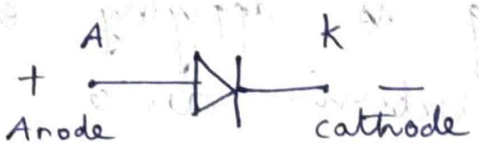
Unit - I

PN junction diode

* A diode is a two terminal device or two electrode device
* A diode is a one way device, offering low resistance when forward biased and as open circuit when reverse bias



* In a piece of semiconductor one half is doped with P-type another with N type
* N-type has high concentration of free electrons
* At the junction there is a tendency of free electrons to diffuse over P side and holes to N side, is called diffusion
* In thermal equilibrium, the region near the junction, there exist a wall of -ve immobile charges on p side and +ve immobile ions on n side
* In this region there are no mobile charge carriers and is called depletion layer
* Due to immobile +ve charges on n-side and -ve charges on p side, there exists an

electric field across the junction called barrier potential or junction potential

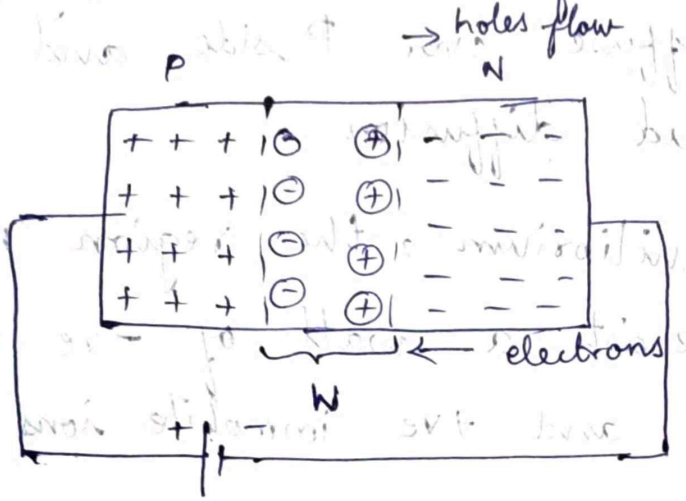
→ It is 0.3V for Ge and 0.7V for Si

→ Barrier potential depends on i) Type of semiconductor 2) donor and acceptor impurity added 3) temp 4) intrinsic concentration

→ Biasing is applying external dc voltage to any electronic device

Forward Bias:

→ when positive terminal of battery is connected to P-type and negative terminal to N-type, The bias applied is known as forward bias



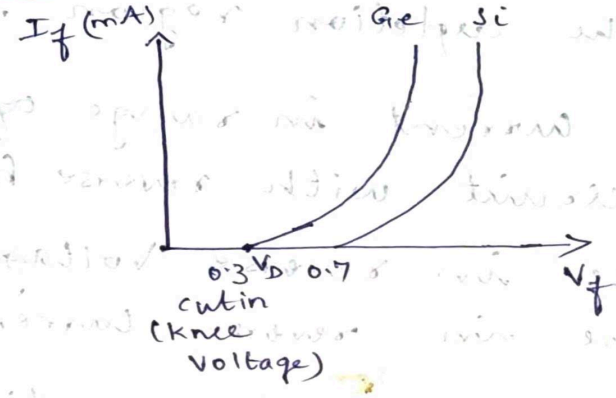
→ If the applied voltage is less than the barrier potential there is no conduction

→ If it exceeds, the +ve potetial repels holes in p region and -ve potential repels electrons in N region and they cross the junction

→ They decreases the width of depletion region

→ The forward current due to majority carriers is large

V-I characteristics

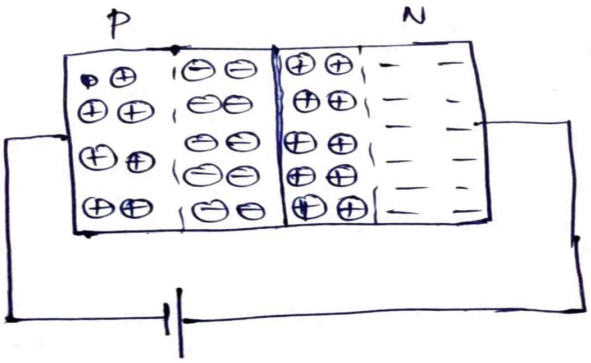


→ As $V_f < V_D$, $I_f = 0$ because the barrier potential prevents the current flow

→ As $V_f > V_B$, potential barrier dissappears and large current flow

Reverse Bias

→ In Reverse bias, N side is connected to positive terminal and P side to negative terminal



→ The majority carriers holes of P side move towards negative terminal

→ The majority carriers electrons of N side move towards Positive terminal

→ The width of the depletion region increases.

→ A very small current in range of microamperes flow in the circuit with reverse bias.

→ With increase in reverse Voltage there is a small increase in reverse current

→ The current increase after particular Voltage called breakdown Voltage

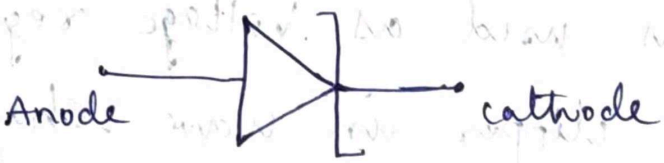
→ At this Voltage the minority carriers get enough kinetic energy and break the junction

Zener Diode

→ The diode is designed to work in reverse breakdown region

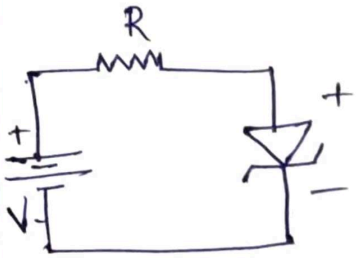
→ It is a reverse biased heavily doped PN junction diode

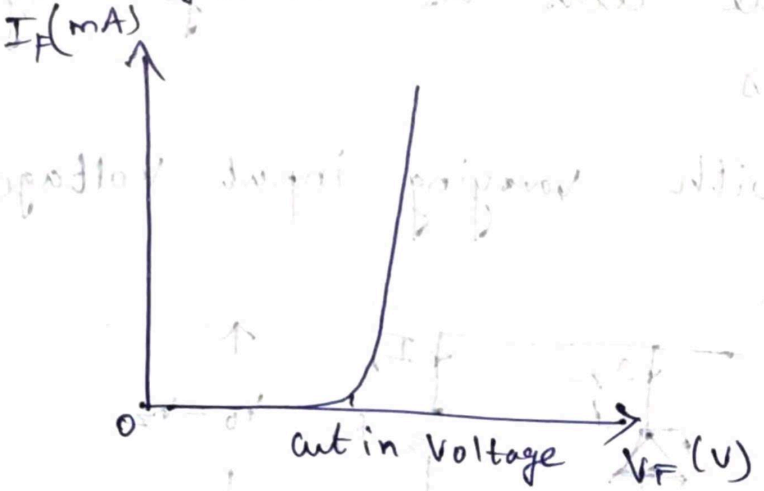
Symbol



Forward bias

→ when forward biased the diode acts as ordinary PN junction diode



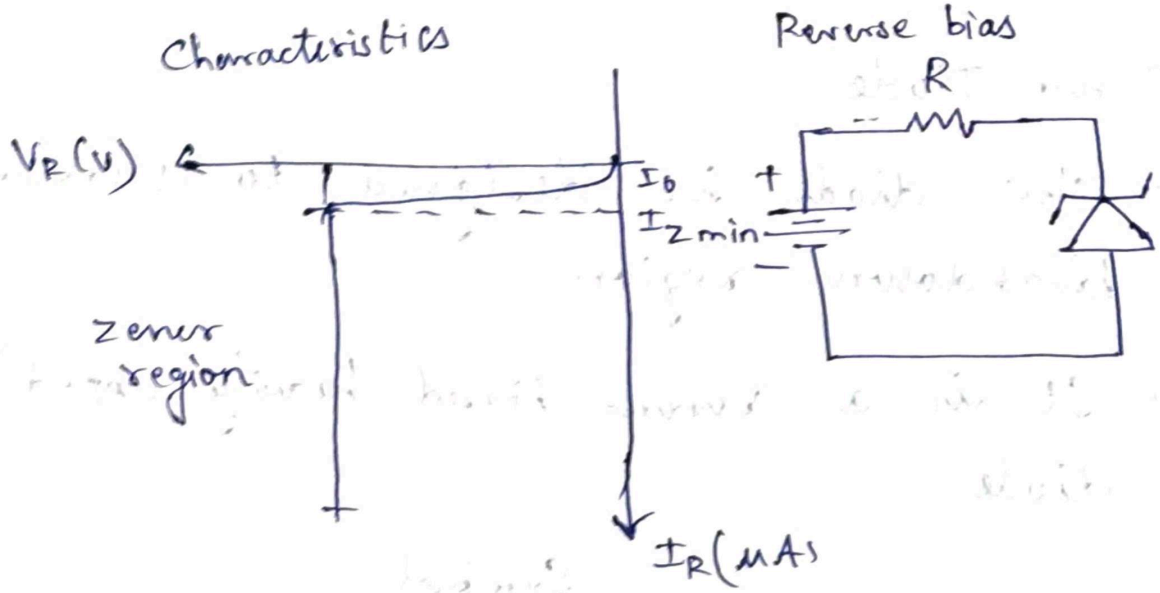


Reverse characteristics

→ As reverse voltage is increased, a small reverse saturation current I_0 in mA will flow

→ At certain value of reverse voltage, the reverse current will increase suddenly & sharply

→ This break down voltage is called zener breakdown



Applications

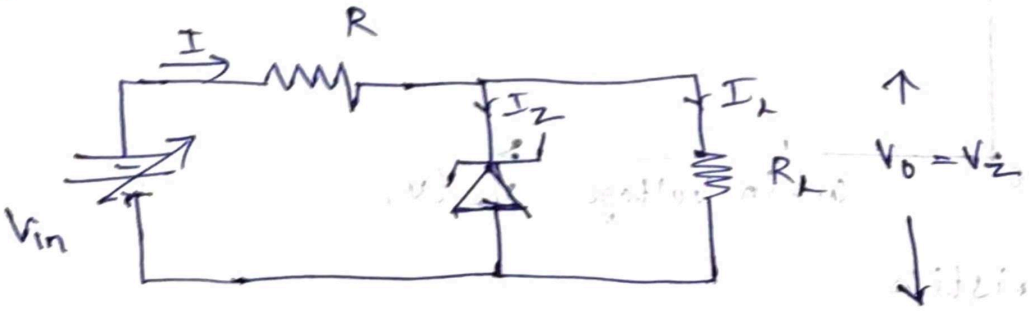
i) Zener diode is used as Voltage regulator

ii) Used as peak clipper in wave shaping circuits

Zener diode as Voltage regulator

→ Zener diode acts as voltage regulator in Reverse bias

Regulation with varying input voltage



→ R_L is fixed and V_{in} varies

$$V_0 = V_Z$$

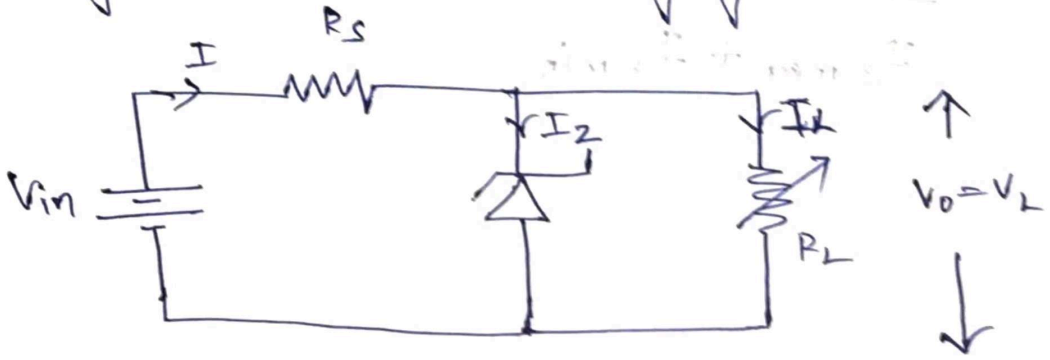
$$I_L = \frac{V_0}{R_L} = \frac{V_Z}{R_L} = \text{constant}$$

$$I = I_Z + I_L$$

→ If V_{in} increases I increases, since I_L is constant I_Z increases

→ If V_{in} decreases I decreases, I_L is constant I_Z decreases

ii) Regulation with Varying load current



→ Assume V_{in} constant but I_L varies

→ The variation of load changes the current through it.

$$I = I_Z + I_L$$

$$I_Z = I - I_L$$

$$I = \frac{V_Z}{R_S}$$

1. when input voltage is minimum, load current is maximum

$$I_{Z\,min} = \frac{V_{in} - V_o}{R_{S\,max}} - I_{L\,max}$$

$$R_{S\,max} = \frac{V_{in(min)} - V_o}{I_{Z\,min} + I_{L\,max}}$$

2. when input voltage is maximum, load current is minimum

$$I_{Z\,max} = \frac{V_{i(max)} - V_o}{R_{S\,min}} - I_{L\,min}$$

$$R_{S\,min} = \frac{V_{in\,max} - V_o}{I_{Z\,max} + I_{L\,min}}$$

Diffusion Capacitance

→ In forward bias condition, width of the depletion region decreases and holes from p side get diffused to n side

→ electrons from n side move to p side, As applied voltage increases, concentration of injected charge particle increases

→ The rate of change of injected charge with applied volt is called diffusion capacitance

$$C_D = \frac{dQ}{dV}$$

$$C_D = \frac{\tau I}{2 V_T}$$

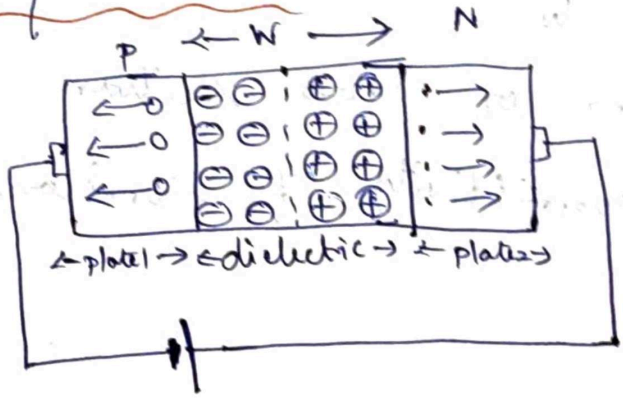
τ - mean life time of holes

→ Diffusion capacitance is proportional to current

It is value in in order of nF

→ C_D is much larger than transition capacitance C_T

Transition Capacitance



→ When a diode in reverse bias, reverse current flows due to minority carriers

→ Majority carriers move away from the junction

→ This increases the width of the depletion region as reverse bias voltage increases

→ As charged particles move away from the junction, there exists a change in charge with respect to applied reverse voltage

→ change in charge with respect to change in voltage is capacitive effect

→ Such capacitance is called Transition Capacitance space charge capacitance, barrier capacitance or depletion layer capacitance

$$C_T = \frac{dQ}{dV}$$

→ If W is width of transition capacitance

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

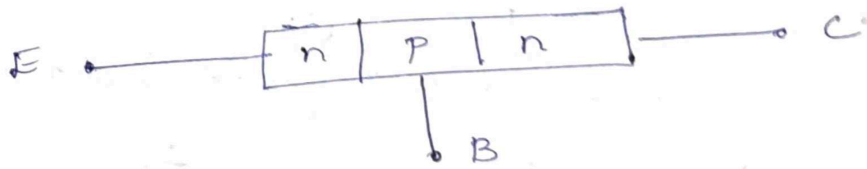
A → Area of cross section

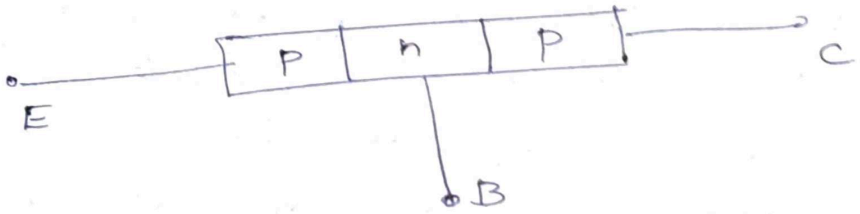
ε → permittivity of semiconductor

Structure of BJT

* Two types of transistor
  i) n-p-n ii) p-n-p

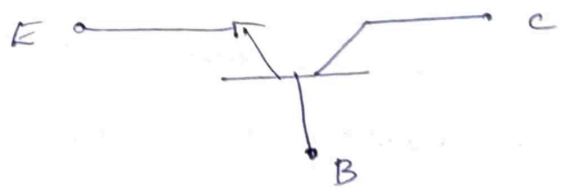
* A transistor is formed by sandwiching single p region between two n-region



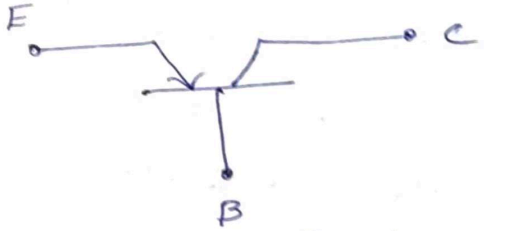


* The middle region is called base of the transistor
* The process by which impurities are added to a pure semiconductor is called doping
* emitter is highly doped and collector is moderately doped

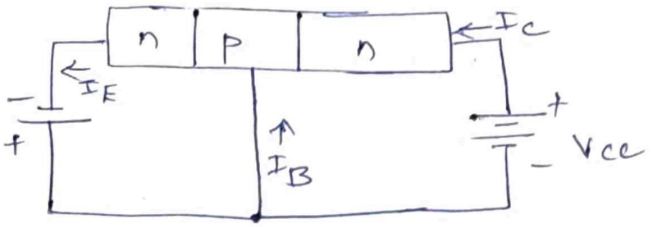
Symbols of npn

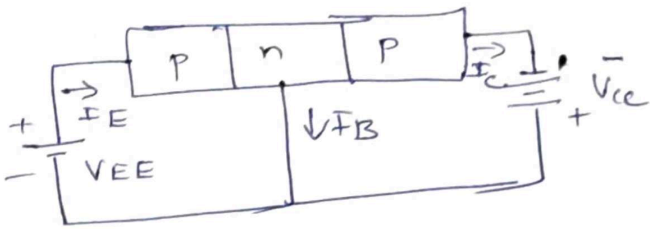


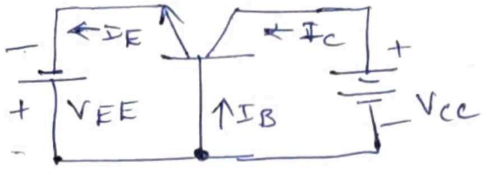
Symbol of pnp

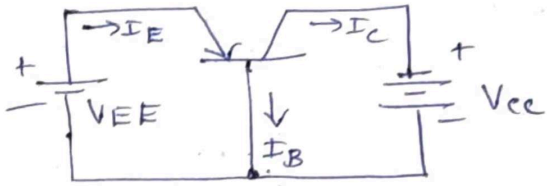


* It has two pn junctions. One junction is between E & B and other junction is between collector and base









→ for an npn transistors electrons are injected into base

→ these e^{-n} constitute the emitter current I_E,

→ Of 100 e^{-n} are injected into base, the base is very thin and recombine with holes and constitute base current

→ the remaining 98 e^{-n} cross the base collector junction and appear on collector side and constitute collector current I_C.

So, $$I_E = I_B + I_C$$

→ As the base current is very small

$$I_C \cong I_E$$

I_{CO}: Reverse saturation current

* when emitter is open, the base and collector acts as a reverse biased diode and collector current acts as reverse saturation current

I_{CBO}: Reverse collector saturation current:

I_{CBO} is nothing but I_{CO} for physical (real, non idealized transistor). It is defined as current

Unit -I

BJT

Biasing of Discrete BJT

* Transistor is a three terminal devile

1) Base
2) emitter
3) collector

* St is operated in 3 configuration

1) common base
2) common emitter
3) common collector

* The amplification of transistor is achieved by passing input current from a region of low resistance to high resistance

Need for Biasing

* The transistor is operated in 3 regions

cutoff, active, saturation.

by applying proper biasing condition

* In order to operate transistor in the desired region external d.c voltages of

Correct polarity and magnitude is applied

Region of operation	Emitter Base function	Collector Base function
Cutoff	Reverse Bias	Reverse Bias
Active	Forward Bias	Reverse Bias
Saturation	Forward Bias	Forward Bias

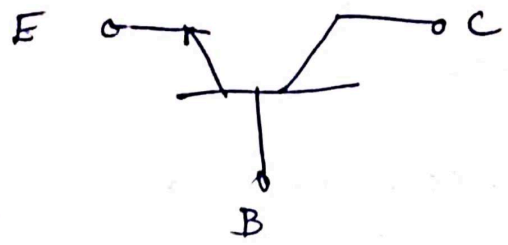
* In transistor amplifier circuits, output signal power is always greater than input signal power

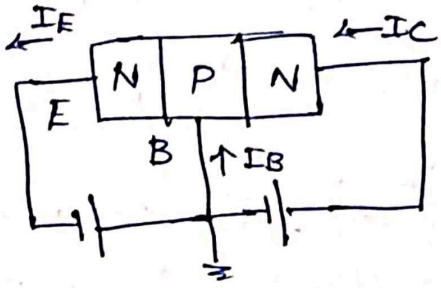
* The dc sources supplies the power to the transistor circuit to get the output signal power greater than input signal power

* When a transistor is biased certain current and voltages condition exists.

* This is known as operating condition; dc operating point, quiescent point or Q point, or bias point

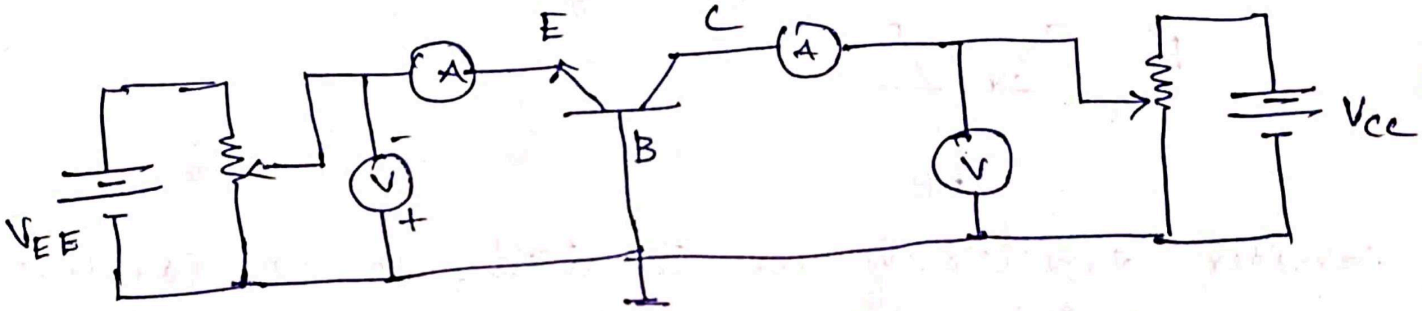
Common base configuration



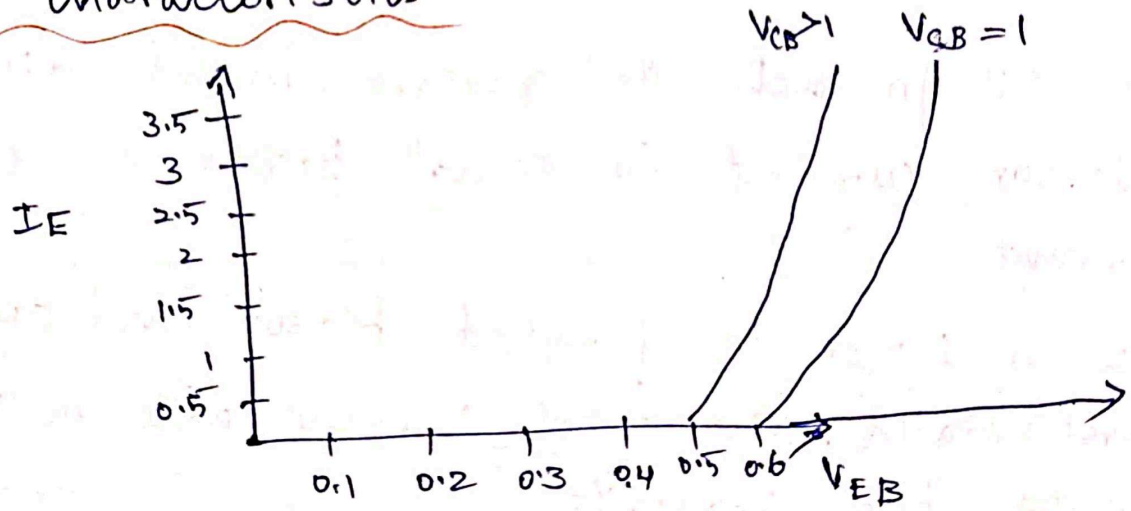


→ the input is taken from emitter and base

→ output is taken from collector and base



Input characteristics



→ when $V_{CB} = 0$ the junction behaves as forward bias diode

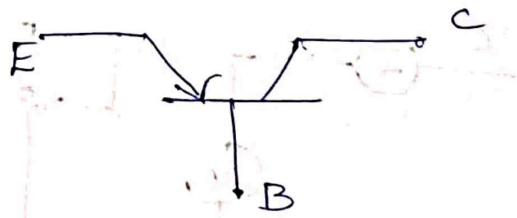
→ when V_{CB} is increased keeping V_{EB} constant I_E increases

when emitter current is zero

→ It constitutes the leakage current flowing around the jn. and across the surfaces

→ so $|I_{CBO}| \gg |I_{CO}|$. Since new carriers are generated by collision in collector jn. leading to avalanche multiplication.

I_{CBO}



→ Consider emitter is open circuited, so no carriers are injected from emitter into base and emitter current is zero.

→ So CB jn acts as reverse biased diode and collector current is equal to reverse saturation current

→ when EB jn is forward biased and CB jn. is reverse biased, the total current will be the sum of two currents

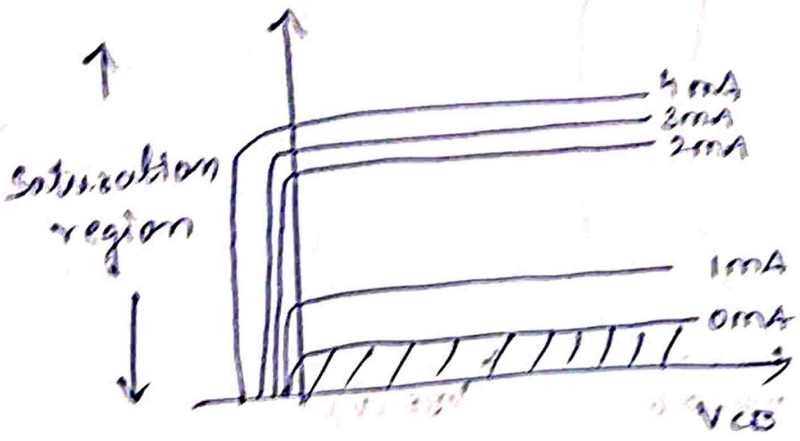
$$I_C = I_{pc} + I_{CBO}$$

I_{pc} ↓ Current due to injected carriers

$$= \alpha_{dc} I_E + I_{CBO}$$

$\left(\because \alpha_{dc} = \frac{I_{pc}}{I_E}\right)$

Output characteristics

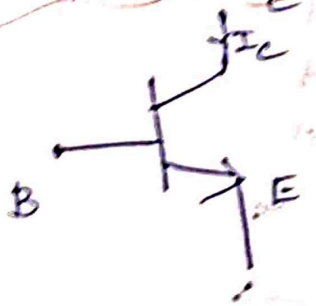


→ For constant value of I_E, I_C is independent of V_{CB}

→ I_C flows even when V_{CB} is equal to 0.

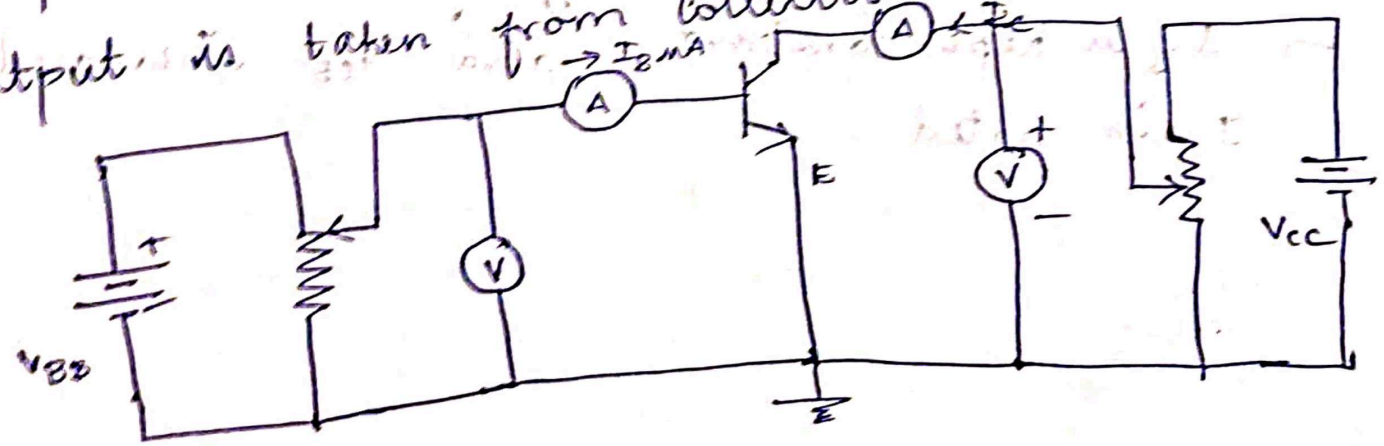
Current Amplification factor $\alpha = \left.\frac{I_C}{I_E}\right|_{V_{CB} \text{ as constant}}$

Common Emitter configuration

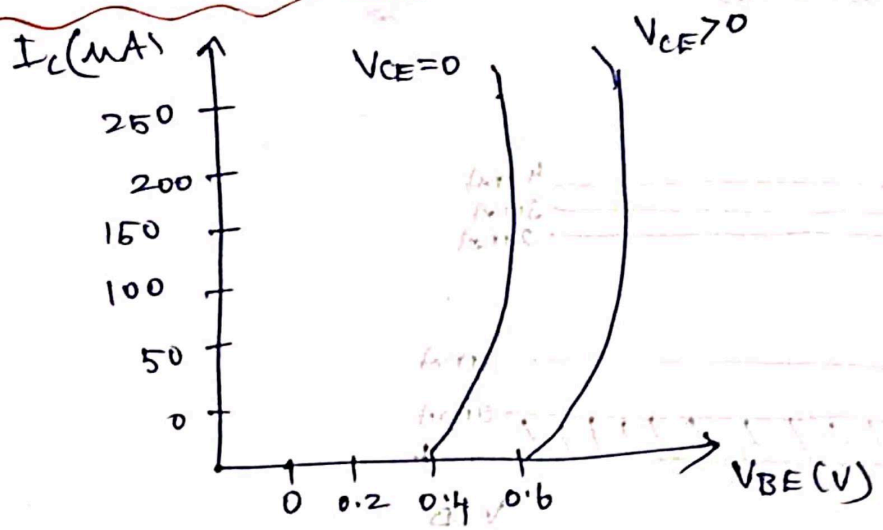


→ Input is applied between base and emitter

→ output is taken from collector and emitter



Input characteristics

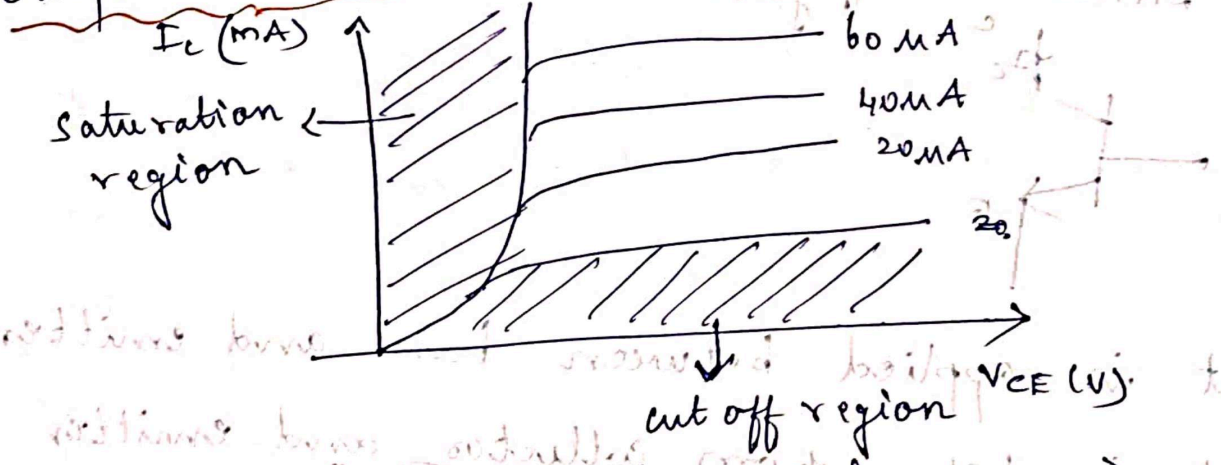


→ When $V_{CE} = 0$, emitter base junction is forward biased and behaves as a diode

→ when V_{CE} is increased width of depletion region decreases.

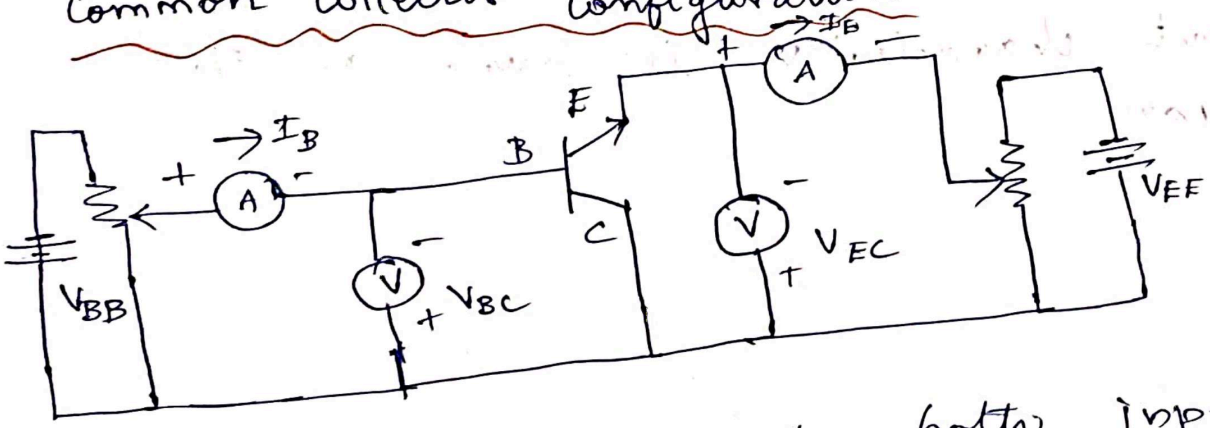
→ this effect causes decrease in base current I_B.

Output characteristics



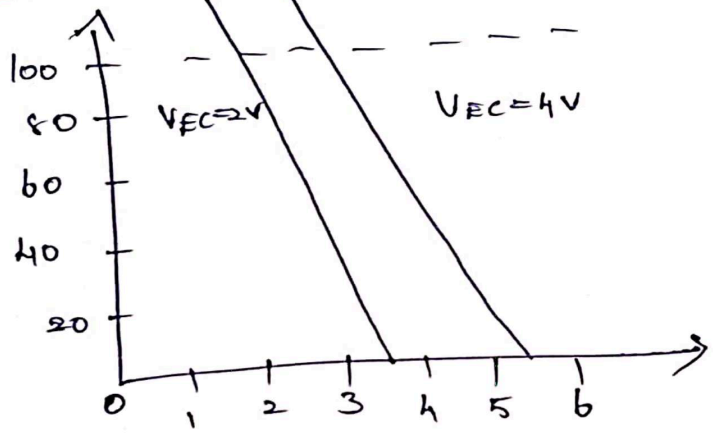
→ I_B is kept constant and V_{CE} increased and I_C is noted

Common collector configuration



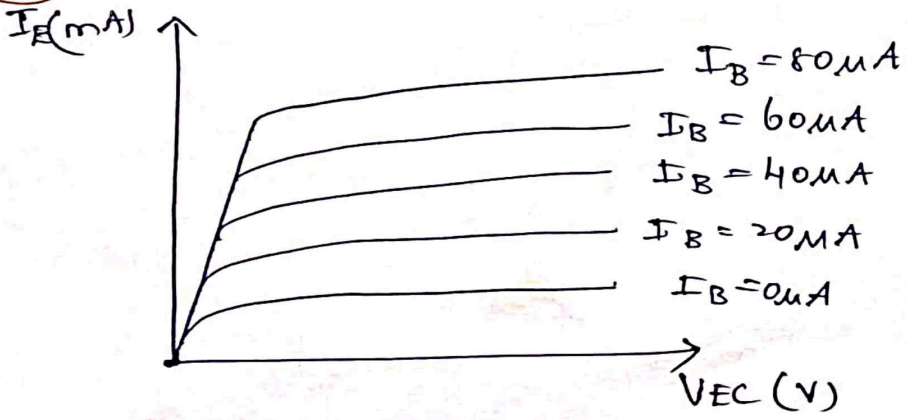
→ collector is common to both input and o/p

⇒ Input characteristics

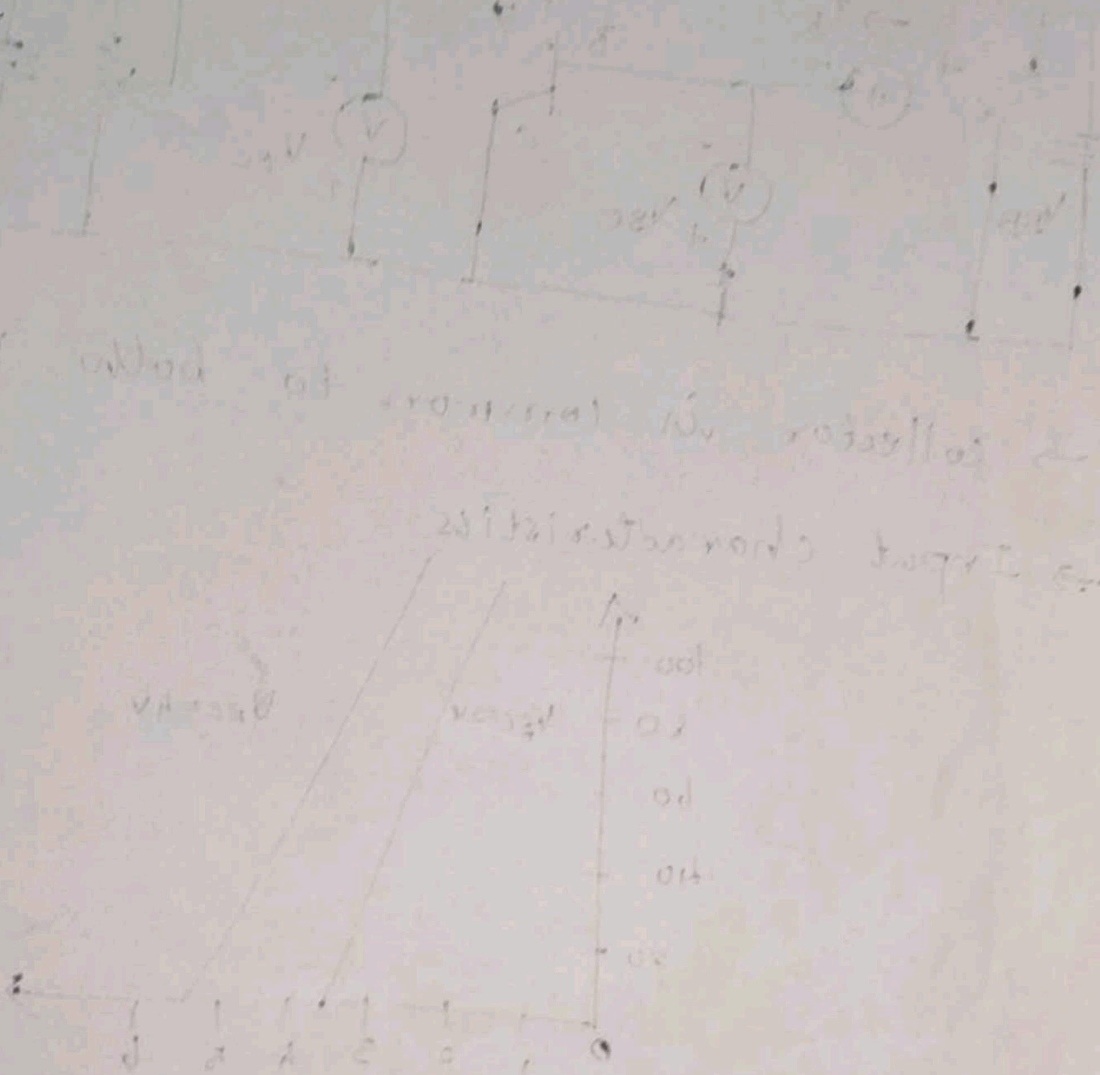


→ By keeping V_{EC} as constant voltage V_{BC} is increased in equal steps and increase in I_B is noted

Output characteristics



The output characteristics is same as CE configuration



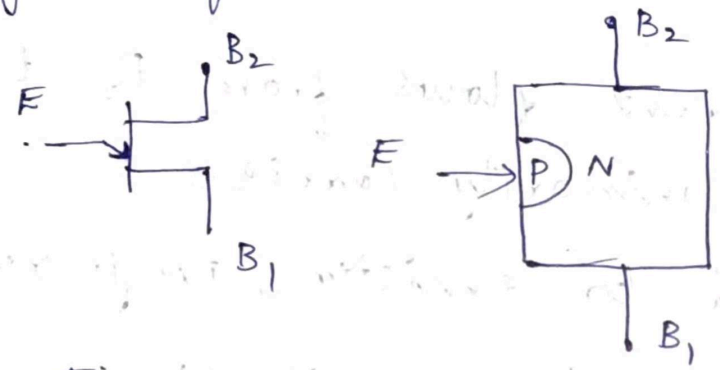
UJT - Unijunction Transistor

→ 3 terminal device - emitter, Base 1 & Base 2

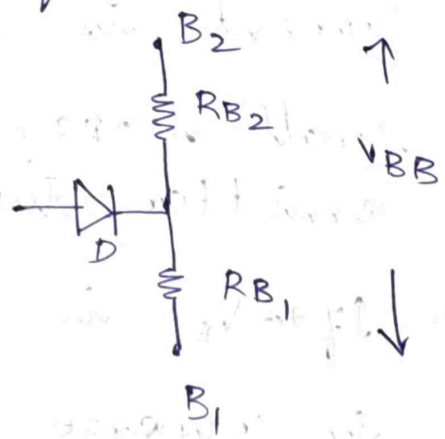
→ only one PN junction

→ Consists of lightly doped silicon bar with heavily doped P type alloyed to its one side closer to B_2 → producing single PN junction

Symbol of UJT



Equivalent ckt



Operation of UJT

Inter base resistance between B_1 & B_2 of silicon bar is

$$R_{BB} = R_{B_1} + R_{B_2}$$

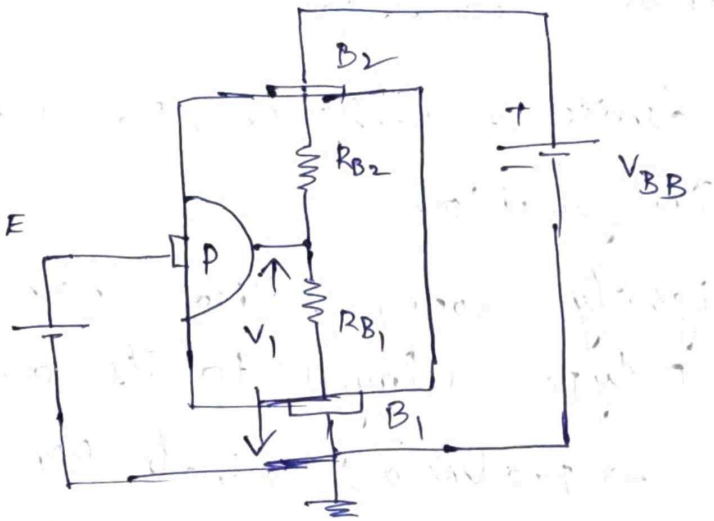
→ with emitter open, V_{BB} is applied between bases

→ Voltage gradient is established between N type bar

→ Voltage drop across R_{B_1} is $V_1 = \eta V_{BB}$

η - intrinsic stand off ratio

$$\eta = \frac{R_{B_1}}{R_B + R_{B_2}}$$



→ V_1 reverse biases the PN junction, so emitter current is cut off

→ Small leakage current flows from B_2 to emitter due to minority carriers

→ If +ve V_E is applied to emitter, PN jn remain in reverse bias as long as $V_E < V_1$

→ If $V_E > V_1$, then diode becomes forward biased, so holes are injected into P type bar

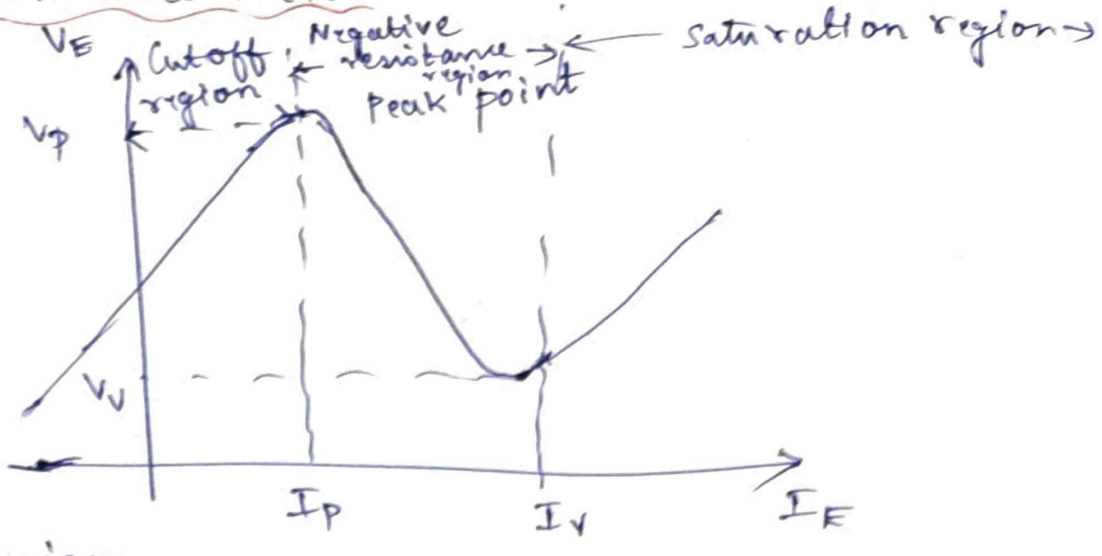
→ These holes are repelled by B_2 and attracted by B_1

→ Accumulation of holes in Emitter to B_1 reduces resistance and I_E increases. and the device is in ON state

→ If negative voltage is applied to emitter PN is reverse biased. I_E is cut off

So device is OFF.

UJT characteristics



1) Cut off region

→ If $V_E < V_P$, PN jn. is RB, so leakage current flows through the device. This is upto peak point

2) Negative Resistance region

→ when $V_E = V_P$, PN jn. is FB, I_E starts flowing

→ The voltage across the device ↓ and current ↑

→ This region is called negative resistance region. it is upto valley point

3) saturation region

→ ↑ in I_E further valley point I_V drive the device in saturation region

→ It is similar to semiconductor diode

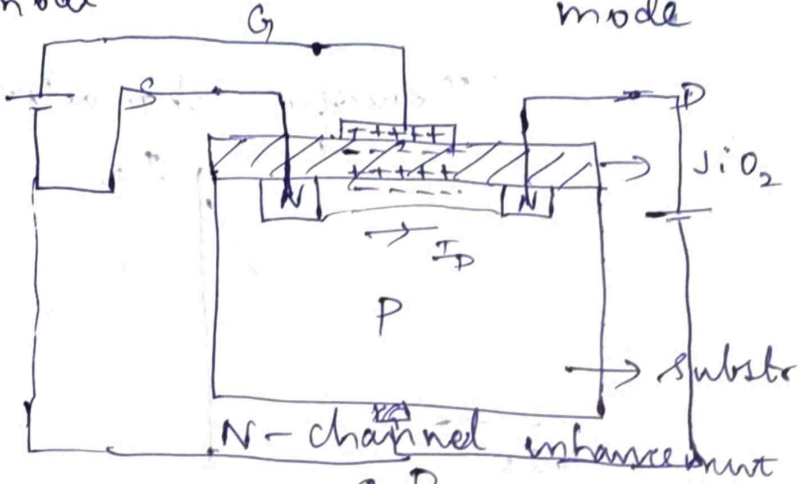
MOSFET

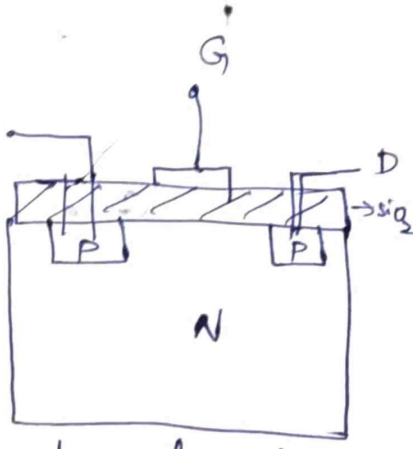
→ Metal oxide semiconductor Field effect Transistor

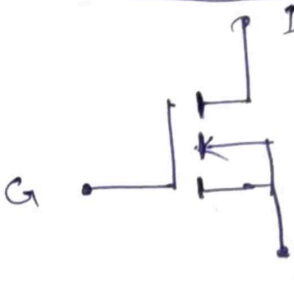
→ Two type

Depletion mode

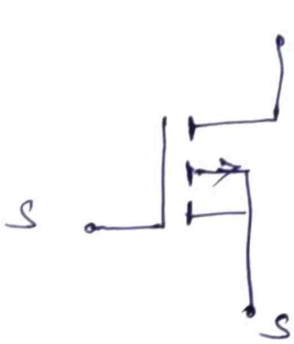
enhancement mode





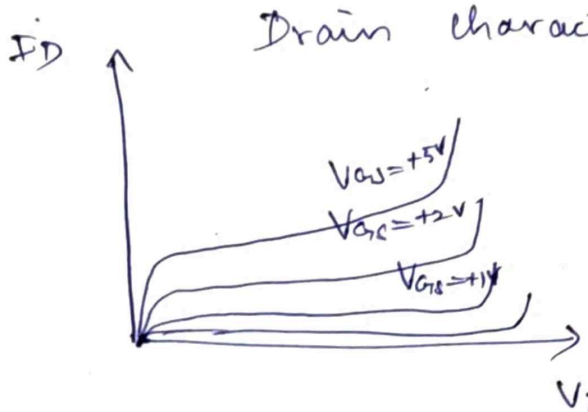


N channel enhancement

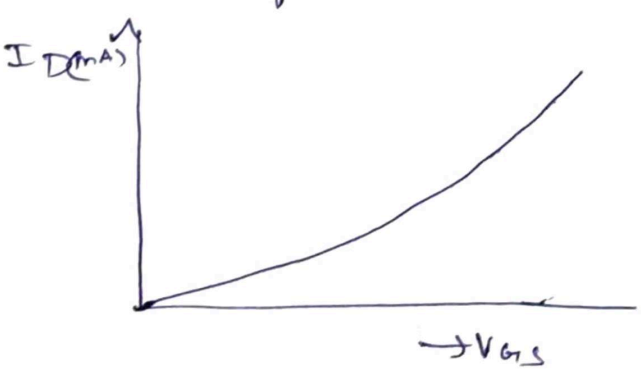


→ Gate is given +ve, V_{GS} increases the I_D current so it is called enhancement mode

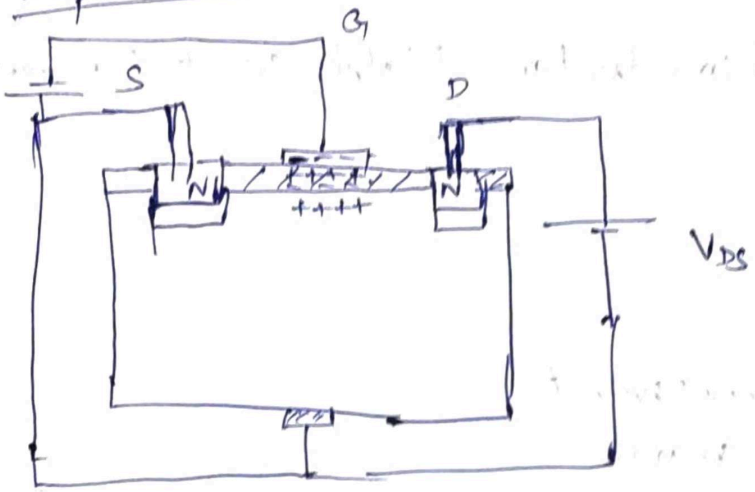
Drain characteristics



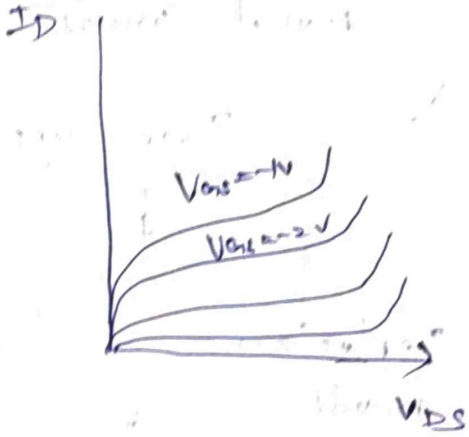
Transfer characteristics

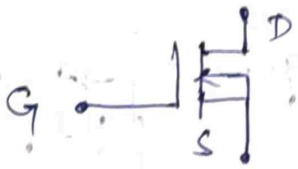


Depletion mode.

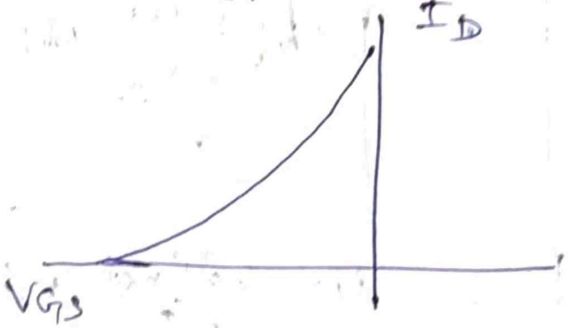


Drain characteristics





Transfer characteristics



MOSFET

→ MOSFET is metal oxide semiconductor field effect transistor.

→ MOSFET is made very small and used to design high density VLSI circuits.

→ The gate of the MOSFET is insulated from channel by a silicon dioxide layer

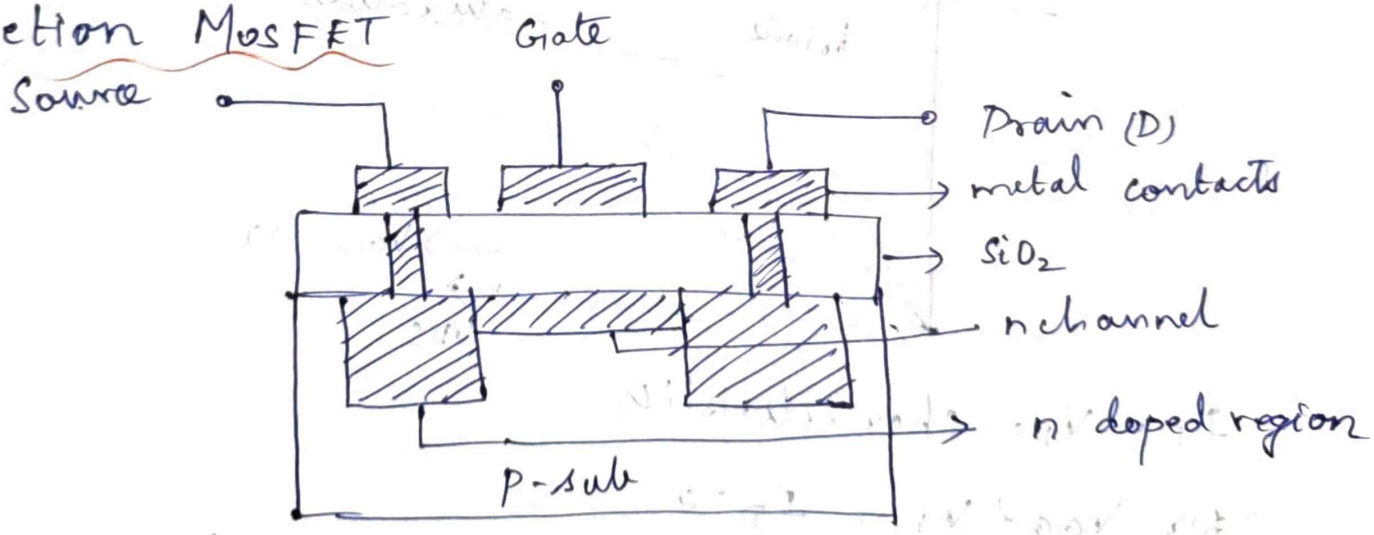
→ Due to this input resistance of MOSFET is very high

Two types of MOSFET

i) Depletion MOSFET

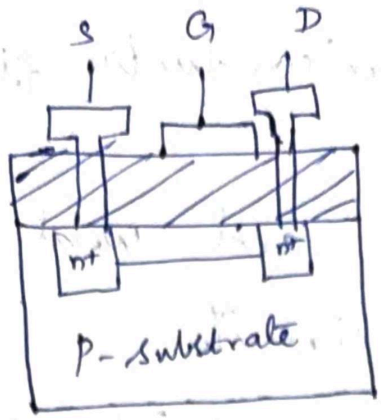
ii) Enhancement MOSFET

Depletion MOSFET

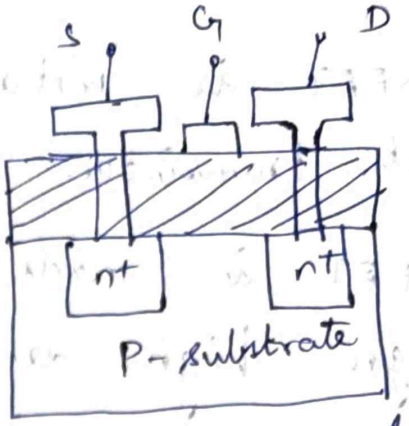


→ A slab of p-type material is formed from a silicon base referred to as substrate

→ The gate is insulated from n-channel by a thin SiO_2 layer



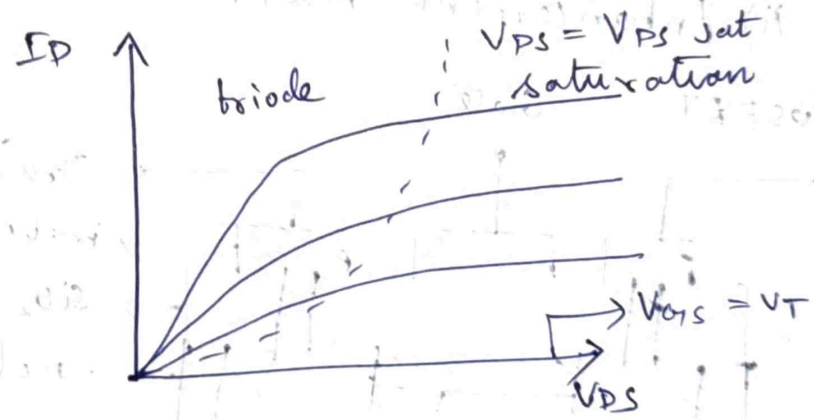
$V_{GS} = 0$



$V_{GS} = -$ ve value

→ In depletion mode, before applying gate voltage the channel is established between source and drain

→ when -ve voltage is applied to gate, it will pressure electron towards P type substrate and attract holes to N-type substrate



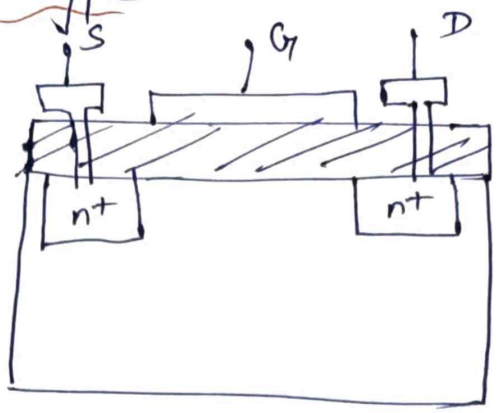
$I_D - V_{DS}$ characteristics

for $V_{GG} < V_T$, $I_D = 0$

→ As V_{DS} increases at a fixed V_{GS}, I_D increases in triode region due to increased in voltage

→ once pinch off is reached, further increase in V_{DS} increase I_D

Enhancement type



→ when V_{DS} & V_{gs} is set at some positive voltage, the drain and gate are positive with respect to source.

→ The positive potential will presure the channd

Rectifiers with Filters

* Rectifier converts ac. to pulsating d.c.
* P.n junction diode conducts in only one direction
* It conducts when forward biased and does not conduct when reverse biased

Important characteristics of a Rectifier

Waveform of load current:

The current through the load determines the waveform of the load voltage

Regulation of output Voltage

As load current changes, load voltages also changes, but load voltage must remain constant. So regulation is to study the effect of change in load current on load voltage

Rectifier efficiency:

It denotes how efficiently the rectifier converts ac power into dc power.

Peak Value of current in rectified circuit

Peak value is the maximum value of alternating current in the rectifier circuit

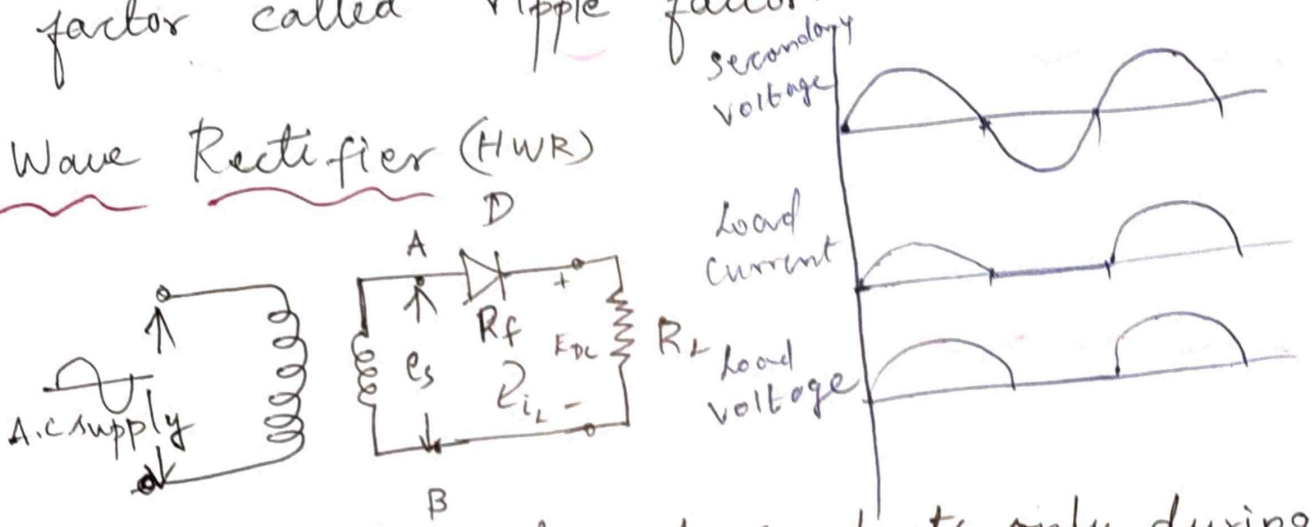
Peak Value of Voltage across rectifier element in reverse direction (PIV)

When diode is not conducting, the reverse voltage gets applied across the diode. the peak value of voltage is the Peak Inverse Voltage (PIV) of the diode

Ripple Factor

The amount of a.c. content is expressed by factor called ripple factor.

Half Wave Rectifier (HWR)



→ In HWR, rectifying element conducts only during positive half cycle of input a.c supply.

→ -ve half cycle are eliminated from the output

→ rectifier circuits are operated from a.c mains supply with frequency 50Hz

→ The transformer decides the peak value of the secondary voltage

N_1 → primary number of turns

N_2 → secondary number of turns

$$\frac{N_2}{N_1} = \frac{E_{sm}}{E_{pm}}$$

E_{sm} - Peak value of secondary a.c voltage

$e_s = E_{sm} \sin \omega t$

$\omega = 2\pi f$

f - supply freq

R_f - forward resistance of the diode

Operation:

* During +ve half cycle diode is forward biased and current flows in clockwise direction
* The current flowing thro' load resistance R_L is I_L (load current). The o/p is called pulsating d.c

Average D.C Load Current (I_{DC})

* The average or d.c value is obtained by integration
* average value is area under one complete cycle divided by base 2π

$i_L = I_m \sin \omega t$ for $0 \leq \omega t \leq \pi$

$i_L = 0$ for $\pi \leq \omega t \leq 2\pi$

I_m = Peak value of load current

$$I_{DC} = \frac{1}{2\pi} \int_0^{2\pi} i_L \, d(\omega t) = \frac{1}{2\pi} \int_0^{2\pi} I_m \sin(\omega t) \, d\omega t$$

$$\frac{N_2}{N_1} = \frac{E_{sm}}{E_{pm}}$$

E_{sm} - Peak value of secondary a.c voltage

$e_s = E_{sm} \sin \omega t$

$\omega = 2\pi f$

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no current flows thrau' π to 2π so limit becomes

$$I_{DC} = \frac{1}{2\pi} \int_0^{\pi} I_m \sin(\omega t)\, d(\omega t) = \frac{I_m}{2\pi} \left[-\cos \omega t \right]_0^{\pi}$$

$$= -\frac{I_m}{\pi} \left[\cos\pi - \cos 0 \right] = -\frac{I_m}{2\pi} \left[-1 - 1 \right] = \frac{I_m}{\pi}$$

$$I_{DC} = \frac{I_m}{\pi} = \text{average value}$$

Applying Kirchoff's voltage law

$$I_m = \frac{E_{sm}}{R_f + R_L + R_s}$$

R_s = Resistance of secondary winding of tranformer

Average DC. Load Voltage (E_{DC})

It is the product of average D.C load current and load Resistance R_L

$$E_{DC} = I_{DC} R_L$$

$$E_{DC} = \frac{I_m}{\pi} R_L = \frac{E_{sm}}{(R_f + R_L + R_s)\pi} \times R_L$$

$R_s + R_f << R_L$

$$E_{DC} = \frac{E_{sm}}{\pi \left[\frac{R_f + R_s}{R_L} + 1 \right]}$$

$\frac{R_f + R_s}{R_L} << 1$, so

$$\boxed{E_{DC} = \frac{E_{sm}}{\pi}}$$

RMS Value of Load Current (I_{RMS})

RMS → squaring, finding mean and then square root

$$I_{RMS} = \sqrt{\frac{1}{2\pi}\int_0^{\pi} (I_m \sin\omega t)^2 \, d(\omega t)} = \sqrt{\frac{1}{2\pi}\int_0^{\pi} I_m^2 \sin^2\omega t \, d(\omega t)}$$

$$= I_m \sqrt{\frac{1}{2\pi}\int_0^{\pi} \frac{1-\cos(2\omega t)}{2} d(\omega t)} = I_m \sqrt{\frac{1}{2\pi}\left\{\frac{\omega t}{2} - \frac{\sin(2\omega t)}{4}\right\}_0^{\pi}}$$

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

$$\boxed{I_{RMS} = \frac{I_m}{2}}$$

D.C Power Output (P_{DC})

$$P_{DC} = E_{DC} I_{DC} = I_{DC}^2 R_L$$

D.C Power output $= I_{DC}^2 R_L = \left(\frac{I_m}{\pi}\right)^2 R_L = \frac{I_m^2}{\pi^2} \cdot R_L$

$$P_{DC} = \frac{I_m^2}{\pi^2} \cdot R_L$$

$$I_m = \frac{E_{sm}}{R_f + R_L + R_s}$$

$$P_{DC} = \frac{E_{sm}^2 \cdot R_L}{\pi^2 \left[R_f + R_L + R_s\right]^2}$$

A.C Power Input (P_{AC})

The power i/p taken from secondary of transformer is power supplied to R_L, R_f & R_s

$$P_{AC} = I^2_{RMS} [R_L + R_f + R_s]$$

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

$$P_{AC} = \frac{I_m^2}{4} [R_L + R_f + R_s]$$

Rectifier Efficiency

* It is defined as ratio of output d.c power to input a.c power

$$\eta = \frac{\text{D.C output power}}{\text{A.C input power}} = \frac{P_{DC}}{P_{AC}}$$

$$\eta = \frac{\frac{I_m^2}{\pi^2} R_L}{\frac{I_m^2}{4} [R_f + R_L + R_s]} = \frac{(4/\pi^2) R_L}{[R_f + R_L + R_s]}$$

$$\eta = \frac{0.406}{1 + \left(\frac{R_f + R_s}{R_L}\right)}$$

If $(R_f + R_s) << R_L$, the theoretical efficiency

$$\% \eta_{max} = 0.406 \times 100 = 40.6\%.$$

40% of a.c power gets converted to d.c, remaining 60% is present in terms of ripples.

Ripple Factor

The output contains pulsating component called ripples.

ripple factor is defined as the ratio of RMS value of a.c. component in the output to average or d.c component present in output

$$\text{Ripple factor } \gamma = \frac{\text{R.M.S. Value of a.c component of output}}{\text{Average or d.c component of output}}$$

I_{ac} = RMS Value of a.c component present in output

I_{DC} = D.C component present in output

I_{RMS} = RMS value of total output current

$$I_{RMS} = \sqrt{I_{ac}^2 + I_{DC}^2}$$

$$I_{ac} = \sqrt{I_{RMS}^2 - I_{DC}^2}$$

$$\text{Ripple factor} = \frac{I_{ac}}{I_{DC}}$$

$$\gamma = \frac{\sqrt{I_{RMS}^2 - I_{DC}^2}}{I_{DC}}$$

$$V = \sqrt{\left(\frac{I_{RMS}}{I_{DC}}\right)^2 - 1}$$

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

$$\gamma = \sqrt{\left[\frac{\frac{I_m}{2}}{\frac{I_m}{\pi}}\right]^2 - 1} = \sqrt{\frac{\pi^2}{4} - 1} = \sqrt{1.4674}$$

$$\boxed{\gamma = 1.211}$$

indicates ripple in the output are 1.211 times d.c component (ie) 121.1 % of d.c component.

ripple factor is minimised using filter along with rectifiers

Load current:

* i_L composed of a.c and d.c components

$$i_L = I_m \left[\frac{1}{\pi} + \frac{1}{2}\sin\omega t - \frac{2}{3\pi}\cos 2\omega t - \frac{2}{15\pi}\cos 4\omega t + \ldots\right.$$

↓ d.c value (average)

↓ second harmonic

Peak Inverse Voltage (PIV)

→ is the peak voltage across the diode in reverse direction

PIV = E_{sm} = maximum value of secondary voltage = πE_{DC} / $I_{DC} = \pi$

Transformer Utilization Factor (T.U.F)

→ factor indicates how much is the utilization of transformer in the Ckt.

A.C power rating of transformer $= E_{RMS} I_{RMS}$

$$= \frac{E_{sm}}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}} = \frac{E_{sm} \cdot I_m}{2}$$

D.C power delivered to the load $= I_{DC}^2 \cdot R_L$

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

$$T.U.F = \frac{\text{D.C power delivered to the load}}{\text{A.C power rating of transformer}}$$

$$T.U.F = \frac{\left(\frac{I_m}{\pi}\right)^2 \times R_L}{\frac{E_{sm} \times I_m}{2}}$$

$E_{sm} = I_m R_L$

$$= \frac{I_m^2}{\pi^2} \times R_L \cdot \frac{2}{I_m^2 R_L} = \frac{2}{\pi^2} = 0.287$$

Voltage Regulation

The secondary voltage should not change with respect to load current. It is the change in output voltage as load changes from no load to full load.

$$\text{Voltage regulation} = \frac{V_{dc(NL)} - V_{dc(FL)}}{V_{dc(FL)}}$$

$$V_{dc(NL)} = \frac{E_{sm}}{\pi}$$

$$V_{dc(FL)} = I_{DC} R_L = \frac{I_m}{\pi} \times R_L = \frac{E_{sm}}{\pi(R_f + R_s + R_L)} \times R_L$$

$$\%R = \frac{\frac{E_{sm}}{\pi} - \frac{E_{sm}}{\pi}\frac{R_L}{(R_f+R_s+R_L)}}{\frac{E_{sm}}{\pi}\frac{R_L}{R_f+R_s+R_L}} \times 100$$

$$= 1 - \frac{\frac{R_L}{R_f+R_s+R_L}}{\frac{R_L}{R_f+R_s+R_L}} = \frac{R_f+R_s}{R_L} \times 100$$

Neglecting winding Resistance

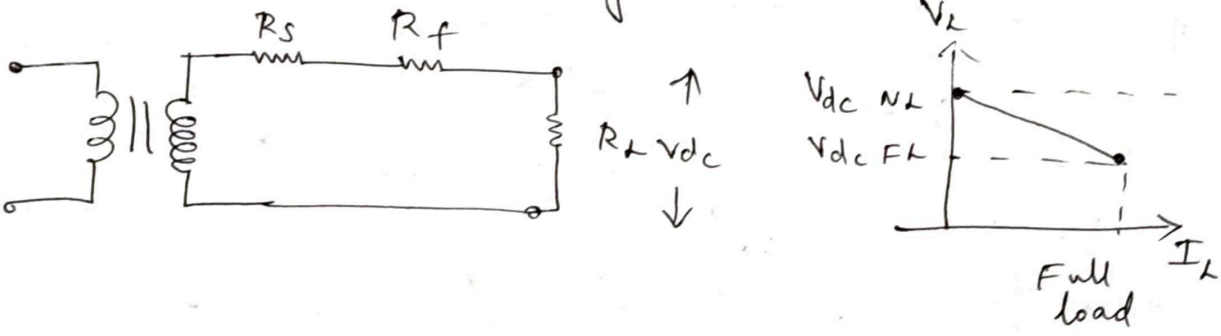
$$\%R = \frac{R_f}{R_L} \times 100$$

Regulation characteristics

→ As load current increases, drop across R_s & R_f goes on increasing, d.c o/p voltage decreases as load change from NL to FL

→ To keep drop across R_s & R_f minimum, R_s & R_f must be small.

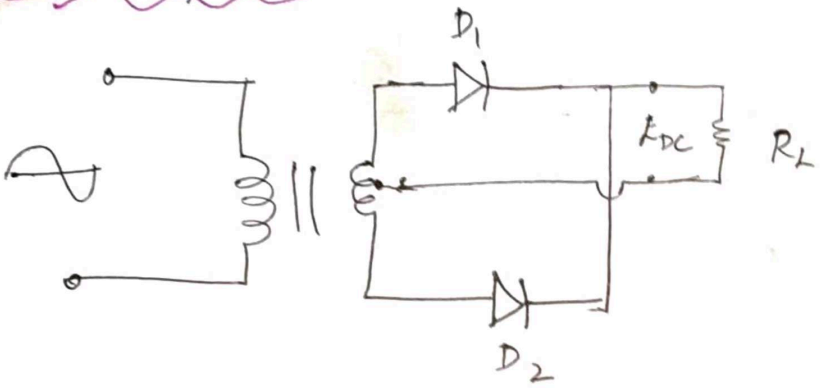
→ the graph of load voltage against load current is called regulation characteristics

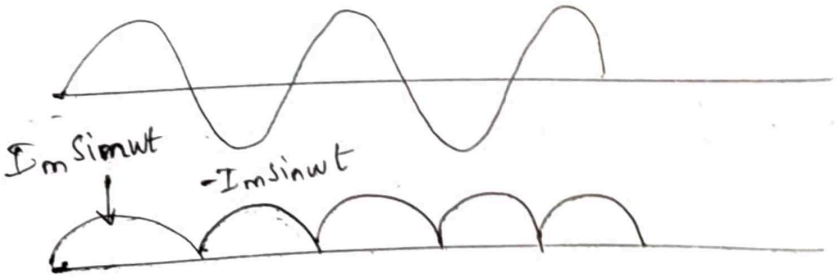


Disadv of HWR

1) V is high
2) η is 40% which is <
3) low T.U.F
4) cause d.c saturation of the core

Full Wave Rectifier:





Average D.C Load Current

$$i_L = I_m \sin\omega t \quad 0 \le \omega t \le \pi$$

$$i_L = -I_m \sin\omega t \quad \pi \le \omega t \le 2\pi$$

$$I_{av} = I_{DC} = \frac{1}{2\pi}\int_0^{2\pi} i_L \, d(\omega t) = \frac{I_m}{2\pi}\int_0^{\pi} \sin\omega t \, d\omega t + \int_{\pi}^{2\pi} -\sin\omega t \, d\omega t$$

$$= \frac{I_m}{2\pi}\left[-\cos\omega t\right]_0^{\pi} - \left(-\cos\omega t\right)_{\pi}^{2\pi}$$

$$= \frac{I_m}{2\pi}\left[-\cos\pi + \cos 0 + \cos 2\pi - \cos\pi\right] = \frac{I_m}{2\pi}\left[-(-1)+1+1-(-1)\right]$$

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

$$E_{DC} = \frac{2E_{sm}}{\pi}$$

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

$$P_{DC} = \frac{4}{\pi^2} \frac{E_{sm}^2}{(R_s + R_f + R_L)^2} \times R_L$$

$$P_{AC} = \frac{E_{sm}^2}{2(R_f + R_s + R_L)}$$

$$\eta = \frac{8}{\pi^2} \times 100 = 81.2\%$$

$$\gamma = 0.48$$

$$I_{d_1} = I_m \left[\frac{1}{\pi} + \frac{1}{2} \sin \omega t - \frac{2}{3\pi} \cos 2\omega t - \frac{2}{15\pi} \cos 4\omega t \right]$$

$$i_2 = I_{d_1} + I_{d_2} = I_m \left[\frac{2}{\pi} - \frac{4}{3\pi} \cos 2\omega t - \frac{4}{15\pi} \cos 4\omega t + \cdots \right]$$

$$PIV = 2E_{sm} - 0.7$$

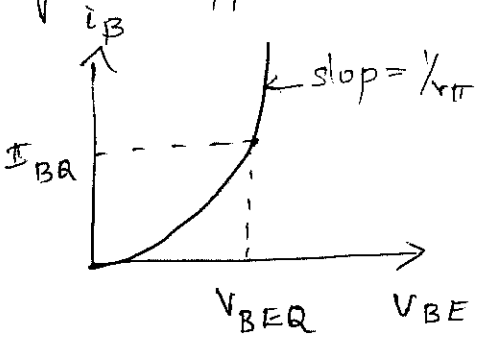
$$TUF = \frac{8}{\pi^2} = 0.812$$

$$\% R = \frac{R_f}{R_L} \times 100$$

Unit - II BJT Amplifiers

Small signal hybrid π equivalent

BJT as two port n/w, small signal i/p base current related to i/p voltage by



$$V_{be} = i_b r_\pi$$

$1/r_\pi$ = slope of $i_B - V_{BE}$ curve

i_b - time varying base current is given by

$$i_b = \left(\frac{I_{BQ}}{V_T}\right) \times V_{be}$$

$$r_\pi = \frac{V_{be}}{i_b} = \frac{V_T}{I_{BQ}} = \frac{\beta V_T}{I_{CQ}}$$

. The resistance r_π is called diffusion resistance or base-emitter input resistance.

The output collector current is a function of base emitter voltage and independent of collector-emitter voltage. The relation is

$$\Delta i_c = \frac{\partial i_c}{\partial V_{BE}}\bigg|_{Q\,pt} \Delta V_{BE}$$

$$i_c = \frac{\partial i_c}{\partial V_{BE}}\bigg|_{Q\,pt} \cdot V_{be}$$

$$i_c = I_S \exp\left(\frac{V_{BE}}{V_T}\right)$$

$$\frac{\partial i_c}{\partial V_{BE}}\bigg|_{Q\,pt} = \frac{1}{V_T} \cdot I_S \exp\left(\frac{V_{BE}}{V_T}\right)\bigg|_{Q\,pt} = \frac{I_{CQ}}{V_T}$$

term $I_S \exp\left(\frac{V_{BE}}{V_T}\right)$ is I_{CQ} at Q pt.

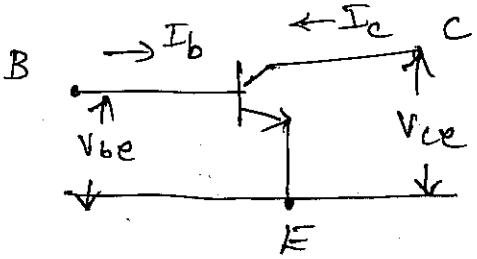
The term $\frac{I_{CQ}}{V_T}$ is conductance, relates collector current to voltage in base emitter ckt. the parameter is called transconductance given by

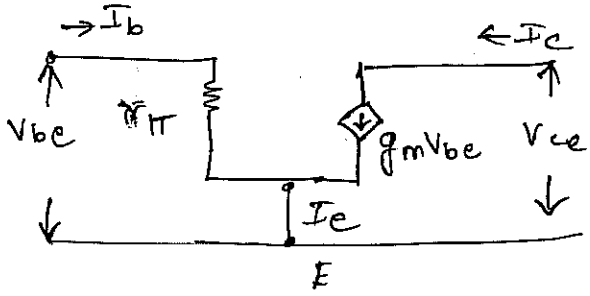
$$g_m = \frac{I_{CQ}}{V_T}$$

r_π, g_m → Q pt. parameters

r_π inv. prop. to I_{CQ}, $g_m \propto I_{CQ}$

Small signal parameters





Common - Emitter current Gain

→ defined as β_{ac} (does not include leakage current)

β_{dc} → includes leakage current $\left(\frac{I_c}{I_B}\right)$

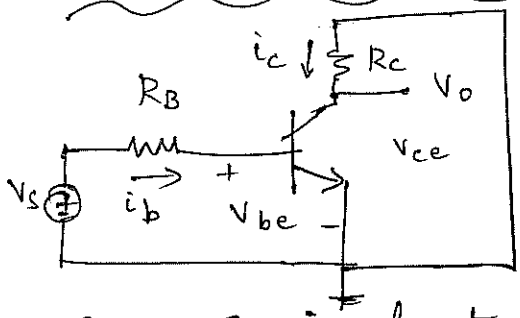
By neglecting leakage current

$$\beta_{dc} = \beta_{ac}$$

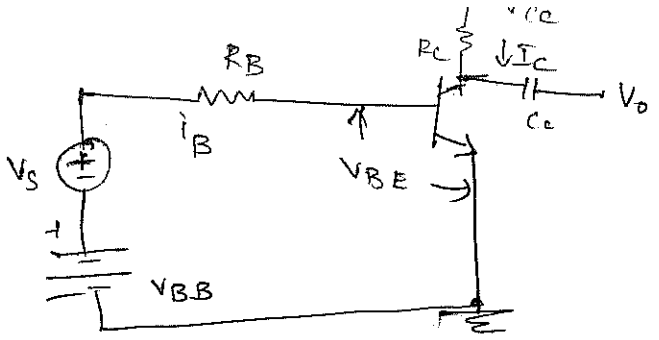
x|y r_π & g_m

$$r_\pi g_m = \left(\frac{\beta V_T}{I_{CQ}}\right)\left(\frac{I_{CQ}}{V_T}\right) = \beta$$

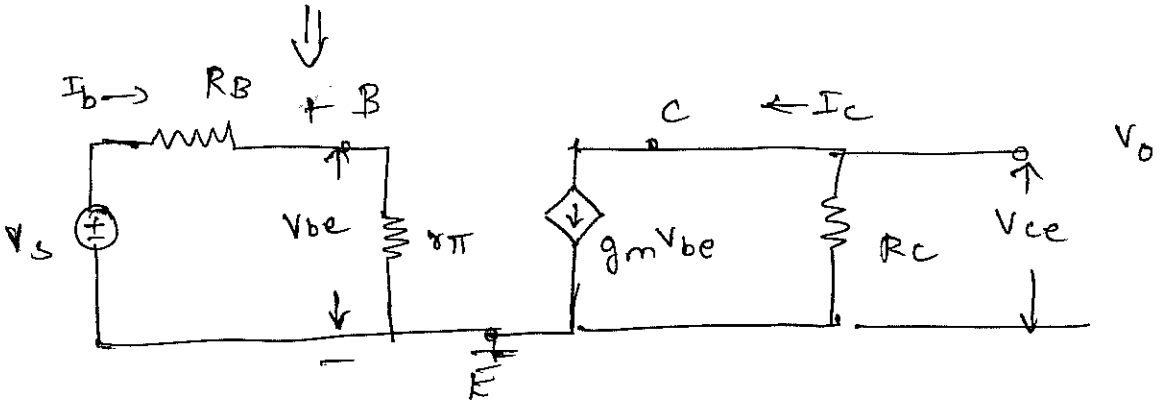
Common Emitter Amplifier



a.c equivalent



CE Amplifier



Small signal voltage gain (A_v)

defined as ratio of output signal voltage to input signal voltage. It is given by

$$A_v = \frac{V_o}{V_s}$$

$$V_o = V_{ce} = -(g_m V_{be}) R_c$$

Applying voltage divider rule

$$V_{be} = \left(\frac{r_\pi}{r_\pi + R_B}\right) \times V_s$$

$$A_v = \frac{V_o}{V_s} = -(g_m R_c) \cdot \left(\frac{r_\pi}{r_\pi + R_B}\right)$$

Hybrid π equivalent ckt including Early Effect

According to Early Effect, I_c vary with V_{ce}.

$$r_o = \left.\frac{\partial V_{CE}}{\partial I_c}\right|_{@pt} = \frac{V_A}{I_{cQ}}$$

$$A_v = \frac{V_o}{V_s} = -g_m(r_o \| R_c)\left(\frac{R_1 \| R_2 \| r_\pi}{R_1 \| R_2 \| r_\pi + R_s}\right)$$

$$A_v = -g_m(r_o \| R_c)\left(\frac{R_i}{R_i + R_s}\right)$$

Output resistance (R_o)

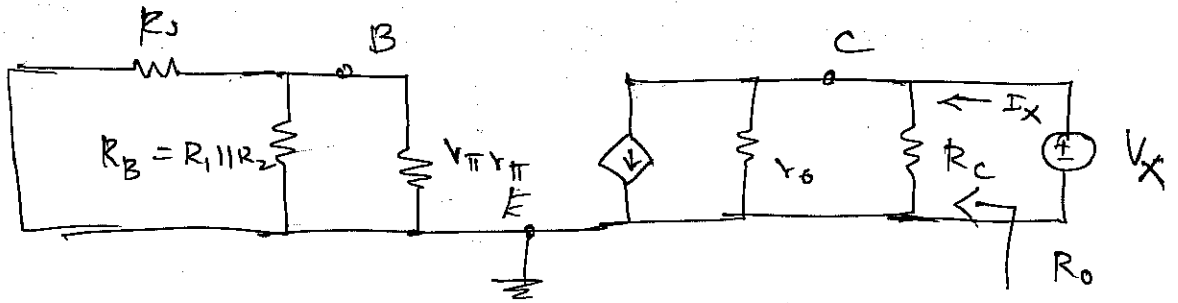
$V_s \rightarrow$ set to 0.

No excitation to i/p portion of circuit $V_\pi = 0$.

$g_m V_\pi = 0$ (o.c).

The o/p resistance looking back into o/p terminals is given by

$$R_o = \frac{V_x}{I_x} = r_o \| R_c$$



Early Effect

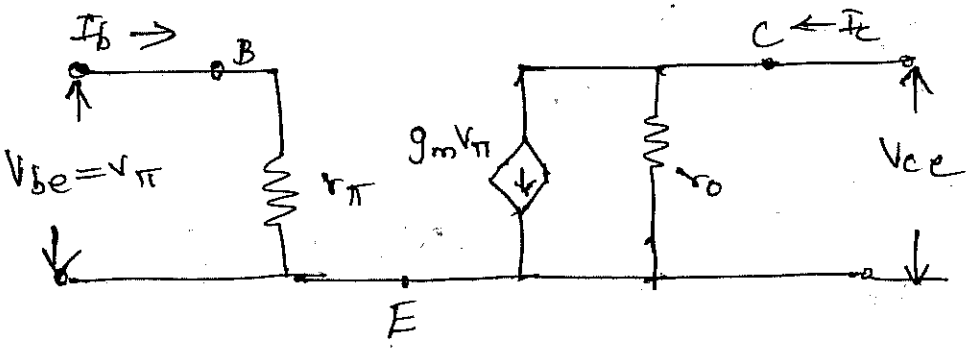
According to Early effect, the collector current does vary with collector emitter voltage. The relation between I_c and V_{CE} is given by

$$r_o = \frac{\partial V_{CE}}{\partial i_c}\Big|_{Q\,pt} = \frac{V_A}{I_{CQ}}$$

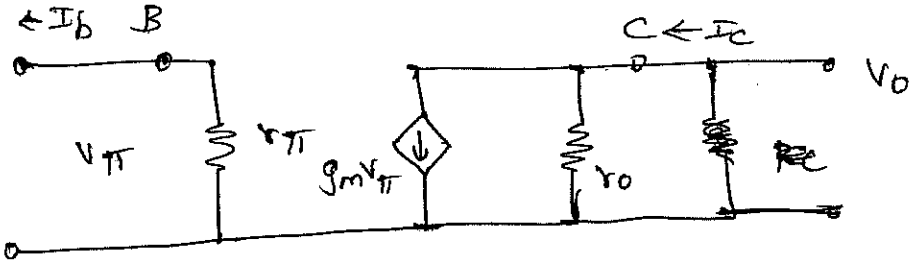
$V_A \rightarrow$ early voltage

This r_o is called small signal transistor o/p resistance

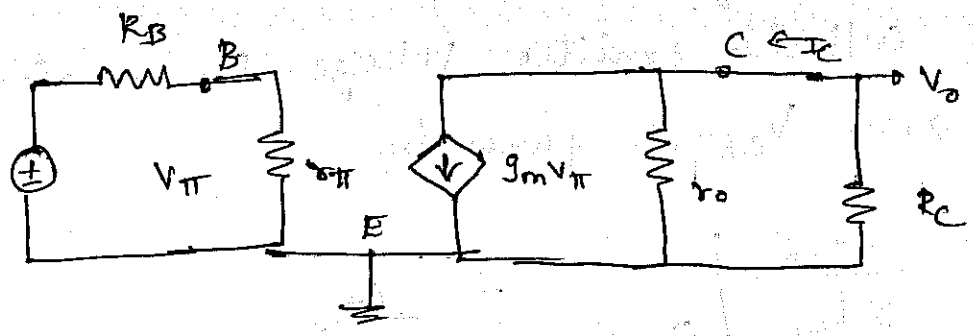
Hybrid π – equivalent (npn)



Hybrid π – equivalent pnp



CE amplifier with early effect



output voltage is

$$V_o = -g_m V_\pi (r_o \| R_C)$$

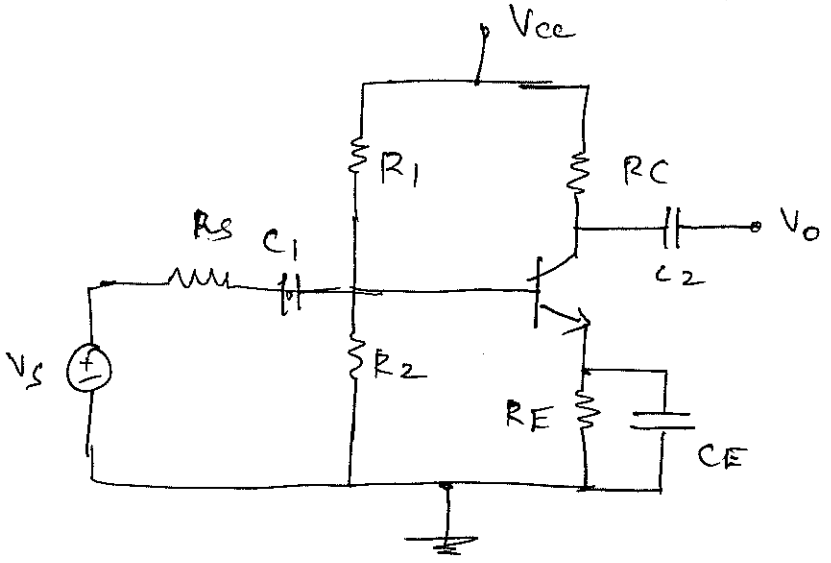
$$V_\pi = \left(\frac{r_\pi}{r_\pi + R_B}\right) V_s$$

$$A_v = \frac{V_o}{V_s}$$

$$= -g_m (r_o \| R_C)\left(\frac{r_\pi}{r_\pi + R_B}\right)$$

$$= \frac{-\beta}{r_\pi + R_B} (r_o \| R_C) \qquad \because g_m r_\pi = \beta$$

Small signal Analysis of common Emitter Amplifier

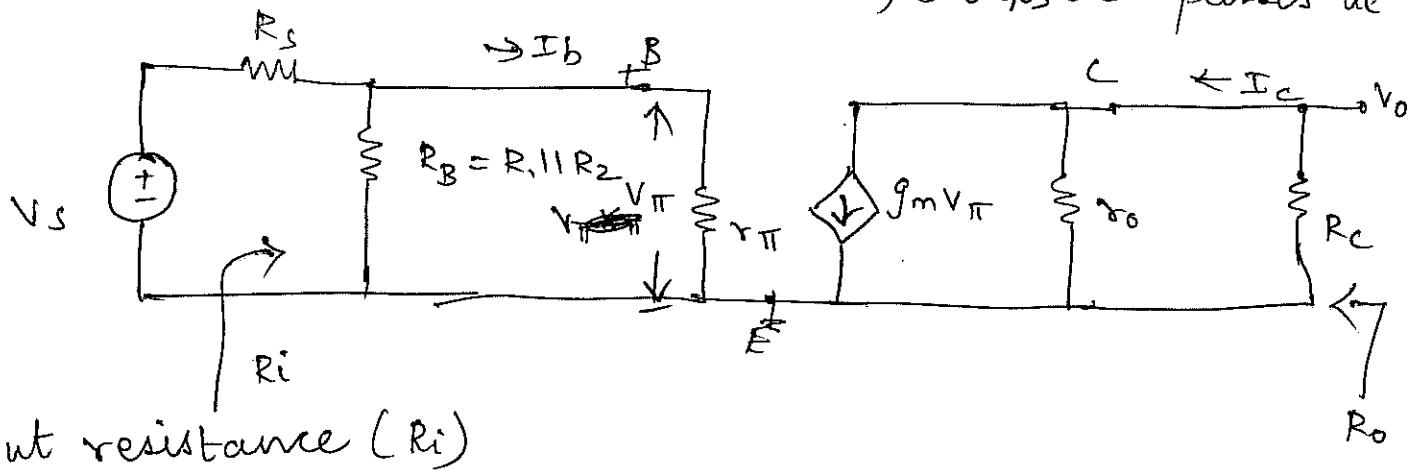


→ Signal source coupled into base by C_c (provides dc isolation

→ R_1, R_2, R_E - dc biasing

→ C_E → provides low reactance path if not inserted produces voltage drop across R_E ↓ V_o, ↓ gain

C_c → couples o/p of amp. to load, blocks dc passes ac



Input resistance (R_i)

$$R_i = R_1 \| R_2 \| r_\pi$$

Voltage gain (A_V)

$$V_o = -g_m V_\pi (r_o \| R_c) \quad \text{——} \textcircled{1}$$

$$V_\pi = \left(\frac{R_1 \| R_2 \| r_\pi}{R_1 \| R_2 \| r_\pi + R_s} \right) V_s \quad - \textcircled{2}$$

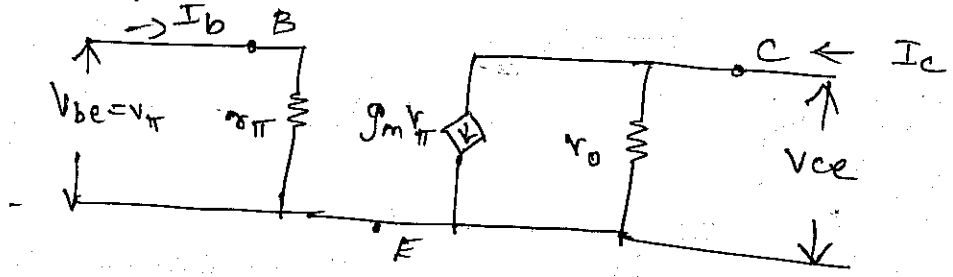
Sub ② in ①

$$V_o = -g_m (r_o \| R_c) \left(\frac{R_1 \| R_2 \| r_\pi}{R_1 \| R_2 \| r_\pi + R_s} \right) V_s$$

V_A = Early Voltage

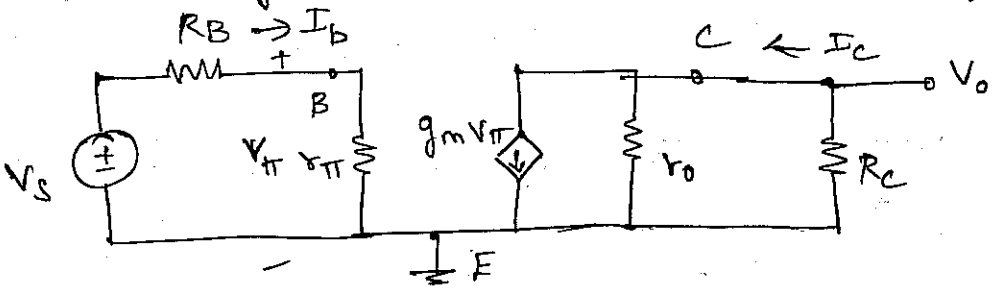
r_o is called small signal transistor o/p resistance.

Considering r_o hybrid-π equivalent ckt is



for pnp transistor current direction are reversed

Small signal equivalent including early effect



the o/p voltage is given by

$$V_o = -g_m V_\pi (r_o \| R_c)$$

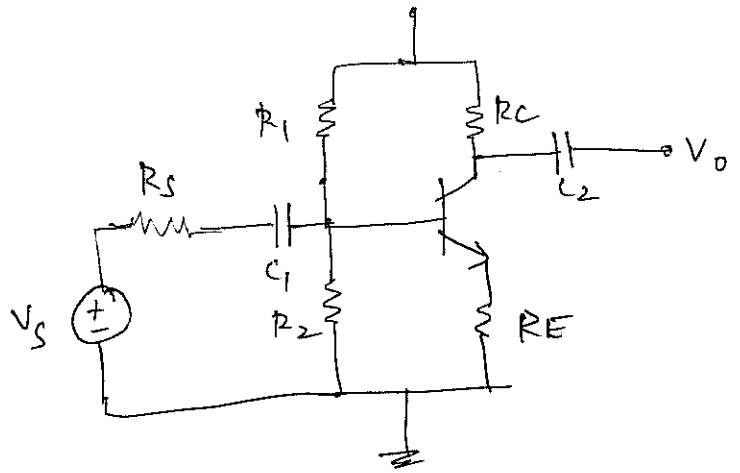
Applying voltage divider rule

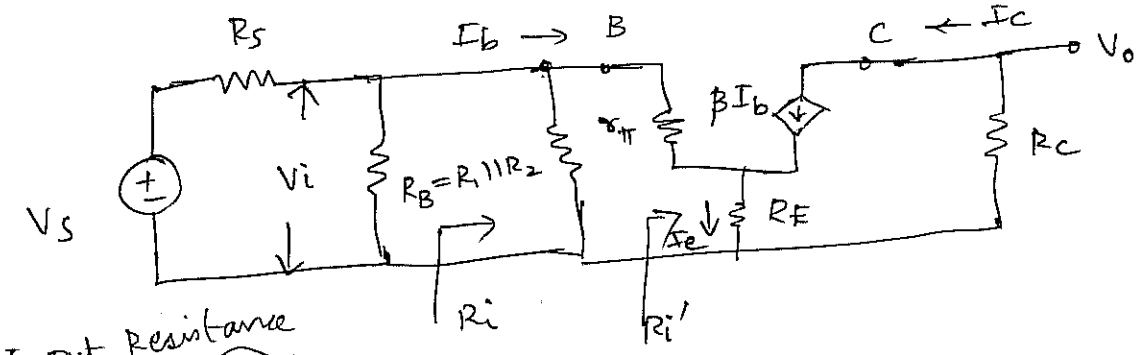
$$V_\pi = \left(\frac{r_\pi}{r_\pi + R_B}\right) V_S$$

$$A_V = \frac{V_o}{V_S} = -g_m (r_o \| R_c) \left(\frac{r_\pi}{r_\pi + R_B}\right)$$

$$= \left(\frac{-\beta}{r_\pi + R_B}\right)(r_o \| R_c) \qquad \because g_m r_\pi = \beta$$

Common Emitter with Unbypassed R_E





Input Resistance

$$V_i = I_b r_\pi + (1+\beta) I_b R_E$$

$$R_i' = \frac{V_i}{I_b} = r_\pi + (1+\beta) I_b R_E$$

$$R_i = R_1 \| R_2 \| R_i' = R_B \| R_i'$$

Voltage gain A_V

$$V_o = -(\beta I_b) R_C$$

$$A_V = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s}$$

$$V_i = I_b R_i' \quad \& \quad \frac{V_i}{V_s} = \frac{R_i}{R_i + R_s}$$

$$A_V = \frac{-\beta (I_b) R_C}{I_b R_i'} \times \frac{R_i}{R_i + R_s}$$

$$= \frac{-\beta R_C}{r_\pi + (1+\beta) R_E} \times \frac{R_i}{R_i + R_s}$$

if $R_i >> R_s$ & if $(1+\beta) R_E >> r_\pi$

$$A_V = \frac{-\beta R_C}{(1+\beta) R_E} \approx \frac{-R_C}{R_E}$$

Adv

1) A_V less independent on current gain (β)
2) R_E unbypassed $R_i >> R_i >> R_s$ & reduces loading effect

Ac. load line

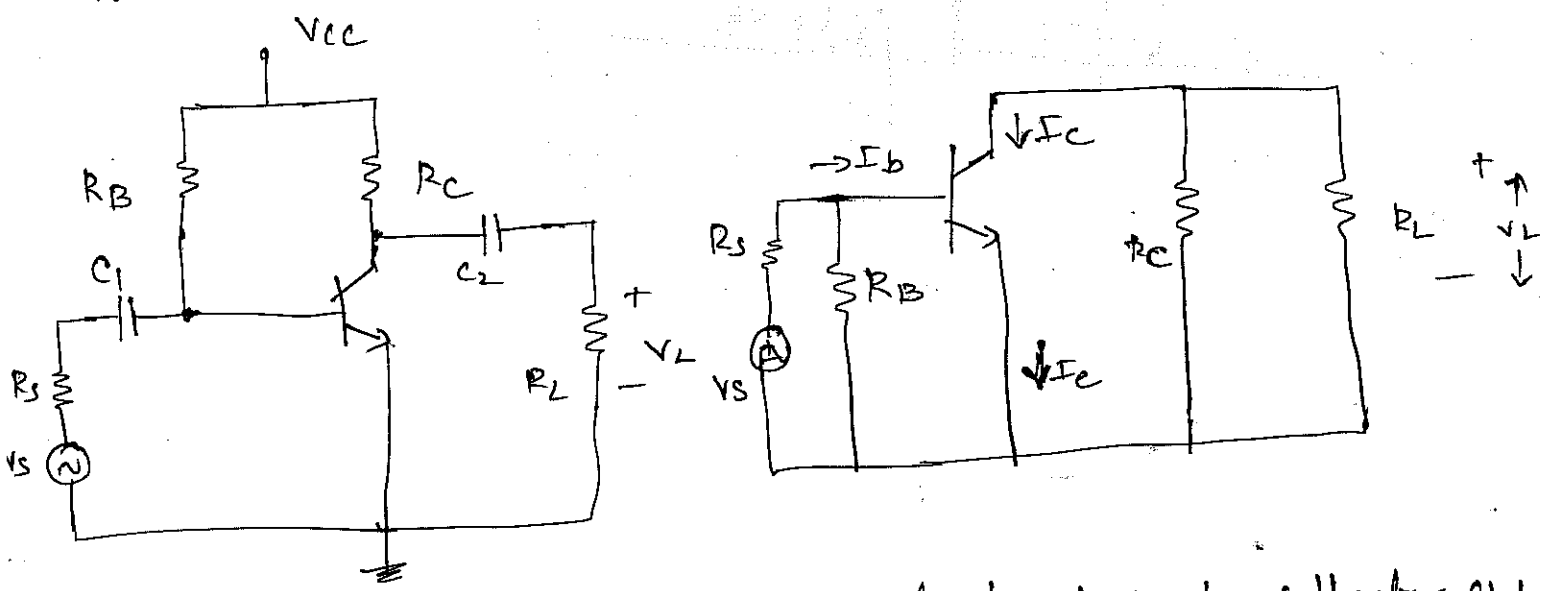
→ DC load line → relation b/w Q pt. & Tr. Characteristics

→ when capacitors are included in transistor, a new effective load line called ac load line exists.
(gives relation b/w small signal response & Tr. characteristics)

For ac. analysis

* capacitors acts as short, dc source by short

The resultant ac. equivalent



The collector ckt resistance seen by dc bias current I_{CQ} is

$$R_{dc} = R_c$$

The collector ckt current sees collector ckt. resistance

$$R_{ac} = R_c \| R_L = \frac{R_c R_L}{R_c + R_L}$$

$$R_{ac} \neq R_{dc}$$

Apply KVL to collector ckt.

$$V_{ce} = I_c R_{ac} \quad — (1)$$

V_{ce} → ac. collector to emitter Volt

I_c : ac collector current

$$I_c = i_c - I_{CQ} \text{ \& } V_{ce} = V_{CEQ} - V_{CE} \quad — (2)$$

Sub (2) in (1)

$$V_{CEQ} - V_{CE} = (i_c - I_{CQ}) R_{ac}$$

i_c → Total instantaneous collector current

V_{CE} → Total instantaneous

Rearranging

$$i_c = \frac{V_{CEQ}}{R_{ac}} - \frac{V_{CE}}{R_{ac}} + I_{CQ} \quad -(3)$$

If $i_c = I_{CQ}$ is substituted in (3) $V_{CE} = V_{CEQ}$.

This indicates a.c load line intersect d.c. load line at Q pt.

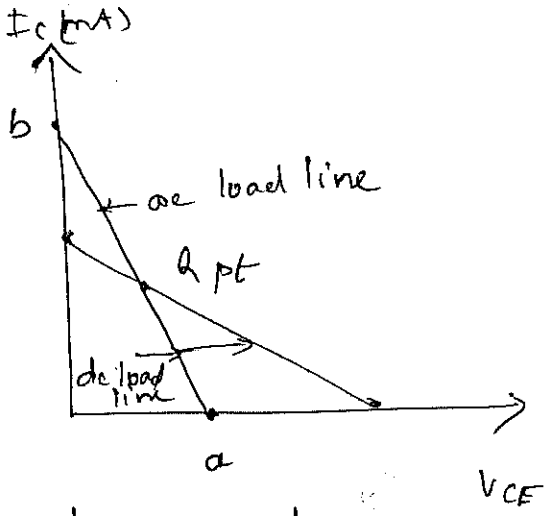
Point of intersection

Point a set $i_c = 0$ in (3)

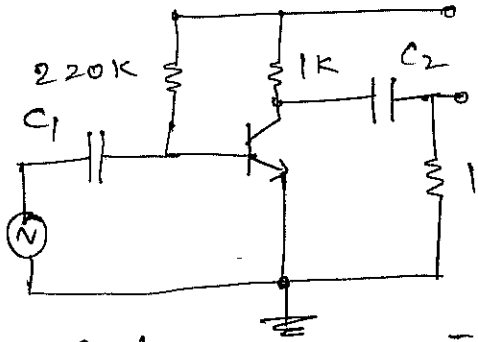
$$V_{CE\,max} = V_{CEQ} + I_{CQ} R_{ac}$$

Point b set $V_{CE} = 0$

$$I_{c\,max} = \frac{V_{CEQ}}{R_{ac}} + I_{CQ}$$



eg: 3.5.1 The ckt. of BJT amp. is shown draw ac & d.c load line. Also find Q pt. Assume $V_{BE} = 0.7$ & $\beta = 100$



$V_{cc} = 12V$

Step 1: obtain I_{CQ} & V_{CEQ} point A & point B

Apply KVL to collecto ckt.

$V_{cc} - I_c R_c - V_{CEQ} = 0$

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

$= 12 - 5.136 \times 10^{-3} \times 1000 = 6.864\,V$

Apply KVL to base ckt.

$V_{cc} - I_B R_B - V_{BE} = 0$

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

$$= \frac{12 - 0.7}{220K} = 51.36\,\mu A$$

$I_{CQ} = \beta I_B = 100 \times 51.36\mu A$

$= 5.136\,mA$

Q pt. is $I_{CQ}, V_{CEQ} = 3.424\,mA, 6.864\,V$

Axes inter section

Point A: $V_{CE} = V_{cc} = 12V$ at $I_c = 0$

point B: $I_C = V_{cc}/R_{dc} = V_{cc}/R_c = \frac{12}{1K} = 12m.$

Step 2: obtain R_{ac}. Point a, b

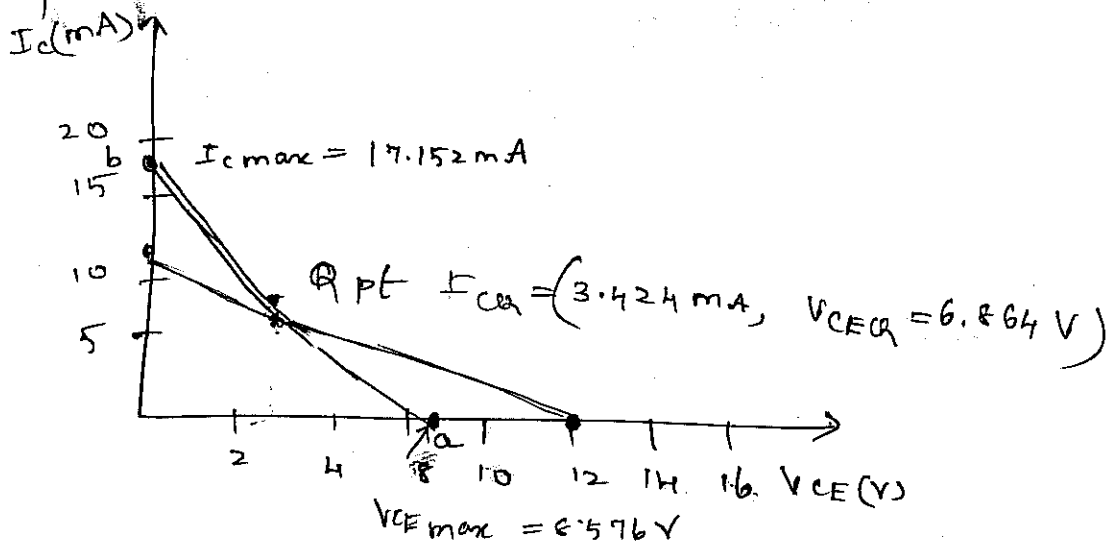
$R_{ac} = R_c \| R_L = 1K \| 1K = 500\,\Omega$

Point a: $V_{CE\,max} = V_{CEQ} + I_{CQ} R_{ac} = 6.864 + 3.424 \times 10^{-3} \times 500$

$= 8.576V$

Point b: $I_{cmax} = \frac{V_{CEQ}}{R_{ac}} + I_{CQ} = \frac{6.864}{500} + 3.424 \times 10^{-3} = 17.132mA$

Step 3:



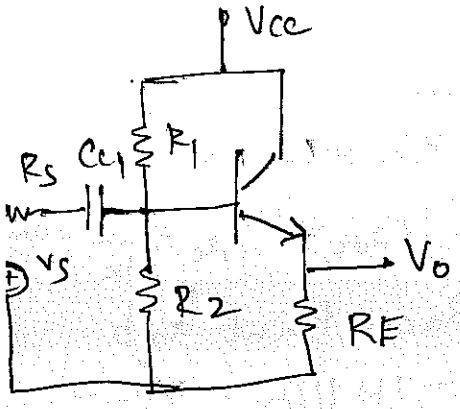
Voltage Swing Limitation

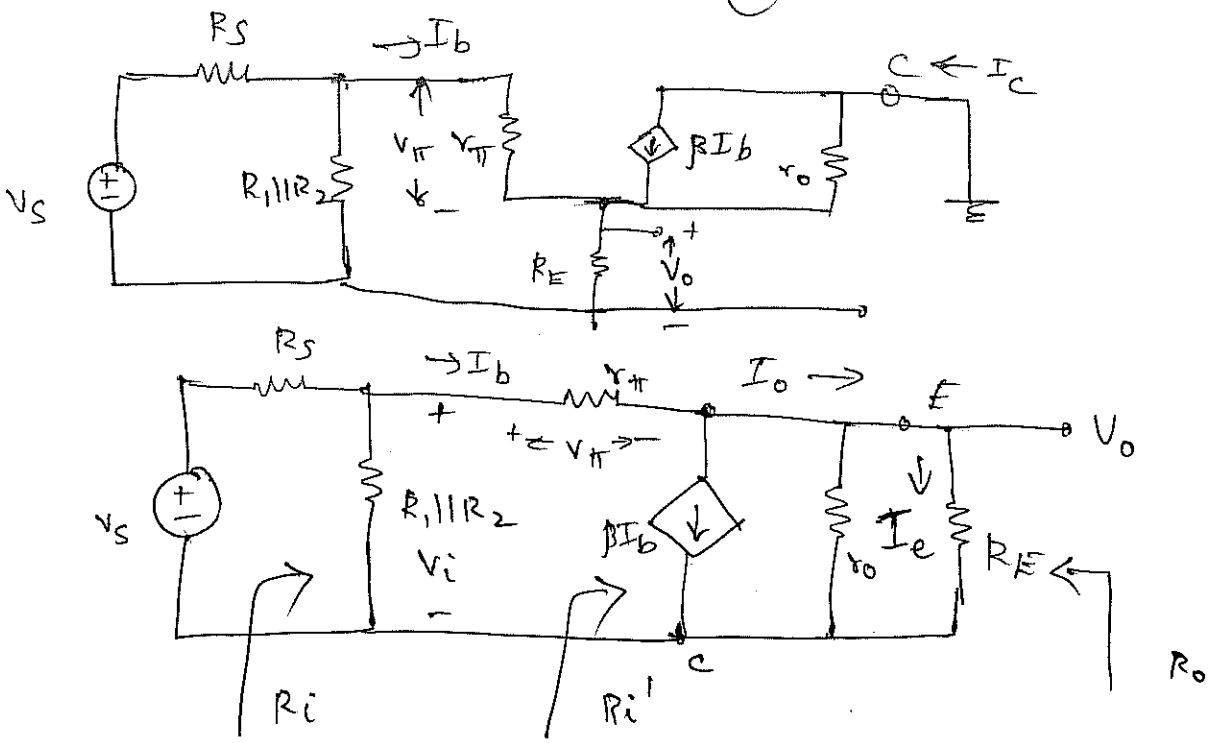
→ Symmetrical sinusoidal signals are applied to i/p get amplified sinusoidal signals at o/p

→ possible to obtain maximum o/p symmetrical ~~signal~~ swing amplifier can provide using ac load line

→ If o/p exceeds this limit, portion of o/p signal clipped resulting signal distortion

Common collector Amplifier Emitter follower.





Input Resistance (R_i)

$$I_o = I_b + \beta I_b = (1+\beta) I_b$$

$$V_o = I_o (r_o \| R_E) = (1+\beta) I_b (r_o \| R_E)$$

Apply KVL to base emitter loop

$$V_i = V_\pi + V_o = r_\pi I_b + (1+\beta) I_b (r_o \| R_E)$$

$$= I_b \left[r_\pi + (1+\beta)(r_o \| R_E) \right]$$

$$R_i' = \frac{V_i}{I_b} = r_\pi + (1+\beta)(r_o \| R_E)$$

$$R_i = R_1 \| R_2 \| R_i'$$

Voltage gain (A_V)

$$A_V = \frac{V_o}{V_S} = \frac{V_o}{V_i} \times \frac{V_i}{V_S} \qquad V_i = \left(\frac{R_i}{R_i + R_S} \right) V_S$$

$$= \frac{I_b (1+\beta)(r_o \| R_E)}{I_b \left[r_\pi + (1+\beta)(r_o \| R_E) \right]} \cdot \frac{R_i}{R_i + R_S}$$

$$A_V = \frac{(1+\beta)(r_o \| R_E)}{r_\pi + (1+\beta)(r_o \| R_E)} \cdot \frac{R_i}{R_i + R_S}$$

Current gain (A_i)

$$A_i = \frac{I_e}{I_i} = \frac{I_e}{I_o} \times \frac{I_o}{I_b} \times \frac{I_b}{I_i}$$

$$\frac{I_e}{I_o} = \frac{r_o}{r_o + R_E} \quad \text{— (1)}$$

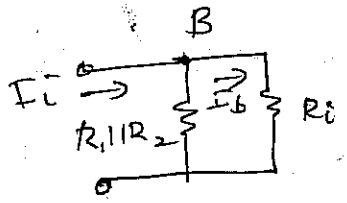
Apply KCL at middle node

$$I_o = (1+\beta)\, I_b$$

$$\frac{I_o}{I_b} = (1+\beta) \quad \text{— (2)}$$

Using collector divider at node B

$$I_b = \frac{R_1 \| R_2}{(R_1 \| R_2) + R_i'} I_i$$



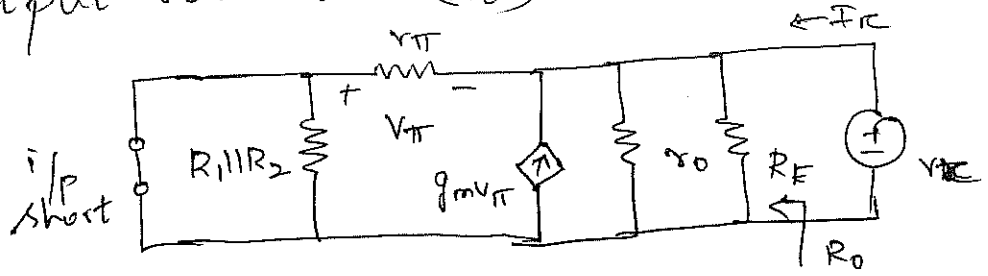
$$\frac{I_b}{I_i} = \frac{R_1 \| R_2}{(R_1 \| R_2) + R_i'} \quad \text{— (3)}$$

$$A_i = \frac{I_e}{I_i} = \frac{I_e}{I_o} \times \frac{I_o}{I_b} \times \frac{I_b}{I_i} = \left(\frac{r_o}{r_o + R_E}\right) \cdot (1+\beta) \left(\frac{R_1 \| R_2}{R_1 \| R_2 + R_i'}\right)$$

If $R_1 \| R_2 >> R_i'$ & $r_o >> R_E$

$$\boxed{A_i = (1+\beta)}$$

Output resistance (R_o)



$V_s = 0$, $R_s = 0$

Apply test voltage V_x to measure I_x.

↓ Test volt

$$R_o = \frac{V_x}{I_x}$$

Apply KVL to outer loop

$$V_\pi = -V_x$$

Apply KCL

$$I_x + g_m V_\pi = \frac{V_x}{r_\pi} + \frac{V_x}{r_o} + \frac{V_x}{R_E}$$

$$I_x - g_m V_x = \frac{V_x}{r_\pi} + \frac{V_x}{r_o} + \frac{V_x}{R_E}$$

$$\frac{I_x}{V_x} = \frac{1}{R_o} = g_m + \frac{1}{r_\pi} + \frac{1}{r_o} + \frac{1}{R_E}$$

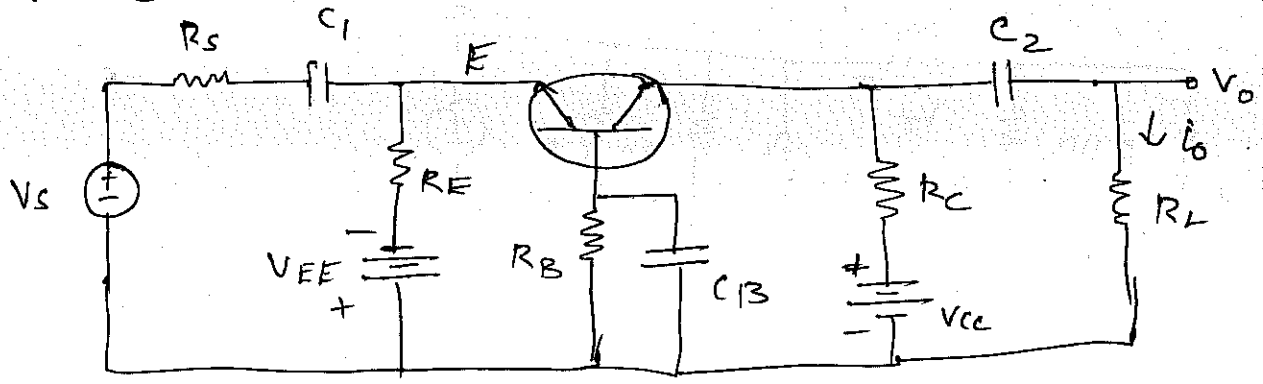
$$R_o = \frac{1}{g_m} \| r_\pi \| r_o \| R_E$$

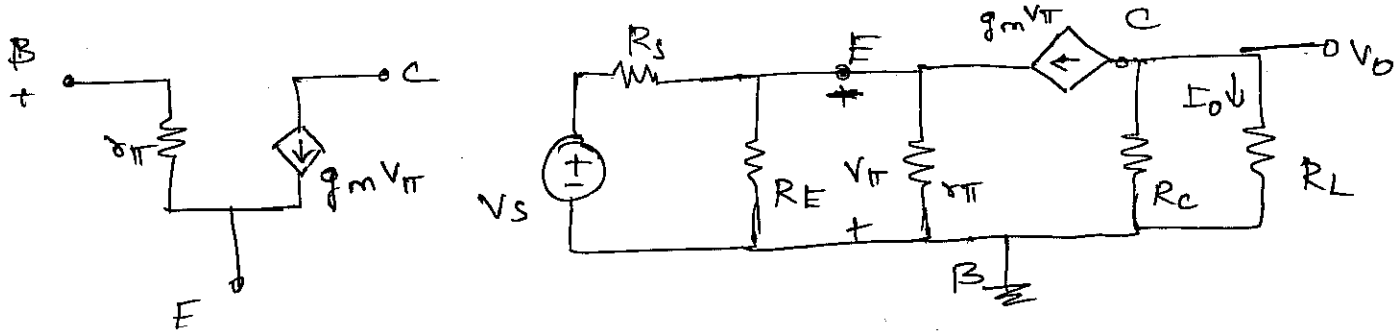
o/p resistance written in

$$\frac{1}{R_o} = \left(g_m + \frac{1}{r_\pi}\right) + \frac{1}{r_o} + \frac{1}{R_E} = \left(\frac{1+\beta}{r_\pi}\right) + \frac{1}{r_o} + \frac{1}{R_E}$$

$$R_o = \frac{r_\pi}{1+\beta} \| r_o \| R_E$$

Common Base Amplifier





Voltage gain :

$$V_o = (-g_m V_\pi)\, R_C \| R_L \qquad \text{——— (1)}$$

Apply KCL to emitter node

$$\frac{V_s - (+V_\pi)}{R_S} + \frac{V_\pi}{R_E} + \frac{V_\pi}{r_\pi} + g_m V_\pi = 0 \qquad \text{——— (2)}$$

$$V_\pi \left(\frac{1}{R_S} + \frac{1}{R_E} + \frac{1}{r_\pi} + g_m \right) = -\frac{V_s}{R_S} \qquad \text{— (3)}$$

$\left[\because \beta = g_m r_\pi \right]$

$$V_\pi \left(\frac{1}{R_S} + \frac{1}{R_E} + \frac{(1+\beta)}{r_\pi} \right) = \frac{V_s}{R_S} \qquad \text{— (4)}$$

$$V_\pi = \frac{-V_s}{R_S} \left[R_S \| R_E \| \frac{r_\pi}{1+\beta} \right] \qquad \text{——— (5)}$$

Sub (5) in (1)

$$V_o = -g_m \left[\frac{-V_s}{R_S} \left[R_S \| R_E \| \frac{r_\pi}{1+\beta} \right] \right] R_C \| R_L$$

$$A_V = \frac{V_o}{V_s} = g_m (R_c \| R_L)$$

Current gain (A_i)

Apply KCL to emitter node

$$I_i + \frac{V_\pi}{R_E} + \frac{V_\pi}{r_\pi} + g_m V_\pi = 0$$

$$V_\pi \left(\frac{1}{R_E} + \frac{1}{r_\pi} + g_m \right) = -I_i$$

$$V_\pi \left[\frac{1}{R_E} + \frac{(1+\beta)}{r_\pi} \right] = -I_i$$

$$V_\pi = -I_i \left[R_E \| \frac{r_\pi}{1+\beta} \right] \quad \text{— (8)}$$

Using current division formula

$$I_o = (-g_m V_\pi) \frac{R_c}{R_c + R_L} \quad \text{— (9)}$$

Sub (8) in (9)

$$I_o = -g_m \left[-I_i \left[R_E \| \frac{r_\pi}{1+\beta} \right] \right] \left[\frac{R_c}{R_c + R_L} \right]$$

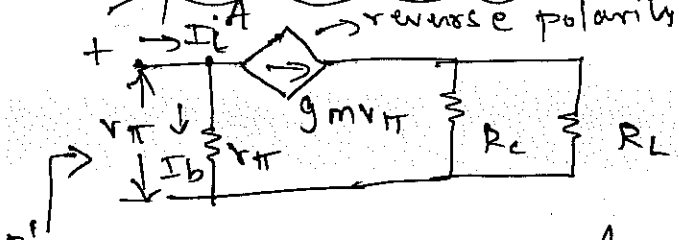
$$A_i = \frac{I_o}{I_i} = \frac{g_m R_c}{R_c + R_L} \left[R_E \| \frac{r_\pi}{1+\beta} \right]$$

$R_E \to \infty \quad R_L \to 0$

$$A_i = \frac{g_m r_\pi}{1+\beta} = \frac{\beta}{1+\beta} = \alpha$$

α – CB current gain of Tr.

Input Resistance



Apply KCL to I/p node A

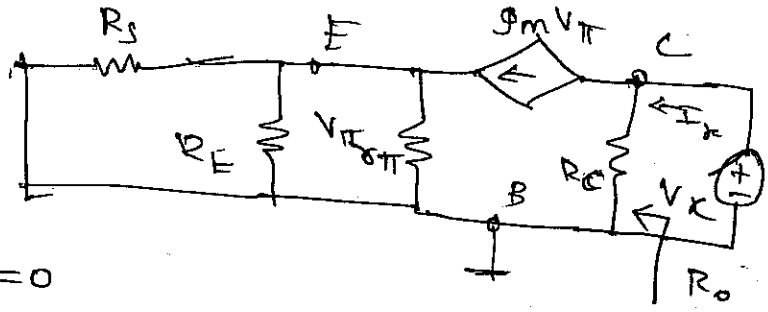
$$I_i = I_b + g_m V_\pi = \frac{V_\pi}{r_\pi} + g_m V_\pi = V_\pi \left(\frac{1 + g_m r_\pi}{r_\pi} \right)$$

$$= V_\pi \frac{(1+\beta)}{r_\pi}$$

$$R_i' = \frac{V_\pi}{I_i} = \frac{r_\pi}{1+\beta} = r_e$$

output Resistance (R_o)

$V_s \rightarrow$ set to 0



$$g_m V_\pi + \frac{V_\pi}{r_\pi} + \frac{V_\pi}{R_E} + \frac{V_\pi}{R_s} = 0$$

$V_\pi = 0$, $\therefore g_m V_\pi = 0$

$$\boxed{R_o = R_c}$$

Unit - II
Amplifiers

BJT Small Signal Model - Analysis of CE, CB, CC amplifiers - Gain & frequency response - MOSFET Small Signal model - Analysis of CS & Source follower - Gain & frequency response - High frequency analysis.

BJT Small Signal Model :-

→ To simplify the analysis of the BJT, its operation is restricted to the linear V-I characteristics around the Q-point. ie active region.

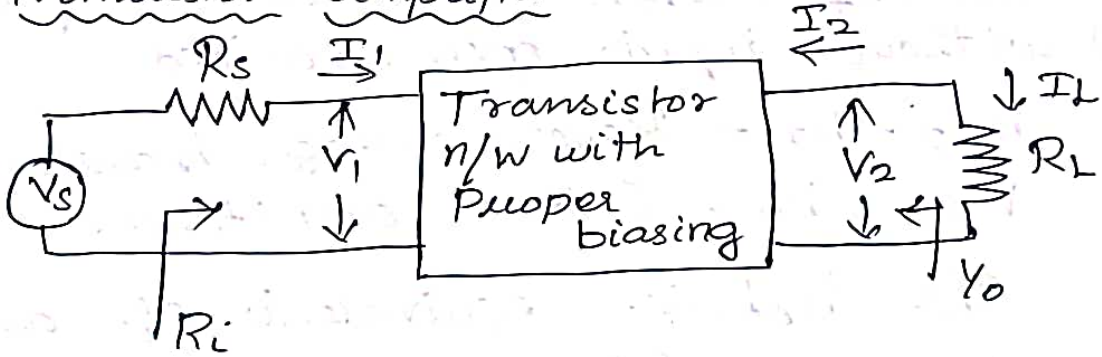
→ This approximation is possible only with small input signal.

→ The term small signal amplifier refers to the use of signal that takes up a relatively small percentage of an amplifier's operational range.

→ with small signal the transistor can be replaced with small signal linear model.

→ This model is also called small signal equivalent circuit.

Basic transistor amplifier:



Hybrid Parameters (or) h parameters:

The parameters are defines as follows,

$h_{11} = \left.\frac{V_i}{I_i}\right|_{V_o=0}$ = input resistance with output short-circuited, in ohms.

$h_{12} = \left.\frac{V_i}{V_o}\right|_{I_i=0}$ = Fraction of output voltage at input with input open circuited.

$h_{21} = \left.\frac{I_o}{I_i}\right|_{V_o=0}$ = Forward current transfer ratio or current gain with output short circuited.

$h_{22} = \left.\frac{I_o}{V_o}\right|_{I_i=0}$ = Output admittance with input open-circuited, in mhos.

→ They have different units. In other words, they are mixture of different units & hence referred to as hybrid parameters.

→ Standard notations can be given as.

$i = 11$ = Input $o = 22$ = output

$f = 21$ = forward transfer $r = 12$ = reverse transfer.

a) with output short circuited:

$h_{11} = h_i$ = Input Resistance

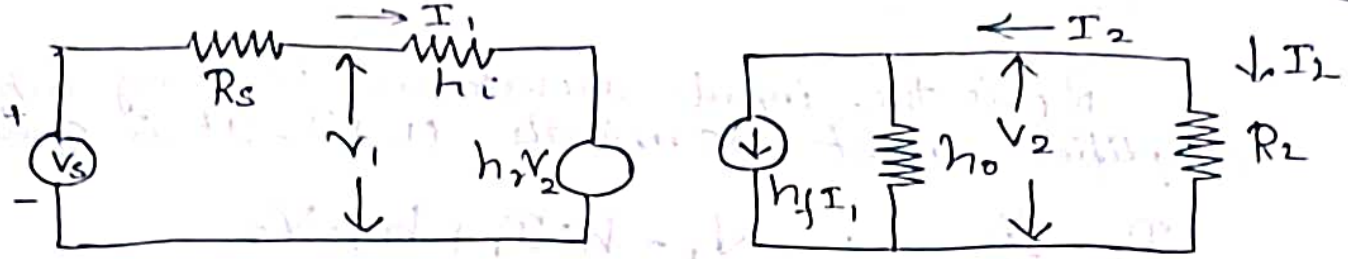
$h_{21} = h_f$ = Short circuit current gain

b) without input open circuited:

$h_{12} = h_r$ = Reverse voltage transfer ratio

$h_{22} = h_o$ = Output admittance.

The hybrid equivalent of amplifier circuit is as follows.



Hybrid parameters for the amplifier circuit is given below,

$$\left.\begin{aligned} V_1 &= h_{11} i_1 + h_{12} V_2 \\ I_2 &= h_{21} i_1 + h_{22} V_2 \end{aligned}\right\} \text{ for two port networks.}$$

1) Current Gain:

The current gain of the amplifier circuit is,

$$A_I = \frac{I_L}{I_1} = -\frac{I_2}{I_1}$$

$$I_2 = h_f I_1 + h_o V_2$$

But $V_2 = I_L R_L = -I_2 R_L$

$$I_2 = h_f I_1 + h_o (-I_2 R_L)$$

$$(1 + h_o R_L) I_2 = h_f I_1$$

$$A_I = \frac{-I_2}{I_1} = -\frac{h_f}{1 + h_o R_L}$$

Using Source resistance R_s,

$$A_{IS} = \frac{-I_2}{I_S} = \frac{-I_2}{I_1} \cdot \frac{I_1}{I_S} = A_I \frac{I_1}{I_S}$$

Current divider eqn,

$$I_1 = I_S \frac{R_S}{R_i + R_S}, \quad A_{IS} = A_I \cdot \frac{R_S}{R_i + R_S}$$

2) Input Resistance (R_i):

R_i is the input resistance looking into the amplifier input terminals (1, 1'). It is given by,

$$R_i = \frac{V_1}{I_1} \quad ; \quad V_1 = h_i I_1 + h_r V_2$$

$$R_i = \frac{h_i I_1 + h_r V_2}{I_1} = h_i + h_r \frac{V_2}{I_1}$$

But $V_2 = -I_2 R_L = A_I I_1 R_L$ — (1)

$$R_i = h_i + h_r A_I I_1 R_L / I_1 = h_i + h_r A_I R_L$$

$$R_i = h_i - \frac{h_r h_f R_L}{1 + h_o R_L} \quad \left[\text{Since } A_I = \frac{-h_f}{1+h_o R_L}\right]$$

$$= h_i - \frac{h_r h_f}{1/R_L + h_o} = h_i - \frac{h_r h_f}{Y_L + h_o}$$

3) Voltage Gain (A_v):

It is the ratio of o/p voltage V_2 to the input voltage V_1. It is given by,

$$A_v = \frac{V_2}{V_1} = \frac{A_I I_1 R_L}{V_1 = I_1 R_i} = \frac{A_I R_L}{R_i}$$

Including source resistance,

$$A_{VS} = \frac{V_2}{V_S} = \frac{V_2}{V_1} \times \frac{V_1}{V_S} = A_V \frac{V_1}{V_S}$$

But $V_1 = \frac{R_i}{R_S + R_i} V_S$

$$A_{VS} = A_V \cdot \frac{R_i}{R_i + R_S} = \frac{A_I R_L}{R_i + R_S} \quad \left[\text{Since } A_V = \frac{R_L A_I}{R_i}\right]$$

4) Output Admittance (Y_o):

It is the ratio of output current I_2 to the output voltage V_2. It is given by,

$$Y_o = \frac{I_2}{V_2} \quad \text{with } V_s = 0$$

$$I_2 = h_f I_1 + h_o V_2$$

$$Y_o = \frac{I_2}{V_2} = \frac{h_f I_1}{V_2} + h_o \quad \text{---(1)}$$

Apply KVL $\quad R_s I_1 + h_i I_1 + h_r V_2 = 0$

$$(R_s + h_i) I_1 = - h_r V_2$$

$$I_1/V_2 = \frac{-h_r}{R_s + h_i} \quad \text{---(1)}$$

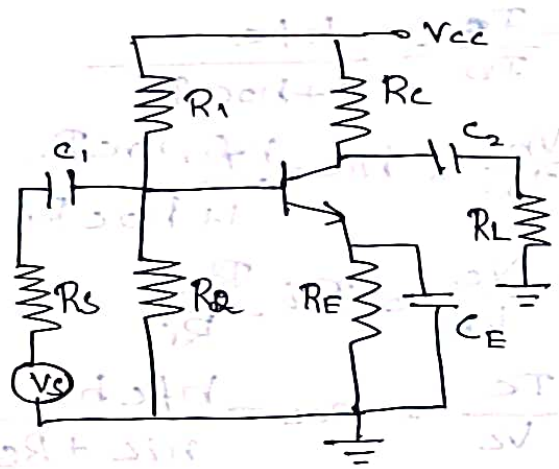
$$Y_o = h_o - \frac{h_f h_r}{h_i + R_s}$$

5) Power Gain (A_P):

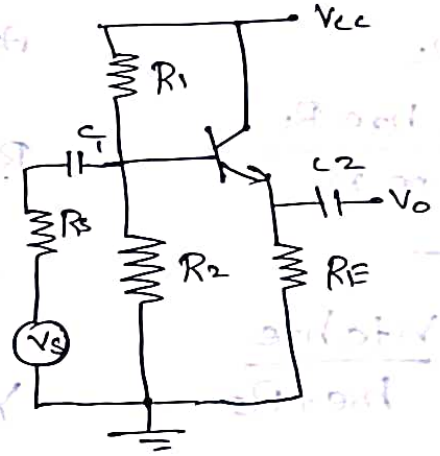
$$P_2 = V_2 I_L = -V_2 I_2 \; ; \quad P_1 = V_1 I_1$$

$$A_P = \frac{P_2}{P_1} = \frac{-V_2 I_2}{V_1 I_1} = A_V A_I = A_I^2 \frac{R_L}{R_i}$$

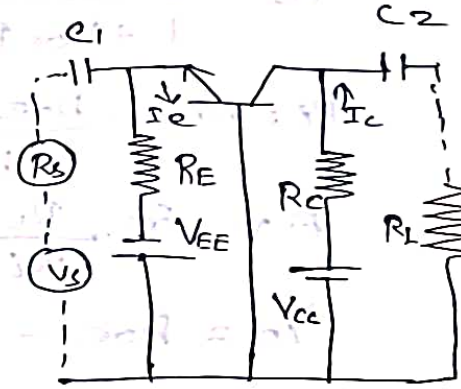
CE Amplifier



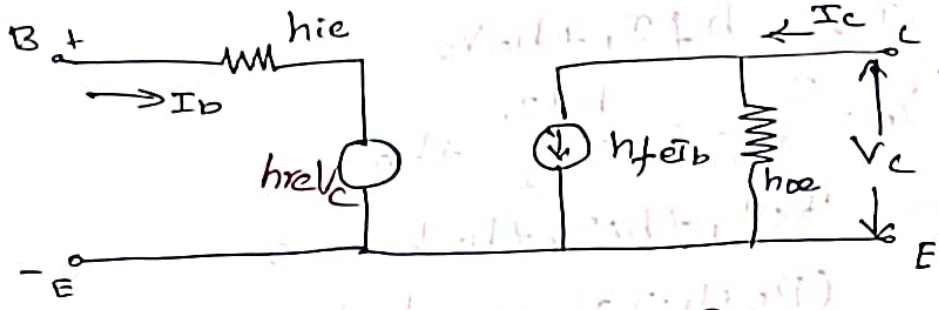
CC Amplifier



CB Amplifier



Hybrid Models of Transistors

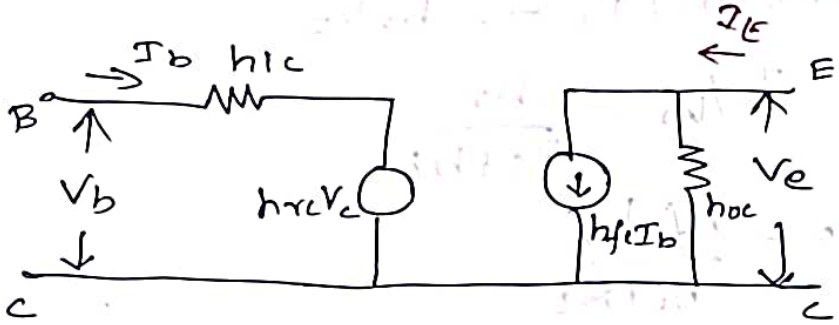


CE

$V_b = h_{ie} I_b + h_{re} V_c$

$I_c = h_{fe} I_b + h_{oe} V_c$

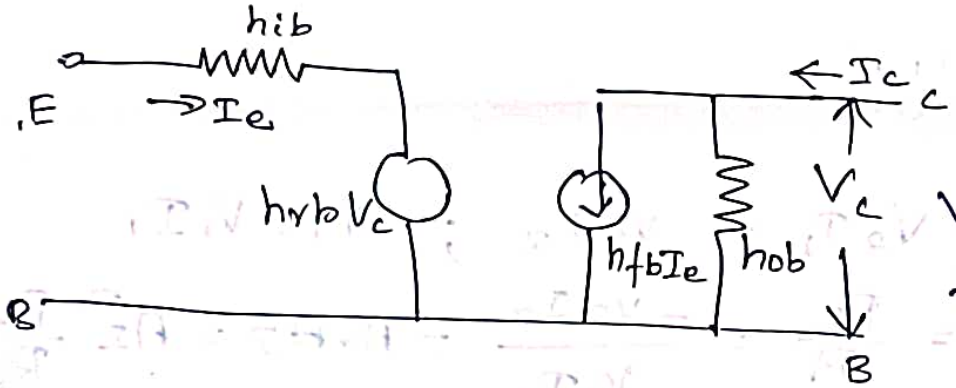
CB



CC

$V_b = h_{ic} I_b + h_{rc} V_c$

$I_e = h_{fc} I_b + h_{oc} V_c$



CB

$V_e = h_{ib} I_e + h_{rb} V_c$

$I_c = h_{fb} I_e + h_{ob} V_c$

CE Amplifier

$$A_I = \frac{-h_{fe}}{1 + h_{oe} R_L}$$

$$R_i = h_{ie} - \frac{h_{fe}\, h_{re} R_L}{1 + h_{oe} R_L}$$

$$A_V = A_I \frac{R_L}{R_i}$$

$$Y_o = h_{oe} - \frac{h_{fe} h_{re}}{h_{ie} + R_s}$$

CC Amplifier (emitter follower)

$$A_I = -\frac{I_e}{I_b} = \frac{-h_{fc}}{1 + h_{oc} R_L}$$

$$R_i = \frac{V_b}{I_b} = h_{ic} - \frac{h_{fc} h_{rc} R_L}{1 + h_{oc} R_L}$$

$$A_V = \frac{V_c}{V_b} = A_i \frac{R_L}{R_i}$$

$$Y_o = \frac{I_e}{V_c} = h_{oc} - \frac{h_{fc} h_{rc}}{h_{ic} + R_s}$$

CB Amplifier:

$$A_I = -\frac{h_{fb}}{1 + h_{ob} R_L} = \frac{-I_c}{I_e}$$

$$R_i = \frac{V_e}{I_e} = \frac{h_{ib}}{1 + h_{ob} R_L} \qquad R_i = h_{ib} + h_{rb} A_I R_L = h_{ib} - \frac{h_{fb} h_{rb} R_L}{1 + h_{ob} R_L}$$

$$A_V = \frac{h_{fb} R_L}{h_{ib}} = \frac{V_c}{V_e} = A_I \frac{R_L}{R_i}$$

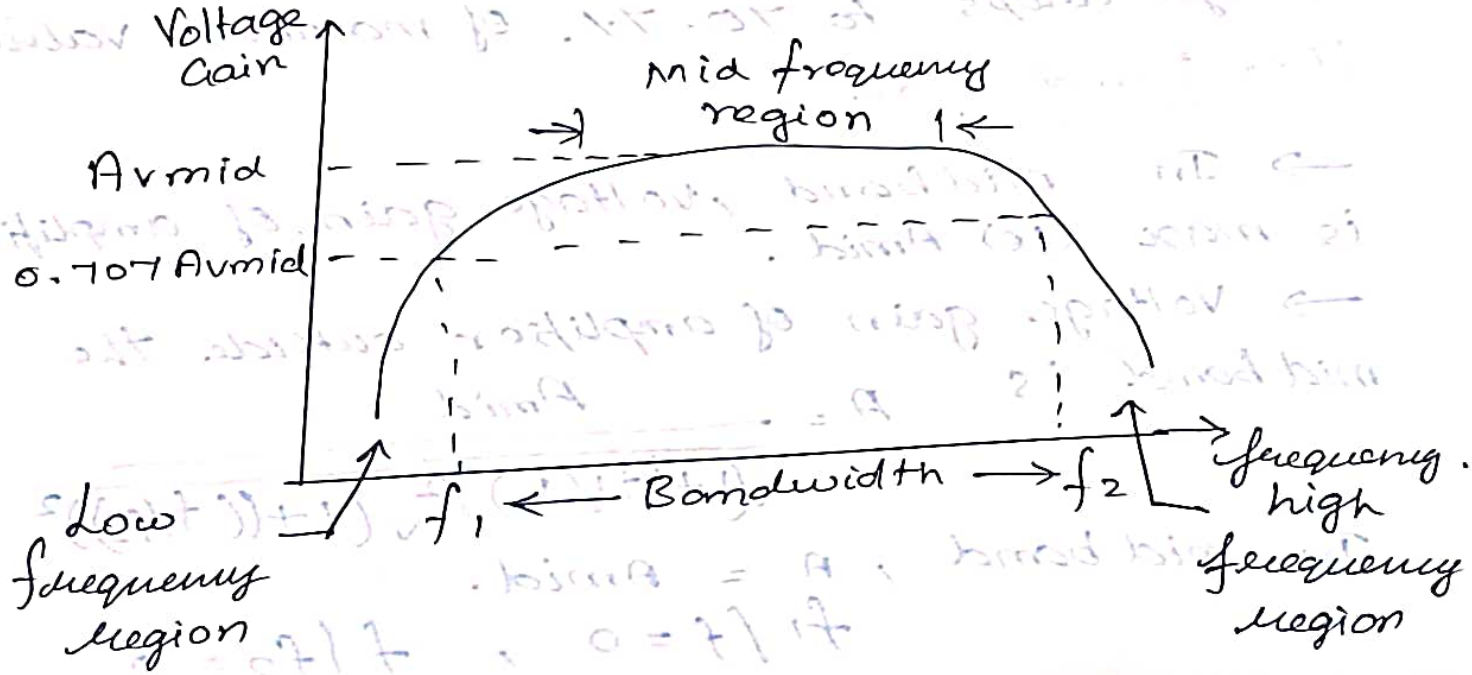
$$Y_o = \frac{I_c}{V_c} = h_{ob} - \frac{h_{fb} h_{rb}}{h_{ib} + R_s}$$

Frequency Response of an amplifier:

→ An amplifier ideally provides the same amplification for all frequencies. The degree to which this is done is usually indicated by a curve called frequency response of amplifier.

→ This curve is plot of voltage gain against frequency of input signal.

Ex: frequency response of RC coupled amplifier



→ To plot this curve, input voltage to amplifier is kept constant & frequency of input is varied & output voltage at each frequency is noted.
→ Gain of amplifier is calculated. Then voltage gain is plotted against frequency.
→ Frequency response is ideal over a mid-frequency. It deviates from ideal characteristics.
→ The decrease in voltage gain with frequency is called roll-off.
→ The two frequencies at which voltage gain starts decreasing below 70.7% are f_1 & f_2. (ie) Lower & upper cut off frequencies respectively.

$$\text{Bandwidth} = f_2 - f_1$$

→ These two frequencies are referred as half power frequencies. Since gain or output voltage drops to 70.7% of maximum value.

Mid Band Gain:

→ In midband, voltage gain of amplifier is max (ie) Amid.
→ Voltage gain of amplifier outside the midband is

$$A = \frac{Amid}{\sqrt{(1+((f_1/f)))^2}\sqrt{(1+((f/f_2)))^2}}$$

In midband, $A = Amid$.

$f_1/f = 0$, $f/f_2 = 0$

Below mid Band, $f/f_2 \approx 0$ $\quad A = \frac{A_{mid}}{\sqrt{1+(f_1/f)^2}}$

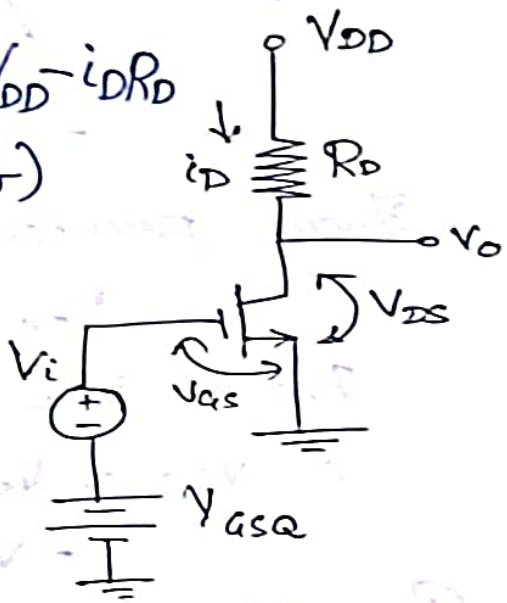
Above mid Band, $f_1/f = 0$ $\quad A = \frac{A_{mid}}{\sqrt{1+(f/f_2)^2}}$

MOSFET Small Signal Model:

→ Small Signal Model is an equivalent circuit which interrelates the incremental changes in i_D, V_{GS}, V_{DS} etc. Since the changes are, the Small-Signal equivalent circuit has linear element only (capacitors, resistors, controlled Sources)

→ Output Voltage $V_{DS} = V_o = V_{DD} - i_D R_D$

But $g_m = \frac{Id}{Vgs} = 2K(V_{GSQ} - V_T)$



Small Signal Parameters:

Instantaneous gate to source voltage is,

$$V_{GS} = V_{GSQ} + V_i = V_{GSQ} + V_{gs}$$

V_{GSQ} → dc component

V_{gs} → ac component

Instantaneous drain current,

$$i_D = K(V_{GS} - V_T)^2$$
$$= K(V_{GSQ} + V_{gs} - V_T)^2$$
$$= K[(V_{GSQ} - V_T) + V_{gs}]^2$$

$$= K(V_{GSQ} - V_T)^2 + 2K(V_{GSQ} V_T) V_{gs} + K V_{gs}^2$$

Dc component — Time varying I_D component — Produces harmonics

$V_{gs} << 2(V_{GSQ} - V_T) \longrightarrow$ To minimize harmonics.

Neglect V_{gs}^2, $i_D = I_{DQ} + i_d$

but $I_{DQ} = K(V_{GSQ} - V_T)^2$ &

$i_d = 2K(V_{GSQ} - V_T) V_{gs}$

$g_m = I_d / V_{gs} = 2K(V_{GSQ} - V_T)$

$= 2\sqrt{K I_{DQ}}$ $\quad \because I_{DQ} = K(V_{GSQ} - V_T)^2$

From previous fig output voltage,

$$V_{DS} = V_D = V_{DD} - i_D R_D$$

But $i_D = I_{DQ} + i_d$

$\therefore V_{DS} = V_o = V_{DD} - (I_{DQ} + i_d) R_D$

$= V_{DD} - I_{DQ} R_D - i_d R_D$

Time varying output $V_o = V_{ds} = -i_d R_D$

But $i_d = g_m V_{gs}$.

Instantaneous values of these equations are

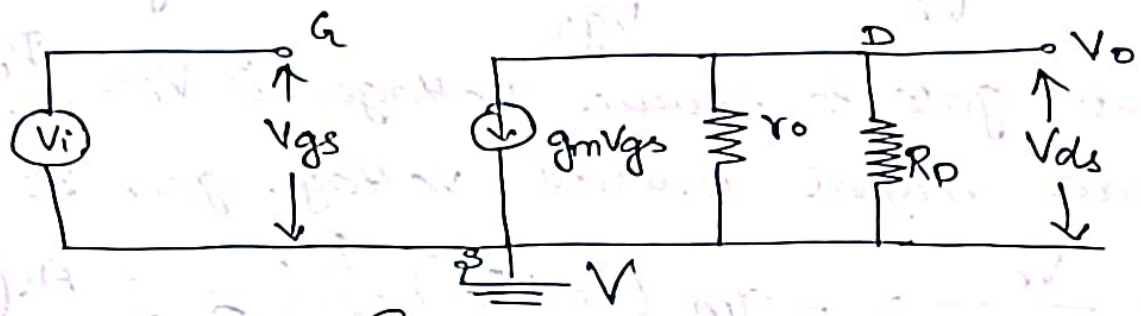
$V_{gs} = V_i$, $V_{ds} = -I_D R_D = -i_d R_D$

$$r_o = \left(\frac{\partial i_D}{\partial V_{DS}}\right)^{-1}$$

$i_D = K[(V_{GS} - V_T)^2 (1 + \lambda V_{DS})]$

$$r_o = \left[k\,(V_{GSQ} - V_T)\,\lambda\right]^{-1}$$

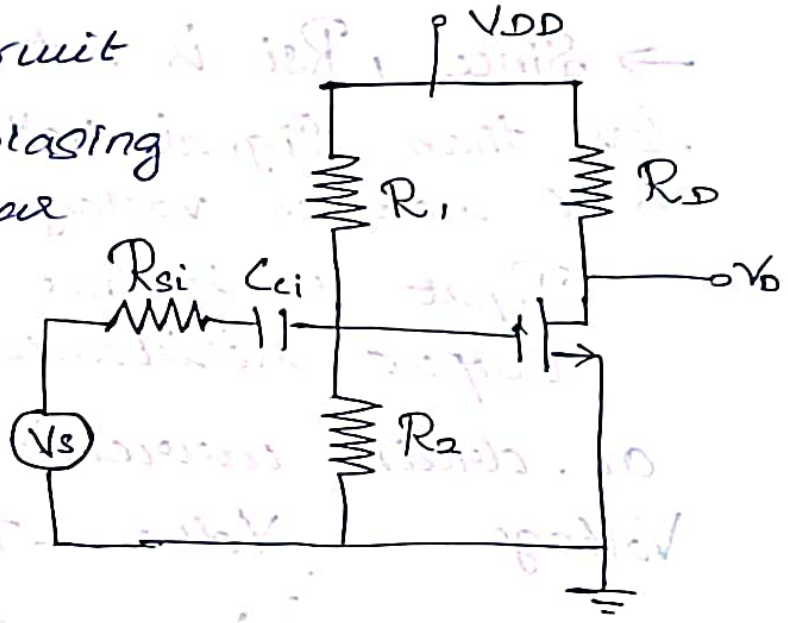
$$r_o = \left[\lambda\, I_{DQ}\right]^{-1}$$



. Small Signal Model of Nmos.

Analysis of CS (Common Source) Amplifier:

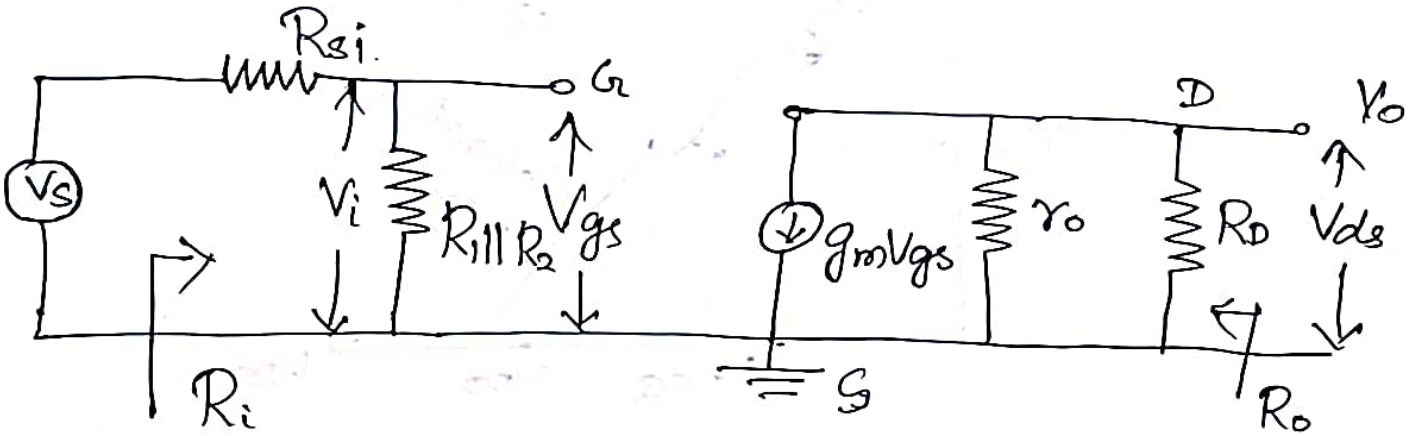
→ Common Source circuit with voltage divider biasing MOSFET is biased near the middle of Saturation region.



$R_{si} \ll R_i = R_1 \| R_2$

→ To minimize loading effects.

Small Signal Model for CS:



$$V_o = -g_m V_{gs} (r_o \| R_D)$$

$$V_i = V_{gs}$$

$$A_v = \frac{V_o}{V_i} = \frac{-g_m V_{gs} (r_o \| R_D)}{V_{gs}} = -g_m (r_o \| R_D$$

→ Input gate to source voltage is $V_i = \frac{R_i}{R_i + R_{si}} V_s$

→ Small signal overall voltage gain is

$$G_v = \frac{V_o}{V_s} = -g_m (r_o \| R_D) \frac{R_i}{R_i + R_{si}} = A_v \left(\frac{R_i}{R_i + R_{si}} \right)$$

$$= \frac{V_o}{V_i} \times \frac{V_i}{V_s}$$

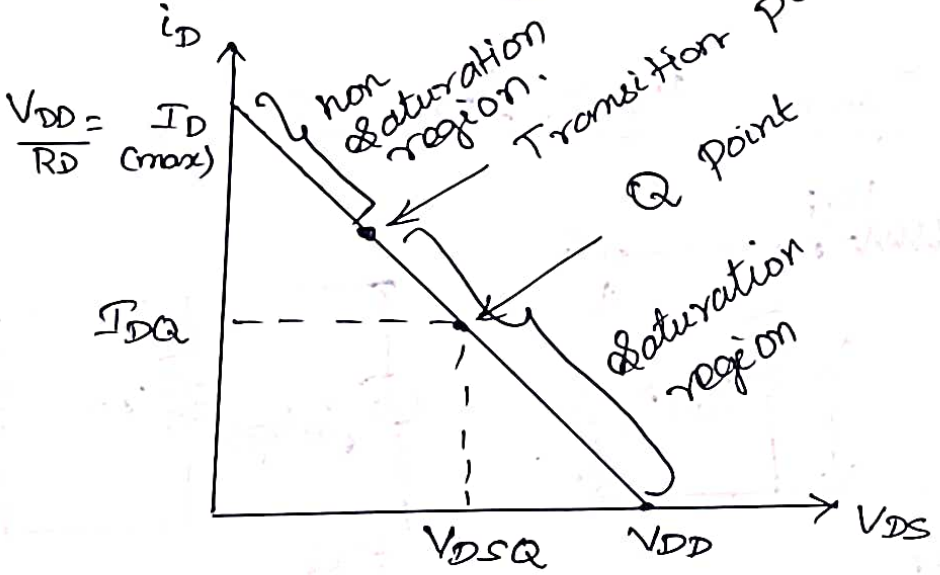
→ Since, R_{si} is not zero, amplifier signal V_i is less than signal voltage. This is loading effect.

→ It reduces voltage gain of amplifiers.

. Input resistance $R_{is} = R_1 \| R_2$

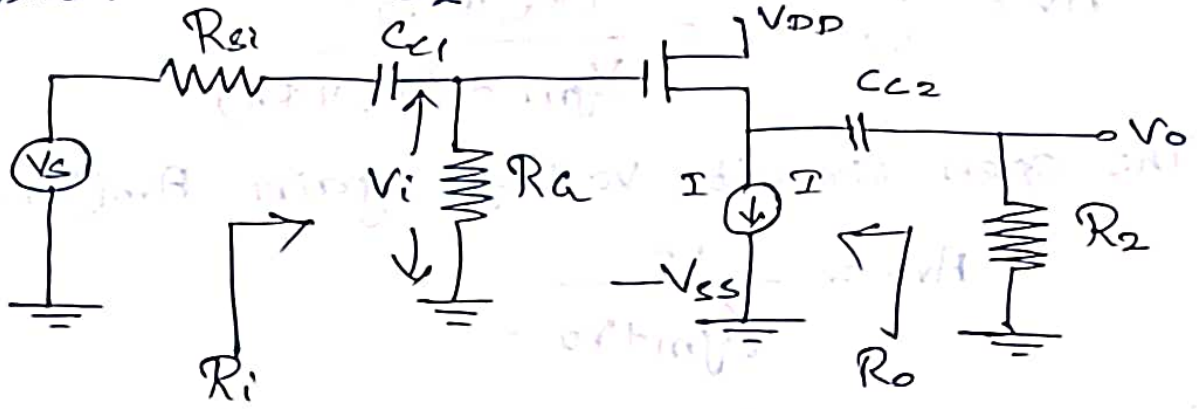
Output resistance $R_o = R_D \| r_o$

ac. drain current to ac drain to source voltage. $V_{ds} = -I_d R_D$



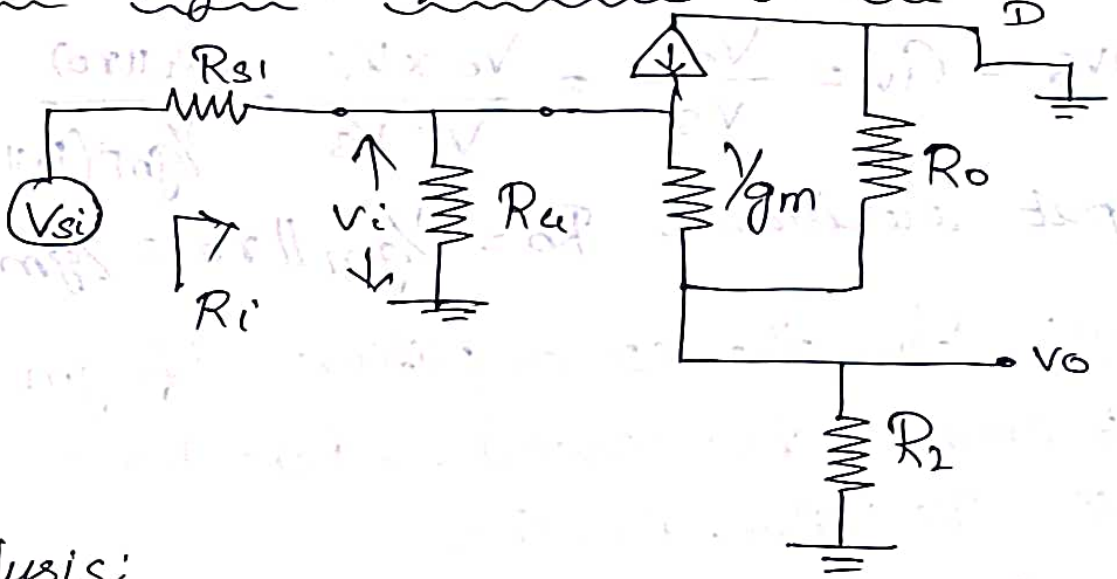
Analysis of CD / Source follower:

→ Input signal is coupled via C_{C1} to MOSFET gate & output signal at MOSFET source is coupled via C_{C2} to a load resistance R_L.



→ Since R_L is in effect connected in series with source terminal of MOSFET, it is more convenient to use MOSFET's T model for the analysis.

Small signal equivalent for CD amplifier.



Analysis:

$$R_i = R_G$$

$$V_i = V_S \left[\frac{R_i}{R_i + R_{si}}\right] = V_S \frac{R_G}{R_G + R_{si}}$$

→ R_L is parallel with r_o & resistance $1/g_m$ in

devices with

V_i appears across total resistance ($R_2 \| r_o$) of voltage divider rule $V_o = \dfrac{V_o \times (R_2 \| r_o)}{1/g_m + (R_2 \| r_o)}$

$$A_v = \frac{V_o}{V_i} = \frac{(R_L \| R_o)}{1/g_m + (R_L \| R_o)}$$

The open circuit voltage gain $A_{vo}(R_2 = \infty)$

$$A_{vo} = \frac{r_o}{1/g_m + r_o}$$

Since $r_o >> 1/g_m$, open circuit voltage gain tends to unity.

usually $R_2 << r_o$ $\therefore A_v = \dfrac{V_o}{V_i} = \dfrac{R_L}{1/g_m + R_L}$ $\left[R_2 << r_o\right]$

Overall volt gain of CD amplifier

$$A_{vs} = G_v = \frac{V_o}{V_s} = \frac{V_o \times V_i}{V_i V_s} = \frac{(R_L \| r_o)}{1/g_m + (R_L \| r_o)} \times \frac{R_G}{R_G + R_{si}}$$

$\therefore$ output resistance $R_o = 1/g_m \| r_o = 1/g_m$.

Problem: For the CD amplifier, if $g_m = 1.2$ mA/V, $R_G = 6.8\,M\Omega$, $r_o = 150\,k\Omega$, $R_{si} = 500\,k$, $R_2 = 10\,k$, calculate R_i, A_{vo}, A_v, G_v.

Given: $R_i = R_G = 6.8\,M\Omega$

$$A_{vo} = \frac{r_o}{1/g_m + r_o} = \frac{150 \times 10^3}{833 + 150 \times 10^3} = 0.9945$$

$$A_V = \frac{R_L \| r_o}{\frac{1}{g_m} + (R_L \| r_o)} = \frac{10K \| 150K}{833 + (10K \| 150K)}$$

$$= \frac{9.375 \times 10^3}{833 + 9.375 \times 10^3} = 0.918$$

9375

$$G_V = \frac{V_o}{V_s} = A_V \times \frac{R_G}{R_G + R_S} = 0.918 \times \frac{6.8}{6.8 \times 0.5}$$

$$G_V = 0.855$$

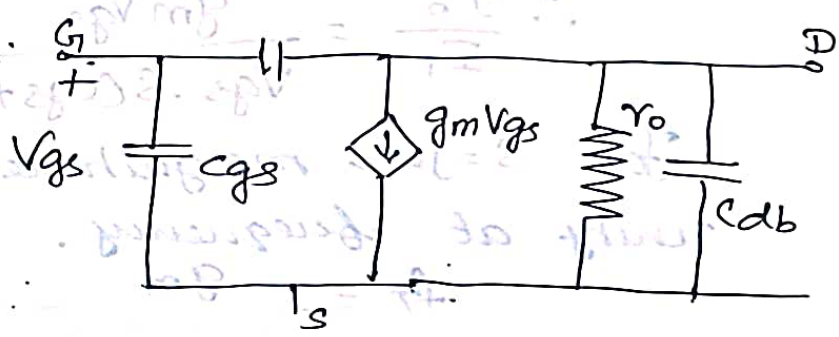
High frequency MOSFET Model:

→ High frequency response of MOSFET affects due to internal capacitances.

→ Types of internal capacitances in MOSFET

* Gate capacitance (C_{ox}): Parallel plate capacitor formed by gate electrode with the channel with the oxide layer acts as a capacitor dielectric.
* Junction capacitance: due to reverse biased PN Junctions formed by n^+ source region & P-type substrate & n^+ region & P-type substrate.

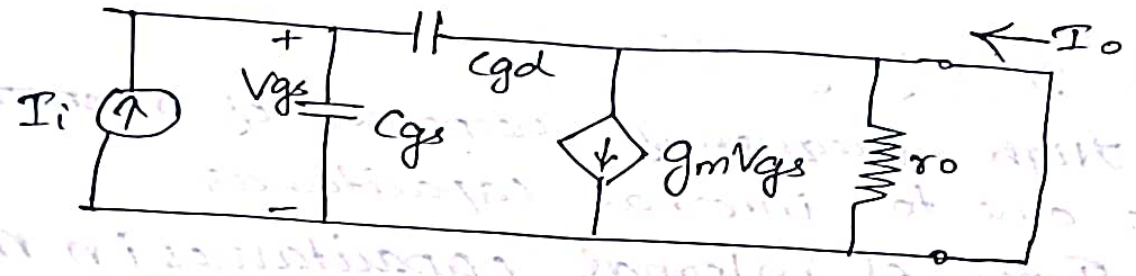
High frequency Model:



Unity Gain Frequency (f_T)

$f_T \rightarrow$ frequency at which the short-circuit current gain of CS MOSFET amplifier becomes unity.

For simplification, neglect C_{db}. So, modified high frequency model to find short circuit gain is given below.



Here input is fed with a circuit source signal I_i & output terminals are shorted.

Shorts circuit current $I_o = g_m V_{gs} - sC_{gd}V_{gs}$

Second term is very small, so neglect this.

$$I_o \approx g_m V_{gs}$$

The V_{gs} in terms of I_i is.

$$V_{gs} = I_i / s(C_{gs} + C_{gd})$$

$$\therefore \frac{I_o}{I_i} = \frac{g_m V_{gs}}{V_{gs} \cdot s(C_{gs}+C_{gd})} = \frac{g_m}{s(C_{gs}+C_{gd})}$$

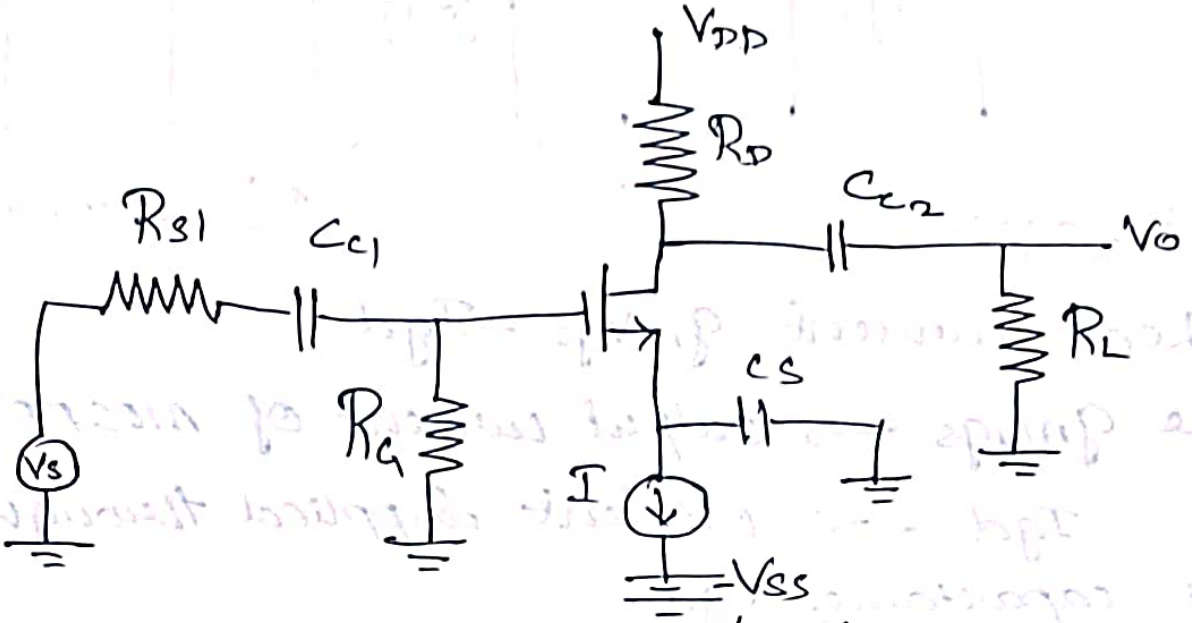
For $s = j\omega$, magnitude of current becomes unity at frequency. $\omega_T = \frac{g_m}{C_{gs}+C_{gd}}$

$$f_T = \frac{g_m}{2\pi C(C_{gs}+C_{gd})}.$$

Frequency response of CS amplifier:

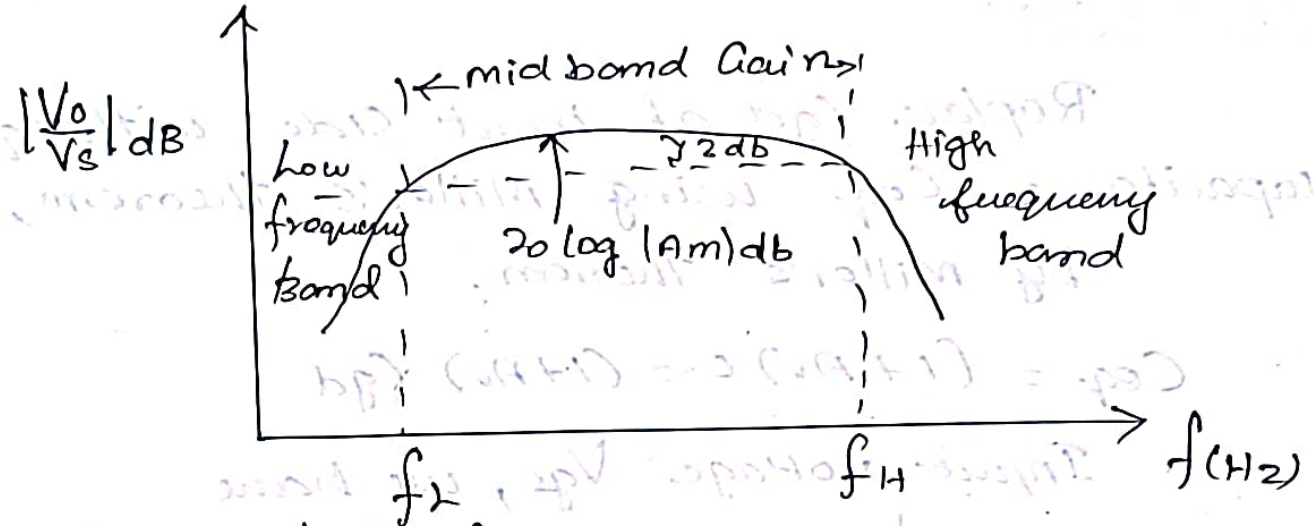
→ Gain falls at low frequency due to the effect of C_{c1}, C_s & C_{c2}.

→ Its gain falls at high frequency due to the effect of C_{gs} & C_{gd}.



CS MOSFET amplifier

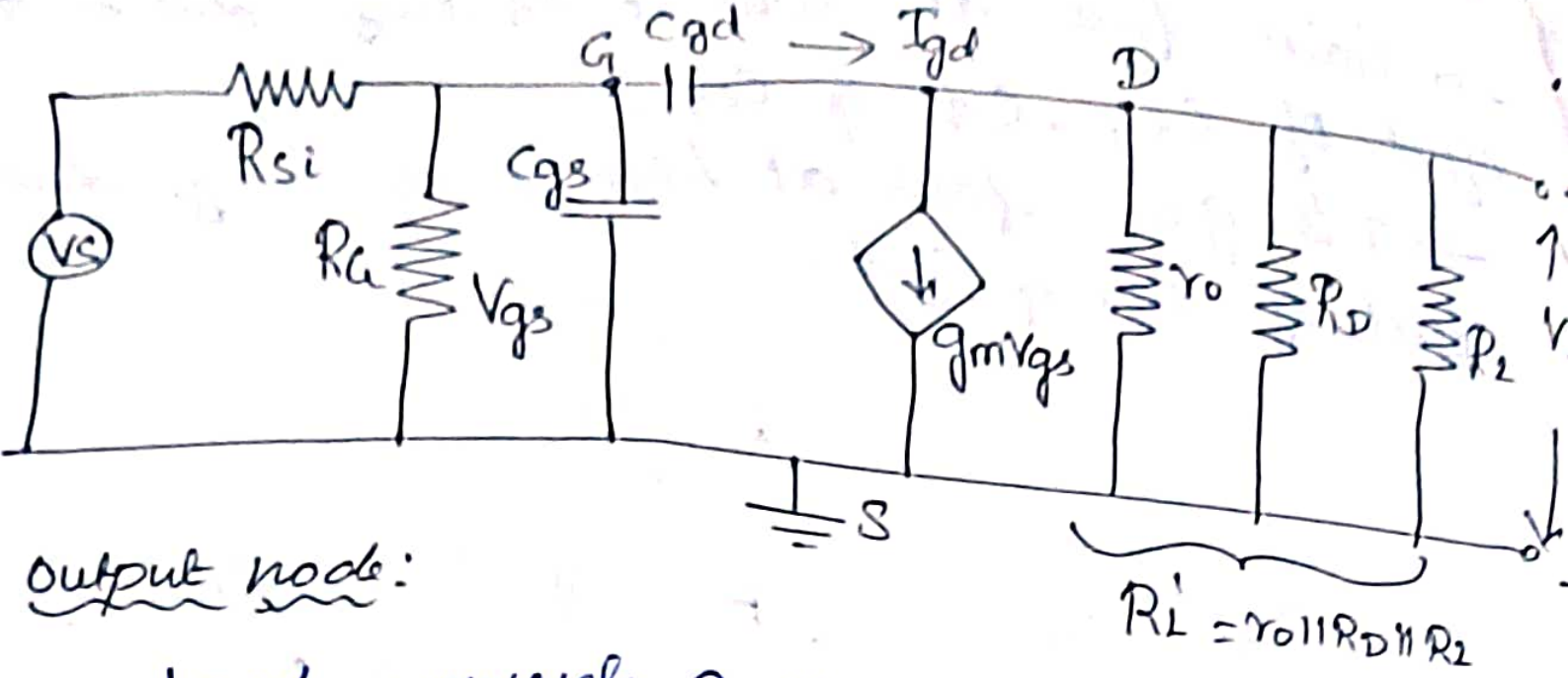
Frequency response of CS MOSFET amplifier.



f_L – Low frequency.

f_H – High frequency.

High frequency response:



Output node:

Load current $g_m V_{gs} - I_{gd}$

where $g_m V_{gs}$ → output current of MOSFET

I_{gd} → current supplied through small capacitance C_{gd}.

At frequency f_H, I_{gd} is small & neglecte

$V_o = -I_L R_L' = -g_m V_{gs} R_L'$ where $R_L' = r_o \| R$

Input node:

Replace C_{gd} at input side with equiva capacitance C_{eq} using Miller's theorem,

By Miller's theorem.

$$C_{eq} = (1 + A_V) C = (1 + A_V) C_{gd}$$

Input voltage V_{gs}, we have

$$A_V = \frac{V_o}{V_i} = \frac{-g_m V_{gs} R_L'}{V_{gs}} = -g_m R_L'$$

$$C_{eq} = (1 + g_m R_L') C_{gd}.$$

Total internal capacitance $C_{in} = C_{gs} + C_{eq}$

$= C_{gs} + (1 + g_m R_L') C_{gd}$

Total resistance $R_{si}' = R_{si} \parallel R_G$

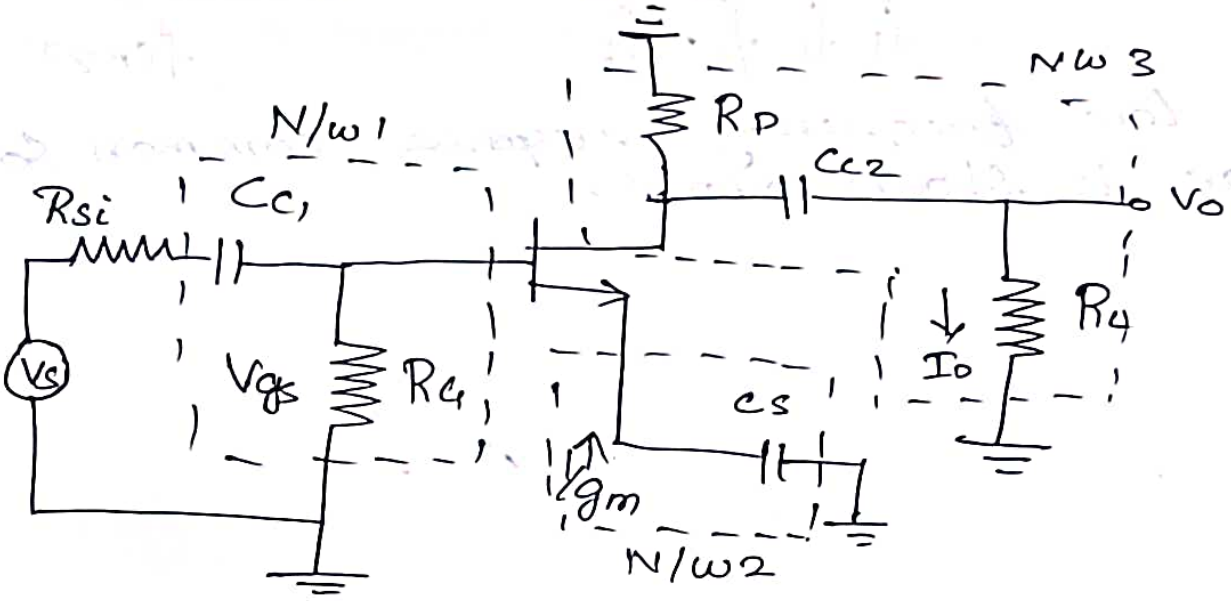
Time constant $\tau = R_c = R_{si}' C_{in}$

$$\omega_H = \omega_0 = \frac{1}{\tau} = \frac{1}{R_{si}' C_{in}}$$

$$f_H = \frac{1}{2\pi R_{si}' C_{in}}$$

Low frequency response:

→ affected by three RC networks.



corner frequencies due to three networks are

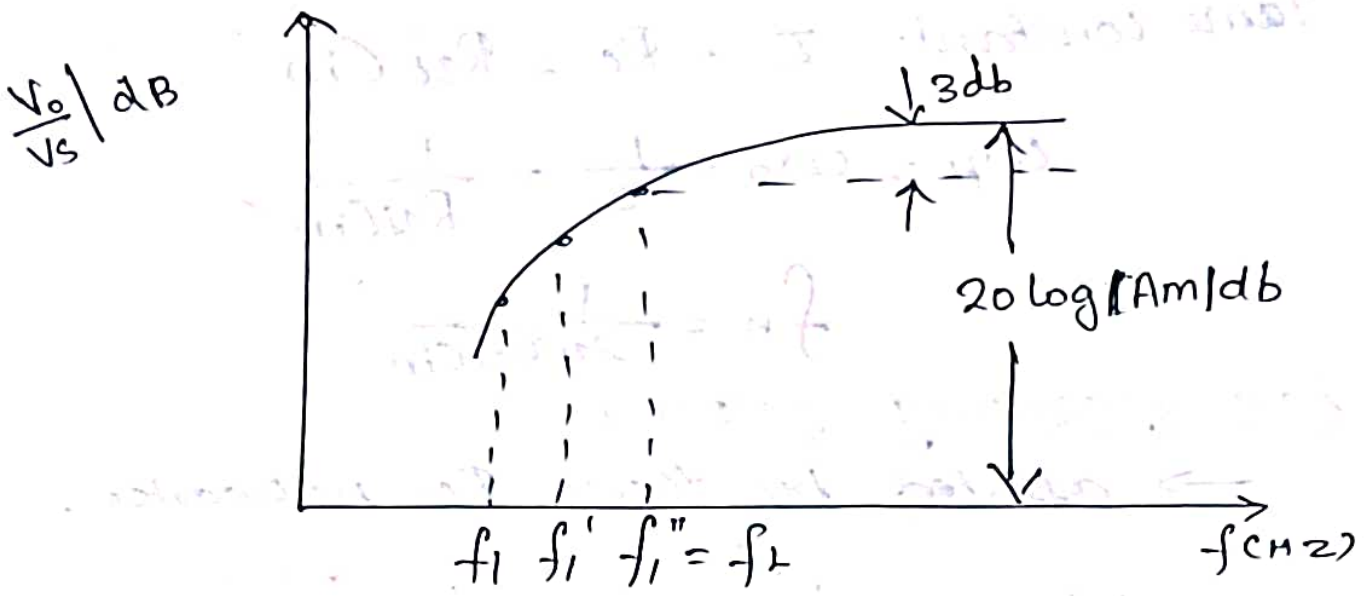
$$f_1 = \frac{1}{2\pi C_{C1}(R_{si} + R_G)}, \quad f_1' = \frac{g_m}{2\pi C_S}$$

$$f_1'' = \frac{1}{2\pi C_{C2}(R_D + R_L)}$$

overall low frequency transfer function of amplifier can be found from above three

frequencies.

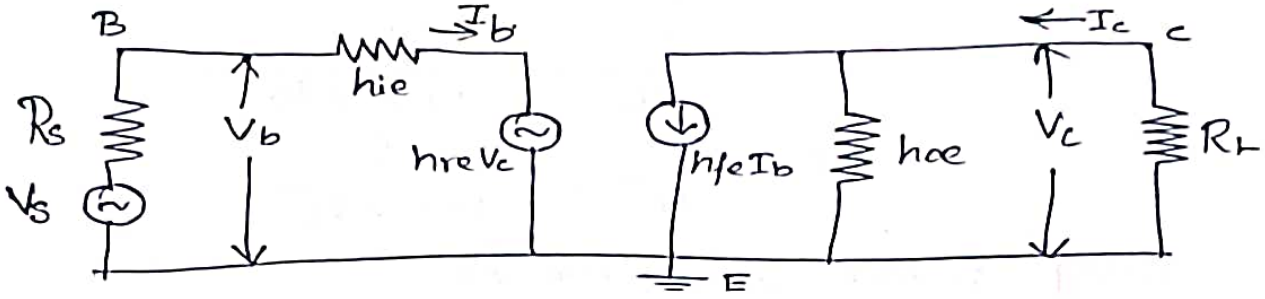
$$\frac{V_o}{V_s} = -\left(\frac{R_G}{R_G + R_{si}}\right)[g_m(R_D \| R_L)]\left(\frac{s}{s+\omega_1}\right)\left(\frac{s}{s+\omega_1'}\right)\left(\frac{s}{s+\omega_1''}\right)$$



The low frequency response common Source amplifier is shown above.

Analysis of Common Emitter Circuit:

Let us consider the h-parameter equivalent circuit for the amplifier.



Let us analyze hybrid model to find the current gain, input resistance, the voltage gain & the output resistance.

Current Gain:

For transistor amplifier A_i is defined as the ratio of output currents. It is given by,

$$A_I = \frac{-I_c}{I_b}$$

From the h-parameter equivalent circuit of the common emitter configuration, we can write,

$$\left.\begin{aligned} V_b &= h_{ie} I_b + h_{re} V_c \\ I_c &= h_{fe} I_b + h_{oe} V_c \end{aligned}\right\} \text{ for two port Network}$$

Substitute I_c in A_I

$$A_I = \frac{-(h_{fe} I_b + h_{oe} V_c)}{I_b}$$

$$A_I = -h_{fe} - h_{oe} \frac{V_c}{I_b}$$

where $V_c = -I_c R_L$

$$I_c = h_{fe} I_b + h_{oe} R_L I_c$$

$$I_c + h_{oe} R_L I_c = h_{fe} I_b$$

$$(1 + h_{oe} R_L) I_c = h_{fe} I_b$$

$$\frac{I_c}{I_b} = \frac{h_{fe}}{1 + h_{oe} R_L}$$

$$\boxed{A_I = \frac{-I_c}{I_b} = \frac{-h_{fe}}{1 + h_{oe} R_L}}$$

using source resistance R_s.

$$A_{IS} = \frac{-I_c}{I_s} = \frac{-I_c}{I_b} \cdot \frac{I_b}{I_s} = A_I \frac{I_b}{I_s}$$

using current divider equation,

$$I_b = \frac{I_s R_s}{R_i + R_s}$$

$$\frac{I_b}{I_s} = \frac{R_s}{R_i + R_s}$$

$$\boxed{A_{IS} = A_I \cdot \frac{R_s}{R_i + R_s}}$$

Input Resistance:

R_i is the input resistance. It is given by

$$R_i = \frac{V_b}{I_b}$$ we know that,

$$V_b = h_{ie} I_b + h_{re} V_c$$

$$Z_i = \frac{h_{ie} I_b + h_{re} V_c}{I_b} = h_{ie} + h_{re} \frac{V_c}{I_b}$$

Substituting $V_c = -I_c R_L = I_b R_L A_I$

$$I_i = h_{ie} + h_{re} \left(\frac{-I_c R_L}{I_b}\right) = h_{ie} + h_{re} A_I R_L$$

$$Z_i = h_{ie} + \frac{h_{re}(-h_{fe} R_L)}{1 + h_{oe} R_L}$$

$$Z_i = h_{ie} - \frac{h_{re} h_{fe} R_L}{1 + h_{oe} R_L}$$

$$Z_i = h_{ie} - \frac{h_{re} h_{fe}}{Y_L + h_{oe}}, \quad Y_L = \frac{1}{R_L}$$

$$\boxed{Z_i = h_{ie} - \frac{h_{re} h_{fe}}{Y_L + h_{oe}}}$$

From this equation we can note that input impedance is a function of the load impedance.

Voltage Gain:

It is a ratio of output voltage V_c to the input voltage V_b. It is given by,

$$\boxed{A_V = \frac{V_c}{V_b} = \frac{A_I I_b R_L}{V_b} = \frac{A_I R_L}{R_b}}$$

using Voltage divider rule

$$A_{VS} = \frac{V_L}{V_S} = \frac{V_c}{V_b} \cdot \frac{V_b}{V_S} = A_V \frac{V_b}{V_S}$$

By voltage divider rule) $V_b = \frac{R_i}{R_S + R_i} V_S$

$$\boxed{A_{VS} = A_V \frac{R_i}{R_S + R_i} = \frac{A_I R_L}{R_i + R_S}} \quad \left[A_V = \frac{R_L A_I}{R_i}\right]$$

Output admittance Y_o:

It is a ratio of output current I_c to the output voltage V_c. It is given by,

$$Y_o = \frac{I_c}{V_c} \quad \text{w.r.t.}$$

$$I_c = h_{fe} I_b + h_{oe} V_c$$

By substituting I_c in Y_o.

$$Y_o = \frac{h_{fe} I_b}{V_c} + h_{oe}$$

Apply KVL,

$$R_s I_b + h_{ie} I_b + h_{re} V_c = 0$$

$$R_s I_b + h_{ie} I_b = -h_{re} V_c$$

$$I_b (R_s + h_{ie}) = -h_{re} V_c$$

$$\frac{I_b}{V_c} = \frac{-h_{re}}{(R_s + h_{ie})}$$

Substitute the value of I_b/V_c in Y_o, we get.

$$Y_o = h_{fe}\left(\frac{-h_{re}}{R_s + h_{ie}}\right) + h_{oe}$$

$$\boxed{Y_o = h_{oe} - \frac{h_{re} h_{fe}}{R_s + h_{ie}}}$$

Power Gain:

It is the ratio of average power gain delivered to the load R_L to the input power. Output power is given by.

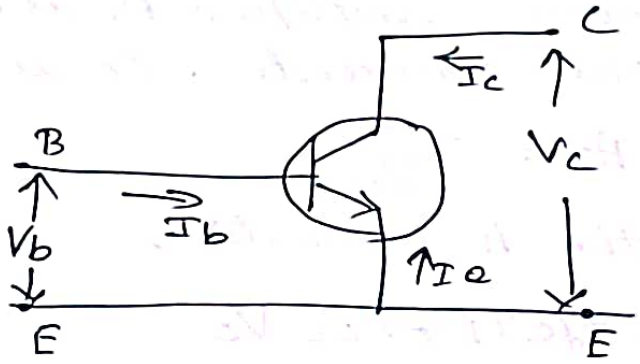
$$P_2 = -V_c I_c \quad , \quad P_1 = V_b I_b .$$

$$A_p = \frac{P_2}{P_1} = \frac{-V_c I_c}{V_b I_b} = A_v A_i = \left[\because A_v = \frac{A_i R_L}{Z_i}\right]$$

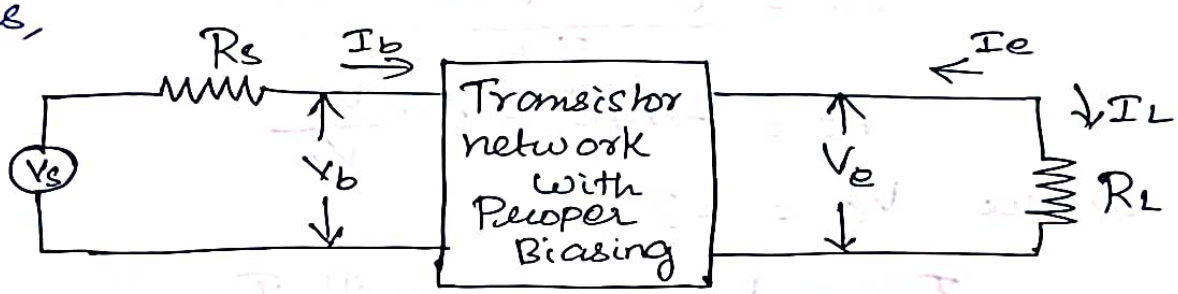
$$\boxed{A_p = A_i^2 \frac{R_L}{Z_i}}$$

Analysis of common Collector Circuit:

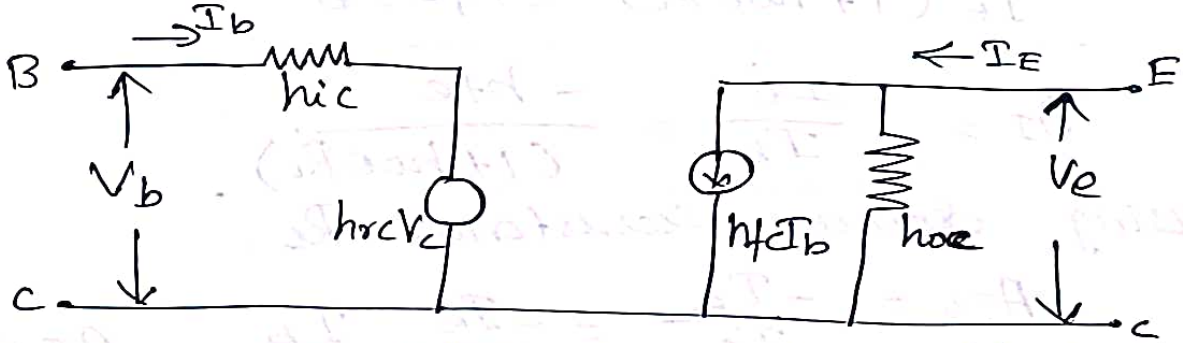
The circuit Diagrams of are valid for either n-p-n or p-n-p transistor and are independent of the type of load or method of biasing.



The basic amplifier circuit is as follows,



The transistor network is replaced with its small signal hybrid model as shown below.



Let us analyze hybrid model to find the current gain, input resistance, the voltage gain & the output admittance.

For two port Network,

$$V_b = h_{ic} I_b + h_{rc} V_c$$

$$I_a = h_{fc} I_b + h_{oc} V_c$$

Current Gain:

For transistor amplifier A_i is defined as the ratio of output currents. It is given by,

$$A_I = \frac{-I_e}{I_b}$$

From the h-parameter,

$$I_e = h_{fc} I_b + h_{oc} V_c$$

Substituting I_e in A_I

$$A_I = \frac{-(h_{fc} I_b + h_{oc} V_c)}{I_b}$$

$$A_I = -h_{fc} - h_{oc} \frac{V_c}{V_b}$$

where $V_c = -I_e R_L$

$$I_e = h_{fe} I_b - h_{oe} R_L I_e$$

$$I_e + h_{oe} R_L I_e = h_{fe} I_b$$

$$I_e (1 + h_{oe} R_L) = h_{fe} I_b$$

$$\boxed{A_I = \frac{-I_e}{I_b} = \frac{-h_{fe}}{(1 + h_{oe} R_L)}}$$

Using source resistance R_s.

$$A_{Is} = \frac{-I_e}{I_s} = \frac{-I_e}{I_b} \cdot \frac{I_b}{I_s} = A_I \cdot \frac{I_b}{I_s}$$

Using voltage devider equation,

$$I_b = \frac{I_s R_s}{R_i + R_s}$$

Cascode Amplifier

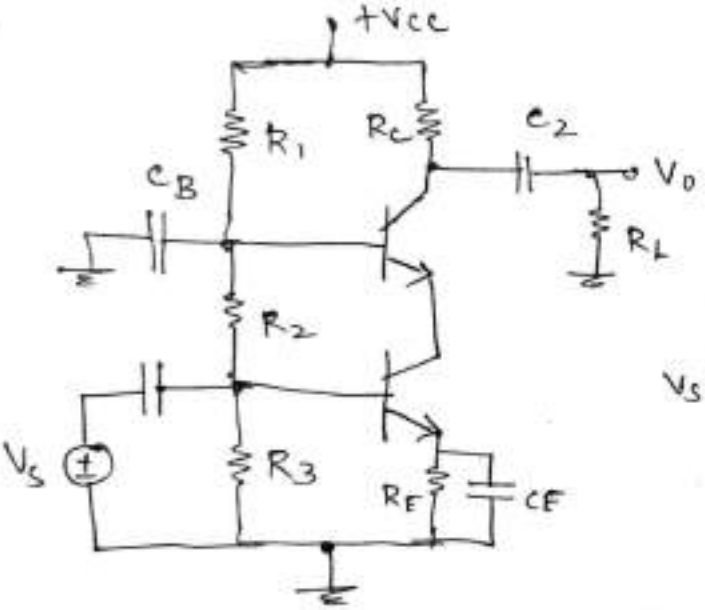
→ consists of common emitter in series with common base

Features:

1. It provides high input impedance, high voltage gain
2. provides improved input-output isolation as there is no direct coupling from o/p to i/p, eliminates miller effect & provides much B.W
3. provides very high o/p resistance, high slew rate, high stability

Disadv.

1) It requires two transistor & high supply voltage



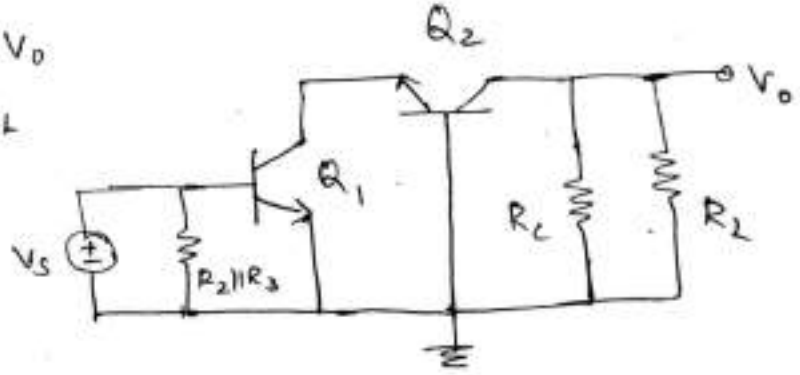
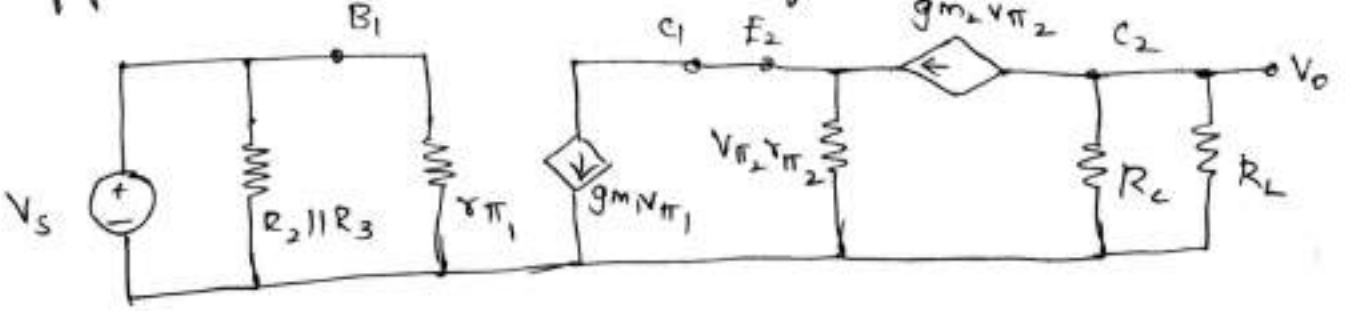


fig. shows a.c equivalent of CE - CB cascode amp.



Input impedance (R_i) = $R_2 \| R_3 \| r_{\pi_1}$

Output Impedance (R_o) = $R_c \| R_L$

Voltage gain (A_v):

$V_s = V_{\pi_1}$

Applying KCL at t_2

$$g_{m_1} V_{\pi_1} = \frac{V_{\pi_2}}{r_{\pi_2}} + g_{m_2} V_{\pi_2}$$

$$g_{m_1} V_s = \frac{V_{\pi_2}}{r_{\pi_2}} + g_{m_2} V_{\pi_2} \qquad \because V_s = V_{\pi_1}$$

$$g_{m_1} V_s = \frac{V_{\pi_2} + g_{m_2} V_{\pi_2} r_{\pi_2}}{r_{\pi_2}} = \frac{V_{\pi_2}(1+\beta_2)}{r_{\pi_2}} \qquad \because g_m r_\pi = \beta$$

$$V_{\pi_2} = \frac{r_{\pi_2}}{(1+\beta_2)} (g_{m_1} V_s) \quad \text{—(8)}$$

Output voltage V_o is given by

$$V_o = -(g_{m_2} V_{\pi_2})(R_c \| R_L) \quad \text{—(9)}.$$

sub (8) in (9)

$$V_o = -g_{m_2}(g_{m_1} V_s)\left(\frac{r_{\pi_2}}{1+\beta_2}\right)(R_c \| R_L)$$

$$\boxed{A_v = \frac{V_o}{V_s} = -g_{m_1} g_{m_2} \left(\frac{r_{\pi_2}}{1+\beta_2}\right)(R_c \| R_L)}$$

$$A_v = -g_{m_1} \left(\frac{g_{m_2} r_{\pi_2}}{1+\beta_2}\right)(R_c \| R_L)$$

$$\boxed{A_v = -g_m (R_c \| R_L)}$$

$\because g_{m_2} r_{\pi_2} = \beta_2$

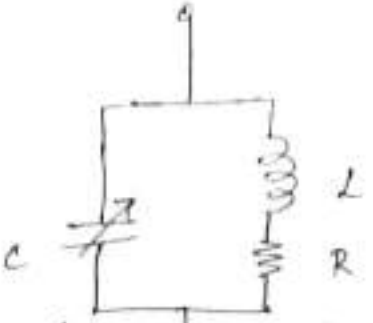
$$\frac{\beta_2}{1+\beta_2} \cong 1$$

Unit - IV

Tuned Amplifiers

To amplify selective range of freq Rc replaced by tuned ckt

→ A tuned parallel LC circuit resonates at a particular freq f_0.



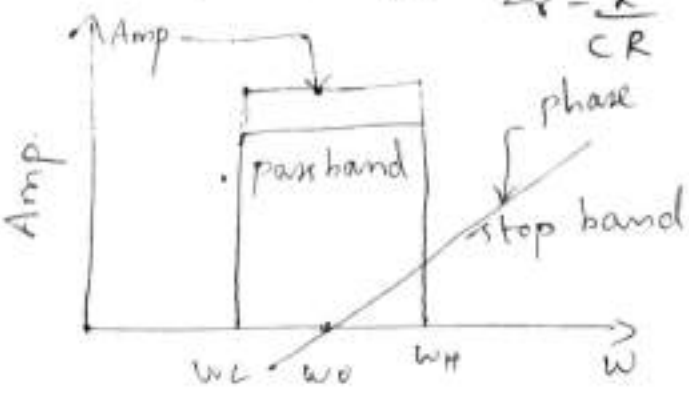
→ The response of tuned amplifiers is max. at resonant freq & falls below below value to

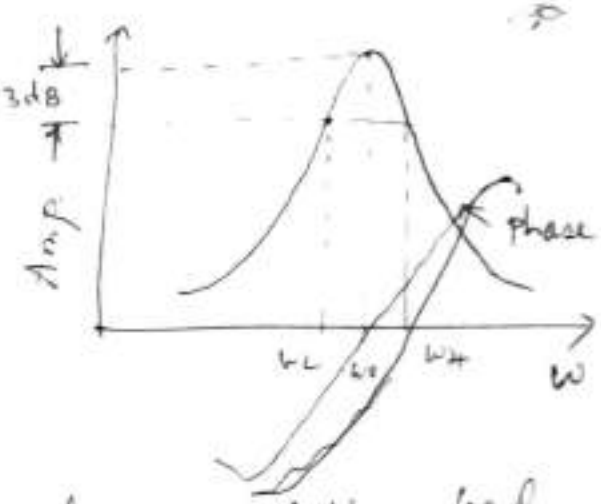
→ Tuned amplifier amplifies signals within narrow freq band centered about f_0.

→ The resonant freq. of the circuit is given as

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \quad \text{or} \quad \omega_0 = \frac{1}{\sqrt{LC}}.$$

$$Z_r = \frac{L}{CR}$$





→ At resonance, tuned ckt is used as resistive load as $X_L = X_C$, and reactance part becomes zero.

→ The ckt acts as pure resistive with V & I in phase

→ for freq. > fo, the ckt. is capacitive, I leads V

→ for freq. < fo, the ckt is inductive, I lags V

→ gain is $\propto Z_L$

→ Tuned amp. are used for amplification of narrow band of freq.

coil losses:

→ Tuned ckt consists of coil

→ coil is not purely inductive → leakage resistance in series with inductor

→ total loss = Cu loss + eddy current loss + hysteresis loss

→ Cu loss ↓ d.c resistance of coil → ½ f

eddy current loss ↓ I flowing in copper or core cased by induction heating within inductors, Cu or core → $\propto f$

hysteresis loss ↓ area of hysteresis loop → rate at which it is traversed indep. of freq. — energy lost as heat in traversing a magnetic material

Q factor:

The lower the value of resistance in inductor better is the Q.

Q factor → ratio of impedance of coil to its resistance

$$Q = 2\pi \times \frac{\text{max. energy stored per cycle}}{\text{Energy dissipated per cycle}}$$

Energy stored in cap. $= \frac{1}{2} C V_{max}^2$ → max value of volt across capacitor

$R << \frac{1}{\omega C}$

$V_{max} = \frac{I_m}{\omega C}$ → max value of I throu' C & R

$= \frac{1}{2} C V_{max}^2 = \frac{1}{2} \frac{I_m^2}{\omega^2 C}$

Energy dissipated per cycle $= \frac{I_m^2 R}{2f}$

$$Q = 2\pi \times \frac{\frac{I_m^2}{2\omega^2 C_s}}{\frac{I_m^2 R_s}{2f}} = 2\pi \times \frac{I_m^2}{2\omega^2 C_s} \times \frac{2f}{I_m^2 R_s}$$

$$= \frac{2\pi \times I_m^2 \times 2f}{2 \times (2\pi f)^2 C_s \times I_m^2 R} = \frac{1}{2\pi f \times C \times R} = \frac{1}{\omega C R}.$$

Unloaded & Loaded Q

→ When tank ckt is not connected to external circuit or load, Q accounts for internal losses and known as unloaded quality factor Q_u.

$$Q_u = 2\pi \times \frac{\text{max. energy stored per cycle}}{\text{Energy dissipated per cycle in tank ckt}}$$

→ when tank circuit is connected to load, the quality factor Q_L is

$$Q_L = 2\pi \times \frac{\text{max. energy stored per cycle}}{\text{Energy dissipated per cycle in tank ckt} + \text{Energy dissipated per cycle due to presence of external load}}$$

I_m is the peak value of current

→ max energy stored per cycle

$$= \frac{1}{2} L I_m^2$$

→ average power dissipated in inductor per cycle

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

Energy = power x time

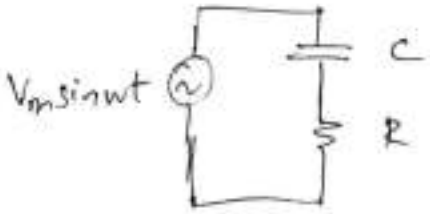
Energy dissip. per cycle = power x periodic time for one cycle

$$= \left(\frac{I_m}{\sqrt{2}}\right)^2 R \times T$$

$$= \frac{I_m^2}{2} R T = \frac{I_m^2}{2} \frac{R}{f}$$

$$Q = 2\pi \times \frac{\frac{1}{2} L I_m^2}{\frac{I_m^2 R}{2f}} = \frac{\omega L}{R} = \frac{X_L}{R}$$

Q factor for capacitor

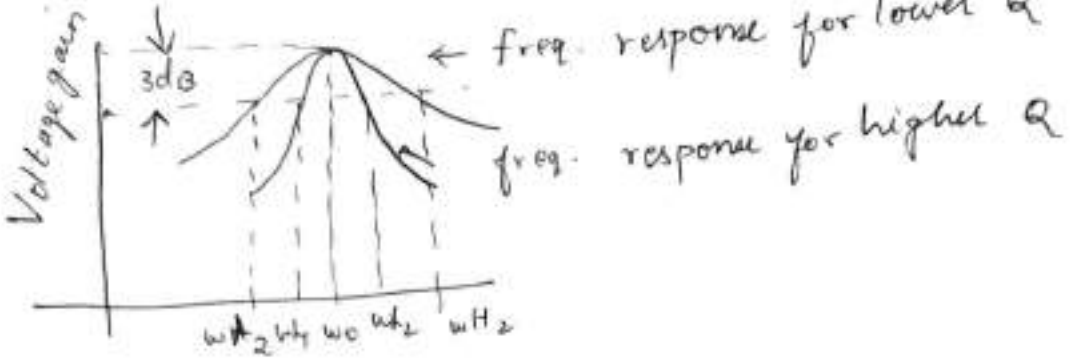


Gain and Frequency response

- determines 3dB Bandwidth for resonant ckt.

$$B.W = \frac{f_r}{Q} \rightarrow \text{represents centre freq.}$$

for higher Q values B.W is reduced.



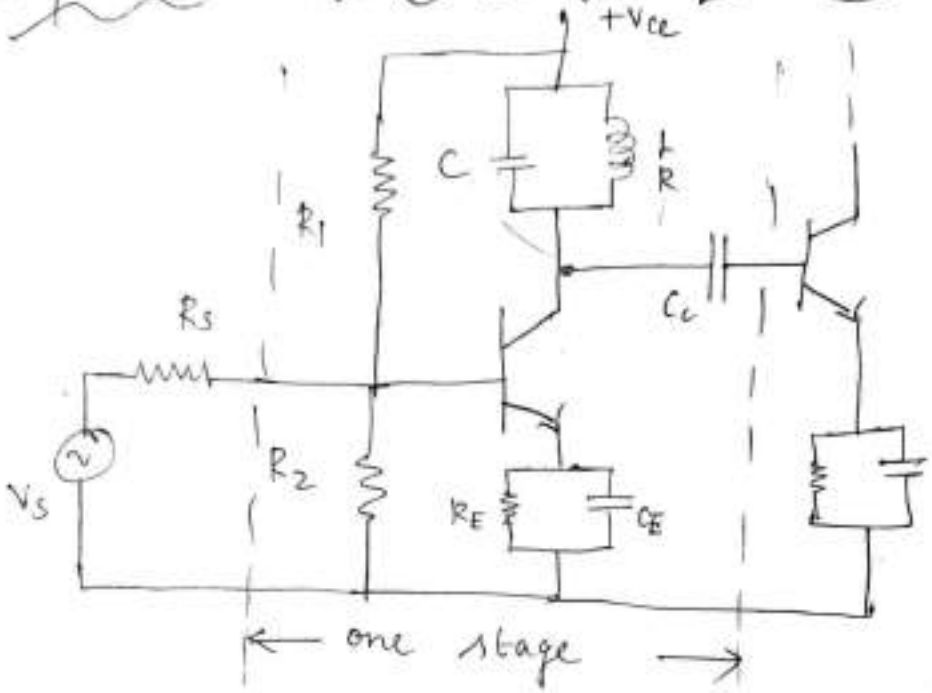
Small signal Tuned Amplifier

→ To obtain large voltage gain, it is required to use number of tuned amplifier stages in cascade

→ Cascade Tuned amplifier

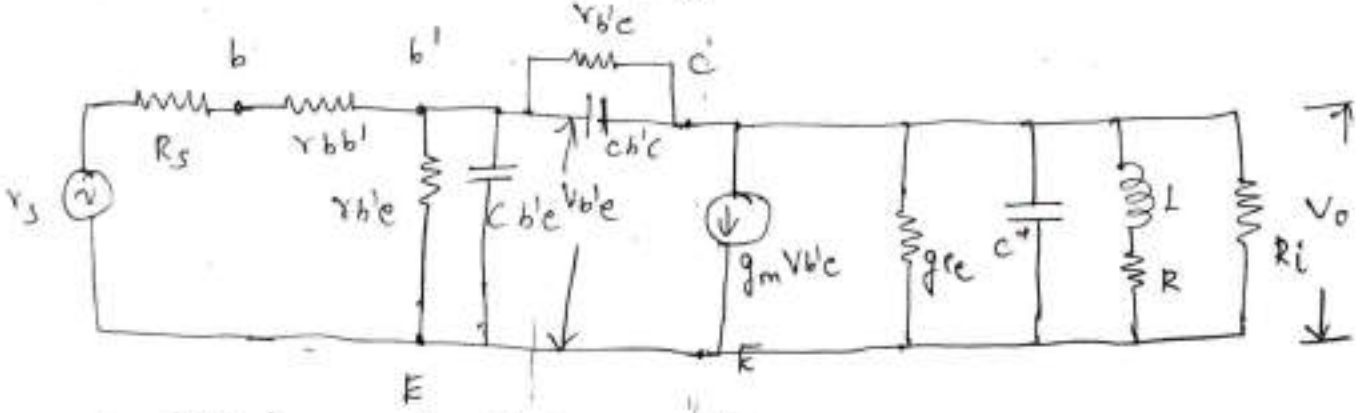
- Single tuned (one parallel resonant as load impedance & all tuned circuits are tuned to same freq)
  - capacitance coupled
  - transformer coupled or inductively coupled
- double tuned (two inductively coupled tuned ckts per stage both tuned to same freq)
- Stagger tuned (no. of single tuned stages in cascade successive tuned ckts tuned to slightly different freq).

Analysis of Capacitance coupled Single Tuned Amplifier

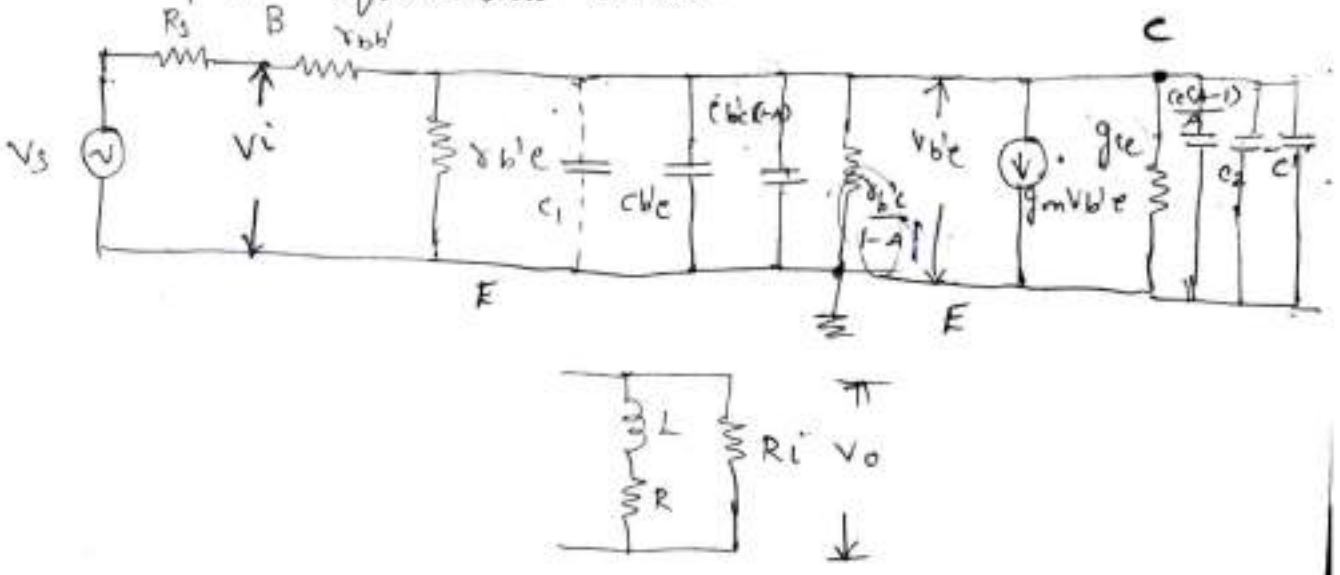


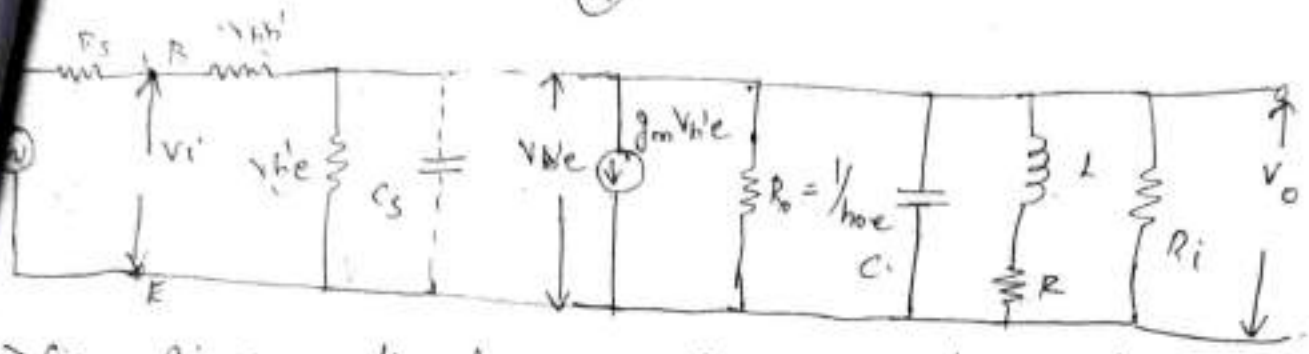
→ o/p across tuned ckt is coupled to next stage thro' C_c.

The equivalent ckt using hybrid π model is,



modified equivalent circuit





→ fig. gives modified equivalent ckt by applying Miller Theorem

→ A voltage gain, C_1, C_2 – stray wiring capacitances

→ all capacitances in i/p can be grouped to form C_s

$$C_s = C_{b'c} + C_1 + C_{b'c}(1-A)$$

→ all capacitances at o/p is grouped to

$$C = C_{b'c}\left(\frac{A-1}{A}\right) + C_2 + C'$$

$$g_{ce} = \frac{1}{r_{ce}} = h_{oe} - g_m h_{re} \Rightarrow h_{oe} = \frac{1}{R_o} \rightarrow \text{o/p resistance}$$

→ The reactance of by pass & coupling capacitor are negligibly small at operating freq and neglected.

Admittance of inductor along with resistor R is,

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

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

$$= \frac{1}{R_p} + \frac{1}{j\omega L_p}$$

where $R_p = \frac{R^2+\omega^2 L^2}{R}$ and $L_p = \frac{R^2+\omega^2 L^2}{\omega^2 L}$

The inductor is represented by R_p & L_p

Q of the coil $Q_0 = \frac{\omega_0 L}{R}$

where $\omega_0 = \frac{1}{\sqrt{LC}}$ is freq of resonance of ckt

Q_0 of coil is large $\omega L >> R$ in freq range of operation.

As $\frac{R}{\omega^2 L^2} << 1$,, $R_p = \frac{R^2 + \omega^2 L^2}{R} = \frac{R\left(R + \frac{\omega^2 L^2}{R}\right)}{R}$ $R_p = \frac{\omega L}{R}$

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

÷ Nr & Dr terms by $\omega^2 L$

$$L_p = \frac{\frac{R^2}{\omega^2 L} + L}{1}$$

$\therefore \omega L >> R$

$$L_p \cong L$$

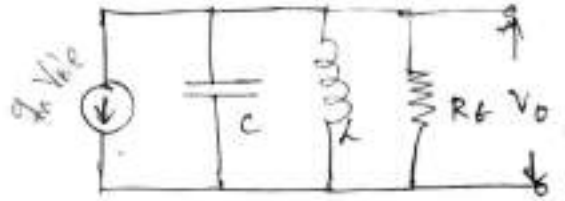
The o/p ckt of amplifier

R_t in parallel combination of R_o, R_p & R_i ie.

$$\frac{1}{R_t} = \frac{1}{R_o} + \frac{1}{R_p} + \frac{1}{R_i}$$

The effective Q factor

$$Q_e = \frac{\text{susceptance of inductance } L \text{ or } C}{\text{conductance of shunt resistance } R_t}$$



$$Q_e = \omega_0 C R_t = \frac{R_t}{\omega_0 L}$$

from o/p

$$V_o = -g_m V_{b'e} Z$$

Z is impedance of C, L and R_t in parallel.

$Y = 1/Z$ is

$$Y = \frac{1}{Z} = \frac{1}{R_t} + \frac{1}{j\omega L} + j\omega C$$

$$= \frac{1}{R_t}\left[1 + \frac{R_t}{j\omega L} + j\omega C R_t\right]$$

y/y Nr. & Dr by ω_0

$$Y = \frac{1}{R_t}\left[1 + \frac{R_t \omega_0}{j\omega L \omega_0} + \frac{j\omega \omega_0 C R_t}{\omega_0}\right]$$

$$\frac{R_t}{\omega_0 L} = \omega_0 C R_t = Q_e$$

$$Y = \frac{1 + jQ_e\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]}{R_t}$$

$$Z = 1/Y = \frac{R_t}{1 + jQ_e\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]}.$$

'δ' indicate fractional freq. Variation

$$\delta = \frac{\omega - \omega_0}{\omega_0} = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega_0} = \frac{\omega}{\omega_0} - 1$$

$$\frac{\omega}{\omega_0} = 1 + \delta$$

$$Z = \frac{R_t}{1 + jQ_e\left[(1+\delta) - \frac{1}{(1+\delta)}\right]}$$

SIMD Instruction set Extension for multimedia

→ graphic system used 8 bits to represent three primary colours

→ By partitioning

$$= \frac{R_t}{1+jQ_e\left[\frac{1+\delta^2+2\delta-1}{(1+\delta)}\right]} = \frac{R_t}{1+j2\delta Q_e\left[\frac{\delta/2+1}{1+\delta}\right]}$$

At any freq ω is close to ω_0, $\delta << 1$ ($\because \omega - \omega_0 << 1$

$$Z = \frac{R_t}{1+j2Q_e\delta}$$

At resonance $\omega = \omega_0$ & $\delta = 0$. The impedance Z becomes

$$Z = R_t = R_0 \| R_p \| R$$

$$R_p = \frac{\omega_0^2 L^2}{R} = \frac{\omega_0 L}{\omega_0 CR} \quad \left(\because \omega_0 L = \frac{1}{\omega_0 C}\right)$$

$$= \frac{L}{CR}$$

At resonance, R_p can be expressed as

$$R_p = \frac{\omega_0^2 L^2}{R}$$

$$Q_0 = \frac{\omega_0 L}{R}$$

$$Q_0^2 = \frac{\omega_0^2 L^2}{R^2}$$

$$R_p = Q_0^2 R = \omega_0 L Q_0$$

where Q_0 is Q of the coil alone at resonance.

from fig neglecting C_s

$$V_{b'e} = V_i \frac{r_{b'e}}{r_{bb'} + r_{b'e}}$$

$$V_o = -g_m V_{b'e} Z$$

$$= -g_m \left[V_i \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right] Z$$

Voltage gain without considering source resistance is

$$A = \frac{V_o}{V_i} = -g_m \left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right) Z \quad R_t$$

$$\frac{}{1 + 2j Q_{eff} \delta}$$

Voltage gain at resonance ($\delta = 0$)

$$A_{res} = -g_m \left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right) R_t$$

$R_t = R_0 \| R_p \| R_{in}$

$$\frac{A}{A_{res}} = \frac{1}{1 + j 2 \delta Q_{eff}}$$

$$\left| \frac{A}{A_{res}} \right| = \frac{1}{\sqrt{1 + (2 \delta Q_e)^2}}$$

Phase angle $\frac{A}{A_{res}}$ is given by

$$\phi = \tan^{-1} (2 \delta Q_e)$$

At freq ω_t below resonant freq

$$\delta = \frac{1}{2 Q_e}$$

$$\left| \frac{A}{A_{res}} \right| = \frac{1}{\sqrt{2}} = 0.707$$

gain A is 3 dB lower than A_{res}.

freq w_1 is lower 3 dB freq

at freq w_2 above w_0,

$$\delta = \pm \frac{1}{2Q_e}$$

$$\left|\frac{A}{A_{res}}\right| = \frac{1}{\sqrt{2}} = 0.707$$

w_2 is upper 3 dB freq.

Δw or $BW = w_2 - w_1$

$$\Delta w = \frac{[(w_2 - w_0) + (w_0 - w_1)]w_0}{w_0}$$

$$= \left[\frac{(w_2 - w_0)}{w_0} + \frac{w_0 - w_1}{w_0}\right] w_0 = [\delta + \delta] w_0$$

$$\Delta w = 2\delta w_0$$

$$\delta = \frac{1}{2Q_e}$$

$$2\delta = \frac{1}{Q_e}$$

$$\Delta w = \frac{w_0}{Q_e}$$

from (13.18) $Q_e = w_0 C R_t = \frac{R_t}{w_0 L}$

$$\Delta w = \frac{w_0}{R_t w_0 C} = \frac{1}{R_t C} \text{ rad/s.}$$

→ In single tuned amplifier 3dB B.W is $\frac{\omega_0}{Q}$

→ In double tuned amplifier B.W exceeds in single tuned by $\sqrt{(b^2-1) \pm 2b}$

→ b is +ve for practical ckt. $b < 0.414$ no real values of B.W

→ increase in b 3dB B.W of double tuned ↑. overshoot also increases

→ adv. large 3dB B.W than STA, & GBW, it provides Gain Vs freq curves steeper than flatter top

Effects of cascading single tuned Amp. on B.W

→ high gain, identical stages can be cascaded

→ overall Av is product of individual stages

→ high Av accompanied by narrow B.W

relative gain w.r.to gain at resonant freq fo

$$\left|\frac{A}{A_{res}}\right| = \frac{1}{\sqrt{1+(\delta Q_e)^2}}$$

for n stage

$$\left|\frac{A}{A_{res}}\right|^n = \left[\frac{1}{\sqrt{1+(2\delta Q_e)^2}}\right]^n$$

gain A is 3 dB lower than

→ 3dB freq for n stage cascade amplifier

by equating $\left|\frac{A}{A_{mid}}\right|^n$ to $\frac{1}{\sqrt{2}}$

$$\left[\frac{1}{\sqrt{1+(2\delta Q_e)^2}}\right]^n = 1/\sqrt{2} \qquad \left[1+(2Q_e\delta)^2\right]^{n} = 2^{1/n}$$

Squaring & Taking nth root $(2\delta Q_e)^2 = 2^{1/n}$

$$\left[1+(2\delta Q_e)^2\right]^{n/2} = 2^{1/2}$$

$$2\delta Q_e = \pm\sqrt{2^{1/n}-1}$$

Sub $\delta = \frac{\omega-\omega_0}{\omega_0}$, fractional freq. variation

$$2\left(\frac{f-f_0}{f_0}\right)Q_e = \pm\sqrt{2^{1/n}-1}$$

$$2(f-f_0)Q_e = \pm f_0\sqrt{2^{1/n}-1}$$

$$f_2 - f_0 = +\frac{f_0}{2Q_e}\cdot\sqrt{2^{1/n}-1}$$

$$f_0 - f_1 = +\frac{f}{2Q_e}\sqrt{2^{1/n}-1}$$

B.W of n stage identical amplifier is

$$B_{1n} = f_2 \sim f_1 = (f_2-f_0) + f_0 - f_1$$

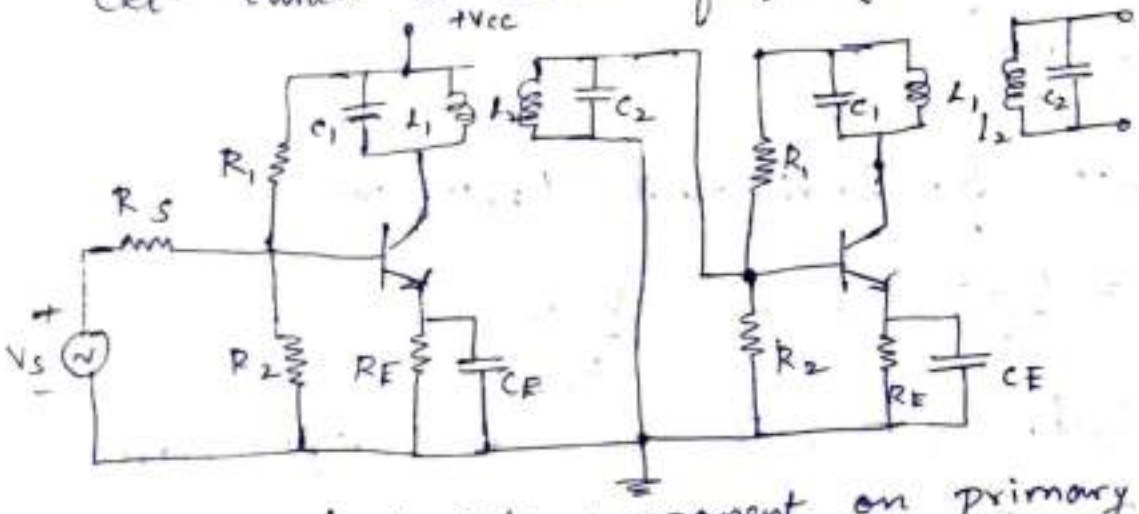
$$= \frac{f_0}{2Q_e}\sqrt{2^{1/n}-1} + \frac{f_0}{2Q_e}\sqrt{2^{1/n}-1}$$

$$= \frac{f_0}{Q_e}\sqrt{2^{1/n}-1}$$

$$= B_1\sqrt{2^{1/n}-1}$$

Double Tuned Amplifier:

→ Voltage developed across tuned circuit is coupled inductively to another tuned circuit. Both Tuned ckt tuned to same frequency.



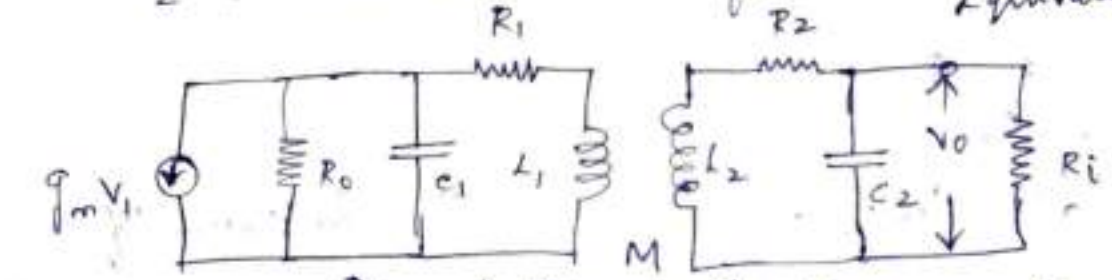
$L_1 C_1$ → tank ckt component on primary side

$L_2 C_2$ → tank ckt components on secondary side

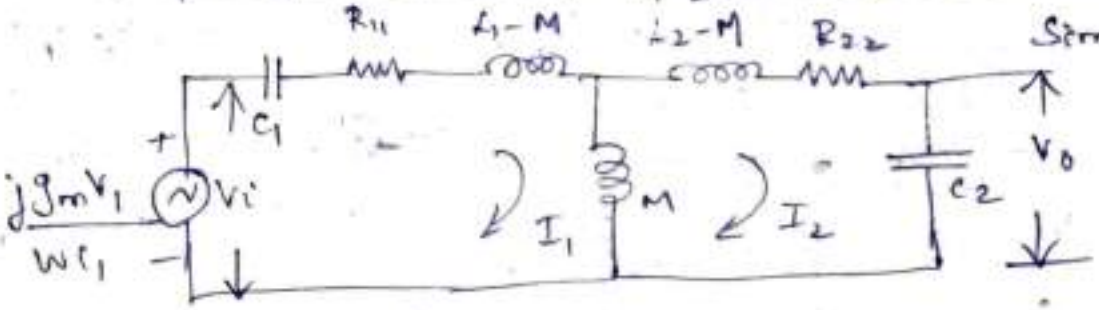
R_1 → series resistance of inductance L_1

R_2 → series resistance of inductance L_2

Equivalent ckt



Simplified Equivalent ckt.



$$R_{11} = \frac{\omega_0^2 L_1^2}{R_0} + R_1$$

$$R_{12} = \frac{\omega_0^2 L_2^2}{R_i} + R_2$$

gain A_i.

In simplified ckt current source replaced by voltage source which is in series with C_1.

→ & also effect of mutual inductance on primary & secondary side

$$Q = \frac{\omega_r L}{R}$$

Q factor for individual tank ckt

$$Q_1 = \frac{\omega_r L_1}{R_{11}} \quad \& \quad Q_2 = \frac{\omega_r L_2}{R_{22}}$$

Q factor for both ckt are kept same

$$Q_1 = Q_2 = Q$$

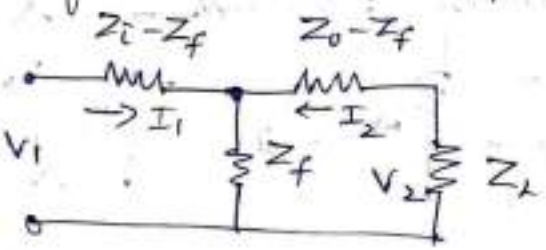
resonant frequency $\omega_r^2 = 1/L_1C_1 = 1/L_2C_2$

The o/p voltage can be given as

$$V_0 = \frac{-j}{\omega_r C_2} I_2$$

To calculate V_0/V_1 → represent I_2 interms of V_1

Find transfer admittance Y_T.



$(Z_i - Z_f) I_1 + Z_f (I_1 - I_2) = V_1$

$Z_i I_1 - Z_f I_2 = V_1$

$Z_i I_1 - Z_f I_2 = V_1$

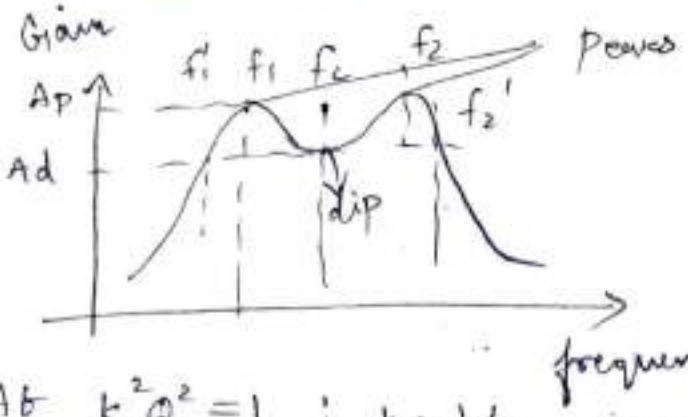
$$Y_T = \frac{I_2}{V_1} = \frac{I_2}{I_1 Z_{11}} = \frac{A_i}{Z_{11}}$$

$$= \frac{Z_f}{Z_f^2 - Z_i(Z_0 + Z_L)}$$

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

$$A_i = \frac{I_2}{I_1} = \frac{-Z_f}{Z_0 + Z_L}$$

The simplified ckt is similar to



At $k^2Q^2 = 1$; $k = 1/Q$, $f_1 = f_2 = f_0$ → is called critical coupling

→ for $k < 1/Q$ peak gain is less than maximum gain & coupling is poor

→ At $k > 1/Q$, circuit over coupled → shows double peak with more B.W

→ gain magnitude at peak is

$$|A_p| = \frac{g_m w_0 \sqrt{L_1 L_2} * Q}{2}$$

→ gain at dip is

$$|A_d| = |A_p| \frac{2kQ}{1+k^2Q^2}$$

ratio of peak & dip gain is γ

$$\gamma = \left|\frac{A_p}{A_d}\right| = \frac{1+k^2Q^2}{2kQ}$$

$$kQ = \gamma + \sqrt{\gamma^2 - 1}$$

The B.W b/w freq at which gain is $|A_d|$ is useful B.W is

$$B.W = 2\delta' = \sqrt{2}\,(f_2 - f_1)$$

At 3dB $V = \sqrt{2}$

gain A is

At 3dB $\gamma = \sqrt{2}$

$$kQ = \gamma + \sqrt{\gamma^2 + 1} = \sqrt{2} + \sqrt{\sqrt{2}^2 + 1} = 2.414$$

3 dB B.W $= 2\delta' = \sqrt{2}(f_2 - f_1)$

$$= \sqrt{2}\left[f_r\left(1 + \frac{1}{2Q}\sqrt{k^2Q^2 - 1}\right) - f_r\left(1 - \frac{1}{2Q}\sqrt{k^2Q^2 - 1}\right)\right]$$

$$= \sqrt{2}\left(\frac{f_r}{Q}\sqrt{k^2Q^2 - 1}\right) = \sqrt{2}\frac{f_r}{Q}\sqrt{(2.414)^2 - 1} = \frac{3.1 f_r}{Q}$$

3 dB BW for single Tuned is $2f_r/Q$

" " " Double " is $3.1 f_r/Q$ > than single tuned.

$$\frac{w - w_0}{w_0} = \delta$$

$$\frac{w}{w_0} - \frac{w_0}{w_0} = \delta$$

$$\frac{w}{w_0} = \delta + 1$$

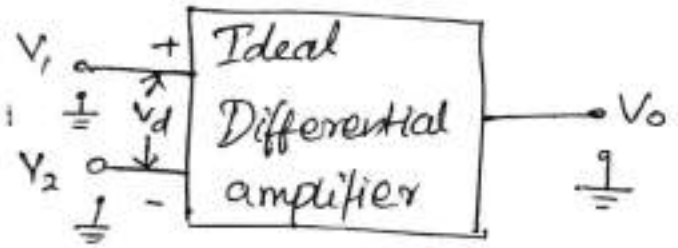
Differential Amplifier :-

→ A device which accepts an input Signal & produces an output Signal proportional to the input, is called an amplifier.

→ An amplifier which amplifies the difference between the two input signal is called differential amplifier.

Basics of Differential Amplifier:

→ consider an ideal differential amplifier



V_d = Difference input Signal

$= V_1 - V_2$

→ An ideal difference amplifier, the output voltage V_o is proportional to the difference between the two i/p Signal. Hence $V_o \propto (V_1 - V_2)$.

i) Differential Gain (A_d):

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

where A_d - constant of proportionality.

→ The A_d is the gain with which the differential amplifier amplifies the difference between two input Signal. Hence it is called differential gain of amplifier.

A_d - Differential Gain.

V_d - Difference Voltage

$V_o = A_d V_d$

$A_d = V_o / V_d$. $20 \log_{10}(A_d)$ in dB

ii) Common Mode Gain (A_c):

→ If we apply two input voltages which are equal in all the respects ie $V_1 = V_2$. then ideally $V_o = (V_1 - V_2) A_d = 0$

→ But practically, the difference amplifier also depends on the average common level of the two inputs.

→ The average level of the two input signal is called common mode signal, V_c

$$V_c = \frac{V_1 + V_2}{2}$$

→ The gain with which the differential amplifiers amplifies the common mode signal to produce the output is called common mode gain of the differential amplifier denoted as A_c

$$V_o = A_c V_c$$

→ The total output of any differential amplifier can be expressed as

$$V_o = A_d V_d + A_c V_c$$

Note: For an ideal differential amplifier

A_d = Infinite

A_c = Zero

$V_o = 0$

Practically

A_d - Very large

A_c - Not Zero (but very small)

iii) Common Mode Rejection Ratio (CMRR)

→ When $V_1 = V_2$, many disturbance signals, noise

signals appear as a common input signal to both input terminals which should be rejected.

→ " The ability of a differential amplifier to reject a common mode signal is expressed by a ratio called common mode rejection ratio, CMRR".

→ CMRR is defined as the ratio of the differential voltage gain A_d to common mode voltage gain A_c.

$$CMRR = \rho = \left|\frac{A_d}{A_c}\right|$$

→ Ideally $A_c = 0$ → CMRR is infinite.

→ Practically A_d - large, A_c - small, So CMRR is very large.

$$CMRR \text{ in } dB = 20 \log \left|\frac{A_d}{A_c}\right| \text{ dB}$$

→ output voltage in terms of CMRR.

$$V_o = A_d V_d + A_c V_c$$

$$= A_d V_d \left[1 + \frac{A_c V_c}{A_d V_d}\right]$$

$$= A_d V_d \left[1 + \frac{(V_c/V_d)}{(A_d/A_c)}\right]$$

$$V_o = A_d V_d \left[1 + \frac{1}{CMRR} \frac{V_c}{V_d}\right]$$

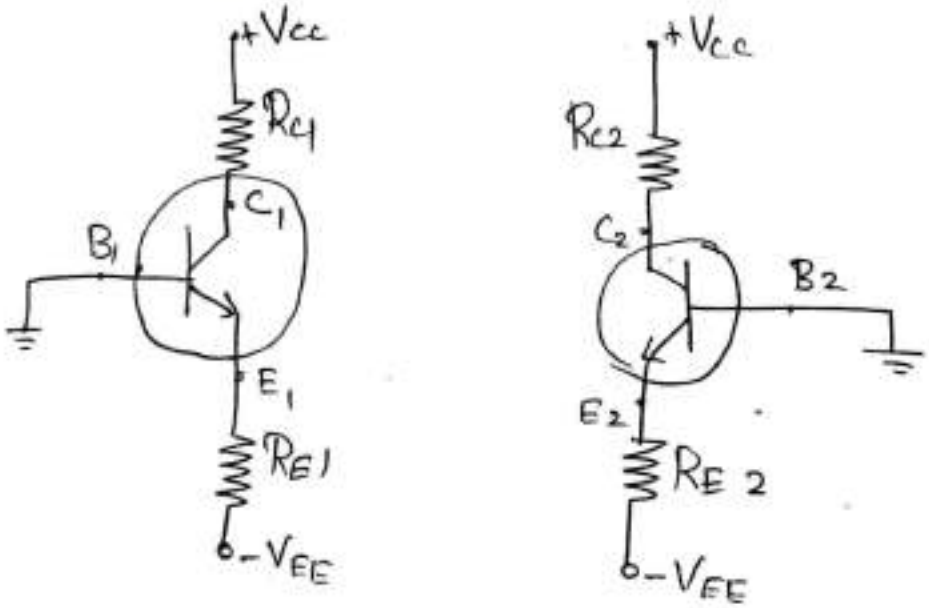
→ CMRR is very large, So if V_c & V_d both are present, the output is proportional to the difference signal only, the common mode component is greatly rejected.

Features of Differential Amplifier:

→ High differential Voltage gain
→ Low common mode gain
→ High CMRR
→ Two input terminals
→ High input impedance
→ Large bandwidth
→ Low offset Voltages & currents
→ Low output impedance

Transistorised Differential Amplifier:

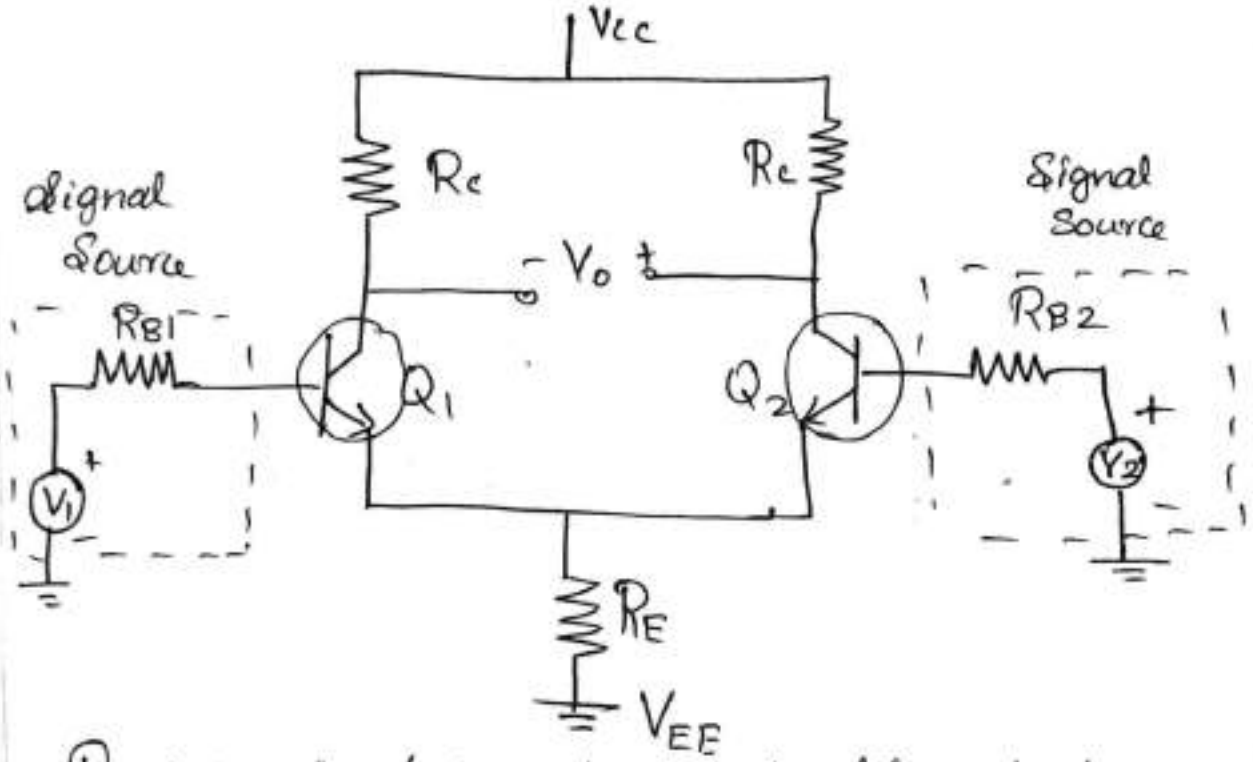
→ The transistorised differential amplifier uses the emitter biased circuits which are identical in characteristics.



→ Q_1 & Q_2 have matched characteristics.
Thus $R_{C1} = R_{C2} = R_C$, $R_{E1} = R_{E2}$, $|V_{CC}| = |V_{EE}|$

→ The differential amplifier can be obtained by connecting emitter of Q_1 to emitter of Q_2, Since now R_{E1} || to R_{E2}, we replace it with R_E.

→ Input V_1 is applied to B_1 & V_2 to B_2. The balanced output is taken across C_1 & C_2.
→ This amplifier is called emitter coupled differential amplifier.
→ Based upon, where the output is taken it is divided into two types.



Dual input, balanced output differential amplifier

Balanced output: double ended output or floating output is the output taken between two collectors, none of them is grounded.

Unbalanced output: or Single ended output is the output taken between any one of the collectors & the ground.

→ The circuit operation is mainly divided into two modes.

* Differential Mode Operation
* Common Mode Operation

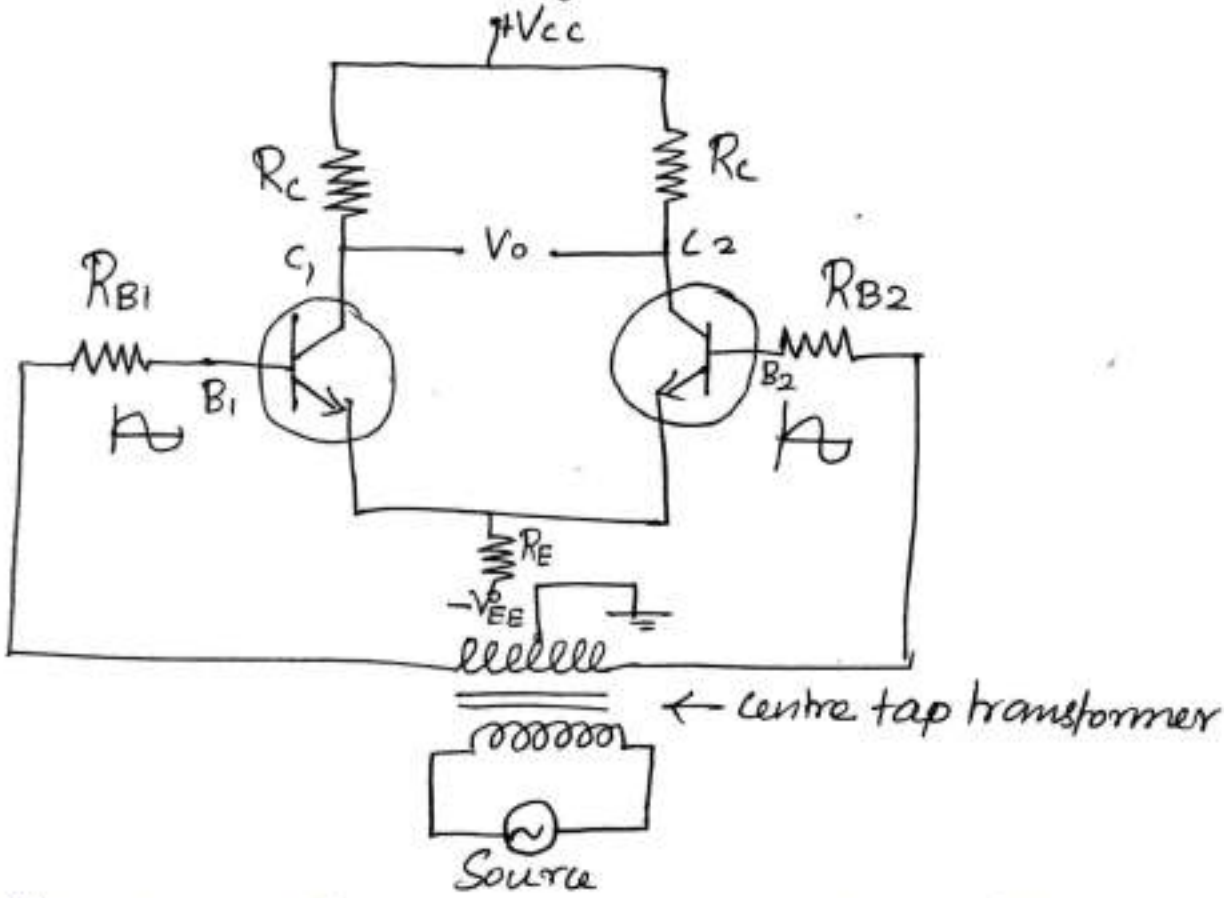
Differential Mode Operation:

→ In this mode, V_1 & V_2 are different from each other.

→ Consider V_1 is 180° out of phase with V_2, which can be obtained using center tap transformer.

→ Assume V_1 is +ve going on B_1 & negative going on Q_2.

→ with a +ve going signal on B_1, an amplified.



negative going signal develops on C_1 & due to +ve going signal, current through R_E also increases & a +ve going signal develops across R_E.

→ Similarly, when -ve going signal is applied at B_2 an amplified +ve going signal develops on C_2 & a -ve going signal develops across R_E.

→ Due to the effect of Q_1 & Q_2, the signal voltages at R_E are equal in magnitude but 180° out of phase.

→ The two signals cancel each other & there is no signal across R_E.

$$V_o = V_1 - V_2 \quad \text{Eg } +10 - (-10) = 20$$

→ Hence, the difference output, V_o is twice as large as signal voltage from either collector to ground.

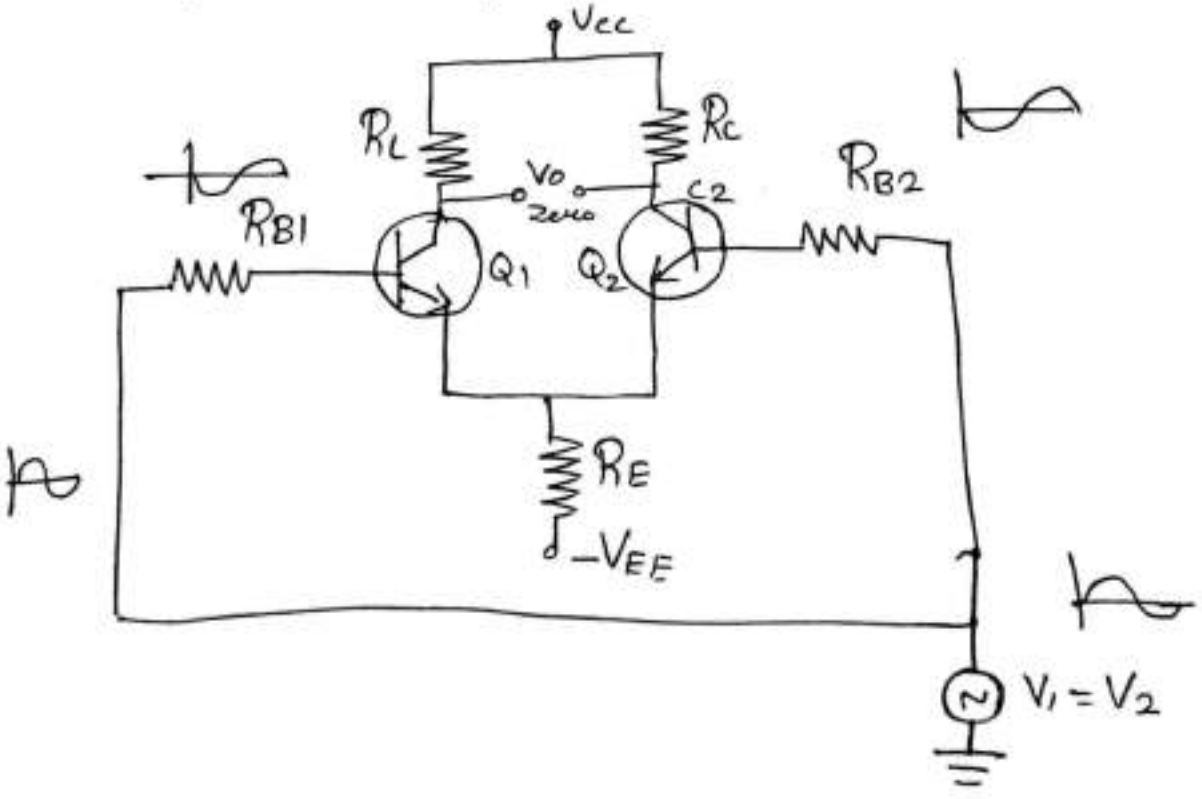
Common Mode Operation:

→ In this mode, $V_1 = V_2$ both in magnitude & phase

→ In phase signal voltages at the bases of Q_1 & Q_2 causes in phase signal voltages to appears across R_E which add together.

→ So, R_E provides negative feedback, which reduces A_c

→ The difference output V_o is almost zero.



Configurations of Differential Amplifier:

→ There are four configurations.

i) Dual input, balanced output differential Amplifier (DIBO)
ii) Dual input, unbalanced output differential Amplifier (DIUO)
iii) Single input, Balanced output differential Amplifier (SIBO)
iv) Sigle input, Unbalanced output differential Amplifier (SIUO)

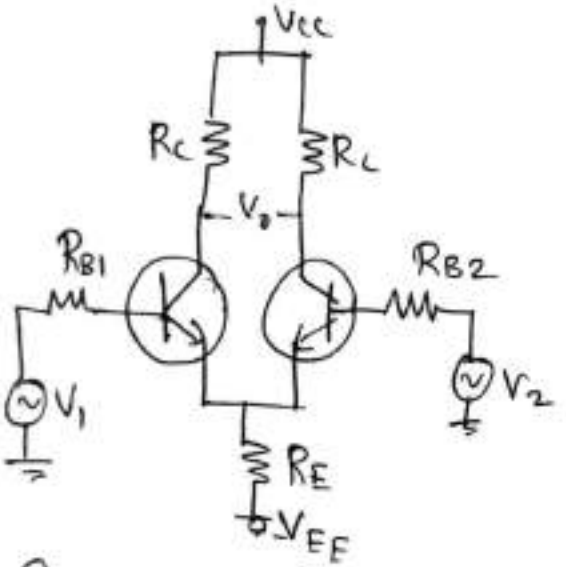
Differential amplifier

Balanced output: O/p taken between two collectors

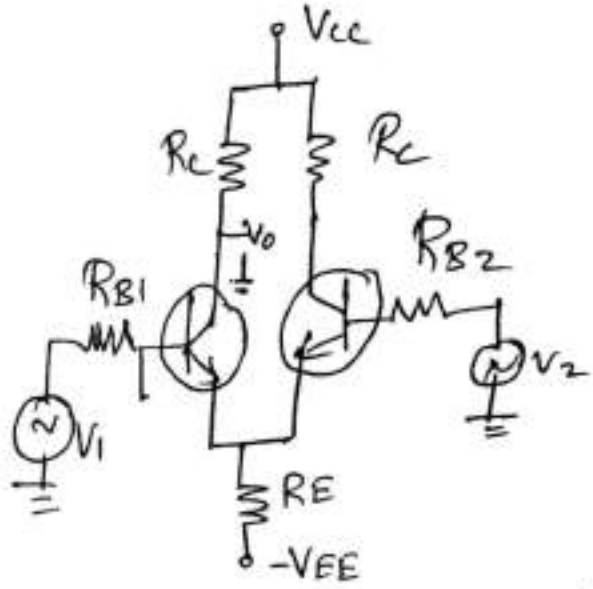
Unbalanced output: O/p taken between on collector with respect to ground.

Single input : Signal is given to one input terminal the other is grounded

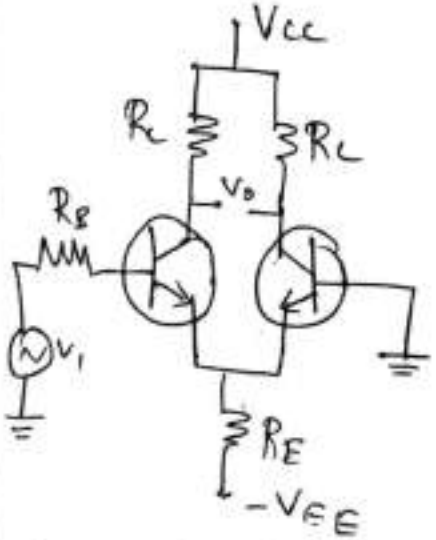
Double ended input : Signal is given to both the input terminal.



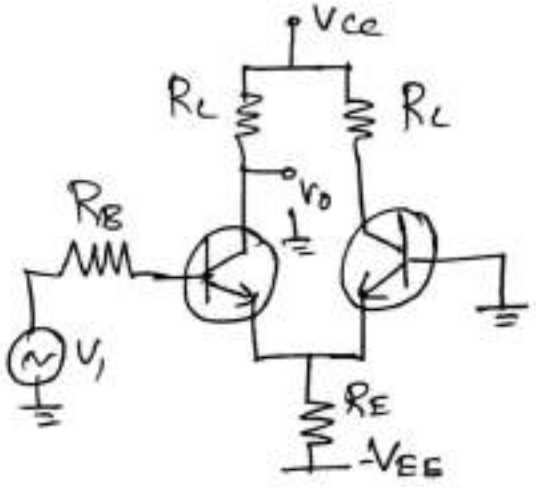
Dual input, balanced output



Dual input unbalanced output



Single input, balanced output



Single input, unbalanced output

→ A multistage amplifier with a desired voltage gain can be formed using a direct connection between successive stages of differential amplifier.

JET Input Stage:

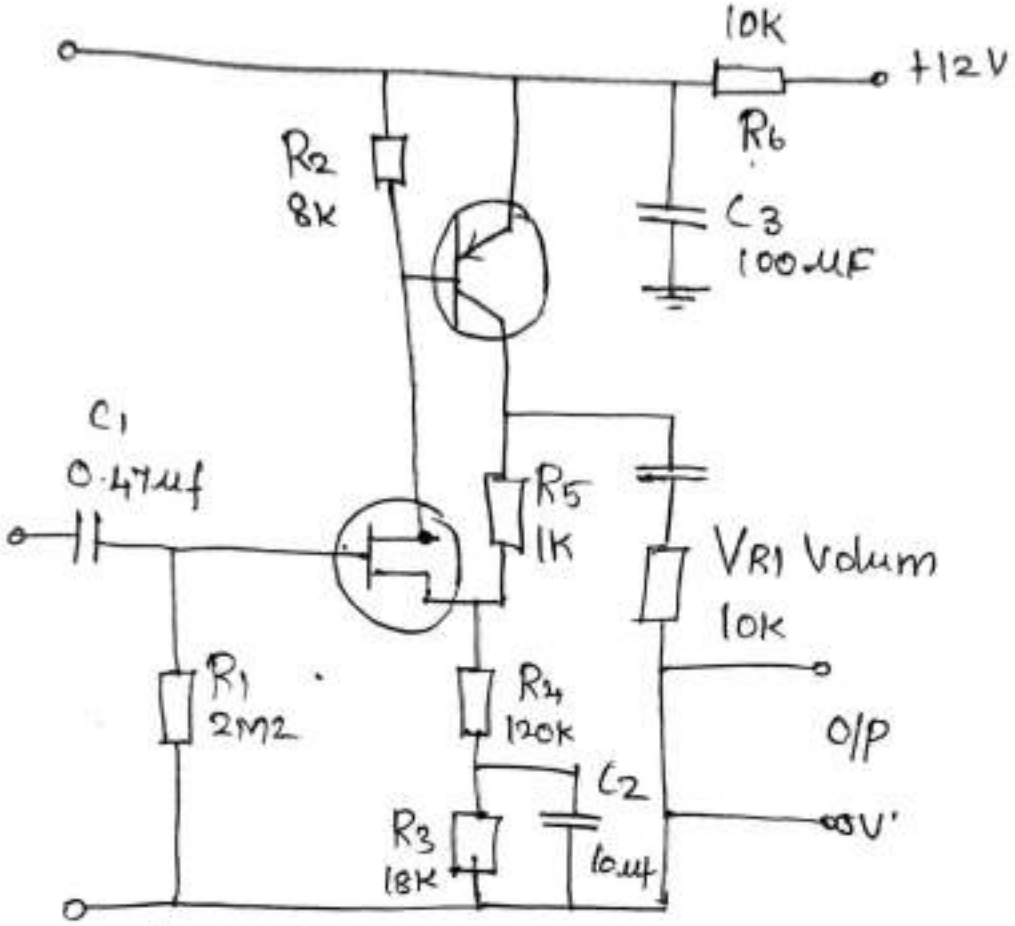


fig: high Impedance JFET Input Stage.

→ where very high impedance & low noise is required in an amplifier input, it is common to use a FET in an amplifier's input stage

→ Very high input impedance is obtainable with JFETs as its gate is voltage, rather than current operated.

→ Therefore the JFET takes hardly any current from the device connected to the amplifier input.

→ Even higher input impedance are available where MOSFETs with insulated gate construction (IGFETs) are used.

→ Although FETs generally have less voltage gain & less bandwidth than BJT transistors they also create much less internally generated noise, which makes them ideally suited for use in the early stages of an amplifier, where good signal to noise ratio is important.

Operation:

→ Because the input resistance of the JFET is extremely high, the input impedance of the circuit is approximately the value of R_1 & as practically no current is flowing into the input there is no potential across R_1, therefore the gate of transistor1 is effectively at zero volts.

→ To operate correctly, the gate of the N channel JFET must be more negative than the Source, this is achieved by making the Source of Transistor1 positive.

→ The biasing of the JFET is set by R_2 & R_3

→ As JFET gain is not particularly high, extra gain is provided by PNP transistor T_2

→ The overall gain of the two-stage amplifier is set at approximately 11 by the negative feedback provided by R_4 & R_5

Decoupling:

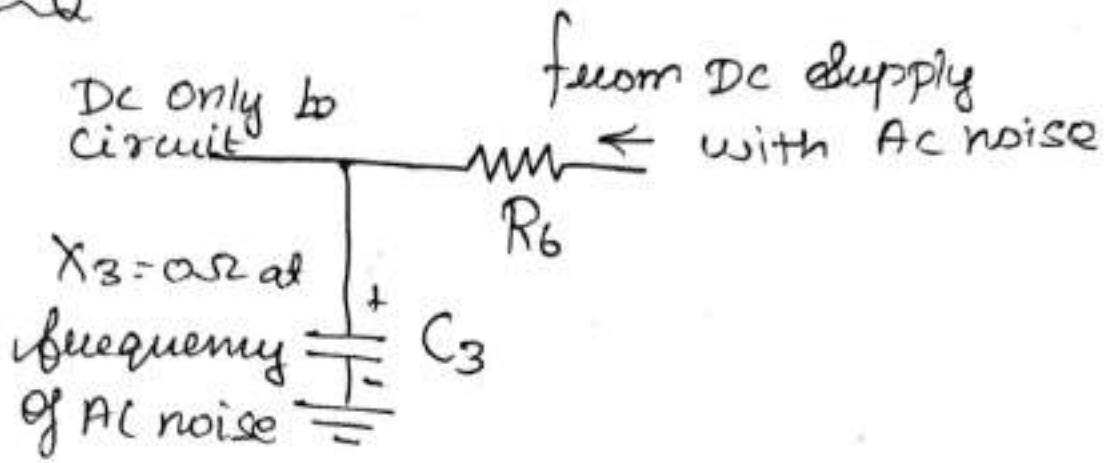


fig : Supply Decoupling

→ In the previous figure R_3 is decoupled by C_2 so that the bottom end of R_3 is effectively at ground potential as far as Ac is concerned, the value of C_2 is not particularly large in this circuit, as the larger the value of electrolytic capacitor the more noise it will produce, & the aim of the circuit is to keep internally generated noise to a minimum.

→ C_1 & C_4 coupling capacitors (also relatively small values) provide isolation from any Dc voltages present on any connected circuits.

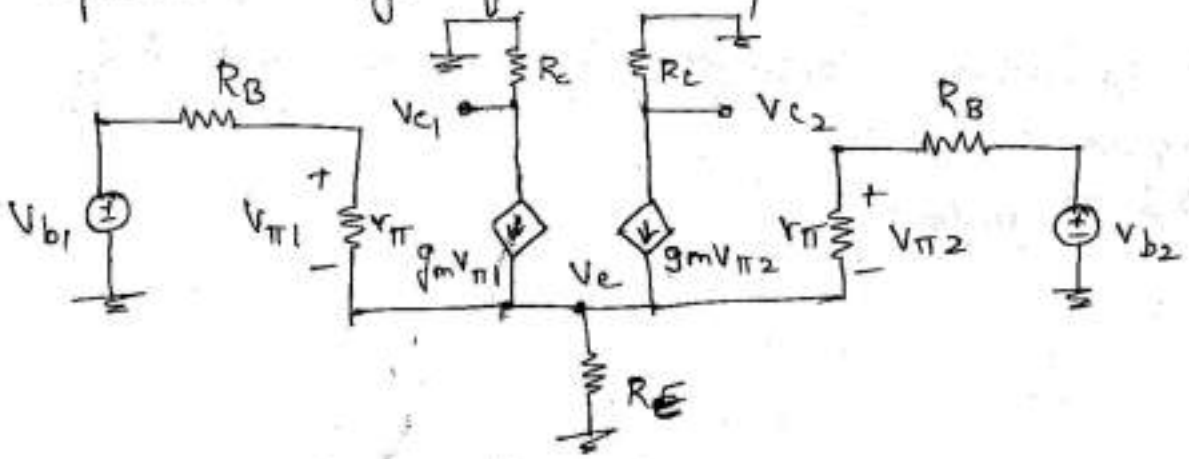
→ Using a very high value for R_1 produces a high input impedance but the higher the value, the more prone the

circuit will be to instability & oscillation.

→ To prevent this possibility, effective decoupling from other circuits & the supply is necessary, decoupling here is provided by R_6 & C_3 as shown in the figure.

Small Signal Analysis of Differential Amplifier

Assume early voltage $V_A = \infty$; $r_o = \infty$
constant current source is not ideal but represented by finite output resistance R_o.



Apply KVL to V_e node

$$\frac{V_{\pi 1}}{r_\pi} + g_m V_{\pi 1} + g_m V_{\pi 2} + \frac{V_{\pi 2}}{r_\pi} - \frac{V_e}{R_E} = 0 \quad \text{—①}$$

$$\frac{V_{\pi 1} + g_m V_{\pi 1} r_\pi}{r_\pi} + \frac{g_m V_{\pi 2} r_\pi + V_{\pi 2}}{r_\pi} = \frac{V_e}{R_E} \quad \text{—②}$$

Sub $g_m r_\pi = \beta$ in ②

$$\frac{V_{\pi 1} + \beta V_{\pi 1}}{r_\pi} + \frac{\beta V_{\pi 2} + V_{\pi 2}}{r_\pi} = \frac{V_e}{R_E}$$

$$V_{\pi 1}\frac{(1+\beta)}{r_\pi} + V_{\pi 2}\frac{(1+\beta)}{r_\pi} = \frac{V_e}{R_E} \quad \text{—③}$$

From the ckt

$$\frac{V_{\pi 1}}{r_\pi} = \frac{V_{b1} - V_e}{r_\pi + R_B} \quad \text{and} \quad \frac{V_{\pi 2}}{r_\pi} = \frac{V_{b2} - V_e}{r_\pi + R_B}$$

$$V_{\pi 1} = \frac{r_\pi (V_{b1} - V_e)}{r_\pi + R_B} \quad \text{and} \quad V_{\pi 2} = \frac{r_\pi (V_{b2} - V_e)}{r_\pi + R_B} \quad \text{— (4)}$$

Sub values of $V_{\pi 1}$ & $V_{\pi 2}$ from (4) in (3)

$$\frac{r_\pi (V_{b1} - V_e)}{r_\pi + R_B}\left(\frac{1+\beta}{r_\pi}\right) + \frac{r_\pi (V_{b2} - V_e)}{r_\pi + R_B}\left(\frac{1+\beta}{r_\pi}\right) = \frac{V_e}{R_E}$$

$$\frac{V_{b1} + V_{b2} - 2V_e}{r_\pi + R_B}(1+\beta) = \frac{V_e}{R_E}$$

Solving for V_e

$$\frac{V_{b1} + V_{b2} - 2V_e}{r_\pi + R_B} = \frac{V_e}{R_E(1+\beta)}$$

$$\frac{V_{b1} + V_{b2}}{r_\pi + R_B} = \frac{V_e}{R_E(1+\beta)} + \frac{2V_e}{r_\pi + R_B}$$

$$\frac{V_{b1} + V_{b2}}{r_\pi + R_B} = \frac{V_e r_\pi + R_B V_e + 2V_e R_E + 2V_e R_E \beta}{R_E(1+\beta)(r_\pi + R_B)}$$

$$\frac{V_{b1} + V_{b2}}{r_\pi + R_B} = \frac{V_e (r_\pi + R_B) + 2V_e R_E (1+\beta)}{R_E(1+\beta)(r_\pi + R_B)}$$

$$= \frac{V_e\left[(r_\pi + R_B) + 2R_E(1+\beta)\right]}{R_E(1+\beta)(r_\pi + R_B)}$$

$$V_e = \frac{V_{b1} + V_{b2}}{(r_\pi + R_B) + 2R_E(1+\beta)} \times R_E(1+\beta)$$

$$V_e = \frac{V_{b1} + V_{b2}}{2 + \frac{r_\pi + R_B}{R_E(1+\beta)}} \quad \text{— (5)}$$

One sided o/p at collector of Q_2 is given by

$$V_o = V_{C_2} = -(g_m V_{\pi_2}) R_C = -g_m r_\pi \left(\frac{V_{\pi_2}}{r_\pi}\right) R_C \quad \text{— (6)}$$

Sub V_{π_2}/r_π from (4) in (6)

$$V_o = -g_m r_\pi \left(\frac{V_{b_2} - V_e}{r_\pi + R_B}\right) R_C = -\beta R_C \left(\frac{V_{b_2} - V_e}{r_\pi + R_B}\right) \quad \because g_m r_\pi = \beta \quad \text{— (7)}$$

Sub (5) in (7)

$$V_o = -\beta R_C \left[\frac{V_{b_2} - \dfrac{(V_{b_1} + V_{b_2})}{2 + \dfrac{r_\pi + R_B}{(1+\beta) R_E}}}{r_\pi + R_B}\right]$$

$$V_o = \frac{-\beta R_C}{r_\pi + R_B} \left[\frac{2V_{b_2} + V_{b_2}\left(\dfrac{r_\pi + R_B}{(1+\beta) R_E}\right) - V_{b_1} + V_{b_2}}{2 + \dfrac{r_\pi + R_B}{(1+\beta) R_E}}\right]$$

$$= \frac{-\beta R_C}{r_\pi + R_B} \cdot \frac{V_{b_2}\left(1 + \dfrac{r_\pi + R_B}{(1+\beta) R_E}\right) - V_{b_1}}{2 + \dfrac{r_\pi + R_B}{(1+\beta) R_E}} \quad \text{— (8)}$$

For ideal current source $R_E = \infty$

$$V_o = \frac{-\beta R_C (V_{b_2} - V_{b_1})}{2(r_\pi + R_B)} \quad \text{— (9)}$$

$$V_o = \frac{\beta R_c}{2(r_\pi + R_B)} \cdot V_d + \frac{\left(V_{cm} - \frac{V_d}{2}\right)\left(\frac{r_\pi + R_B}{(1+\beta) R_E}\right)}{2 + \frac{r_\pi + R_B}{(1+\beta) R_E}}$$

$$V_o = \frac{\beta R_c}{2(r_\pi + R_B)} \cdot V_d - \frac{g_m R_c}{1 + \frac{2(1+\beta) R_E}{r_\pi + R_B}} \cdot V_{cm} \quad \text{— (1)}$$

$$V_o = A_d V_d + A_{cm} V_{cm} \quad \text{— (2)}$$

comparing (1) & (2)

$$A_d = \frac{\beta R_c}{2(r_\pi + R_B)} \quad \text{and} \quad A_{cm} = \frac{-g_m R_c}{1 + \frac{2(1+\beta) R_o}{r_\pi + R_B}}$$

Considering Source resistors $R_B = 0$;

$$A_d = \frac{\beta R_c}{2 r_\pi} = \frac{g_m R_c}{2} = \frac{I_{CQ} R_c}{2 V_T}$$

$\because g_m = \frac{\beta}{r_\pi}$ & $g_m = \frac{I_{CQ}}{V_T}$

$$A_d = \frac{I_Q R_c}{4 V_T}$$

$$I_{CQ} = \frac{I_Q}{2}$$

$$A_{cm} = \frac{-g_m R_c}{1 + \frac{2(1+\beta) R_o}{r_\pi}} = \frac{-\frac{I_{CQ} R_c}{V_T}}{1 + \frac{2(1+\beta) R_o g_m}{\beta}}$$

$$r_\pi = \frac{\beta}{g_m}$$

$$= \frac{-\frac{I_{CQ} R_c}{V_T}}{\frac{1 + 2(1+\beta) R_E I_{CQ}}{V_T \beta}} = \frac{-\frac{I_Q R_c}{2 V_T}}{1 + \frac{2(1+\beta) R_E I_Q}{2 V_T \beta}}$$

$$CMRR = \rho = \frac{A_d}{A_{CM}}$$

$$= \frac{\frac{I_{CQ} R_c}{2 V_T}}{\frac{I_{CQ} R_c}{\frac{V_T}{1 + (1+\beta) R_E I_Q}{V_T \beta}}} = \frac{1}{2}\left[1 + \frac{(1+\beta) R_E I_Q}{V_T \beta}\right]$$

$$A_d = \frac{V_o}{V_d} = \frac{-\beta R_c (V_{b_2} - V_{b_1})}{2 V_d (r_\pi + R_B)}$$

$$A_d = \frac{\beta R_c}{2(r_\pi + R_B)} \qquad \left[\because V_d = (V_{b_1} - V_{b_2}),\ V_{b_2} - V_{b_1} = -V_d\right]$$

If $R_B = 0$

$$A_d = \frac{\beta R_c}{2 r_\pi} = \frac{g_m r_\pi R_c}{2 r_\pi} \qquad \left[\because g_m r_\pi = \beta\right]$$

$$\boxed{A_d = \frac{g_m R_c}{2}}$$

is identical to voltage transfer characteristics

$$V_{b_1} = V_{cm} + \frac{V_d}{2} \quad ; \quad V_{b_2} = V_{cm} - \frac{V_d}{2} \qquad \text{---(10)}$$

[V_{cm} - common mode voltage]

Sub (10) in (8)

$$V_o = -\beta R_c \left[\frac{\left(V_{cm} - \frac{V_d}{2}\right)\left[1 + \frac{r_\pi + R_B}{(1+\beta) R_E}\right] - \left(V_{cm} + \frac{V_d}{2}\right)}{2 + \frac{r_\pi + R_B}{(1+\beta) R_o}} \right] \qquad \text{---(11)}$$

Unit - IV

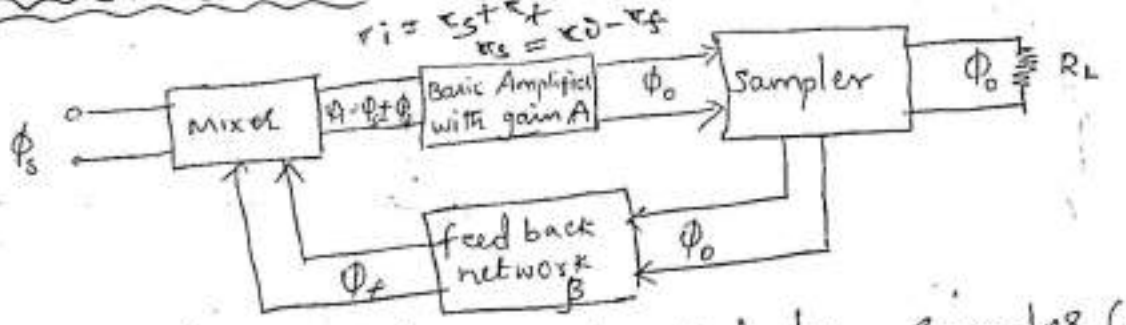
Feedback Amplifiers

Concepts

* In large signal amplifiers and measuring instruments distortion should be avoided.
* gain must be independent of variation in voltage of dc supply, and values of circuit components.
* This is achieved by feedback, portion of output signal is combined with normal input signal and feedback is accomplished.

GAIN WITH FEED BACK

BLOCK DIAGRAM!



* output quantity is sampled by sampler (voltage, current) and fed to the feedback network.
* output of feedback network is combined with external source signal ϕ_s throu' mixer and fed to basic amplifier.
* mixer is of two types series mixer and shunt mixer.

$$A - \text{gain} = \frac{\phi_o}{\phi_i}$$

$$\beta - = \frac{\phi_f}{\phi_o}$$

A_f - gain of feedback Amplifier $= \frac{\phi_o}{\phi_s}$

ϕ_s - a.c signal in the input side

ϕ_f - feedback signal

Positive feedback

* If the feedback signal ϕ_f is in phase with i/p signal ϕ_s, the net effect will increase the input signal. $\phi_i = \phi_s + \phi_f$.

* The input voltage applied to basic amplifier is increased by increasing ϕ_o exponentially.

* This type of feedback is positive or regenerative feedback.

* Gain of the amplifier is,

$$A_f = \frac{\phi_o}{\phi_s} = \frac{\phi_o}{\phi_i - \phi_f} = \frac{1}{\frac{\phi_i}{\phi_o} - \frac{\phi_f}{\phi_o}} = \frac{1}{\frac{1}{A} - \beta} = \frac{A}{1 - A\beta}$$

$|A_f| > |A|$. loop gain $= A\beta$

* product of open loop gain and feed back factor is called loop gain.

* If $|A\beta| = 1$, then $A_f = \infty$. the gain of amplifier with positive feedback is infinite and amplifier gives a.c o/p without ac i/p. and acts as an oscillator.

* +ve fdb. increases instability and reduces the bandwidth and increases distortion and noise

Negative feedback

* If the feedback ϕ_f is out of phase with input signal then $\phi_i = \phi_s - \phi_f$.

* The input voltage is decreased and the output is decreased

* Voltage gain is reduced and feedback is negative or degenerative feedback.

* Gain of amplifier with negative feedback

$$A_f = \frac{\phi_o}{\phi_s} = \frac{\phi_o}{\phi_i + \phi_f} = \frac{1}{\frac{\phi_i}{\phi_o} + \frac{\phi_f}{\phi_o}} = \frac{1}{\frac{1}{A} + \beta}$$

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

* $|A_f| < |A|$. If $|A\beta| \gg 1$. $A_f = 1/\beta$,

β is feedback ratio.

* -ve feedback improves the performance of electronic amplifiers.

* -ve feedback increase bandwidth, decrease distortion and noise

Effects of negative feedback on gain stability

Gain $A_f = \frac{A}{1+A\beta}$ $\quad \frac{dA_f}{dA} = \frac{vu' - uv'}{v^2}$

Diff. w.r.to A

$$\frac{dA_f}{dA} = \frac{1}{(1+A\beta)^2} = \frac{A_f}{A}\frac{1}{1+A\beta}$$

* $\therefore \quad \frac{dA_f}{A_f} = \frac{dA}{A} \quad \frac{1}{1+A\beta}$

* $\frac{dA_f}{A_f}$ represents fractional change in amp Volt. gain with feedback

* $\frac{dA}{A}$ denotes fractional change in Voltage gain without feedback.

* $\frac{1}{1+A\beta}$ is called sensitivity.

* Sensitivity is defined as ratio of % change in voltage gain without feedback

$$\text{Sensitivity} = \frac{\left(\frac{dA_f}{A_f}\right)}{\left(\frac{dA}{A}\right)} = \frac{1}{1+A\beta}$$

* reciprocal of the term Sensitivity is called desensitivity.

$$\text{desensitivity} = 1+A\beta$$

Cutoff Frequencies:

* Bandwidth of an amplifier is the difference between the upper cutoff f_2 and lower cutoff frequency f_1.

* The product of Voltage gain and B.W of an amplifier without feedback and with feedback remains the same (ie) $A_f \times BW_f = A \times BW$

* Decreased Noise

* many sources of noise depending upon active devices
* Using negative feedback with feedback ratio β, the noise N, can be reduced by factor $\frac{1}{1+A\beta}$
* The noise with feedback is given by

$$N_f = \frac{N}{1+A\beta}$$

Increased in Input Impedance:

* Amplifier should have high input impedance, so that it will not load the preceding stage
* It is achieved by negative series voltage feedback
* Input impedance with feedback is given by

$$Z_{if} = Z_i (1+A\beta)$$

Decrease in Output Impedance

* amplifier with low o/p impedance is capable of delivering power to load without loss
* It is achieved by negative series voltage feedback in an amplifier
* The output impedance is

$$Z_{of} = \frac{Z_o}{1+A\beta}$$

* output impedance is reduced by factor $1+A\beta$

* The Voltage gain of feedback amplifier reduces by the factor $1/1+A\beta$

* Bandwidth is increased by $1+A\beta$.

$$BW_f = BW(1+A\beta)$$

$A \rightarrow$ midband gain without feedback.

* Due to negative feedback upper cutoff frequency f_{2f} is increased by factor $(1+A\beta)$.

* lower cutoff freq f_{1f} is decreased by factor $(1+A\beta)$.

* upper and lower 3dB freq. with negative feedback is given by

$$f_{2f} = f_2(1+A\beta) \quad \text{and} \quad f_{1f} = \frac{f_1}{(1+A\beta)}$$

Decreased Distortion

* An amplifier with an open loop Voltage gain and a total harmonic distortion D,

* The introduction of negative feedback with feedback ratio β, the distortion will reduce to

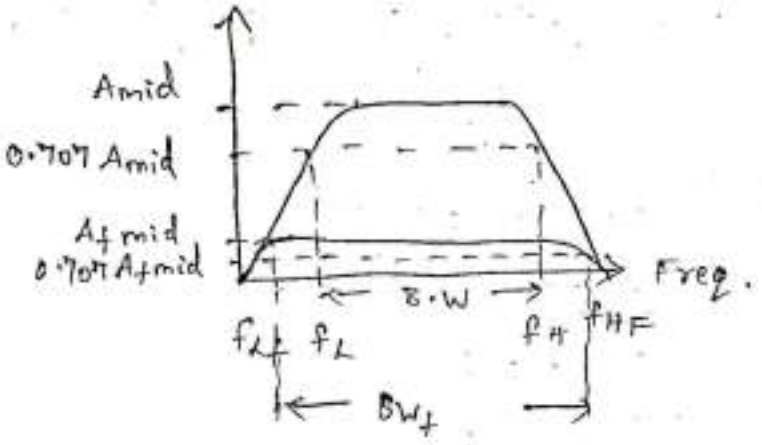
$$D_f = \frac{D}{1+A\beta}$$

Bandwidth

BW = Upper cutoff frequency − Lower cut off freq.

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

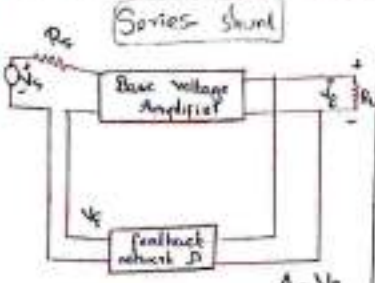
* $(f_{Hf} - f_{Lf}) > (f_H - f_L)$ ∴ hence B.W of amplifier with feedback is greater than B.W of amplifier without feedback



Voltage Series

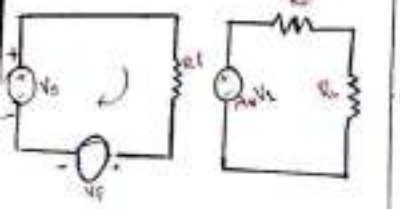
(voltage mixing voltage sampling)

(Series - Shunt)



$A = \frac{V_o}{V_f}$

Input Impedance :-



Apply KVL to input circuit

$V_s - I_i R_i - V_f = 0$ $\quad A_v = \frac{V_o}{V_i}$

$V_s = I_i R_i + V_f$ $\quad A_v\beta = 1$, $\beta = \frac{V_f}{V_o}$

$= I_i R_i + \beta V_o \rightarrow (1)$ $\quad R = \frac{V_i}{I_i}$

$V_o = A_v V_i \left(\frac{R_o}{R_o + R_o}\right) = A_f V_i$

$\Rightarrow A_{vf} I_i R_i \rightarrow (2)$ $\quad A_{vf} = \frac{V_o}{V_f}$

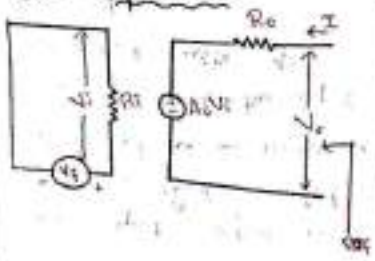
Subs (2) in (1).

$V_s = I_i R_i + \beta A_{vf} I_i R_i$

$V_s = I_i R_i (1 + \beta A_{vf})$

$$R_{if} = \frac{V_s}{I_i} = R_i (1 + \beta A_{vf})$$

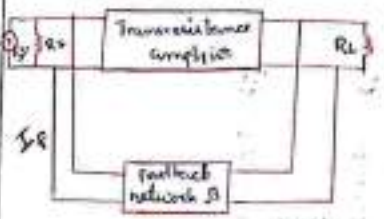
Output Impedance



Voltage Shunt

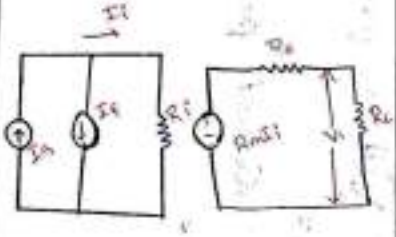
(current mixing voltage sampling)

(Shunt - Shunt)



$A = \frac{V_o}{I_f}$

Input Impedance



Darlington

$R_m = \frac{V_o}{I_i}$

$V_o = R_m I_i$

Apply KCL at input node,

$I_f = \beta V_o$

$I_s = I_i + I_f$

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

The output voltage V_o is given by,

$V_o = R_m I_i \times \frac{R_L}{R_L + R_o}$

$= R_{mf} I_i \rightarrow (2)$

Subs (2) in (1)

$I_s = I_i + \beta R_{mf} I_i$

$I_s = I_i (1 + \beta R_{mf})$

Input resistance with feedback is R_{if}.

$R_{mf} = \frac{R_m R_L}{R_L + R_o}$

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

$R_{if} = \frac{V_i}{I_i (1 + \beta R_{mf})}$

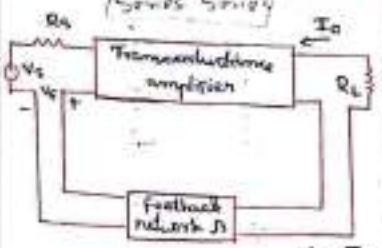
$$R_{if} = \frac{R_i}{1 + \beta R_{mf}}$$

$R_i = \frac{V_i}{I_i}$

Current series

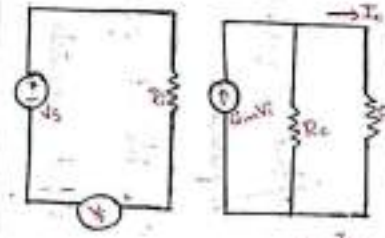
(voltage mixing current sampling)

(Series - Series)



$A = \frac{I_o}{V_f}$

Input Impedance



$G = \frac{I_o}{V_i}$ $\quad \beta = \frac{V_f}{I_o}$

Apply KVL to input side

$V_s - I_i R_i - V_f = 0$

$V_s = I_i R_i + V_f$

$= I_i R_i + \beta I_o \rightarrow (1)$

From output circuit,

$I_o = G_m V_i \left(\frac{R_o}{R_o + R_L}\right)$

$= G_{mf} V_i \rightarrow (2)$ $\quad G_{mf} = \frac{G_m R_o}{R_o + R_L}$

Subs (2) in (1),

$V_s = I_i R_i + \beta G_{mf} V_i$

$= I_i R_i + \beta G_{mf} I_i R_i$

$= I_i R_i (1 + \beta G_{mf})$

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

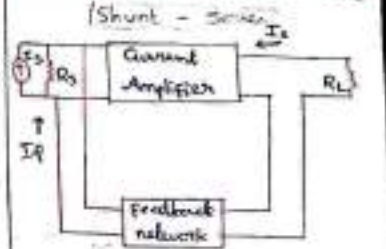
$= R_i (1 + \beta G_{mf})$

$$R_{if} = R_i (1 + \beta G_{mf})$$

Current Shunt

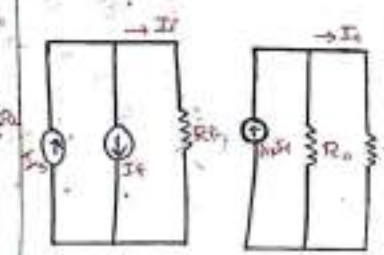
(current mixing current sampling)

(Shunt - Series)



$A = \frac{I_o}{I_f}$

Input Impedance :



$A_I = \frac{I_o}{I_i}$

$I_o = A_I I_i$

Apply KCL to input side,

$I_s = I_i + I_f$

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

$I_o = \frac{A_I I_i R_o}{R_o + R_L}$

$I_o = A_{If} I_i \rightarrow (2)$

$\therefore A_{If} = \frac{A_I R_o}{R_o + R_L}$

Subs (2) in (1),

$I_s = I_i + \beta A_{If} I_i$

$I_s = I_i (1 + \beta A_{If})$

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

$= \frac{V_i}{I_i (1 + \beta A_{If})}$

$R_i = \frac{V_i}{I_i}$

$$R_{if} = \frac{R_i}{(1 + \beta A_{If})}$$

For output impedance, input is shed circuited and resistance is measured by looking into the input from output with R_L disconnected.

Apply KVL to output side,

$A_v V_i + I R_o - V_o = 0$ $\quad \beta = \frac{V_f}{V_o}$, $V_f = \beta V_o$

$\therefore I = \frac{V_o - A_v V_i}{R_o} \rightarrow (1)$

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

Subs (2) in (1)

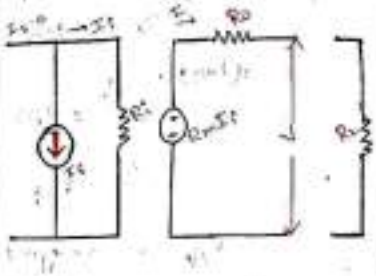
$I = \frac{V_o + A_v \beta V_o}{R_o}$

$I = \frac{V_o (1 + A_v \beta)}{R_o}$

$R_{of} = \frac{V}{I} = \frac{R_o}{1 + A_v \beta}$

This is the output impedance.

Output Impedance :



Apply KVL to output side

$R_m I_f + I R_o - V = 0$

$I = \frac{V - R_m I_f}{R_o} \rightarrow (1)$

$I_f = -I_f = -\beta V \rightarrow (2)$

Subs (2) in (1)

$I = \frac{V + R_m \beta V}{R_o}$

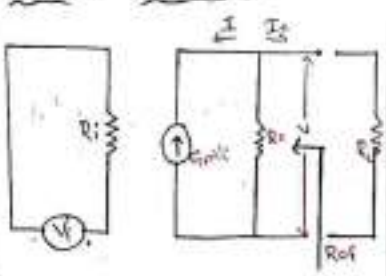
$I = \frac{V (1 + \beta R_m)}{R_o}$

Similarly,

$R_{of} = \frac{V}{I} = \frac{R_o}{1 + \beta R_m}$

This is the output impedance.

Output Impedance



Apply KCL to output circuit

$I = \frac{V}{R_o} - G_m V_i \rightarrow (1)$

Input voltage,

$V_i = -V_f$

$= -\beta I_o$

$= \beta I \rightarrow (2)$

Subs (2) in (1)

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

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

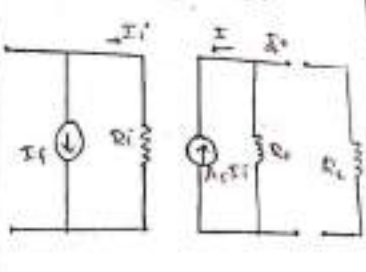
$\frac{V}{R_o} = I (1 + G_m \beta)$

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

$R_{of} = R_o (1 + G_m \beta)$

This is the output impedance.

Output Impedance



$I = \frac{V}{R_o} - A_I I_i \rightarrow (1)$

Input current,

$I_i = -\beta I_o$

$I_i = \beta I \rightarrow (2)$

Subs (2) in (1)

$I = \frac{V}{R_o} - A_I \beta I$

$\frac{V}{R_o} = I + A_I \beta I$

$= I (1 + A_I \beta)$

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

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

$R_{of} = R_o (1 + \beta A_I)$

This is the output impedance.

MUST STUDY

Unit - IV Oscillators

Concepts

→ The circuit which is used to generate periodic voltage without a.c input signal is called an oscillator

→ The circuit is supplied with energy from a d.c source

→ If the output is sine wave function of time the oscillator is called sinusoidal or harmonic oscillator

Classification of Oscillators

① According to wave form

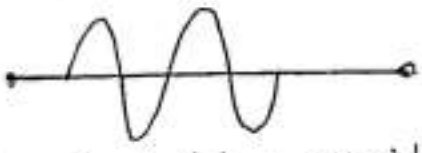
i) Sinusoidal oscillator

ii) Relaxation oscillator

Sinusoidal oscillator - generates sinusoidal voltages or current

Relaxation oscillator - generates voltage or current which vary abruptly one or more times in a cycle of oscillation

Wave form sinsoidal



Waveform for relaxation oscillator

② According to fundamental mechanism

i) negative resistance oscillator uses negative resistance of amplifying device to neutralize positive resistance of oscillation

ii) Feedback oscillator → uses positive feedback in feedback amplifier to satisfy Barkhausen criterion

③. According to frequency generated

i) Audio Frequency Oscillator – upto 20 KHz

ii) Radio Frequency oscillator – 20 KHz to 30 MHz

iii) VHF – 30 MHz to 300 MHz – Very High frequency

iv) UHF – 300 MHz to 3 GHz – Ultra High frequency

v) Microwave frequency – above 3 GHz

④ According to type of circuit used

a) LC tuned oscillator b) RC phase shift oscillator

Barkhausen criterion

→ The oscillator circuit is set to oscillation by random variation caused in base current due to noise component

→ The noise components (small random electrical voltages and currents are present in any conductor, tube or transistor

→ Even when no external signal is applied, noise will cause some signal at output of amplifier

→ amplifier tuned at particular frequency the output signal caused by noise is predominant at f_0.

→ If amplifier has gain of more 1/p, output increases and feedback signal becomes larger and goes on increasing

→ As signal level increases, gain decreases

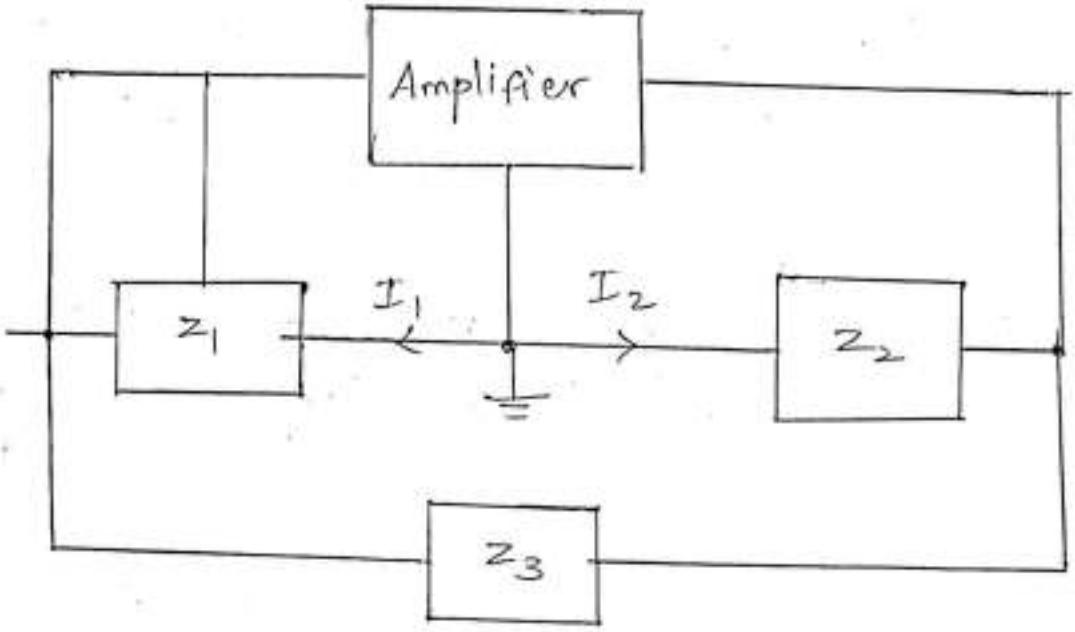
→ The gain is decreased at particular value of output. gain

→ The output voltage remains constant at f_0 called (frequency of oscillation)

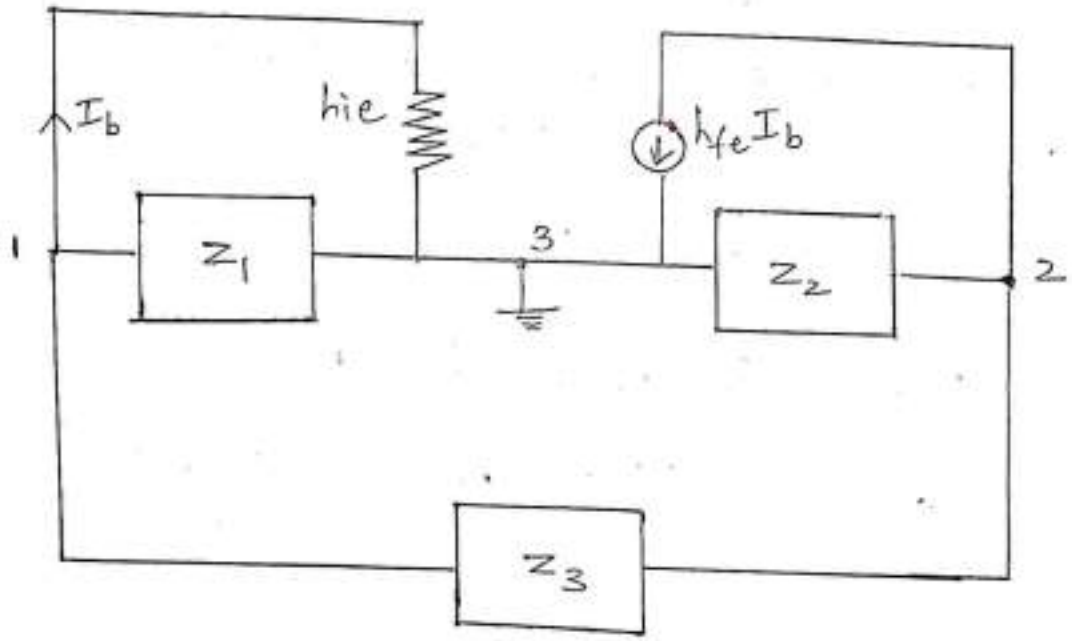
conditions for maintaining oscillation

1. $|A\beta| = 1$, loop gain must be unity
2. The total phase shift around closed loop is 0° or 360°

General form of an LC oscillator



Equivalent circuit



Z_1, Z_2, Z_3 → reactive elements inductor or capacitor of feedback tank circuit determines frequency of oscillation

Z_1, Z_2 → acts as voltage divider for output voltage and feedback signal

→ Voltage across Z_1 is feedback signal

Frequency of oscillation of LC oscillator is

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

Load impedance

Z_1 and hie are parallel.

→ Their equivalent impedance $Z' = Z_1 \parallel hie$

$$Z' = \frac{Z_1\, hie}{Z_1 + hie}$$

→ The load impedance Z_L between output terminals 2 and 3 is the impedance of Z_2 parallel with $Z' + Z_3$

$$\therefore Z_L = Z_2 \parallel Z' + Z_3 = \frac{Z_2(Z' + Z_3)}{Z_2 + Z' + Z_3} = \frac{Z_2 Z' + Z_2 Z_3}{Z_2 + Z' + Z_3}$$

$$= \frac{Z_2\left(\frac{Z_1\, hie}{Z_1 + hie}\right) + Z_2 Z_3}{Z_2 + \left(\frac{Z_1\, hie}{Z_1 + hie}\right) + Z_3} = \frac{Z_1 Z_2 hie + Z_1 Z_2 Z_3 + Z_2 Z_3 hie}{Z_2 Z_1 + Z_2 hie + Z_1 Z_3 + Z_3 hie + Z_1 hie}$$

$$= \frac{hie\,(Z_1 * Z_2 + Z_2 Z_3) + Z_1 Z_2 Z_3}{hie\,(Z_1 + Z_2 + Z_3) + Z_1 Z_2 + Z_1 Z_3} = \frac{hie\, Z_2\left[(Z_1 + Z_3) + Z_1 Z_3\right]}{hie(Z_1 + Z_2 + Z_3) + Z_1 Z_2 + Z_1 Z_3}$$

Voltage gain without feed back

$$A_{ve} = \frac{-h_{fe} Z_L}{hie}$$

Feed back fraction β

Output voltage in terms of current I_1 between terminals 3 & 1

$$V_o = -I_1 (Z' + Z_3) = -I_1 \left(\frac{Z_1 hie}{Z_1 + hie} + Z_3 \right)$$

$$= -I_1 \left(\frac{hie(Z_1 + Z_3) + Z_1 Z_3}{Z_1 + hie} \right)$$

Voltage feedback to input terminals 3 & 1 is given by

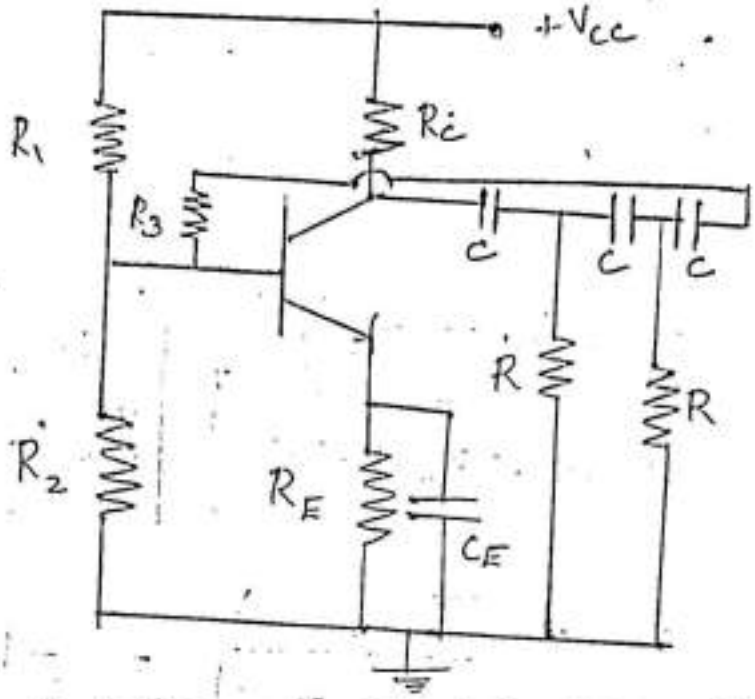
$$V_{fb} = -I_1 Z' = I_1 \left(\frac{Z_1 hie}{Z_1 + hie} \right)$$

$$\beta = \frac{V_{fb}}{V_o} = I_1 \left(\frac{Z_1 hie}{Z_1 + hie} \right) \left(\frac{Z_1 + hie}{hie(Z_1 + Z_3) + Z_1 Z_3} \right) I_1$$

$$\beta = \frac{Z_1 hie}{hie(Z_1 + Z_3) + Z_1 Z_3}$$

RC - Phase shift Oscillator

→ RC oscillator operate at low frequency



→ Common Emitter amplifier is followed by 3 RC section
→ The output of the last Rc network is fed to input
→ To make 3 RC sections identical, R_3 is chosen as $R_3 = R - R_i$

R_i — input impedance

→ Phase shift is $\phi = \tan^{-1}\left(\frac{1}{\omega CR}\right)$

→ The Resistor 'R' is adjusted to give $\phi = 60°$

→ For a given frequency fo the RC ladder network produces a total phase shift of 180° between input and output

→ The total phase shift from base of transistor around circuit to base is 360°

Equation for oscillation

$$A_{ve}\beta = 1$$

$$\left(\frac{-h_{fe} z_1}{h_{ie}}\right)\left[\frac{z_1 h_{ie}}{h_{ie}(z_1+z_3)+z_1 z_3}\right] = 1$$

$$\left\{\frac{h_{fe} z_2 \left[\cancel{h_{ie}(z_1+z_3)+z_1 z_3}\right]}{h_{ie}(z_1+z_2+z_3)+z_1 z_2+z_1 z_3}\right\}\left[\frac{z_1}{\cancel{h_{ie}(z_1+z_3)+z_1 z_3}}\right] = -1$$

$$\left(\frac{h_{fe} z_2 z_1}{h_{ie}(z_1+z_2+z_3)+z_1 z_2+z_1 z_3}\right) = -1$$

$$h_{ie}(z_1+z_2+z_3)+z_1 z_2+z_1 z_3 = -h_{fe} z_1 z_2$$

$$\boxed{h_{ie}(z_1+z_2+z_3)+z_1 z_2(1+h_{fe})+z_1 z_3 = 0}$$

is the general equation for oscillator

The frequency of oscillation

$$f_0 = \frac{1}{2\pi RC\sqrt{6}}$$

→ At this frequency the feedback factor $|\beta| = 1/29$

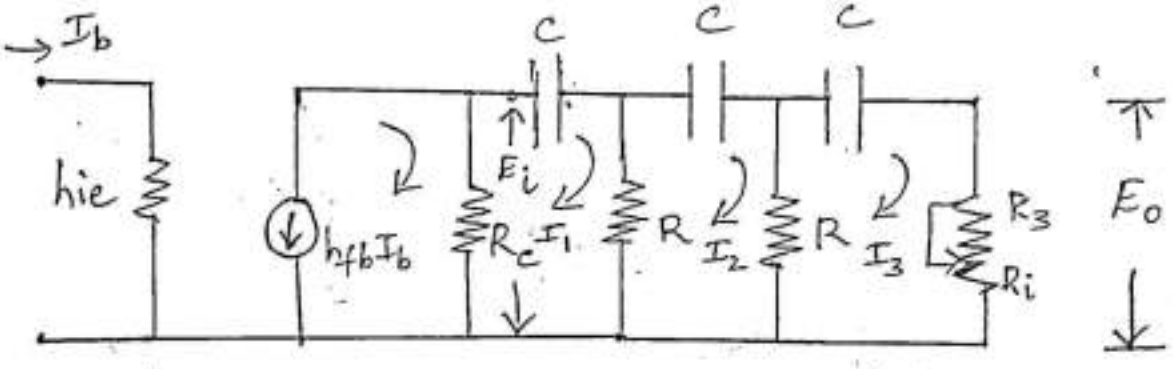
$A\beta > 1$, $|A| > 29$ for operation

→ It is suitable for audio frequency only

Draw back

1. the 3 RC network must be changed simultaneously to change frequency.
2. It is difficult to control amplitude of oscillation without affecting frequency of oscillation.

Analysis



From the fig.

loop1 : $R(I_1 - I_2) + \frac{I_1}{j\omega C} = E_i$

$$I_1\left(R + \frac{1}{j\omega C}\right) - I_2 R = E_i$$

loop 2: $R(I_2 - I_1) + \frac{I_2}{jwC} + R(I_2 - I_3) = 0$

$$-I_1 R + I_2\left(2R + \frac{1}{jwC}\right) - I_3 R = 0$$

loop 3: $R(I_3 - I_2) + \frac{I_3}{jwC} + (R_3 + R_i) I_3 = 0$

$$-I_2 R + I_3\left(2R + \frac{1}{jwC}\right) = 0$$

Replacing jw by s & writing equation

$$\begin{bmatrix} R + \frac{1}{jwC} & -R & 0 \\ -R & 2R + \frac{1}{jwC} & -R \\ 0 & -R & 2R + \frac{1}{jwC} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} = \begin{bmatrix} E_i \\ 0 \\ 0 \end{bmatrix}$$

Applying cramers rule

$$I_3 = \frac{\Delta_3}{\Delta} \quad \text{---} \quad (1)$$

$$\Delta_3 = \begin{bmatrix} R + \frac{1}{jwC} & -R & E_i \\ -R & 2R + \frac{1}{jwC} & 0 \\ 0 & -R & 0 \end{bmatrix} = E_i R^2 \quad \text{---} \quad (2)$$

$$\Delta = \begin{bmatrix} R+\frac{1}{jwc} & -R & 0 \\ -R & 2R+\frac{1}{jwc} & -R \\ 0 & -R & 2R+\frac{1}{jwc} \end{bmatrix}$$

$$= \left(R+\frac{1}{jwc}\right)\left[\left(2R+\frac{1}{jwc}\right)^2 - R^2\right] - R^2\left(2R+\frac{1}{jwc}\right)$$

$$= \frac{(jwcR+1)(2Rjwc+1)^2}{j^3w^3c^3} - \frac{R^2(jwcR+1)}{jwc} - \frac{\left[R^2(2jwcR+1)\right]}{jwc}$$

$$= \left[\frac{(jwcR+1)(-4R^2w^2c^2+1+4Rjwc)}{j^3w^3c^3}\right] - \left[\frac{jwcR^3+R^2+2jwcR^3+R^2}{jwc}\right]$$

$$= \left[\frac{-4jR^3w^3c^3 - 4R^2w^2c^2 + jwcR + 1 - 4R^2w^2c^2 + 4Rjwc}{j^3w^3c^3}\right] - \left[\frac{3jwcR^3+2R^2}{jwc}\right]$$

$$= \frac{-4jR^3w^3c^3 - 4R^2w^2c^2 + jwcR + 1 - 4R^2w^2c^2 + 4Rjwc + 3jwcR^3 - 2j^2R^2w^2c^2}{j^3w^3c^3}$$

$$= \frac{(-4jR^3w^3c^3 - 8R^2w^2c^2 + 5jwcR + 1 + 3jw^3c^3R^3 + 2R^2w^2c^2)}{j^3w^3c^3}$$

$$= (-jR^3w^3c^3 - 6R^2w^2c^2 + 5jwcR + 1) / j^3w^3c^3$$

$$= \left(R^3\omega^3C^3 - 6jR^2\omega^2C^2 - 5\omega CR + j\right) / \omega^3C^3$$

Replace ωCR by $1/\alpha$

$$= \left(\frac{1}{\alpha^3} - \frac{6j}{\alpha^2} - \frac{5}{\alpha} + j\right) / \left(\frac{1}{\alpha R}\right)^3$$

$$= \left(1 - 6j\alpha - 5\alpha^2 + j\alpha^3\right) / \left(1/R^3\right)$$

$$\Delta = R^3\left[\left(1 - 5\alpha^2\right) + j\alpha\left(\alpha^2 - 6\right)\right] \quad \text{—③}$$

Sub ② and ③ in ①

$$I_3 = \frac{E_i R^2}{R^3\left[\left(1 - 5\alpha^2\right) + j\alpha\left(\alpha^2 - 6\right)\right]}$$

$$I_3 = \frac{E_i}{R} \cdot \frac{1}{\left(1 - 5\alpha^2\right) + j\alpha\left(\alpha^2 - 6\right)}$$

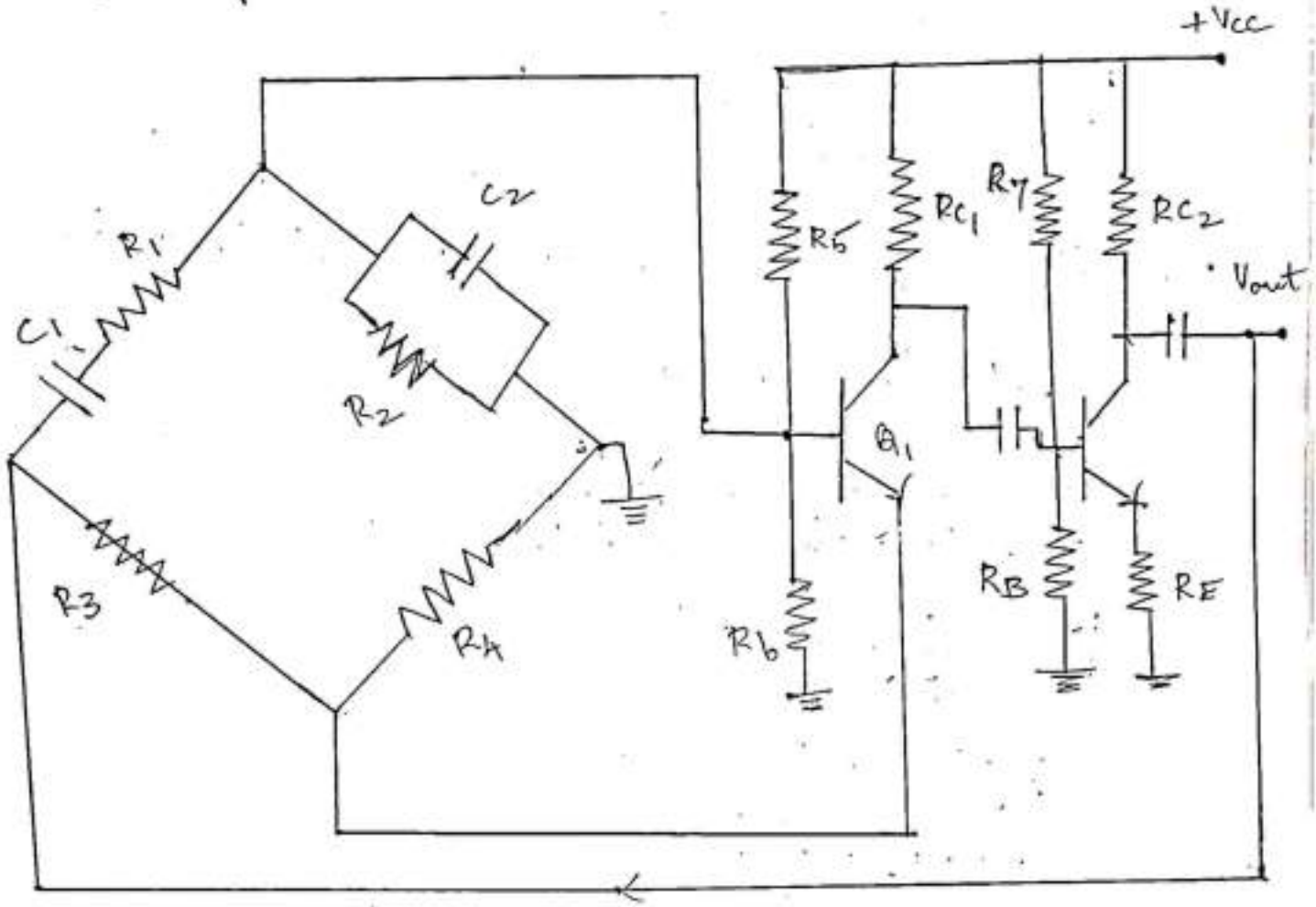
$$E_0 = I_3 R = E_i \times \frac{1}{\left(1 - 5\alpha^2\right) + j\alpha\left(\alpha^2 - 6\right)}$$

$$\beta = \frac{E_0}{E_i} = \frac{1}{1 - 5\alpha^2 + j\alpha\left(\alpha^2 - 6\right)} \quad \text{—④}$$

frequency of oscillation is determined by equating imaginary part to 0

$$\alpha\left(\alpha^2 - 6\right) = 0$$

Wein Bridge oscillator



→ Consists of two stage RC coupled amplifier provides phase shift of 360°

→ Balanced bridge is used as feedback network and provides no phase shift

→ The feedback network consists of lead-lag network R_1-C_1 and R_2-C_2 and voltage divider R_3-R_4

$$\alpha = \frac{1}{\omega_r CR} = \sqrt{6}$$

$$f_r = \frac{1}{2\pi RC\sqrt{6}}$$

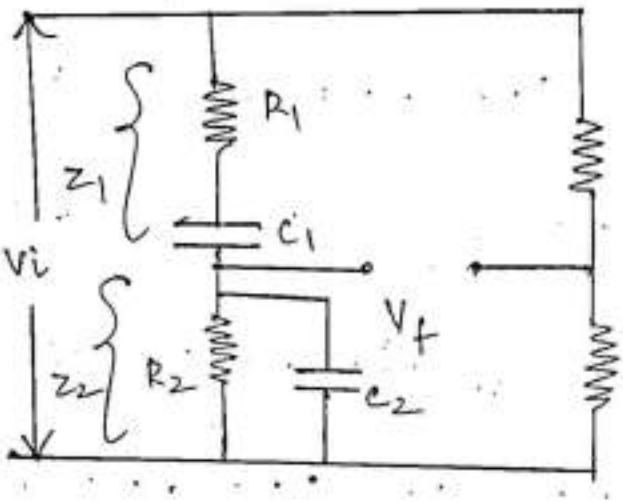
condition for oscillation is obtained by sub $\alpha = \sqrt{6}$ in (4)

$$\beta = \frac{1}{1-5\times 6} = \frac{-1}{29}$$

$$|\beta| = \frac{1}{29} \angle 180^\circ$$

$$|A\beta| = 1$$

$$\boxed{|A| > 29}$$



$V_f = V_1 - V_2$

$$V_1 = \frac{Z_2}{Z_1+Z_2} \times V_i \quad ; \quad V_2 = \frac{R_4}{R_3+R_4} \times V_i$$

$$V_f = \left[\left(\frac{Z_2}{Z_1+Z_2}\right) - \left(\frac{R_4}{R_3+R_4}\right)\right] V_i$$

$$Z_1 = R_1 + \frac{1}{jwC_1} = R_1 + \frac{1}{sC_1} = \frac{1+sC_1R_1}{sC_1}$$

$$Z_2 = R_2 \| \frac{1}{jwC_2} = R_2 \| \frac{1}{sC_2} = \frac{R_2 \times \frac{1}{sC_2}}{R_2 + \frac{1}{sC_2}} = \frac{R_2}{sC_2R_2+1}$$

$$\frac{Z_2}{Z_1+Z_2} = \frac{R_2/1+sC_2R_2}{\frac{1+sC_1R_1}{sC_1} + \frac{R_2}{sC_2R_2+1}} = \frac{R_2 \times sC_1}{(1+sC_2R_2)(1+sC_1R_1)+R_2sC_1}$$

$$= \frac{sC_1R_2}{1+sC_1R_1+sC_2R_2+s^2C_1C_2R_1R_2+R_2sC_1}$$

Sub $s = jw$

$$= \frac{jw\,C_1R_2}{(1-w^2C_1C_2R_1R_2)+jw(C_1R_1+C_2R_2+C_1R_2)}$$

$$= \frac{jwC_1R_2\left[1-w^2C_1C_2R_1R_2-jw(C_1R_1+C_2R_2+C_1R_2)\right]}{(1-w^2C_1C_2R_1R_2)^2+w^2(C_1R_1+C_2R_2+C_1R_2)^2}$$

$$= \frac{jwC_1R_2(1-w^2C_1C_2R_1R_2)+w^2(C_1R_1+C_2R_2+C_1R_2)}{(1-w^2C_1C_2R_1R_2)^2+w^2(C_1R_1+C_2R_2+C_1R_2)^2}$$

$$V_f = \left[\frac{w^2(C_1R_1+C_2R_2+C_1R_2)}{(1-w^2C_1C_2R_1R_2)^2+w^2(C_1R_1+C_2R_2+C_1R_2)^2} + \frac{jwC_1R_2(1-w^2C_1C_2R_1R_2)}{(1-w^2C_1C_2R_1R_2)^2+w^2(C_1R_1+C_2R_2+C_1R_2)^2} - \frac{R_3}{R_3+R_4}\right]V_i$$

Equating imaginary part to 0

$$jwCR_2(1-w^2C_1C_2R_1R_2)=0$$

$$w^2C_1C_2R_1R_2=1$$

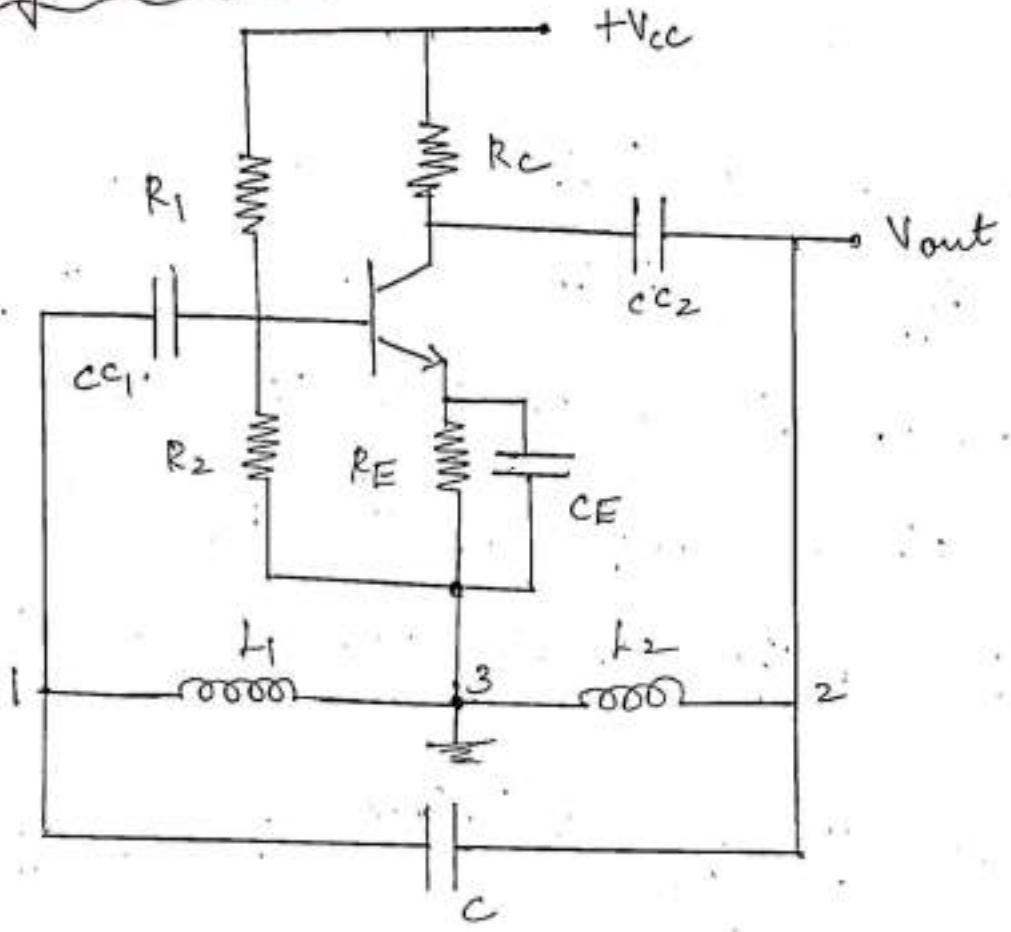
$$w=\frac{1}{\sqrt{C_1C_2R_1R_2}}$$

$$f_r=\frac{1}{2\pi\sqrt{C_1C_2R_1R_2}}$$

$$\boxed{f_r=\frac{1}{2\pi RC}} \qquad \left[\because\ R=R_1=R_2,\ C=C_1=C_2\right]$$

$R_3/R_4 > 2$ → provide sufficient gain for the circuit to osillate at desired frequency

Hartley Oscillator



→ Z_1, Z_2 → inductors, Z_3 → capacitor

→ R_1, R_2 and R_E → provides dc bias to transistor

→ C_E is bypass capacitor

→ CC_1, CC_2 are coupling capacitor

→ The feedback network consists of L_1, L_2 and C which determines the frequency of oscillator

→ when V_{cc} is ON transient current is produced in tank circuit and damped harmonic oscillation are set up in circuit

→ oscillatory current in tank circuit produces ac voltages across L_1 and L_2

→ If 3 is ground, 1 is +ve, 2 is -ve

→ Phase difference between 1 & 2 is always 180°

→ In CE, transistor provides 180° phase shift

→ Total phase shift is 180° + 180° = 360°

→ If $A\beta = 1$, the circuit acts as oscillator

frequency of oscillation $f_0 = \frac{1}{2\pi\sqrt{LC}}$

where $L = L_1 + L_2 + 2M$

M → mutual inductance between coils L_1 & L_2

→ condition for sustained oscillation

$$h_{fe} \geq \frac{L_1 + M}{L_2 + M}$$

Analysis.

$$Z_1 = j\omega L_1 + j\omega M$$

$$Z_2 = j\omega L_2 + j\omega M$$

$$Z_3 = \frac{1}{j\omega C} = \frac{-j}{\omega C}$$

Sub the values in general equation of oscillator

we get

$$j\omega hie\left[L_1+L_2+2M-\frac{1}{\omega^2 C}\right]-\omega^2(L_1+M)\left[(L_2+M)(1+hfe)-\frac{1}{\omega^2 C}\right]=0 \quad \text{——— (1)}$$

freq. f_0 is determined by equating imaginary part to 0

$$\left[L_1+L_2+2M-\frac{1}{\omega^2 C}\right]=0$$

$$L_1+L_2+2M=\frac{1}{\omega_0^2 C}$$

$$\frac{1}{\omega_0^2}=C\left[L_1+L_2+2M\right]$$

$$\omega_0^2=\frac{1}{C\left[L_1+L_2+2M\right]}$$

$$\frac{\omega_0}{2\pi}=\frac{1}{2\pi\sqrt{(L_1+L_2+2M)C}}=f_r \quad \text{— (2)}$$

→ condition for maintenance of oscillation is obtained by Subs. (2) in (1)

imaginary part becomes 0 & hence

$$(L_2+M)(1+h_{fe}) - \frac{1}{\omega_0^2 C} = 0 \quad \text{— (3)}$$

Sub (2) in above equn. we get

$$\therefore \frac{1}{\omega_0^2 C} = (L_2+M)(1+h_{fe})$$

$$L_1+L_2+2M = (L_2+M)(1+h_{fe})$$

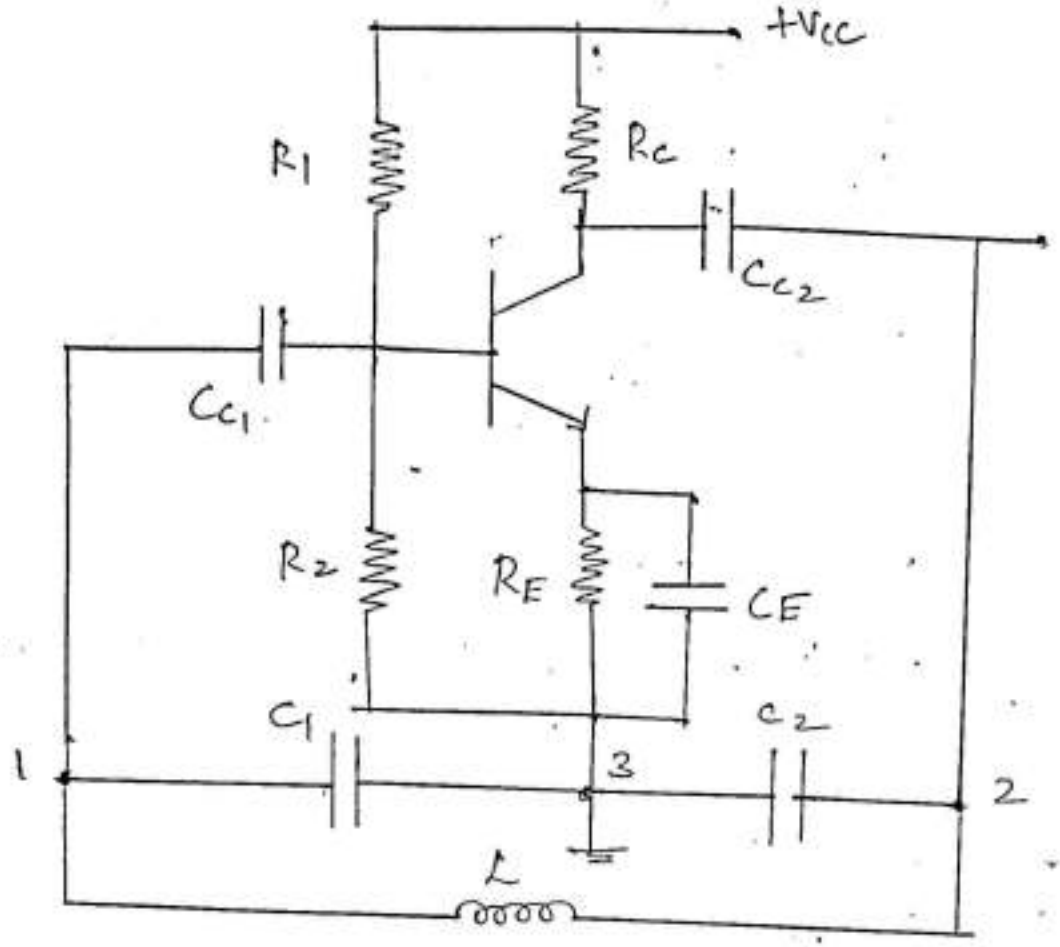
$$L_1+L_2+2M = L_2 + L_2 h_{fe} + M + M h_{fe}$$

$$L_1+L_2+2M - L_2 - M = h_{fe}(L_2+M)$$

$$L_1+M = h_{fe}(L_2+M)$$

$$\boxed{h_{fe} = \frac{L_1+M}{L_2+M}}$$

Collpitts Oscillator



Z_1, Z_2 → are capacitors

Z_3 is inductor

CC_1, CC_2 are coupling capacitors

→ feedback network C_1, C_2, L determines frequency of oscillation

→ when V_{cc} is switched ON, transient current is produced in tank circuit

→ Oscillatory current in tank circuit produces a.c voltages across C_1 & C_2

→ 3 is earth, terminal 1 is positive and terminal 2 is negative

→ Phase difference between 1 & 2 provides 180° and transistor provides 180° phase difference

→ Total phase difference is $180° + 180° = 360°$

→ The necessary condition for sustained oscillation is satisfied

→ The feedback is adjusted so that loop gain

$A\beta = 1$

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

$$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} \qquad \text{(or)} \quad C = \frac{C_1 C_2}{C_1 + C_2}$$

Analysis

$$Z_1 = \frac{1}{jwC_1} = \frac{-j}{wC_1}$$

$$Z_2 = \frac{1}{jwC_2} = \frac{-j}{wC_2}$$

$$Z_3 = jwL$$

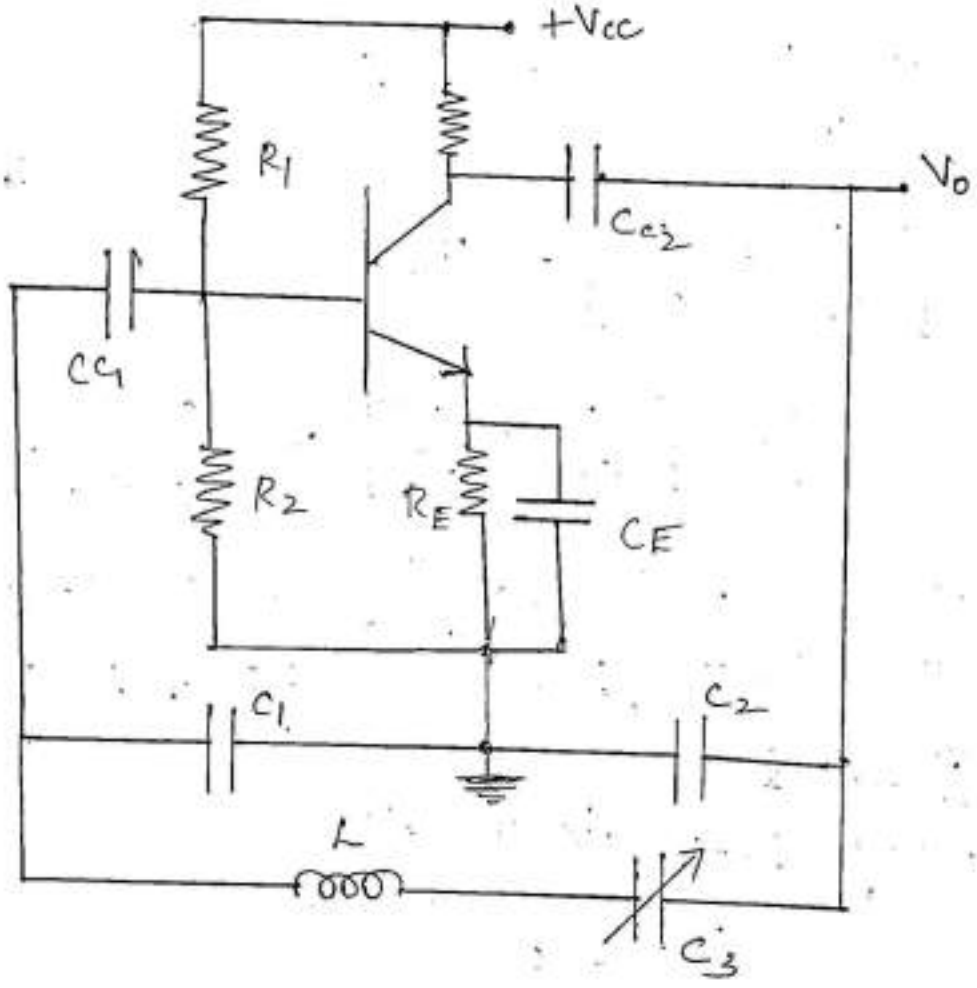
Sub. these in standard equation

$$h_{fe} C_1 = C_2$$

$$h_{fe} = C_2/C_1$$

$$\boxed{h_{fe} = \frac{C_2}{C_1}}$$

clapp oscillator



Z_1 & Z_2 are capacitors

Z_3 is series combination of L and C_3

C_3 improves frequency stability

$$j\,hie\left(\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L\right)+\left(\frac{1+h_{fe}}{\omega^2 C_1 C_2}-\frac{L}{C_1}\right)=0 \quad -(2)$$

frequency of oscillation, f_o is found by equating imaginary part to 0

$$f_o=\frac{\omega_0}{2\pi}=\frac{1}{2\pi}$$

$$\frac{1}{\omega_0 C_1}+\frac{1}{\omega_0 C_2}-\omega_0 L=0$$

$$\frac{1}{\omega_0}\left[\frac{1}{C_1}+\frac{1}{C_2}\right]=\omega_0 L$$

$$\omega_0^2=\frac{1}{L}\left[\frac{1}{C_1}+\frac{1}{C_2}\right]=\frac{C_2+C_1}{L C_1 C_2}$$

$$f_o=\frac{\omega_0}{2\pi}=\frac{1}{2\pi}\sqrt{\frac{C_2+C_1}{L C_1 C_2}}=\frac{1}{2\pi}\sqrt{\frac{C_2+C_1}{L C_1 C_2}} \quad -(3)$$

Sub (3) in (2)

imaginary part is 0 so

$$\frac{1+h_{fe}}{\omega_0^2 C_1 C_2}-\frac{L}{C_1}=0$$

$$\frac{1+h_{fe}}{\frac{C_2+C_1}{L C_1 C_2}\times C_1 C_2}=\frac{L}{C_1}$$

$$L(1+h_{fe})C_1=L(C_2+C_1)$$

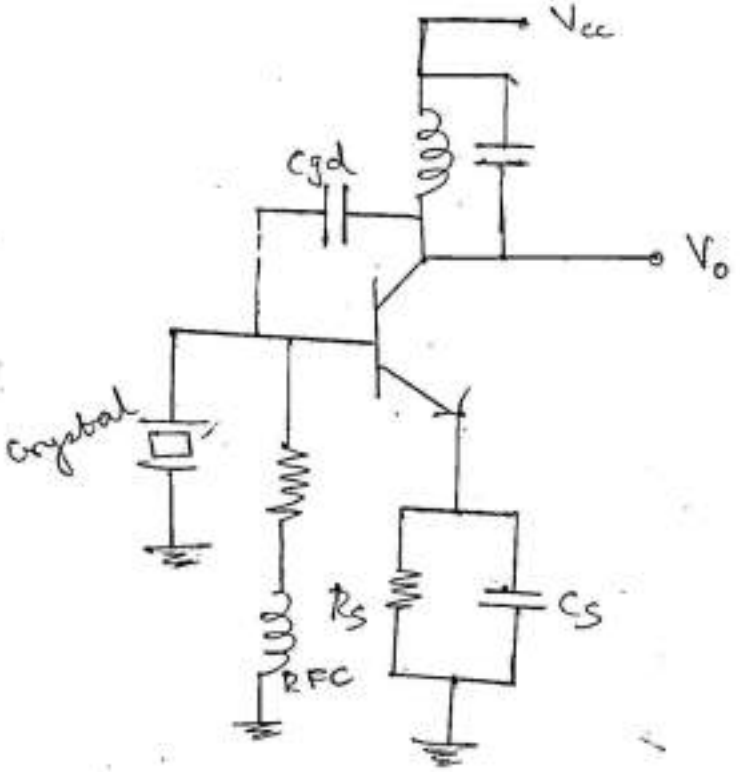
$$C_1+h_{fe}C_1=C_2+C_1$$

Crystal oscillator

→ crystal is thin slice of piezo electric material such as quarty, tourmaline rochelle salt which exhibits piezo electric effect

→ piezo electric effect is that the crystal reacts to any mechanical stress by producing an electric charge in reverse effect. electric field results in mechanical strain

→ high degree of frequency stability is obtained in crystal oscillator



$$\omega L_1 - \frac{1}{\omega C_1} = \frac{1}{\omega C_2} + \frac{1}{\omega C_3}$$

$$\omega L_1 = \frac{1}{\omega C_1} + \frac{1}{\omega C_2} + \frac{1}{\omega C_3}$$

$$\omega^2 L_1 C_1 = 1 + \frac{C_1}{C_2} + \frac{C_1}{C_3}$$

$$f_o = \frac{1}{2\pi\sqrt{L_1 C_1}} \sqrt{1 + \frac{C_1}{C_2} + \frac{C_1}{C_3}}$$

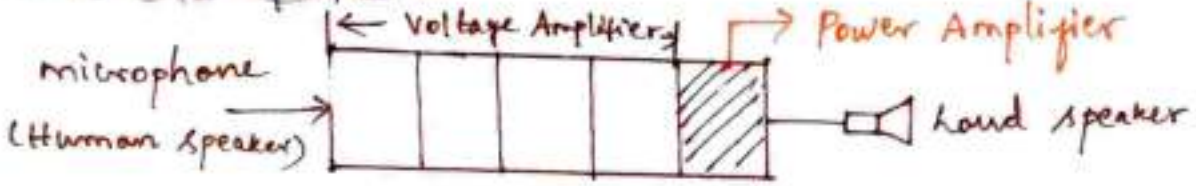
→ It uses a particular combination of an inductor and three capacitors. to set the oscillator frequency

→ the oscillator has positive feedback.

Unit-5

Power Amplifiers and DC convertors

Power Amplifiers



→ In a public address system, there are many stages connected in cascade

→ The input and intermediate stages are small signal amplifiers

→ The last stage gives output to load like loud speaker.

→ A stage which develops sufficient power to the load like loud speaker is called large signal Amplifier or Power Amplifier

Power Amplifiers finds application in public address systems, radio receivers, Industrial control system, tape players, T.V receivers and CRT.

Types

→ the position of Q point on load line decides the class of operation of power amplifier

→ The various classes of power Amplifier are

i) class A ii) class B iii) class C and iv) class AB

class A amplifier

→ In class A amplifier Q point is selected at the centre of load line

→ The output flows for full input cycle

→ The transistor remains in active region for full cycle

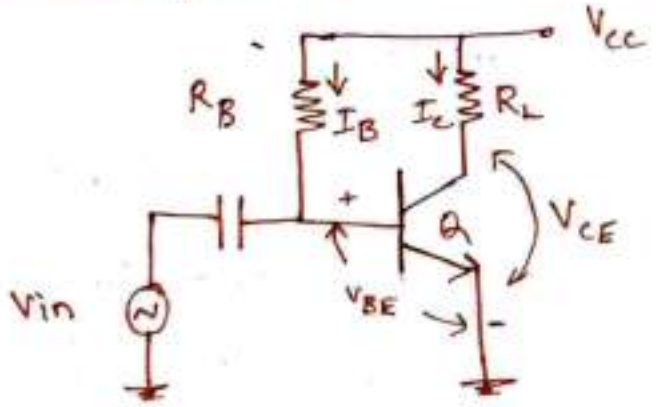
→ The collector current flows for 360°

Types of class A

i) Directly coupled class A amplifier (series fed)

ii) Transformer coupled class A

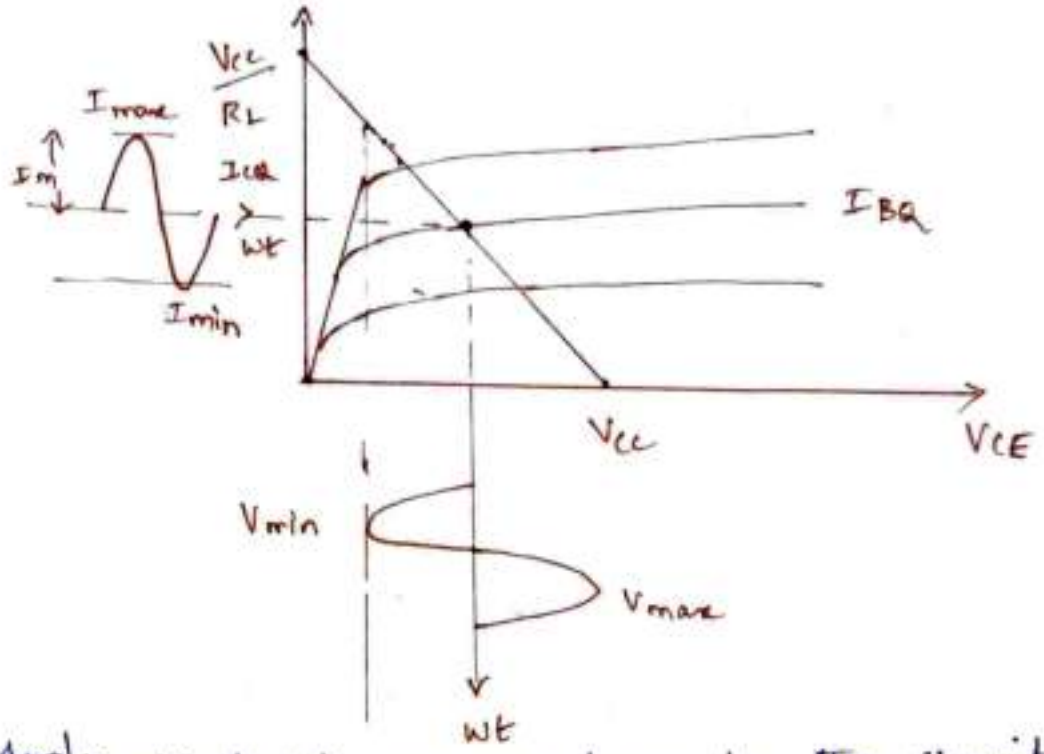
Series fed, Directly coupled class A



→ In directly coupled, the load is directly connected in collector circuit

→ The transistor is used is power transistor

→ The value of R_B is selected such a way that Q pt. lies in centre of d.c load line.



→ Apply Kirchoff's voltage law to the circuit

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

$$I_c R_L = -V_{CE} + V_{cc}$$

$$I_c = \left[-\frac{1}{R_L}\right] V_{CE} + \frac{V_{cc}}{R_L}$$

→ The slope of load line is $-\frac{1}{R_L}$ and y interapt is $\frac{V_{cc}}{R_L}$

DC operation

→ The collector supply voltage V_{cc} and resistance R_B devides dc bias current I_{BQ}

→ Applying KVL to base emitter loop

$$I_{BQ} = \frac{V_{cc} - 0.7}{R_B}$$

$$I_{CQ} = \beta I_{BQ}$$

$$V_{CEQ} = V_{cc} - I_{CQ} R_L$$

DC power input

→ DC power input is

$$P_{DC} = V_{cc} \cdot I_{CQ}$$

AC power output

$$V_m = \frac{V_{PP}}{2} = \frac{V_{max} - V_{min}}{2}$$

V_m → amplitude of a.c output voltage

V_{max} → maximum instantaneous value of collector voltage

V_{min} → minimum instantaneous value of collector voltage

V_{PP} → peak to peak value of a.c output voltage

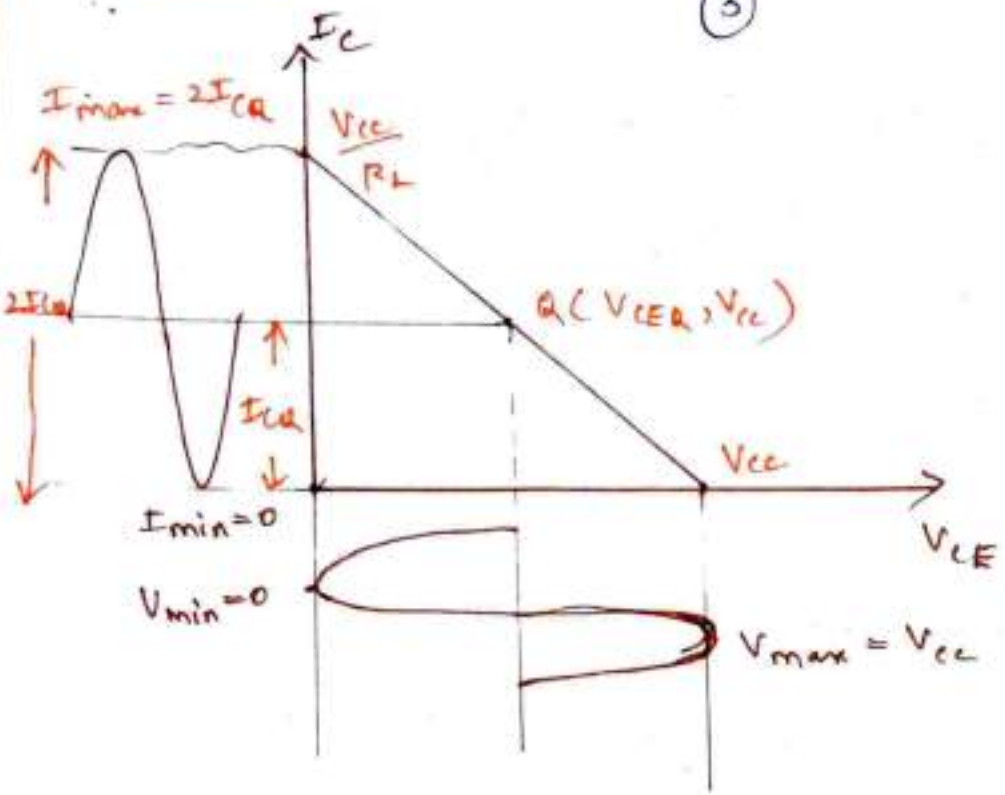
$$P_{ac} = \frac{V_m I_m}{2} \quad \frac{\frac{V_{PP}}{2} \times \frac{I_{PP}}{2}}{2}$$

$$P_{ac} = \frac{V_{PP} \times I_{PP}}{8}$$

$$P_{ac} = \frac{I_{PP}^2 R_L}{8}$$

$$P_{ac} = \frac{V_{PP}^2}{8R_L}$$

$$P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}$$



$V_{max} = V_{cc}$ & $V_{min} = 0$

$I_{max} = 2I_{CQ}$ & $I_{min} = 0$

$$\%\eta = \frac{(V_{cc}-0)(2I_{CQ}-0)}{8 V_{cc} I_{CQ}} \times 100 = \frac{2 V_{cc} I_{CQ}}{8 V_{cc} I_{CQ}} \times 100$$

$$= 25\%$$

Power Dissipation

$$P_d = P_{DC} - P_{ac}$$

→ maximum power dissipation occurs when there is no a.c input signal

$$P_{d\,max} = V_{cc} \times I_{CQ}$$

Advantages:

→ circuit is simple to design
→ load is connected directly no transformer is necessary
→ less no. of components is required

Disadvantages

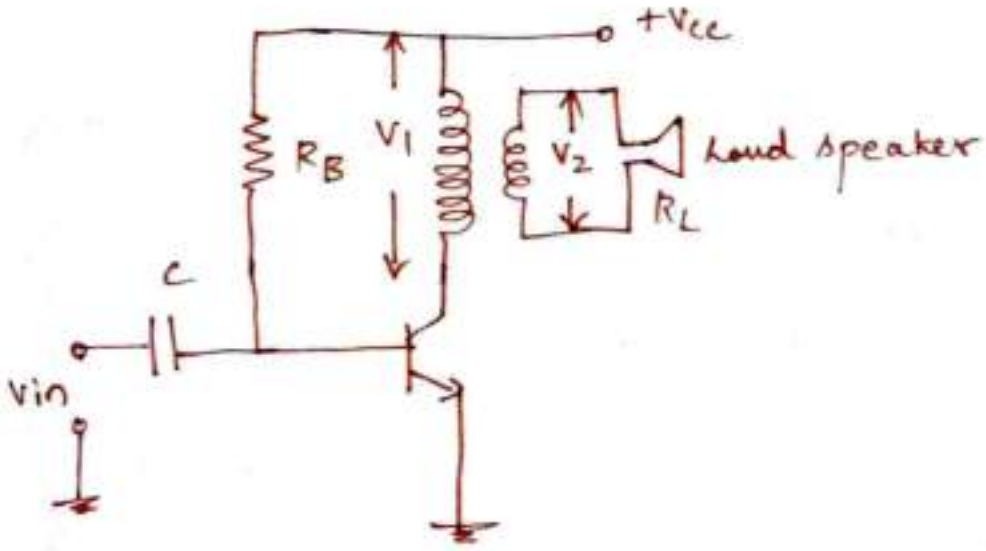
1. Since load resistance is directly connected in collector there is wastage of power
2. Power dissipation is more
3. the output impedance is high
4. The efficiency is very poor.

Transformer coupled class A Amplifier

→ For maximum power transfer to load, impedance matching is necessary

→ the loudspeaker has low output impedance which has to be matched with class A amplifier of high output impedance

→ This is eliminated by using transformer to deliver power to load.



Dc operation

Apply KVL to collector circuit

$$V_{cc} - V_{CE} = 0$$

$$V_{cc} = V_{CE}$$

$$\boxed{V_{CEQ} = V_{cc}}$$

Dc operation power input

$$P_{DC} = V_{cc} I_{CQ}$$

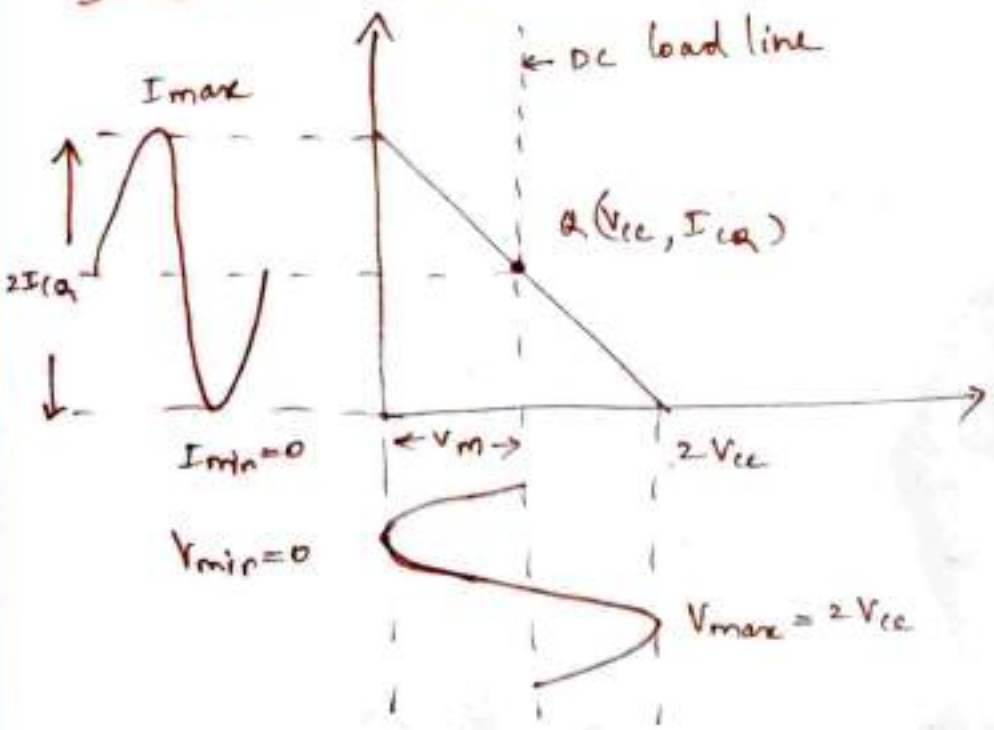
Ac output power

$$P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}$$

Efficiency

$$\% \eta = \frac{P_{ac}}{P_{dc}} \times 100 = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8 V_{cc} I_{CQ}} \times 100$$

maximum Efficiency



$V_{min} = 0$ and $V_{max} = 2V_{cc}$

$I_{min} = 0$ and $I_{max} = 2I_{CQ}$

$$\% \eta_{max} = \frac{(2V_{cc} - 0)(2I_{CQ} - 0)}{8 V_{cc} I_{CQ}} \times 100$$

$$= \frac{4 V_{cc} I_{CQ}}{8 V_{cc} I_{CQ}} \times 100 = 50\%$$

Power Dissipation

$$P_d = P_{DC} - P_{ac}$$

→ when input signal is larger, more power is delivered to load and less is power dissipation

→ when there is no input signal, the entire dc input power gets dissipated in form of heat

$$P_{d\,max} = V_{cc} \times I_{CQ}$$

Advantages:

1. the efficiency is higher
2. Impedance matching is possible

Disadvantages

1. Due to transformer, the circuit is bulkier
2. The circuit is complicated
3. frequency response is poor.

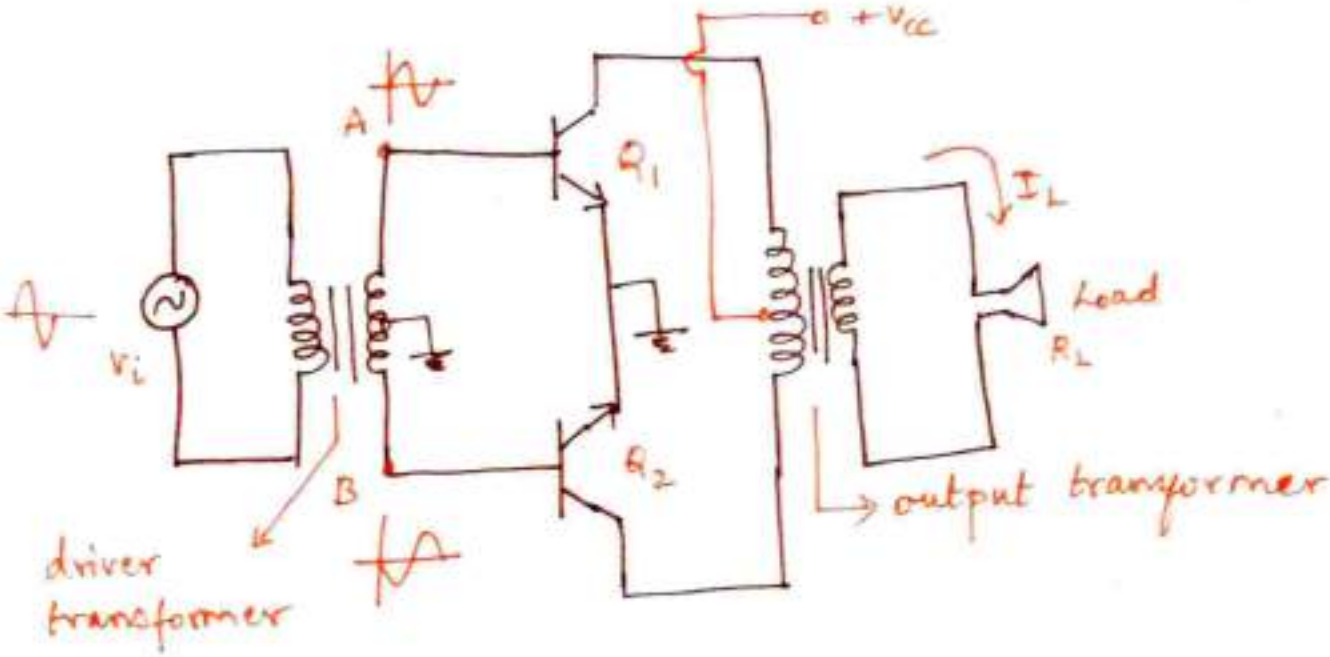
Analysis of class B

→ The Q point is located on x-axis itself.
→ The collector current flows only for half cycle
→ To get full cycle across load, two transistors conduct in alternate half cycle

Two types of class B
1) class B push pull
2) complementary Symmetry class AB

class B push pull

→ It has two transformers, input and output transformer
→ It has two transistors both of same type either npn or pnp.
→ The input signal is applied to primary of driver transformer
→ The centre tap on secondary of driver transformer is grounded
→ The voltage of secondary of driver transformer is equal with opposite polarity
→ The input to base of Q_1 and Q_2 will be 180° out of phase



Operation :

→ for positive half cycle of input A is positive and B is negative

→ Transistor Q_1 conducts and Q_2 is cut off

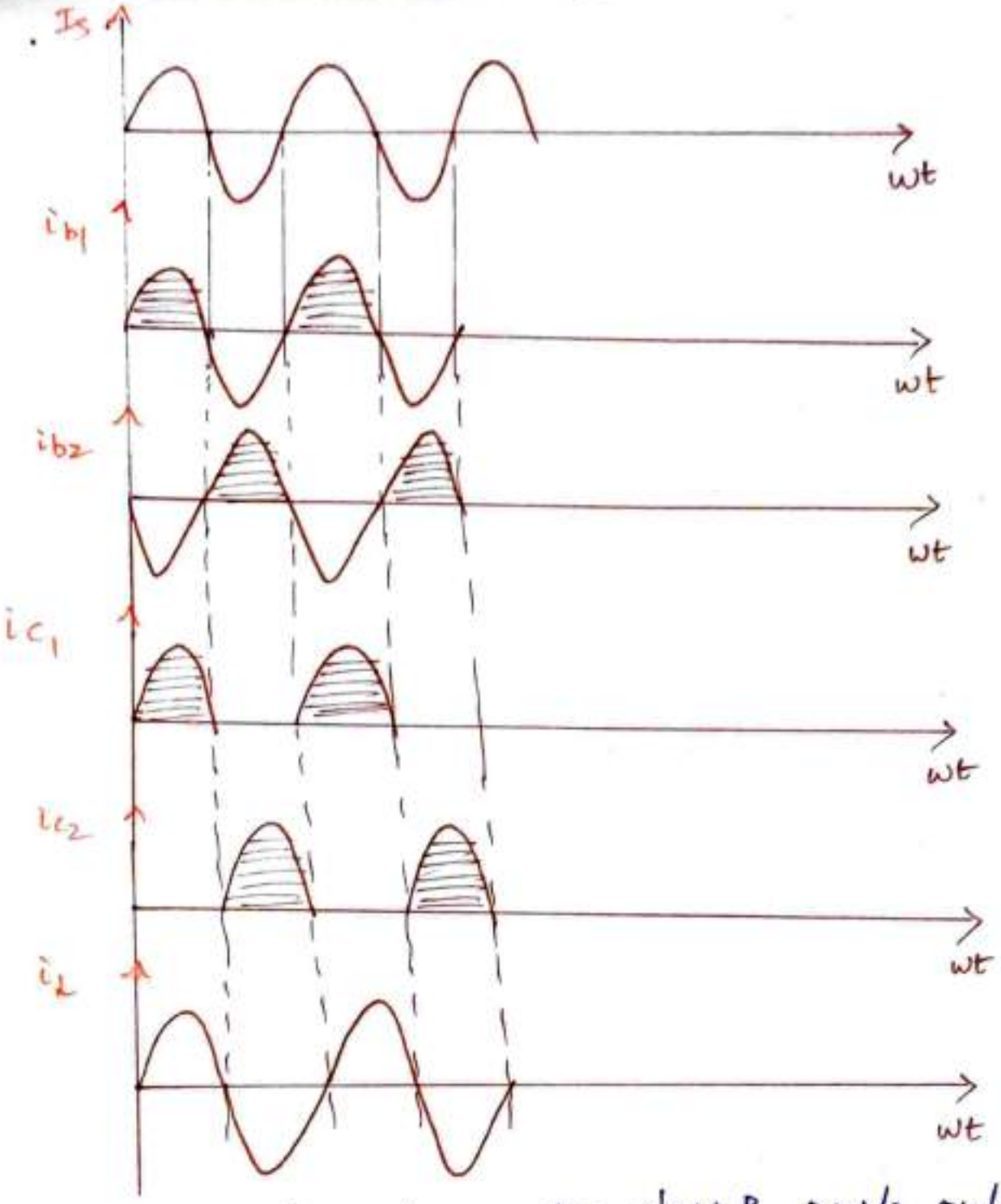
→ So i_{b_1} flows and $i_{b_2} = 0$, and i_{c_1} flows for upper part of primary

→ For negative half cycle, B is positive and A is negative

→ Transistor Q_2 conducts and Q_1 is cutoff.

→ So i_{b_2} flows and $i_{b_1} = 0$, and i_{c_2} flows through lower part of primary

→ Thus full cycle is obtained across the load



Wave form for class B push pull

DC power input

$$P_{dc} = V_{cc} \times I_{dc}$$

$$I_{dc} = \frac{I_m}{\pi} + \frac{I_m}{\pi} = \frac{2I_m}{\pi} \quad (\because \text{two transistor})$$

$$\boxed{P_{dc} = \left(\frac{2I_m}{\pi}\right) \times V_{cc}}$$

Ac power output

$$P_{ac} = \frac{V_m I_m}{2}$$

Efficiency :

$$\% \eta = \frac{P_{ac}}{P_{dc}} \times 100 = \frac{\left(\frac{V_m I_m}{2}\right)}{\frac{2}{\pi} V_{cc} \times I_m} \times 100$$

$$\% \eta = \frac{\pi}{4} \times \frac{V_m}{V_{cc}} \times 100$$

max efficiency :

maximum value of V_m is possible if

$$V_m = V_{cc}$$

$$\% \eta_{max} = \frac{\pi}{4} \times \frac{V_{cc}}{V_{cc}} \times 100 = 78.5\%$$

$$\boxed{\% \eta = 78.5\%}$$

Power dissipation

$V_m = \frac{2}{\pi} V_{cc}$ - for max. power dissipation

$$P_d = P_{dc} - P_{ac} = \frac{4}{\pi^2} \frac{V_{cc}^2}{R_L'} - \frac{2}{\pi^2} \frac{V_{cc}^2}{R_L'}$$

$$P_{d\,max} = \frac{2}{\pi^2} \frac{V_{cc}^2}{R_L'}$$

$$\boxed{P_{d\,max} = \frac{2}{\pi^2} \frac{V_{cc}^2}{R_L'}}$$

Advantages:

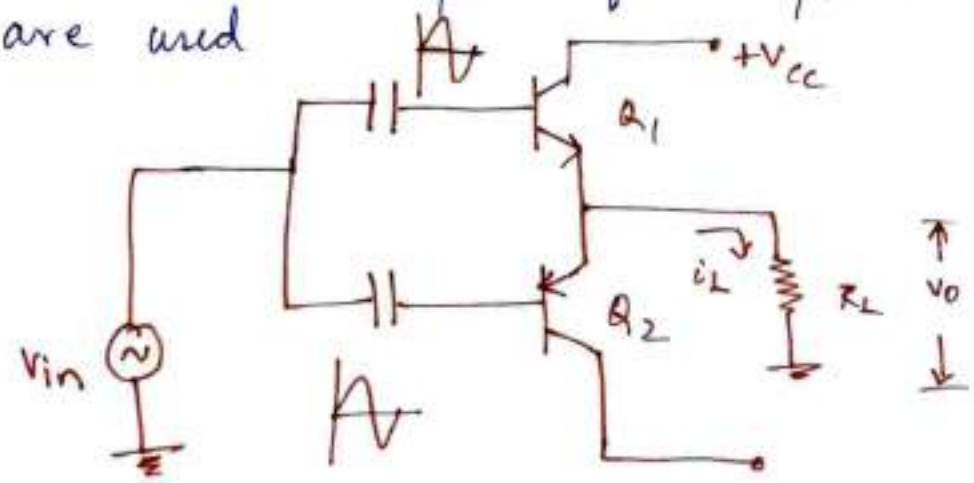
1) Efficiency is much higher than class A
2) Due to transformer impedance matching is possible

Disadv:

1) 2 centre tap transformer are necessary
2) Transformer make the circuit bulky and costlier
3) Frequency response is poor

Complementary symmetry class B Amplifier

→ instead of same type of transistor (npn or pnp) one npn and one pnp is used.
→ so this circuit is transformer less circuit
→ It is diffiudt to match the output impedance
→ Hence matched pair of Complementary transistors are used



→ The circuit is driven from dual supply of $+V_{cc}$

→ Q_1 is npn & Q_2 is pnp

→ During positive half cycle Q_1 is driven to active and Q_2 is cutoff

→ So current flows through R_L

→ During negative half cycle Q_2 conducts and Q_1 is cutoff.

→ Hence Q_2 conducts during negative half cycle of input and current flows thro' R_L.

Analysis is same as class B pushpull

Advantages:

1. As circuit is transformer less, weight, size and cost are less
2. Due to common collector impedance matching is possible

Disadvantage

1. It needs two separate voltage supplies
2. Output is distorted to cross over distortion.

Cross over Distortion

→ For transistors to be in active region the base emitter junction must be forward bias

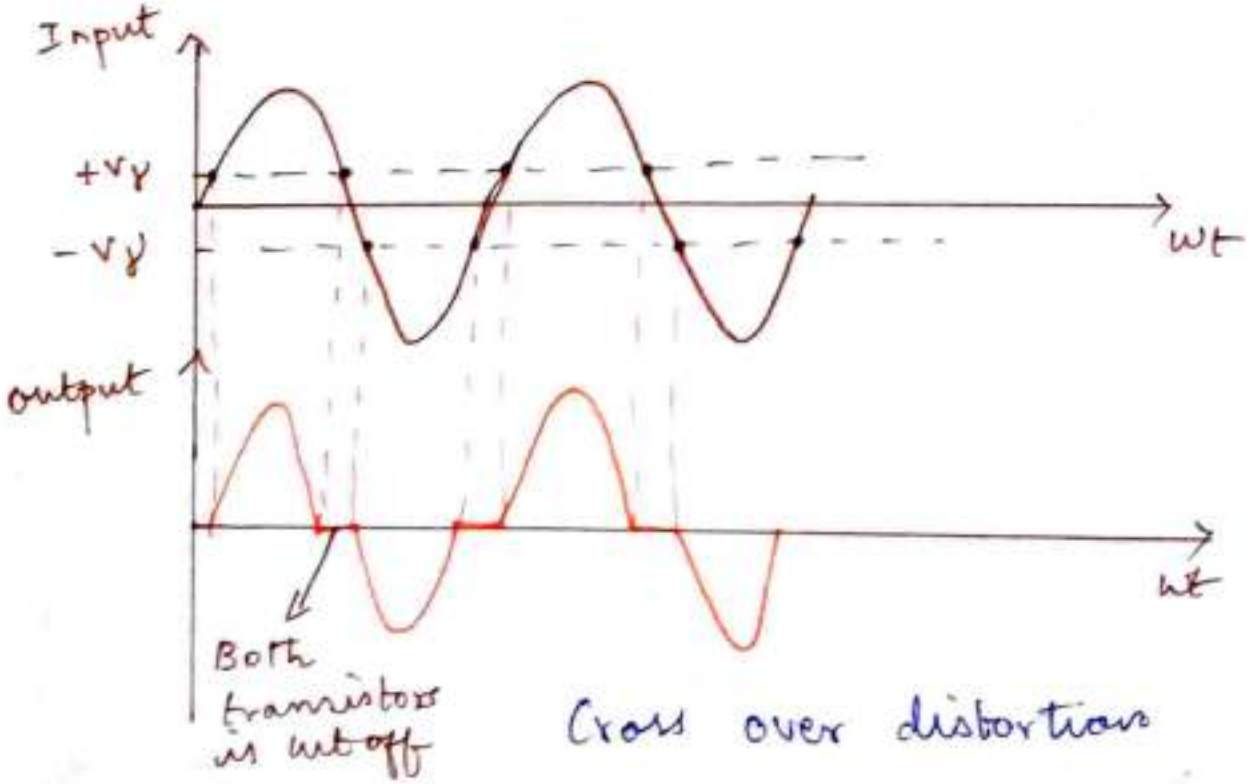
→ the junction is made forward bias till the voltage applied becomes greater than cut in voltage (0.7V) for si

→ when magnitude of input is less than 0.7 the collector current remains zero and transistor remains in cut off.

→ There is a period between crossing of half cycles, none of transistor conducts and output is zero.

→ Hence the nature of output is distorted and do not remains same.

→ Such a distortion is called cross over distortion



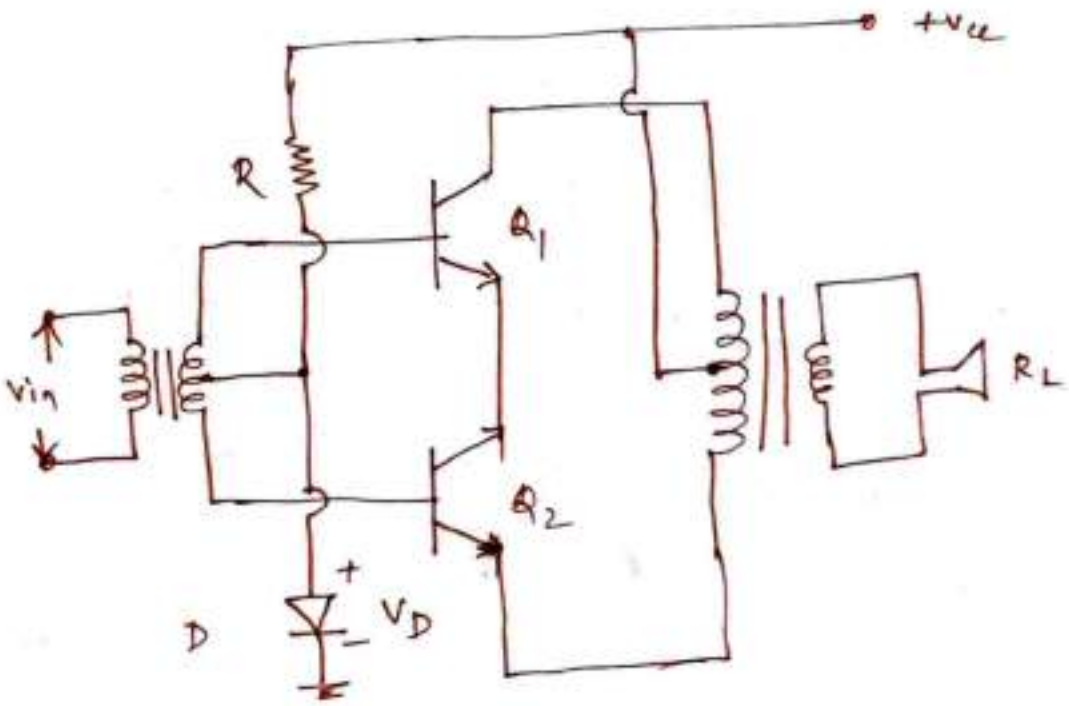
Cross over distortion

Elimination of cross over distortion

To eliminate cross over distortion, a small forward bias is applied to transistor

Push pull class AB

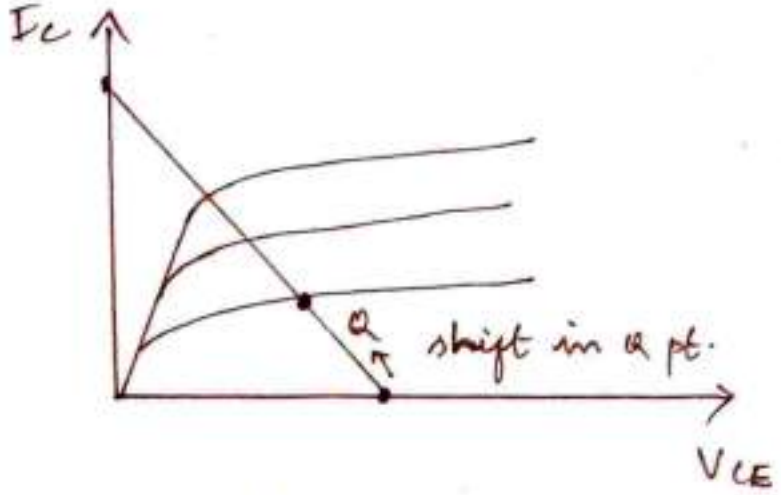
→ The forward bias across the transistor is provided by using diode



→ The drop across diode D is equal to cut in voltage of base emitter junction of transistor

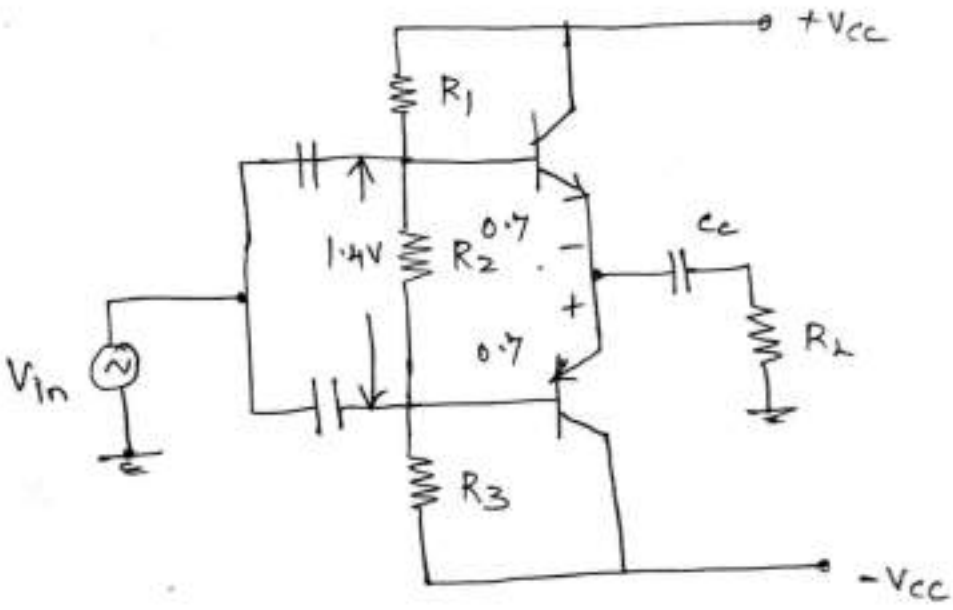
→ Hence both transistor conducts for full cycle

→ The Q point shifts upward and the operation becomes class AB



Complementary Symmetry Class AB

→ In complementary Symmetry, base emitter junction of both Q_1 and Q_2, are required to provide fixed bias



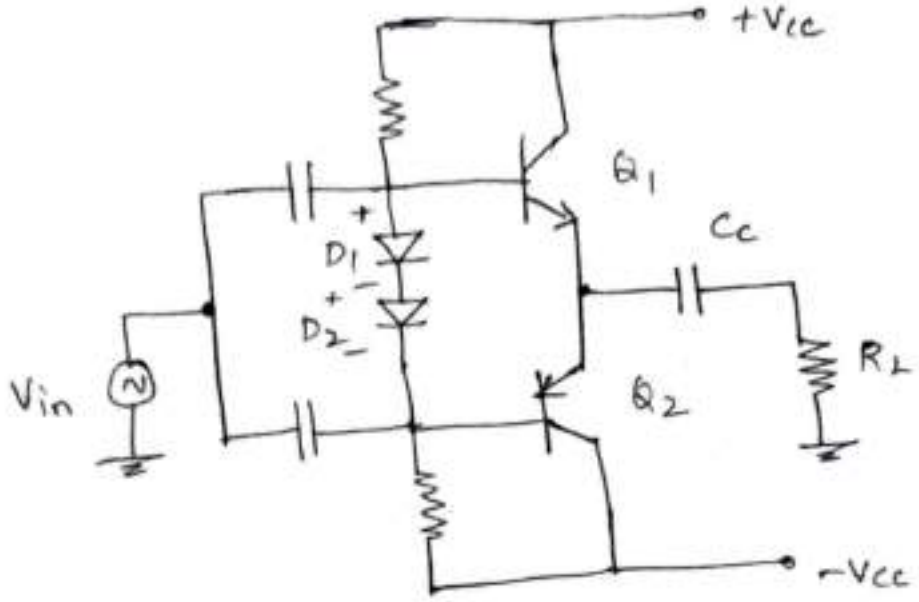
→ For silicon Transistor fixed bias of 0.7+0.7 = 1.4V is required

→ This is achieved by using potential divider arrangement

→ As the junction cut in voltage changes with temperature, there is still possibility of

distortion as temperature changes

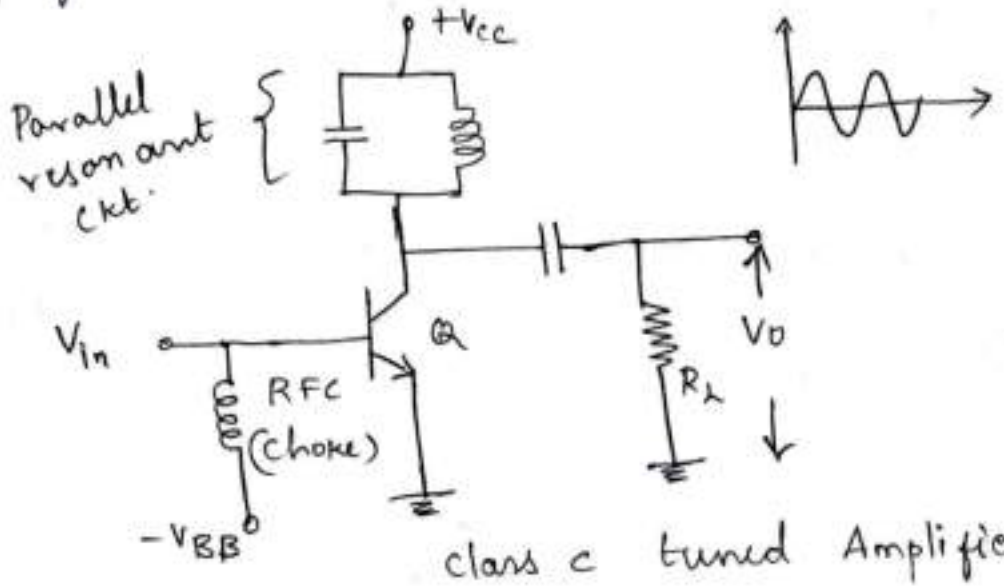
→ Hence instead of R_2, two diodes is used to provide required fixed bias



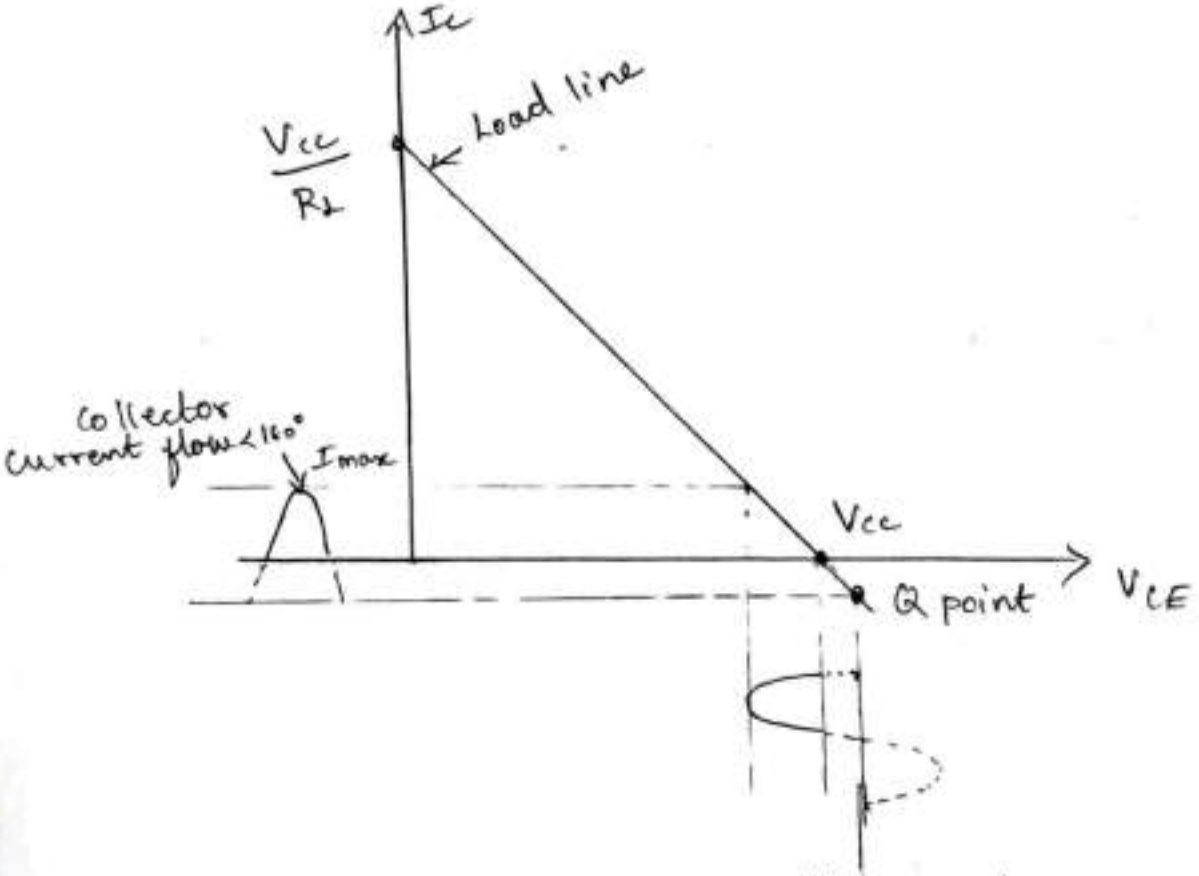
→ As temperature changes along with junction characteristics, the diode get charged and maintain necessary biasing to overcome distortion

Class c operation

→ In class c, resonating circuit is used as load. So most of class c amplifiers are tuned amplifiers



class c tuned Amplifier



waveform representing class c

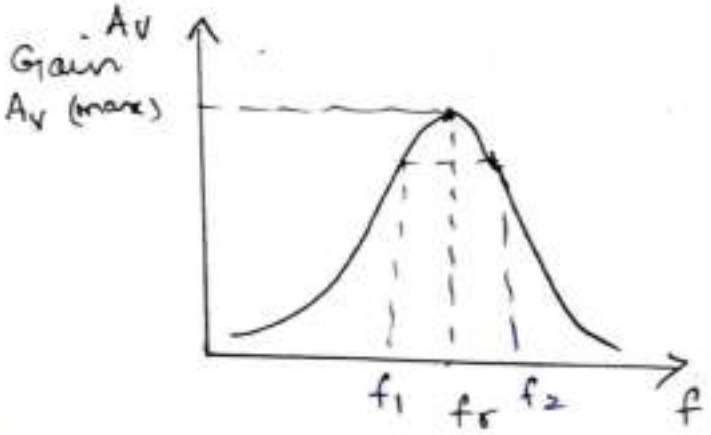
→ A parallel resonant circuit acts as load impedance

→ The collector current flows for less than half a cycle and consists of series of pulses. with harmonics

→ The resonant frequency is given by

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

→ The output Voltage is maximum at resonant frequency. The gain drops on either side



Bandwidth

$$BW = \frac{f_r}{Q}$$

$$B.W = f_2 - f_1$$

Q – Quality factor

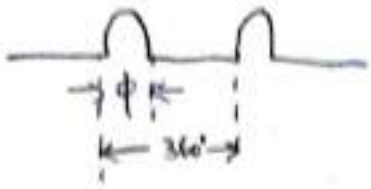
Duty cycle

$$D = \frac{W}{T}$$

W – width of pulse, T – period of pulse

→ Interms of conduction angle

$$D = \frac{\phi}{360^\circ}$$



Output Power

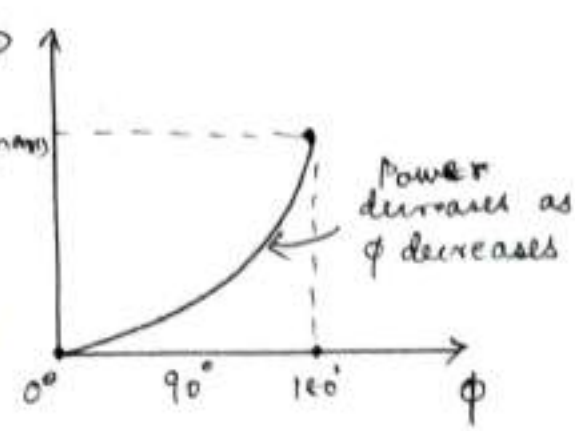
$$P_{out} = \frac{V_{rms}^2}{R_L}$$

$$V_{pp} = 2V_m = 2\sqrt{2}\, V_{rms}$$

$$P_{out} = \frac{(V_{pp}/2\sqrt{2})^2}{R_L} = \frac{V_{pp}^2}{8R_L}$$

Transistor Dissipation

$$P_{D(max)} = \frac{V_{pp\,max}^2}{40\,r_c}$$

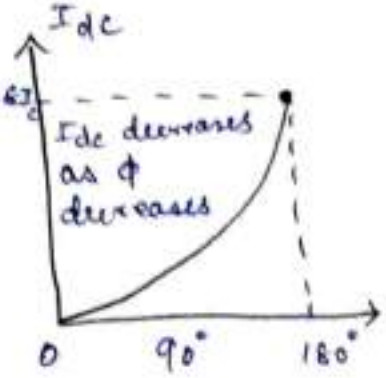


→ Power dissipation depends on conduction angle ϕ

→ less ϕ, less dc power, and less is transistor dissipation

DC Input Power

$$I_{dc} = \frac{I_{c(sat)}}{\pi} = 0.318\, I_{c(sat)}$$



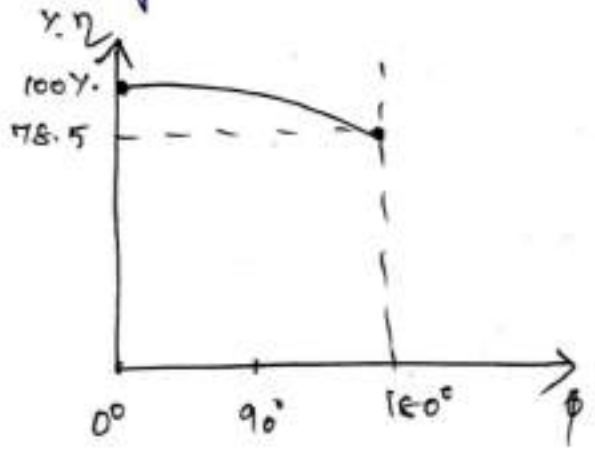
→ If conduction angle is 180°, current is half wave of rectified waveform

Efficiency

→ efficiency is given by ratio of a.c power output to dc power input

$$\%\eta = \frac{P_{out}}{P_{DC}} \times 100 = \frac{P_{out}}{V_{cc} \times I_{dc}} \times 100$$

100% is efficiency is achieved at very small conduction angle



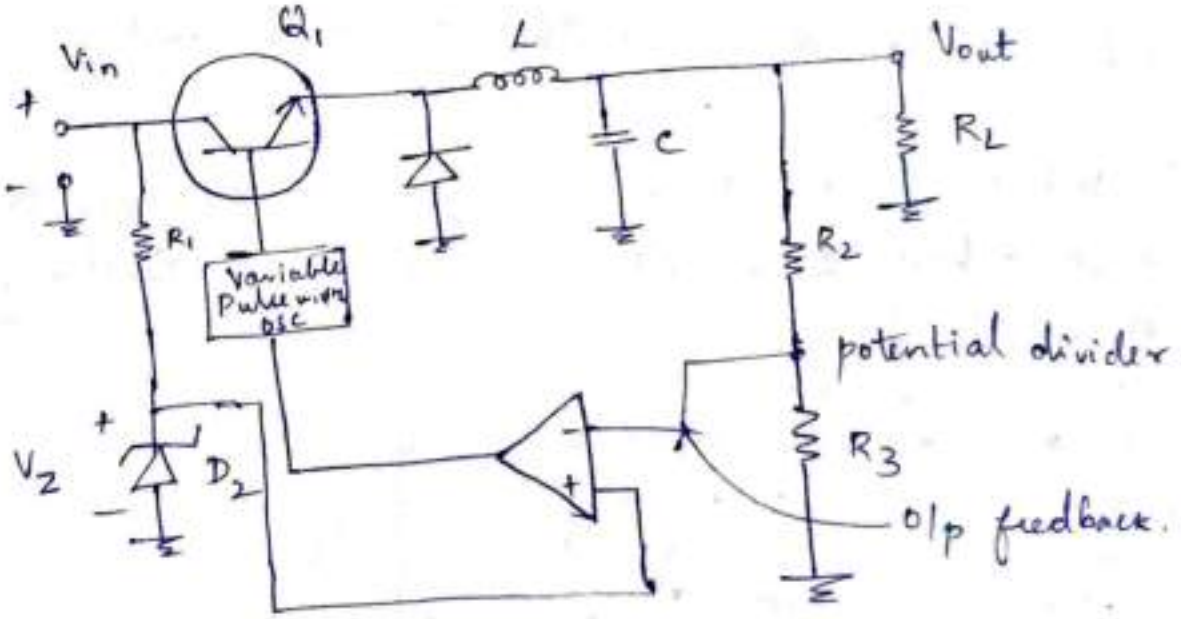
DC/Dc convertors

→ The switch mode regulator is used to describe a ckt which takes dc i/p & provides single d.c o/p

There are three basic configuration of switching regulator

1. Step down or Buck switching regulator
2. Step up or Boost switching regulator
3. Inverting type (Buck-Boost switching regulator)

Step down Switching Regulator (Buck)



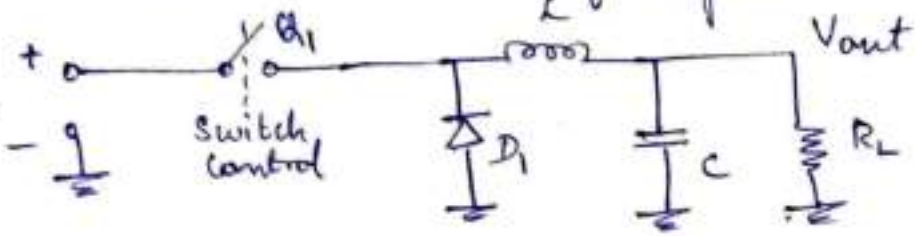
→ It uses an inductor L and series transistor Q_1 which acts as a switch

→ The reference for error amplifier it provided by zener voltage V_Z.

→ The o/p is fed back to error amplifier through potential divider

→ The pulse width oscillator controls the operation of Q_1 as on or off, depending on load requirements

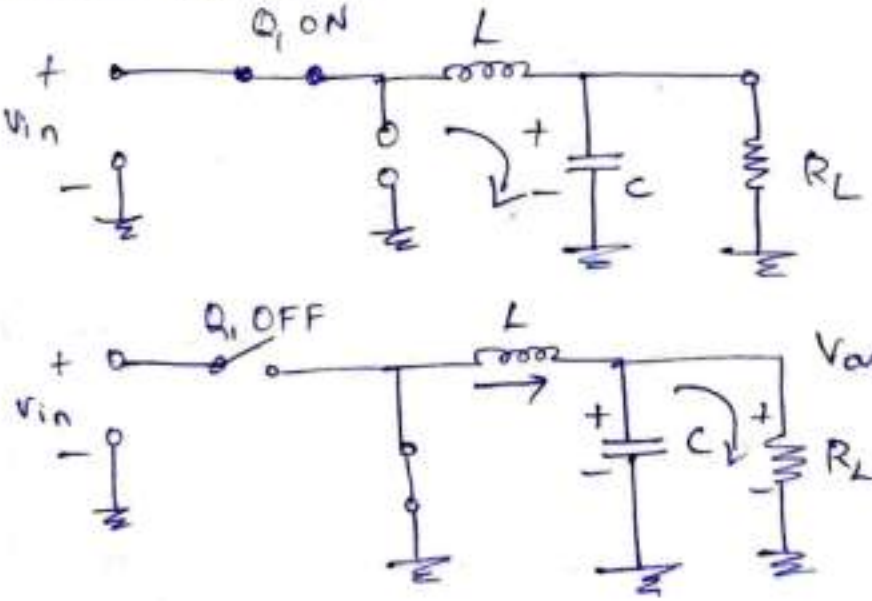
→ The equivalent ckt. of regulator is shown in fig.



→ Q_1 is used for switching the i/p voltage for the required period of time

→ LC filter averages the switched voltage

→ when Q_1 is ON, the capacitor charges thro' it and when Q_1 is OFF, the capacitor discharges thro' load



→ The Variable pulse width oscillator controls ON/OFF periods of Q_1

→ when ON time is more than OFF time, the capacitor charges more, increasing o/p voltage.

→ when OFF time is more, than ON time, the capacitor discharges more, reducing o/p voltage.

→ By adjusting duty cycle $\delta = \frac{t_{on}}{t_{on} + t_{off}}$ of Q_1, the o/p voltage can be regulated

→ If o/p volt ↓, the voltage across R_3 ↓, the error across error amplifier is more

→ this produces pulse of higher width. This increases the charging of capacitor, producing more o/p voltage, Thus the decreased voltage gets compensated

→ If o/p volt ↑, the voltage across R_3 ↑, which produces pulse of smaller width, which reduces t_{on} for Q_1

→ This makes the capacitor to discharge more which increases the o/p voltage

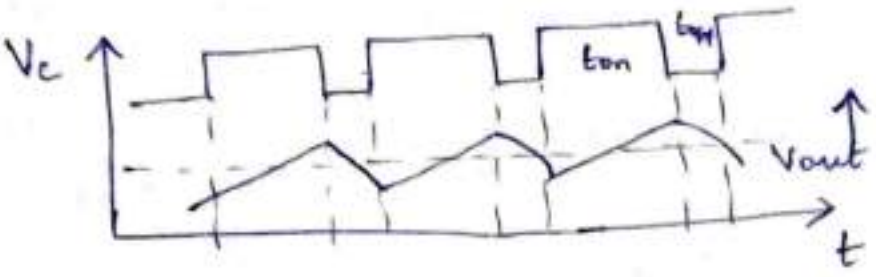
→ Thus o/p voltage is maintained constant by controlling duty cycle of Q_1

→ the o/p voltage is given by

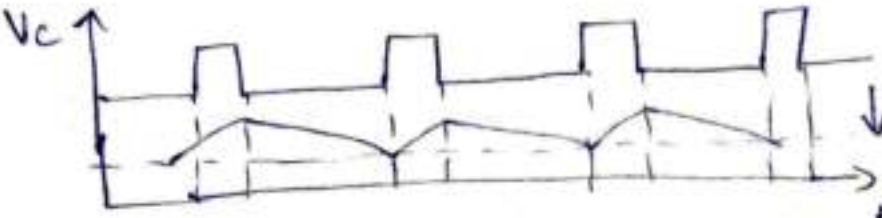
$$\boxed{V_{out} = \delta V_{in}}$$

where $\delta = \frac{t_{on}}{t_{on} + t_{off}} = \frac{t_{on}}{T} = t_{on}(f)$

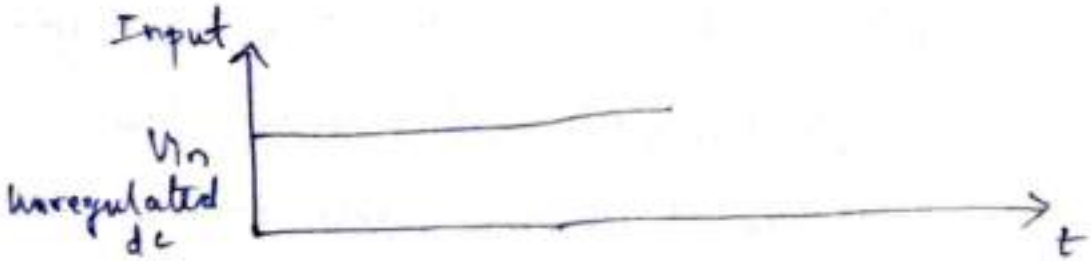
f = frequency

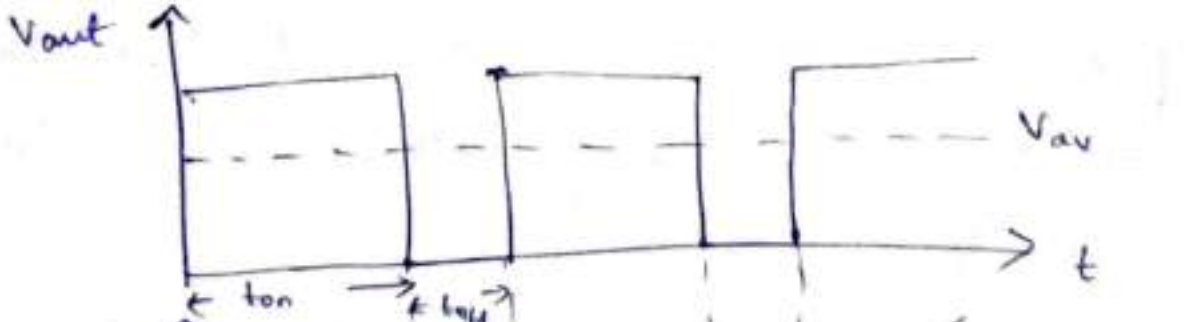


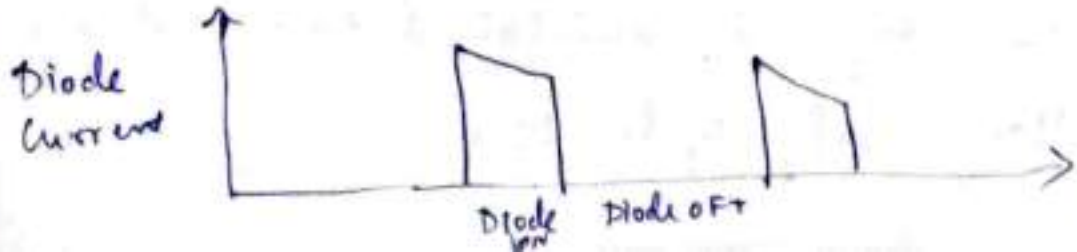
a) $t_{on} > t_{off}$



b) $t_{on} < t_{off}$







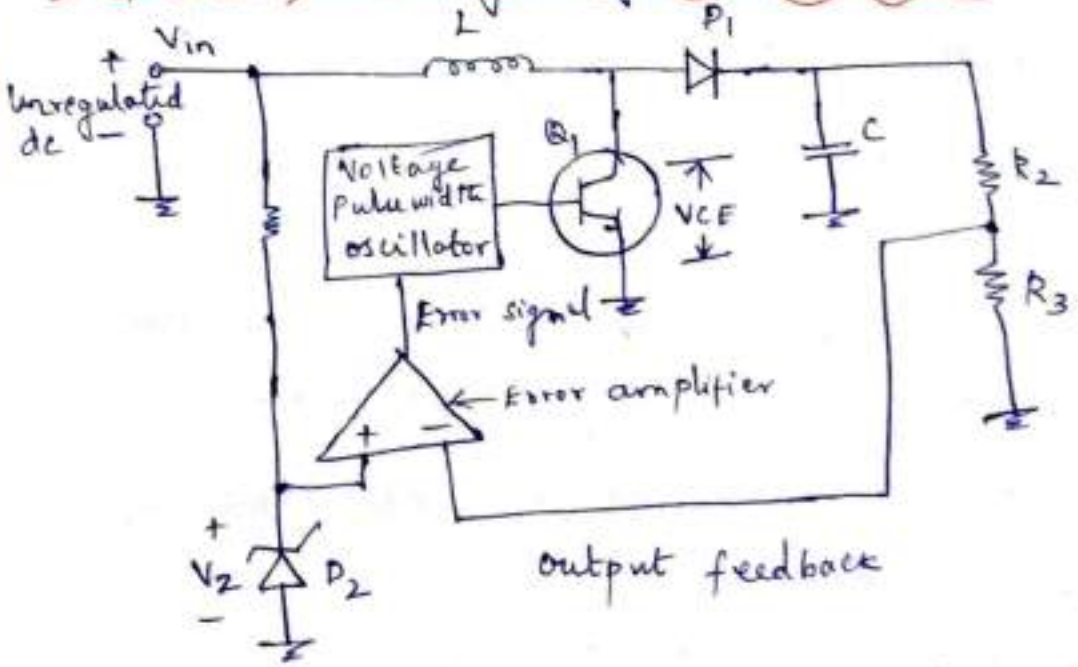
Advantage:

1. High efficiency
2. Simple to design
3. Low ripple content
4. Small output filter.

Disadvantage:

1. Single o/p
2. High i/p ripple
3. No Isolation b/w i/p & o/p
4. Slow transient response

Step Up Switching Regulator (Boost)



→ In Q_1 acts as ON/OFF switch

→ when Q_1 is driven into saturation, V_{CE} is very very small and acts as short circuit.

$V_{in} - V_{CE(sat)}$

Case (1): Let Q_1 is ON (saturation)

→ when Q_1 is ON, V_{CE} is denoted as $V_{CE}(sat)$

→ The voltage across L becomes $[V_{in} - V_{CE(sat)}]$
this expands the magnetic field around the inductor

→ During ON time of Q_1, the voltage across the inductor starts decreasing exponentially from $[V_{in} - V_{CE(sat)}]$

Case :2 Let Q is OFF (cutoff)

→ when Q_1 is OFF, the magnetic field of inductor L, collapses and its polarity gets reversed, since inductor current cannot change instantly.

→ The shorter the ON period, the greater is V_1.

→ The longer the ON time, the smaller the inductor voltage V_1 & less voltage get added to V_{in}, decreasing o/p voltage.

→ when o/p voltage decrease due to increase in load current Then ON time of Q_1 gets reduced

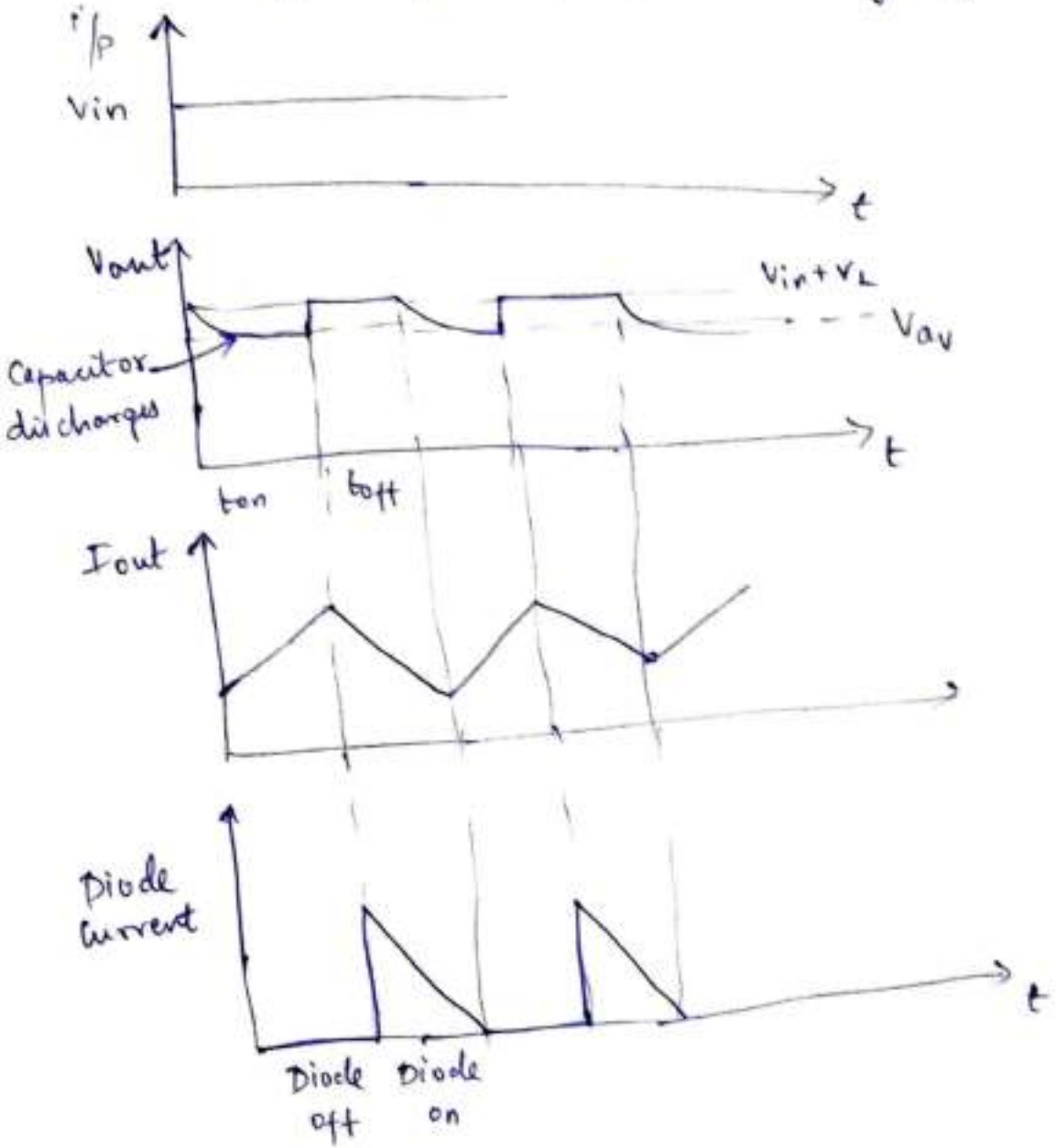
→ when o/p voltage increase, the ON time of Q_1 gets increased.

the output voltage is given by

$$V_{out} = \frac{V_m}{\delta}$$

$$\delta = \frac{t_{on}}{T}.$$

Waveform for step up switching regulator



Advantage

1. o/p voltage is higher than i/p voltage
2. η is high
3. Low i/p ripple
4. Simple in design

Disadvantage:

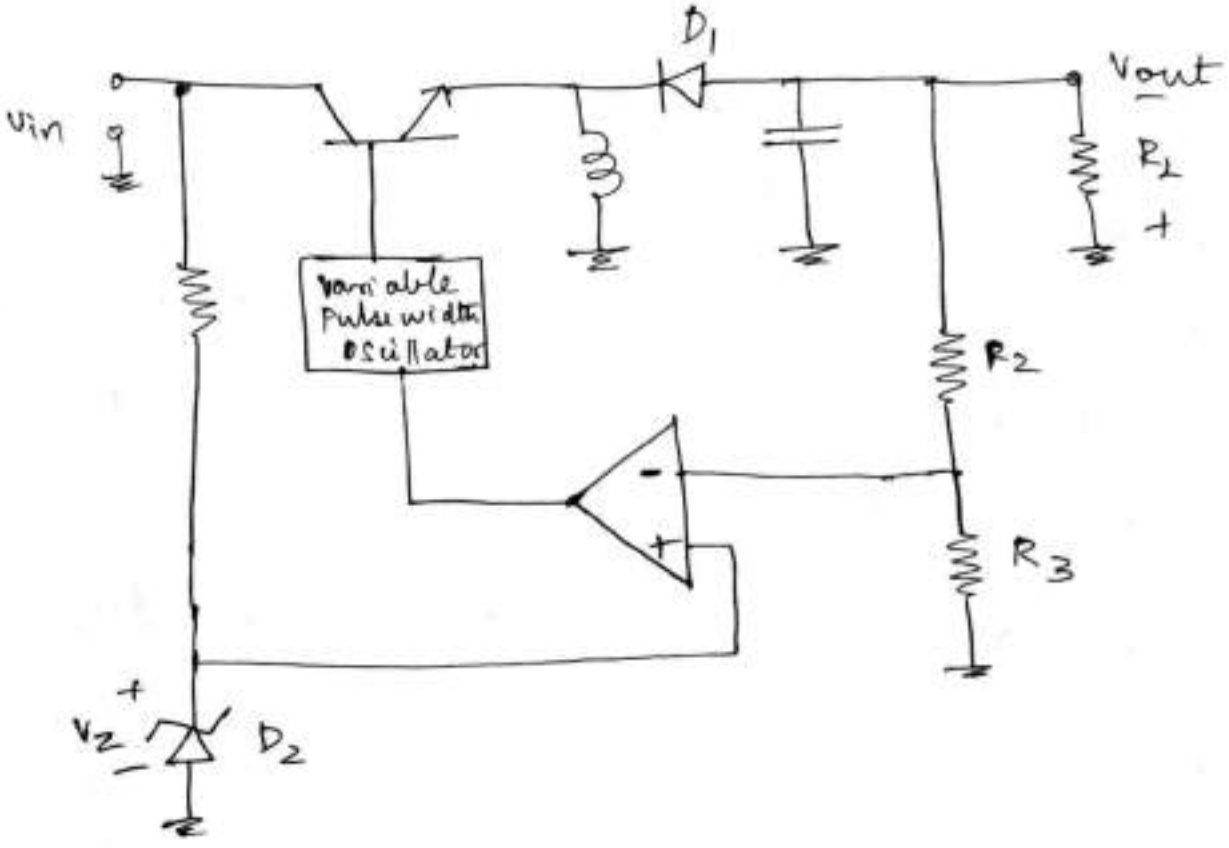
1. It provides single o/p
2. Duty cycle limited to 50%.
3. No isolation b/w i/p & o/p.

Buck-Boost or Voltage Inverter Type Switching Regulator

→ It produces output voltage having polarity opposite to that of input voltage

→ Any change in output produces error which gets amplified by op amp error amplifier

→ this control on/off period of Q_1 to regulate the output, through variable pulse width oscillator



Case: 1 Let Q_1 is switched ON

→ Q_1 goes into saturation and Voltage across it drops to $V_{CE(sat)}$ about 0.3V

→ Due to this voltage across inductor

suddenly rises to $[V_{in} - V_{CE(sat)}]$ and magnetic field around it suddenly expands

→ the inductor value starts exponentially decreasing from initial value $[V_{in} - V_{CE(sat)}]$

Case 2 : let Q_1 is OFF

→ As Q_1 OFF, the magnetic field across L gets collapsed

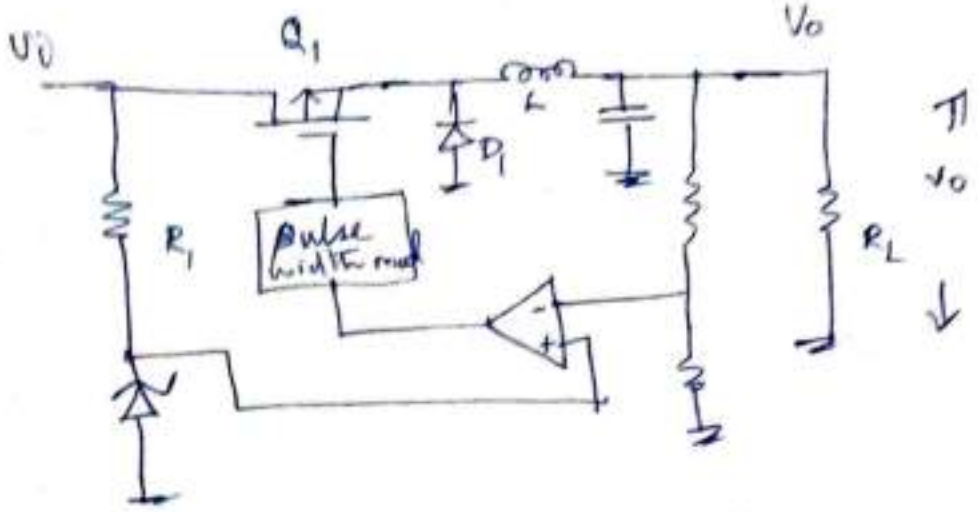
→ Voltage reverses its polarity

→ Due to reverse V_1, diode D_1 is now forward biased

→ The capacitor charges through D_1 producing output voltage of opposite polarity to that of Vin

→ Hence regulator is voltage inverter type

SMPS (MOSFET)



→ when switch is on close

- $V_L = V_i - V_o$ & $V_D = V_L$

→ inductor current will rise

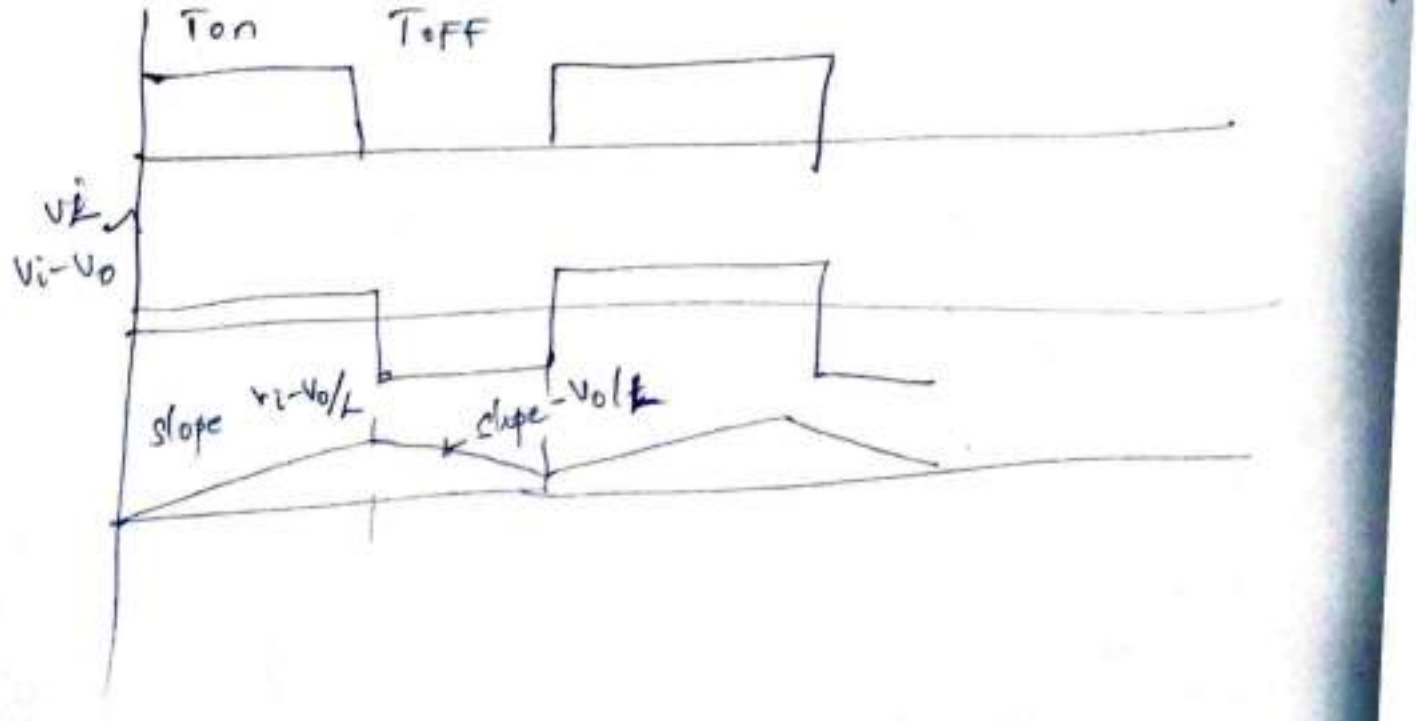
→ Diode D is reverse bias & does not conduct

→ when switch is off current still flows in inductor and into load

$V_L = -V_o$ & $V_D = 0.7$

→ inductor current decreases at rate of $-V_o/L$

→ Diode D is forward biased and conducts $I_L = I_{diode}$



→ R_2/R_1+R_2 is fed back to inverting input of error amplifier

→ difference is given to comparator inverting terminal

→ oscillator generates triangular waveform at fixed freq.

→ o/p of comparator is high when triangular waveform is above the level of error amplifier o/p.

→ when Q_1 is on entire input appears at point A

→ when Q_1 is off, L_1 → supply current to load

→ the diode provides return path for current

→ capacitor C_1 smooths out the voltage at o/p.

$$V_o = \frac{t_{on}}{T} \times V_{in} = \delta V_{in}$$

→ low switch freq improve efficiency and reduce noise but require filtering

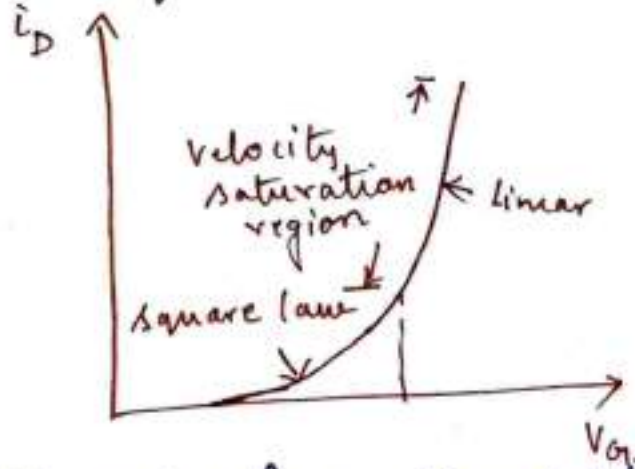
Power MOSFET

→ The operation of power MOSFET is same as conventional MOSFET but power handling of conventional MOSFET is less than 1W.

Features of Power MOSFET

1. Power handling is more than 100W
2. Current handling is in ampere range
3. Large forward conductance
4. As input impedance is very high, large currents can be switched with very small control currents.

characteristics of Power MOSFET



→ power MOSFET have threshold voltages in range of 2V to 4V.

→ In saturation the drain current is related to V_{GS} by square law characteristics

→ The linear portion of characteristics is as a

result of high electric field along short channel causing velocity to reach upper limit, known as velocity saturation.

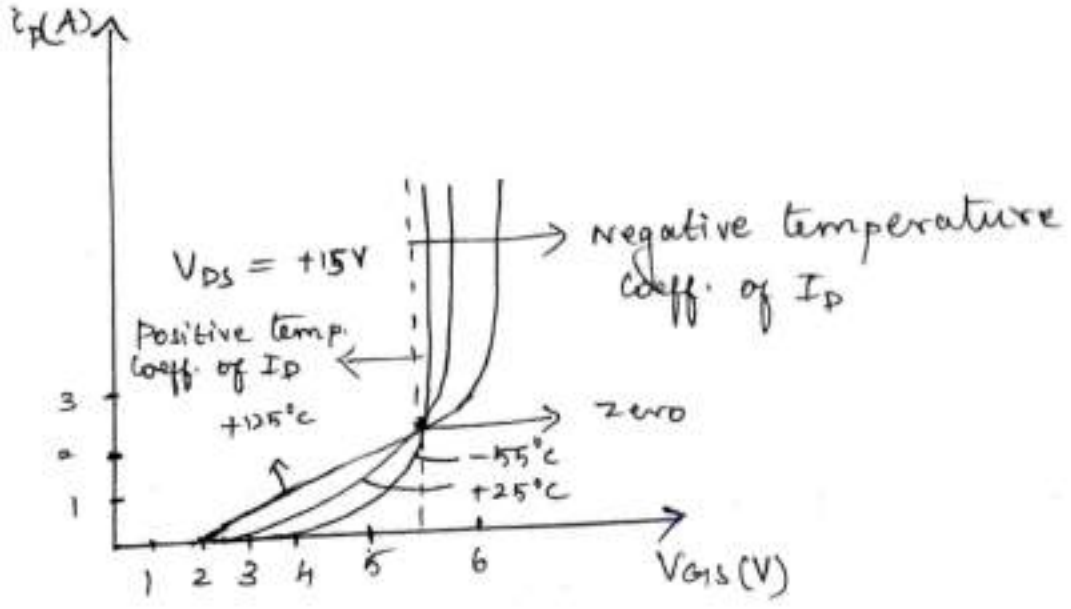
→ The linear i_D-V_{GS} relation implies constant g_m

→ MOSFET is driven in cutoff by applying $V_{GS} < V_{GS(th)}$

→ In ohmic region, MOSFET conducts heavily.

→ In power applications, MOSFET is never operated in active region

MOSFET Temperature effects



→ It is observed that there is value of V_{GS} in the range of 4V to 6V at which temp coeff of i_D is zero

→ At higher values of V_{GS}, i_D exhibits negative temperature coefficient

→ At i_D lower than zero temperature coeff, the temperature coefficient of i_D is positive and MOSFET suffer from thermal runaway

→ the reason for positive temperature coeff. of i_D at low current is that $V_{OV} = V_{GS} - V_t$ is relatively low

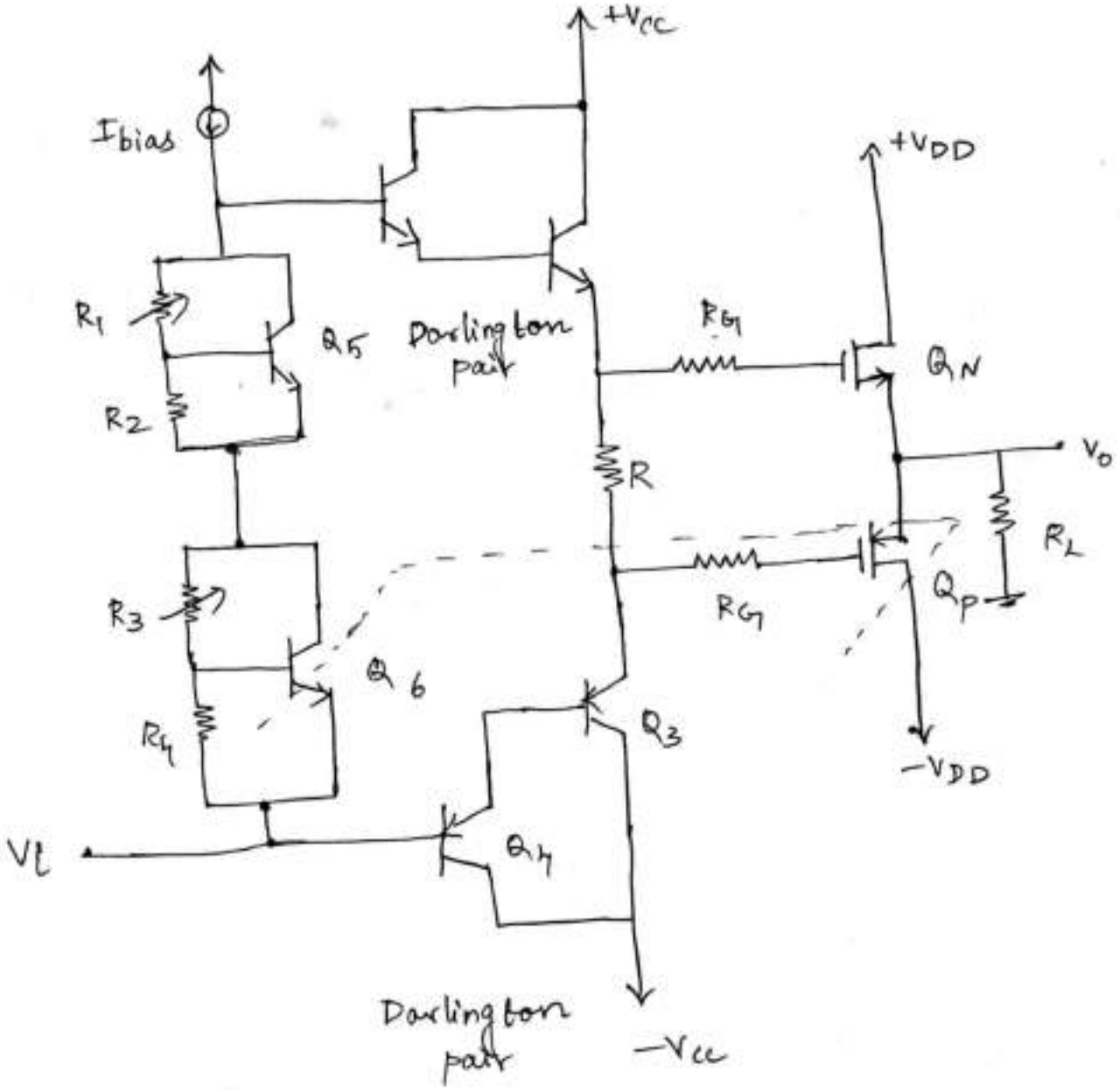
→ Temperature dependence is dominated by negative temperature coefficient of V_t which causes V_{OV} to rise with temperature

class AB power amplifier using power MOSFET

→ The output stage of class AB has a pair of complementary MOSFET

→ BJT are used for biasing and as a driver

→ Complementary Darlington emitter follower formed by Q_1 through Q_4 provide low output resistance necessary for driving output MOSFET at high speeds



class AB push pull amplifier output stage

⇒ The bias circuit uses two V_{BE} multipliers formed by Q_5 and Q_6

→ By adjusting the V_{BE} multiplication factor of Q_6 the bias voltage V_{GG} can be made to decrease with temperature at same rate

$$\star \quad V_{GG} = \left(1+\frac{R_3}{R_4}\right) V_{BE_6} + \left(1+\frac{R_1}{R_2}\right) V_{BE_5} - 4V_{BE}$$

The value of V_{GG} is adjusted to get the desired Q point.