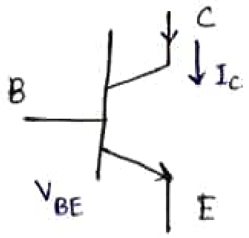
सम्मितीयवैद्युतकपरिपथाः

ANALOG ELECTRONIC CIRCUITS

1-08-2023

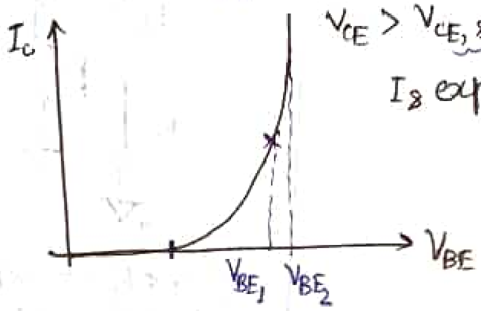
BIPOLAR JUNCTION TRANSISTOR (BJT)



↳ Two terminals control the other two.

Input Characteristic:

↳ what happens at B-E decide C-current.



$V_{CE} > V_{CE,sat}$ ⇝ voltage at which the device goes into forward active mode ($I_c \approx$ const.)

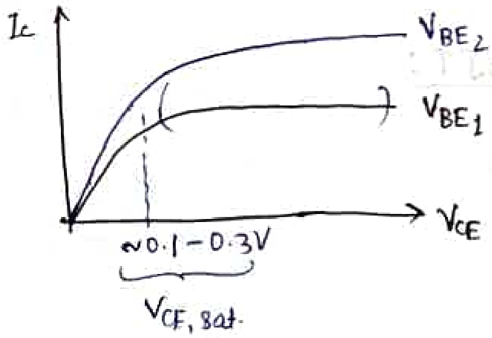
$I_s \exp\left(\frac{V_{BE}}{nV_T}\right)$

$I_C = \beta I_B$

$\alpha I_E = I_c$

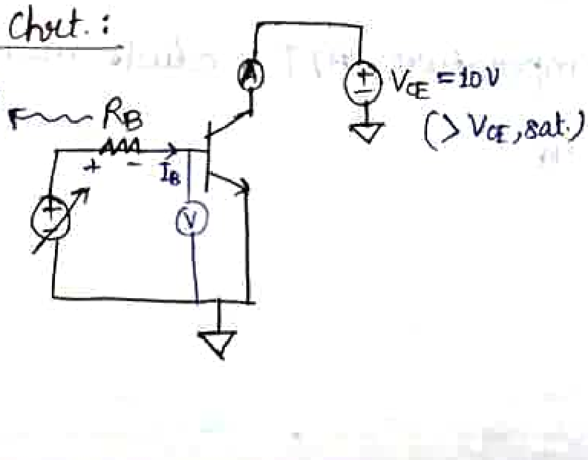
$I_E = (\beta+1) I_B$

$I_E = I_B + I_c$

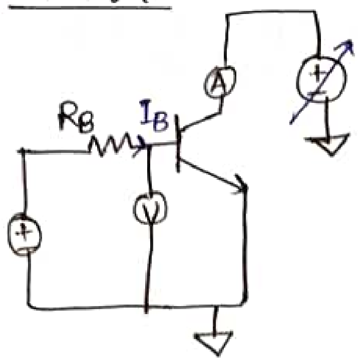


Setup for Input Chrt.:

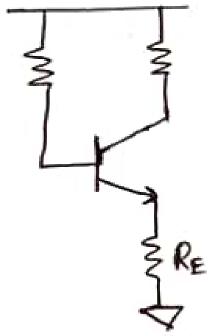
large value of R_B for limit I_B; otherwise transistor may burn



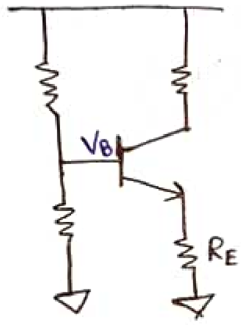
Setup for Output Chrt.:



Fixed Bias



Voltage Divider Bias

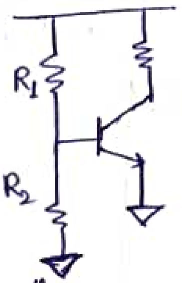


For same ΔV_{BE}, I_c changes by large amount,

$$V_B - V_{BE} - I_E R_E = 0$$

Introducing R_E decreases change in V_{BE}.

What if R_E is not there?



→ is theoretically as good as V.D.B circuit

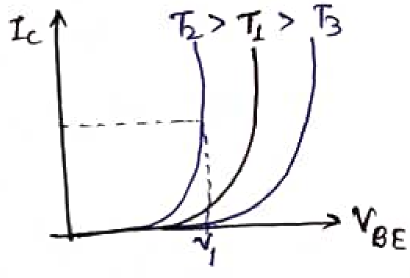
⇓

But R_1 & R_2 may vary due to tolerance

⇓

(R_1 & R_2 are such that V_{BE} (& hence I_B) is same as the previous)

Temperature Dependence of BJT:



At higher temperature, BJT conducts more current (higher I_c) for same V_{BE}.

"Thermal runaway" Problem:

↳ Device goes into a viscous cycle where more heating produces more current and more current produces more heating (Joule's heating), which may damage the device.

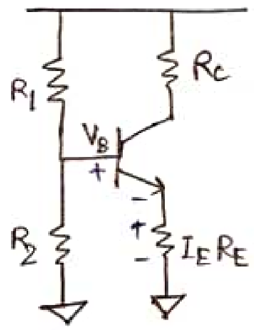
↳ As temp. ↑, for const. current, V_{BE} ↓ by 2.1 mV/°C,
or V_{BE} changes by −2.1 mV/°C.

Eg. For 1mA
@ 27°C
$V_{BE} = 0.695\,V$

If internally temp. changes to 37°C,
$V_{BE} = 0.695 - 2.1\,mV/°C \times 10°C$

→ BJT is a temp. sensitive device.

→



$T\uparrow \Rightarrow I_C, I_E\uparrow \Rightarrow \underset{(V_E)}{I_E R_E}\uparrow \Rightarrow V_{BE}\downarrow$

$T\downarrow \Rightarrow V_{BE}\uparrow$

→ Introducing R_E gives control over V_{BE}.

↳ "Negative feedback action"
↳ Thermal bias stabilization/stability } all b/c of R_E

→ $V_B - V_{BE} - I_E R_E = 0$

for ΔT change in temp., → $R_E\uparrow \Rightarrow$ more stable current

I_E changes by $-2.1\,mV \times \Delta T$,
giving a ΔI_E.

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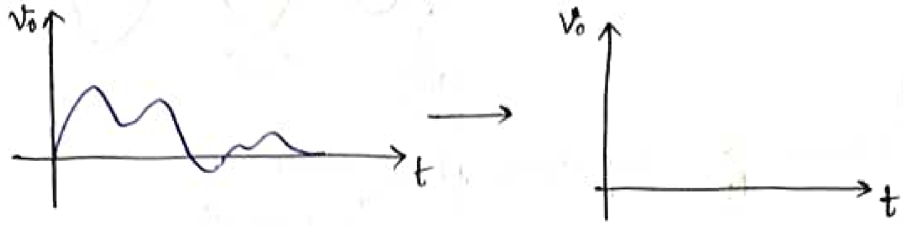
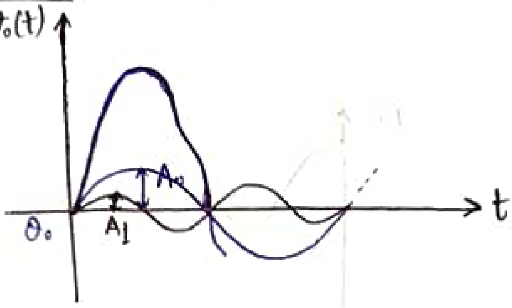
Filters



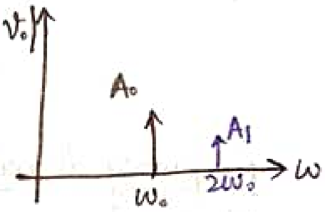
Sine waves:

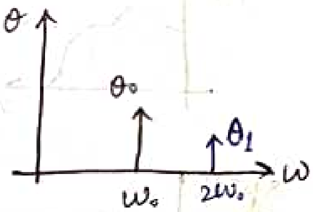


$V_0(t) = \sin(\omega_0 t) \xrightarrow{\text{generalise}} V_0(t) = \sin(\omega_0 t - \theta_0) + \sin(2\omega_0 t - \theta_1)$

ω_0 : frequency in rad/sec.

Wrt. ω :



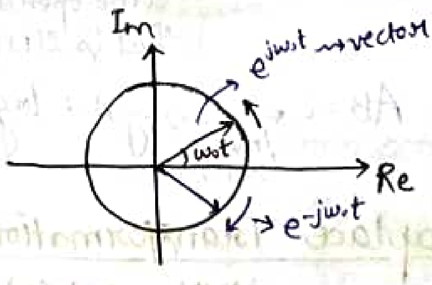


↳ Frequency domain Representation

Spectrum Analyser

vu meter: measures loudness

$\rightarrow \dfrac{e^{j\omega_0 t} - e^{-j\omega_0 t}}{2j}$



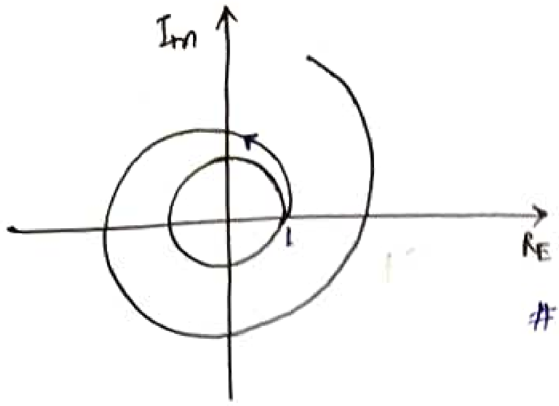
In frequency domain representation, complex frequency content are

$j\omega_0, -j\omega_0$

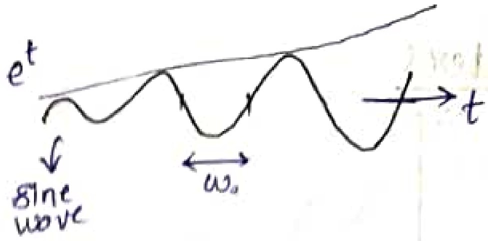
$e^{j\omega_0 t} \longrightarrow e^{(\sigma + j\omega_0)t}$

$= e^{\sigma t} \cdot e^{j\omega_0 t}$

$= e^{t} \cdot e^{j\omega_0 t}$ (Assume $\sigma = 1$)



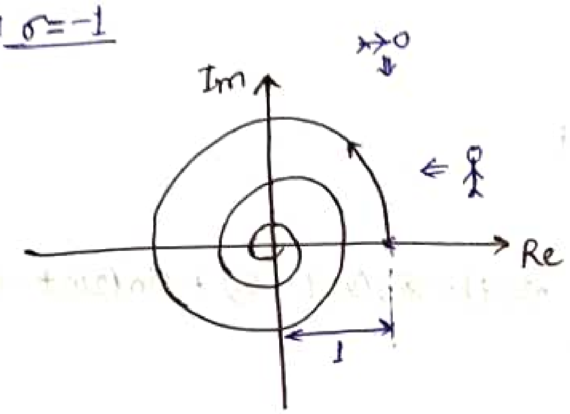
Side view:

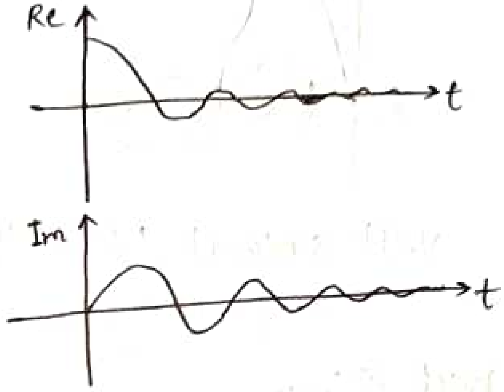


Top view: phase shift
⇓ cosine wave

sin & cos are projections of time-varying $e^{(\sigma + j\omega_0 t)}$

For $\sigma = -1$





→ Frequency $= \sigma + j\omega$ (now onwards)
$= 's'$: complex frequency

Fourier series

time domain signal $\xrightarrow{\text{transform}}$ frequency domain

$V_o(t) \longrightarrow V_o(\omega)$

$\log(AB)$

A $\xrightarrow{\text{transform}}$ $\log(A)$

B ⟶ $\log(B)$

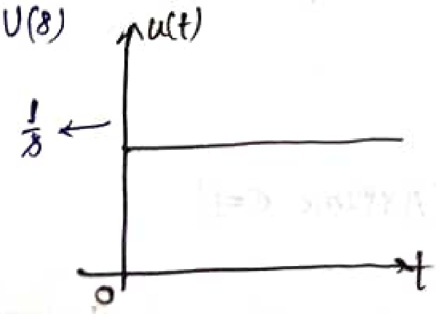
↳ some operation that is simpler

$AB = C \xleftarrow[\text{transform}]{\text{inv.}} \log A + \log B$

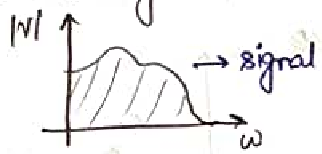
Laplace Transformation

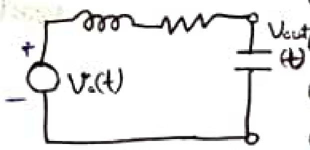
$V_o(t) \longrightarrow V_o(s)$

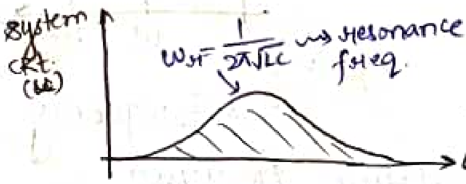
$u(t) \rightarrow U(s)$



Frequency domain wave/fn:

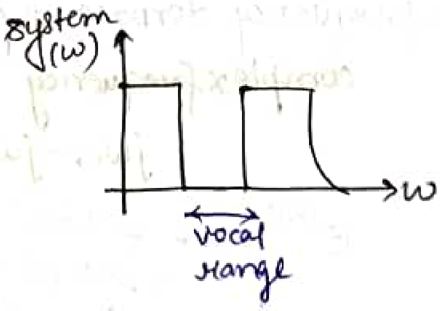






→ Responce

$Out(\omega) = Input(\omega) \times \underbrace{system(\omega)}_{\text{filter}}$



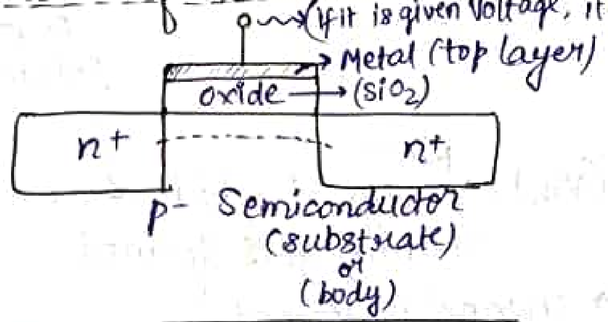
MOSFETs

(Metal Oxide Semiconductor Field Effect Transistor)

→ Transistor
- → Bipolar Junction Transistor ⤳ Junction (p-n) controls the current through the conductor
- → Field Effect Transistor ⤳ Electric field applied controls the transistor current

MOSFET

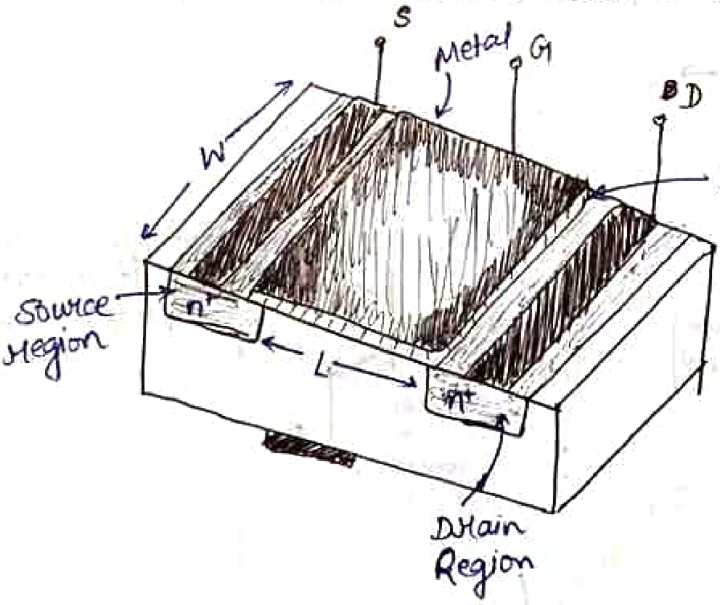
Cross-section of the device:

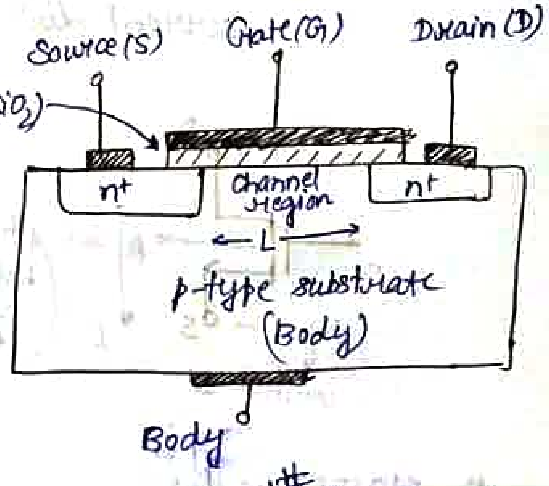


p^- → lightly doped
p^+ → highly doped

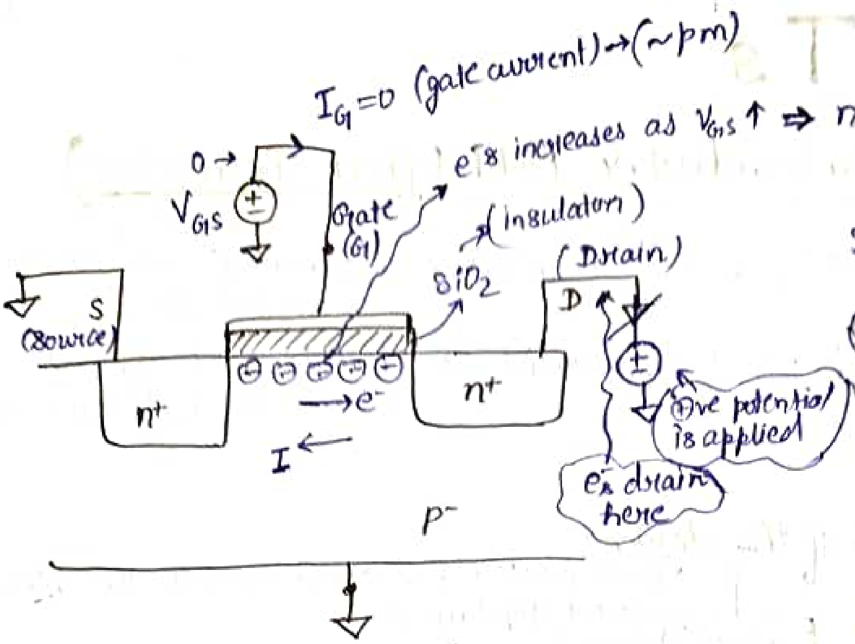
- → N-Mos (p^- substrate) → semiconductor is p-doped
- → PMos (n^- substrate) → semiconductor is n-doped

↳ 4-terminal device (But we can get 3 terminals also).





9mm semiconductor → the length 'L' b/w 2 n^+ regions



⇒ nature at the top of p^- semiconductor changes

⇓

Semiconductor property has changed (p type to n type) (only at the vicinity of oxide layer)

"Inversion layer"

⇓

channel (b/w 2 n^+s) is formed (or channel region is inverted)

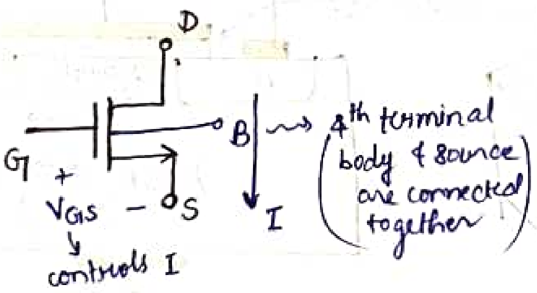
→ Threshold voltage (V_{Th}) : Minm Voltage at which ~~cannot g~~ channel gets formed

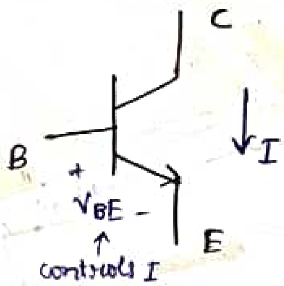
$V_{GS} \geq V_{Th}$ ⇒ channel is formed

→ MOSFET is a Symmetric device.
↳ can interchange D & S.

→ Higher potential → Drain (attracts e^-s)
Lower potential → Source (source e^-s)

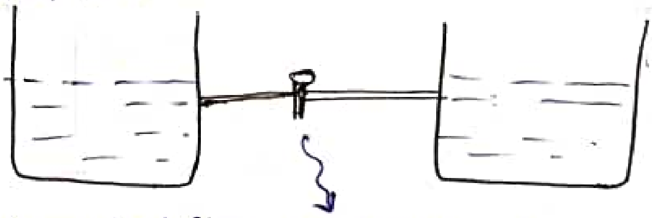
current dirn: D → S





⇒ cannot work if reversed (not a symmetric device)

MOSFET is like:

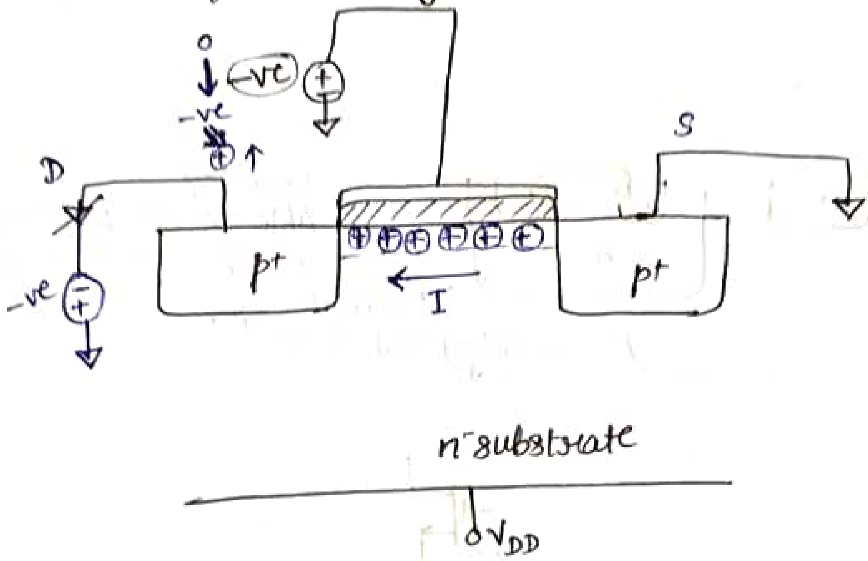


→ for current flow (water flow):

① $V_{GS} < V_{TH}$ (cut off region) ⇝ Tap ~~open~~ off

② $V_{GS} \geq V_{TH}$ & $V_{DS} = 0$ ⇝ Some ~~different~~ water levels in the two tanks.

→ CMOS - complementary MOS



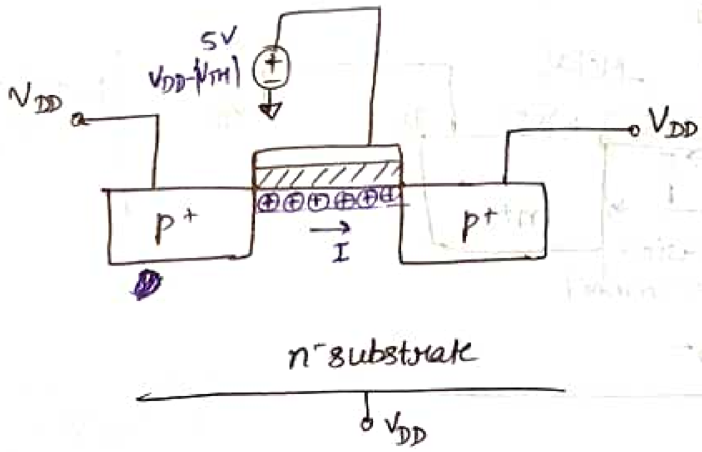
→ p type Inversion layer

⇓

p type channel

$V_{GS} < -V_{TH_p}$

But, we take +ve voltage.

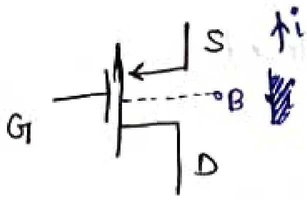


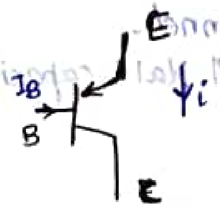
For 'holes' charge carriers:

higher potential → S

lower potential → D

NMOS	PMOS
$V_{GS} > V_{TH}$	$V_{SG} > \lvert V_{TH} \rvert$
	$S_{SD} > 0$





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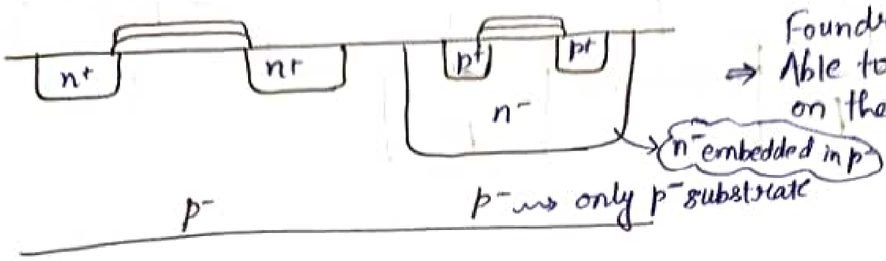
→ CMOS:



Foundry
⇒ Able to do NMOS & PMOS on the same p− substrate.

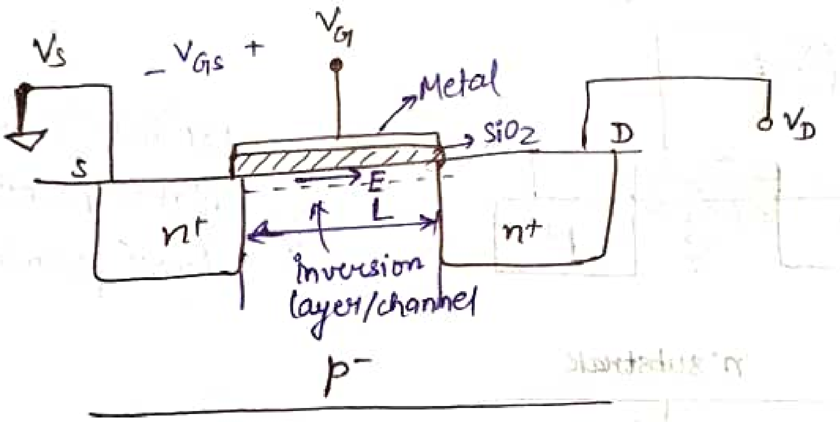
Modes of Operation

① Cut-off Region

↳ controls I_D & I_S.

$V_{GS} < V_{TH}$

$I_D = 0$ (no channel b/w D & S)



For channel formation: $V_{GS} > V_{TH}$

② $V_{GS} \geq V_{TH}$ & $V_{DS} = 0$

$\Rightarrow I_D = 0$

③ $V_{DS} > 0$ & $V_{GS} \geq V_{TH}$

$I_D = ?$

$Q = CV$, Q: charge in channel.

C: capacitance (‖ plate capacitor formed)

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

C_{ox} = oxide capacitance / unit area

$\therefore C = C_{ox} \cdot WL$

↳ Total capacitance

$V = V_{GS} - V_{TH} = V_{OV}$: Over drive voltage

$\therefore Q = CV = C_{ox} \cdot WL(V_{GS} - V_{TH})$

Drain current:

I_D = charge / unit length × velocity of electron

$$= \frac{Q}{L} \times \mu_n E$$

drift velocity (V_d) $= \mu_n E$ — mobility, electric field

$$E_{DS} = \frac{V}{L} = \frac{V_{DS}}{L}$$

Electric field in channel (b/w D & S)

$$\therefore I_D = C_{ox} \cdot W (V_{GS} - V_{TH}) \mu_n \frac{V_{DS}}{L}$$

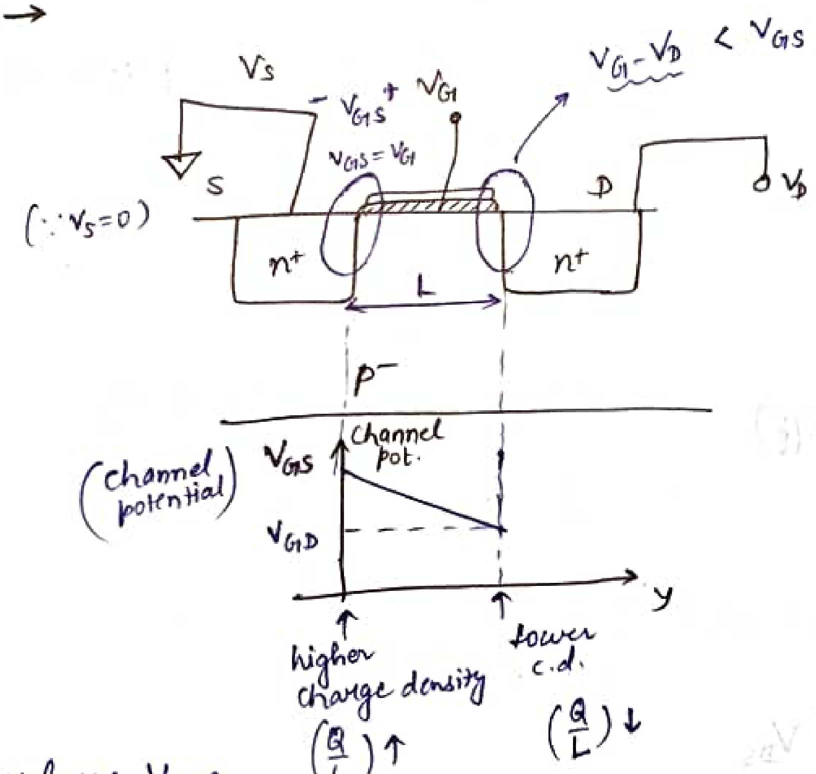
$$\Rightarrow \boxed{I_D = \mu_n C_{ox} \left(\frac{W}{L}\right) (V_{GS} - V_{TH}) V_{DS}} \quad -(1)$$

→ only V_{GS} & V_{DS} can be controlled

$$I_D \propto (V_{GS} - V_{TH}) = V_{OV}$$

$$I_D \propto V_{DS}$$

→



For large V_{DS}:

Average potential ($V_{OH,eff}$) $= \dfrac{(V_{GS} - V_{TH}) + (V_{GD} - V_{TH})}{2}$, where $V_{GD} = V_{GS} - V_{DS}$

$$= \frac{(V_{GS} - V_{TH}) + (V_{GS} - V_{DS} - V_{TH})}{2}$$

$$\therefore I_D = C_{ox} \cdot W \left(\frac{(V_{GS} - V_{TH}) + (V_{GS} - V_{DS} - V_{TH})}{2}\right) \mu_n \frac{V_{DS}}{L}$$

$$= \mu_n C_{ox} \left(\frac{W}{L}\right) \left[(V_{GS} - V_{TH}) - \frac{V_{DS}}{2}\right] V_{DS}$$

$$\Rightarrow \boxed{I_D = \mu_n C_{ox} \left(\frac{W}{L}\right) \left[(V_{GS} - V_{TH}) V_{DS} - \frac{1}{2} V_{DS}^2\right]} \quad -(2)$$

If we keep increasing V_D:

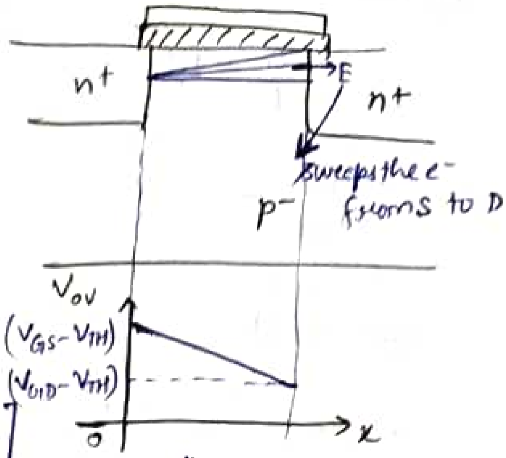
$V_{GD} \leq V_{TH}$

$\Rightarrow V_{GS} - V_{DS} \leq V_{TH}$

$\Rightarrow \boxed{V_{DS} \geq V_{GS} - V_{TH} = V_{OV}}$ → channel gets pinched off

⇓

V_{DS} has no control over I_D



⇓

Channel & E still exist.

⇓

Similar to BC reverse biased in BJT (current flows despite of reverse bias)

For $V_{DS} = (V_{GS} - V_{TH})$,

$$I_D = \mu_n C_{ox}\left(\frac{W}{L}\right)\left[(V_{GS}-V_{TH})(V_{GS}-V_{TH}) - \frac{1}{2}(V_{GS}-V_{TH})^2\right]$$

$$\Rightarrow \boxed{I_D = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)(V_{GS}-V_{TH})^2} \quad -③$$

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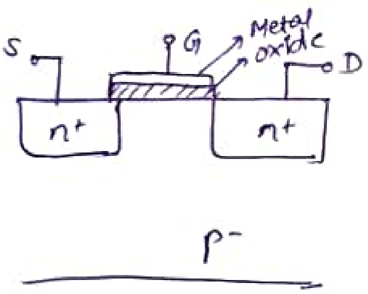
→



(I) $V_{GS} < V_{TH}$, $I_D = 0$ (cutoff)

(II) $V_{GS} \geq V_{TH}$ & $V_{DS} = 0$

$I_D = 0$

For small V_{DS}:

$$I_D = \underbrace{\mu_n C_{ox}\left(\frac{W}{L}\right)}_{K_n}(V_{GS}-V_{TH})V_{DS}$$

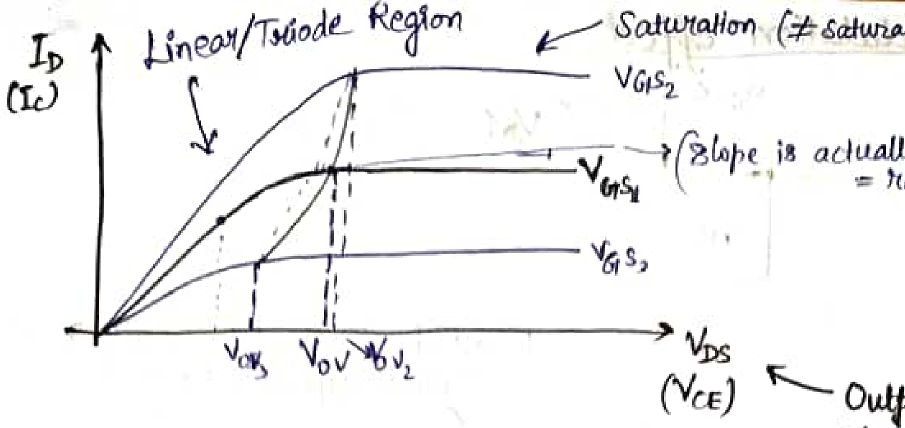
$$I_D = K_n(V_{GS}-V_{TH})V_{DS} \quad -①$$

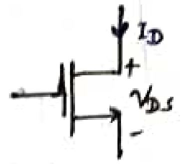
For little larger V_{DS}:

$$I_D = K_n\left[(V_{GS}-V_{TH})V_{DS} - \frac{1}{2}V_{DS}^2\right] \quad -②$$

For $V_{DS} \geq V_{GS} - V_{TH} = V_{OV}$,

$$I_D = \frac{K_n}{2}(V_{GS}-V_{TH})^2 \quad -③$$





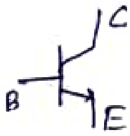
← Output Characteristics

$V_{GS_2} > V_{GS_1} > V_{GS_3}$

→ $\underline{V_{OV}}$: well-defined point (unlike in BJT) where the device moves from linear to saturation region or vice versa.

Input Characteristics:

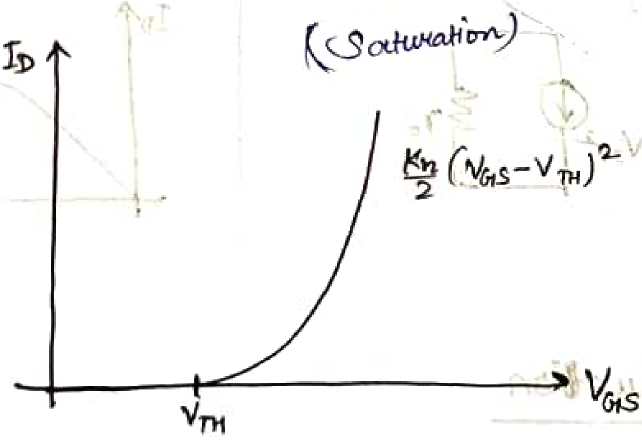
BJT		MOSFET
E	⟶	S
B	⟶	G
C	⟶	D



Differences b/w BJT & MOSFET:

(A) $I_G = 0$ vs. $I_B \neq 0$
↳ (As gate doesn't get current due to oxide dielectric)

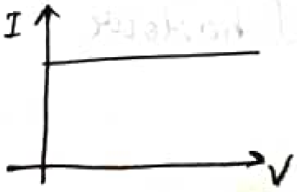
(B) $I_D = I_S$ vs. $I_E = I_C + I_B$



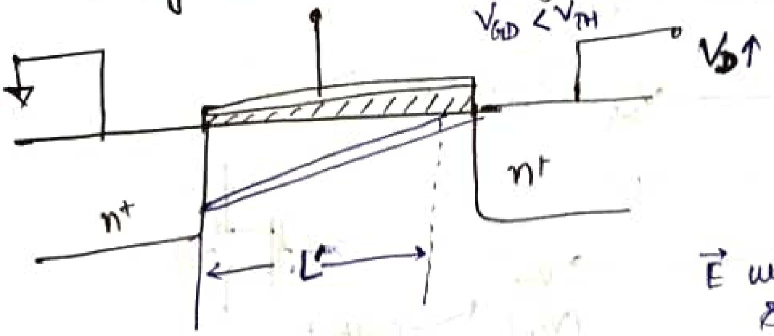
⇒ BJT ↓ Exponential curve; MOSFET ↓ quadratic curve

← Input Characteristics

→ There are no ideal current source.



Channel Length Modulation Effect



$\vec{E}$ will be present but only in the small effective region

$$\uparrow I_D = \frac{\mu_n \cdot C_{ox}}{2}\left(\frac{W}{L\downarrow}\right)(V_{GS} - V_{TH})^2 \quad (\because L \rightarrow L')$$

→ on increasing V_{GS}, effective channel length decreases.

$$I_D = \frac{K_n}{2}(V_{GS} - V_{TH})^2(1 + \lambda V_{DS})$$

, λ : channel length modulation coefficient
↳ constant.

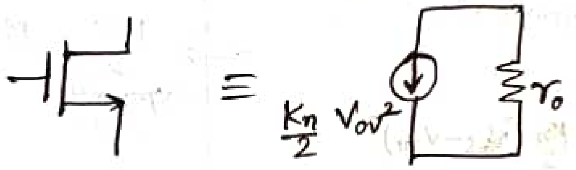
$$\left.\frac{\partial I_D}{\partial V_{DS}}\right|_{V_{GS} = \text{const.}} = \lambda \cdot \frac{K_n}{2}(V_{GS} - V_{TH})^2 \approx \lambda I_D$$

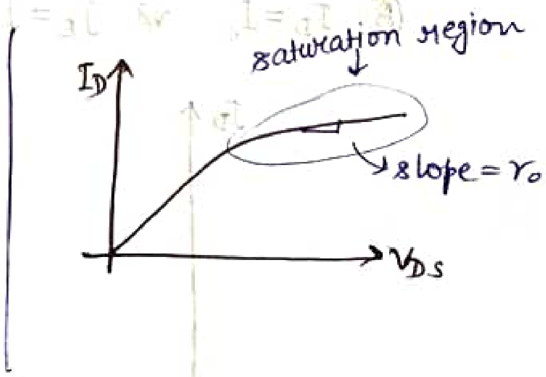
↳ conductance

$\therefore$ $g_{ds} \cong \lambda I_D$ → Output conductance

$$r_o = \frac{1}{g_{ds}} \cong \frac{1}{\lambda I_D}$$ → Output impedence

In the saturation region,



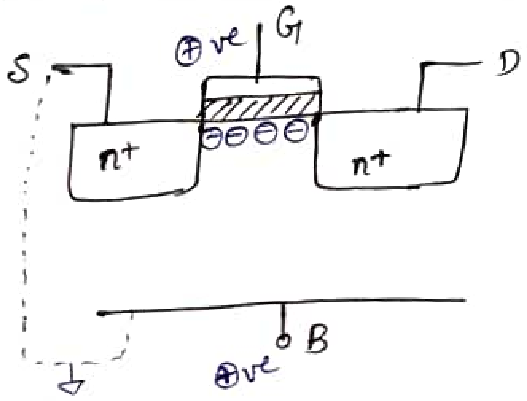


① Sub-threshold conduction

↳ Small I_D exists for $V_{GS} < V_{TH}$.

↳ Exponential behaviour.

② Body-effect

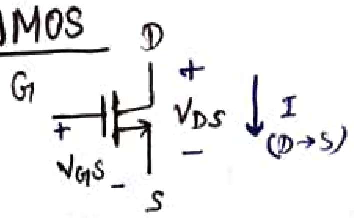


→ Normally, S & B are connected together.
(body is grounded)

→ Using body voltage, threshold voltage can be altered.
↳ also called "Back Gate".

14-08-2023

NMOS



$V_{GS} \geq V_{THN}$

Triode:

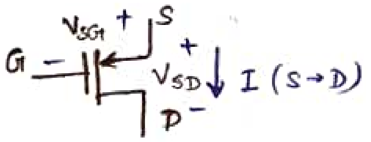
$$I_D = \mu_n C_{ox}\left(\frac{W}{L}\right)(V_{GS} - V_{THN})V_{DS}$$

$$= \mu_n C_{ox}\left(\frac{W}{L}\right)\left[(V_{GS} - V_{THN})V_{DS} - \frac{1}{2}V_{DS}^2\right]$$

Saturation: $V_{DS} \geq V_{GS} - V_{TH}$

$$I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_{THN})^2 = \frac{K_n}{2}(V_{GS} - V_{THN})^2$$

PMOS



For PMOS to turn ON,

$V_{SG} > |V_{THP}|$

$$I_D = \mu_p C_{ox}\left(\frac{W}{L}\right)(V_{SG} - |V_{THP}|)V_{SD}$$

($V_{GS} \xrightarrow{flip} V_{SG}$)

$$I_D = \mu_p C_{ox}\left(\frac{W}{L}\right)\left[(V_{SG} - |V_{THP}|)V_{SD} - \frac{1}{2}V_{SD}^2\right]$$

hole mobility < e^- mobility

Saturation region: $V_{SD} \geq V_{SG} - |V_{THP}|$

$$I_D = \frac{\mu_p C_{ox}}{2}\left(\frac{W}{L}\right)(V_{SG} - |V_{THP}|)^2$$

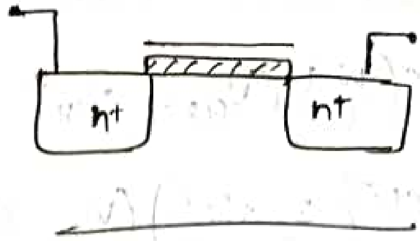
$$K_n = \mu_n C_{ox}\left(\frac{W}{L}\right)$$

or

$$K_n' = \mu_n C_{ox}.$$

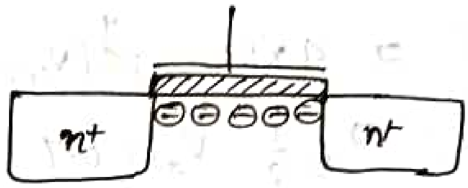
Enhancement MOSFET

↦ Normal MOSFETs.



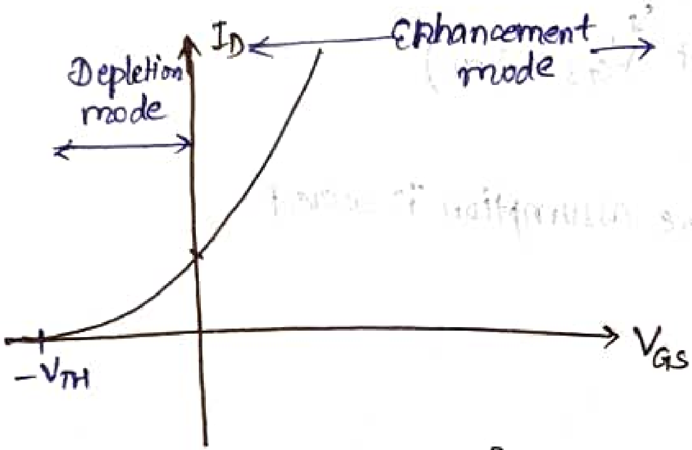
Depletion type MOSFET

↦ channel is already embedded.

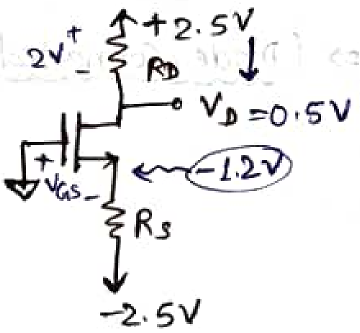


- Even for $V_{GS} = 0$, channel is already formed.
- To turn off the device,

$$V_{GS} \rightarrow \ominus ve.$$



Q) Design the circuit such that you get $I_D = 0.4mA$, $V_D = +0.5V$, $V_{TH} = 0.7V$, $\mu_n C_{ox} = 100\ \mu A/V^2$, $W = 32\ \mu m$, $L = 1\ \mu m$, $\lambda = 0$. (find R_D & R_S)



Soln: Voltage drop across R_D = $2.5 - 0.5 = 2V$

$R_D I_D = 2V$

$\Rightarrow R_D = \frac{2}{0.4m} = 5K\Omega$

Using $I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_{TH})^2$ → ~~from~~ for saturation region
then check if
$V_{DS} \geq (V_{GS} - V_{TH})$.

$\Rightarrow 0.4 \times \cancel{10^{-3}}^{10} = \frac{\cancel{100 \times 10^{-6}}}{2}\left(\frac{32 \times 10^{-6}}{1 \times 10^{-6}}\right)(V_{GS} - 0.7)^2$

$\Rightarrow \cancel{4 \times 2} = \cancel{32}^{4}(V_{GS} - 0.70)^2$

$\Rightarrow \pm \frac{1}{2} = V_{GS} - 0.7$

$\Rightarrow V_{GS} = 0.5 + 0.7$ or $V_{GS} = -0.5 + 0.7$

$\Rightarrow V_{GS} = 1.2\ V$ or $V_{GS} = 0.2\ V$

$\Rightarrow V_{GS} = 1.2 V$ (Neglect $V_{GS} < V_{TH}$)

Here,
$V_{DS} = 1.7V$
$V_{GS} - V_{TH} = 0.5$
} hence, the assumption is correct.

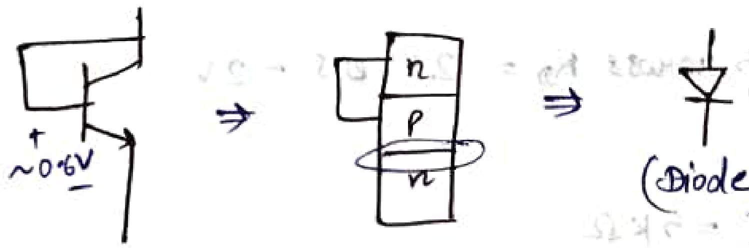
Now, $V_G = 0 \Rightarrow V_S = -1.2V$

$\frac{V_S + 0.2}{R_S} = I_S = I_D$

$\Rightarrow R_S = \frac{1.2 + 0.2}{0.4 m} = \frac{1.4}{0.4 m} = 3.5\ k\Omega$

→ 
← always in saturation when it is conducting. ⇒ (Diode-Connected Transistor)

$V_{DS} = V_{GS}$
$> V_{GS} - V_{TH}$

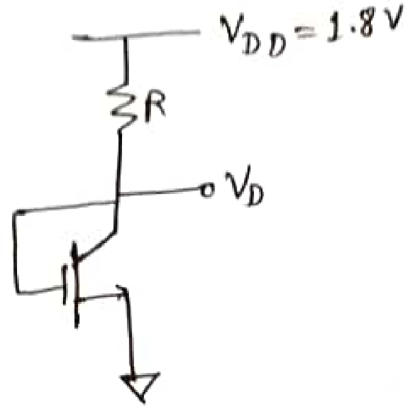
→ 

"Diode Connected Transistor"

Q) Find R such that $V_D = \cancel{0.5V}\ 0.7V$.

$\mu_n C_{ox} = 0.4 mA/V^2$

$\frac{W}{L} = \frac{0.72\ \mu m}{0.18\ \mu m}$, $V_{TH} = 0.5V$



Soln: $I_D = \frac{V_{DD} - V_D}{R}$

$= \frac{1.8 - 0.7}{R} = \frac{1.1}{R}$

and, $I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{\cancel{GS}}^{DS} - V_{TH})^2$

$\Rightarrow \frac{1.1}{R} = \frac{0.4 \times 10^{-3}}{\cancel{2}}\left(\frac{0.\cancel{72}}{0.\cancel{18}}\right)(0.7 - 0.5)^2$

$\Rightarrow \frac{1.1}{R} = 8 \times 10^{-3}(0.04)$

$\Rightarrow \frac{1.1}{R} = 32 \times 10^{-5}$

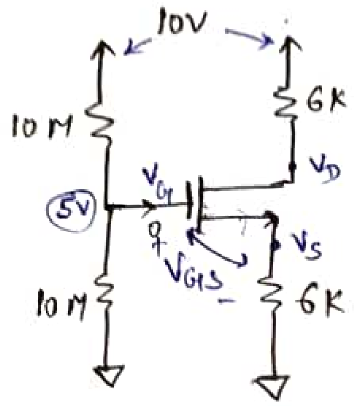
$\Rightarrow R = \frac{11 \times 10^5}{8 \times 4}$

$\Rightarrow R = 34.375\ k\Omega$

Q) Determine all node voltages & branch current.

$V_{TN} = 1V$ $(= V_{TH})$

$K_n = 1 mA/V^2$



Soln: $V_G = 5V$

Using $I_D = \underbrace{\frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)}_{K_n \times 2}(V_{GS} - V_{TH})^2$

$\Rightarrow I_D = \frac{K_n}{2}(V_{GS} - V_{TN})^2$

$= \frac{1 \times 10^{-3}}{2}(5 - V_S - 1)^2$

Using $5 - V_{GS} - I_D R_S = 0$

$\Rightarrow \cancel{5} = (\cancel{5} - V_S) + \frac{10^{-3}}{2}(4 - V_S)^2 \cdot 6 \times 10^3$

$\Rightarrow V_S = \frac{(16 + V_S^2 - 8V_S)}{\cancel{2}} \cdot 3$

$\Rightarrow 3V_S^2 - 25V_S + 48 = 0$

$\Rightarrow V_S = 3, \frac{16}{3}$

As $V_{GS} > V_{TH}$, $V_S \neq \frac{16}{3}$.

$\therefore V_S = 3V$

$\Rightarrow I_D = \frac{1}{2}(4-3)^2 m = 0.5 mA$

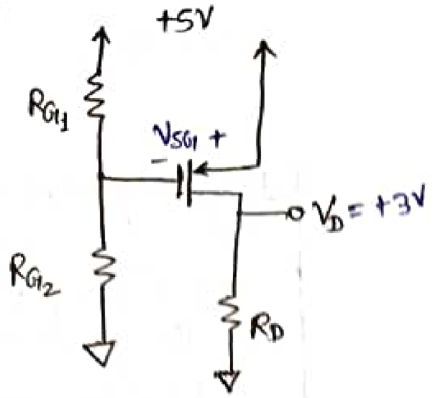
$V_D = 10 - (6K \times 0.5m) = 10 - 3 = 7V.$

$I_D = 0.5 mA$

$5 - I_D R_S - V_{GS} = 0$

21-08-2023

Q)



$I_D = 0.5\,mA$

$V_D = +3V$

$V_{TP} = -1V$

$k_p'\left(\frac{W}{L}\right) = 1\,mA/V^2$

① What is the value of R_D and V_G?

② What is the largest value of R_D that can be used while keeping the transistor in saturation region?

Sol^n: ① $R_D = \dfrac{3V - 0}{0.5mA} = 6\,K\Omega.$

$R_D = 6K\Omega$
$V_G = 3V$

and,

Using $I_D = \frac{k_p'}{2}\left(\frac{W}{L}\right)(V_{SG} - |V_{TP}|)^2$

$\Rightarrow \cancel{\tfrac{1}{2}} \times 10^{\cancel{-3}} = \cancel{\tfrac{1}{2}} \times 10^{\cancel{-3}} (V_{SG} - 1)^2$

$\Rightarrow V_{SG} - 1 = \pm 1$

$\Rightarrow V_{SG} = 2 \quad (\because V_{SG} \neq 0)$

and, $V_{SG} = V_{DD} - V_G$

$\Rightarrow V_G = 5 - 2 \quad (\because V_{DD} = 5V)$ supply voltage

$= 3V$

② For saturation region,

$\cancel{V_{GS}} \; V_{SD} \geq V_{SG} - |V_{TP}|$

$5 - V_D \geq 5 - 3 - 1$

$\Rightarrow 5 - V_D \geq 1$

$\Rightarrow V_D \leq 4$

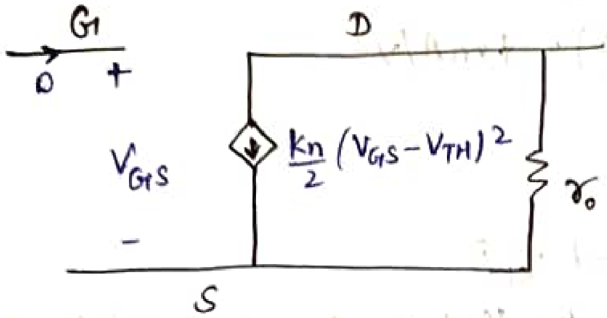
$\therefore R_D = \dfrac{V_D - 0}{I_D}$

$\leq \dfrac{4}{0.5m} = 8\,K\Omega$

$R_D = 8K\Omega$

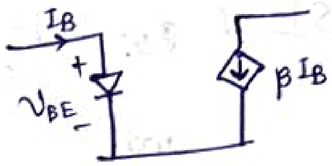
$\Rightarrow$ Largest possible $R_D = 8\,K\Omega.$

Large Signal Model of MOSFET

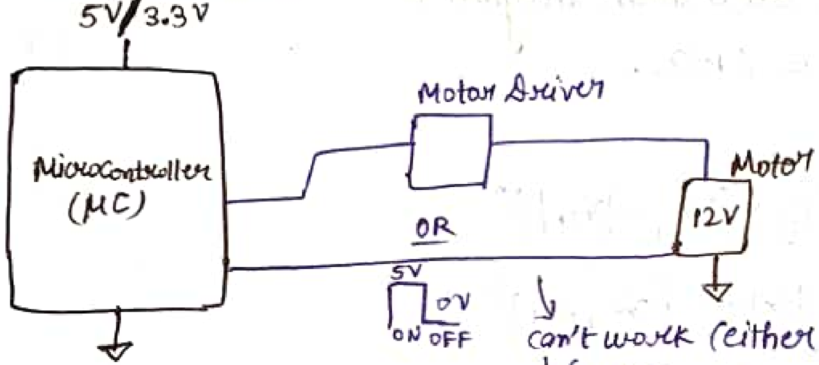


Large Signal Model
↳ can be used for both AC & DC
Small Signal Model
↳ can only be used for ac

BJT:

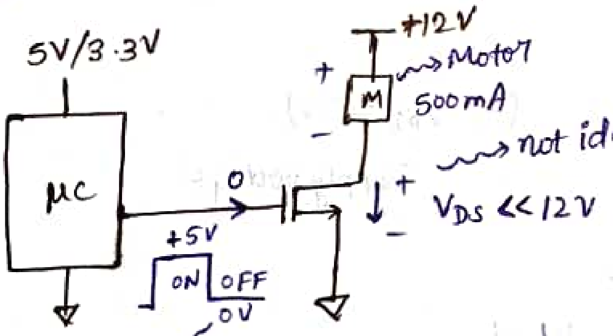


→ MOSFET as Switch:



can't work (either burn μC or not work)
↳ (as μC can't provide enough I)
even if motor is 5V

We can make it work even without a motor driver.



→ not ideal switch

OFF

+ 0V −

+ 12V −

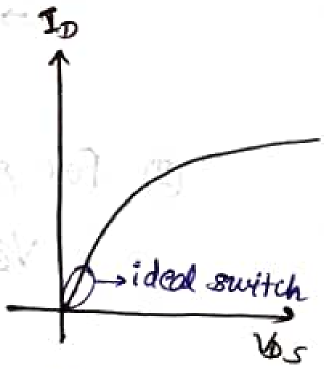
MOSFET

$V_{DS,max} > V_{supply}$

Motor won't rotate

$P_{diss.,M} = V_{DS} \times I_L$

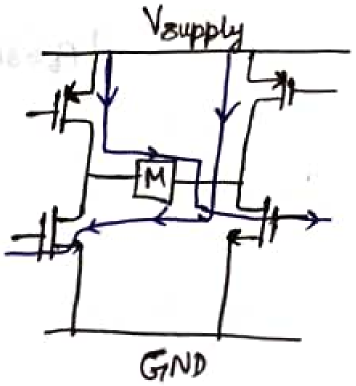
$= I_L^2 R_{ON}$



$I_D = k_n (V_{GS}-V_{TH}) V_{DS}$

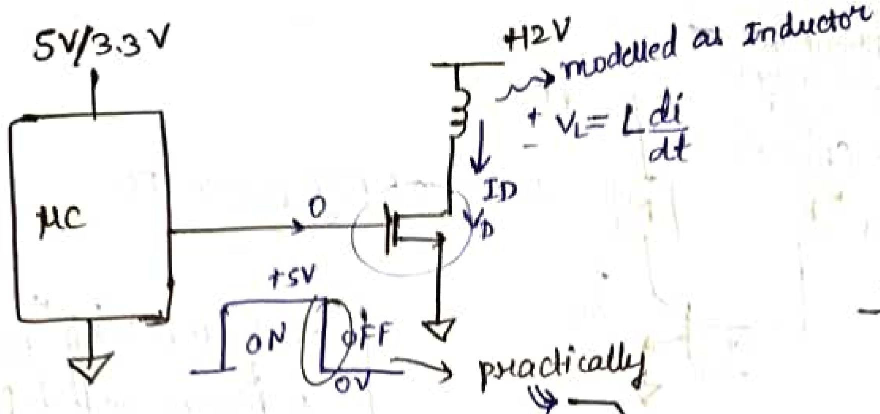
$\frac{V_{DS}}{I_D}$ $R_{ON} = \frac{1}{k_n(V_{GS}-V_{TH})}$

#

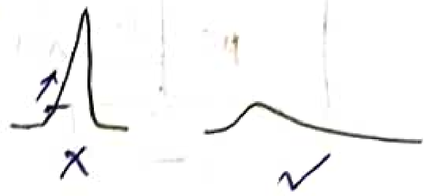


← Motor Driver (L293D)

→



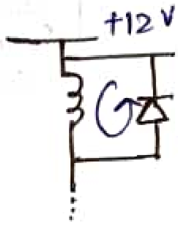
→ practically



At turn OFF

$\frac{di}{dt}$ → large -ve value ⇒ device may damage

To protect the device,

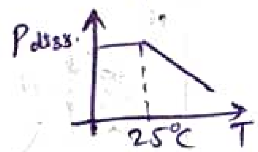


→ current circulates

22-08-2023

→ with T↑, power dissipation derates (↓) (above ambient temp., T_A).

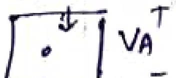
Eg.



→ Thermal Resistance :

Junction T = Ambient T + Thermal resistance ($R_{\theta JA}$)
(Room T)

Eg $T_A = 30°C$



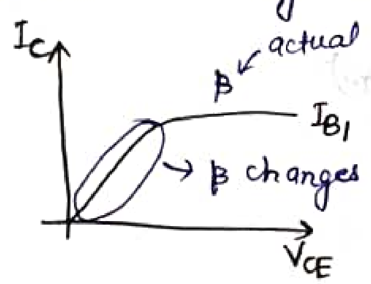
$P_{dissipated} = 10mW$

If $R_{\theta JA} = 150\ °C/W$,

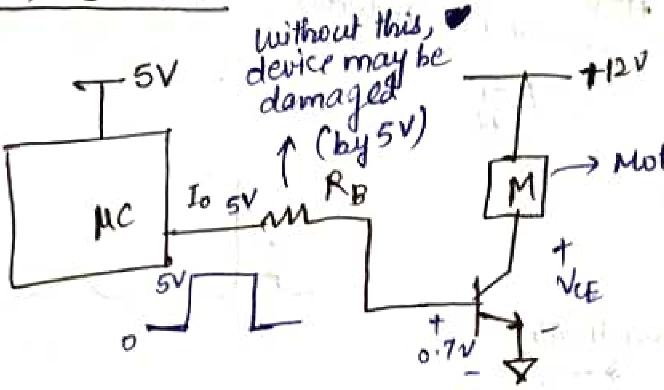
Junction $T = 30°C + 0.15°C$
$= 30.15°C$.

DATASHEET
BS170
MOSFET

→ In the saturation region of BJT, β decreases.



BJT as Switch

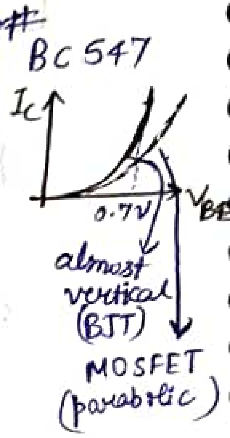


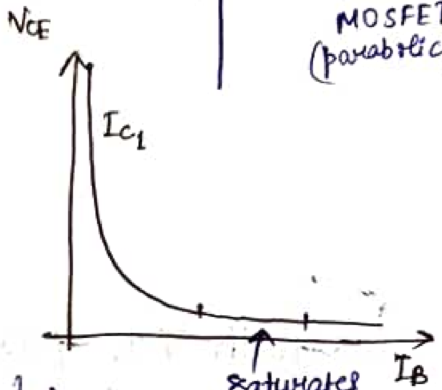
If $I_C = 50mA$ (say) has to be passed — $\beta_{forced} \approx 50$ (β gets changed from the original)
$I_B \approx 1mA$
$R_B = ?$

$$R_B = \frac{(5-0.7)V}{1mA} = 4.3 K\Omega$$

V_{CE} has to be less ↓ to have more voltage for $\boxed{M}$ & for power ↓

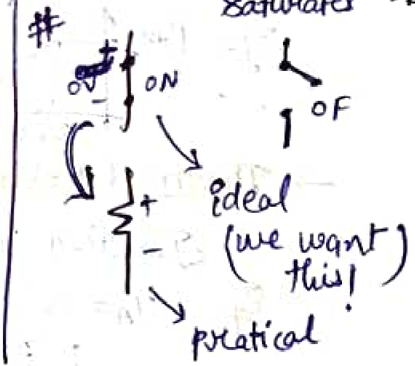
BC 547





- If $V_{in} = 0V$, $V_{BE} = 0$; cutoff region (turns OFF)

#



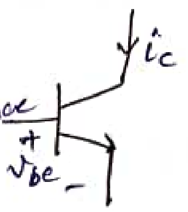
Small Signal Operation

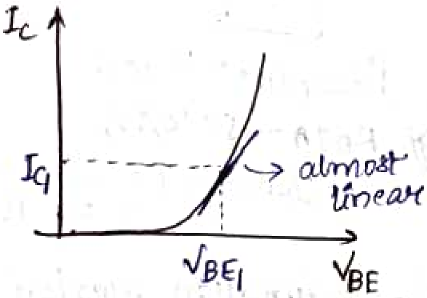
Condition (approximation): $|v_{ac}| << V_B$ ↳ dc voltage

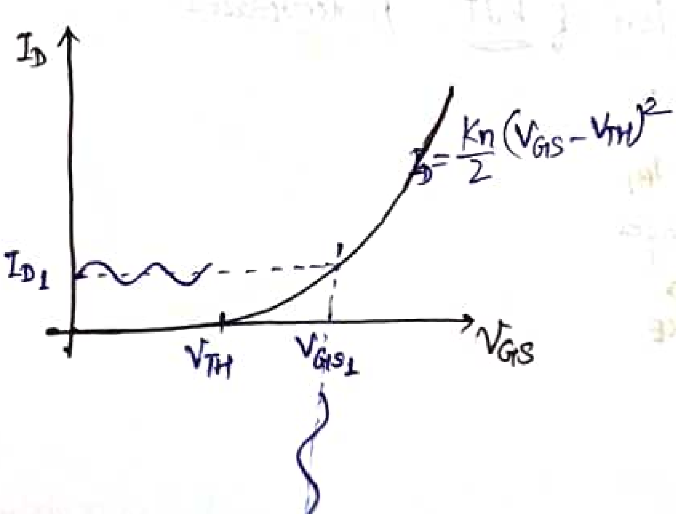
$$V_{in} = V_{BE_1} + v_{ac}$$

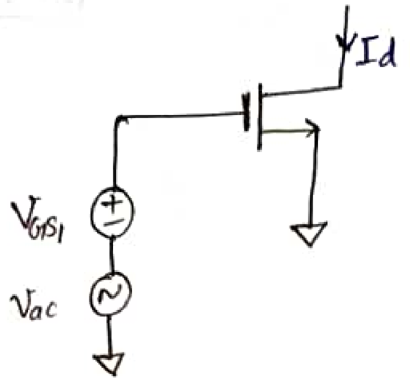
$$\frac{\partial I_C}{\partial V_{BE}} = g_m$$ ↳ transconductance

$$I_C = I_{C_1} + g_m v_{ac}$$









$$\boxed{V_{in} = V_{GS_1} + V_{ac}}$$

$$\boxed{g_m = \frac{\partial I_D}{\partial V_{GS}} = K_n (V_{GS} - V_{TH})}$$

$$\left[I_D = \frac{K_n}{2} (V_{GS} - V_{TH})^2 \right]$$

$$\boxed{\underbrace{I_d}_{\text{total current (ac+dc)}} = I_{D_1} + g_m V_{ac}}$$

$$\boxed{g_m = \frac{\partial I_D}{\partial V_{GS}} = \frac{2 I_D}{(V_{GS} - V_{TH})} = \sqrt{2 I_D K_n}}$$

As, $g_m = K_n (V_{GS} - V_{TH})$

$$= \frac{2 I_D}{(V_{GS} - V_{TH})^2} (V_{GS} - V_{TH})$$

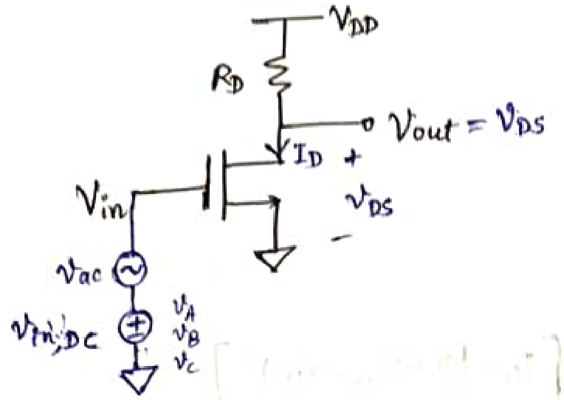
$$= \frac{2 I_D}{(V_{GS} - V_{TH})}$$

& $g_m = K_n \cdot \sqrt{\dfrac{2 I_D}{K_n}} = \sqrt{2 I_D K_n}$

28-08-2023

Common Source Amplifier

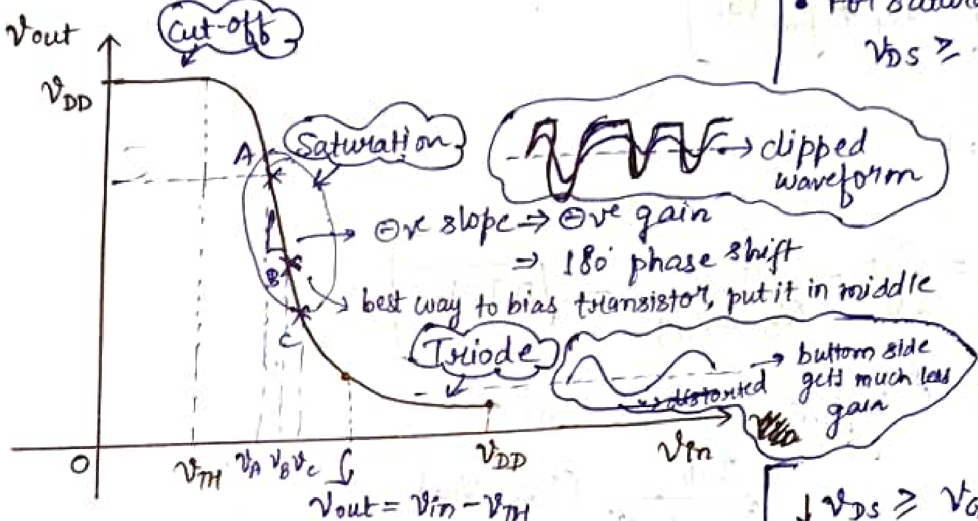
↳ Similar to Common Emmitter Amplifier of BJT



$$V_{out} = V_{DD} - I_D R_D$$

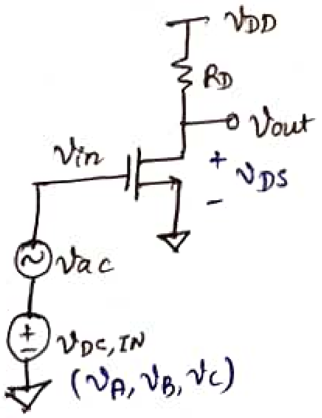
- $V_{in} = 0 \Rightarrow I_D = 0$
  ↳ cutoff region
  $\Rightarrow V_{out} = V_{DD}$
- For saturation,
  $V_{DS} \geq V_{GS} - V_{TH}$

Voltage Transfer Curve →



$[\downarrow V_{DS} \geq V_{GS}\uparrow - V_{TH}]$

→



- Since V_A is near V_{DD}, on high ac wave input, output will be clipped b/c V_{out} cannot be $> V_{DD}$.
- At V_C, since ⊖ve output cycle is near Triode region & hence that would not be amplified much due to less slope.
- So, the best biasing point is V_B, i.e., middle point V_{DD} and GND $\left(\frac{V_{DD}}{2}\right)$ for amplification.

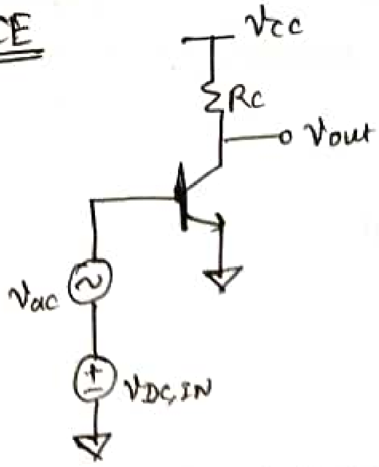
$$V_{out} = V_{DD} - I_d R_D$$
$$= V_{DD} - (I_{DC} + g_m V_{ac}) R_D$$

where $I_{DC} = \frac{K_n}{2} (V_{DC,IN} - V_{TH})^2$

$$g_m = K_n (V_{DC,IN} - V_{TH})$$

$$V_{out,DC} = V_{DD} - I_{DC} R_D$$

CE



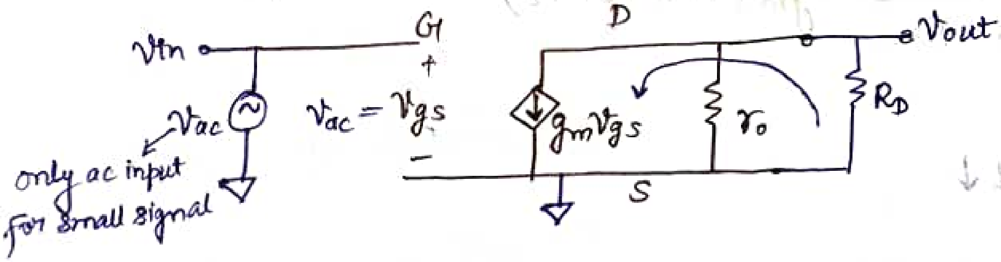
$$V_{out} = V_{cc} - I_{c} R_C$$
$$= V_{cc} - (I_{DC} + g_m V_{ac}) R_C$$

where,
$$I_{DC} = I_S \exp\left(\frac{V_{DC,IN}}{V_T}\right) \quad \& \quad g_m = \frac{I_C}{V_T}$$

CS: $\text{Gain} = -g_m R_D = \dfrac{-g_m V_{ac} R_D}{V_{ac}}$ $\left(\text{CS gain} = \dfrac{V_{out,dc}}{V_{in}} = -\dfrac{\overbrace{g_m . V_{ac}}^{I_{ac}} . R_D}{V_{ac}} \right)$

CE: $\text{Gain} = -g_m R_C$

Small Signal Model of MOSFET

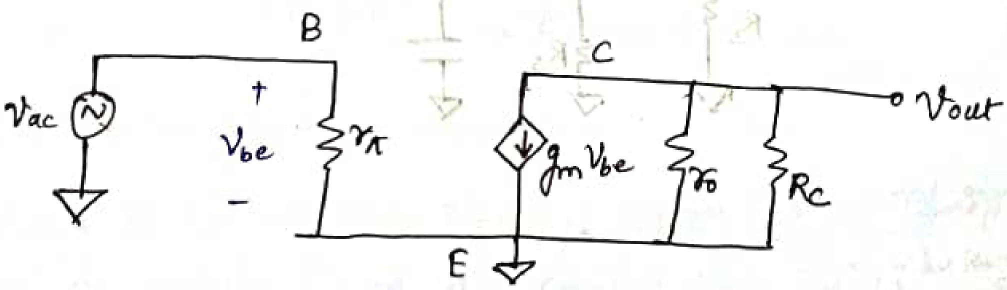


• DC voltages are AC grounds.

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

$$\text{Gain} = \frac{V_{out}}{V_{in}} = -g_m (r_o \| R_D) \approx -g_m R_D \quad (\text{for } R_D \ll r_o)$$

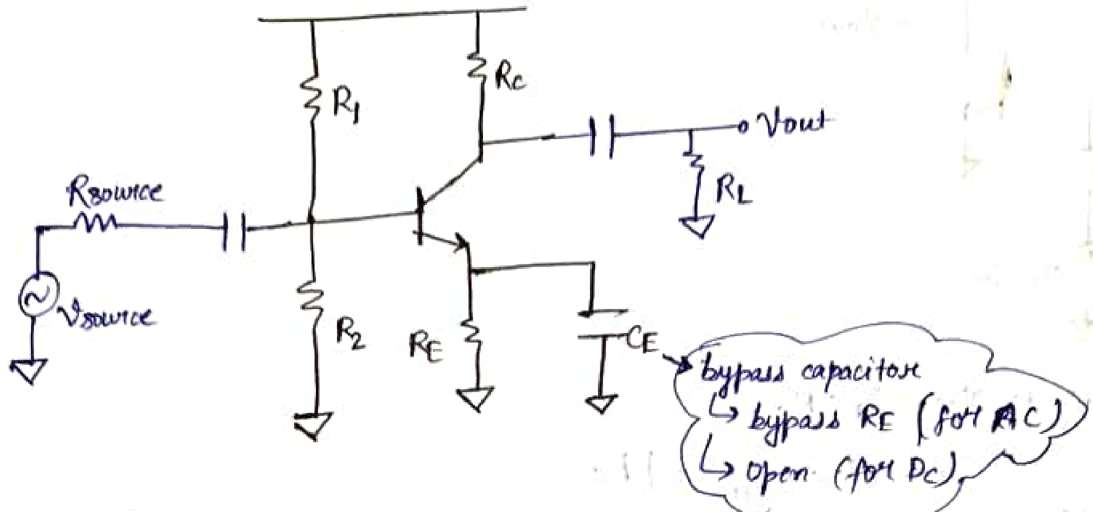
CE:

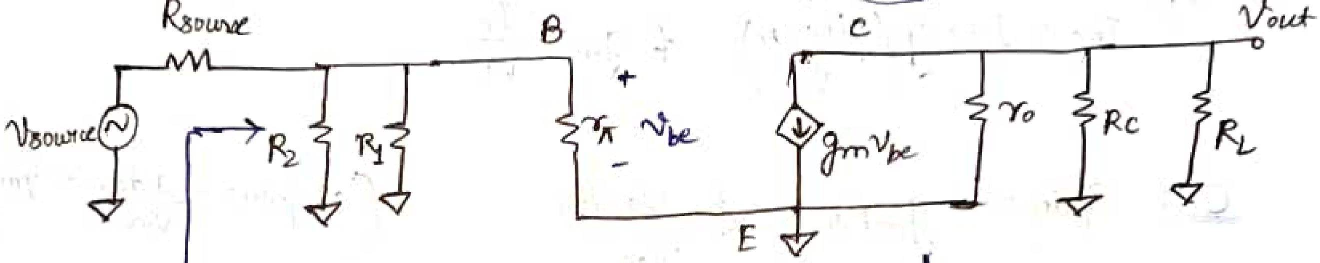


$$V_{out} = -g_m V_{be} (r_o \| R_C)$$

$$\text{Gain} = \frac{V_{out}}{V_{in}} = -g_m (r_o \| R_C) \approx -g_m R_C$$

Common Emitter Amplifier





$R_{in} = R_1 \parallel R_2 \parallel r_\pi$

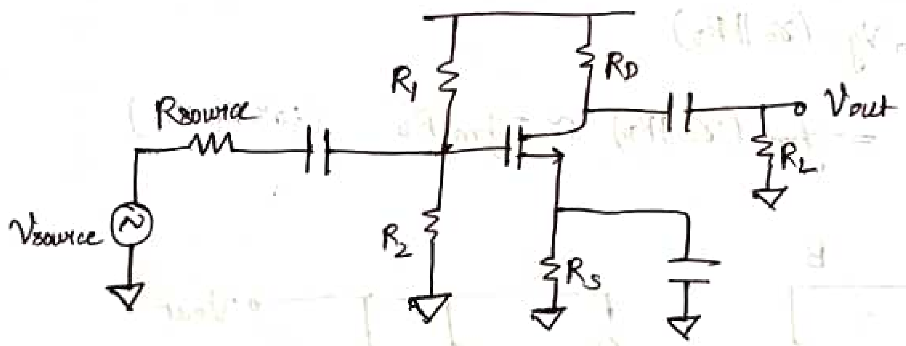
In ac analysis, treat capacitors to be short circuited.

$$V_{out} = \frac{V_{source} \cdot R_{in}}{R_{in} + R_{source}} \cdot g_m (r_o \parallel R_c \parallel R_L)$$

$$G_v = \frac{-R_{in}}{R_{in} + R_{source}} \cdot g_m (r_o \parallel R_c \parallel R_L)$$

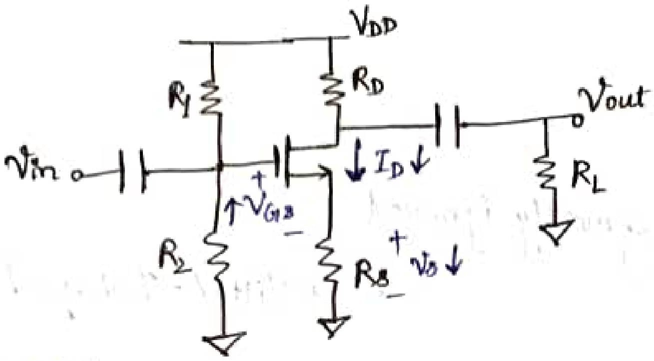
CS:

\# Temp. ↑ ⇒ current ↓

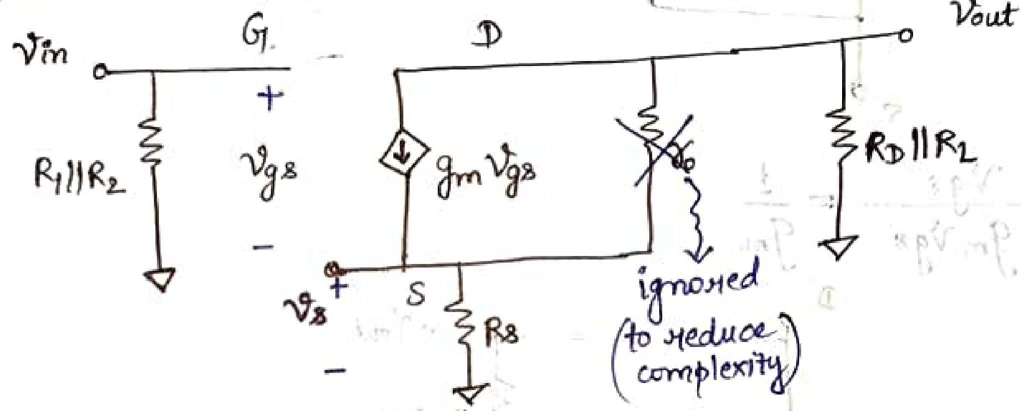


1-09-2023

Common Source Amplifier with Source Degeneration



SSM :



$R_{in} = R_1 \| R_2$

$v_{in} = v_{gs} + v_s$

$v_s = g_m v_{gs} R_S$

$v_{out} = -g_m v_{gs} (R_D \| R_L)$

- BJT and MOSFET have opposite behaviour on temperature dependence
- $T \uparrow \Rightarrow$ Mosfet current $\downarrow$

$$\frac{v_{out}}{v_{in}} = \frac{-g_m (R_D \| R_L)}{1 + g_m R_S}$$

⇝ reduced by $(1 + g_m R_S)$ wrt. if R_S is bypassed.
⇝ easier to control gain

If R_S is bypassed,

$$A_{CS} = -g_m (R_D \| R_L)$$ ⇝ harder to control gain

→ $I_D \downarrow \Rightarrow v_S \downarrow \Rightarrow V_{GS} \uparrow \Rightarrow$ Current $\uparrow$

→ R_S tries to correct the signal (Eg, in sinusoidal signal) (but cannot do fully), hence decreasing the gain (by factor of $\frac{1}{1 + g_m R_S}$).

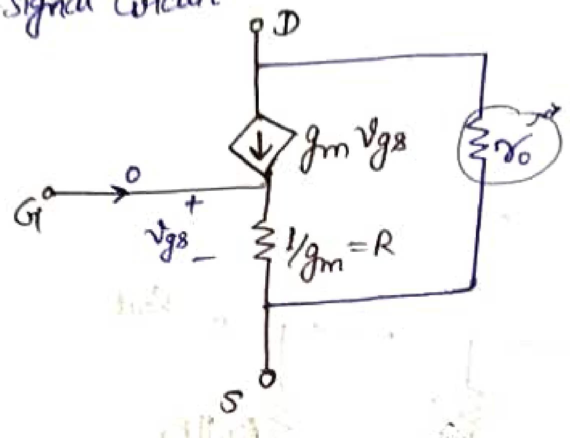
$$\text{Gain} = \frac{v_{out}}{v_{in}} = \frac{-g_m (R_D \| R_L)}{1 + g_m R_S} = \frac{-(R_D \| R_L)}{1/g_m + R_S} \approx -\frac{(R_D \| R_L)}{R_S}$$

$(\because R_S \gg \frac{1}{g_m})$

$\rightarrow A_v \approx \dfrac{\text{Total resistance @ Drain}}{\text{Total resistance @ Source}}$

T-Model of MOSFET

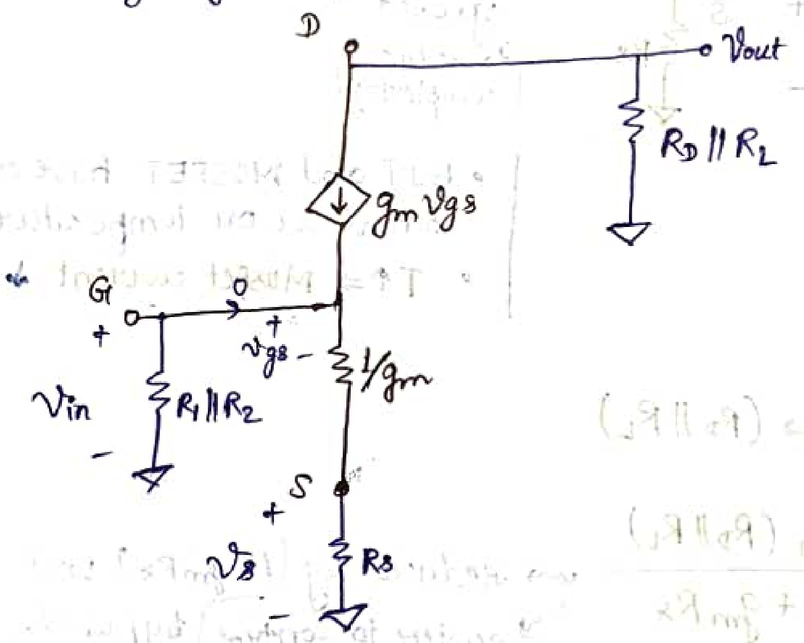
↳ Small-Signal Circuit



→ mostly ignored

⟶ Tilted-T shaped

$$R = \frac{v_{gs}}{g_m v_{gs}} = \frac{1}{g_m}$$

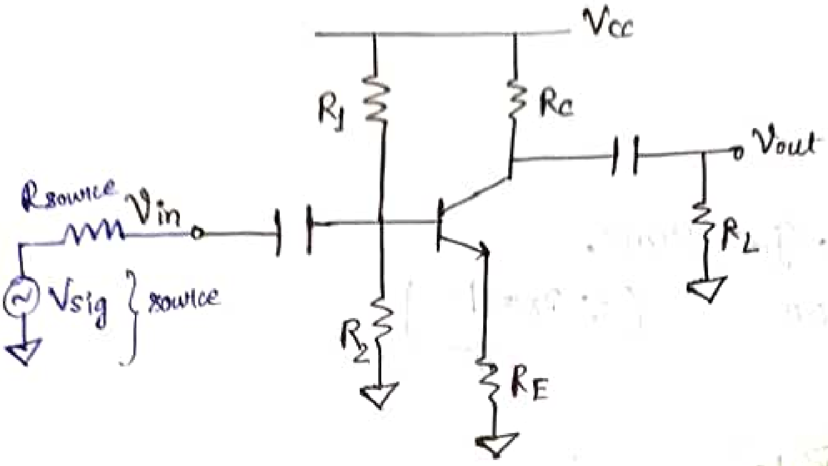


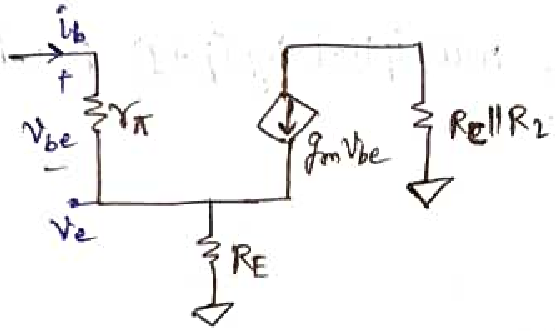
$$V_{out} = -g_m V_{gs} (R_D \| R_L)$$

$$V_{gs} = \frac{\cancel{R_S}\ 1/g_m\, V_{in}}{R_S + 1/g_m}$$

$$\frac{V_{out}}{V_{in}} = \frac{-g_m (R_D \| R_L) \cancel{V_{in}}}{1 + g_m R_S}$$

CE with Emitter Degeneration

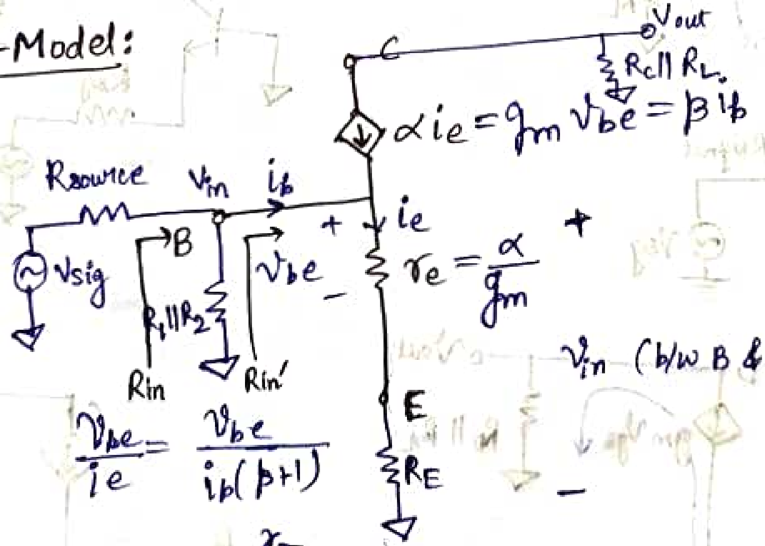




$$v_e = \left(\frac{v_{be}}{r_\pi} + g_m v_{be}\right) R_E$$

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

T-Model:



v_{in} (b/w B & GND)

$$\frac{v_{be}}{i_e} = \frac{v_{be}}{i_b(\beta+1)} = \frac{r_\pi}{\beta+1}$$

$$A_v = \frac{v_{out}}{v_{in}} \approx \frac{-g_m (R_c \parallel R_L)}{(1+g_m R_E)}$$

$$\approx -\frac{R_c \parallel R_L}{r_e + R_E}$$

$$i_e = \frac{v_{in}}{r_e + R_E}$$

$$i_b = \frac{i_e}{\beta+1}$$

$$R_{in} = \frac{v_{in}}{i_b}$$

$$i_b = \frac{v_{in}}{(\beta+1)(r_e+R_E)}$$

$$v_{out} = -\beta i_b (R_c \parallel R_L)$$

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$$\frac{v_{in}}{i_b} = R_{in}' = \frac{i_e(r_e + R_E)}{i_b}$$

$$\Rightarrow R_{in}' = (\beta+1)(r_e + R_E)$$
$$= r_\pi + (\beta+1)R_E$$

For $\beta = 100$, $R_E = 500\,\Omega$, $g_m = 40\,mS$,

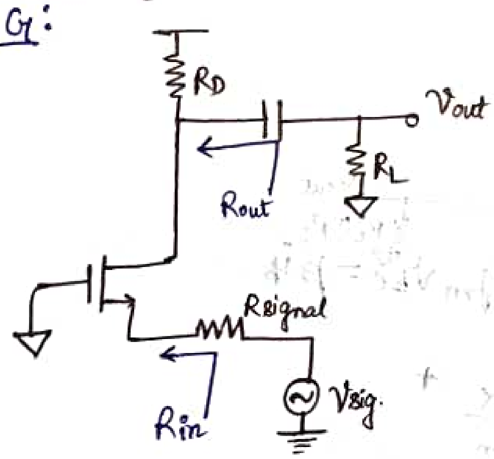
$R_{in}' = 2500 + 50500 \qquad \left(\because r_\pi = \frac{\beta}{g_m}\right)$

"Impedance Reflection Rule"

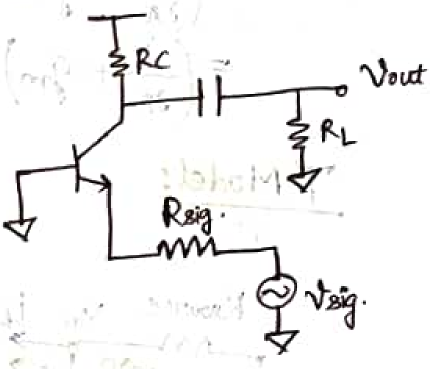
↳ Anything we put at emitter, multiplied by $(\beta+1)$ is what appears at the input side.

Common Gate / Common Base

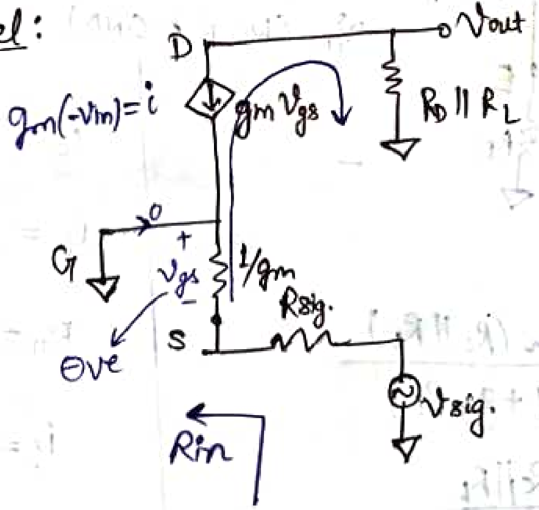
CG:

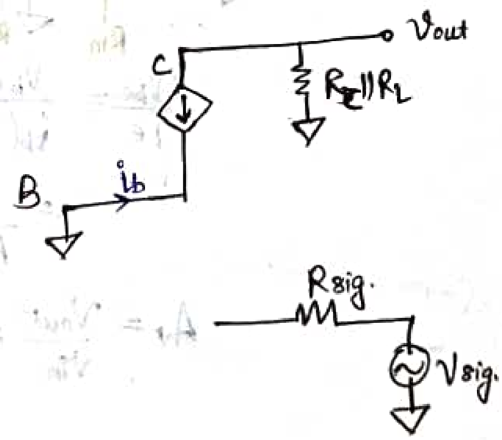


GB:



T-Model:





Common Gate

$R_{in} = 1/g_m$ ⇝ very low Input impedance (e.g., 25Ω)

$V_{out} = i\,(R_D \| R_L)$

$= \dfrac{V_{in}}{1/g_m}(R_D \| R_L)$

$A_v = \dfrac{V_{out}}{V_{in}} = g_m (R_D \| R_L)$

$G_v = \dfrac{V_{out}}{V_{sig}} = \dfrac{1/g_m}{1/g_m + R_{sig}} \times g_m (R_D \| R_L)$

Overall gain

$= \dfrac{R_D \| R_L}{1/g_m + R_{sig}} \cdot R_{in}$

Common Base

$R_{in} = \dfrac{V_e}{i_e} = \dfrac{\alpha}{g_m} \approx \dfrac{1}{g_m}$

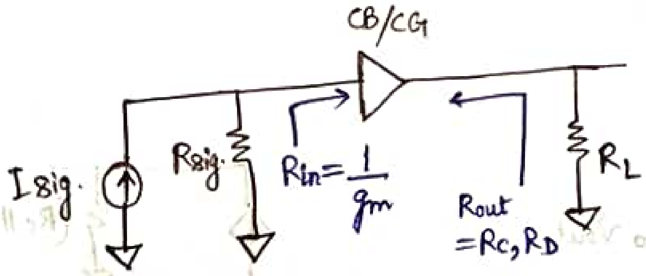
$V_{out} = \dfrac{V_{in}}{\alpha/g_m}(R_C \| R_L)$

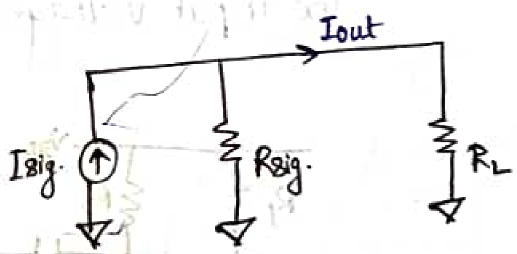
$A_v \approx g_m (R_C \| R_L)$

Amplifier used with today's transmission line (having low chrts. impedance, $Z_0 \sim 50\text{-}70\Omega$) needs to have low input impedance, i.e., CG/CB amplifier.

Output Impedance, $R_{out} = R_D$ (∵ $R_L \rightarrow$ load impedance)

Current Buffer





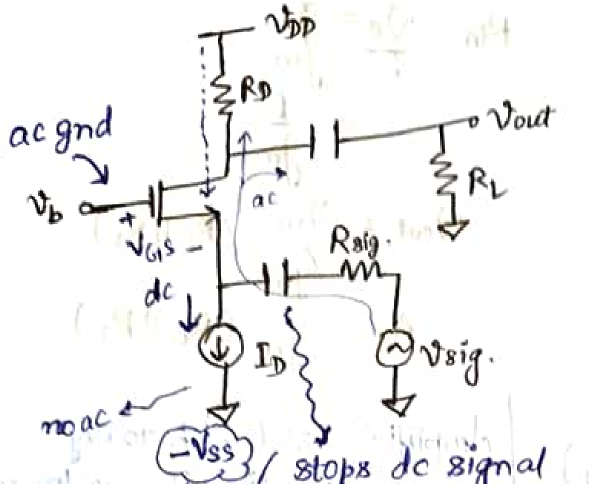
$I_{in} = \dfrac{R_{sig}}{R_{sig} + 1/g_m} \cdot I_{sig}$

$I_{out} = \dfrac{R_D}{R_D + R_L} \times I_{in}$

$I_{out} = I_{sig} \cdot \dfrac{R_{sig}}{R_{sig} + R_L}$

08-09-2023

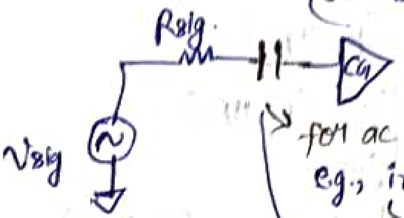
CG/CB Amplifier:



transistor will work even with this

(stops dc signal
dc current may damage the sensors)

ac coupling → only ac signal is coupled in the ~~transis~~ amplifier (dc blocked)

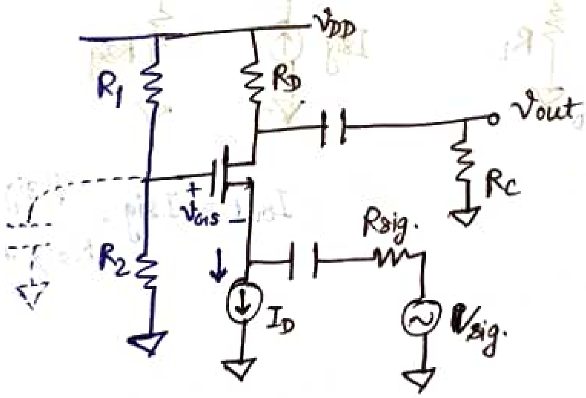


→ for ac coupling
e.g., in microphone
↳ dc has no info
→ not required in temperature sensor (which doesn't change much)

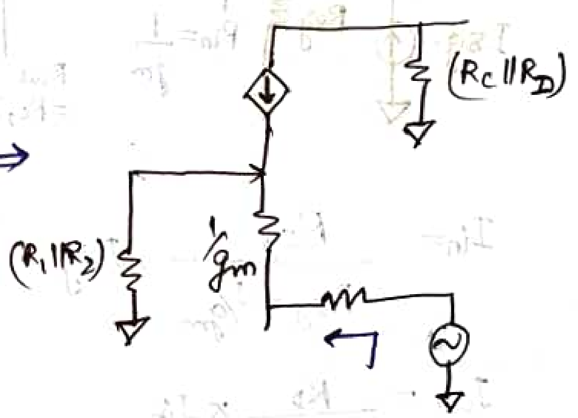
Quiscent Condition

↳ Current consumption (dc voltages, etc.) of the ~~circu~~ circuit when no input is provided.

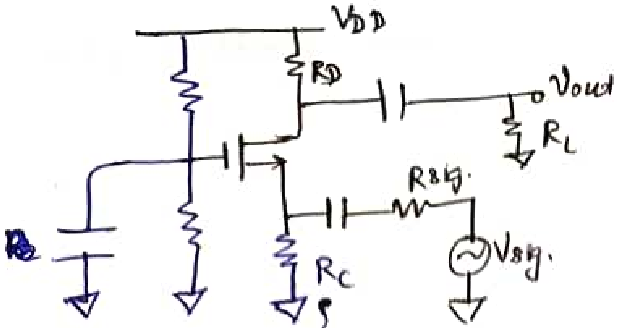
→ For the transistor to work, it must be biased (like engine turned ON in a car) (having certain g_m), no matter what the input voltages are.



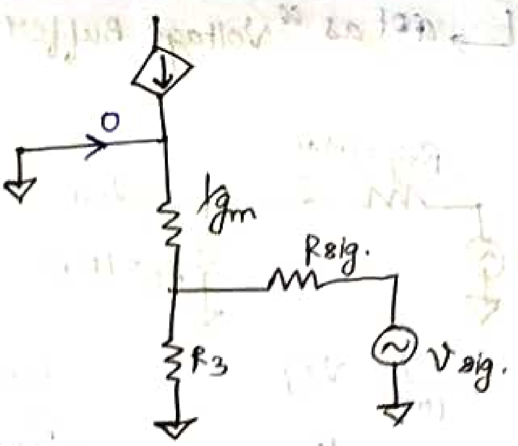
⇒



For CG amplifier,
G → ac gnd
↳ Transistor won't work.



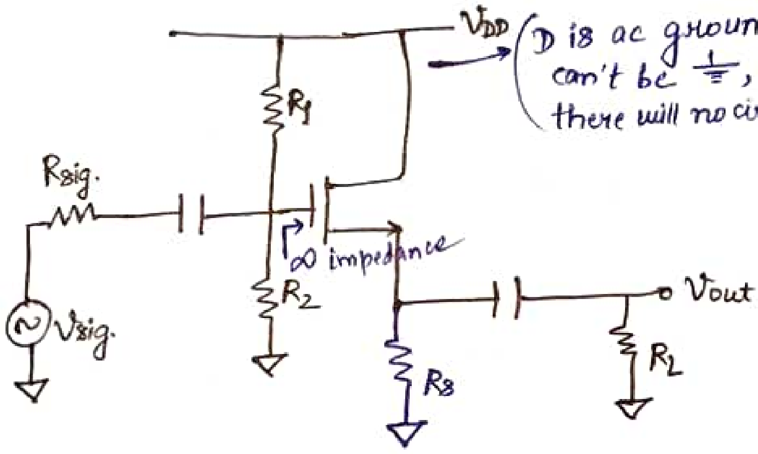
$\Rightarrow$



(Instead of current source, a resistor or a saturated transistor can be used) $\rightsquigarrow$ (large value of R doesn't allow current to change fast)

$$R_{in} = \frac{1}{g_m} \,\|\, R_3$$

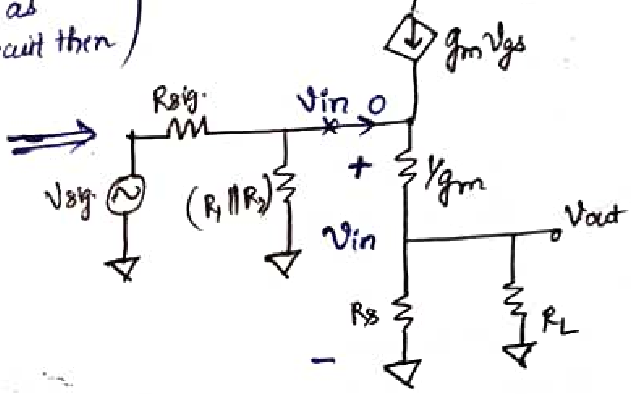
Common Drain / Common Collector Amplifier



(D is ac grounded can't be $\frac{1}{\mp}$, as there will no circuit then)

$\Longrightarrow$

T-Model:



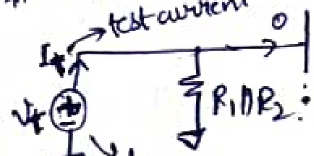
$$R_{in} = R_1 \,\|\, R_2$$

$$A_v = \frac{V_{out}}{V_{in}} = \frac{R_s \,\|\, R_L}{R_s \,\|\, R_L + 1/g_m}$$

$$G_v = \frac{V_{out}}{V_{sig}}$$

$$= \frac{R_1 \,\|\, R_2}{R_1 \,\|\, R_2 + R_{sig}} \cdot \frac{R_s \,\|\, R_L}{R_s \,\|\, R_L + 1/g_m}$$

Rin calculation:



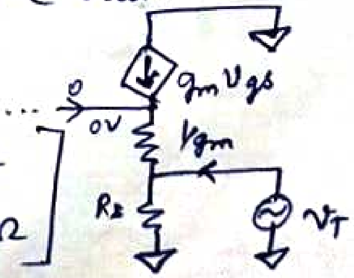
$$I_T = \frac{V_T}{R_1 \,\|\, R_2}$$

$$R_{in} = \frac{V_T}{I_T} = R_1 \,\|\, R_2$$

To find Rout.

$V_{sig} = 0$ & find $\frac{V_T}{I_T}$ by connecting V_T @ Vout.

$$R_{out} = R_s \,\|\, 1/g_m \approx 1/g_m \quad (\because R_s \gg 1/g_m)$$

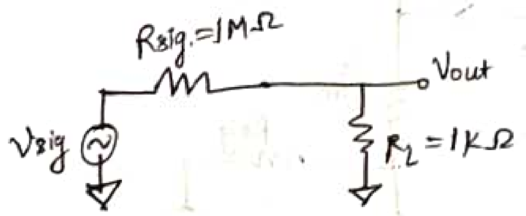


$$A_v = \frac{R_s \,\|\, R_L}{R_s \,\|\, R_L + 1/g_m} \approx 1$$

$$\left[\begin{array}{l} R_s, R_L \sim k\Omega \\ 1/g_m \sim 10's\ \Omega \end{array}\right]$$

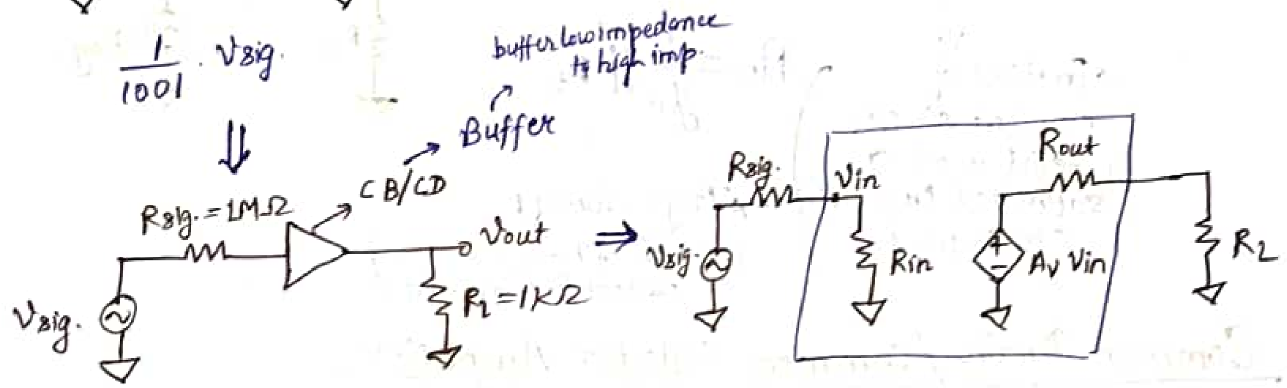
$(A_v < 1)$

↳ act as "Voltage Buffer".



$\frac{1}{1001} \cdot V_{sig}$

⇓



Common Collector (CC):

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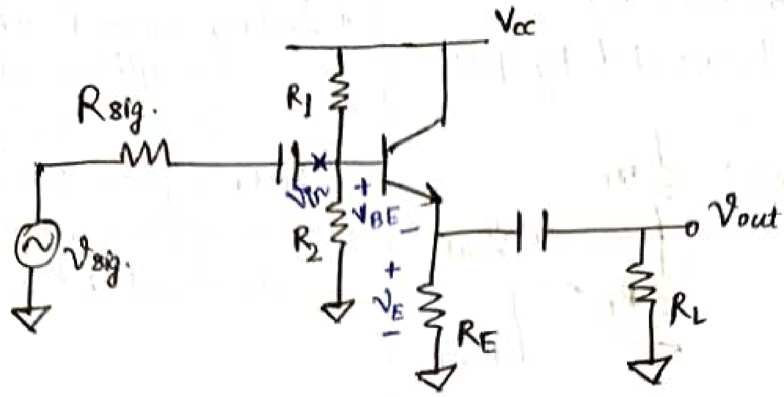
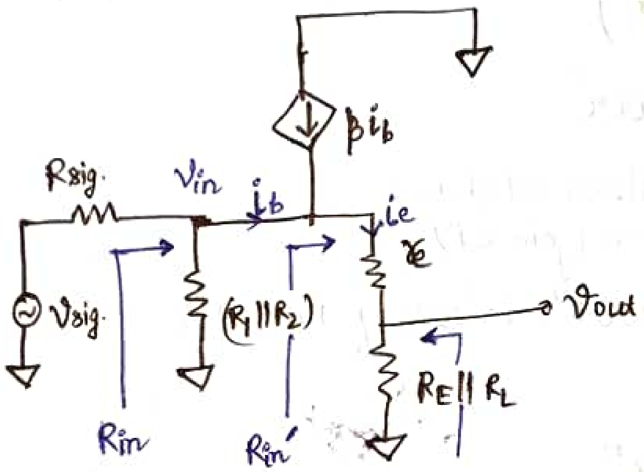
CC:
input = B
output = E



T-Model:



$$R_{in}' = \frac{v_{in}}{i_b}$$

$$i_e = \frac{v_{in}}{r_e + (R_E \| R_L)}$$

$$i_e = (\beta + 1)\, i_b$$

... Impedance Reflection rule

$$R_{in}' = (\beta+1)(r_e + R_E \| R_L)$$

$$v_{out} = v_{in} \frac{R_E \| R_L}{r_e + (R_E \| R_L)} \quad \text{(voltage divider rule)}$$

$$\Rightarrow A_v = \frac{v_{out}}{v_{in}} = \frac{R_E \| R_L}{r_e + (R_E \| R_L)}$$

$$\approx 1$$
$$< 1$$

$$R_{in} = R_1 \| R_2 \| (\beta+1)(r_e + (R_E \| R_L))$$

$$G_v = \frac{v_{out}}{v_{sig}} = \frac{R_{in}}{R_{in} + R_{sig}} \cdot \frac{R_E \| R_L}{r_e + (R_E \| R_L)}$$

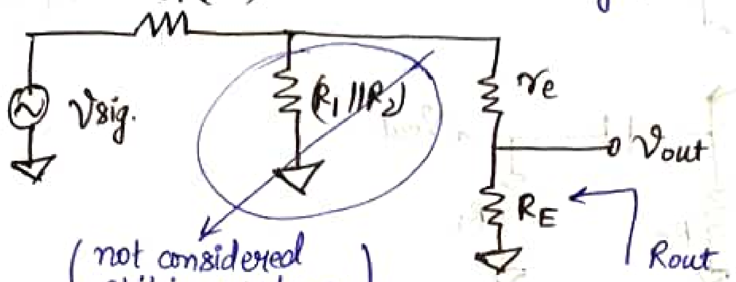
Set $v_{sig} = 0$

$$R_{out} \approx r_e \| R_E \approx r_e$$

Thevenin Equivalent Circuit:

(As seen from ~~output~~ Emitter side)

$R_{sig}/(\beta+1)$ ⇝ downscaled by $(\beta+1)$



(not considered as it is very large)

$v_{sig} = 0$

$$R_{out} = R_E \,\|\, \left(r_e + \frac{R_{sig}}{\beta+1}\right)$$

↳ "Emitter Follower"

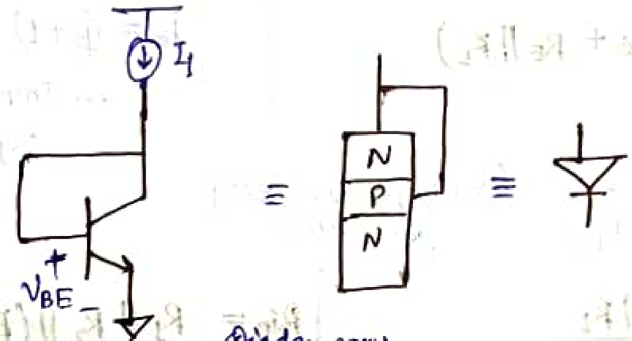
↳ Emitter follows whatever is at base (gain ≈ 1)

"Source Follower" in case of MOSFET

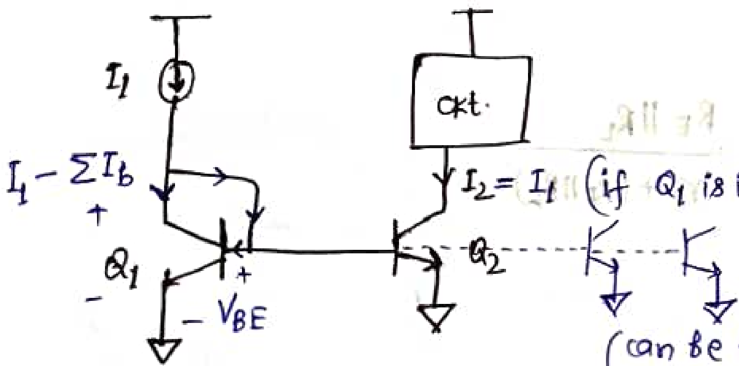
In BJT,

- Looking from B-side, everything at E-side is scale by $(\beta+1)$
- Looking from E-side, everything at B-side is downscaled by $(\beta+1)$.

→ "Diode connected transistor"



~~Diode conv~~ Transistor converted to diode.



$I_2 = I_1$ (if Q_1 is identical to Q_2)

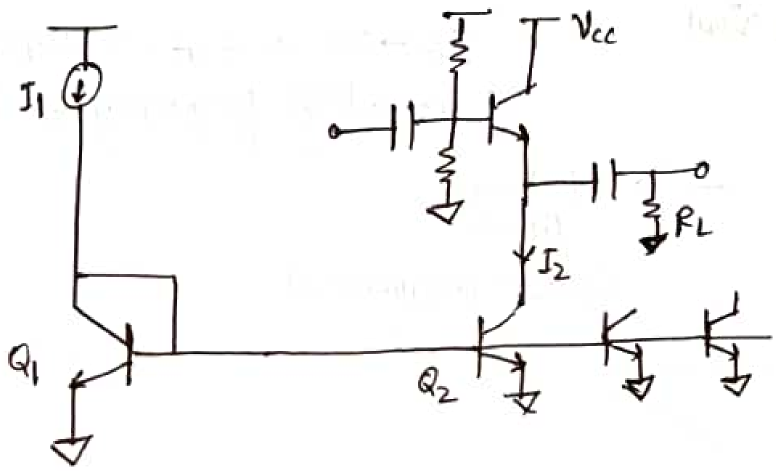
(can be mirrored multiple times) → As BJT needs I_B current getting mirrored is $(I_1 - \Sigma I_B)$.

↳ no such case in MOSFET

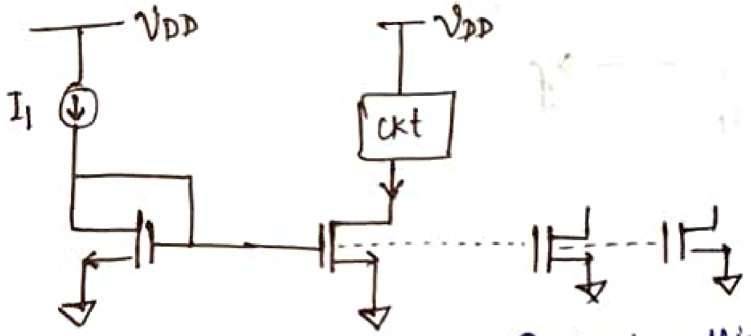
↳ "Current Mirror"

$V_{CE} = V_{BE} \approx 0.6 - 0.7 V$
$> 0.3 V$

#



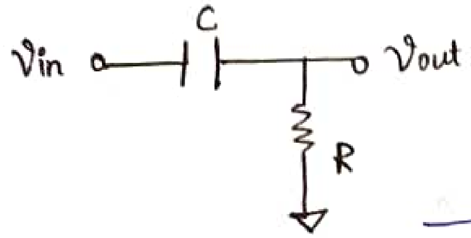
→



Current getting mirrored = I_1.

12-09-2023

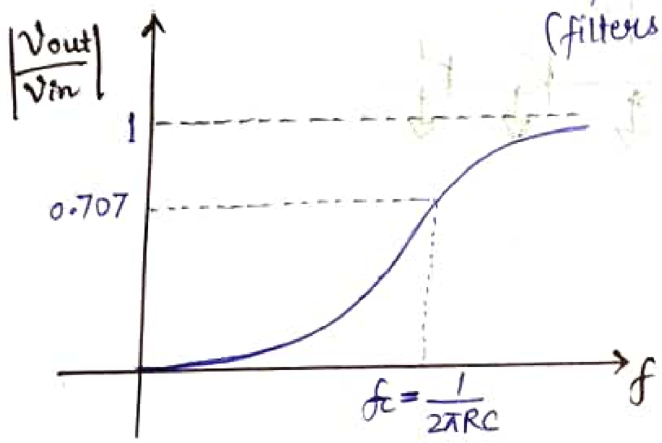
FREQUENCY RESPONSE OF AMPLIFIERS



$V_{in} \rightarrow DC \Rightarrow V_{out} = 0$ (open)

$V_{in} \rightsquigarrow$ high frequency $\Rightarrow V_{out} = V_{in}$
(short)

$\longrightarrow$ Electronic filter
(filters frequencies)



$$V_{out} = \frac{R}{R + \frac{1}{j\omega C}} V_{in}$$

$$\frac{V_{out}}{V_{in}}(j\omega) = \frac{j\omega RC}{1 + j\omega RC}$$

$$\left|\frac{V_{out}}{V_{in}}\right| = \frac{\omega RC}{\sqrt{1 + (\omega RC)^2}}$$

$|\ \omega = 2\pi f$

$|\ \cancel{\omega = \frac{1}{RC}}$

$\hookrightarrow$ High Pass Filter

$\hookrightarrow$ Passes high frequency

$$\left|\frac{V_{out}}{V_{in}}\right| = \frac{1}{\sqrt{2}} \quad \text{when } \omega = \frac{1}{RC}$$

$$\Rightarrow \frac{P_{out}}{P_{in}} = \left|\frac{V_{out}}{V_{in}}\right|^2 = \frac{1}{2}$$

"Bandwidth"

→ Quantity in dB $= 10 \log P$ or $10 \log \frac{P_{out}}{P_{in}}$

$$\frac{V_{out}}{V_{in}} = 100$$

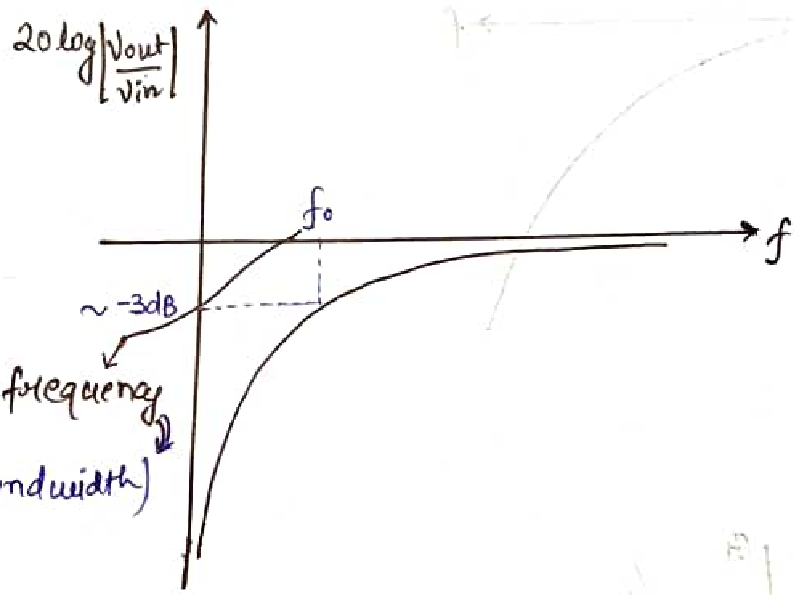
$$\frac{P_{out}}{P_{in}} = 10^4$$

$$10 \log 10^4 = 40\,dB$$

OR

$$20 \log \frac{V_{out}}{V_{in}}$$

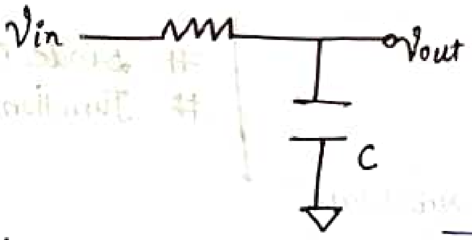
→



Corner frequency

(3dB bandwidth)

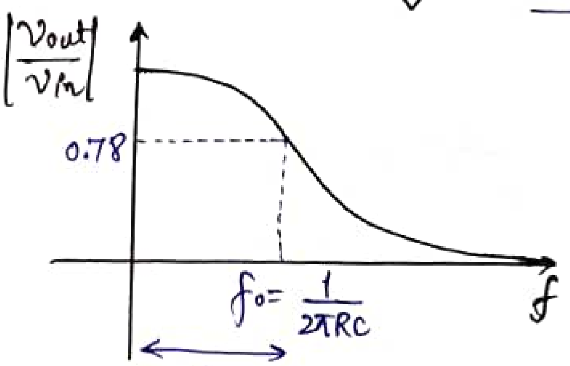
→



at DC, $V_{out} = V_{in}$

@ $f >>$, $V_{out} = 0$

→ Allow low frequencies

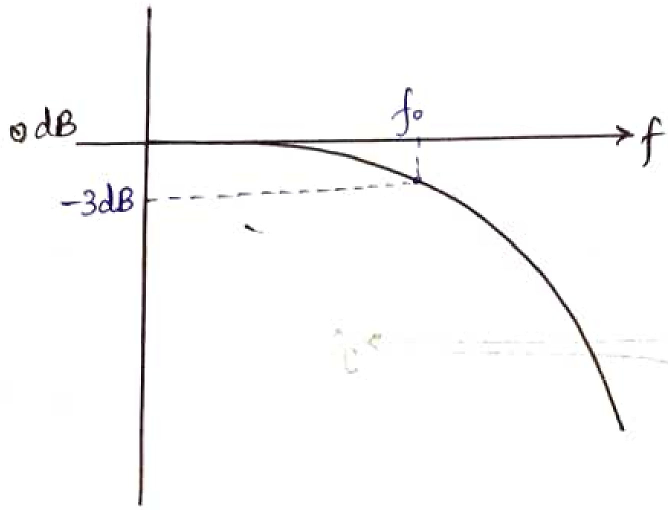


↳ bandwidth: frequency range over which signal is passed.

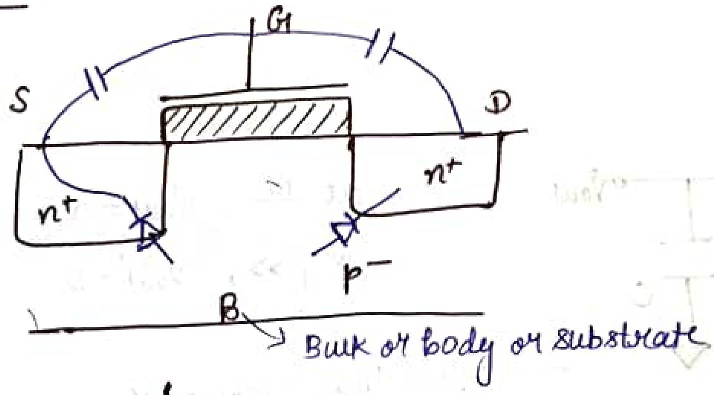
$$\frac{V_{out}}{V_{in}}(j\omega) = \frac{\frac{1}{j\omega C}}{R + \frac{1}{j\omega C}} = \frac{1}{1 + j\omega RC}$$

$$\Rightarrow \left| \frac{V_{out}}{V_{in}}(j\omega) \right| = \frac{1}{\sqrt{1 + (\omega RC)^2}}$$

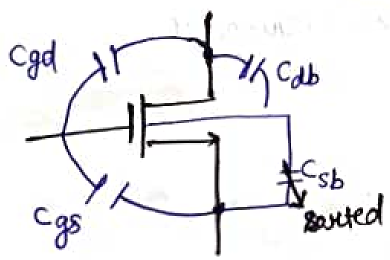
$$\Rightarrow \left| \frac{V_{out}}{V_{in}}(j\omega) \right|_{\omega = \frac{1}{RC}} = \frac{1}{\sqrt{2}}$$



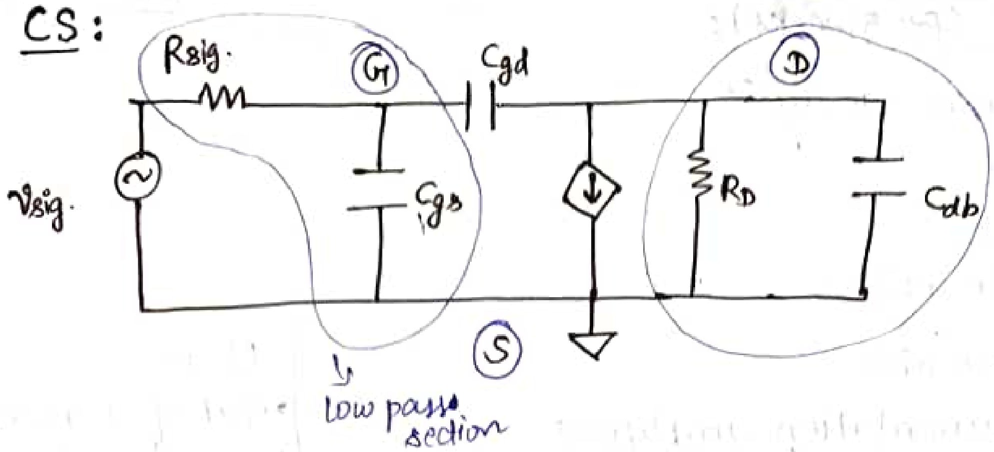
MOSFET:



\# Diode capacitance
\# Junction capacitance

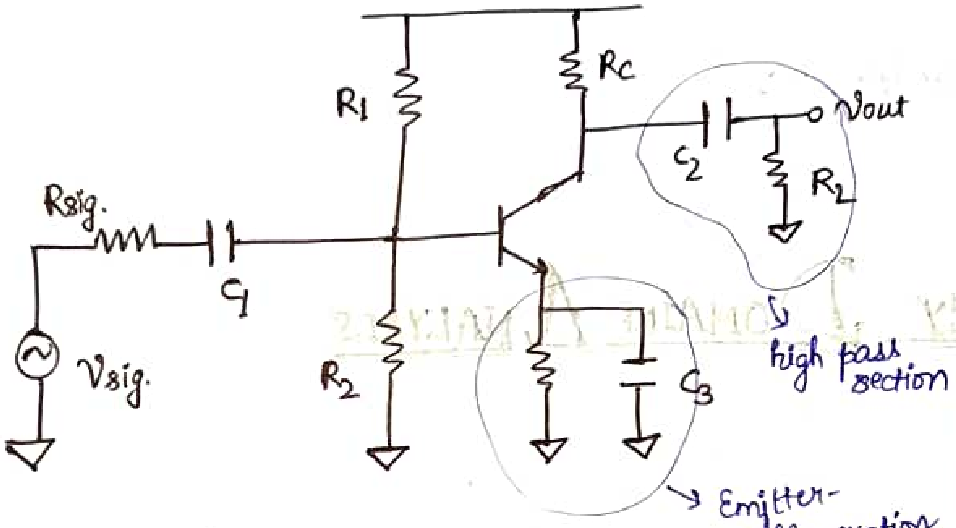


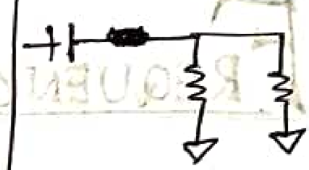
CS:



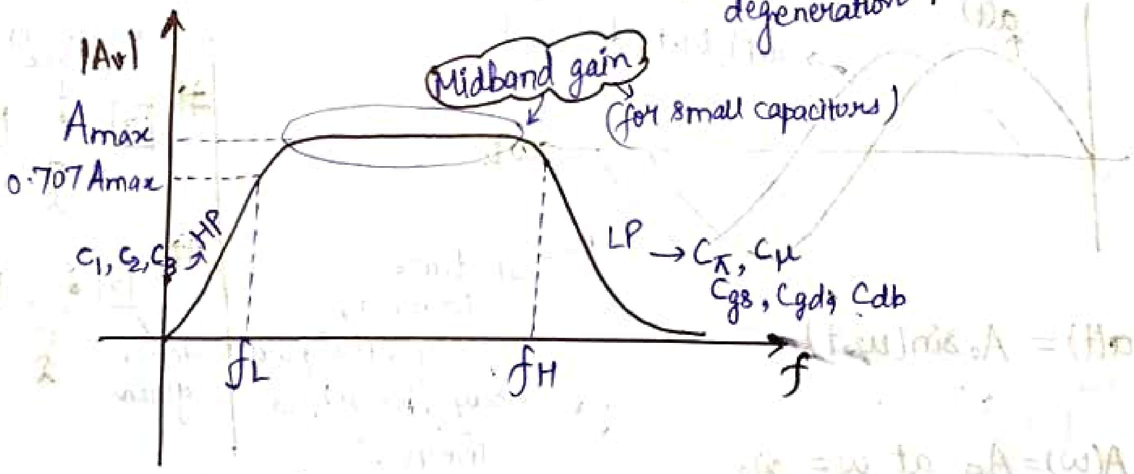
→ Low pass behaviour

• No device has infinite bandwidth





↳ High pass



↳ Band of frequency, over which it amplifies as intended (amplifier functions)

↳ "Bandpass"

$$\boxed{f_H - f_L = BW}$$

18-09-2023

Topics till Quiz 2 (for 5 weeks):

① Frequency domain analysis

② Opamps
- Circuits
- Non-linearities

③ Precision electronics
- Instrumentation amplifiers

④ Filters
- RC
- 1st order

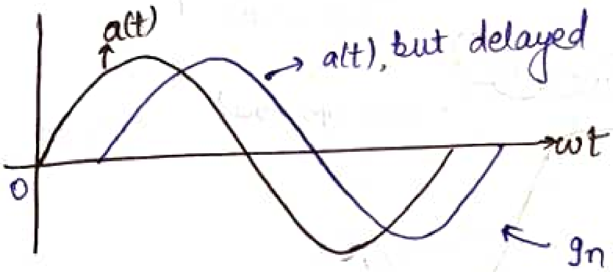
⑤ Oscillators

Book:
- Art of Electronics
- Boylestad

FREQUENCY DOMAIN ANALYSIS

Review

①



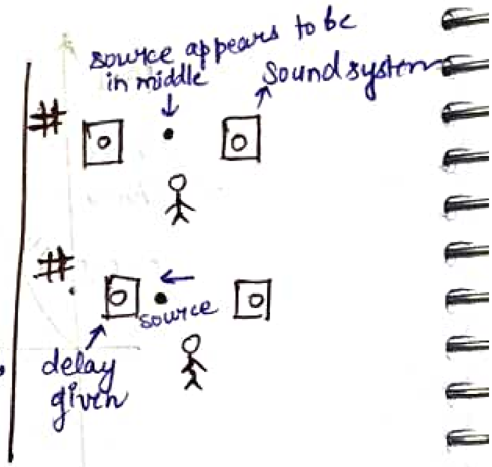
← In time domain, only real part, say Re(A), is there.

$a(t) = A_0 \sin(\omega_0 t)$

or

$A(\omega) = A_a$ at $\omega = \omega_0$

phase $< \theta_0 = 90^\circ$



② Frequency as complex no.:

$$s = \sigma_0 + j\omega_0$$

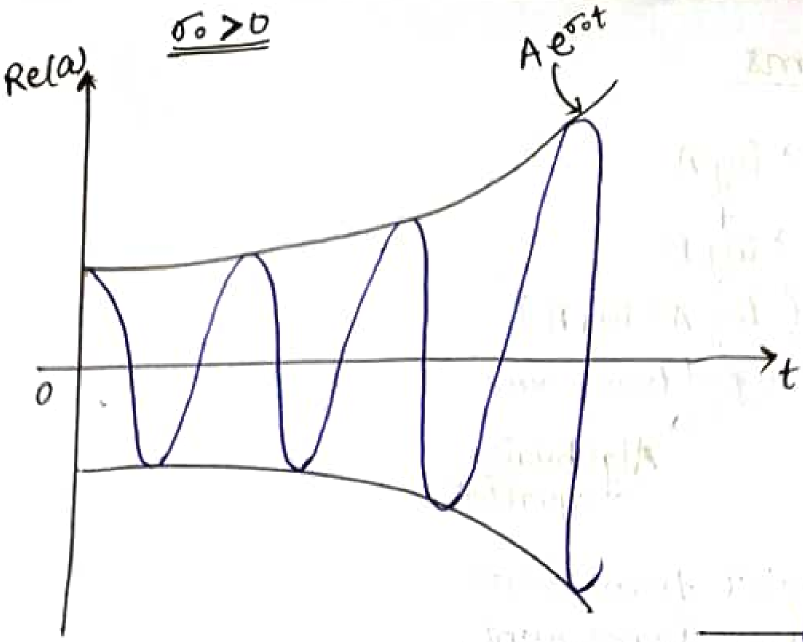
σ_0: rate of increase/decay of amplitude → constitutes damping

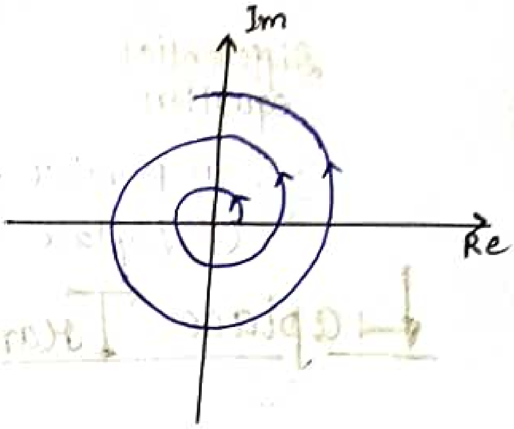
ω_0: frequency → natural freq.

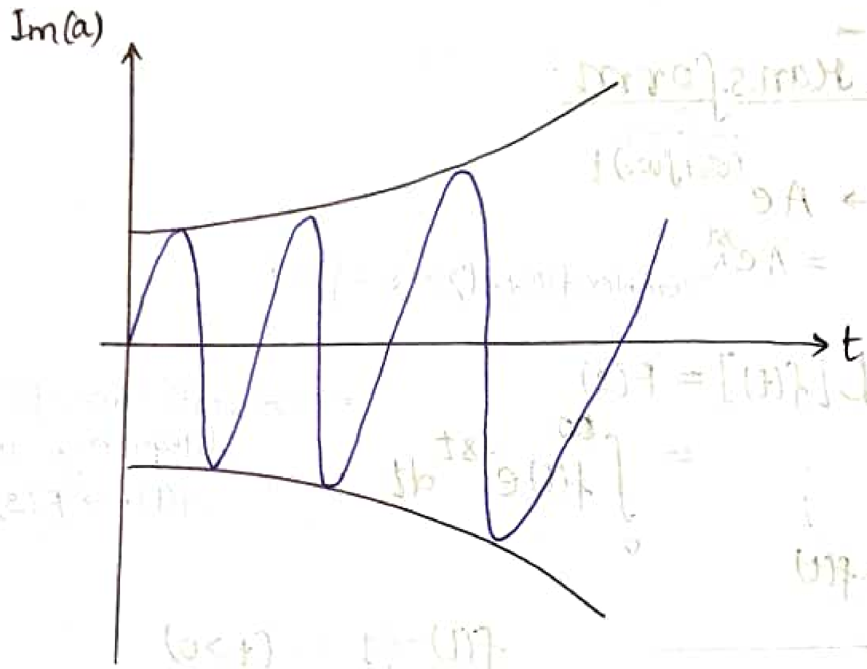
$$a(t) = A\, e^{\overbrace{(\sigma_0 + j\omega_0)}^{\text{complex freq.}} t}$$

← Frequency domain
- both Re & Im part
- take only Re part to get time-domain

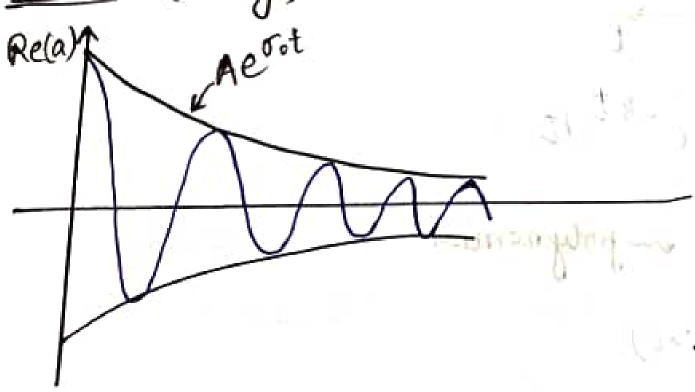
$\underline{\sigma_0 > 0}$



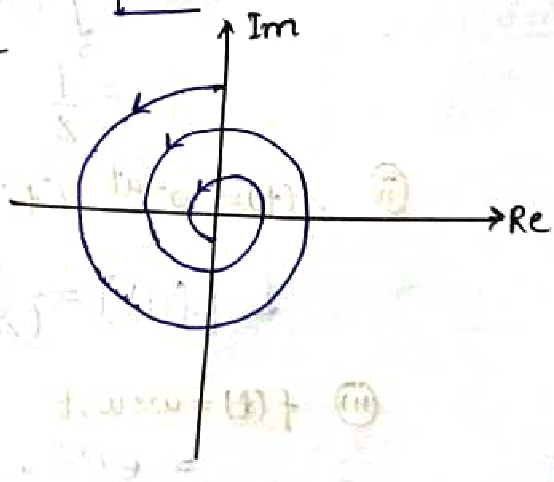


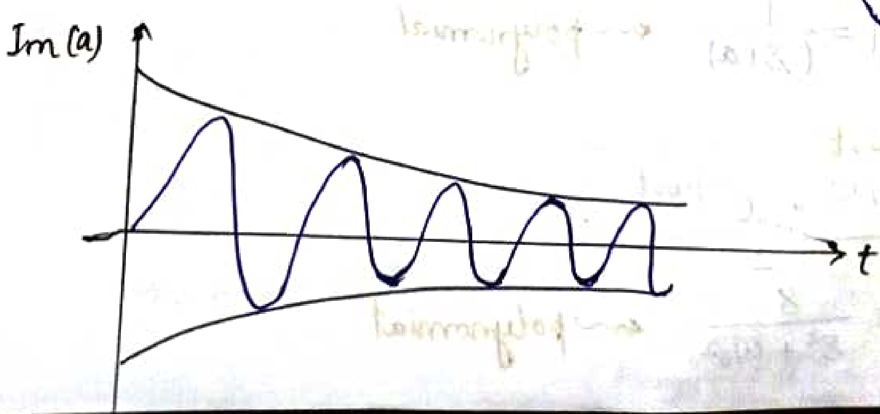


$\underline{\sigma_0 < 0}$ (decay)



Overlapping Re(a) & Im(a)
↓
Helix





Use of Transforms

$$A \longrightarrow \log A$$
$$+$$
$$B \longrightarrow \log B$$
$$A*B \longleftarrow (\log A + \log B)$$

Integral transforms

Differential equation $\longrightarrow$ Algebraic equation

Eg. ① Fourier transform
② Laplace transform.

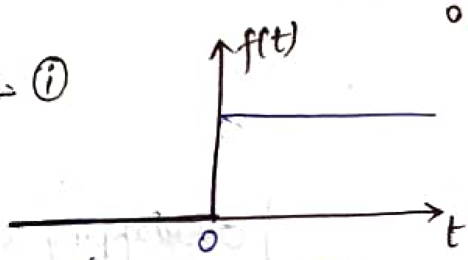
Laplace Transform

$$a(t) \longrightarrow Ae^{\overbrace{(\sigma_0 + j\omega_0)}^{s}t}$$
$$= Ae^{st}$$ complex freq. $(s = \sigma_0 + j\omega_0)$

$$f(t) \longrightarrow \mathcal{L}[f(t)] = F(s)$$
$$= \int_0^{\infty} f(t)e^{-st}\,dt$$

→ converts time function to frequency function $f(t) \longrightarrow F(s)$

Eg. (i)



$$f(t) = \begin{cases} 1 & (t>0) \\ 0, & \text{otherwise} \end{cases}$$

$$F(s) = \int_0^{\infty} 1 . e^{-st}\,dt$$
$$= \frac{1}{s}$$ ~ polynomial

(ii) $f(t) = e^{-at}$ $(t>0)$

$$\mathcal{L}[f(t)] = \frac{1}{(s+a)}$$ ~ polynomial

(iii) $f(t) = \cos\omega_0 t$

$$= \frac{e^{j\omega_0 t} + e^{-j\omega_0 t}}{2}$$

$$\mathcal{L}[f(t)] = \frac{s}{s^2 + \omega_0^2}$$ ~ polynomial

$\mathcal{L}[f(t)] = \frac{\text{Num.}}{\text{Den.}}$

roots of (Num)

roots of (Den)

For $\frac{1}{s} \rightarrow$ root of den $= 0 \rightarrow$ pole — (root of num → doesn't exist)

For $\frac{s}{s^2+\omega_0^2} \rightarrow$ root of num $= 0 \rightarrow$ zero

In amplifiers, we take pole nowhere near to zero.
In oscillators, we take pole $= 0$.

19-09-2023

Properties of Laplace Transforms

Linear

① $f(t) \longrightarrow F(s)$

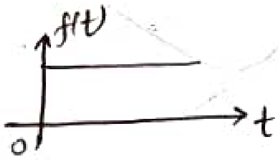
$af(t) \longrightarrow aF(s)$

② $f(t) \longrightarrow F(s)$

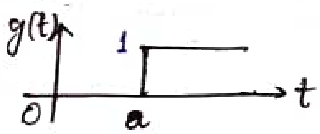
$g(t) \longrightarrow G(s)$

$af(t) + bg(t) \longrightarrow aF(s) + bG(s)$

Eg., $f(t) = 1 \quad (t>0)$



$g(t) = 1 \quad (t>0)$



$f(t) \downarrow$

$\rightarrow F(s) = \frac{\text{Num}(s)}{\text{Den}(s)}$

roots of Num(s) ⟶ Zeros of F(s)

roots of Den(s) ⟶ Poles of F(s).

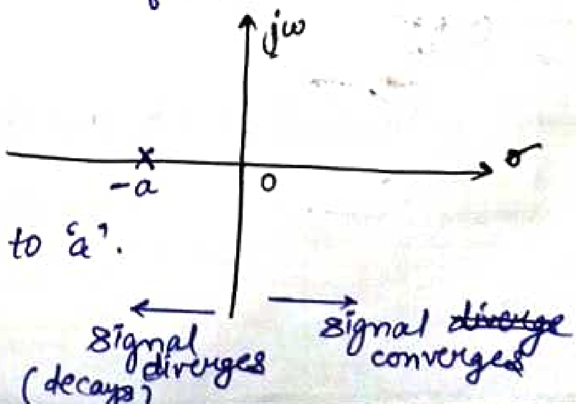
Representation:
o → Zero
x → Pole

Eg., $F(s) = \frac{1}{s+a}$

$f(t) \nearrow$

roots: $\boxed{s=-a}$

$\sigma = -a$
$j\omega = 0$



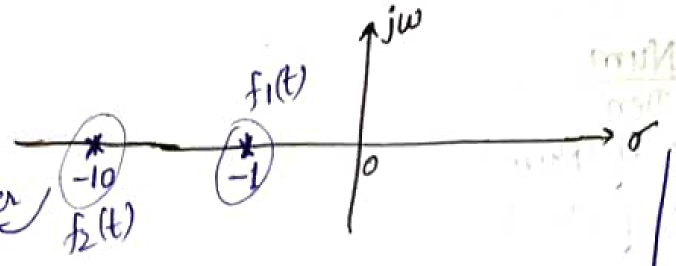
f(t) has a decay related to 'a'.

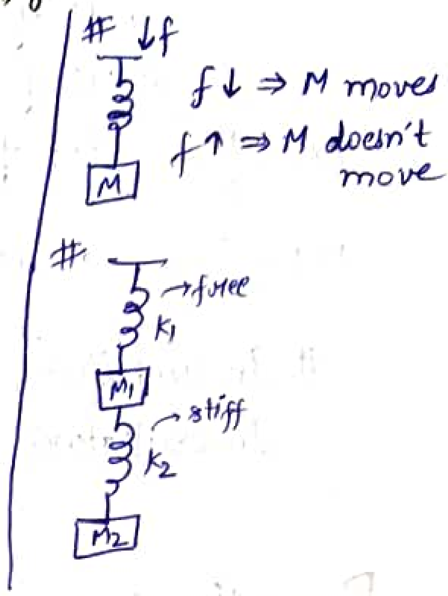
↳ $f(t) = e^{-at}, \ t>0$

Eg $f_1 = e^{-t}$

$f_2 = e^{-10t}$

↳ decays faster

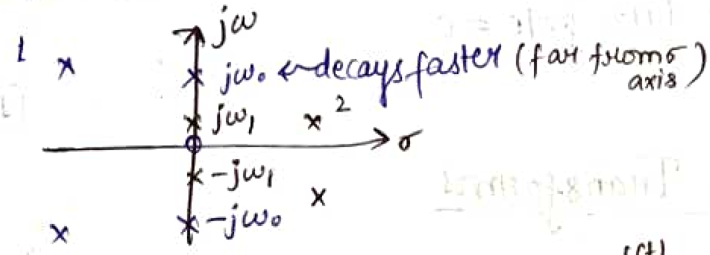


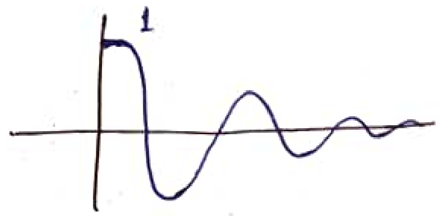


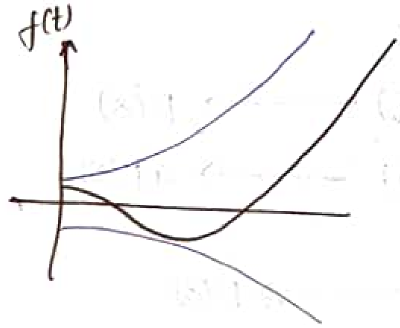
Eg: $F(s) = \dfrac{\omega_0^2}{s^2+\omega_0^2}$ || $F(s) = \dfrac{s}{s^2+\omega_0^2}$

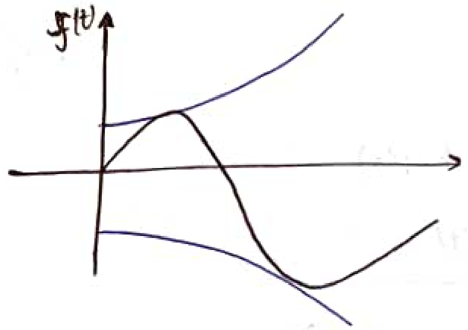
poles: $s = \pm j\omega_0$

$f(t) = \cos\omega_0 t$





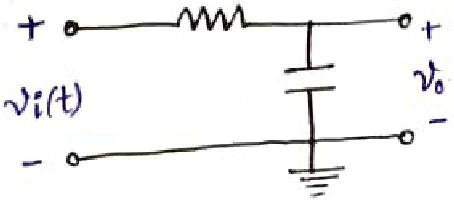


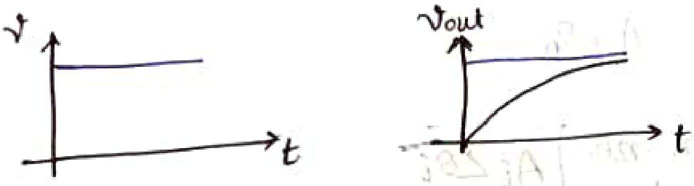


22-09-23

Laplace Transform

$$f(t) \longrightarrow F(s)$$



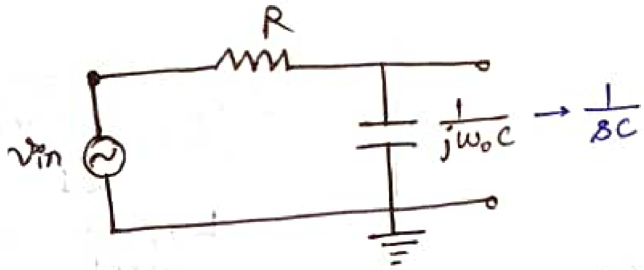


v_i transfer v_{out} function

Transfer function,

$$TF = \frac{v_{out}}{v_{in}}$$ ⇝ a function in s-plane.

$$V_{out}(s) = T.F. \times V_{in}(s)$$



$v_{in} = A\sin(\omega_0 t)$

$$v_{out} = v_{in} \times \left(\frac{\frac{1}{j\omega_0 C}}{R + 1/j\omega_0 C} \right)$$

$\downarrow$

$$\frac{V_{out}(s)}{V_{in}(s)} = \underbrace{\frac{1/sC}{R + 1/sC}}_{TF}$$

→ $s = \sigma + j\omega$

We are interested in the response for a particular frequency @ steady state.

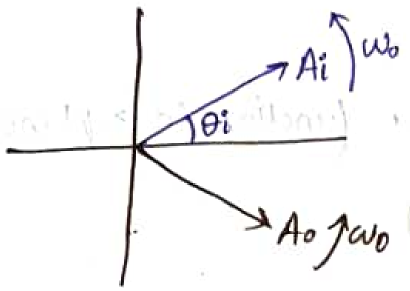
⇒ Replace $s = \sigma + j\omega$ by $s = j\omega_0$.

⇒ If we want to know the response for all ω, replace $s = j\omega$. → Frequency response

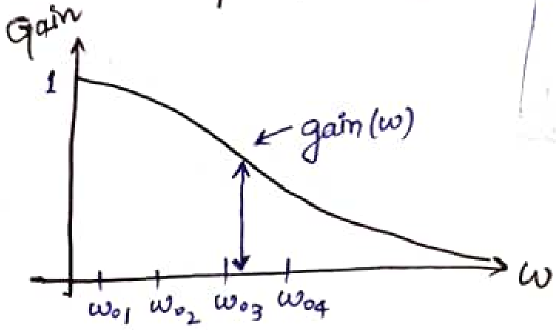
$$TF = \frac{1/sC}{R + 1/sC} \; ; \quad FR = \frac{1/j\omega C}{R + 1/j\omega C}$$

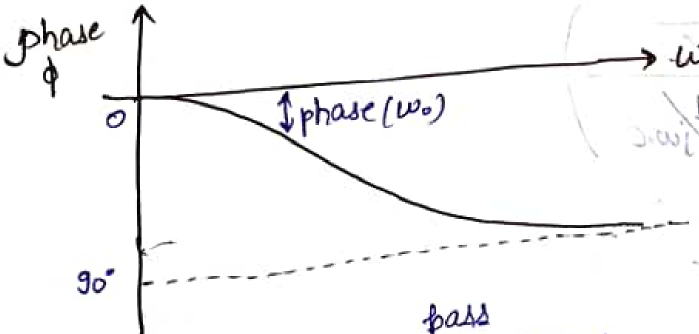
Input	Output
$A_i \angle \theta_i$	$A_o \angle \theta_o$

$$\rightarrow \underbrace{A_o \angle \theta_o}_{@\omega_o} = \left(\begin{array}{c} \text{complex no.} \\ FR|_{\omega_o} \end{array} \right) A_i \angle \theta_i$$



→ Change in magnitude + change in direction





↳ Low ~~path~~ pass filter

For very small frequency,
$1/sC \rightarrow$ very large
≈ 1
↳ phase = 0
of $FR|_{\omega_o}$

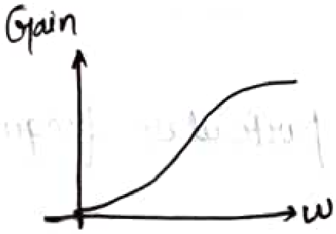
$$FR = \frac{1}{j\omega RC + 1}$$

as $\omega \rightarrow \infty$,
$|FR| \rightarrow 0$

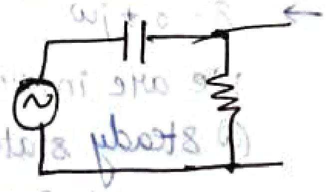
$$FR = \frac{1}{j\omega RC + 1} \cdot \frac{1 - j\omega RC}{1 - j\omega RC}$$

$$= \frac{1 - j\omega RC}{1^2 + \omega^2 R^2 C^2}$$

as $\omega \rightarrow \infty$,
phase $\rightarrow 90°$

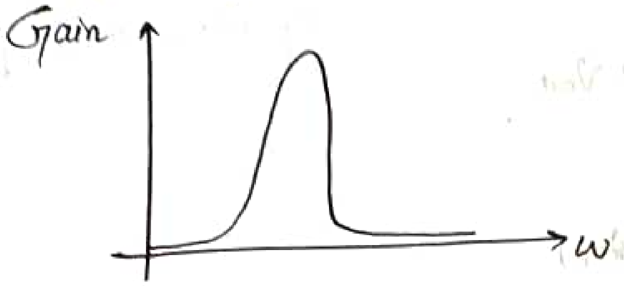


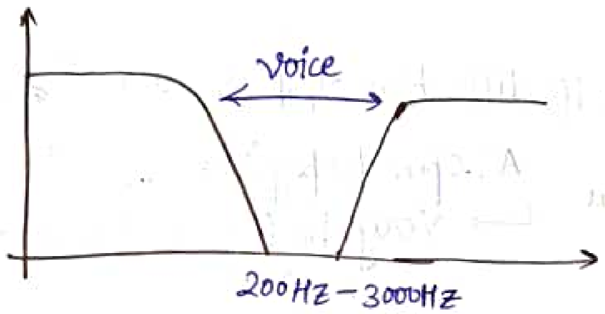
→ High pass filter →



→ Cascading (cascading one with other')

↳ connecting input of high pass filter to the output of low pass filter (or vice versa).



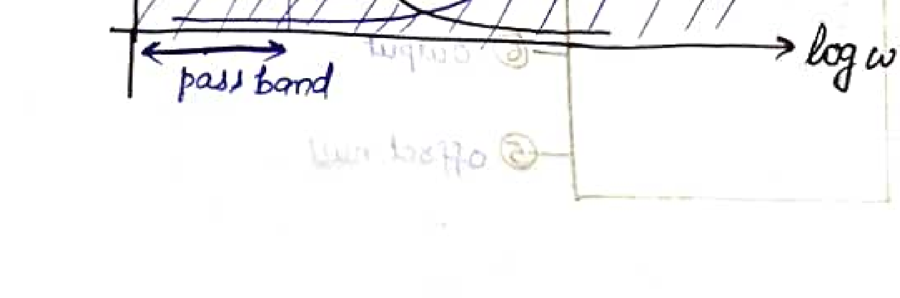


- Analysis → Generating curve from problem
- System design → Generating function from curve.
- Synthesis

→ in → [1] → [2] → out

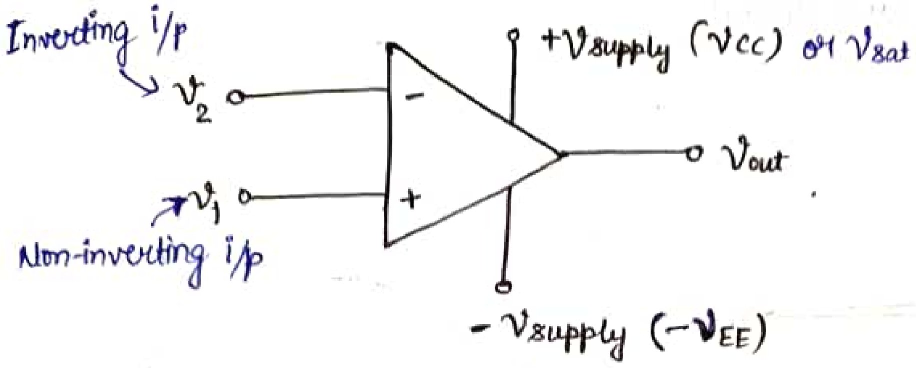
TF1 TF2

Out = In × TF1 × TF2



29-09-2023

OPERATIONAL AMPLIFIER

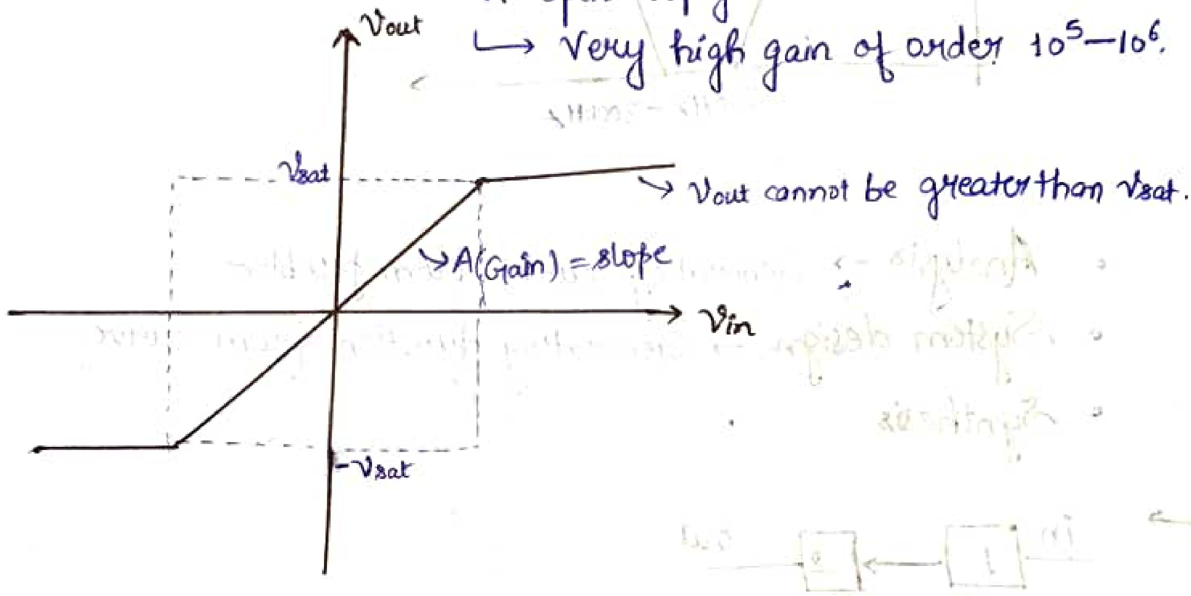


→ Used to do mathematical operations in the past.

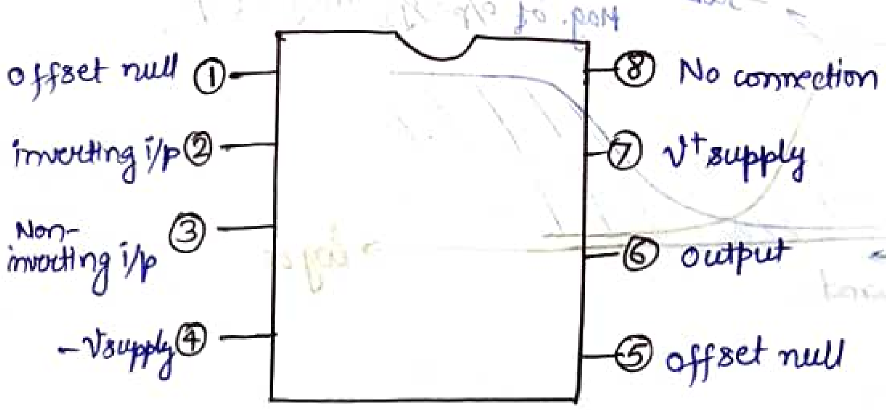
Gain: $10^5 - 10^6$

$\boxed{V_{out} = (v_1 - v_2) A}$ → if diff. b/w v_1 & v_2 ~ 10^{-5} V.

A: Open-loop gain
↳ Very high gain of order $10^5 - 10^6$.



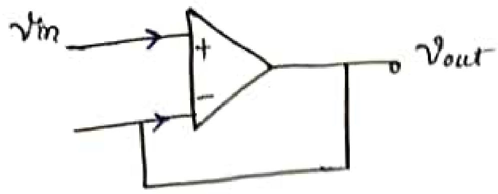
→ Opamps come in ICs.
eg. LF411



Golden Rules:

① Virtual Ground Rule :

LDR → light dependant resistor

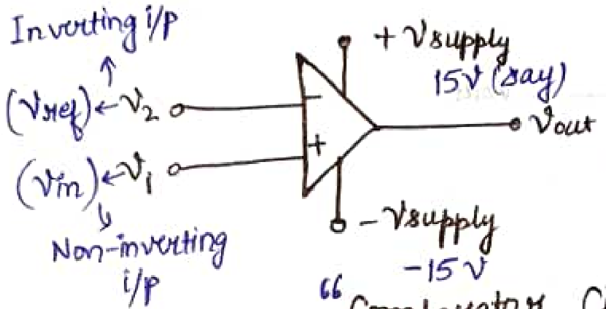


→ voltage at the inverting terminal is equal to the voltage at the non-inverting terminal.

↳ "Voltage Follower circuit"

Opamps try to make $v_1 = v_2$.

② $I_{in} = 0$. (Input draws no current).



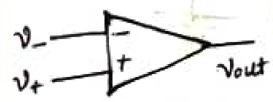
"Comparator Circuit"

If $(v_{ref} > v_{in})$,

$v_{out} = -v_{sat}$

If $(v_{ref} < v_{in})$,

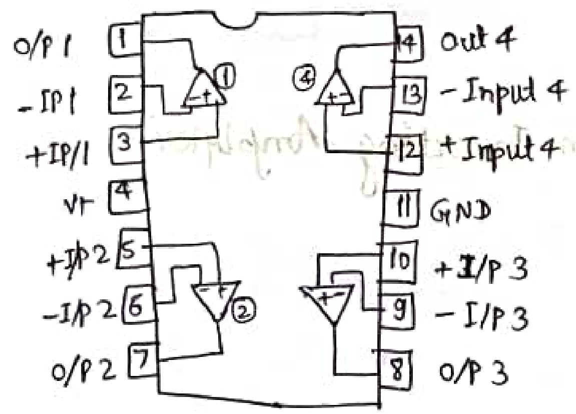
$v_{out} = v_{sat}$.

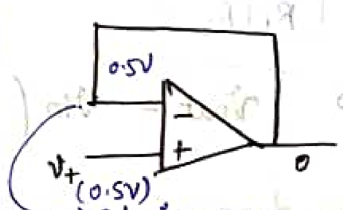


• if $(V_+ > V_-)$,
↳ V_{out} ↗
↳ increases very suddenly (10^6 V/s)

• if $(V_+ < V_-)$,
↳ V_{out} ↘
↳ decreases suddenly

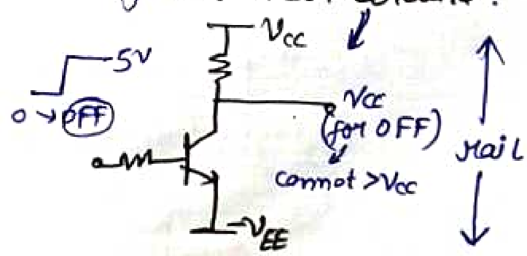
LM324A





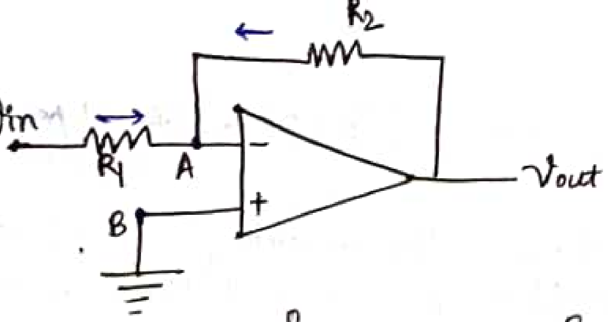
→ V_- increases until it is 0.5V

→ V_{out} is limited by a range of voltages ($+V_{supply}$ — $-V_{supply}$)
↳ cannot be > V_{supply}.
↳ for "Linear Circuits".



→ Rail-rail operation
→ Opamp is an extreme e.g. of differential amplifier.

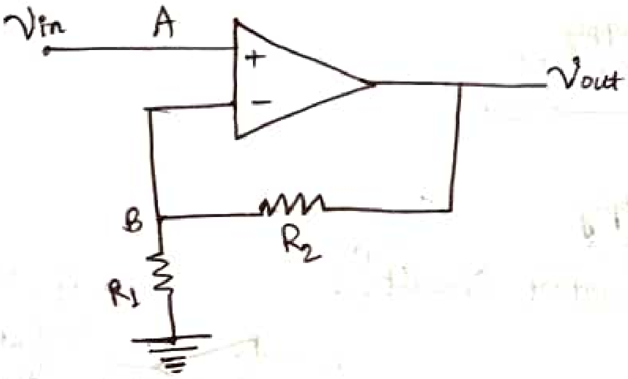
Eg.



$$\frac{v_{in} - \cancel{v_A}^{0}}{R_1} + \frac{v_{out} - \cancel{v_A}^{0}}{R_2} = 0$$

$\therefore$ $\boxed{V_{out} = -v_{in}\left(\frac{R_2}{R_1}\right)}$ $\rightarrow$ "Inverting Amplifier"

Eg.



$$v_B = \left(\frac{R_1}{R_1+R_2}\right) v_{out} \; ; \; v_A = v_{in}$$

$$\left(\frac{R_1}{R_1+R_2}\right) v_{out} = v_{in}$$

$\Rightarrow$ $\boxed{v_{out} = v_{in}\left(1 + \frac{R_2}{R_1}\right)}$

↳ "Non-Inverting Amplifier"

3-10-2023

Golden Rules

① $V_{INV} = V_{non\text{-}inv}$

② $I_{in} = 0$.

→ Basic Opamp Circuits:

→ ① Inverting amplifier

② Non-inverting amplifier

③ Voltage follower

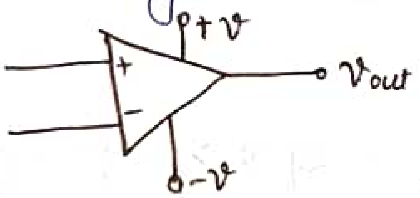
④ Comparator circuit.

Basic Cautions:

→ Negative feedback

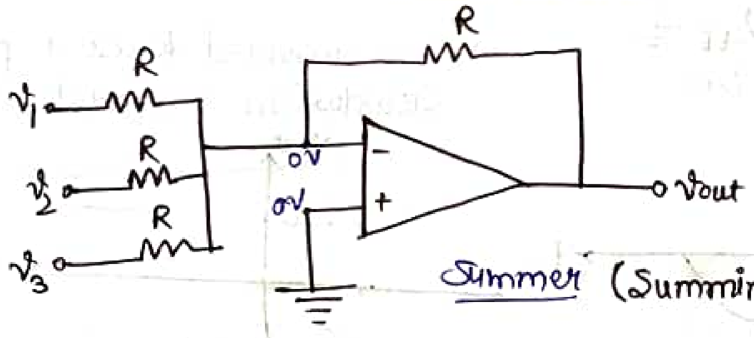
→ Transistor → Active region
Opamp should work in

→ Do not give voltage more than specified differential input.



Summer/Adder

→

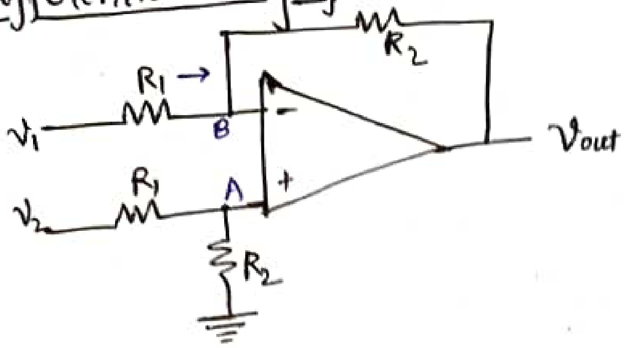


Summer (Summing Amplifier)

$$\frac{V_1}{R} + \frac{V_2}{R} + \frac{V_3}{R} + \frac{V_{out}}{R} = 0$$

$$\Rightarrow \boxed{V_{out} = -(V_1 + V_2 + V_3)}$$

Differential Amplifier



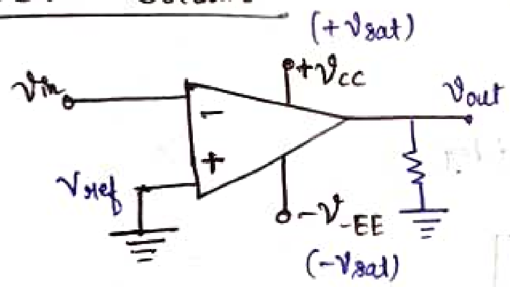
$$v_A - \left(\frac{R_2}{R_1+R_2}\right)v_2 = v_B \qquad \text{(VG Rule)}$$

KCL at point B :

$$\frac{v_1 - v_B}{R_1} + \frac{v_{out} - v_B}{R_2} = 0$$

$$\Rightarrow \boxed{v_{out} = (v_2 - v_1)\frac{R_2}{R_1}}$$

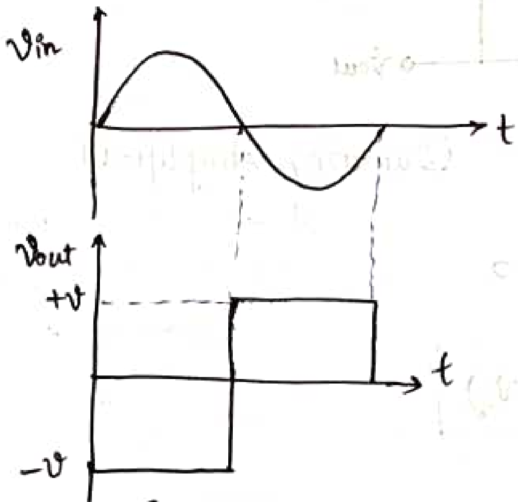
Comparator Circuit

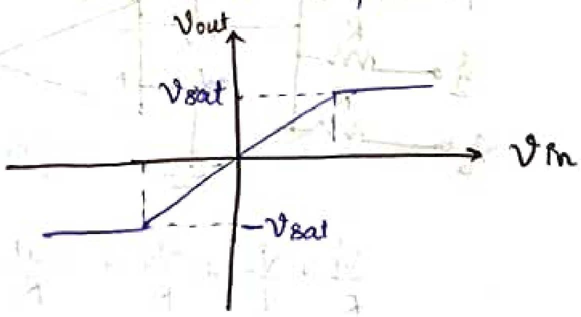


If $v_i > v_{ref} \Rightarrow \boxed{v_{out} = +v_{sat}}$

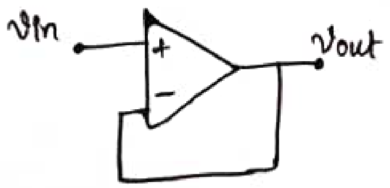
If $v_i < v_{ref} \Rightarrow \boxed{v_{out} = -v_{sat}}$

→ It is required to use a pull-up resistor in comparator circuits.



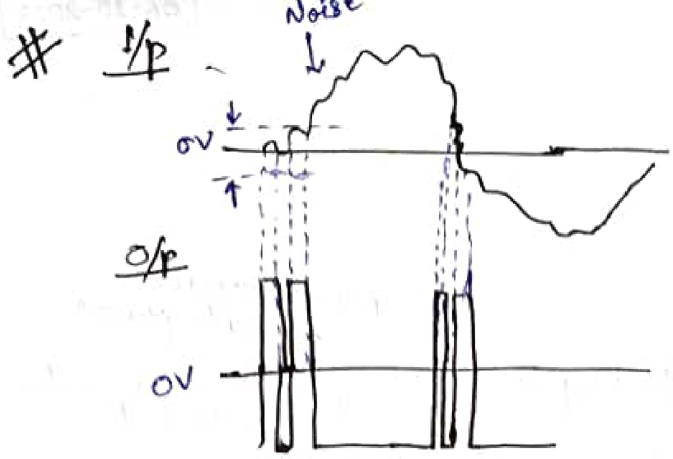


Voltage Follower Circuit



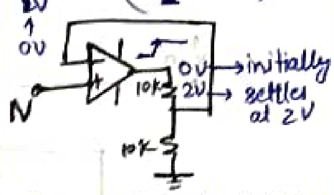
$\boxed{v_{out} = v_{in}}$

→ Also called as the buffer.

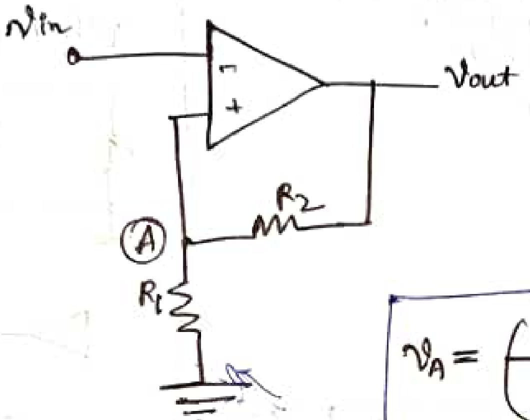


→ Negative feedback

$V_{out} \uparrow \Rightarrow (V_- - V_+) \downarrow$



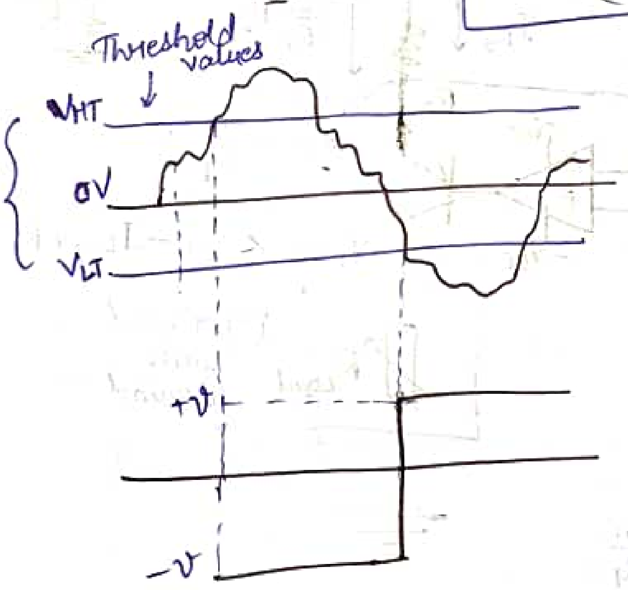
Increase in V_{out} causes diff. $(V_+ - V_-)$ decrease



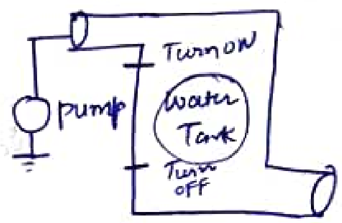
→ "Schmitt Trigger Circuit"

$$\boxed{V_A = \left(\frac{R_1}{R_1 + R_2}\right) V_{out}}$$

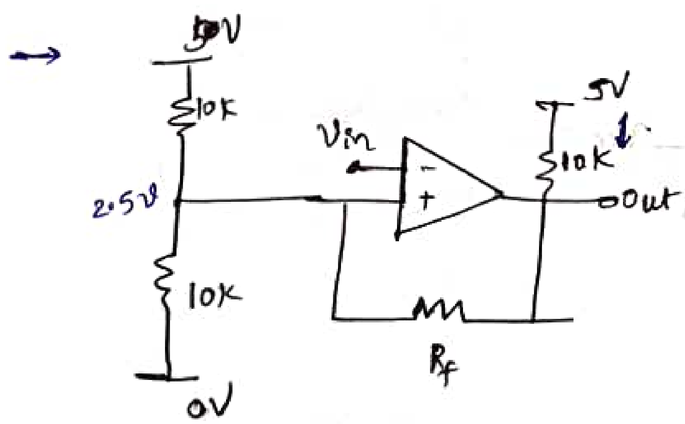
$$\boxed{\begin{aligned} V_H &= V_A \\ V_L &= -V_A \end{aligned}}$$



#



→



$V_+ (set) = 2.5V$

$V_{in} = 2.4V$

$\Rightarrow V_{out} = 5V$ (high)

$V_+ (modified) =$ Superpose $(2.7V)$

$V_{in} \to 2.8$

$V_{out} = 0$

$V_+ (mod) = 2.3V$

$\Rightarrow$ used to automate (gains memory)

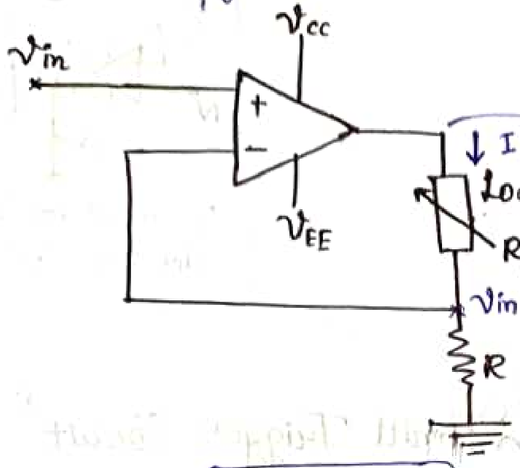
⇓

Responds only to i/p, not noise

06-10-2023

Current Source

↳ Using Opamps



Vout swings up such that Vin → steady state value

$\downarrow \frac{V_{in}}{R}$

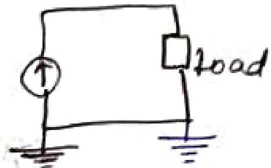
Floating load ⇝ one of the terminal is not grounded

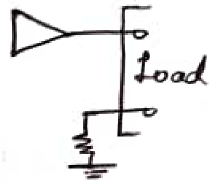
↑ not referenced with ground

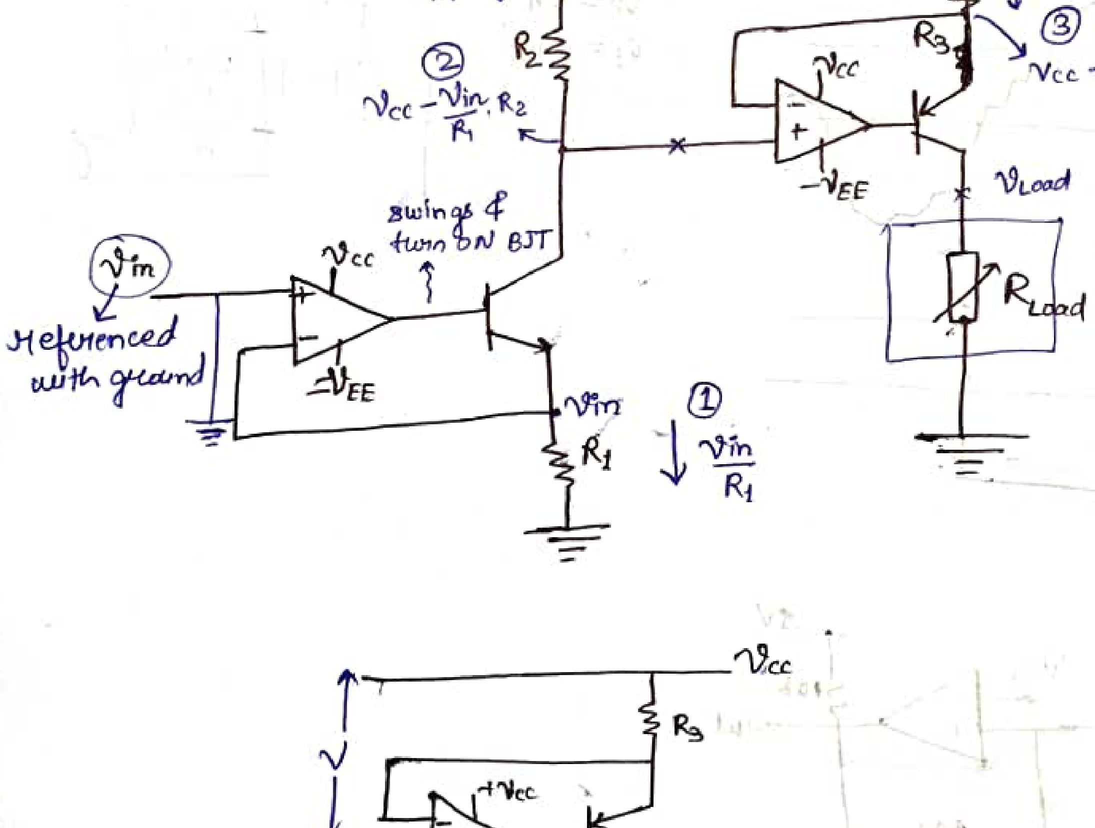
↳ Inconvenient

$$\boxed{I_{Load} = \frac{V_{in}}{R}}$$

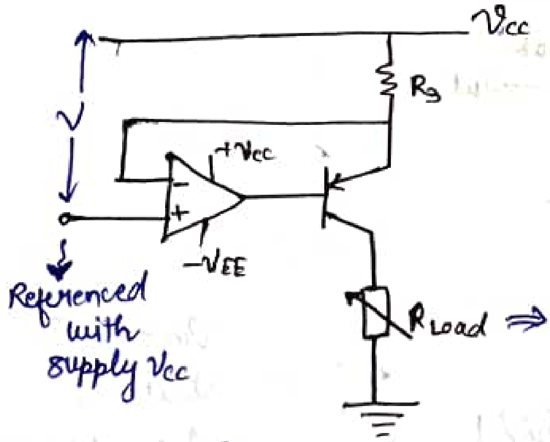
What we want



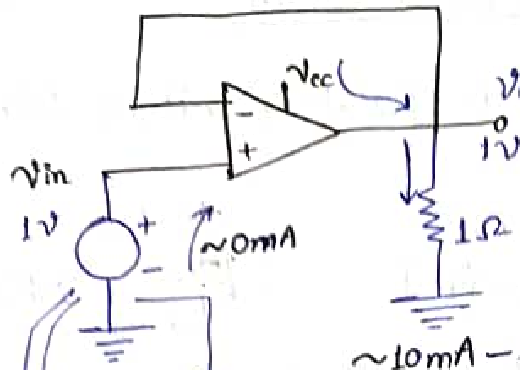




Referenced with ground



R_{Load} ⇒ Passive Load



Vin ⇝ can drive current

~10mA – 50mA : Typical value

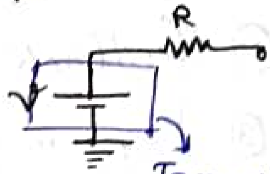
Unaffected, whatever we connect at the output

If we try to drive current, voltage will collapse.

- Used in circuits with a weak voltage source (not capable of drawing current from the voltage source)
- Current is drawn from the biasing voltage applied.

voltage follower circuit
⇓
Increases current delivering capability

⇒ Application:



To measure V, we give zero current

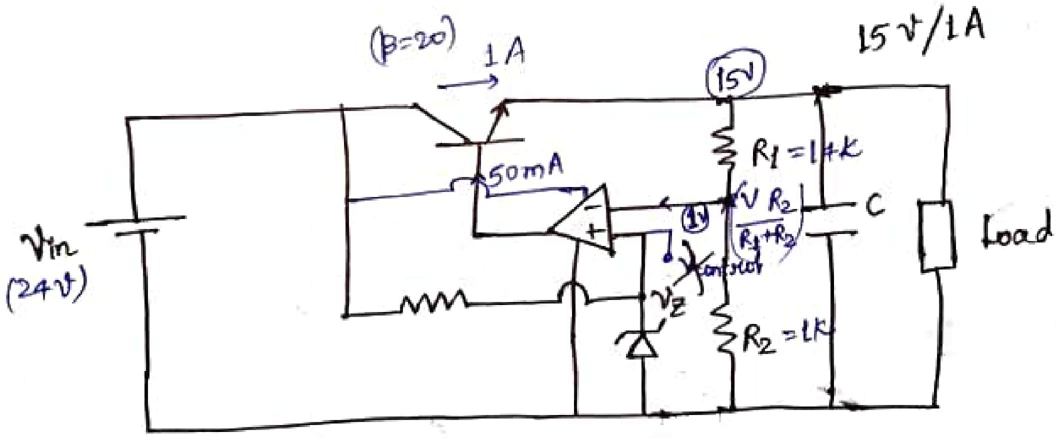
Power Booster / Voltage Regulator

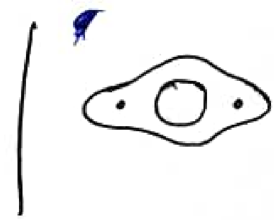
Control parameters: ~~R_1/R_2~~, ~~Vcontrol~~

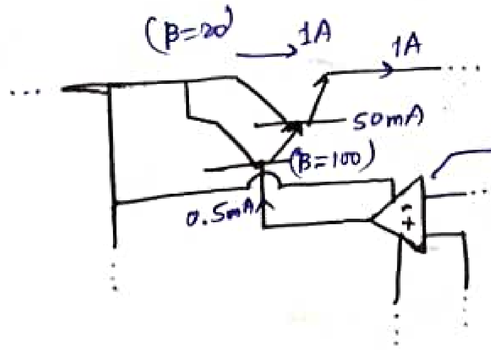
① Ratio of R_1 & R_2 [$R_1+R_2 \to$ very large $\approx 100k\Omega$]

② $V_{control} \to V_Z$

15V/1A
Voltage whatever is Load — max. current







→ Should be turned ON

If MOSFETs were used ⇒ LDO
Low Drop Out regulator

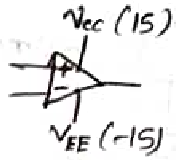
Darlington pairs are used to make sure opamp is capable of driving the transistor.

09-10-2023

Basic Cautions when using Op-amps in a Circuit

① Opamp follows 2 "golden rules" when operating within bias voltages.

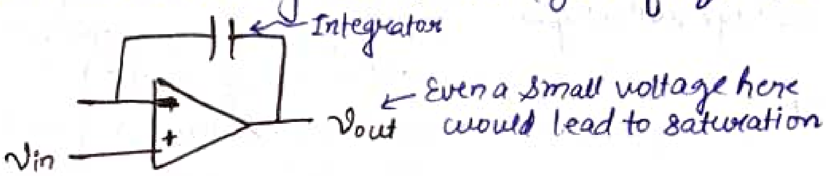
Typically: $v_{cc} = -1.5v$ | Rail-Rail
$v_{EE} = +1.5v$ | $(v_{cc}, -v_{EE})$



±15v

② Feedback must be arranged so that it is negative.

③ Negative feedback should be present at (zero freq) ac.

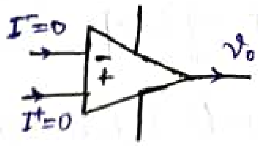


④ Opamp tries to reduce the differential voltage $(v_{in+} - v_{in-})$.
The circuit should be such that it allows this.

Departure from Ideal Opamp Characteristics

Ideal Characteristics:

① Input impedance $= \infty$.



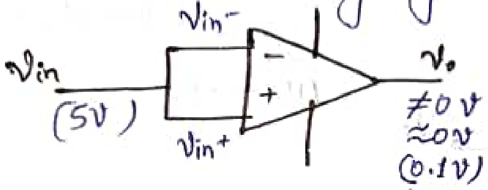
② Output impedance $= 0$.

↳ Limit on how much current can be drawn

③ Open loop voltage gain $= \infty$.

$$gain = \frac{v_o}{(v_+ - v_-)}$$

④ Common mode voltage gain $= 0$.



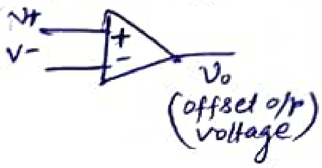
⇓ Due to non-ideality

$\frac{v_o}{v_{in}} = 0$.

$$\boxed{\frac{v_{in-} + v_{in+}}{2} = v_{CM}}$$

$$A = \frac{v_o}{v_+ - v_-}$$

$(v_+ \neq v_-)$

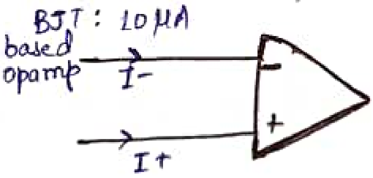


⑤ Opamp's output swings instantaneously.

↳ slew rate $= 10^6$ V/sec. (Rate at which o/p swings)

\# JFET: 10 pA to 10 fA ✓ → choose this input current

BJT based opamp: 10 μA



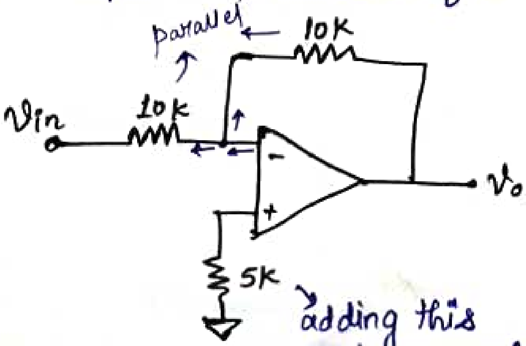
Input offset current] → Critical

↳ affects output

① If input current should be low, choose JFET opamps ($I_{in} < 100$ pA).

② Input offset current will affect the output → Difference in the input current of both terminal of the opamp.

→ Try to make input impedances equal.



→ has little bit resilliance towards input offset current.

adding this resistor provides better resillience for input offset current.

→ Better than the previous circuit.

③ Input impedance: (If the opamp circuit, to be used is sensitive to input impedance)

JFET opamp (good)

BJT opamp (not good).

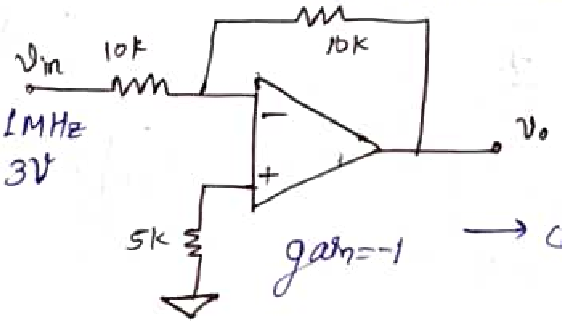
④ Common mode input range.

⑤ Slew rate

Lower slew rate → Slow opamp

⇒ Low Bandwidth (BW)

⇝ opamp should be chosen with adequate bandwidth based on the application.

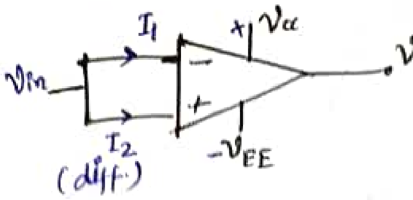


Rate:
1μs → 1V

→ Cannot work with the given conditions

10-10-2023

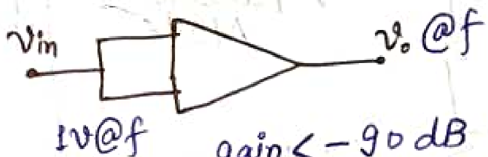
① Input Bias Current



$v_o \approx 0$
minimal voltage
output effect voltage

Slew rate: How fast the output swings (V/μs typically).
- 1V per μs.
- → How fast the output swings.

① CMRR – Common Mode Rejection Ratio



gain < −90 dB

$\boxed{-100\,dB}$

Say, gain < -100dB

$20\log(x) = -100dB$

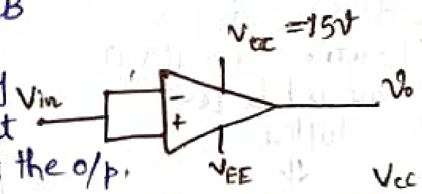
$\Rightarrow \boxed{x = 10^{-5}}$ → gain applied to common mode voltage

At the o/p, the effect of CM voltage is reduced by a factor of 10^{-5}.

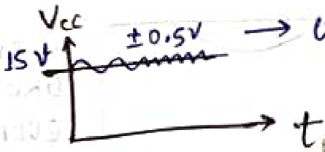
② PSRR – Power Supply Rejection Ratio

≈ −50dB

gain seen by (disturbance at the supply) at the o/p.



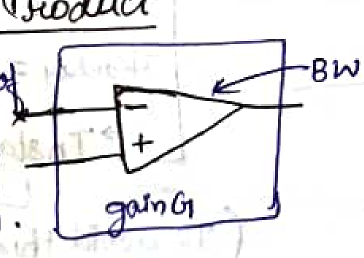
- Having a high PSRR helps in avoiding the transfer of disturbance from the supply to the output.



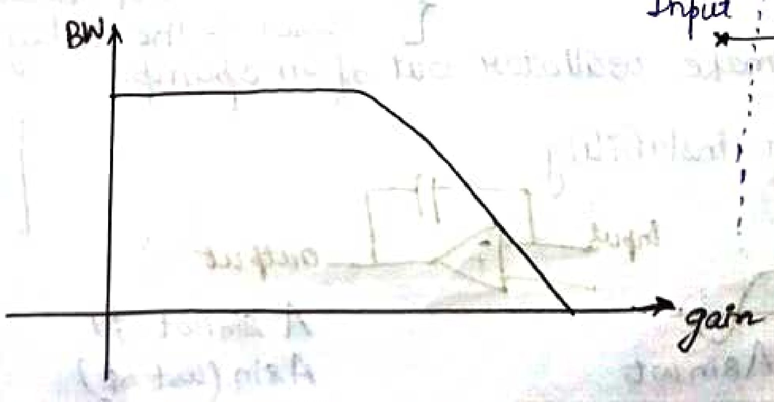
→ we get 15.5 to 14.5 V

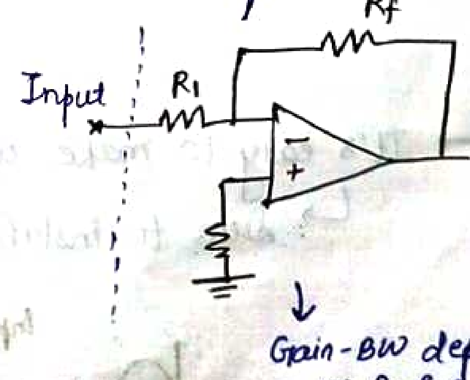
③ Gain-Bandwidth Product

→ For higher gain, BW of opamp may not be same when compared to BW for lower gain.



Slew Rate↑ ⇒ BW↑



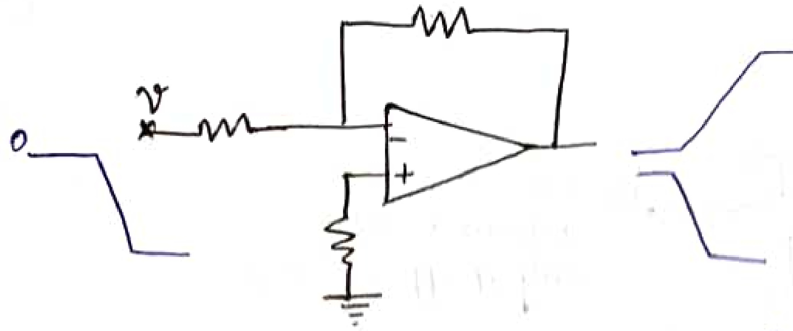


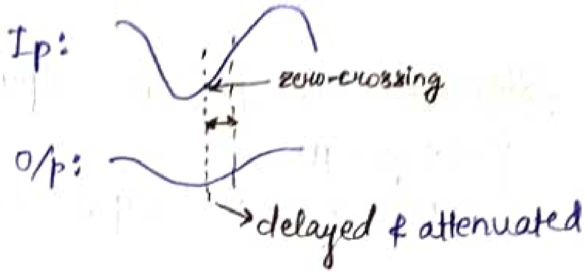
↓

Gain-BW depends on R_1 & R_f also.

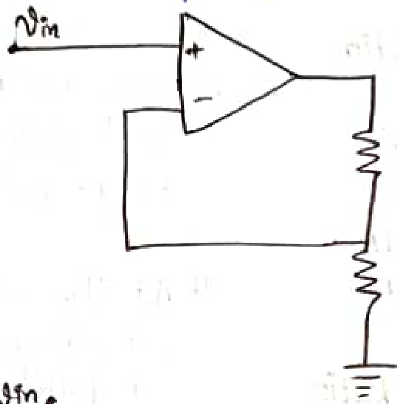
10 GHz ⇒ g=100

1 THz ⇒ g=1





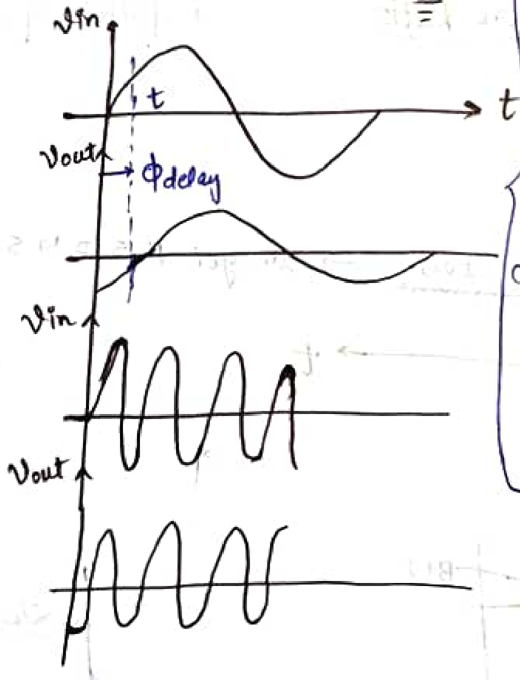
Delayed output voltage depending upon the slew rate of the opamp. ⇐



: Failure Mode ⟹ Ways to damage Opamp:
- Give high voltage to V_{cc} & VEE or Invert Vcc & VEE voltages
- Frequency → X
  - ↳ Opamp will underperform

If the freq. of Vin is increased to a specific value, the o/p would be delayed by the same amount & it would be 180° shifted
⇓
cause circuit to behave like ⊕ve feedback even when ckt. is ⊖ve feedback ⇒ saturation

Slew Rate
- Property of opamp
- doesn't depend on i/p



[$\phi_{delay} \approx 0$: good situation

$\phi_{delay} = 180°$

↳ Instability of an opamp]

↳ Saturation

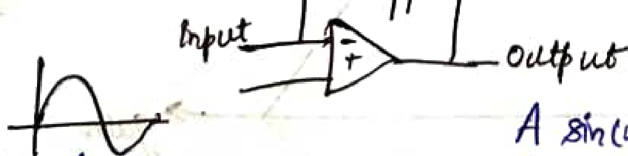
{ To avoid this, i/p should not be given closer to the delay frequency. }

⇓

[This concept can be used to make oscillator]

→ It's easy to make oscillator out of an opamp.

↳ Due to instability.

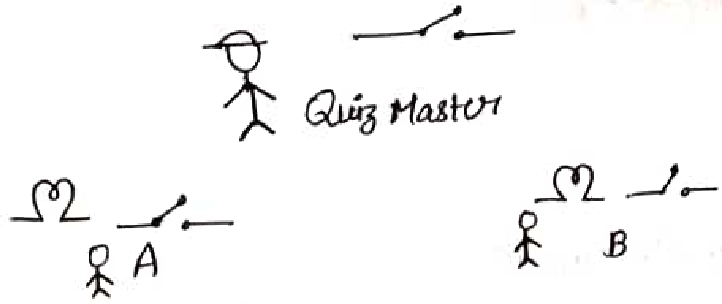


$A \sin \omega t$ $\quad$ $A \sin(\omega t - \phi)$

$A \sin(\omega_0 t - \phi)$

At ω_0 → particular freq. $\quad = -A \sin \omega_0 t$ ⇒ Oscillator

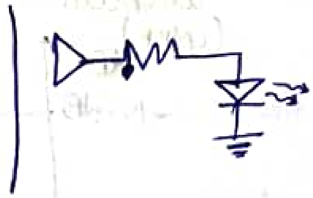
Q)



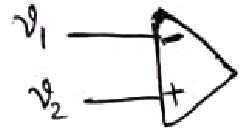
If team A presses switch, its bulb will glow only. Even if B tries to press its switch, there will be no change in state of both the switches, and vice versa.

Quiz Master ~~also~~ has a master switch to reset ~~the~~ both the switches.

Design an opamp-circuit for this functionality.



13-10-2023



$$CMV = \frac{v_1+v_2}{2}$$

$$DMV = v_1 - v_2 \rightarrow \text{Differential mode}$$

① CMRR is a function of frequency

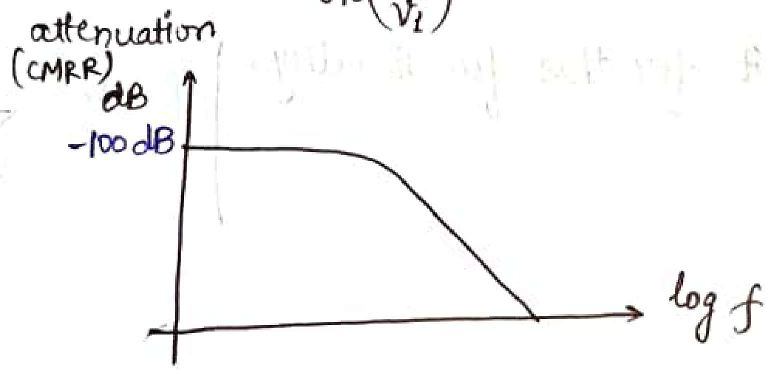
CMRR → attenuation (100 dB)

frequency → Common mode

$$20\log_{10}\left(\frac{v_o}{v_1}\right)$$

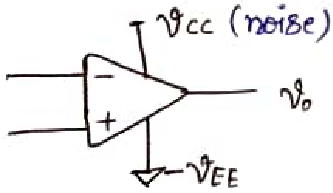
20 dB ↓ 1 decay

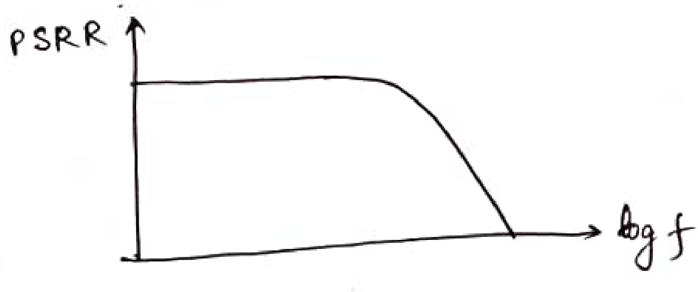
1V CMV ⟶ 10^{-5} V output



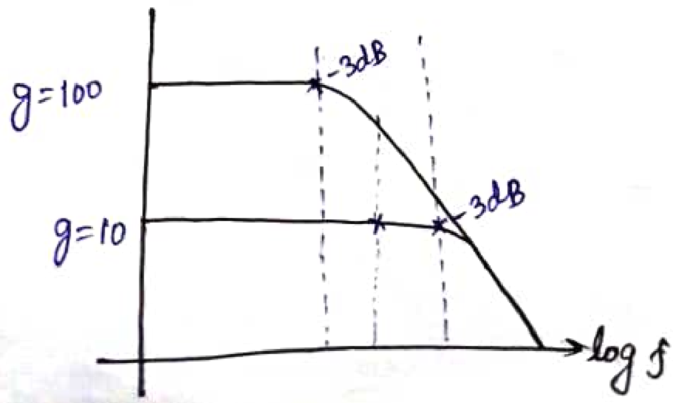
② PSRR

↳ 70 to 90dB





③ Gain-BW Product



→ Electronics manufacturing process (in B2B) → Business-to-Business

Schematic - Circuit design

↓

Prototype (10 to 15)

↓

Manufacture (Millions)

⇓

involves

⇓

① Obsolescene ⇝ Electronics stop working after a desired time.
(pre-planned)

② Not selling enough

③ Lost Schematics

④ Manufacturer out of Business.
↳ e.g., LTSpice (Linear Technology)

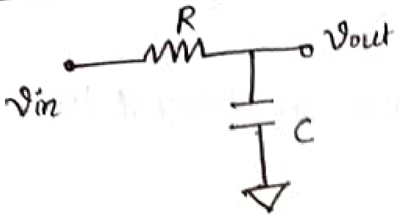
Lead time
↳ typically 52 weeks

Traiterous Eight

16-10-2023

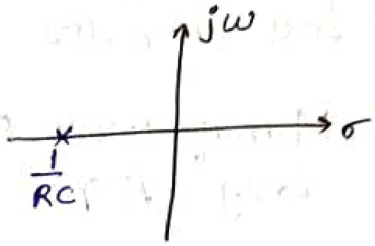
Passive / Active FILTERS

RC Filter in frequency domain: (Passive Filter)



$$Z_C(s) = \frac{1}{sC}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{1/sC}{R + \frac{1}{sC}} = \frac{1}{1+sRC} = \frac{1}{1+ s/\omega_p}$$



$$s_p = -\frac{1}{RC}$$

$$\omega_p = |s_p| = \frac{1}{RC}.$$

$$H(s) = \frac{1}{1+sRC}$$ ⇝ low pass filter

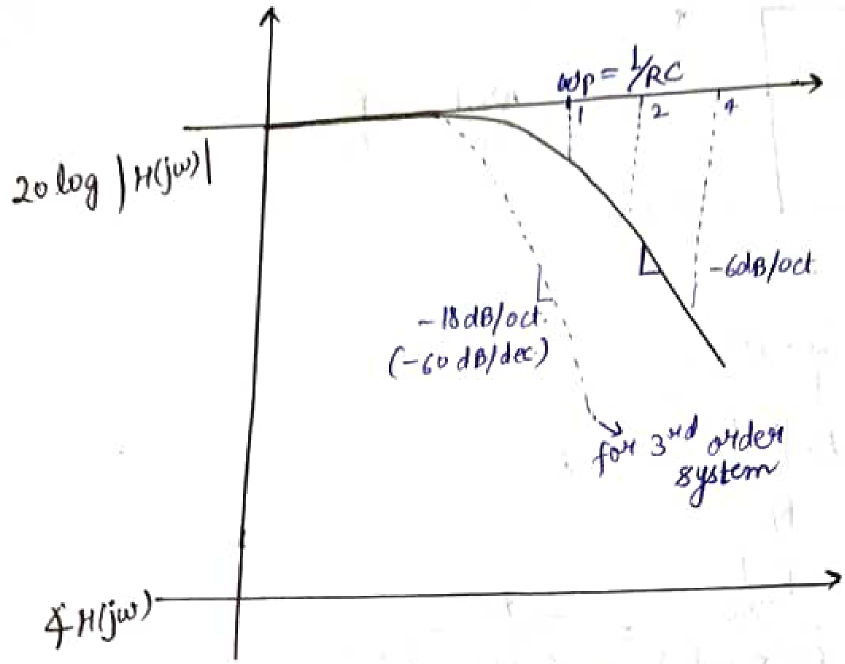
$s = j\omega$ (to look at steady state response)

$$H(j\omega) = \frac{1}{1+j\omega RC}$$

$$|H(j\omega)| = \frac{1}{\sqrt{1+(\omega RC)^2}}$$

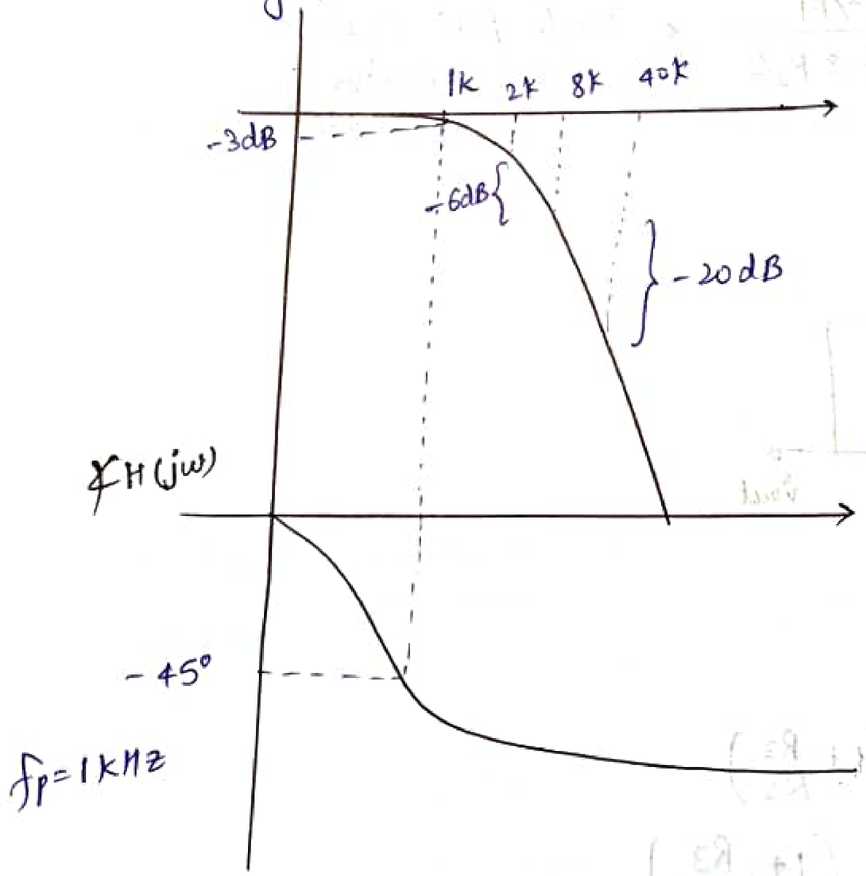
$$\angle H(j\omega) = -\tan^{-1}(\omega RC)$$

	$\lvert H(j\omega)\rvert$	$20\log\lvert H(j\omega)\rvert$	$\angle H(j\omega)$
$\omega = 0$:	1	0	0
$\omega = \frac{1}{RC}$:	0.707	−3.01	−45
$\omega = \frac{2}{RC}$:	0.447	−6.99	−63
$\omega = \frac{4}{RC}$:	0.243	−12.30	
$\omega = \frac{6}{RC}$:	0.164	−15.66	
$\omega = \frac{20}{RC}$:	0.05	−26.03	−87
$\omega = \frac{40}{RC}$:	0.025	−32.04	
$\omega = \frac{200}{RC}$:	0.005	−46.02	−89.71



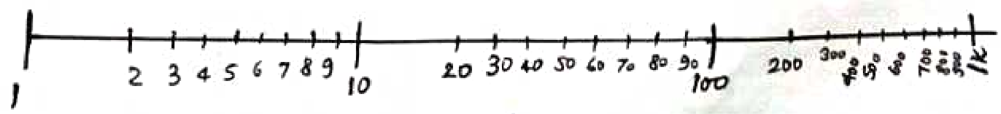
For doubling of freq (> f_p) — pole freq.
↓
decrease of 6dB
[-6dB / octave] ↓ For 1st order system
↓ OR
-20dB/decade.

→ Every pole adds a -20dB of slope per decade.

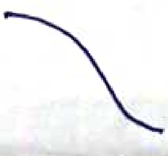


→ To accomodate a large frequency with a good resolution at lower frequencies : Use a semilog graph (logarithmic scale).

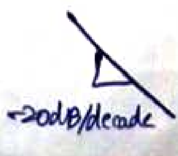
Logarithmic Scale:



On linear scale:

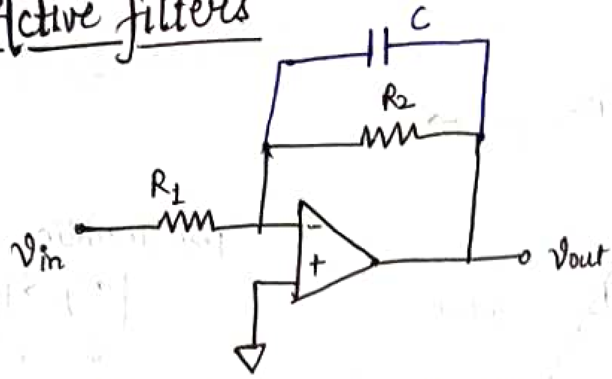


On log scale:



Active filters

→



$$\frac{V_{out}}{V_{in}} = -\frac{R_2}{R_1}$$

After adding the capacitor,

$$\frac{V_{out}}{V_{in}}(s) = \frac{-R_2 \parallel \frac{1}{sC}}{R_1}$$

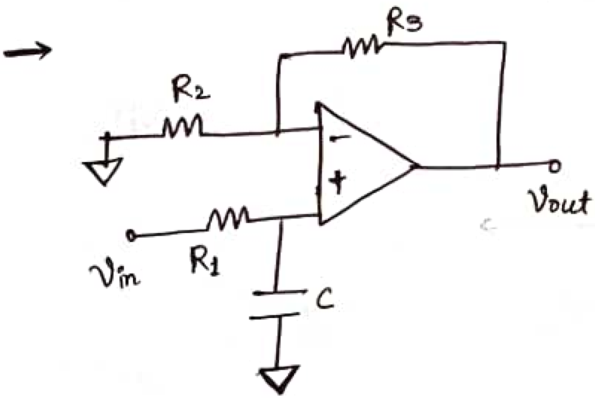
$$= -\frac{R_2 \frac{1}{sC}}{R_1\left(R_2 + \frac{1}{sC}\right)} = \frac{-R_2}{R_1(1+sCR_2)}$$

$$= \frac{-R_2/R_1}{1+s\,R_2C}$$ ← single pole system (−6dB/octave, −20dB/decade)

$$\omega_p = \frac{1}{R_2C}$$

$$s_p = -\frac{1}{R_2C}$$

→



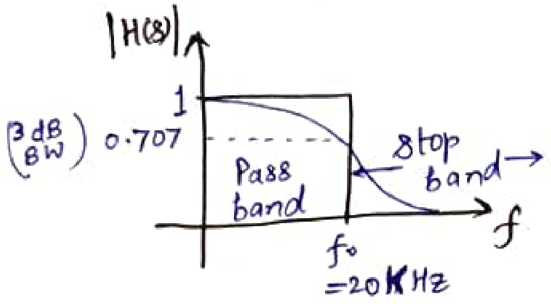
$$\frac{V_{out}}{V_{in}}(s) = V_+(s)\left(1+\frac{R_3}{R_2}\right)$$

$$= \frac{1}{1+sR_1C}\left(1+\frac{R_3}{R_2}\right)$$

17-10-2023

Low Pass Filter [Ideal LPF]

$H(s)$: filter transfer function



$f < f_0 \Rightarrow |H(s)| = 1$ (filter allows the signals to pass through)

$f > f_0 \Rightarrow |H(s)| = 0$ (filter rejects signals $f > f_0$)

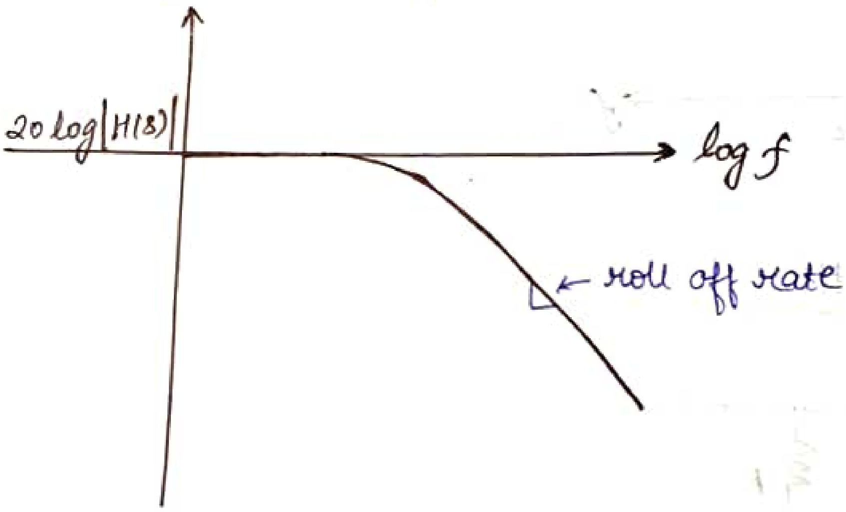
"Brick-wall Response"

↳ cannot be implemented in practice

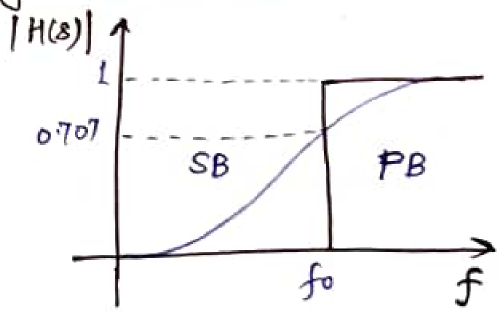
→ Abrupt change from pass band to stop band, ideally.

→ Non-idealities:

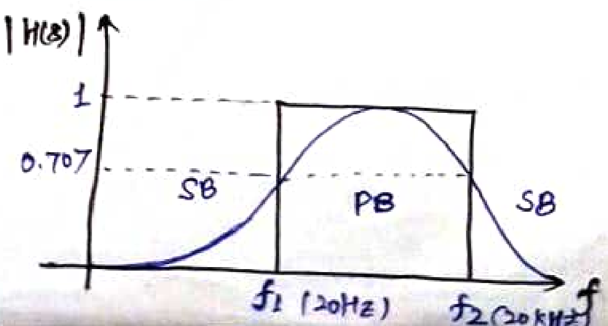
1. Pass band is not flat
2. No abrupt transition from pass band to stop band. (gentle transition)



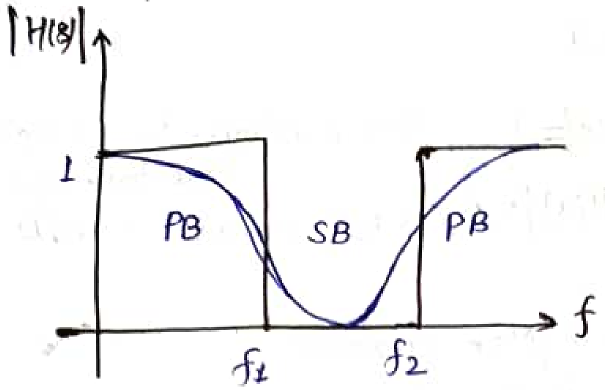
High Pass Filter (HPF)



Band Pass Filter

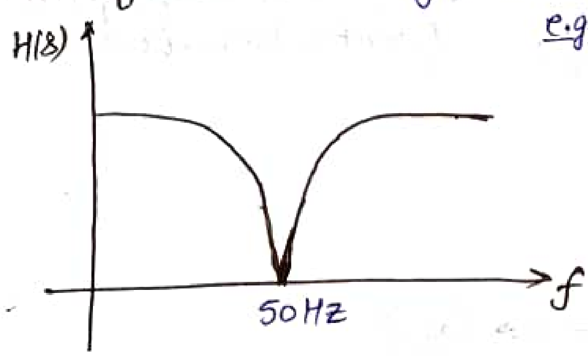


Band stop Filter

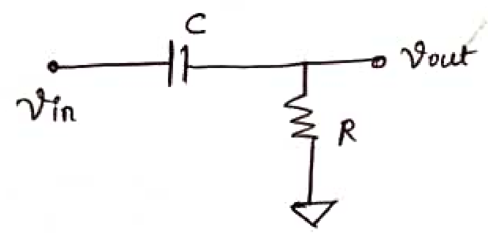


Notch filter

→ To get rid of a particular frequency noise
e.g., to get rid of 50 Hz signal (electrical signal) in medicals.



1st order Passive HPF



$$\frac{V_{out}}{V_{in}}(s) = \frac{R}{R+\frac{1}{sC}} = \frac{sRC}{1+sRC}$$

$= H(s)$

Poles $= 1$, $s_p = -\frac{1}{RC}$

Zeroes $= 1$, $s_z = 0$ (DC)

$|H(s)|_{\omega=0} = 0$

Eg. $G(s) = 1 + s/\omega_z$

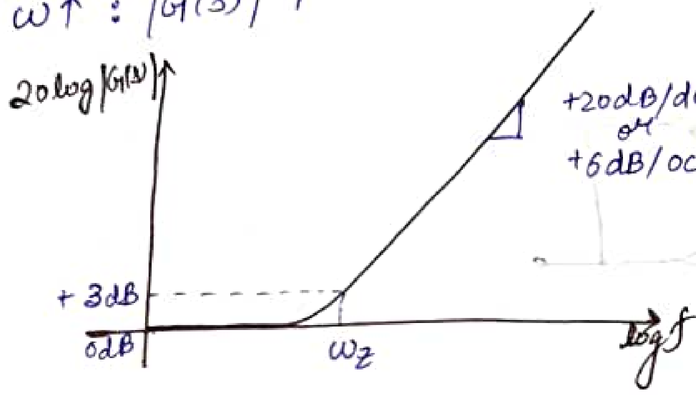
$$|G(s)| = \sqrt{1 + \left(\frac{\omega}{\omega_z}\right)^2}$$

$\omega = 0$: $|G(s)| = 1$ $\quad \measuredangle = 0$

$\omega = \omega_z$: $|G(s)| = \sqrt{2}$ $\quad \measuredangle = +45$

$\omega \uparrow$: $|G(s)| \uparrow$

$\omega \to \infty$: $\measuredangle \to +90^\circ$



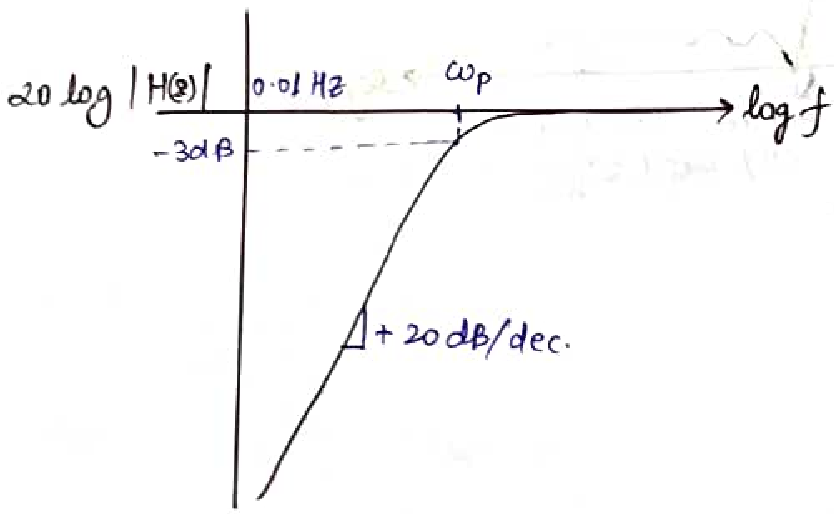
	Pole	Zero
Slope :	−20dB/dec −6dB/oct	+20dB/dec +6dB/oct
$\omega = \omega_0$:	−3dB	+3dB
$\measuredangle\ \omega = \omega_0$:	-45°	$+45^\circ$
$\measuredangle\ \omega \to \infty$:	-90°	$+90^\circ$

HPF :

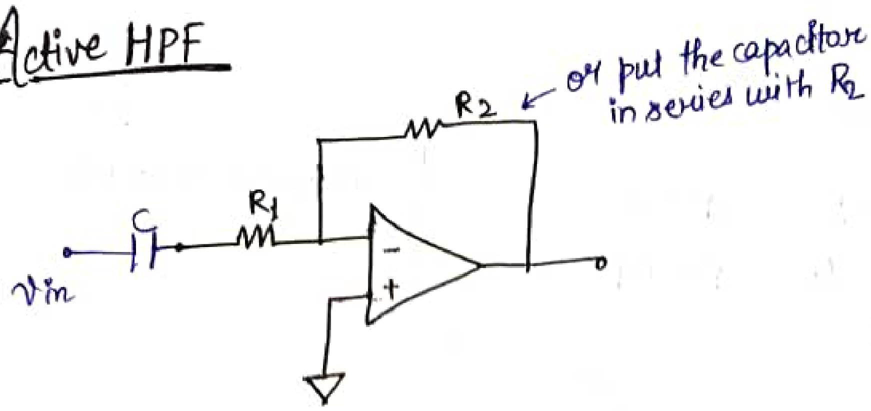
$$H(s) = \frac{sRC}{1 + sRC}$$

$\omega_z = 0$

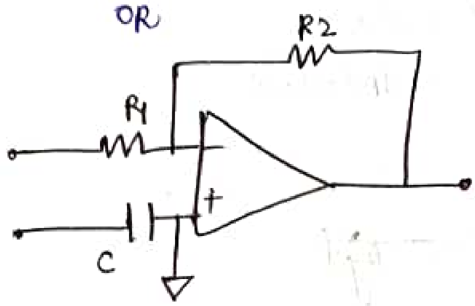
$\omega_p = \frac{1}{RC}$ $(= 1\text{KHz})$



Active HPF



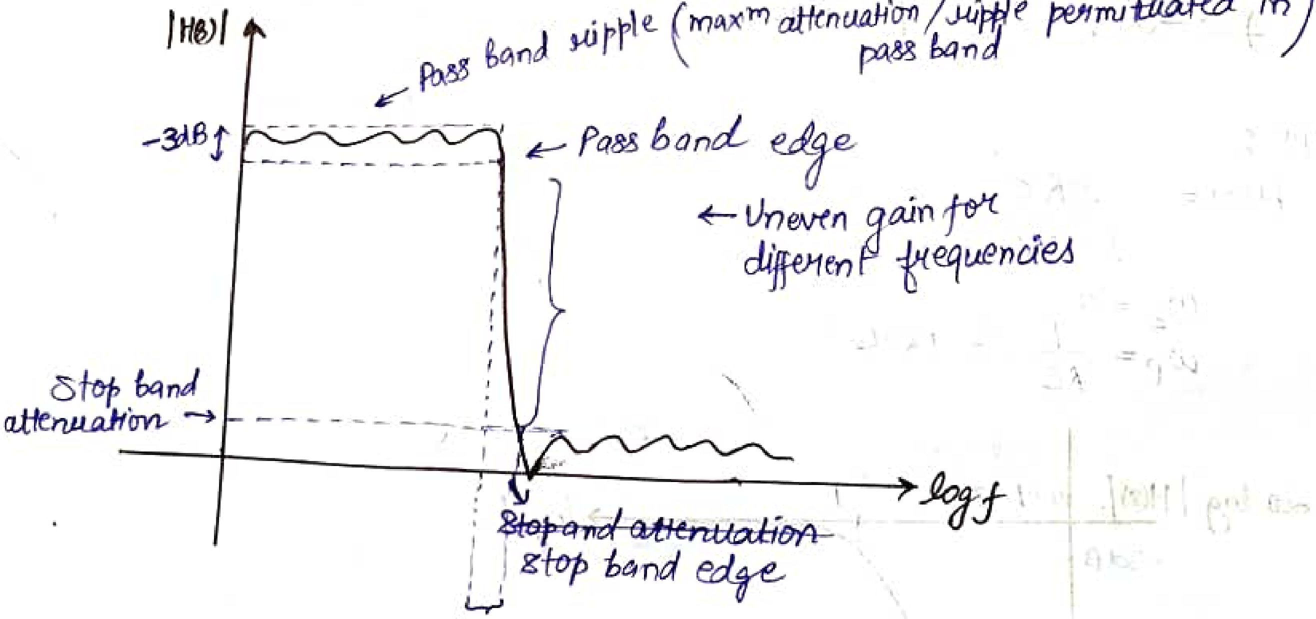
OR



LPH

$$H(s) = \frac{s^2 + (\ )s + (\ )}{s^2 + (\ )s + (\ )}$$

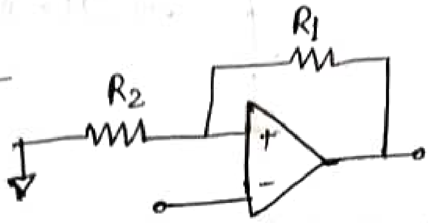
- Types of transfer functions: Butterworth

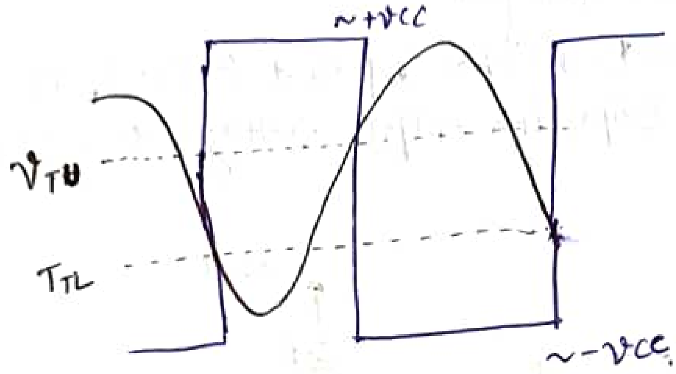


30-10-2023

MULTIVIBRATORS

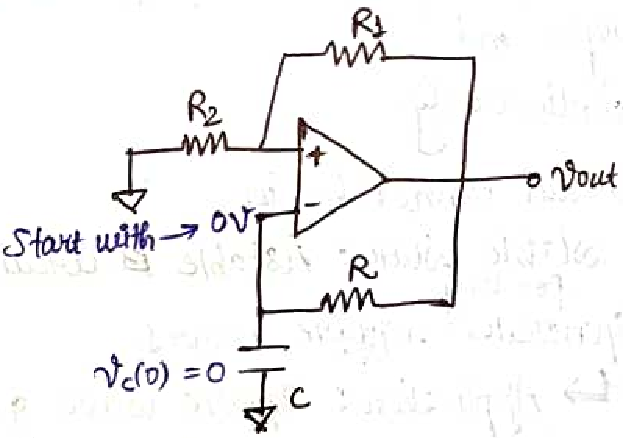
Smith Trigger:





This smith trigger is called Bistable multivibrator

Astable Multivibrator:



⇓

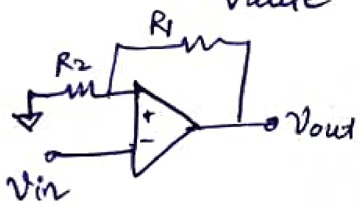
v_{out} : $+v_{sat}$ or $-v_{sat}$
↳ Indetermined

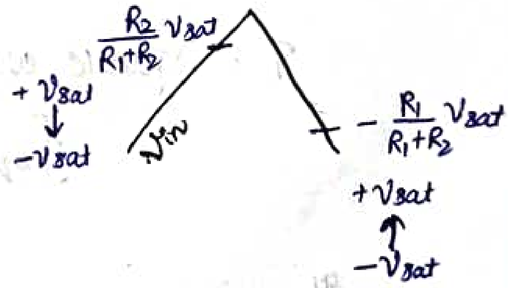
Start with $v_{out} = +v_{sat}$

$+v_{sat}$: Highest saturation value (close to vcc)

$-v_{sat}$: lowest saturation value

#





$$v_{TU} = \frac{R_2}{R_1+R_2} v_{sat} = \beta v_{sat}$$

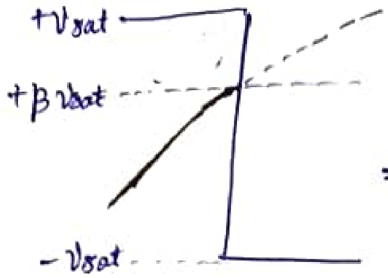
$$v_{TL} = -\frac{R_2}{R_1+R_2} v_{sat} = -\beta v_{sat}$$

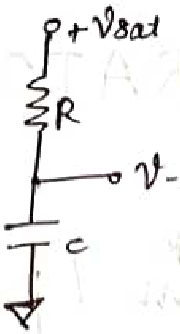
Initially,

$v_C^{(0)} = 0V$

$v_{out}(0) = +v_{sat}$

$v_+ = +\beta v_{sat}$





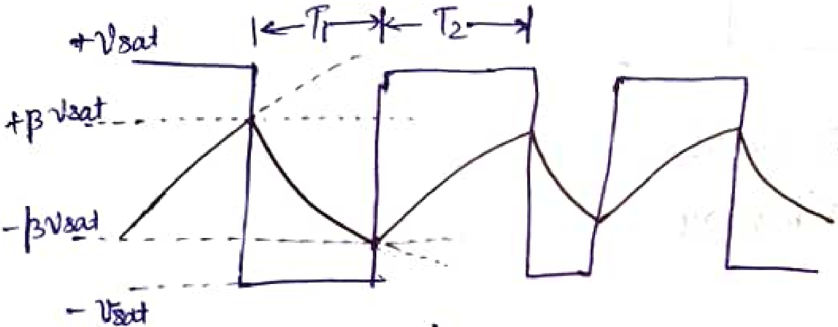
$v_-(t) = v_c(t) = +v_{sat}\,(1-e^{-t/\tau})$

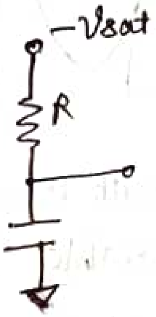
$\tau = RC$

⇒ Capacitor charges.

When v_c reaches $+\beta v_{sat}$ $(= v_{TU})$, it trips the output voltage to $-v_{sat}$.

Afterwards,





↳ Capacitor discharges

↳ capacitor charges and discharges continuously.

↳ The circuit cannot be in one stable state for long: Astable circuit

↳ Generates square waves.

↳ Application: Square wave generator.

T : Square wave time period $= T_1 + T_2$

$$f_{sq} = \frac{1}{T_1 + T_2}$$

If,

$v_i = v_{initial}$

v_f : Voltage to which the capacitor is being charged/discharged.

then,

$$v_c(t) = v_f - (v_f - v_i)\,e^{-t/\tau}$$

↳ Instantaneous.

For $0 \rightarrow A$

$A - (A-0)e^{-t/\tau}$

$= A(1-e^{-t/\tau})$

But, here $v_{initial} \neq 0$.

$\underline{T_1}$:

$v_f = -v_{sat}$

$v_i = +\beta v_{sat}$

[The capacitor is ~~alwa~~ actually getting charged towards v_f $(-v_{sat})$; in the meanwhile it is getting changed]

$$v_c(t) = -v_{sat} - (-v_{sat} - \beta v_{sat})\, e^{-t/\tau}$$

At $t = T_1$, $v_c(t) = -\beta v_{sat} = -v_{sat} - (-v_{sat} - \beta v_{sat})\, e^{-T_1/\tau}$

$$\Rightarrow -\beta v_{sat} = -v_{sat} + (v_{sat} + \beta v_{sat})\, e^{-T_1/\tau}$$

$$\Rightarrow e^{-T_1/\tau} = \frac{v_{sat} - \beta v_{sat}}{v_{sat} + \beta v_{sat}} = \frac{1-\beta}{1+\beta}$$

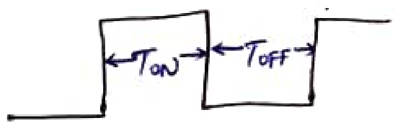
$$\Rightarrow -\frac{T_1}{\tau} = \ln\left(\frac{1-\beta}{1+\beta}\right)$$

$$\Rightarrow T_1 = RC \ln\left(\frac{1+\beta}{1-\beta}\right)$$

Similarly, $T_2 = RC \ln\left(\frac{1+\beta}{1-\beta}\right)$

R_1, R_2 should be in range of the current supplying capability of the opamp.

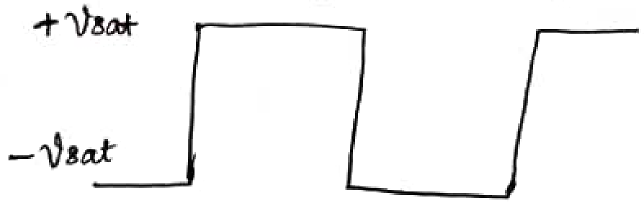
$$T = T_1 + T_2 = 2\tau \ln\left(\frac{1+\beta}{1-\beta}\right)$$



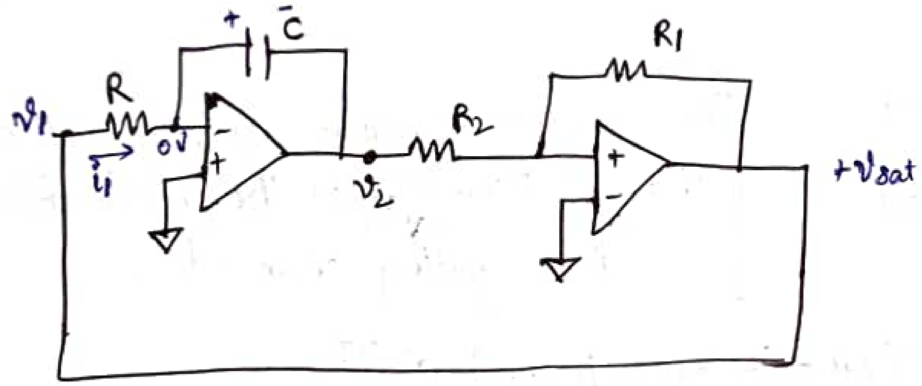
$$D = \frac{T_{ON}}{T_{ON} + T_{OFF}} \times 100\%$$

→ Duty cycle

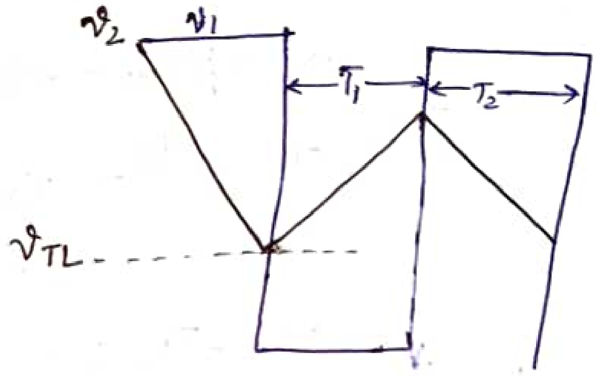
To get triangle wave from square wave:



[Astable MV] → [Integrator] : Add an integrator circuit



$$i_1 = \frac{v_1}{R}$$



$$v_c(t) = \frac{1}{C}\int i\,dt$$

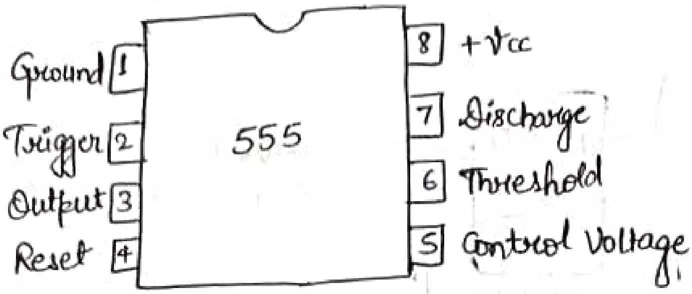
$$v_2 = -\frac{1}{C}\int \frac{v_1}{R}\,dt$$

$$= \frac{1}{RC}\int v_1\,dt$$

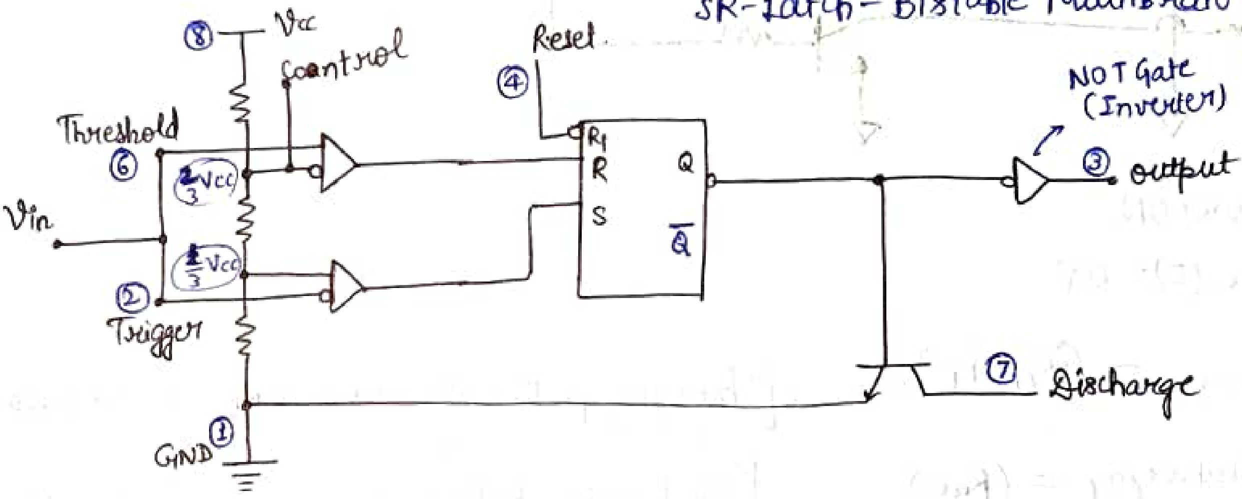
$$+\frac{v_{sat}}{RC}\cdot t \;\Big|\; -\frac{v_{sat}}{RC}\cdot t$$

555 Timer IC

(NE555)



SR-Latch - Bistable Multibrator



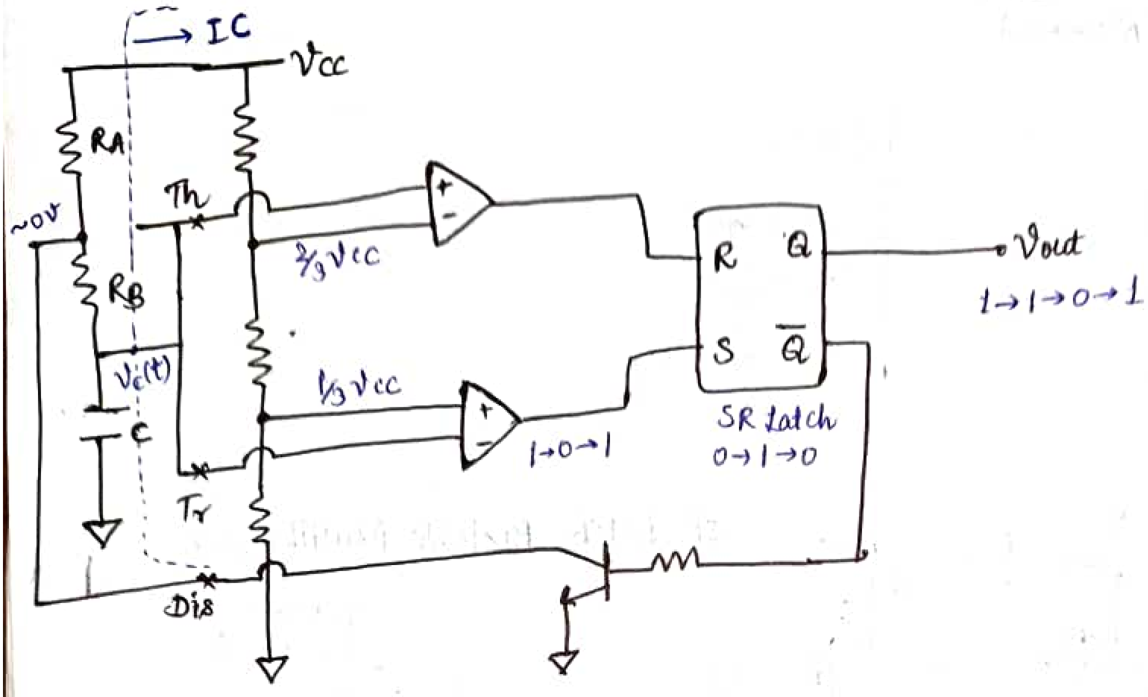
$V_{in} < \frac{1}{3}V_{cc}$: Trigger : ON

$\frac{1}{3}V_{cc} < V_{in} < \frac{2}{3}V_{cc}$: Thres.: OFF, Trig.: OFF

$V_{in} > \frac{2}{3}V_{cc}$: Thres.: ON

→ Connect Reset ($\overline{RESET}$) to V_{cc}
↳ Used as Enable Pin

→ $Q=0, \overline{Q}=1$: Transistor (BJT) : ON
↳ Discharge → GND

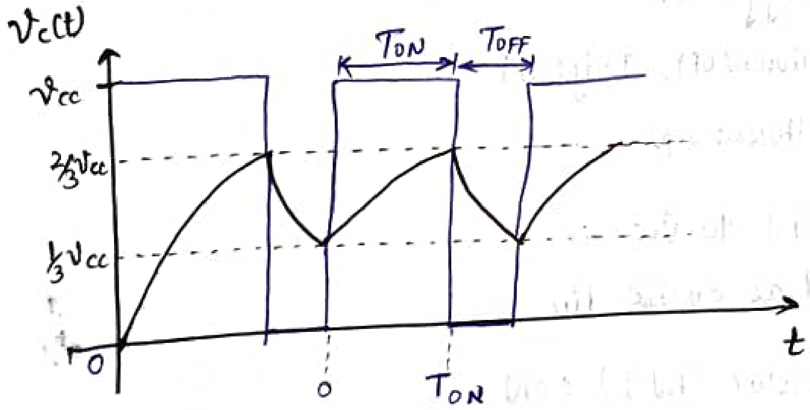


At power ON,

$v_c(0) = 0V$

$\tau_{charging} = (R_A + R_B)C$ [Charging path : $V_{CC} \to R_A \to R_B \to C \to GND$]

$\tau_{discharging} = (R_B C)$ [Discharging path : $C \to R_B \to GND\ (\sim 0V)$]



Charging phase :

$$v_c(t) = v_f - (v_f - v_i)\, e^{-t/\tau}$$

$$v_c(t) = v_{cc} - \left(v_{cc} - \frac{1}{3} v_{cc}\right) e^{-t/\tau_{ch}}$$

At $t = T_{ON}$, $v_c(t) = \frac{2}{3} v_{cc}$

$$\Rightarrow \frac{2}{3} V_{cc} = V_{cc} - \frac{2}{3} V_{cc}\, e^{-\frac{t_{ON}}{\tau_{ch}}}$$

$$\Rightarrow \frac{2}{3} V_{cc}\, e^{-t_{ON}/\tau_{ch}} = \frac{1}{3} V_{cc}$$

$$\Rightarrow \frac{t_{ON}}{\tau_{ch}} = \ln 2$$

$$\Rightarrow T_{ON} = \tau_{ch} \ln 2 \cong 0.69\, C\,(R_A + R_B)$$

At $t = T_{OFF}$, $v_c(t) = \frac{1}{3} V_{cc}$,

$$v_c(t) = \frac{1}{3} V_{cc} = 0 - \left(0 - \frac{2}{3} V_{cc}\right) e^{-\frac{t_{OFF}}{\tau_{dis}}}$$

$$\Rightarrow \frac{V_{cc}}{3} = \frac{2}{3} V_{cc}\, e^{-\frac{T_{OFF}}{\tau_{dis}}}$$

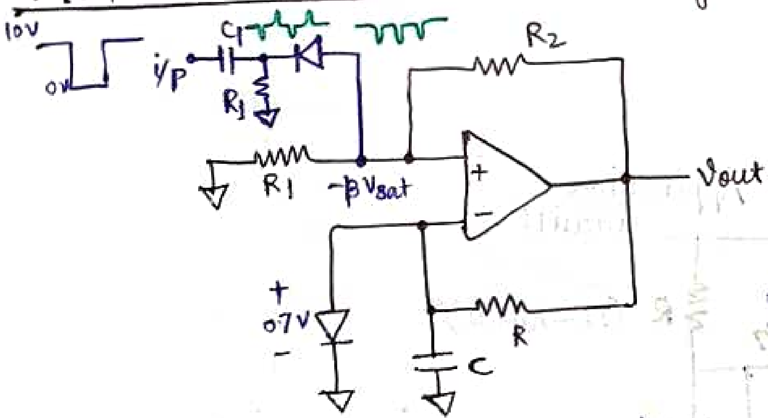
$$\Rightarrow T_{OFF} = \tau_{dis} \ln 2 \cong 0.69\, C R_B$$

Dutycycle, $D.C. = \dfrac{R_A + R_B}{R_A + 2R_B}$

$$T = T_{ON} + T_{OFF}$$

$$\Rightarrow \boxed{T = \frac{1}{f} = 0.69\, C\,(R_A + 2R_B)}$$

Monostable Multivibrator (using Opamp)



$$\beta = \frac{R_1}{R_1 + R_2}$$

Using $v_c(t) = v_f - (v_f - v_i) e^{-t/\tau}$,

Time period,

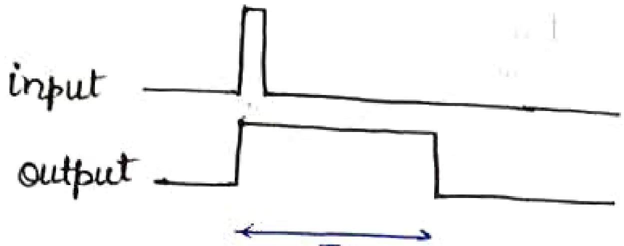
$$\boxed{T = RC \ln\left(1 + \frac{R_1}{R_2}\right)}$$

- $v_i = 0.7V$, $v_f = -V_{sat}$
  $v_c = -\beta V_{sat}$
  $\tau_{ch} = RC \ln\left(\dfrac{V_{cc} + \beta(-V_{cc})}{V_{cc} - V_D}\right)$
- $v_i = 0$, $v_f = V_{cc}$
  $v_c(t) = 2/3\, V_{cc}$

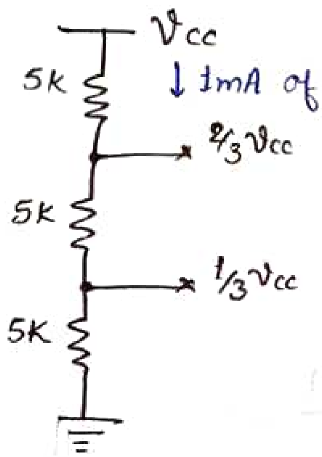
03-11-2023

Relaxation Oscillators

Multivibrators
- astable } 555 IC
- bi-stable → flip-flop
- mono-stable } 555 ICs



↳ determined by RC

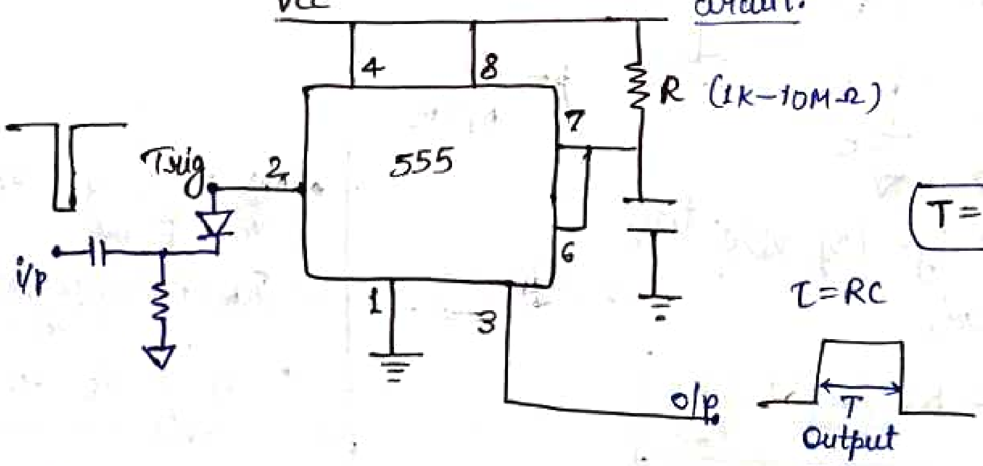


↓ 1mA of current leak ⇒ Too much

↳ Aim: To minimize the current.

Monostable Multivibrator

Application circuit:



$\boxed{T = 1.1RC}$

$\tau = RC$

Functional Block: Opens up the internal circuit.

2nd Order Filter

↳ Sallen-Key filter
- → Allows us to use only one opamp.
- → Single ended supply.

Generally,

1st order filter $\xrightarrow{\text{use}}$ one opamp
2nd order filter $\xrightarrow{\text{use}}$ 2 opamps
3rd order filter $\xrightarrow{\text{use}}$ 3 opamps.

Review :

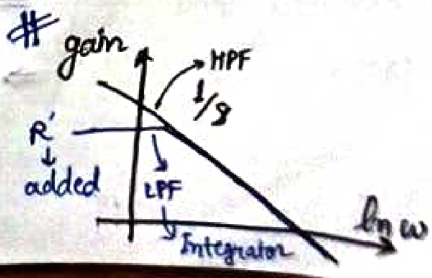
* Opamps
— 2 golden rules
- Comparators
  ↳ Schmidt Trigger
- Summer
- Inv., Non-inv. amplifiers
- Differential amplifiers
- Integrator / Differentiator

* Frequency Domain "s"
Plots "jw"
- Stability

* Non-ideal
- input — bias, impedance
- output impedance
- G-BW product
- Slew

* Active filters.

[Integrator : Modified / limiting case of LPF.
Differentiator : Limiting case of HPF.]



07-11-2023

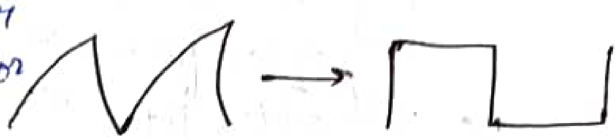
OSCILLATORS

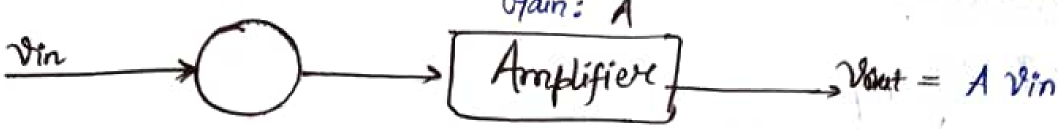
↳ Creates frequency (like sine wave).

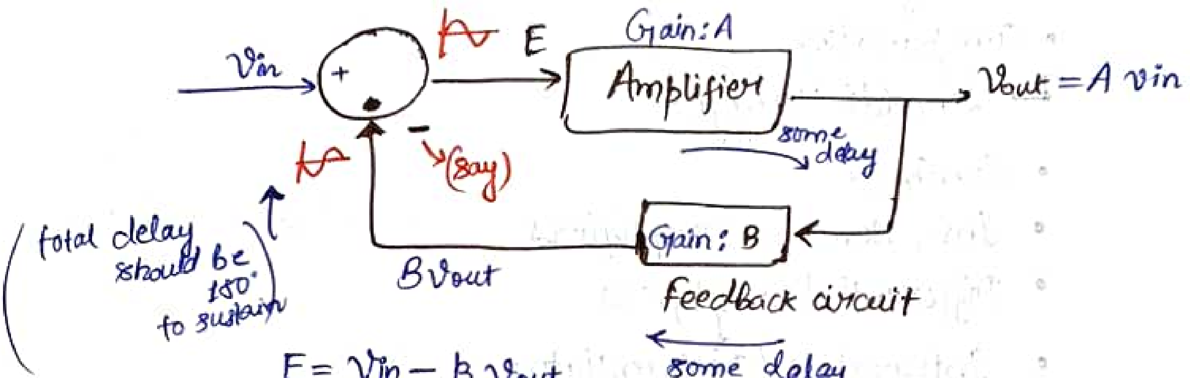
: Sine wave oscillators

Relaxation oscillator :

- ↦ Astable multivibrator
- ↦ Bistable multivibrator
- ↦ Monostable multivibrator







$$E = V_{in} - \beta V_{out}$$

$$V_{out} = AE = A\,(V_{in} - \beta V_{out})$$

$$\boxed{\frac{V_{out}}{V_{in}} = \frac{A}{1 + AB}}$$ → Overall gain

→ For V_{out} to exist for $V_{in} = 0$, the circuit should sustain V_{out} even without any input.

$$1 + AB \to 0$$

$$\Rightarrow \quad AB = -1$$

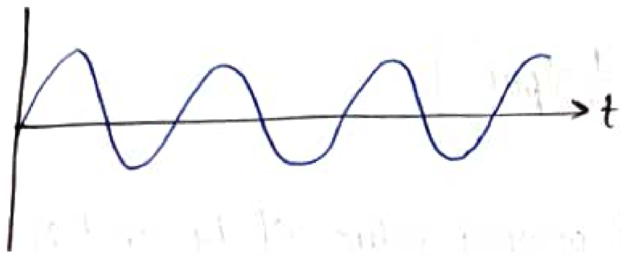
$$\Rightarrow \quad |AB| = 1$$

$\Rightarrow$ Phase shift from loop $AB = 180°$

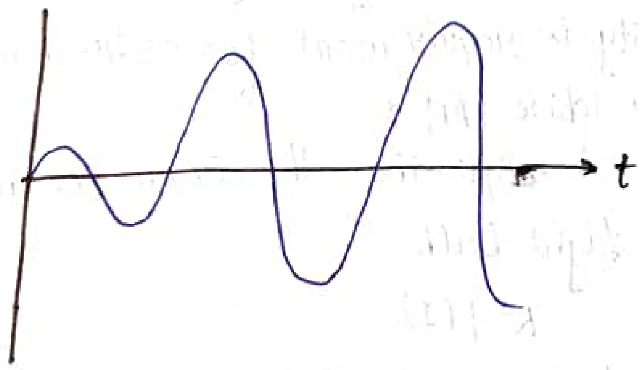
Case-1 : $AB < 1$

Getting an oscillator oscillates is difficult, as getting a particular gain and a phase shift of 180° at the same time can be challanging.

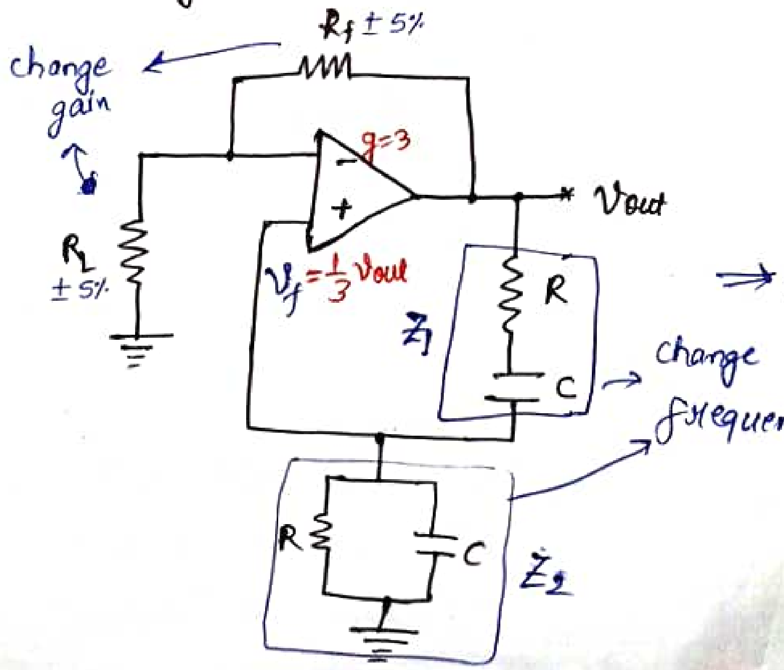
Case-2 : $AB = 1$



Case-3 : $AB > 1$



① Wein Bridge Oscillator



$$\boxed{f = \frac{1}{2\pi RC}}$$

→ linear

To sustain oscillation, we need non-linear network
⇓
To make $|AB| = 1$.

$$v_f = \frac{j\omega CR}{1 + 3RCj\omega - C^2R^2\omega^2} v_{out}$$

To ensure 0° phase shift at input,

$$1 - C^2R^2\omega^2 = 0$$

$$\Rightarrow \omega = \frac{1}{RC} \Rightarrow \frac{1}{2\pi RC} = f$$

$$\Rightarrow \boxed{v_f = \frac{1}{3} v_{out}}$$

→ It's difficult to get correct value of R_f and R_L while designing

↳ Use some tricks to gave $g=3$.

Non-linearity is requirement for oscillation (to sustain):

↳ To achive $|A\beta| = 1$.

↳ After that, the circuit becomes linear.

① Light bulb

$R = f(I)$

② Anti-parallel diodes

10-11-2023

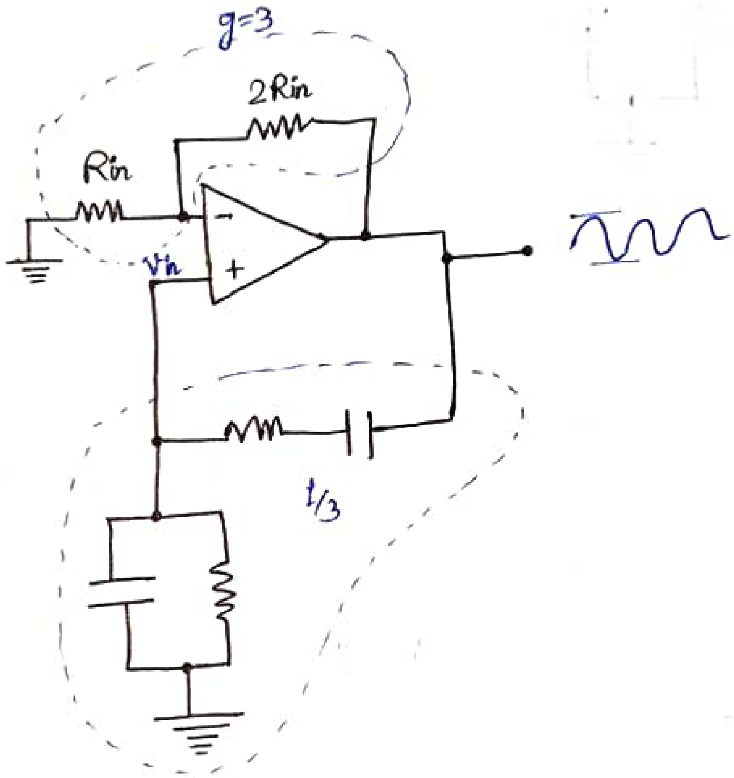
Conditions for oscillation

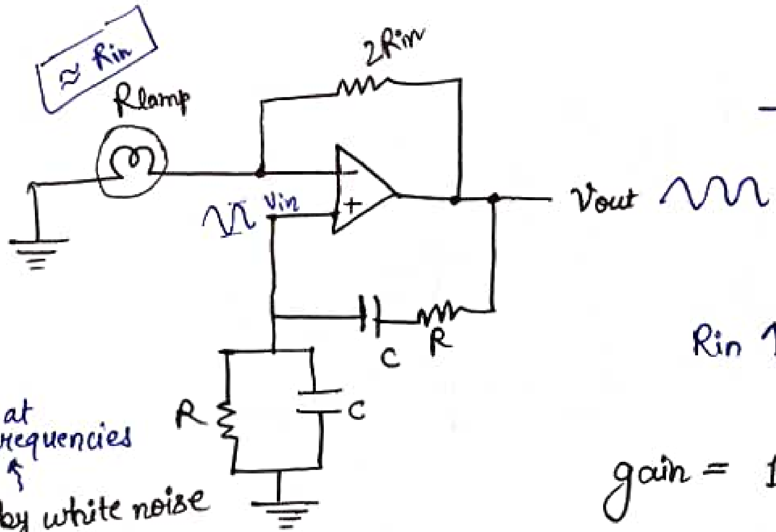
① Feedback Reinforcing Mech.

② |forward gain × feedback gain| = 1
or, $|AB| = 1$

③ Phase of feedback signal should be 180°.

} Barkhausen Criteria





→ Wein Bridge Oscillator

present at all frequencies by white noise

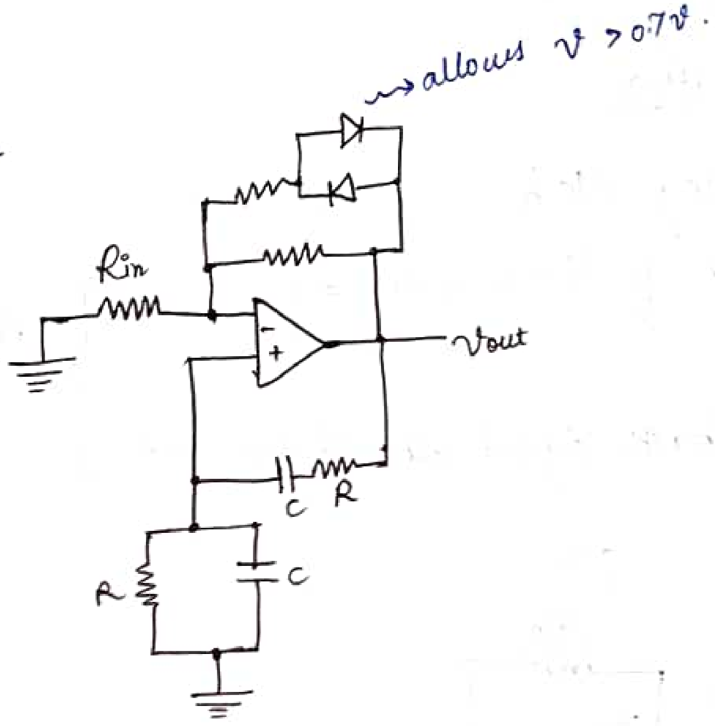
Rin ↑ when current ↑

$$gain = 1 + \frac{2R_{in}}{R_{lamp}}$$

Say $V_{out}\uparrow$, $V_{in}\uparrow$, $V_{lamp}\uparrow$, $I_{lamp}\uparrow$, $R_{lamp}\uparrow$, gain ↓, $V_{out}\downarrow$.

[Make sure than Rin is fixed or has lower thermal coefficient than that of Rlamp.]

II



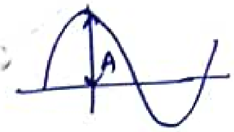
14-11-2023

Power Amplifiers

↳ Deal with high output power.

For 100 W of output power,

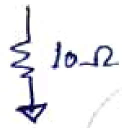
$$100\,W = \frac{A^2}{2R_L} \rightsquigarrow P_{RMS} = \left(\frac{A}{\sqrt{2}}\right)^2 \frac{1}{R_L}$$



$R_L = 10\,\Omega$

$\Rightarrow A = \sqrt{2000}$

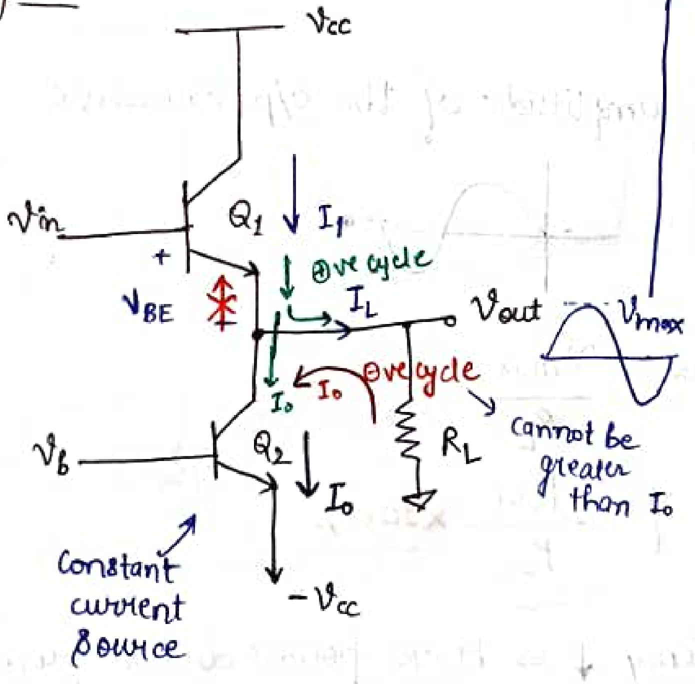
$= 44.72\,V$



$\Rightarrow I_{max} = 4.4\,A$

↳ Power amplifiers come with heat sink.
  ↳ Have lower β.

Class A amplifier:



Positive half cycle:

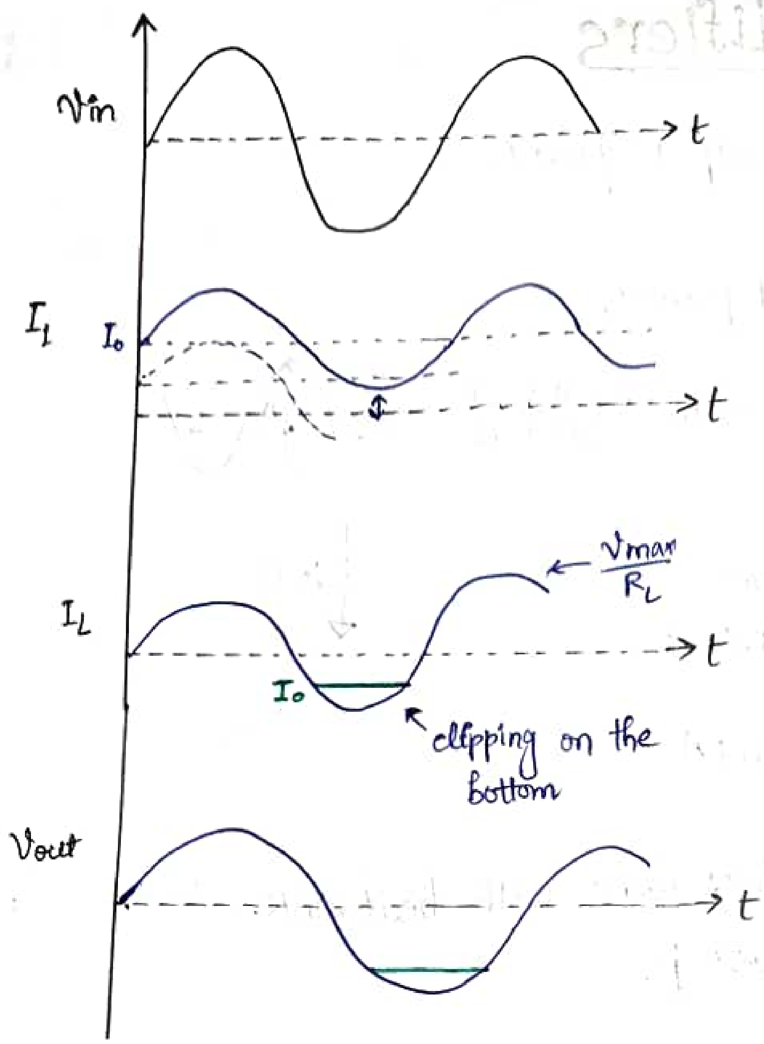
$V_{in} \uparrow \Rightarrow V_E(Q_1) \uparrow$

$\Rightarrow V_{BE} \uparrow$

$\Rightarrow I_1 \uparrow$

$I_0 \rightarrow$ constant

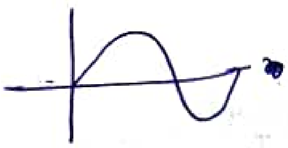
$$I_1 = I_L + I_0$$



$$I_o \geq \frac{V_{max}}{R_L}$$

$$I_o < \frac{V_{max}}{R_L}$$

V_{max} : Max amplitude of the o/p sinewave



$$I_{max} = \frac{V_{max}}{R_L}$$

Efficiency : $\eta = \frac{P_{out}}{P_{DC}} \times 100\%$

Efficiency $\downarrow \Rightarrow$ More power drawn from the source.

$$P_{out} = \frac{V_{out}^2}{2R_L}$$

$$P_{DC} = 2 \times V_{cc} \times I_o$$

$$\eta = \frac{V_{out}^2}{2R_L} \cdot \frac{1}{2V_{cc} \times I_0} \times 100\%$$

$$= \frac{1}{4} \cdot \frac{V_{out}^2}{V_{cc} \times I_0 R_0} \times 100\%$$

$\Rightarrow$ Output voltage $\uparrow \Rightarrow \eta \uparrow$

$$\boxed{V_{out_{max}} = I_0 R_L = V_{cc}}$$

$\hookrightarrow \eta_{max} = 25\%$ $\quad (\because I_0 R_L = V_{max})$

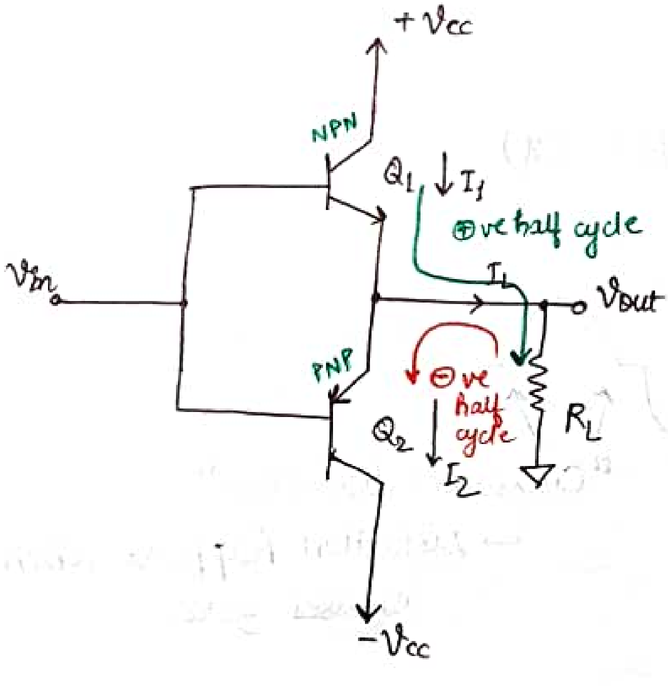
$\hookrightarrow$ Theoretical limit

$\eta_{max} \rightarrow V_{max}$
$\downarrow$
V_{cc}

→ Conduction angle = 360° (2π)

$\hookrightarrow$ Duration (of time) of sine wave for which each of the transistor is conducting current ~~working~~.

Class B amplifier:

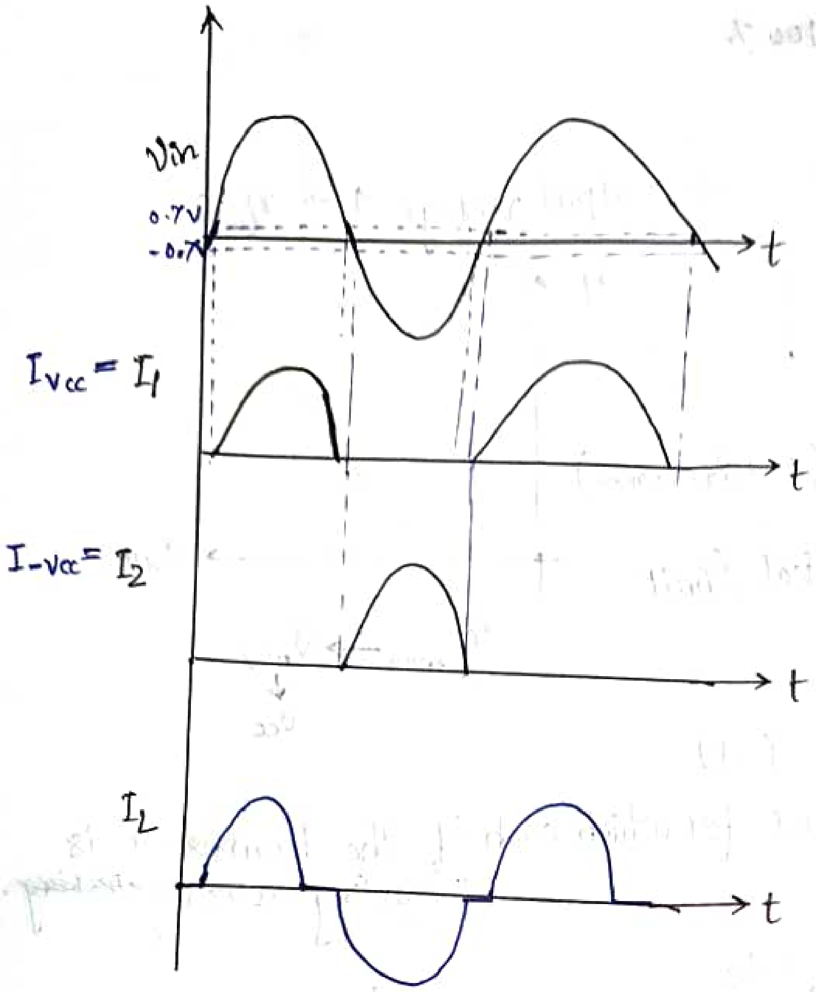


Positive half cycle:

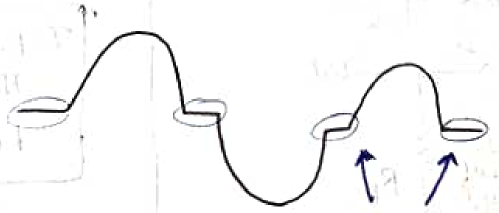
Q_1: ON

Q_2: OFF (pnp)

To turn ON,
NPN: $V_{BE} > 0.7$
PNP: $V_{BE} < -0.7$



Conduction angle = 180° (π)



"Crossover distortion"
↳ Distortion happens when Vin crosses zero.

$$P_{out} = \frac{V_{out}^2}{2R_L}$$

$$P_{DC} = \frac{2 \times V_{out}}{\pi R_L} \times V_{cc}$$ (using Fourier series)

For η_{max}, $V_{out} = V_{cc}$.

$$\eta_{max} = \frac{\pi}{4} \times 100\% \Rightarrow \boxed{\eta_{max} = 78.4\%}$$

17-11-2023

Class AB Power Amplifier

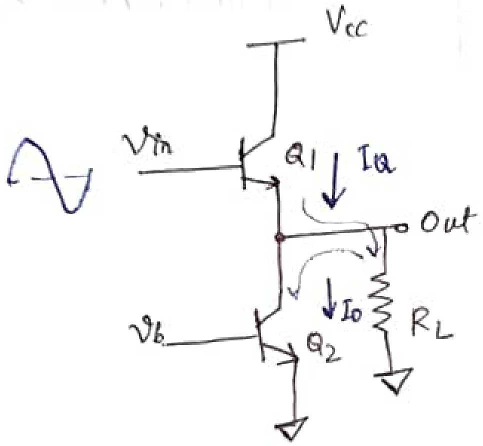
→ Conduction angle $> 180°\ (\pi)$
$< 360°\ (2\pi)$

→ Small Quiscient current (I_Q)

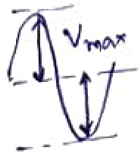
Class A : $I_Q \geq I_{max}$

Class B : $I_Q = 0$

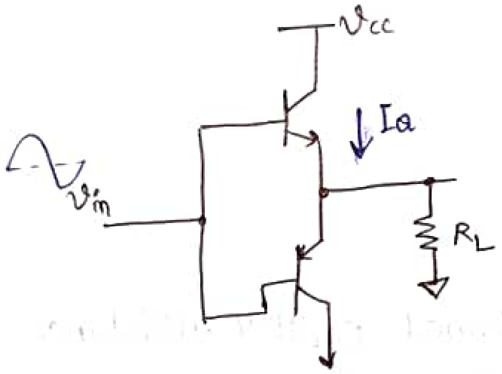
Class AB: $I_Q > 0$ and $I_Q \ll I_{max}$



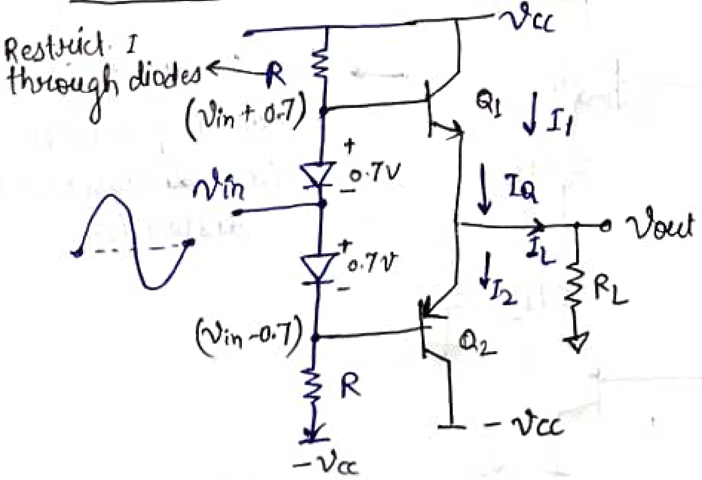
Pout 10 W, $R_L = 10\Omega$
↘ I_{max}



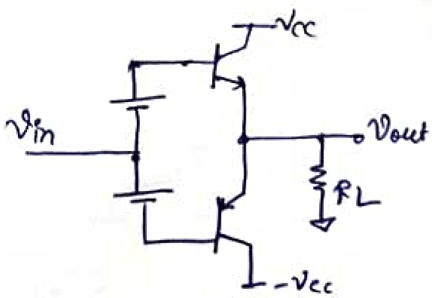
$$\frac{V_{max}}{R_L} = I_{max}$$



Class AB PA :



When $V_{in} = 0$, the transistors should just turn on.



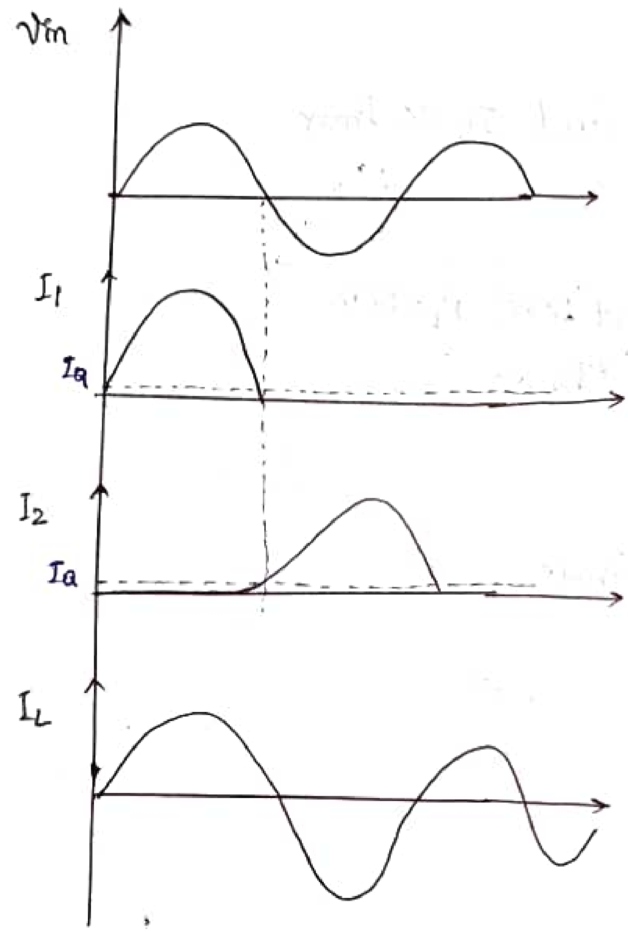
← Something like this
← But this can't be put into practice

By using this bias, crossover distortion can be avoided, while maintaing a good efficiency.

Choose I_Q such that

$I_Q \downarrow \rightarrow$ crossover distortion ~~can be~~ avoided

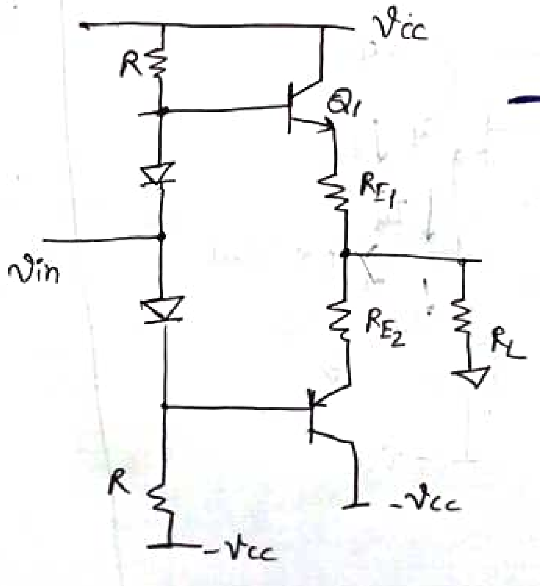
$I_Q \uparrow \rightarrow$ Reduce efficiency $\rightarrow$ maintain efficiency.



Class C PA

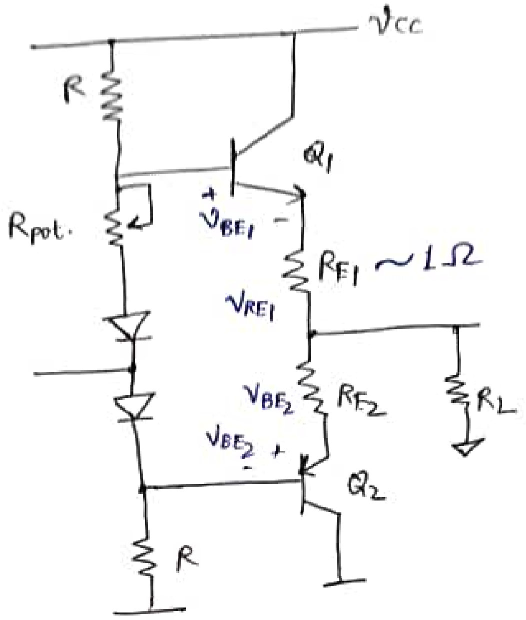
↳ Conduction Angle $< 180°$ (π)

→ To avoid thermal runaway problem, connect emitter resistance to both the resistors (these provide thermal stability).



→ But we again get into class B problem, i.e., crossover distortion.

Now, connect a potentiometer, as we need more than 2 diode drops.



$(2V_D + I R_{pot.})$

Short-circuit Protection:

If output is connected to ground by mistake, a large current will flow through both the transistors and will permanantely damage them.

To detect the short-circuit, look for the voltage drop across R_E's.

To prevent the short-circuit, reduce the current.

As supply current is difficult to reduce, so add two transistors.