276

(276)

8

-: HAND WRITTEN NOTES:-

OF (1)

ELECTRICAL ENGINEERING

-: SUBJECT:-

ANALOG ELECTRONICS

2

Topic →

1) Semi conductor physics. 3

2) P N jn diode.

Special diodes
- ✓ Zener diode
- ✓ Tunnel diode
- → Schottky diode
- → photo diode

Applications →
- Rectifiers
- filters

Diode ckts →
- Ideal diode problems.
- Practical diode problems
- clippers
- clampers.

BJT —
- BJT device analysis
- BJT biasing (DC)

Analog ckts —
- Small signal analysis
- low freq Analysis
- High freq Analysis
- freq response of an amplifier.
- large signal Amplifiers. (power amplifiers)
- multistage amplifiers

Feedback —
- Feedback amplifiers
- Oscillators

FET / MOSFET :-

4

- Fet biasing
- Small signal analysis
- low freq analysis
- High freq analysis

} Problems.

OP-Amps -

- Differential amplifiers.
- Op amp applications.
- 555 timer
- Active filters

Semi conductor physics -

Introduction :→

$1s^2\ 2s^2\ 2p^6$

Cu - 29

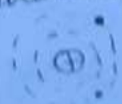
(S)

valence Shell

Unstable atom.

Mg - $1s^2\ 2s^2\ 2p^6\ 3s^2$

→ conductors.



Al - 13

d - 10

f - 14.

Q. The max no. of e^- that can be filled in the valence shell of an atom will be

a) $4e^-$, b) $6e^-$ c) $19e^-$ d) None [$8e^-$].

Ans: Acc. to Aufbau principle, s, p, d, s, p

Si - 14

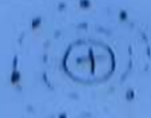
Ge - 32

T=0K → Insulators }→ S.C.

T≠0K → conductors

As - 33

Br - 35



← Insulators

Ne - 10. $1s^2\ 2s^2\ 2p^6$.

→ Stable atom :

* The no. of e^- that can be filled in the valence shell of a semiconductor will be ——

Ans. 4 e^-

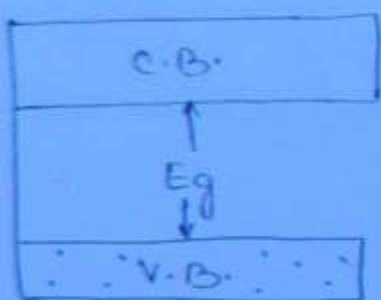
6

Examples of Semiconductor :—

$\underbrace{\text{Si, Ge}}_{\text{Single crystalline}}$; $\underbrace{\text{GaAs}}_{\text{compound semi conductors}}$ [Gallium Arsenic]

Q. Why Si and Ge are generally preferred compare to Gallium Arsenid (GaAs)?

Ans.



C.B. → conduction band
Eg → en. gap
V.B. → valence band.

* At t = 0K,

E_{g0}

Ge – 0.785 eV

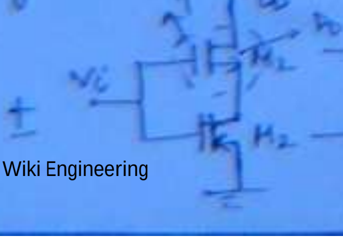
Si – 1.21 eV

GaAs – 1.58 eV.

As the en. gap value for Silicon and germanium is less compare to ~~gall~~ GaAs, we expect more conduction is possible in case of Silicon and Germanium.

Q. Why GaAs is used in 'CMOS' technology?

Ans.



M_1 — saturation (ON) → cut off (OFF)

M_2 → cut off (OFF) → saturation (ON)

} switching application.

movement of carrier (majority) is called mobility. -

1) The mobility of charge (μ) carriers in case of GaAs > Si and Ge.

2) The temp with standing capability is more for GaAs (Eg is more)

→ Typical values of temp. for which the device can withstand with :-

Ge → 100°C

Si → 200°C

GaAs > 200°C

Q. Why $\mu_e > \mu_h$ μ → mobility

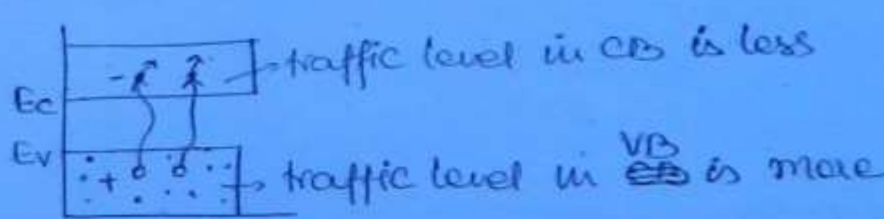
(a) The trafic level in CB < in V.B.

b) The trafic level in CB > VB

c) The trafic level in CB = VB

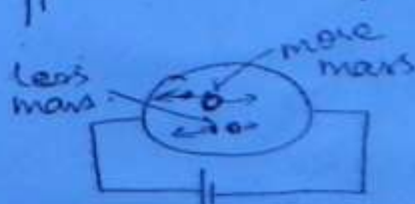
d) None

Ans.



2nd reason, when things are in motion

The effective mass of a hole is always greater than effective mass of an e⁻.

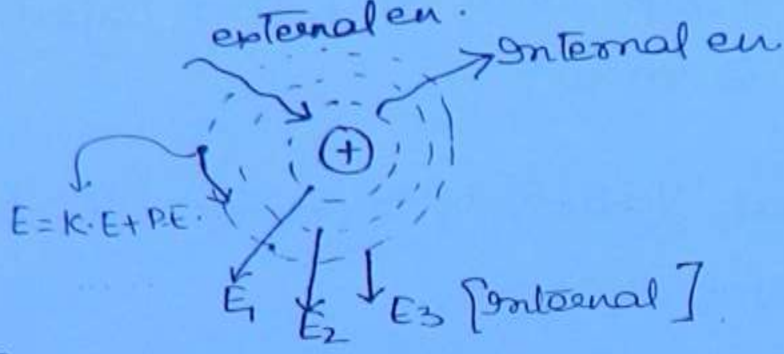


Traffic level is more in Ge, Si compare to GaAs.

mobility of Ge, Si < mobility of GaAs.

Atomic structure of an atom :—

Q. What is the relation b/w E_1, E_2, E_3.



a) $E_1 < E_2 < E_3$
b) $E_1 > E_2 > E_3$.
c) $E_1 = E_2 = E_3$
d) none.

Ans. E_1, E_2 and E_3 are internal orbit energies.

Int. en.

$E_n \rightarrow e^-$ revolving around the ~~molecules~~ nucleus.

$$= \frac{-13.56}{n^2} \text{ eV}$$

$$E_1 = \frac{-13.56}{1^2} = -13.56 \text{ eV}$$

$$E_2 = \frac{-13.56}{2^2} = -3.5 \text{ eV}$$

$$\vdots$$

$$E_\infty = \frac{-13.56}{\infty^2} = 0 \text{ eV}.$$

$n = \infty$ - - - - - - - - - 0 eV

$n = 3$ - - - - - -1.5 eV

$n = 2$ - - - - - - -3.5 eV

$n = 1$ - - - - - - - - -13.56 eV

en. diagram for an isolated atom

* The valance cell of an atom will have the highest en. level.

Ions →

There are two types of ions :—

1) +ive ions

2) -ive ions

+ive ions →

When the atom looses an e^-, the atom is characterised as +ive ion.

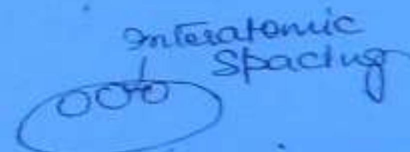
-ive ion :—

When the atom gains an e^-, the atom is characterised as -ive ion.

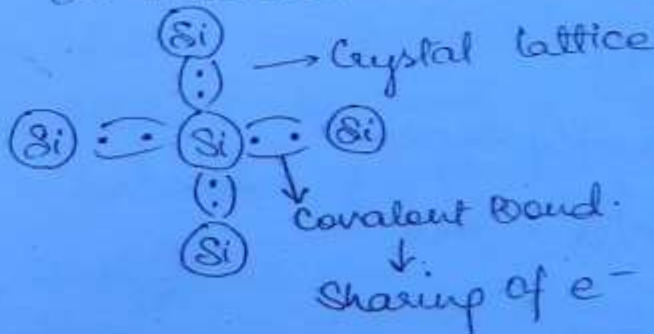
NOTE -

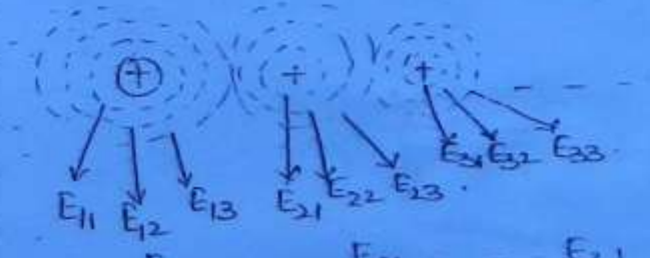
Ions are indirectly atoms which are immobile in nature.

Energy band diagram concept :—



Ex :— Si material





ignore E_{11}, E_{21}, E_{31} - -

E_{12}, E_{22}, E_{32} - -

because outer levels interact with each other.

E_{11} - - - E_{21} - - - E_{31}

E_{12} - - - E_{22} - - - E_{23}

E_{13} - - - E_{23} - - - E_{33}

CONCLUSIONS :—

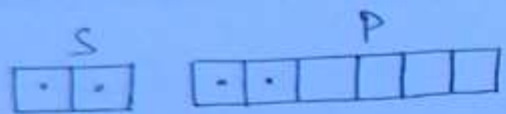
1) The above en. diagram is valid for isolated atoms.

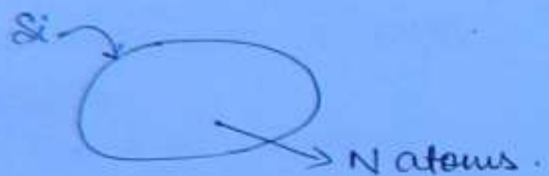
2) This type of diagram is not valid for material concept because the interatomic spacing b/w the atoms is very less.

10

Si = 14.

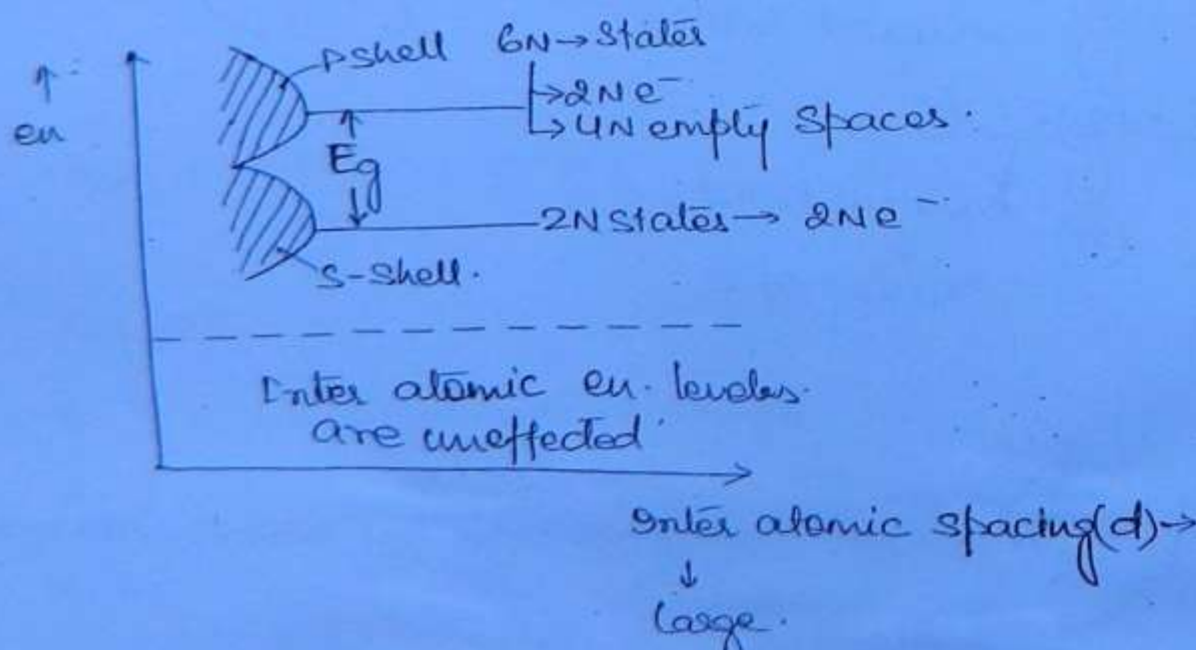
$1s^2 2s^2 2p^6 3s^2 3p^2$



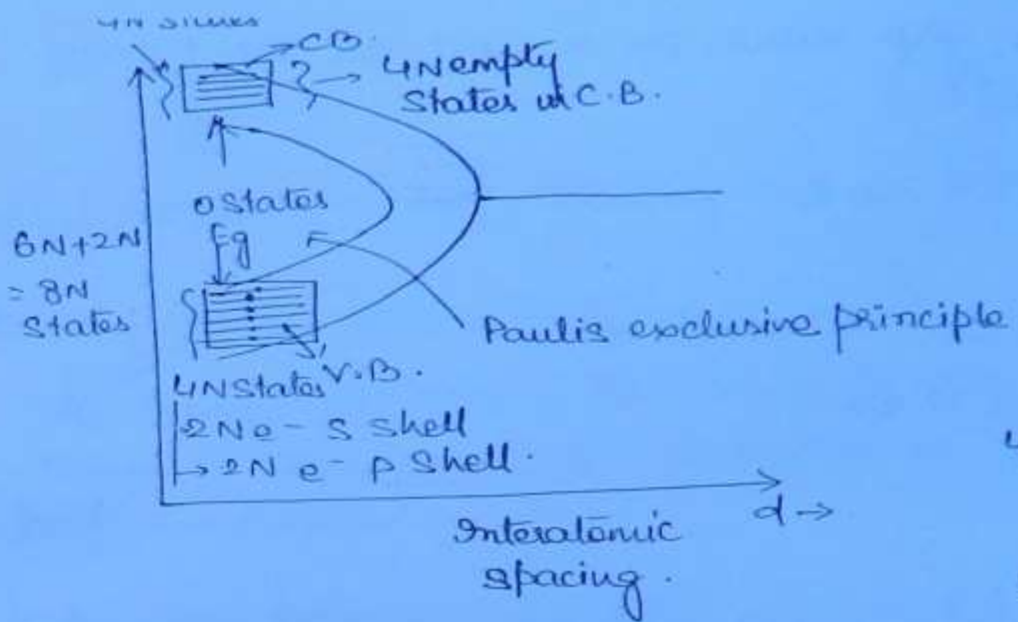


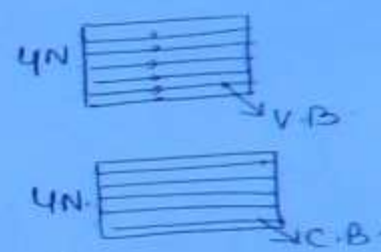
2N States → 2N e^-
(S Shell)

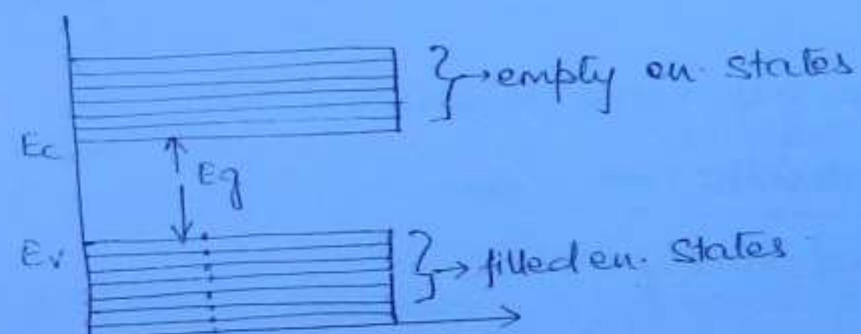
6N States → 2N e^-
→ 4N empty states.
(P Shell)



Millman *
Neamen.







$E_c \rightarrow$ lowest en. level in C.B.

$E_V \rightarrow$ highest en. level in V.B.

$E_g \rightarrow$ Forbidden gap or en. gap.

energy gap Vs temp :-

* $\boxed{E_g(\text{at any temp in K}) = E_{go} - \beta T}$

E_g ↓ en. gap at any temp in K.

E_{go} ↓ en. gap at 0K.

Ge − 0.785 eV.
Si − 1.21 eV.

$\beta \rightarrow$ const for a material

$\beta(Si) = 3.6 \times 10^{-4}$
$\beta(Ge) \rightarrow 2.23 \times 10^{-4}$

$T \rightarrow$ temp in K.

Q: Cal. the en. gap value for Si and Ge at room temp. [T = 27°C]

Ans. $T = 273 + 27 = 300\ K$.

Si

$$E_g(300K) = 1.21 - 3.6\times10^{-4}\times300$$
$$= 1.12\ ev.$$

Ge

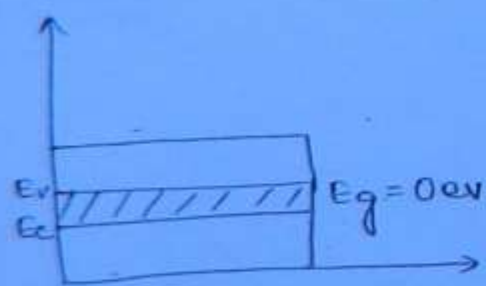
$$E_g(300K) = 0.785 - 2.23\times10^{-4}\times300$$
$$= 0.72\ ev.$$

NOTE →

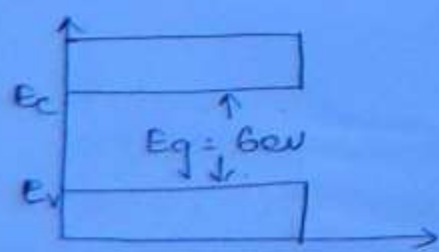
As temp increases, en. gap value decreases.

Classification of materials :→

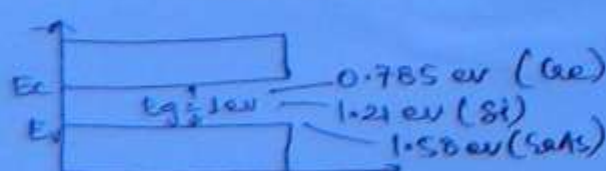
- based on en. band theory.



Conductor.



Insulator.



Semiconductor.

C → 2ev going toward Insulator.

$1s^2 \underbrace{2s^2 2p^2}_{IV}$ $E_g = 2ev$.

Q) Why Carbon is not behaving like a semiconductor?

Ans. E_g for carbon = 2ev.

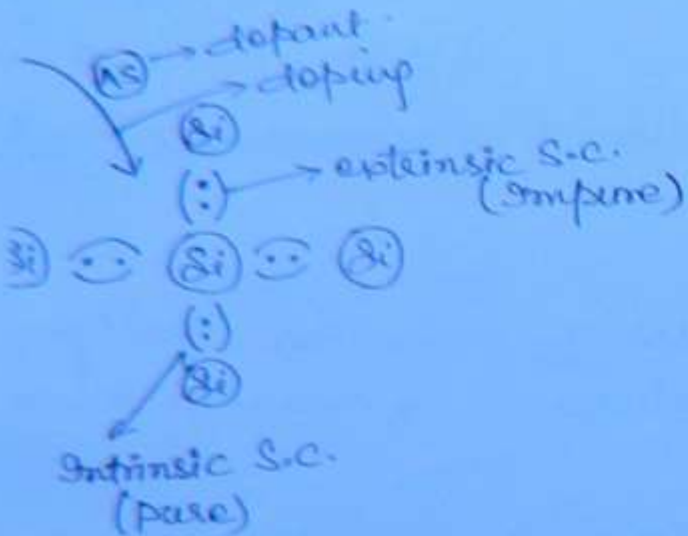
As energy gap value for carbon is more than the typica value of semiconductor. Sometimes it is behaving like a perfect insulator.

* Differences between conductors, Insulators and semiconduc

Property	Conductors	Insulators	Semi-conductors.
1) Resistivity	less $10^{-8}\ \Omega cm$	Highest $10^{12}\ \Omega cm$	10^{-4} to $10^{3}\ \Omega cm$
2) type of bonding	Metallic (free)	ionic & covalent	covalent
3) Energy gap	0ev	6ev	1ev
4) temp coeff. of resistance	positive	—	negative.
5) Charge Carriers	e^-	—	electron & holes

Properties of semiconductors →

1) Resistivity of a semiconductor is more than that of a conductor but less than of an insulator.
2). It is having -ive temp co-eff. of Resistance property.
3) Doping is possible in it.



When an impurity is added to an intrinsic S.C., its conducting property changes.

Note → Doping is a process which improves the conductivity of a semiconductor.

Unit of en :—

J

$1 ev = 1.6 \times 10^{-19}\,C \times 1V$

$= 1.6 \times 10^{-19}\,J$

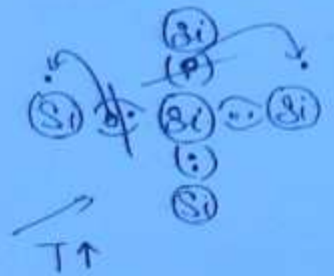
1 ev

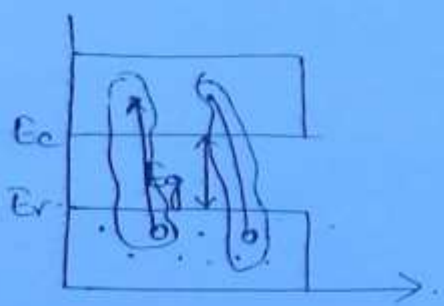
The amount of en. required for an e^- to fall through a pot diff. of 1V is called as '1 ev.'

Semiconductors →

- → Intrinsic S.C. (pure
- → Extrinsic S.C. (impure)

Intrinsic S.C.
(pure)





$n = P$

n = conc. of e^-/cm^3

P = conc. of e^- holes$/cm^3$.

$n = P = n_i$

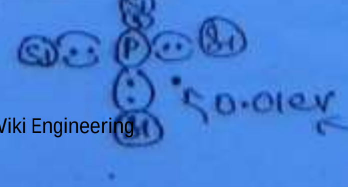
$n_i \rightarrow$ Intrinsic conc. (e-h) $/cm^3$

Extrinsic S.C.
(Impure)

Intrinsic S.C. + Other material = Extrinsic S.C.

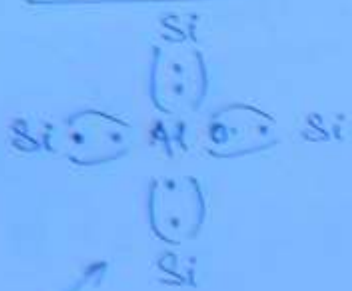
Other material → dopants
- → Pentavalent impurity – Arsenic, Antimony, Bismuth, Phospho etc.
- → Trivalent Impurity – eg – Gallium, Boron, Al, Indium.

Pentavalent



maj carriers → e^-
minority carriers → holes.
n type S.C.

Trivalent –



maj. carriers – holes
min. carriers – e^-
P type S.C.

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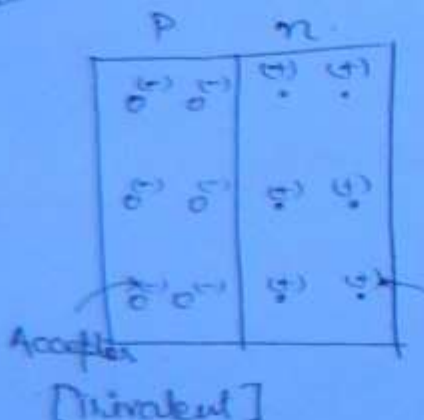
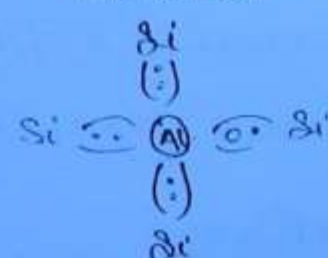


Pentavalent

Trivalent



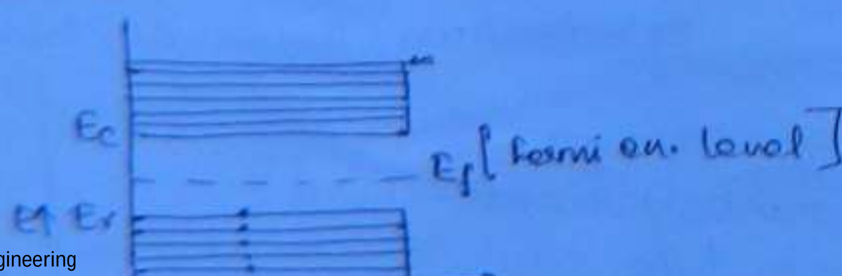
En. band diagram in a semiconductor :→

Fermi Dirac probability f^n :→

"In energy band diagram, the probability that the en. level of e^- is given by a f^n called as fermi f^n defined as

$$F(E) = \frac{1}{1 + e^{(E-E_f)/KT}}$$

At t = 0K :



T = 0K,

$F(E) = 1 \qquad E < E_f$

$\quad = 0 \qquad E > E_f$

$\quad = X \qquad E = E_f$

indeterminate

12

$$\boxed{F(E) = \frac{1}{1+e^{(E-E_f)/KT}}}$$

At $E = E_F$,
T = 0K

$F(E) = \frac{1}{1+e^{0/0}} \rightarrow$ Indeterminate

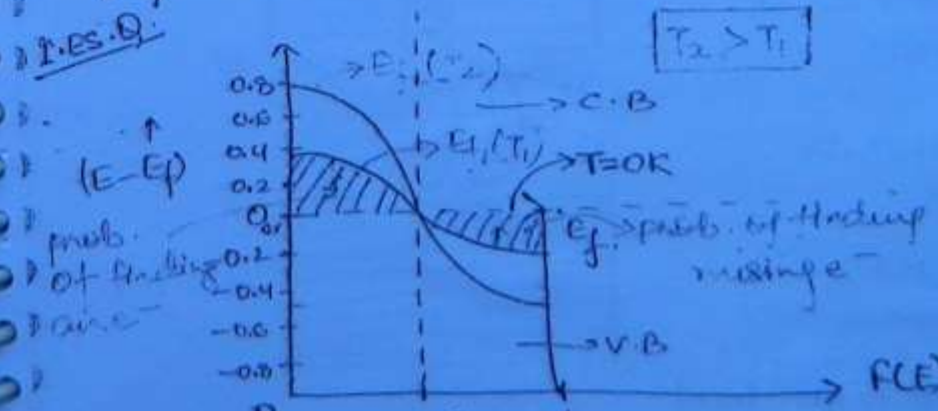
Empty states

$\rightarrow E_f$

filled states

Fermi level :→
It is the ref. en. level which separates filled states and empty states ~~at 0K~~.

I.ES.Q.



$T \neq 0K$,

$F(E) = 0.5 \quad E = E_f$

$$F(E) = \frac{1}{1+e^{(E-E_f)/KT}}$$

$$= \frac{1}{1+1}$$

$$= 0.5$$

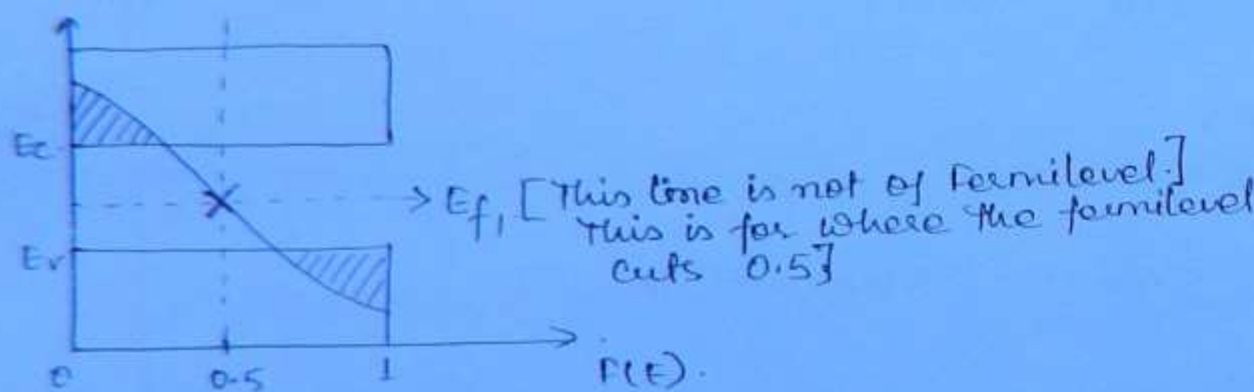
1) At t=0K, the fermilevel line will be shown like a flat line.

2) At t≠0K, the fermilevel will be represented like a curvature nature.

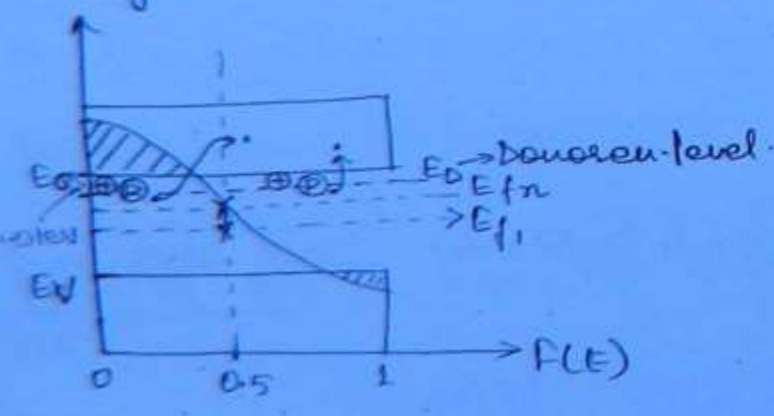
• Fermilevel is not a const level and depends on doping.

Energy band diagrams :-

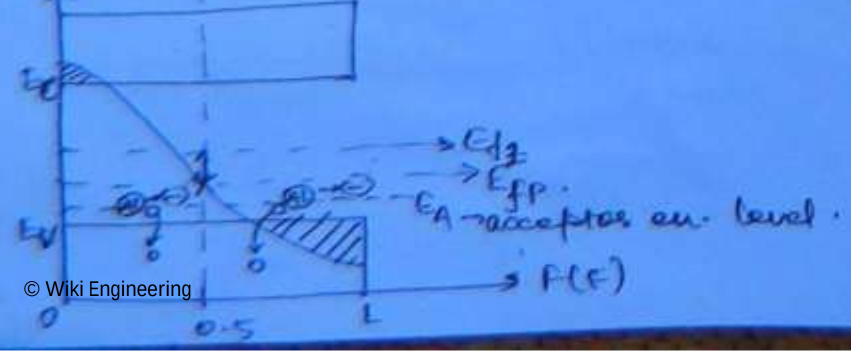
Intrinsic S.C. →



N-type



P-type →



Mathematical Analysis →

$$n = N_c e^{-(E_c - E_f)/KT}$$
$$P = N_v e^{-(E_f - E_v)/KT}$$
→ Fermi Dirac prob. f^n.

19

n → conc. of e^-

P → conc. of holes.

N_c → effective density of states in C.B.

N_v → Eff. density of states in V.B.

E_c → lowest en. level in C.B.

E_v → highest en. level in V.B.

E_F → Fermi level

K → Boltzmann const $= 1.38 \times 10^{-23}$ J/K.

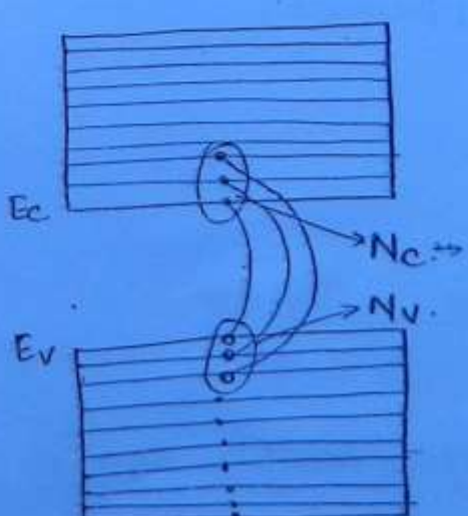
T → Temp in Kelvin.

$$\frac{K}{q} = \frac{1.38 \times 10^{-23} \text{ J/K}}{1.6 \times 10^{-19} \text{ C}}.$$

$$\frac{K}{q} = 8.65 \times 10^{-5} \text{ eV/K}.$$

N_c & N_v →

$T \neq 0K$.



N_c → The density of en. states where the e^- are filled.

N_v → The density of en. states where the e^- are missing

$N_c \approx N_v$ → intrinsic S.C. [material perfect]

$N_c \neq N_v$ → intrinsic S.C. [material imperfect]

$\boxed{N_c = N_v \rightarrow \text{Ideal condition}}$

$N_c = 4.82 \times 10^{15} \left(\frac{m_n}{m}\right)^{3/2} T^{3/2} /cm^3$

$N_V = 4.82 \times 10^{15} \left(\frac{m_p}{m}\right)^{3/2} T^{3/2} /cm^3$.

If $\boxed{m_n = m_p} \Rightarrow N_C = N_V$.

$m_n \rightarrow$ Effective mass of e^- in C.B.

$m_p \rightarrow$ Effective mass of proton [i.e holes]

$m \rightarrow$ Rest mass of an e^-

$= 9.18 \times 10^{-31}$ kg.

Expression for fermilevel in case of Intrinsic Semiconductor :→

$n = P$.

$N_C e^{-(E_C - E_f)/KT} = N_V e^{-(E_f - E_V)/KT}$

$\Rightarrow KT \ln \frac{N_C}{N_V} = E_C + E_V - 2E_f$

$\Rightarrow E_f = \frac{E_C + E_V}{2} - KT \ln\left(\frac{N_C}{N_V}\right) \Rightarrow N_C \neq N_V$ [Imperfect]

$E_f = \frac{E_C + E_V}{2} \Rightarrow N_C \approx N_V$ [perfect]

if $N_C > N_V$, $E_f' < E_f [0.5]$.

if $N_C < N_V$, $E_f' > 0.5$.

Expression for fermilevel in Extrinsic S.C. :→

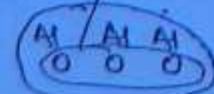
N type →

$n \approx N_D$ [Donor conc]

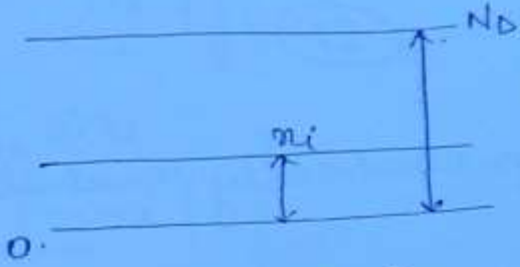
eg:-

P type →

$P = N_A$ [Acceptor conc].



mini conc. level in S.C. = $n = P = n_i$ (intrinsic S.C).



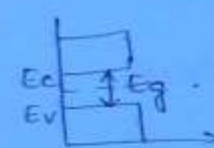
$N_D >> n_i$

Law of mass Action :→ [Applied for all type of. S.C.]

$\boxed{n \cdot P = \text{const}}$

$n = N_c e^{-(E_c - E_f)/KT}$

$P = N_v e^{-(E_f - E_v)/KT}$



$n_0 P_0 = N_c N_v e^{-(E_c - E_v)/KT}$

$= N_c N_v e^{-E_g/KT}$

$= n_i^2$

$$N_c N_v e^{-E_g/KT} = \underbrace{4.82 \times 10^{15} \left(\frac{m_n}{m}\right)^{3/2}}_{C} T^{3/2} \underbrace{4.82 \times 10^{15} \left(\frac{m_p}{m}\right)^{3/2}}_{C} T^{3/2} \times e^{-E_{g0}/KT} \underbrace{e^{\beta/K}}_{C}$$

$= A_0 T^3 e^{-E_{g0}/KT}$

$= n_i^2$

Intrinsic S.C. :—

$n \cdot P = n_i^2$

$n = P$

n type →

$n \cdot P = n_i^2$

majority minority

P type

$n \cdot P = n_i^2$

minority → majority

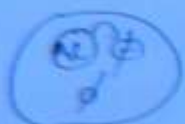
Law of mass action is valid for all types of s.c.
[Intrinsic, n-type, P-type]

(22)

Charge neutrality equation :→

"Any part of a S.C. bar is always electrically neutral"


⇒ neutral.


⇒ neutral.

Net charge density = 0.

Total +ve charge density = total -ve charge densities

n	P	N_D	N_A
$-$	$+$	$\oplus$	$\ominus$

$$\boxed{\underset{+}{P} + \underset{\oplus}{N_D} = \underset{-}{n} + \underset{\ominus}{N_A}}$$

N - type →

$$P + N_D = n + N_A$$

$$n >> P \; ; \; N_A \approx 0$$

$$\frac{n_i^2}{n} + N_D = n + 0$$

$$\Rightarrow n^2 - N_D n - n_i^2 = 0$$

$$\Rightarrow n = \frac{N_D}{2} \pm \sqrt{\left(\frac{N_D}{2}\right)^2 + n_i^2}$$

$$n = \frac{N_D}{2} + \sqrt{\left(\frac{N_D}{2}\right)^2 + n_i^2}$$

$$N_D >> n_i \Rightarrow \boxed{n \approx N_D}$$

Similarly, for P-type →

$$P = \frac{N_A}{2} + \sqrt{\left(\frac{N_A}{2}\right)^2 + n_i^2}$$

$N_A \gg n_i$

$$\boxed{P \simeq N_A}$$

N-type

$$n = N_c e^{-(E_c - E_f)/KT}$$

$n \to N_D$

$$KT \ln \frac{N_c}{N_D} = E_c - E_{fn}$$

$$\boxed{E_{fn} = E_c - KT \ln\left(\frac{N_c}{N_D}\right)}$$

P-type

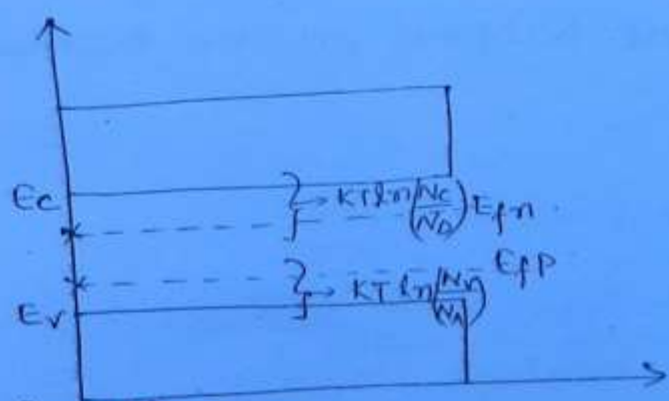
$$P = N_v e^{-(E_f - E_v)/KT}$$

$P \to N_A$

$$N_A = N_v e^{-(E_f - E_v)/KT}$$

$$KT \ln\left(\frac{N_v}{N_A}\right) = E_{fp} - E_v$$

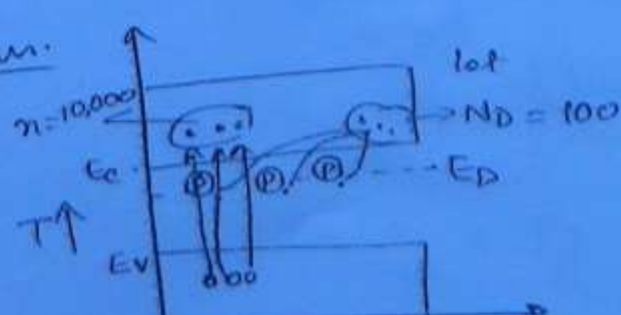
$$\boxed{E_{fp} = E_v + KT \ln \frac{N_v}{N_A}}$$



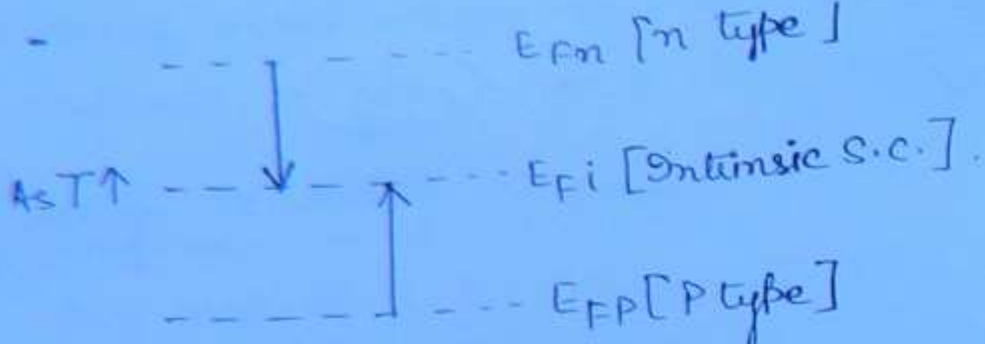
Questions →

Q1. As temp inc. & in an extrinsic S.C. what happens to the position of fermi level?

Ans.



As temp inc, n also increases but on N_D, temp has no effect. So, at high temp, extrinsic S.C behaves as intrinsic S.C.



The fermilevel moves toward the intrinsic S.C. in case of n type and P-type.

* At very very high temp, extrinsic S.C. will behave like an intrinsic S.C.

Q2. As doping conc. increases, in an extrinsic S.C what happens to fermi level?

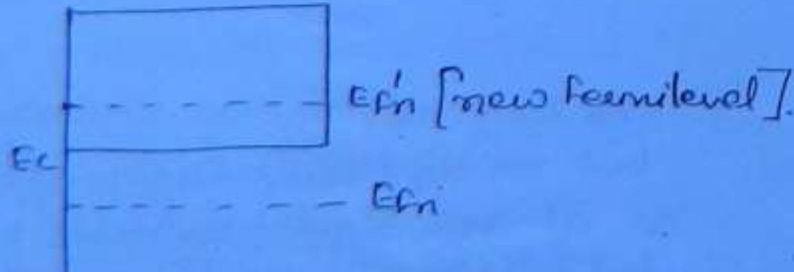
Ans. Fermilevel moves towards the C.B in case of n-type and moves towards the V.B. in case of P-type.

Q3. If $N_D >> N_C$ in n type what happens to the position of fermilevel?

Ans.
$$E_{Fn} = E_C - KT\ln\left(\frac{N_C}{N_D}\right)$$

$N_D >> N_C$.

$E_{Fn} = E_C + KT\ln N_D$.



Q. If $N_A >> N_V$ in p-type, what happens to the fermi level?

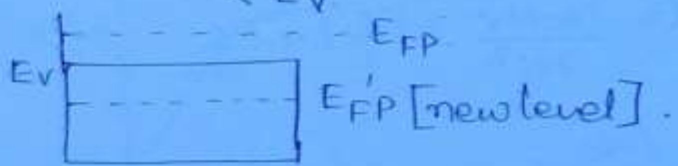
25

Ans. $$E_{FP} = E_V \pm KT \ln\left(\frac{N_V}{N_A}\right)$$

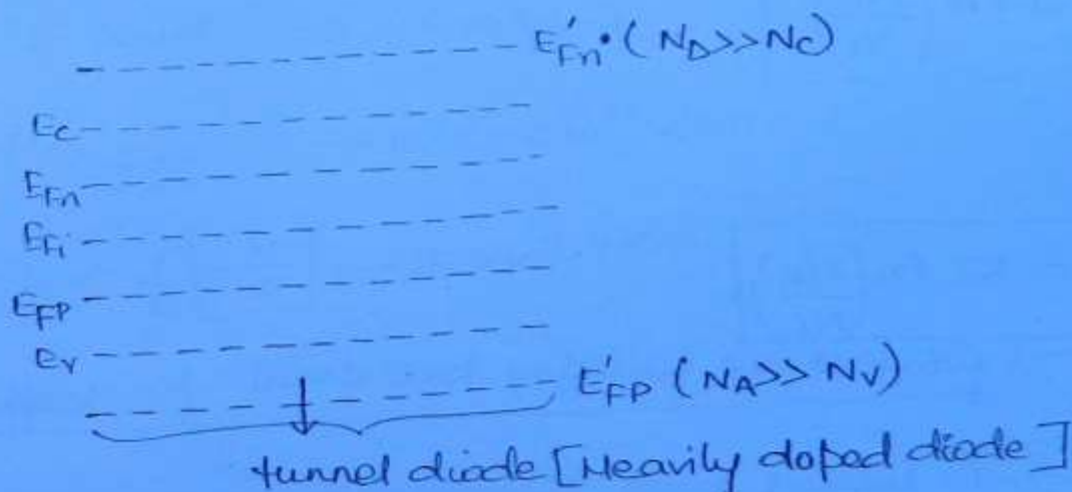
$N_A >> N_V$

$$E_{FP} = E_V - KT \ln N_A$$

$< E_V$



Conclusion :→



Shift in the fermilevel in case of n-type :—

w.r.t intrinsic fermilevel.

$\{E_{Fn} - E_{fi}\}$

$$= E_C - KT\ln\left(\frac{N_C}{N_D}\right) - \left\{\frac{E_C+E_V}{2} - \frac{KT}{2}\ln\frac{N_C}{N_V}\right\}$$

$$= E_C - KT\ln\left(\frac{N_C}{N_D}\right) - \left\{\frac{E_C+E_C-E_g}{2} - \frac{KT}{2}\ln\frac{N_C}{N_V}\right\}$$

$$= -KT\ln\frac{N_C}{N_D} + \frac{E_g}{2} + \frac{KT}{2}\ln\left(\frac{N_C}{N_V}\right) \quad —(i)$$

$n \cdot p = n_i^2$

$n_i^2 = N_c N_v e^{-(E_c - E_v)/KT}$

$E_c - E_v = E_g$

$$\boxed{E_g = KT \ln \frac{N_c N_v}{n_i^2}}$$

From (1)

$$= -KT \ln \frac{N_c}{N_D} + \frac{KT}{2} \ln \frac{N_c N_v}{N_i^2} + \frac{KT}{2} \ln \frac{N_c}{N_v}$$

$$= KT \ln \frac{N_D}{N_c} \frac{\sqrt{N_c N_v}}{n_i} \frac{\sqrt{N_c}}{\sqrt{N_v}}$$

$$\boxed{E_{Fn} - E_{Fi} = KT \ln \left(\frac{N_D}{N_i}\right)}$$

Similarly,

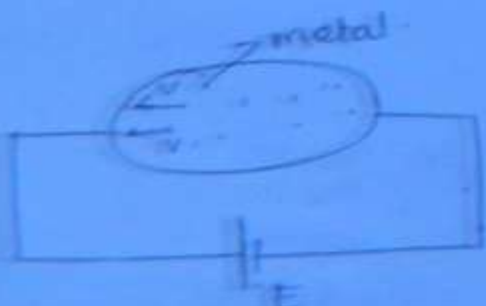
$$\boxed{E_{Fi} - E_{Fp} = KT \ln \left(\frac{N_A}{N_i}\right)}$$

For P-type

Drift current →

It occurs in metals and S.C.

eg



$v \propto E$

$\boxed{v = \mu E}$

$$\text{Mobility} = \frac{\text{Drift velo}}{\text{Electric fld}} = \frac{m/\text{sec}}{V/m}$$

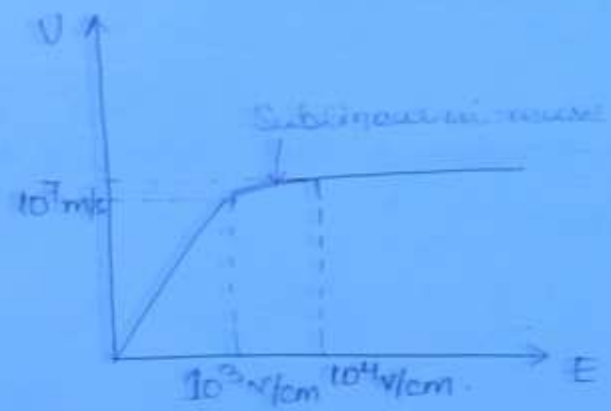
$$= m^2/V\text{-sec}.$$

* The current is produced due to the drifting of free e is called as <u>drift current</u> [drifting means movement]

(27)

<u>Mobility</u> →

<u>Effect of E-field on mobility</u> :—

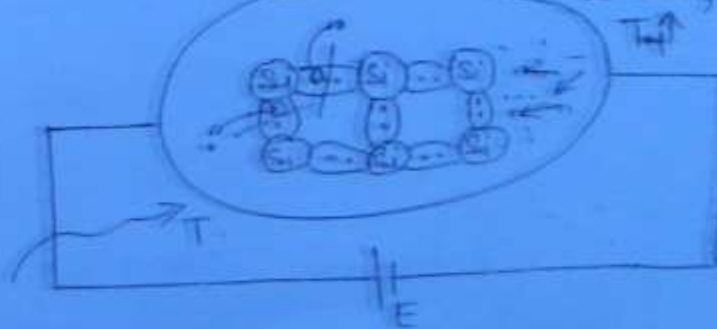


$\mu = \text{const} \rightarrow E < 10^3$ V/cm

$\mu \propto \frac{1}{\sqrt{E}} \longrightarrow 10^3 < E < 10^4$ V/cm

$\mu \propto \frac{1}{E}\left[=\frac{V}{E}\right] \longrightarrow E > 10^4$ V/cm

<u>Effect of temp and impurity on mobility</u> :—



* when, crystal vibration occurs [crystal lattice effect more and doping effect less]
⇒ when T↓, crystal vibration disappears [lattice effect less and doping eff]
⇒ effect of E on crystal lattice is negligible.

The e⁻ and hole mobility are influenced by two scattering phenomena. One is lattice scattering effect and other is impurity scattering

Lattice scattering :→

As temp increases, there will be vibration in the crystal lattice which produces the mobility. 28

$$\boxed{\mu \propto T^{-m}}$$

Impurity scattering →

As temp decreases, Impurity scattering will become more dominant then mobility reduces.

$$\boxed{\mu \propto T^{m}}$$

Current density :→

$$J = \frac{I}{A}$$

$$I = \frac{N.e}{T} \qquad T \to \text{time}$$

$$J = \frac{N.e}{T.A}$$

$$\text{Time} = \frac{\text{dist}}{\text{Velo}}$$

$$T = \frac{L}{v}$$

$N \to$ no. of e^-

$$\boxed{J = \frac{Nev}{LA}}$$

$$n = \frac{N}{LA}$$

$$\boxed{J = nev}$$

$\rho = n \times e$ = charge density

$$\boxed{J = \rho v}$$

metals

$J = n e v$

29

$v = \mu E$

$J = n e \mu E$

$\boxed{J = \sigma E}$ $\boxed{\sigma \rightarrow \text{conductivity}}$

$\boxed{\sigma = n e \mu}$

S.C.

$J_{s.c.} = J_n + J_p$

$J_n = n q \mu_n E$

$J_p = p \times q \times \mu_p E$

$J_{sc} = n q \mu_n E + p q \mu_p E$

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

where $\sigma_{sc} = (n \mu_n + p \mu_p) q$

Expression for conductivity in case of ~~extr~~ intrinsic S.C.:-

$\sigma_{sc} = (n \mu_n + p \mu_p) q$

$n = p = n_i$

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

where $\boxed{n_i^2 = A_0 T^3 e^{-E_{g0}/KT}}$

$\boxed{n_i \propto T^{3/2}}$

For n-type :—

$\sigma_{sc} = (n \mu_n + p \mu_p) q = n \mu_n q$ $\quad p \mu_p q \rightarrow$ can be neglected

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

$\approx N_D \mu_n q$

For P-type →

$\sigma_{P\text{type}} \approx n \mu_p q$

$\approx N_A \mu_p q$.

9/12/11

Parameter	Ge	Si
1) E_g at 0K	0.785 eV	1.21 eV
2) E_g at 300K	0.72 eV	1.12 eV.
3) n_i/cm^3 at 300K	$2.5 \times 10^{13}/cm^3$	$1.5 \times 10^{10}/cm^3$.
4). μ_n $cm^2/V sec$	3,800	1300
5) μ_h $cm^2/V sec$.	1800	500.
6) no. of atoms/cm^3	$4.4 \times 10^{22}/cm^3$	$5.0 \times 10^{22}/cm^3$.

Diffusion current :→

It occurs in a non uniformly doped S.C.

eg.

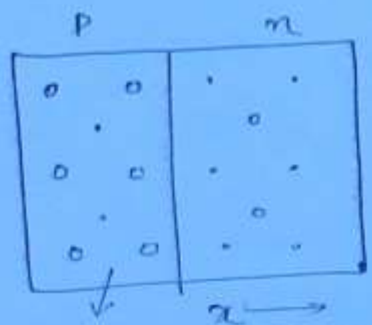
→ uniformly doped S.C.

diffusion $I = 0$.

→ non uniformly doped S.C

$I_{diff} \neq 0$.

(31)



The best eg of non uniformly doped S.C. is p-n j^n.

The rate of change of conc. w.r.t. dist. x. is called as diffusion current.

NOTE →

Diffusion current mechanism can also be called as conc. gradient $\left[\frac{dn}{dx}\right]$.

Drift current mechanism can also be called as pot. gradient $\left[\frac{dv}{dx}\right]$.

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

$$J_n = qD_n\frac{dn}{dx}$$

D_n → diffusion const of e^-

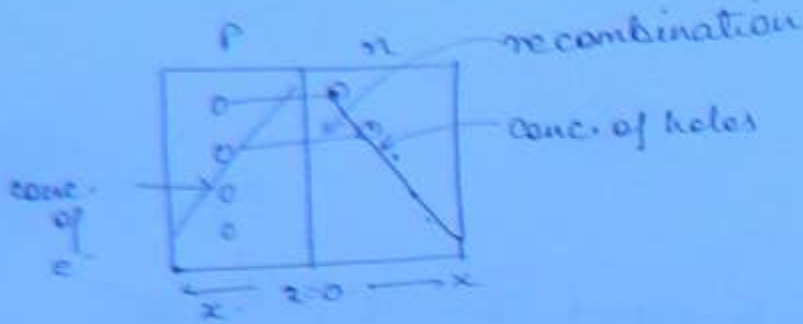
$$I_n = AqD_n\frac{dn}{dx}$$

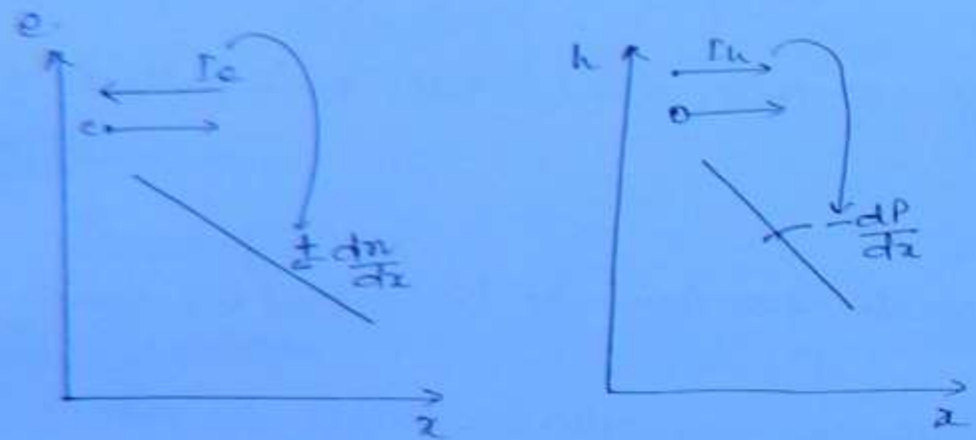
$$J_p \propto q\frac{dP}{dx}$$

$$\boxed{J_P = qD_P\frac{dP}{dx}} \to X$$

D_P → Diffusion const of holes

$$\boxed{I_P = AqD_P\frac{dP}{dx}} \to X$$





$\Rightarrow J_p = -qD_p\frac{dP}{dx}$

$\Rightarrow I_p = -AqD_p\frac{dP}{dx}$.

Total current in a S.C. :→

$I_{sc} = I_{drift} + I_{diff}$.

$= nq\mu_nEA + Pq\mu_pEA + AqD_n\frac{dn}{dx} - AqD_p\frac{dP}{dx}$.

There is an imp. relation b/w diffusion const. and mobility.

$D \propto \mu$.

$D = V_T\mu$.

$V_T \to$ volt equivalent temp or Thermal voltage.

$V_T = \frac{KT}{q} = \bar{K}T = \frac{T}{11600}$

1.38×10^{-23} J/K $\quad$ 8.65×10^{-5} eV/K.

Q. Cal the volt. eq. temp V_T at room temp.

Ans. $T = 27^\circ C$

$= 300 K$

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

$$= 0.026 V$$

or

$$\boxed{V_T = 26 mV}$$

Units

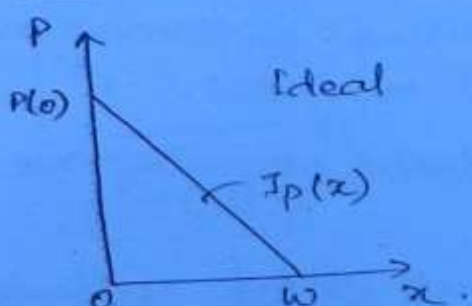
$\mu \rightarrow m^2/v\text{-}sec$

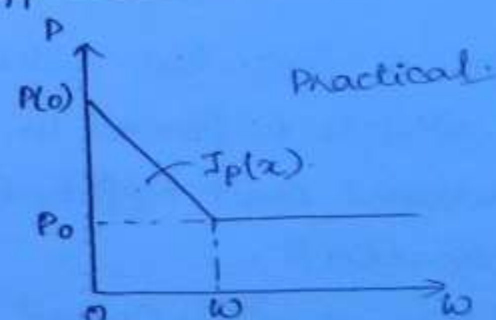
$V_T \rightarrow v$

$D \rightarrow m^2/sec$

$$\boxed{\frac{D_n}{\mu_n} = \frac{D_p}{\mu_p} = V_T} \rightarrow \text{Einstein relation}$$

Problems based on diffusion current :-





$$J_p(x) = -qD_p \frac{dP(x)}{dx}$$

$$= -qD_p\left(\frac{P(0) - 0}{0 - W}\right)$$

$$= qD_p \frac{P(0)}{W}$$

$$J_p(x) = -qD_p \frac{dP(x)}{dx}$$

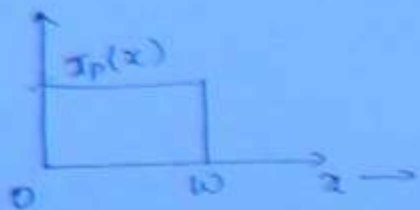
$$= -qD_p\left(\frac{P(0) - P_0}{0 - W}\right)$$

$$= \frac{qD_p\{P(0) - P_0\}}{W}$$

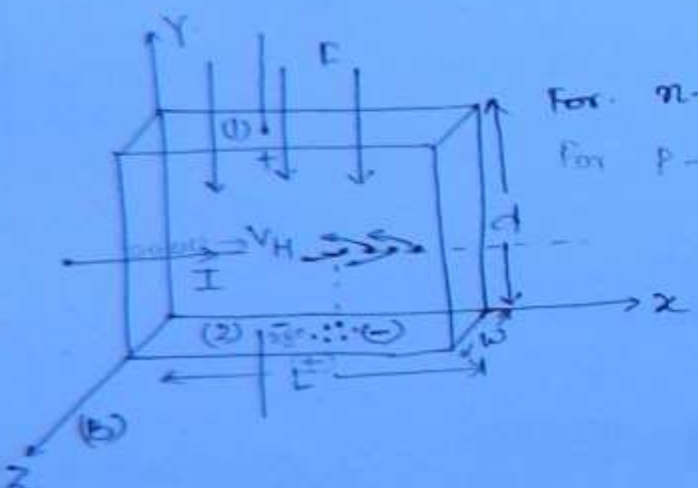
$$J_p(x) = \frac{qD_p\{P(0) - P_0\}}{W} \quad 0 < x < W.$$

34

$= 0$ else



<u>Hall effect</u> →



For n-type, side 2 (−).
For p-type, side (1).

1) let us consider a S.C. bar whose current is flowing in the +ve X-dir^n which is placed in a ⊥^{r} mag. fld.

2) Due to the current and M-fld, an Electro magnetic force (EMF) exist in the (−)Y-dir^n.

3) Due the the EMF, a P.d. called as hall voltage is developed in the Y-dir^n

The main application of hall effect is to find the S.C. bar whether it is n-type or P-type.

Side (2) "−" → n type
"+" → P type

35

Other application of hall effect :—

1) We can determine the mobility of charge carriers in a S.C. bar.

2) We can also determine the conductivity of charge carri

3) It can be used as a voltage multiplier $[V_H \propto B \times I]$

Expression for hall voltage (V_H), hall co-efficient (R_H), mobilit (μ), conductivity (σ) :—

$F_E \rightarrow$ Force due to E-fld.

$F_B \rightarrow$ " " " M-fld.

$$F_E = F_B$$

$$qE = Bqv$$

$v \rightarrow$ velo.

$V' \rightarrow$ volt. $= V_H$.

$$\frac{V'}{d} = Bv$$

$$V' = Bvd$$

$$= B\frac{I}{\rho}d \quad [I = \rho v]$$

$$= \frac{BI}{A\rho}d$$

$$= \frac{BI}{w\cancel{d}\rho}\cancel{d}$$

$$\boxed{V_H = \frac{BI}{w\rho}}$$

$$\boxed{R_H = \frac{1}{\rho}}$$ $\rho \rightarrow$ charge density

$$V_H = \frac{BI}{w}R_H.$$

$$\boxed{R_H = \frac{V_H w}{BI}}$$

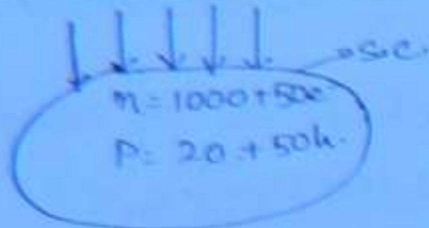
$$\sigma = \underbrace{n \times e}_{\rho} \times \mu$$

$$\sigma = \rho\mu$$

$$\boxed{\mu = \frac{\sigma}{\rho} = \sigma R_H.}$$

Generation and Recombination of charge carriers :—

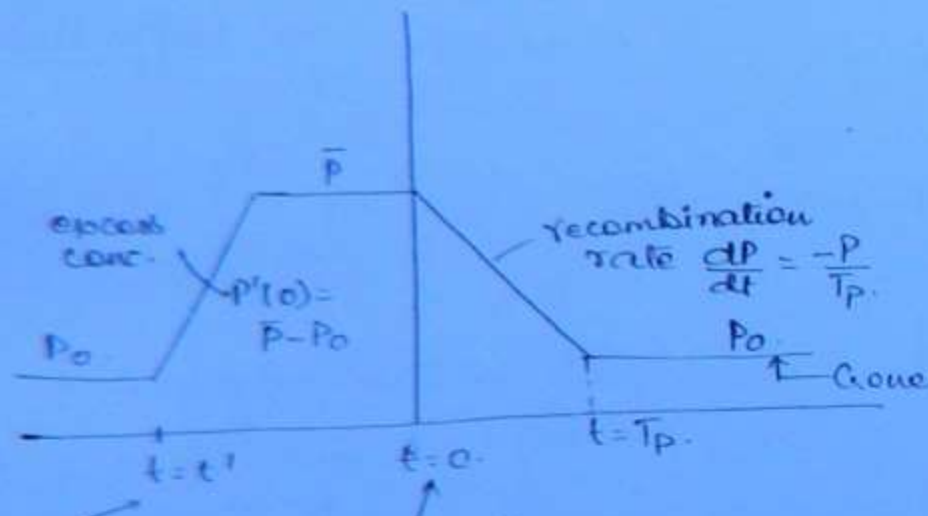
eg

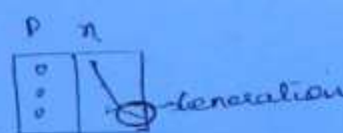


rate of incr. e⁻ is less as compared to rate of incr of holes. ⇒ 250%.
5%

For the analysis of S.C. bar, we have to conc. on minority carriers but not majority carrier.

(36)





o - - - - o
↓
T_P [carrier life time]

$T_P \rightarrow$ Carrier life time for holes.

It is a time taken for a hole to exist before recombination is called as carrier life time for holes.

1) When light radiation is applied to a S.C. bar, two mechanism will be happen.

Recombination rate →

$$\frac{dP}{dt} = \frac{-P}{T_P}$$

Generation rate →

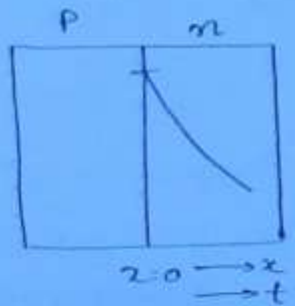
$$\frac{dP}{dt} = \frac{P_0}{T_P}$$

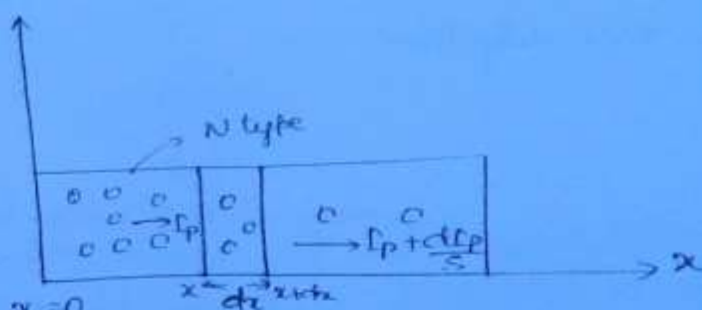
2) Generation rate always occurs after recombination rate.

Continuity equation →

1) For the analysis of a diode or a transistor we have to determine the conc. of charge carriers w.r.t time and spac

2) For such analysis, we have a mathematical eqn called as continuity eqn





Factors affecting the conc. P of thin slice of thickness dx and area of X sec A is

1) **Diffusion** →

$$\frac{dP}{dt} = \frac{\text{charge taken away}}{\text{Unit charge} \times \text{Vol.}}$$

$$= \frac{-dI_p}{q \times A \times dx} \quad -(1)$$

2) when light radiation is applied →

$$G = \frac{dP}{dt} = \frac{P_0}{\tau_p} \quad -(2)$$

$$R = \frac{dP}{dt} = \frac{-P}{\tau_p} \quad -(3)$$

(1)+(2)+(3) ⇒

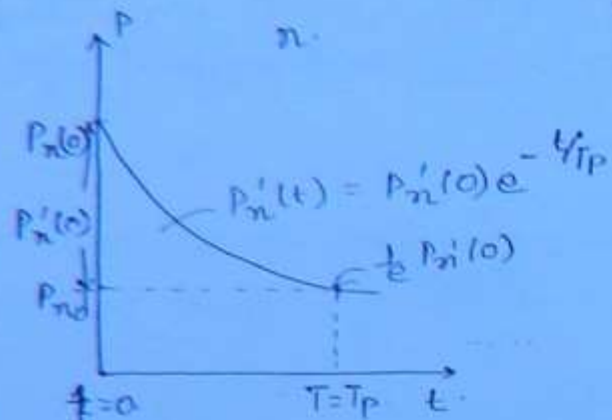
$$\boxed{\frac{dP}{dt} = \frac{-dI_p}{q \times A \times dx} + \frac{P_0}{\tau_p} - \frac{P}{\tau_p}}$$

Case 1 →
w.r.t time →

P_{no} → thermal equilibrium

$P_n'(0)$ → excess charge carriers.

(38)



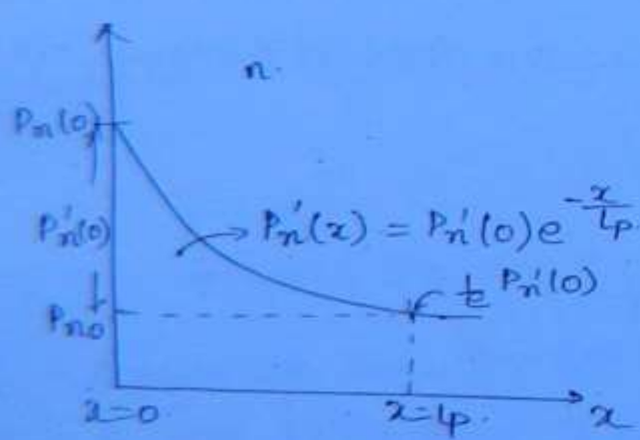
$P_n(0)$ → Injected minority Carrier conc. [on the j^n always e^- or hole → minority]

P_{no} → Thermal equilibrium minority conc.

$P_n'(0)$ → excess conc.

Case 2 →
w.r.t Space →

L_p → mini dist traveled by hole before recombination = diffusion length for holes.

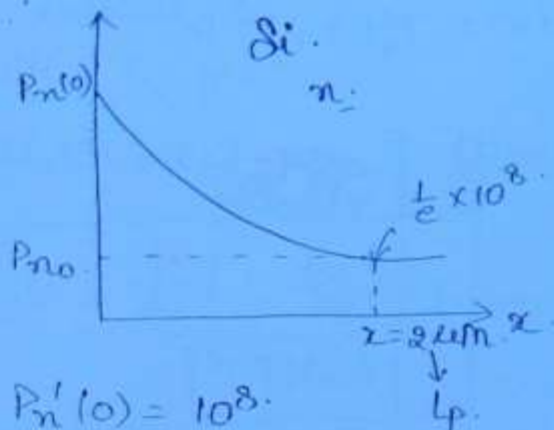


L_p →

It is the mini distance travelled by a hole before recombination is called as diffusion length for a hole. [L_p]

$$\boxed{L_p^2 = D_p T_p}$$

Q.



P9

$P_n'(0) = 10^8$

$P_n(0) = 1\times10^8$

$P_n(0) = 2\times10^8$

Cal.

a) e^- conc.

b) $L_p = \cancel{\mu m}$

c) $T_p = \frac{L_p^2}{D_p} = ?$

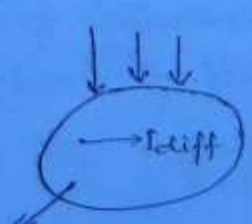
Ans. a) $n = \frac{n_i^2}{P_{no}} = \frac{(1.5\times10^{10})^2}{1\times10^8}$

b) $L_p = 2\mu m = 2\times10^{-6} m$

c) $T_p = \frac{L_p^2}{D_p} =$

$D_p = \mu_p V_T = 0.05 \times 0.026$

Total current in an illuminated O.C. S.c. bar :—



non uniformly doped S.c.

$I_{diff} = I_n(x) + I_p(x)$

$$I_p(x) = -Aq D_p \frac{d}{dx} P(x) \qquad P(x) = P_n'(0)\, e^{-x/L_p}$$

$$= \frac{+Aq D_p P_n'(0) e^{-x/L_p}}{+L_p} \quad -(1)$$

40

$$I_n(x) = -\frac{Aq D_n P_n'(0) e^{-x/L_n}}{L_n} \quad -(11)$$

When light is ON.

1) $P_n'(0) = n_p'(0)$

2) $L_p = L_n$.

$$\therefore I_n(x) = \frac{-Aq D_n P_n'(0)\, e^{-x/L_p}}{L_p} \quad -(3)$$

$(1)+(3) \Rightarrow$

$$I_{diff} = \frac{Aq D_p\left(1 - \frac{D_n}{D_p}\right) P_n'(0)\, e^{-x/L_p}}{L_p}$$

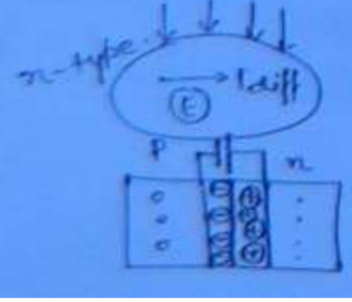
$$= I_p\left[1 - \frac{D_n}{D_p}\right]$$

As $D_n > D_p$

$I_{diff} = -$ive.

Conclusion:→

In a O.C. S.C. bar, net current should be equal to 0 that means there is an E-fld generated which produces drift current that cancels the diff: current



net current $= 0$.

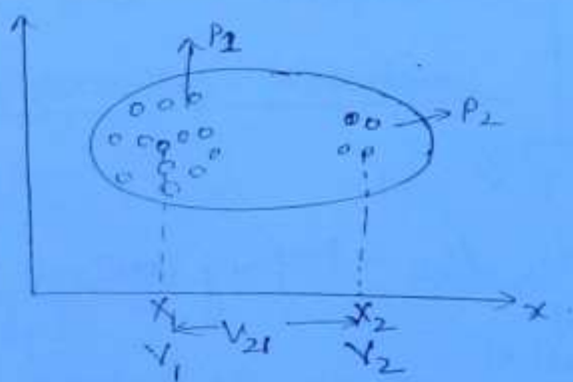
$I_{drift} + I_{diff} = 0$.

$nq\mu_n E A + I_p\left[1 - \frac{D_n}{D_p}\right] = 0$.

41

$$\boxed{E = \frac{\left|I_p\left[1 - \frac{D_n}{D_p}\right]\right|}{|nq\mu_n A|}}$$

<u>Potential variation in a O.C. S.c. bar :-</u>



$J = 0$

$\underbrace{I_{drift}} + \underbrace{I_{diff}} = 0$.

$I_n + I_p + I_n + I_p = 0$

$I_p = 0$. We are taking about holes. So, ignore I_n.

$J_{P drift} + J_{P diff} = 0$

$Pq\mu_p E - qD_p\frac{dP}{dx} = 0$.

$E = \frac{D_p}{\mu_p} \frac{1}{P} \frac{dP}{dx}$.

$\Rightarrow -\int_{V_1}^{V_2} \frac{dv}{dx} dx = V_T \int_{P_1}^{P_2} \frac{1}{P}\frac{dP}{dx} dx$

$V_2 - V_1 = V_T(-\ln P)_{P_1}^{P_2}$.

$V_{21} = V_T \ln\left(\frac{P_1}{P_2}\right)$

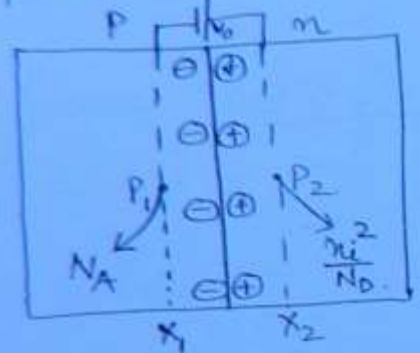
$$P_1 = P_2 e^{+V_{21}/V_T}$$
$$P_2 = P_{e_1} e^{-V_{21}/V_T}$$

$\Big\}\rightarrow$ Boltzman relation

Kinetic gas theory.

30/12/11

Applying the Boltzmann relation in a o.c. pn $j^{\underline{n}}$ to find the barrier potential V_0



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

$N_A, N_D \rightarrow$ Doping conc.

$V_0(Ge) = 0.1$ to $0.3V$

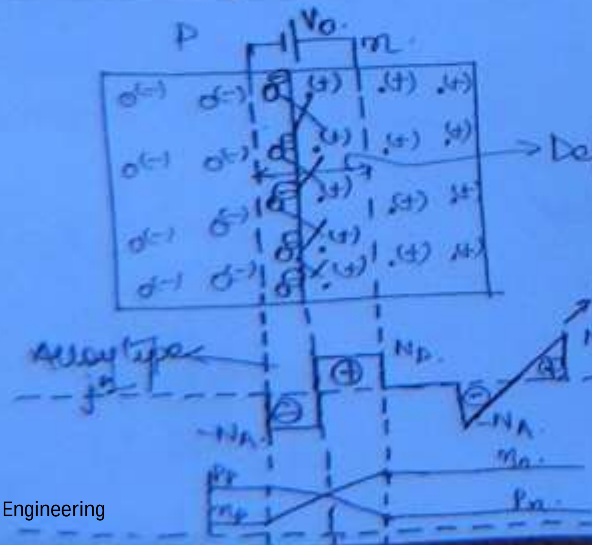
$V_0(Si) = 0.5$ to $0.7V$

$n_i(Ge) = 2.5 \times 10^{13} / cm^3$.

$n_i(Si) = 1.5 \times 10^{10} / cm^3$

P-N Junction Diode :→

open ckt P-n $j^{\underline{n}}$:→

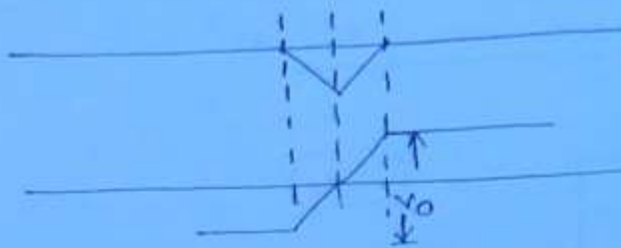


→ Depletion region or space charge region

→ Grown pn type. First doped with holes then with e⁻

Distribution of charge.

Diffusion process.



E-fld $E = -\frac{dV}{dx}$.

(43)

barrier pot.
$V_0 = -\int E\,dx$.

Mobile charge carriers

- e^-
- holes.

Immobile charge carriers

- donor ions
- acceptor ions

Pn j^n formation techniques

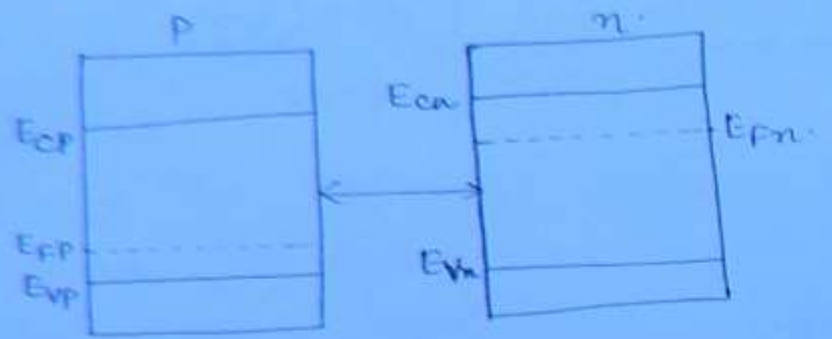
$W \propto \sqrt{V_j}$
Alloy type j^n.
↓
Step graded j^n.
↓
Sudden change or abrupt change.

$W \propto \sqrt[3]{V_j}$
Grown j^n type
↓
to linearly graded j^n
↓
diffused type.

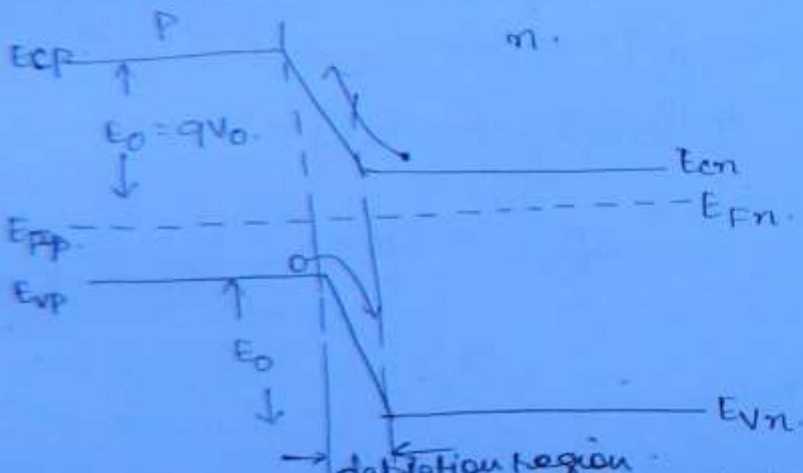
V_0 or V_j or V_r →
Barrier potential
'or'
Built in voltage
'or'
Cut in voltage
or
contact potential.
or
Depletion voltage
or
space charge voltage
or
transition voltage.

* Electric fld will be more conc. on the j^n

Energy band diagram in a u.b. P-n Jn.



For eg →



E_0 → potential en for an e^-

V_0 → Barrier pot.

Note → Any two materials are joined fermi levels will be in aligned position.

Q) Derive an expression for pot. en. for an e^-, E_0 from En. band theory concept.

Ans.

$E_0 = E_{cp} - E_{cn}$

'or'

$= E_{vp} - E_{vn}$

45

'or'

$= E_1 + E_2$

$E_{FP} - E_{VP} = \frac{1}{2} E_g - E_1 \quad \text{—(1)}$

$E_{cn} - E_{Fn} = \frac{1}{2} E_g - E_2 \quad \text{—(2)}$

$E_0 = E_1 + E_2$

$= E_g - \{E_{FP} - E_{VP}\} - \{E_{cn} - E_{Fn}\}$

$= KT \ln\left(\frac{N_c N_V}{n_i^2}\right) - KT \ln\left(\frac{N_V}{N_A}\right) - KT \ln\left(\frac{N_c}{N_D}\right)$

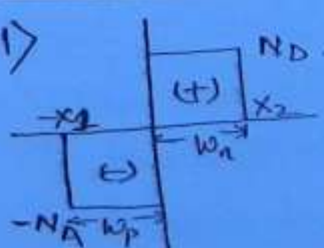
$= KT \ln\left(\frac{N_c N_V}{n_i^2} \frac{N_A}{N_V} \frac{N_D}{N_c}\right)$

$$\boxed{E_0 = KT \ln\left(\frac{N_A N_D}{n_i^2}\right) \text{ eV}}$$

$$\boxed{V_0 = \frac{KT}{q} \ln\left(\frac{N_A N_D}{n_i^2}\right) \text{ eV}}$$

Formulas based on O.C. P-N jn.

1)



Net charge densities $= 0$.

Positive charge densities $=$ -ive charge densities.

$q N_D W_n = q N_A W_p$

$\boxed{\frac{N_D}{N_A} = \frac{W_p}{W_n}}$ — neutrality concept

2) $\frac{\partial^2 V}{\partial x^2} = \frac{-\rho}{\epsilon}$

3) ρ : qN_D $\quad 0 < x < x_2$

$-qN_A$ $\quad -x_1 < x < 0$

0 $\quad$ else

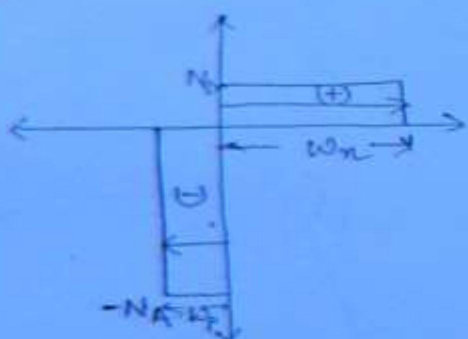
46

4). $E = \left|\frac{qN_D W}{\epsilon}\right| = \left|\frac{qN_A W}{\epsilon}\right|$

$V_0 = \left|\frac{qN_D W^2}{2\epsilon}\right| = \left|\frac{qN_A W^2}{2\epsilon}\right|$

$\boxed{W \propto \sqrt{V_0}}$

5) P^+N.



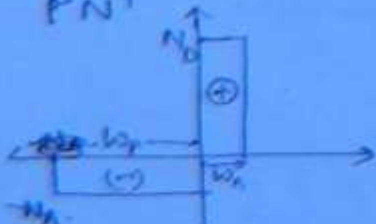
$W_n + W_p = W$

$W \simeq W_n$

$E = \left|\frac{qN_D W_n}{\epsilon}\right|$

$V_0 = \left|\frac{qN_D W_n^2}{2\epsilon}\right|$

6) PN^+



$W \simeq W_p$.

$E = \left|\frac{qN_A W_p}{\epsilon}\right|$

$V_0 = \left|\frac{qN_A W_p^2}{2\epsilon}\right|$

Note →

1) When the material is lightly doped one, the penetration of the depletion region will be more.

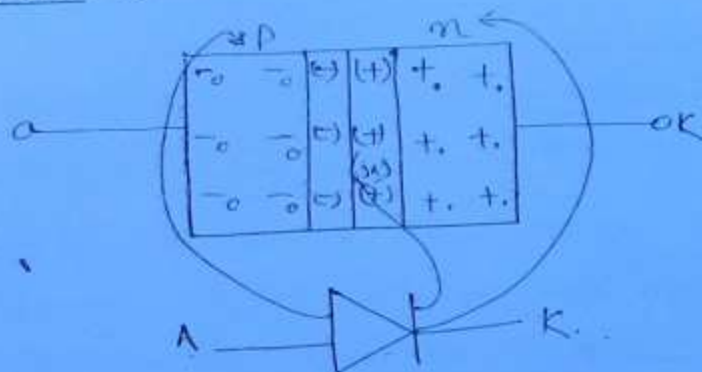
2) When material is heavily doped, the penetration of the depletion region will be narrower. (47)

$$W = \left[\frac{2\epsilon_0\epsilon_r V_j}{q}\left(\frac{1}{N_A}+\frac{1}{N_D}\right)\right]^{1/2}$$

$$W_n = \left\{\frac{2\epsilon_0\epsilon_r V_j}{q}\,\frac{1}{N_D}\right\}^{1/2}$$

$$W_p = \left\{\frac{2\epsilon_0\epsilon_r V_j}{q}\,\frac{1}{N_A}\right\}^{1/2}$$

Bias →



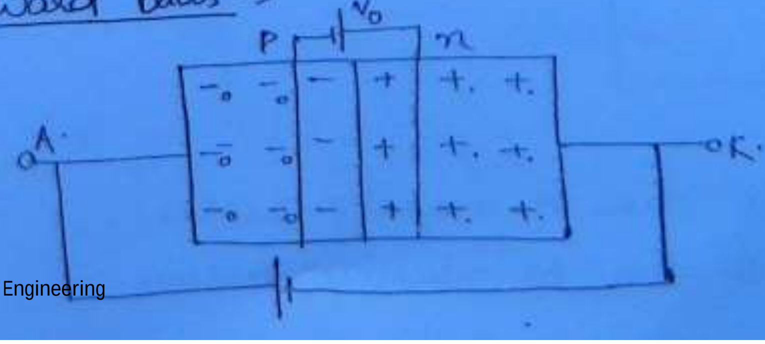
– Adding an ext. dc voltage to a p-n jn diode is called 'Bias'

→ There are two types of Bias –

1) Forward bias

2) Reverse bias.

Forward Bias →



1) If $V_f > V_0$, current conduction is possible.

2) The current is possible due to majority carriers. and that majority current is represented by

$I_f \rightarrow mA$.

3) The current will be zero due to minority carriers.

4) The depletion region becomes narrower or negligible.

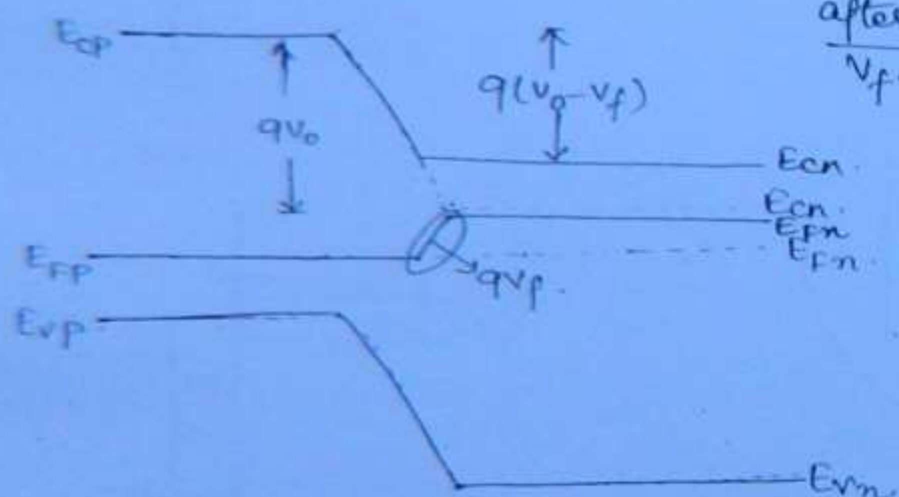
5) In en. band diagram, barrier height decreases.

6) Resistance offered by the diode is very less.

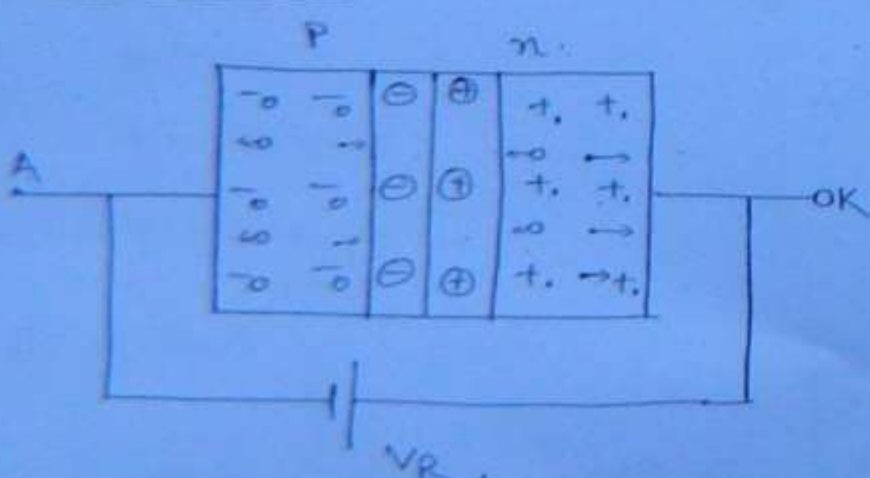
Energy band diagram in forward bias :—

after V_f is applied.

$V_f < V_0$.



Reverse bias →



1) The current will be 0 due to majority carriers.

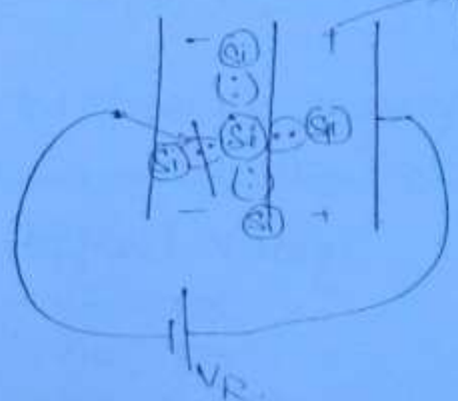
2) The current is possible due to minority carriers.

3> Minority current can also be called as reverse saturation current. [I_0 or I_s]

μA (Ge) $\quad$ nA (Si)

Q: Why the reverse saturation current for Ge is in 'μA' and for Si it is in 'nA'?

Ans.



→ Uncovered charges [Immobile ions] always in depletion region [because of e or holes are not with ions]

→ In the depletion region, there will be some stable atoms along with uncovered charges.

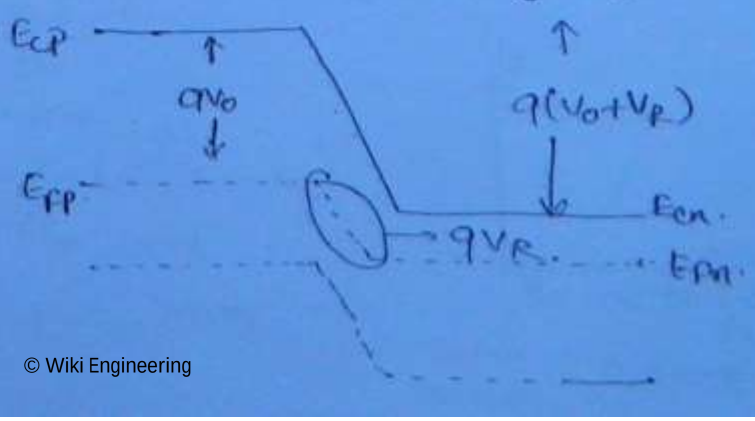
→ As the reverse bias voltage inc, the velo. of the minority charge carrier increases which collides with the stable atom in the depletion region.

→ As the en. gap value for Ge is less compare to Silicon we expect I_0 to will be in μA for Ge and nano amp. in Silicon.

→ Depletion region becomes wider in reverse bias.

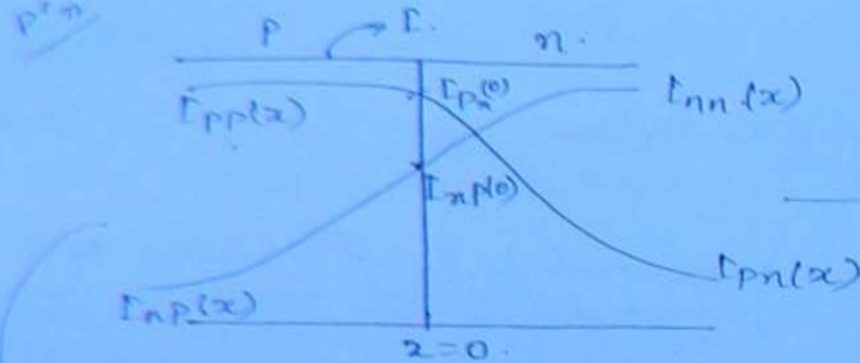
→ In en. band diagram, barrier height increases.

→ Resistance offered is very high in the order of $M\Omega$.



Current components in P-N diode :—

(50)



$\longrightarrow I_{nn}(x) + I_{pn}(x)$.

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

$I_{pp}(x) + I_{np}(x)$

At $x=0$,

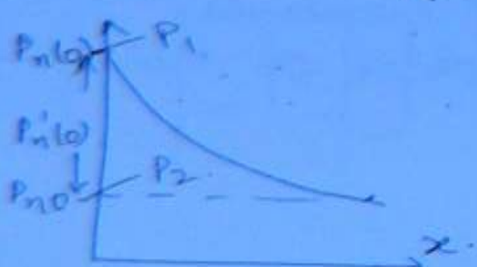
$I = \underbrace{I_{pn}(0)}_{\text{minority}} + \underbrace{I_{np}(0)}_{\text{minority}}$

$$I_{pn}(x) = -Aq D_p \frac{dP(x)}{dx} \quad \longrightarrow P_n'(0)\, e^{-x/L_p}$$

$$= \frac{-Aq D_p P_n'(0) e^{-x/L_p}}{-L_p}$$

at $x=0$.

$$I_{pn}(0) = \frac{Aq D_p P_n'(0)}{L_p} = Aq D_p P_{n0}\left(e^{V/V_T} - 1\right)$$



$$\boxed{P_n(0) = P_{n0}\, e^{V/V_T} \rightarrow \text{Law of jn}}$$

$$P_n'(0) = P_n(0) - P_{n0}.$$
$$= P_{n0}\left(e^{V/V_T} - 1\right)$$
$$P_1 = P_2\, e^{V_{21}/V_T}.$$

$$I = \underbrace{\left[\frac{-Aq D_p P_n(0)}{L_p} + \frac{Aq D_n n_{p0}}{L_n}\right]}_{I_0}\left(e^{V/V_T} - 1\right)$$

51

31/12/11

Diode current equation :→

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

I → Total current through diode

I_0 → Reverse saturation current

V → Applied Voltage

→ +ive FB

→ -ive RB

η → Idealised factor

= 1 (Ge)

= 2 (Si).

V_T → Volt eq temp.

$= \frac{T}{11,600}$

At 300K, V_T → 26mV.

Diode current eqn can also be called as V-I char. eqn.

VI characteristics of P-n diode :→

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

Case 1. →

F.B.

V → +ive

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

$$e^{\frac{V}{\eta V_T}} \gg 1$$

$$I = I_0\, e^{V/\eta V_T}.$$

V_γ = 0.1 to 0.3 V. (Ge)

Cut in Voltage = 0.5 to 0.7 V (Si).

Case 2.

RB.

V → -ve

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

$e^{-V/\eta V_T} << 1$

$I = -I_0$

Temp dependence on V-I char. of P-n diode :→

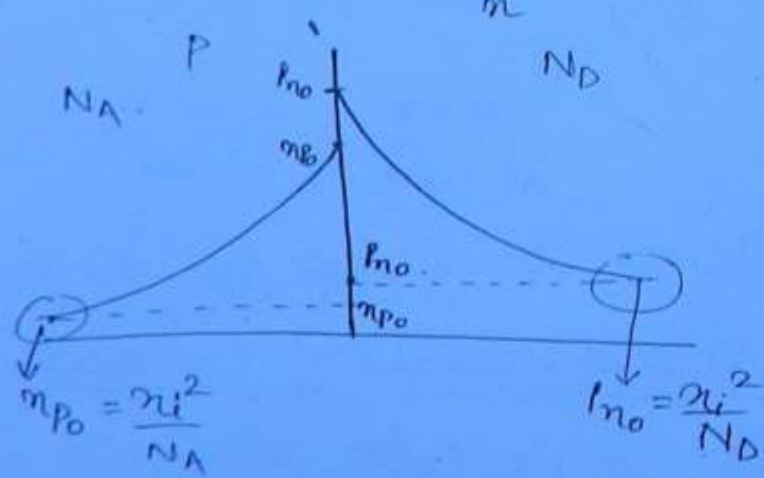
$$I = I_0\left(e^{\frac{V}{\eta V_T}} - 1\right)$$

I_0 v/s T →

$$I_0 = \left[\frac{A q D_p P_{n0}}{L_p} + \frac{A q D_n n_{p0}}{L_n}\right]$$

$$= \left[\frac{D_p}{L_p N_D} + \frac{D_n}{L_n N_A}\right] A q n_i^2$$

$$= \cancel{\frac{D_p}{L_p N_D}} + D$$



Property →

Si

$I_0 \propto n_i$

$$\boxed{n_i^2 = A_0 T^3 e^{-E_{g0}/KT}}$$

Ge

$D \propto \frac{1}{T}$

For Si →

$$I_0 = K_1 T^{3/2} e^{\frac{-E_{g0}}{2KT}}$$

For Ge,

$$I_0 = K_2 T^2 e^{-\frac{E_{go}}{KT}}$$

(53)

$$I_0 = K' T^m e^{\frac{-E_{go}}{\eta KT}}$$

	m	E_{go}	η
Si	1.5	1.21	2
Ge	2	0.785	1

Relative temp co-eff of I_0

$$\frac{\partial I_0}{I_0} \Big/ \partial T$$

$$\ln I_0 = \ln K' + m \ln T - \frac{E_{go}}{\eta KT}$$

diff w.r.t T

$$\frac{1}{I_0}\frac{dI_0}{dT} = 0 + \frac{m}{T} + \frac{E_{go}}{\eta KT^2}$$

Ge,

$$\frac{\partial I_0}{I_0} \Big/ dT = \left[\frac{2}{300} + \frac{0.785}{1 \cdot 0.026 \times 300}\right] = 10.7\% / {}^\circ C$$

Si,

$$\frac{\partial I_0}{I_0} \Big/ dT = \frac{1.5}{300} + \frac{1.21}{2 \cdot 0.026 \cdot 300} = 8.2\% / {}^\circ C$$

$\rightarrow 7\% / {}^\circ C$

$$(1.07)^{10} = 2$$

→ For every 10°c rise in temp, I_0 will be double

→ for every 1°c " " ", I_0 will be 7%.

1) I_0 increases by 7% for every degree centigrade rise in temp. & I_0 doubles for every 10°c rise in temp.

Applied Voltage V/S temp :→

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

54

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

F.B.

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

$$\Rightarrow V = \eta V_T \ln\left(\frac{I}{I_0}\right)$$

diff. w.r.t T.

$$\frac{dV_D}{dT} = -0.0019V \; ; \; -0.0017V \rightarrow (-2mV/^\circ C) \rightarrow 0.5\ mV$$

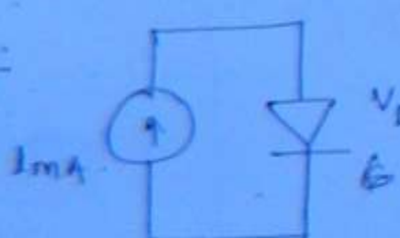
$$= -1.9mV \; , \; -1.7\ mV$$

↓ Si , ↓ Ge.

$$\frac{dV_D}{dT} \overset{F.B.}{=} -2.5\ mV/^\circ C \; , \; -2mV/^\circ C$$

↓ Or 0.7 mV , ↓ Or 0.5 mV

eg.



$V_D = 700mV$ $\quad T_1 = 20^\circ C$

at $T_2 = 24^\circ C$ $\quad V_D = ?$

Ans.

$$\frac{dV}{dT} = -2.5\ mV$$

$$\therefore V_D = 0.7\ V - 4 \times 2.5$$

$$= 700\ mV - 10\ mV$$

$$= 690\ mV.$$

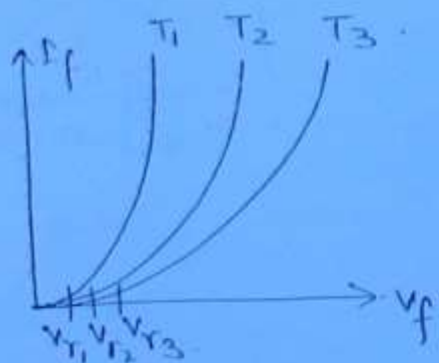
As T ↑, I_0 ↑ and V_D ↓

V_T v/s Temp →

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

T ↑, V_T ↑

Q.



T ↑, V_γ ↓

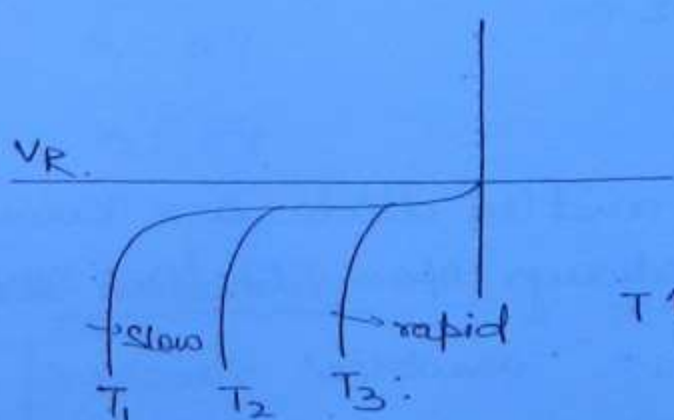
a) $T_1 < T_2 < T_3$

b) $T_1 > T_2 > T_3$ ✓

c) $T_1 = T_2 = T_3$

d) None.

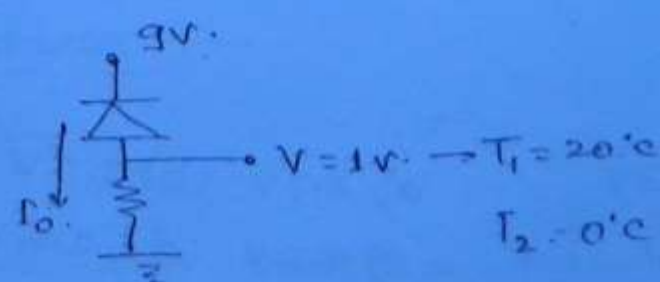
Q.



T ↑, I_0 ↑.

$T_3 > T_2 > T_1$

Q



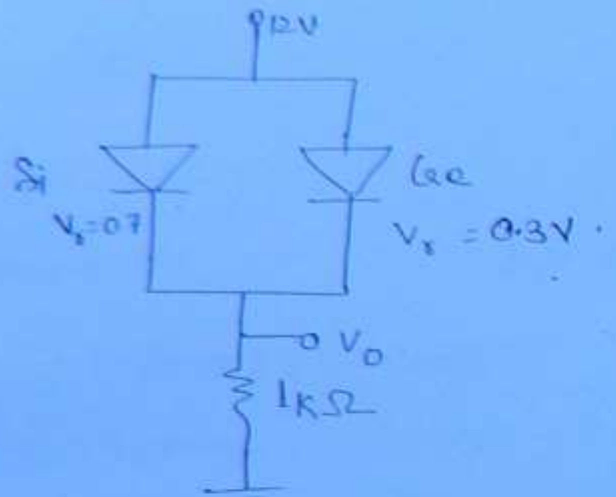
0°c 10°c 20°c 30° 40

$\frac{I_0}{4}$ $\frac{I_0}{2}$ I_0 $2I_0$ $4I_0$

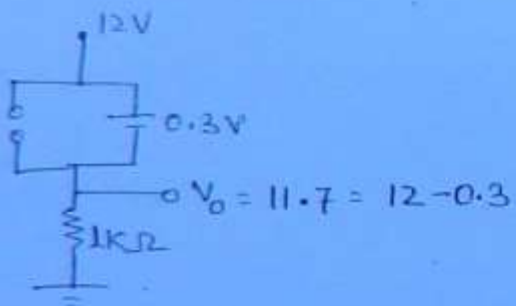
56

at $T_2 = 0°C$, $\boxed{V_t = 0.25V}$

Q.

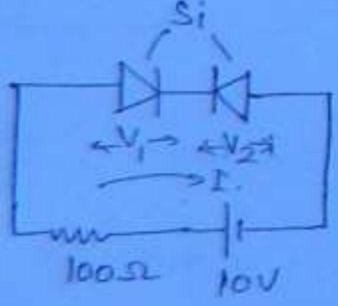


Ans.



- When a practical Silicon and Ge diode are connected in parallel, Silicon will be always open ckt (OFF state)

Q.



$V_1 =$

a) 0.7 b) 9.3V

c) 8.6V d) 2.6V

$V_2 =$

a) 0.7V b) 9.3V

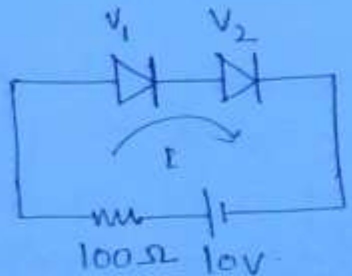
c) 8.6V d) 2.6V

$I =$

a) 3mA b) $\frac{3}{2}$ mA

c) 6.2mA d) I_0

2)

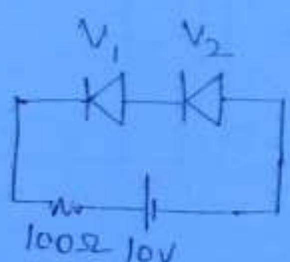


$V_1 = 0.7$

$V_2 = 0.7$

$I = 86\text{mA} = \frac{10-0.7-0.7}{100} = 86 \times 10^{-3}\text{A}$

3)

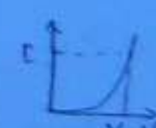
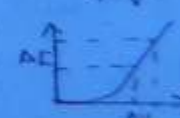


$V_1 = 5V$

$V_2 = 5V$

$I = I_0$.

Diode resistance →

- → Static Resistance DC  $R = \frac{V}{I}$ 
- → Dynamic Resistance  $r = \frac{\partial v}{\partial I}$ Ac Res. 

Static resistance :→

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

F.B.

$R = \frac{V}{I} = \frac{0.7V}{100mA} = 7\,\Omega$

RB →

$$R = \frac{V}{I} = \frac{10V}{1\mu A} = 10\,M\Omega$$

Dynamic resistance → (58)

$$r = \frac{\partial V}{\partial I}$$

$$g = \frac{\partial I}{\partial V}$$

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

$$= \frac{I_0\, e^{\frac{V}{\eta V_T}}}{\eta V_T}$$

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

V_T v/s temp

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

\# $$\boxed{g = \frac{I + I_0}{\eta V_T}}$$

For F.B. →

$$g = \frac{I}{\eta V_T}$$

Typically

$\eta = 1$, $I = 100\,mA$, $V_T = 26\,mV$

$g >> 1$

$r \rightarrow$ less

For R.B. →

$$g = \frac{2 I_0}{\eta V_T} < 1$$

$r \rightarrow$ high

Junction capacitance :→

→ transition capacitance → R.B.

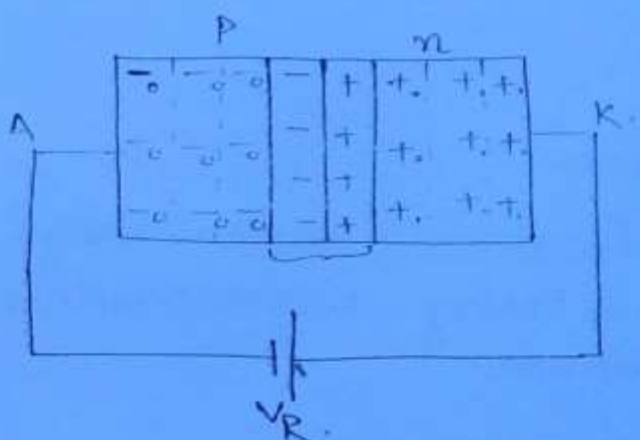
→ Diffusion capacitance → F.B.

59

Transition means depletion region. Depletion region exist only in R.B.

Diffusion is a process occurs in Forward bias because of negligible depletion region.

Transition capacitance :— C_s → Space charge
(C_T)



The rate of change of charge with the applied voltage — capacitance

The rate of change of immobile charge in the depletion region with the applied R.B. Voltage is a capacitive effect called as transition capacitance.

$$C = \frac{\epsilon A}{d}$$

$$C_T = \frac{\epsilon A}{W}$$

$W \propto \sqrt{V_j}$ $W \propto \sqrt[3]{V}$ } → open ckt

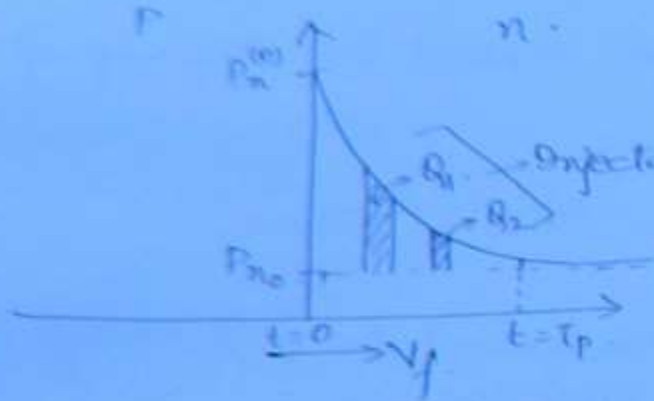
↓ (Alloy type)   ↓ (Grown J^n)

$W \propto \sqrt{V_j + V_R}$
$W \propto \sqrt[3]{V_j + V_R}$ } → R.B.

$C_T \propto \frac{1}{\sqrt{V_R}}$ → alloy type

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

Diffusion Capacitance →



$P_n^{(0)}$ – injected minority carrier conc.

The rate of change of injected minority carrier charge with the applied forward bias voltage is a capacitive effect called as diffusion capacitance.

$$Q = I_p \times t_p$$

$$C_D = \frac{dQ}{dV} = T_p \frac{\partial I_p}{\partial V} \rightarrow \frac{I_p + I_{p0}}{\eta V_T}$$

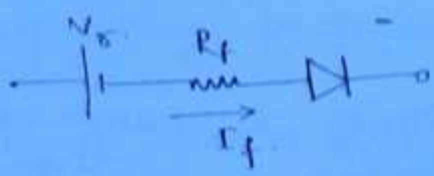
$$C_D = \frac{T_p I_p}{V_T}$$

2/01/12 **Diode equivalent models →**

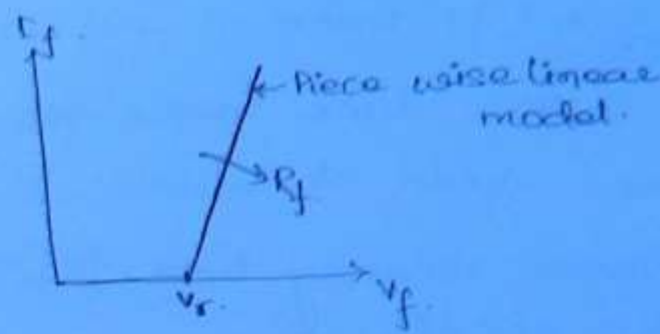
$V_f \geqslant V_\gamma + I_f R_f$

Case 1 →

R_f = const.

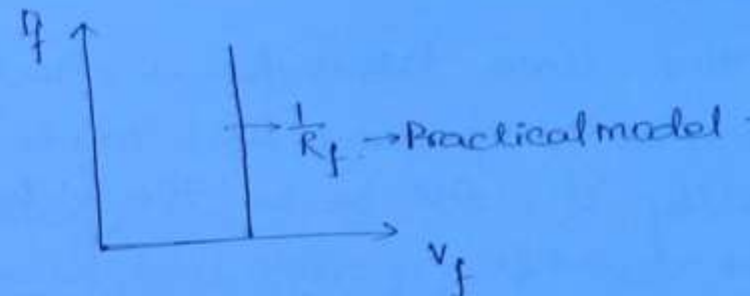


61



Case 2 →

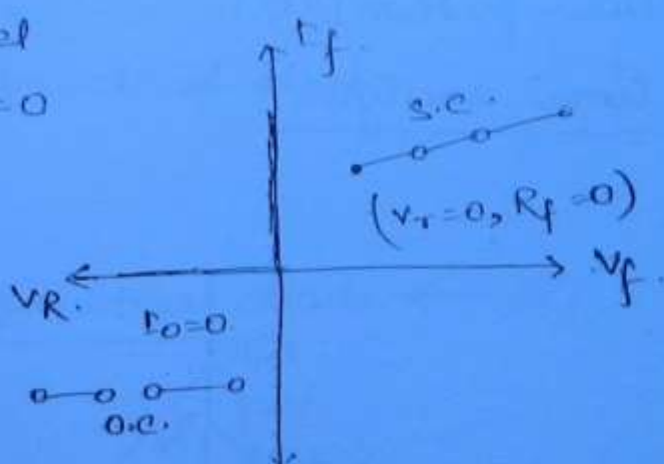
$R_f = 0$



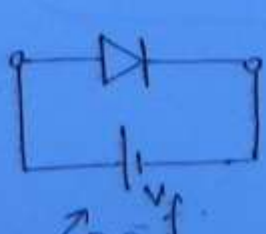
Case 3 →

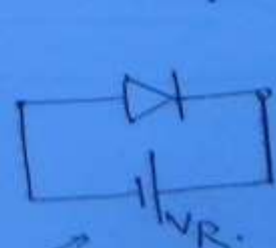
Ideal model

$\cdot V_\gamma = 0, R_f = 0$



Diode switching times :→





R.B. "t_{rr}" reverse Recovery time

F.B. "t_{fr}" forward Recovery time

t_{rr} :—

It is the time taken for a diode to change forward biased mode to reverse biased mode.

Reverse recovery time practically, it will be in the order of μs.

(62)

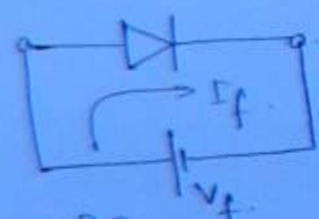
t_{fr} :—

It is the time taken for a diode to change reverse bias mode to forward bias mode.

Practically it will be in the order of Pico secs.

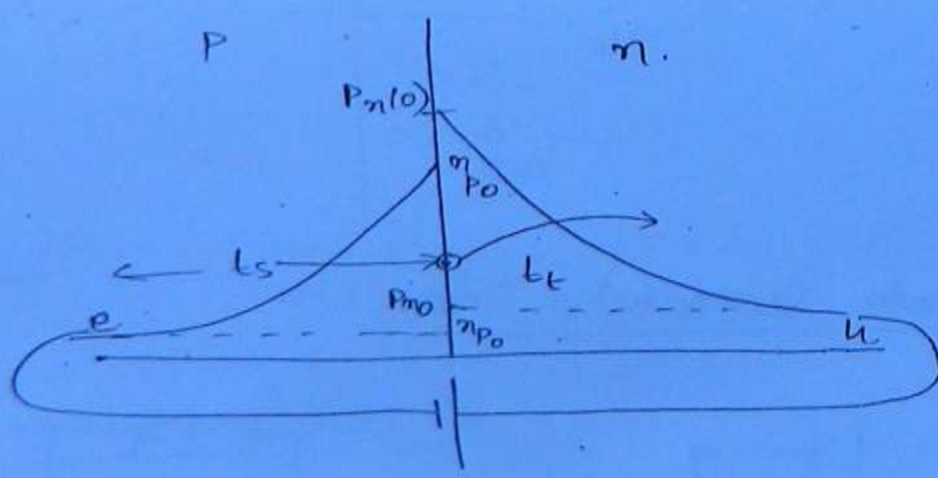
In high freq. applications, forward recovery time is neglected in the analysis (Pico secs) but reverse recovery time becomes a serious problem (μs)

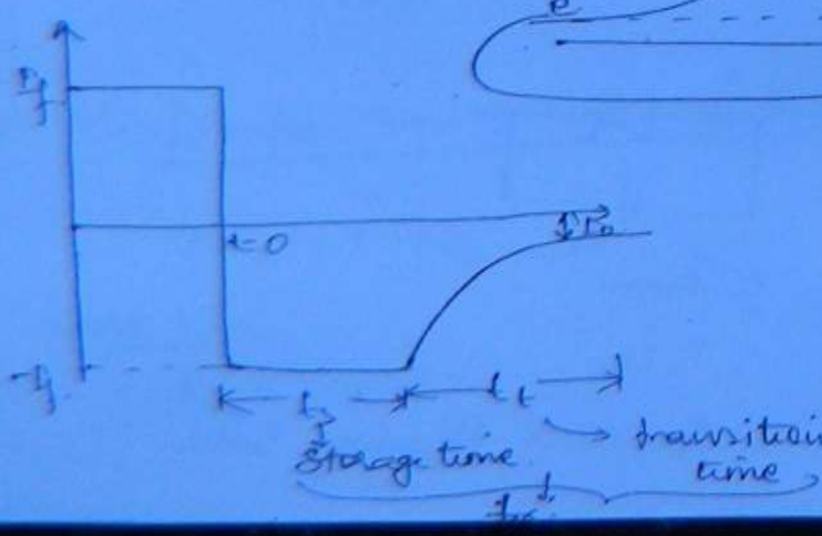
Reverse Recovery time analysis :→



R.B

t = 0

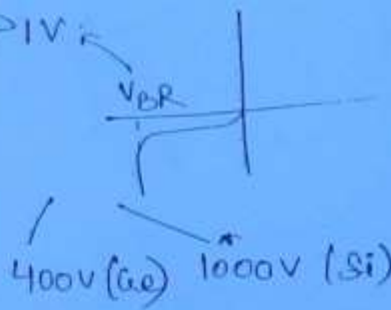




Q Why Si is preferred over Ge in most of the applications?..

Ans. 1) Temp - 200°C (Si)
- 100°C (Ge)

2) PIV :-



PIV (Peak Inverse Voltage) →
It is the max reverse bias voltage applied to a dio

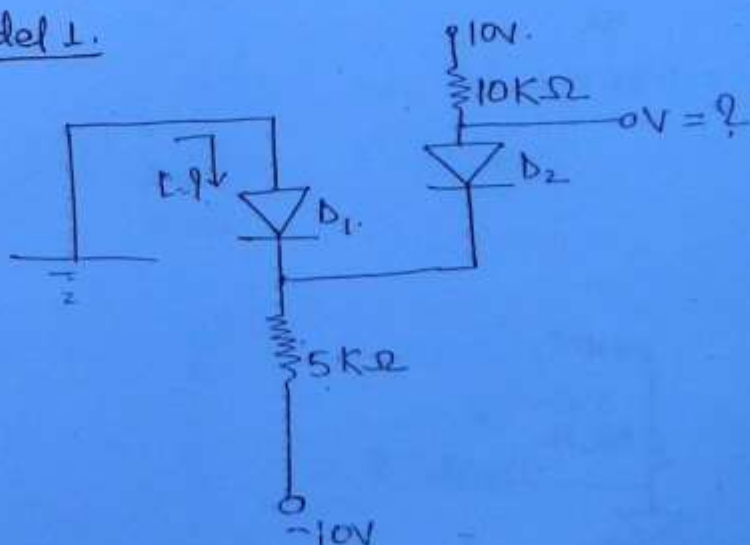
The main drawback of silicon is cut-in voltage.

Diode problems →

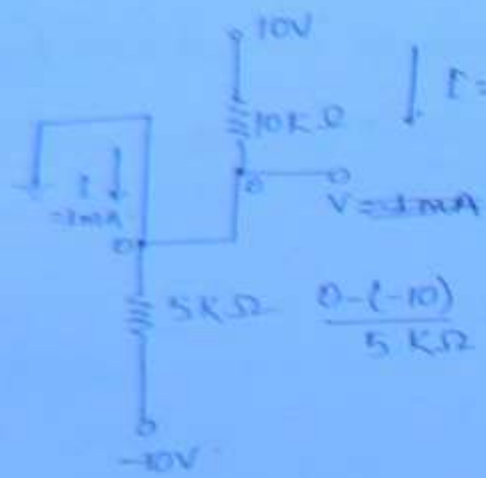
→ Ideal and practical diode
→ clippers
→ clampers.

Ideal and practical diode :—

Model 1.



Assume D_1 and D_2 FB.

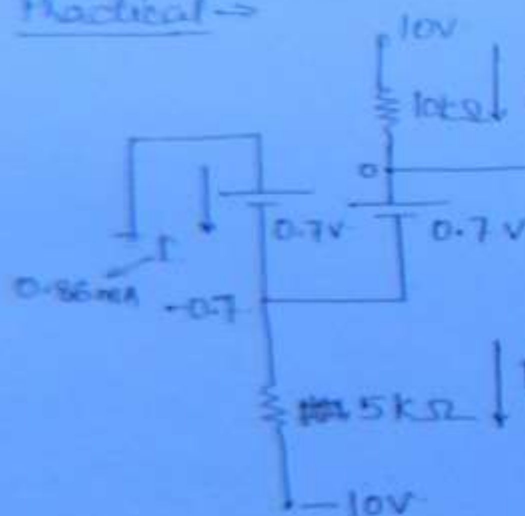


$I = \frac{10-0}{10k} = 1mA$

$\frac{0-(-10)}{5k\Omega} = 2mA$

$V = 0V$
$I = 1mA$

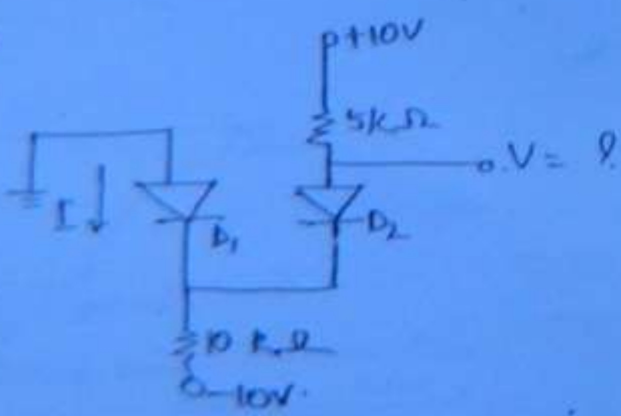
Practical →



$I = \frac{-0.7+10}{5k}$
$= 1.86mA$

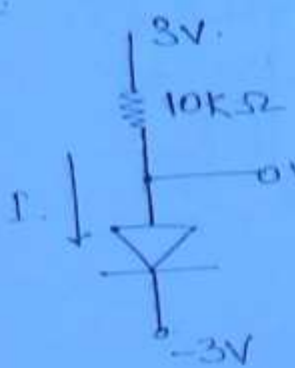
$V = 0V$
$I = 0.86mA$

model 2

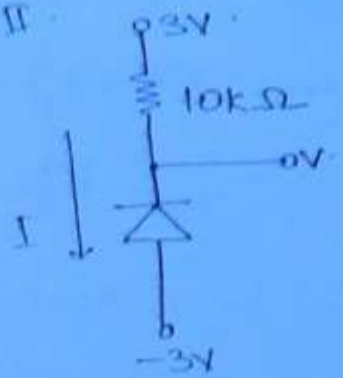


Single branch problem. → Model-2

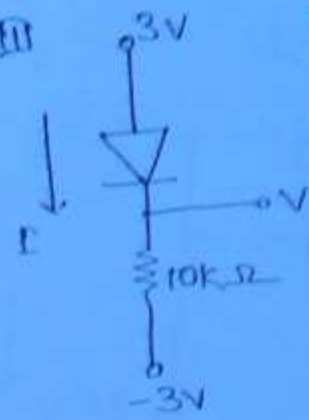
I.



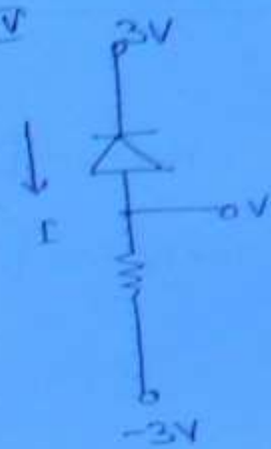
II



III

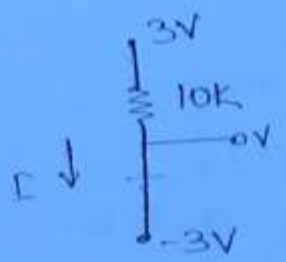


IV



65

I.

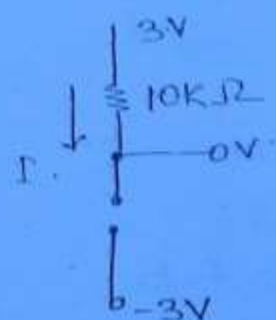


$$I = \frac{3-(-3)\ V}{10K}$$

$$= 0.6\ mA$$

$$V = -3V$$

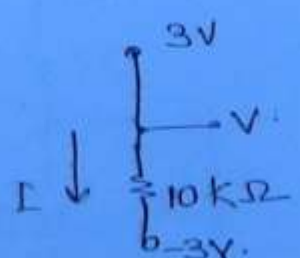
II



$I = 0$

$V = 3V$

III.



$V = 3V$

$I = 0.6mA$

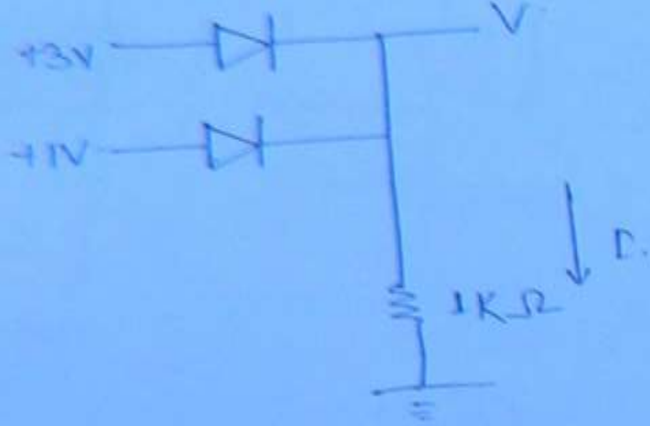
IV

$V = -3V$

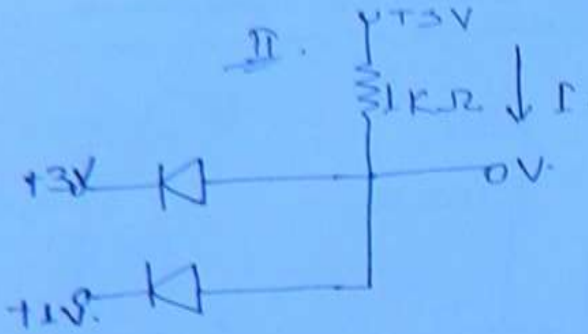
$I = 0$

model - 3. →

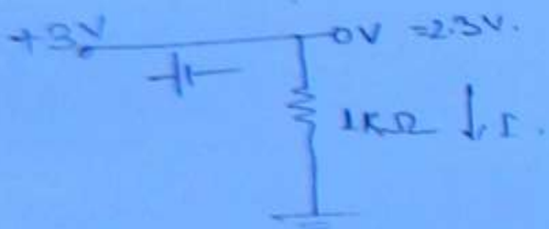
I.



II.



I. OR - logic.

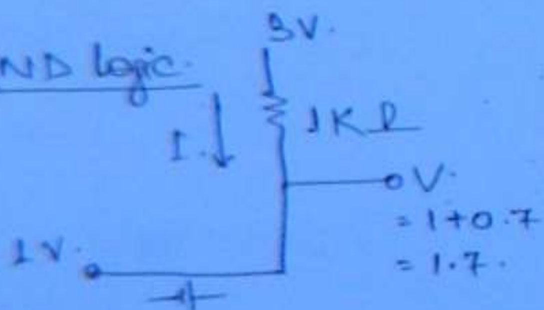


highest voltage is F.B.

V = 3V

I = 3mA

II. AND logic.



lowest voltage is F.B.

I = 2mA.

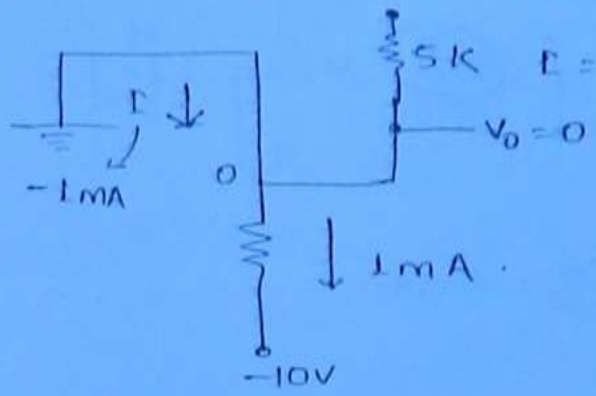
V = 1V.

⇒ In OR logic problems highest I/P voltage applied to the diode will be the O/P.

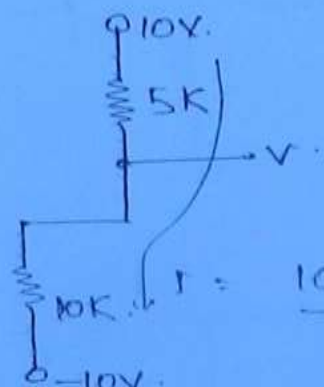
⇒ In AND logic problems, lowest I/P voltage applied to the diode will be the O/P.

ideal →

Assume D_1 and D_2 are F.B.



$I = 2mA$

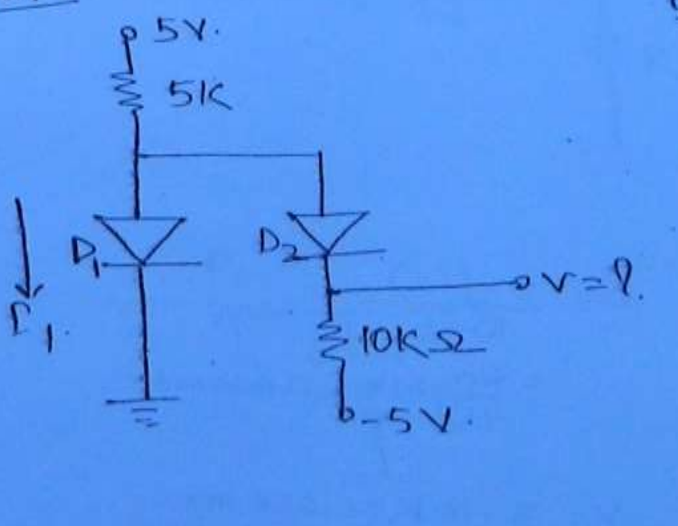


$D_1 \rightarrow R.B.$
$D_2 \rightarrow F.B.$

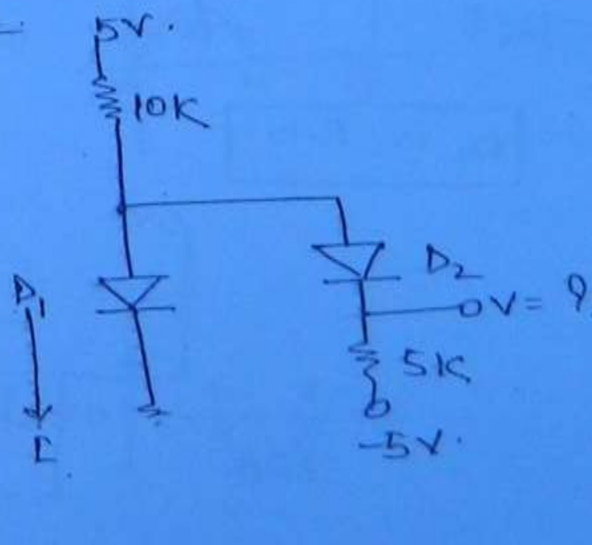
$$I = \frac{10-(-10)}{15K} = 1.33\ mA$$

$$V = 10K \times 1.33mA - 10$$
$$= 13.3 - 10$$
$$= 3.3V$$

ckt 1.



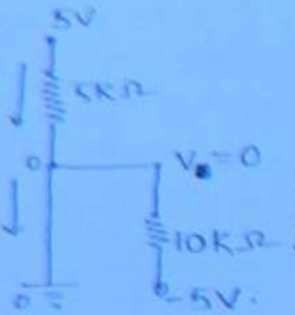
Ckt-2.



Ckt 1.

Assume D_1 and D_2 are F.B.

$I = \frac{5-0\,V}{5K\Omega} = 1mA$



$\frac{1}{2}mA = I$

$I = \frac{0-(-5)}{10K\Omega} = \frac{1}{2}mA$

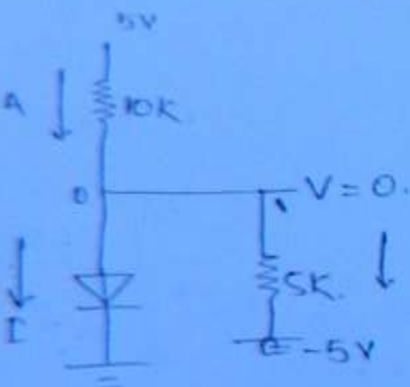
Assumption is true

∴ D_1 and D_2 are F.B.

Ckt 2.

Assume D_1 and D_2 are F.B.

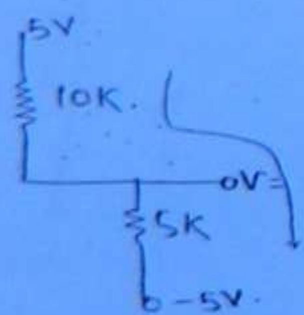
$\frac{1}{2}mA$



$-\frac{1}{2}mA = I$

$I = \frac{0-(-5)}{5K} = 1mA$

$\boxed{D_1 \text{ is R.B}}$



$I = 0.$

$I = \frac{5-(-5)}{10+5}$

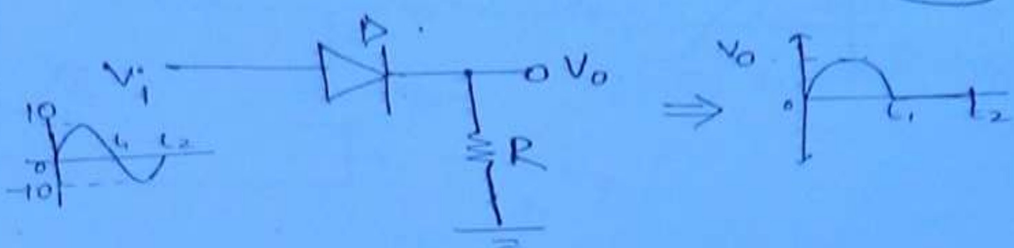
$= \frac{\cancel{10}^{2}}{\cancel{15}_{3}} mA = 0.666mA.$

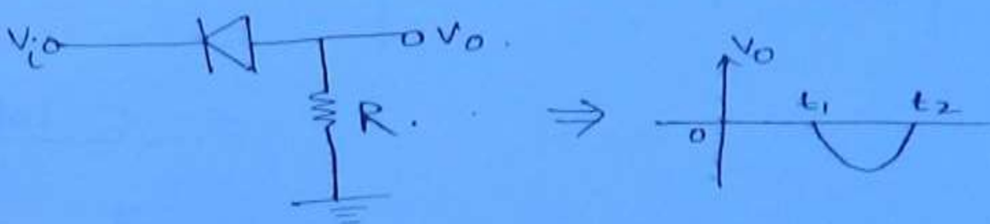
$V = 5 - 10K \times 0.666\,mA$

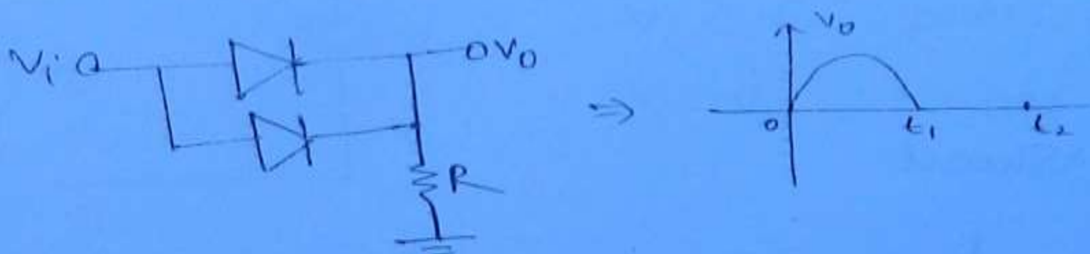
$= -1.67\,V.$

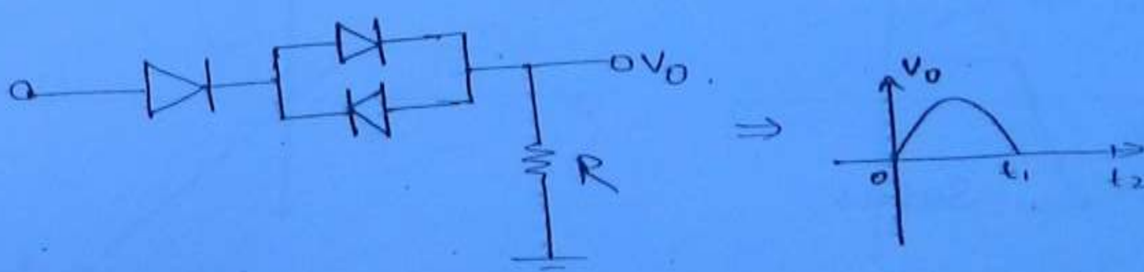
Model 4:

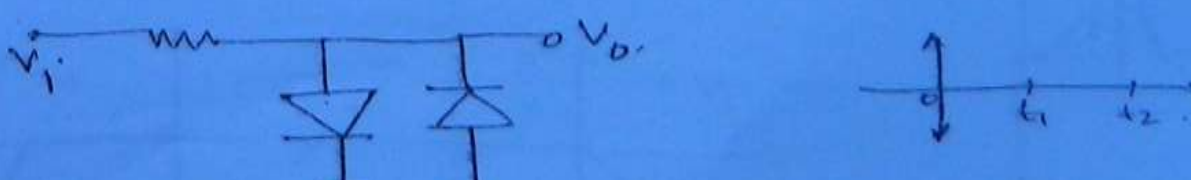
AC Analysis.

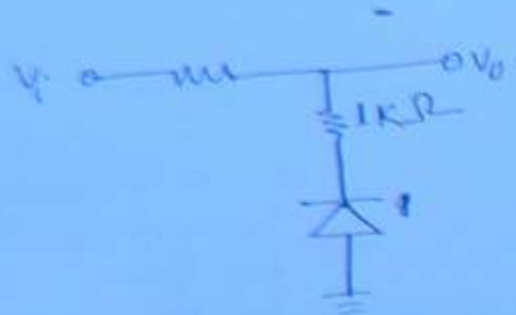


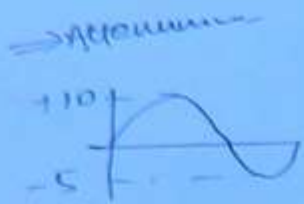




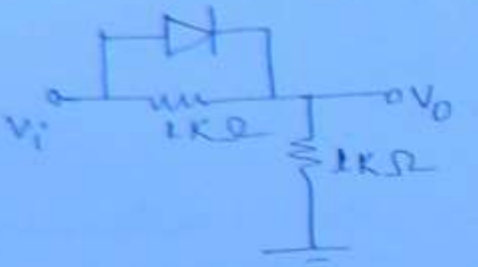




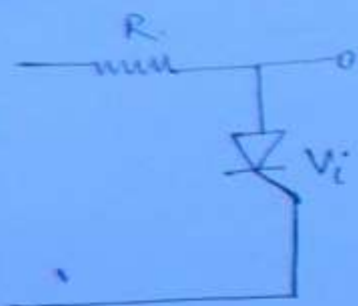




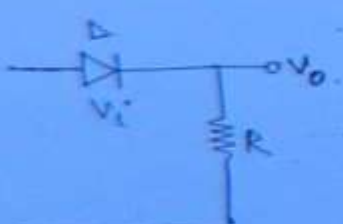
70



Clippers →

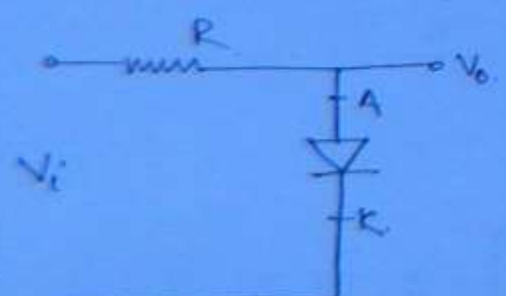


→ Shunt



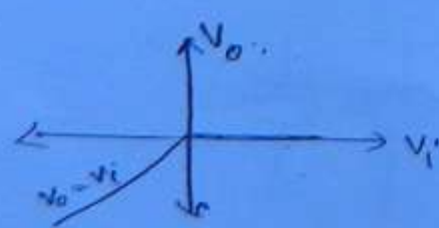
→ Series

Shunt clippers →



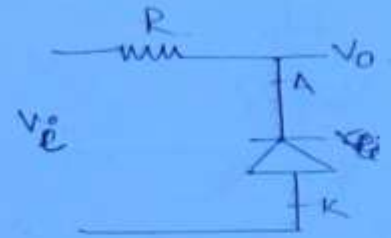
$V_i < 0 \rightarrow D \rightarrow$ OFF $\quad V_0 = V_i$

$V_i > 0 \rightarrow D \rightarrow$ ON. $\quad V_0 = 0$.



21

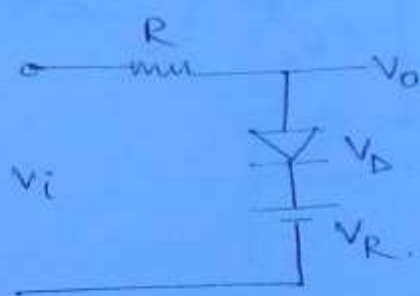
Model 1. →



$V_i < 0$, D-ON $V_o \to 0$.

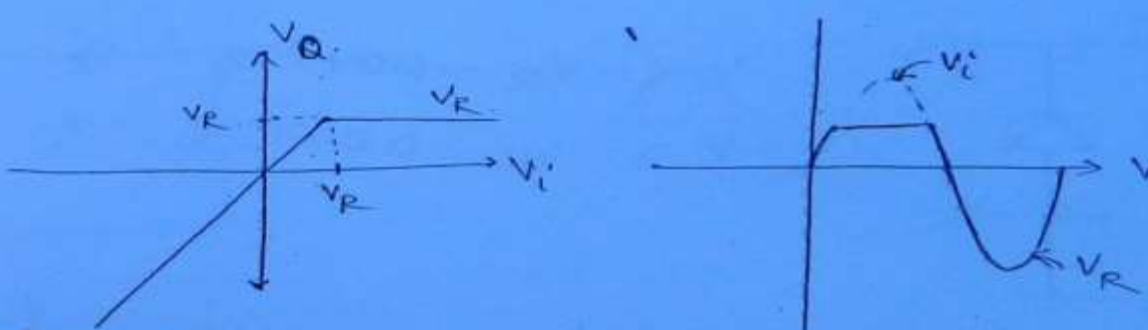
$V_i > 0$, D-OFF $V_o = V_i$

Model-3 →

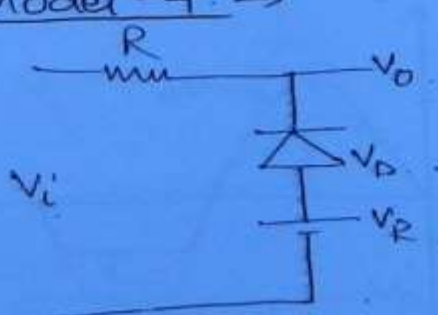


$V_i < V_R$, D-OFF, $V_o = V_i$

$V_i > V_R$, D-ON, $V_o = \cancel{V_D} + V_R$.

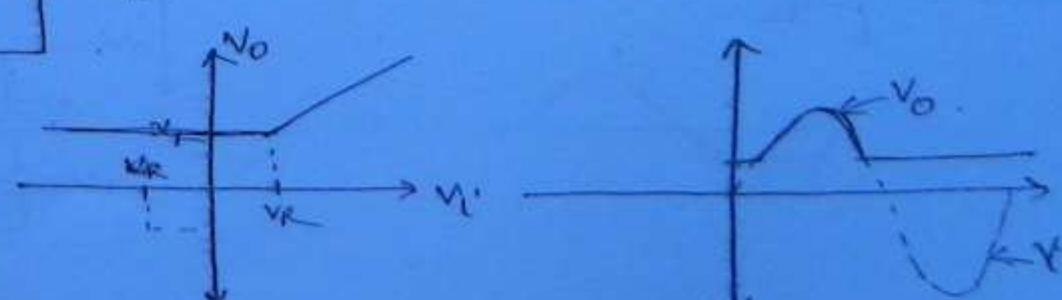


model-4 →

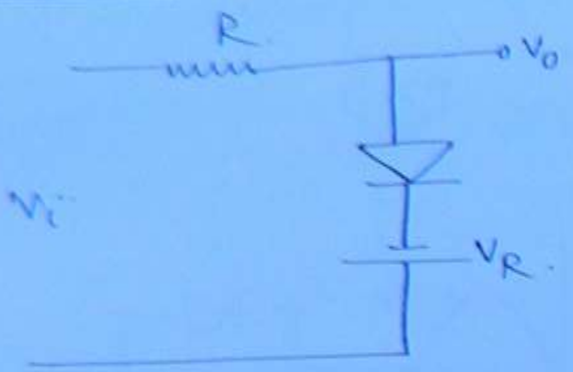


$V_i < V_R$, D-ON, $V_o = V_R$

$V_i > V_R$, D-OFF, $V_o = V_i$

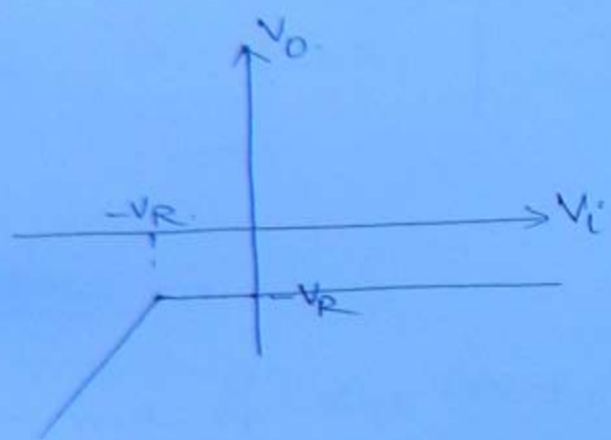


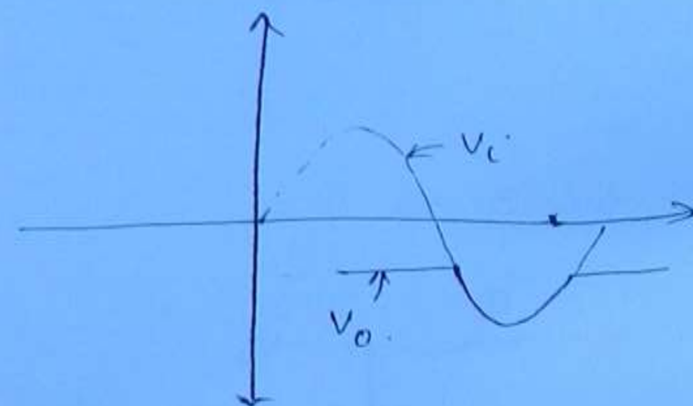
model 5 →



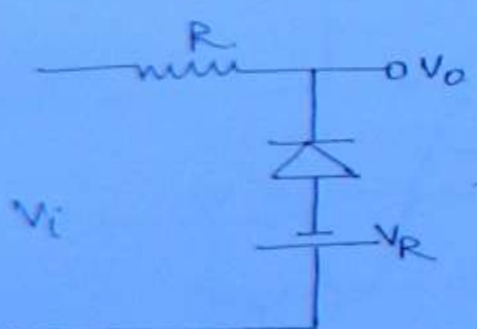
$V_i < -V_R \rightarrow$ D OFF $\quad V_o = V_i$

$V_i \geq -V_R \rightarrow$ D ON $\quad V_o = -V_R$



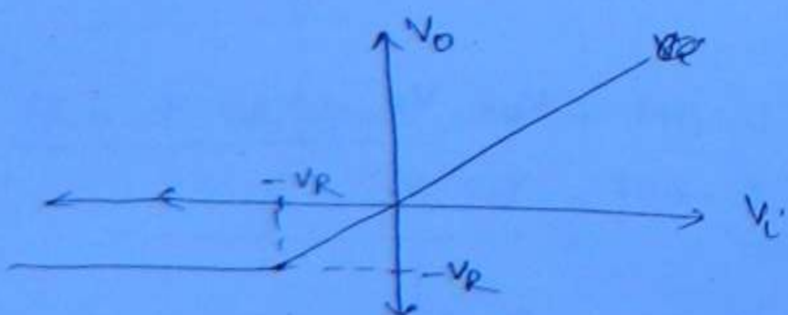


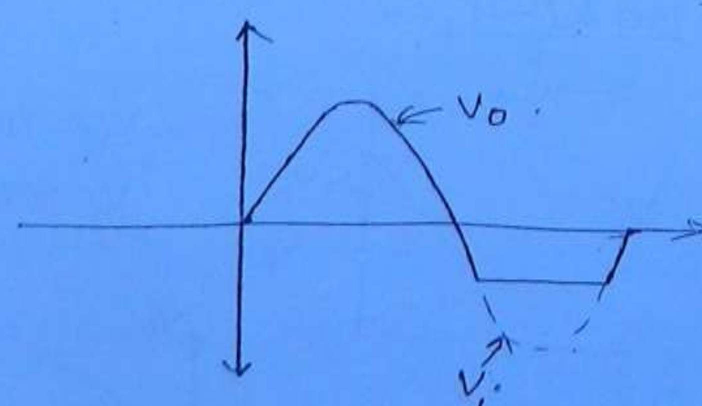
model 6 —

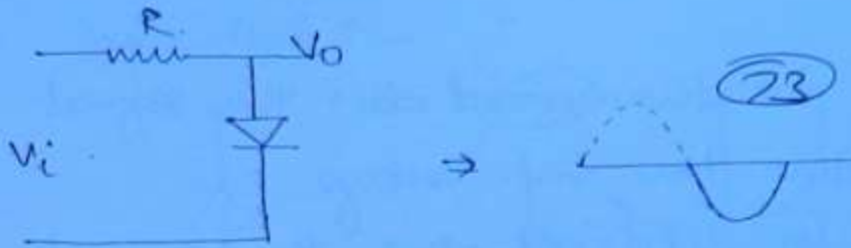


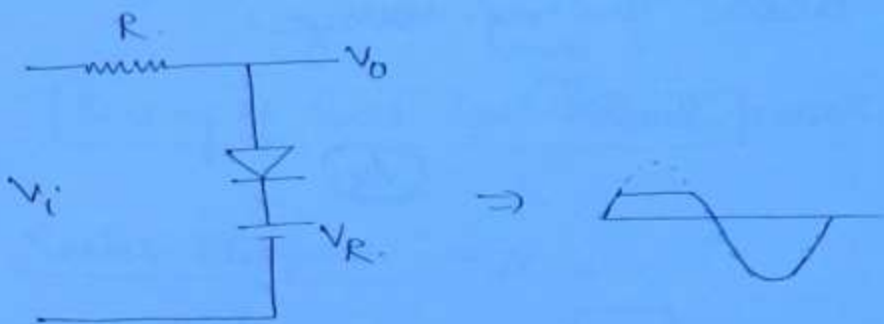
$V_i < -V_R \rightarrow$ D ON $\quad V_o = -V_R$

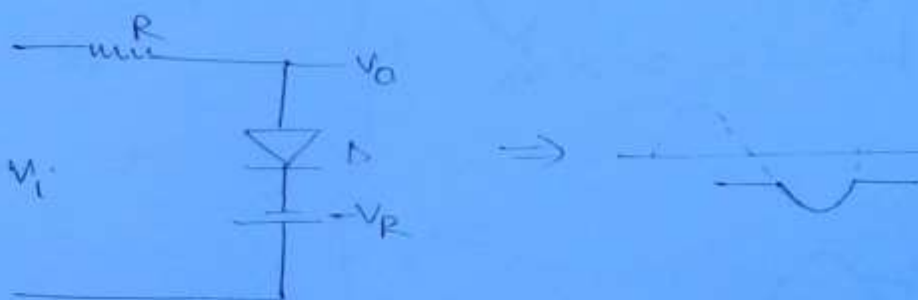
$V_i > -V_R \rightarrow$ D OFF $\quad V_o = V_i$

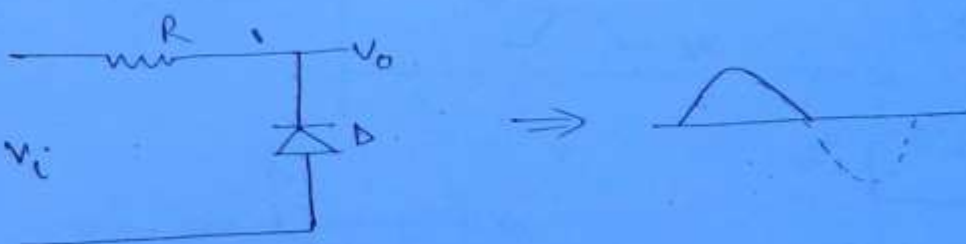





R
Vo
Vi
⇒


R
Vo
Vi
VR
⇒


R
Vo
Vi
D
VR
⇒


R
Vo
Vi
D
⇒


R
Vo
Vi
VR
⇒

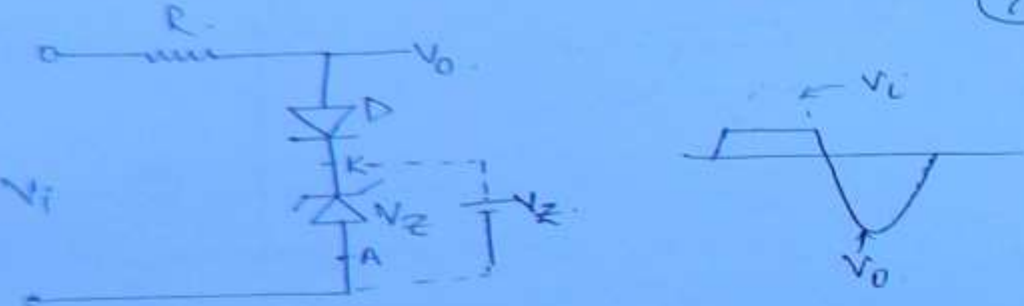

R
Vo
Vi
D
VR
⇒

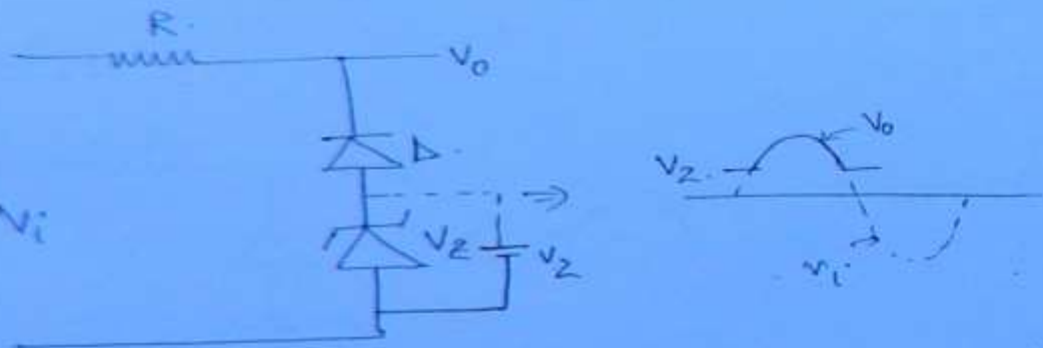
Conclusion :—

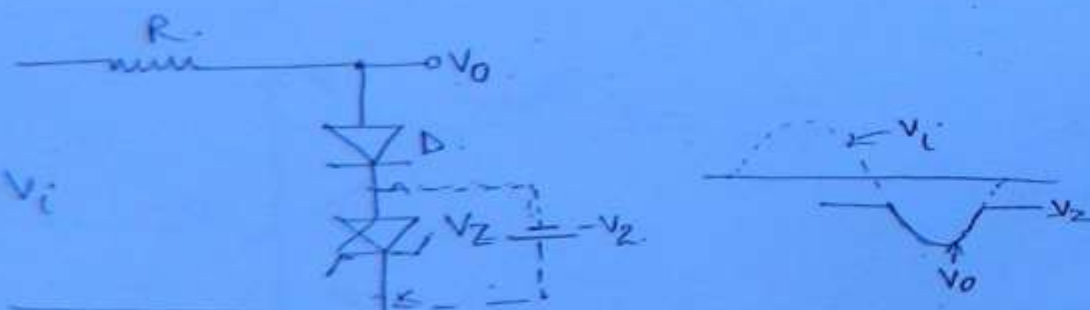
1) When the diode is in downward dirⁿ the signal will be transmitted below the ref. voltage.

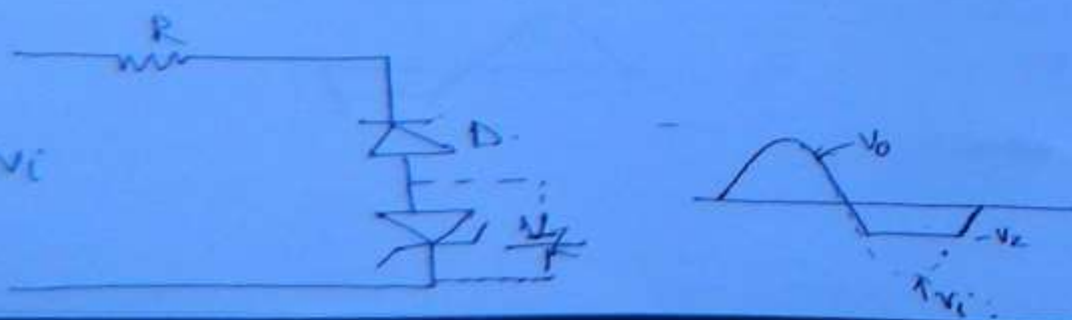
2) When the diode is in upward dirⁿ the signal will be transmitted above the ref. voltage.

Typical models → • Zener is ON.

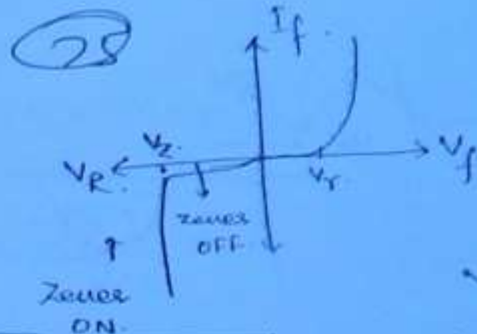








Zener diode →



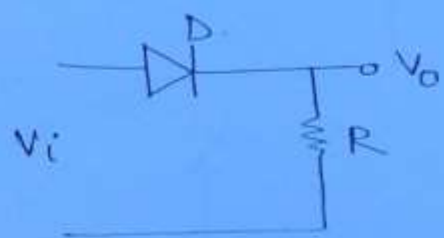
$V > V_Z \rightarrow$ ON. (K —|⊢— A, V_Z)

$V < V_Z \rightarrow$ OFF (O.C.)

$V \rightarrow V_f \rightarrow$ FB (S.C)

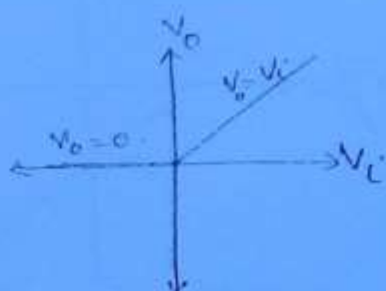
$\boxed{\text{Best eg of const volt. source}}$

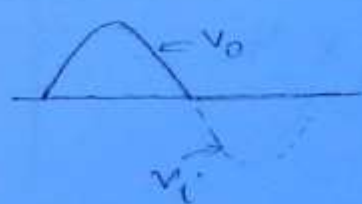
Series clippers →

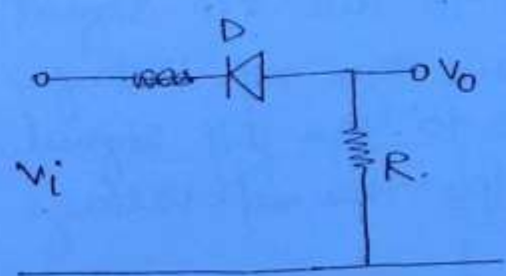


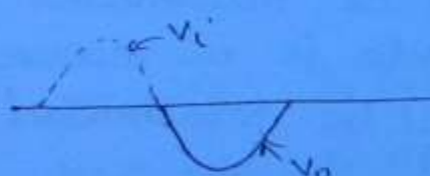
$V_i < 0 \rightarrow D - \text{OFF} \rightarrow \text{O} \quad V_o = 0$

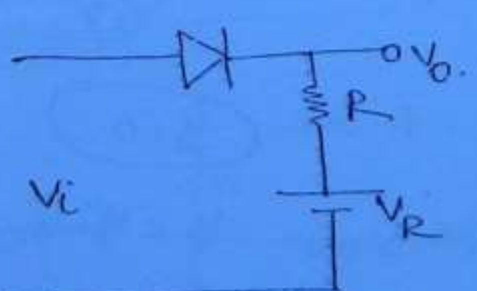
$V_i > 0 \rightarrow D \rightarrow \text{ON} \quad V_o = V_i$

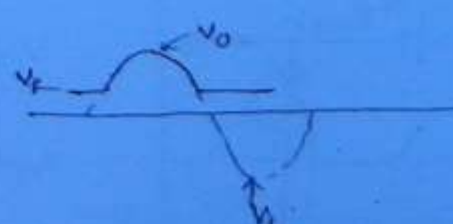


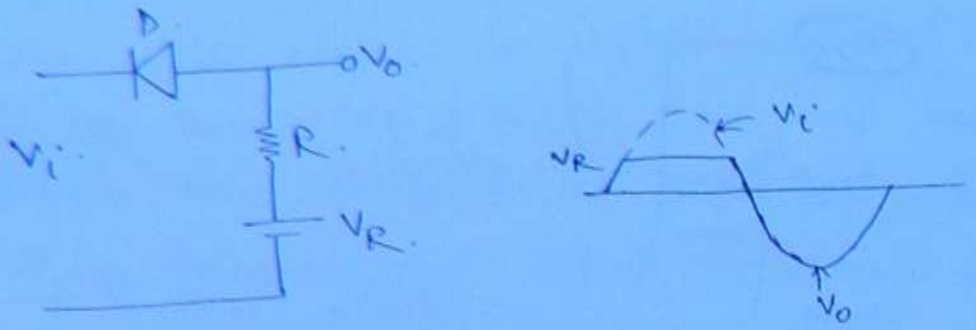


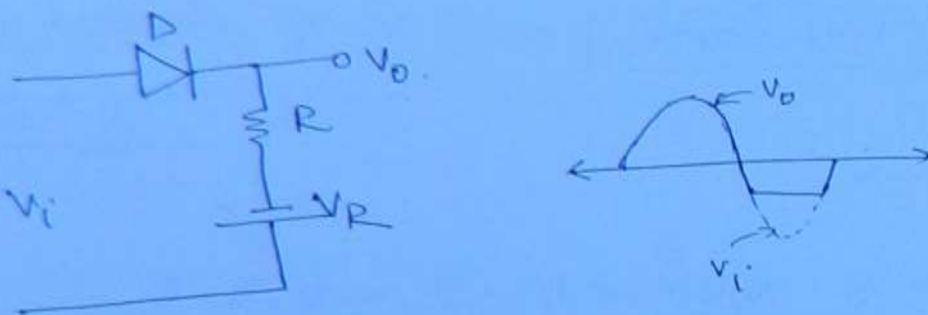


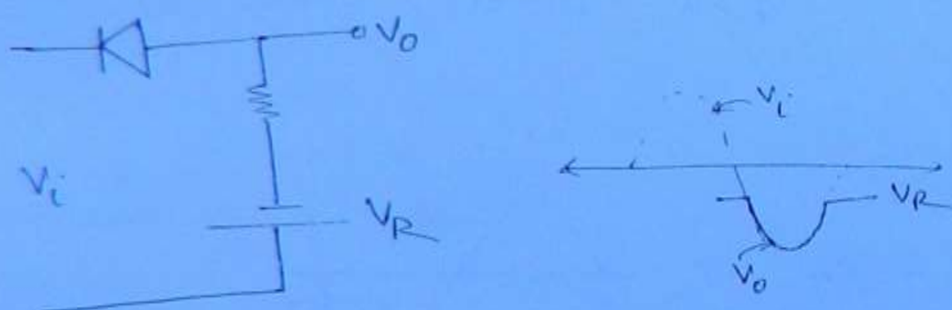










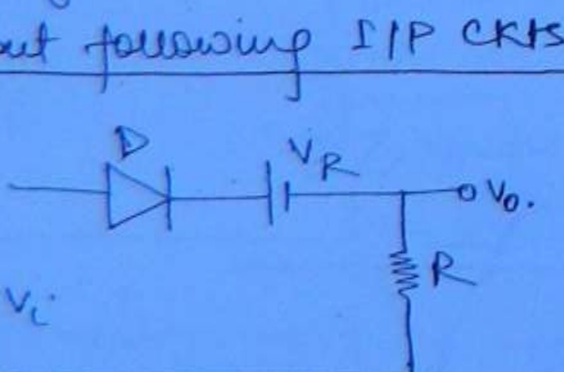


<u>Conclusions →</u>

> When the diode is in forward dirⁿ to the I/P signal, the signal will be transmitted above the ref. voltage.

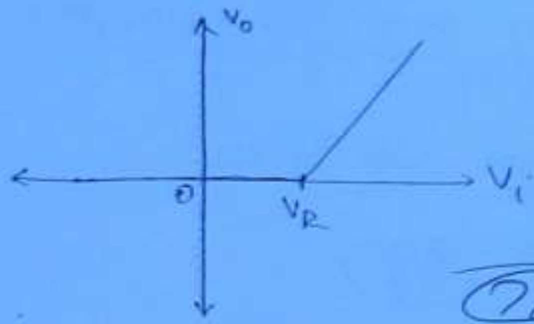
→ When the diode is in reverse dirⁿ to the I/P signal, the signal will be transmitted below the ref. voltage

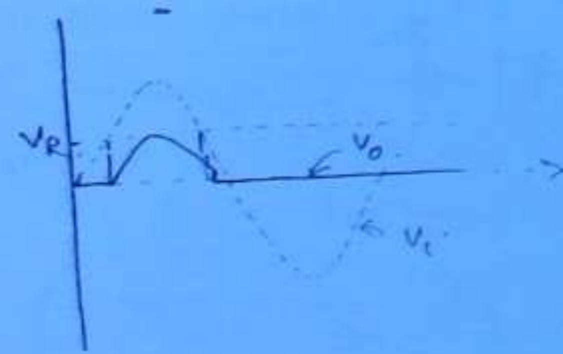
<u>Output following I/P ckts →</u>



$V_i < V_R$. D OFF $V_o = 0$.

$V_i > V_R$ D ON $V_o = V_i - V_R$





77

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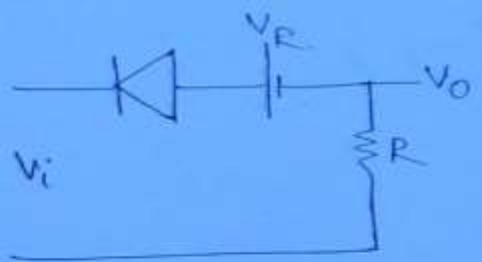
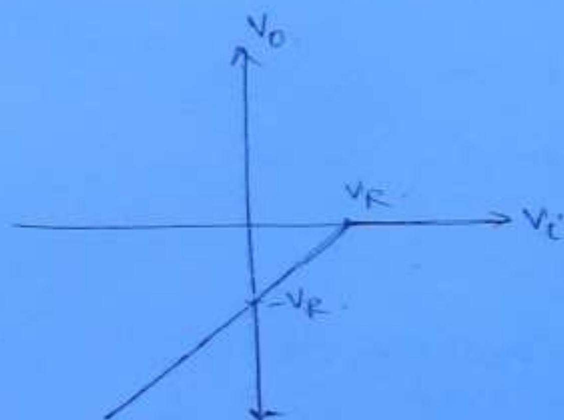


$V_i < V_R$, D-ON , $V_o = V_i - V_R$

$V_i > V_R$, D-OFF , $V_o = 0$



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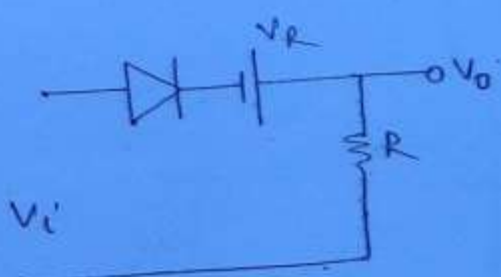
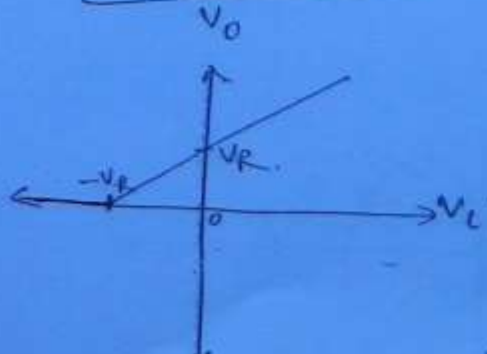
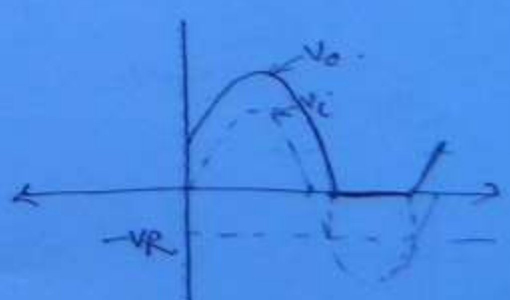
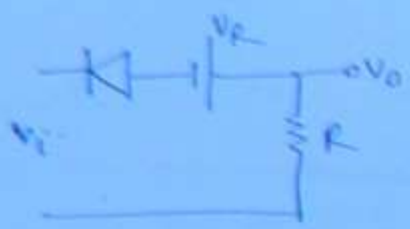


$V_i > -V_R$ D ON $V_o = \cancel{V_o}\ V_i + V_R$

$V_i < -V_R$ D OFF $V_o = 0$

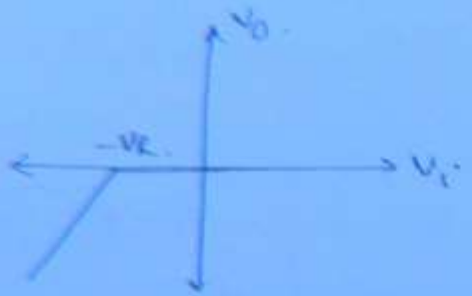


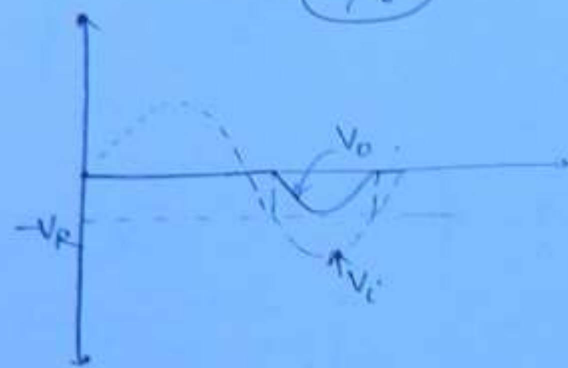




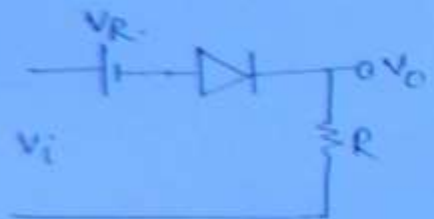
$V_i > -V_R$. D OFF. $V_o = 0$

$V_i < -V_R$ D ON $V_o = V_i + V_R$





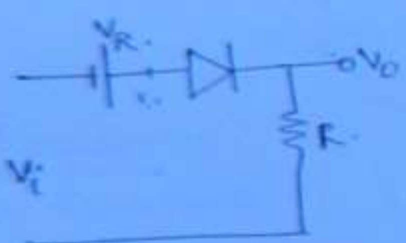
Q1.



$V_i > V_R$ D → ON $V_o = V_i - V_R$

$V_i < V_R$ D → OFF $V_o = 0$

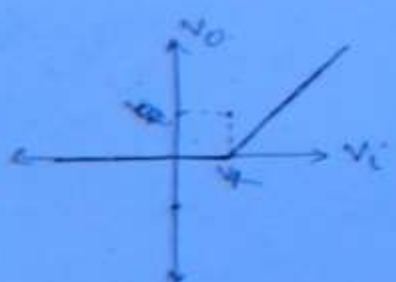
2.

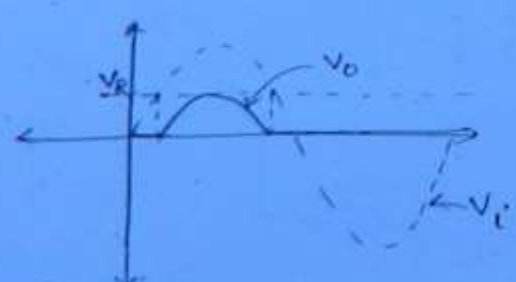


$V_i > -V_R$ D ON $V_o = V_i + V_R$.

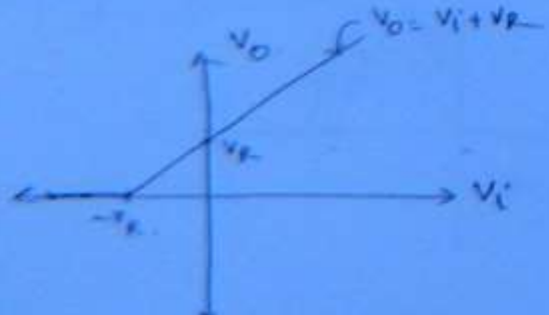
$V_i < -V_R$ D OFF $V_o = 0$.

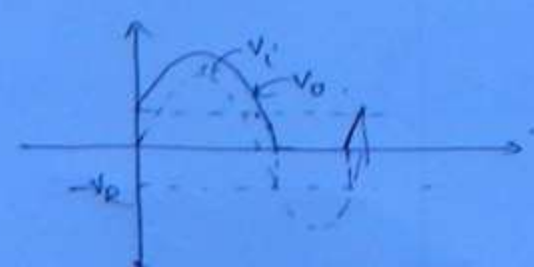
Sol(1)





Sol(2)

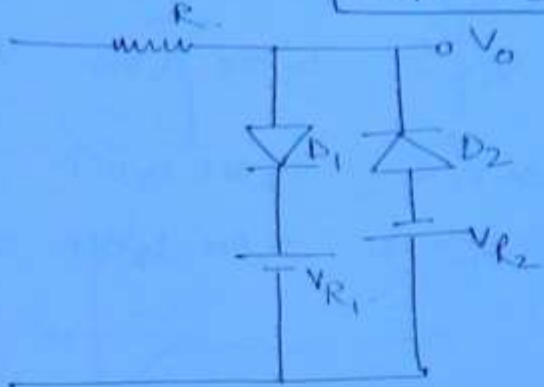


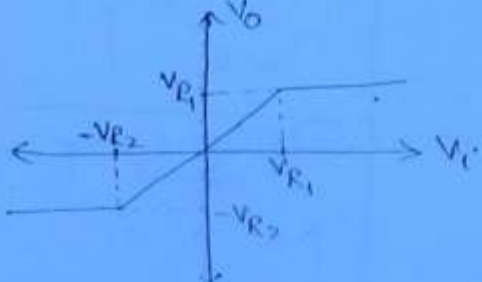


Clipping at two independent levels :→

$\boxed{V_{R_1} > V_{R_2}}$ ← always.

29





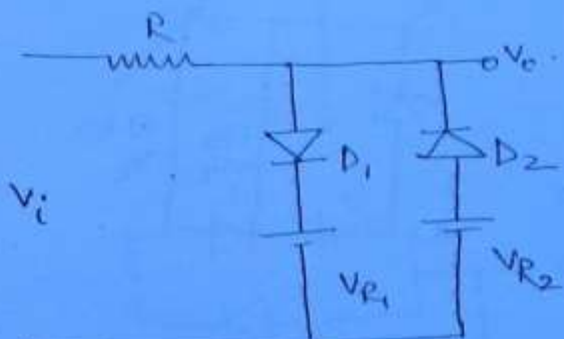
$V_i < -V_{R_2}$ D_1 OFF D_2 ON $V_o = -V_{R_2}$

$-V_{R_2} < V_i < V_{R_1}$ D_1 OFF D_2 OFF $V_o = V_i$

$V_i > V_{R_1}$ D_1 ON D_2 OFF $V_o = V_{R_1}$.

3/01/12.

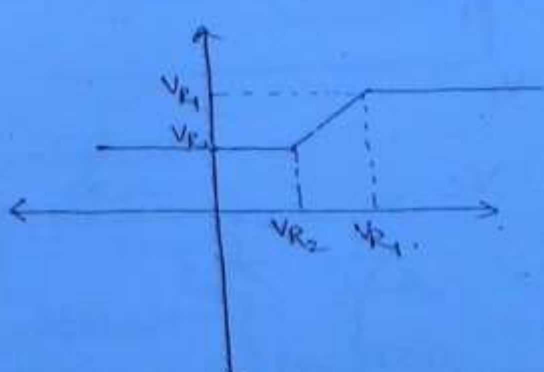
$V_{R_1} > V_{R_2}$

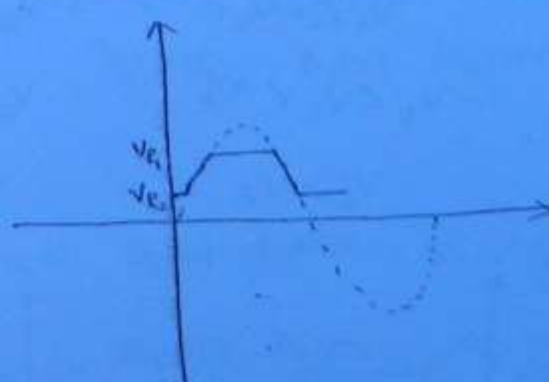


$V_i < V_{R_2}$. D_1 OFF D_2 ON. $V_o = V_{R_2}$.

$V_{R_2} < V_i < V_{R_1}$ D_1 OFF D_2 OFF $V_o = V_i$

$V_i > V_{R_1}$ D_1 ON D_2 OFF $V_o = V_{R_1}$





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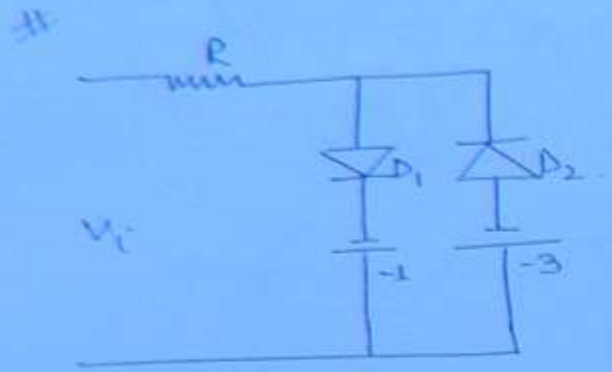
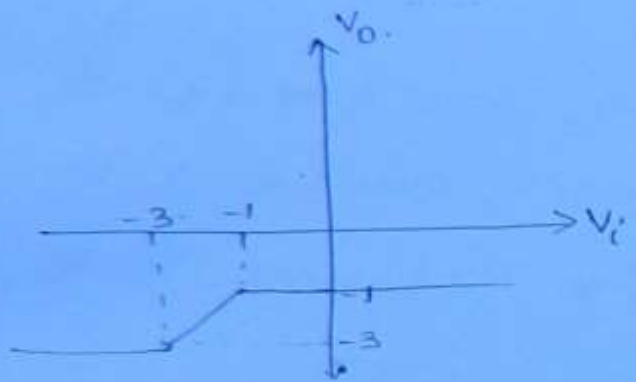
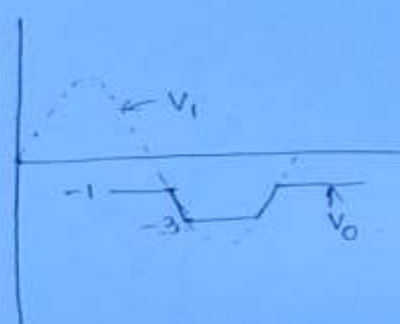


$V_i < -3$ D_1 OFF D_2 ON $V_0 = -3V$.

$-3 < V_i < -1$ D_1 OFF D_2 OFF $V_0 = V_i$

$V_i > -1$. D_1 ON D_2 OFF. $V_0 = -1V$.





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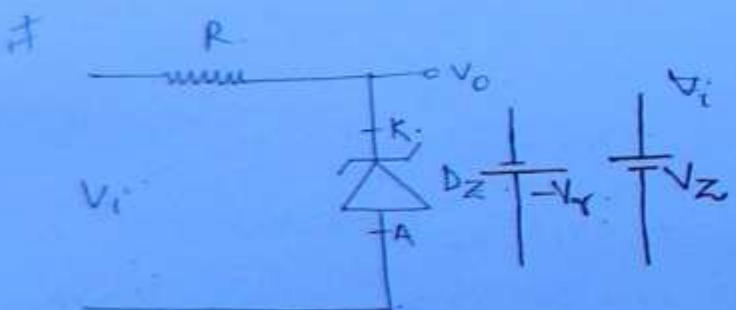
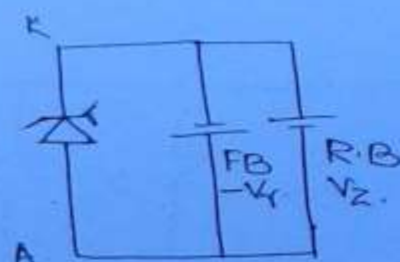
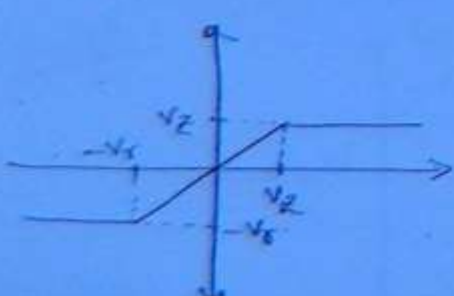


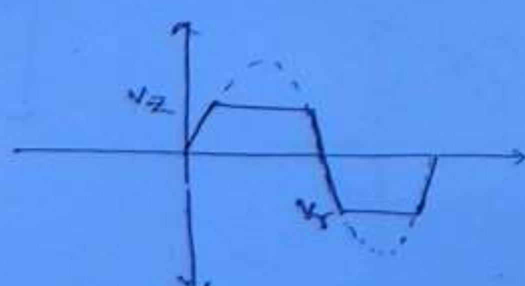


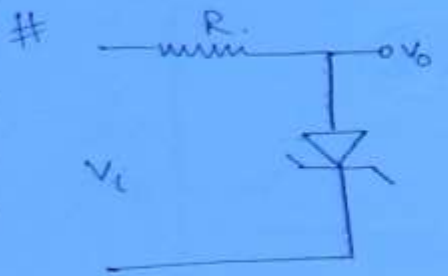
$V_i < -V_r$ $V_0 = -V_r$.

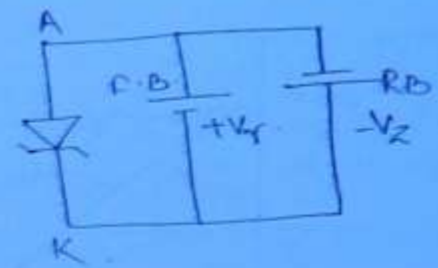
$-V_r < V_i < V_Z$ $V_0 = V_i$

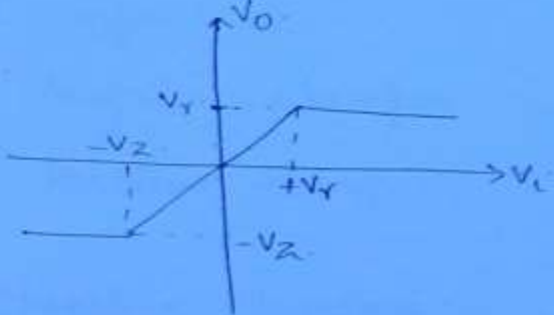
$V_i > V_Z$. $V_0 = V_Z$.



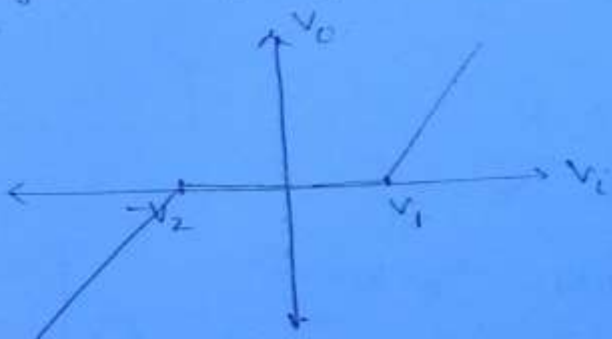




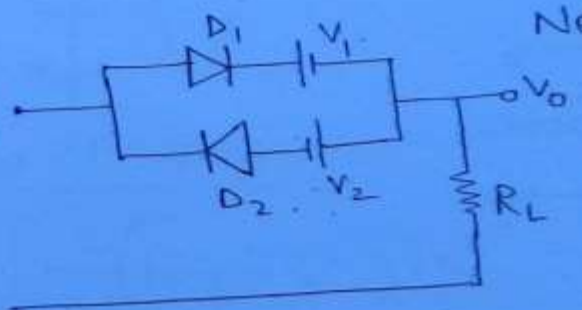




Design a clipping ckt for the transfer char. shown below.



Ans.



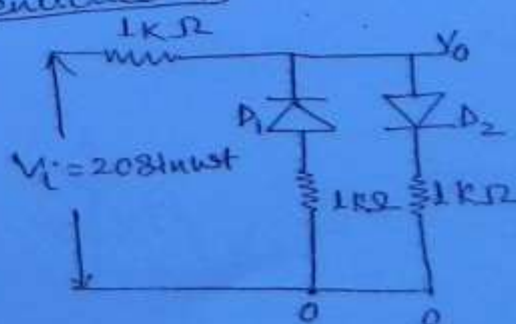
Noise clippers design.

$V_i < -V_2$ D_1 OFF D_2 ON. $V_o = V_i + V$

$-V_2 < V_i < V_1$ D_1 OFF D_2 OFF $V_o = 0$.

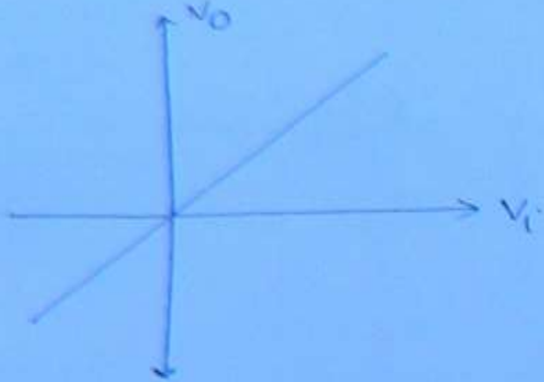
$V_i > V_1$ D_1 ON D_2 OFF $V_o = V_i - V$

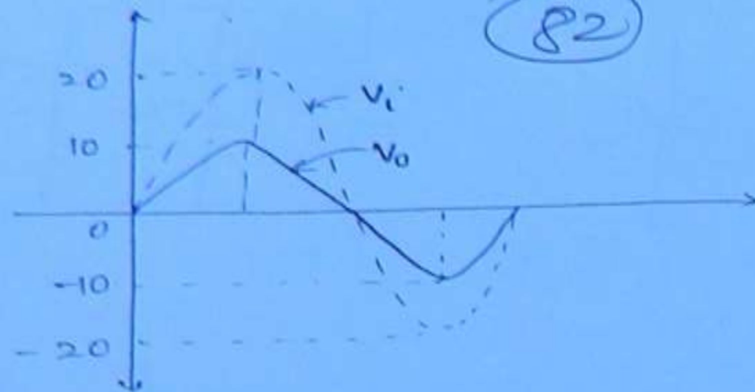
Attenuated ckt



$V_i < 0$. D_1 ON D_2 OFF $V_o = \frac{V_i}{2}$

$V_i > 0$ D_1 OFF D_2 ON $V_o = \frac{V_i}{2}$.





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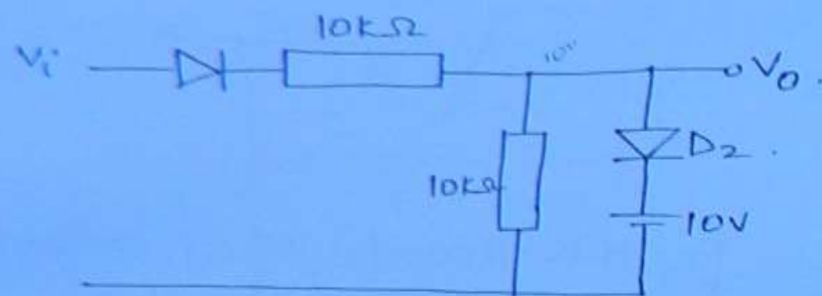
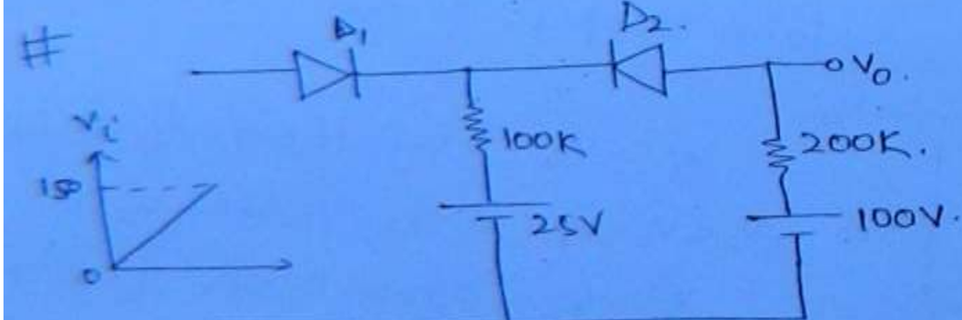


$V_i < 0$ D_1 OFF $V_o = 0$

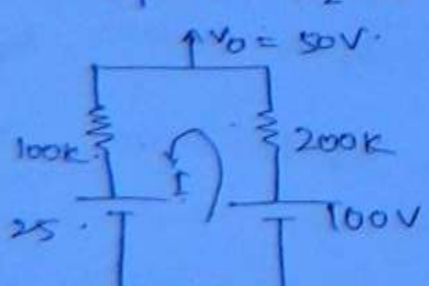
$0 < V_i < 20V$ D_1 ON D_2 OFF $V_o = \frac{V_i}{2}$

$V_i > 20V$ D_1 ON D_2 ON. $V_o = 10V$.

#



At $V_i = 0V$, D_1 OFF D_2 ON.



$I = \frac{75}{300} = 0.25V$

$V_i = 0$ D_1 OFF D_2 ON. $V_0 = 50V$.

$50 < V_i < 100$ $V_0 = V_i$

$V_i > 100V$ $V_0 = 100V$

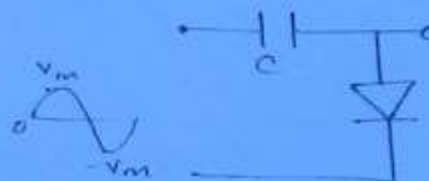
83

Clampers →

Negative clamper

Positive clamper.

Negative clampers :–

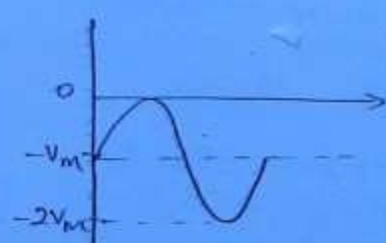


$V_0 = V_i - V_m$

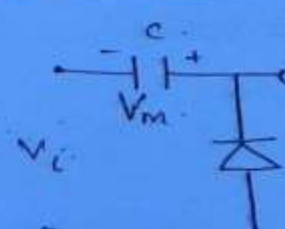
$V_i = V_m$ $V_0 = 0$

$V_i = 0$ $V_0 = -V_m$

$V_i = -V_m$ $V_0 = -2V_m$.



Positive clampers –

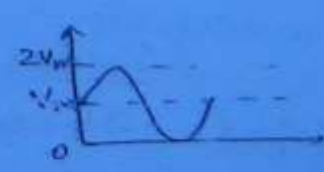


$V_0 = V_i + V_m$.

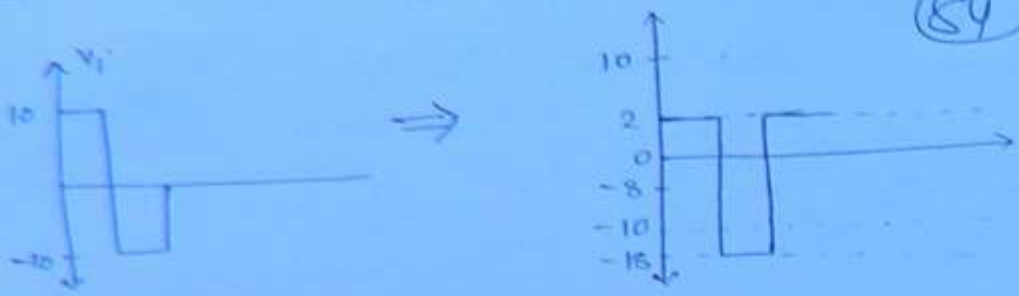
$V_i = V_m$ $V_0 = 2V_m$

$= 0$ $V_0 = V_m$

$= -V_m$ $V_0 = 0$.



CW-2 8-3



(d)

Conclusions →

1) When the diode is in downward dir^n, the total signal will be clamped below the ref. voltage.

2) When the diode is in upward dir^n, the total signal will be clamped above the ref. voltage.

Special diodes →

Zener diode →

10^8 S.C. atoms → 1 impurity → ordinary diode.

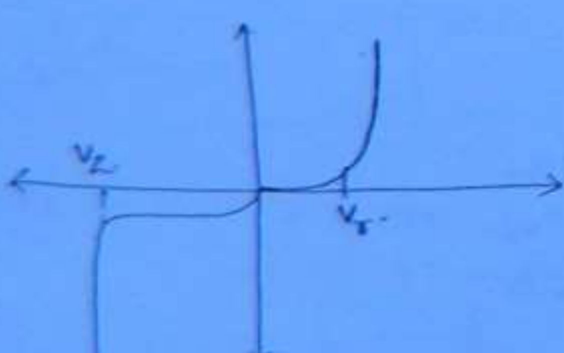
10^6 S.C. atoms → 1 impurity → Zener diode

10^3 S.C. atoms — 1 impurity → Tunnel diode.

Breakdown mechanism →

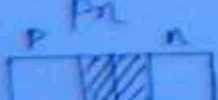
two types of breakdown.

- Avalanche breakdown
- zener breakdown.



Avalanche breakdown.

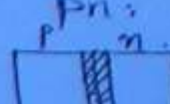
It occurs in a lightly doped pn



$\boxed{V_Z > 6}$

Zener breakdown.

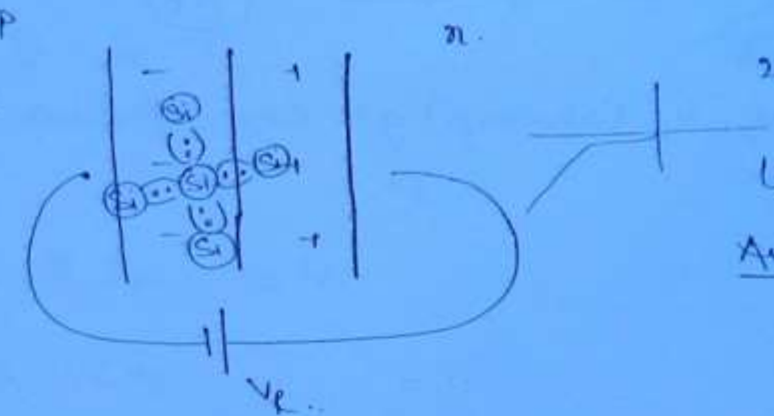
It occurs in a heavily doped pn.



$\boxed{V_Z < 6V}$

Mechanism →

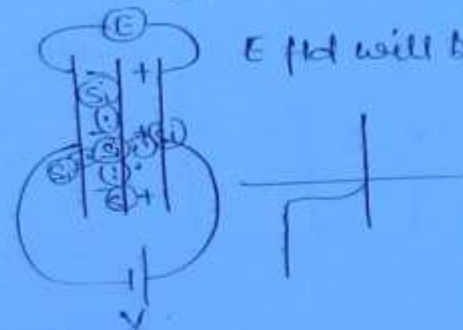
I.



Linear breakdown

Avalanche breakdown

II



E fld will be stronger.

instant breakdown

Zener breakdown

I. This type of breakdown occurs because of the velocity of minority charge carriers colliding with the stable atom in the depletion region.

II Due to heavily doped P-n j^n depletion region becomes narrower. When small reverse bias voltage is applied. Strong int. E-fld will be generated bcoz of immobile ions.

Temp co-eff. of V_{BR}

Zener breakdown →

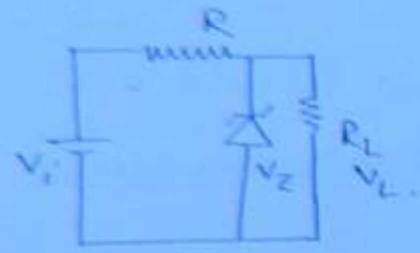
$T\uparrow, V_{BR}\downarrow$ ∴ negative temp coeff.

Avalanche breakdown:—

$T\uparrow, V_{BR}\uparrow$ ∴ +ive Temp coeff.

As T inc, vibrational effect inc., so we require more en.

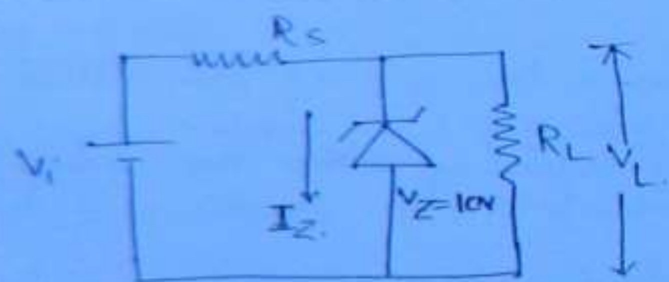
Zener diode as a voltage regulator →



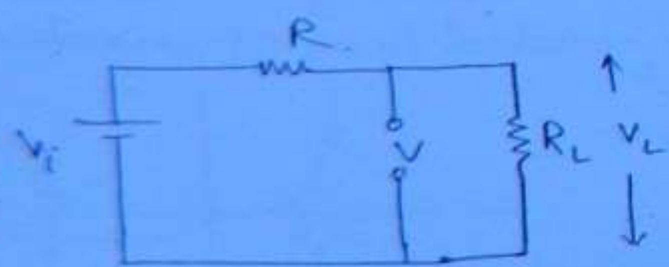
$V_Z = V_L$ (always) at any condition.

1) V_i & R_L fixed.
2) V_i fixed and R_L variable
3) V_i variable and R_L fixed.
4) V_i & R_L variable

I V_i & R_L fixed problem.

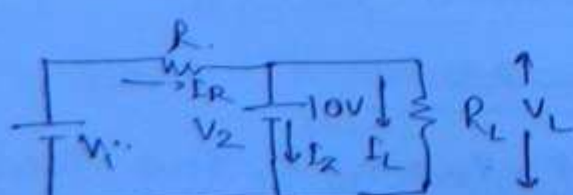


cal. V_L, I_Z, P_Z.



$$V = V_L = \frac{V_i \times R_L}{R + R_L}.$$

if $V > V_Z$,



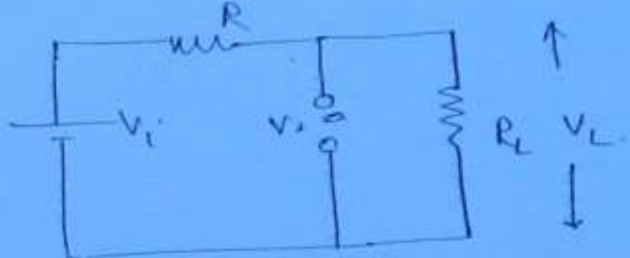
1) $V_Z = V_L = 10V$.

(2) $I_Z = I_R - I_L$.

$$= \frac{V_i - V_Z}{R} - \frac{V_Z}{R_L}.$$

(3) $P_Z = V_Z I_Z$

When $V < V_Z$.

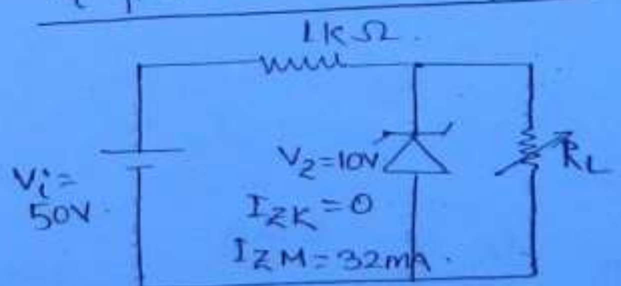


1) $V_L \cdot V$

2) $I_Z = 0$

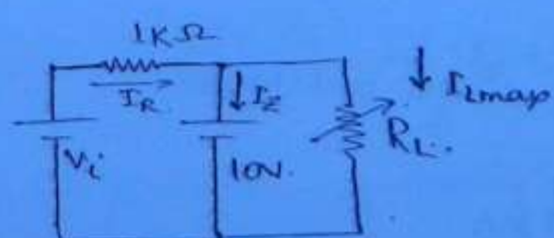
3) $P_Z = 0$

II V_i fixed and R_L variable →



B/w R_{Lmin} and R_{Lmax}, Zener diode is ON.

Cal. R_L min, R_{Lmax}, I_L min, I_{Lmax}.



$$I_{L\,max} = I_R - I_{ZK}$$

$$= \frac{50-10}{1K} - 0$$

$$= 40mA$$

$$R_{Lmin} = \frac{V_Z}{I_{Lmax}} = \frac{10}{40mA} = 250\Omega.$$

$I_{Lmin} = I_R - I_{ZM}$

$= 40mA - 32mA$

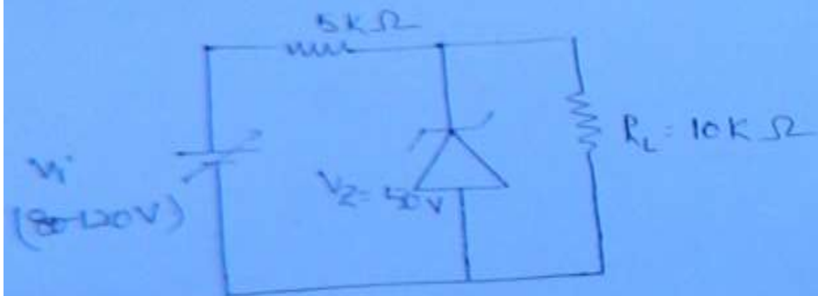
$= 8mA$

$R_{Lmax} = \frac{V_Z}{I_{Lmin}}$

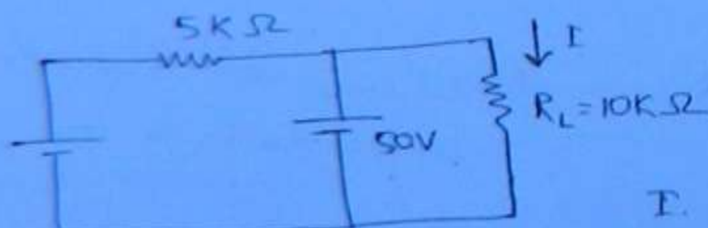
$= \frac{10}{8mA}$

$= 1.25 K\Omega$

B. <u>V_i variable and R_L fixed :—</u>



Cal. $I_{Z min}$ and $I_{Z max}$.



$I = \frac{50 V}{10 K\Omega} = 5mA.$

<u>I_{Zmax},</u>

$I_{R max} = \frac{120 - 50}{5K\Omega} = \frac{70}{5} = 14 mA.$

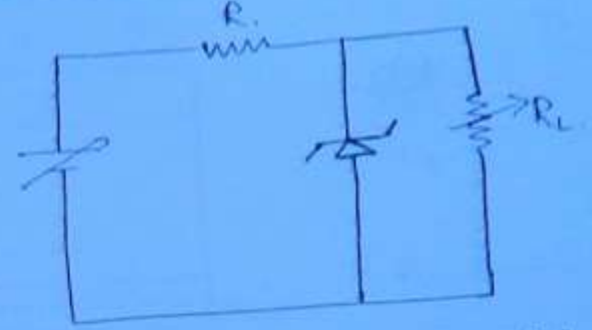
$I_{Zmax} = (14 - 5) mA$

$= 9 mA.$

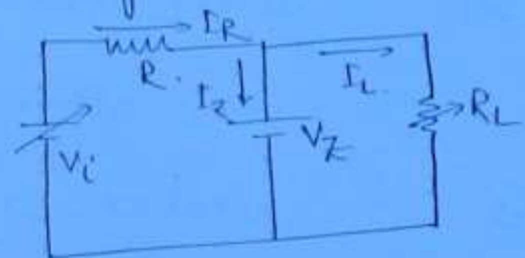
<u>I_{Zmin},</u> $I_{R min} = \frac{80 - 50}{5K\Omega} = \frac{30V}{5K\Omega} = 6 mA$ $I_{Zmin} = (6 - 5) mA$

$= 1 mA.$

IV V_i and R_L variable →



Cal. the dynamic range of series resistance R.



$$R_{min} = \frac{V_{i\,max} - V_Z}{(I_R)_{max}}$$

$$R_{max} = \frac{V_{i\,min} - V_Z}{(I_R)_{min}}.$$

$$I_R = I_Z + I_L.$$

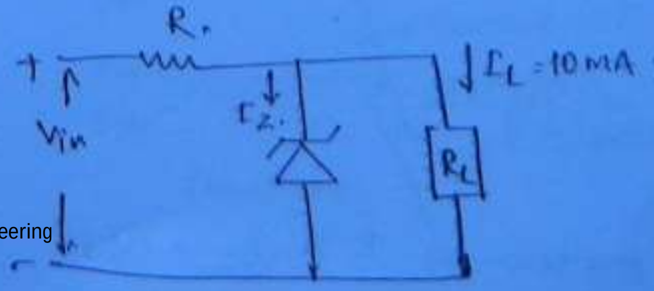
$$\frac{V_i - V_Z}{R} = I_Z + I_L.$$

NOTE →

For variable problems in zener diode, we can have an eqⁿ

$$\frac{V_i - V_Z}{R} \geq I_Z + I_L.$$

Q-15.
Ch-1.



Given

$V_{in} = 30V$ to $50V$.

$I_L = 10mA$

$V_0 = V_Z = 10V$.

$I_{ZK} = 1mA$.

$I_R \geq I_Z + I_L$.

$$\frac{V_i - V_Z}{R} \geq I_Z + I_L$$

$V_{in} = 30V$

$$\frac{30-10}{R} \geq 1mA + 10mA.$$

$$\boxed{R \leq \frac{20}{11mA} = 1818\Omega} \longrightarrow \text{Ans}$$

$V_{in} = 50V$.

$$\frac{50-10}{R} \geq 11mA.$$

$$R \leq \frac{40}{11mA} = 3636\Omega$$

16).

$$\frac{V_i - V_Z}{R} \geq I_Z + I_L \qquad \cancel{R_L = 10}$$

$I_L = 100mA$.

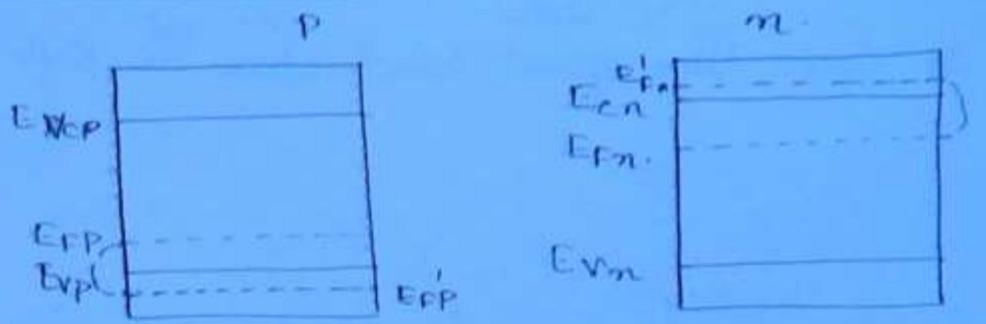
$$\frac{12-5}{R} \geq 0 + 100$$

$$R \leq \frac{7}{100mA} = 70\Omega$$

$I_L = 500mA$.

$$\frac{12-5}{R} \geq 0 + 500 \rightarrow R \leq \frac{7}{500} = 14\Omega.$$

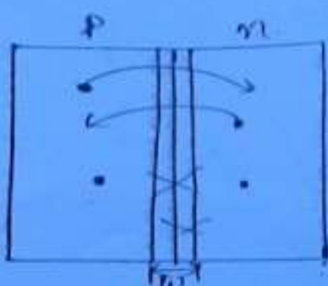
Tunnel diode →



$E_{FP} = E_V + KT \ln \frac{N_V}{N_A}$ $\qquad$ $E_{Fn} = E_C - KT \ln \frac{N_C}{N_D}$

$N_A >> N_V$ $\qquad$ $N_D >> N_C$

$E_{FP} = E_V - KT \ln N_A$ $\qquad$ $E_{Fn} = E_C + KT \ln N_D$



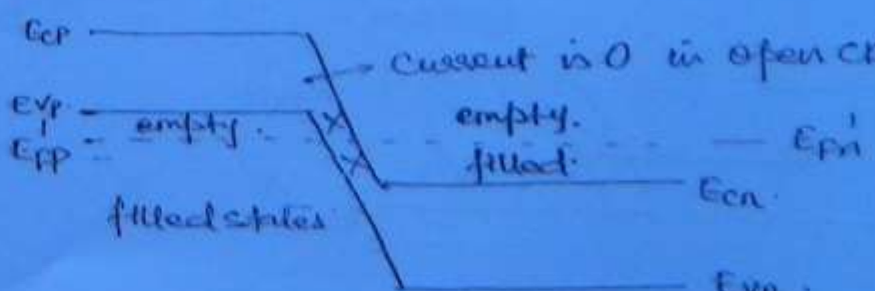
$W = 100\,\text{Å}$ $\qquad$ $1\,\text{Å} = 10^{-8}\,\text{cm}$

↓

$\frac{1}{15^{th}}$ wavelength of light.

Case 1. →

(1) open ckt p-n jn.



(ii) Reverse bias

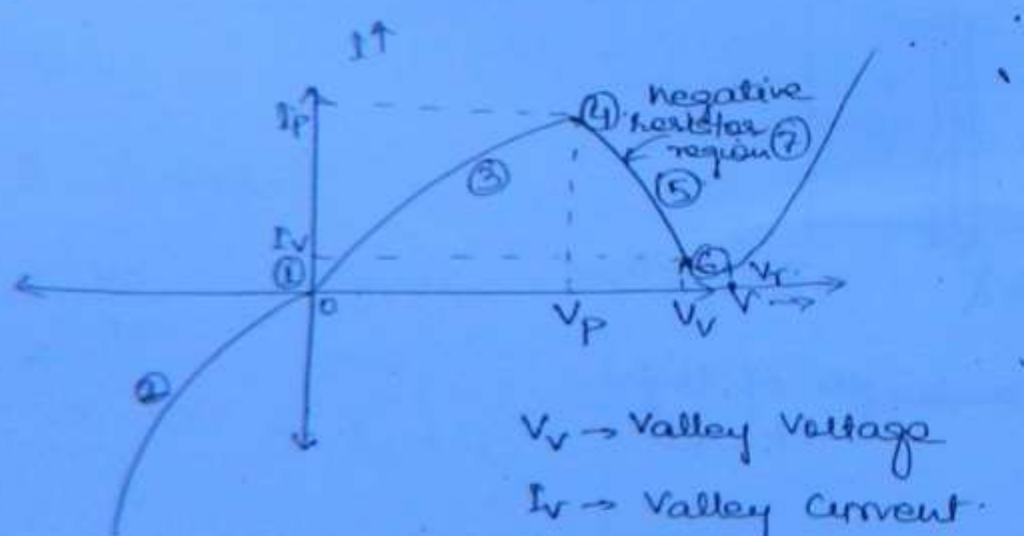
It acts like good conductor in R.B

Case 3.

Small FB (mV)

Zener → Si

tunnel → Ge

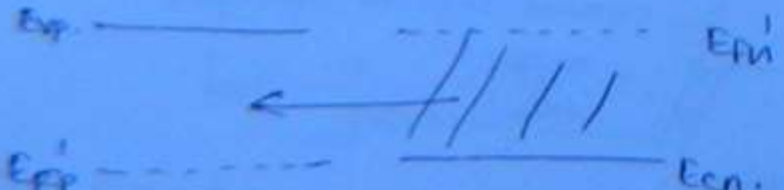


V_v → Valley Voltage

I_v → Valley current.

V_γ → Cut in Voltage $\approx 0.3V$.

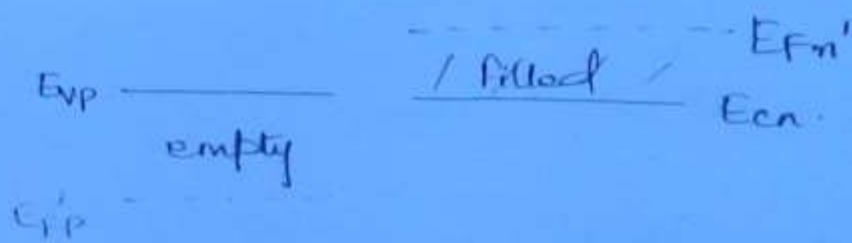
Case 4.



Case 5 →

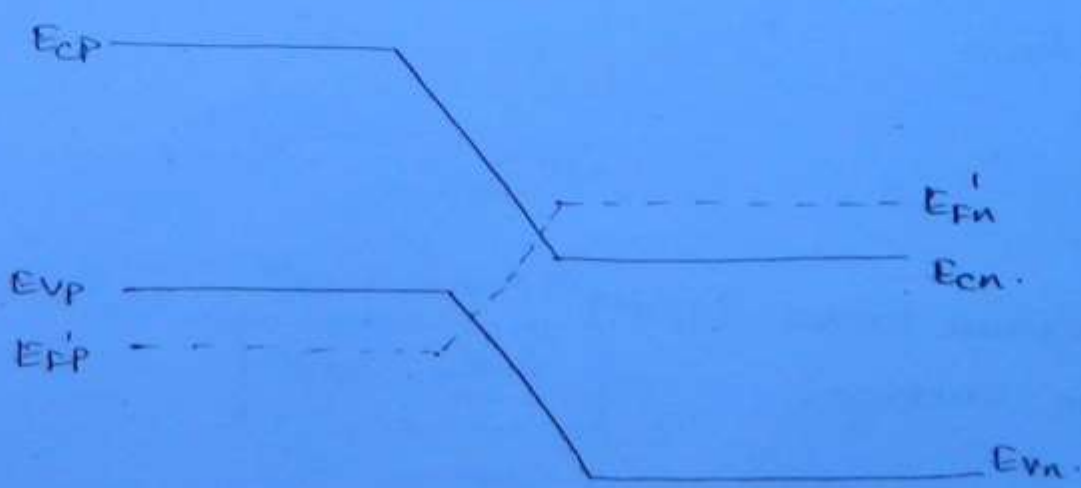
$V_f > V_p$

Case 6 →



Case 7 →

$V_f > V_v$



* $\frac{I_p}{I_v}$ ratio = 15 (GaAs)

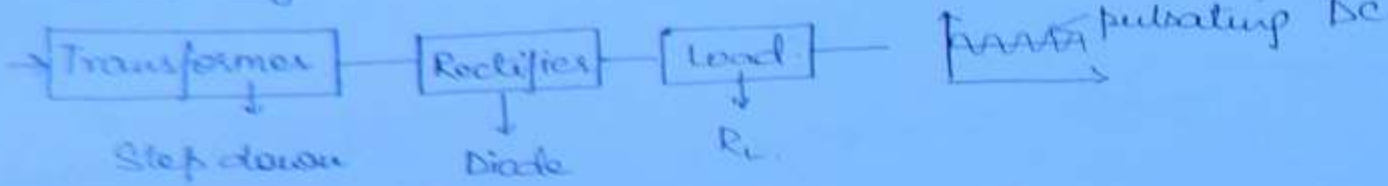
= 8 (Ge)

= 3 (Si)

4/1/17 **Rectifiers →**

AC

Block diagram



It is a ckt which converts AC to pulsating DC.

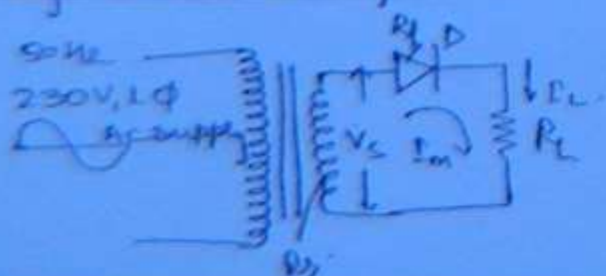
Practical rectifiers designs :-

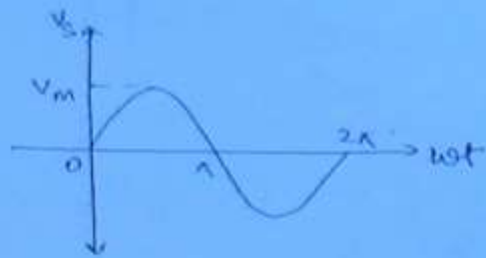
1) half wave rectifier →
2) Full wave rectifier →
3) Bridge rectifiers —

Rectifier parameters →

i) Av current I_{DC}
ii) DC output voltage V_{dc}.
iii) Rms load current I_{rms}.
iv) Ripple factor
v) voltage regulation
vi) Rectification η
vii) transformer utilisation factor (TUF)
viii) PIV (Peak inverse Voltage)
ix) F.F.
x) Peak factor

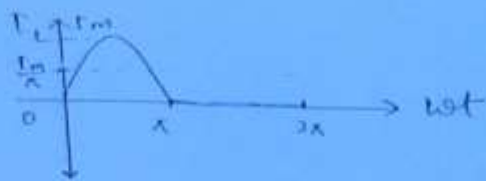
Half wave Rectifier →





$V_s = V_m \sin \omega t$

95



$I_L = I_m \sin \omega t \qquad 0 < \omega t < \pi$

$\quad = 0 \qquad \pi < \omega t < 2\pi$

Rectifier parameters :—

1) Av. current I_{DC} :—

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

$$\frac{1}{2\pi} \int_0^{2\pi} I \, d(\omega t)$$

$$= \frac{I_m}{2\pi} \int_0^{\pi} \sin \omega t \, d(\omega t)$$

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

2) DC O/P voltage V_{DC} :—

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

$$\boxed{V_{dc} = \frac{I_m}{\pi} R_L}$$

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

$$= \frac{V_m}{\pi} \frac{1}{\frac{R_f + R_s}{R_L} + 1}$$

When $R_L \to \infty$.

$$\boxed{(V_{dc})_{\text{No load}} = \frac{V_m}{\pi}}$$

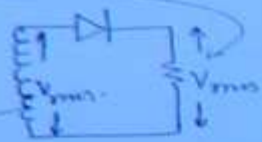
3) RMS load current (I_{Rms}) →

RMS ←

$$I_{rms} = \sqrt{\frac{1}{2\pi}\int I^2 d(\omega t)}$$

$$\frac{V_m}{\sqrt{2}} \quad \frac{I_m}{2}$$

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



$$= \frac{V_m}{\sqrt{2}}$$

4) Ripple factor →

$$= \frac{\text{RMS of alternating component}}{\text{Av. Value}} = \frac{I'_{rms}}{I_{dc}}$$

Assume →

I' = Ac current

I_{dc} = dc current

I — total current

I'_{rms} = rms value of ac current

$$I'_{rms} = \left[\frac{1}{2\pi}\int_0^{2\pi} I'^2 d\alpha\right]^{1/2}$$

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

$$I'_{rms} = \left[\frac{1}{2\pi}\int_0^{2\pi} (I - I_{dc})^2 d\alpha\right]^{1/2}$$

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

$$= I_{rms}^2 - 2 I_{dc}^2 + I_{dc}^2$$

$$I_{rms}^2 - I_{dc}^2$$

$$r.f. = \frac{I_{rms}}{I_{dc}} = \frac{\sqrt{I_{rms}^2 - I_{dc}^2}}{I_{dc}}$$

$$\boxed{R.f. = \sqrt{\frac{I_{rms}^2}{I_{dc}^2} - 1}}$$

HWR

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

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

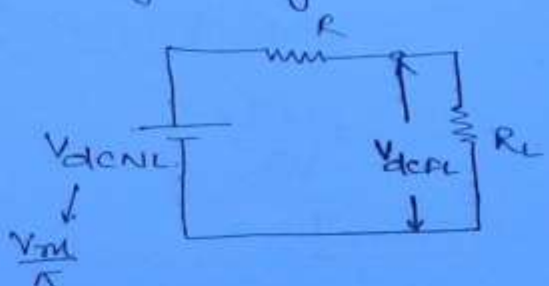
$$= 1.21$$

$$= 121\%$$

Conclusion →

Ripple factor value is 1.21 for half wave rectifier i.e 121% of ac comp. is present in dc value.

5> Voltage Regulation →



$$\%VR = \frac{V_{dcNL} - V_{dcFL}}{V_{dcFL}} \times 100\%$$

HWR.

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

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

$$\underbrace{I_{dc}(R_d + R_s)}\ + \underbrace{I_{dc}R_L}_{V_{dcFL}} = \underbrace{\frac{V_m}{\pi}}_{V_{dcNL}}$$

$$\%VR = \frac{I_{dc}(R_d + R_s)}{\frac{V_m}{\pi} - I_{dc}(R_d + R_s)} \times 100\%$$

ideally the V.R. should be 0. Practically it should be mini in value.

98

6) Rectification η :-

$$\eta = \frac{\text{Dc power delivered to the load.}}{\text{ac I/P power}}$$

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

$$= \frac{I_{dc}^2}{I_{rms}^2} \frac{1}{\frac{R_f + R_s}{R_L} + 1}$$

$$\frac{R_f + R_s}{R_L} << 1.$$

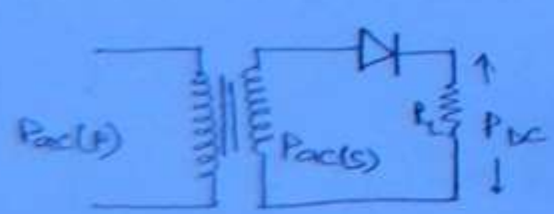
$$\eta \quad \frac{I_{dc}^2}{I_{rms}^2}$$

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

$$= \frac{4}{\pi^2} = 0.406.$$

$$= 40.6\%.$$

7) Transformer utilisation factor :→



$$TUF = \underbrace{\frac{(TUF)_p + (TUF)_s}{2}}_{\text{for full wave}}$$

for utilisation of 360°

$$(TUF)_p = \frac{P_{DC}}{(P_{ac})_p}.$$

$$(TUF)_s = \frac{P_{DC}}{(P_{ac})_s}.$$

$$TUF = (TUF)_S$$

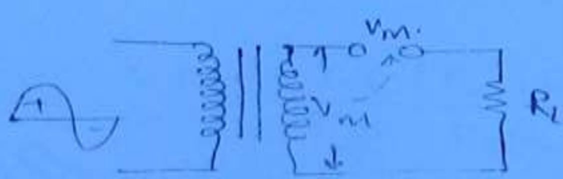
$$= \frac{\text{DC power delivered to load}}{\text{Pac rated of sec}}$$

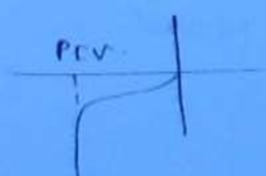
$$= \frac{I_{dc}^2 R_L}{V_{rms} \times I_{rms}}$$

$$V_{rms} \downarrow \frac{V_m}{\sqrt{2}} \qquad I_{rms} \downarrow \frac{I_m}{2}$$

$$\boxed{TUF = 28\%}$$

8) PIV (Peak Inverse Voltage) →





PIV = V_m [negative half cycle]

9) Form factor →

$$\frac{RMS}{AV}$$

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

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

$$= \pi/2$$

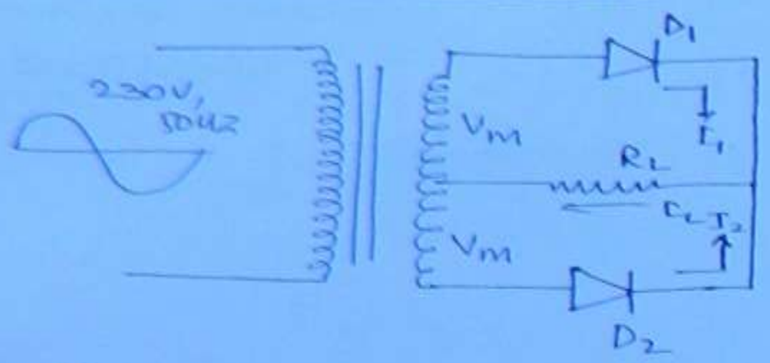
10) Peak factor →

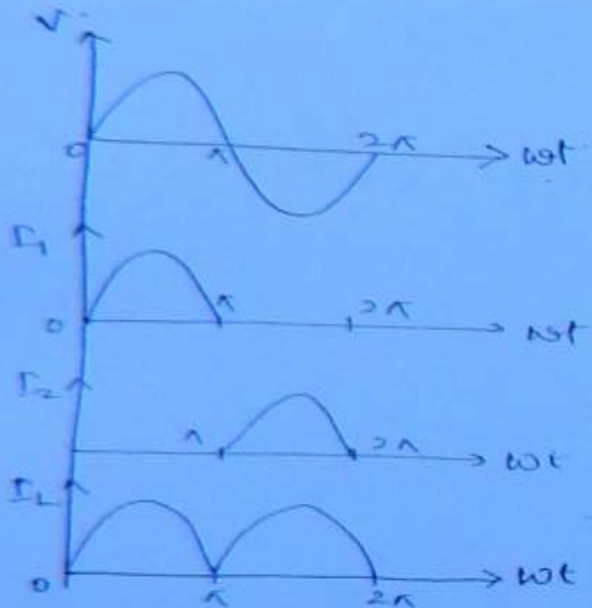
$$\frac{Peak}{RMS}$$

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

$$\boxed{= 2}$$

Full wave Rectifier →





Rectifier parameters →

1) Av. current I_{DC} →

$$I_{DC} = \frac{2I_m}{\pi}$$

2) DC Output voltage V_{DC} →

$$V_{DC} = \frac{2V_m}{\pi}$$

3) rms I_{rms} →

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

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

4) $R.f. = \sqrt{\dfrac{I_{rms}^2}{I_{dc}^2} - 1}$

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

$$= \sqrt{\dfrac{\pi^2}{4\times 2} - 1}$$

$$= 48\%$$

5) Voltage Regulation :—

$$\underbrace{I_{dc}(R_d + R_s) + I_{dc}R_L}_{} = V_{dc\,NL}$$

$$I_{dc}R_L \downarrow V_{dcFL} \qquad V_{dc\,NL} \downarrow \dfrac{2V_m}{\pi}$$

$$V.R. = \dfrac{I_{dc}(R_d + R_s)}{\dfrac{2V_m}{\pi} - I_{dc}(R_d + R_s)}$$

6) $\eta = \dfrac{P_{DC}}{P_{ac}}$

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

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

$$= \dfrac{8}{\pi^2}$$

$$= 81.2\%$$

7) TUF →

$$= \dfrac{(TUF)_p + (TUF)_s}{2}$$

$$(TUF)_s = \dfrac{P_{DC}}{P_{ac} \text{ rated sec one half} - V_{rms}I_{rms} + P_{ac} \text{ rated sec other half} - V_{rms}I_{rms}}$$

$V_{rms} = \dfrac{V_m}{\sqrt{2}}$, $I_{rms} = \dfrac{I_m}{2}$

$$= 57.3\%$$

$(TUF)_P = \eta = 81.2\%$

$$TUF = \frac{81.2 + 57.3}{2}$$

$$= 69.3\%$$

8) PIV →

$2V_M$

9) F.F. $= \frac{Rms}{Av}$

$$= \frac{I_{Rms}}{I_{dc}}$$

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

$$= \frac{\pi}{2\sqrt{2}}$$

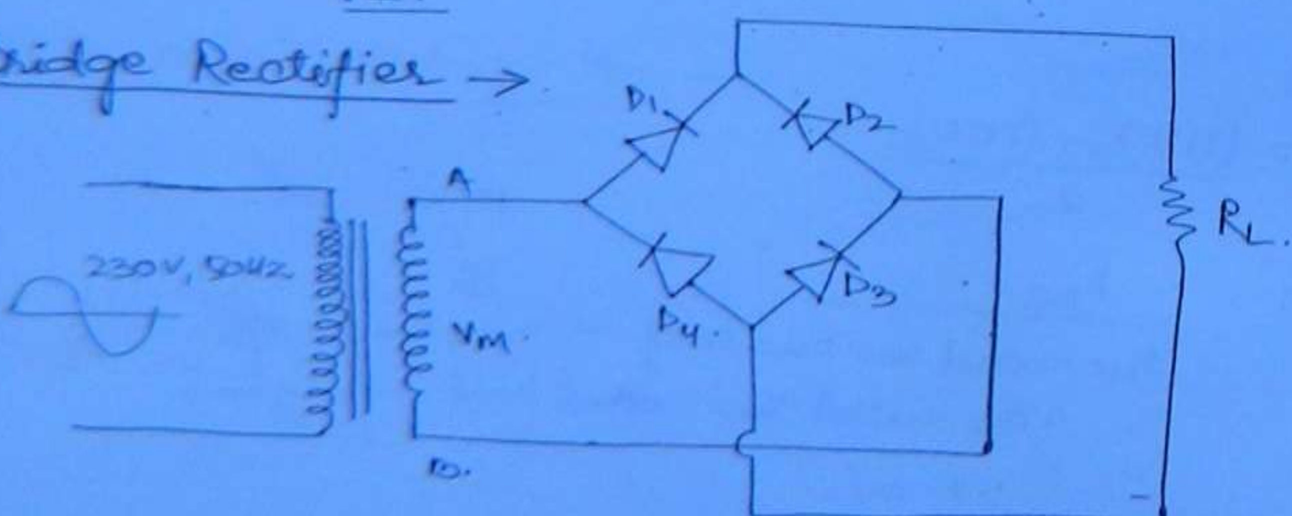
10) Peak factor =

$$\frac{Peak}{RMS}$$

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

$$= \sqrt{2}.$$

Bridge Rectifier →



+ive half cycle →

$A\,D_1\,R_L\,D_3\,B$

$D_1\,D_3 \rightarrow F.B$

$D_2\,D_4 \rightarrow R.B.$

-ive half cycle

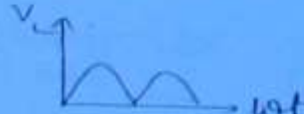
$B\,D_2\,R_L\,D_4\,A.$

$D_2\,D_4 \rightarrow F.B.$

$D_1\,D_3 \rightarrow R.B.$

Common parameters as full wave Rectifier :—

$$I_{DC} = \frac{2I_m}{\pi}$$



$$V_{DC} = \frac{2V_m}{\pi}$$

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

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

$R.f. = 48\%$.

~~V.R.~~

$\eta = 81.2\%$.

Other parameters →

Voltage Regulation =

$$\underbrace{I_{dc}(2R_d + R_s) + I_{dc}R_L}_{F.L.} = \underbrace{\frac{2V_m}{\pi}}_{N.L.}$$

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

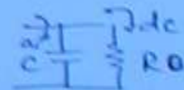
$$= 81.2\%.$$

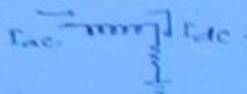
$PIV = V_m.$

Filters →

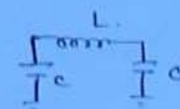
R.f. (HWR) → 1.21.
(FWR) → 0.48.
(BR) → 0.48.

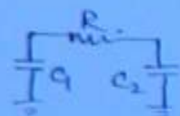
Filters →

→ Capacitive filters 

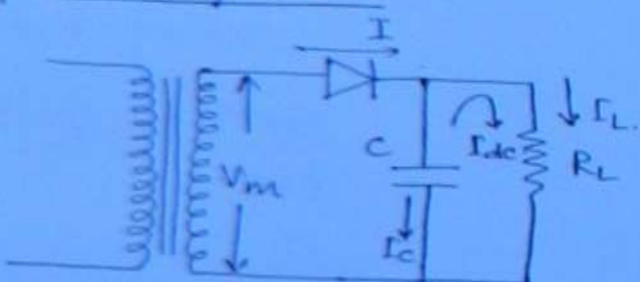
→ Inductor filter 

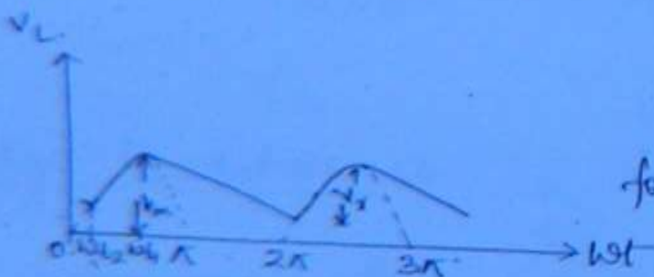
→ L section filter 

→ π Section filter [not for heavier loads] 

→ Rc filter 

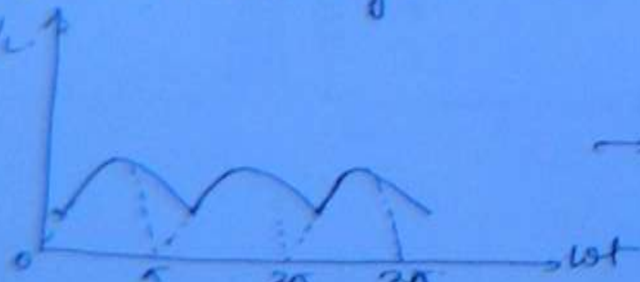
Capacitive filters →





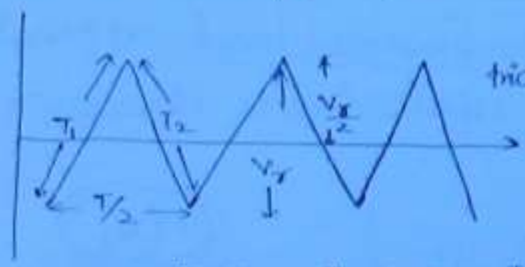
for half wave.

ωt_2 → cut in angle.
ωt_1 → cut out angle.



→ for full wave.

Expression for ripple factor for full wave →



triangular waveform (ripple)

$$V_{rms} = \frac{V_m}{2\sqrt{3}} = \frac{V_r/2}{\sqrt{3}}$$

$$T_1 + T_2 = \frac{T}{2} \Rightarrow T_2 = \frac{T}{2}$$

charge lost at time T_2 = charge lost by the capacitor

$$I_{dc} \times T_2 = C \times V_r$$

$$V_r = \frac{I_{dc} \times T}{2C}$$

$$= \frac{I_{dc}}{2fC}$$

$$= \frac{V_{dc}}{2fCR_L}$$

$$R.f. = \frac{V_{rms}}{V_{dc}}$$

$$= \frac{V_r}{2\sqrt{3}} \Big/ V_{dc}$$

$$= \frac{V_{dc}}{4\sqrt{3}fCR_L} \Big/ V_{dc}$$

$$\boxed{R.f. = \frac{1}{4\sqrt{3}fCR_L}}$$

→ for high load and capacitive value must be high for low R.f.

ωt_2 → Cut in angle →

It is the angle at which the diode starts conducting or capacitor gets charged.

ωt_1 → Cut out angle →

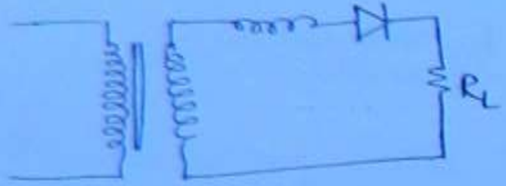
It is the angle at which the diode stops conducting or the capacitor gets discharged.

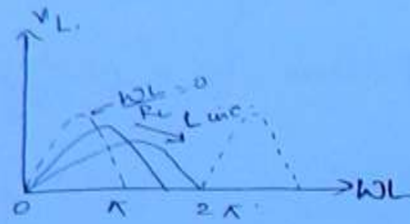
continue...

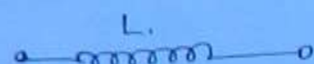
- Capacitive filters are used for heavier load application.
- As the capacitive value inc., ripple factor decreases.

Inductive filters :→

HWR.







It can not block 2nd harmonics.
It block all the harmonic components except 2nd.
because In 2nd, $X_L = 2\pi f' L$

$f' = 2 \times 50$
$= 100Hz$

not very high, unable to block.

$$I(t) = \underbrace{\frac{I_m}{\pi}}_{dc} - \underbrace{\frac{2I_m}{3\pi}\cos 2wt}_{2nd\ harmonic}$$

FWR.

$$I(t) = \frac{2I_m}{\pi} - \underbrace{\left(\frac{4I_m}{3\pi}\right)}_{I_m'}\cos 2wt$$

$$I_{dc} = \frac{2I_m}{\pi} = \frac{2V_m}{\pi R_L}.$$

$$(I_{rms})_{2nd} = \frac{I_m'}{\sqrt{2}} = \frac{4I_m}{3\pi\sqrt{2}}$$

$$= \frac{4V_m}{3\pi\sqrt{2}\,(R_L + j2wL)}$$

$$= \frac{4V_m}{3\pi\sqrt{2}\sqrt{R_L^2+4\omega^2L^2}}$$

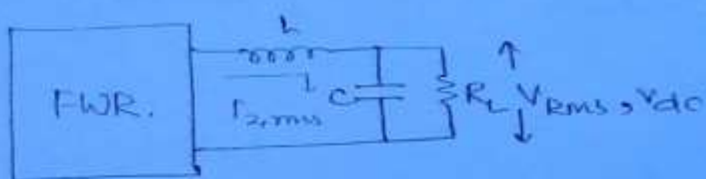
$$R.f. = \frac{I_{rms}}{I_{dc}}$$

$$= \frac{\sqrt{2}}{3}\frac{R_L}{2\omega L}$$ → for lighter loads ($R_L\downarrow$) high L.

Conclusion →

1) Inductor filter designs are used for lighter load application [R_L]
2) If the inductor value is more, ripple factor will be less.

L-section filter → [for varying load application]



$$(I_{rms})_{2nd} = \frac{4V_m}{3\pi\sqrt{2}}\frac{1}{X_L} \quad (X_L = 2\omega L).$$

$$= \frac{2}{3\sqrt{2}}\underbrace{\left(\frac{2V_m}{\pi}\right)}_{V_{dc}}\frac{1}{X_L}$$

$$= \frac{\sqrt{2}}{3}\frac{V_{dc}}{X_L}$$

$$V_{rms} = \frac{\sqrt{2}}{3}\frac{V_{dc}}{X_L}X_C \qquad X_C = \frac{1}{j2\omega C}$$

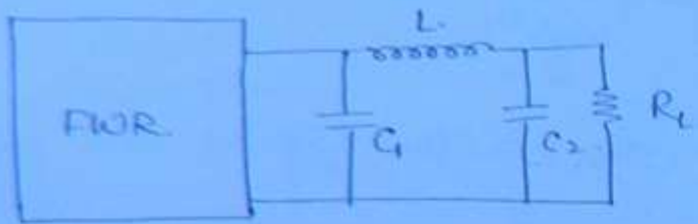
$$r.f. = \frac{V_{rms}}{V_{dc}} = \frac{\frac{\sqrt{2}}{3}\frac{V_{dc}}{X_L}X_C}{V_{dc}}$$

$$\boxed{R.f. = \frac{\sqrt{2}}{3}\frac{X_C}{X_L}}$$ ← free from load.

$$\boxed{X_L >> X_C}$$

L-section filters are used for varying load applications

π Section filter →



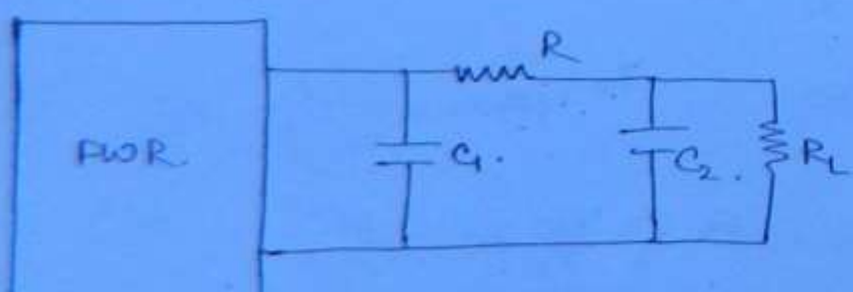
(108)

$$r.f. = \frac{\sqrt{2}\, X_{C_1} X_{C_2}}{X_L \cdot R_L}$$

$\frac{\sqrt{2}}{3} \frac{X_C}{X_L}$

Parameter	L section	π section
V_{DC}		better
r.f	better	
%VR	better	
PIV		

RC filter :→



$R \leq 10 X_{C_2}$

$$r.f. = \frac{\sqrt{2}\, X_{C_1} X_{C_2}}{R R_L}$$

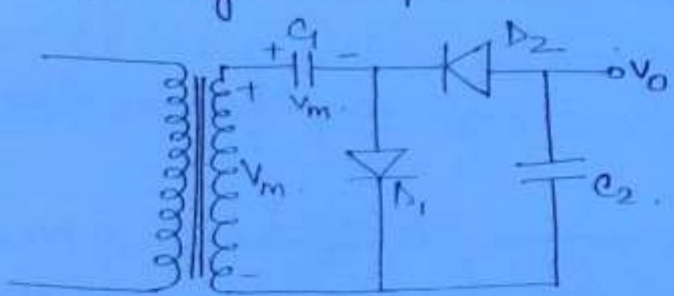
The main drawback of π section filter is it can not be used for heavier load application because of ind. property.

In practical applications, ind. is becoming more costlier. To reduce that cost effectiveness, we replace inductor by resistor.

~~Voltage~~

5/11/12

Voltage multiplier :→



→ Half wave Voltage doubler

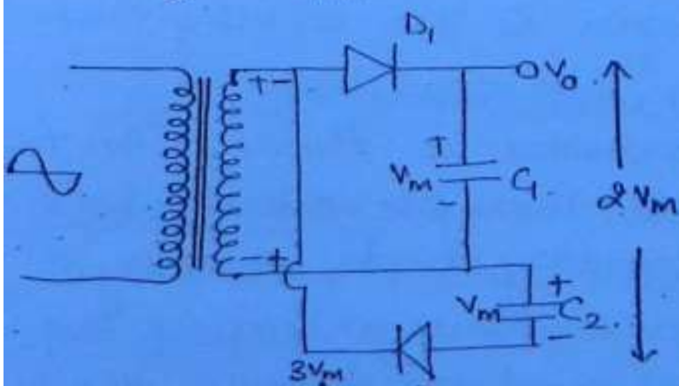
+ive half cycle,

D_1 ON D_2 OFF

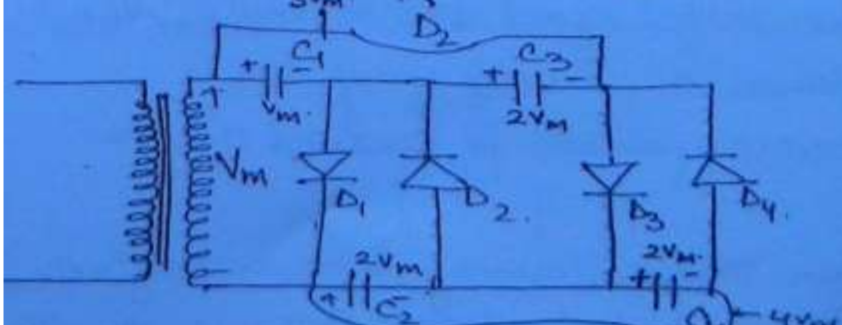
-ive half cycle,

D_1 OFF D_2 ON.

$V_0 = -2V_m$.



→ full wave voltage doubler

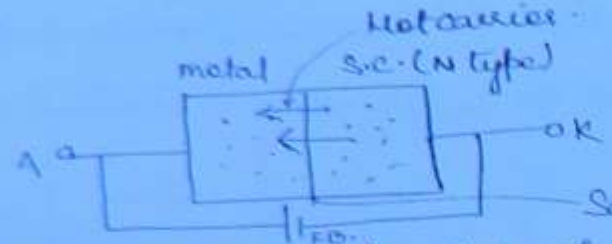


Special diode —

Schottky diode →

$f > 10MHz$

(110)



holes are not present. depletion region is not present.

Conclusions →

→ Reverse recovery time is a problem in ordinary p-n jn diode which cannot be operated for high freq greater than 10 MHz.

- To reduce the effect of reverse recovery time above 10 MHz range special diode called as schottky diode was imple-mented.

- Schottky diode construction details are diff. from ordinary diode.

- Schottky diode is a unipolar device where as ordinary diode is a bipolar device.

- There will be no depletion region in the schottky diode because of missing of immobile ions and holes.

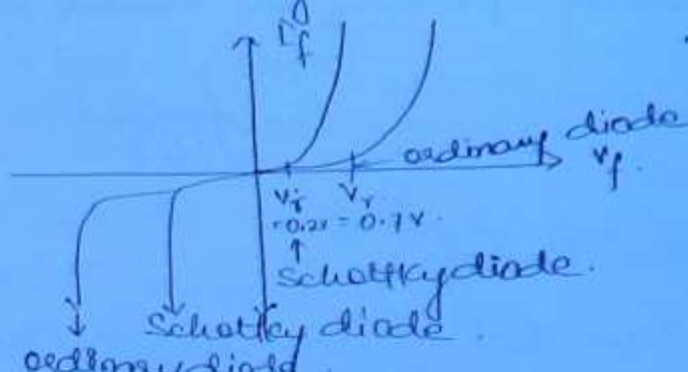
→ As the e⁻ en. levels in semiconductor is always less than ~~en~~ e⁻ en. levels in metal side. Therefore current is not possible in open ckt for schottky diode [Schottky barrier]

⇒ As the contact area b/w metal and s.c. is more, the forward and reverse resistance ~~or~~ is very very less.

→ A cut in voltage of schottky diode is around 0.2 to 0.25 V.

→ In reverse bias condition, reverse recovery time is neglected therefore there will be rapid leakage current and loss

Breakdown voltage-



NOTE →
When Schottky diode is operated in FWD bias the e⁻ from N type will penetrate towards the metal side such carri are called as hot carriers. That is the reason we called it as hot carrier diode.

Simple representation →

Application →

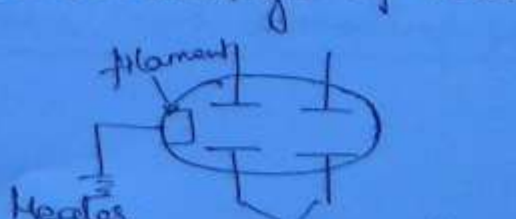
1) mainly used to rectify the signals above 300 MHz.
2) In digital comp., we already use this diode.
3) Digital ckts ~~(Special TTL diodes)~~
4) TTL high speed logic family)

BJT →

(Bipolar j^n transistor)

Introduction →

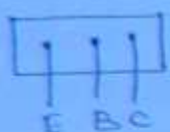
Before 1957, vacuum tubes are the devices which solve the ~~per~~ purpose of amplification.
The main disadvantage of vacuum tubes are



It is having an internal filament or heater which requires more power.
b) It takes more space.

c) Its life span is less.
d) Its power dissipation is high

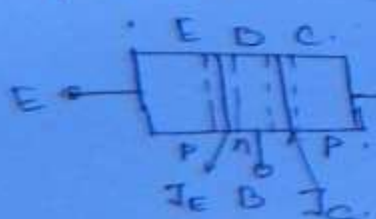
After 1957 BJT was invented which can also the serve the purpose of amplification.
The main advantage of BJT compare to vacuum tubes are



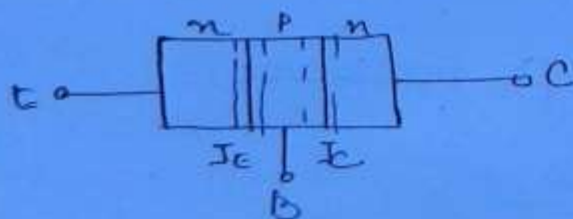
1) It is not having any internal filament or heater which requires more power
2) It takes very less space
3) Its life span is more.
4) Its power dissipation is very less in the order of ~~MW~~. mW (milli watt)

Types of BJT :→

PnP



npn.



For proper working of BJT :→

conditions →

1) Doping level

$E > C > B$.

2) $W_B < L_p$ or L_n.

i) width of the base should be narrower.

~/ Transfer + resistor = transistor.

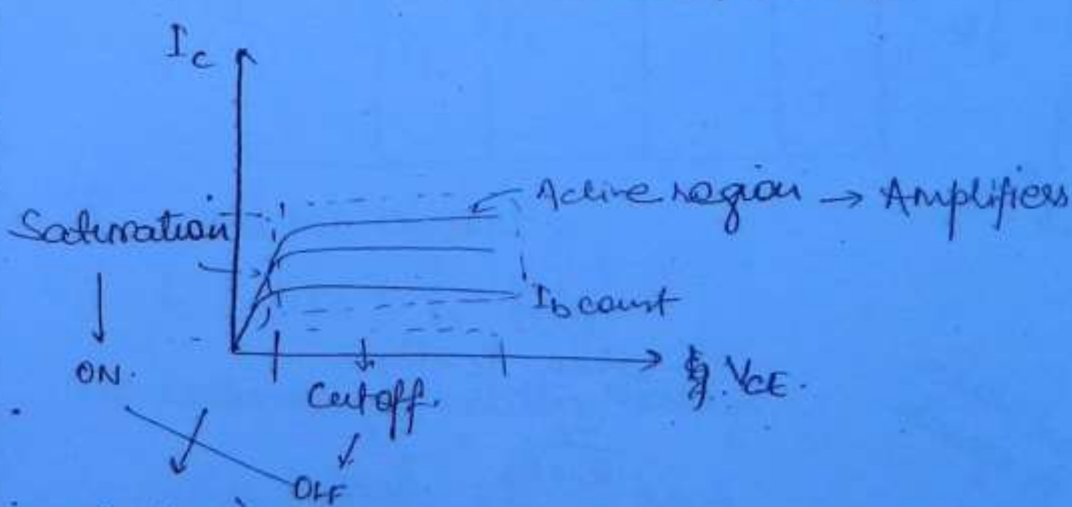
$Z_i \rightarrow$ low high
$Z_o \rightarrow$ high low

~~Transf~~ Transfer of I from one place to another imped. is called transistor.

Types of operating region →

Region	I_E	I_C
Active Region	FB	RB
Saturation region	FB	FB.
Cutoff region	R.B.	RB.
Reverse Active region	RB.	FB.

→ BJT is current controlled device

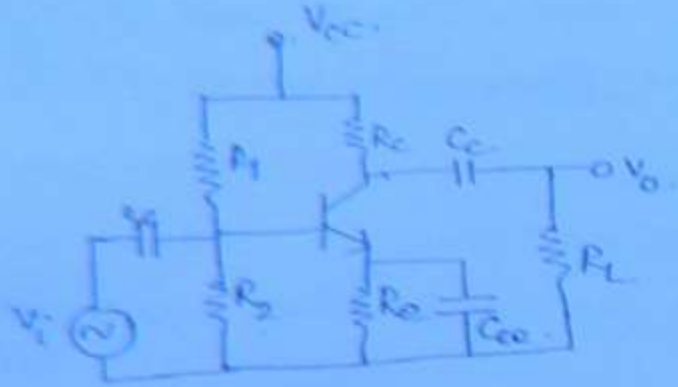


Switching.

DC load line and operating pt:

DC load line and operating pt :—

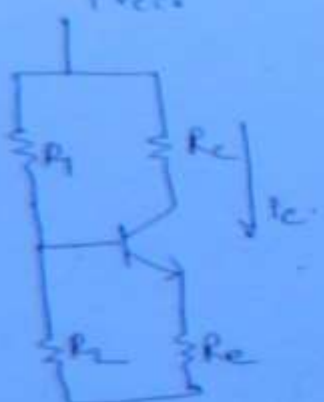
CE amplifier →

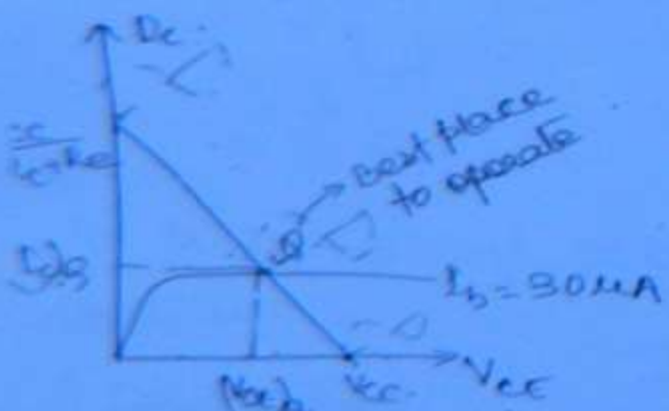


DC analysis →

i) ac should be grounded.

ii) $X_C \propto \frac{1}{f \to 0} = \infty$





When $I_c = 0$,

$$\frac{V_{CC}}{R_c + R_e} = \frac{V_{CE}}{R_c + R_e}$$

$$\Rightarrow V_{CE} = V_{CC}.$$

O/P loop → $V_{CC} = I_C (R_c + R_e) + V_{CE}$.

$$I_C = \frac{V_{CC}}{R_c + R_e} - \frac{V_{CE}}{R_c + R_e}$$

Conclusion :—

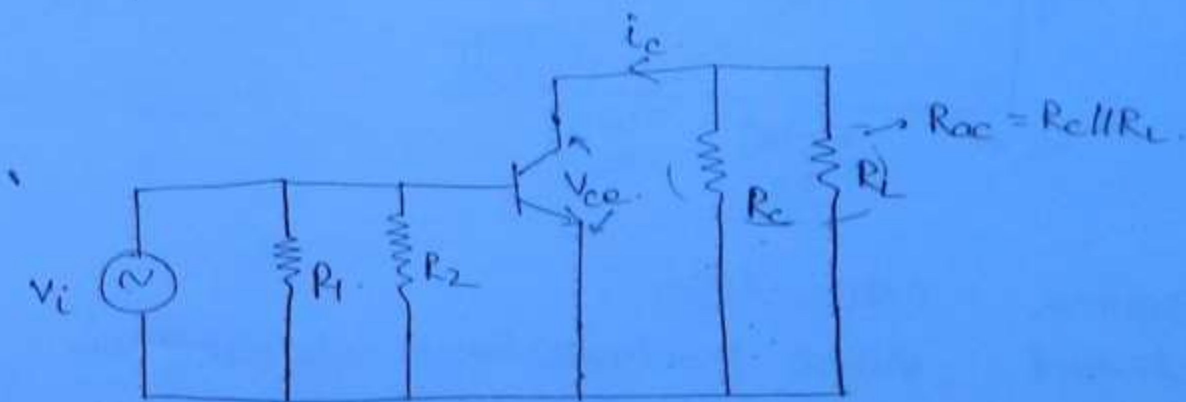
- For any analysis of BJT, we always prefer O/P char of BJT [It is a current controlled device] (115)
- Dc load line → The locus of all the Q pts are conc. on a line is called as Dc load line.
- Operating pt :—
  (Quiescent pt)
  ↓
  Inactive

  It is the intersection pt b/w sample graph of I_C w.r.t dc load line

Ac Analysis →

Dc should be grounded.

$X_C \propto \frac{1}{f \to \infty} = 0$



Ac collector current =

$$i_c = \Delta I_C = i_C - (I_C)_Q.$$

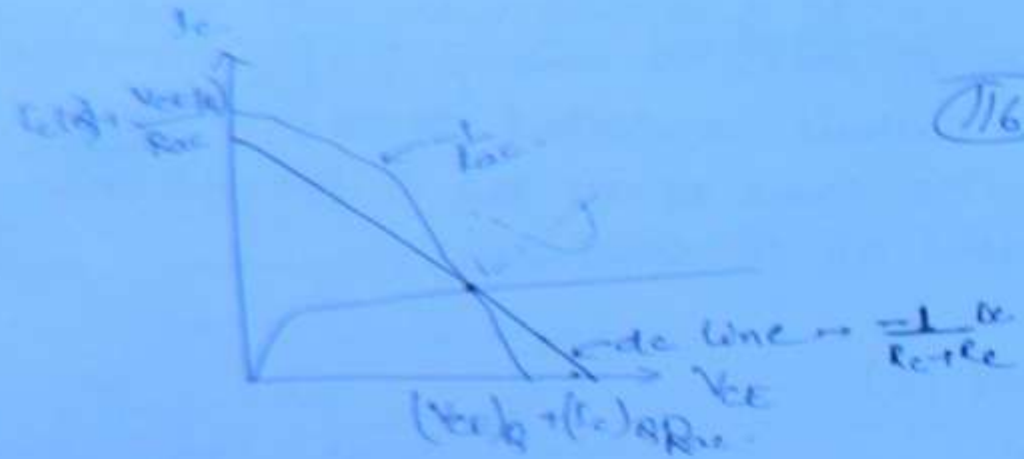
ac collector to emitter voltage

$$v_{ce} = \Delta V_{CE} = V_{CE} - (V_{CE})_Q.$$

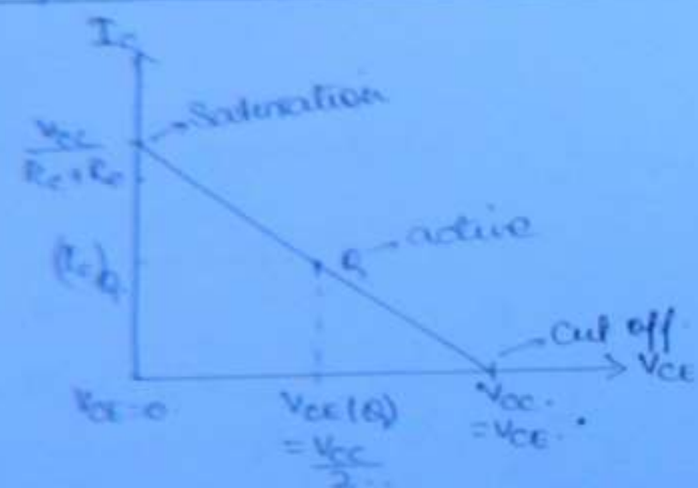
Apply KVL to O/P loop

$$v_{ce} + i_c(R_{ac}) = 0.$$

$$V_{CE} - (V_{CE})_Q + (i_C - (I_C)_Q)\, R_{ac} = 0.$$

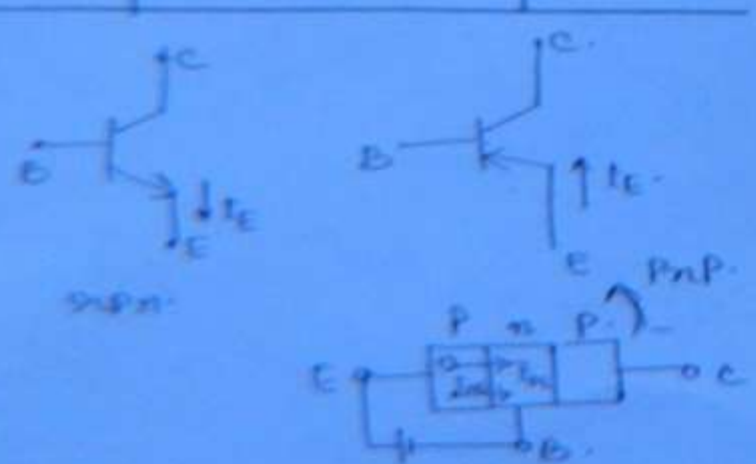


Condition for V_{CE} in saturation, active and cut off :—

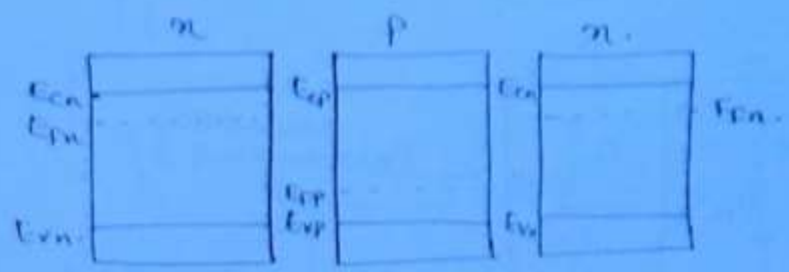


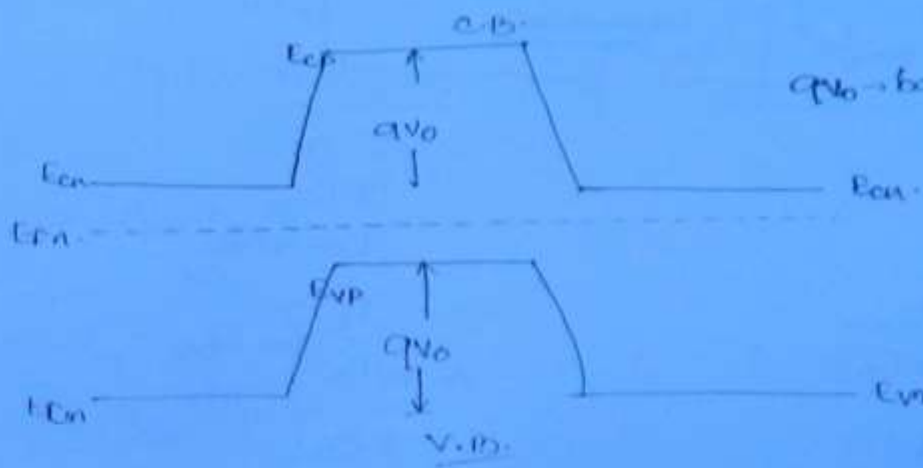
Active Source	Active device
Power delivered	which produces some internal gain.
eg → Volt. Source or current Source	eg → BJT, FET, MOSFET, OPAMP.

Simple representation of BJT :→



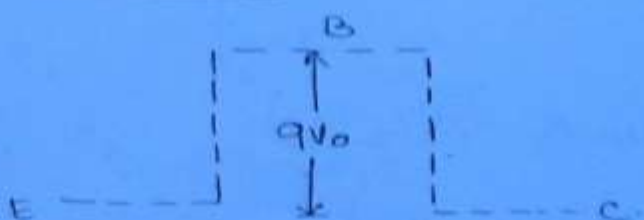
En. band diagram concept in a U.B. BJT :-





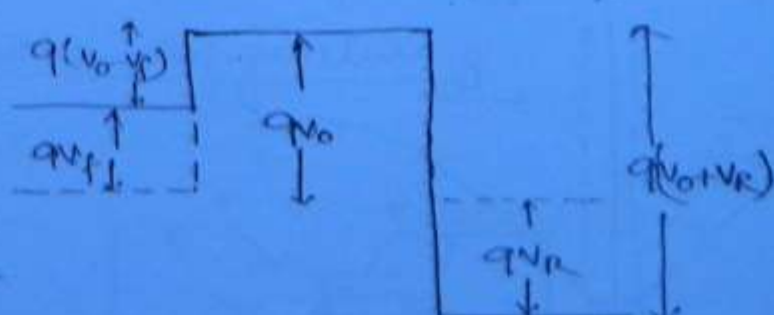
qV_0 - barrier height

Conclusion →

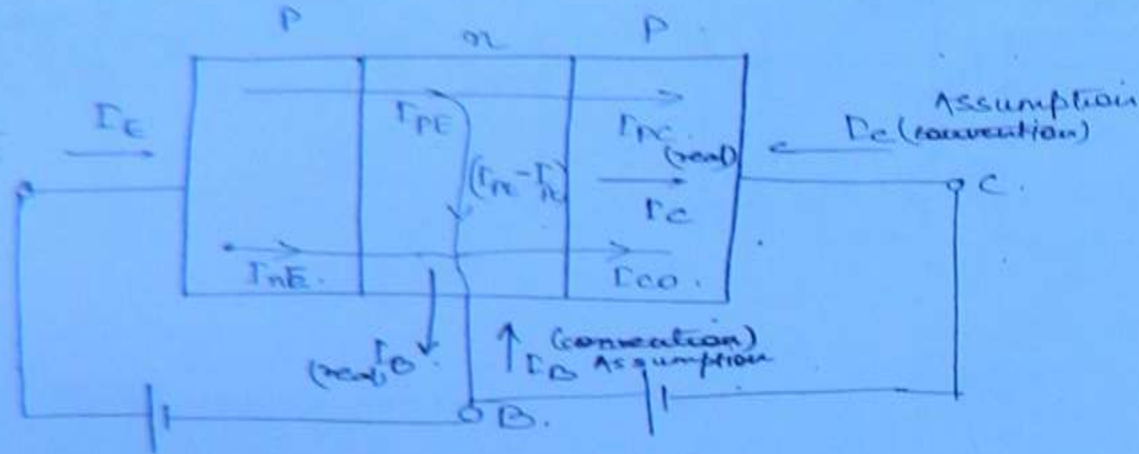


Note → Any no. of materials are joined, fermi levels will be in aligned position.

Biased condition →



Transistor current components →



1) emitter η :—

$$\gamma = \frac{I_{PE}}{I_{PE} + I_{nE}}$$

2) Transport factor :—

$$\beta^* = \frac{I_{PC}}{I_{PE}}$$

PnP

I_E +ive

I_B, I_C → –ive

nPn

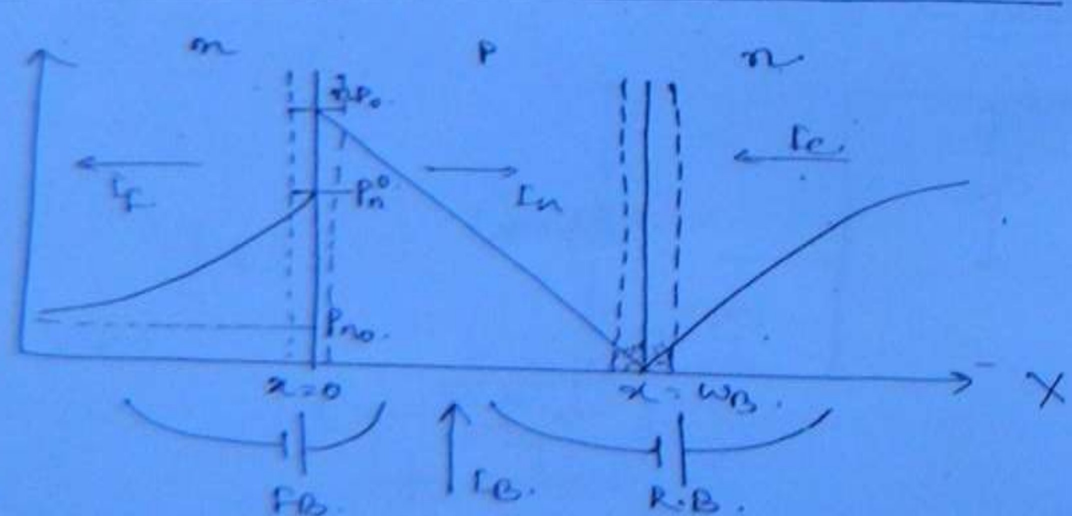
I_E –ive

I_B, I_C → +ive

3) large ~~ac~~ signal current gain $\alpha = \frac{-(I_C - I_{CO})}{I_E - 0}$

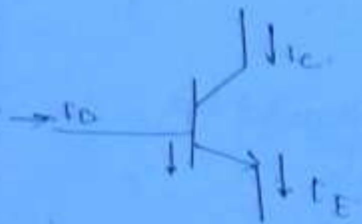
$$\boxed{\alpha = \beta^* \gamma}$$

Carrier Concentration V/s distance in BJT :—



npn →

(119)



$$I_n = AqD_B \frac{dn(x)}{dx}$$

$$= AqD_B \frac{n_p(0) - 0}{0 - W_B}$$

$$= -\frac{AqD_B n_{p_0}}{W_B}$$

$$I_C = -I_n \cdot$$

$$= \underbrace{\left\{\frac{AqD_B n_{p_0}}{W_B}\right\}}_{I_S} e^{\frac{V_{BE}}{BEV_T}}$$

$$n_{p(0)} = n_{p_0} e^{\frac{V_{BE}}{V_T}}$$

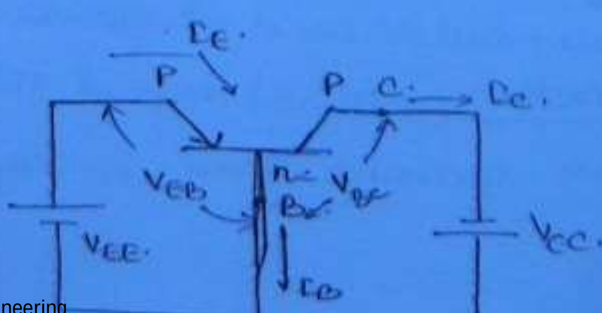
↳ Law of jn.

$$\boxed{I_C = I_S e^{V_{BE}/V_T}}$$

Where $I_S = \dfrac{AqD_B n_{p_0}}{W_B}$

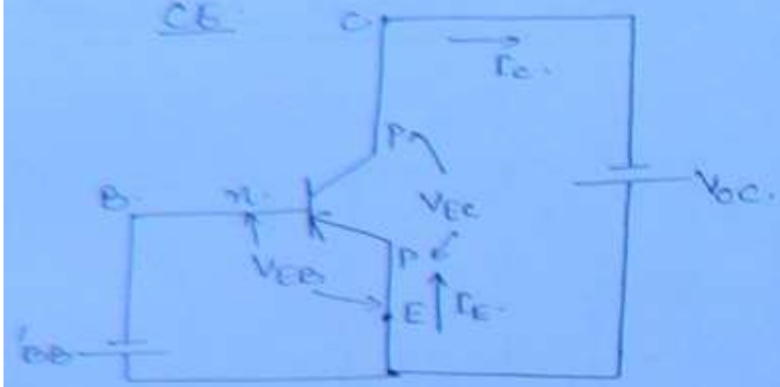
Transistor Configuration →

CB.



Current gain = $\dfrac{I_C}{I_E} = \alpha$

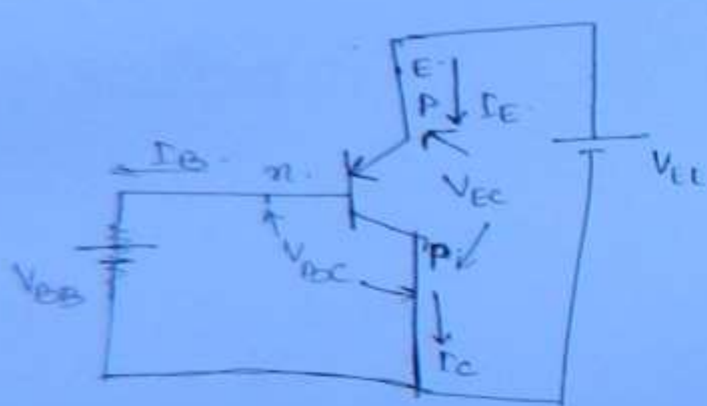
CE



collector should be always R.B.

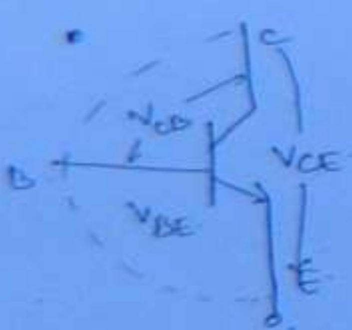
current gain = $\frac{I_C}{I_b} = \beta$.

CC

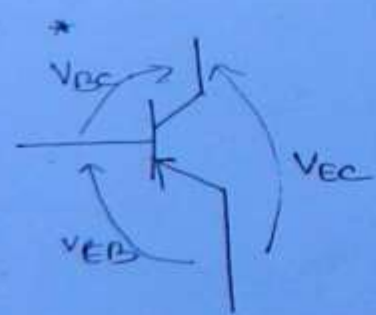


$\frac{I_E}{I_b} = \gamma$

npn



PnP



3/1/12.

α, B, γ Relations →

current amplification factor:—

1) In CB configuration the current gain is represented by α.

$\alpha = \frac{I_C}{I_E}$ Practically it lies b/w 0.9 to 0.99

2) In CE configuration, the current gain is represented by β

$$\beta = \frac{I_C}{I_b} \Rightarrow 20 \text{ to } 500.$$

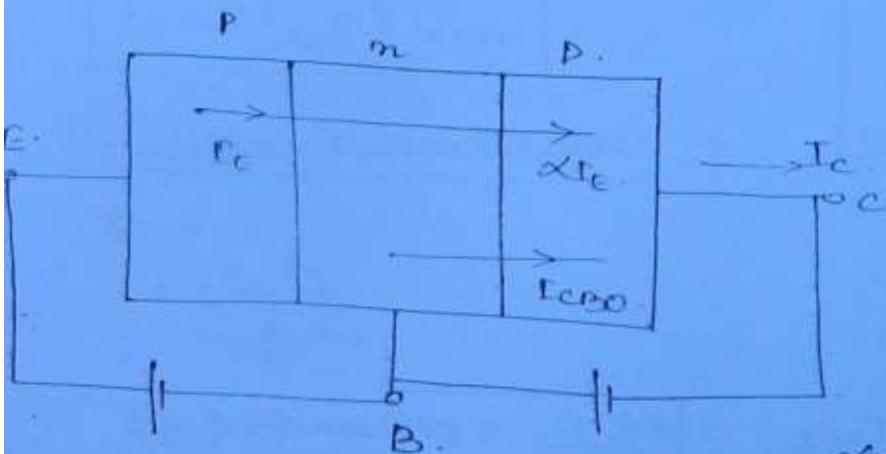
121

3) In CC configuration, the current gain is represented by γ.

$$\gamma = \frac{I_E}{I_b} \Rightarrow 20 \text{ to } 500$$

Total O/P current in CB:—

* magnitude



$$\alpha = \frac{I_C}{I_E}.$$

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

$$I_E = I_B + I_C.$$

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

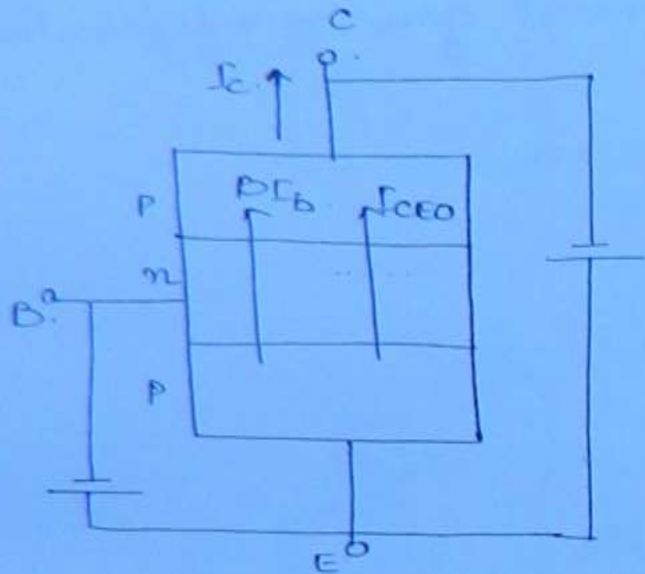
$$\Rightarrow I_C (1-\alpha) = \alpha I_B + I_{CBO}.$$

$$\boxed{I_C = \frac{\alpha}{1-\alpha} I_B + \frac{I_{CBO}}{1-\alpha}.}$$

I_{CBO} → collector to base value reverse saturation current when I/P is open (emitter)

Total O/P current in CE →

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



$$I_c = \beta I_b + I_{CEO}$$

$$I_{CEO} > I_{CBO}$$

$$\beta = \frac{I_c}{I_b} = \frac{I_c}{I_E - I_c} = \frac{I_c/I_E}{1 - I_c/I_E} = \boxed{\frac{\alpha}{1-\alpha} = \beta}$$

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

$$I_c = \frac{\alpha}{1-\alpha} I_b + \frac{1}{1-\alpha} I_{CBO}$$

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

$$\Rightarrow I_c = \beta I_b + (1+\beta) I_{CBO}$$

$$\boxed{I_{CEO} = (1+\beta) I_{CBO}}$$

Total O/P current in CC →

$$I_E = I_B + I_C$$

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

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

$$I_E = \frac{I_B}{1-\alpha} + \frac{I_{CBO}}{1-\alpha}.$$

$$\boxed{I_E = (1+\beta) I_B + (1+\beta) I_{CBO}}$$

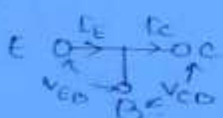
$$\gamma = \frac{I_E}{I_b} = \frac{I_E}{I_E - I_C} = \frac{1}{1 - I_C/I_E} = \frac{1}{1-\alpha} = 1+\beta.$$

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

$$\boxed{I_E = \gamma I_b + \gamma I_{CBO}}$$

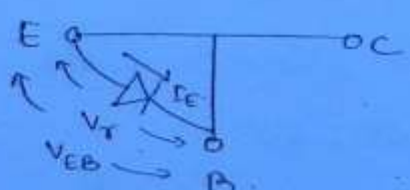
Transistor characteristics →

CB →



I/P parameters = I_E, V_{EB}

O/P parameters = V_{CB}, I_C.



$V_{EB} \to 0.5$ to 0.7 V.

$I_E \to 0$ mA to 140 mA.

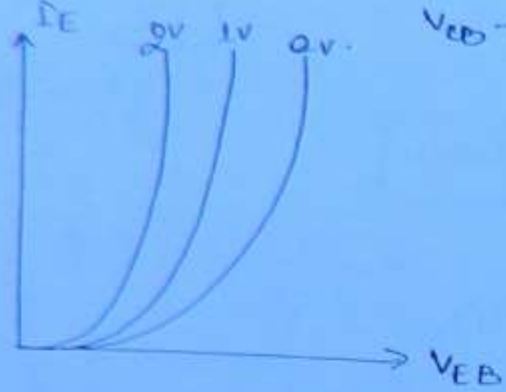
conclusion →

1) This is To get the I/P char., O/P voltage should be const [O/P current I_C never change the I/P current I_E]

2) To get the O/P char., I/P current should be constant. [I/P voltage will have less dynamic range].

3) BJT is current controlled device. That means the O/P char are controlled by I/P current but not voltage.

I/P char. in CB:→

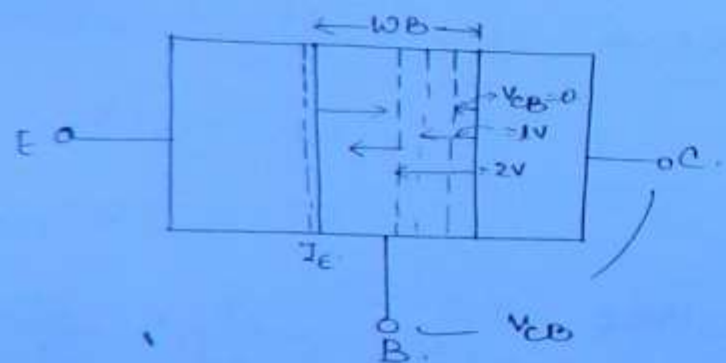


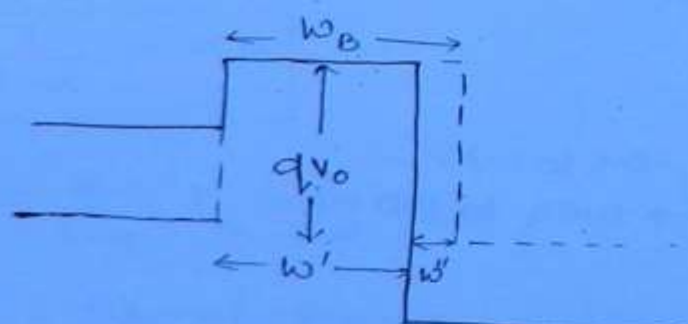
V_{CB} → const.

$V_{CB}\uparrow$, $W_B\downarrow$ $I_b\downarrow$ I/P FB↑,
$I_E\uparrow$.

~~Early~~

Early effect 'or' Base width modulation :—



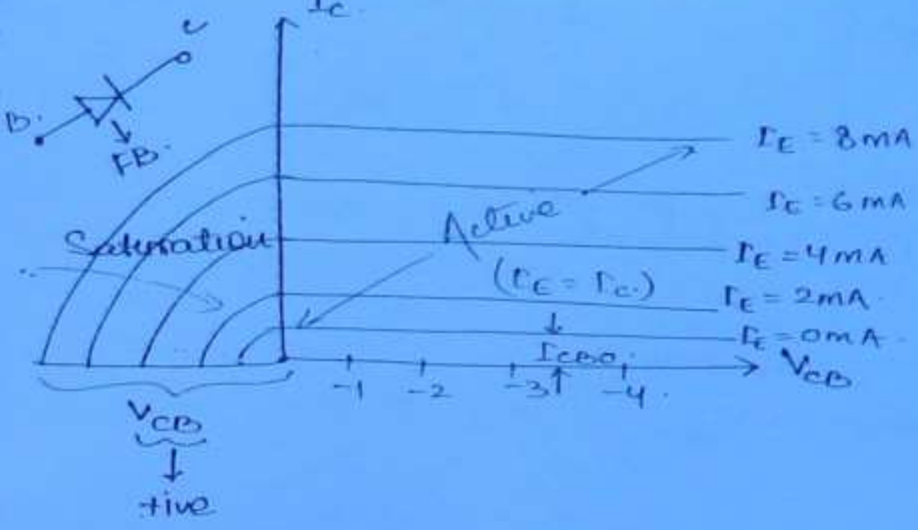


$W_B = W'\downarrow + W''\uparrow$

Physical width of base; effective base width; ← penetration width.

$W_B = W''$, $W' = 0$ } → punch through 'or' reach through.

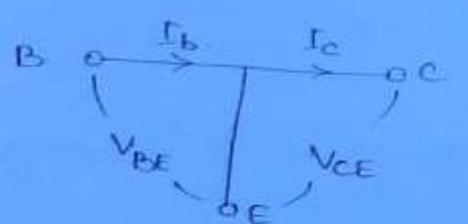
C.B. O/P characteristics :—



$I_E = I_B + I_C$
mA μA mA

$I_E \simeq I_C$

C.E. I/P characteristics :—

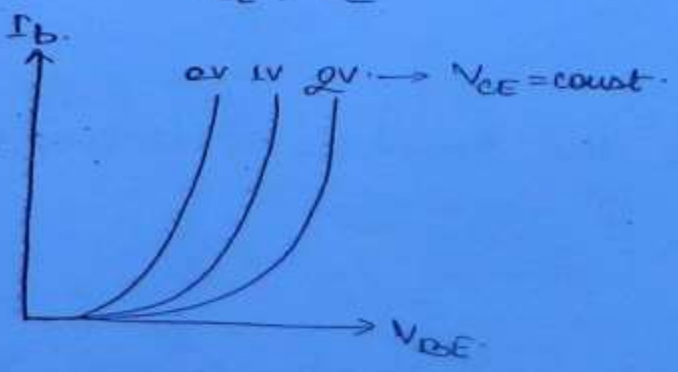


I/P parameters →

V_{BE}, I_b.

O/P parameters →

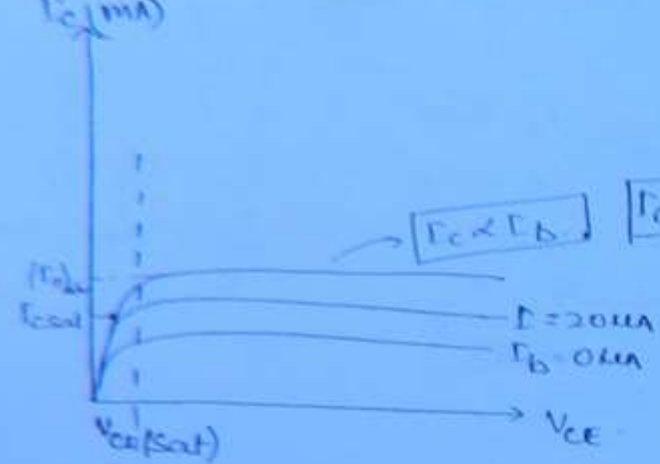
V_{CE}, I_C



$V_{CE}\uparrow, W_B\downarrow, I_b\downarrow$

early effect is possible in CE because collector terminal is R.B.

CE O/P Characteristics



Q. In an npn Si transistor if $V_{CE} = 0.3V$ then the transistor is acting in <u>Active</u> region

Q. If $I_C = \beta I_b$ then the transistor is acting in <u>Active</u> region.

Q. If $I_b > \frac{I_C}{\beta}$, then the transistor is acting in <u>Saturation</u> region.

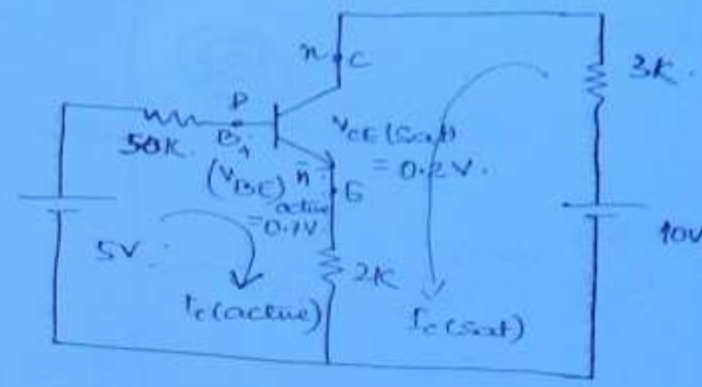
Q. If $(I_C)_{active} > I_C(sat)$, then the transistor is acting in <u>Saturation</u> region.

Q. If $(I_C)_{active} < I_C(sat)$ then the transistor is acting in <u>Active</u> region.

npn	$V_{CE}(sat)$	$V_{BE}(sat)$	$V_{BE}(active)$	V_{BE} (cut-in)	V_{BE} (cut off)
Si	0.2 V	0.8V	0.7V.	0.5V	0.0V
Ge	0.1 V	0.3V	0.2 V	0.1V	−0.1V.

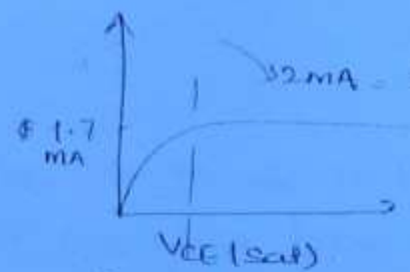
Pg - 22

Ans 5

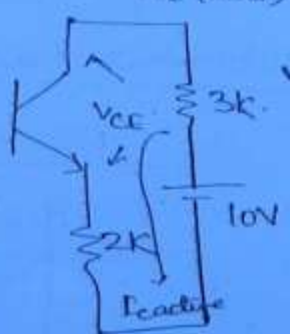


$$I_c(\text{active}) = \frac{5-0.7}{2K} = 1.7\ mA$$

$$I_c(\text{sat}) = \frac{10-0.2}{2K+3K} = \frac{10}{5K} = 2\ mA$$



So active region.



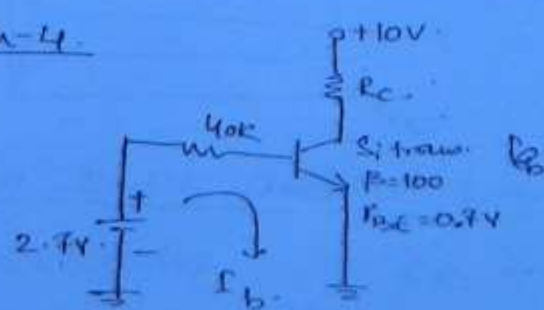
$V_{CE} = 10 - 5K \times I_c(\text{active})$

$\simeq 1.5V$

	I/P	O/P
Active	F.B.	R.B.
Sat	F.B.	F.B.

Pg-21 Ch-4

Q1.



In active region only

$$I_b = \frac{2.7-0.7}{40K}$$

$$= \frac{2}{40K} = \frac{1}{20}\ mA$$

$I_c = \beta I_b$

$= 100 \times \frac{1}{20}\ mA = 5mA$

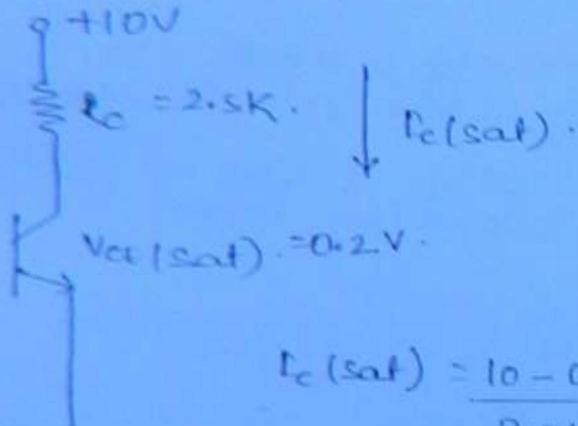
$V_{CE} = 10 - 2.5 \times 5\ mA$

$= -2.5\ V$

$V_{EB} + V_{BE} \simeq V_{CE}$

$V_{EB} = -2.5 - 0.7$

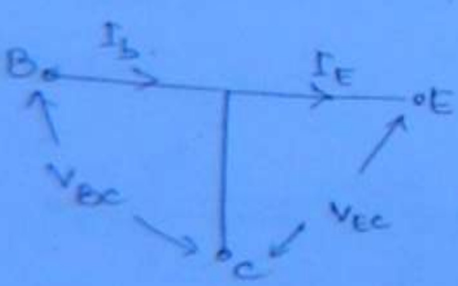
$= -3.2V$



$$I_C(sat) = \frac{10 - 0.2}{2.5K}$$

$$= 3.9mA$$

Common Collector characteristics →

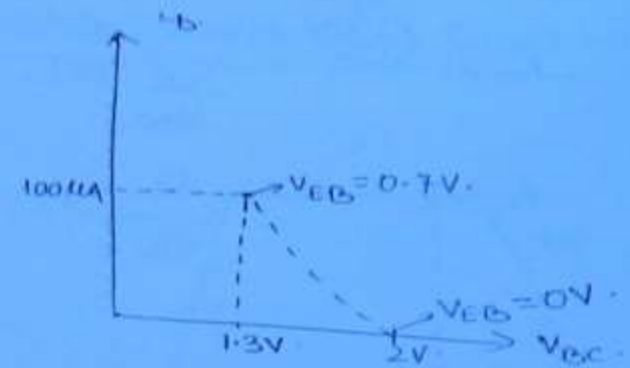


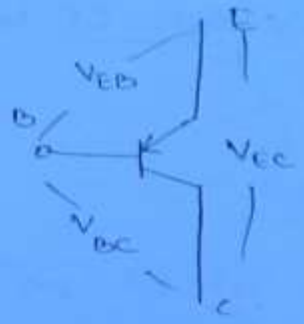
I/P parameters —

V_{BC}, I_b

O/P parameters —

V_{EC}, I_E





$V_{EC} = V_{EB} + V_{BC}$

$V_{BC} = V_{EC} - V_{EB}$.

Assume $V_{EC} = 2V$.

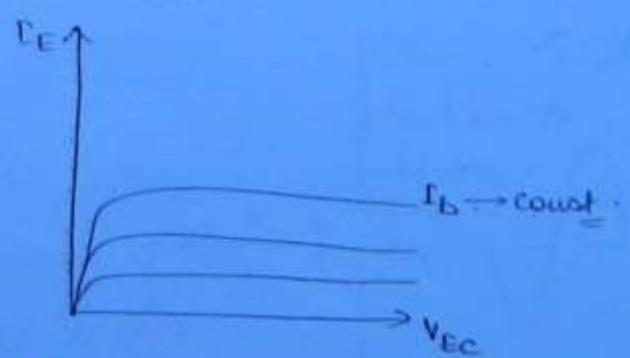
$V_{BC} = 2 - 0.7$

$= 1.3V$.

V_{EB} can create a base current I_b in the transistor. V_B can also create a base current.

common collector O/P characteristics :→.

C.C. O/P



Diff. b/w CB, CE, and C.C. :→

characteristics	CB	CE	CC
1) Z_i	$\frac{V_{EB}}{I_E} = \frac{0.7}{100mA} = 7\Omega$	moderate	$\frac{V_{BC}}{I_B} = \frac{10V}{1\mu A} > 10$
2) Z_o	$\frac{V_{CB}}{I_C} = \frac{10V}{10mA} = 1k\Omega$	moderate	$\frac{V_{EC}}{I_E} = \frac{10V}{[illegible]} = 1k$

3) A_V $\quad \frac{V_{CB}}{V_{EB}} > 1 \quad A_V > 1 \quad \frac{V_{EC}}{V_{BC}} \approx 1.$

4) A_I $\quad \frac{I_C}{I_E} = 1 \quad A_I > 1. \quad \frac{I_E}{I_B} > 1.$

Conclusions →

1) In most of the practical application, we choose CE because power gain will be high. $A_V > 1$, $A_I > 1$

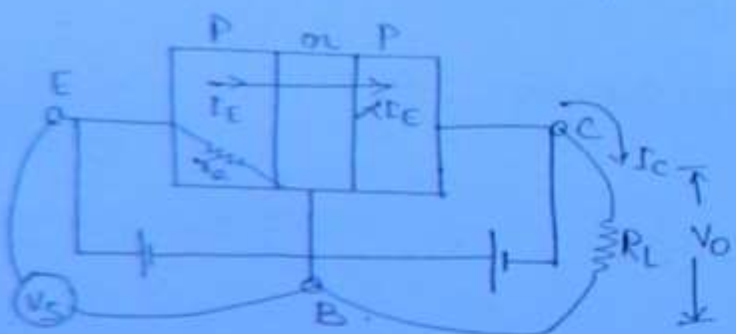
2) In CC, voltage gain = unity

3) In CB, current gain will be unity.

3/1/12

BJT Applications →

1) Transistor as an amplifier →



$$A_V = \frac{V_o}{V_i} = \frac{I_C R_L}{I_E r_e}$$

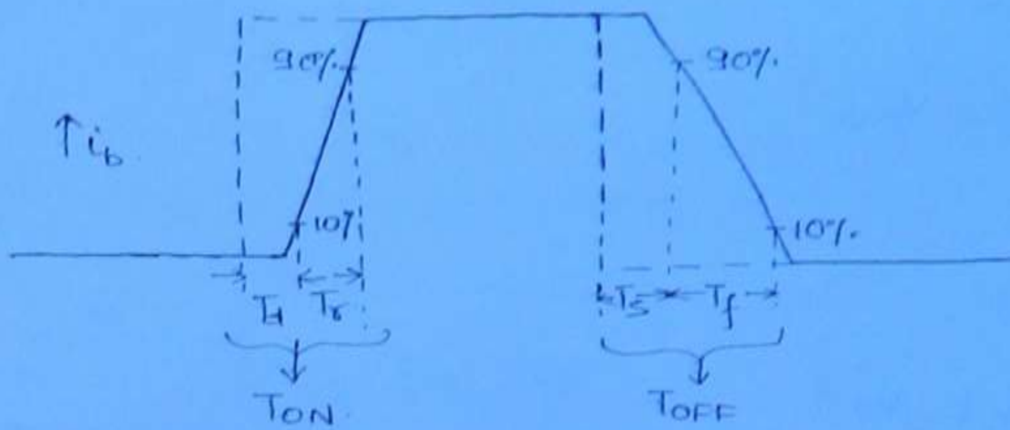
$$= \frac{\alpha I_E R_L}{I_E r_e}$$

$$= \frac{\alpha R_L}{r_e}$$

Assume $\alpha = 1$, $R_L = 3k\Omega$, $r_e = 30\Omega$

$$A_V = \frac{\alpha R_L}{r_e} = \frac{1 \times 3000}{30} = 100$$

2) Transistor switching times →



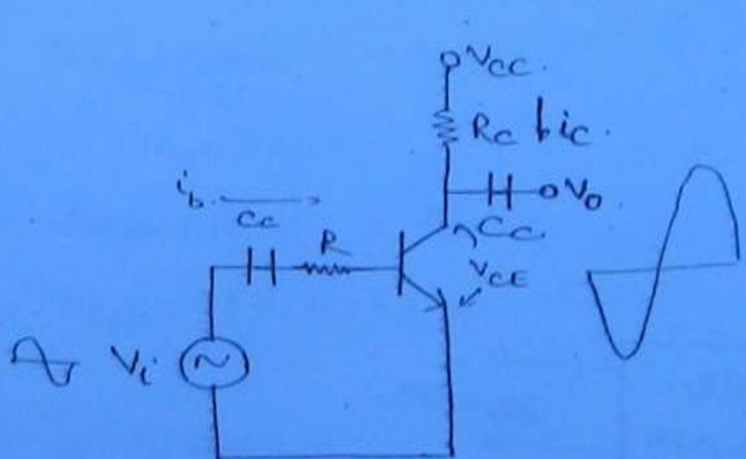
$T_{ON} = T_d + T_r$

delay time (for T_d), rise time (for T_r)

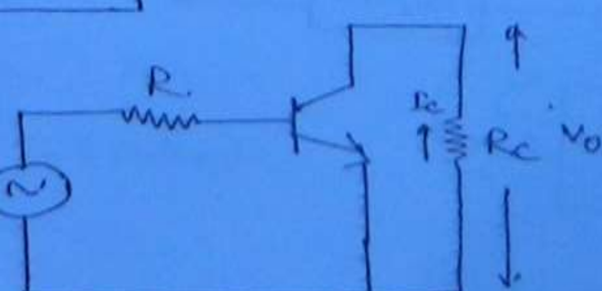
$T_{OFF} = T_s + T_f$

Storage time (for T_s), fall time (for T_f)

CE as 180° phase shift :→



Ac Analysis.



$$\boxed{V_0 = -i_c R_c}$$

+ive half cycle →

$i_b\uparrow$, $i_c\uparrow$, $i_cR_c\uparrow$, $V_0\downarrow$

$$\boxed{i_c = \beta i_b}$$

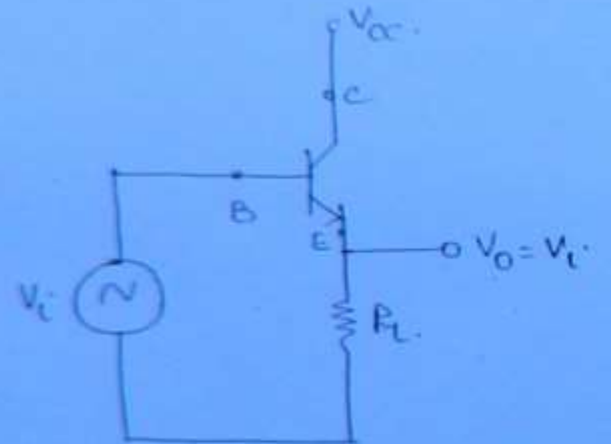
-ive half cycle →

$i_b\downarrow$, $i_c\downarrow$, $i_cR_c\downarrow$, $V_0\uparrow$

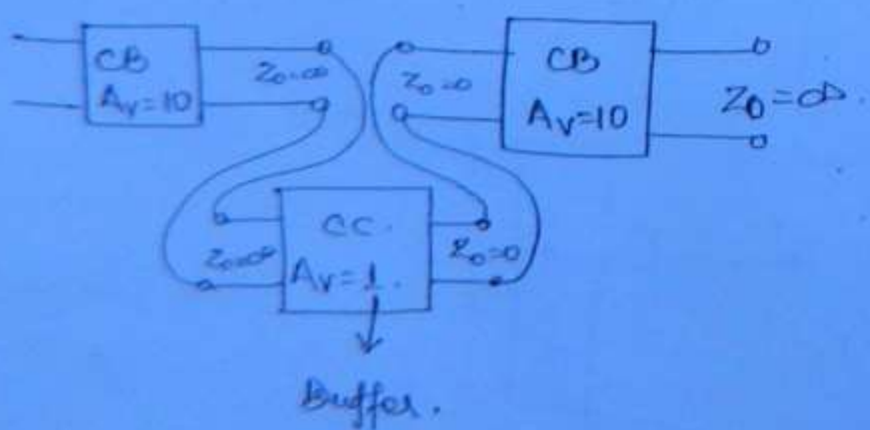
CC in impedance matching →

CC as emitter follower.

* $A_v = 1$

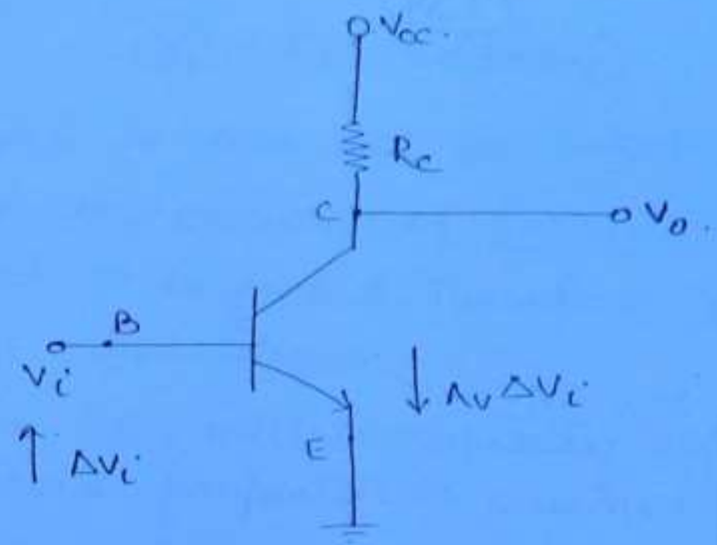


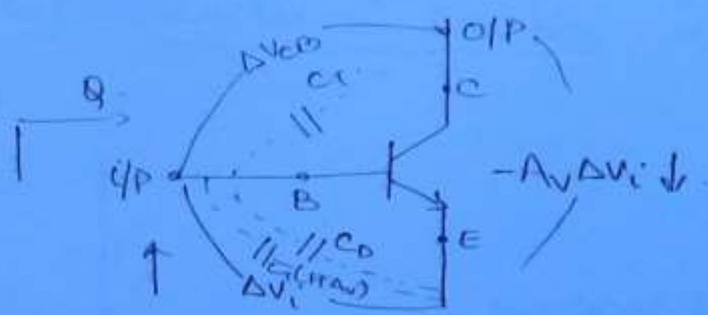
Impedance matching →



CB in high freq. application →

CE.





$$\Delta V_{CB} + \Delta V_i = -A_V \Delta V_i$$

$$V_B > V_C$$

$$-\Delta V_{CB} + \Delta V_i = -A_V \Delta V_i$$

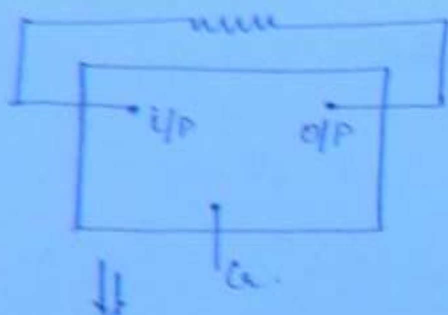
$$\Delta V_{CB} = (1 + A_V)\Delta V_i$$

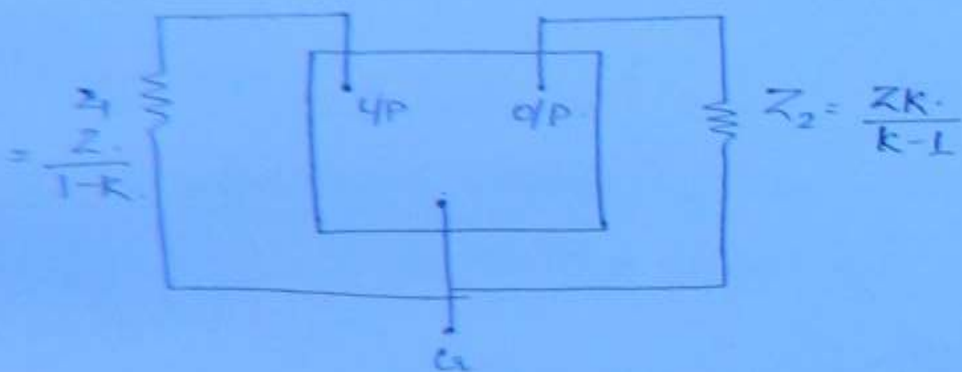
$$Q = C_T \Delta V_{CB}$$

$$= C_T(1 + A_V)\Delta V_i$$

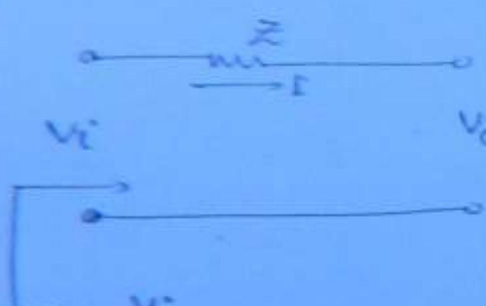
$$C_{in}' = C_D + C_T(1 + A_V)$$

Millers theorem :-





Proof →



$$Z_1 = \frac{V_i}{I}$$

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

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

$$Z_1 = \frac{V_i}{I} = \frac{Z}{1 - V_o/V_i}$$

$$\boxed{\frac{V_o}{V_i} = K}$$

Conclusion →

1) In CE, because of 180° phase shift property, millers capacitive effect is existing at the I/P as

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

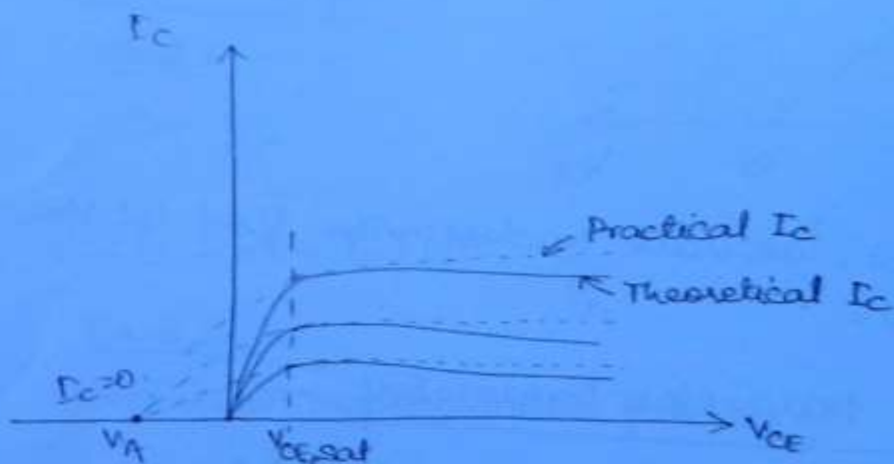
135

Hence It can not be used for high freq. applications.

2) In CB, Because of inphase property, millers capacity effect is neglected. Therefore it can be used for high freq. applications.

3) In CC, millers capacity effect is neglected because of in phase property. It can not be used for high freq. because voltage gain is unity.

i_c dependendence on V_{CE} →



$$\boxed{\uparrow I_C = I_S\, e^{V_{BE}/V_T}}$$

where $A_q \uparrow I_S = \dfrac{A q D_B P_{n_0}}{W_B \downarrow}$

early effect →

$V_{CE}\uparrow,\ W_B\downarrow$

$$I_C' = I_S e^{V_{BE}/V_T}\left[1 + \frac{V_{CE}}{V_A}\right]$$

if $V_A = \infty$

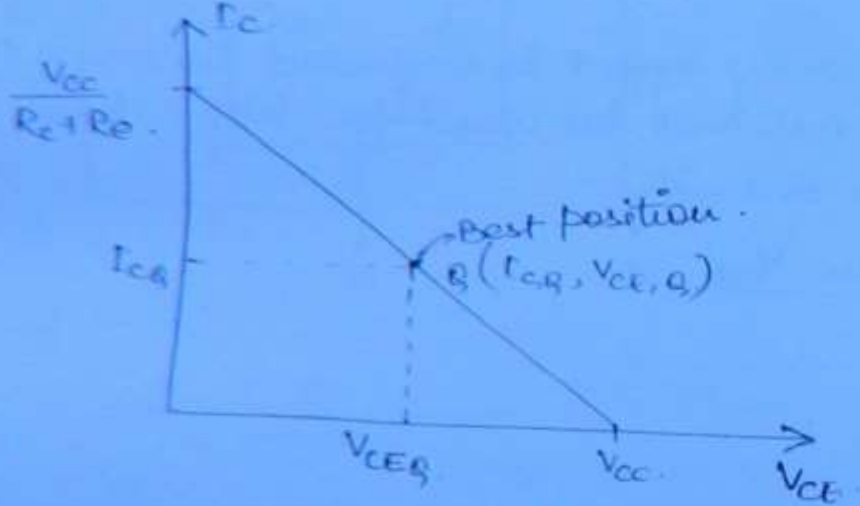
$\boxed{I_c' = I_c}$

if $V_A = 0$,

$\boxed{I_c' = \infty}$

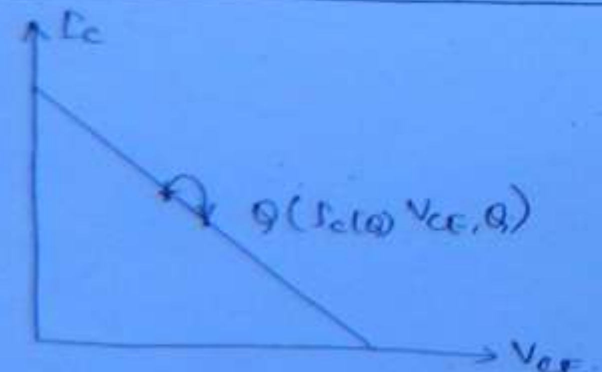
BJT biasing →

Location of Q point on DC load line.



For amplifier analysis, we have to design the Q pt at the centre of dc load line.

Temp dependence on transistor parameters :—



$I_{CQ} \rightarrow \underbrace{I_{CO}, V_{BE}, \beta}_{\text{Temp. dependence}}$

If $V_A = \infty$

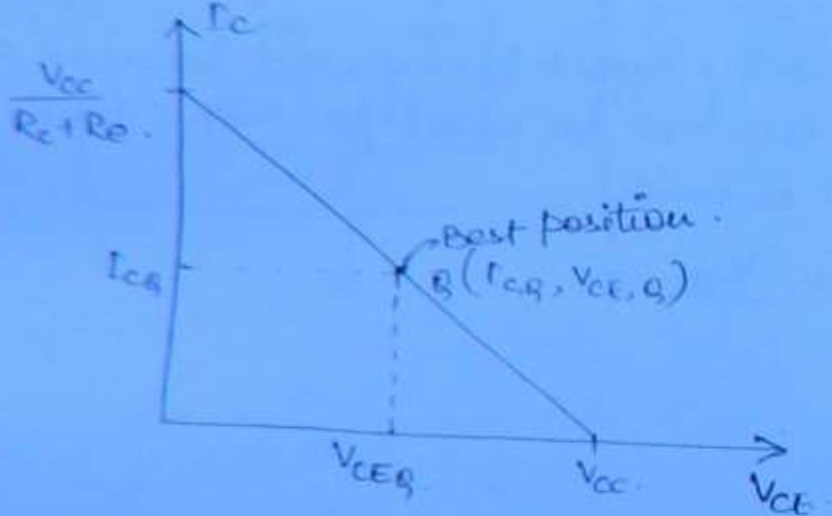
$$\boxed{I_c' = I_c}$$

If $V_A = 0$,

$$\boxed{I_c' = \infty}$$

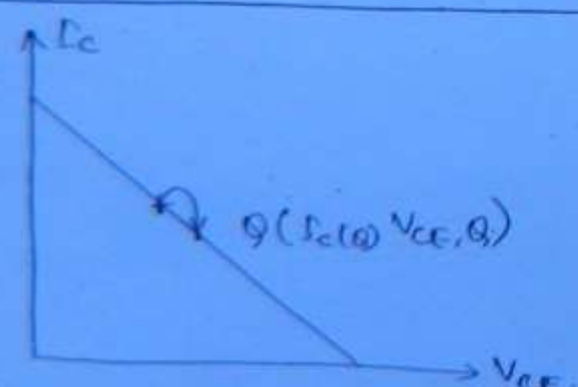
BJT biasing :→

Location of Q point on DC load line.



For amplifier analysis, we have to design the Q pt at the centre of dc load line.

Temp dependence on transistor parameters :—



$I_{CQ} \rightarrow \underbrace{I_{CO}, V_{BE}, \beta}$

↓

Temp. dependence.

I_{co} v/s temp —

I_{co} increases by 7% for every °c rise in temp or doubles for every 10°c rise in temp.

(137)

$T\uparrow, I_{co}\uparrow$

· As temp inc, I_{co} will increase.

V_{BE} v/s temp —

$$\frac{dV_{BE}}{dT} = -2.5\ mV/°c$$

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

β v/s temp →

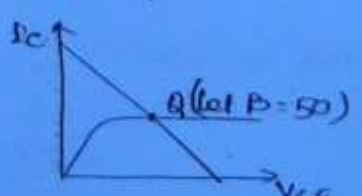
Case 1 —

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

$T\uparrow, I_{co}\uparrow, I_c\uparrow, \uparrow\beta = \frac{I_c}{I_b}$

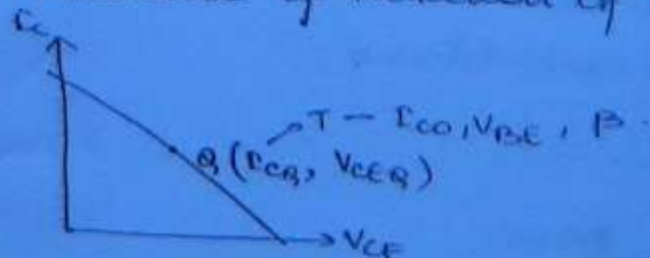
Case 2 →

transistor replacement problem.



Stability factor → "S"

It is a measure of variation of operating pt w.r.t. temp



$S' = \frac{\partial I_c}{\partial I_{co}}$ keeping $\beta, V_{BE} \to$ const

$S'' = \frac{\partial I_c}{\partial V_{BE}}$ $\beta, I_{co} \to$ const.

$S''' = \frac{\partial I_c}{\partial \beta}$ $I_{co}, V_{BE} \to$ const.

Stability factor must be less.

NOTE :—

The most dominant parameter w.r.t temp is I_{co}.
Ideally the stability factor should be 0.
Practically the stability factor should be mini in value.

Expression for stability factor :—

$$I_c = \beta I_b + (1+\beta) I_{co}$$

diff w.r.t I_c,

$$1 = \beta \frac{\partial I_b}{\partial I_c} + (1+\beta)\left(\frac{\partial I_{co}}{\partial I_c}\right)$$

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

Biasing techniques →

Biasing →

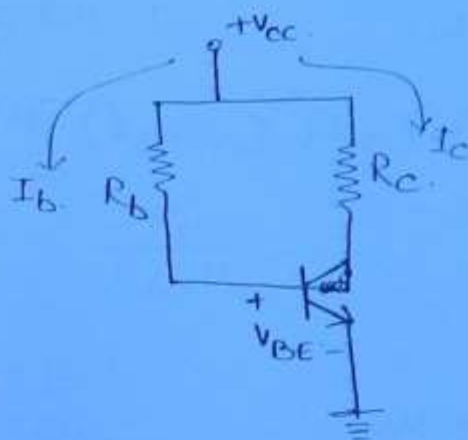
1) Stabilization techniques
→ Fixed bias
→ Collector to base bias.
→ Voltage divider bias
'or'
Self bias.
or
Emitter base
or
Universal bias

2) Compensation Techniques →

- Through diode
- Through Thermistor
- Through Sensistor

139

Fixed bias :—



I/P loop →

$$V_{CC} = I_b R_b + V_{BE}$$

$$I_b = \frac{V_{CC} - V_{BE}}{R_b}$$

$$V_{CC} >> V_{BE}$$

$$I_b = \frac{V_{CC}}{R_b}$$

$I_c = \beta I_b$ → depend on T & β replacement

↓ fixed ↓ fixed

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

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

$$I_b \approx \frac{V_{CC}}{R_b}$$

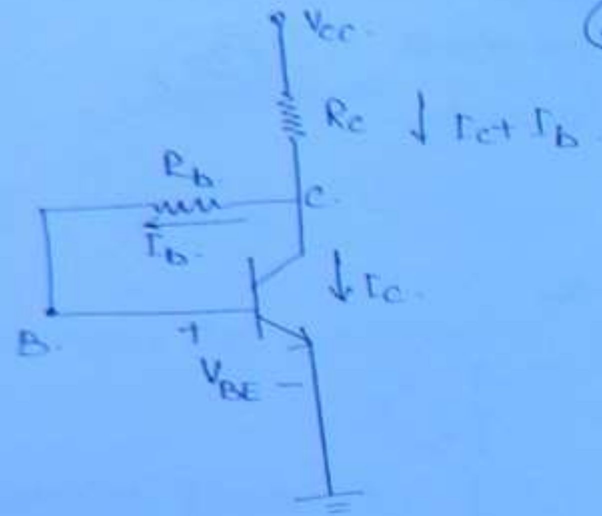
Assume $\beta = 100$

diff. w.r.t I_c

$\boxed{S = 101}$ → so high so not preferred fixed biasing

$$\frac{dI_b}{dI_c} = 0$$

Collector to Base bias :—



I/P loop →

$$V_{cc} = (I_C + I_b) R_C + I_b R_b + V_{BE}$$

$$V_{cc} = I_b (R_b + R_C) + I_C R_C + V_{BE}$$

$$I_b = \frac{V_{cc} - I_C R_C - V_{BE}}{R_b + R_C}$$

O/P loop →

$$V_{cc} = (I_C + I_b) R_C + V_{CE}$$

$$V_{cc} - I_C R_C = I_b R_C + V_{CE}$$

$$\therefore \boxed{I_b = \frac{V_{CE} + I_b R_C - V_{BE}}{R_b + R_C}}$$

Analysis →

$T\uparrow$, $I_{co}\uparrow$, $I_C\uparrow$.

$(I_C + I_b) R_C \uparrow$, $V_{CE}\downarrow$, $I_b\downarrow$, $I_C\downarrow$

$$\boxed{\downarrow I_C = \beta I_b \downarrow}$$

Stability factor in collector to base biasing.

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

$$V_{cc} = I_b(R_b+R_c) + I_cR_c + V_{BE}$$

diff w.r.t. I_c

$$0 = \frac{\partial I_b}{\partial I_c}(R_b+R_c) + R_c + 0$$

$$\frac{\partial I_b}{\partial I_c} = \frac{-R_c}{R_b+R_c}$$

$$S = \frac{1+\beta}{1+\beta\frac{R_c}{R_b+R_c}}$$

$$\Rightarrow S = \frac{1+\beta}{1+\frac{\beta}{\frac{R_b}{R_c}+1}}$$

when $\frac{R_b}{R_c} << 1$. $\boxed{R_c >> R_b}$

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

$$\boxed{S = 1}$$

Conclusions –

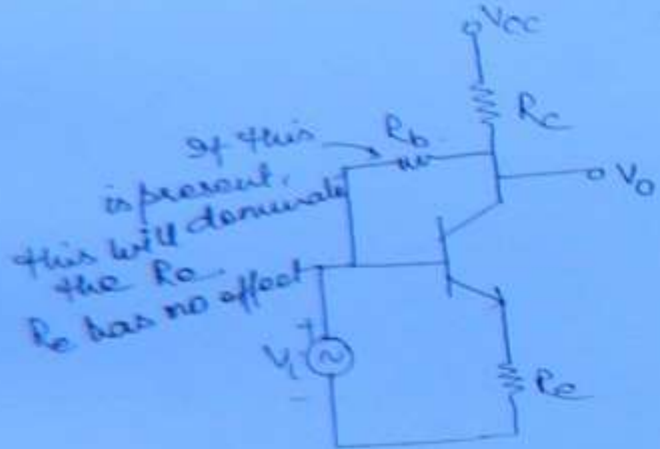
1) If $\frac{R_b}{R_c} << 1$, The stability factor approaches to unity.

2) If R_b is less, base current drawn from the battery will be more that too means it reduces the battery life time

3) If R_c is more, power dissipation losses increases and it may disturb the Q pt also.

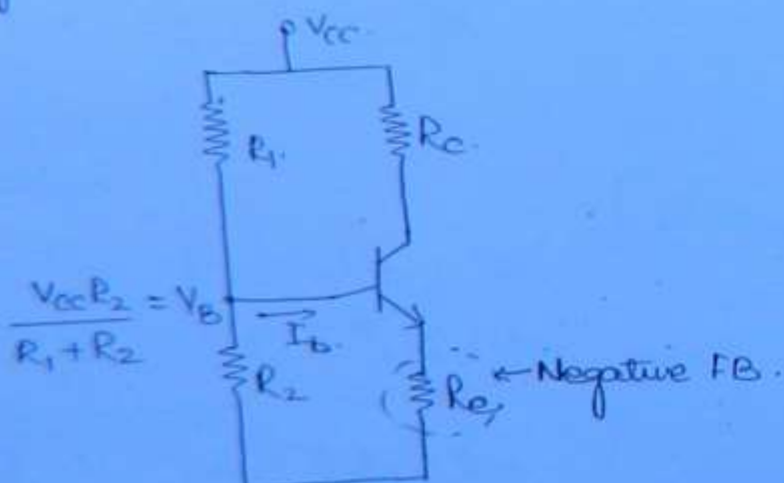
Voltage divider bias :—

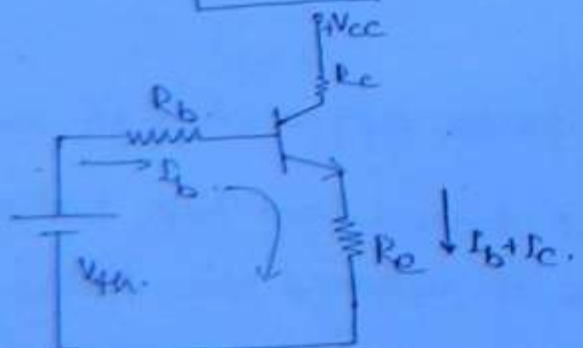
Feedback
- +ive feedback $A_v \uparrow$, $S \downarrow$ → oscillators
- -ive feedback. $A_v \downarrow$, $S \uparrow$ → Amplifier



* If we have two feedbacks in the ckt, we take the feedback which is nearest to the Vcc.

Voltage divider bias ckt :—





$$V_{Th} = \frac{V_{cc} \times R_2}{R_1 + R_2}$$

$$R_b = R_1 \parallel R_2$$

I/P loop :—

$$V_{Th} = I_B R_B + V_{BE} + \cancel{I_B I_C}(I_B + I_C) R_e$$

$$V_{Th} = I_B (R_B + R_e) + V_{BE} + I_C R_e$$

diff w.r.t. I_C

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

$$\Rightarrow \boxed{\frac{\partial I_B}{\partial I_C} = \frac{-R_e}{R_B + R_e}}$$

$$\boxed{S = \frac{1+\beta}{1+\beta \frac{R_e}{R_B + R_e}}}$$

$$= \frac{1+\beta}{1+\beta \frac{1}{R_B/R_e + 1}}$$

if $\frac{R_B}{R_e} << 1$

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

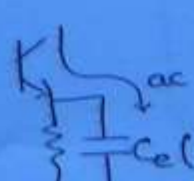
$$\boxed{S = 1}$$

Conclusions →

1) If $\frac{R_B}{R_e} << 1$, Stability factor reaches to unity

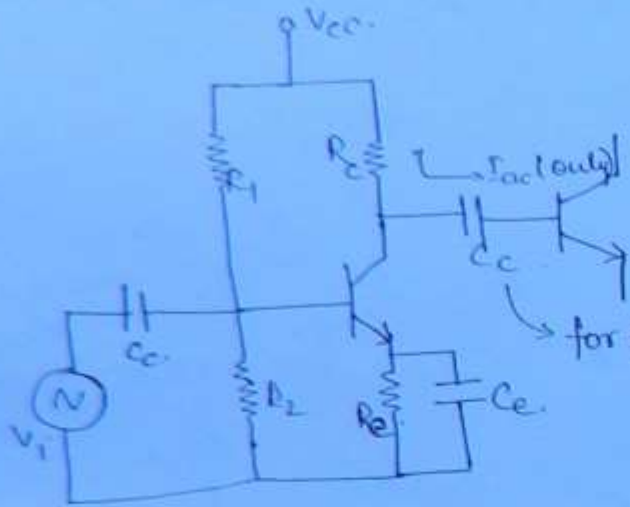
2) R_B value should be less that means R_1 and R_2 resistors should be less. If R_1 is less, it reduces the battery life time. So, take $R_1 > R_2$ in the design.

3)

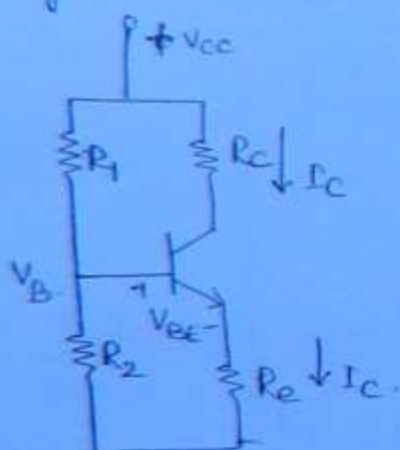


C_e (by pass capacitor) to make the gain stable

If Re is more, it is increasing the -ve feedback nature of the ckt that means the gain be affected. This problem can be solved by keeping a shunt capacitor C_e (by pass capacitor) to resistor R_e. Bypass



Analysis →



$T\uparrow,\ I_{co}\uparrow,\ I_c\uparrow$

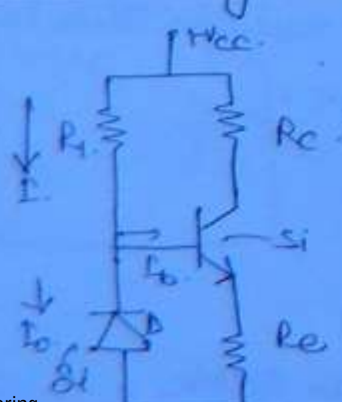
$I_cR_e\uparrow,\ V_{BE}\downarrow,\ I_b\downarrow$

$I_c\downarrow$

$$\boxed{V_B = \downarrow V_{BE} + I_cR_e\uparrow}$$

compensation techniques →

compensation through diode →



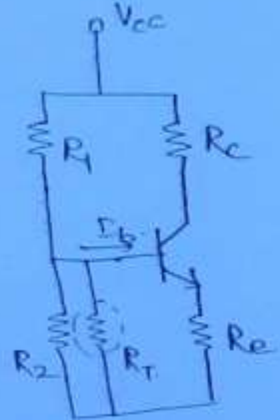
$T\uparrow, I_{co}\uparrow, I_c\uparrow, I_D\uparrow, I_b\downarrow, I_c\downarrow$

Compensation through Thermistor →

Thermistor →

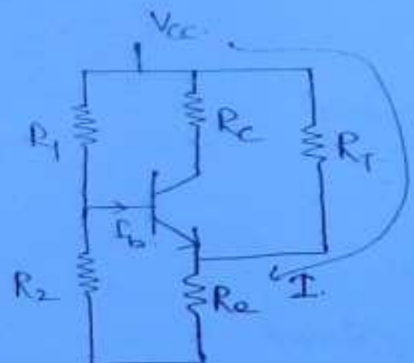
(145)

$T\uparrow$, $R_T\downarrow$, -ive temp coeff.



$T\uparrow$, $I_{CO}\uparrow$, $I_C\uparrow$, $R_T\downarrow$, $R_2 \| R_T\downarrow$

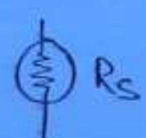
$V_B\downarrow$, $I_b\downarrow$, $I_C\downarrow$



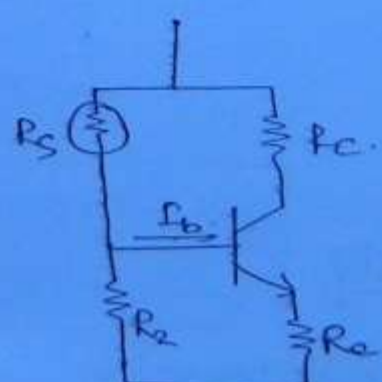
$T\uparrow$, $I_{CO}\uparrow$, $I_C\uparrow$, $R_T\downarrow$, $I_e\uparrow$, $I_eR_e\uparrow$,

$V_{BE}\downarrow$, $I_b\downarrow$, $I_C\downarrow$

Compensation through Sensistor →



$T\uparrow$, $R_S\uparrow$.



$T\uparrow$, $I_{CO}\uparrow$, $I_C\uparrow$,

$R_S\uparrow$, $I\downarrow$, $I_b\downarrow$, $I_C\downarrow$

10/1/12

Thermal runaway :—

$T\uparrow$, $I_{CO}\uparrow$, $I_C\uparrow$, $P_D = I_C^2 R_C$, (146)

Thermal resistance (θ)

$T_J - T_A \propto P_D$

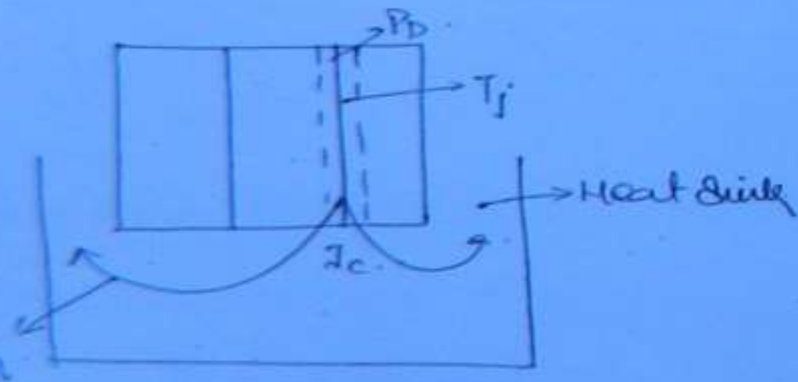
$= \theta P_D$ $\quad\theta \to$ thermal resistance

$T_J \to j^n$ temp

$T_A \to$ Ambient temp

$P_D \to$ power dissipation

eg



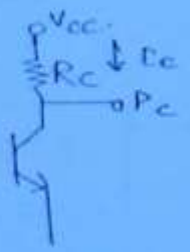
$$\theta = \frac{T_J - T_A}{P_D} \quad {}^\circ C/mW$$

condition for Thermal stability

$$\frac{\partial P_C}{\partial T_J} < \frac{\partial P_D}{\partial T_J} \to \text{safe operation}$$

The rate of heat generated at the collector j^n should not exceed the heat dissipated at the collector j^n.

147



P_C = Power dissipated generated at C-j^n.

$$T_j - T_A = \theta P_D$$

diff w.r.t T_j

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

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

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

$P_C =$

$$\frac{\partial P_C}{\partial T_j} = \frac{\partial P_C}{\partial I_C} \frac{\partial I_C}{\partial T_j}$$

$$P_C = V_{CC} I_C - I_C^2 R_C$$

diff. w.r.t I_C,

$$\frac{\partial P_C}{\partial I_C} = V_{CC} - 2 I_C R_C$$

$$\frac{\partial I_C}{\partial T_j} = \frac{\partial I_C}{\partial I_{CO}} \frac{\partial I_{CO}}{\partial T_j}$$

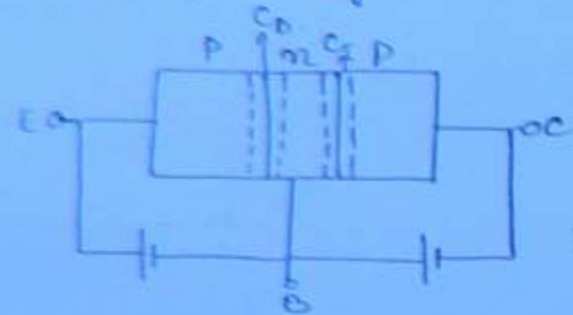
$$= S \times 0.07 I_{CO}$$

$$\boxed{(V_{CC} - 2 I_C R_C) S \, 0.07 \, I_{CO} < \frac{1}{\theta}}$$

Small Signal Analysis :—

BJT :—

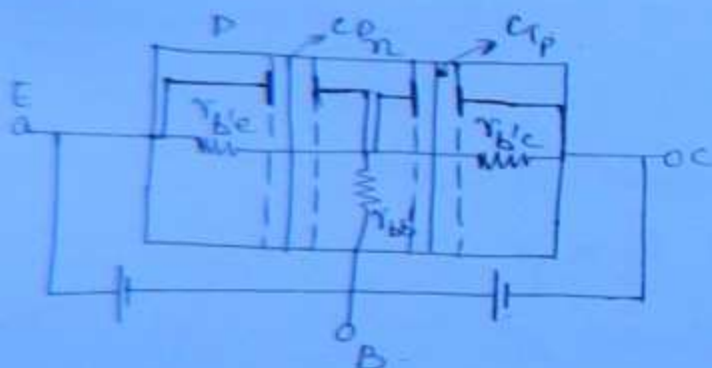
- Low freq. Analysis



$X_C \propto \frac{1}{f}$, $f \to 0$, $X_C \to \infty$

Internal capacitance should be open in low freq.

- High freq. Analysis.



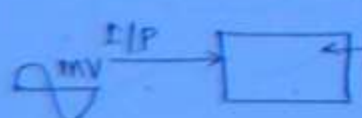
$X_C \propto f$, $f \to \infty$, $X_C \to 0$

∴ Internal capacitance are taken into account

We can not short ckt the capacitance as if it happens all resistance will be gone and no action as BJT.

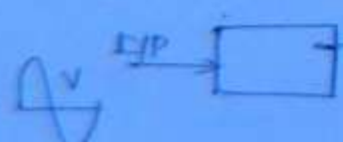
Q. What is the diff. b/w small signal and large signal in amplifiers?

Ans.

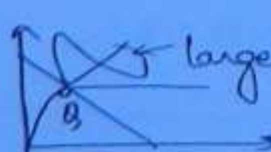


Amplifier → $\boxed{\text{mathematical model}}$ → two port N/W.

(V.A) → Voltage Amplifier.



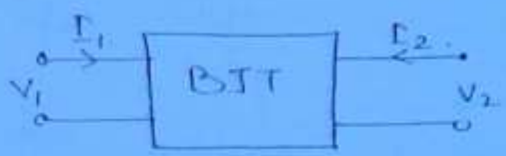
Amplifier
(P.A)
↑
Power Amplifier.



Low freq. Analysis →

Two port n/w →

(149)



$V_1 = h_{11} I_1 + h_{12} V_2$

$I_2 = h_{21} I_1 + h_{22} V_2$

Case 1. →

$V_2 = 0$

$h_{11} = \frac{V_1}{I_1}$ → I/P impedance.

$h_{21} = \frac{I_2}{I_1}$ → forward current gain.

Case 2 →

$I_1 = 0$

$h_{12} = \frac{V_1}{V_2}$ → Reverse Voltage gain.

$h_{22} = \frac{I_2}{V_2}$ → O/P admittance.

$$\boxed{\begin{aligned} V_1 &= h_i I_1 + h_r V_2 \\ I_2 &= h_f I_1 + h_o V_2 \end{aligned}}$$

For amplifier analysis, we require general char. like

1) Input impedance Z_i

2) O/P impedance Z_o.

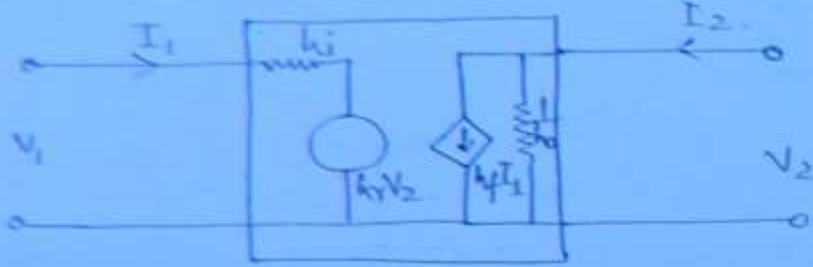
3) Voltage gain A_V

4) Current gain A_I

All these char. are efficiently explained by h-parameter model.

h-parameter model →

low freq. analysis.



CE	CB	CC
hie	hib	hic
hfe	hfb	hfc
hre	hrb	hrc
hoe	hob	hoc

Typical Values of h-parameters :→

Parameter	CE	CC	CB
h_i	1100Ω	1100Ω	22Ω
h_{fe}	50	-51 $\frac{I_E}{I_b}$	-0.98 $\frac{I_c}{I_E}$
h_r	2.4×10^{-4}	≈ 1	2.9×10^{-4}
h_o	24×10^{-6} A/V	25×10^{-6} A/V	0.49×10^{-6} A/V

Conversion Techniques →

CE to CB.

$$h_{ib} = \frac{h_{ie}}{1+h_{fe}}$$

$$h_{fb} = \frac{-h_{fe}}{1+h_{fe}}$$

$$h_{ob} = \frac{h_{oe}}{1+h_{fe}}$$

$$h_{rb} = \frac{h_{ie}h_{oe}}{1+h_{fe}} - h_{re}$$

CB TO CE :→

(157)

$$h_{ie} = \frac{h_{ib}}{1+h_{fb}}$$

$$h_{fe} = \frac{-h_{fb}}{1+h_{fb}}$$

$$h_{re} = \frac{h_{ib}h_{ob}}{1+h_{fb}} - h_{rb}$$

$$h_{oe} = \frac{h_{ob}}{1+h_{fb}}$$

CE to CC :→

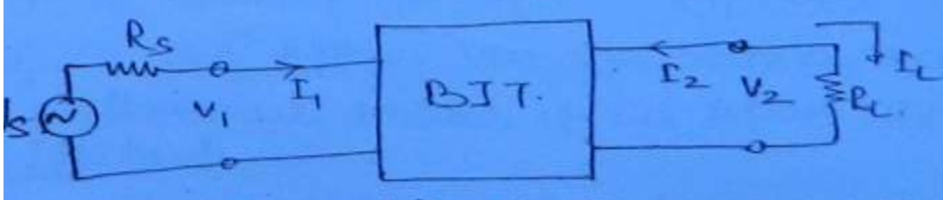
$$h_{ic} = h_{ie}$$

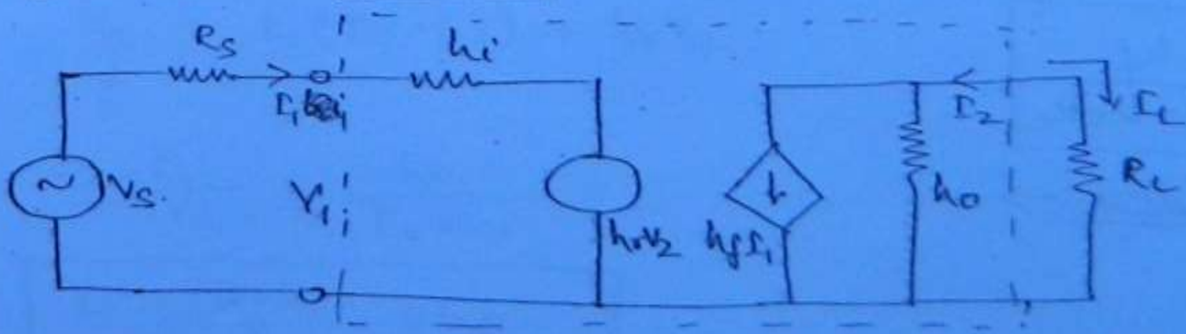
$$h_{fc} = -(1+h_{fe})$$

$$h_{rc} = 1$$

$$h_{oc} = h_{oe}$$

Transistor Amplifier Analysis :→
[using exact model]





Characteristics of Amplifier →

1) current gain A_I

2) I/P impedance Z_i

152

3) voltage gain A_V

4) O/P impedance Z_o.

5) voltage amplification A_{VS}

6) Current Amplification A_{IS}

7) effective I/P impedance Z_i'

8) effective O/P impedance Z_o'.

Current gain →

$$V_1 = h_i I_1 + h_r V_2$$

$$I_2 = h_f I_1 + h_o V_2$$

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

$$I_2 = h_f I_1 + h_o V_2 \qquad V_2 = I_L R_L = -I_2 R_L.$$

$$I_2 = h_f I_1 + h_o(-I_2 R_L)$$

$$\Rightarrow I_2(1 + h_o R_L) = h_f I_1$$

$$\Rightarrow \frac{I_2}{I_1} = \frac{h_f}{1 + h_o R_L}$$

$$\boxed{A_I = -\frac{I_2}{I_1} = \frac{-h_f}{1 + h_o R_L}}$$

Input Impedance →

$$Z_i = \frac{V_1}{I_1}.$$

$$V_1 = h_i I_1 + h_r V_2$$

$$\frac{V_1}{I_1} = h_i + h_r \frac{V_2}{I_1} = h_i + h_r \frac{(-I_2 R_L)}{I_1} = h_i + A_I h_r R_L.$$

$$\boxed{Z_i = h_i + h_r A_I R_L}$$

Voltage gain :— 153

$$A_V = \frac{V_2}{V_1}$$

$$= \frac{-\frac{I_2 R_L}{I_1}}{V_1/I_1}$$

$$\boxed{A_V = \frac{A_I R_L}{Z_i}}$$

Output Impedance →

$$Z_0 = \frac{V_2}{I_2}$$

$$Y_0 = \frac{I_2}{V_2}$$

$$= h_f\left(\frac{I_1}{V_2}\right) + h_o\left(\frac{V_2}{V_2}\right)$$

$$= h_f\left(\frac{I_1}{V_2}\right) + h_o .$$

To cal. the O/P impedance, two conditions should be follow

1) R_L should be open. ($R_L = \infty$)

2) If the I/P is voltage source, make it as short ckt otherwise make it as open ckt if it is a current source

I/P loop →

$$\underset{\substack{\downarrow \\ 0}}{V_s} = I_1(R_s + h_i) + h_r V_2 .$$

$$\Rightarrow \frac{I_1}{V_2} = \frac{-h_r}{R_s + h_i}$$

$$\boxed{Y_0 = \frac{-h_f h_r}{R_s + h_i} + h_o}$$

$$\boxed{Z_0 = \frac{1}{Y_0}}$$

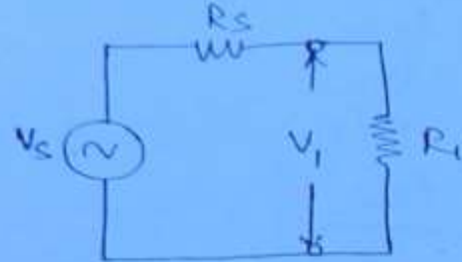
Voltage Amplification →

$$A_{VS} = \frac{V_2}{V_s}$$

$$= \frac{V_2}{V_1} \times \frac{V_1}{V_s}$$

$$= A_v\left(\frac{V_1}{V_s}\right)$$

$$\boxed{A_{VS} = A_v \frac{R_i}{R_i+R_s}}$$



$$V_1 = \frac{R_i}{R_i+R_s} V_s$$

$$\frac{V_1}{V_s} = \frac{R_i}{R_i+R_s}$$

When $R_s = 0$ [Ideality].

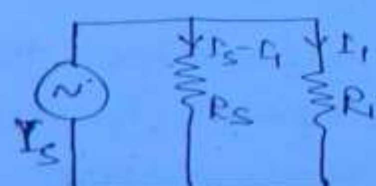
$$\boxed{A_{VS} = A_V}$$

Current Amplification →

$$A_{IS} = \frac{I_L}{I_s} = \frac{I_L}{I_1} \times \frac{I_1}{I_s}$$

$$= A_I \frac{I_1}{I_s}$$

$$= A_I \frac{R_s}{R_i+R_s}$$



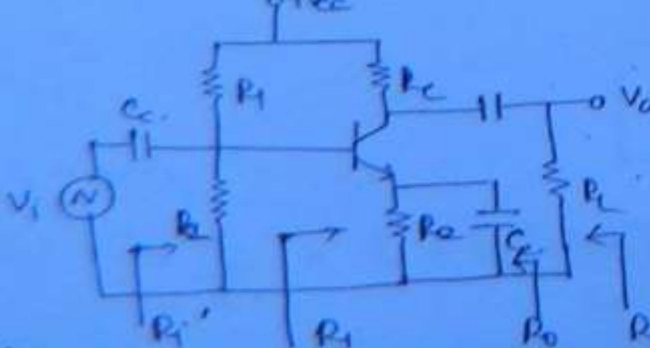
$$(I_s - I_1)R_s = I_1 R_i$$

$$\frac{I_1}{I_s} = \frac{R_s}{R_i+R_s}$$

when $R_i = 0$,

$$\boxed{A_{IS} = A_I}$$

Effective I/P and O/P Imp :→



$R_i = h_{ie}$

$R_i' = R_1 \| R_2 \| h_{ie}$

$R_o = \frac{1}{h_{oe}}$ (155)

$R_o' = \frac{1}{h_{oe}} \| R_C \| R_L$

Conclusion →

Exact analysis is valid for all the 3 types of configuratio like common emitter, CB, CC.

For eg →

$$(A_I)_{CB} = \frac{-h_{fb}}{1+h_{ob}R_L}$$

$$(A_I)_{CC} = \frac{-h_{fc}}{1+h_{oc}R_L}$$

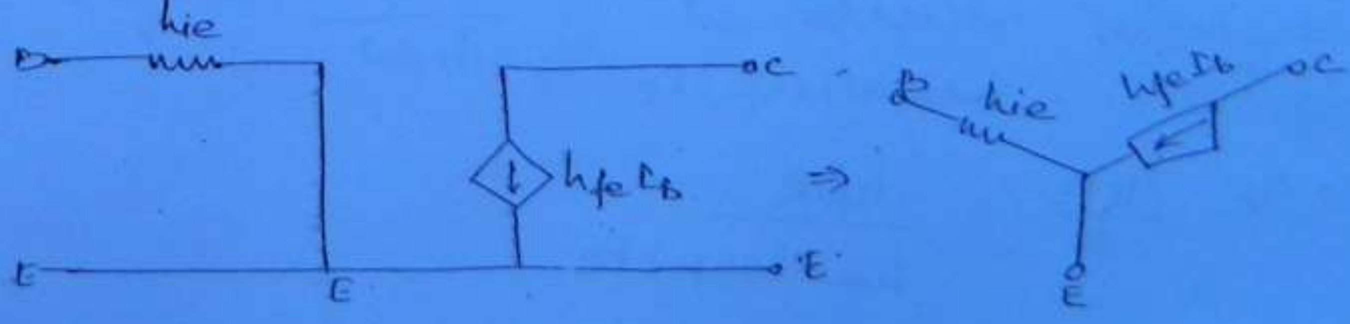
Approximate model Analysis :→

1) $h_{oe}R_L \leq 0.1$

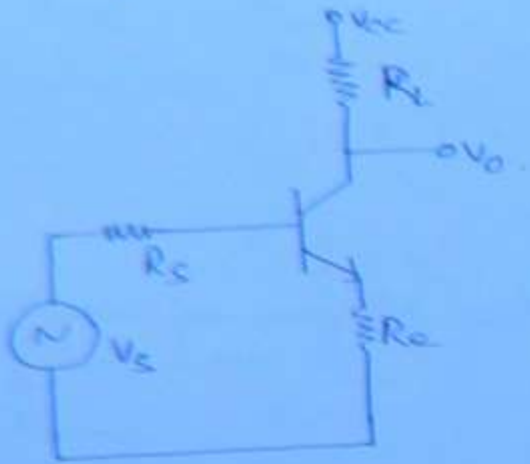
eg. $$A_I' = \frac{-h_{fe}}{1+h_{oe}R_L} = \frac{-h_{fe}}{1+0.1} = -.91\,h_{fe}$$

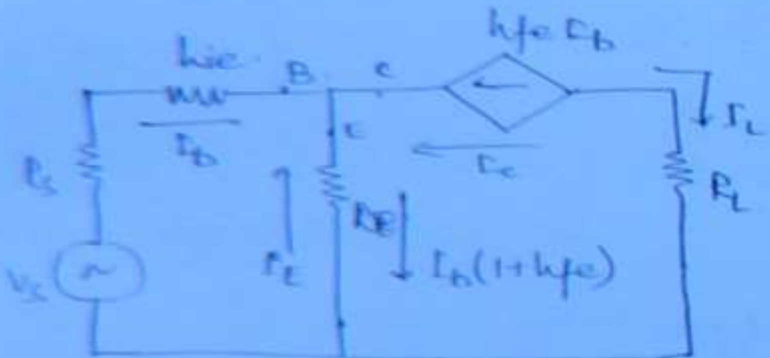
2) h_{re} and h_{oe} should be neglected

3) This is valid in CE mode only.



CE unbypass amplifier :-





1) $A_I = \frac{I_L}{I_b} = \frac{-h_{fe} I_b}{I_b}$

$= -h_{fe}$

or $\frac{-h_{fe}}{1+h_{oe}R_L}$ $h_{oe} \rightarrow$ neglected

$= -h_{fe}$

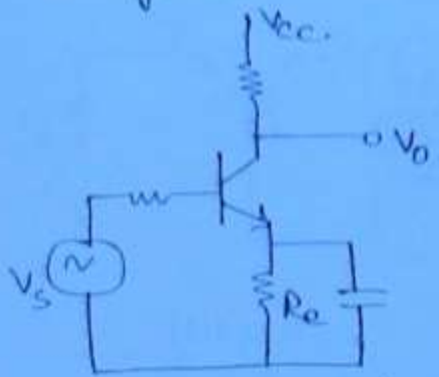
2) $R_i = \frac{V_1}{I_b} = \frac{h_{ie} I_b + (1+h_{fe}) I_b R_e}{I_b}$

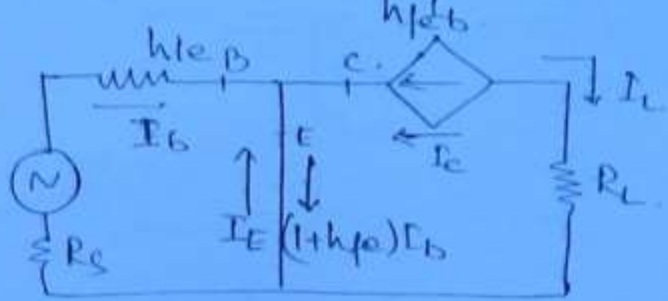
$= h_{ie}\cancel{I_b} + (1+h_{fe}) R_e$

3) $A_V = \frac{A_I R_L}{R_i} = \frac{-h_{fe} R_L}{h_{ie} + (1+h_{fe}) R_e}$

4) $R_o = \frac{V_2}{I_2} \Rightarrow \infty$

CE bypass →





$$A_I = \frac{I_L}{I_b} = -h_{fe}$$

$$R_i = h_{ie}$$

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

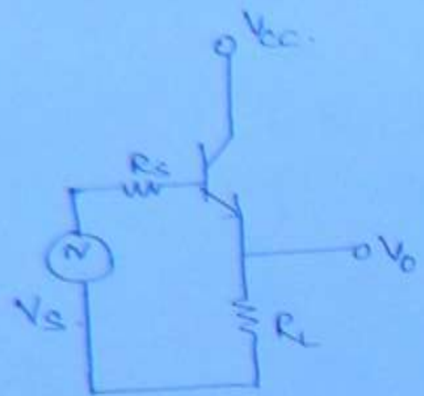
$$R_o = \frac{V_2}{I_c \to 0} \Rightarrow \infty.$$

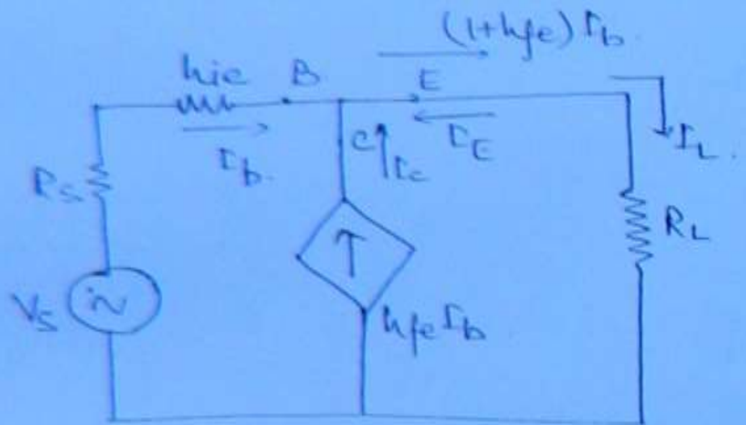
Table →

	A_V	R_i	Feedback
CE bypass	$-h_{fe}R_L$	h_{ie}	No
CE Unbypass	$\frac{-h_{fe}R_L}{h_{ie}+(1+h_{fe})R_e}$	$h_{ie} + (1+h_{fe})R_e$	Yes (Negative)

CC →

(158)





$$A_I = \frac{I_L}{I_b} = \frac{(1+h_{fe}) I_b}{I_b} = 1+h_{fe}.$$

$$R_i = \frac{V_1}{I_b} = \frac{h_{ie} I_b + (1+h_{fe}) I_b R_L}{I_b}$$

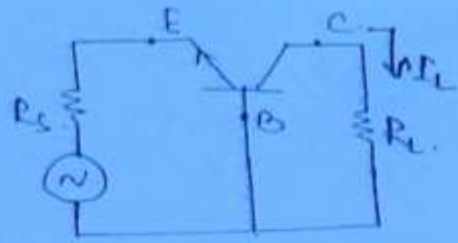
$$= h_{ie} + (1+h_{fe}) R_L.$$

$$A_V = \frac{(1+h_{fe}) R_L}{h_{ie} + (1+h_{fe}) R_L}.$$

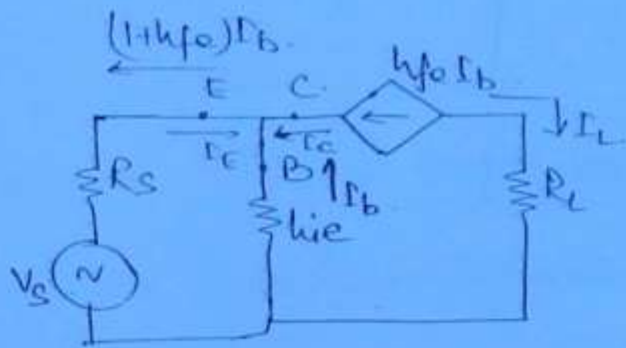
$$R_o = \frac{1}{h_{oc} - \frac{h_{rc} h_{fc}}{h_{ic} + R_s}} \Rightarrow ?$$

$$= \frac{h_{ie} + R_s}{1+h_{fe}}$$

CB. →



(159)



$$A_I = \frac{I_L}{I_E} = \frac{-h_{fe} I_b}{(1+h_{fe}) I_b}$$

$$= \frac{h_{fe}}{1+h_{fe}}$$

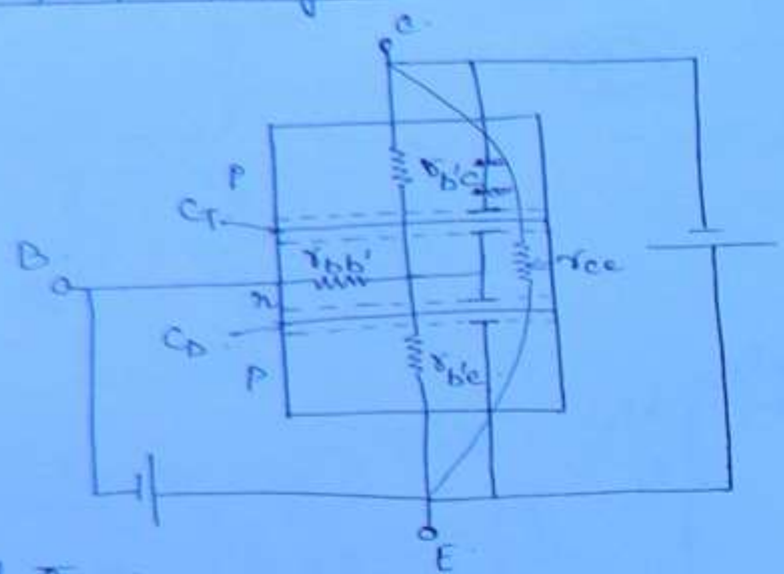
$$R_i = \frac{V_1}{I_E} = \frac{-h_{ie} I_b}{-(1+h_{fe}) I_b}$$

$$= \frac{h_{ie}}{1+h_{fe}}$$

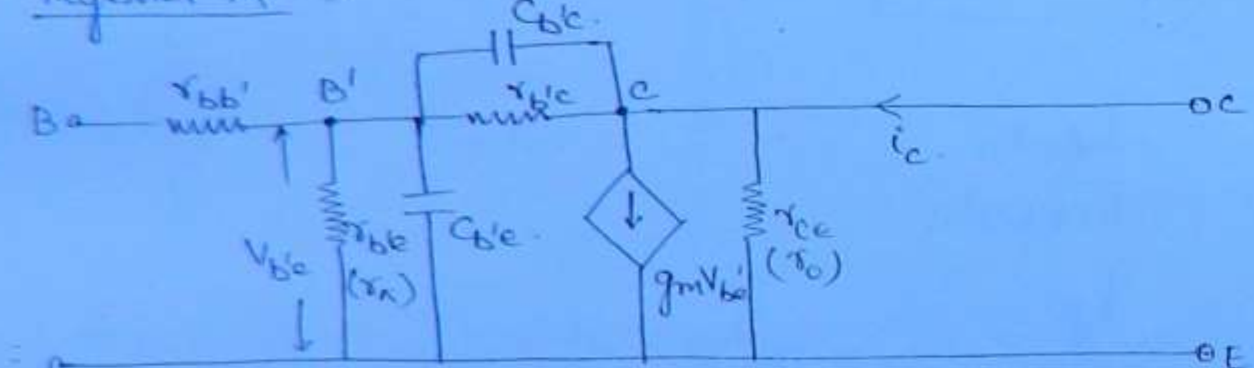
$$A_V = \frac{h_{fe} R_L}{h_{ie}}$$ [no minus sign beause of same phase shif

$$R_0 = \frac{V_2}{I_C} \Rightarrow \infty .$$

High freq. analysis of BJT :—

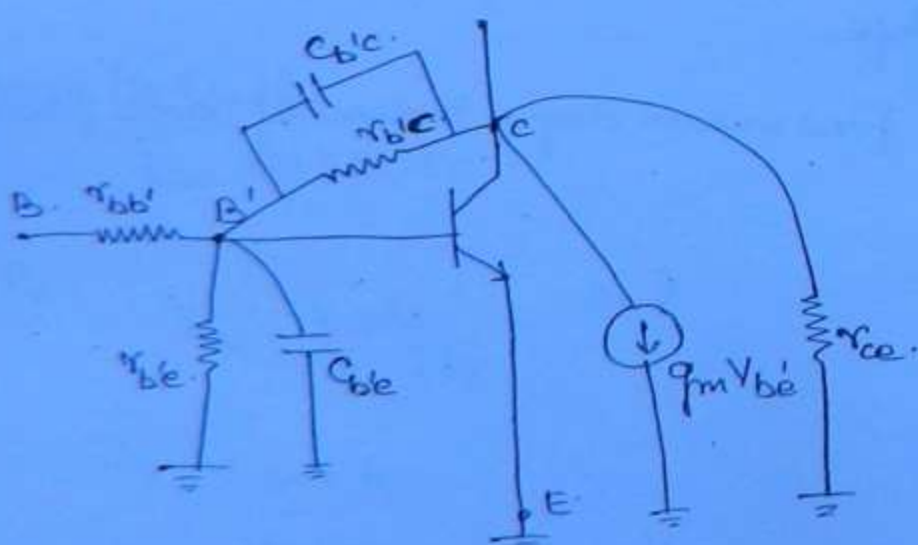


Hybrid π →



$$g_m V_{b'e} = \frac{i_c}{V_{b'e}} \times V_{b'e}$$

$$= i_c$$



$g_m \rightarrow$ transconductance $= 50\,mA/V$

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

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

$r_{bb'}$ — base spread resistance $= 100\Omega$

$r_{ce} \rightarrow$ O/P resistance $= 80K\Omega$

$C_{b'e} \rightarrow$ Diffusion Capacitance $= 100pF$

$C_{b'c} \rightarrow$ Transition Capacitance $= 3pF$

Conclusion :→

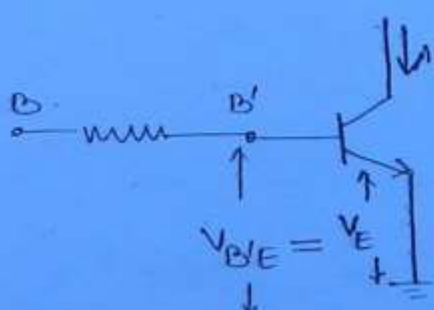
Hybrid π parameters are depending on 3 parameters

1) Collector current I_C

2) temp T

3) Collector to Emitter Voltage V_{CE}.

Transconductance g_m :→



$$g_m = \frac{\partial I_C}{\partial V_{B'E}}$$

$$I_C = \alpha I_E$$

$$\partial I_C = \alpha \partial I_E$$

$$g_m = \alpha \frac{\partial I_E}{\partial V_{B'E}} \qquad \boxed{V_{B'E} = V_E}$$

$$g = \frac{I + I_0}{\eta V_T}$$

F.B.

$$g = \frac{I}{\eta V_T}$$

$$g = \frac{\partial I_E}{\partial V_E} = \frac{I_E}{\eta V_T}$$

$$= \frac{\alpha I_E}{\eta V_T}$$

$$= \left|\frac{I_C}{V_T}\right| \quad \eta = 1$$

I_C → dc current

$I_C \uparrow,\ g_m \uparrow$

$T \uparrow,\ g_m \downarrow$

$V_{CE} \uparrow,\ g_m$ – const

[illegible]

1) As collector current I_c increases, transconductance g_m also ~~is~~ increases.

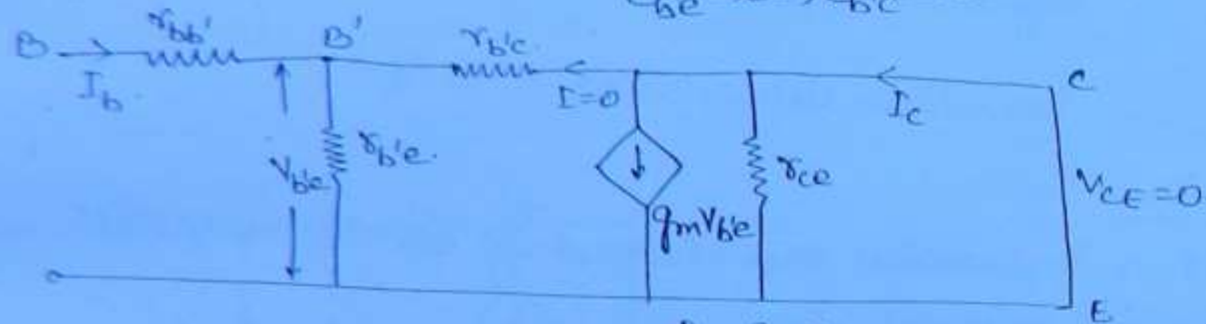
2) As temp increases, transconductance $g_m \downarrow$. $\boxed{g_m = \frac{I_c}{T} \times 11600}$

162

$r_{b'e} \rightarrow$

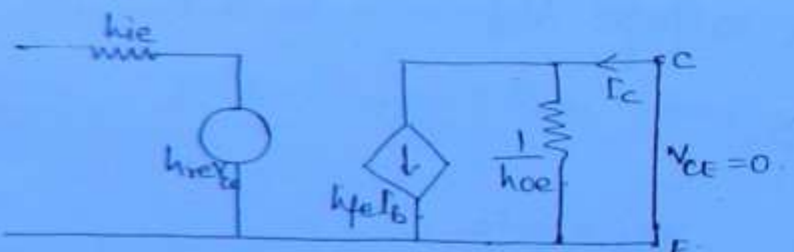
High freq. model to low freq.

$C_{b'e} \rightarrow \infty, C_{b'c} \rightarrow \infty$



$I_c = g_m V_{b'e}$

low freq. model :-



$I_c = h_{fe} I_b$

$$\boxed{\frac{I_c}{I_b} = h_{fe}}$$

High to low $\rightarrow$

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

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

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

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

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

Conclusion . . .

- As I_c increases, $r_{b'e}$ value decreases
- As temp increases, $r_{b'e}$ value increases.

$r_{b'c}$ →

163

O/P parameter

$I_b = 0$

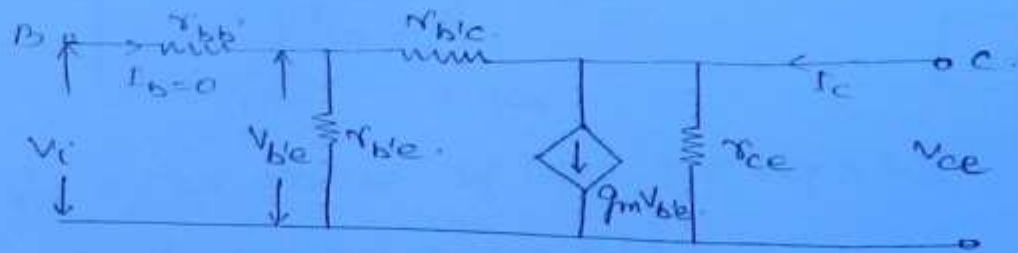
low freq model →

$V_i = h_{ie} I_b + h_{re} V_{ce}$.

$I_b = 0$,

$V_i = h_{re} V_{ce}$.

high freq model



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

$$V_{b'e} = V_i = h_{re} V_{ce}.$$

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

$r_{b'e}$ is very small.

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

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

Conclusions :—

1) As I_c current ↑, $r_{b'c}$ decrease.

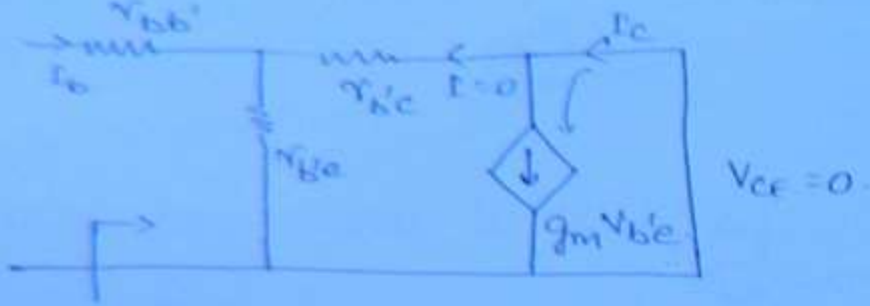
2) As temp increases, $r_{b'c}$ increases.

$r_{bb'}$

I/P parameter

$\boxed{V_{CE} = 0}$

(164)



$Z_i = r_{bb'} + r_{b'e}$.

low freq. model →

$Z_i = h_{ie}$

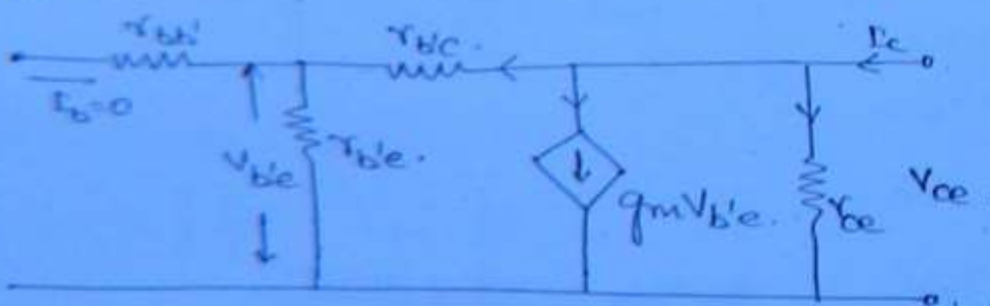
$h_{ie} = r_{bb'} + r_{b'e}$

$\Rightarrow \boxed{r_{bb'} = h_{ie} - r_{b'e}}$

r_{ce} : →

O/P parameter

$I_b = 0$.



low freq →

$$h_{oe} = \frac{I_c}{V_{ce}}$$

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

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

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

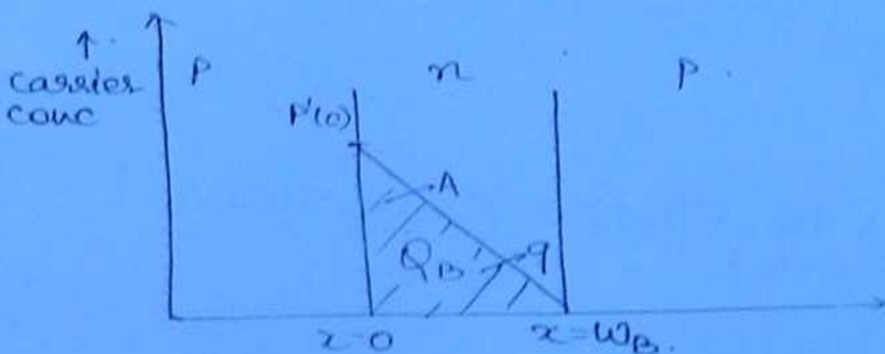
$$h_{oe} = g_{ce} + h_{fe}\, g_{b'c}$$

165

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

$$\boxed{r_{ce} = \frac{1}{g_{ce}}}$$

$C_{b'e} \rightarrow$



$$\therefore Q_B = \text{av. value of conc.} \times A \times q \times W_B$$

$$= \frac{1}{2} p'(0) \times A \times q \times W_B$$

$$I = -AqD_B \frac{dp(x)}{dx}$$

$$= -\frac{AqD_B\{p'(0)-0\}}{0-W_B}$$

$$I = AqD_B \frac{p'(0)}{W_B}$$

$$\therefore Q_B = \frac{1}{2D_B} I W_B^2$$

$$C_{b'e} = \frac{dQ_B}{dV} = \frac{1}{2D_B} W_B^2 \left(\frac{dI}{dV}\right) \leftarrow g_m$$

$$\boxed{C_{b'e} = \frac{1}{2D_B} W_B^2\, g_m} \quad \boxed{C_{b'e} = \frac{1}{2D_B} W_B^2 \frac{1}{r_e}}$$

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

$C_{b'c}$:→

transition capacitance.

$C_T \propto \frac{1}{\sqrt{V_R}}$ → Alloy type.

$C_T \propto \frac{1}{\sqrt[3]{V_R}}$ → Grown j^n.

$V_R \to V_{CE}$

$C_{b'c} \propto \frac{1}{(V_{CE})^n}$.

$n = \frac{1}{2} \to$ Alloy

$= \frac{1}{3} \to$ grown j^n.

Q1. As $V_{CE}\uparrow$, what happens to diffusion capacitance?

Ans. early effect

$V_{CE}\uparrow$, $W_B\downarrow$, $C_{b'e}\downarrow$

$C_{b'e} = \frac{1}{2D_B} W_B^2 g_m$

In all the hybrid π parameter which parameter is independent of I_C and temp?

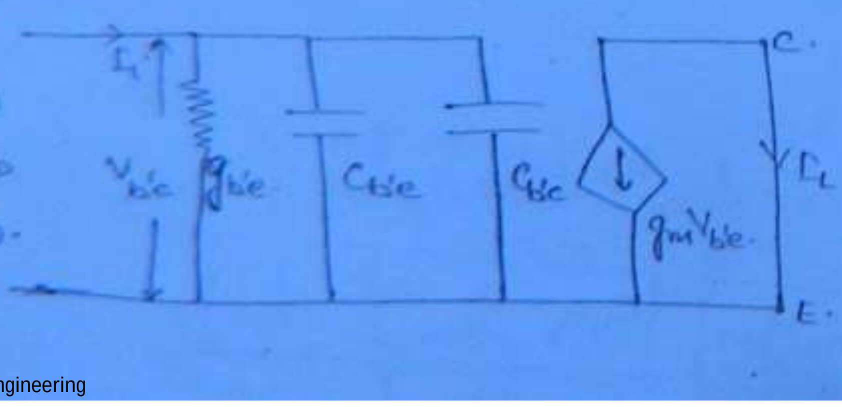
$C_{b'c}$.

Short ckt current gain :→

$r_{bb'} = 0$

$r_{b'c} \to \infty$

$r_{ce} \to 0$.



$$-A_I = \frac{I_L}{I_i}$$

$$= \frac{-g_m V_{b'e}}{V_{b'e}\left(g_{b'e} + j\omega(C_{b'e} + C_{b'c})\right)}$$

167

$$= \frac{-g_m}{\frac{g_m}{h_{fe}} + j2\pi f}$$

$$f_T = \frac{1}{2\pi RC} \qquad \frac{1}{R} \to g_m$$

$$= \frac{g_m}{2\pi C_{b'e}}$$

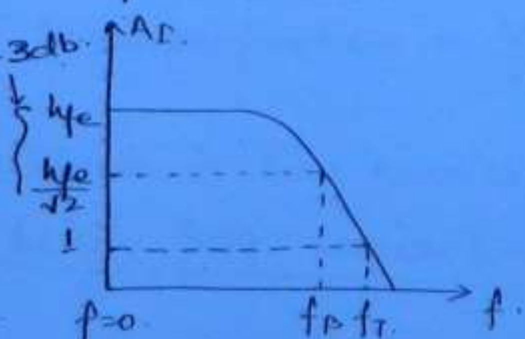
$$= \frac{-g_m}{\frac{g_m}{h_{fe}} + j2\pi f\frac{g_m}{2\pi f_T}}$$

$$= \frac{-g_m}{\frac{g_m}{h_{fe}}\left[1 + j\frac{f}{(f_T/h_{fe})}\right]}$$

$$\therefore f_\beta = \frac{f_T}{h_{fe}}$$

$$\boxed{A_I = \frac{-h_{fe}}{1 + j(f/f_\beta)}}$$

freq. Response :→



$$A_I = \frac{-h_{fe}}{1 + j(f/f_\beta)}$$

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

Beta cut off freq :—
f_β →

It is the freq. at which C.E. short ckt current gain reduces by 3dB of its value

Unity gain band width freq(f_T) :—

It is the freq. at which CE short ckt current gain reduces to unity

f_α $\boxed{f_T = h_{fe} f_\beta}$

α cut off freq :—

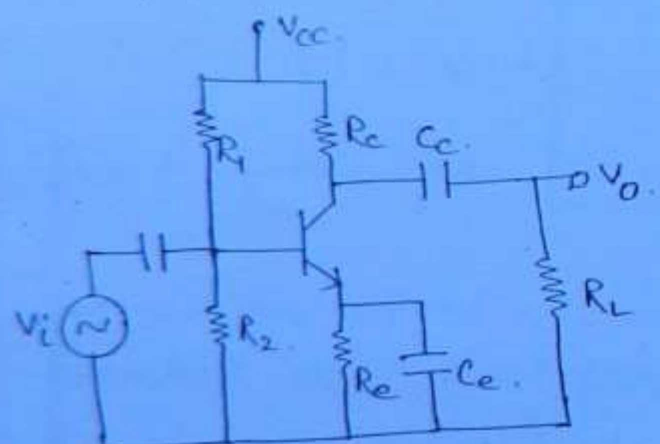
It is the freq. at which CB short ckt current gain reduces by 3dB of its value.

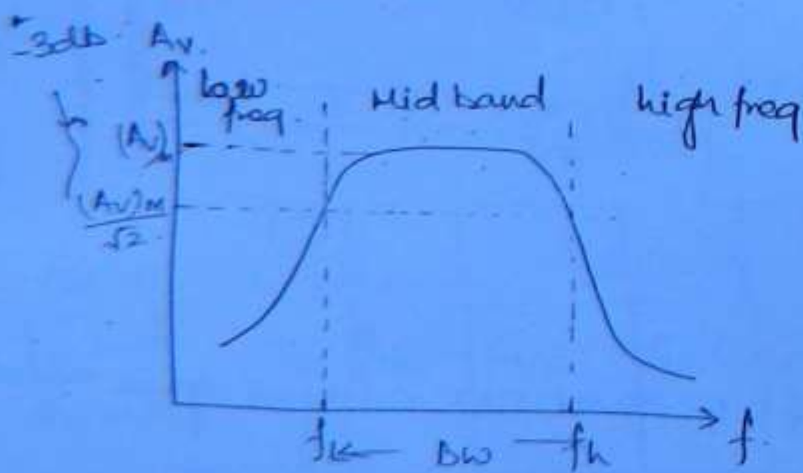
$$\boxed{\begin{aligned} f_\alpha &= (1+\beta) f_\beta \\ &= \frac{1}{1-\alpha} f_\beta \end{aligned}}$$

CB is used for high freq. than CE as bandwidth is higher.

Frequency response of an amplifier :— BJT

CE amplifier →





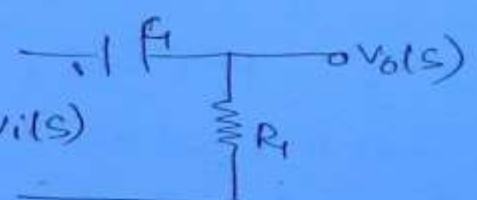
Capacitance →

Internal Capacitance
- → C_D (Diffusion Capacitance)
- → C_T (transition Capacitance)

External Capacitance
- → C_c (coupling)
- → C_E — (by pass)

Low freq region →

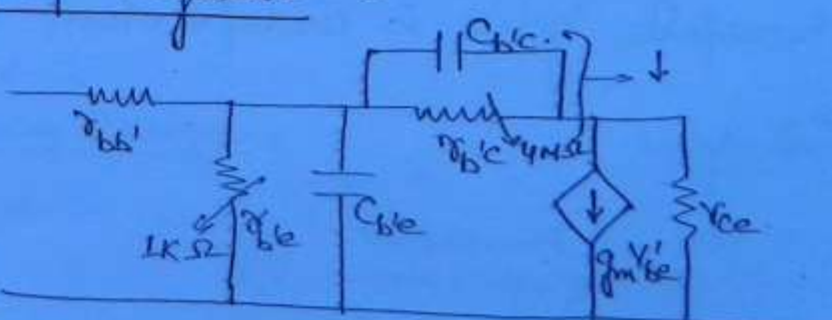
- At low freq. Analysis, internal capacitors are neglected which are parallel to the jn of transistors [open]
- If ext. Capacitors are open, the signal will not pass into the transistors that means C_c and C_E are affecting sys. response.
- The System response is considered as high pass response.



$$\text{transfer function} = \frac{V_0(s)}{V_i(s)} = \frac{S}{S + \frac{1}{R_1C_1}}$$

There is one zero and one pole is considered.

High freq. region: →



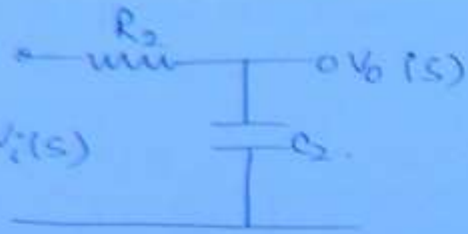
- At high freq. range, ext. capacitors are not affecting the System response because they are short ckt.
- When the int capacitors are short ckt, it will affec

The '$r_{b'e}$' and '$C_{b'c}$' value.

That means int. Capacitors are affecting the sys. response.

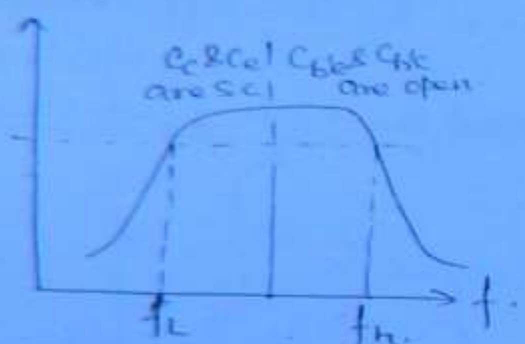
If the system response is considered as low pass response

(170)



$$T.F. = \frac{V_0(s)}{V_i(s)} = \frac{1}{s + \frac{1}{R_2 C_2}}.$$

Mid band Range :→



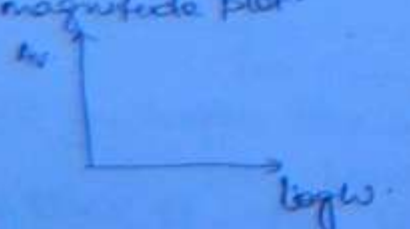
At low frequencies, mid band Range freq. is considered as high freq, therefore ext. capacitors are neglected. At high freq. range mid band range freq. are considered as low freq. Therefore int. capacitors are neglected.

• In the mid band range, all the internal and ext. capacitors are neglected. Therefore gain is independent of freq.

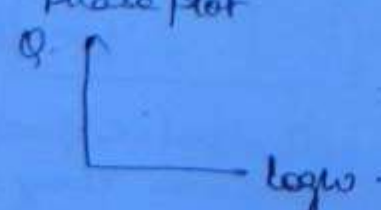
-3dB concept :→

Bode plot →

magnitude plot.



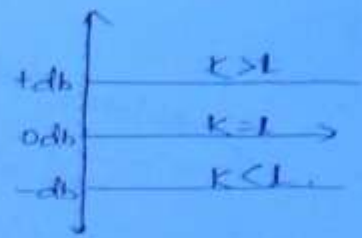
Phase plot



$$T.F = \frac{KS}{1+ST}$$

$G(s) = K$

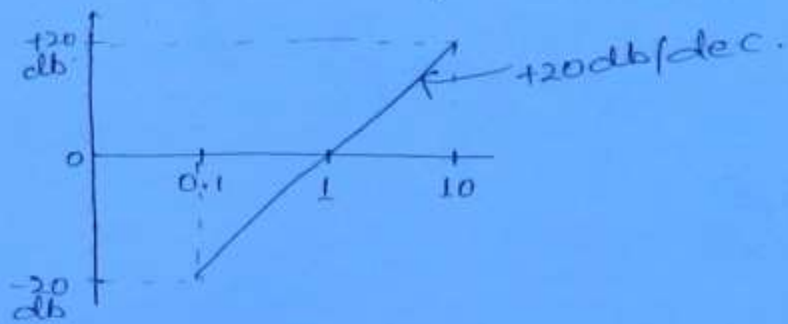
$|G(j\omega)|_{db} = 20 \log K$



171

$\Rightarrow G(s) = S$

$|G(j\omega)|_{db} = 20 \log \omega$



$$\Rightarrow G(s) = \frac{1}{1+ST}$$

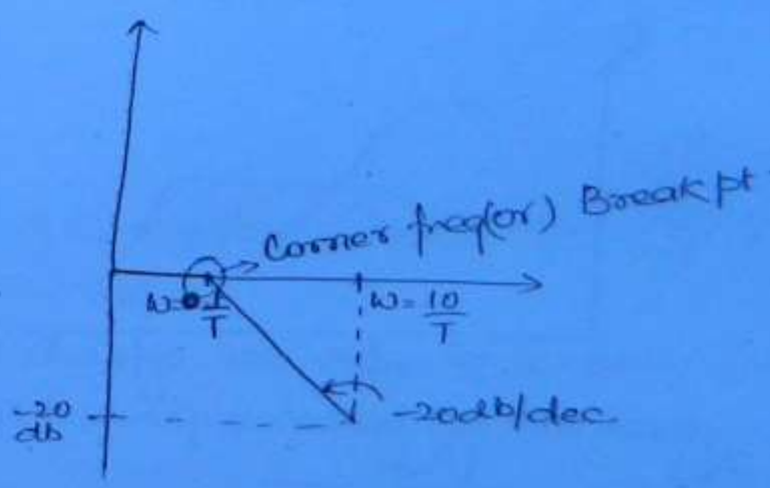
$|G(j\omega)|_{db} = -20 \log \sqrt{1+\omega^2 T^2}$

$\omega T << 1$, low freq.

$|G(j\omega)| = -20 \log 1 = 0db$

$\omega T >> 1$, high freq.

$|G(j\omega)| = -20 \log \omega T$



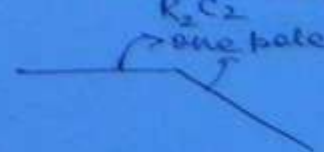
Bode plot →

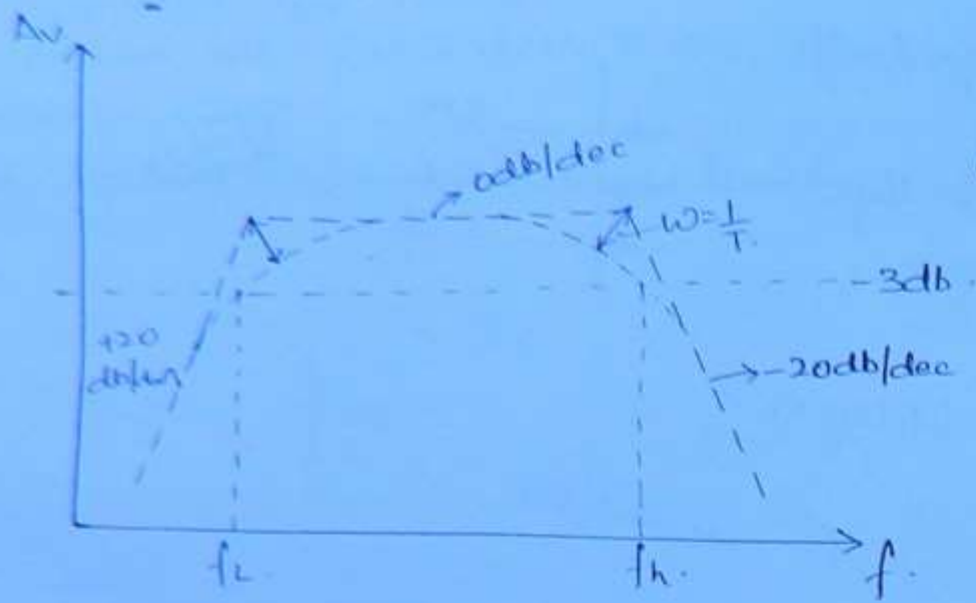
Low freq Range →

$$T.F. = \frac{S \rightarrow \text{one } 0}{S + \frac{1}{R_1 C_1} \rightarrow \text{one pole}}$$

High freq. Range :—

$$T.F. = \frac{1}{S + \frac{1}{R_2 C_2}} \rightarrow \text{one pole}$$





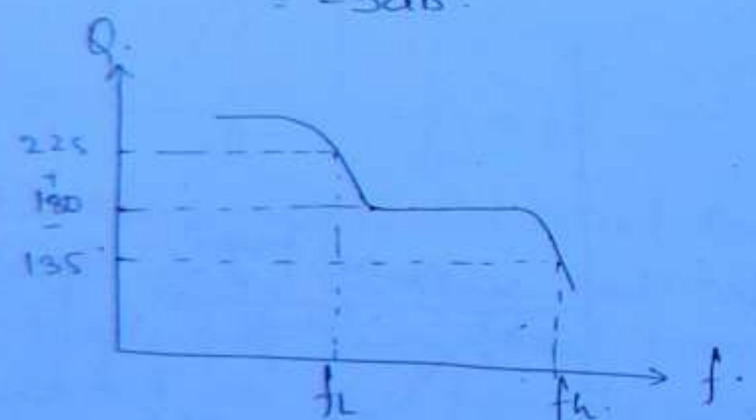
$$|G(j\omega)|_{db} = -20\log\sqrt{1+\omega^2T^2}$$

$$\omega = \frac{1}{T}$$

$$= -20\log\sqrt{1+1}$$

$$= -20\log\sqrt{2}$$

$$= -3db$$



$$(A_v)_L = \frac{(A_v)_m}{1-j\left(\frac{f_L}{f}\right)}$$

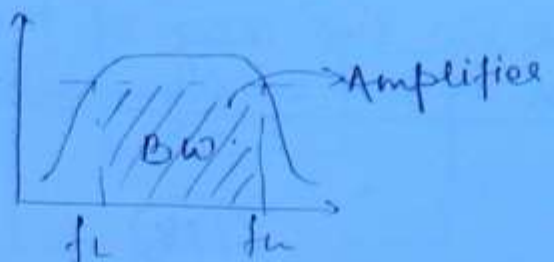
$$(A_v)_H = \frac{(A_v)_m}{1+j\left(f/f_H\right)}$$

$$\theta_L = \tan^{-1}\left(\frac{f_L}{f}\right)$$

$$\theta_H = \tan^{-1}\left(f/f_H\right)$$

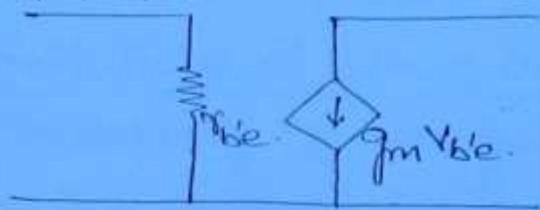
<u>Conclusion</u> →

Concept of Amplifier problems.



(123)

Amplifier →



$C_{b'e} \to \infty$

$C_{b'c} \to \infty$

$r_{bb'} \to 0$

$r_{b'c} \to \infty$

$r_{ce} \to \infty$

Ch-2

Ch-2		
Q-4. Q-4	160	(d).

Pg-34

Q10 b

12/1/12.

Feedback Amplifier :—

→ Amplier The basic char. of an amplifier are input imp., O/P imp., Voltage gain, BW. etc.

→ Suppose if we want to change the char. of a basic amplifier, we have to use a technique called as Feedback.

Feedback →

"When a part or a fraction of the O/P feedback to I/P, feedback is said to be exist."

It classified into two ways :—

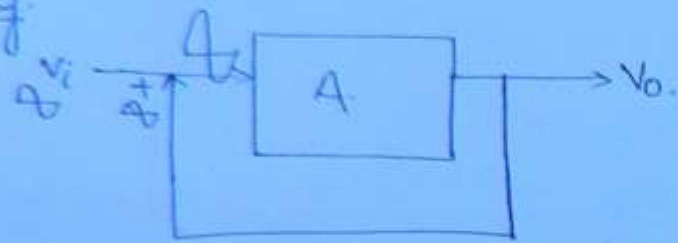
1) +ive feedback

2) -ive feedback.

Positive feedback :→

If the net effect of the feedback inc. the mag. of the I/P signal, It is called as +ive or direct or regenerative feedback.

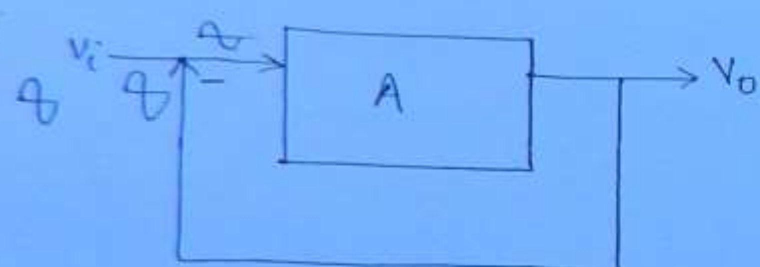
eg.



negative feedback :→

If the net effect of the feedback decreases the mag. of the I/P signal, It is called as -ive or Inverse or degenerative feedback.

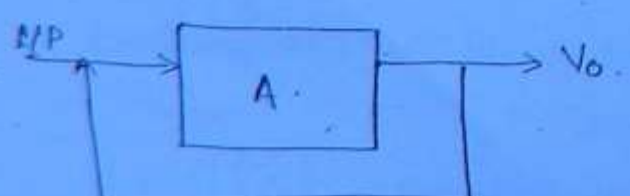
eg



Feedback can also be classified as

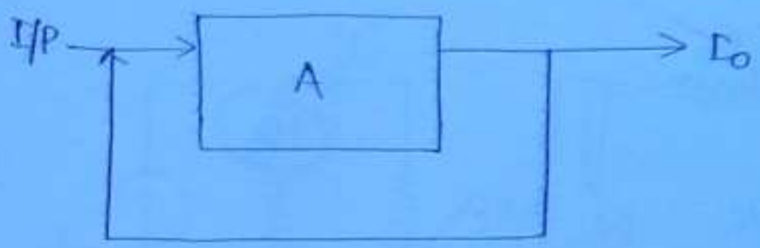
1) voltage feedback.
2) current feedback.

Voltage feedback :—



feedback signal $\propto V_o$.

Current feedback →



(125)

feedback signal $\propto I_o$.

Limitations of basic amplifiers :→

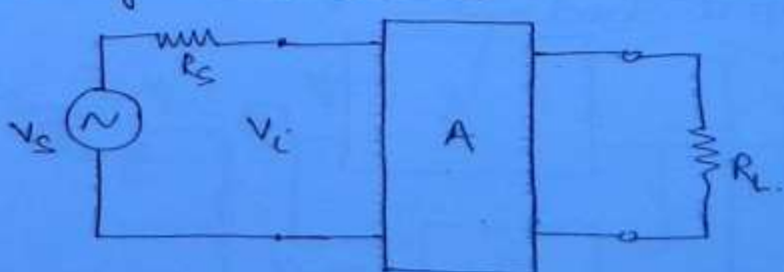
1) Instability of ac gain.
   - Due to power supply variation
   - Due to change in h parameter.
   - age of the device.
2) I/P Impedance is low.
3) Output Impedance is high.
4) Noise level is high.
5) Distortion is more
6) BW is not sufficiently large.

Types of basic amplifiers →

based on Z_i and Z_o →

1) Voltage amplifier
2) Current amplifier.
3) Transconductance amplifiers.
4) Transresistance amplifiers.

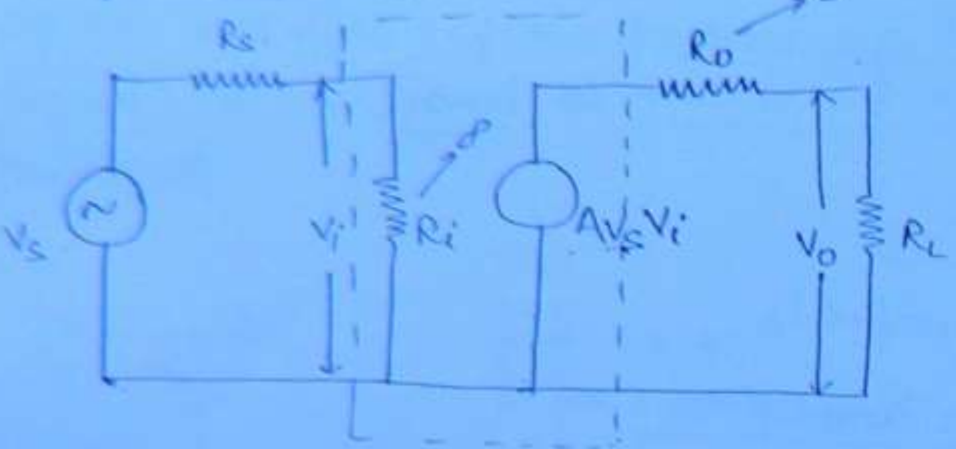
Voltage amplifiers :→



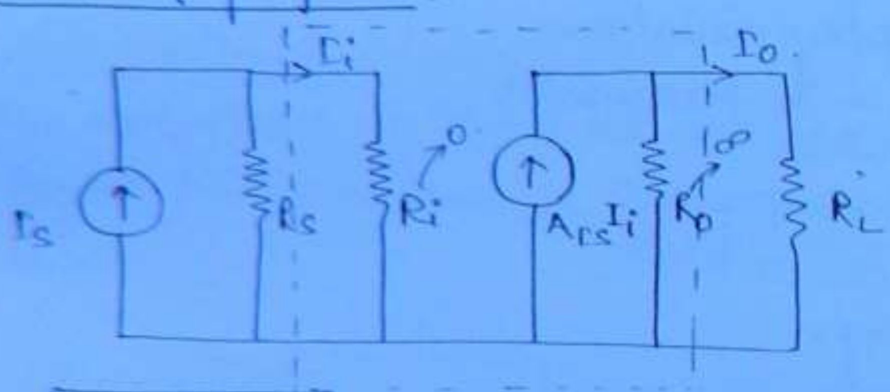
Voltage gain $A_V = \frac{V_o}{V_i}$. Voltage Amplification $A_{VS} = \frac{V_o}{V_s}$

$\boxed{V_o = A_{VS} V_s}$

126

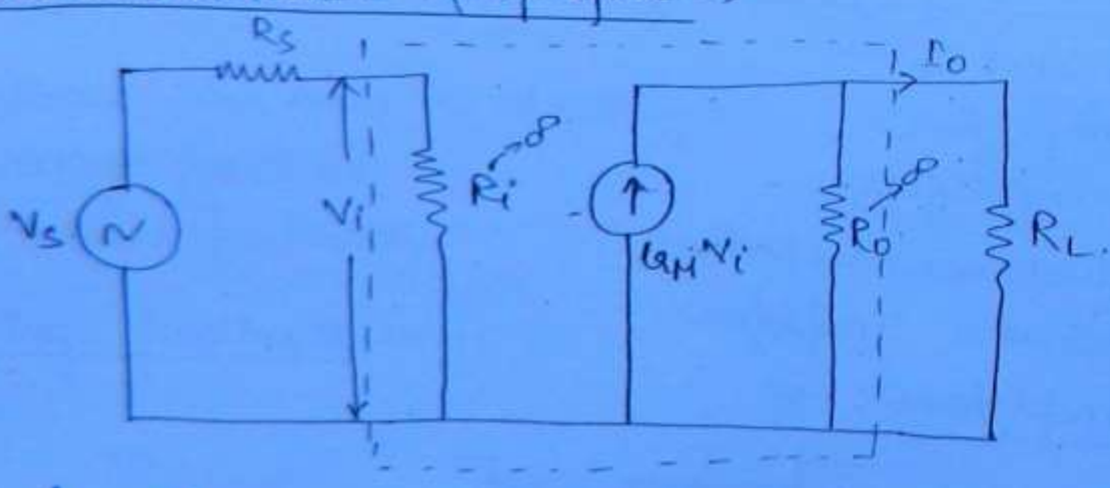


Current Amplifier →



$\boxed{I_o = A_{IS} I_s}$

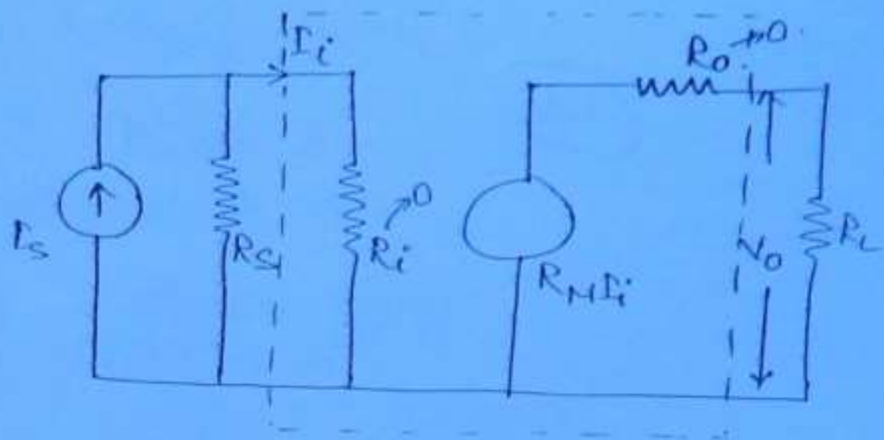
Transconductance Amplifier :→



$$G_M = \frac{I_o}{V_s}$$

$$I_o = G_M V_s.$$

Transresistance amplifier :—

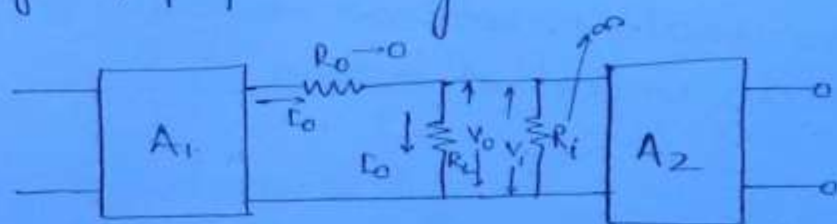


$R_M = \frac{V_o}{I_s}$

$\boxed{V_o = R_M I_s}$

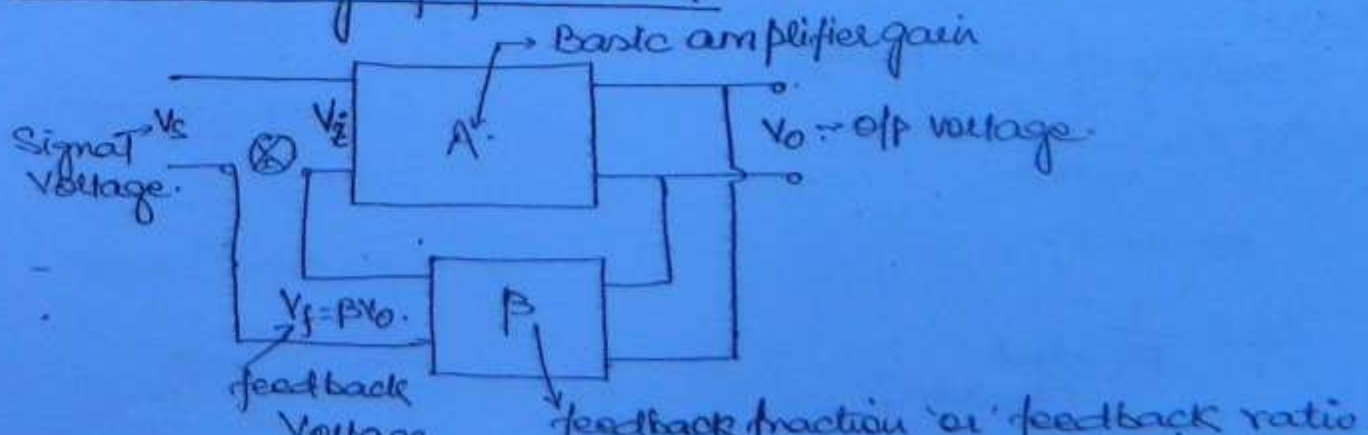
#Q.) In most of the practical applications why we choose Voltage Amplifiers only?

Ans.



For any ideal amplifier, I/P impedance should be ∞ and O/P impedance should be 0.
Therefore voltage amplifier specification are matching with ideal amplifier specification. That is the reason, most of the practical applications, voltage amplifiers are used.

General theory of feedback: →



Positive FB

$V_o = AV_i$

$V_i = V_s + V_f$

$V_o = A(V_s + \beta V_o)$

$\Rightarrow \frac{V_o}{V_s} = \frac{A}{1-A\beta}$

(178)

Negative FB :-

$V_o = AV_i$

$V_i = V_s - V_f$

$V_o = A(V_s - \beta V_o)$

$\frac{V_o}{V_s} = \frac{A}{1+A\beta}$

Conclusions →

> $A_{pf} > A > A_{nf}$

> $A_{nf} = \frac{V_o}{V_s} = \frac{A}{1+A\beta}$

$A_{nf} \downarrow, S \uparrow$

Always for amplifier analysis we use negative feedback because stability is an important criteria.

3). $A_{pf} = \frac{V_o}{V_s} = \frac{A}{1-A\beta}$

$|A\beta| = 1.$

$A_{pf} \to \infty, S \to 0.$

+ive feedback is always be applied for unstable sys' like oscillators.

Negative FB amplifiers →

Adv →

1) Stability of ac gain.

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

$A\beta >> 1.$

$A_f = \frac{1}{\beta} \to$ fixed (Resistor)

resistance of an amplifier then the fractional change of gain with feedback is.

(179)

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

diff. w.r.t. A.

$$\frac{\partial A_f}{\partial A} = \frac{(1+A\beta) - A(\beta)}{(1+A\beta)^2}$$

$$= \frac{1}{(1+A\beta)^2}$$

$$\Rightarrow \frac{\partial A_f}{A_f} = \frac{1}{(1+A\beta)^2} \frac{\partial A}{A}$$

$\frac{\partial A}{A} \rightarrow$ w/o Feedback fractional change of gain.

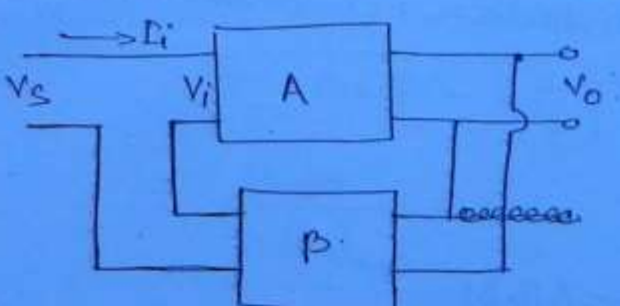
$\frac{\partial A_f}{A_f} \rightarrow$ with Feedback fractional change of gain

$$\boxed{\frac{\partial A_f}{A_f} < \frac{\partial A}{A}}$$

$\Rightarrow \frac{1}{1+A\beta} \rightarrow$ Sensitivity

$\Rightarrow 1+A\beta \rightarrow$ Desensitivity

<u>Increase in I/P impedance</u> $\rightarrow$ (Z_i)



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

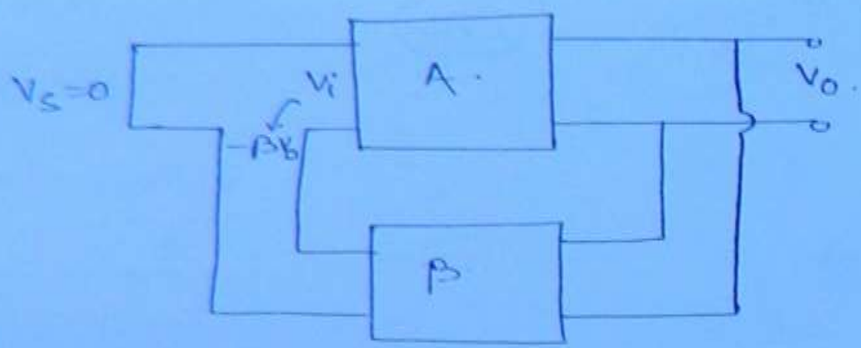
$$Z_i = \frac{V_i}{I_i}$$

$$V_i = V_s - \beta V_o$$

$$= V_s - \beta A V_i$$

$$(1+A\beta)\frac{V_i}{I_i} = \frac{V_s}{I_i}$$

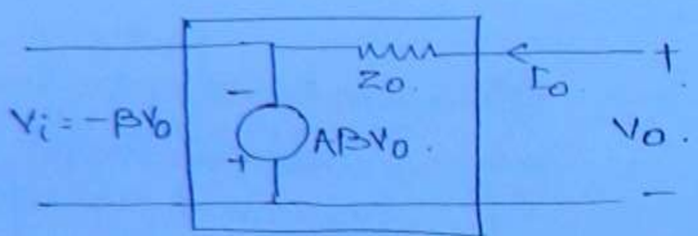
$$\boxed{(1+A\beta) Z_i = Z_{if}}$$



$V_s =$

$V_i = V_s - \beta V_o$, $V_s = 0$

$$\boxed{V_i = -\beta V_o}$$



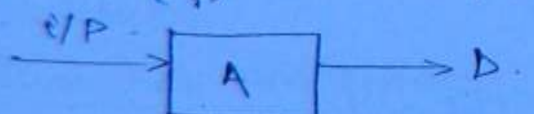
$$V_o + A\beta V_o = I_o Z_o$$

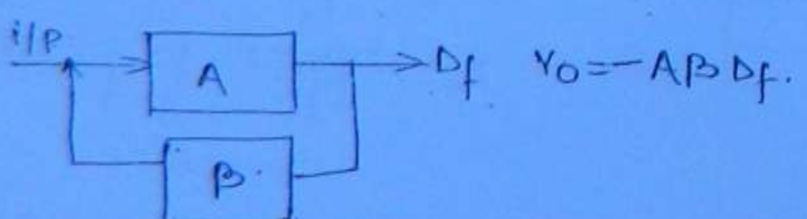
$$V_o(1 + A\beta) = I_o Z_o$$

$$\boxed{\frac{V_o}{I_o} = Z_{of} = \frac{Z_o}{1 + A\beta}}$$

Reduction in distortion and noise :→

Net distortion (D_f) = Original distortion (D) + distorted O/P.





$V_o = -A\beta D_f$.

$$D_f = D - A\beta D_f$$

$$D_f(1 + A\beta) = D$$

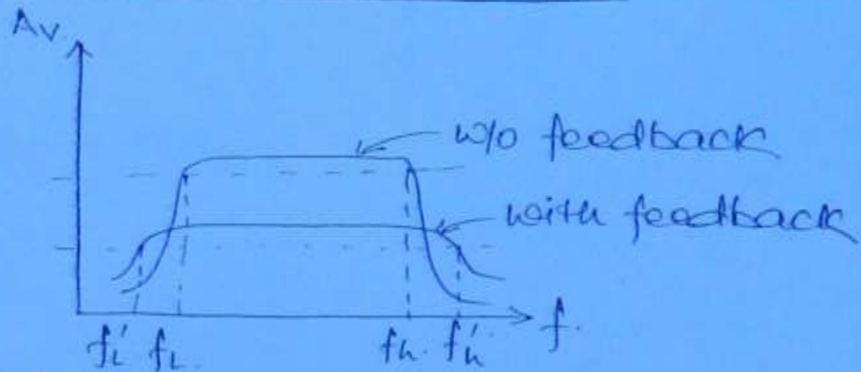
$$\boxed{D_f = \frac{D}{1 + A\beta}}$$

Similarly,

$$N_f = \frac{N}{1+A\beta}$$

(8)

Increase in Bandwidth :→



lower cutoff freq →

$$(A_v)_l = \frac{(A_v)_m}{1-j(f_L/f)}$$

$$= \frac{(A_v)_l}{1+(A_v)_l\beta}$$

$$= \frac{(A_v)_m}{1-j(f_L/f)} \Bigg/ 1+\frac{(A_v)_m\,\beta}{1-j\left(\frac{f_L}{f}\right)}$$

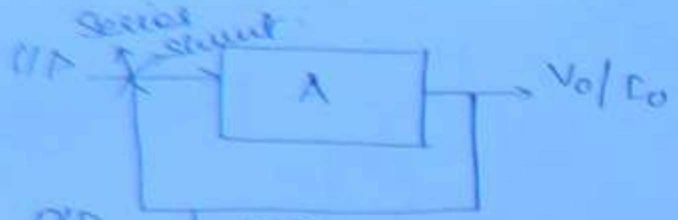
$$(A_v)_{lf} = \frac{A_{vm}}{1+A_{vm}\beta - j\left(\frac{f_L}{f}\right)}$$

$$= \frac{\dfrac{(A_v)_m}{1+(A_v)_m\beta}}{1-j\dfrac{f_L}{f(1+(A_v)_m\beta)}}$$

$$= \frac{(A_v)_{mf}}{1-j(f_L'/f)}$$

$$f_L' = \frac{f_L}{1+(A_v)_m\beta} \Rightarrow f_h' = f_h(1+(A_v)_m\beta)$$

Topology :→

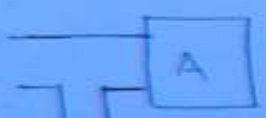


O/P	I/P
Voltage	Series
Voltage	Shunt
Current	Series
Current	Shunt

Topology →
- → Block diagram Analysis.
- → Practical ckt Analysis.

Block diagram Analysis :→

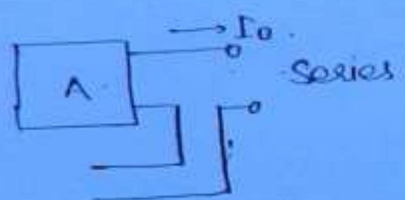
i/P.

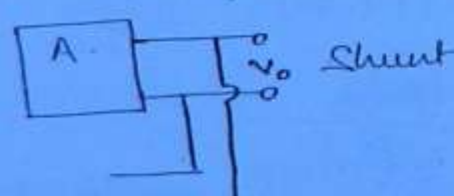


⇒ series

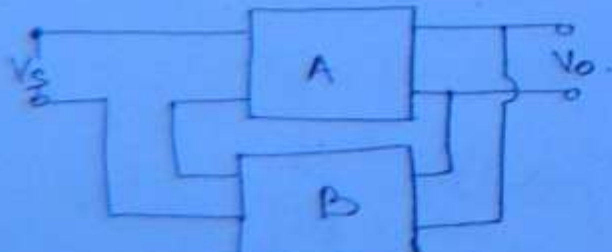
⇒ Shunt

O/P Side.



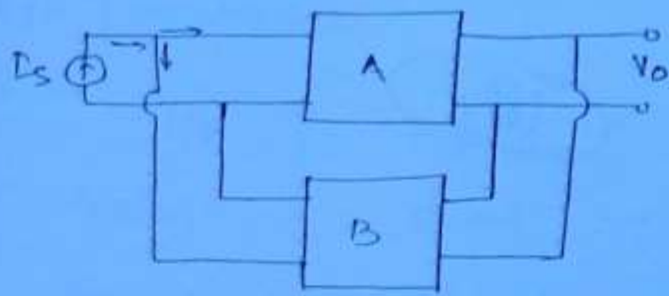


Voltage Series →



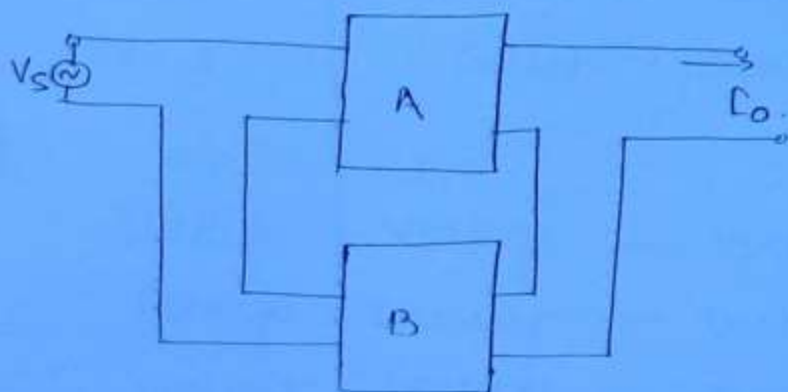
→ Voltage amplifier.

Voltage shunt



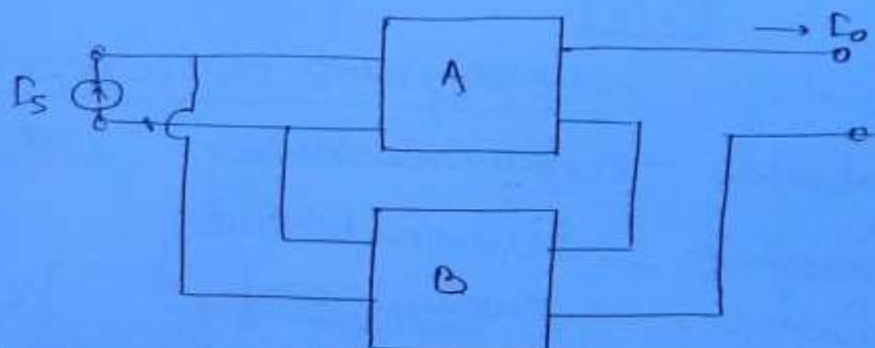
→ transresistance amplifier

Current Series



→ transconductance amplifier.

Current Shunt



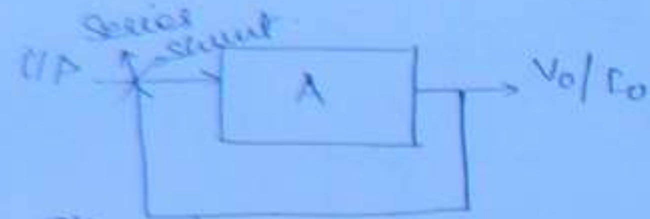
→ current amplifier

I
(1) In Voltage Series ________ amplifier is used.
2) In Voltage shunt ________ amplifier is used.
3) In current series ________ amplifier is used.
4) In current Shunt ________ amplifier is used.

II
1) In I/P is series connection Voltage source is taken
2) In I/P is Shunt Connection Current source is taken

III
1) +ve VCVS
NCIS
ICVS

Topology :→



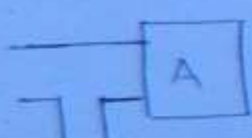
O/P	I/P
Voltage	Series
Voltage	Shunt
Current	Series
Current	Shunt

Topology →

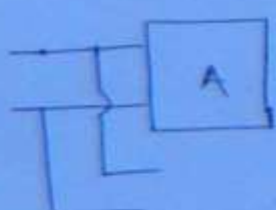
- → Block diagram Analysis.
- → Practical ckt Analysis.

Block diagram Analysis :→

I/P

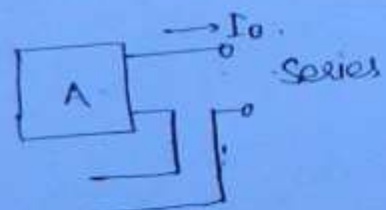


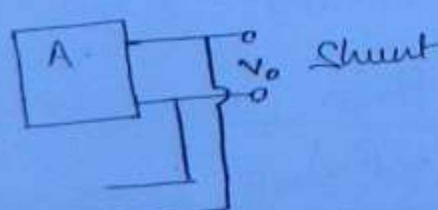
⇒ series



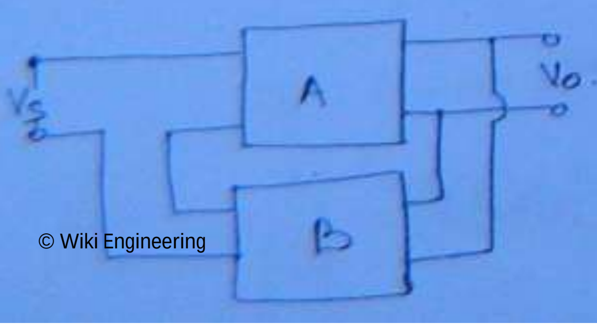
⇒ Shunt

O/P side.



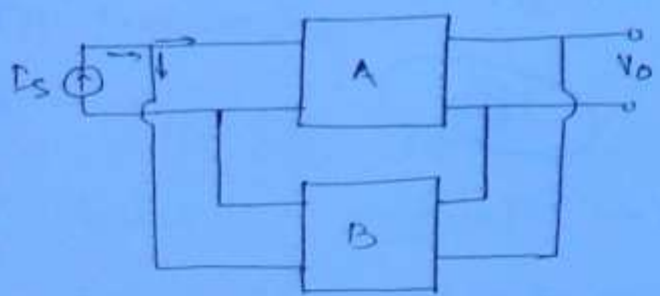


Voltage series →



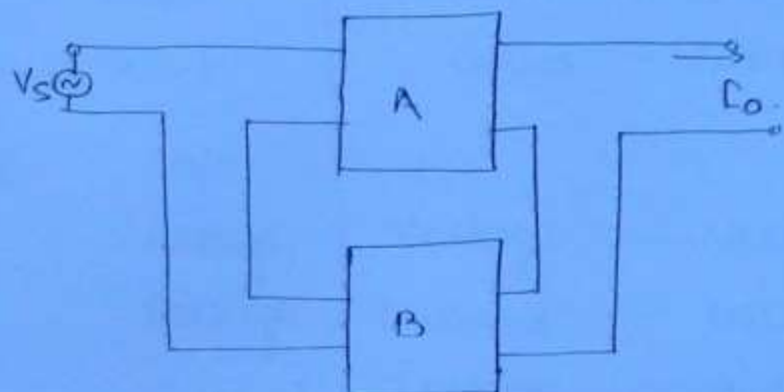
→ Voltage amplifier.

Voltage shunt



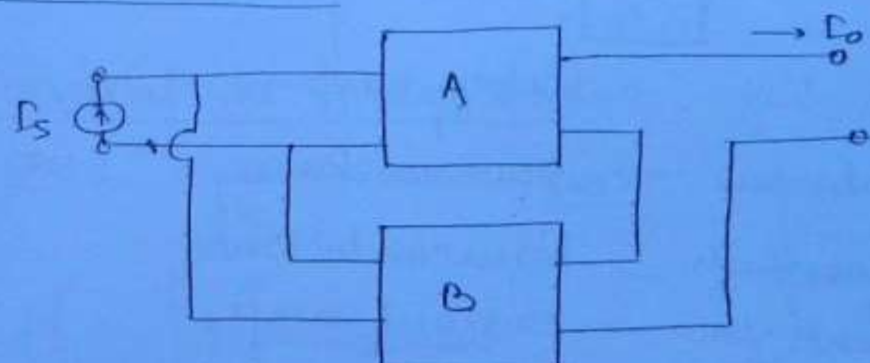
→ transresistance amplifier

Current Series



→ transconductance amplifier.

Current Shunt



→ current amplifier

I

1) In Voltage Series ____________ amplifier is used.
2) In Voltage Shunt ____________ amplifies is used.
3) In Current series ____________ amplifier is used.
4) In current Shunt ____________ amplifier is used.

II

1) In I/P is series connection <u>Voltage</u> source is taken
2) In I/P is Shunt Connection <u>Current</u> source is taken

III

1) ~~Vc~~ V C V S
N C I S
I C V S

1<1S

IV

Series shunt

Shunt Shunt

Series Series

Shunt Series

V Voltage Voltage

Voltage current

Current Voltage

Current current.

184.

Answers →

I

Z_o	Z_i	$\boxed{Z_i Z_o}$ →
low ← Voltage 0	Series → high ∞	→ voltage amp.
low ← Voltage 0	Shunt → low 0	→ transresistance
high ← current ∞	Series → high ∞	→ transconductance
high ← Current ∞	Shunt → low 0	→ current amp.

Opamp is a voltage control device → $V_o \propto V_d$

FET is a voltage control device → $I_d \propto V_{GS}$

BJT is a current control device → $I_C \propto I_b$

III

I/P	O/P	
Series V C	V S	→ ser voltage series
Series V C	I S	→ current series.
Shunt I C.	V S	→ voltage shunt
Shunt I C	I S	→ Current Shunt.

Any device [illegible] of the I/P parameter.

IV

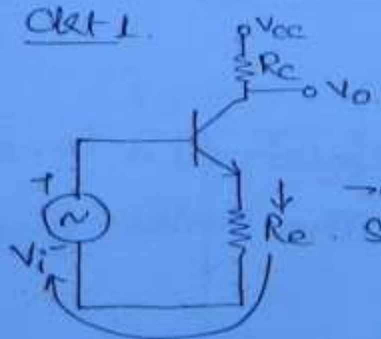
I/P	O/P	
Series	Shunt	→ Voltage Series
Shunt	Shunt	→ Voltage Shunt
Series	Series	→ Current Series
Shunt	Series	→ Current Shunt

V

O/P	I/P	
Voltage	Voltage	→ Voltage Series
Voltage	Current	→ Voltage Shunt
Current	Voltage	→ Current Series
Current	Current	→ Current Shunt

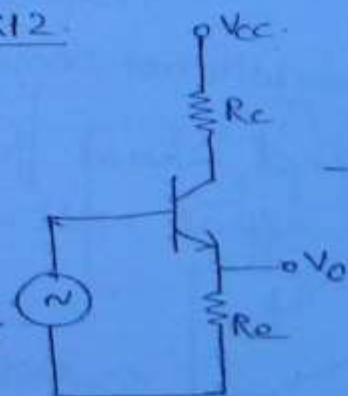
Practical Ckt Analysis :→

Ckt 1.



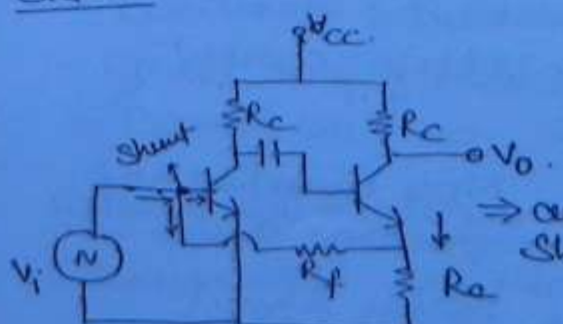
→ current Series.

Ckt 2.



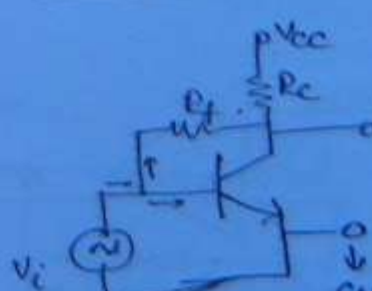
– Voltage Series

Ckt 3.



⇒ current Shunt

Ckt 4.

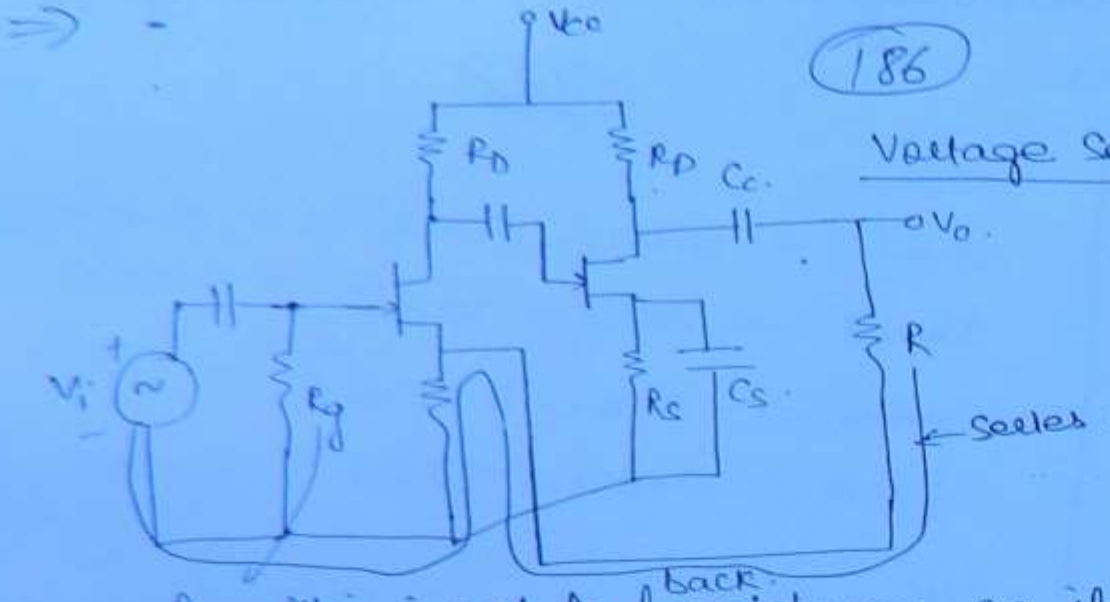


⇒ Voltage Shunt.

current Shunt.

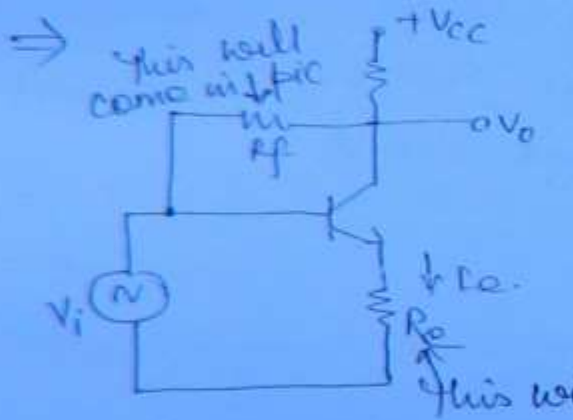
⇒ -

Voltage Series.



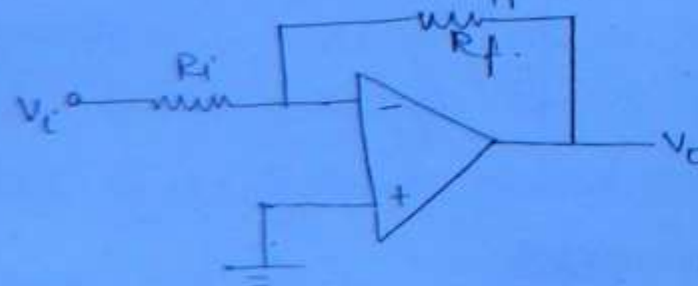
Rg → this is not feedback resistance. so, it is series connection.

⇒

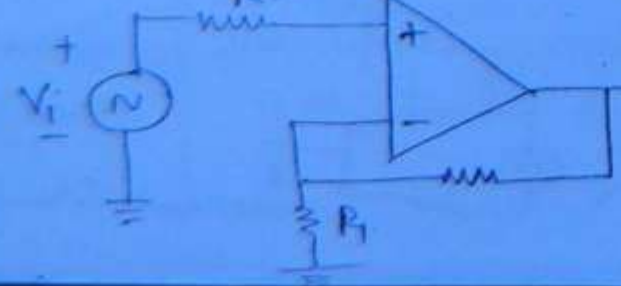


⇒ Voltage shunt

Whenever series and shunt feedback exist at a time in the ckt, shunt effect will dominate the series effect.



inverting amplifier.
Voltage shunt
Shunt Shunt
Voltage current



Non inverting amp.
Voltage series
Series shunt
Voltage Voltage.

Calculation of i/p impedance and o/p impedance of diff. topology :->

	Voltage Series ($Z_o = 0$, $Z_i = \infty$)	Voltage Shunt ($Z_o = 0$, $Z_i = 0$)	Current Series ($Z_o = \infty$, $Z_i = \infty$)	Current Sh. ($Z_o = \infty$, Z_i)
Z_{if}	$Z_i(1+A\beta)$	$Z_i/1+A\beta$	$Z_i(1+A\beta)$	$Z_i/1+A\beta$
Z_{of}	$Z_o/1+A\beta$	$Z_o/1+A\beta$	$Z_o(1+A\beta)$	$Z_o(1+A\beta)$

OSCILLATORS ->

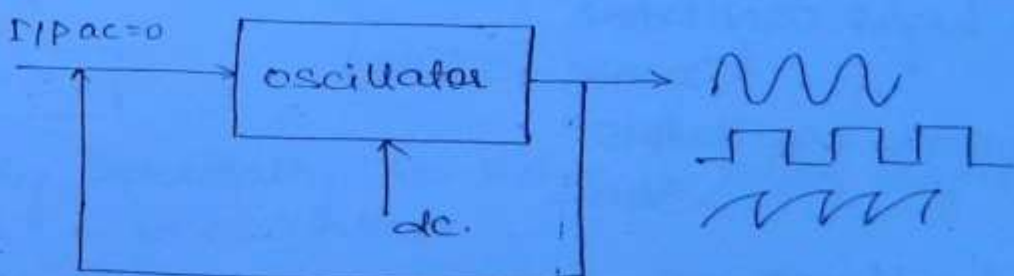
Definitions ->

-> It is a ckt which converts dc en. into ac en. at any required freq.

-> It is an electronic ckt of alternating current and voltage having sine, square, sawtooth waveform etc.

-> It is a ckt which generates ac o/p signal w/o requiring any ac I/P signal.

-> It is an unstable amplifier.



Classification of oscillators ->

- Based on nature of wave generated
- Based on fundamental mechanism involved.
- Based on feedback
- Based on freq. response.

Based on nature of wave generated ->

-> Sinusoidal oscillations
eg tunnel diode, RC, LC oscillation.

eg - UJT, Astable multivibrator using opamp.

Based on fundamental mechanism involved :-

- Negative resistance region
  - tunnel diode, UJT.
- feedback
  - RC oscillator
  - LC oscillator.

Based on freq. response :-

- Audio freq oscillators (20Hz - 20kHz)
- Radio freq oscillators. (30kHz to 30MHz)
- Very high freq. oscillators. (30MHz - 300MHz).
- Ultra high oscillators (300MHz - 3 GHz)
- Microwave oscillators > 3GHz

Based on feedback →

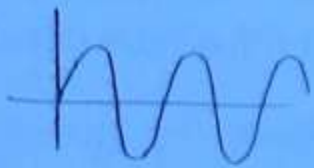
→ RC oscillators
- RC phase shift oscillators
- Wein bridge oscillators.

→ LC oscillators.
- Hartley
- Colpitts
- Clapp

→ Crystal oscillators

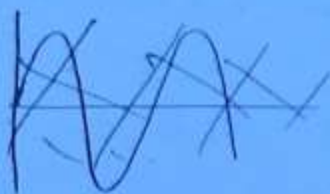
Effect of $|A\beta|$ on oscillations :→

$|A\beta| = 1$.



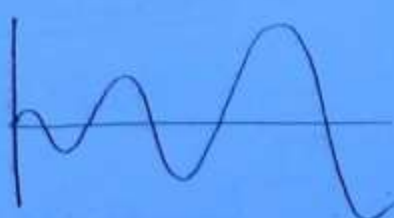
→ Undamped

$|A\beta| < 1$.



→ Underdamped.

$|A\beta| > 1$



→ overdamped.

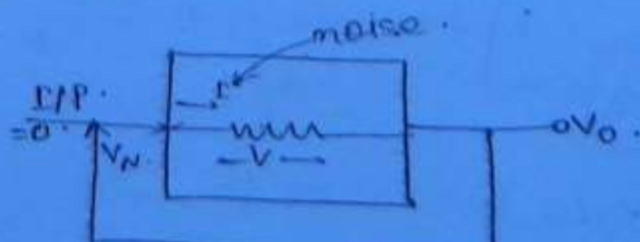
$\boxed{A\beta \geq 1}$

Q. Why practically $A\beta$ value is taken slightly greater than 1. in oscillators.

Ans. Because of the non linearities in the ckt, we take $A\beta$ slightly greater than 1.

Q. W/o giving any input how an oscillator can produce an O/P?

Ans.

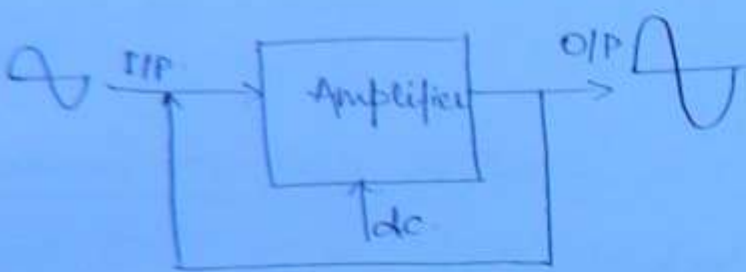


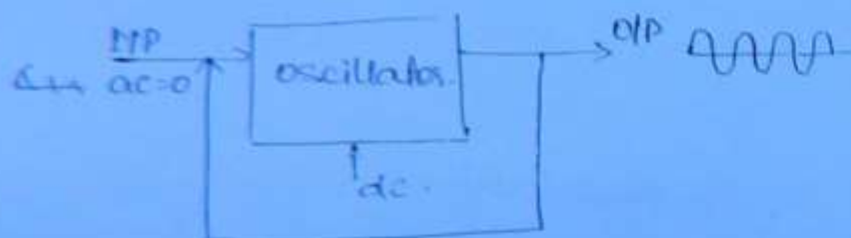
V_N → Noise voltage.

Any electrical circuit will behave like a resistor because of the random moment of e^- in an electronic device a noise voltage will be produce. Such noise voltage will be feedback to the I/P which will be a regenerative action, that means there is no requirement of giving an ext I/P.

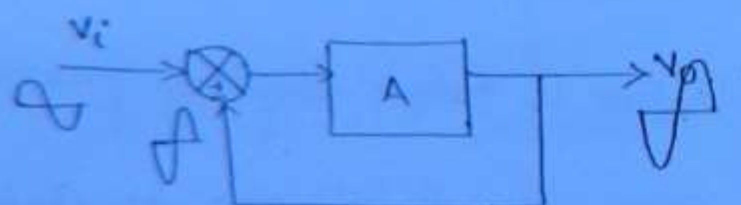
3/1/12

Basic diff b/w Amplifier and oscillator :→ (190)

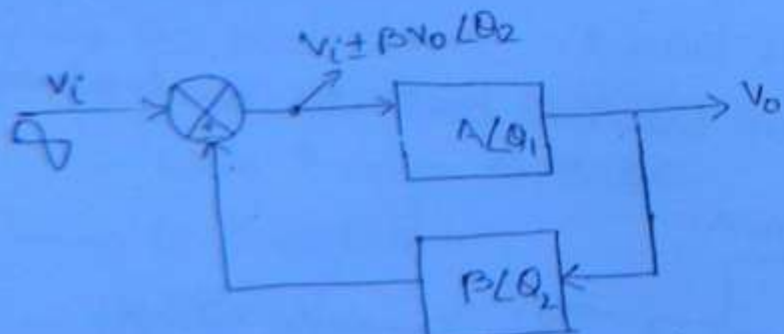




Feedback oscillators —



→ CE give 180° phase shift at O/P for I/P. So phase shift is taken at A and β.



$$V_0 = A\angle\theta_1 \{ V_i \pm \beta V_0 \angle\theta_2 \}$$

for positive feedback.

$$V_o = A\angle\theta_1 \{ V_i + \beta V_o \angle\theta_2 \}$$

$$= A\angle\theta_1 V_i + A\beta V_o \angle(\theta_1+\theta_2)$$

$$\Rightarrow V_o(1 - A\beta\angle(\theta_1+\theta_2)) = A\angle\theta_1 V_i$$

$$\Rightarrow \frac{V_o}{V_i} = \frac{A\angle\theta_1}{1 - A\beta\angle\theta_1+\theta_2}$$

(191)

For Oscillators,

for $V_i = 0$,

$$1 - A\beta\angle\theta_1+\theta_2 = 0.$$

Magnitude

$$|A\beta| = 1$$

Phase,

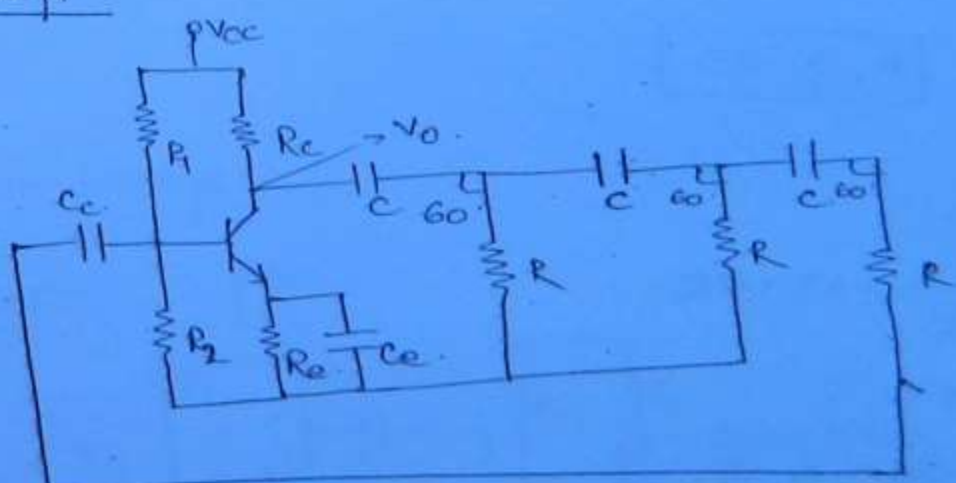
$$\theta_1+\theta_2 = 0 \text{ or multiple of } 2\pi$$

θ_1

→ Barkhausen criteria.

RC oscillators →

Phase shift :—



conclusions :—

1) Q + C R → 90° → ideal.

↓

360

3 sets of RC = 360°

L180°

2> $f = \frac{1}{2\pi RC\sqrt{4k+6}}$

where $k = \frac{R_c}{R}$

192

for $k=1$,

$f = \frac{1}{2\pi RC\sqrt{10}}$

for $k \ll 1$,

$f = \frac{1}{2\pi RC\sqrt{6}}$

3> $-h_{fe}\frac{R_c}{R} = -29 - 23k - 4k^2$

$\downarrow$

$\boxed{A_v = -29}$ $k \ll 1$

$\beta = \frac{+1}{A_v}$ as $|A\beta| = 1$.

$= \frac{1}{-29}$

$A_v \not< -29$ $\boxed{A_v \geq -29}$

4> $(h_{fe})_{min}$

$h_{fe} = \frac{29}{k} + 23 + 4k$.

diff. w.r.t. k.

$\frac{\partial h_{fe}}{\partial k} = \frac{-29}{k^2} + 4 = 0$

$k = 2.7$.

$\boxed{h_{fe\,min} = 44.5}$

Drawbacks of RC phase shift oscillator

for 3 KHz – 20KHz

(193)

$$f = \frac{1}{2\pi RC\sqrt{6+4K}}$$

We cannot make it for dynamic ~~freq~~ frequency. as R and C can't be change as phase is affected

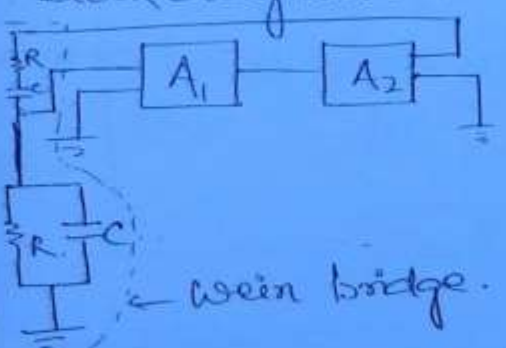
1) RC phase shift oscillator is used for single freq. operation

2) Due to thermal variation Resistance tolerance value may change which will disturb the phase relations

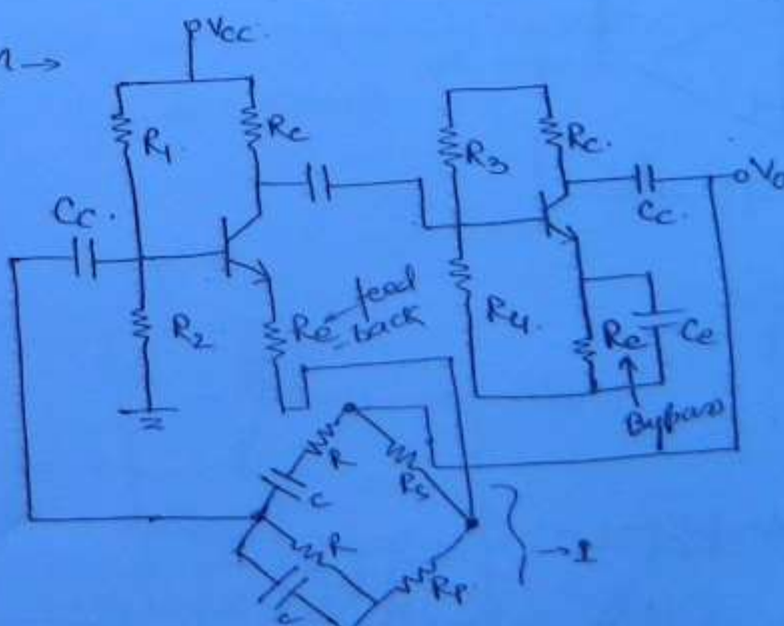
3) This type of ckt can be used for low freq oscillation [audio range] [20 Hz – 20 KHz] because Q-factor of RC n/w is poor

Wien bridge oscillator: →

Block diagram: →



Ckt diagram →



1) $Q_1 + Q_2 +$ Wien bridge
$\downarrow$ $\downarrow$ $\downarrow$
$180°$ $180°$ $0°$ $\rightarrow 360°$

(194)

2) $f = \dfrac{1}{2\pi\sqrt{C_1C_2R_1R_2}}$ → In I.

$R_1 = R_2 = R$, $C_1 = C_2 = C$.

$f = \dfrac{1}{2\pi RC}$.

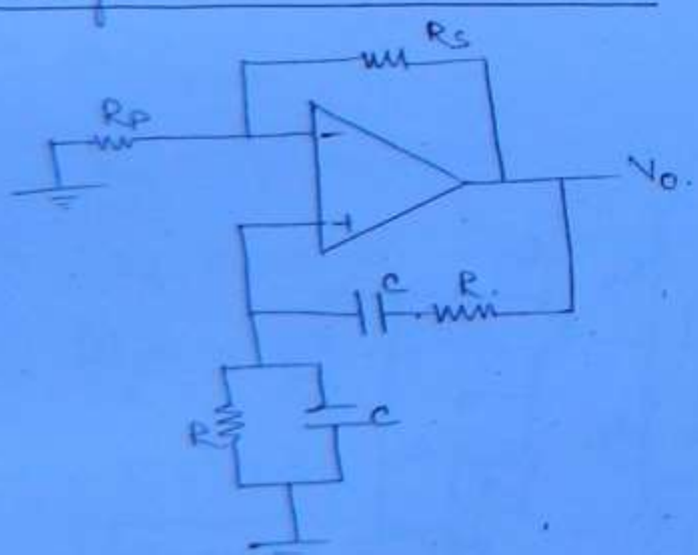
3) $A_V = -3$.

$\beta = -\dfrac{1}{3}$.

Gain stabilization →

In wien bridge oscillator, the gain requirement is 3 but because of cascading of two transistors the gain is becoming very high.
To reduce the gain, a -ive feedback is implemented across the transistor Q_1 by R_e resistor.

Bridge balance condition :—



Non inverting Amplifier →

$A_V = 1 + \dfrac{R_f}{R_1} = 1 + \dfrac{R_S}{R_P} = 3. \Rightarrow \boxed{\dfrac{R_S}{R_P} = 2}$

$\boxed{R_S = 2R_P}$

Importance of Wien bridge :—

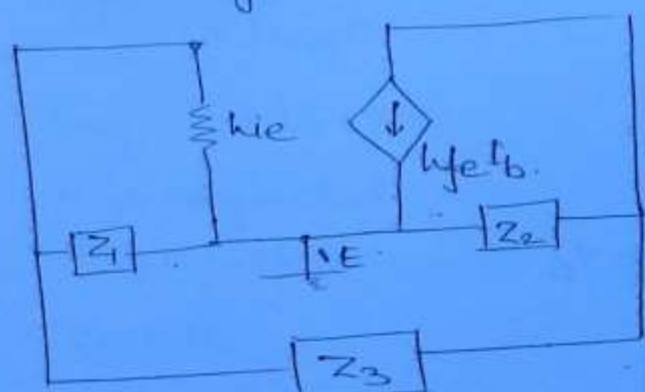
- ~~Induced~~ It good freq. stability.
- It can be used for dynamic range of freq. because wein bridge do not depend on phase.

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

- Drawbacks →
- ckt complexity is more.
- It can not be used for high freq. because Q-factor of RC n/w is poor.

LC Oscillators :→

Block diagram :—

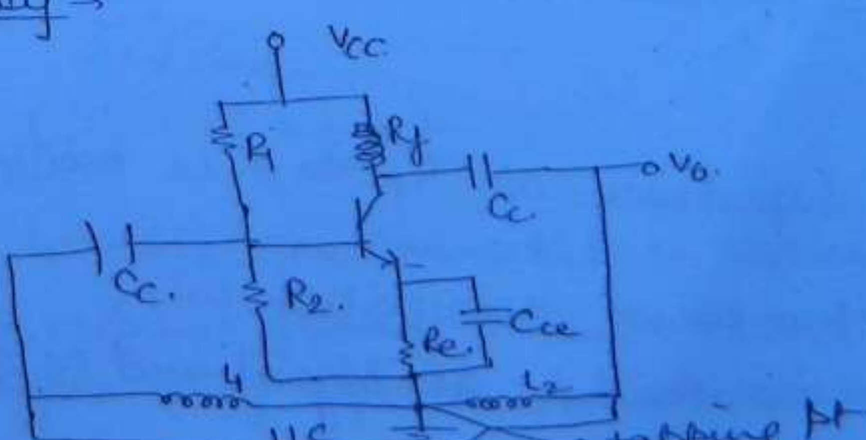


condition :—

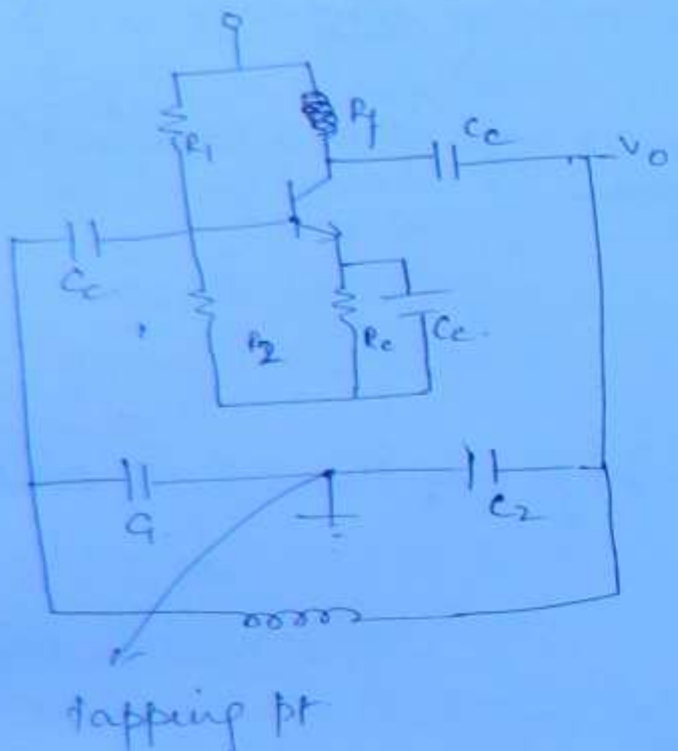
Z_1 and Z_2 are same reactive comp.

Z_3 is opp. reactive comp.

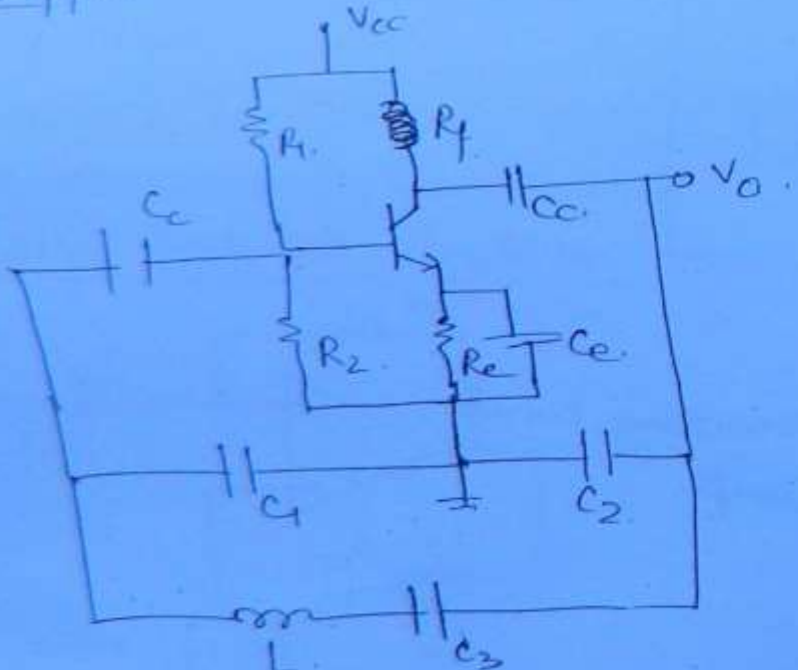
Hartley →



Colpitts



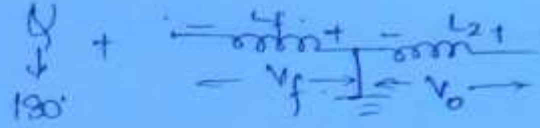
Clapp →



Analysis →

Q1 What is the importance of Rf as in LC oscillatory cct.

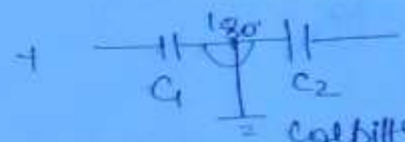
Ans 1) To reduce the power dissipation loss,
2) gives an isolation b/w ac signal and DC supply.



— Hartley.

(or)

197



Colpitts.

Tapping pt is an adjustable moving pt which can contro the +ive feedback nature in the ckt.

→ Explain the freq. of oscillation in the LC oscillators?

Ans. $f = \frac{1}{2\pi\sqrt{LC}}$

Hartley →

$$f = \frac{1}{2\pi\sqrt{L_{eq} C}}$$

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

M is neglected

$L_{eq} = L_1 + L_2$.

Colpitts

$$f = \frac{1}{2\pi\sqrt{L C_{eqv}}}$$

$$= \frac{1}{2\pi\sqrt{L \frac{C_1 C_2}{C_1 + C_2}}}$$

Clapp.

$$f = \frac{1}{2\pi\sqrt{L C_{eq}}}.$$

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

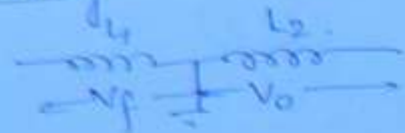
Q. Why L_1 and L_2 are connected in series in LC oscillators

Ans. The current flowing through L_1 and L_2 are same. Therefore they are connected in series.

design a condition for gain in LC oscillators.

Hartley →

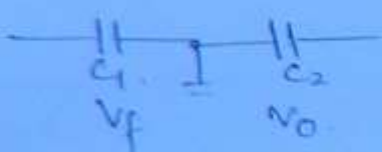
98



$$\beta = \frac{V_f}{V_o} = \frac{IX_{L_1}}{IX_{L_2}} = \frac{L_1}{L_2}$$

$$A = \frac{L_2}{L_1}$$

Colpits



$$\beta = \frac{V_f}{V_o} = \frac{IX_{C_1}}{IX_{C_2}} = \frac{C_2}{C_1}$$

$$\boxed{A = \frac{C_1}{C_2}}$$

→ Why LC oscillators can not be designed for low freq. of operation.

$$\downarrow f = \frac{1}{2\pi\sqrt{L\uparrow C\uparrow}}$$

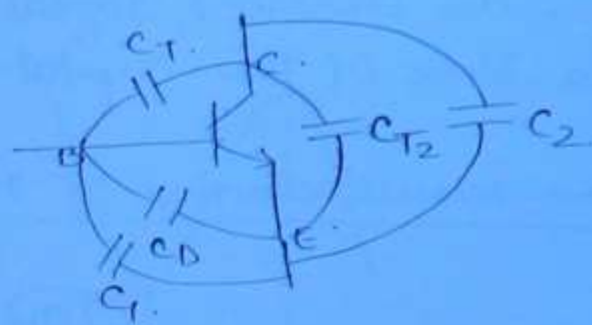
Size of equipment become large.

As the freq. of design reduces, the comps L and C values will increase. Practically it is impossible to design for very low freq.

⇒ LC oscillators are used for high freq design because the Q-factor is more.

→ What is the imp. of clapp oscillators?

colpitts → :



$$f = \frac{1}{2\pi\sqrt{LC_{eqv}}}$$

$$C_{eq} = \frac{C_1 C_2}{C_1 + C_2}.$$

$$f = \frac{1}{2\pi\sqrt{L C_{eqv}}}.$$

$$\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

$C_1 >> C_3$
$C_2 >> C_3$

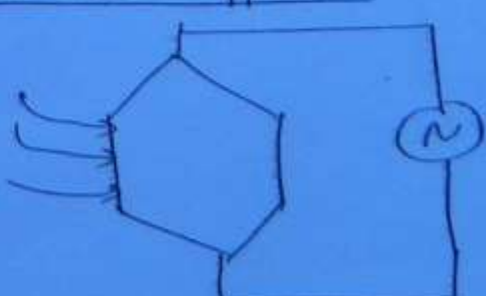
$$\boxed{\frac{1}{C_{eqv}} = \frac{1}{C_3}}$$

$f = \frac{1}{2\pi\sqrt{LC_3}}$ → independent of internal capacitors.

Crystal oscillators →.

whenever high freq. stability is required.

Piezzo electric effect :-



when a mechanical force/pressure is applied on one phase of the crystal, an Electrical signal will be generated on the other side of the crystal and vice versa.

Substances used for the manufacturing of the crystal:—

1) Rochelle Salts
2) Quartz
3) tourmaline.

203

Which Substance will have more piezoelectric effect?

Ans. Rochelle Salts > Quartz > tourmaline.

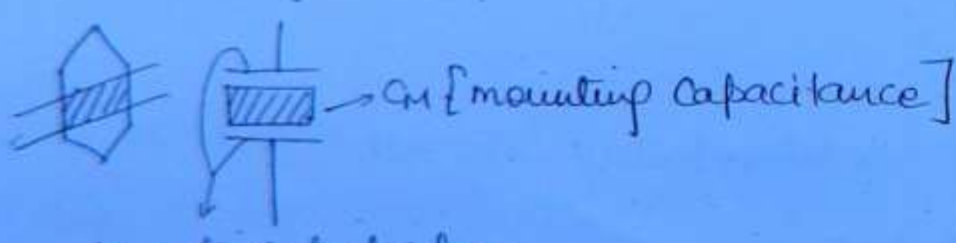
Q. Which substances is mechanically strongest?

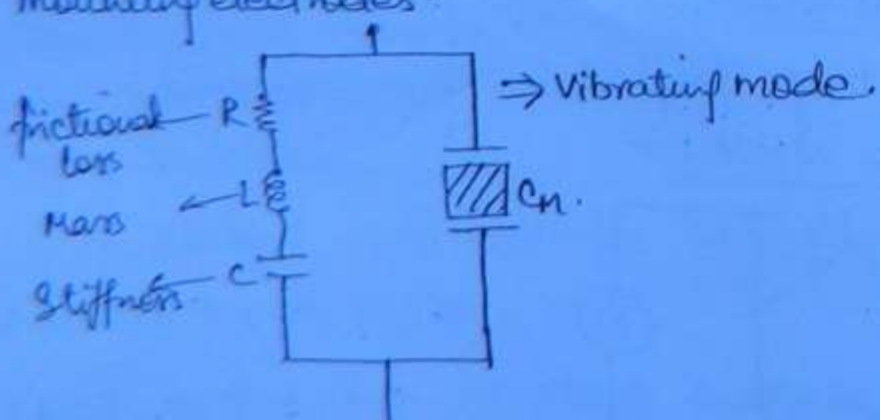
Ans. tourmaline > Quartz > Rochelle Salt

Quartz is a compromise b/w Rochelle Salts and tourmaline as easily available in nature.

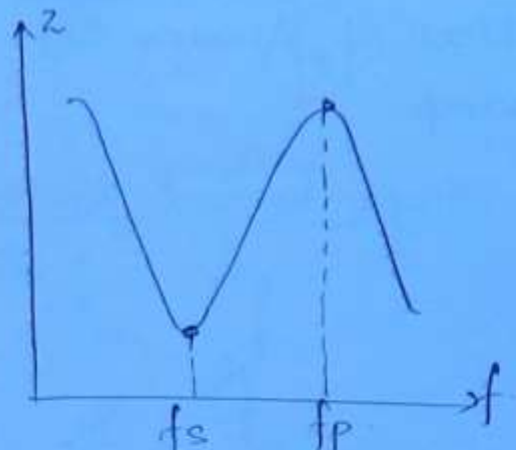
Crystal equivalent model:—

non vibrating mode →.





Crystal operating mode.



fs (zero imp. freq.) fp (Infinite impedance freq)

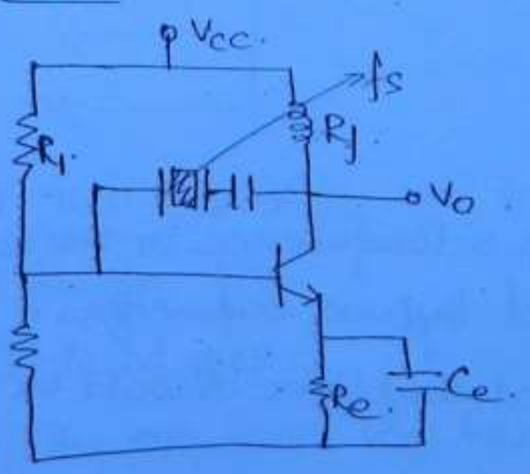
$$f_s = \frac{1}{2\pi\sqrt{LC}}$$

$$f_p = \frac{1}{2\pi\sqrt{LC}}\sqrt{1+\frac{C}{C_M}}$$

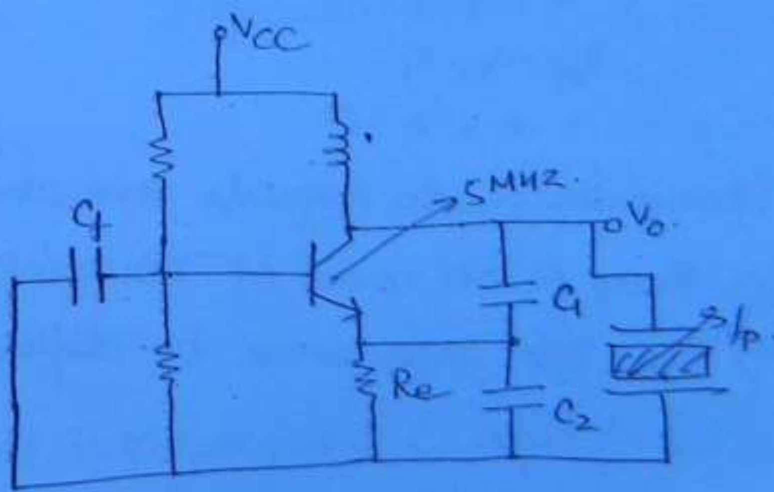
* $C_M >> C$.

$f_p \approx f_s$

Ckt 1.



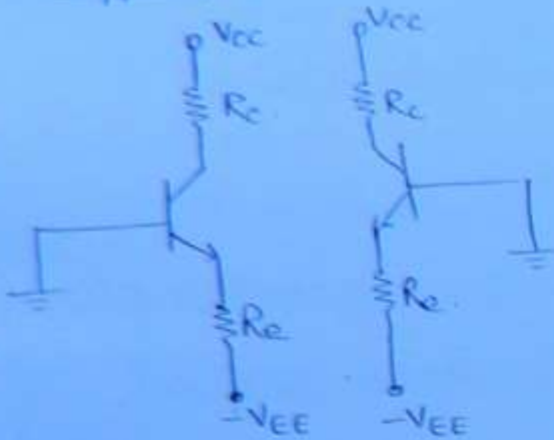
Ckt 2.

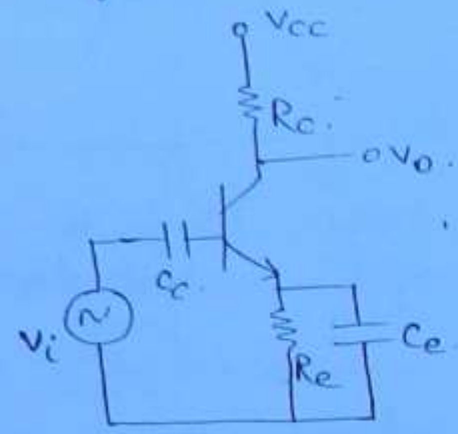


Differential Amplifiers →

It is the basic Building block of Analog ckts.
It is the I/P stage of opamp.

Diff. Amp → | **Single ended Amplifier**





The imp. of differential amplifiers compare to single ended amplifiers are

1) It is having best noise rejection capability.

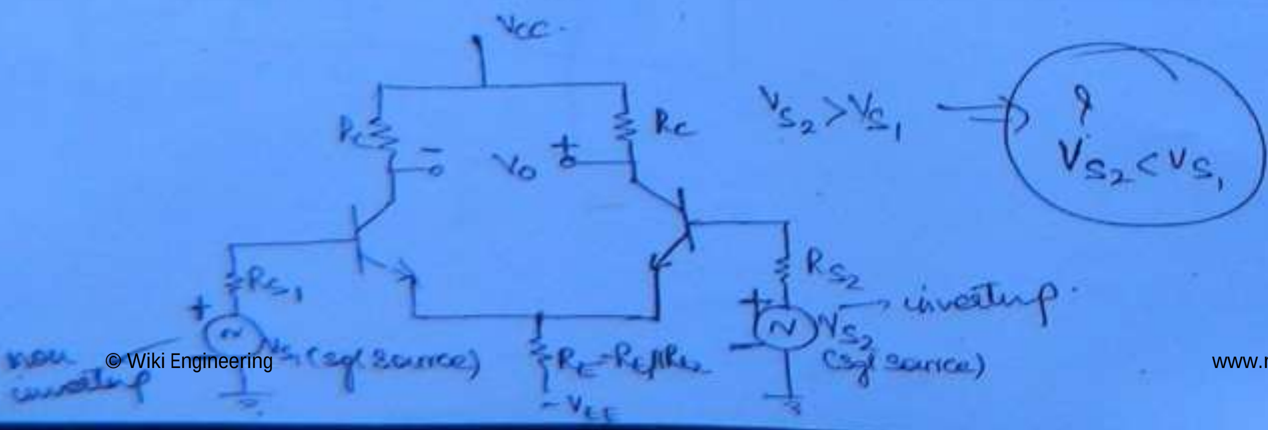
eg –

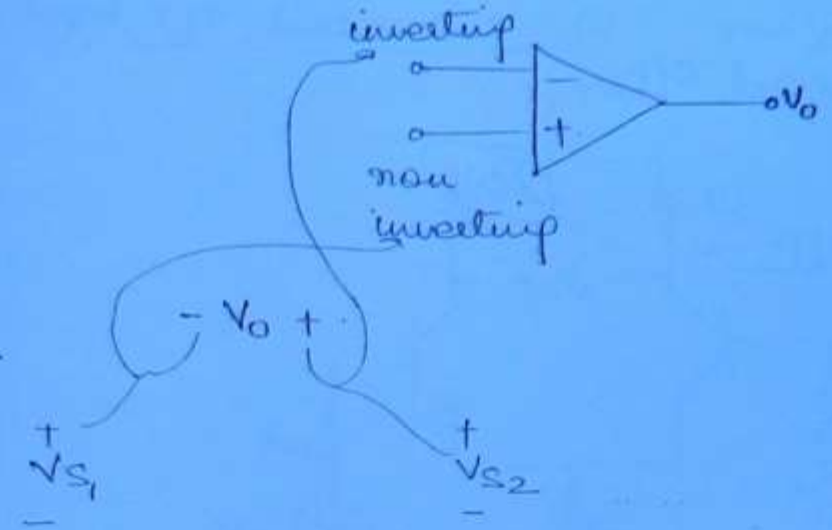
$V_d = V_2 - V_1$
$= 2 - 2 = 0$

common mode signals are always interference or noise signals

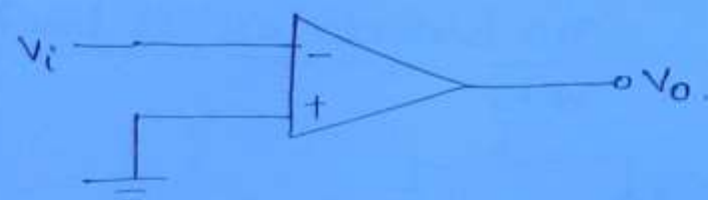
2) There is no need of coupling and bypass capacitors.

3) The I/P impedance of differential amplifier should be high.





Ckt Configuration :→



inverting amp.

non inverting Amp.

Differential amp.

At the I/P side, we have,

1> Single I/P.

2> dual I/P.

At the O/P Side, we have,

1> balanced Output [Output is taken across C_1 and C_2]

2> Unbalanced output [Output is taken across C_1 to ground or C_2 to ground].

Acc. to this we have four types of configuration.

1> Single I/P balanced O/P.

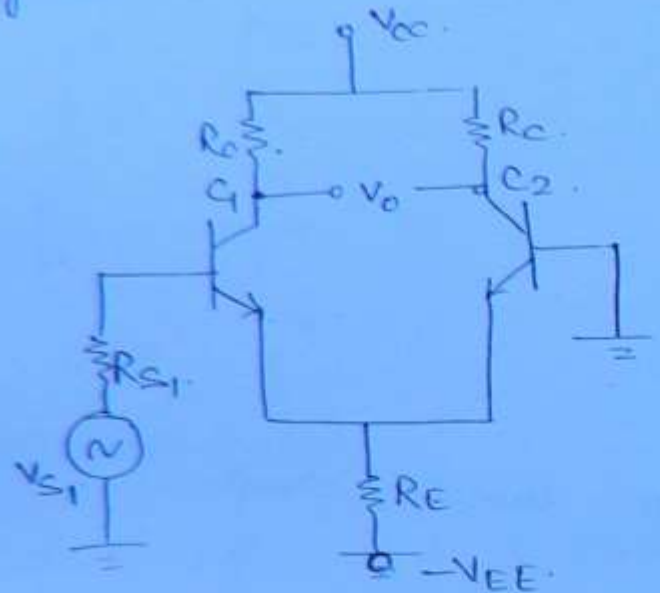
2> Single I/P Unbalanced O/P.

3) Dual I/P balanced O/P.
4) Dual I/P Unbalanced O/P.

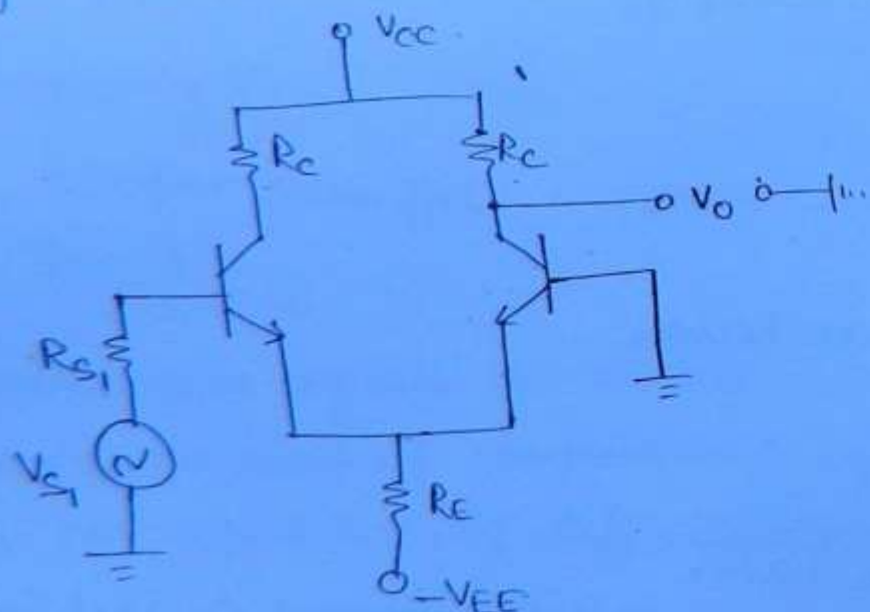
CKt diagrams →

204

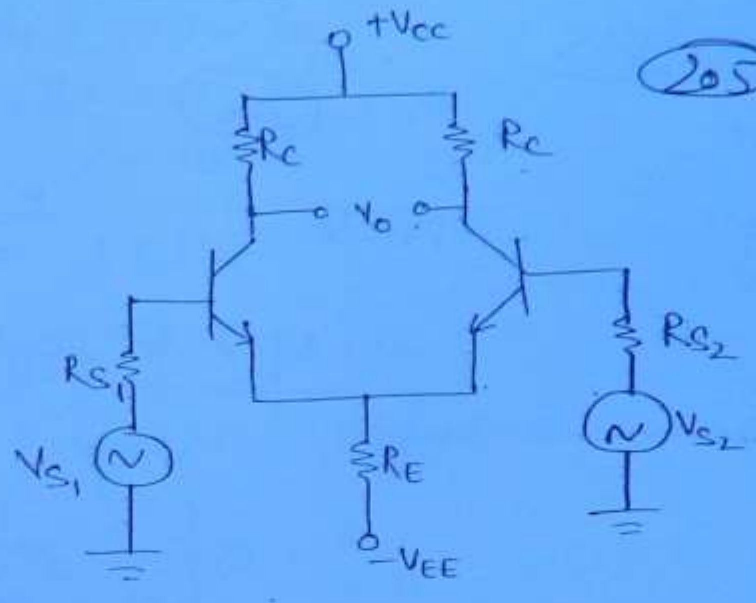
Single I/P balanced O/P →



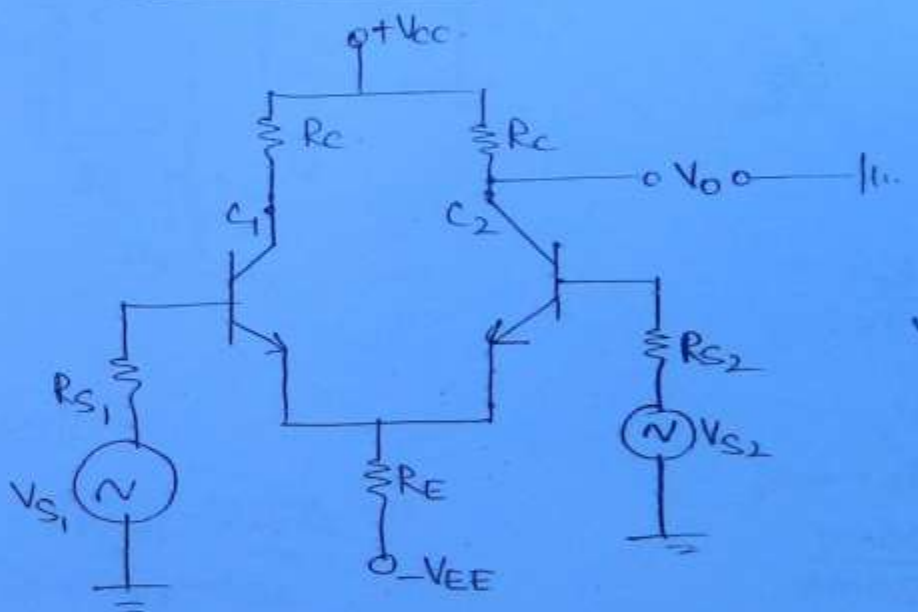
Single I/P Unbalanced O/P →



Dual I/P Balanced O/P.



Dual I/P unbalanced O/P.



Problems →

1> Cal. voltage gain of a single I/P balanced O/P. or dual I/P balanced O/P. differential amplifier ckt with $R_C = 1K\Omega$ and $I_E = 26mA$.

2) Cal. Voltage gain for a single I/P a unbalanced O/P or dual I/P unbalanced O/P differential amplifier ckt with $R_C = 1K\Omega$ and $I_E = 26mA$.

3) Cal. I/P impedance of a differential amplifier ckt with $\beta_{ac} = 50$, $I_E = 26mA$.

Q. cal. O/P impedance of ... ...ff amplifier ckt with $R_E =$ $R_C = 1K\Omega$

(206)

	A_V	R_i	R_o
Single I/P balanced O/P	$\frac{R_C}{r_e}$	$2\beta_{ac} r_e$	R_C
Single I/P Unbalanced O/P	$\frac{1}{2}\frac{R_C}{r_e}$	"	R_C
Dual I/P balanced O/P	$\frac{R_C}{r_e}$	"	"
Dual I/P Unbalanced O/P	$\frac{R_C}{2r_e}$	"	"

Answers →

1> $A_V = \frac{R_C}{r_e}$

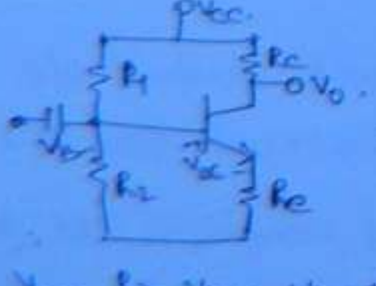
$r_e = \frac{V_T}{I_E \to dc}$

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

$= 1\,\Omega$

$R_C = 1K\Omega$

$A_V = \frac{1K\Omega}{1\Omega} = 1000$ → for both ckt



→ $A_V = \frac{R_C}{r_e} = \frac{R_C \cdot I_C}{V_T}$

$V_B = \frac{R_2}{R_1+R_2} V_{CC}$. $\frac{V_B - V_{BE}}{R_2+R_E} = I_C$

2) $A_V = 500$

3) $R_i = 2\beta_{ac} r_e$

$= 2 \times 50 \times 1$

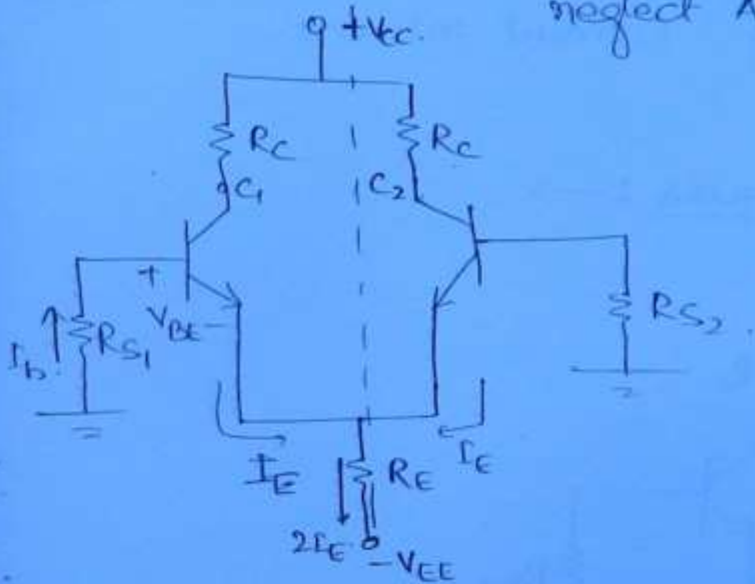
$= 100\ \Omega$

4) $R_o = 1\ K\Omega$

(207)

Calculation of Q-pt:→

neglect Ac concept



Q_1

I/P loop→

$$V_{EE} - I_b R_{S_1} - 2I_E R_E \underset{-V_{BE}}{} = 0$$

$$I_E = \frac{V_{EE} - V_{BE}}{2R_E + \frac{R_{S_1}}{1+\beta}}$$

$1+\beta >> R_{S_1}$

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

$$\boxed{(I_c)_Q = I_E = \frac{V_{EE} - V_{BE}}{2R_E}}$$

$$\boxed{(V_{CE})_Q = V_{cc} - (I_c)_Q R_c + V_{BE}}$$

$(V_{CE})_Q = V_c - V_E$

$V_c = V_{cc} - (I_c)_Q R_c \Rightarrow V_E = -V_{BE}$

Design problems in differential amplifiers

I> 1> $A_V = \frac{R_C}{r_e}$ — gain is not stable.

2> $R_i = 2\beta_{ac} r_e = 2 \times 50 \times 1$
$= 100\,\Omega$

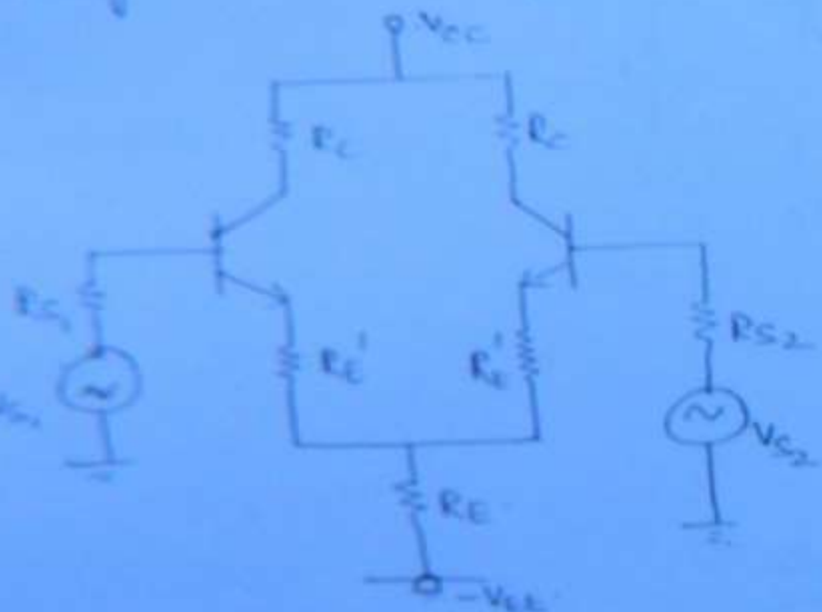
Rectify R techniques
→ Swamping resistor techniques

208

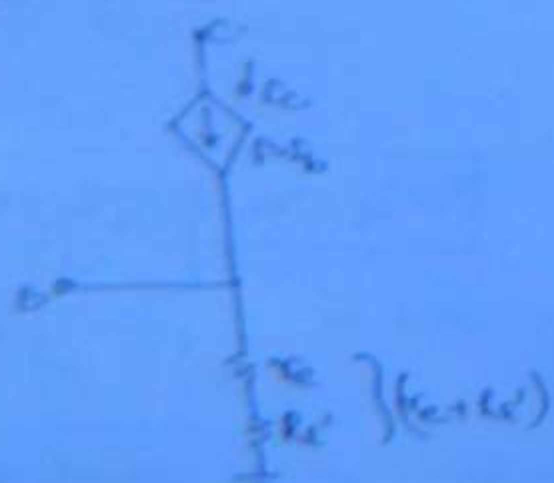
II> $Q((I_C)_Q, (V_{CE})_Q)$
$\downarrow$
$I_E = \frac{V_{EE} - V_{BE}}{2R_E}$

Rectify techniques
→ constant current Source
current mirrors.

Swamping resistor techniques :→



r_e model



$A_V = \frac{R_C}{r_e}$
↓ swamping

$A_V = \frac{R_C}{r_e + R_E'}$

$= \frac{R_C}{R_E'}$ $\quad R_E' \gg r_e$.

→ gain is stable.

$M = 4\beta_{ac} r_{ce}$.

$\downarrow$ swamping

$R_i = 2\beta_{ac}(r_e + R_E')$

Assume $R_E' = 100K$.

$R_i = 2 \times 50 \times 100K$

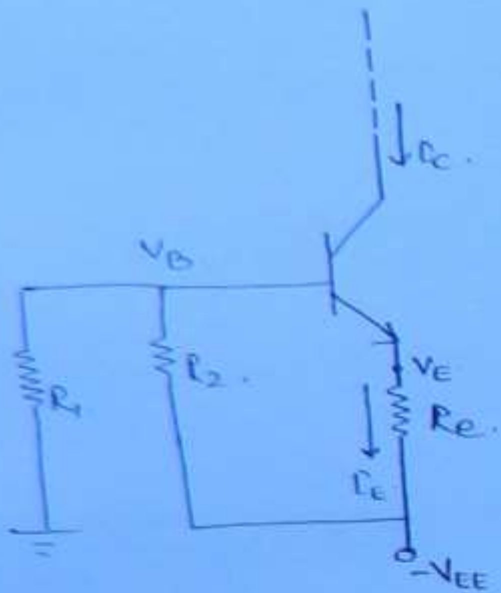
$= 10\ M\Omega$.

6/1/12

209

Const current source →

Ckt 2.



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

$$V_E = V_B - V_{BE}$$

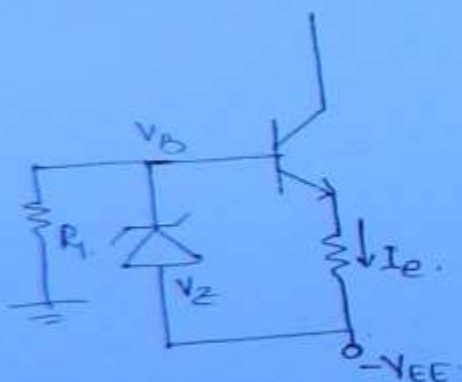
$$V_B = \frac{-V_{EE} \times R_1}{R_1 + R_2}$$

$$I_E = \frac{V_B - V_{BE} + V_{EE}}{R_E}$$

$\Downarrow$

the main drawback is I_E is dependending on V_{EE} supply

Ckt 3



$$I_E = \frac{V_Z - V_{EE} - V_{BE} + V_{EE}}{R_E}$$

$$V_B = V_Z - V_{EE}$$

Current mirror →

Ckt 1.

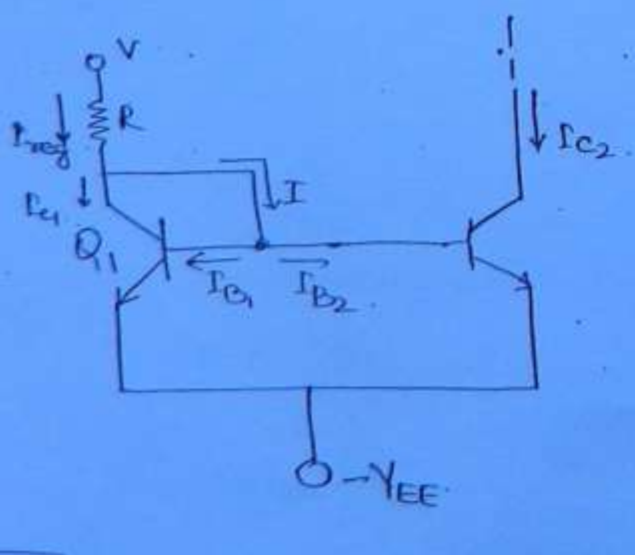
$I_{ref} = I_{C_1} + I$

$= I_{C_1} + I_{b_1} + I_{b_2}$

$= I_{C_2} + 2 I_{b_2}$

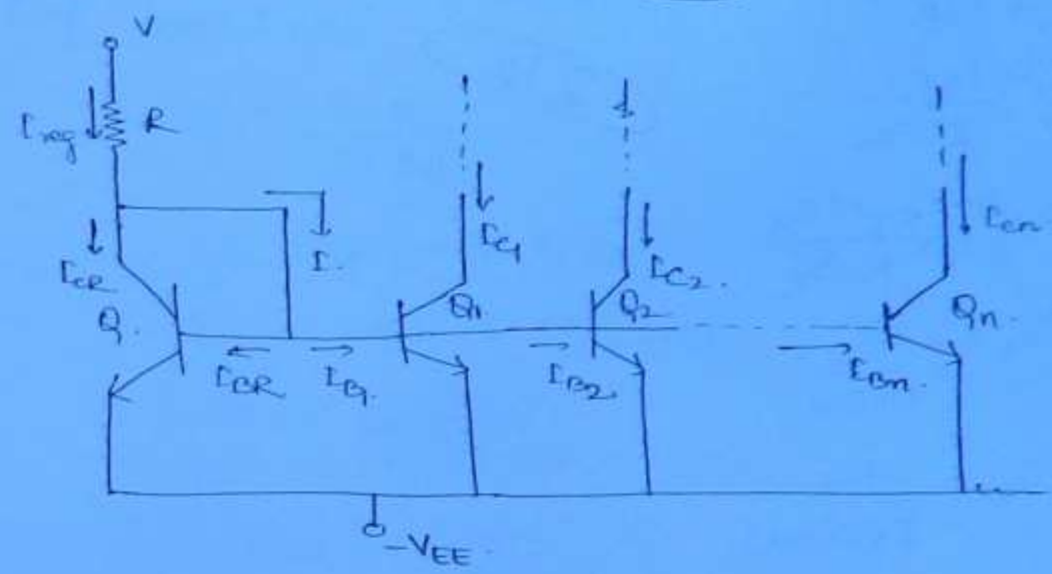
$= I_{C_2} + 2 \frac{I_{C_2}}{\beta}$

$= I_{C_2} \left[1 + \frac{2}{\beta} \right]$



$$\boxed{I_{C_2} = \frac{I_{ref}}{1 + 2/\beta}}$$

n stage → ckt2.



$$I_{C_1} = I_{C_2} = \cdots = I_{Cn} = I_{CR}$$
$$= I_{Ci} \quad (i = 1, 2, 3 \ldots n)$$

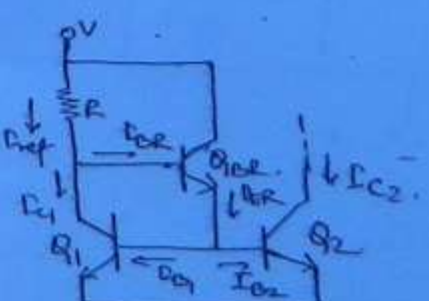
$$I_{BR} = I_{B_1} = I_{B_2} \cdots = I_{Bn} = I_{Bi}$$

$$I_{reg} = I_{CR} + I$$
$$= I_{CR} + (I_{BR} + I_{B_1} + I_{B_2} \cdots + I_{Bn})$$
$$= I_{CR} + I_{BR}(1+N)$$
$$= I_{CR} + \frac{I_{CR}}{\beta}(1+N)$$
$$= I_{CR}\left[1 + \frac{1+N}{\beta}\right]$$

$$I_{CR} = \frac{I_{reg}}{1 + \frac{(1+N)}{\beta}}$$

N should be less for less error.

Ckt 3 →



$$I_{ref} = I_{C1} + I_{BR}$$

$$= I_{C1} + \frac{I_{ER}}{1+\beta_R}$$

$$= I_{C2} + \frac{I_{B1}+I_{B2}}{1+\beta_R}$$

$$= I_{C2} + \frac{2I_{B2}}{1+\beta_R}$$

$$= I_{C2} + \frac{2I_{C2}}{\beta(1+\beta_R)}$$

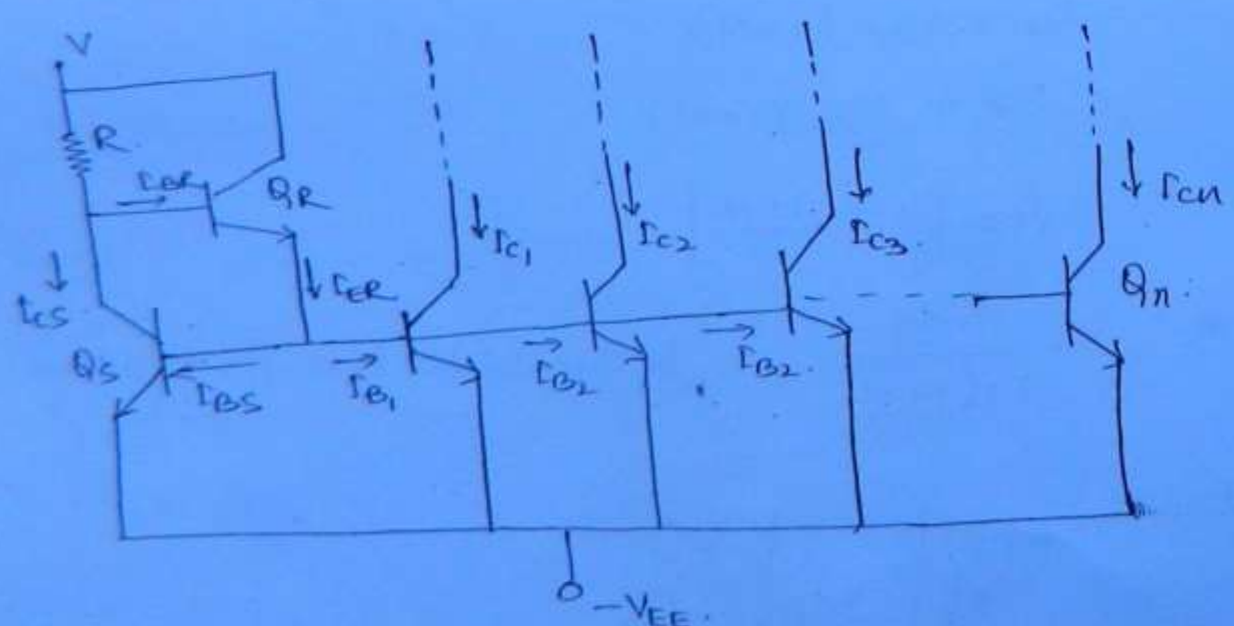
2-2

$$I_{C2} = \frac{I_{ref}}{1+\frac{2}{\beta(1+\beta_R)}}$$

$$I_{C2} \approx I_{ref}$$

$$\boxed{\text{as } \beta(1+\beta_R) >> 2}$$

<u>n stage</u> → (ckt-4)



$$I_{C1} = I_{C2} = \dots = I_{Cn} = I_{CS} = I_{C1}$$

$$I_{BS} = I_{B1} = I_{B2} = \dots = I_{Bn} = I_{B1}$$

(213)

$I_{ref} = I_{CS} + I_{BR}$

$= I_{CS} + \dfrac{I_{ER}}{1+\beta_R}$

$= I_{CS} + (I_{BS} + I_{B_1} + I_{B_2} + \cdots I_{Bn})/1+\beta_R$

$= I_{CS} + I_{BS}(1+N)/1+\beta_R$

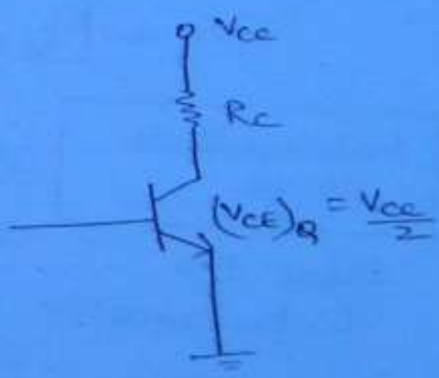
$= I_{CS} + \dfrac{I_{CS}(1+N)}{\beta(1+\beta_R)}$

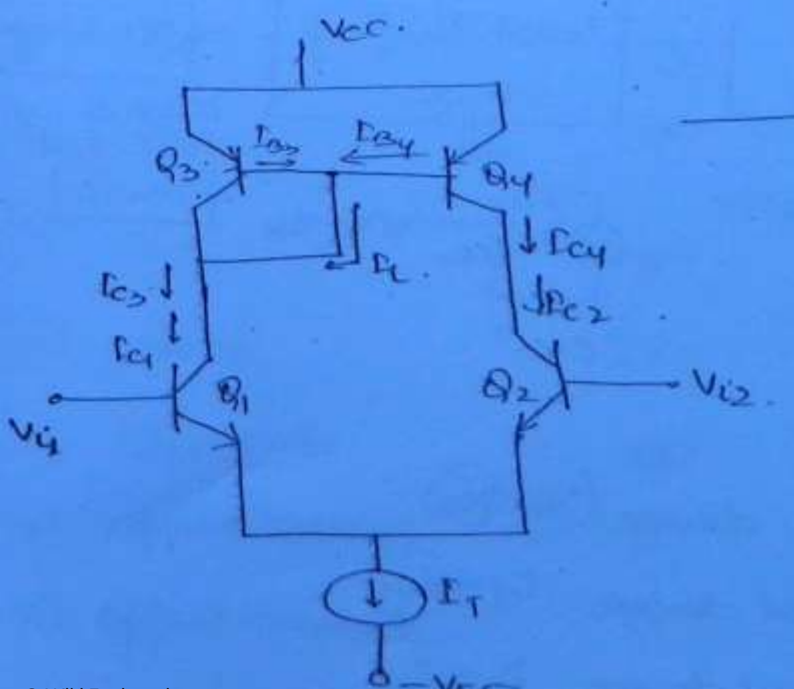
$= I_{C_2}\left[1 + \dfrac{1+N}{\beta(1+\beta_R)}\right]$

$$I_{C_2} = \frac{I_{ref}}{1 + \dfrac{(1+N)}{\beta(1+\beta_R)}}$$

Active load in differential amplifiers →

$\uparrow A_d = \dfrac{R_C \uparrow}{r_e}$





Assume Q_1, Q_2, Q_3, Q_4 are identical.

$$I_i = I_{B3} + I_{B4}$$
$$= \frac{1}{\beta}(I_{C3} + I_{C4})$$
$$= \frac{1}{\beta}(I_{C1} + I_{C2})$$
$$= \frac{1}{\beta} I_7 \approx 0$$

Active load $= \infty$

$$CMRR \uparrow = \frac{A_d \uparrow}{A_c}$$

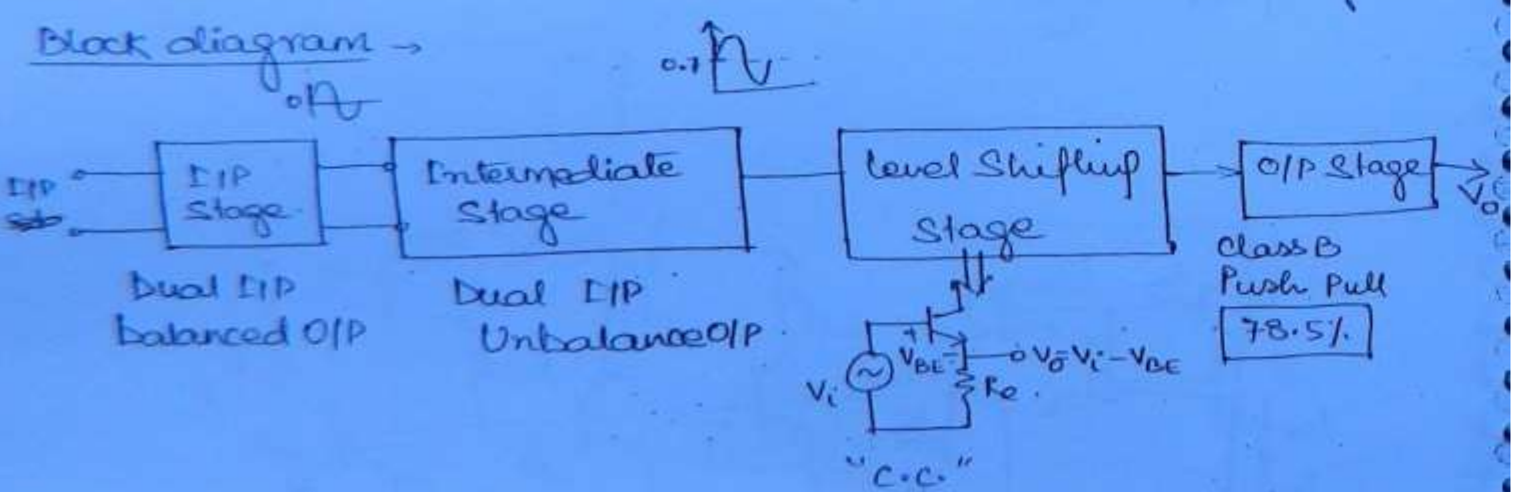
Active loads are used in differential amplifiers to improve

1) Differential gain A_d.
2) CMRR (common mode rejection ratio)

$$= \frac{A_d}{A_c}$$

OPAMP →

Block diagram →



Imp. pts →

1) Opamp is a voltage control device $(V_o \propto V_d)$
2) FET is a Voltage control device $I_d \propto V_{GS}$
3) BJT is a Current control device $I_C \propto I_b$

$I_{D,P}$

II Match 1 (215) Match 2.

2) VCVS ...... 1) BJT

3) VCCS ... 2) Opamp

ICVS ..... 3) FET.

1) ICCS

III Increasing order of $Z_i \rightarrow$

BJT, Opamp, FET, MOSFET

Typical Values $\rightarrow$

BJT $\rightarrow$ $1K\Omega$

OPAMP $\rightarrow$ $10^6 \Omega$

FET $\rightarrow$ 10^7 to 10^{10}

MOSFET $\rightarrow$ 10^{11} to $10^{14} \Omega$

Characteristics of OP-AMP:$\rightarrow$

	Characteristics	Ideal	Practical.
1)	Z_i	∞	10^6
2)	Z_o	0.	100Ω
3)	A_v	∞	10^6
4)	BW	∞	10^6 HZ
5)	CMRR	∞	10^6 or 120dB
6)	Slew rate	∞	80V/μs
7)	Offset Voltage	0	min
8)	Offset current	0.	min

CMRR →

$$CMRR = \frac{A_d}{A_c}$$

216

$$(CMRR)_{dB} = 20 \log\left(\frac{A_d}{A_c}\right)$$

Ex

It is a linear IC. So, superposition theorem is applied.

$V_o \propto V_d$

$V_o = A_1V_1 + A_2V_2$ —(1)

$V_d = V_2 - V_1$

$V_c = \frac{V_1 + V_2}{2}$

$2V_c = V_1 + V_2$

$V_1 = V_c + V_d/2$

$V_2 = V_c - V_d/2$

Substitute in eqⁿ (1)

$$V_o = A_1(V_c + V_d/2) + A_2(V_c - V_d/2)$$

$$= \underbrace{(A_1 + A_2)}_{A_c} V_c + \underbrace{\left(\frac{A_1 - A_2}{2}\right)}_{A_d} V_d$$

$V_o = A_cV_c + A_dV_d$

$$V_o = A_dV_d\left[1 + \frac{A_c}{A_d}\frac{V_c}{V_d}\right]$$

$$\frac{A_c}{A_d} \to \frac{1}{CMRR}$$

Q. A differential Amplifier has I/P $V_1 = 1050\,\mu V$, $V_2 = 950\,\mu V$.

CMRR = 1000.

217

diff. error O/P = ?

Ans. $\%E = \dfrac{1}{1000} \times \dfrac{V_c}{V_d}$

$V_1 = V_c + V_d/2$

$1050 = V_c + V_d/2$

$950 = V_c - V_d/2$

$2V_c = 2000$

$V_c = 1000$

$V_d = 50 \times 2$

$= 100$

$\cancel{1000 \times 100}$

diff error o/p $= \dfrac{1}{1000} \times \dfrac{1000}{100} \times 100 = 1\%$.

SLEW RATE →

$$SR = \left.\frac{dV_o}{dt}\right|_{\mu s}.$$

$$= \frac{\Delta V_o}{\Delta V_i} \cdot \frac{\Delta V_i}{\Delta t}$$

$$\boxed{S.R. = A_{CL} \frac{\Delta V_i}{\Delta t}}$$

$$\omega_{max} = \frac{S.R.}{K} = \frac{S.R.}{A_{CL}\Delta V_i}$$

$$f_{max} = \frac{S.R.}{2\pi A_{CL}\Delta V_i}$$

$$f_{max} = \frac{S.R.}{K}$$

$\omega_{input} < \omega_{max}$ → O/P will be replica of I/P

$\omega_{input} > \omega_{max}$ → O/P will be distorted.

It is a measure of variation of ... I/P in ms.

218

OFF SET :—

$V_{io} = V_B - V_A$

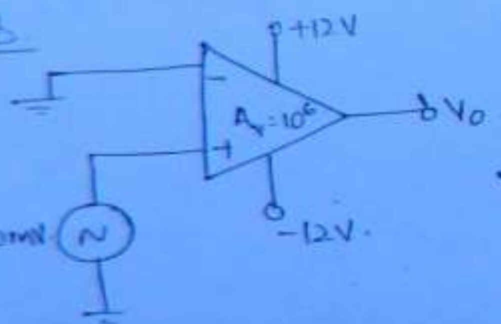
↓

I/P OFFset Voltage.

used to make $\boxed{V_0 = 0}$

W/o giving any I/P to the terminals of the opamp, output is not 0 [small magnitude] the reason is I/P stage is differential amplifier where the two transistor should be identical. Practically it is not possible. To make the O/P V_0 to 0 we use offset condition at the I/P of the OP AMP.

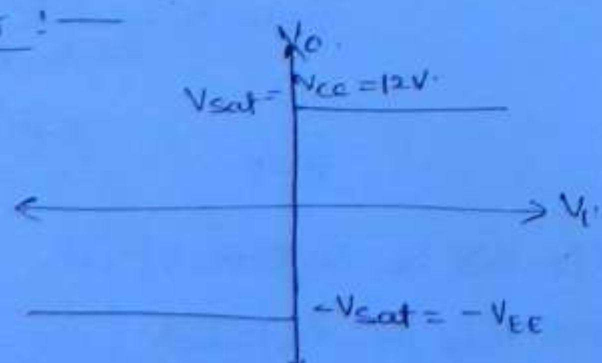
Prob



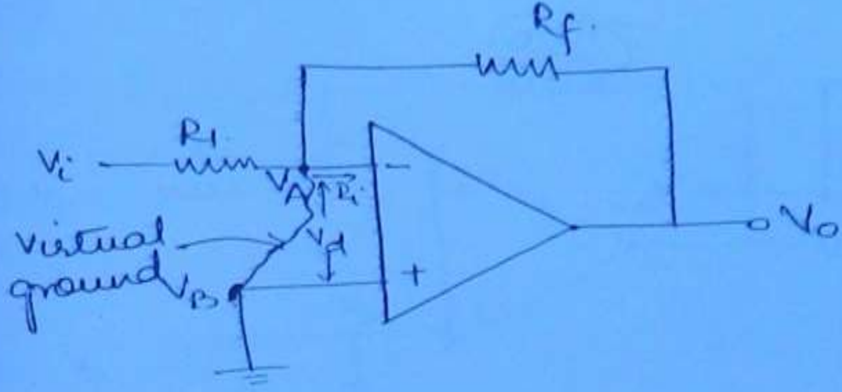
$V_0 = ?$

$V_0 = 12V$ irrespective of supply because +12 V and −12V is the sat value of this opamp.

transfer char :—



Virtual ground :-



$$V_d = V_B - V_A$$

Ideal →

$R_i \to \infty$

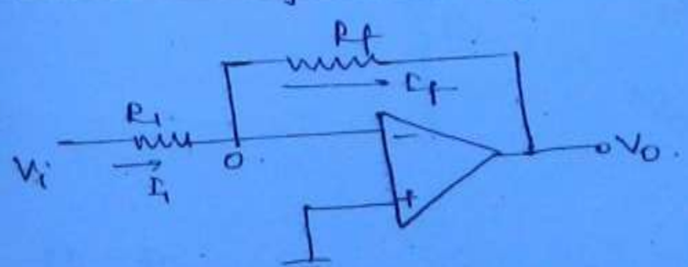
$I_i \to 0$

$V_d \to 0$

$$0 = V_B - V_A$$

$$\Rightarrow \boxed{V_B = V_A}$$

Inverting Amplifier →



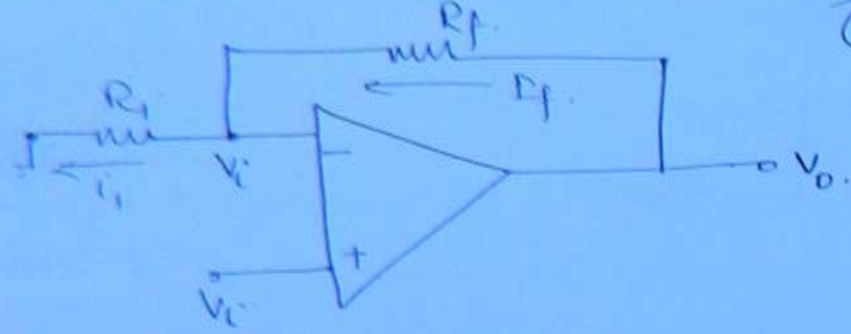
$$I_1 = I_f$$

$$\frac{V_i - 0}{R_1} = \frac{0 - V_o}{R_f}$$

$$\frac{V_o}{V_i} = -\frac{R_f}{R_1}$$

Non inverting amplifier

220

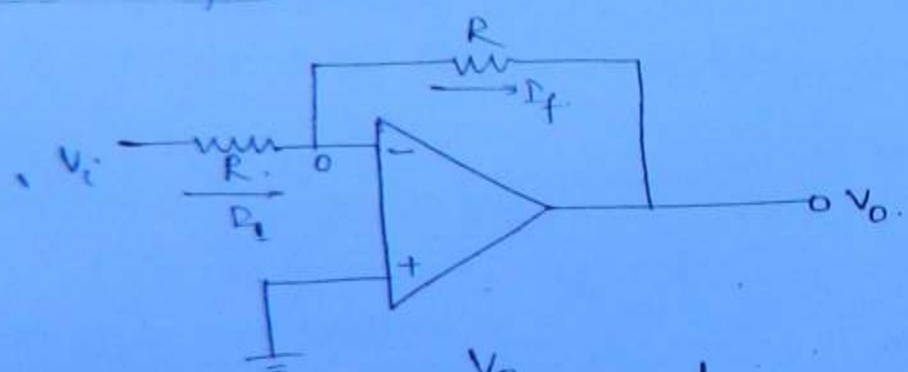


$I_f = I_1$

$$\frac{V_o - V_{o1}}{R_f} = \frac{V_o - 0}{R_1}$$

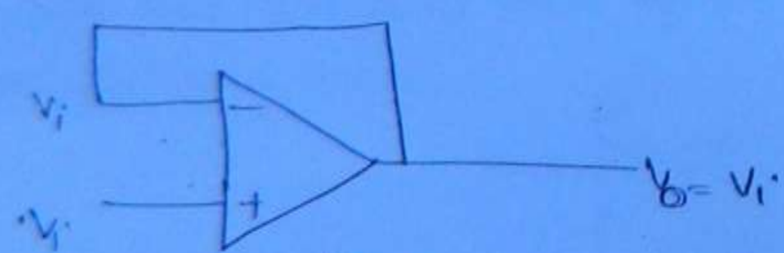
$$\Rightarrow \boxed{\frac{V_o}{V_i} = 1 + \frac{R_f}{R_1}}$$

Phase shifter →



$$\frac{V_o}{V_i} = -1$$

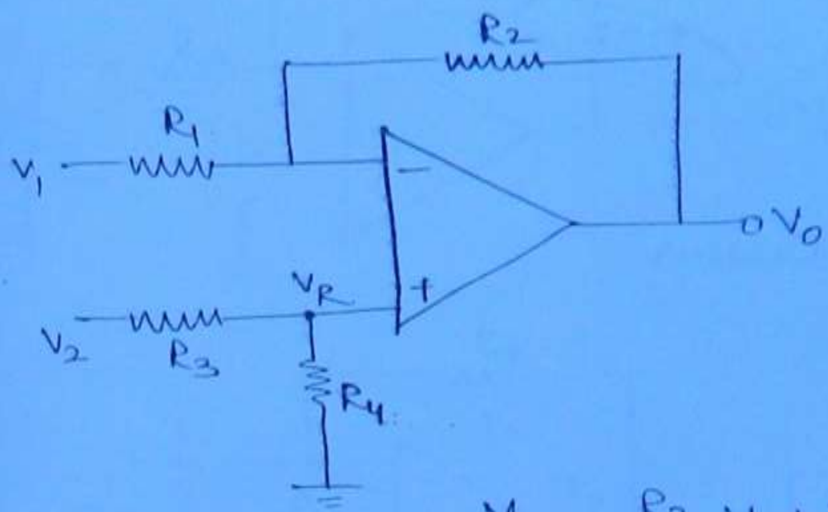
Voltage follower →



$R_f = 0$

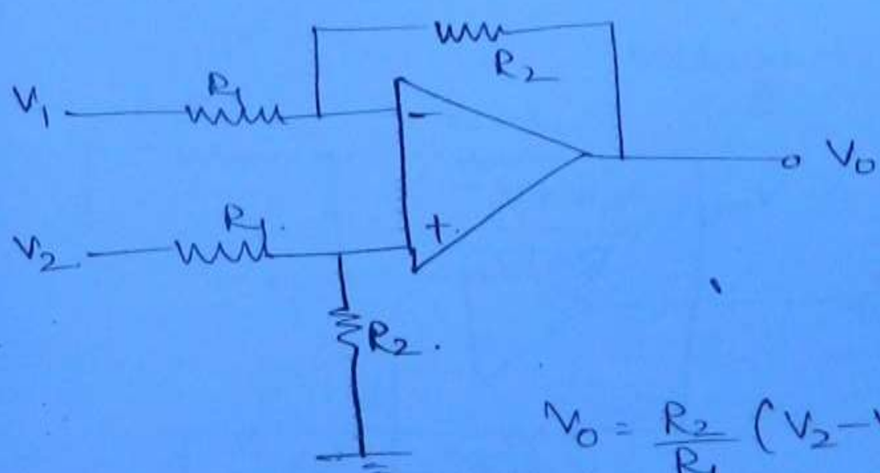
$R_1 = \infty$

Differential Amplifier →



$$V_o = -\frac{R_2}{R_1} V_1 + \left(1 + \frac{R_2}{R_1}\right) V_R$$

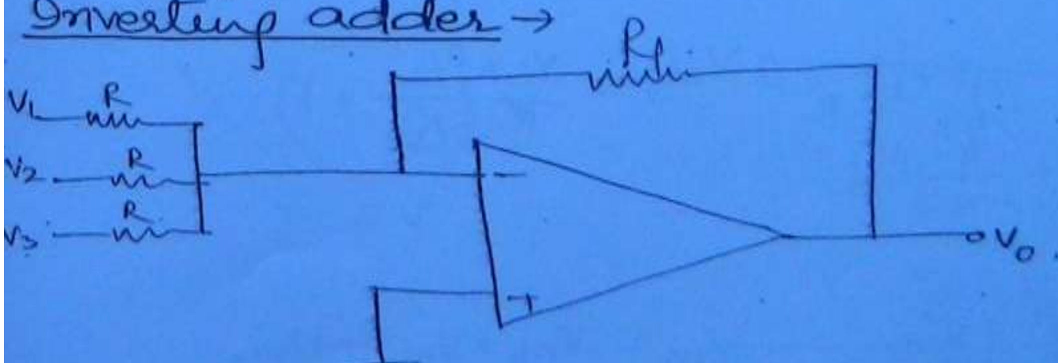
$$V_R = \frac{V_2 \times R_4}{R_3 + R_4}$$



$$V_o = \frac{R_2}{R_1}(V_2 - V_1)$$

if $R_2 = R_1$, $V_o = V_2 - V_1$ → Subtractor.

Inverting adder →



$$I = I_f$$

$$I_1 + I_2 + I_3 = I_f$$

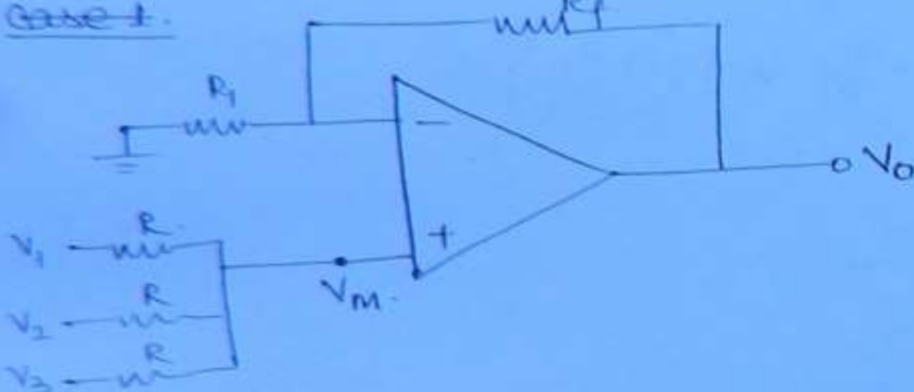
222

$$\Rightarrow \frac{(V_1 + V_2 + V_3)}{R} = \frac{-V_0}{R_0}$$

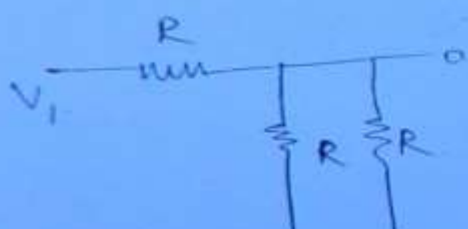
$$\Rightarrow \boxed{V_0 = -(V_1 + V_2 + V_3)}$$

Non inverting adder:→

~~Case 1~~.

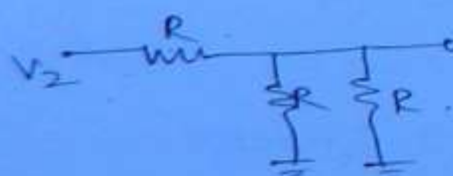


Case 1



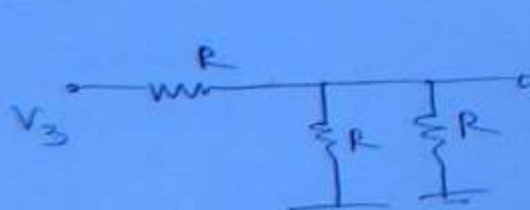
$$V_{m_1} = \frac{V_1 \times R/2}{R + R/2}$$

Case 2



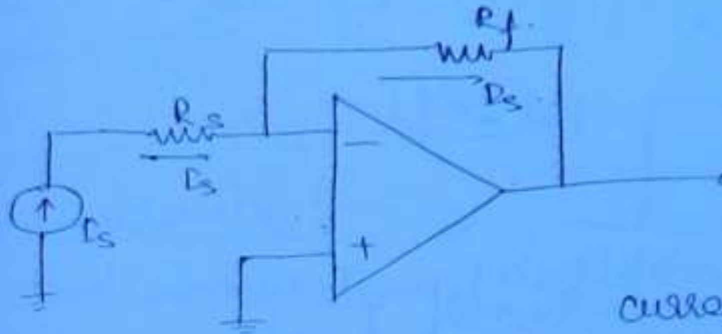
$$V_{M_2} = \frac{V_2 \times R/2}{R + R/2}$$

Case 3



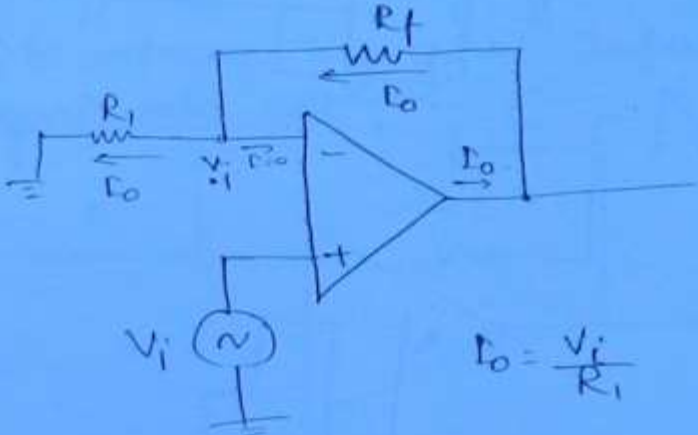
$$V_{m_3} = \frac{V_3 \times R/2}{R + R/2}.$$

$$V_0 = \left(1 + \frac{R_f}{R_1}\right) V_M \qquad V_M = V_{M_1} + V_{M_2} + V_{m_3}$$



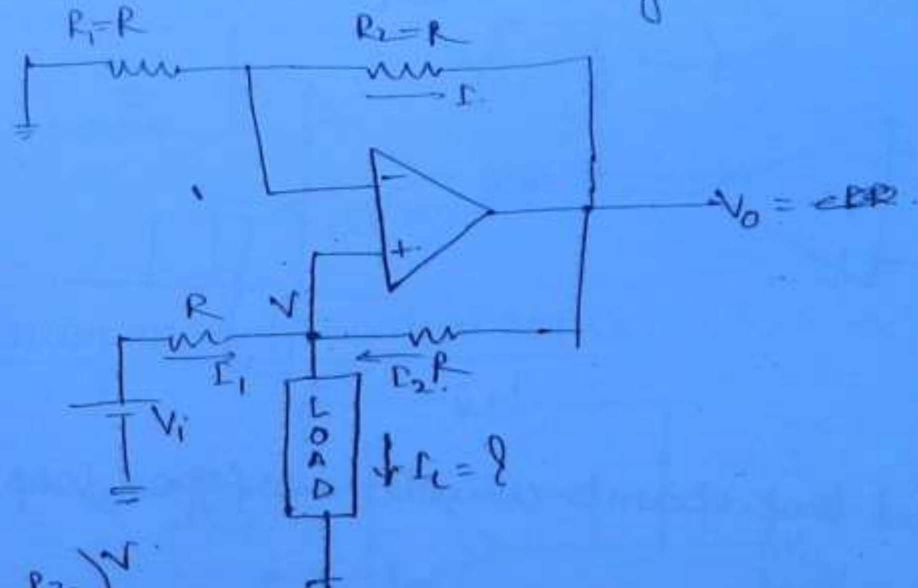
$V_0 = -I_s R_f$

current to voltage converter



$I_0 = \frac{V_i}{R_1}$

Voltage to current converter



$V_0 = \left(1 + \frac{R_2}{R_1}\right) V$

$V_0 = \left(1 + \frac{R}{R}\right) V$

$= 2V$

$I_L = I_1 + I_2$

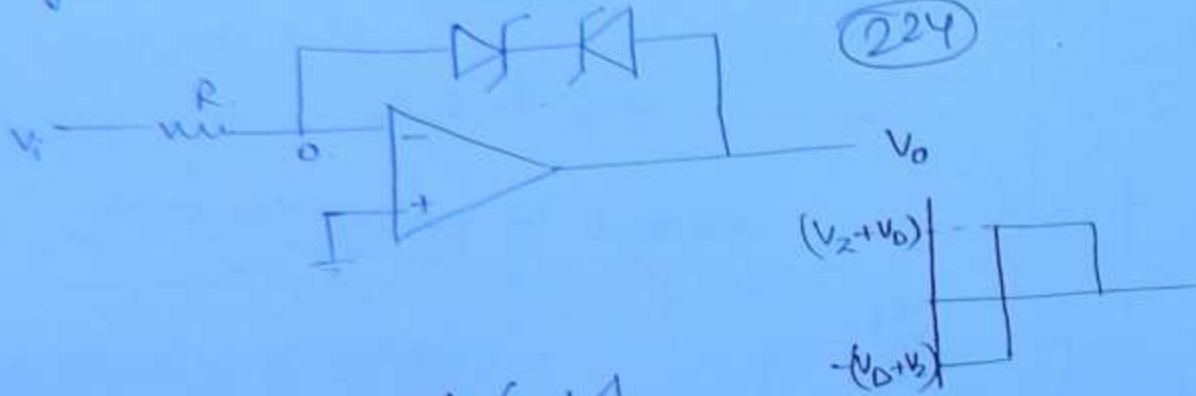
$= \frac{V_i - V}{R} + \frac{V_0 - V}{R}$

$= \frac{V_i}{R} - \frac{V}{R} + \frac{2V - V}{R}$

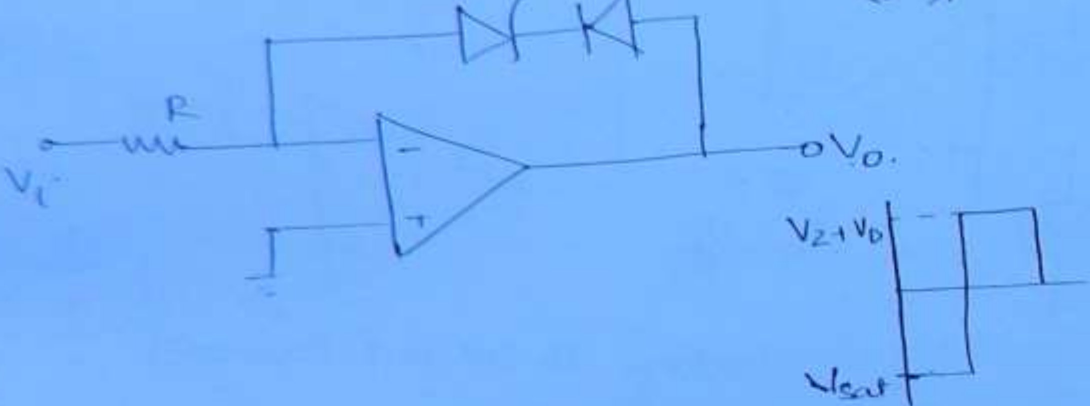
$$\boxed{I_L = \frac{V_i}{R}}$$

Voltage limiters :—

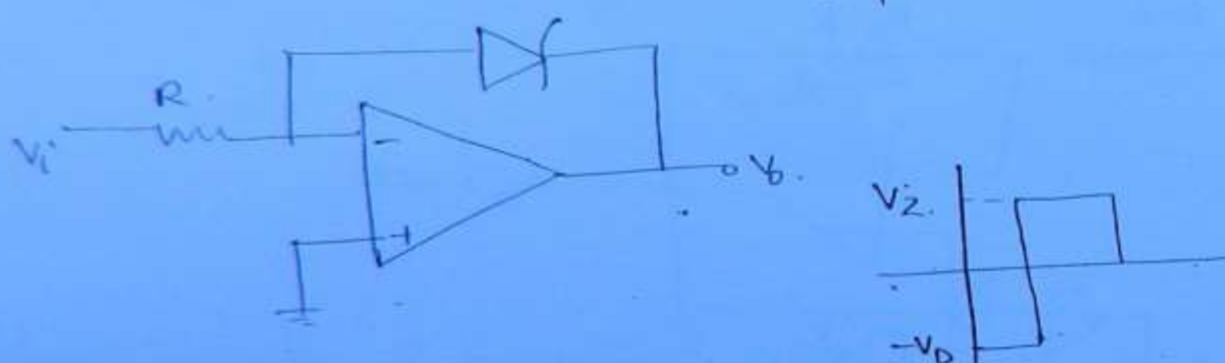
I



II

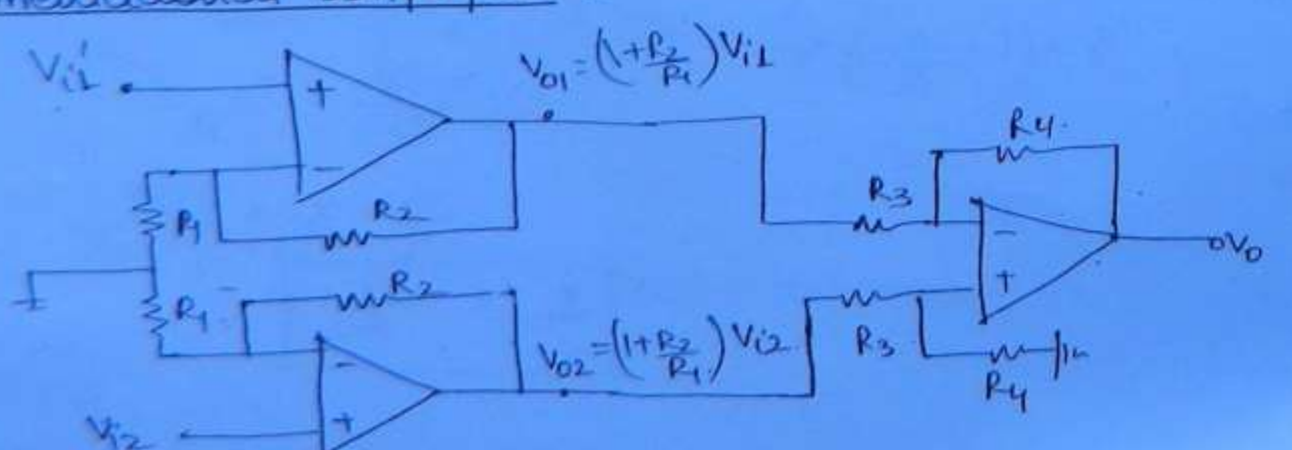


III



whenever a closed loop opamp converts into open loop o/p will be $+V_{sat}$ or $-V_{sat}$

Instrumentational amplifier →



$$-V_O = \frac{R_4}{R_3}\left(1+\frac{R_2}{R_1}\right)(V_{i2}-V_{i1})$$

$$V_{i2} - V_{i1} = V_d$$

225

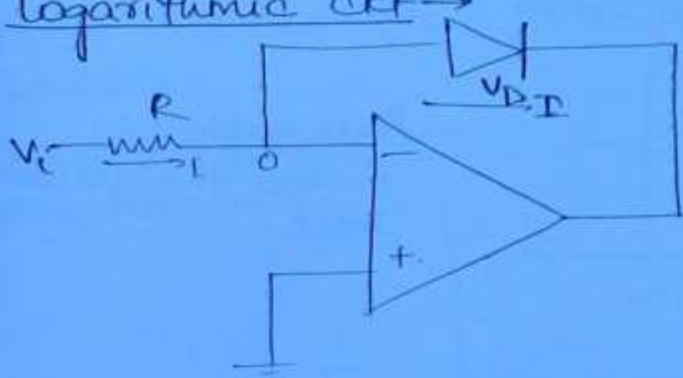
conclusion :—

Instruementational amplifiers are used to improve

1> Differential voltage gain.

2> to improve the input impedance of differential amplifie

logarithmic ckt →



$$V_O = -V_D$$

$$0 - V_O = V_D$$

$$\boxed{V_O = -V_D}$$

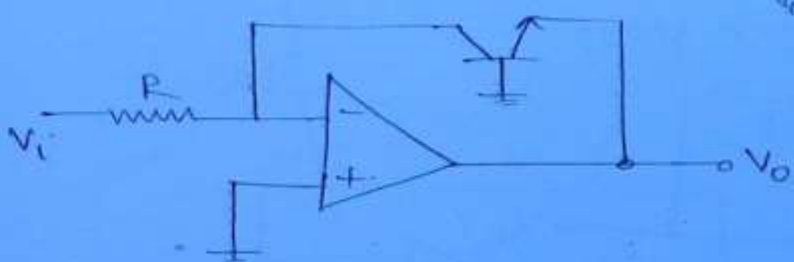
$$I = I_O\left(e^{V_D/\eta V_T} - 1\right)$$

But FB,

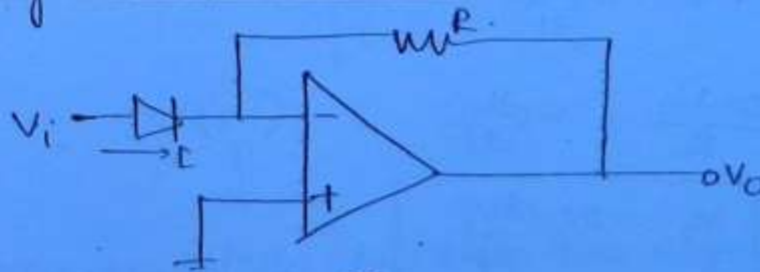
$$V_D = \eta V_T \ln\left(\frac{I}{I_O}\right)$$

$$V_O = -\eta V_T \ln\left(\frac{I}{I_O}\right)$$

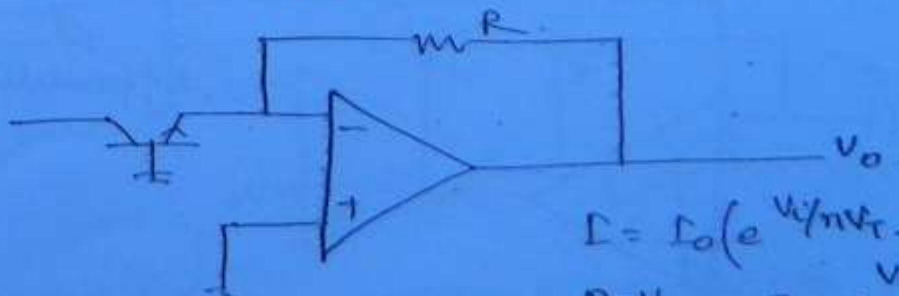
$$V_O = \eta V_T \ln\left(\frac{I_O R}{V_i}\right) \quad I_o = \frac{V_i}{R}.$$



Antilog or exponential ckt →



or

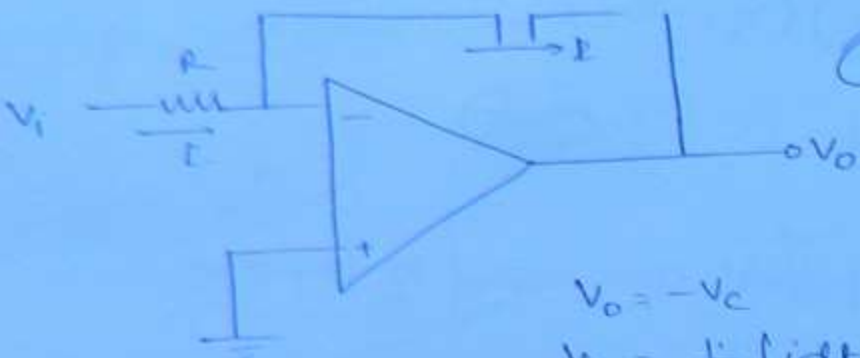


$$I = I_O\left(e^{V_i/\eta V_T} - 1\right)$$

$$\frac{0 - V_O}{R} = I_O\, e^{V_i/\eta V_T}$$

$$V_O = -I_O R\, e^{V_i/\eta V_T}.$$

⇒ Integrator

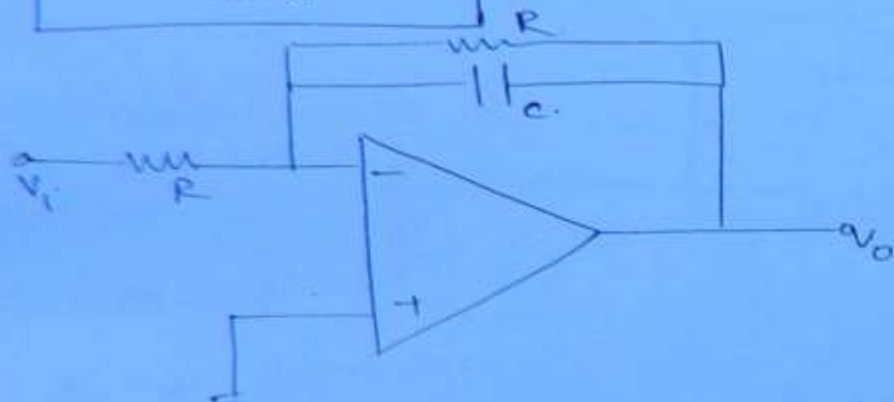


$$V_o = -V_c$$

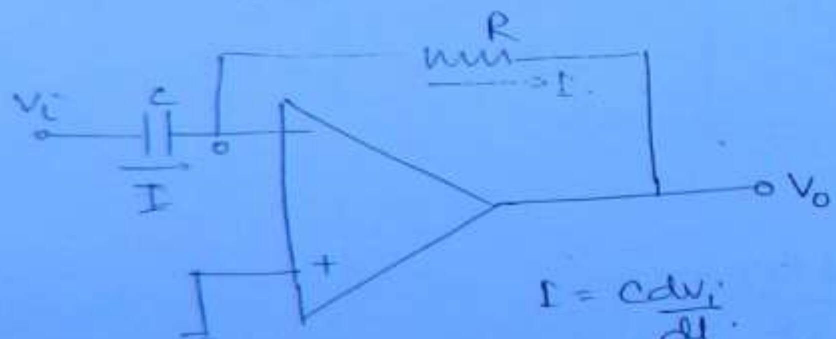
$$V_c = \frac{1}{C}\int i\,dt$$

$$\boxed{V_o = -\frac{1}{RC}\int V_i\,dt}$$

RC → time const.



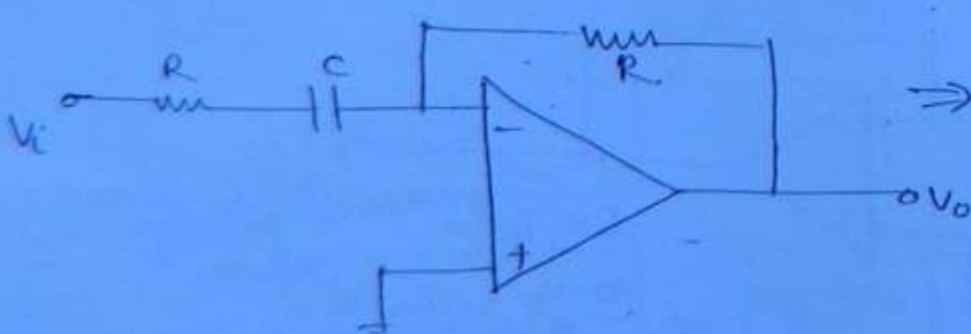
R is added to make the stability more.



$$I = C\frac{dV_i}{dt}$$

$$\frac{0 - V_o}{R} = C\frac{dV_i}{dt}$$

$$\boxed{V_o = -RC\frac{dV_i}{dt}}$$



⇒ differentiator.

Integrator →

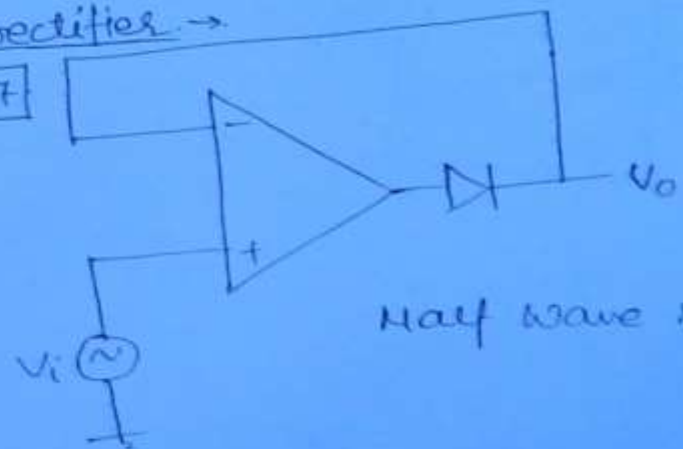
For practical Integrator, we take resistor shunt to the capacitor because at low freq, capacitor becomes O.C. which make the opamp to work in open loop system.

Differentiator:—

At the I/P resistor is in the series with the capacitor because at high freq capacitor acts as a S.C. high current flow through the device which creates a problem.

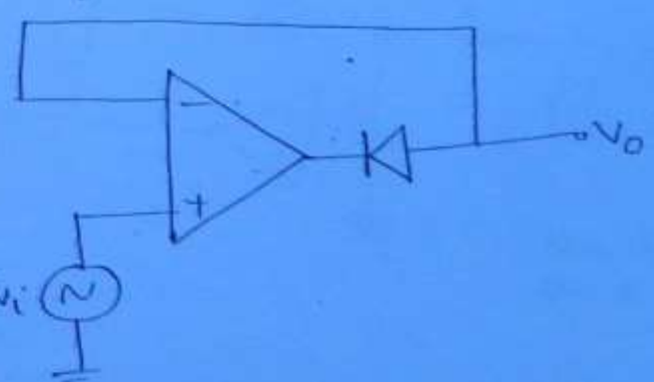
Precision rectifier →

when $\boxed{V_i < 0.7}$



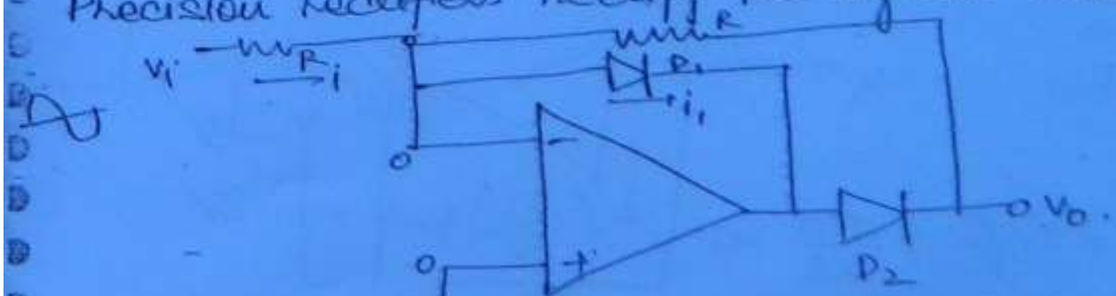
+ve half cycle, D is FB.

Half wave rectifier.



for -ve half cycle, D is R.B.

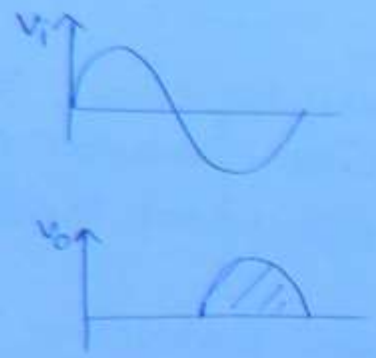
Precision rectifiers rectify the signals below 0.7 Volts.

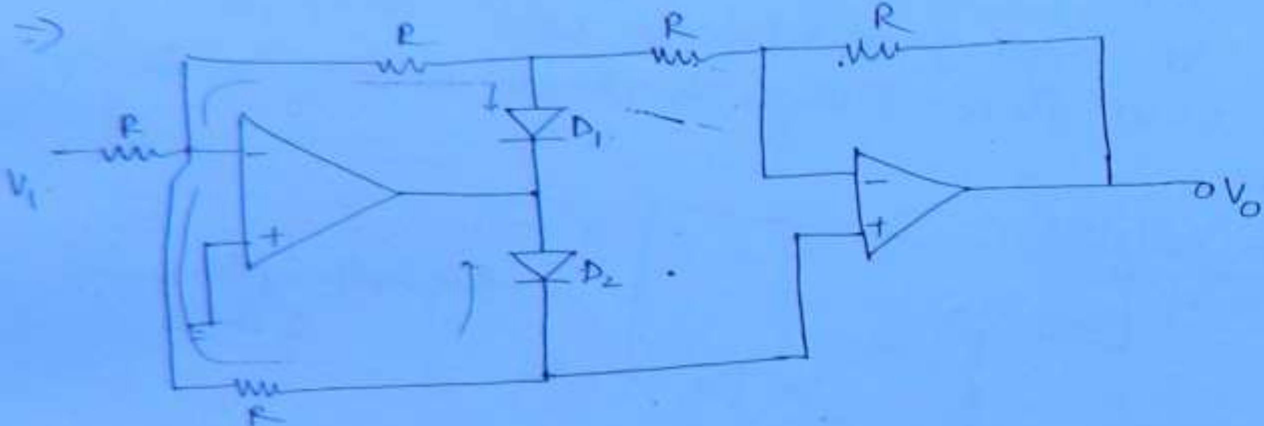


for +ve half, ~~of Vo~~ D_1 FB, D_2 RB, $V_o = 0$

fo -ve half, D_1 RB D_2 FB → $V_o = -V_i$

228

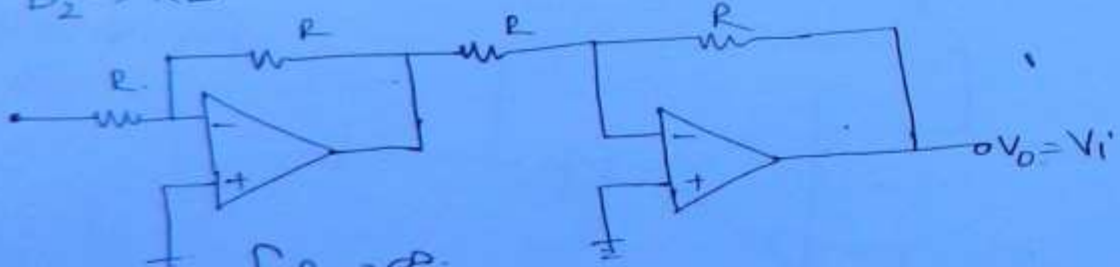




+ve half cycle →

$D_1 \to F.B.$

$D_2 \to RB.$

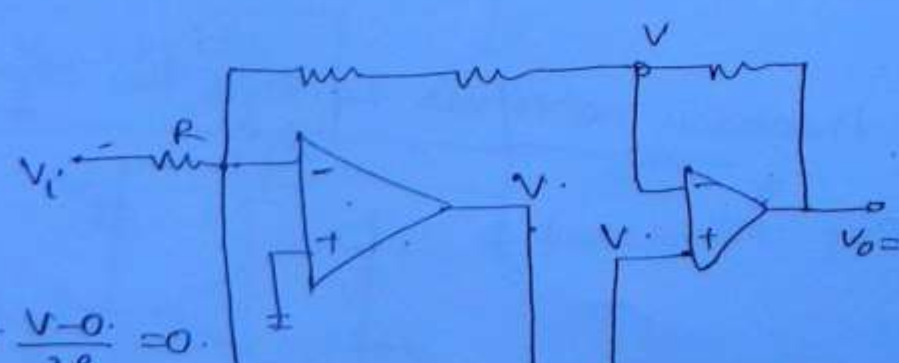


$R \to \infty$.
$I = 0$.

-ve half cycle →

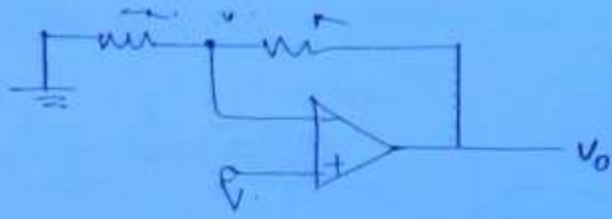
$D_1 \to R.B$

$D_2 \to FB$

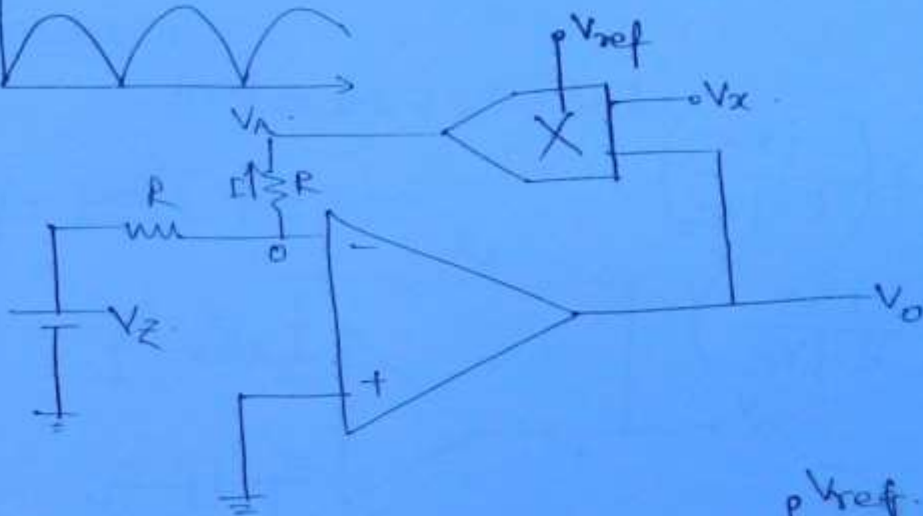


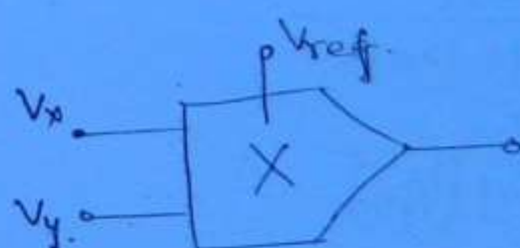
$$\frac{V_i - 0}{R} + \frac{V - 0}{R} + \frac{V - 0}{2R} = 0$$

$$V = -\frac{2}{3} V_i$$



$$V_0 = \left(1+\frac{R}{2R}\right) V$$
$$= \frac{3}{2} V$$
$$= \frac{3}{2} \times \left(\frac{-2}{3}\right) V_i$$
$$= -V_i$$





$$V_0 = \frac{V_x V_y}{V_{ref}}$$

$$V_A = \frac{V_x V_0}{V_{ref}}$$

$$\Rightarrow -IR = \frac{V_x V_0}{V_{ref}}$$

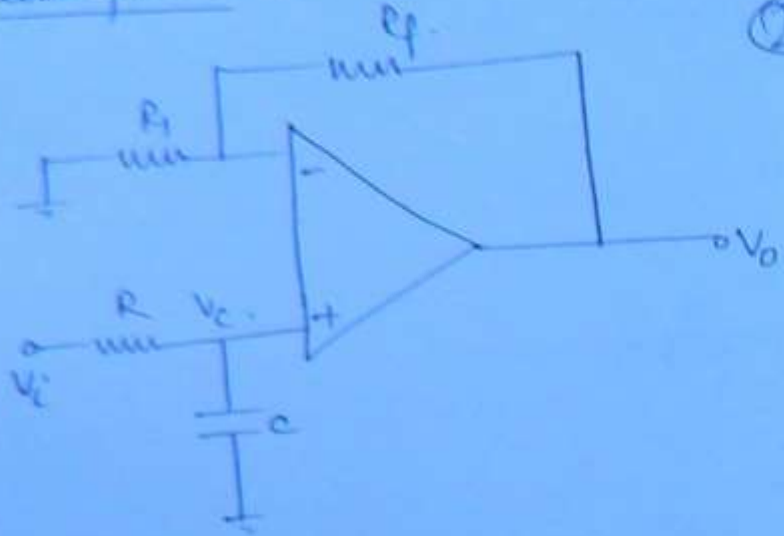
$$\Rightarrow -V_Z = \frac{V_x V_0}{V_{ref}}$$

$$V_0 = -V_{ref}\left(\frac{V_Z}{V_x}\right)$$

Active filters →

low pass filters →

230



$$V_o = \left(1+\frac{R_f}{R_1}\right) V_c$$

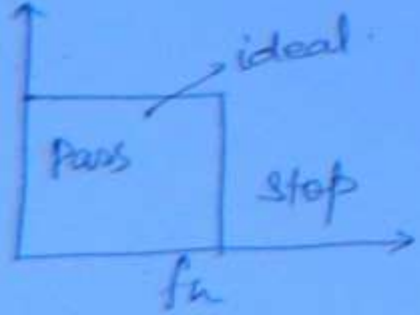
$$= \left(1+\frac{R_f}{R_1}\right) V_i \times \left(\frac{-jX_c}{R-jX_c}\right)$$

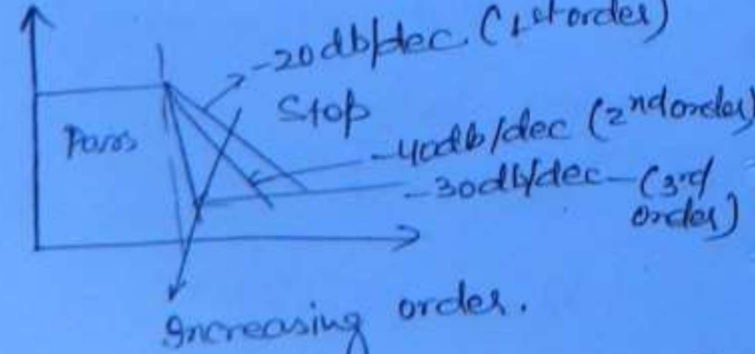
$$\frac{V_o}{V_i} = \left(1+\frac{R_f}{R_1}\right)\left(\frac{-jX_c}{R-jX_c}\right)$$

$$\frac{V_o}{V_i} = \frac{A_{CL}}{1+j2\pi fRC}$$

$$= \frac{A_{CL}}{1+j\,(f/f_H)}$$

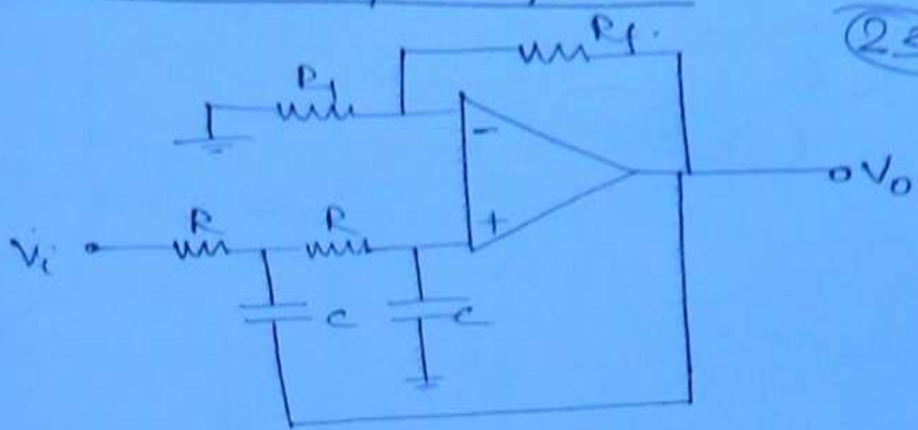
$$f_H = \frac{1}{2\pi RC}$$





2nd order low pass filter →

(231)



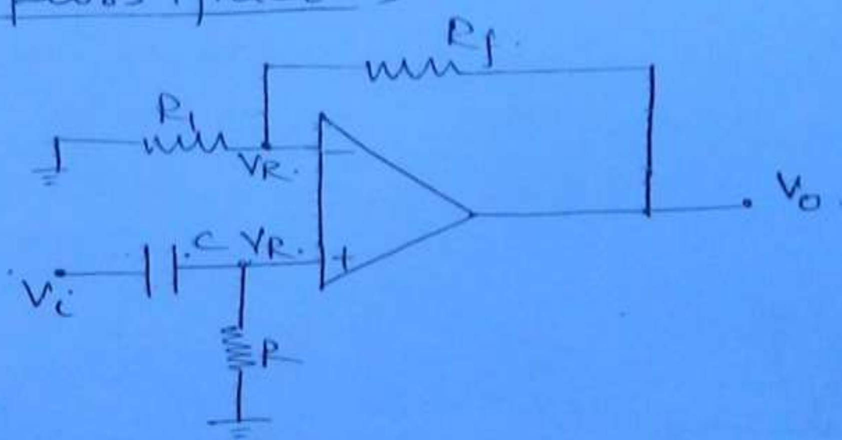
3rd order →

2nd order + 1st order.

4th order →

2nd order + 2nd order.

High pass filter →

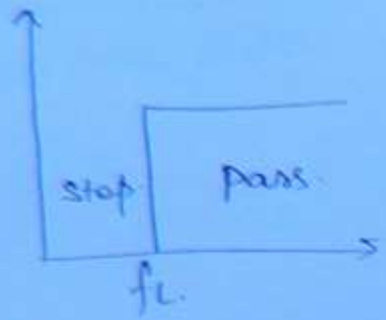


$$V_o = \left(1+\frac{R_f}{R_1}\right) V_R$$

$$= \left(1+\frac{R_f}{R_1}\right) V_i \left(\frac{R}{R-jX_C}\right)$$

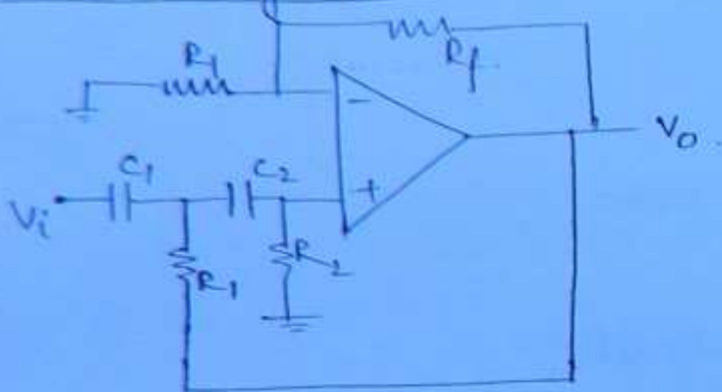
$$= \frac{A_{CL}}{1-j(f_L/f)}$$

$$f_L = \frac{1}{2\pi RC}.$$

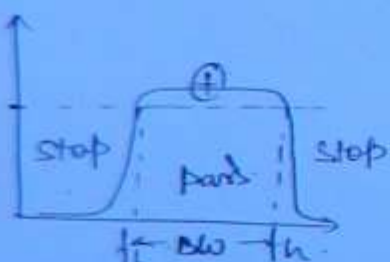


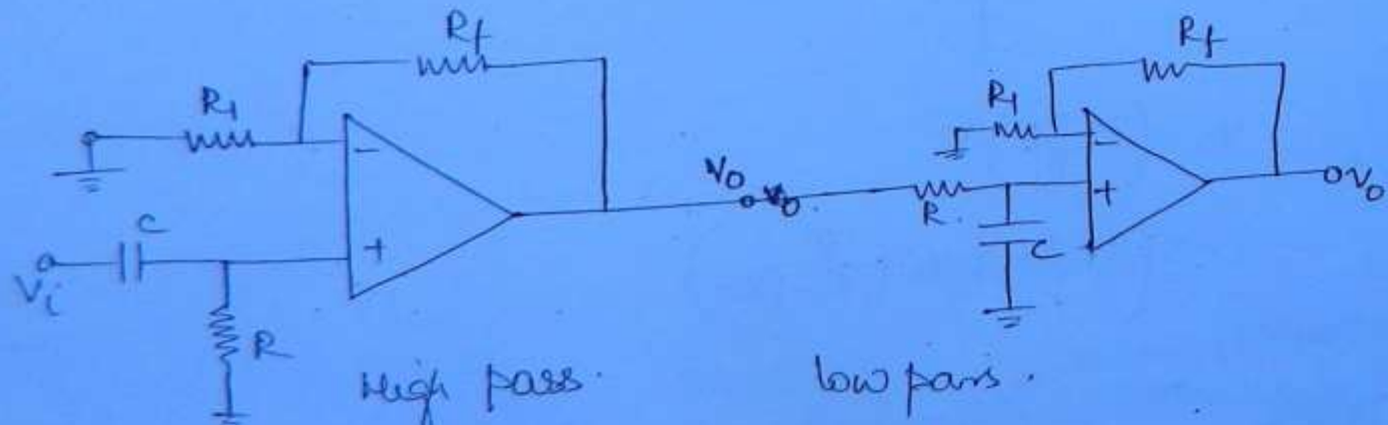
(232)

2nd order high pass →



Band pass →



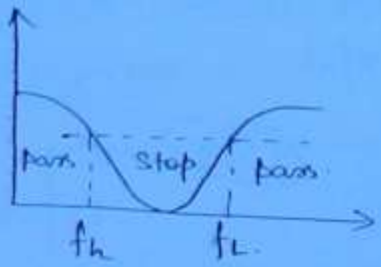


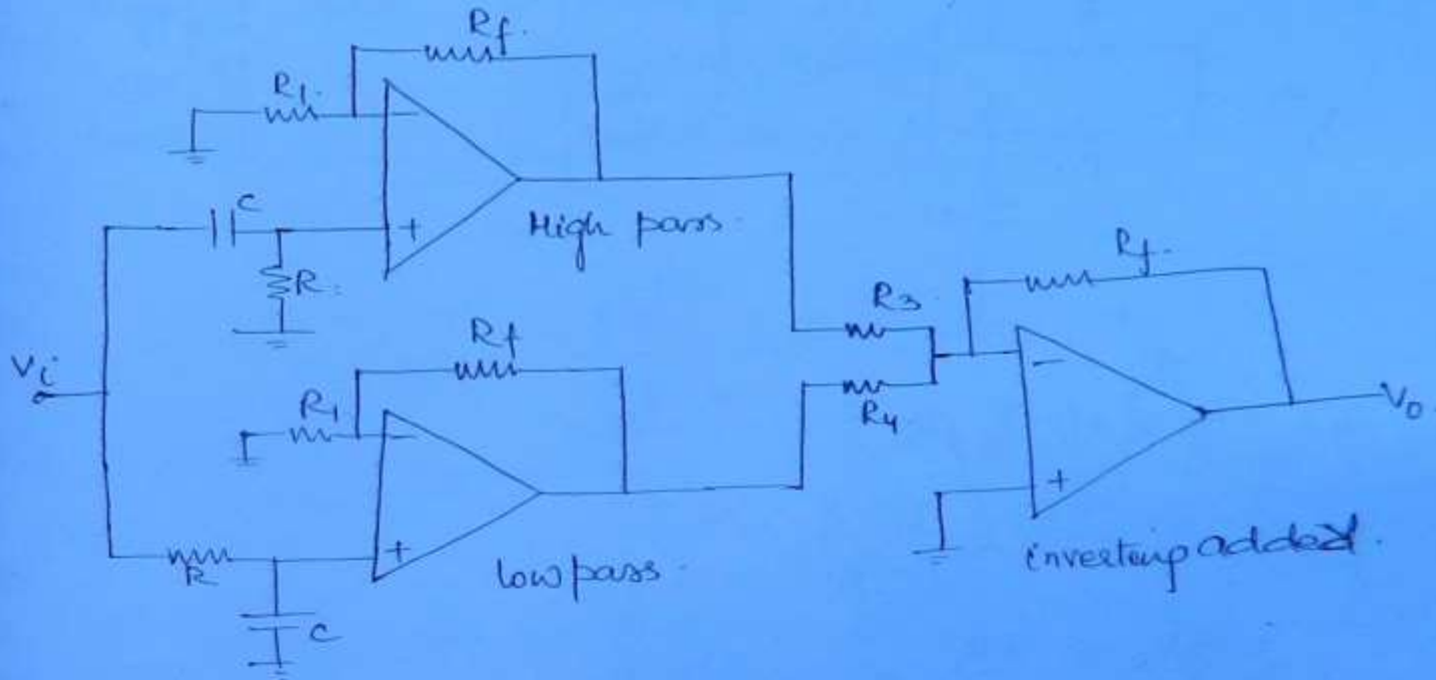
High pass.

low pass.

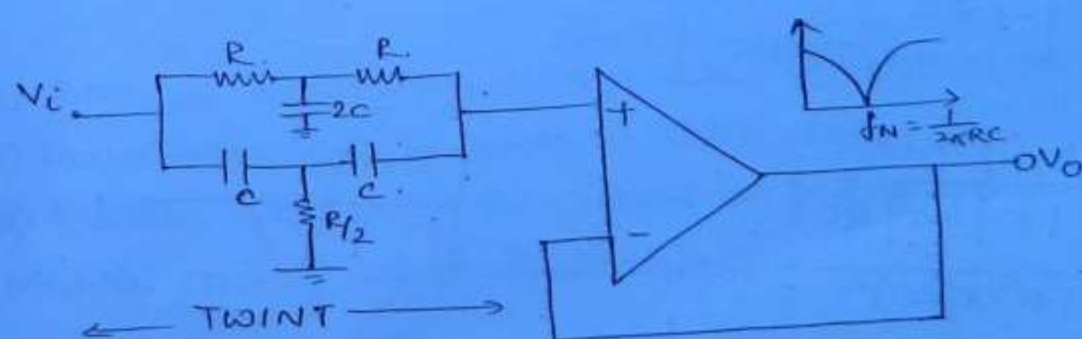
Band reject →

233

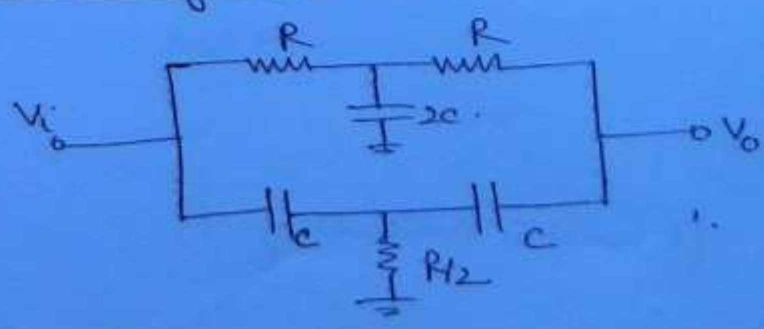


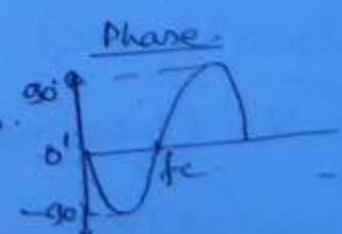


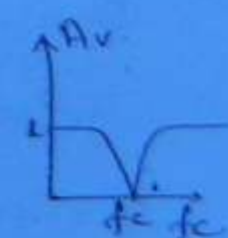
Notch filter →



Ckt diagram →

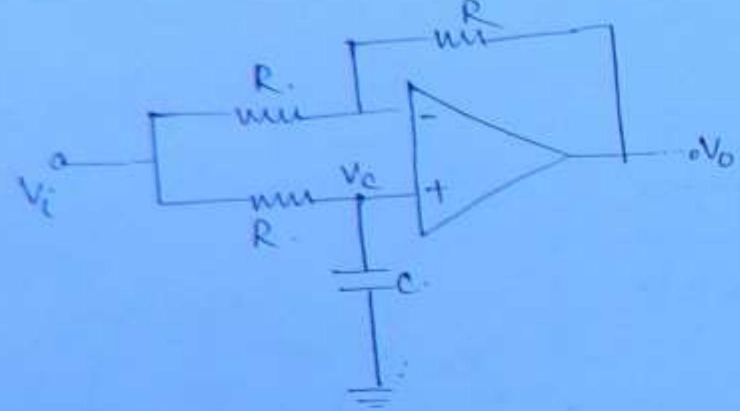






> A voltage follower n/w is added to the front n/w because to improve the quality factor of RC n/w.

All pass filter →



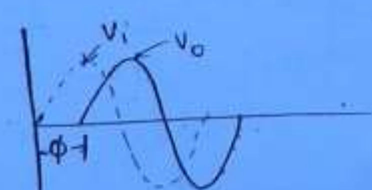
$$V_o = -\frac{R}{R} V_i + \left(1+\frac{R}{R}\right) V_c$$

$$= -V_i + 2V_i \left(\frac{-jX_c}{R-jX_c}\right)$$

$$= -V_i + \frac{2V_i}{1+j2\pi fRC}$$

$$= V_i \left[-1 + \frac{2}{1+j2\pi fRC}\right]$$

$$\frac{V_o}{V_i} = \frac{1-j2\pi fRC}{1+j2\pi fRC}$$



$$\left|\frac{V_o}{V_i}\right| = \frac{\sqrt{1^2+(2\pi fRC)^2}}{\sqrt{1^2+(2\pi fRC)^2}} = 1$$

$$\phi = -2\tan^{-1}(2\pi fRC)$$

↓

V_o lags V_i (low pass)

$\phi = +2\tan^{-1} 2\pi fRC$. (R and C is interchanged)

↓

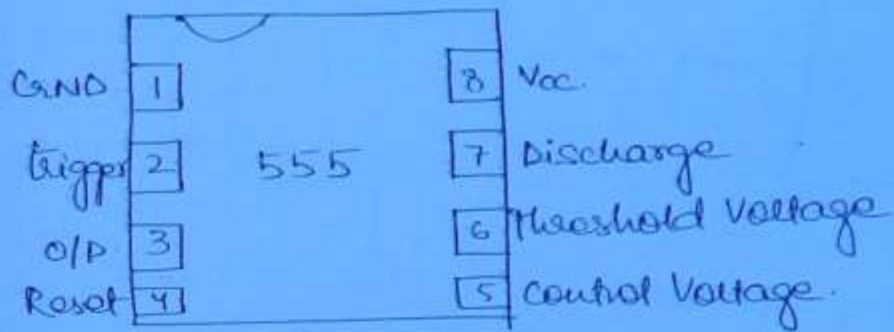
V_o leads V_i (high pass)

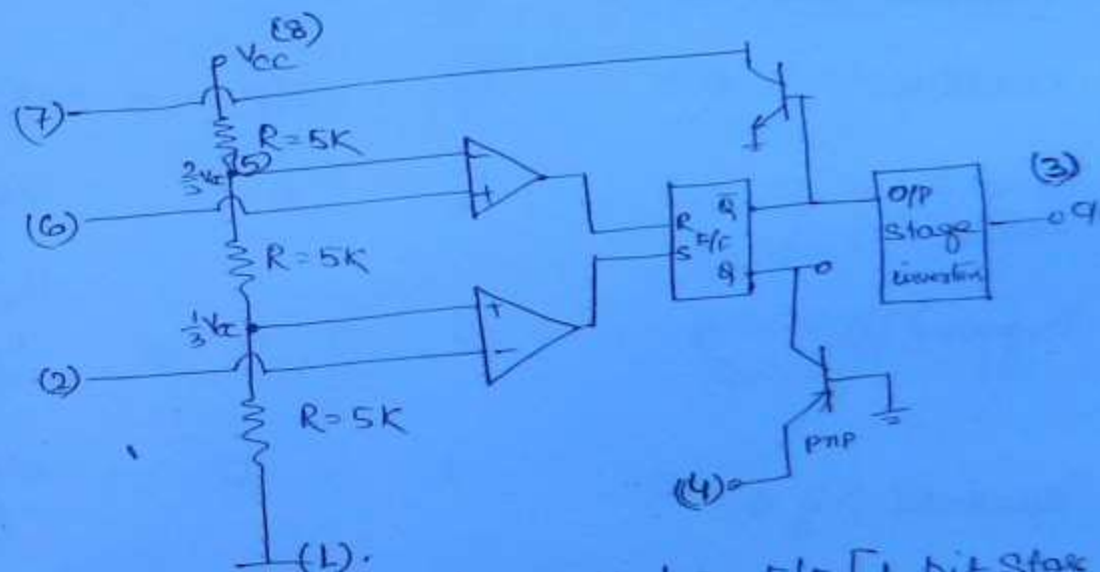
555 timer.

Whenever time delay is required.

(235)

Pin diagram →





1) Bistable multivibrator can also be called as F/F [1 bit Stor

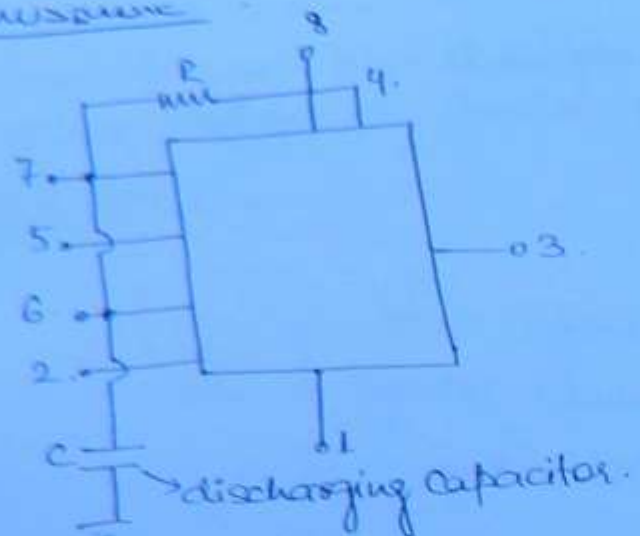
2) Monostable can generate a pulse O/P.

3) Astable can also be called as ~~p~~ free running oscillator which can generate oscillations.

555 timer operating modes :—

- monostable mode
- Astable mode

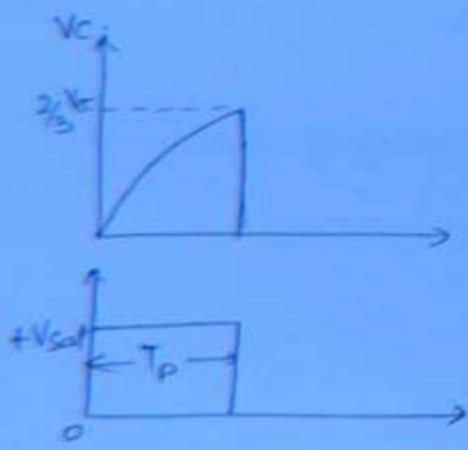
Monostable



(236)

tabular form →

condition	Q	$\bar{Q}$	o/p	capacitor position
stand by mode	0	1	0	discharging (Stable) [because transistor is in sat]
Trigger $< \frac{1}{3}V_{cc}$	1	0	1	Charging (Quasi).
threshold $> \frac{2}{3}V_{cc}$	0.	1	0.	discharging (Stable).



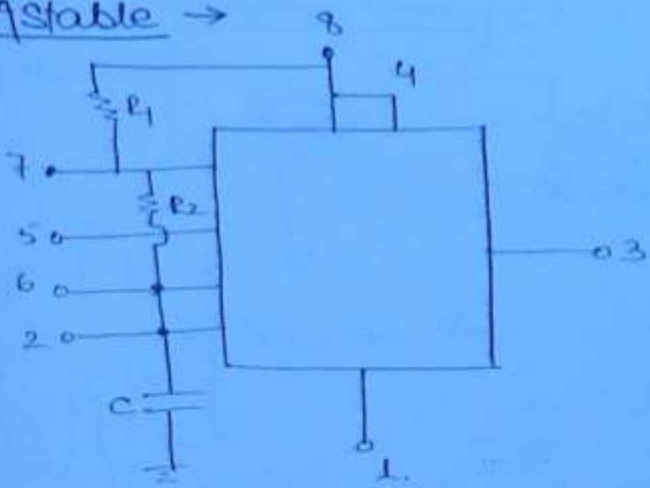
$$V_c = V(1-e^{-t/RC})$$

$$\frac{2}{3}V_{cc} = V_{cc}(1-e^{-t_p/RC})$$

R will be fixed always = ?.

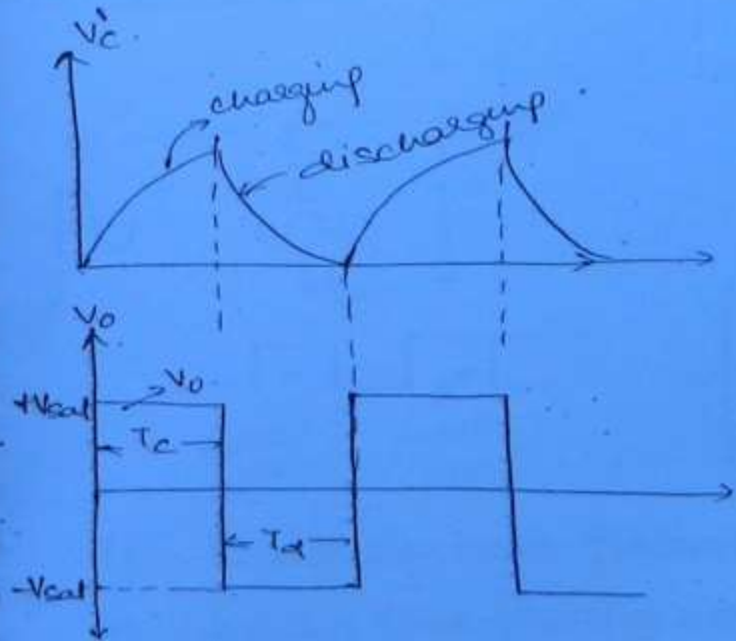
$$\boxed{t_p = 1.1RC}$$

Astable →



tabular form →

condition	Q	$\bar{Q}$	O/P	capacitor position
standby	0	1	0	discharging (stable)
trigger $< \frac{1}{3} V_{cc}$ (capacitor)	1	0	1	charging (Quasi)
threshold $> \frac{2}{3} V_{cc}$	0	1	0	discharging (Quasi)



$T_c = 0.69(R_1+R_2)C.$

$T_d = 0.69 R_2 C.$

$T = T_c + T_d = 0.69(R_1+2R_2)C.$

Duty cycle $= \frac{T_c}{T_c + T_d}$

238

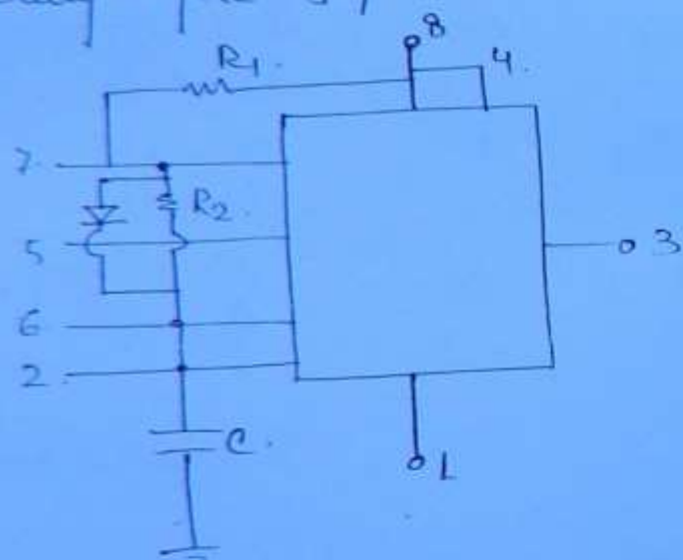
$$= \frac{0.69\,(R_1+R_2)C}{0.69\,(R_1+2R_2)C}$$

$$= \frac{R_1+R_2}{R_1+2R_2}$$

$$f = \frac{1}{T} = \frac{1}{0.69(R_1+2R_2)C}$$

Q Design a stable multivibrator to generate square wave oscillation?

for duty cycle = 50%.



$$D.C. = \frac{R_1}{R_1+R_2}$$

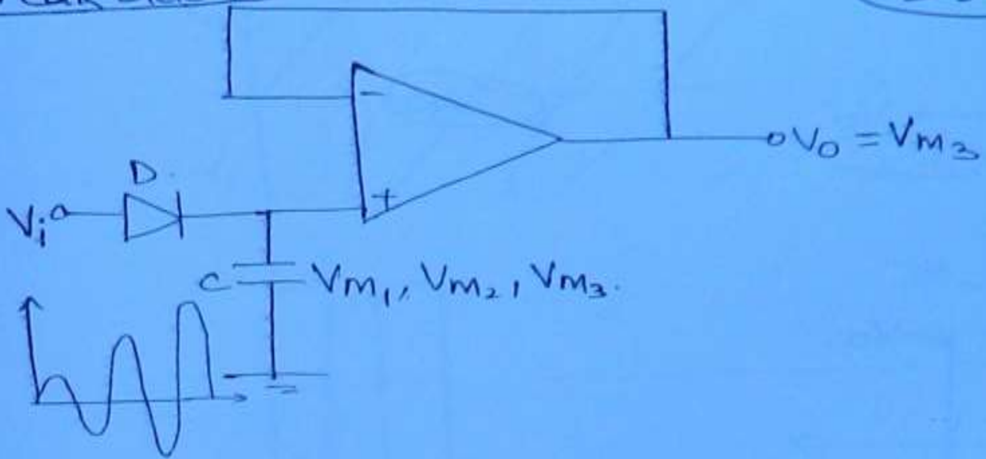
$R_1 = R_2 = R$

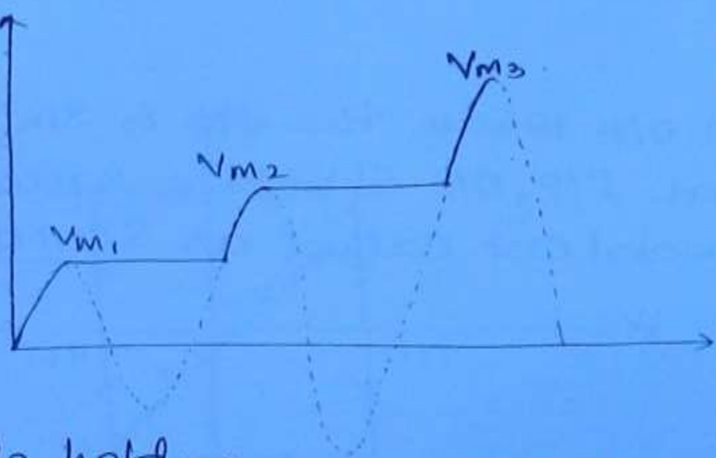
D.C. = 50%.

Special opamp design.

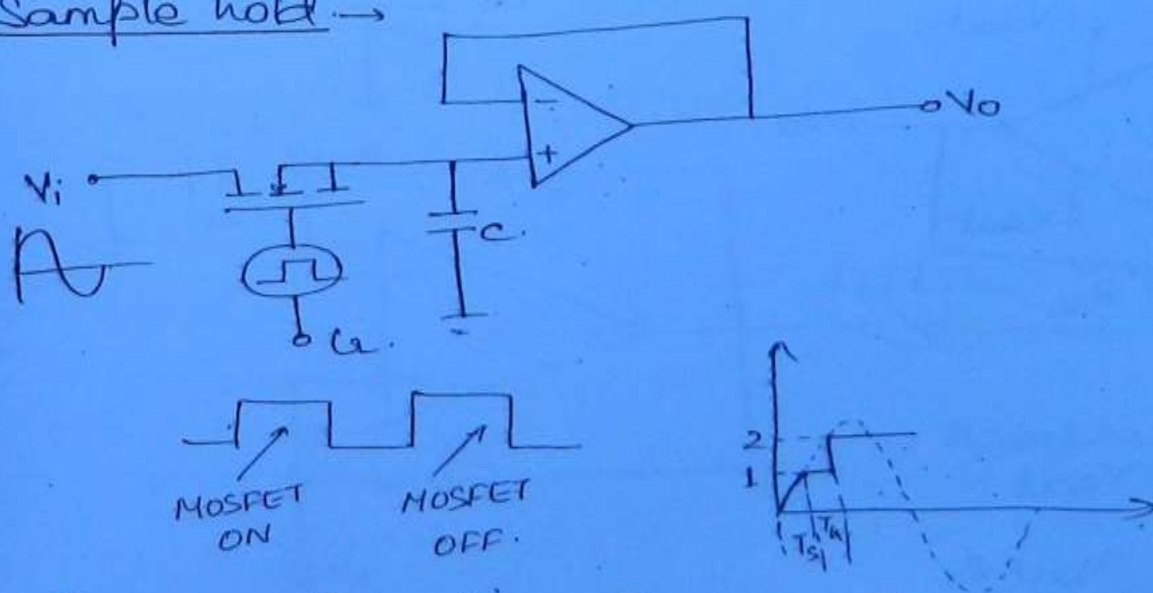
239

Peak detector





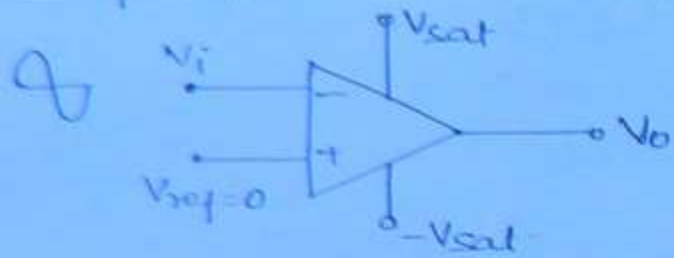
Sample hold →



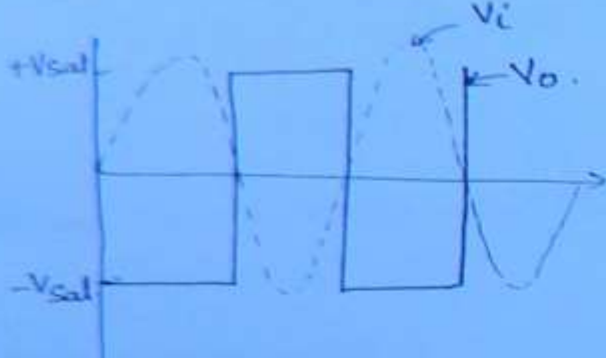
T_S → sampling time [MOSFET ON]

T_h → holding time [MOSFET OFF].

Comparator → ZCD → Zero crossing detection

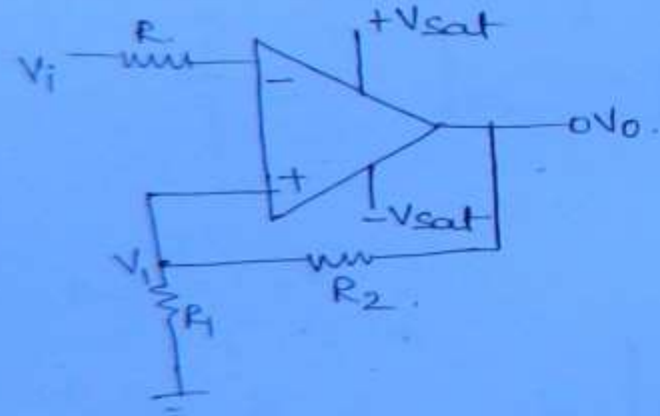


(240)



ZCD can give square wave O/P when the I/P is sinusoidal signal. But for any irregular I/P, O/P should be square wave. It is possible only by special ckt called as SCHMITT TRIGGER.

SCHMITT TRIGGER →



Assume $V_o = +V_{sat}$.

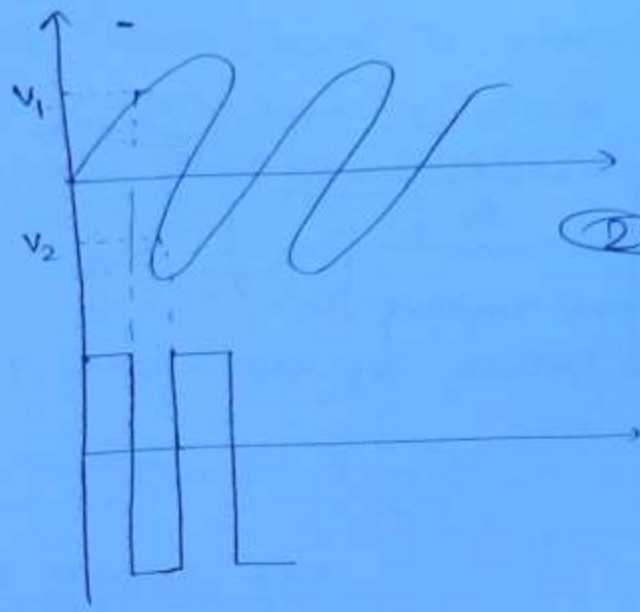
[UTP] Upper trigger pt design

$$V_1 = \frac{V_{sat} R_1}{R_1 + R_2}$$

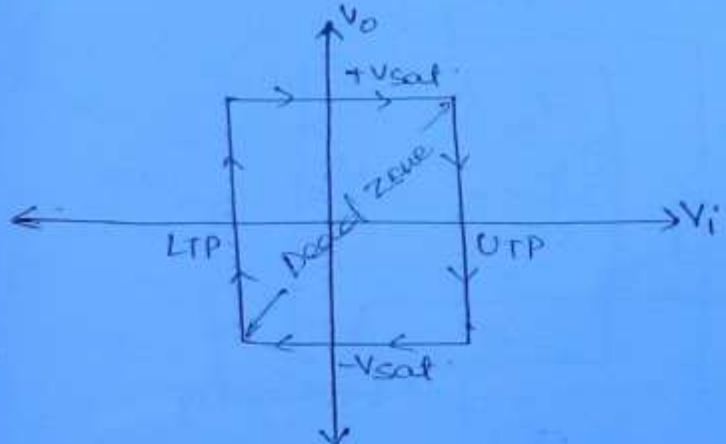
$$V_o = -V_{sat}$$

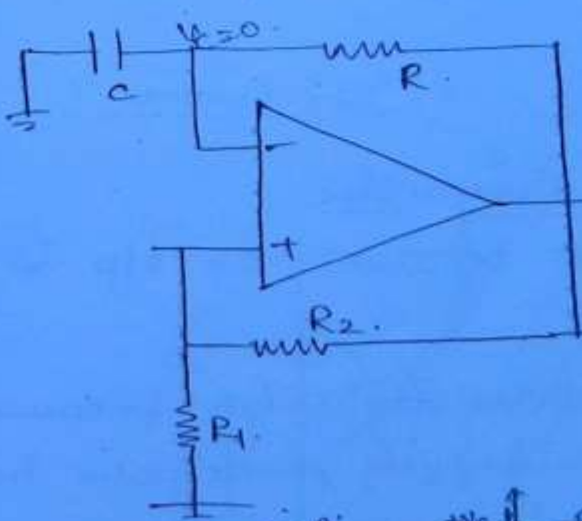
LTP

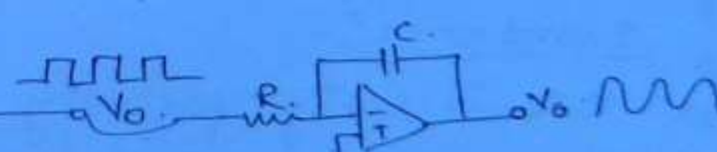
$$V_2 = \frac{-V_{sat} \times R_1}{R_1 + R_2}$$



(241)

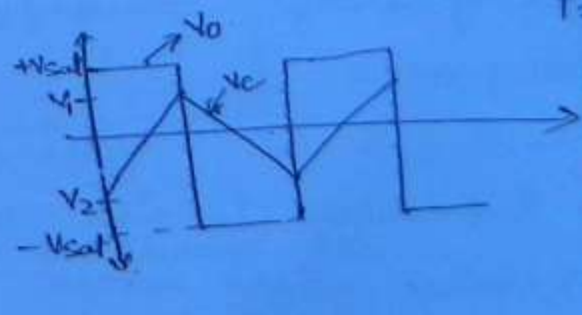






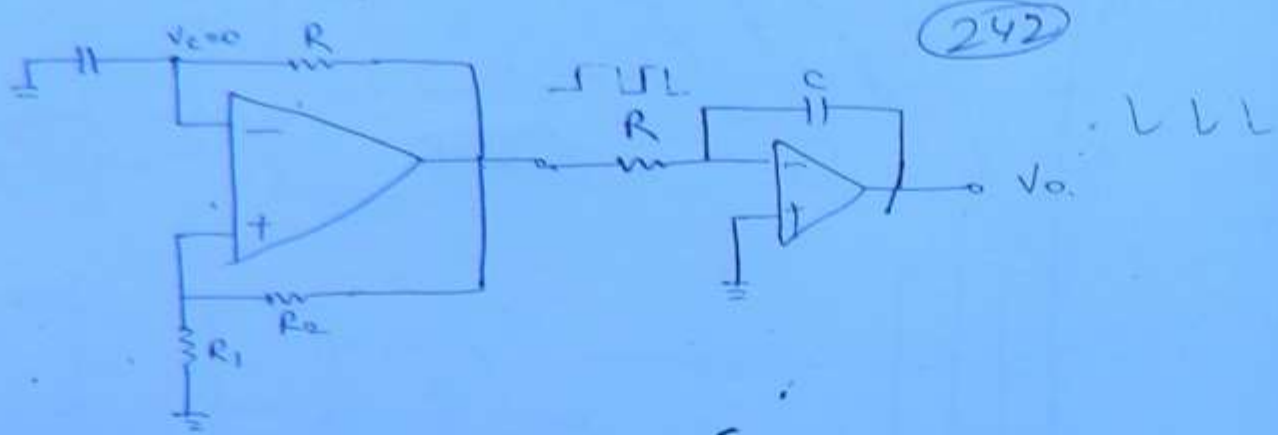
⇒ Astable multivibrator.

$V_C > UTP$, discharging



$$T = 2RC \ln\left(\frac{2R_1 + R_2}{R_2}\right)$$

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



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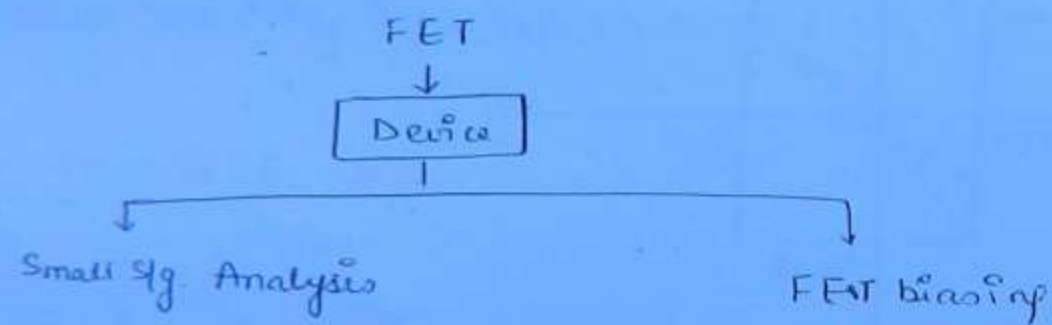
FET :-



Introduction :-

BJT is having following disadvantage.

i). The I/p impedence is low because the I/p is forward bias.

ii). The noise level is comparatively high because both type of charge carriers will participates in the conduction process.

The advantages of FET compared to BJT :-

- The I/p impedence is high because the I/p is always operated with reverse bias.

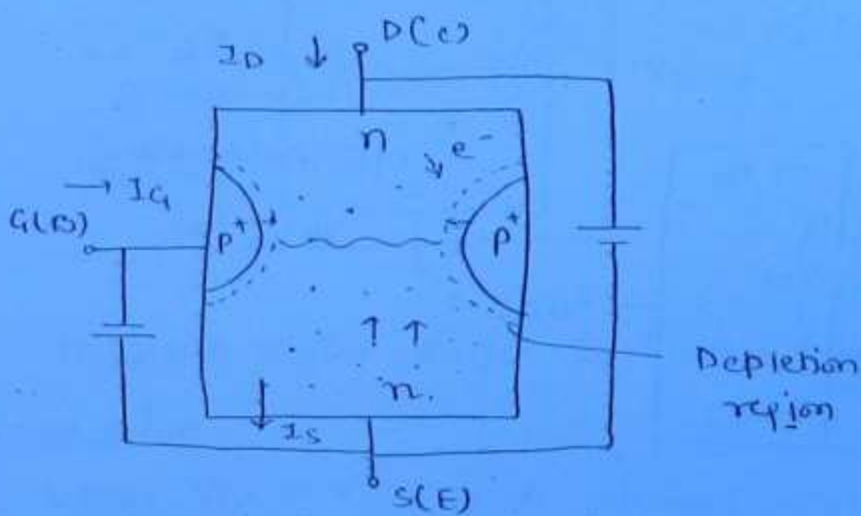
ii). The noise level is comparatively low because only one type of charge carrier will participate in the conduction process.

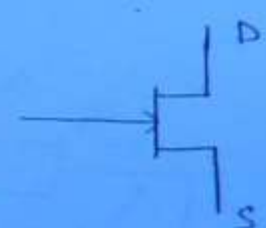
Note:-

BJT is a current control device whereas FET is a voltage control device.

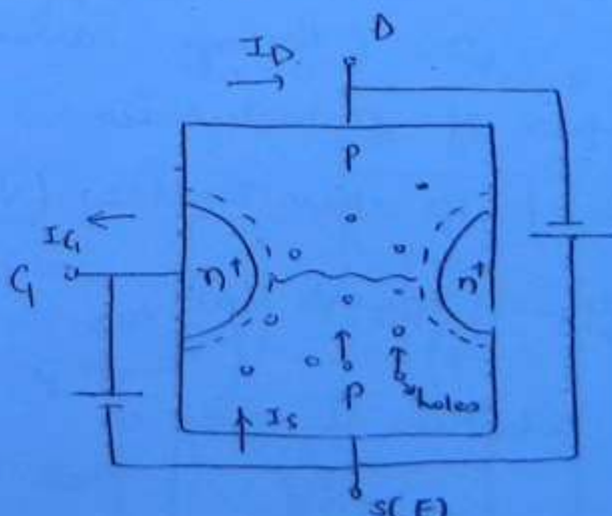
Types of FET:-

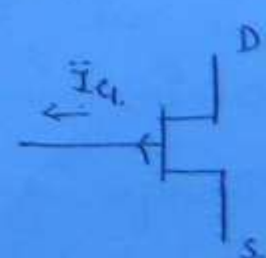
1) n channel FET.





2) P channel FET:-



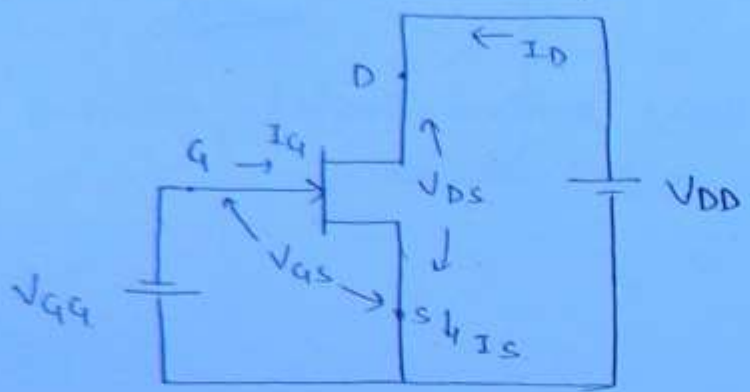


Note:- Drain should be always RB.

Circuit symbols and notations :-

1). <u>N- Channel FET</u>

(244)



$V_{GG} = -ve$

$\left.\begin{matrix} I_D \\ V_{DS} \end{matrix}\right\} = +ve.$

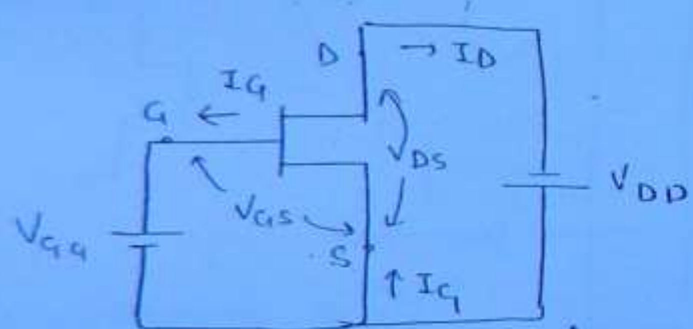
I/p parameters -

V_{GS}, $I_G = 0$.

O/p parameter

V_{DS}, I_D.

2). <u>P-Channel FET :-</u>



<u>Conclusion :-</u>

1). There will be no I/p characteristics in FET because I_G is approximately $= 0$. (leakage current)

2). There are two types of characteristics.

i). Drain characteristics / o/p characteristics $(V_{DS}\ V_s\ I_D)|_{V_{GS}}$.

ii). Transfer characteristics $(V_{GS}\ V_s\ I_D)|_{V_{DS}}$.

In N-channel the polarities are given as.

$V_{GS} = -ve$

$\left.\begin{matrix} I_D \\ V_{DS} \end{matrix}\right\} = +ve.$

In P- Channel the polarities are given as-

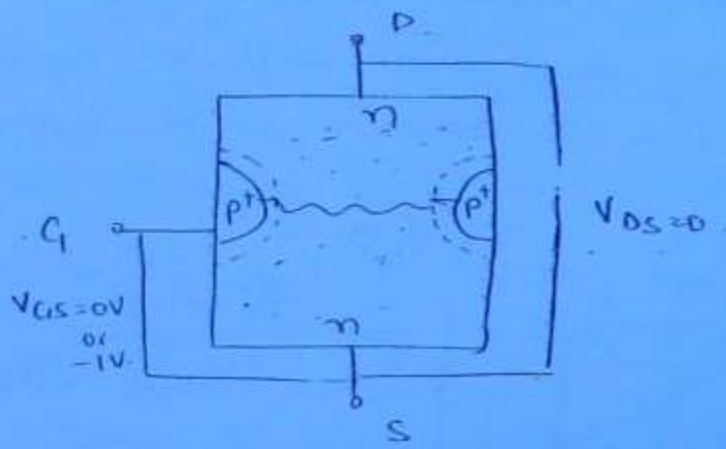
$V_{GS} = +ve$ (245)

$\left.\begin{array}{l} I_D \\ V_{DS} \end{array}\right\} = -ve.$

Working principle of FET :-

Case-1 :-

$V_{GS} = 0V, \quad V_{DS} = 0\,V.$



Depletion region layer is ↑. then region ↓

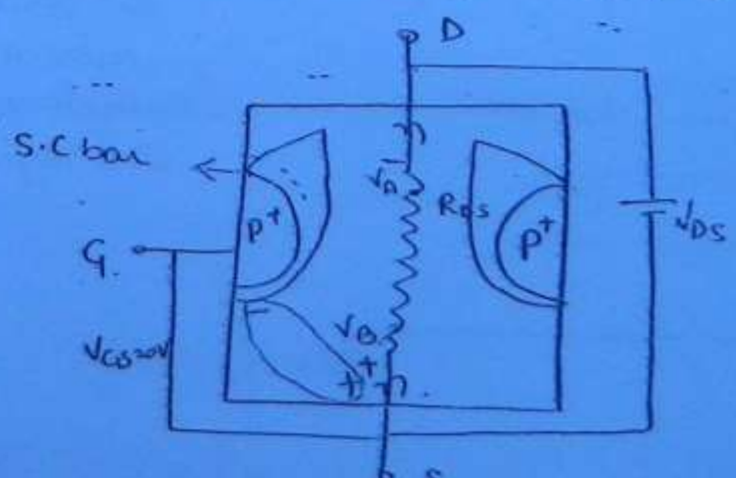
Conclusion :-

1) When $V_{GS} = 0V$ max. channel can be achieved

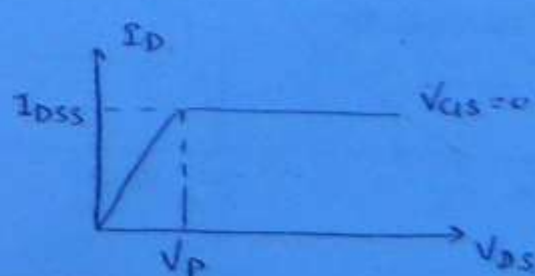
2) when $V_{DS} = 0V$ Drain current I_D becomes 0

Case-2 :-

$V_{GS} = 0\,V, \quad V_{DS} \neq 0\,V.$



$V_{DS} \propto I_D$



I_{DSS} - Short ckt Drain current

Conclusions:

1) When $V_{GS} = 0V$ max. channel can be achieved.

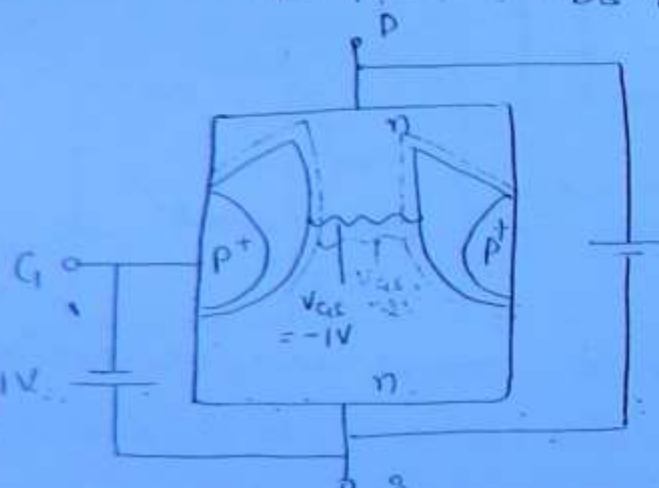
2) When $V_{DS} \neq 0V$ the S.C bar will act like a resistor. ($V_{DS} \propto I_D$)

(246)

3) The basic mechanism of FET is top side of the drain is covered with R. Bias & bottom side of the source is covered with F. Bias.

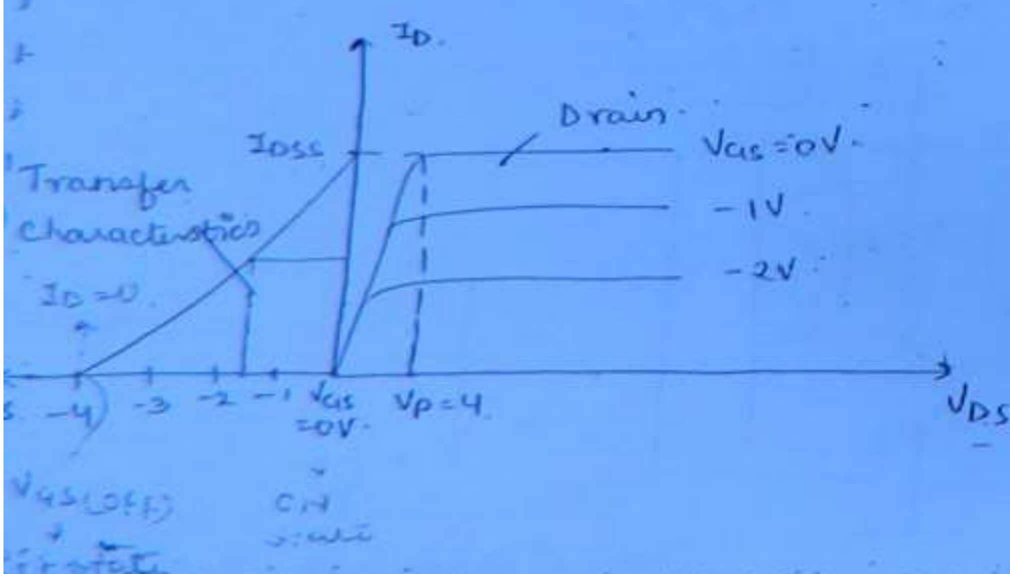
4) Due to the internal Reverse bias of gate to source the depletion region penetration will be more at the top side & less penetration at the bottom side.

Case-3:- $V_{GS} \neq 0$. $V_{DS} \neq 0$



Q point-

$(V_{GS})_Q$, $(I_D)_Q$



$(V_{GS}) = off$ means no channel current is 0. Both penetration will together

Transfer characteristics

$$\boxed{I_D = I_{DS}\left[1 - \frac{V_{GS}}{V_P}\right]^2}$$

247

$|V_{GS}|_{off} = |V_P|$

V_P = pinch off voltage & it depends on dopping concentration.

⇒ For FET device analysis we have to concentrate on transfer characteristics but not drain characteristic.

⇒ In FET Biasing Q point is defined as $[(V_{GS})_Q \ (I_D)_Q]$

FET is always thermally stable that means I_D is independent of temperature.

FET Parameters :-

ΔV_{GS}, ΔV_{DS}, ΔI_D

1). Ac drain resistance

$$r_d = \left.\frac{\Delta V_{DS}}{\Delta I_D}\right|_{V_{GS}}$$

2). Transconductance

$$g_m = \left.\frac{\Delta I_D}{\Delta V_{GS}}\right|_{V_{DS}}$$

3). Amplification factor

$$\mu = \left.\frac{\Delta V_{DS}}{\Delta V_{GS}}\right|_{I_D}$$

$$\boxed{\mu = g_m r_d}$$

$\boxed{r_d > R_D}$

Expression for transconductance:

$$g_m = \frac{\Delta I_D}{\Delta V_{GS}}$$

248

$$= \frac{\partial I_{DSS}}{\partial V_{GS}}\left[1 - \frac{V_{GS}}{V_P}\right]^2$$

$$g_m = -\frac{2 I_{DSS}}{V_P}\left[1 - \frac{V_{GS}}{V_P}\right]$$

$$\boxed{g_m = g_{mo}\left[1 - \frac{V_{GS}}{V_P}\right]}$$

$$g_{mo} = -\frac{2 I_{DSS}}{V_P}.$$

↓

max. transconductance.

Prove $g_m = \frac{2}{|V_P|}\sqrt{I_D - I_{DSS}}$.

Sol:-

$$g_m = -\frac{2 I_{DSS}}{V_P}\left[1 - \frac{V_{GS}}{V_P}\right]$$

$$\sqrt{\frac{I_D}{I_{DSS}}} = \left[1 - \frac{V_{GS}}{V_P}\right]$$

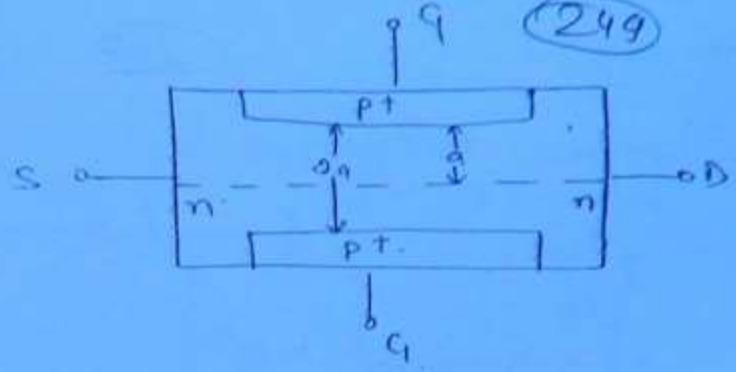
$$g_m = -\frac{2 I_{DSS}}{V_P}\sqrt{\frac{I_D}{I_{DSS}}}$$

$$g_m = \frac{2}{|V_P|}\sqrt{I_D - I_{DSS}}$$

Pinch off voltage:

$$V_P = \left|\frac{a^2 q N_A}{2\epsilon}\right| = \left|\frac{a^2 q N_D}{2\epsilon}\right|$$

↓ P channel ↓ n channel.



a = half channel height

2a = full channel height

Problems:

Small S/g Analysis of FET

Low freq Analysis

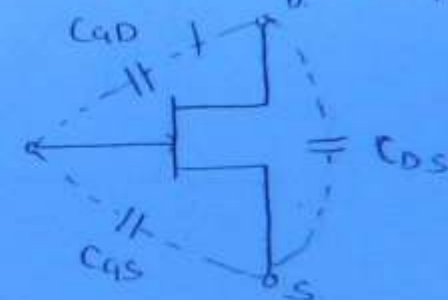
$I_D = f(V_{GS}, V_{DS})$

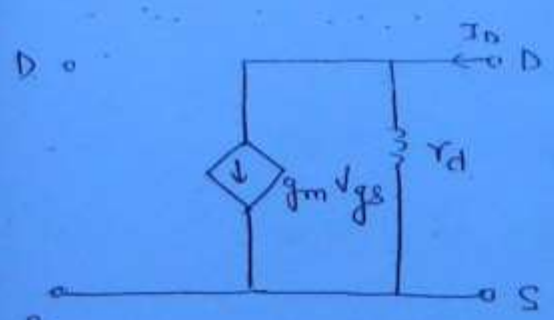
$I_D = \frac{\partial I_D}{\partial V_{GS}} + \frac{\partial I_D}{\partial V_{DS}} \cdot V_{DS}$

$= g_m V_{GS} + \frac{V_{DS}}{r_d}$

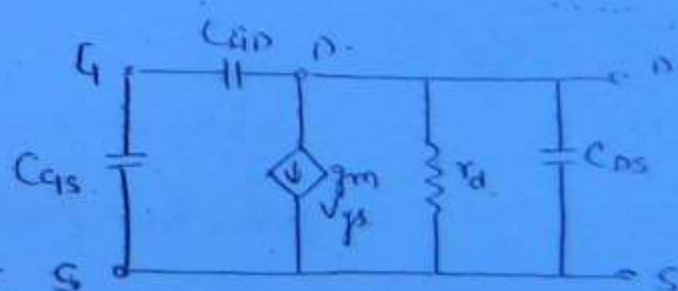
High freq Analysis

interelectrode Capacitance





Low freq.



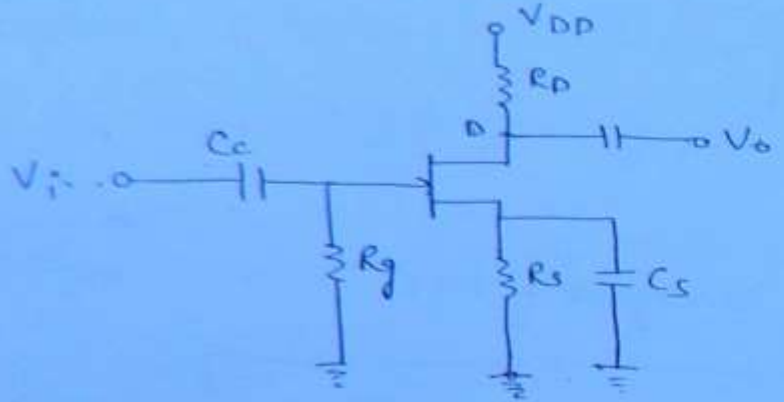
High freq.

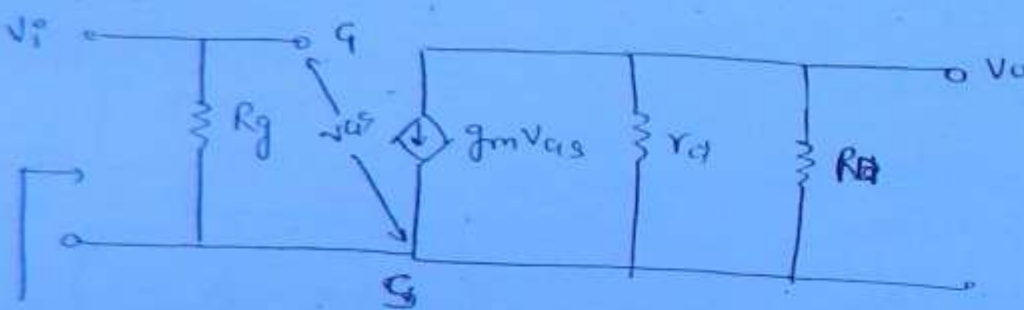
Conclusion :-

⇒ Ampr analysis should be done in mid band range therefore the high freq. model of FET when it is represented in the mid band range it converts

like a low freq model. (All the Interelectrode capacitances are open).

Common Source Bypass Amplifier:





Z_o:

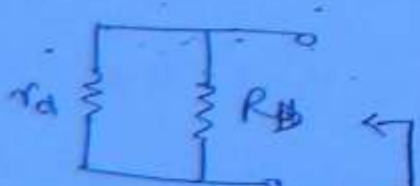
$V_i = 0$

$\downarrow$

$V_{gs} = 0$

$\downarrow$

$g_m V_{gs} = 0$ (open)

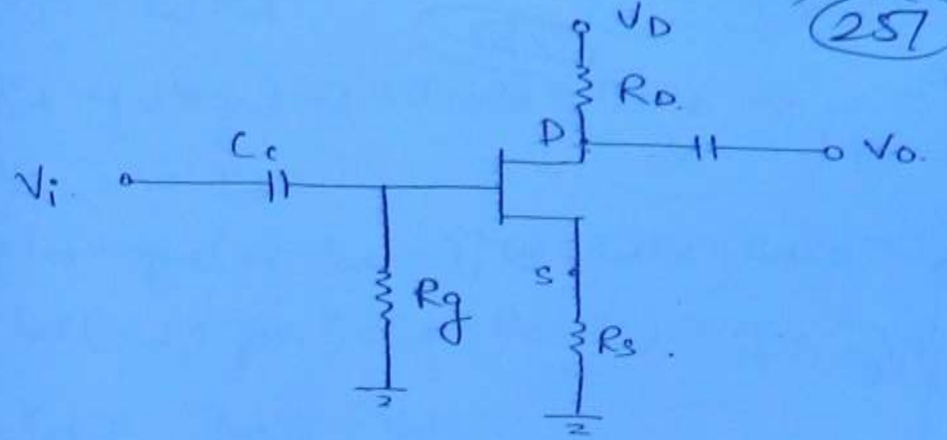


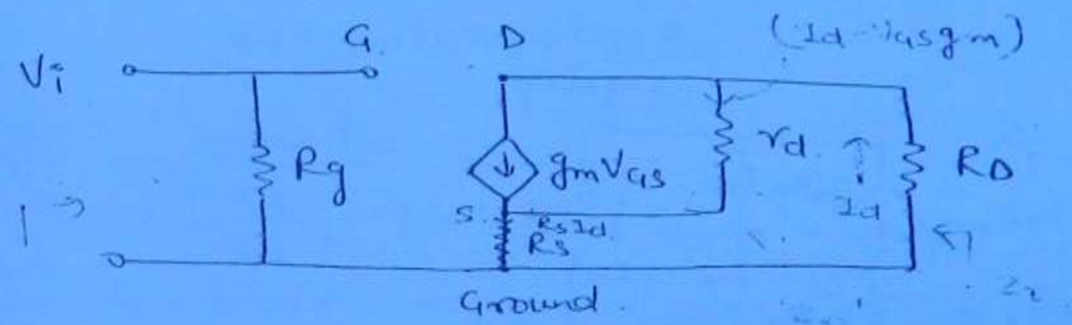
$Z_o = r_d \,||\, R_D$

$A_v = \frac{V_o}{V_i}$

$V_o = -g_m V_{gs}\, r_d \,||\, R_D \Rightarrow \boxed{\frac{V_o}{V_i} = -g_m\, r_d \,||\, R_D}$

Common source unbypass type :-





$Z_i = R_g$

Z_o :-

$$V_i = V_{GS} + I_D R_s$$

$$V_i = 0$$

$$V_{GS} = - I_D R_s$$

$$V_o = (I_d - V_{GS} g_m) r_d + R_s I_d$$

$$V_o = \{ I_d - g_m(-I_D R_s)\} r_d + R_s I_d$$

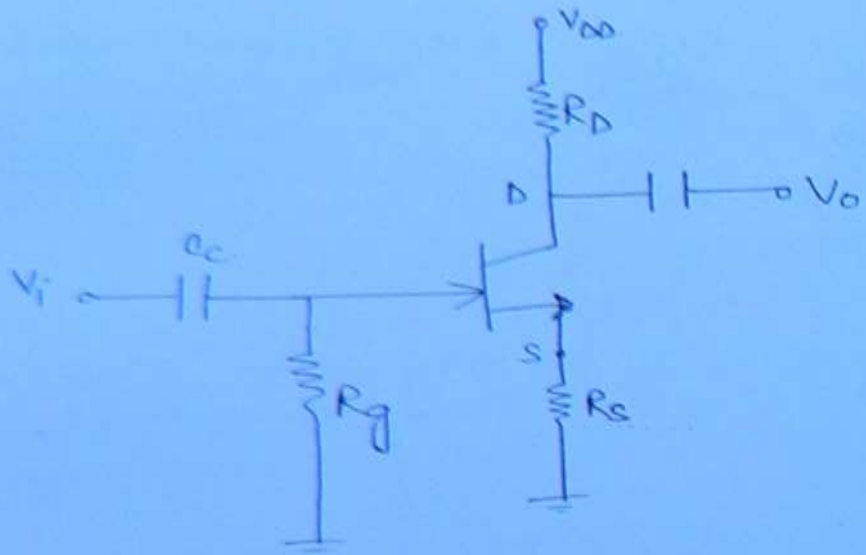
$$= I_d (r_d + g_m R_s r_d + R_s)$$

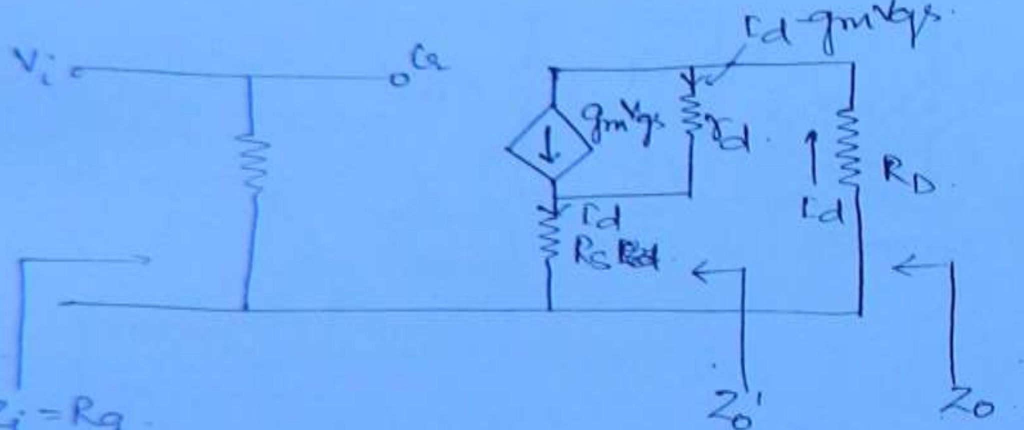
$$\boxed{Z_o' = \frac{V_o}{I_d} = r_d (1+\mu) R_s}$$

$$Z_o = Z_o' \| R_D$$

$$= \{ r_d + (1+\mu) R_s \} \| R_D$$

CS Unbypass →





$Z_i = R_g$

Z_o →

$V_i = V_{GS} + I_D R_S$

$V_i = 0$

$V_{GS} = -I_D R_S$

$V_o = (I_d - g_m V_{gs}) r_d + I_d R_s$

$= \{I_d - g_m(-I_D R_S)\} r_d + I_d R_s$

$= I_d \{(r_d + g_m R_s r_d) + R_s\}$

$$\boxed{Z_o' = \frac{V_o}{I_d} = r_d + (1+\mu) R_s}$$

$$\boxed{Z_o = Z_o' \parallel R_D}$$

$A_v =$

$V_o = -I_d R_d$ 253

$(I_d - g_m V_{gs}) r_d + I_d R_s + I_d R_D = 0$

$V_{GS} = V_i - I_d R_s$

$\{I_d - g_m(V_i - I_d R_s)\} r_d + I_d R_s + I_d R_D = 0$

$\Rightarrow I_d (r_d + g_m R_s r_d + R_s + R_D) = g_m V_i r_d$

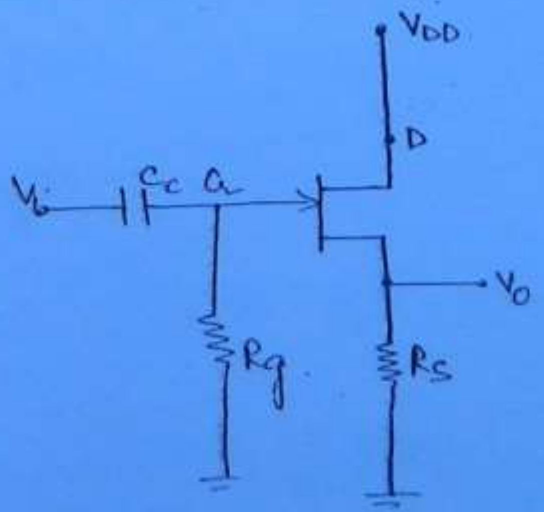
$$\Rightarrow I_d = \frac{g_m V_i + r_d}{r_d + g_m R_s r_d + R_s + R_D}$$

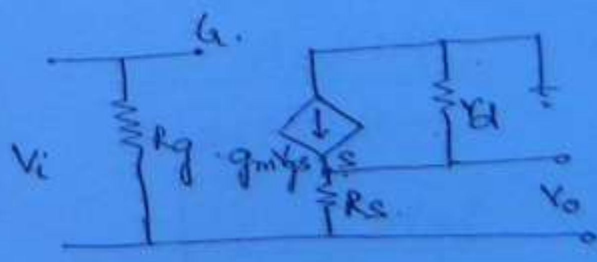
$$\frac{V_o}{V_i} = \frac{-g_m R_D}{\frac{r_d}{r_d} + g_m R_s \frac{r_d}{r_d} + \frac{R_s + R_D}{r_d}} \leftarrow \text{neglected}$$

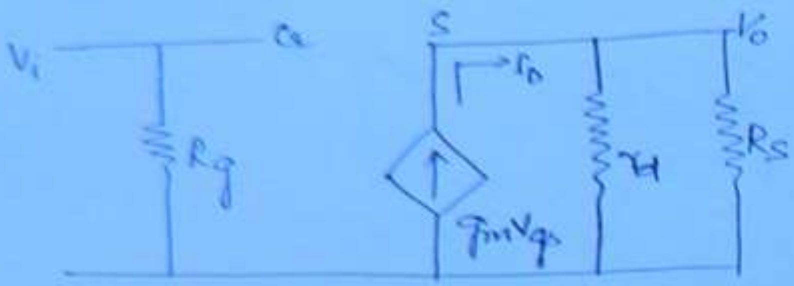
$r_d >> R_s + R_D$

$$\boxed{A_v = \frac{-g_m R_D}{1 + g_m R_s}}$$

Common drain amplifier :—







254

1) $Z_i = R_g$

2) Z_o :-

$V_i = -V_{GS} + V_O$

$\Rightarrow V_i + V_{GS} = V_O$

if $V_i = 0$

$V_{GS} = V_O$ source is at higher potential than gain.

$I_d = g_m V_{gs} \rightarrow V_o$

$Z_o' = \frac{V_o}{I_d} = \frac{1}{g_m}$

$Z_o = Z_o'' // R_S$

$= \frac{1}{g_m} // R_S$

$A_V = \frac{V_o}{V_i}$

$V_o = -g_m V_{gs} r_d // R_S$

$V_i = -V_{gs} + V_o$

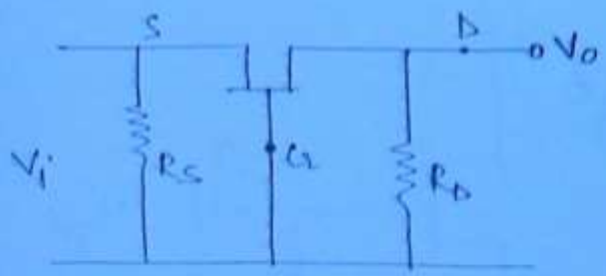
$= -V_{gs} - g_m V_{gs} r_d // R_S$

$\frac{V_o}{V_i} = \frac{-g_m V_{gs} R_s}{-V_{gs}(1 + g_m R_s)}$

$$\boxed{A_V = \frac{g_m R_s}{1 + g_m R_s}}$$

Common gate amplifier.

255

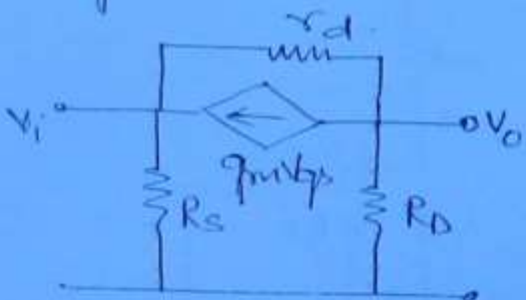


$R_i = \frac{1}{g_m} \| R_S$

$R_o = R_D \| r_d$

~~$R_o =$~~

$A_V = g_m R_D$



Conclusion →

	R_i	R_o	A_V
CS bypass	R_g	$r_d \| R_d$	$-g_m R_D$
CS unbypass	R_g	$r_d + (1+\mu) R_d \| R_d$	$\frac{-g_m R_D}{1 + g_m R_g}$
CD	R_g	$\frac{1}{g_m} \| R_S$	$\frac{g_m R_S}{1 + g_m R_S}$
CG	$\frac{1}{g_m} \| R_S$	$r_d \| R_d$	$g_m R_D$

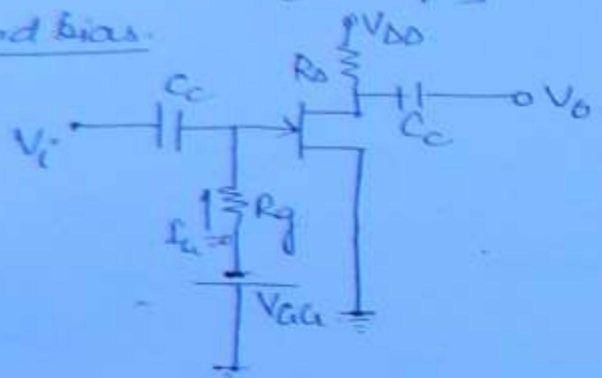
FET Biasing →

256

Condition

1) $I_G = 0$
2) $I_D = I_S$
3) $I_D = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right]^2$

fixed bias.



Calculate

$$(V_{GS})_Q = -V_{GG}$$

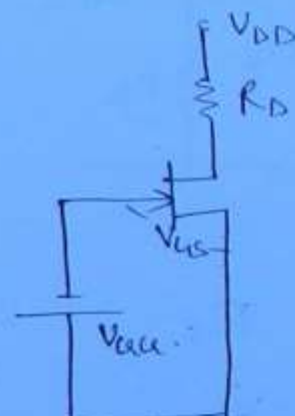
$$(I_D)_Q = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right]^2$$

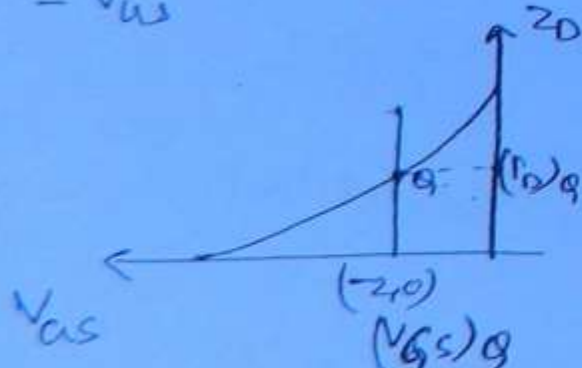
$$V_{DS} = V_{DD} - I_D R_D$$

$$V_D = V_{DS}$$

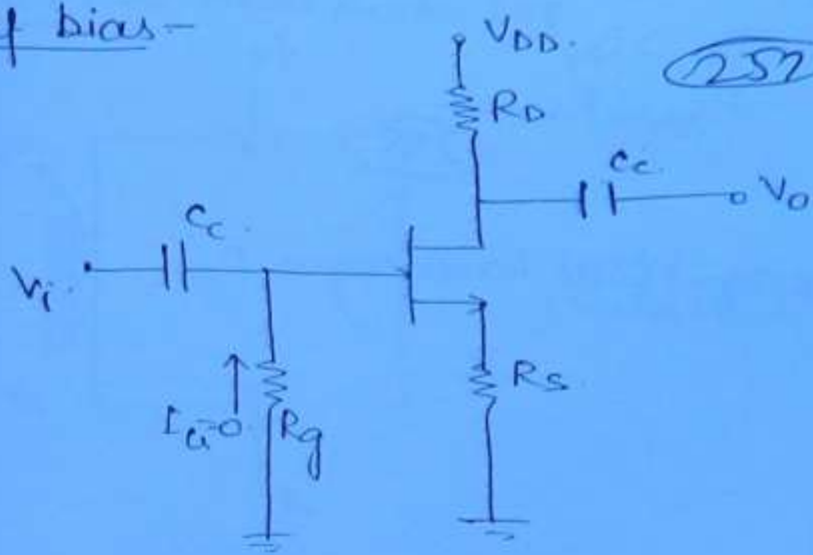
$$V_S = 0$$

$$V_G = -V_{GS}$$

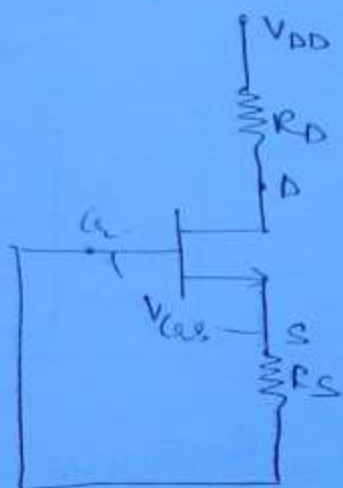




Self bias -

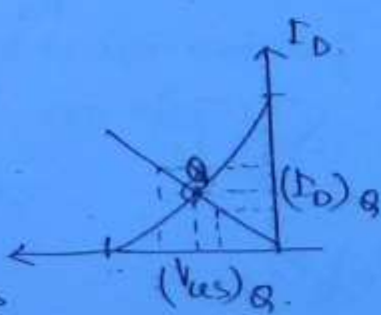


$V_{GS} = -I_D R_S$



$V_{GS} = -1V$, $R_S = 1K\Omega$ } $I_D = 1mA$

$V_{GS} = -2V$, $R_S = 1K\Omega$ } $I_D = 2mA$

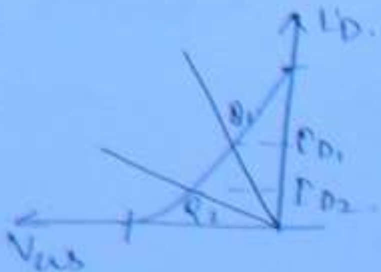


$$(I_D)_Q = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right] \qquad V_{GS} = -I_D R_S$$

$$(V_{GS})_Q = (I_D) R_S$$

$$V_{DS} = V_{DD} - I_D(R_D + R_S)$$

$$V_D = V_{DD} - I_D R_D \quad , \quad V_S = I_D R_S$$



$I_{D_1} > I_{D_2}$

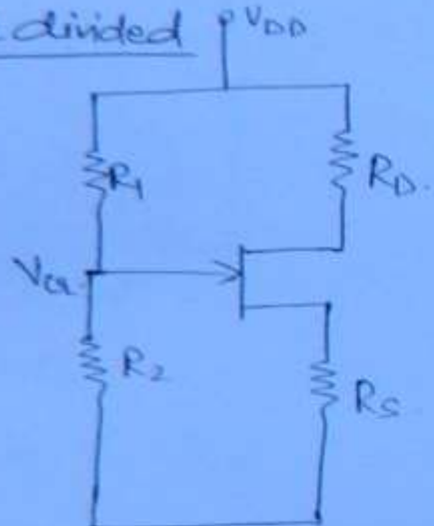
$R_{S_1} < R_{S_2}$

258

Q which transistor design will have more R_S

Ans- $R_{S_2} \rightarrow Q_2$

Voltage divided bias



$V_G = V_{GS} + I_D R_S$

$V_{GS} = V_G - I_D R_S$

$V_{GS} = 0$, $I_D = \frac{V_G}{R_S}$

$I_D = 0$, $V_{GS} = V_G$

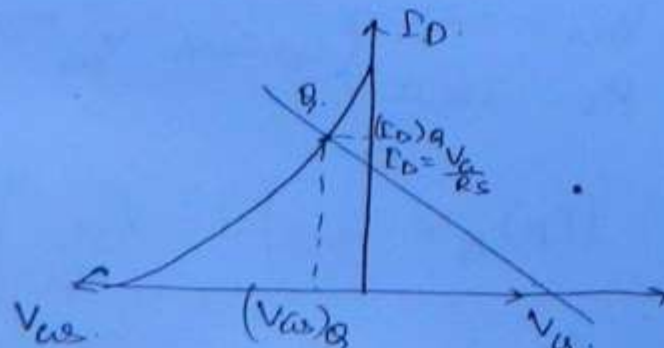
$$(I_D)_Q = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right]^2$$

$$(V_{GS})_Q = V_G - (I_D)_Q R_S$$

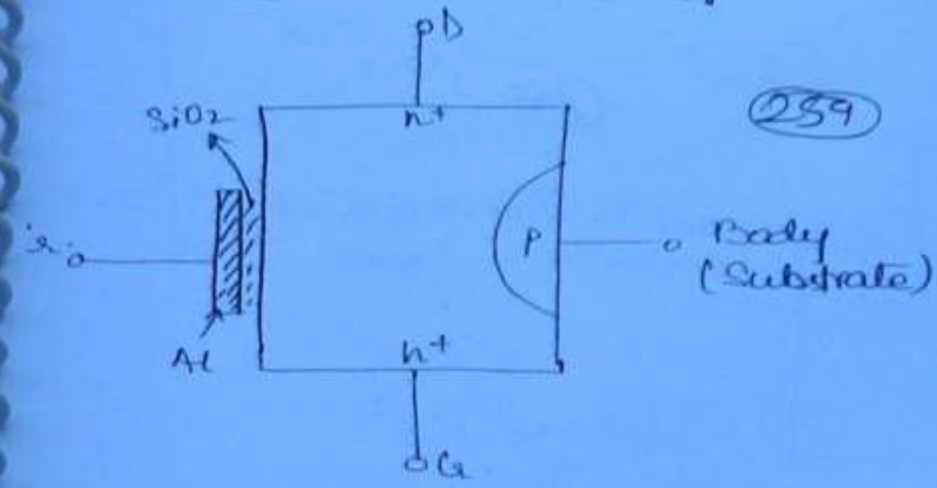
$$V_{DS} = V_{DD} - I_D(R_D + R_S)$$

$$V_D = V_{DD} - I_D R_D$$

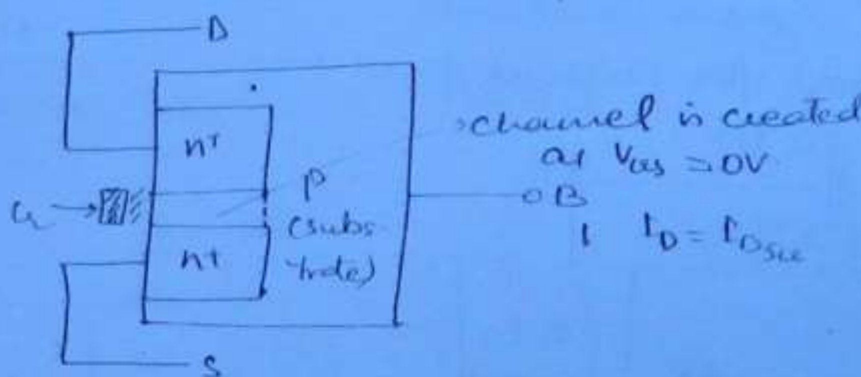
$$V_S = I_D R_S$$



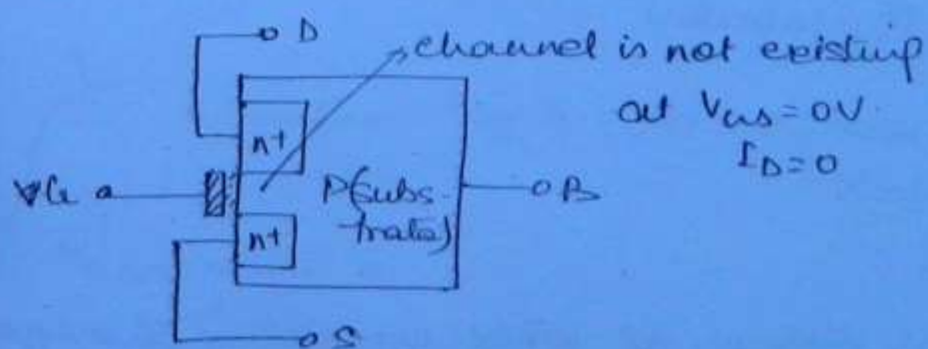
MOSFET → Metal oxide FET.



Depletion MOSFET →



Enhancement MOSFET →



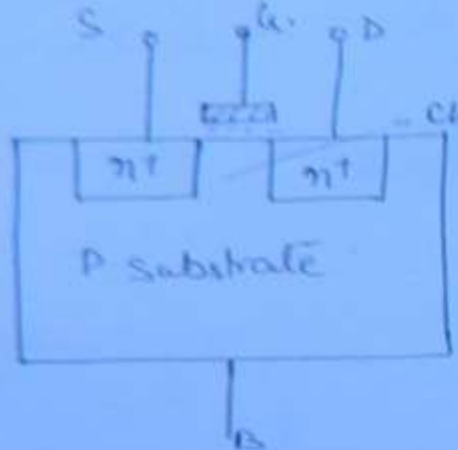
1) Depletion MOSFET analysis is same as FET analysis because channel is map. at $V_{GS} = 0\,V$.
2) In enhancement MOSFET Channel is not existing at $V_{GS} = 0V$, $I_D = 0$.
3) To analyse the MOSFET we always concentrate on enhancement mode. [ON state]

N channel enhancement MOSFET analysis.

Case 1. →

When $V_{GS} = 0V$.

260

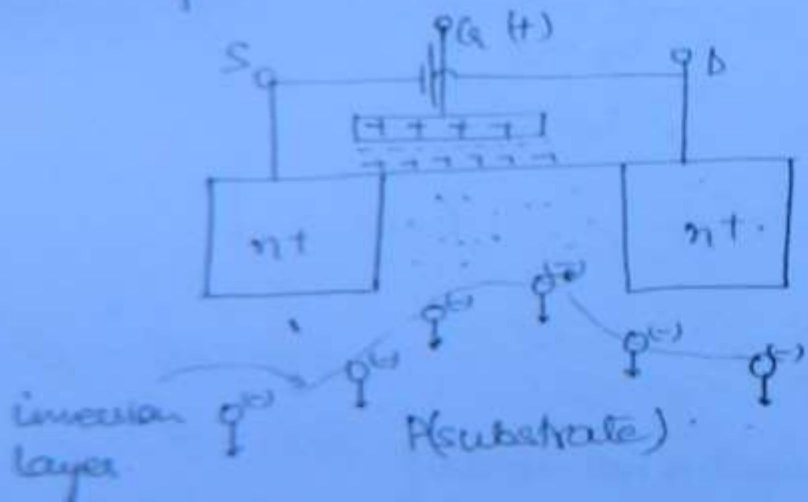


channel is not existing at $V_{GS} = 0V$

$I_D = 0$.

Case 2

Creating a channel for current flow :—



V_T → threshold voltage = 0.5

= Gate Voltage.

Threshold voltage →

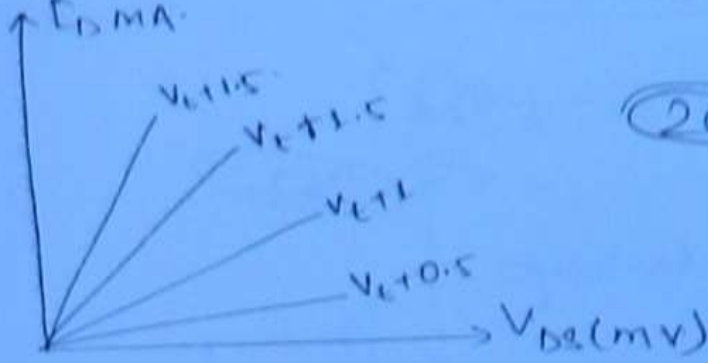
It is the mini V_{GS} Voltage at which conducting channel will be formed.

Practically it lies b/w 0.5 to 1 V.

Case 3

Small V_{DS} (mV)

$V_{DS} \propto I_D$

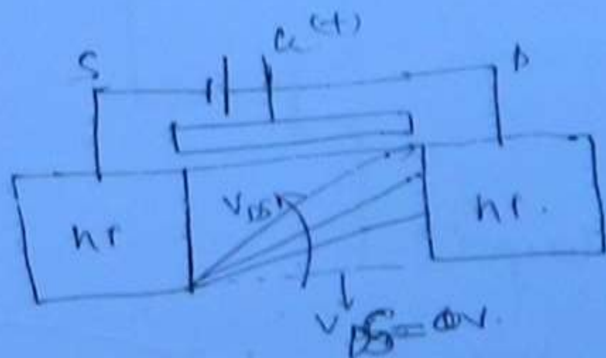


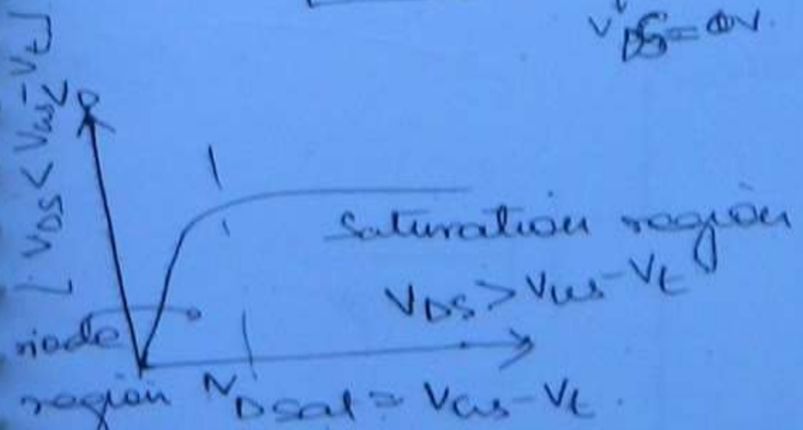
$V_{DS} \propto I_D$

$I_D \propto \underbrace{(V_{GS} - V_T)}_{\text{effective voltage or overdrive voltage}}$

Case 4.

large V_{DS} (V) is applied.





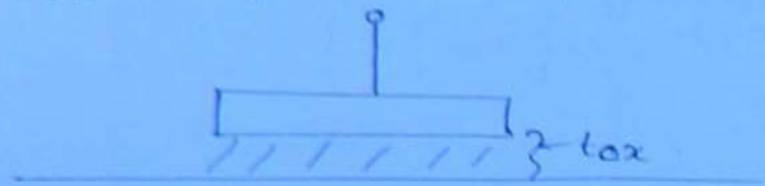
$\boxed{V_{DS} < V_{GS} - V_t}$

V-I characteristics eq. of MOSFET.

$$I_D = \mu_n C_{ox}\left(\frac{W}{L}\right)\left\{(V_{GS}-V_t)V_{DS} - \frac{1}{2}V_{DS}^2\right\}$$

$\mu_n \to$ mobility of e^-

$C_{ox} \to$ capacitance of oxide layer.



$$C_{ox} = \frac{\epsilon_{ox}}{t_{ox}}$$

$\epsilon_{ox} \to$ permittivity of oxide layer.

$\epsilon_{ox} \to 3.9\,\epsilon_0$

$\epsilon_0 \to 8.85\times10^{-12}$ F/m.

$\epsilon_{ox} \to 3.9\times8.85\times10^{-12}$

$= 3.45\times10^{-11}$ F/m

$t_{ox} \to$ thickness of oxide layer.

$\frac{W}{L} \to$ aspect ratio

$V_{GS}-V_t \to$ effective voltage 'or' overdrive voltage.

Triode region $\to$

$V_{GS} \geq V_t$

$V_{DS} < V_{GS} - V_t$.

$$I_D = \underbrace{\mu_n C_{ox}\left(\frac{W}{L}\right)}_{k_n'}\left\{(V_{GS}-V_t)V_{DS}\right\}$$

Saturation region $\to$

$V_{GS} \geq V_t$.

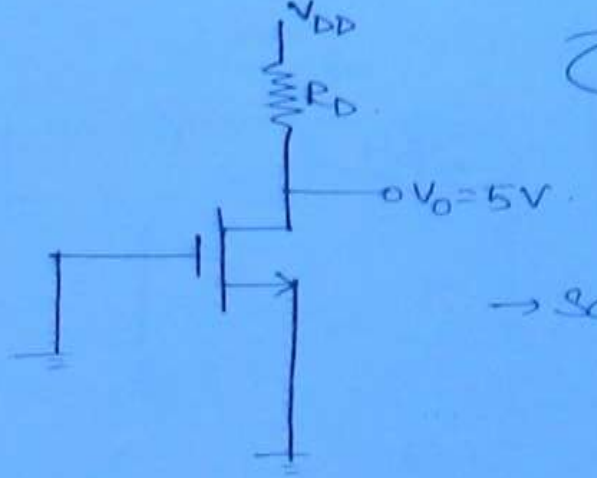
$V_{DS} \geq V_{GS} - V_t$.

$V_{D\,sat} = \underbrace{V_{GS} - V_t}_{\to k_n}$

$$I_D = \frac{1}{2}\mu_n \epsilon_0 C_{ox}\left(\frac{W}{L}\right)(V_{GS}-V_t)^2$$

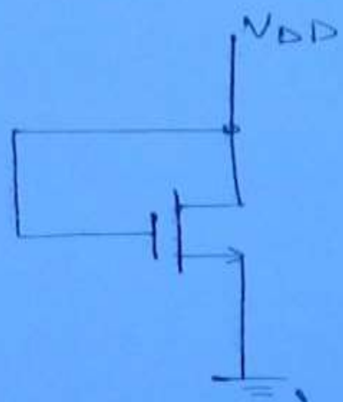
Ckt-1. →

263



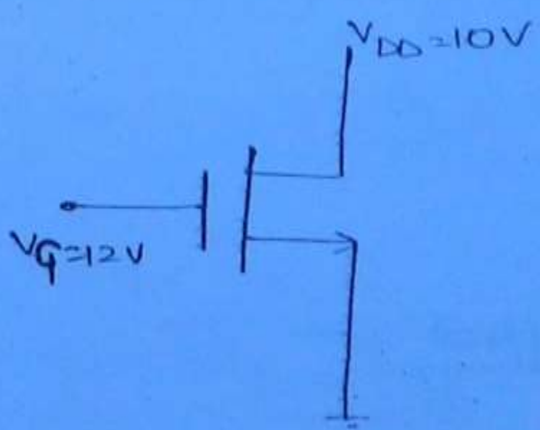
→ Saturation

Ckt 2 →



→ Saturation

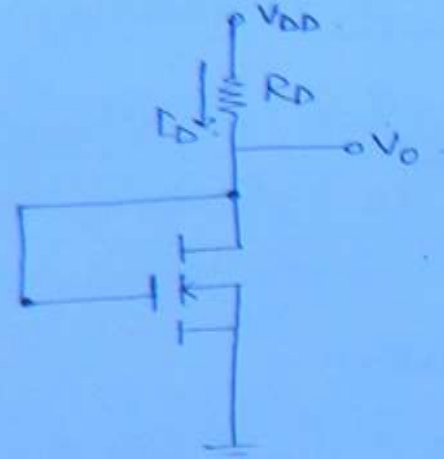
Ckt 3 →



— triode

$$\left.\begin{array}{l} V_D > V_G \\ V_D = V_G \end{array}\right\} \rightarrow \text{saturation}$$

$$V_D < V_G \} \rightarrow \text{triode}$$



264

Given data →

I_{D_1} Cal. Vo and Rd

$\left(\frac{W}{L}\right)$

μ_n

C_{ox}

$\lambda = 0$

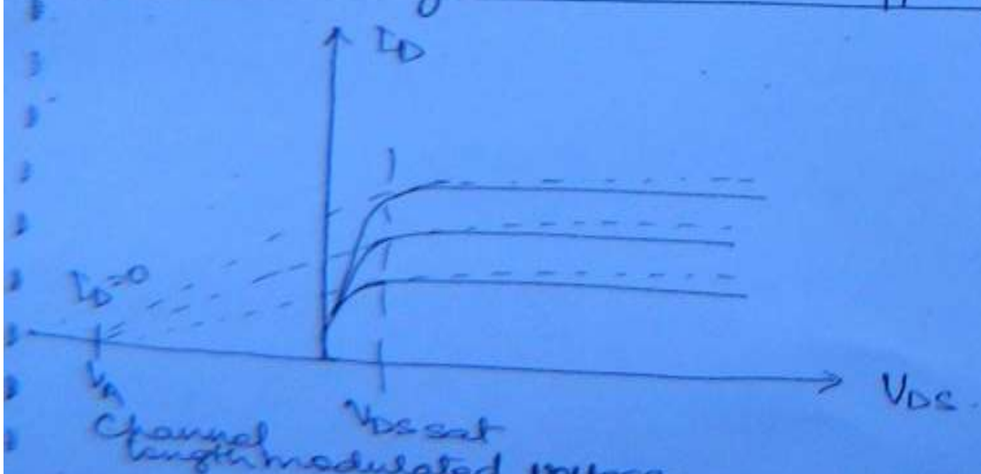
this is in Saturation Region.

$$I_D = \frac{1}{2} \mu_n C_{ox} \left(\frac{W}{L}\right) (V_{GS} - V_t)^2$$

$$V_{GS} = V_O.$$

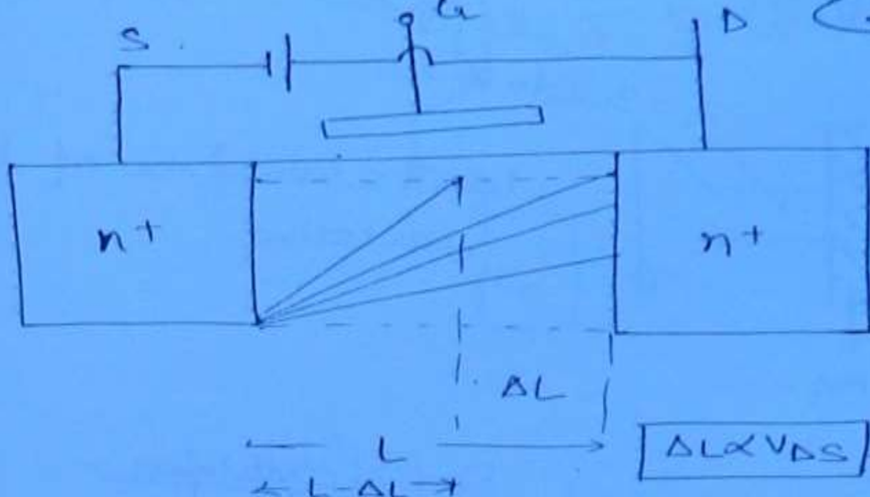
$$R_D = \frac{V_{DD} - V_O}{I_D}.$$

Channel length modulation effect :—



$$I_D = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)(V_{GS} - V_t)^2$$

285



$$\boxed{\Delta L \propto V_{DS}}$$

$$I_D = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L-\Delta L}\right)(V_{GS} - V_t)^2$$

$$= \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left[1 + \frac{\Delta L}{L}\right](V_{GS} - V_t)^2$$

$\Delta L \propto V_{DS}$

$\downarrow \lambda'$

$$I_D = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)\left[1 + \left(\frac{\lambda'}{L}\right)V_{DS}\right](V_{GS} - V_t)^2$$

$$I_D = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)(V_{GS} - V_t)^2(1 + \lambda V_{DS})$$

so for $I_D = 0$

$$1 + \lambda V_{DS} = 0.$$

$$\boxed{V_{DS} = -\frac{1}{\lambda} = V_A}$$

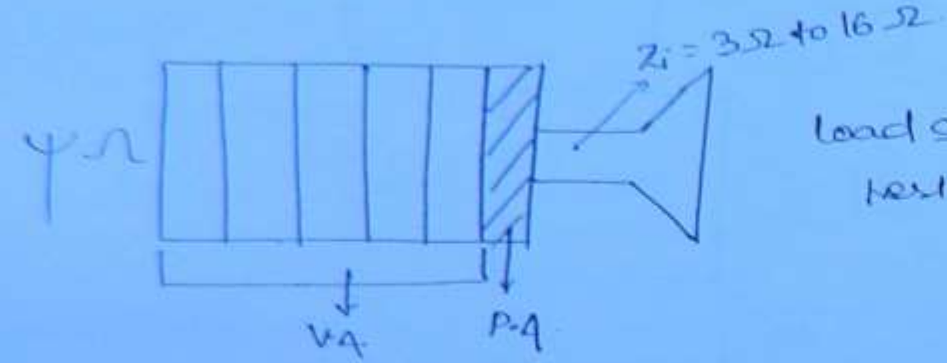
$$|V_A| = \frac{1}{\lambda}$$

$\lambda = 0, \quad V_A = \infty.$

$$I_D = k_n (V_{GS} - V_t)^2\left(1 + \frac{V_{DS}}{V_A}\right)$$

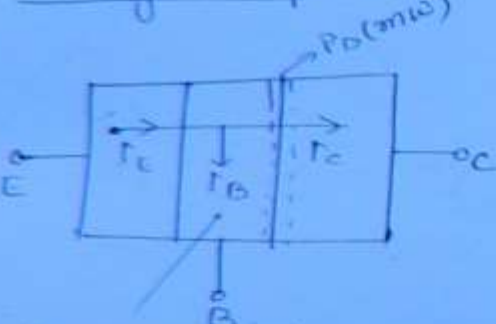
Power Amplifier :-

PAS → Public addressing System.



Load should be of less resistive.

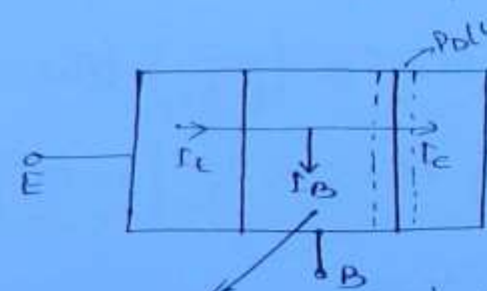
Voltage amplifier :-



Base region is narrower

$\uparrow \beta = \frac{I_C \uparrow}{I_b}$

Power amplifier -



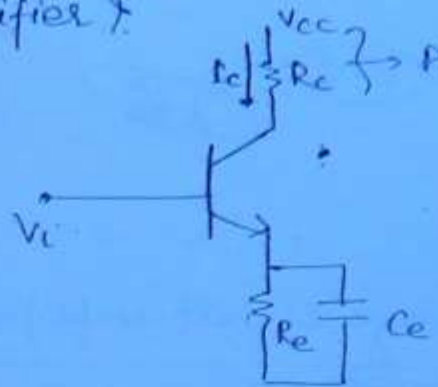
Base region is wider

$\beta \downarrow = \frac{I_C}{I_b \downarrow}$

Diff. b/w power ~~amp or~~ amplifier and Voltage amplifier :-

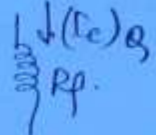
Parameter	V.A.	P.A
1) β	more	less
2) R_C	more	less
3) V_{in}	Small	Large
4) P_o	less	more
5) I_C	less	more
6) coupling.	Resistive coupling is used.	Inductive or X-mer coupling is best

Q. Why a Voltage amplifier can not be used as a power amplifier ?

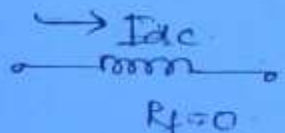


$\Rightarrow P_D = (I_C)_Q^2 R_C$

(267)



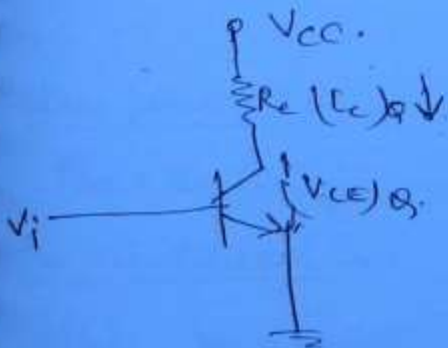
$P_D = (I_C)_Q^2 R_f$.



$\Rightarrow P_D = 0$.

For power amplifier, power dissipation should be neglible. So, in case of resistive load inductive load is preffered.

Power diagram analysis :—



P_{DC}

↓ $(P_{RC})_{DC}$ → $(I_C)_Q^2 R_C$ → R_f

↓ $(P_T)_{DC}$ → $(P_o)_{ac}$; $(P_C)_{dc}$ → Heat Sink.

internal power loss.

$P_{DC} = P_{ac} + P_{losses}$

o/p + int

P_D.

268

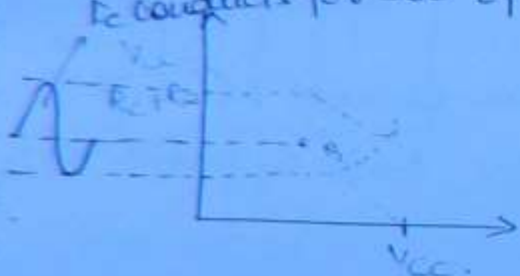
Definitions :—

- Power amplifiers are the large signal amplifiers which rises the power level of the signals.
- It is a device which converts dc power into ac power. and whose action is controlled by ac I/P signal.

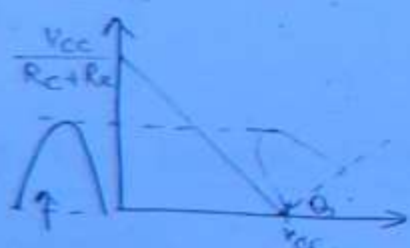
Classes →.

Class A →

I_c conducts for 360° of I/P signal.

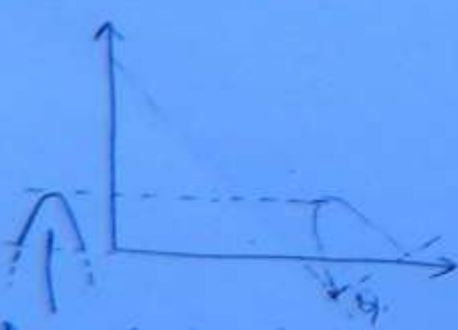


Class B →



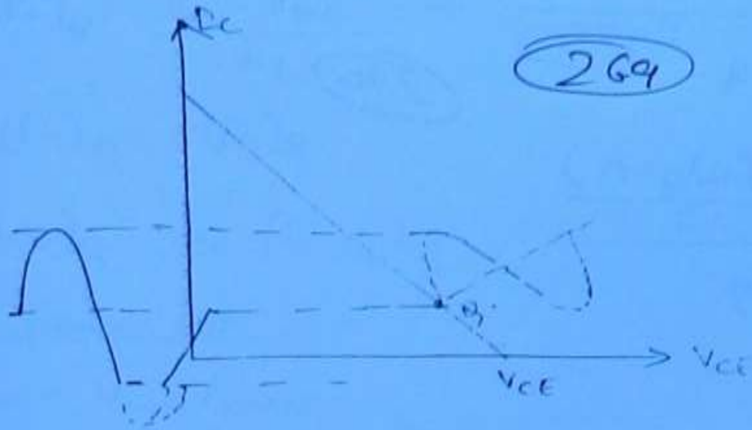
I_c Conducts for 180° of I/P.

Class C



I_c conducts for less than 180°

Class AB.



Class A series fed power amplifier :—

Dc power I/P →

$P_{DC} = V_{CC}(I_C)_Q$.

Dc conditions →

$$(I_b)_Q = \frac{V_{CC} - V_{BE}}{R_b}$$

$$(I_C)_Q = \beta (I_b)_Q.$$

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

ac power O/P.

$P_{ac} = V_{rms} I_{rms}$.

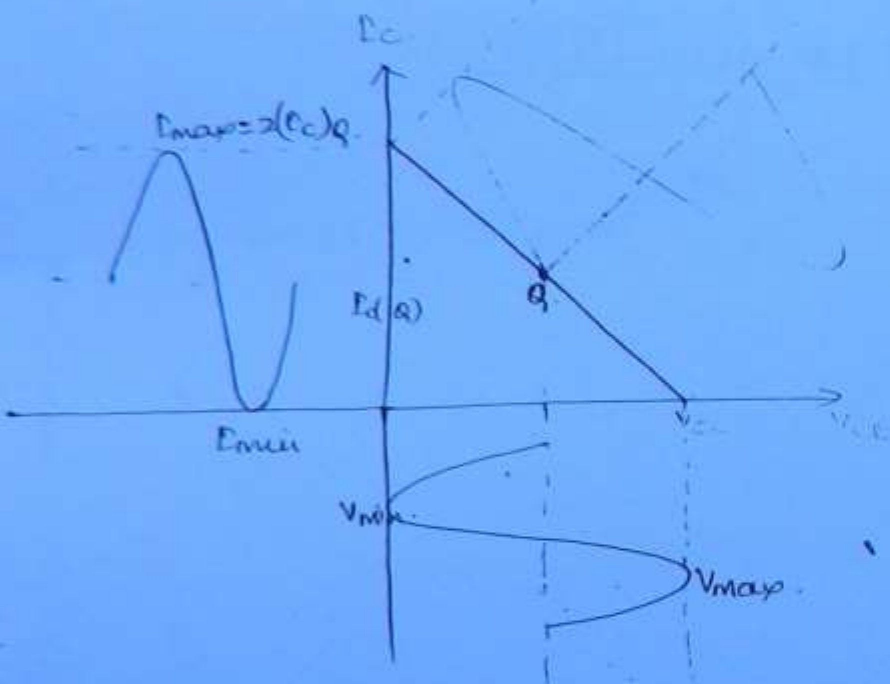
$$= \frac{V_M}{\sqrt{2}} \times \frac{I_M}{\sqrt{2}} = \frac{V_M I_M}{2}.$$

$$\frac{P_{ac}}{P_{DC}} = \frac{\frac{1}{2}\left(\frac{V_{max}-V_{min}}{2}\right)\left(\frac{I_{max}-I_{min}}{2}\right)}{V_{CC}(I_C)_Q}$$

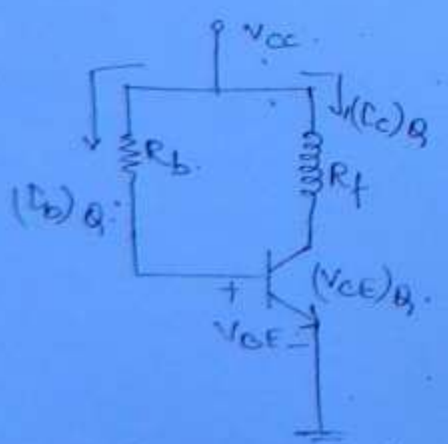
270

$$= \frac{\frac{1}{2}\left(\frac{V_{CC}-0}{2}\right)\left(\frac{2(I_C)_Q-0}{2}\right)}{V_{CC}(I_C)_Q}$$

$$= \frac{1}{4} = 25\%$$



Class A x-mer coupled power amplifier :-



→ I_{dc}

$R_f = 0$

DC power I/P →

$$P_{DC} = (V_{CC})(I_C)_Q.$$

Dc condition →

271

$$(I_b)_Q = \frac{V_{CC} - V_{BE}}{R_b}$$

$$(I_c)_Q = \beta (I_b)_Q$$

$$(V_{CE})_Q = V_{CC} - (I_c)_Q R_f$$

ac power O/P →

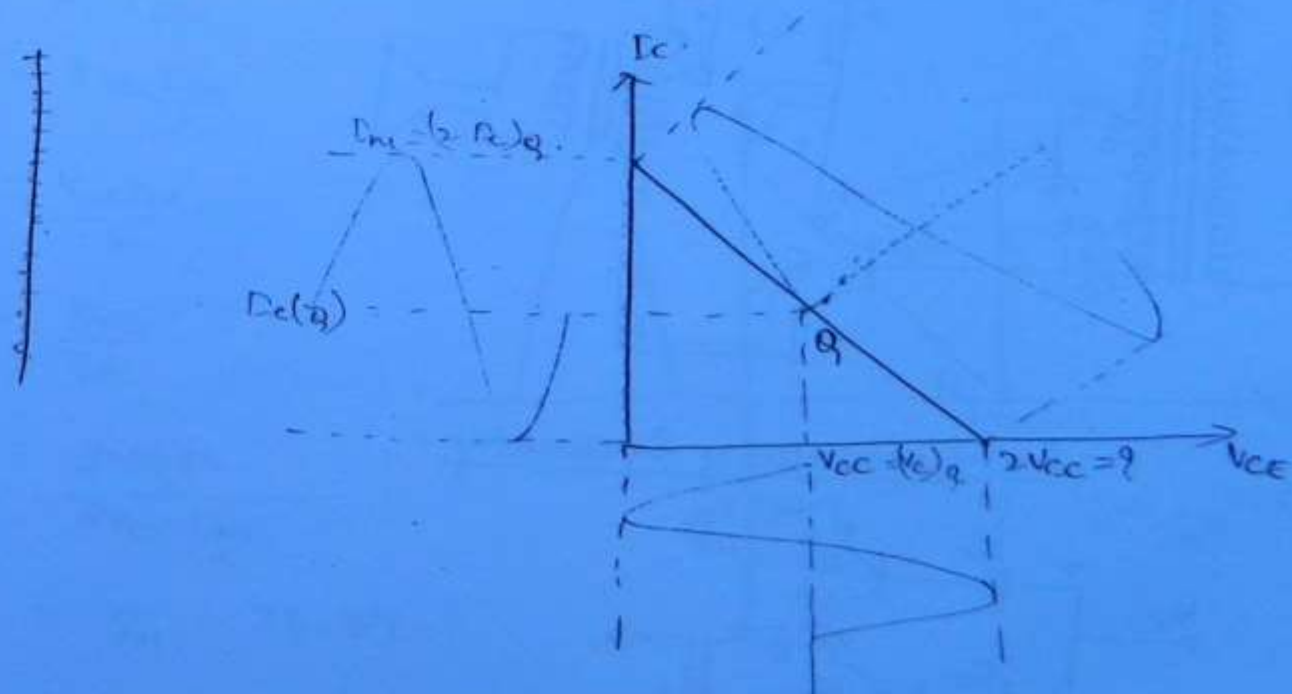
$$P_{ac} = V_{rms} I_{rms}$$

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

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

$$\frac{P_{ac}}{P_{dc}} = \frac{\frac{1}{2}\left(\frac{\overbrace{V_{max}}^{2V_{CC}} - V_{min}}{2}\right)\left(\frac{\overbrace{I_{max}}^{2(I_c)_Q} - I_{min}}{2}\right)}{V_{CC}(I_c)_Q}$$

$$= \frac{1}{2} = 50\%$$



Power dissipation in class A:

$$P_{DC} = P_{ac} + P_D$$

$$P_D = P_{DC} - P_{ac}$$

$$= V_{cc}(I_c)_Q - \frac{V_m I_m}{2}$$

(272)

At zero signal conditions :—

$$V_m = 0, \; I_m = 0$$

$$P_D = P_{DC}$$

When signal is applied,

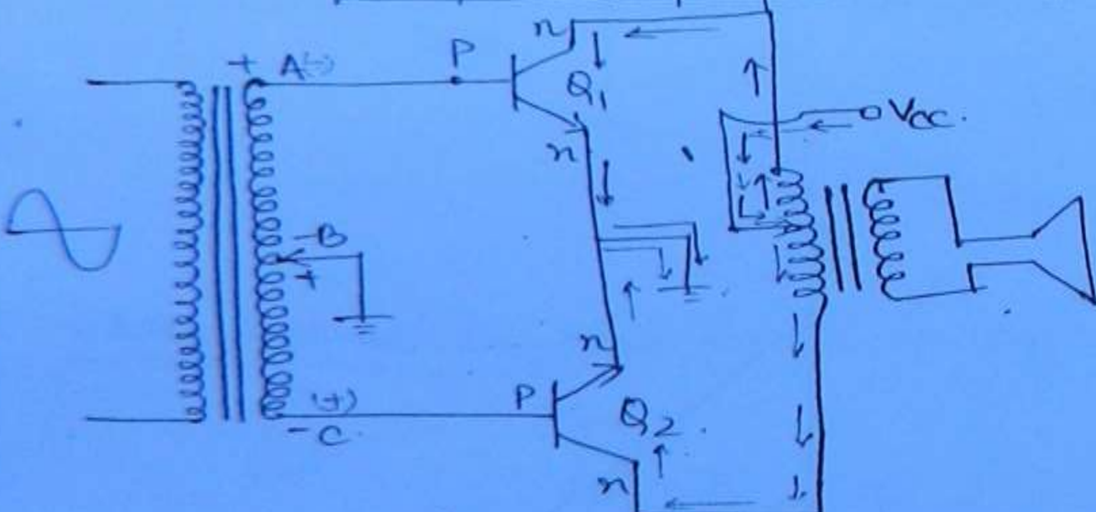
$$P_D = V_{cc}(I_c)_Q - \frac{V_m I_m}{2}$$

$\downarrow$

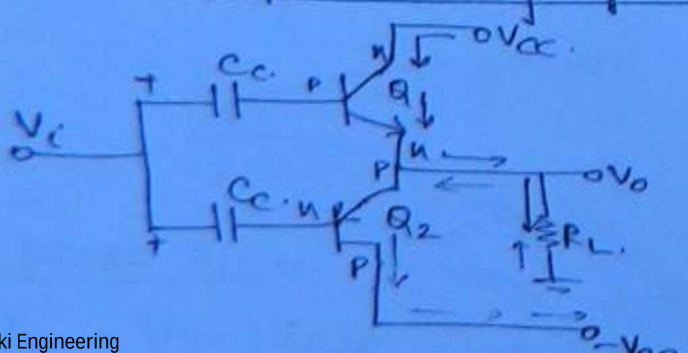
P_D decreases.

Class B Power amplifier :—

— pushpull amplifier.

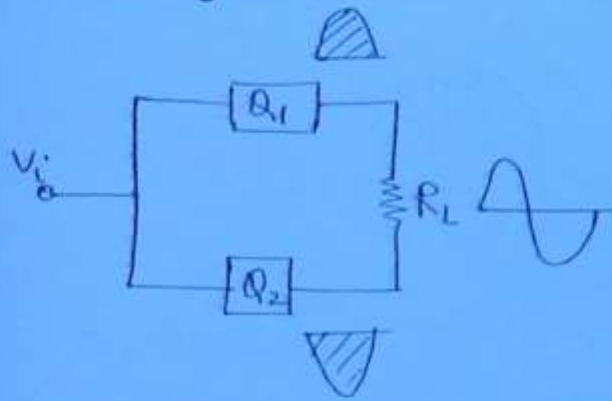


— complementary Symmetry →

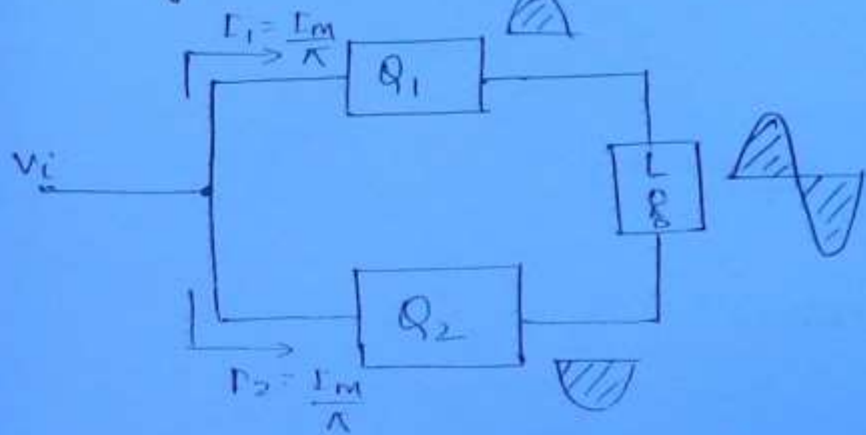


Block diagram →

223



Efficiency cal. in ta class B :—



$$P_{DC} = V_{cc} I_{dc}$$

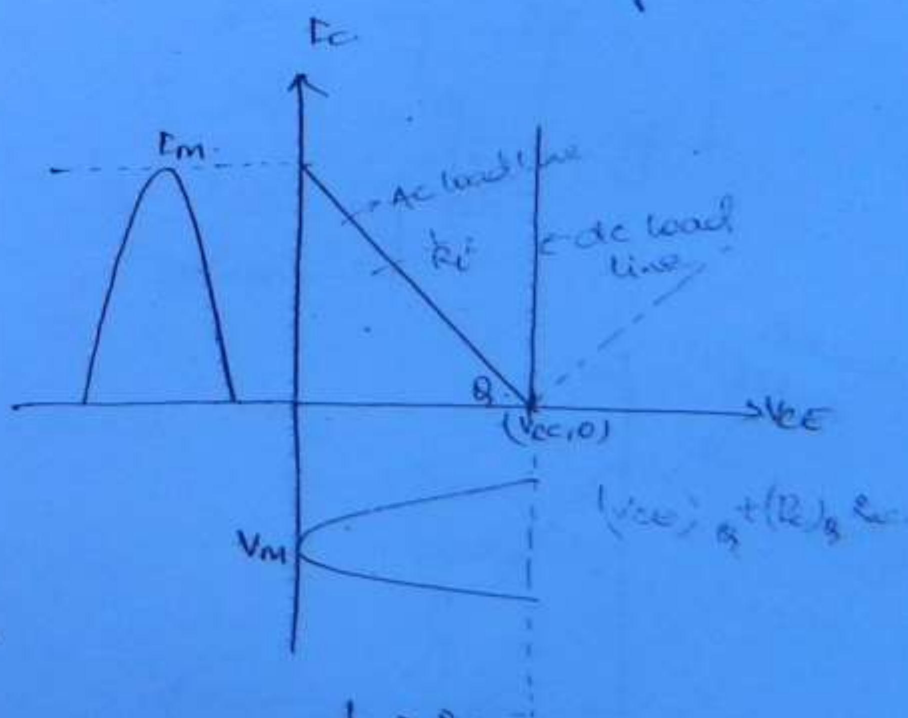
$$= \frac{2 V_{cc} I_m}{\pi}$$

$$P_{ac} = \frac{V_m I_m}{2}$$

$$\% \eta = \frac{P_{ac}}{P_{dc}}$$

$$= \frac{\frac{V_m I_m}{2}}{2V_{cc} \frac{I_m}{\pi}}$$

$$= \frac{\pi}{4} = 78.5\%$$



$$\frac{1}{R'} = \frac{I_m}{V_m}$$

Power dissipation in class B :-

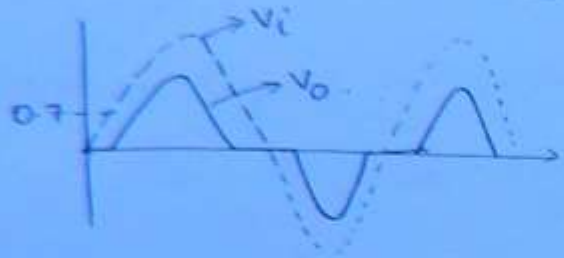
$$P_D = P_{DC} - P_{ac}$$

$$= \frac{2V_{cc}I_m}{\pi} - \frac{V_m I_m}{2}$$

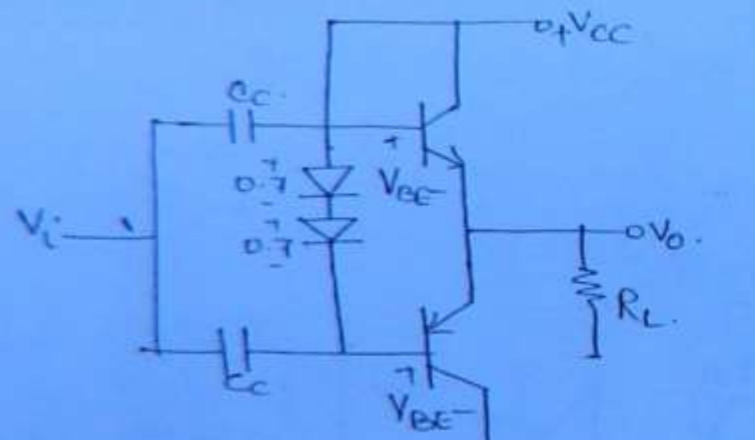
At 0 signal conditions

$$\boxed{P_D = 0}$$

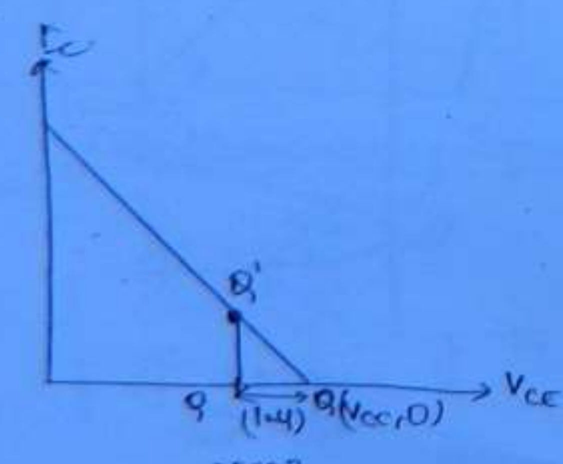
Cross over distortion :-



Class B complementary Symmetry :-



Class AB Amplifier ⇓

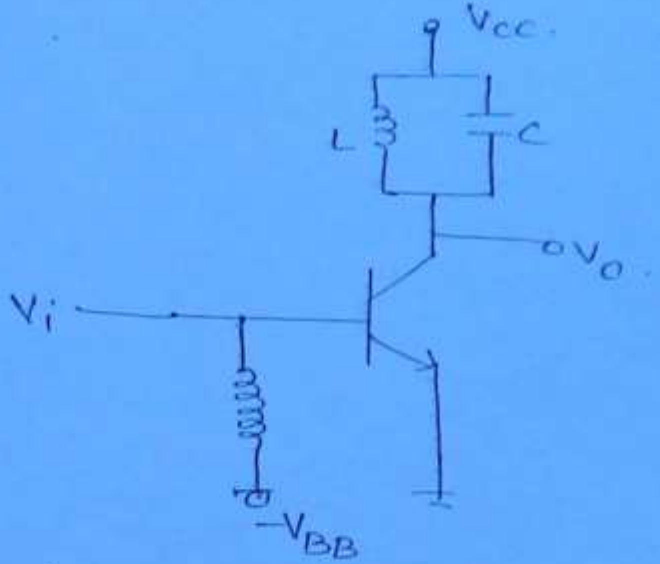


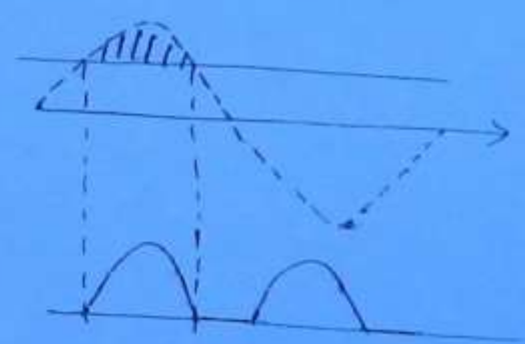
$(V_{cc} - 1.4)$

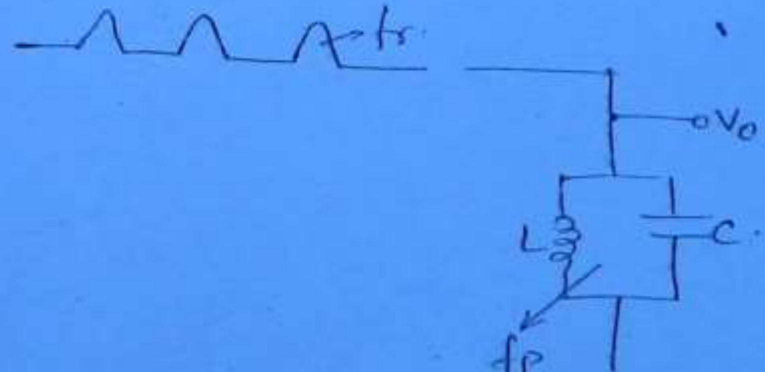
if 1 transistor is there then $(V_{cc} - 0.7)$

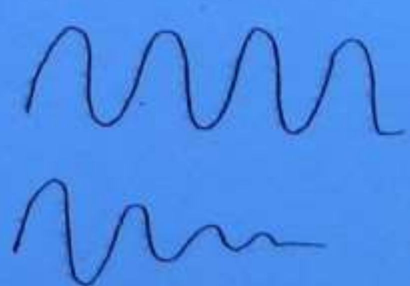
ass C power amp →

225









→ $f_r = f_p$

$f_r \neq f_p$

→ tunner amplifier