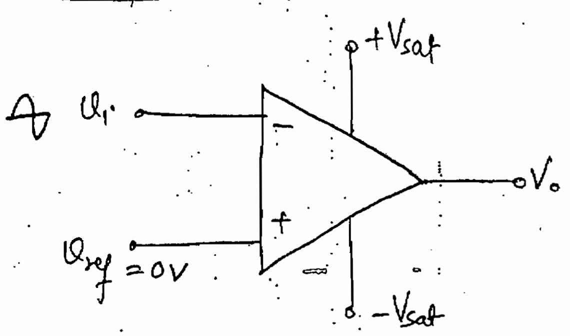
$$\theta = +2\tan^{-1}(2\pi f RC)$$

1^{st} order APF $\rightarrow$ can correct $\pm 180^\circ$.

II^{nd} order APF $\rightarrow$ can correct $\pm 360^\circ$.

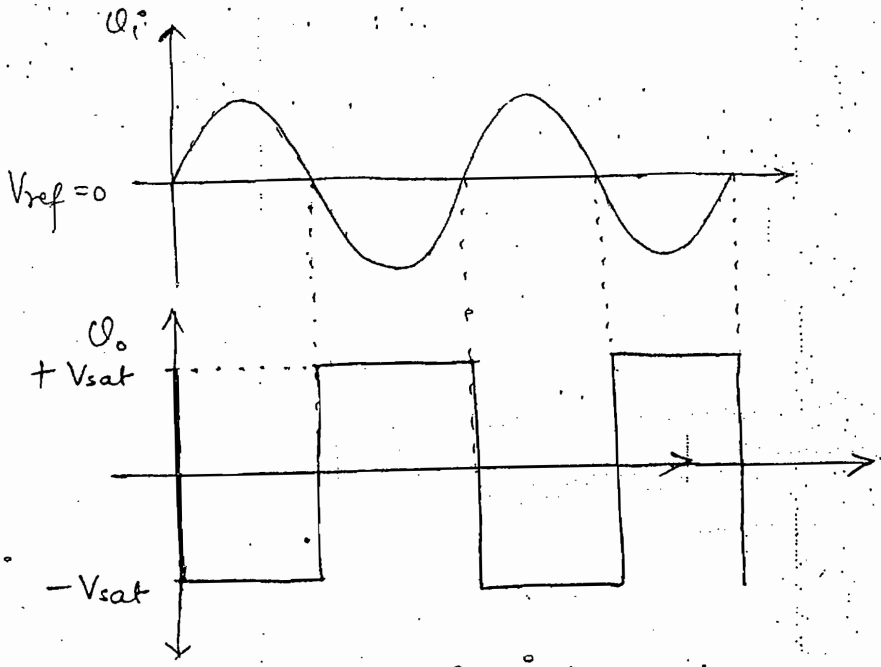
Non-linear applications:

Comparator: Zero crossing detector. (ZCD)

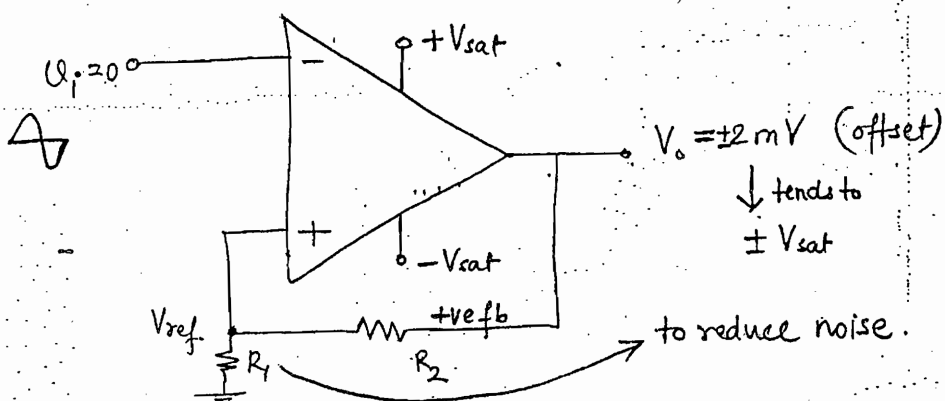


$v_i > 0 \Rightarrow v_d = -ve \Rightarrow V_o = -V_{sat}$

$v_i < 0 \Rightarrow v_d = +ve \Rightarrow V_o = +V_{sat}$



when V_{ref} is not given : Schmitt trigger



$V_{io} \rightarrow$ i/p offset voltage "nV"

$I_{io} \rightarrow$ i/p offset current "nA"

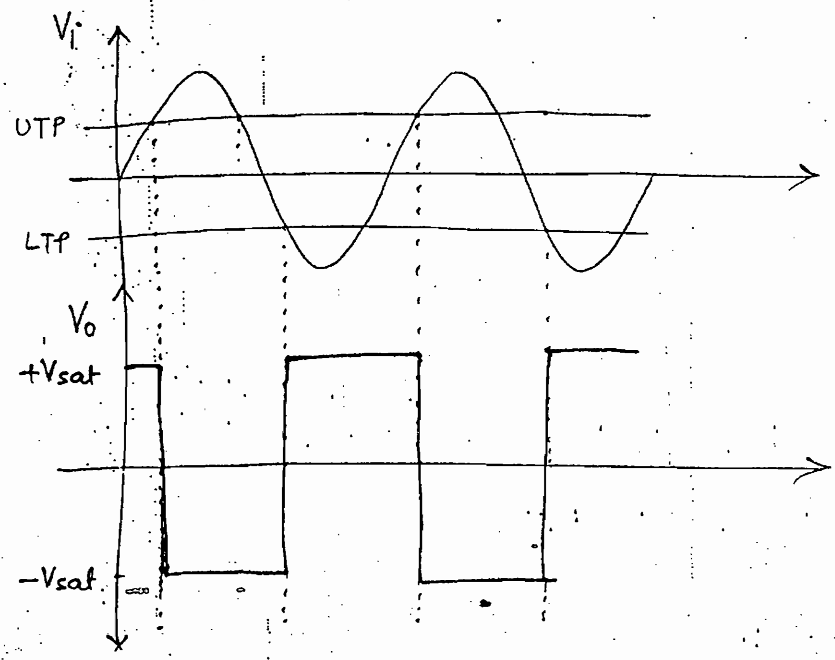
At $V_i = 0$, $\quad V_{io} \times 10^6 = V_o = nV \times 10^6 = mV$.

when, $V_o = +V_{sat} \Rightarrow V_{ref_1} = \dfrac{R_1}{R_1 + R_2} V_{sat}$

when, $V_o = -V_{sat} \Rightarrow V_{ref_2} = \dfrac{R_1}{R_1 + R_2} \cdot -V_{sat}$

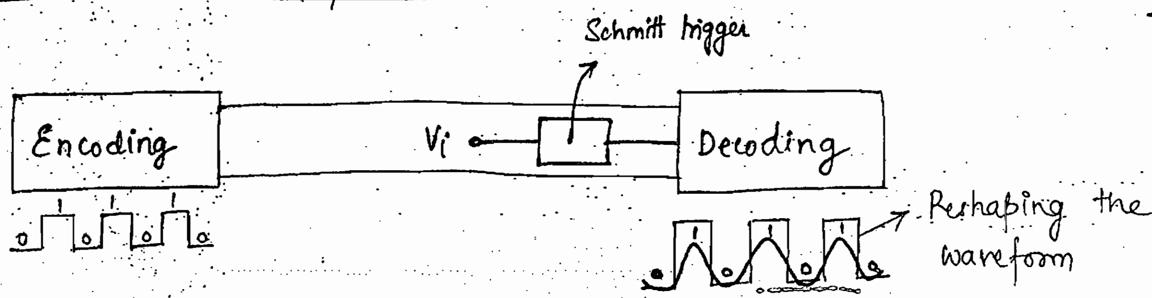
upper trigger point, $UTP = +V_{sat} \cdot \dfrac{R_1}{R_1 + R_2}$

Lower trigger point, $LTP = -V_{sat} \cdot \dfrac{R_1}{R_1 + R_2}$

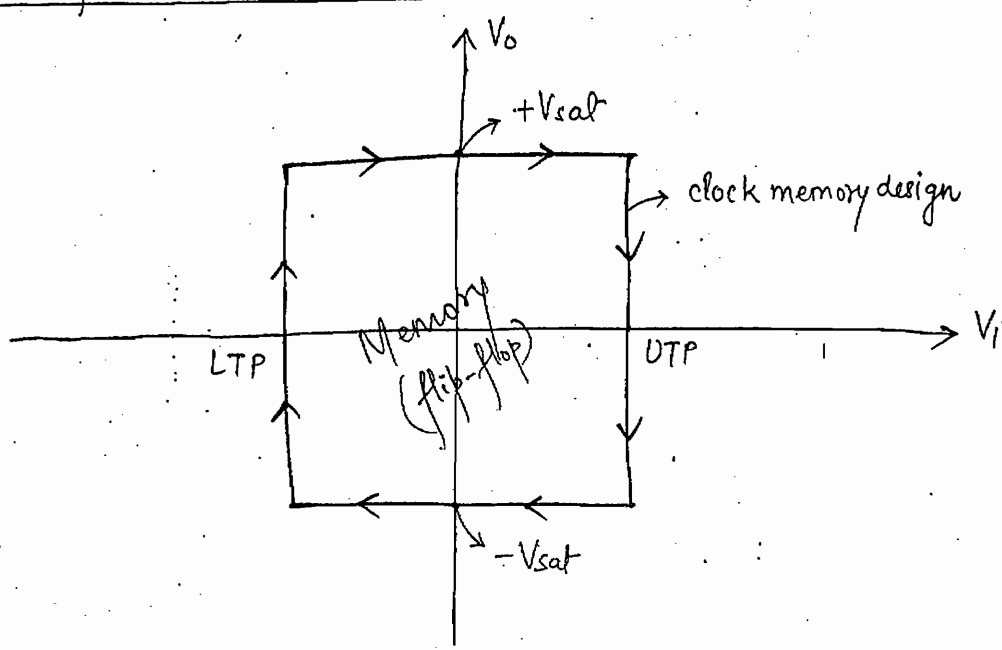


Applications of Schmitt trigger :

Communication system :

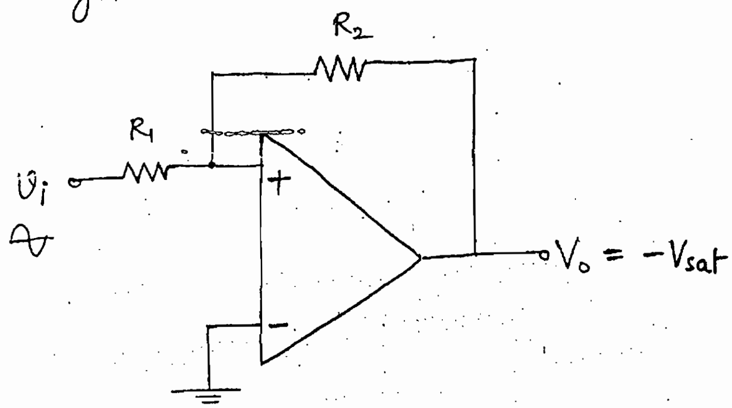


Transfer characteristics :



Q. Design a schmitt trigger which gives anti clock memory design.

Sol.



When $V_o = -V_{sat}$

$$v_i \cdot \frac{R_2}{R_1+R_2} + V_o \cdot \frac{R_1}{R_1+R_2} = 0$$

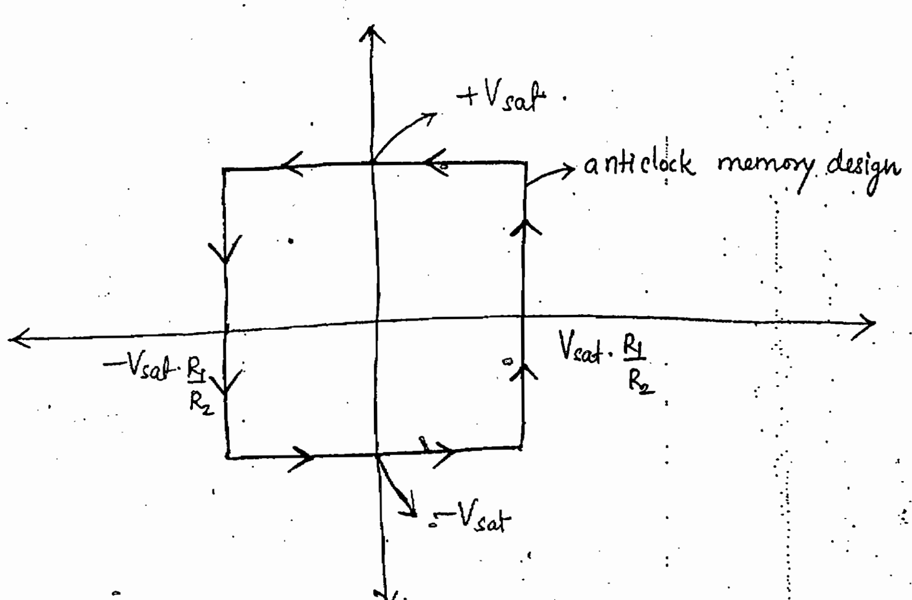
$$V_i \cdot \frac{R_2}{R_1+R_2} - V_{sat} \cdot \frac{R_1}{R_1+R_2} = 0$$

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

When $V_o = +V_{sat}$

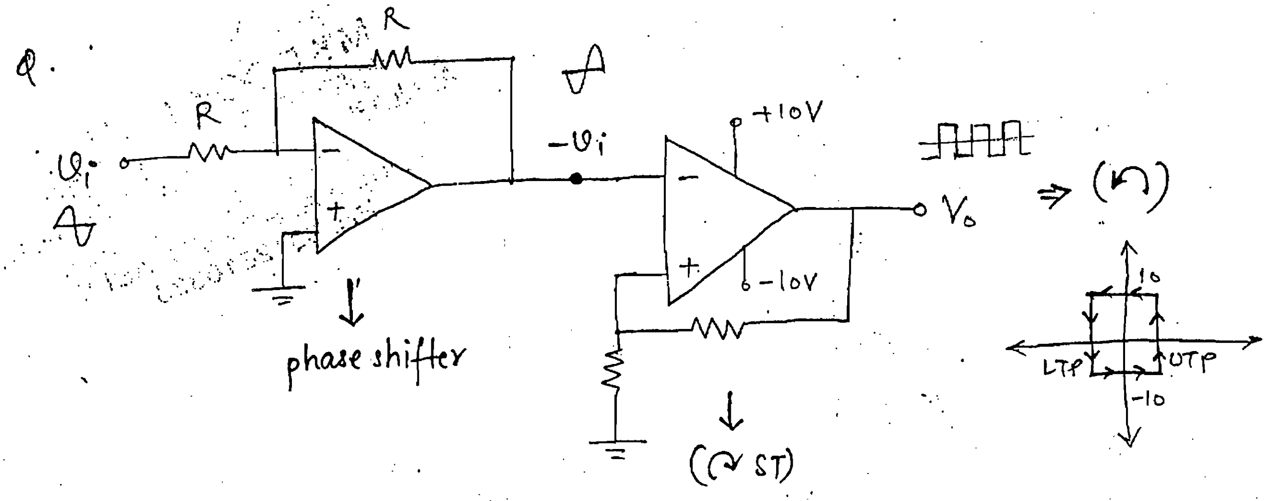
$$v_i \frac{R_2}{R_1+R_2} + V_o \frac{R_1}{R_1+R_2} = 0$$

$$V_i = -V_{sat} \cdot \frac{R_1}{R_2}$$

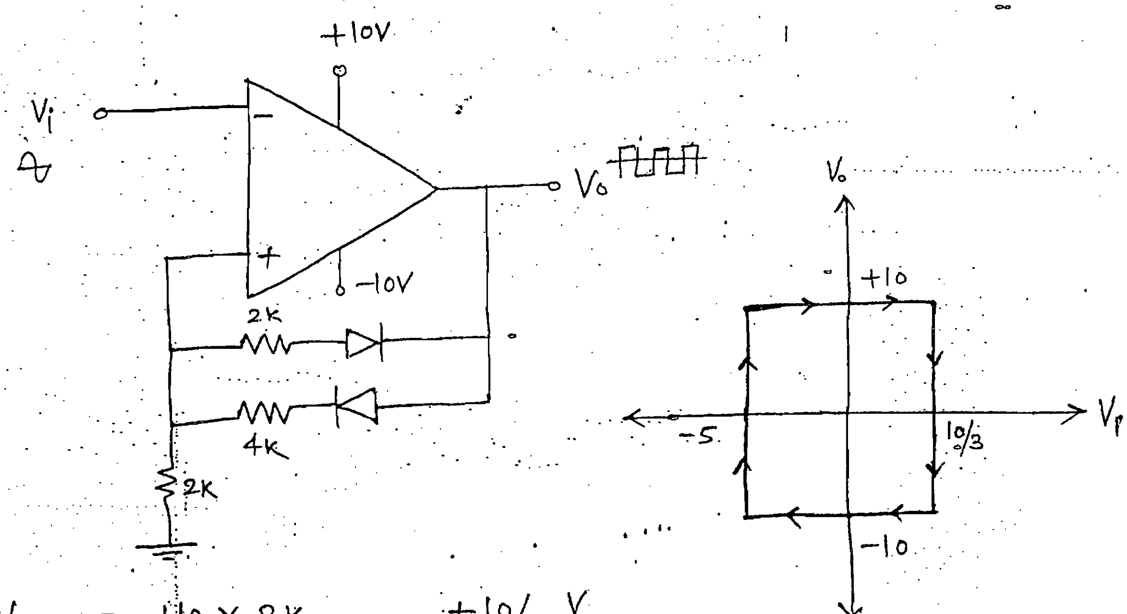


Problems :

Q.



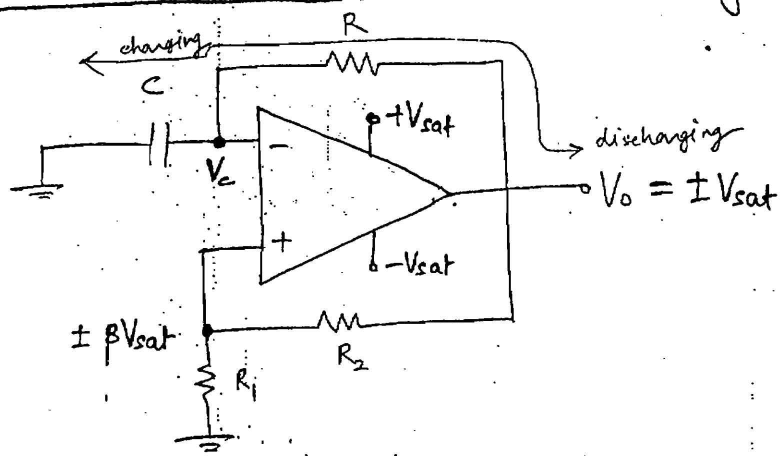
Q.



$$V_{UTP} = +10 \times \frac{2K}{2K+4K} = +10/3 \text{ V}$$

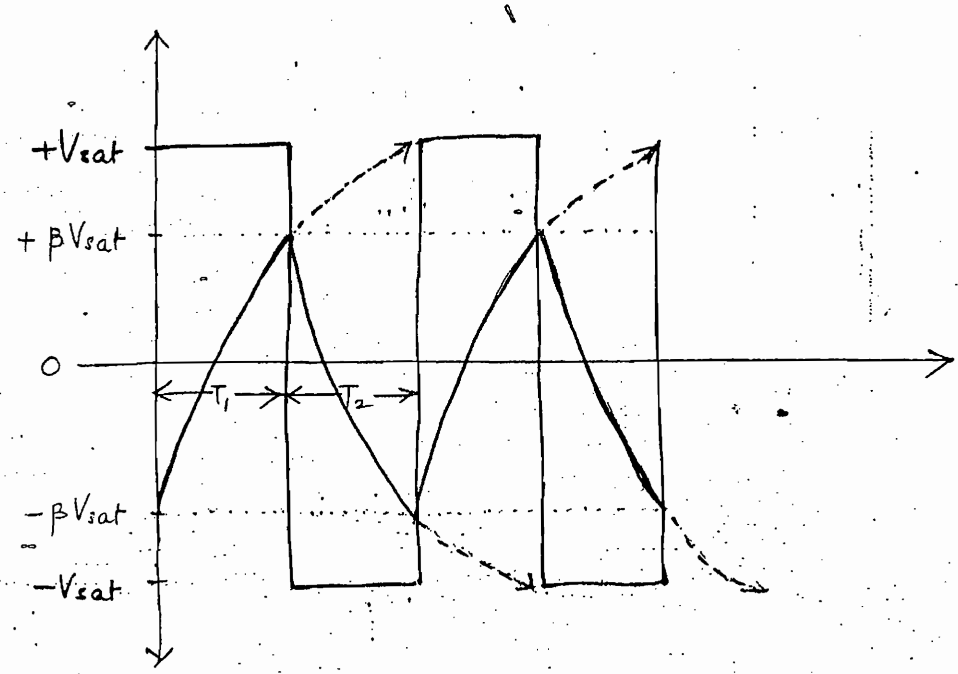
$$V_{LTP} = -10 \times \frac{2K}{2K+2K} = -5V$$

Astable Multivibrator (or) free running oscillator :



$$V_{UTP} = +V_{sat} \cdot \frac{R_1}{R_1+R_2} = \beta V_{sat}$$

$$V_{LTP} = -V_{sat} \cdot \frac{R_1}{R_1+R_2} = -\beta V_{sat}$$



$$V_c = V_{final} - (V_{final} - V_{initial})e^{-t/RC}$$

At $t = T_2$,

$V_c = -\beta V_{sat}$; $V_{final} = -V_{sat}$; $V_{initial} = +\beta V_{sat}$

$$-\beta V_{sat} = -V_{sat} + (1+\beta) V_{sat} \cdot e^{-T_2/RC}$$

$$(1-\beta) V_{sat} = (1+\beta) \cdot V_{sat}\, e^{-T_2/RC}$$

$$T_2 = RC \ln\left(\frac{1+\beta}{1-\beta}\right)$$

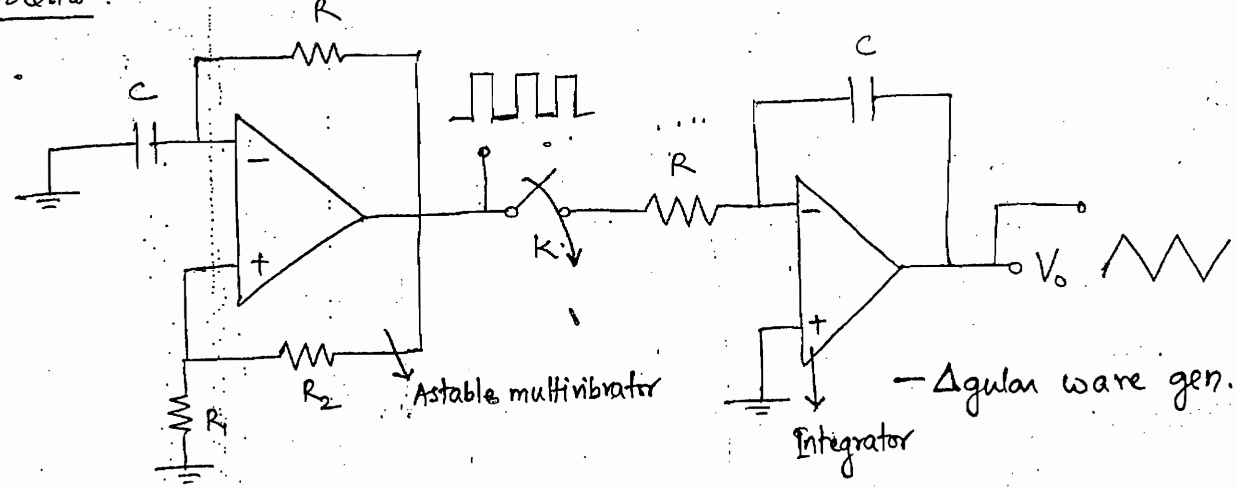
$$T_1 = RC \ln\left(\frac{1+\beta}{1-\beta}\right)$$

$$T = T_1 + T_2 = 2RC \ln\left(\frac{1+\beta}{1-\beta}\right)$$

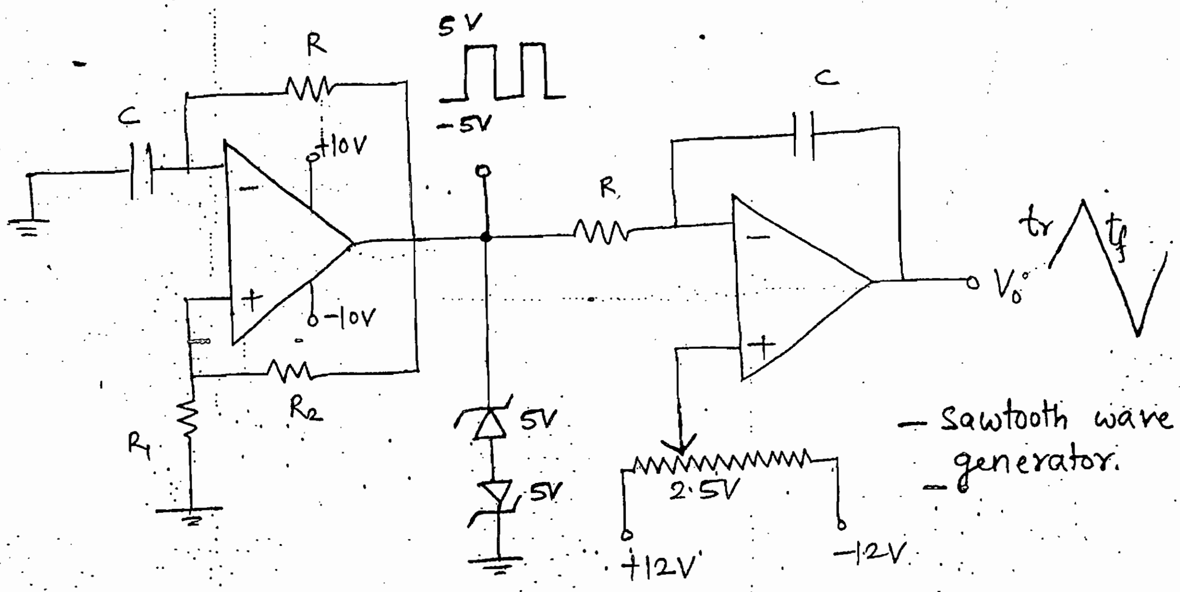
$$\boxed{f = \frac{1}{2RC \ln\left(\frac{1+\beta}{1-\beta}\right)}}$$

Problems:

Q.



Q.



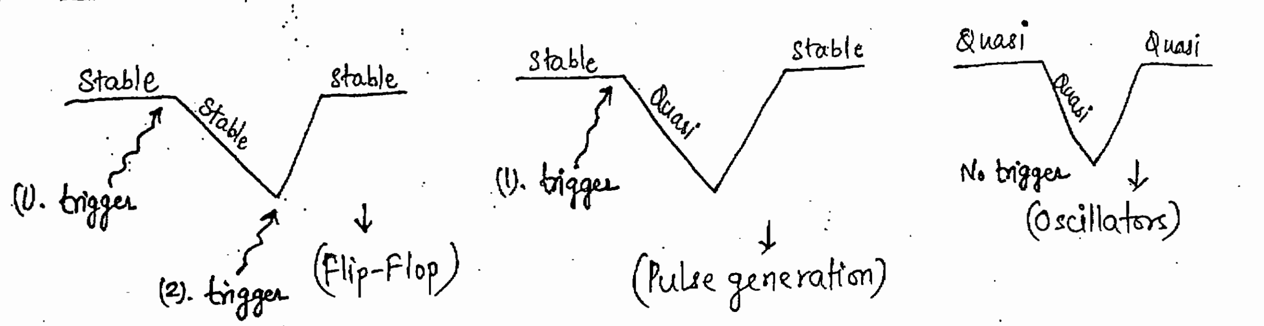
When the wiper moves towards +ve side. $t_f > t_r$

When the wiper moves towards −ve side. $t_r > t_f$

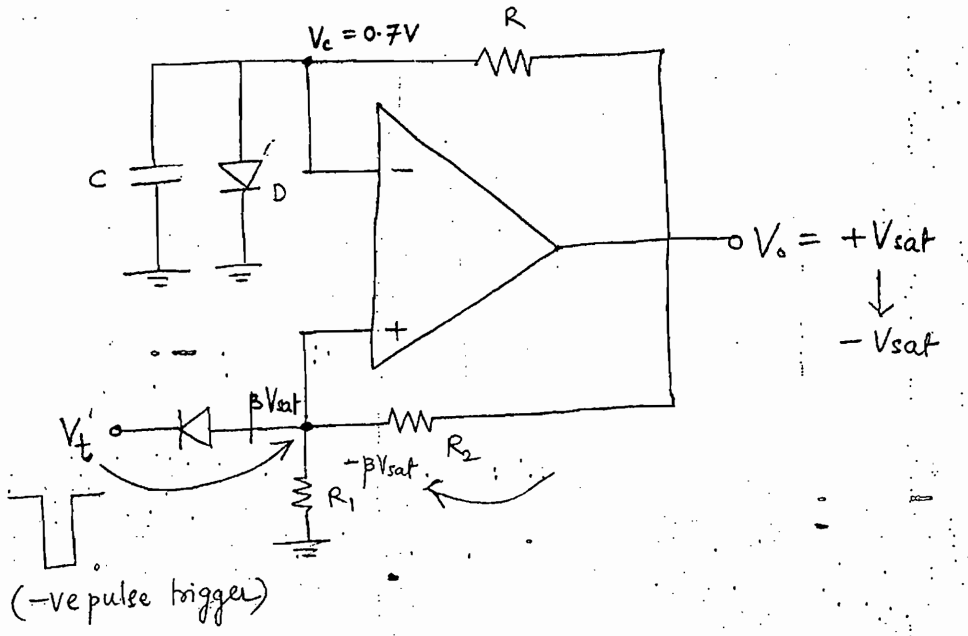
Mono stable multivibrator:

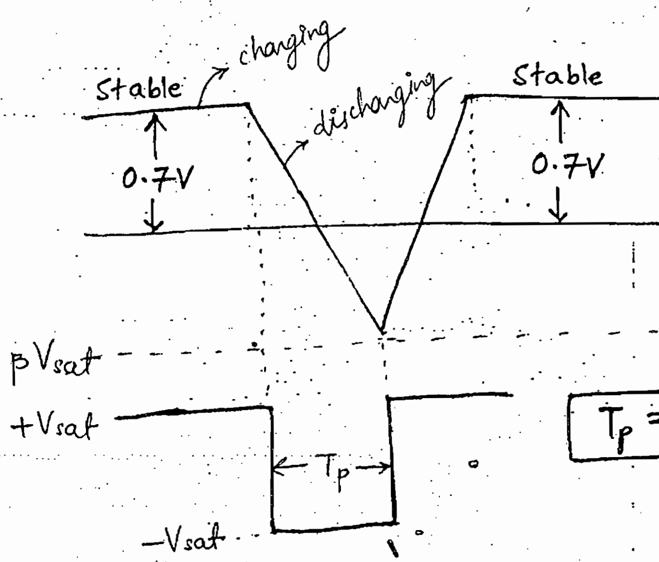
Without capacitor, there is no multivibrator.

Note: Bistable multivibrator — Monostable multivibrator — Astable multivibrator



Monostable Multivibrator :

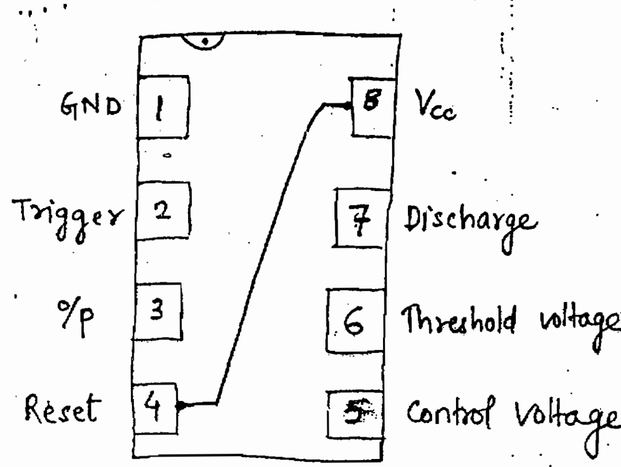


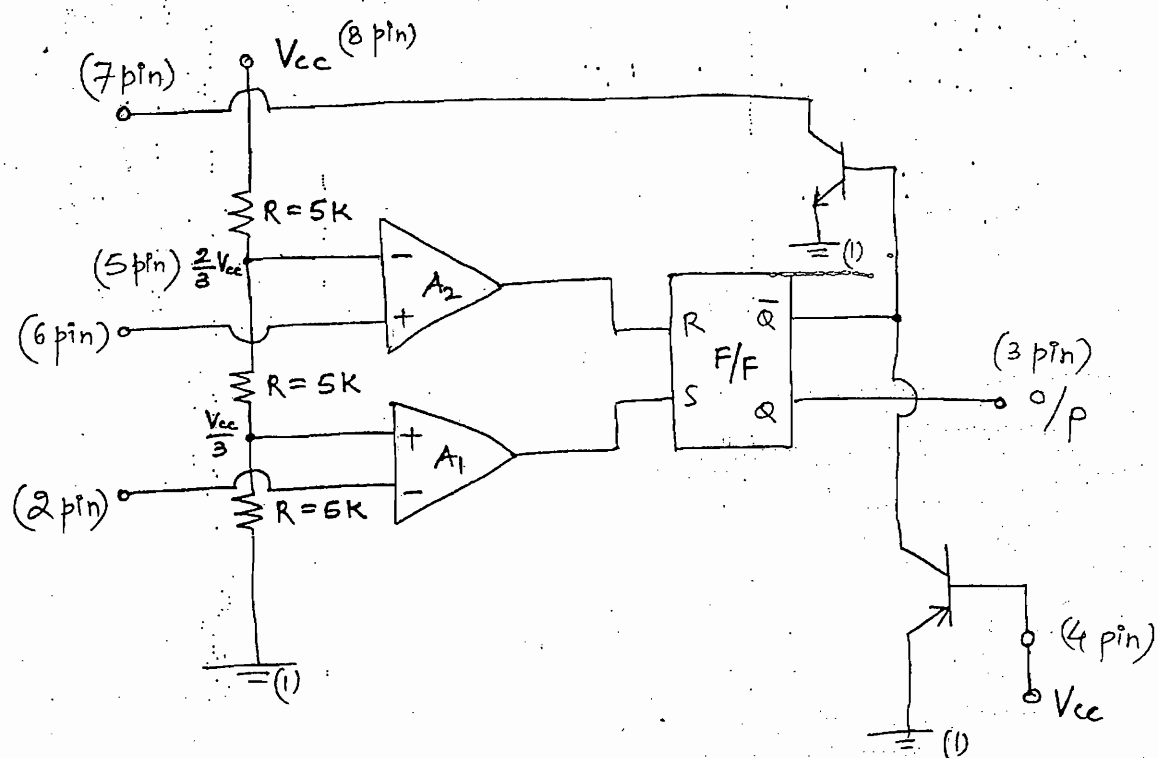


$$\boxed{T_p = 0.693\,RC}$$

555 timer

Pin diagram →

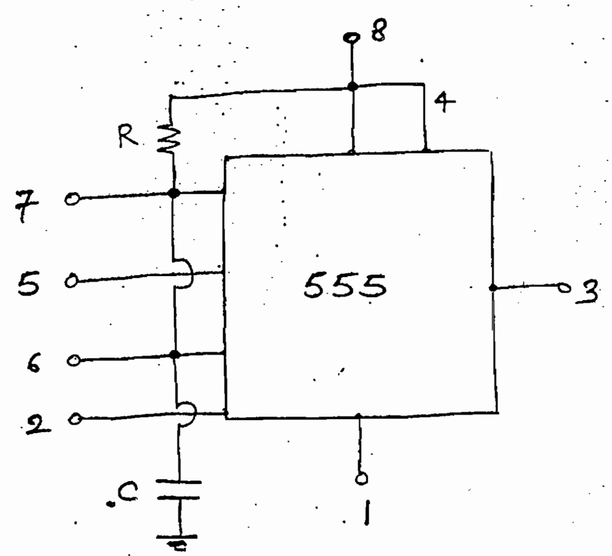




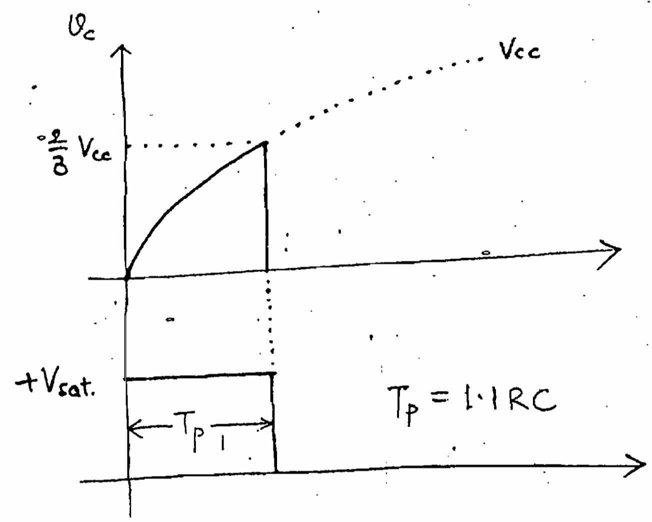
555 timer operating modes :

→ Monostable mode → Astable mode

(1). Monostable 555 timer :

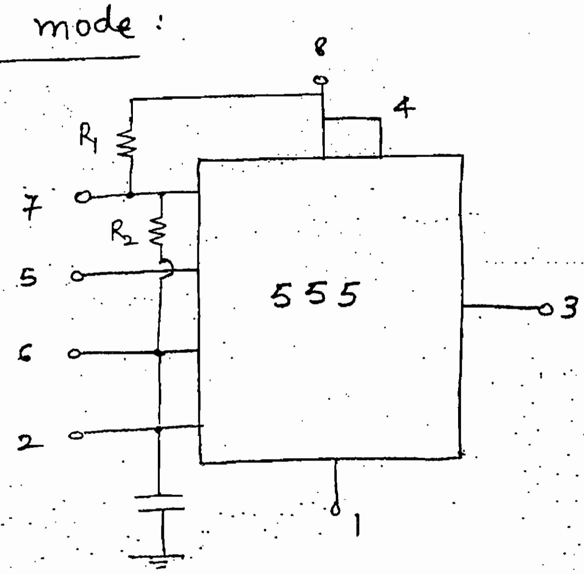


Condition	Q	$\bar{Q}$	o/p	Capacitor position
Stand by	0	1	0	discharging – stable ($V_c=0$)
trigger $< \frac{1}{3}V_{cc}$	1	0	1	charging – quasi ($V_c=\frac{2}{3}V_{cc}$)
$V_c > \frac{2}{3}V_{cc}$	0	1	0	discharging – stable ($V_c=0$)

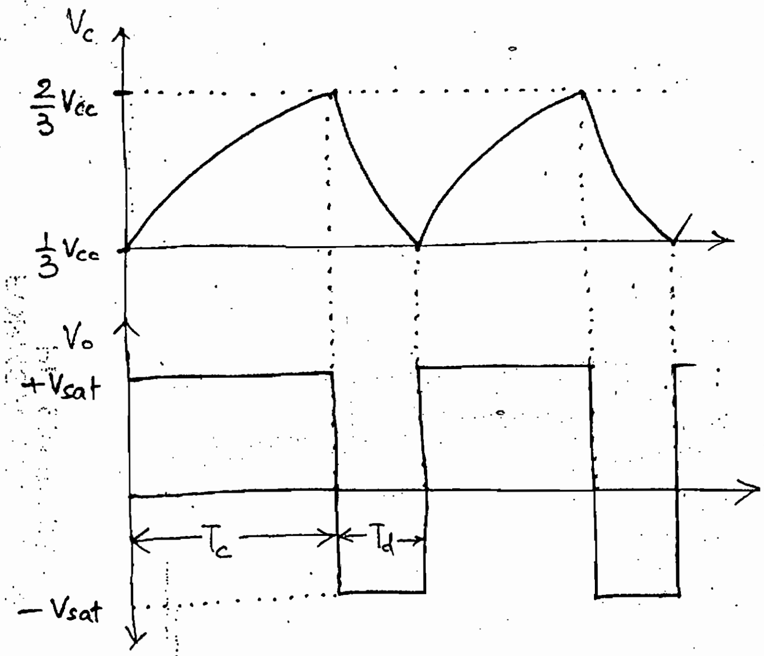


$T_p = 1.1RC$

(2). Astable mode :



Condition	Q	$\bar{Q}$	o/p	Capacitor position
Stand by	0	1	0	dischanging, $V_c = \frac{1}{3} V_{cc}$
$V_c < \frac{1}{3} V_{cc}$	1	0	1	changing, $V_c = \frac{2}{3} V_{cc}$
$V_c > \frac{2}{3} V_{cc}$	0	1	0	dischanging, $V_c = \frac{1}{3} V_{cc}$
⋮	⋮	⋮	⋮	⋮



$$T = T_c + T_d$$

$$T_c = 0.693\,(R_1+R_2)\,C$$

$$T_d = 0.693 \cdot R_2 C$$

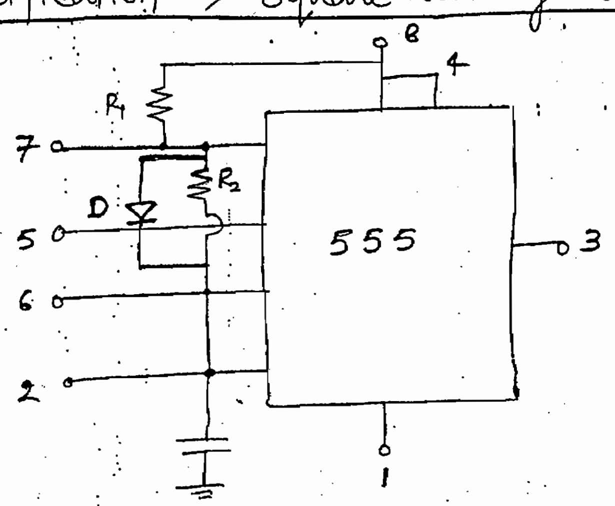
$$T = 0.693\,(R_1+2R_2)\,C$$

$$\text{Duty cycle} = \frac{T_c}{T_c+T_d}$$

$$DC = \frac{0.693\,(R_1+R_2)\,C}{0.693\,(R_1+2R_2)\,C}$$

$$DC = \frac{R_1+R_2}{R_1+2R_2}$$

Modification → Square wave generator



$$T_c = 0.693\,R_1 C$$

$$DC = \frac{R_1}{R_1+R_2}$$

if $R_1 = R_2 = R$

$$DC = \frac{R}{R+R} = 0.5$$

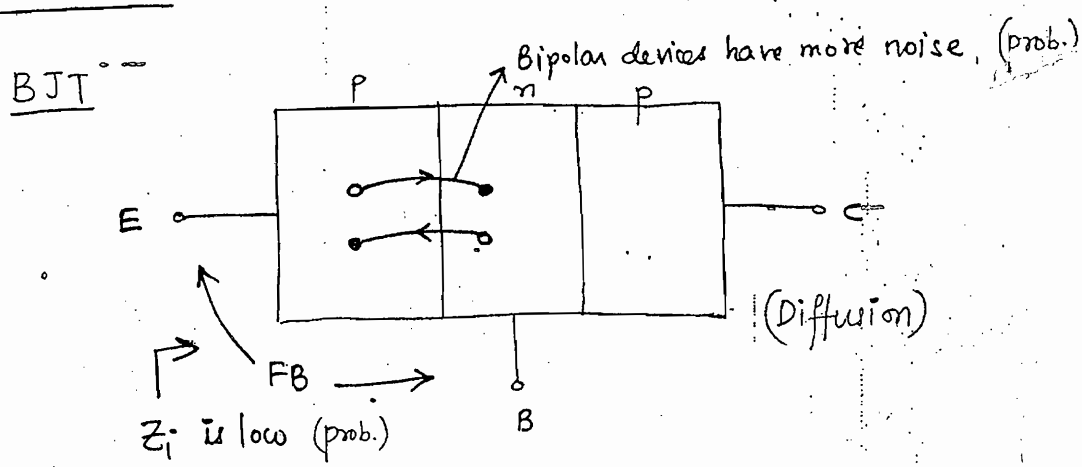
↓

Square wave

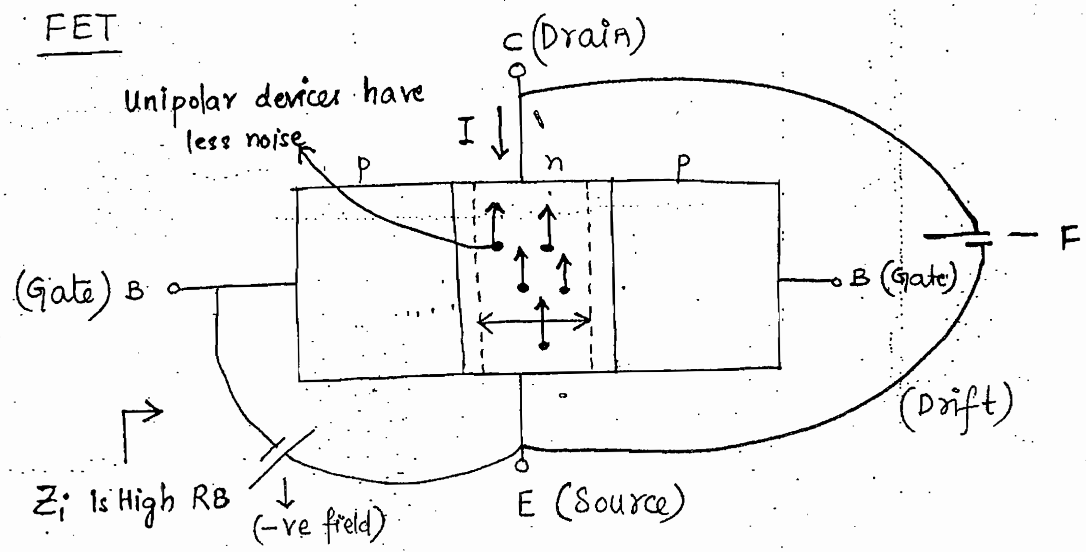
FET – Field Effect Transistor

Introduction :

BJT



FET



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 impedence is low.

(2). The noise level is high.

Field Effect transitor is having following advantages compared to BJT :

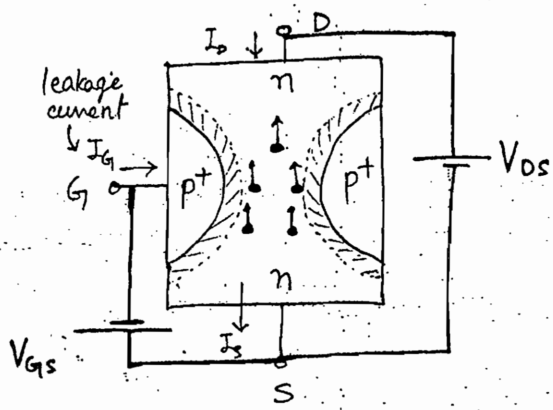
(1). The I/p impedence is high becoz the I/p is always RB

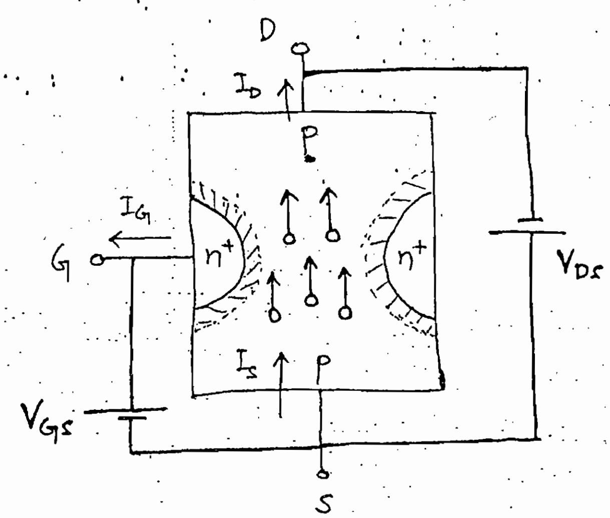
(2). The noise level is comparatively low becoz it is a unipolar device

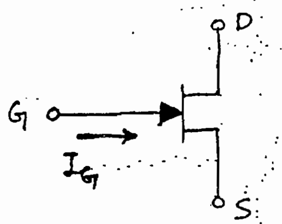
<u>Note:</u> BJT is a current controlled device whereas FET is a voltage controlled device.

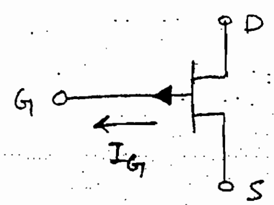
Types of FET are :

FET
- N-channel
- P-channel



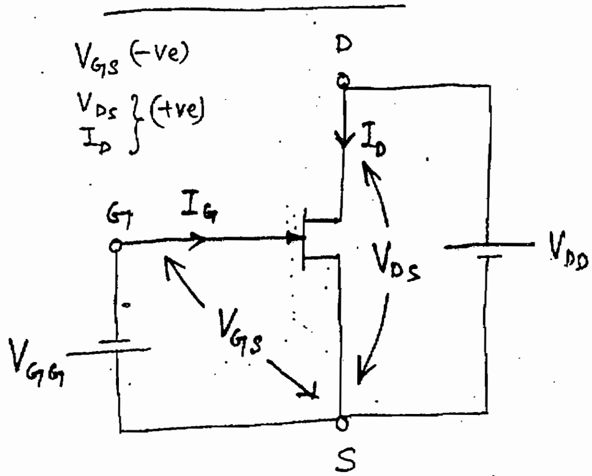




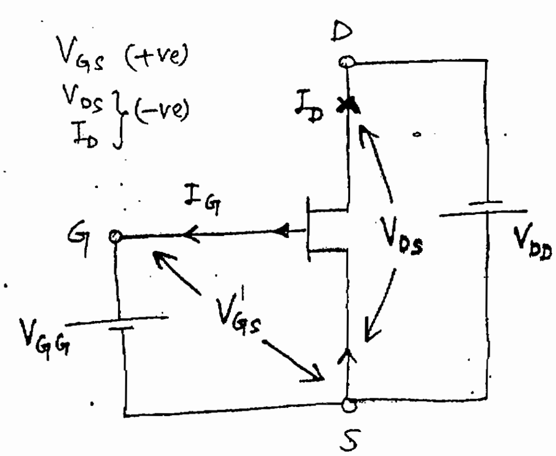


Important points in FET :

n-channel FET



p-channel FET



I/p parameters → V_{GS}, I_G

O/p parameters → 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:

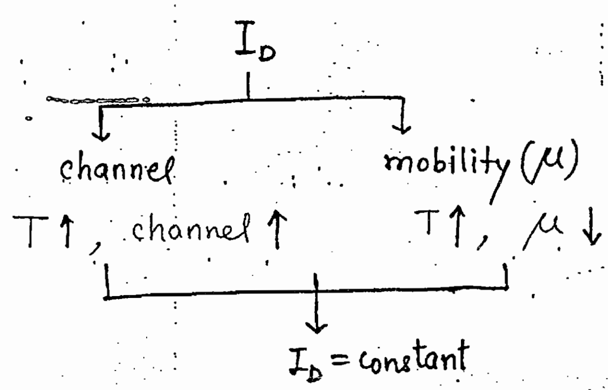
(i). Drain characteristics (V_{DS} Vs I_D at constant V_{GS})

(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 $Q\left[(I_D)_Q, (V_{GS})_Q\right]$.

(6) FET is always thermally stable.



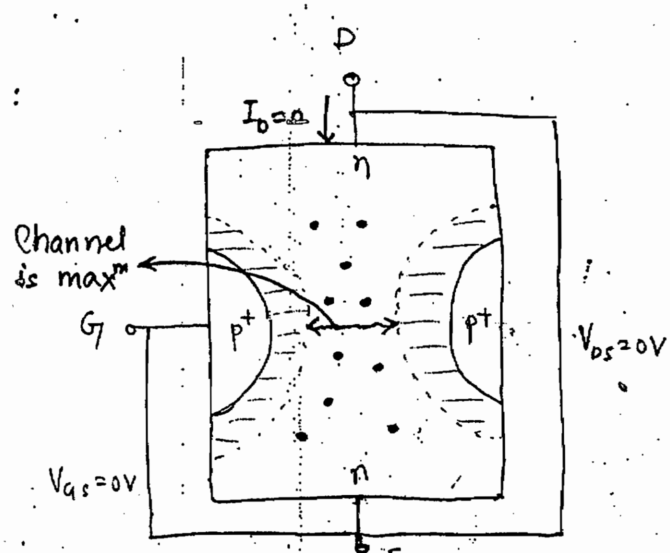
(7). For FET biasing, we have to fix V_{GS} voltage but not drain current I_D.

<u>Working principle of FET</u>:

<u>Case – (1)</u>:

$V_{GS} = 0V$

$V_{DS} = 0V$



Conclusions :

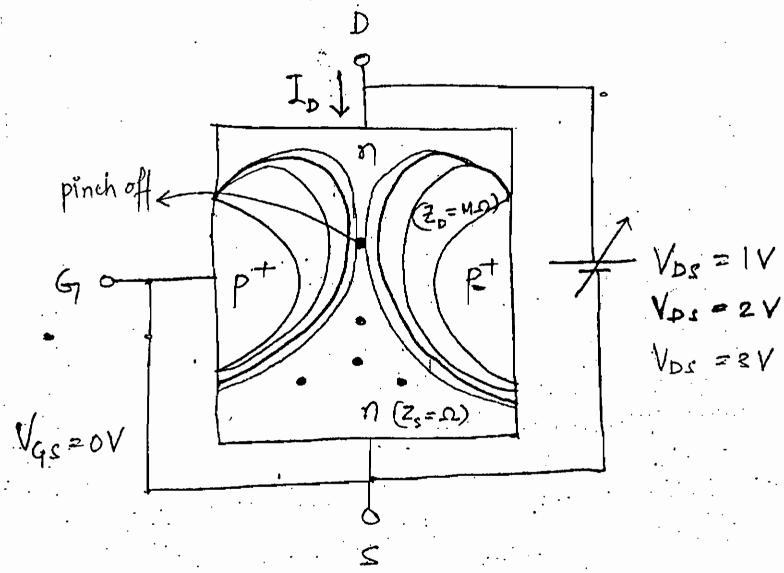
(1). When $V_{GS} = 0V$, max^m channel can be achieved

(2). When $V_{DS} = 0V$, the drain current I_D becomes zero.

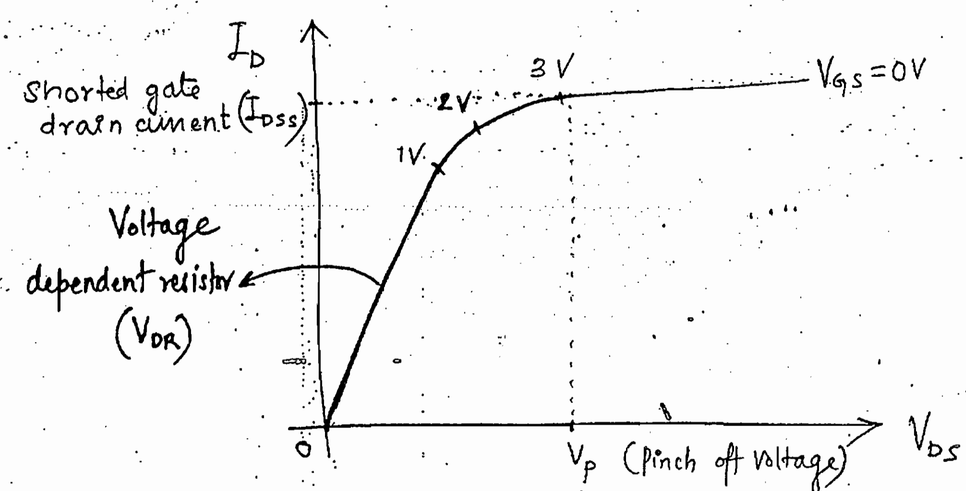
Case – (2):

$V_{GS} = 0V$

$V_{DS} \neq 0V$



Characteristics (V_{DS} Vs I_D) :

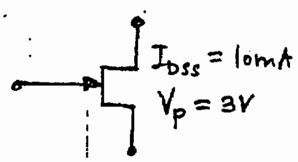


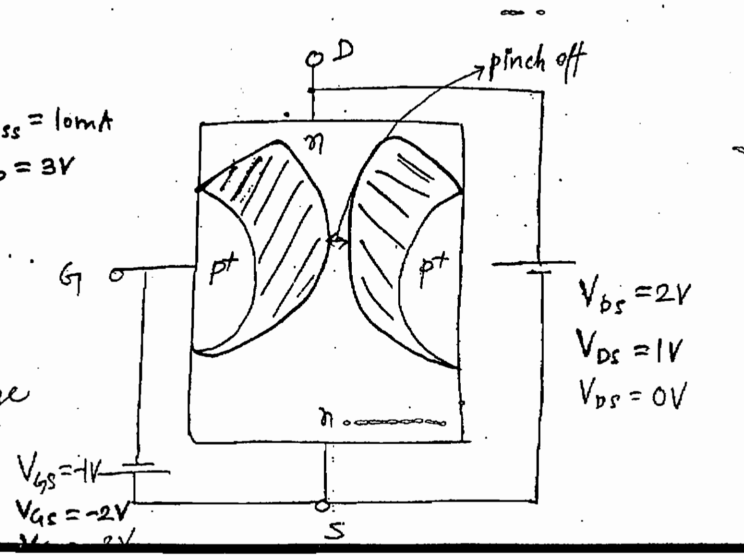
Case – (3) :

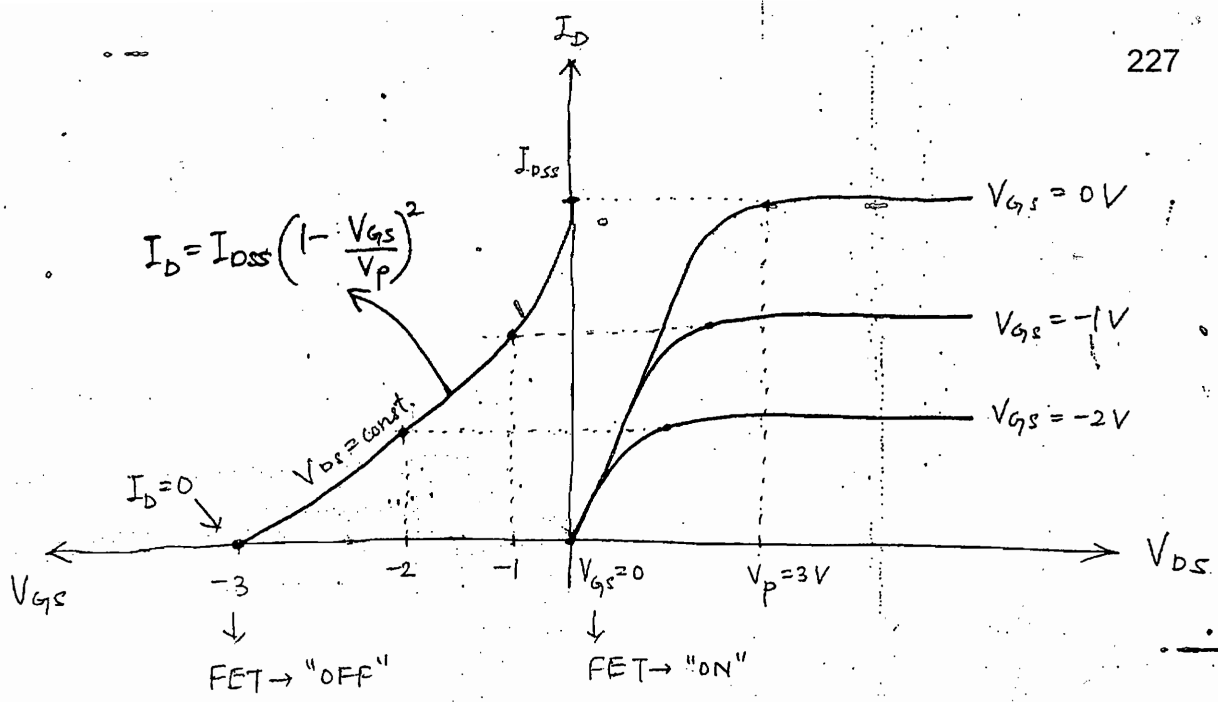
$V_{GS} \neq 0V$

$V_{DS} \neq 0V$

When $V_{DS} = 0V$; $V_{GS} = -3V$ → pinch off voltage





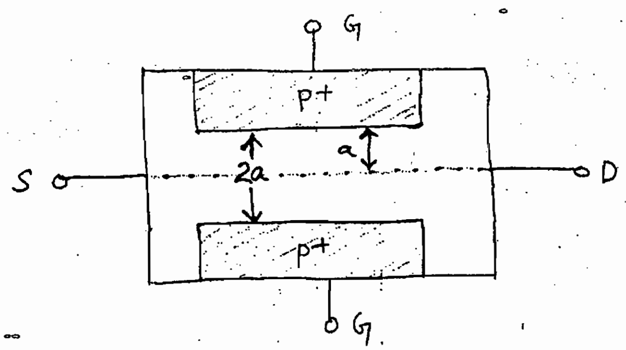


(1). Transfer characteristic eqn is

$$I_D = I_{DSS}\left(1-\frac{V_{GS}}{V_p}\right)^2$$

(2). $|V_{GS}|_{off} = |V_p|$

(3). $V_p = \frac{a^2 q N_A}{2\epsilon}$ (p-channel) (or) $\frac{a^2 q N_D}{2\epsilon}$ (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). Ac drain resistance, $r_d = \left.\frac{\Delta V_{DS}}{\Delta I_D}\right|_{V_{GS} = \text{const.}}$

(2). Transconductance, $g_m = \left.\frac{\Delta I_D}{\Delta V_{GS}}\right|_{V_{DS} = \text{const.}}$

(3). 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:

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

$$\frac{dI_D}{dV_{GS}} = -\frac{2I_{DSS}}{V_P}\left(1 - \frac{V_{GS}}{V_P}\right)$$

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = -\frac{2I_{DSS}}{V_P}\left[1 - \frac{(V_{GS})_Q}{V_P}\right]$$

$$\boxed{g_m = g_{m0}\cdot\left[1 - \frac{(V_{GS})_Q}{V_P}\right]}$$

where, $g_{m0} = \frac{-2I_{DSS}}{V_P}$ (max^m transconductance)

And,

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2 \Rightarrow \left(1 - \frac{V_{GS}}{V_P}\right) = \sqrt{\frac{I_D}{I_{DSS}}}$$

$$g_m = \frac{-2I_{DSS}}{V_P}\left(1 - \frac{V_{GS}}{V_P}\right).$$

$$= \frac{-2I_{DSS}}{V_P}\cdot\sqrt{\frac{I_D}{I_{DSS}}}$$

$$\therefore\ g_m = \frac{-2}{V_P}\sqrt{I_{DSS}\cdot(I_D)_Q} = \frac{2}{|V_P|}\sqrt{I_{DSS}\cdot(I_D)_Q}$$

FET Biasing :

Conditions :

(1). $I_G \approx 0$

(2). $I_D \approx I_S$

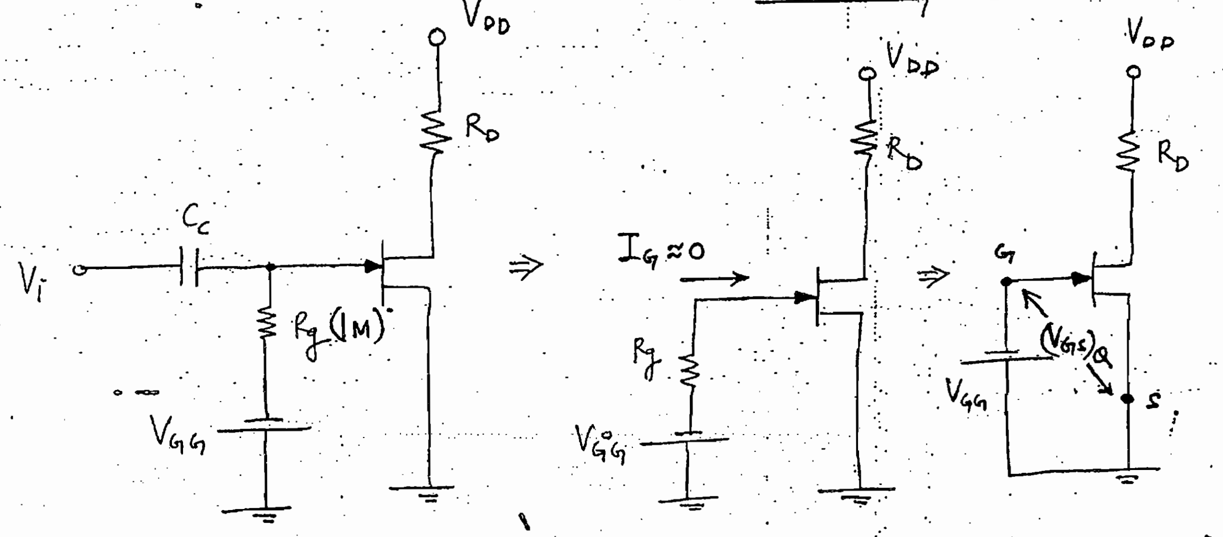
(3). $I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$

(4). Q 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 s/g only

(1). $(V_{GS})_Q = -V_{GG}$ (fixed) — Biasing by keeping a battery.

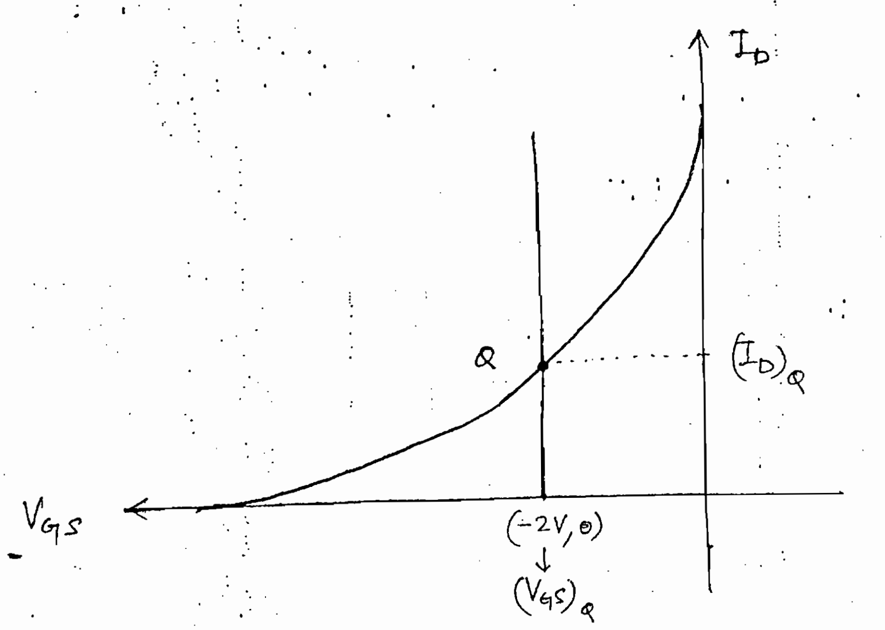
(2). $(I_D)_Q = I_{DSS}\left[1 - \frac{(V_{GS})_Q}{V_P}\right]^2$

(3). $(V_{DS})_Q = V_{DD} - (I_D)_Q \cdot R_D$

(4). $(V_D)_Q = (V_{DS})_Q$

(5). $(V_S)_Q = 0$

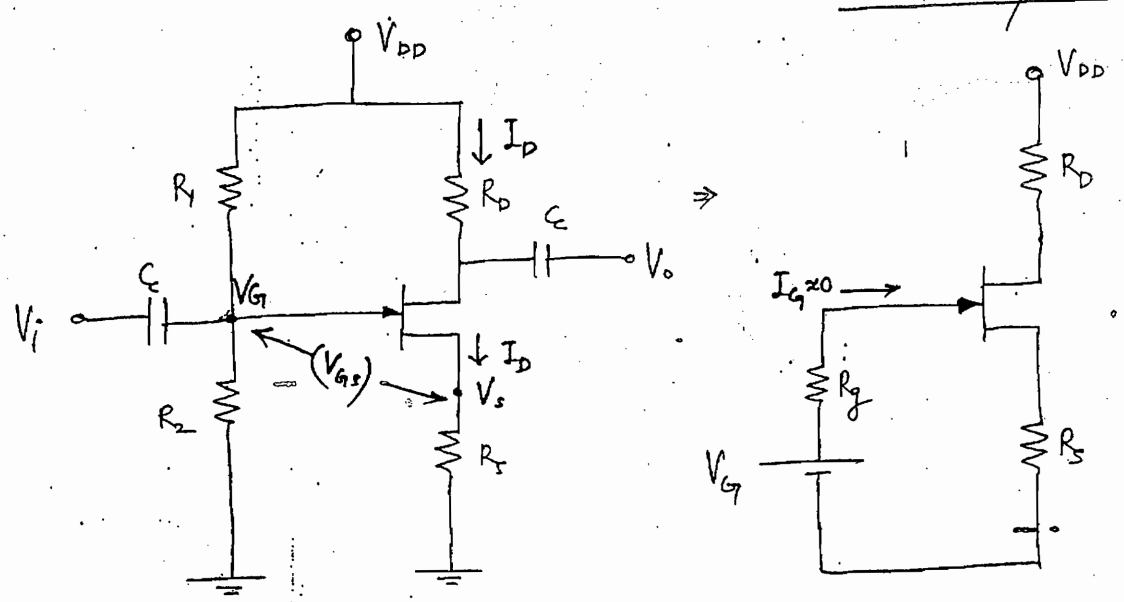
(6). $(V_G)_Q = (V_{GS})_Q$

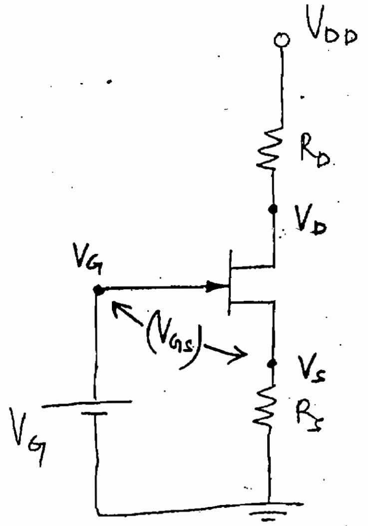


This biasing is not good becoz battery can vary the Q point as there are fluctuations in battery voltage. This is not a stable biasing. Hence, some modifications are required. Bypass amplifiers are always unstable.

(2) Voltage Divider Biasing :

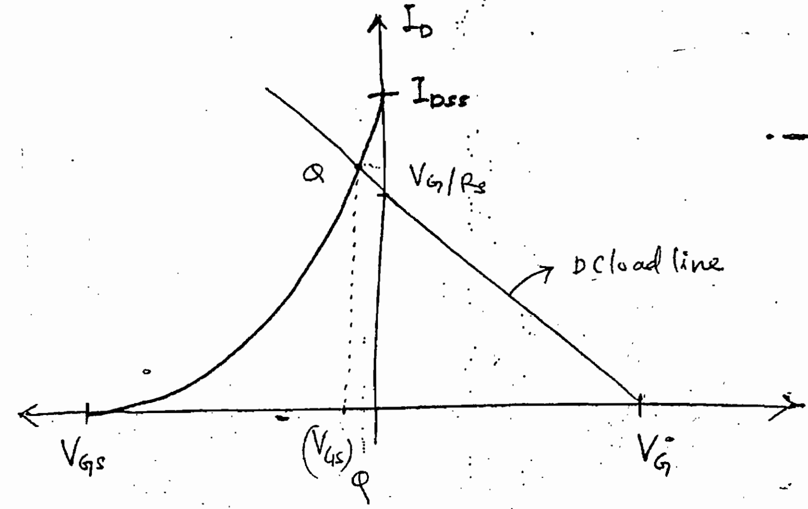
DC analysis





$$V_G = V_{GS} + I_D \cdot R_S$$

$$V_{GS} = V_G - I_D R_S$$



$$V_G = \frac{V_{DD} \times R_2}{R_1 + R_2}$$

$$R_G = R_1 // R_2$$

(1) $\therefore I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$ and $V_{GS} = V_G - I_D R_S$

Solving we get,

$$(I_D)_Q = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

+ve root (n-channel)

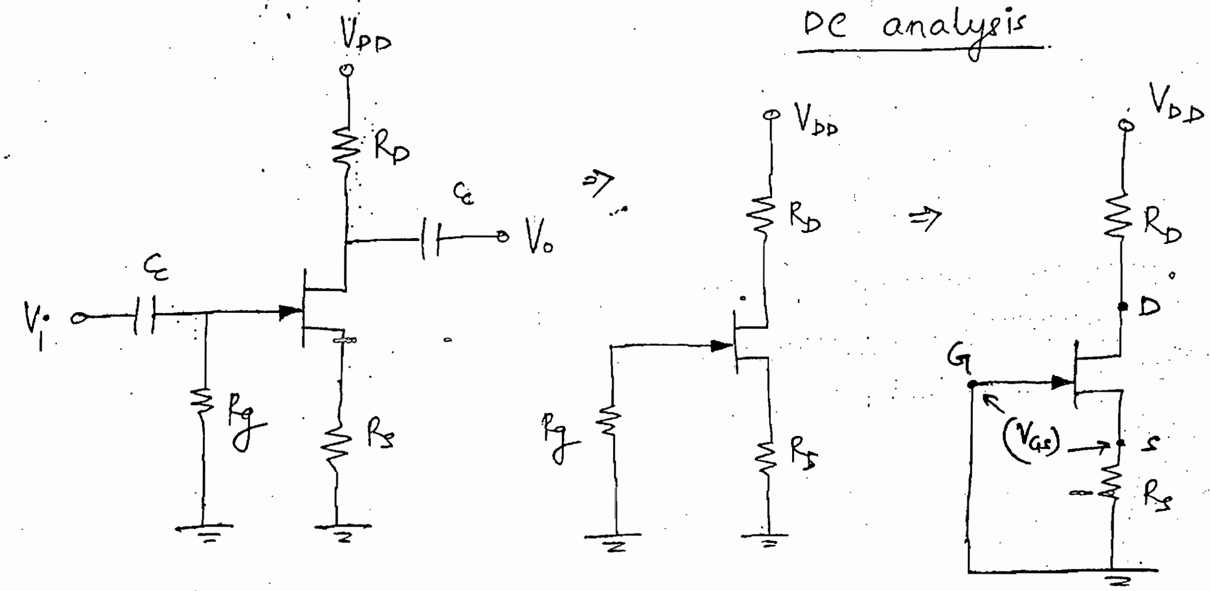
−ve root (p-channel)

(2). $(V_{GS})_Q = V_G - (I_D)_Q \cdot R_S$

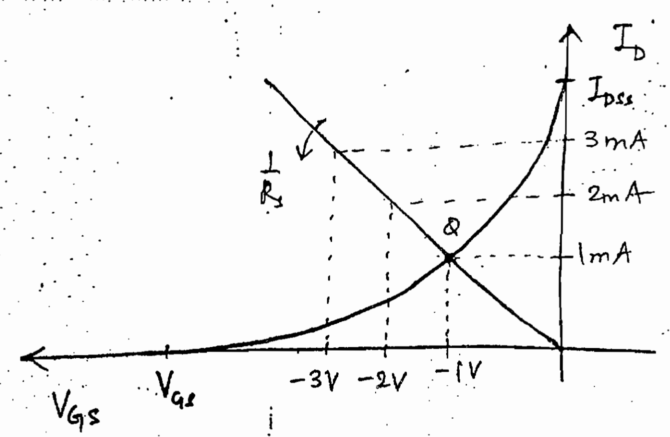
- Fixed biasing uses 2 batteries whereas this biasing is using single battery so it is an improved biasing but due to the presence of battery, it is also not much stable. Hence, more modification is required.

(3). Self Biasing (Stable Biasing): Common source unbypass

DC analysis



$V_{GS} = -I_D R_s \rightarrow$ Ohm's law.



(1). $I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$ and ... $V_{GS} = -I_D R_s$

$(I_D)_Q = \frac{-b \pm \sqrt{b^2-4ac}}{2a}$, < +ve root (n-channel) / -ve root (p-channel)

(2). $(V_{GS})_Q = -(I_D)_Q \cdot R_s$

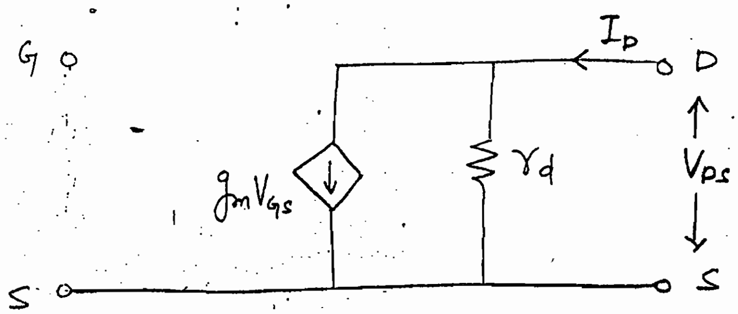
- Self biasing is a stable biasing and improved biasing as it does not depend upon battery voltage.
- UnBypass amplifiers are always stable.

FET model :

$$I_D = f\,(V_{GS}, V_{DS})$$

$$I_D = \frac{\partial I_D}{\partial V_{GS}} \cdot V_{GS} + \frac{\partial I_D}{\partial V_{DS}} \cdot V_{DS}$$

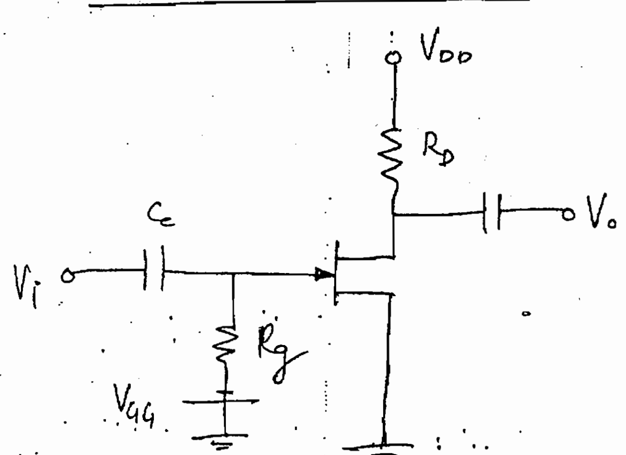
$$I_D = g_m \cdot V_{GS} + \frac{V_{DS}}{r_d}$$

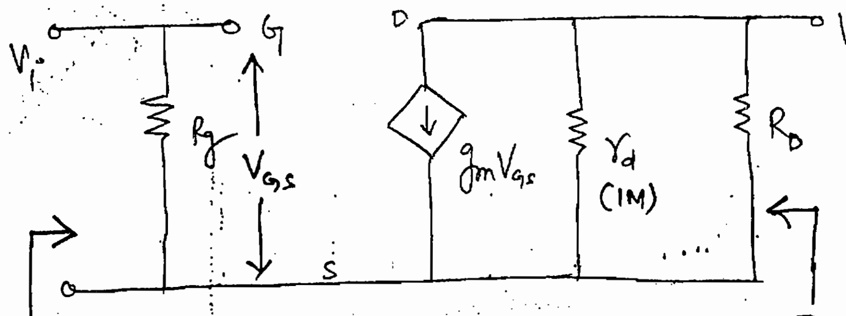


FET amplifiers :

(1). CS bypass amplifier
(2). CS unbypass amplifier
(3). CD amplifier
(4). CG amplifier

(1). CS bypass amplifier :





$\boxed{r_d \geq 10 R_D}$

$Z_o = r_d // R_D \approx R_D$

$Z_i = R_g$

Voltage gain:

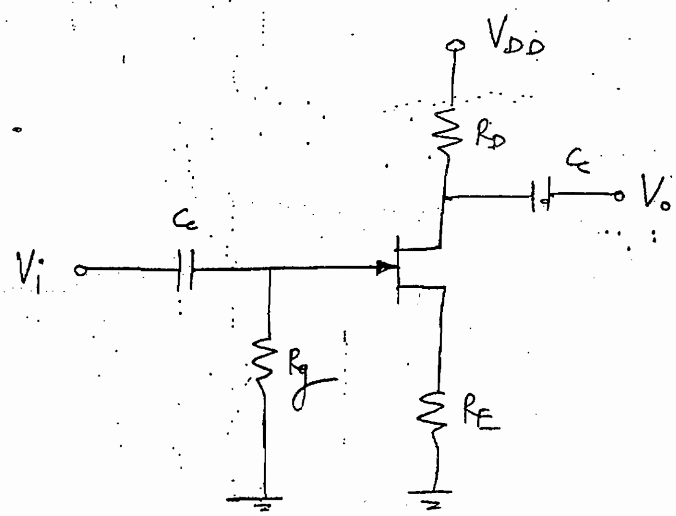
$V_o = -g_m \cdot V_{GS} \cdot (r_d // R_D)$

$V_{GS} = V_i$

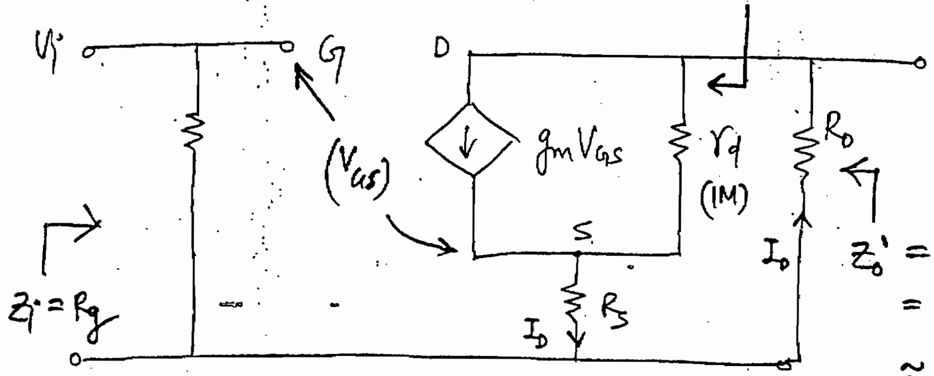
$V_o = -g_m V_i R_D$

$A_V = \dfrac{V_o}{V_i} = -g_m R_D$

CS unbypass amplifier:



$Z_o = r_d + (1+\mu) R_s$



$Z_i = R_g$

$Z_o' = R_D // Z_o$

$= R_D // [r_d + (1+\mu) R_s]$

$\approx R_D$

Voltage gain:

$$V_o = -I_D R_D \quad \text{——(1)}$$

$$I_D R_D + (I_D - g_m V_{GS}) r_d + I_D R_S = 0 \quad \text{——(2)}$$

$$V_i = V_{GS} + I_D R_S \quad \text{——(3)}$$

$$V_{GS} = V_i - I_D R_S$$

$$I_D R_D + [I_D - g_m (V_i - I_D R_S)] r_d + I_D R_S = 0$$

$$I_D R_D + I_D r_d - g_m r_d (V_i - I_D R_S) + I_D R_S = 0$$

$$I_D (R_D + r_d + R_S + g_m r_d R_S) = g_m r_d V_i$$

$$I_D = \frac{g_m r_d V_i}{(R_D + R_S + r_d + g_m r_d R_S)}$$

$$V_o = -I_D . R_D .$$

$$V_o = \frac{-g_m r_d R_D . V_i}{(R_D + R_S + r_d + g_m r_d R_S)}$$

$$\frac{V_o}{V_i} = \frac{-g_m r_d R_D}{(R_D + R_S + r_d + g_m r_d R_S)}$$

$$= \frac{-g_m R_D}{\left(\underbrace{\frac{R_D}{r_d}}_{\text{neglected}} + \underbrace{\frac{R_S}{r_d}}_{\text{neglected}} + 1 + g_m R_S\right)}$$

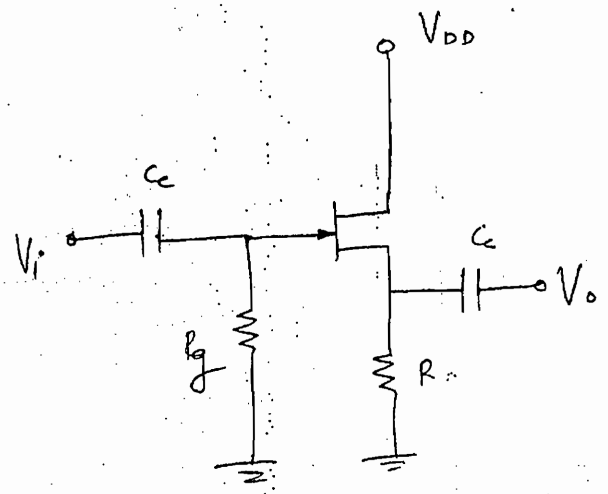
$$\boxed{A_V = \frac{V_o}{V_i} = \frac{-g_m R_D}{(1 + g_m R_S)}}$$

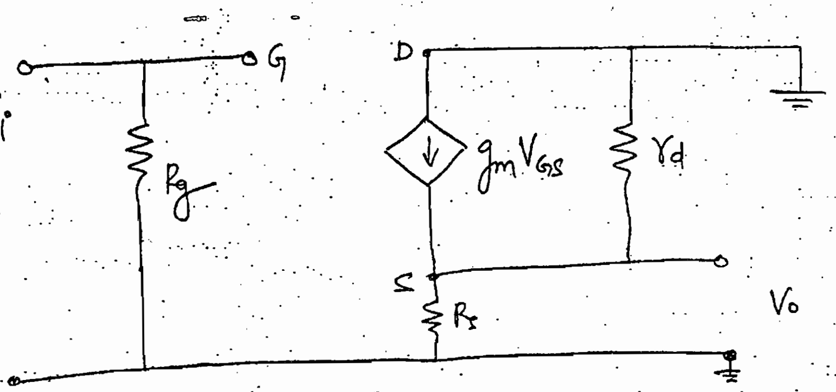
$$A_V = \frac{-g_m R_D}{1 + g_m R_S}$$

if $g_m R_S >>> 1$, then

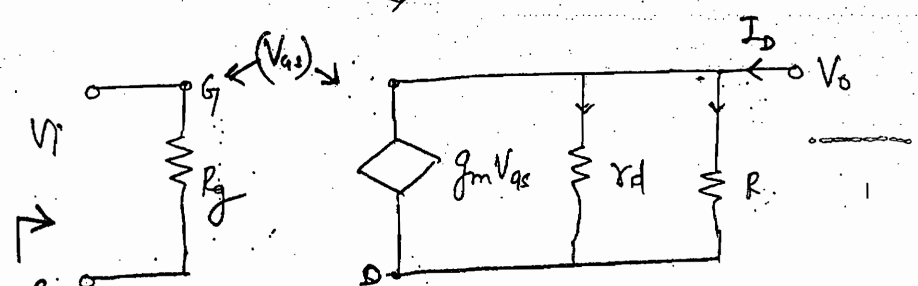
$$\boxed{A_V = -\frac{R_D}{R_S}} \rightarrow \text{stable}$$

(3) Common Drain amplifier :





$\Downarrow$



$Z_i = R_g$

$$I_D + g_m V_{GS} = \frac{V_o}{r_d} + \frac{V_o}{R}$$

$$V_i = V_{GS} + V_o \Rightarrow V_o = -V_{GS} \Rightarrow V_{GS} = -V_o$$

$\downarrow$
0

$$I_D - g_m V_o = \frac{V_o}{r_d} + \frac{V_o}{R}$$

$$I_D = \left(g_m + \frac{1}{r_d} + \frac{1}{R}\right) V_o$$

O/p Impedence :

$$Z_o = \frac{V_o}{I_D} = \frac{1}{\underset{\downarrow \text{neglect}}{\frac{1}{r_d}} + \frac{1}{R} + g_m} \approx \frac{1}{\underset{\downarrow \text{neglect}}{\frac{1}{R}} + g_m} \approx \frac{1}{g_m}$$

(Impedence matching).

$$\approx \left(R_s // \frac{1}{g_m}\right)$$

Voltage gain :

$$V_o = g_m V_{GS} \cdot (r_d // R_s)$$

$$V_i = V_{GS} + V_o$$

$$V_i = V_{GS} + g_m V_{GS} \cdot (r_d // R)$$

$$V_{GS} = \frac{V_i}{1 + g_m (r_d // R)}$$

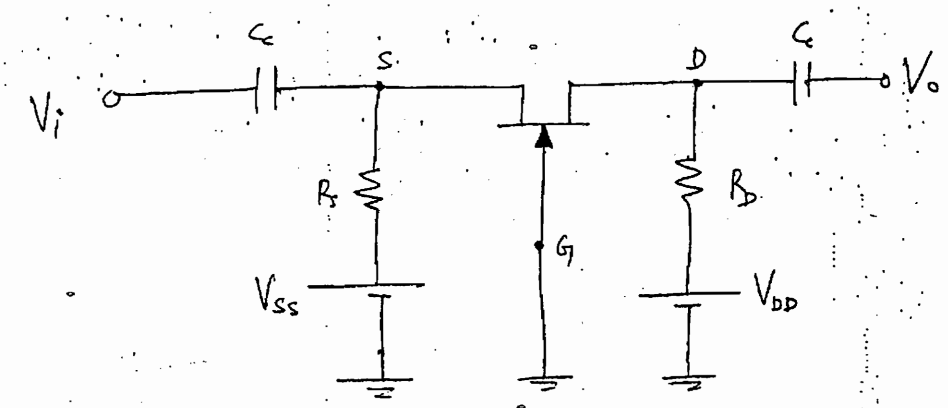
$$V_o = \frac{g_m \cdot V_i \; (r_d // R)}{1 + g_m (r_d // R)}$$

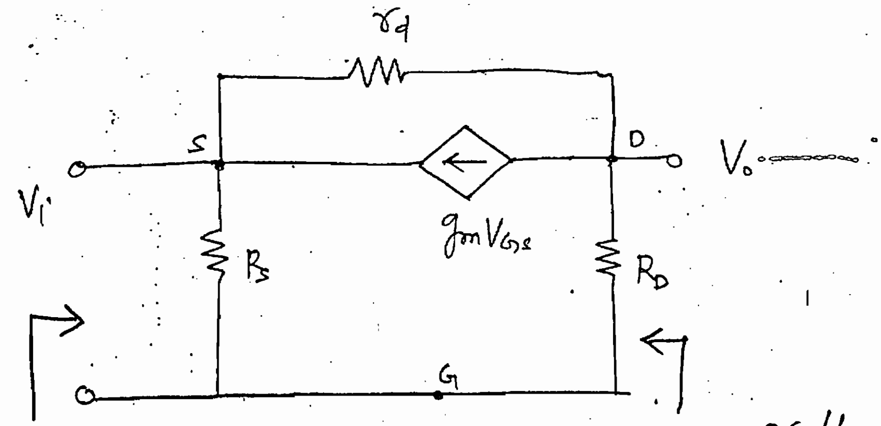
$$A_V = \frac{V_o}{V_i} = \frac{g_m \cdot R_s}{(1 + g_m R_s)}$$

If $g_m R_s >>> 1$, then

$$\boxed{A_V = \frac{V_o}{V_i} \approx 1} \rightarrow \text{Voltage Buffer}$$

(4). Common gate amplifier :





$$Z_o = r_d // R_D \approx R_D$$

$$Z_i = R_s // \frac{1}{g_m}$$

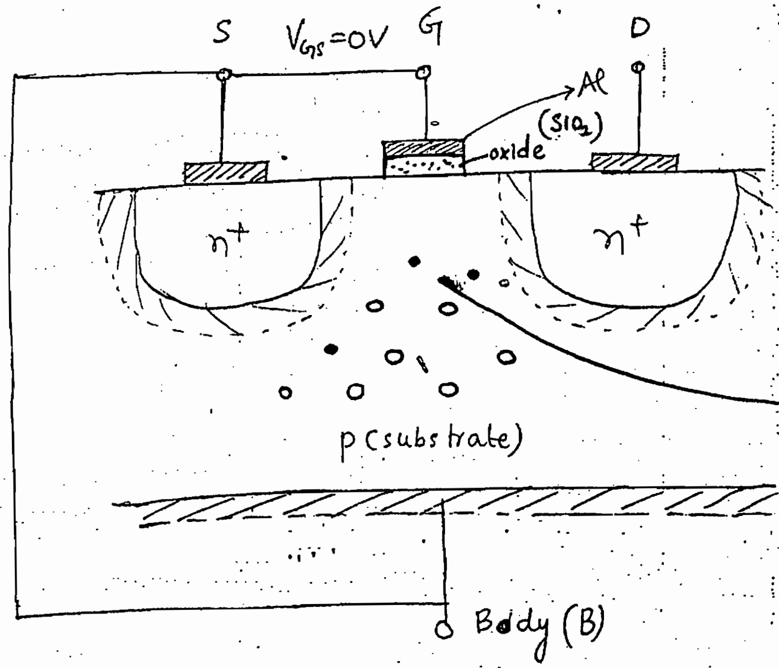
$$\boxed{A_V = g_m R_D}$$

Conclusions:

		Z_i	Z_o	A_V
(1).	CS bypass	R_g	$r_d // R_D$	$-g_m R_D$
(2).	CS unbpass	R_g	$r_d + (1+\mu) R_s // R_D$	$\dfrac{-g_m R_D}{(1+g_m R_s)}$
(3).	CD	R_g	$R_s // \dfrac{1}{g_m}$	$\dfrac{g_m R_s}{(1+g_m R_s)}$
(4).	CG	$R_s // \dfrac{1}{g_m}$	$r_d // R_D$	$+g_m R_D$

MOSFET

n-channel Enhancement MOSFET (NMOS) :



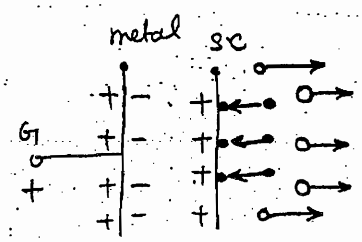
Holes → majority charge carriers

Electrons → minority charge carriers

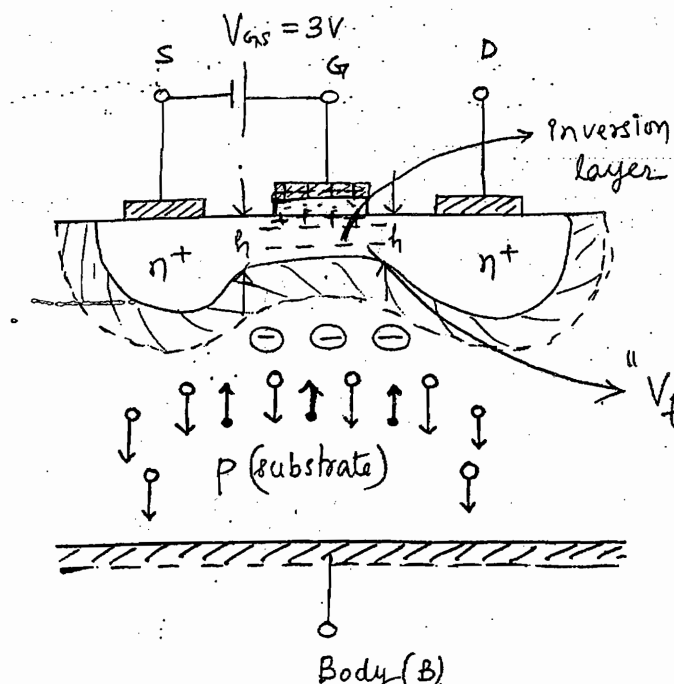
→ channel is not existing at $V_{GS}=0V$

$\therefore I_D = 0$

Mos capacitor



$h \propto V_{GS}$



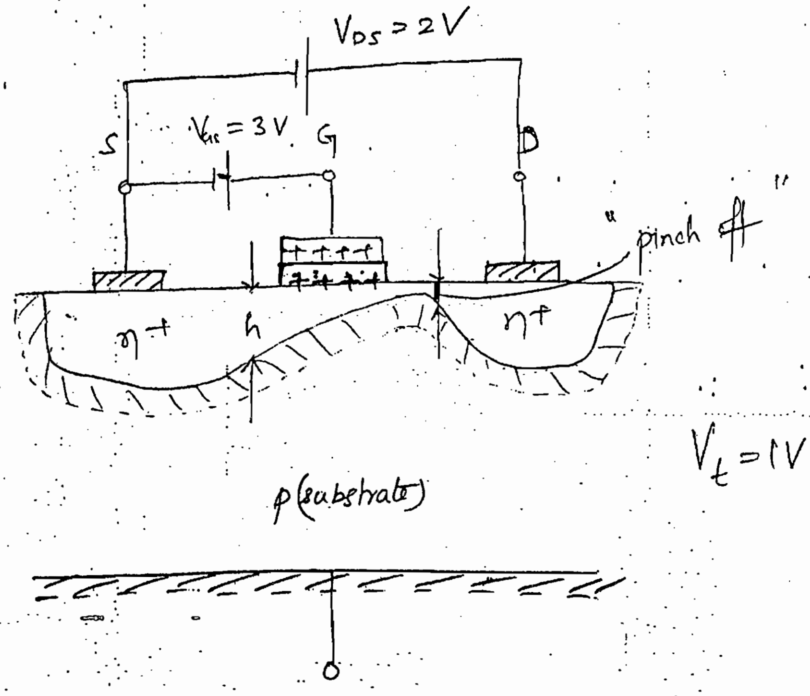
"V_t" → threshold voltage (0.5V to 1V)

$(V_{GS}-V_t)$ → Effective voltage (or) over drive voltage

- V_t is the min^m^ V_{GS} voltage at which conducting channel will be formed.

$$V_{DG} = V_{DS} + V_{SG}$$
$$= V_{DS} - V_{GS}$$
$$V_{GD} = V_{GS} - V_{DS}$$

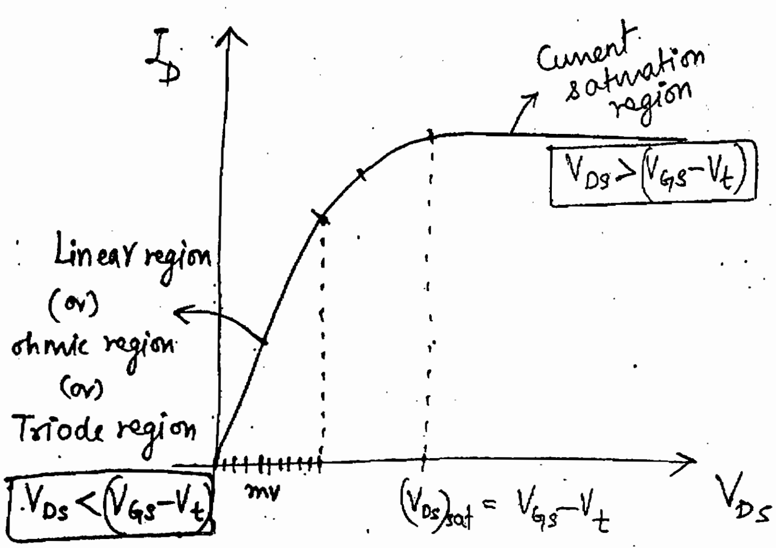
Here, $V_{DS} = 0 \Rightarrow V_{GD} = V_{GS}$



$V_t = 1V$

$$V_{GD} = V_{GS} - V_{DS}$$
$$= 3V - 2V$$
$$= 1V = V_t$$

$$(V_{DS})_{sat.} = V_{GS} - V_t$$



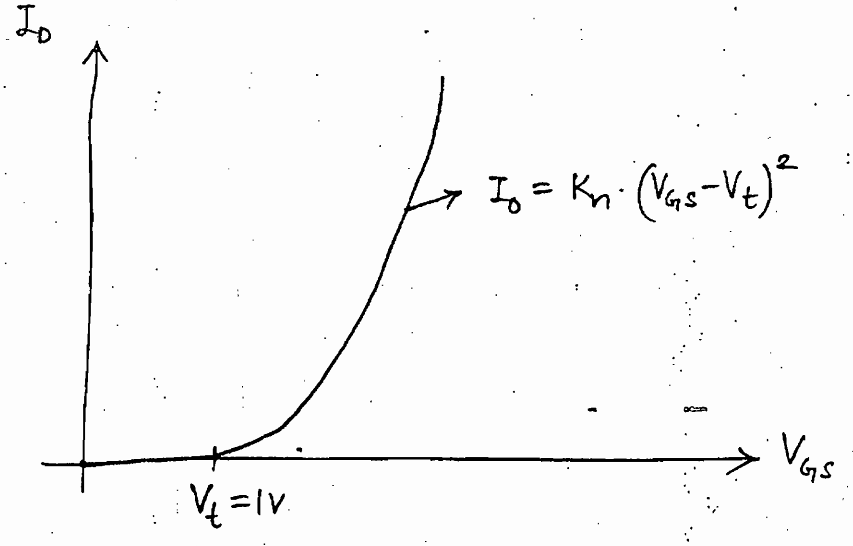
Similarity

$\underbrace{\text{Triode} \rightarrow \text{vacuum}}_{}$

voltage control

(FET, MOSFET)

Transfer characteristics:



V-I characteristic eqn of MOSFET :

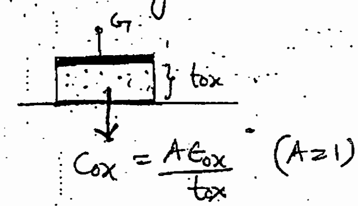
"NMOS"

$$I_D = \mu_n C_{ox} \left(\frac{W}{L}\right) \left\{ (V_{GS} - V_t) V_{DS} - \frac{1}{2} V_{DS}^2 \right\}$$

where, $\mu_n \rightarrow$ mobility of electrons

$C_{ox} \rightarrow$ capacitance of oxide layer

$$\boxed{C_{ox} = \epsilon_{ox} / t_{ox}}$$

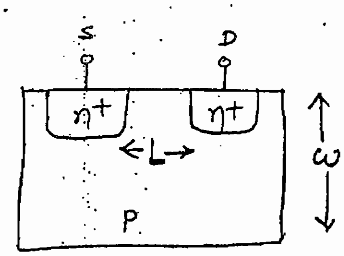


& $\epsilon_{ox} = 3.9\, \epsilon_0$

$= 3.9 \times 8.85 \times 10^{-12}$ F/m

$= 3.45 \times 10^{-11}$ F/m.

$\left(\frac{W}{L}\right) \rightarrow$ aspect ratio



$\therefore \left(\frac{W}{L}\right)$ 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:-

$V_{DS} < V_{GS} - V_t$, if V_{DS} (mV)

$$I_D = \mu_n C_{ox} \left(\frac{W}{L}\right) \cdot (V_{GS} - V_t) \cdot V_{DS}$$

$$R_{DS} = \frac{V_{DS}}{I_D} = \frac{1}{\mu_n \left(\frac{W}{L}\right)(V_{GS} - V_t)}$$

Current Saturation:

$$V_{DS} \geq V_{GS} - V_t$$

$$(V_{DS})_{sat} = V_{GS} - V_t$$

$$I_D = \underbrace{\frac{1}{2} \mu_n C_{ox} \left(\frac{W}{L}\right)}_{K_n} (V_{GS} - V_t)^2$$

K_n (Transconductance parameter) mA/V^2

* 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.

$$\boxed{I_D = K_n \cdot (V_{GS} - V_t)^2}$$

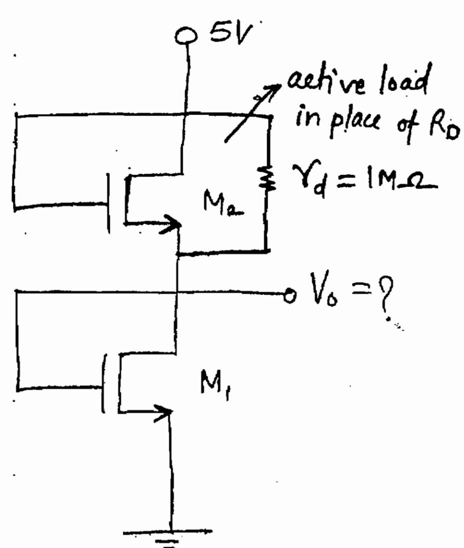
$$\therefore g_m = \frac{\partial I_D}{\partial V_{GS}} = 2K_n (V_{GS} - V_t)$$

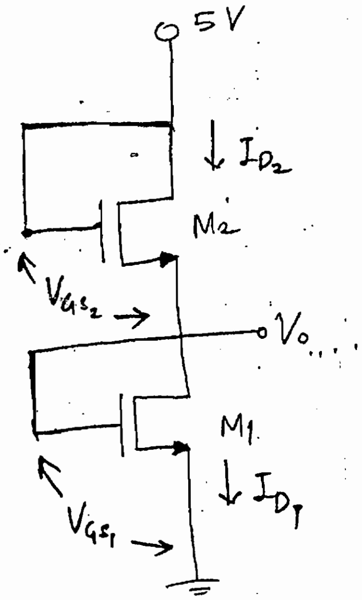
Active load MOSFET problems:

Q. given : $K_1 = 9\ mA/V^2$

$K_2 = 36\ mA/V^2$

$V_t = 1V$





$$I_{D_1} = I_{D_2}$$

$$K_1 (V_{GS_1} - V_t)^2 = K_2 (V_{GS_2} - V_t)^2$$

$$9 (V_o - 1)^2 = 36 (5 - V_o - 1)^2$$

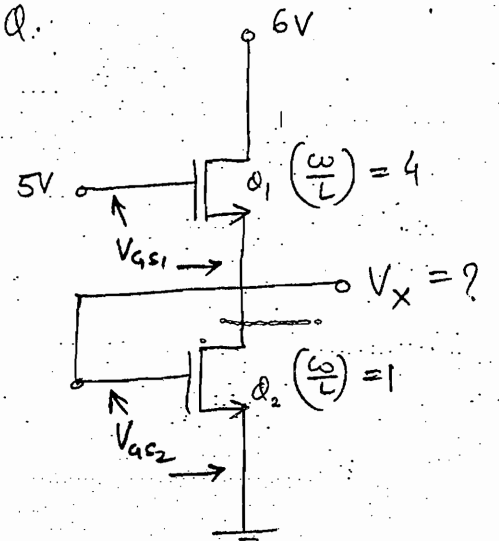
$$V_o^2 + 1 - 2V_o = 4 (V_o^2 + 16 - 8V_o)$$

$$V_o^2 - 2V_o + 1 = 4V_o^2 - 32V_o + 64$$

$$3V_o^2 \ 3\ V_o + 63 = 0$$

$V_o = 3$ Volts, $7V$ (X)

Q.



Q_1 & $Q_2 \Rightarrow \mu_n C_{ox} = 100\mu A/V^2$

$V_t = 1V$

$$I_{D_1} = I_{D_2}$$

$$\cancel{\frac{1}{2}\mu_n C_{ox}} \left(\frac{W}{L}\right)_1 (V_{GS_1} - V_t)^2 = \cancel{\frac{1}{2}\mu_n C_{ox}} \left(\frac{W}{L}\right)_2 (V_{GS_2} - V_t)^2$$

$$4 (5 - V_X - 1)^2 = (V_X - 1)^2$$

$$4 (V_X - 4)^2 = V_X^2 + 1 - 2V_X$$

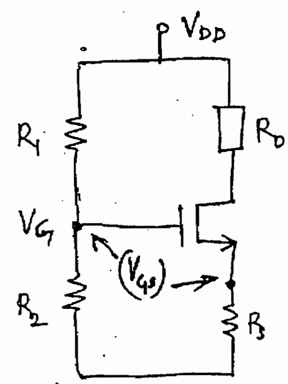
$$4V_X^2 + 64 - 32V_X = V_X^2 - 2V_X + 1$$

$$3V_X^2 - 30V_X + 63 = 0$$

$\therefore V_X = 3$ volts

Biasing:

MOSFET Biasing:



$$I_D = K_n (V_{GS} - V_t)^2$$

$\downarrow$

should be independent of K_n and V_t

$$V_G = \underbrace{V_{GS}}_{\text{small} \rightarrow \text{neglected.}} + I_D R_S \Rightarrow V_G \approx I_D R_S$$

$$\Rightarrow \boxed{(I_D)_Q = \frac{V_G}{R_S}}$$

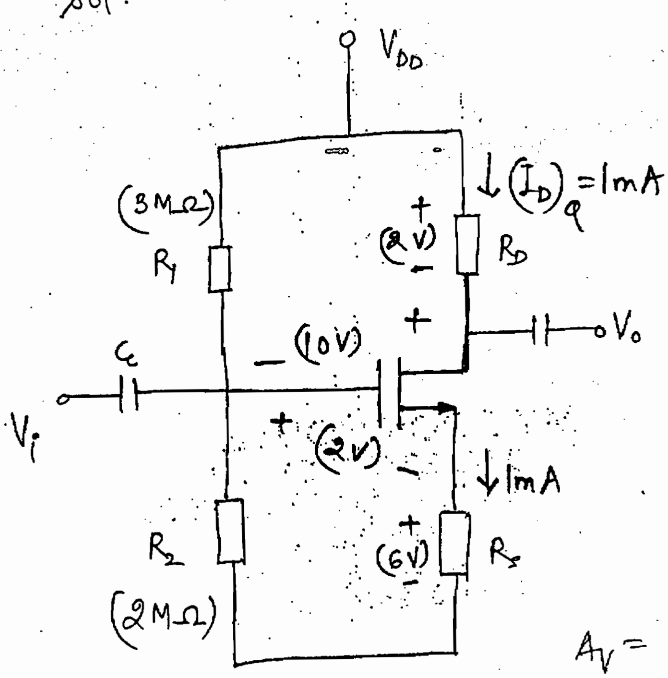
& $\boxed{(V_{GS})_Q = \sqrt{\frac{(I_D)_Q}{K_n}} + V_t}$

* 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:
$V_{DD} = 20\,V$, $(I_D)_Q = 1\,mA$, $K_n = 1\,mA/V^2$, $V_t = 1V$
$(V_{DS})_Q = 10V$.
Calculate the voltage gain of NMOS amplifier.

Sol.



$$(V_{GS})_Q = \sqrt{\frac{1}{1}} + 1 = 2V$$

$$R_S = \frac{6V}{1mA} = 6K$$

$$R_D = \frac{2V}{1mA} = 2K$$

$$V_G = V_{DD} \cdot \frac{R_2}{R_1 + R_2}$$

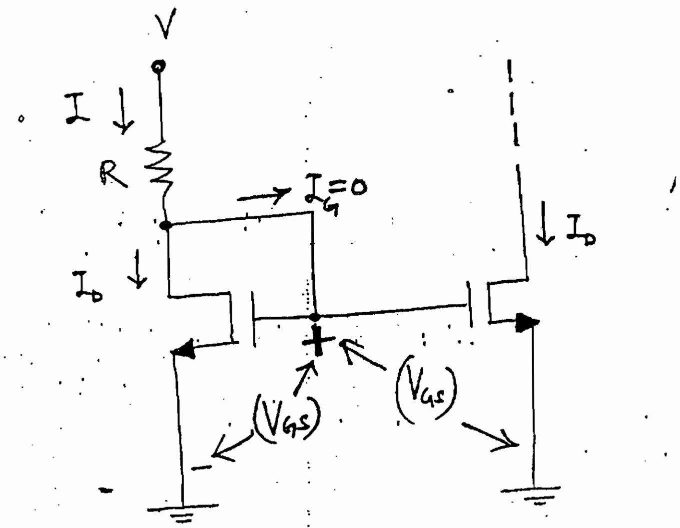
$$\frac{R_2}{R_1+R_2} = \frac{8}{20} = \frac{2}{5}$$

$$A_V = -g_m R_D$$

$$g_m = 2K_n (V_{GS} - V_t)^2$$
$$= 2(1)(2-1)^2$$
$$= 2\,m℧$$

$$A_V = -2 \times 2$$
$$= -4$$

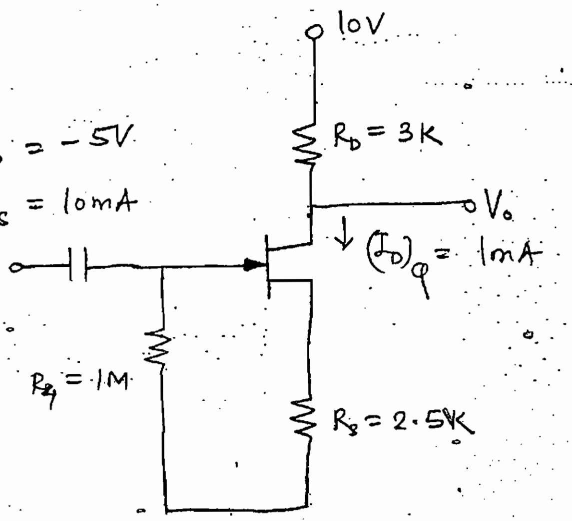
Current mirror biasing :



Q11.

$V_p = -5V$

$I_{DSS} = 10mA$

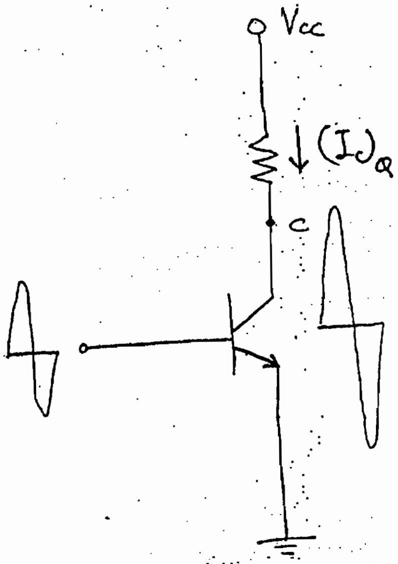


$$(V_{GS})_Q = -(I_D)_Q \cdot R_s$$
$$= -1mA \cdot 2.5k\Omega$$
$$= -2.5$$

$$g_m = -\frac{2 I_{DSS}}{V_p}\left(1 - \frac{(V_{GS})_Q}{V_p}\right]$$
$$= -\frac{2\times 10}{-5}\left[1 + \frac{2.5}{5}\right]$$
$$= 2$$

$$A_v = -g_m R_D = -2 \times 3 k\Omega = -6$$

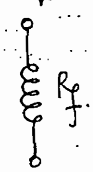
Power Amplifiers



Power diagram

$P_{DC} = V_{cc} \cdot (I_c)_Q$

- P_{R_c} DC
  - $(I_c)_Q^2 \cdot R_c$
    - $\underbrace{\text{External power dissipation}}$



- P_{tr} DC
  - P_{ac}
  - $(P_c)_{DC}$
    - $(V_{CE})_Q \cdot (I_c)_Q$
      - $\underbrace{\text{Internal power dissipation}}_{\text{Heat Sink}}$

$$\% \eta = \frac{P_{ac}}{P_{DC}}$$

$P_{DC} = P_{ac} + \text{losses}$

$P_{DC} = P_{ac} + P_D$

if $P_D \to 0 \Rightarrow \eta \to 100\%$

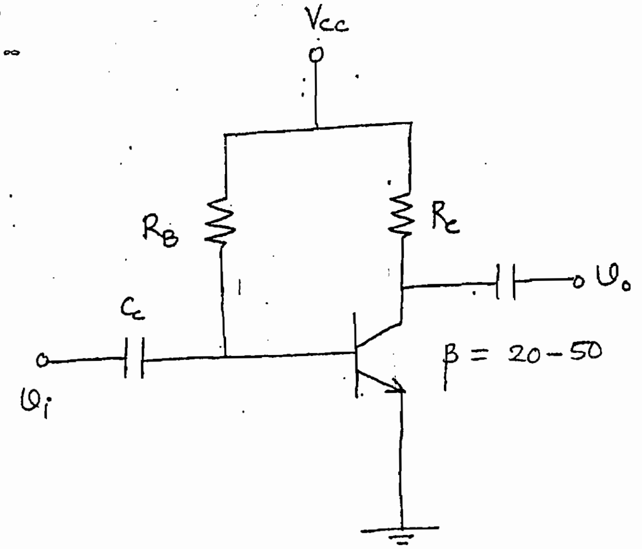
Power amplifier is a device which converts DC power to ac power and whose action is controlled by the i/p s/g.

Class A amplifier :

(1). Class A series fed power amplifier

(2). Class A X^r coupled power amplifier

(1). Class A series fed power amplifier :



Base width is high, so recombination high, I_B is high so β ~~is less~~.

DC i/p → $P_{DC} = V_{cc} \cdot (I_c)_Q$

AC o/p → $P_{ac} = V_{rms} \cdot I_{rms}$

$$= \frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}}$$

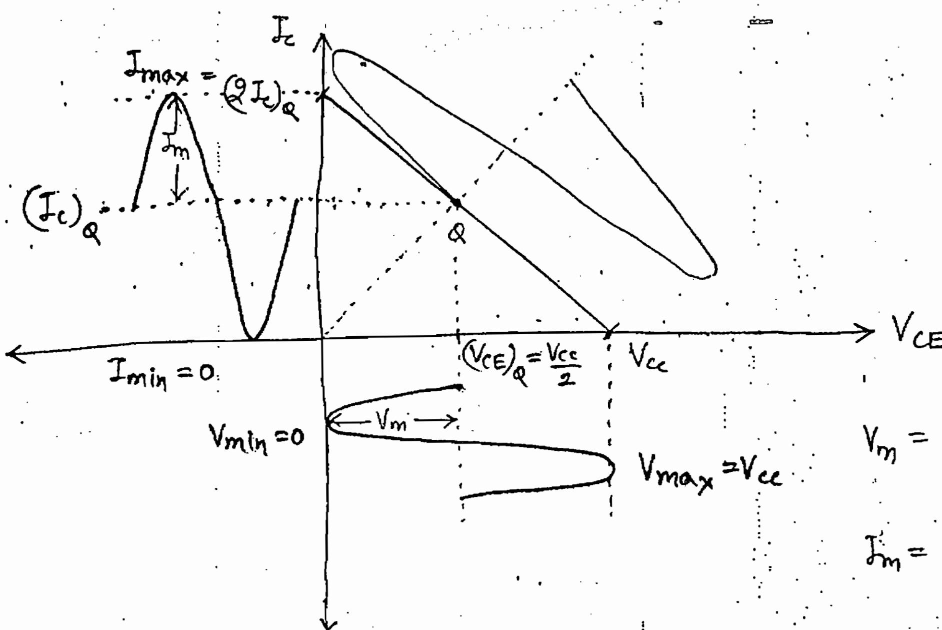
$$= \frac{V_m I_m}{2}$$

DC conditions :

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

$$(I_c)_Q = \beta (I_B)_Q$$

$$(V_{CE})_Q = V_{CC} - (I_c)_Q \cdot R_c$$



$$V_m = \frac{V_{max} - V_{min}}{2}$$

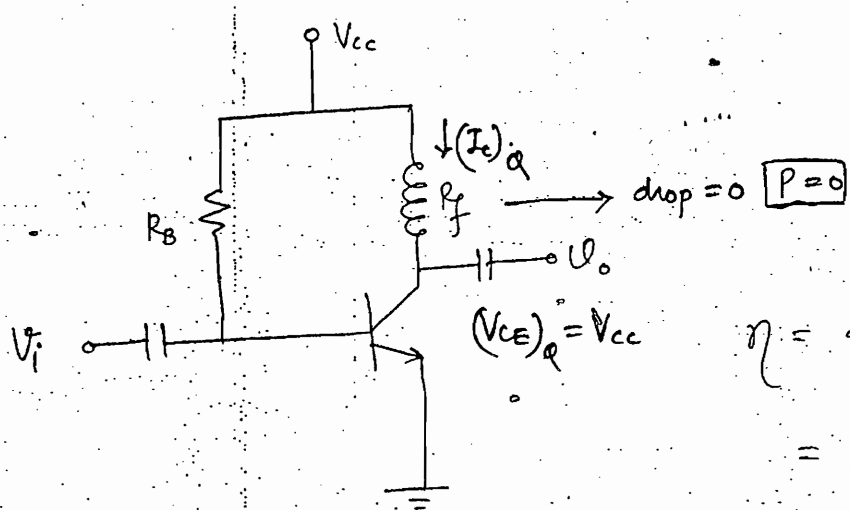
$$I_m = \frac{I_{max} - I_{min}}{2}$$

$$\eta = \frac{P_{ac}}{P_{DC}} = \frac{1}{8} \cdot \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{V_{ce} \cdot (I_c)_Q}$$

$$= \frac{1}{8} \cdot \frac{(V_{cc} - 0)[(2 I_c)_Q - 0]}{V_{ce} \cdot (I_c)_Q}$$

$$= \frac{1}{8} \cdot 2 = \frac{1}{4} = 25\%$$

(2). X^r coupled :



$$\eta = \frac{2 V_{cc} \times 2 (I_c)_Q}{8 \times V_{cc} \cdot (I_c)_Q} = \frac{1}{2} = 50\%$$

Conclusion :

In class A series fed power amplifier the efficiency max^m is 25%. To improve this efficiency, replace the resistor R_c with an inductor R_f where the efficiency goes to 50%.

Power dissipation –

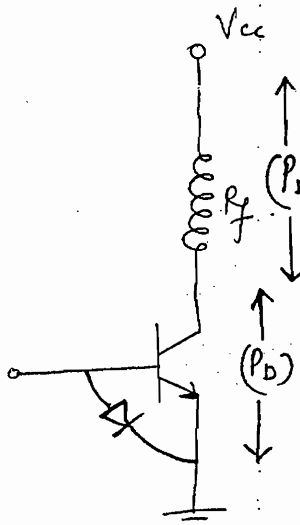
$$P_{DC} = P_{AC} + P_D$$

$$P_D = P_{DC} - P_{AC}$$

$$= V_{ce} \cdot (I_c)_Q - \frac{V_m I_m}{2}$$

At zero s/g conditions, $P_D = P_{DC}$ $\quad (\because P_{ac} = 0)$

Class B power amplifiers :



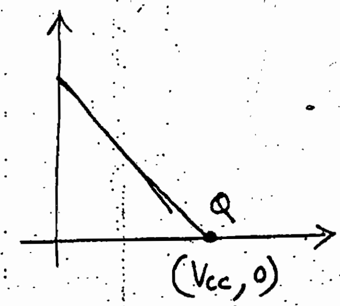
$(P_D)_{external} \approx 0$

$(P_D)_{internal} = (V_{CE})_Q \cdot (I_C)_Q \neq 0$

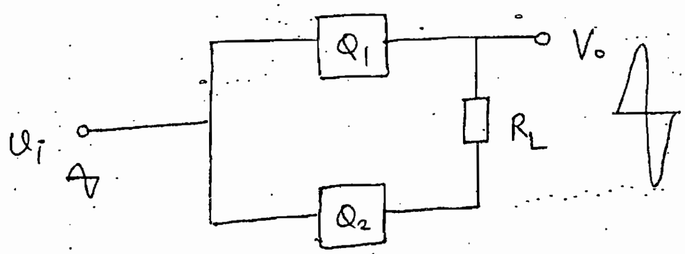
$(V_{CE})_Q \neq 0 \quad (\because V_{cc})$

$(I_C)_Q = 0$

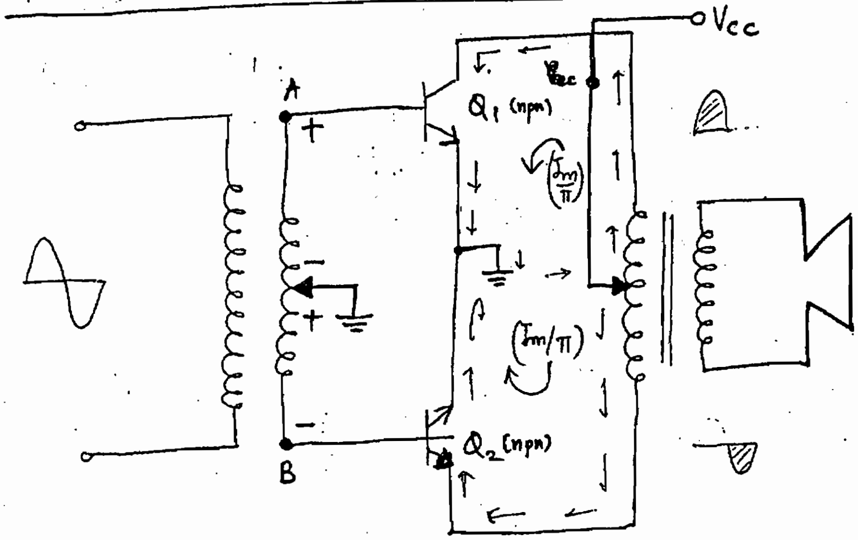
Device should sleep when there is no s/g.



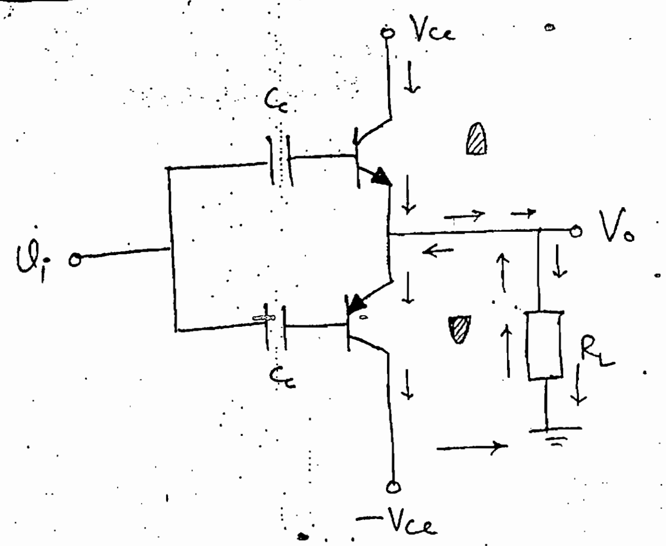
Respond to only +ve s/g



Push pull class B power amplifier : (Single Battery)



Complementary symmetry class B : (2 Batteries)



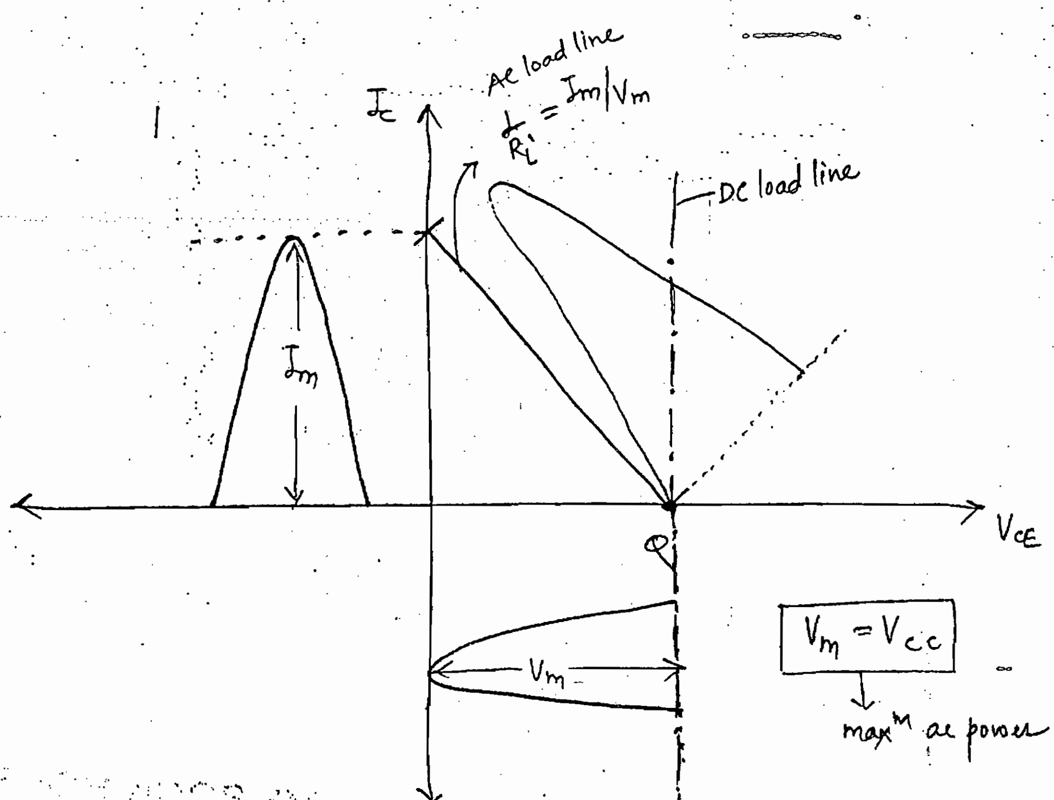
$(I_c)_Q = 0$

$P_{DC} = V_{ce} \cdot I_{dc}$

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

Efficiency : $P_{ac} = \frac{V_m I_m}{2}$

$$\% \eta = \frac{P_{ac}}{P_{DC}} = \frac{V_m I_m / 2}{V_{ce} \cdot 2 I_m / \pi} = \frac{V_m \pi}{4 V_{cc}}$$



$$\% \eta = \frac{V_{ce} \cdot \pi}{4 \, V_{ce}} = \frac{\pi}{4} = 78.5\%$$

Power dissipation in class B :

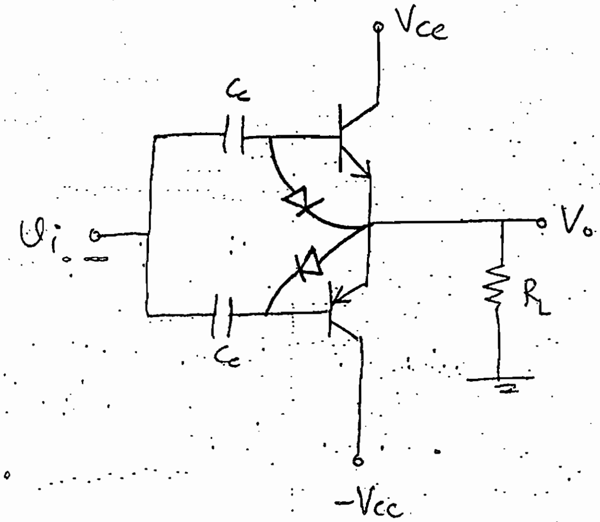
$$P_D = P_{DC} - P_{ac}$$

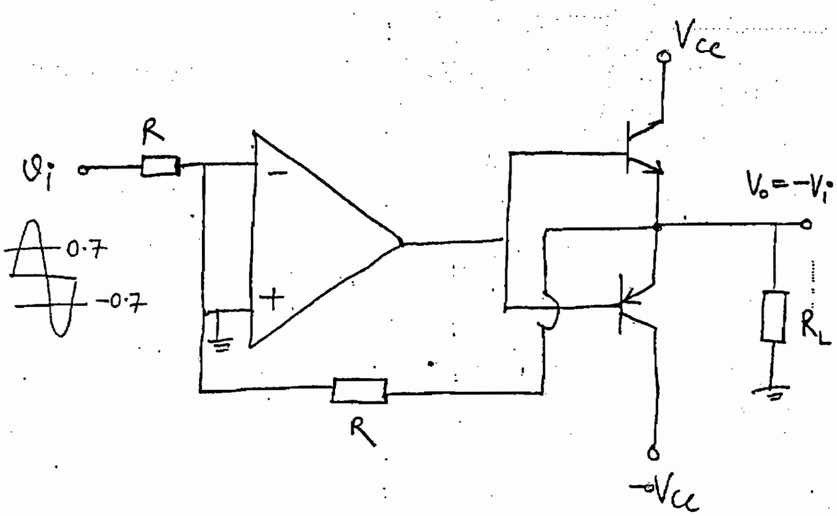
$$= V_{ce} \cdot \frac{2 I_m}{\pi} - \frac{V_m I_m}{2}$$

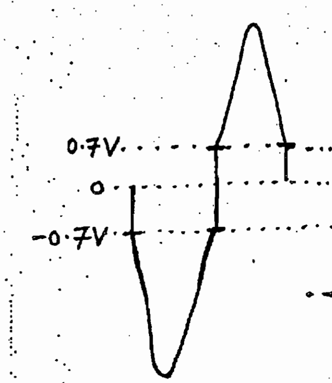
At zero s/g conditions, $V_m = 0$, $I_m = 0$.

$P_D = 0$ (Sleeping mode)

Cross over distortion:







Oscillators

$$\frac{V_{o1}}{V_i} = \frac{2(1-\alpha)\,S^2R^2C^2}{S^2R^2C^2 + 2\alpha SRC + 1} \quad \text{(H.P.)}$$

$$\frac{V_{o2}}{V_i} = \frac{-2(1-\alpha)\,SRC}{S^2R^2C^2 + 2\alpha SRC + 1} \quad \text{(B.P.)}$$

$$\frac{V_{o3}}{V_i} = \frac{2(1-\alpha)}{S^2R^2C^2 + 2\alpha SRC + 1} \quad \text{(L.P.)}$$

Band-pass ⇒

$$\frac{V_{o2}}{V_i} = \frac{-2(1-\alpha)\,SRC}{S^2R^2C^2 + 2\alpha SRC + 1}$$

$$= \frac{-2(1-\alpha)\,SRC}{-\omega^2 R^2C^2 + 2\alpha SRC + 1} \qquad \boxed{\omega_0 = \frac{1}{RC}}$$

$$= \frac{-2(1-\alpha)\,SRC}{2\alpha SRC}$$

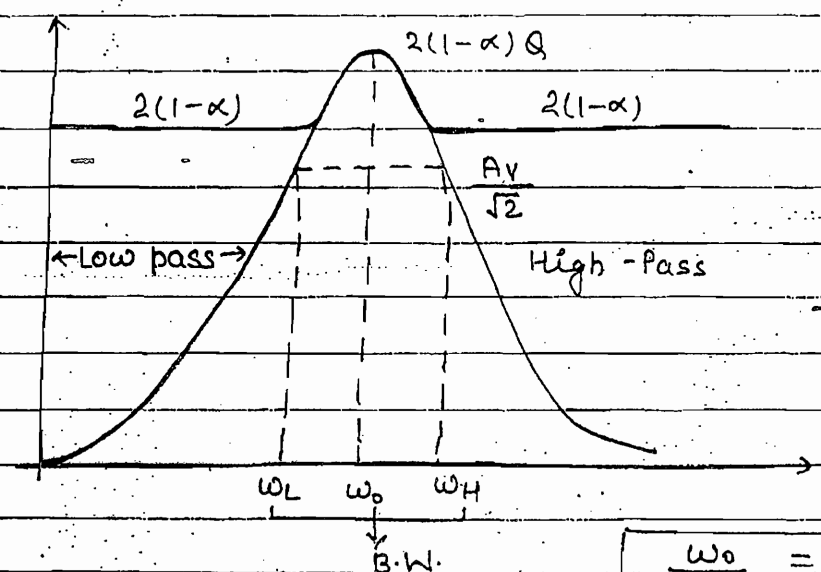
$$= \frac{1-\alpha}{\alpha} = -2(1-\alpha)\,Q$$

$Q \rightarrow$ Quality factor ; $\left\{ Q = \frac{1}{2\alpha} \right\}$ pole Q

$$\omega_o = \frac{1}{RC} \quad \ldots \text{ Pole Frequency.}$$

$$\frac{V_{o1}}{V_i} = \frac{2(1-\alpha)\dfrac{s^2}{\omega_o^2}}{\dfrac{s^2}{\omega_o^2} + \dfrac{s}{\omega_o Q} + 1}$$

$$\frac{V_{o3}}{V_i} = \frac{2(1-\alpha)}{\dfrac{s^2}{\omega_o^2} + \dfrac{s}{\omega_o Q} + 1}$$



$$\boxed{\frac{\omega_o}{B.W.} = Q}$$

Pole- Equation ⇒

$$\frac{s^2}{\omega_o^2} + \frac{s}{\omega_o Q} + 1 = 0 \qquad \text{Assume } \frac{s}{\omega_o} = x$$

$$\rightarrow \quad x^2 + \frac{x}{Q} + 1 = 0$$

$$\frac{s_{1,2}}{\omega_o} = \frac{-\dfrac{1}{Q} \pm \sqrt{\left(\dfrac{1}{Q}\right)^2 - 4(1)\times(1)}}{2 \times 1}$$

$$\frac{s_{1,2}}{\omega_o} = \frac{-\dfrac{1}{Q} \pm \sqrt{\dfrac{1}{Q^2} - 4}}{2}$$

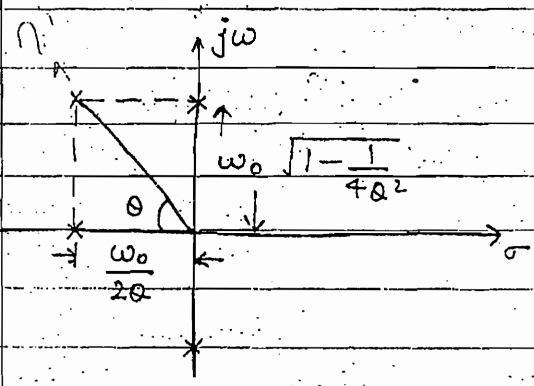
Real-part $\Rightarrow \frac{1}{Q^2} > 4$, $Q < \frac{1}{\sqrt{4}}$; $\boxed{Q < \frac{1}{2}}$

Imajinary part $\Rightarrow$ $\boxed{Q > 1/2}$

$$\Rightarrow \frac{S_{1,2}}{\omega_0} = \frac{-\frac{1}{Q} \pm j\sqrt{4 - \frac{1}{Q^2}}}{2}$$

$$\Rightarrow \frac{S_{1,2}}{\omega_0} = \frac{-1}{2Q} \pm j\sqrt{1 - \frac{1}{4Q^2}}$$

$$\Rightarrow S_{1,2} = \frac{-\omega_0}{2Q} \pm \omega_0 j\sqrt{1 - \frac{1}{4Q^2}}$$



$$\theta = \tan^{-1} \frac{\omega_0\sqrt{1 - 1/4Q^2}}{\omega_0/2Q}$$

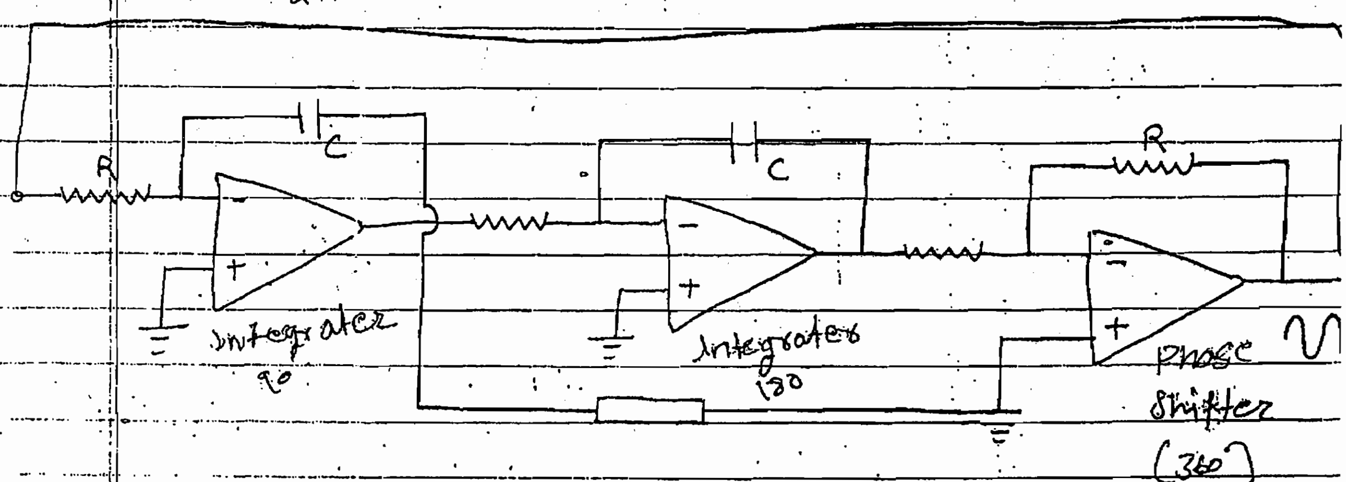
$$\theta = \tan^{-1} 2Q\sqrt{1 - 1/4Q^2}$$

$Q \to \infty$.

$\theta = \tan^{-1} \infty$

$\boxed{\theta = 90^\circ}$

$Q = \frac{1}{2\alpha} = \infty$, $\alpha = 0$



Harmonic Oscillator

(or)

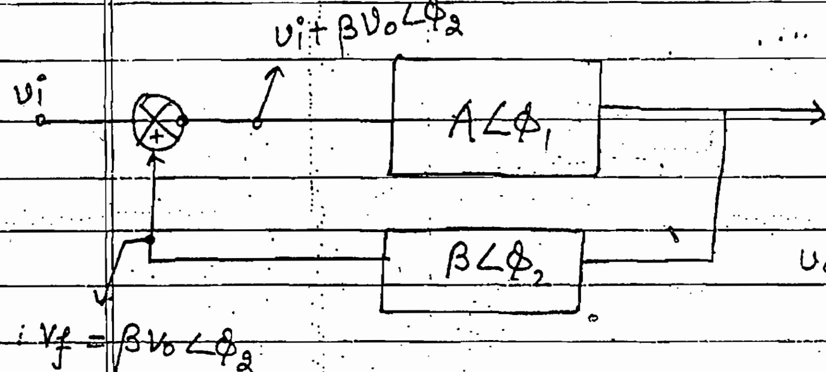
Quadrature Oscillator [α should be zero then pole will come on img axis] quadrature

$$\frac{\partial^2 v_o}{\partial t^2} + \underbrace{\frac{\alpha \partial v_o}{\partial t}}_{0} + \gamma v_o = \underbrace{V_i}_{0}$$

$$\boxed{\frac{\partial^2 v_o}{\partial t^2} + \gamma v_o = 0}$$

harmonic Oscillators :

* Positive feedback Concept ⇒



$$V_o = A\angle\phi_1 \{V_i + \beta V_o \angle \phi_2\}$$
$$= A V_i \angle \phi_1 + A\beta V_o \angle \phi_1 + \phi_2$$

$$V_o(1 - A\beta \angle \phi_1 + \phi_2) = A\angle\phi_1 V_i$$

$$\frac{V_o}{\underset{0}{V_i}} = \frac{A\angle\phi_1}{\underbrace{1 - A\beta\angle\phi_1+\phi_2}_{0}} \qquad 1 - A\beta\angle\phi_1+\phi_2 = 0$$

conditions ⇒ $|A\beta| = 1$

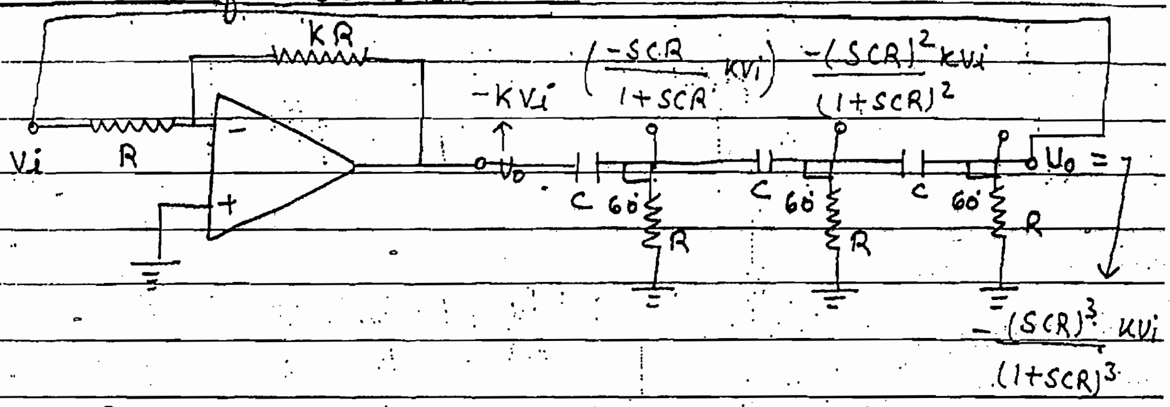
$$A = \frac{V_o}{V_i}, \qquad \beta = \frac{V_f}{V_o}$$

$$A\beta = \frac{\cancel{V_o}}{V_i} \cdot \frac{V_f}{\cancel{V_o}} = \frac{V_f}{V_i} = 1 \qquad \boxed{V_f = V_i}$$

$\angle\phi_1 + \phi_2 = 0^\circ$ (or) multiples of 2π.

Barkhusen criteria.

* RC- Phase Shift Oscillators ⇒



$$-\frac{(SCR)^3}{(1+SCR)^3} K\cancel{V_i} = \cancel{V_i}$$

→ $(1+SCR)^3 + (SCR)^3 K = 0$

→ $1 + 3s^2C^2R^2 + 3SCR + (SCR)^3(K+1) = 0$

→ $1 - 3\omega^2C^2R^2 + 3j\omega CR - j\omega^3R^3C^3(K+1) = 0$

$1 - 3\omega^2C^2R^2 = 0$ (Real part)

$$\left\{ \omega_0 = \frac{1}{\sqrt{3}\,RC} \right\}$$

$3\omega CR - \omega^3R^3C^3(k+1) = 0.$

→ $3 - \omega^2R^2C^2(k+1) = 0.$ ← at gain = 8 will get on img. axis (poles)

→ $\omega_0 = \frac{1}{\sqrt{3}RC}$

$3 - \frac{1}{3}(k+1) = 0$, $\boxed{K = 8}$

Conclusions ⇒

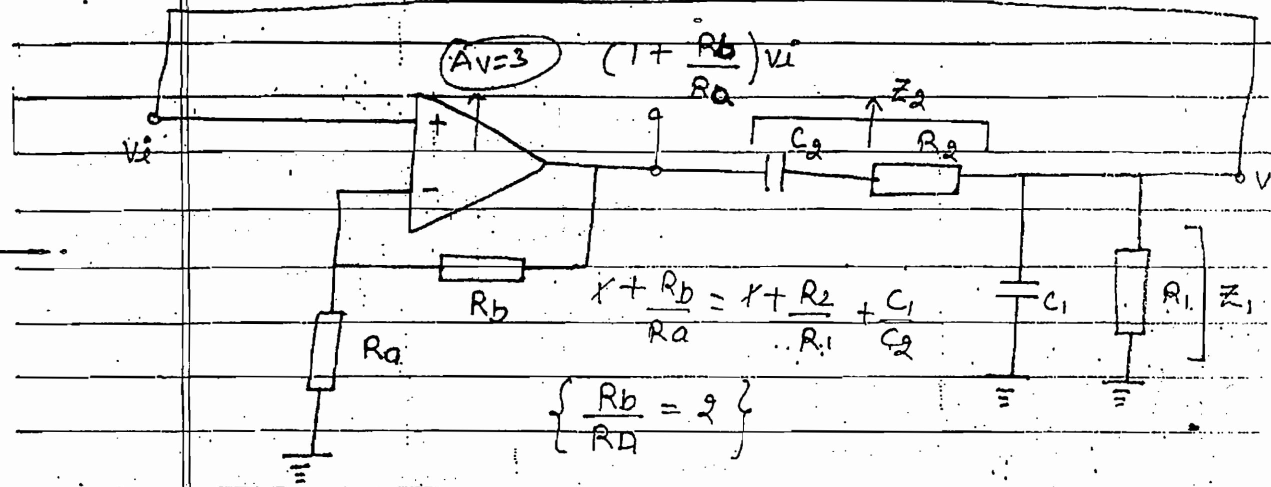
- RC- phase shift osci. are used for audio freq. ranges (20 Hz - 20 KHz).
- Quality factor for RC- n/w will be in multiples of 10.
- RC- phase shift Oscill. can't be design for dynamic range of frequency.

classmate

Date
Page

$$\left(1+\frac{R_b}{R_a}\right)\left(1+\frac{R_2}{R_1}+\frac{C_1}{C_2}\right)=1$$

* Wein-Bridge Oscillator ⇒



$$1+\frac{R_b}{R_a}=1+\frac{R_2}{R_1}+\frac{C_1}{C_2}$$

$$\left\{\frac{R_b}{R_a}=2\right\}$$

$$V_0=\frac{\left(1+\frac{R_b}{R_a}\right)V_i\cdot Z_1}{Z_1+Z_2}\Rightarrow\frac{\left(1+\frac{R_b}{R_a}\right)V_i}{1+Z_2Y_1}$$

$$V_0=\frac{\left(1+\frac{R_b}{R_a}\right)V_i}{1+\left(R_2+\frac{1}{sC_2}\right)\left(\frac{1}{R_1}+sC_1\right)}\Rightarrow\frac{\left(1+\frac{R_b}{R_a}\right)V_i}{1+\frac{R_2}{R_1}+\frac{C_1}{C_2}+\frac{1}{sC_2R_1}+sC_1R_2}$$

$$V_0=\frac{\left(1+\frac{R_b}{R_a}\right)V_i}{1+\frac{R_2}{R_1}+\frac{C_1}{C_2}+j\left[\frac{-1}{\omega C_2R_1}+\omega C_1R_2\right]}$$

j-term ⇒

$$\frac{-1}{\omega C_2R_1}+\omega C_1R_2=0$$

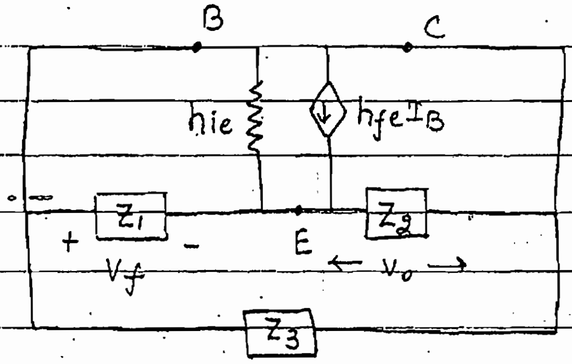
$$\omega_0=\frac{1}{\sqrt{R_1R_2C_1C_2}}$$

$R_1=R_2=R$, $C_1=C_2=C$.

$$\omega_0=\frac{1}{RC}$$

$$f_0=\frac{1}{2\pi RC}$$

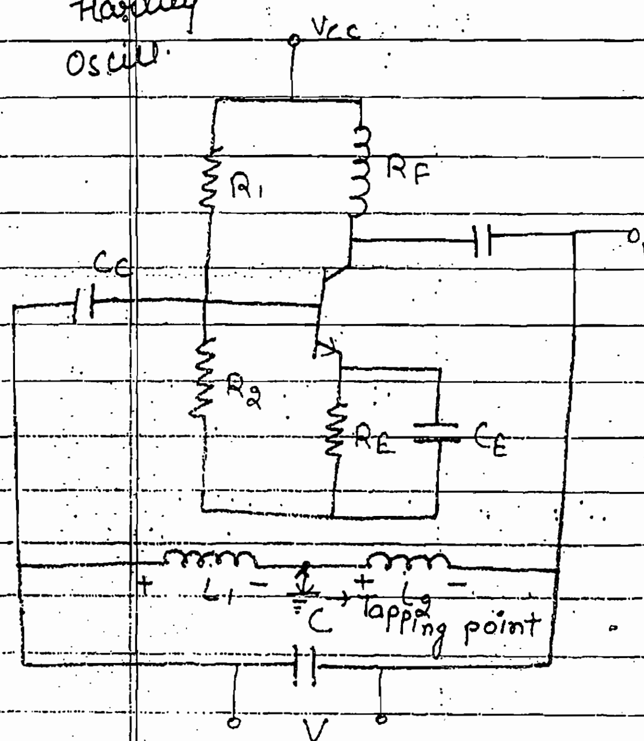
3). LC - Oscillators ⇒



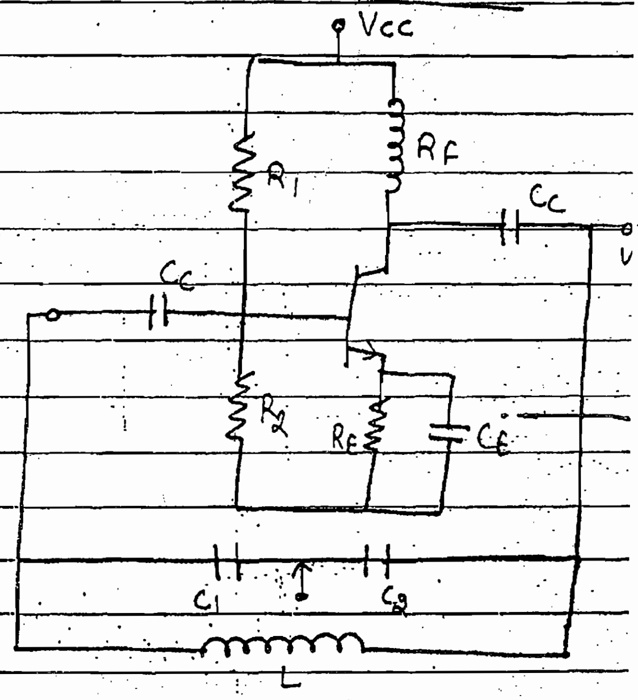
Condition

Z_1 & Z_2 are same reactive components. Z_3 is oppositive compo.(reactive only).

Hartley Oscill.



Colpitts Oscill.



Analysis of LC - oscillators ⇒

Q:1 What is the imp. of inductor (R_F) in LC oscill. ckt?

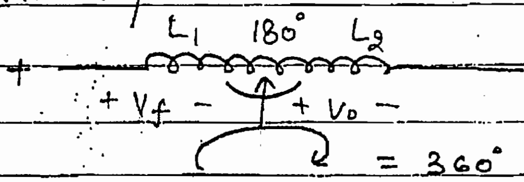
Ans: Inductor R_F is used as an isolation b/w dc - supply & ac-signal.

Q:2 Explain phase relation in LC oscill. ?

Ans:

Solu^n ⇒ Hartley ⇒

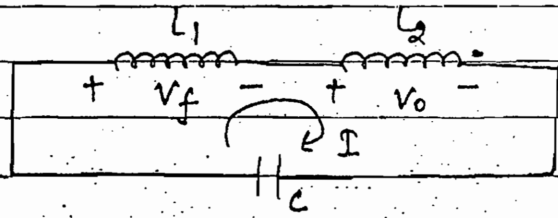
Q ↓ 180°



$= 360°$

Ques: 3 Explain conditions for gain A and β for Hartley Osci.

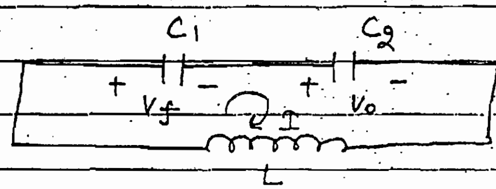
Solu^n ⇒



$\beta = \frac{V_f}{V_o} = \frac{I X_{L_1}}{I X_{L_2}} = \frac{L_1}{L_2}$

$\Rightarrow \left\{ A = \frac{L_2}{L_1} \right\}$

Colpitts ⇒



$\beta = \frac{V_f}{V_o} = \frac{I X_{C_1}}{I X_{C_2}} = \frac{C_2}{C_1}$

$\Rightarrow \boxed{A = \frac{C_1}{C_2}}$

Ques: 3 Explain expressions for freq. of oscill. in LC ckt?

Solu^n ⇒ $f = \frac{1}{2\pi\sqrt{L_T C}} = \frac{1}{2\pi\sqrt{L_T C}}$

$L_T = L_1 + L_2 + 2M$, $M \rightarrow$ negligible

$L_T = L_1 + L_2$

Colpitts ⇒ $f = \frac{1}{2\pi\sqrt{L C_T}}$; $\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2}$.

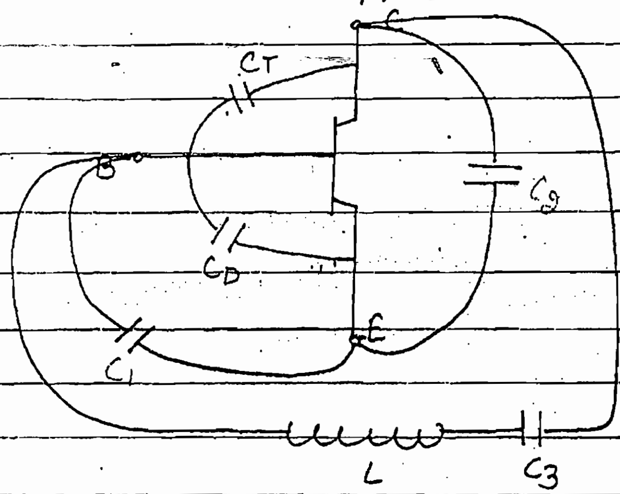
Ques: Why LC oscilla. can't be use for low freq. of oscilla.

Solu^n ⇒ $f\downarrow = \frac{1}{2\pi\sqrt{LC}\uparrow}$; $f \rightarrow$ less, LC compo. become high.

* RC never use for high freq.

Ques Explain about clapp oscillator design ?

Soln ⇒



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

$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

$C_1 >> C_3$, $C_2 >> C_3$

$$\frac{1}{C_T} = \frac{1}{C_3}$$

C_1, C_2 are getting effect by high freq. [C_D and C_T Problem] but C_3 is free from them

$$\left\{ f = \frac{1}{2\pi\sqrt{LC_3}} \right\}$$

- A quality factor of an LC-n/w is in multiples of 100. This LC oscill. can stablized the freq. upto 5MHz.

- modify version of colpitts is called clapp oscilla. which can stablized freq. upto 10 MHz.

- To establized freq. above 10 MHz crystal design will play a role which is known as piezoelectric effect principle. when a mechanical force is applied one face of crystal, a electrical signal will be generate 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 amplifier

O/P Roqp

- differential Amplifier
- op amp. application ...