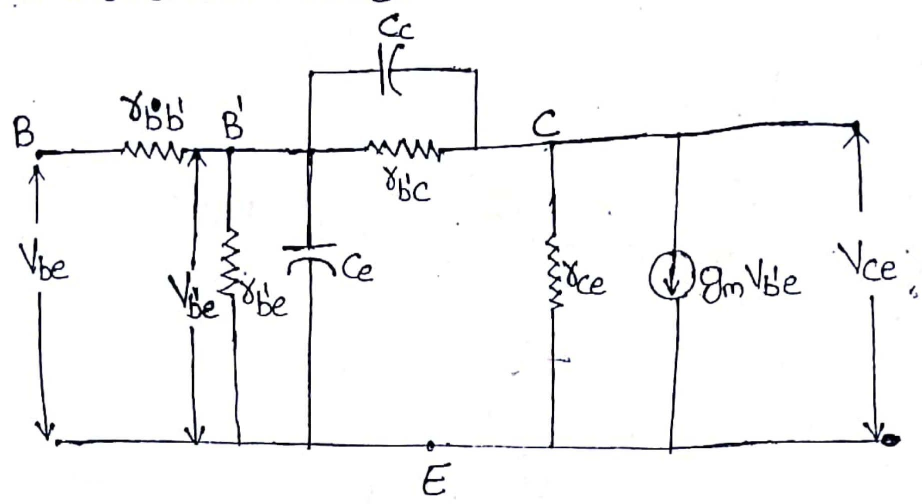
23/11/18
Friday.

Transistor High Frequency Model (or) Hybrid π model.
(or)
Giocoletto Model.

The h-parameter model is valid at low and mid frequenci at high frequencies the junction capacitors of the transis-tor begin to dominate with frequency response. Giocoletto proposed a model for high frequencies. This model is called High frequency model (or) Hybrid π model or Giocoletto model.



$r_{bb'}$ = base spreading resistance.

$g_m V_{b'e}$ = This represents collector current which is a result of excess carries injected by emitter to the base. This current is directly propotional to V_{be}

$r_{b'e}$ = This represents resistance across the base emitter junction

$r_{b'c}$ = This represents a resistance across the collector base junction. It is result of base width modula-tion. $r_{b'c}$ is very large.

r_{ce} = This represents a effective resistance between collector and emitter.

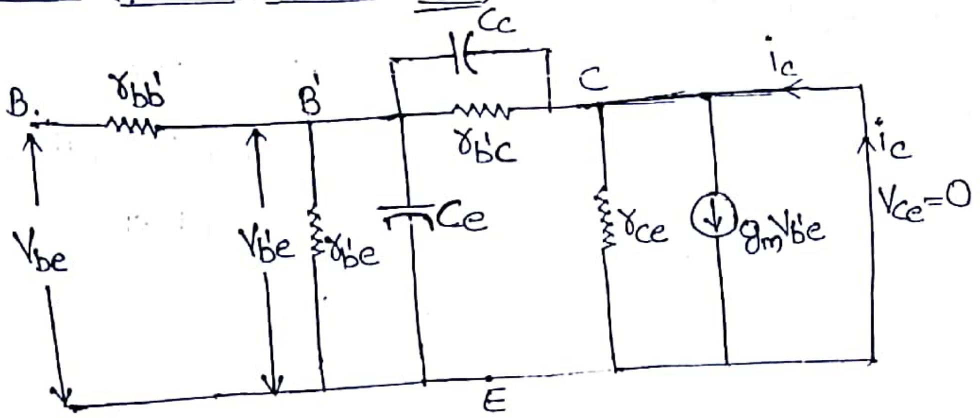
$C_e (C_\pi)$ = This represents emitter base junction capacitance (For-ward bias capacitance & diffusion capacitance)

$C_c (C_\mu)$ = This represents collector base junction capacitance. (Reverse bias capacitance or Transition capacitance)

g_m = Trans conductance.

Determination of hybrid - π - conductances :

Trans conductance (g_m) :- (short circuit Trans conductance)



In the above circuit $V_{ce}=0$.

∴ The short circuit collector current is given by

$$i_c = g_m V_{b'e} \Big|_{V_{ce}=0}$$

$$g_m = \frac{i_c}{V_{b'e}}\Big|_{V_{ce}=0} = \frac{\Delta I_C}{\Delta V_{B'E}}\Big|_{\Delta V_{CE}=0}$$

$V_{B'E} \simeq V_{BE}$ ($r_{bb'}$ is very small).

$$g_m = \frac{\partial I_C}{\partial V_{B'E}}\Big|_{V_{CE}} = \frac{\partial I_C}{\partial V_{BE}}\Big|_{V_{CE}}$$

$$I_C = -\alpha I_E + I_{C0}.$$

$$\frac{\partial I_C}{\partial V_{BE}} = -\alpha \frac{\partial I_E}{\partial V_{BE}}$$

$$\frac{\partial I_C}{\partial V_{BE}} = \frac{-\alpha}{\left(\frac{\partial V_{BE}}{\partial I_E}\right)}$$

$\frac{\partial V_{BE}}{\partial I_E}$ is the dynamic resistance of the base emitter

junction = $\frac{\eta V_T}{\ }$

$$g_m = \left.\frac{\partial I_c}{\partial V_{BE}}\right|_{V_{CE}} = \frac{-\alpha}{\frac{\eta V_T}{I_E}} = \frac{-\alpha I_E}{\eta V_T}$$

$\therefore$ $\eta = 1$ for silicon.

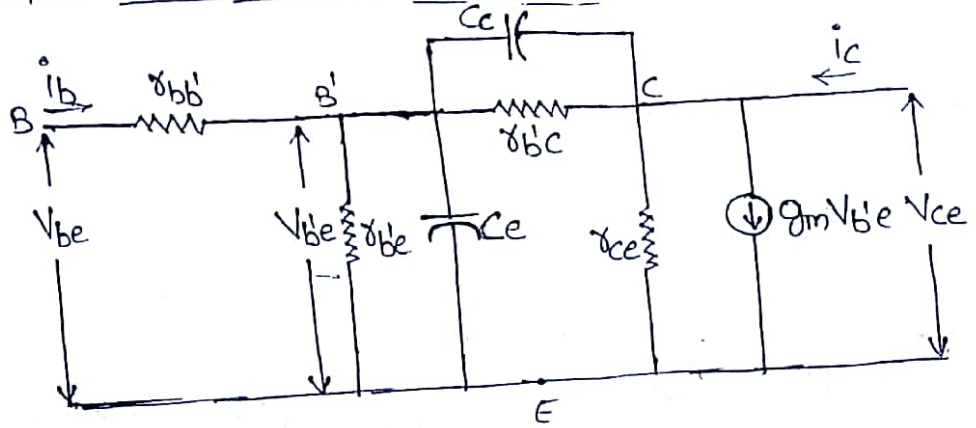
$$g_m = \frac{-\alpha I_E}{V_T}$$

$$I_c \simeq -\alpha I_E$$

$$\boxed{g_m = \left|\frac{I_c}{V_T}\right|}$$

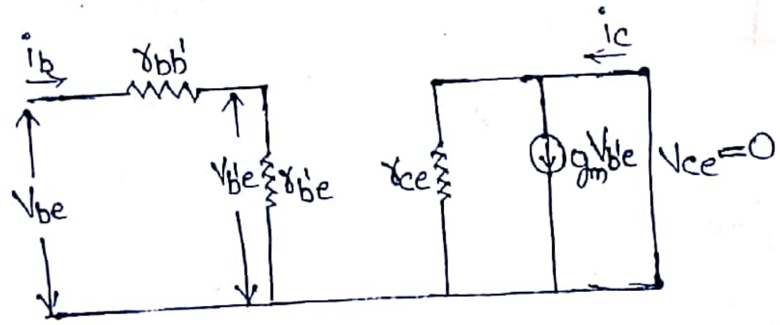
24/11/18
Saturday.

Input conductance $g_{b'e}$ $(1/r_{b'e})$:- (short circuit input conductance)



Consider the hybrid π model at low frequencies. The capacitors C_e, C_c can be replaced with open circuit because their impedance will be very high at low frequencies. $r_{b'c}$ is a very large resistor so that it can be replaced with open circuit.

Set $V_{ce} = 0$ to find i_c

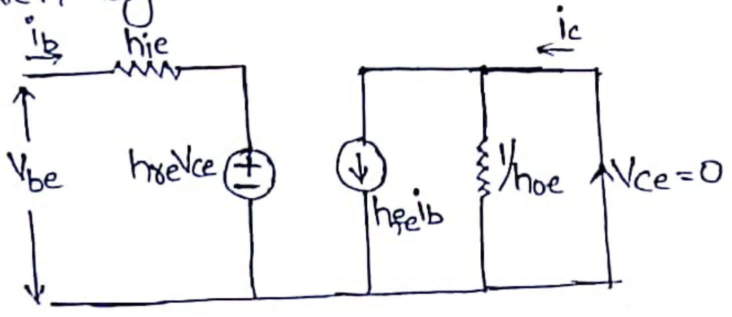


$$i_c = g_m V_{b'e}$$

$$V_{b'e} = i_b r_{b'e}.$$

$$i_c = (g_m r_{b'e})(i_b)$$

low frequencies the h-parameter model with $V_{ce}=0$ is
ven by.



From the above circuit.

$$i_c = h_{fe}\, i_b.$$

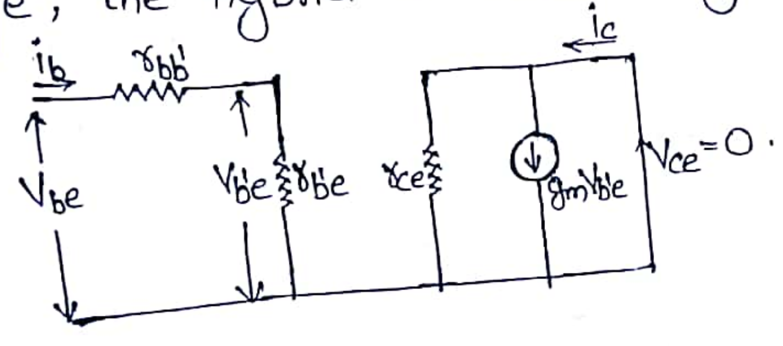
$$\therefore\ g_m\, r_{b'e}\, i_b = h_{fe}\, i_b$$

$$r_{b'e} = \frac{h_{fe}}{g_m}$$

$$\boxed{g_{b'e} = \frac{1}{r_{b'e}} = \frac{g_m}{h_{fe}}.}$$

Base spreading Resistance ($r_{bb'}$):

At low frequencies with $V_{ce}=0$ and $r_{b'c}$ being very
large, the hybrid π model is given by.

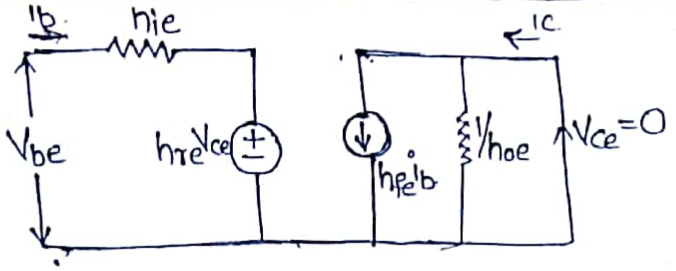


In the above circuit

$$V_{be} = i_b\, r_{bb'} + i_b\, r_{b'e}$$

$$V_{be} = i_b(r_{bb'} + r_{b'e})$$

The h-parameter model with $V_{ce}=0$ is given by



In the above circuit.

$$V_{be} = i_b h_{ie} + h_{re}V_{ce} . \quad (\because V_{ce}=0)$$

$$V_{be} = i_b h_{ie}$$

$$\therefore i_b(r_{bb'} + r_{b'e}) = i_b h_{ie}.$$

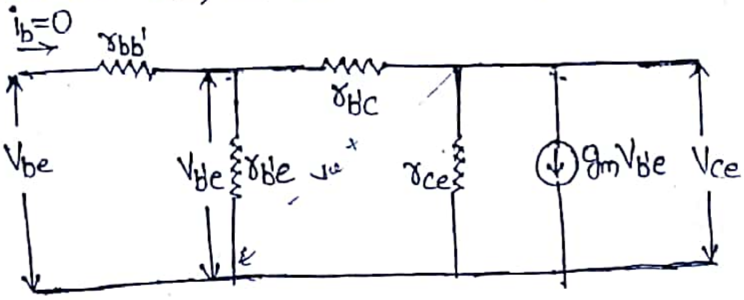
$$r_{bb'} + r_{b'e} = h_{ie}.$$

$$\boxed{r_{bb'} = h_{ie} - r_{b'e}.}$$

26/11/18
Monday.

Feed back conductance ($g_{b'c}$) : (Open circuit feedback conductance) ($1/r_{b'c}$)

If the hybrid π model is used at low frequencies, the capacitors C_e, C_c can be replaced with open circuits.



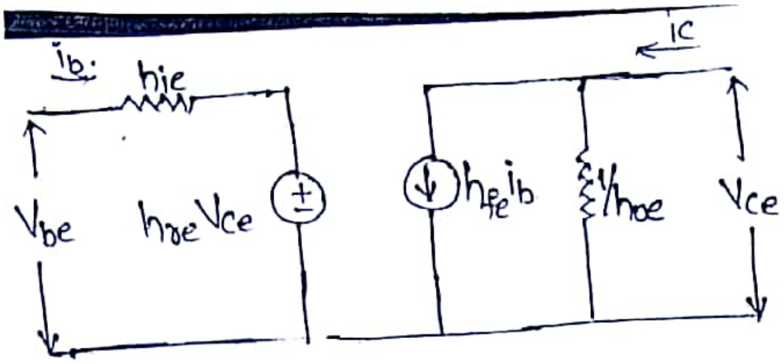
If $i_b=0$, $r_{b'e}$, $r_{b'c}$ will be in series.

voltage across $r_{b'e}$ is

$$V_{b'e} = \frac{V_{ce}\, r_{b'e}}{r_{b'e} + r_{b'c}}$$

$$V_{be} = V_{b'e} = \frac{V_{ce}\, r_{b'e}}{r_{b'e} + r_{b'c}}.$$

The h-parameter model at low frequencies is



$$V_{be} = h_{ie} i_b + h_{re} V_{ce}$$

if $i_b = 0$.

$$V_{be} = h_{re} V_{ce}$$

∴ From the above two circuits we have

$$\frac{V_{ce}\, r_{b'e}}{r_{b'e} + r_{b'c}} = h_{re} V_{ce}.$$

$$\frac{r_{b'e}}{r_{b'e} + r_{b'c}} = h_{re}.$$

$$r_{b'e} = h_{re}\, r_{b'e} + h_{re}\, r_{b'c}$$

$$h_{re}\, r_{b'c} = r_{b'e}(1 - h_{re})$$

$$h_{re} << 1$$

$$h_{re}\, r_{b'c} = r_{b'e}.$$

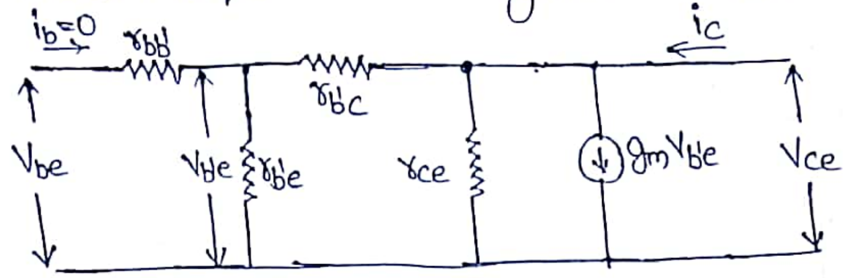
$$r_{b'c} = \frac{r_{b'e}}{h_{re}}$$

$$\boxed{g_{b'c} = \frac{1}{r_{b'c}} = \frac{h_{re}}{r_{b'e}} = g_{b'e}\, h_{re}}$$

28/11/18
Wednesday.

Output conductance (g_{ce}) :- (Open circuit output conductance) ($1/r_{ce}$)

At low frequencies the hybrid π model is given by.



set $i_b = 0$, $r_{b'e}$, $r_{b'c}$ will be in series.

By applying KCL

$$i_c = g_m V_{b'e} + \frac{V_{ce}}{r_{ce}} + \frac{V_{ce}}{r_{b'c} + r_{b'e}}$$

$$\frac{1}{r_{ce}} = g_{ce}.$$

$$g_{b'c} = \frac{1}{r_{b'c}}.$$

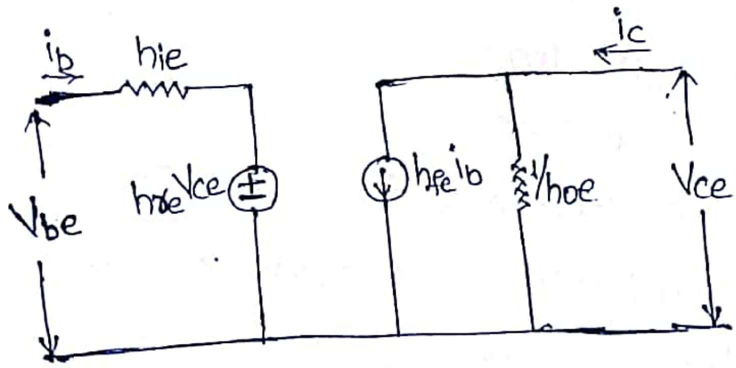
$$V_{b'e} = \frac{V_{ce} \cdot r_{b'e}}{r_{b'c} + r_{b'e}}$$

$$i_c = \frac{g_m r_{b'e}}{r_{b'c} + r_{b'e}} V_{ce} + V_{ce} g_{ce} + \frac{V_{ce}}{r_{b'c}} \quad [\because r_{b'c} >> r_{b'e}]$$

$$i_c = g_m h_{re} V_{ce} + V_{ce} g_{ce} + V_{ce} g_{b'c} \quad \left[\because h_{re} = \frac{r_{b'e}}{r_{b'e} + r_{b'c}}\right]$$

$$\frac{i_c}{V_{ce}} = g_m h_{re} + g_{ce} + g_{b'c}. \longrightarrow (1)$$

The h-parameter model at low frequencies is given by



By applying KCL at o/p node.

$$i_c = h_{fe} i_b + \frac{V_{ce}}{1/h_{oe}}$$

if $i_b = 0$, $i_c = h_{oe} V_{ce}$

$\frac{i_c}{V_{ce}} = h_{oe} \longrightarrow ②$

quating the above equations ① & ② we get

$h_{oe} = g_m h_{re} + g_{ce} + g_{b'c}$

$g_{ce} = h_{oe} - g_m h_{re} - g_{b'c}$

But $r_{b'c}$ is very large.

$\therefore g_{b'c}$ is very small.

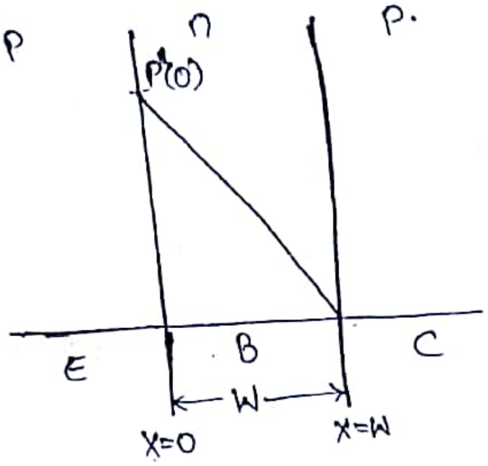
$$\boxed{g_{ce} = h_{oe} - g_m h_{re}}$$

Derivation of hybrid π capacitances:-

Diffusion capacitance (C_e) & (C_π) & ($C_{b'e}$):-

(or)

Base Emitter Junction capacitance:-



The above figure BE junction is forward biased and BC junction is reverse biased.

$P'(0)$ is the excess hole concentration at $x=0$

The excess hole concentration at $x=W$ will be zero because all the holes will be swept by the collector.

$\therefore$ The average excess hole concentration in the base is given by $\frac{1}{2} P'(0)$

~~Total charge~~

The excess charge in the base region is given by.

$$Q = \frac{1}{2} p'(0) \times q \times AW$$

The diffusion hole current density is given by

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

$$p' = p - p_0.$$

$$\frac{dp'}{dx} = \frac{dp}{dx} - 0$$

$$J_p = -q D_p \frac{dp'}{dx}$$

Current $I = A J_p = -A q D_p \frac{dp'}{dx}$

$$I = -A q D_p \left(\frac{0 - p'(0)}{W - 0} \right)$$

$$I = \frac{A q D_p p'(0)}{W}$$

$$p'(0) = \frac{I W}{A q D_p}$$

$$\therefore Q = \frac{1}{2} \frac{I W}{\cancel{A} \cancel{q} D_p} \times \cancel{q} \times \cancel{A} W$$

$$Q = \frac{I W^2}{2 D_p}.$$

The diffusion capacitance $C_e = \frac{dQ}{dV} = \frac{d}{dV}\left(\frac{I W^2}{2 D_p}\right)$

$$= \frac{W^2}{2 D_p} \cdot \frac{dI}{dV}$$

$$= \frac{W^2}{2 D_p} \frac{1}{\left(\frac{dV}{dI}\right)}$$

$\frac{dV}{dI} = r_e =$ dynamic resistance of BE Junction.

$$\therefore C_e = \frac{1}{2} \frac{W^2}{D_p} \frac{1}{r_e}.$$

W.K.T. $r_e = \frac{\eta V_T}{I_E}$, $\eta = 1$ for Si

$$r_e = \frac{V_T}{I_E}$$

$$\therefore C_e = \frac{1}{2} \frac{W^2}{D_p} \cdot \frac{I_E}{V_T}$$

~~IIIly~~ $I_C = \alpha I_E$; $\alpha \simeq 1$

$$I_C \simeq I_E$$

$$C_e = \frac{1}{2} \frac{W^2}{D_p} \left(\frac{I_C}{V_T}\right)$$

$$\boxed{C_e = \frac{g_m W^2}{2 D_p}}$$

Transition capacitance (C_c) or (C_μ) :-

(or)

Collector-Base Junction capacitance ($C_{b'c}$) :-

Collector Base junction ~~capacitance~~ is a Reverse biased junction.

The capacitance at this junction is given by

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

where $W \rightarrow$ width of the depletion Region.

$$W \propto \sqrt{V_j}$$

where $V_j = V_0 + V_d$

The General expression for the Reverse biased capacitance is given by

$$C_c = \frac{C_0}{\left(1 + \frac{V_d}{V_0}\right)^{1/n}}$$

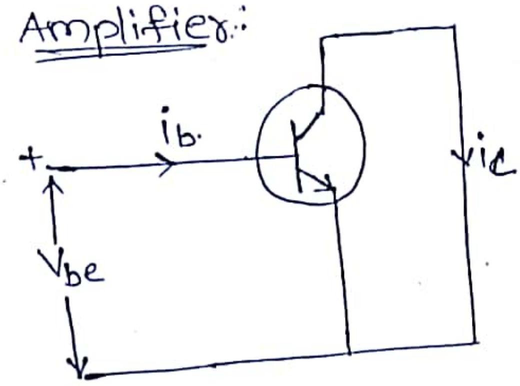
$$C_c = \frac{C_0}{\left(1 + \frac{V_d}{V_0}\right)^{1/n}}$$

where $n = 2$ or 3

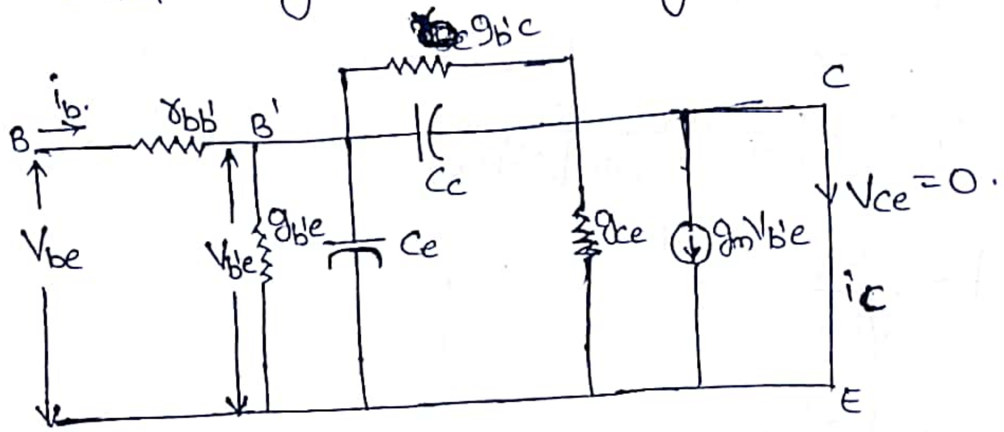
29/11/18
Thursday.

CE short circuit current gain (or) β cut-off frequency (f_β)
(or) CE short circuit unity gain frequency (f_T) (or)
CE short circuit unity gain Bandwidth product (f_T)
(or) Derive β, α cut off frequencies for a short circui

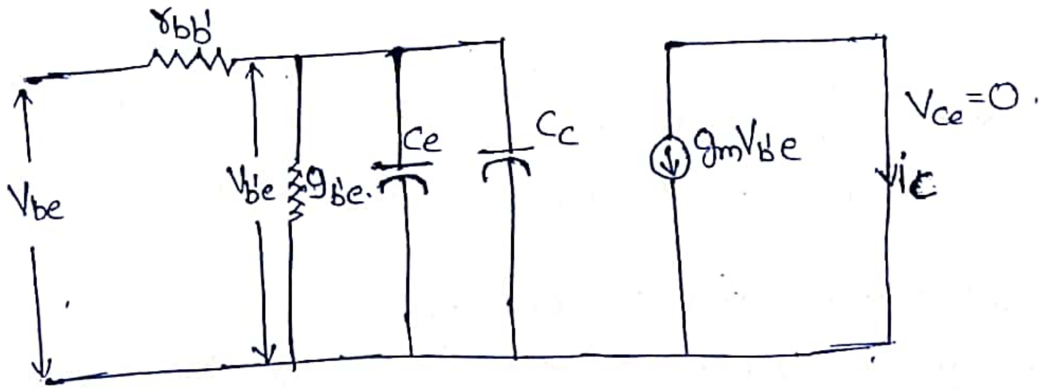
CE Amplifier:



The above circuit shows a CE Amplifier with short circuited output. Replacing the transistor with hybrid π model to get the following circuit.



$r_{b'c}$ is very large & $r_{bb'}$ is very small.
g_{ce} is short circuited. The modified circuit is given below.



Let $C_e + C_c = C_i$

KCL at input node

$$i_b = V_{b'e}\, g_{b'e} + \frac{V_{b'e}}{\frac{1}{j2\pi f C_i}}$$

$$i_b = V_{b'e}\left[g_{b'e} + j2\pi f C_i\right]$$

current gain $A_I = \frac{i_c}{i_b} = \frac{-g_m \cancel{V_{b'e}}}{\cancel{V_{b'e}}\left[g_{b'e} + j2\pi f C_i\right]}$

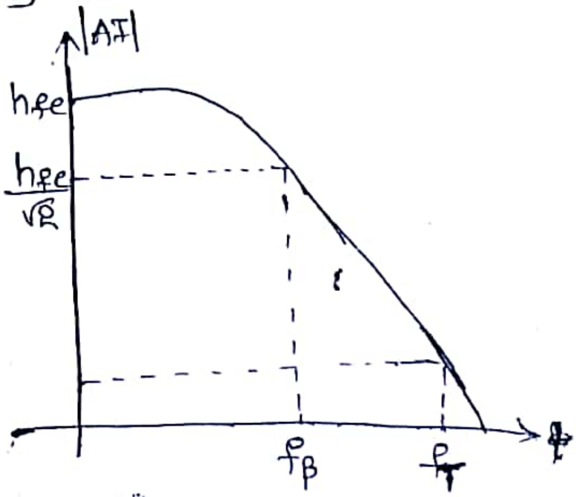
$$A_I = \frac{-g_m}{g_{b'e} + j2\pi f C_i}$$

$$A_I = \frac{-g_m / g_{b'e}}{1 + j\frac{2\pi f C_i}{g_{b'e}}}$$

$$\boxed{A_I = \frac{-h_{fe}}{1 + j\left(f/f_\beta\right)}}$$

where $f_\beta = \frac{g_{b'e}}{2\pi C_i} = \frac{g_{b'e}}{2\pi\left[C_c + C_e\right]}$

$$|A_I| = \frac{h_{fe}}{\sqrt{1 + \left(f/f_\beta\right)^2}}$$



at $f = 0$, $|A_I| = h_{fe}$

at $f = f_\beta$, $|A_I| = \frac{h_{fe}}{\sqrt{2}}$.

f_β is upper cut off frequency and bandwidth.

f_β is also called as β- cut off frequency.

* h_{fe} is the maximum current gain.

Let f_T be the unity gain frequency. At this frequency

$$|A_I| = 1$$

$\therefore$ At $f = f_T$, $|A_I| = 1$

$$\Rightarrow 1 = \frac{h_{fe}}{\sqrt{1 + (f_T/f_\beta)^2}}$$

As $f_T >> f_\beta$, $\Rightarrow \frac{f_T}{f_\beta} >> 1$

$$1 = \frac{h_{fe}}{\sqrt{\left(\frac{f_T}{f_\beta}\right)^2}}$$

$$\frac{f_T}{f_\beta} = h_{fe}.$$

$$f_T = h_{fe} f_\beta \doteq (\text{max gain})(\text{bandwidth}).$$

* For a given amplifier, gain bandwidth product is constant.

<u>α-cut off frequency (f_α):-</u>

f_α is cut off frequency for a short circuit CB amplifier. The short circuit current gain is given by.

$$|A_I| = \frac{h_{fb}}{\sqrt{1+(f/f_\alpha)^2}} \quad ; \quad A_I = \frac{-h_{fb}}{1+j(f/f_\alpha)}$$

At $f = f_\alpha$; $|A_I| = \frac{h_{fb}}{\sqrt{2}}$.

$\therefore f_\alpha$ is upper cut off frequency and bandwidth.

Let f_T be the unity gain frequency.

at $f = f_T$, $|A_I| = 1$, $f_T >>> f_\alpha$.

$$1 = \frac{h_{fb}}{\sqrt{1+(f_T/f_\alpha)^2}} \simeq \frac{h_{fb}}{\sqrt{(f_T/f_\alpha)^2}}$$

$$\therefore f_T = h_{fb} \cdot f_\alpha.$$

$$f_T = (\text{gain})(\text{band width})$$

For a given amplifier; f_T is constant.

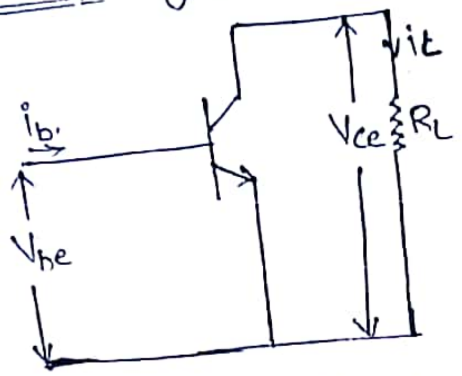
$\therefore h_{fe} f_\beta = h_{fb} f_\alpha$.

$$f_\alpha = \left(\frac{h_{fe}}{h_{fb}}\right) f_\beta.$$

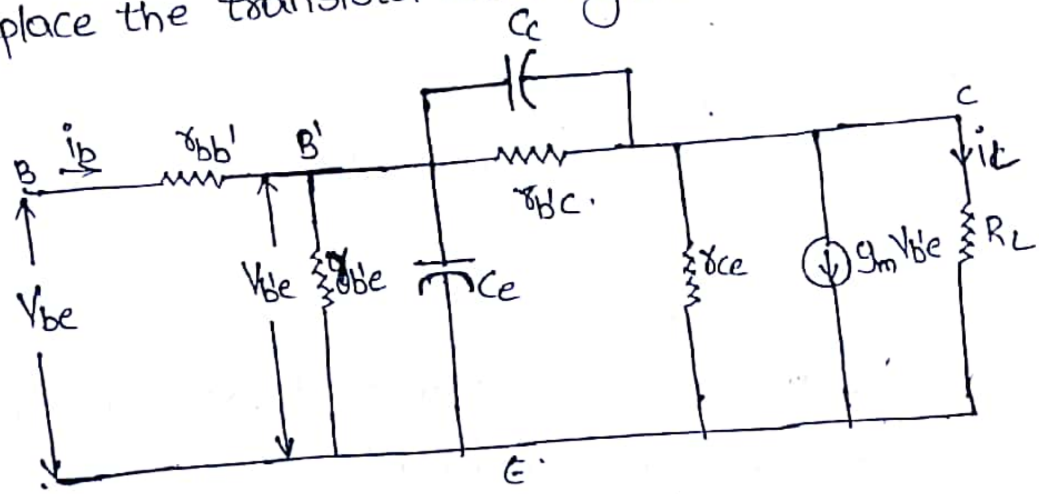
$$h_{fb} = \frac{h_{fe}}{1+h_{fe}} \Rightarrow \frac{h_{fe}}{h_{fb}} = 1+h_{fe}.$$

$$\therefore f_\alpha = (1+h_{fe}) f_\beta$$

CE current gain with resistive load :-

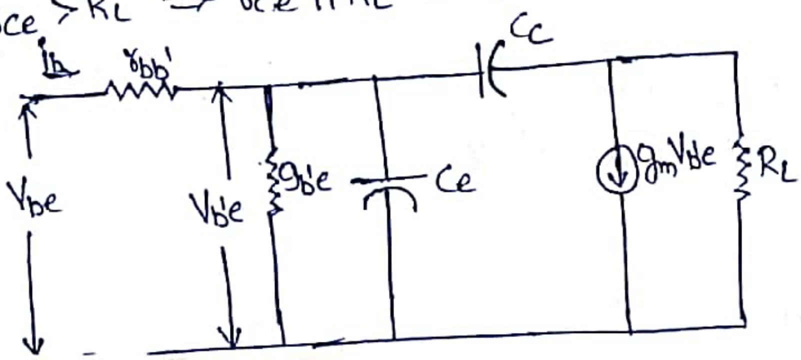


Replace the transistor with hybrid π model



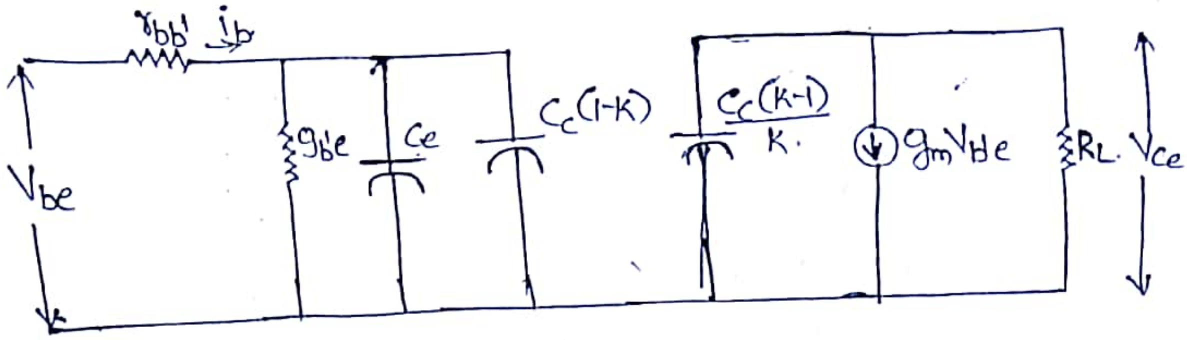
$r_{b'c}$ is very large → open circuit.

$r_{ce} > R_L \Rightarrow r_{ce} \| R_L \simeq R_L$



The above circuit is modified using miller's theorem.

Let $\frac{V_{ce}}{V_{b'e}} = K$ then the above circuit is given as.

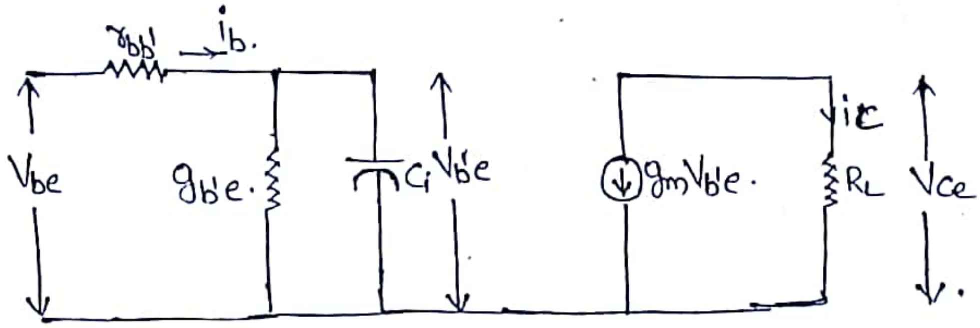


$$C_c\left(\frac{K-1}{K}\right) = C_c\left(1 - 1/K\right)$$

K is voltage gain of CE amplifier and K is a large negative number.

$$\therefore C_c(1 - 1/K) \simeq C_c$$

$$(-jX_c) \,||\, R_L \simeq R_L \qquad (C_c || R_L \simeq R_L)$$



where $C_i = C_e + C_c$;

$$K = \frac{V_{ce}}{V_{b'e}} = \frac{i_c R_L}{V_{b'e}} = \frac{-g_m \cancel{V_{b'e}} . R_L}{\cancel{V_{b'e}}} = -g_m R_L$$

$$i_c = -g_m V_{b'e}.$$

$$i_b = V_{b'e}(g_{b'e}) + \frac{V_{b'e}}{\frac{1}{j2\pi f C_i}}$$

$$i_b = V_{b'e}\left(g_{b'e} + j2\pi f C_i\right)$$

$$\text{Gain} = \frac{i_c}{i_b} = \frac{-g_m V_{b'e}}{(g_{b'e} + j2\pi f C_i) V_{b'e}}$$

$$A_I = \frac{-g_m}{(g_{b'e} + j2\pi f C_i)}$$

$$= \frac{-g_m / g_{b'e}}{1 + j\frac{2\pi f C_i}{g_{b'e}}}$$

$$A_I = \frac{-h_{fe}}{1 + j(f/f_\beta)}$$

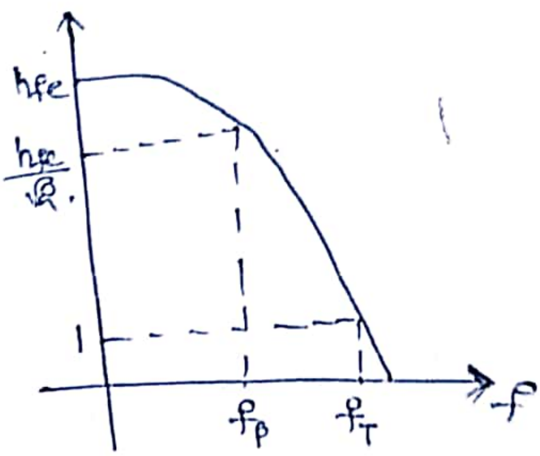
where $f_\beta = \frac{g_{b'e}}{2\pi C_i} = \frac{g_{b'e}}{2\pi\left[C_e + C_c(1+g_m R_L)\right]}$.

$$|A_I| = \frac{h_{fe}}{\sqrt{1+(f/f_\beta)^2}}.$$

at $f = f_\beta$, $|A_I| = \frac{h_{fe}}{\sqrt{2}}$.

at $f = f_T$, $|A_I| = 1$

$$1 = \frac{h_{fe}}{\sqrt{1+\left(\frac{f_T}{f_\beta}\right)^2}}$$



$\therefore \frac{f_T}{f_\beta} >>> 1$

$$1 = \frac{h_{fe}}{f_T/f_\beta}.$$

$$\frac{f_T}{f_\beta} = h_{fe}$$

$$\boxed{f_T = h_{fe} \cdot f_\beta.}$$

Typical values of Hybrid π model parameters:-

$g_m = 50mA/V = 5$ mhos.

$r_{bb'} = 100\Omega$

$r_{b'e} = 1K$.

$r_{ce} = 80K$

$r_{b'c} = 4M\Omega$

$C_e = 100pf$

$C_c = 3pf$.

Formulas:-

* Transconductance $g_m = \frac{I_C}{V_T}$

* Input conductance $g_{b'e} = \frac{g_m}{h_{fe}}$.

* Base spreading Resistance $r_{bb'} = h_{ie} - r_{b'e}$

* Feed back conductance $g_{b'c} = g_{b'e} h_{re}$

* Output conductance $g_{ce} = h_{oe} - g_m h_{re}$.

* $f_\beta = \frac{g_{b'e}}{2\pi[C_e + C_c]}$

* $f_T = h_{fe} f_\beta = \frac{h_{fe} g_{b'e}}{2\pi[C_e + C_c]}$

we know that

$g_m = g_{b'e} h_{fe}$.

$\therefore f_T = \frac{g_m}{2\pi(C_e + C_c)}$

etermine the trans conductances of a hyb transistor of hybrid π model with $\beta = 1$, $h_{ie} = 4K$, $h_{re} = 2.5 \times 10^{-4}$, $h_{fe} = 224$. $h_{oe} = 25 \mu A/V$ and the biasing current $i_c = 2mA$ at the room temperature

A. Given.

$\beta = 1$

$h_{ie} = 4K$

$h_{re} = 2.5 \times 10^{-4}$

$h_{fe} = 224$

$h_{oe} = 25 \mu A/V$

$i_c = 2mA$.

$V_T = 26mV$

$g_m = \frac{i_c}{V_T} = \frac{2\ mA}{26\ mV} = 0.0769 mhos = 76.9\ m.\ mhos$

$g_{b'e} = \frac{g_m}{h_{fe}} = \frac{76.9 \times 10^{-3}}{224} = 0.343 m\ mhos.$

$r_{b'e} = \frac{1}{g_{b'e}} = 2.9 K\Omega$

$r_{bb'} = h_{ie} - r_{b'e} = 4K - 2.9K = 1.08 K.$

$g_{b'c} = g_{b'e} \cdot h_{re} = 0.343 \times 10^{-3} \times 2.5 \times 10^{-4} = 85.8 n\ mhos.$
$= 0.0858\ \mu mhos.$

$r_{b'c} = \frac{1}{g_{b'c}} = 11.66\ M\Omega.$

$g_{ce} = h_{oe} - g_m h_{re} = 25\mu - 76.9m \times 2.5 \times 10^{-4}$
$= 5.77\ \mu mhos.$

21/12/18
Saturday.

2. Following are a low frequency parameters for a transistor at room temperature. $|I_c| = 5mA$, $h_{ie} = 1K\Omega$, $h_{fe} = 100$, $h_{oe} = 4\times10^{-5}$ $h_{re} = 10^{-4}$, $C_{ob} = 2pf$, $f_T = 10MHz$. compute various high parameters

A. Given.

$|i_c| = 5mA$

$h_{ie} = 1K\Omega$

$h_{fe} = 100$

$h_{oe} = 4\times10^{-5}$

$h_{re} = 10^{-4}$

$f_T = 10MHz.$

$C_{cb} = 2PF$

$V_T = 26mV$

$g_m = \frac{i_c}{V_T} = \frac{5mA}{26mA} = 192.30$ m.mhos.

$g_{b'e} = \frac{g_m}{h_{fe}} = \frac{192.30\times10^{-3}}{100} = 1.92$ m mhos.

$r_{b'e} = \frac{1}{g_{b'e}} = \frac{1}{1.92\times10^{-3}} = 520\,\Omega.$

$r_{bb'} = h_{ie} - r_{b'e} = 1K - 520\Omega = 480\,\Omega.$

$g_{b'c} = g_{b'e}h_{re} = 1.92\times10^{-3}\times10^{-4} = 192$ n.mhos.

$r_{b'c} = \frac{1}{g_{b'c}} = 5.20\,M\Omega.$

$g_{ce} = h_{oe} - g_m h_{re} = 4\times10^{-5} - 192.30\times10^{-3}\times10^{-4} = 20.77\ \mu$mhos.

we know that

$$f_T = \frac{g_m}{2\pi(C_e+C_c)}$$

$$10\times10^{6} = \frac{192.30\times10^{-3}}{2\pi(C_e+2\times10^{-12})}$$

$$2\times10^{-12} + C_e = \frac{192.30\times10^{-3}}{2\pi\times10\times10^{6}}.$$

$$C_e = 3.06\times10^{-9} - 2\times10^{-12}$$

$$C_e = 3.05\,nF$$

5. Following are the measurements of a transistor at room temperature, $i_c = 5mA$, $V_{ce} = 10V$, $h_{fe} = 100$, $h_{ie} = 600\Omega$, $|A_I| = 10$ at $10MHz$., $C_c = 3pF$. Find $r_{b'e}$, $r_{bb'}$, f_β, f_T, C_e, f_α.

A. Given

$i_c = 5mA$

$V_{ce} = 10V$

$h_{fe} = 100$

$h_{ie} = 600\Omega$

$|A_I| = 10$ at $10MHz$

$C_c = 3pF$.

$$g_m = \frac{i_c}{V_T} = \frac{5mA}{26mA} = 192.30\ m\ mhos.$$

$$g_{b'e} = \frac{g_m}{h_{fe}} = \frac{192.30\,m\,mhos}{100} = 1.92\ m\ mhos.$$

$$r_{b'e} = \frac{1}{g_{b'e}} = 520\,\Omega$$

$$r_{bb'} = h_{ie} - r_{b'e} = 600 - 520 = 80\Omega$$

we know that

$$|A_I| = \frac{h_{fe}}{\sqrt{1+(f/f_\beta)^2}}.$$

at $f = 10MHz$, $|A_I| = 10$.

$$10 = \frac{100}{\sqrt{1+\left(\frac{10\times10^6}{f_\beta}\right)^2}}$$

$$\sqrt{1+\left(\frac{10\times10^6}{f_\beta}\right)^2} = \frac{100}{10}.$$

$$1+\left(\frac{10\times10^6}{f_\beta}\right)^2 = 100.$$

$$\left(\frac{10\times10^6}{f_\beta}\right)^2 = 99$$

$$\frac{(10\times10^6)^2}{99} = f_\beta^2.$$

$$f_\beta = \sqrt{\frac{(10\times10^6)^2}{\cancel{8}.99}}$$

$$f_\beta = 1\ MHz.$$

$$f_T = h_{fe}\, f_\beta.$$

$$= 100\times 1M$$

$$= 100MHz.$$

$$f_T = \frac{g_m}{2\pi\,[C_e + C_c]} \Rightarrow 100\times10^6 = \frac{192.30\times10^{-3}}{2\pi\,[C_e + 3\times10^{-12}]}$$

$$C_e + 3\times10^{-12} = \frac{192.30\times10^{-3}}{2\pi\times100\times10^6}$$

$$\cancel{C_e = 3\times10}$$

$$C_e = 306.05\times10^{-12} - 3\times10^{-12}$$

$$C_e = 303.05\ pF.$$

$$f_\alpha = (1+\beta)\, f_\beta. \quad (\beta \simeq h_{fe})$$

$$f_\alpha = (1+h_{fe})\, f_\beta$$

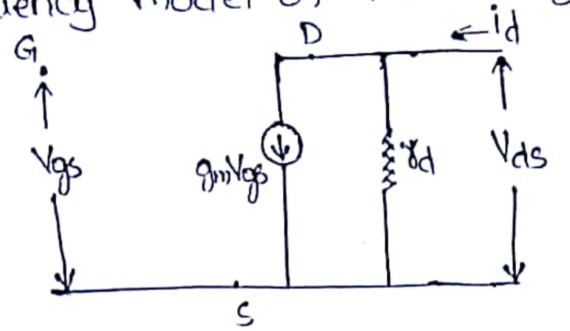
$$f_\alpha = (1+100)\ 1\times10^6.$$

$$= 101MHz.$$

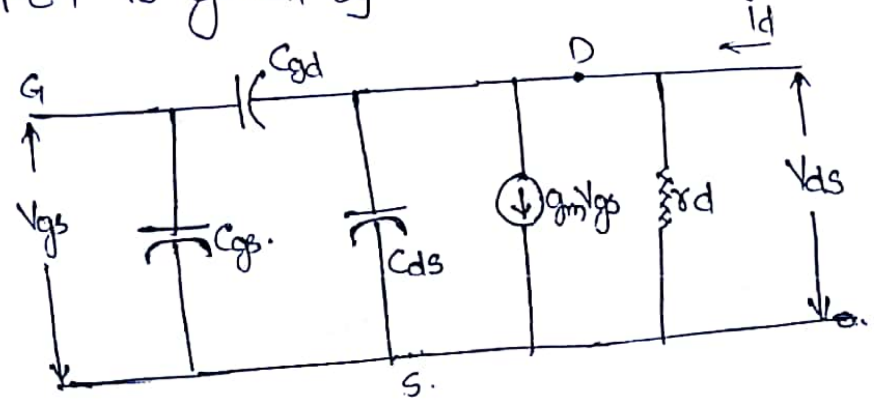
11B. Monday

ET high frequency Model:

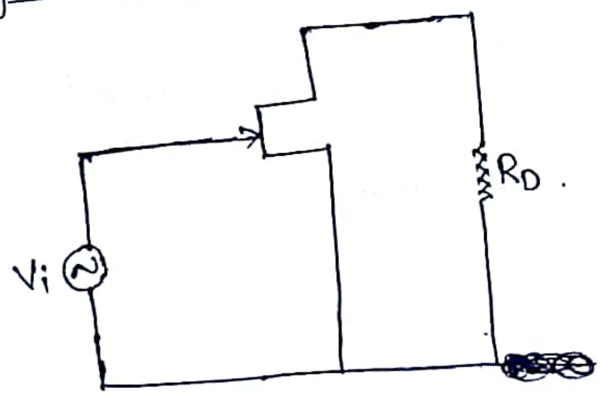
ow frequency model of FET is given by



t high frequencies junction capacitors will begin to dominate ne frequency Response. so the high frequency model of he FET is given by

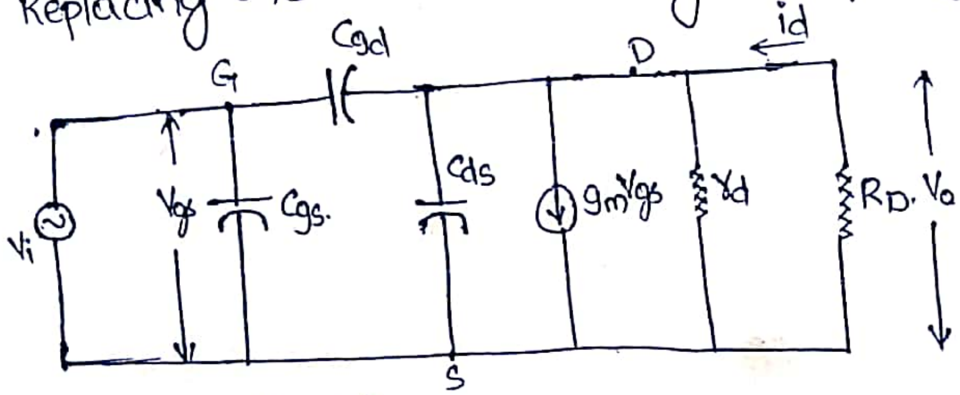


Analysis of common source Amplifier at High frequencies:



Voltage Gain:-

Replacing the FET with its high frequency model.



$V_{gs} = V_i$

Let Y_{ds}; Y_{gd}, Y_{gs} are the admittances of the capacitors.

By writing KCL at the output node

$$i_d = \frac{V_o}{r_d} + g_m V_{gs} + Y_{ds} V_o + (V_o - V_{gs}) Y_{gd}$$

Replace $V_{gs} = V_i$, $i_d = \frac{-V_o}{R_D}$

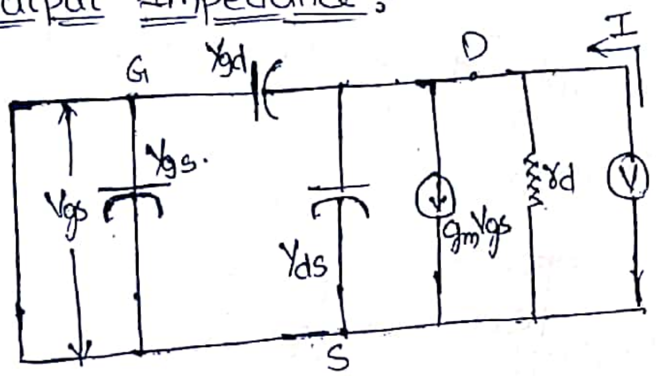
$$\frac{-V_o}{R_D} = \frac{V_o}{r_d} + g_m V_i + Y_{ds} V_o + (V_o - V_i) Y_{gd}$$

$$V_i (Y_{gd} - g_m) = V_o \left(\frac{1}{R_D} + \frac{1}{r_d} + Y_{ds} + Y_{gd}\right)$$

$$\frac{V_o}{V_i} = \frac{(Y_{gd} - g_m)}{\left(\frac{1}{R_D} + \frac{1}{r_d} + Y_{ds} + Y_{gd}\right)}$$

$$\boxed{AV = \frac{V_o}{V_i} = \frac{-g_m + Y_{gd}}{\frac{1}{R_D} + \frac{1}{r_d} + Y_{ds} + Y_{gd}}}$$

Output Impedance :-

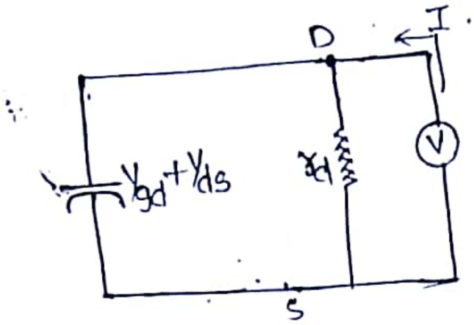


To find the output impedance set $V_i = 0$, disconnect R_D and connect source V

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

$$\Rightarrow Y_o = \frac{I}{V}$$

Rearranging the above circuit and observing that $V_{gs} = 0$, we get the following circuit.



By applying KCL at output node.

$$I = \frac{V}{r_d} + V\left(Y_{ds} + Y_{gd}\right)$$

$$\boxed{\frac{I}{V} = Y_o = \frac{1}{r_d} + \left(Y_{ds} + Y_{gd}\right)}$$

Output Admittance including R_D.

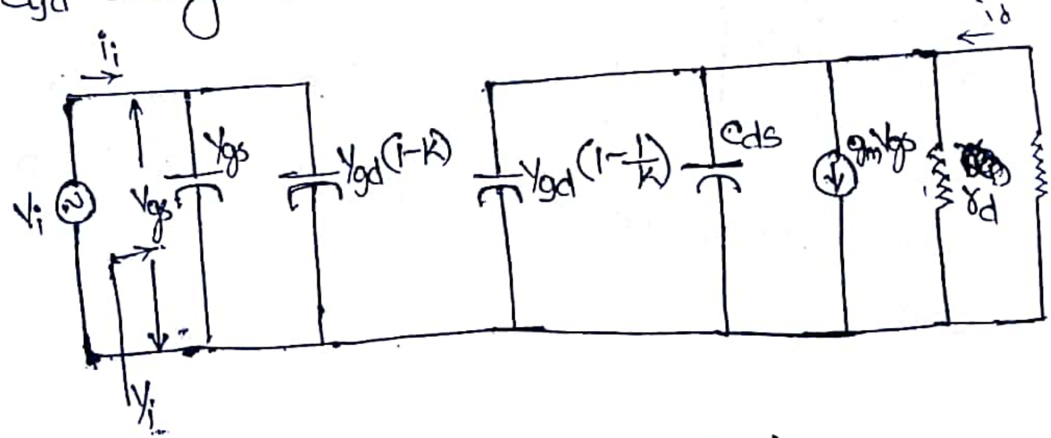
$$Y_o' = Y_o \,||\, 1/R_D$$

$$= Y_o + 1/R_D$$

$$\boxed{Y_o' = \frac{1}{R_D} + \frac{1}{r_d} + Y_{gd} + Y_{ds}}$$

Input Impedance/admittance :-

To find the input impedance split the feed back capacitor C_{gd} using miller's theorem.



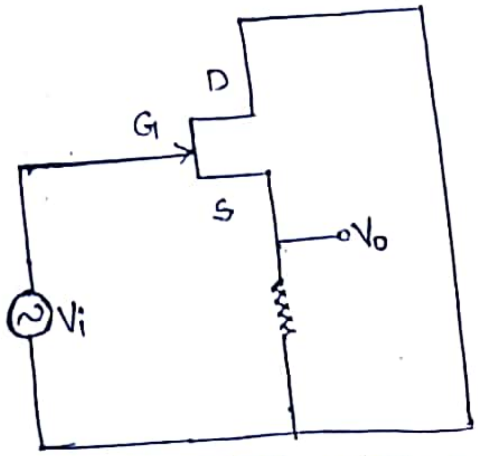
Input admittance $Y_i = Y_{gs} + Y_{gd}(1-K)$ where $K = A_V$.

Net Input capacitance $\boxed{C_i = C_{gs} + C_{gd}(1-K)}$

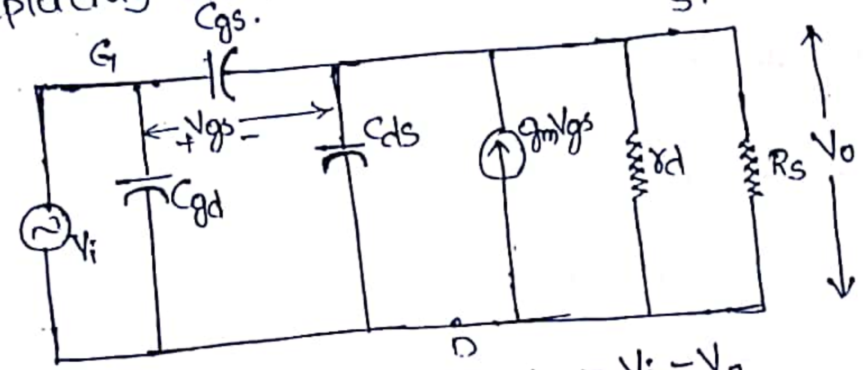
$$A_I = \frac{-i_d}{i_i} = \frac{\frac{V_o}{R_D}}{\frac{V_i}{R_i}} = \frac{V_o}{V_i} \cdot \frac{R_i}{R_D} = A_V \cdot \frac{1}{Y_i} \cdot \frac{1}{R_D}$$

05/12/18
Wednesday

Common drain Amplifier analysis at High frequencies:



Replacing the FET with high frequency model.



In the above circuit $V_{gs} = V_i - V_o$

Voltage Gain:-

Applying KCL at output node.

$$g_m V_{gs} = \frac{V_o}{r_d} + \frac{V_o}{R_S} + V_o Y_{ds} + (V_o - V_i) Y_{gs}.$$

$$g_m (V_i - V_o) = V_o \left[\frac{1}{r_d} + \frac{1}{R_S} + Y_{ds} + Y_{gs}\right] - V_i Y_{gs}.$$

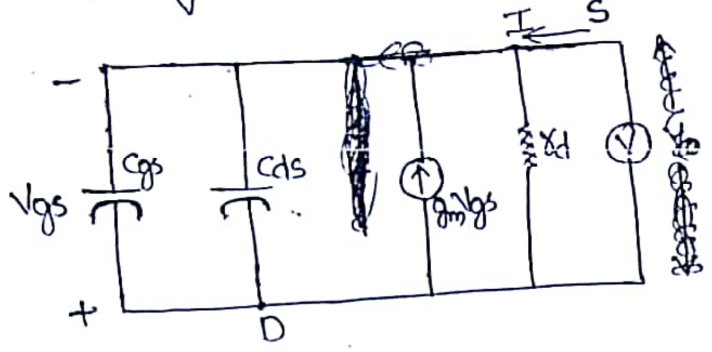
$$V_o \left[\frac{1}{r_d} + \frac{1}{R_S} + Y_{ds} + Y_{gs} + g_m\right] = V_i \left[g_m + Y_{gs}\right]$$

$$\boxed{A_V = \frac{V_o}{V_i} = \frac{g_m + V_{gs}}{\frac{1}{r_d} + \frac{1}{R_S} + Y_{ds} + Y_{gs} + g_m}}$$

Output admittance:-

To find output admittance set $V_i = 0$ and disconnect R_s and connect voltage source V

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



KCL gives

$$I + g_m V_{gs} = \frac{V}{r_d} + V Y_{ds} + V Y_{gs}.$$

In the above circuit $V_{gs} = -V$

$$I - g_m V = \frac{V}{r_d} + V Y_{ds} + V Y_{gs}.$$

$$I = V\left[g_m + \frac{1}{r_d} + Y_{ds} + Y_{gs}\right]$$

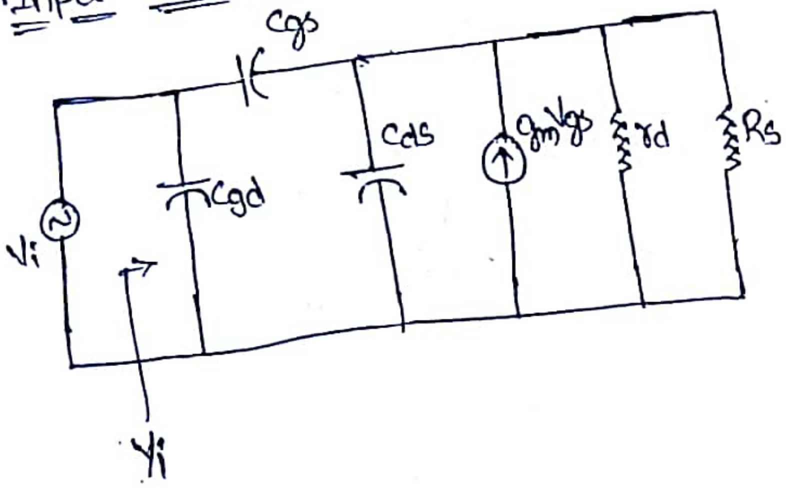
$$\boxed{Y_o = \frac{I}{V} = g_m + \frac{1}{r_d} + Y_{ds} + Y_{gs}.}$$

$$Y_o' = Y_o \parallel \frac{1}{R_s} = Y_o + \frac{1}{R_s}.$$

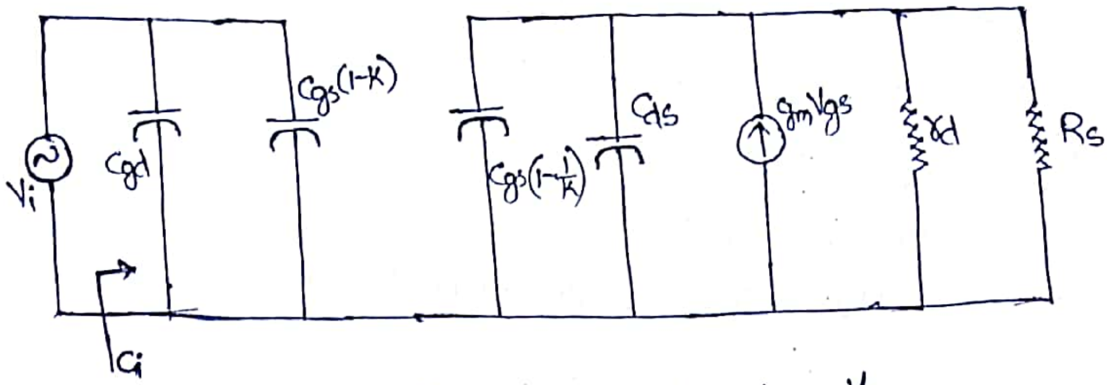
$$\boxed{Y_o' = g_m + \frac{1}{r_d} + Y_{ds} + Y_{gs} + \frac{1}{R_s}.}$$

~~Input admit~~.

Input admittance:-



To find the input admittance the capacitance C_{gs} is split in two parts using miller's theorem as shown in the follow figure.



$$C_i = C_{gd} + C_{gs}(1-K) \quad \text{where } K = A_V = \frac{V_o}{V_i}$$

$$C_i = C_{gd} + C_{gs}(1 - A_V)$$

$$\boxed{Y_i = Y_{gd} + Y_{gs}(1 - A_V)}$$

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Unit-2.

Multi-stage Amplifiers.

Amplifier:-

It is an electronic device which is used to amplify a ~~small~~ weak signal into a sufficiently strong signal.

Multi-stage Amplifiers:-

If the system contains a several amplifiers' cascaded together that system is called multi-stage Amplifier.

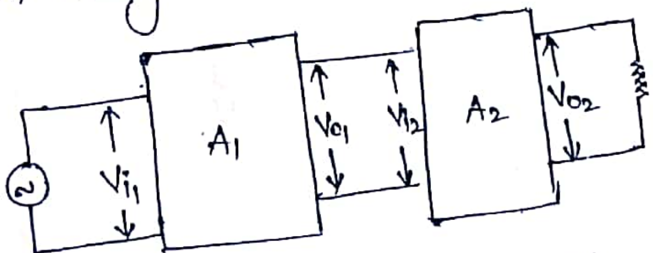
The process of connecting output of one amplifier, to the input of another amplifier is called cascading.

Need for Multi-stage Amplifiers:-

Reason-1

To achieve larger gain:-

If the gain of a single stage is not sufficient when the output of this stage can be connected to another stage and the overall gain will be the product of both stages.



The overall gain $A_I = \frac{V_{o2}}{V_{i_1}} = \frac{V_{o2}}{V_{i_2}} \times \frac{\cancel{V_{i_2}}}{\cancel{V_{o_1}}} \times \frac{V_{o1}}{V_{i_1}}$ [But $V_{o_1} = V_{i_2}$]

$$A_I = \frac{V_{o2}}{V_{i_2}} \times \frac{V_{o1}}{V_{i_1}}$$

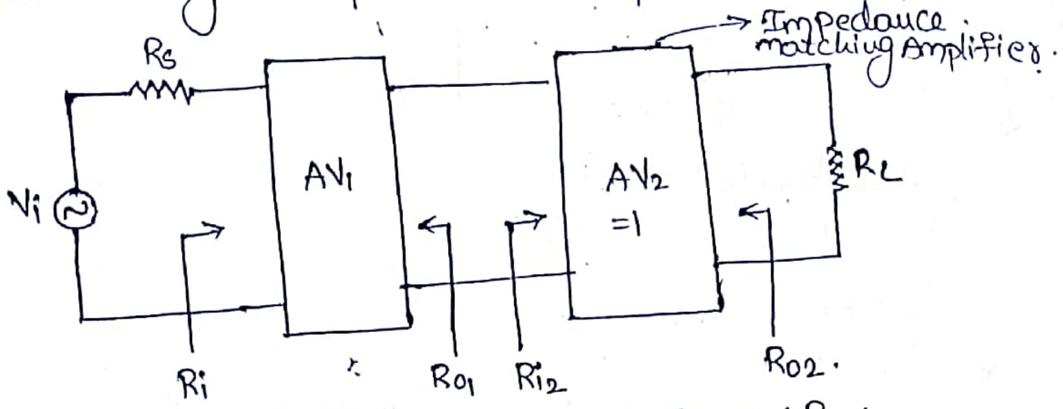
$$\boxed{A_I = A_2 A_1}$$

06/12/18
Thursday.

Reason - 2:-

To provide Impedance Matching:

If the output impedance of an amplifier is not properly match with load impedance R_L, then we need an extra stage amplifier that provides impedance matching.



$R_{o_1} >> R_L$

$R_{i_2} >> R_{o_1}$

$R_{o_2} << R_L$

⇒

Distortion :-

If the output of the amplifier is not an exact replica of the input then we say that the amplifier has distortion.

Types of Distortion

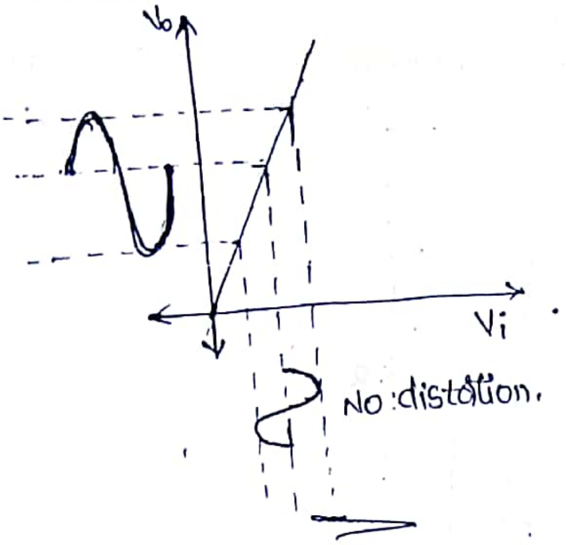
1) Non-linear Distortion (or) Harmonic Distortion (or) Amplitude distortion
2) Frequency Distortion
3) Phase Distortion

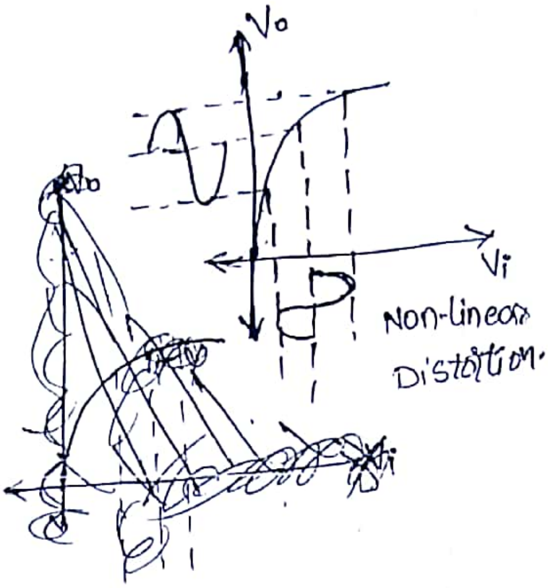
⇒

1) Non-Linear Distortion (or) Harmonic Distortion (or) Amplitude Distortion

If the dynamic transfer curve of the amplifier is not a straight line non-linear distortion occurs.

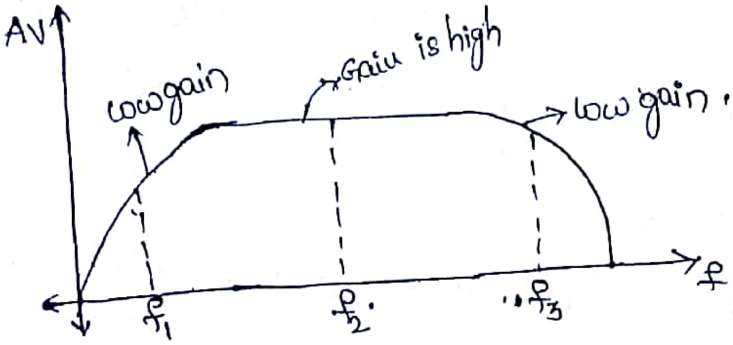
In this type of distortion output contains signals with fundamental frequency and signals which are harmon of the fundamental frequency.





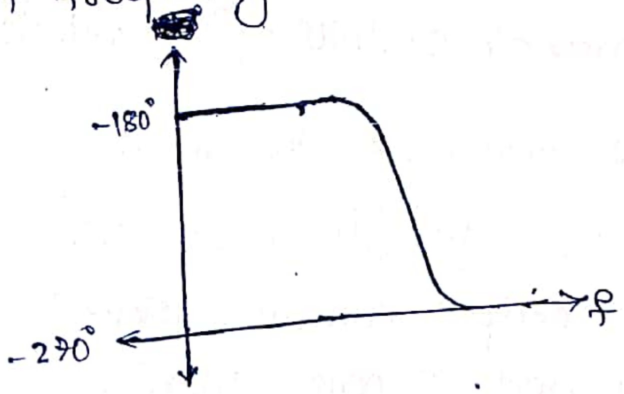
Frequency Distortion:-

If the amplifier provides different gains at different frequencies then we say that the amplifier has frequency distortion.



Phase Distortion:-

If the phase shift provided by the amplifier changes with frequency then their exist phase distortion.



Methods of coupling in Multi stage Amplifiers:

There are 3 types of coupling mechanisms used in multi-stage amplifier, depending upon the type of coupling.

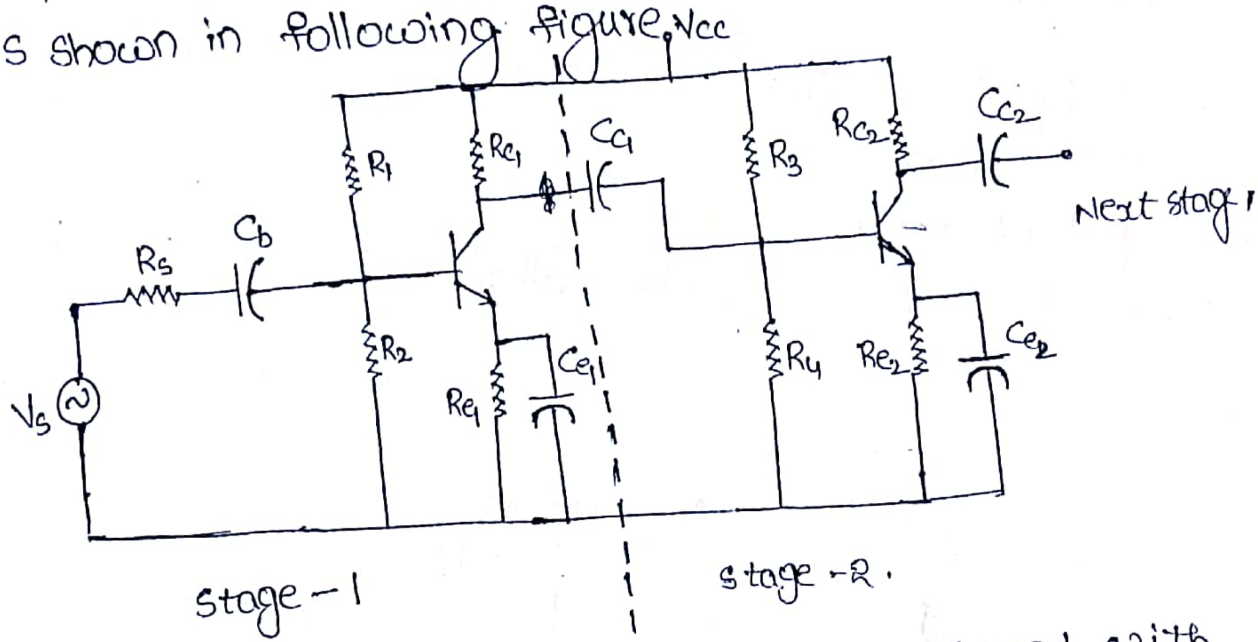
1) RC coupled Amplifiers

2) Transformer coupled Amplifier.

3) Direct coupled Amplifier.

1) RC coupled Amplifier:-

In this type of Amplifier the output of previous stage is coupled to the next stage through coupling capacitor and resistive load. This type of coupling is shown in following figure.



In the above figure capacitor C_{c1} connects stage-1 with stage-2. This capacitor is called coupling capacitor.

At low frequencies impedance of C_c is high

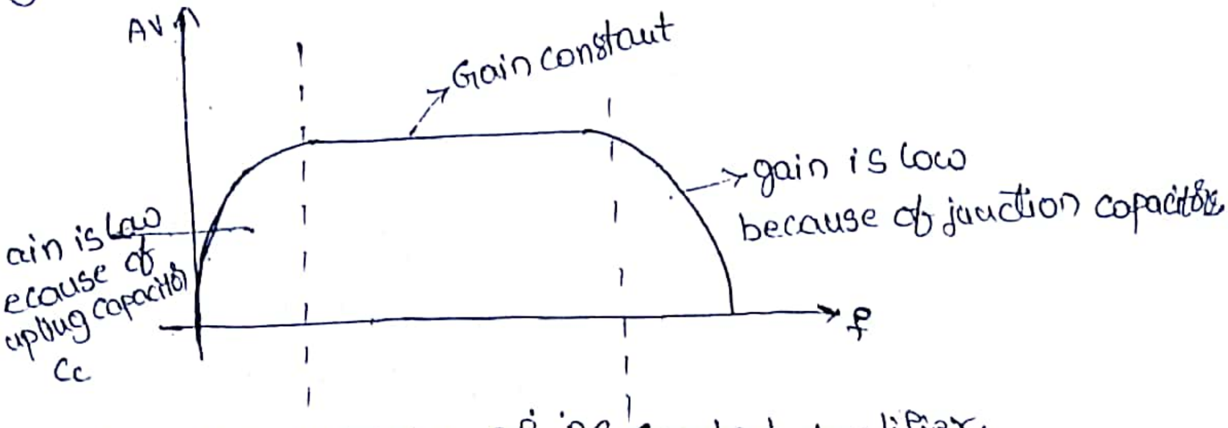
∴ Amount of signal reaching the next stage will be low

As frequency increases impedance of C_c decreases and more signal reaches the next stage, so gain increases.

At mid frequencies, C_c is almost a short circuit.

∴ Gain will be maximum and constant.

high frequencies junction capacitors will appear parallel to the signal path. These capacitors will pass the AC signal and hence gain decreases at high frequencies.

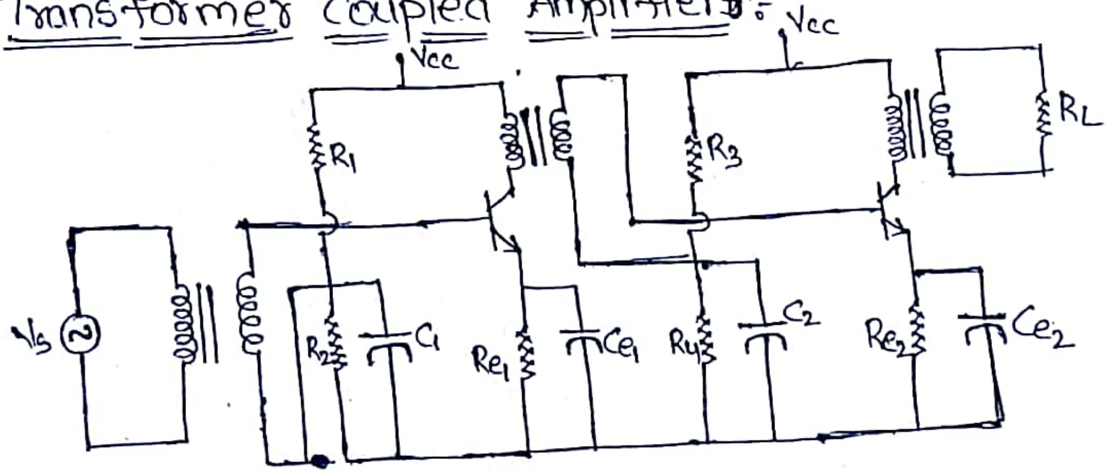


Frequency response of RC coupled Amplifier.

For DC signals coupling capacitor C_c is a open circuit. The DC signal of one stage will not interact with the DC signal of next stage.
Hence this capacitor is also called as Blocking capacitor.

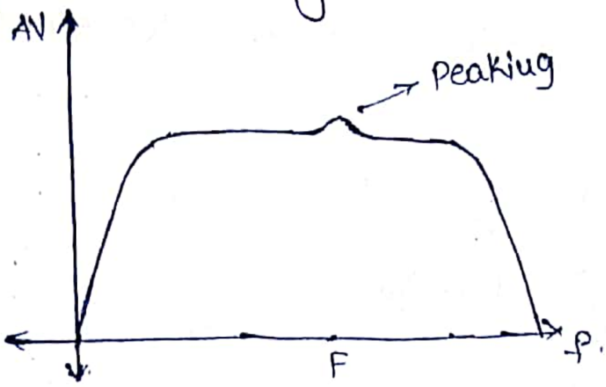
17/12/18
Friday.

Transformer Coupled Amplifiers:-



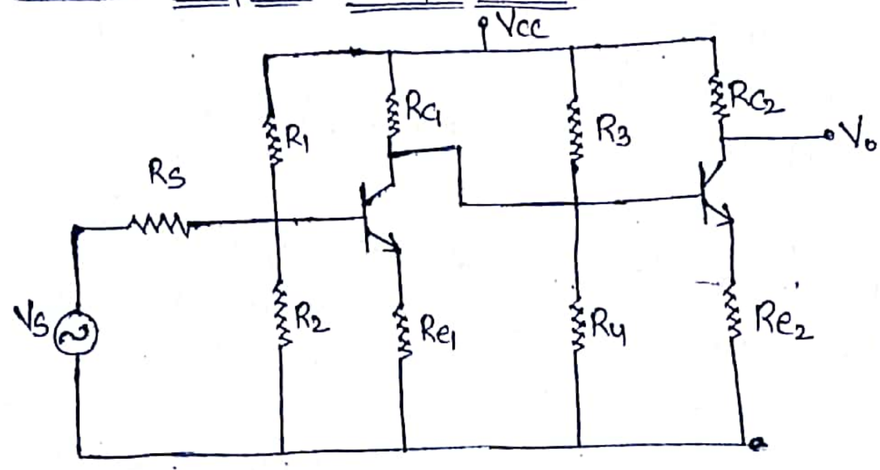
In the above figure output of one stage is connected to the input of another stage through a transformer since transformer couples only AC signals and blocks DC signals.

∴ DC signals of Individual sources are isolated.
The frequency response of transformer coupled amp may have peaking

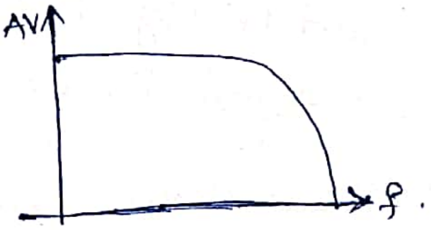


⟹

Direct coupled Amplifiers :-



The above figure shows a direct coupled amplifier. Because of direct coupling the low frequency response will be very good.
At high frequencies the gain will be falls due to junction capacitors.
The DC signals of one stage interact with DC signals of next stages.
∴ care must be taken in the design.
The Frequency response of a direct coupled Amplifier is.



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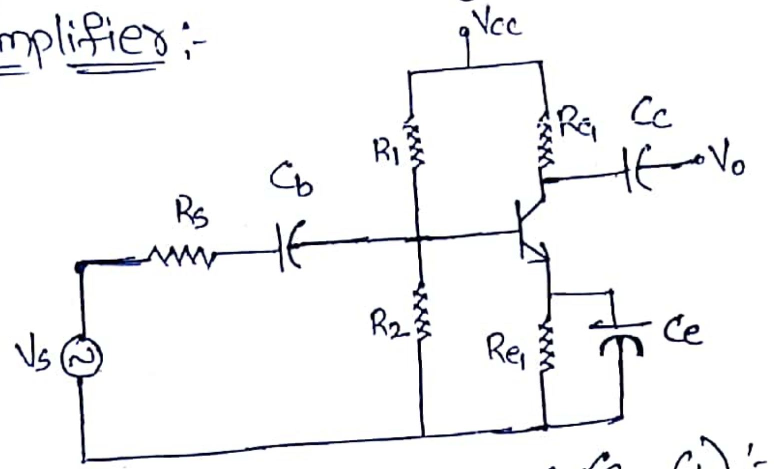
Effect of coupling, Bypass, Junction capacitors on the frequency response of RC coupled Amplifier :-

(or)

Explain the frequency response curve of RC coupled Amplifier :-



Coupling & Blocking capacitor (C_c, C_b) :-

* These capacitors will block the DC signals and allow only AC signal.
* At low frequencies the impedance of the capacitor is high. It allows less amount of AC signal to pass through and hence gain will be low.
* At mid frequencies the impedance of C_c, C_b will be very low and hence they act as short circuit and gain will be high.

Bypass capacitor (C_e) :-

If bypass capacitor Ce is not present, both AC & DC signals will pass through Re. This will provide negative feed back for AC signal. and gain decreases.

To avoid this by pass capacitor Ce must be connected across Re

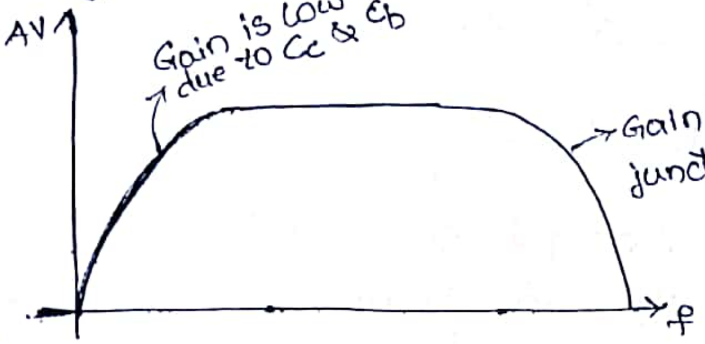
Now AC signals with Ce and DC signals will pass through Re. and hence AC signal feed back is avoided.

Junction capacitor:-

These capacitors are also called as parasitic capacitors. At low & mid frequencies they act as open circuits because they are very very small capacitors. Hence there is no impact on frequency response at low & mid frequencies.

At high frequency the impedance of these capacitors begins to fall.

∴ The gain also begins to fall.

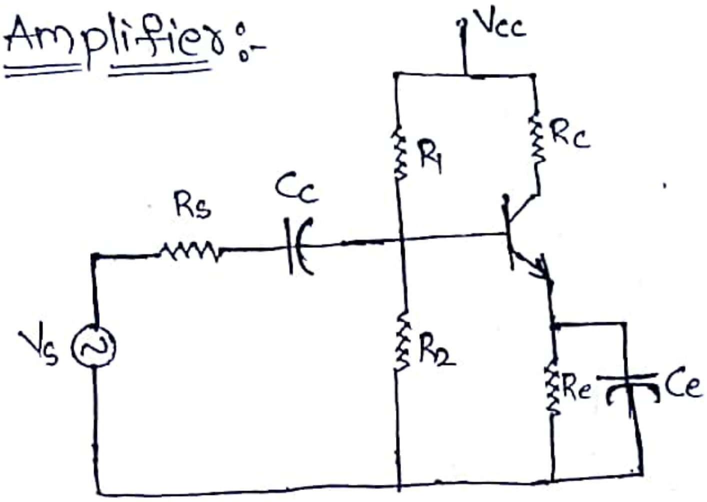


08/12/18
Saturday.

Effect of coupling capacitor on the low frequency Response:

(or)

Lower cut off frequency due to Cc in a RC coupled Amplifier:-

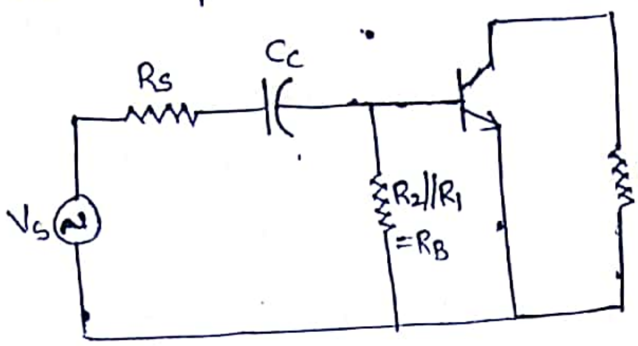


The above figure shows RC coupled amplifier.

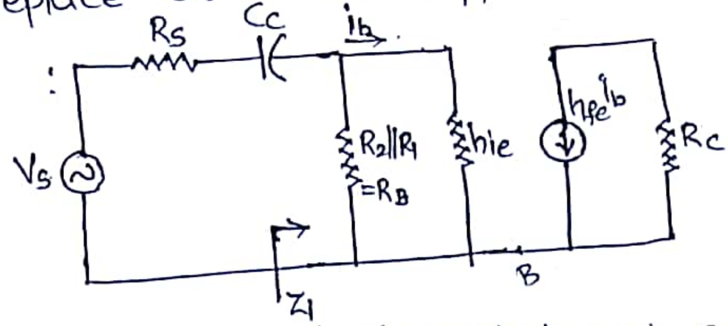
Let $C_e \gg C_c$

∴ The impact of C_e can be neglected and it can be replaced with a short circuit.

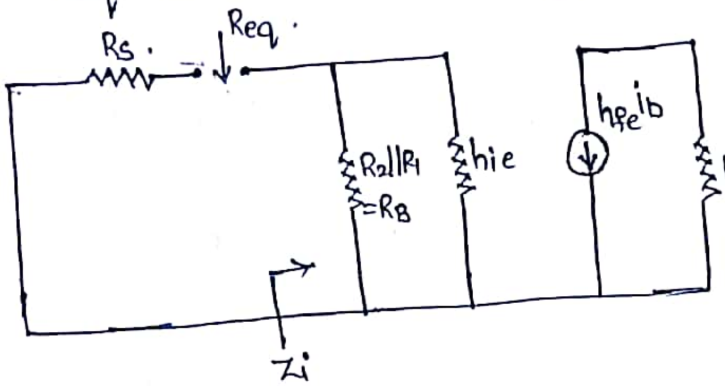
The AC equivalent circuit is given by

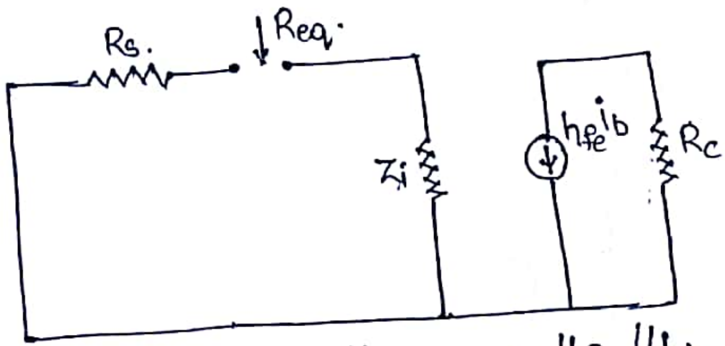


Replace BJT with approximate h-parameter model.



To find equivalent resistance seen by C_c. set $V_s=0$ and find R_{eq} from the following circuit.





where $Z_i = R_B || h_{ie} = R_1 || R_2 || h_{ie}$

$$R_{eq} = Z_i + R_s$$

The cut off frequency due to C_c is given by

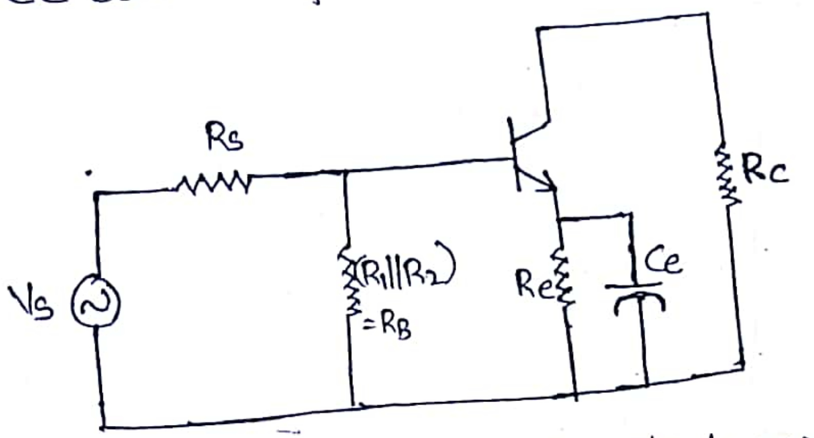
$$\therefore f_L = \frac{1}{2\pi R_{eq} C_c}$$

$$\therefore \boxed{f_L = \frac{1}{2\pi(R_s+Z_i)C_c}}$$

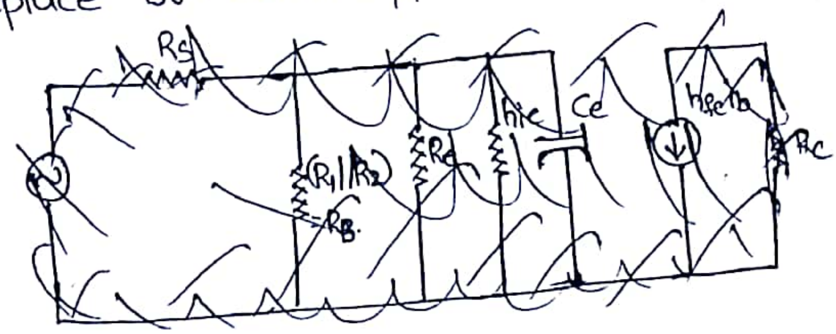
Impact of Bypass capacitor (C_e) on the low frequency Response (or) Lower cut off frequency of the RC coupled Amplifier due to bypass capacitor (C_e):

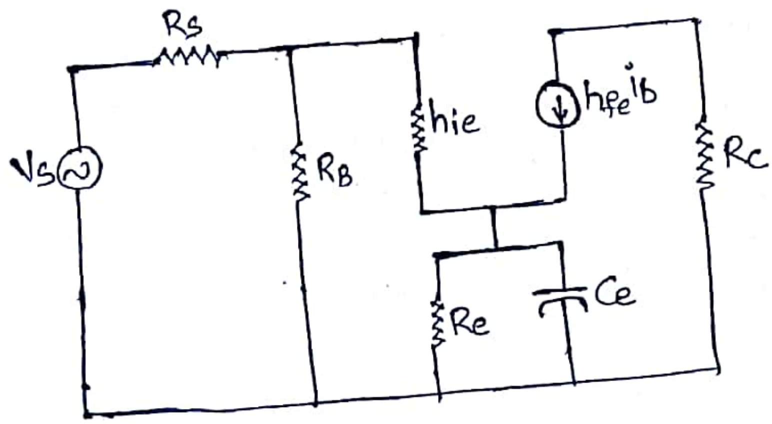
Assume $C_c >> C_e$

$\therefore$ The lower cut off frequency is decided by C_e and C_c can be replaced with a short circuit.



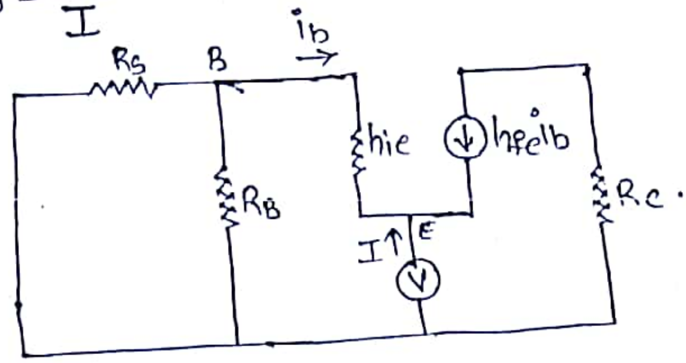
Replace BJT with approximate h-parameter model.



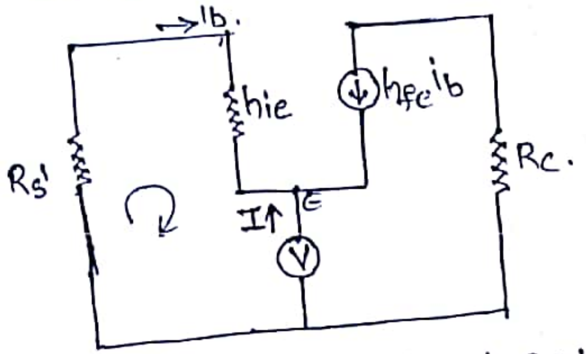


To find the cut off frequency due to C_e. we must find the equivalent resistance seen by C_e.

set $V_s = 0$, Remove R_e, C_e.
assume independent voltage source V. let I be the current.
let $R_o = \frac{V}{I}$.



Let $R_S || R_B = R_S'$



By Applying ~~KVL~~ KCL at emitter node.

$$i_b + I + h_{fe} i_b = 0.$$

$$i_b(1+h_{fe}) = -I$$

$$I = -(1+h_{fe}) i_b.$$

By applying KVL to the input loop

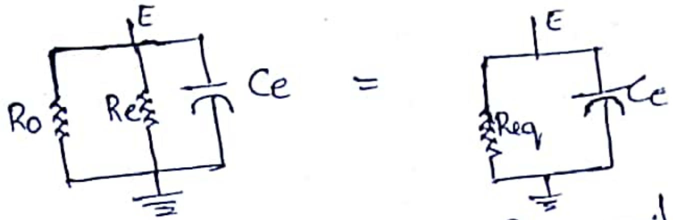
$$R_s' i_b + h_{ie} i_b + V = 0.$$

$$V = -(R_s' + h_{ie}) i_b$$

$$R_o = \frac{V}{I} = \frac{\cancel{-}(R_s' + h_{ie}) \cancel{i_b}}{\cancel{-}(1+h_{fe}) \cancel{i_b}}$$

$$\therefore R_o = \frac{V}{I} = \left(\frac{R_s' + h_{ie}}{1+h_{fe}}\right)$$

$\therefore$ The effective circuit between the emitter and ground is.



∴ cut off frequency $f_L = \dfrac{1}{2\pi R_{eq} C_e}$

$$\boxed{\therefore\ f_L = \frac{1}{2\pi\left[\frac{(R_s'+h_{ie})}{(1+h_{fe})} \,\|\, R_e\right] C_e}}$$

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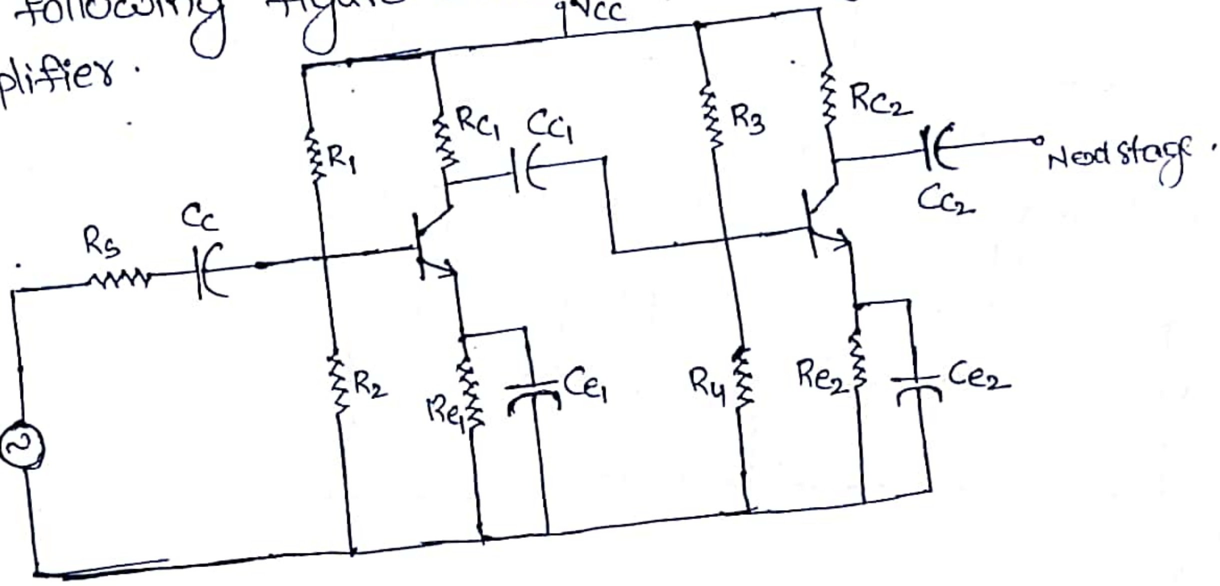
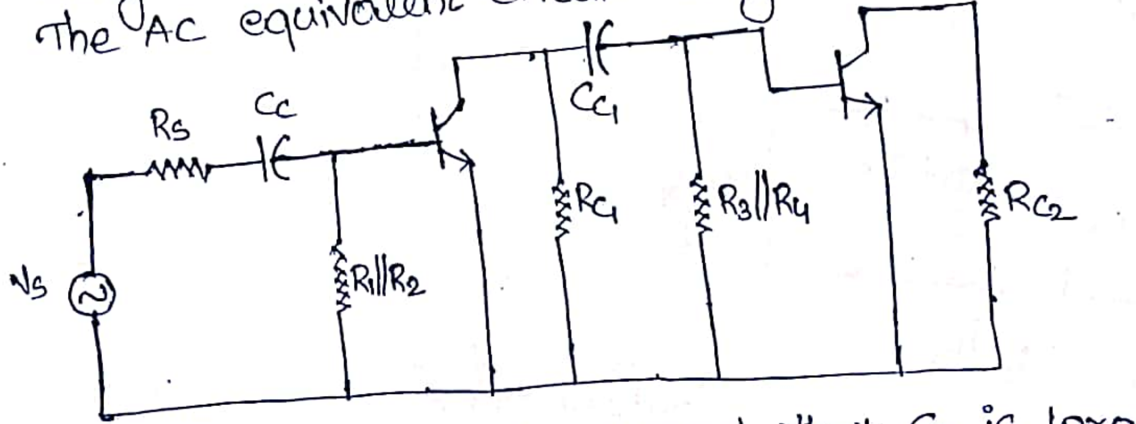
Wednesday.

Analysis of RC coupled Amplifier at low, mid, high frequency Range:-

The following figure shows a two stage RC coupled Amplifier.



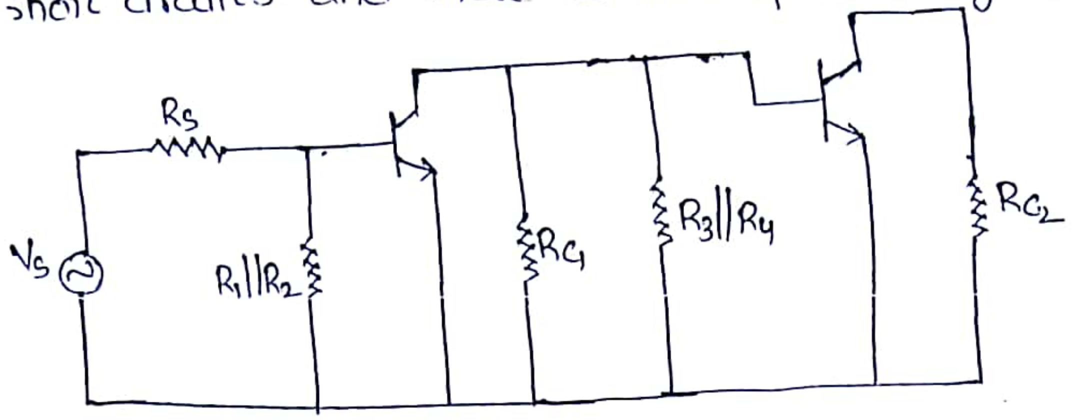
~~Analysis~~ The AC equivalent circuit is given by



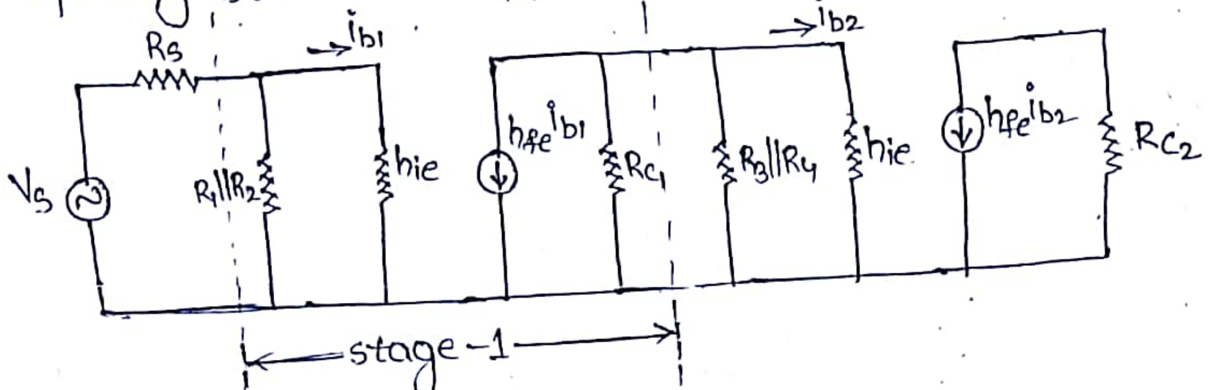
In the above figure, we assumed that Ce is large.

Analysis in the mid frequency Range (Mid frequency Gain):-

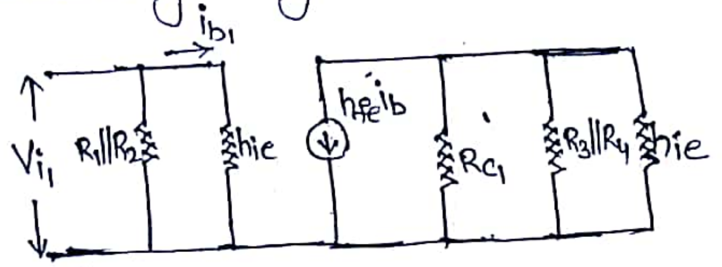
At mid frequencies coupling capacitors can be replaced with short circuits and there is no impact of junction capacitors



Replacing BJT with approximate H-parameter model.

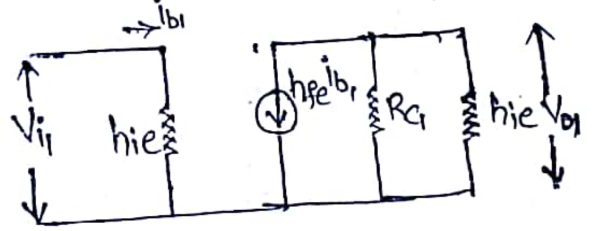


Isolating stage-1, to calculate the gain.



Normally
$R_1, R_2 >> h_{ie}$.
$R_3, R_4 >> h_{ie}$

$\therefore R_1 || R_2 || h_{ie} \simeq h_{ie}$, $R_3 || R_4 || h_{ie} \simeq h_{ie}$



$$V_{i_1} = h_{ie}\, i_{b_1}$$

By writing the KCL at output node.

$$h_{fe} i_{b_1} + \frac{V_{01}}{R_{C1}} + \frac{V_{01}}{h_{ie}} = 0$$

$$V_{o1}\left[\frac{1}{R_{C1}}+\frac{1}{h_{ie}}\right] = -h_{fe}\, i_{b1}$$

$$V_{o1} = \frac{-h_{fe}\, i_{b1}}{\frac{1}{R_{C1}}+\frac{1}{h_{ie}}} = \frac{-h_{fe}\, i_{b1}\, R_{C1}\, h_{ie}}{R_{C1}+h_{ie}}$$

$\therefore$ Mid frequency Gain

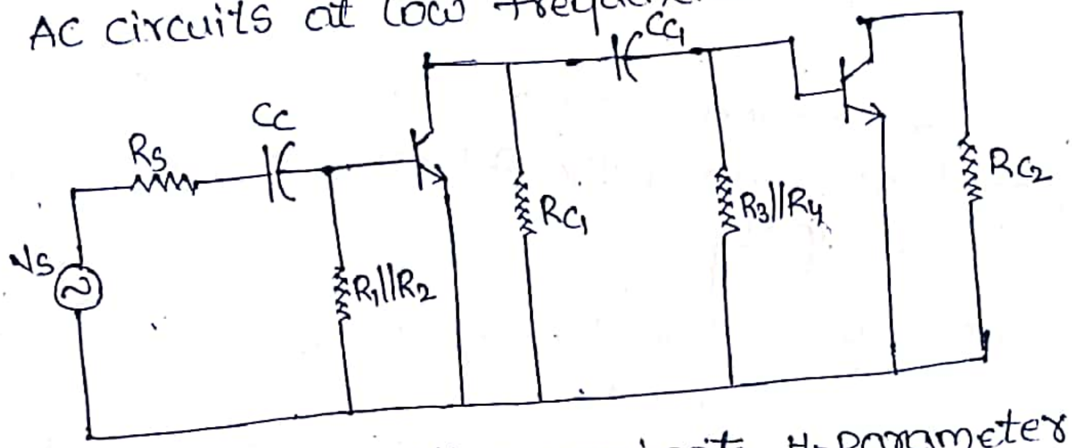
$$A_m = \frac{V_{o1}}{V_{i1}} = \frac{\frac{-h_{fe}\, i_{b1} R_{C1}\, h_{ie}}{R_{C1}+h_{ie}}}{h_{ie}\, i_{b1}}$$

$$\boxed{A_m = \frac{-h_{fe} R_{C1}}{R_{C1}+h_{ie}}}$$

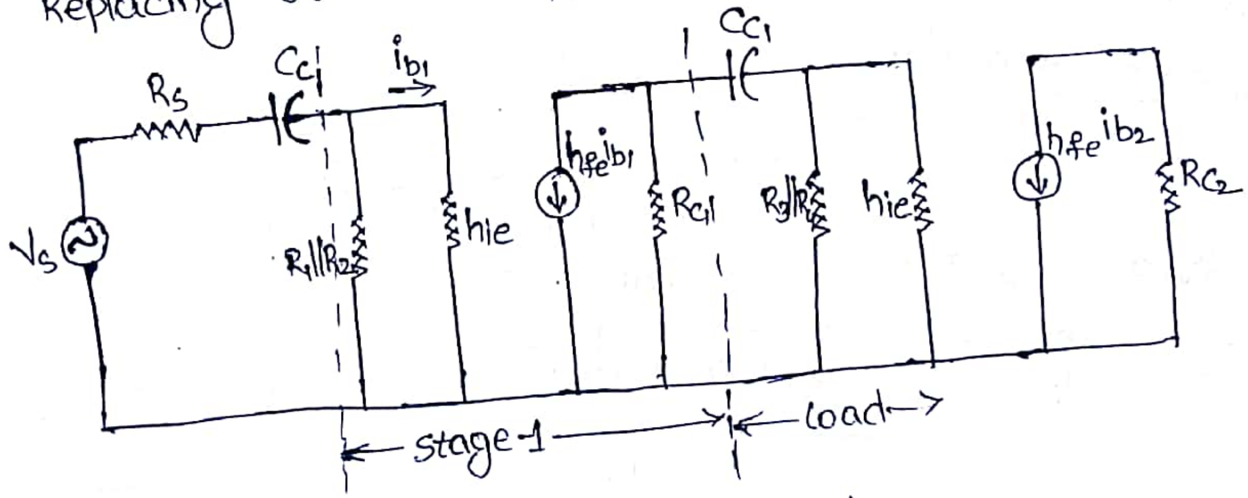
Analysis in the low frequency Range (low frequency gain):-

Assume $C_e >> C_c$. so that the low frequency response is dominated by C_c

AC circuits at low frequencies is



Replacing BJT with approximate H-parameter model.



Isolating stage-1 ~~load~~ including load.

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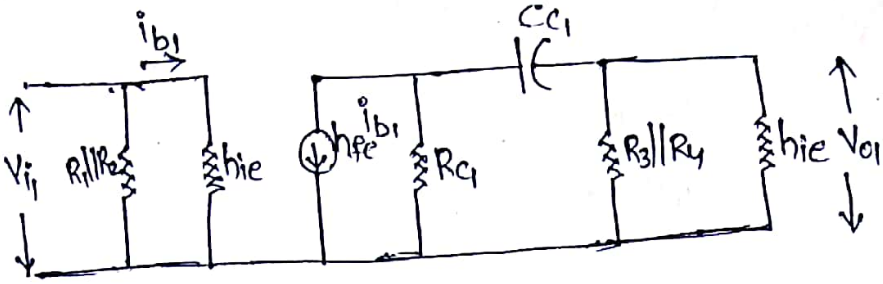
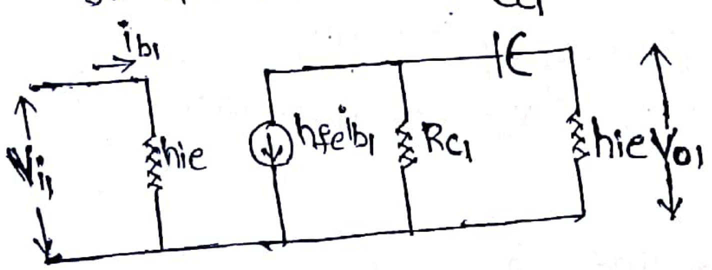


Assume hie is very small.

$R_1, R_2, R_3, R_4 >> h_{ie}$

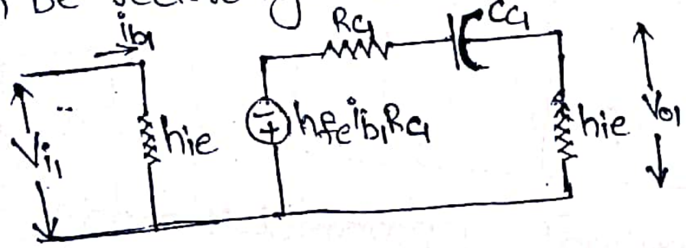
$R_1 || R_2 || h_{ie} \simeq h_{ie}$

$R_3 || R_4 || h_{ie} \simeq h_{ie}$.



$V_{i_1} = h_{ie}\, i_{b1}$

Using source transformation technique the above circuit can be rearranged as shown in the following figure.



From the above circuit

$$V_{01} = \frac{-h_{fe}\, i_{b1} R_{C1} \times h_{ie}}{h_{ie} + R_{C1} + \frac{1}{j2\pi f C_{C1}}}$$

Low frequency Gain.

$$A_L = \frac{V_{01}}{V_{i_1}} = \frac{\dfrac{-h_{fe}\, i_{b1} R_{C1} \cdot h_{ie}}{h_{ie} + R_{C1} + \frac{1}{j2\pi f C_{C1}}}}{h_{ie}\, i_{b1}}$$

$$A_L = \frac{-h_{fe} R_{C1}}{R_{C1} + h_{ie} + \frac{1}{j2\pi f C_{C1}}}$$

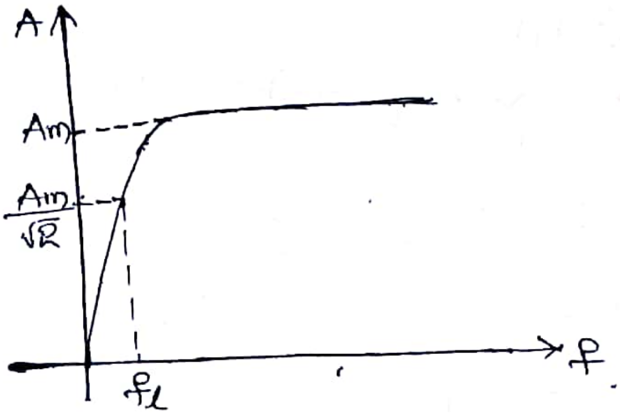
$$A_l = \frac{\dfrac{-h_{fe} R_{C_1}}{R_{C_1} + h_{ie}}}{1 - j\dfrac{1}{2\pi f C_{C_1}(R_{C_1} + h_{ie})}}$$

$$A_l = \frac{A_m}{1 - j\dfrac{f_l}{f}} \Rightarrow |A_l| = \frac{A_m}{\sqrt{1 + \left(\dfrac{f_l}{f}\right)^2}}$$

at $f = f_l$; $|A_l| = \dfrac{A_m}{\sqrt{2}}$

$\therefore f_l$ is lower cut off frequency.

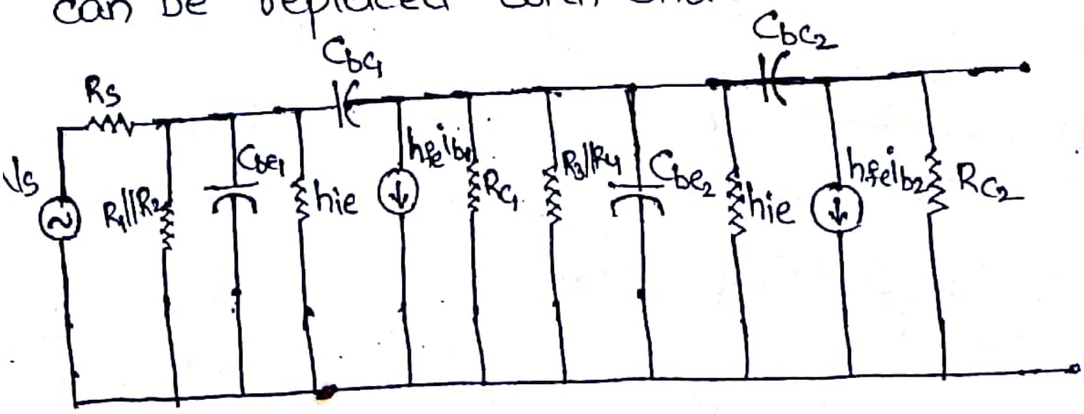
$$f_l = \frac{1}{2\pi C_{C_1}(R_{C_1} + h_{ie})}$$



13/12/18
Thursday

Analysis in the high Frequency Range:

At high frequencies junction capacitors dominate the frequency Response. coupling capacitor and bypass capacitor can be replaced with short circuit.



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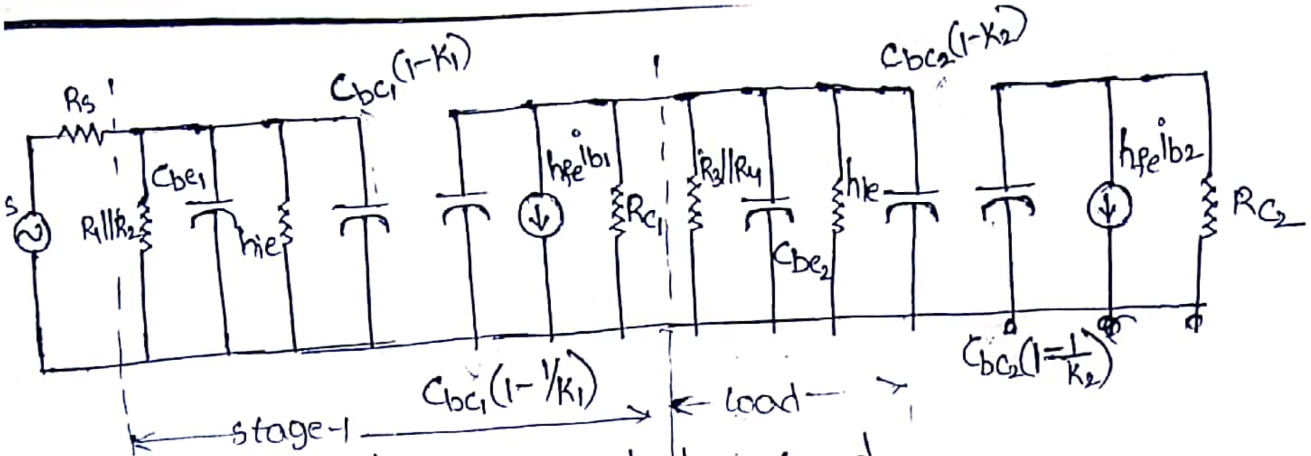
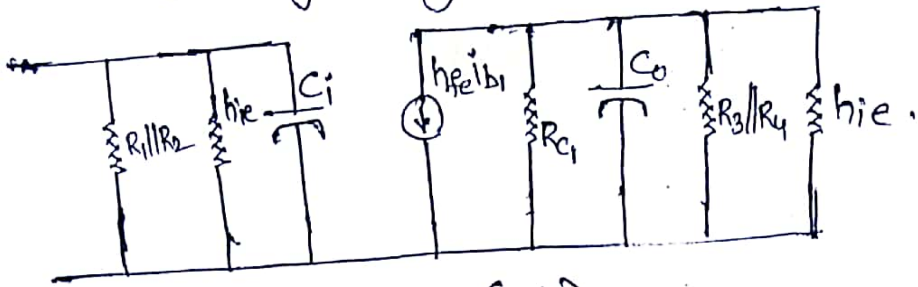


separating single stage including load

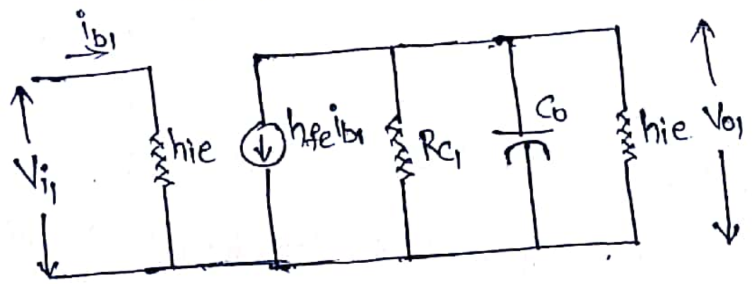


where $C_i = C_{be_1} + C_{bc_1}(1-K_1)$

$$C_o = C_{be_2} + C_{bc_1}(1-1/K_1) + C_{bc_2}(1-K_2)$$

Assume $R_1 \| R_2 \| h_{ie} \| X_{ci} \simeq h_{ie}$

$$R_3 \| R_4 \| h_{ie} \simeq h_{ie}$$



$$V_{i_1} = h_{ie}\, i_{b_1}$$

By writing KCL at the output loop.

$$h_{fe}\, i_{b_1} + \frac{V_{o1}}{R_{c_1}} + \frac{V_{o1}}{\frac{1}{j2\pi f C_o}} + \frac{V_{o1}}{h_{ie}} = 0.$$

$$V_{o1}\left[\frac{1}{R_{c_1}} + \frac{1}{h_{ie}} + j2\pi f C_o\right] = -h_{fe}\, i_{b_1}$$

$$V_{o1} = \frac{-h_{fe}\, i_{b_1}}{\frac{1}{R_{c_1}} + \frac{1}{h_{ie}} + j2\pi f C_o}$$

$$V_{o1} = \frac{-h_{fe} i_{b1} R_{C1} \cdot h_{ie}}{R_{C1} + h_{ie} + j2\pi f C_o R_{C1} h_{ie}}$$

$$A_h = \frac{V_{o1}}{V_{i1}} = \frac{\dfrac{-h_{fe} i_{b1} \cdot R_{C1} \cdot h_{ie}}{R_{C1} + h_{ie} + j2\pi f C_o R_{C1} h_{ie}}}{h_{ie}\, i_{b1}}$$

$$= \frac{-h_{fe} R_{C1}}{R_{C1} + h_{ie} + j2\pi f C_o R_{C1} h_{ie}}$$

$$= \frac{\dfrac{-h_{fe} \cdot R_{C1}}{R_{C1} + h_{ie}}}{1 + j\dfrac{2\pi f C_o R_{C1} h_{ie}}{R_{C1} + h_{ie}}}$$

$$A_h = \frac{A_m}{1 + j(f/f_h)}$$

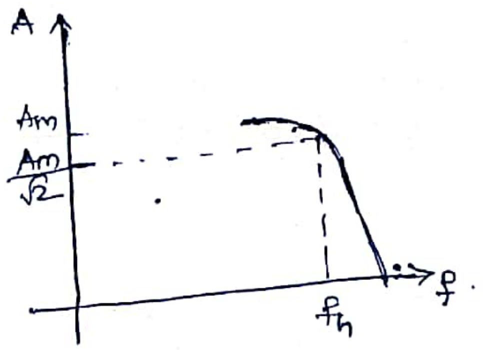
$$\therefore f_h = \frac{R_{C1} + h_{ie}}{2\pi C_o R_{C1} h_{ie}} = \frac{1}{2\pi C_o \left(\dfrac{R_{C1} h_{ie}}{R_{C1} + h_{ie}}\right)}$$

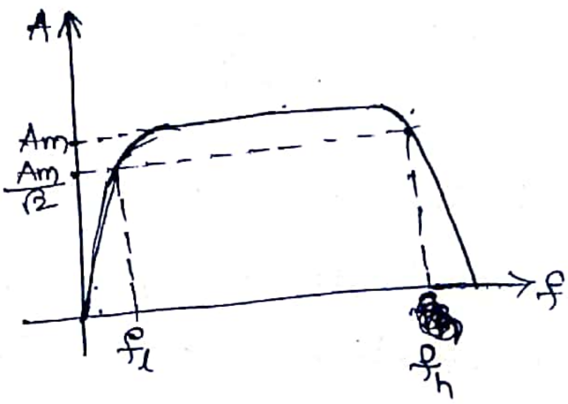
$$f_h = \frac{1}{2\pi C_o (R_{C1} \| h_{ie})}$$

$$\therefore |A_h| = \frac{|A_m|}{\sqrt{1 + (f/f_h)^2}};$$

At $f = f_h$, $|A_h| = \dfrac{A_m}{\sqrt{2}}$

$\therefore f_h$ is upper cut off frequency.





1. Derive the relationship between mid frequency gain and low frequency gain for an RC-coupled Amplifier.

1. Derive A_m and A_L.

2. Derive the relationship between mid frequency gain and high frequency gain for an RC-coupled Amplifier.

A. Derive A_m & A_h

2. Impact of coupling capacitor on gain

A. Derive A_L

<u>Effect of cascading on gain and bandwidth</u>:-

<u>Effect on gain</u>:-

Cascading increases gain, the overall gain of a cascaded amplifier is the product of individual gains.

<u>Effect on Bandwidth</u>:-

Cascading decreases the band width.

<u>Effect of cascading on lower cut off frequency (or) Lower cut off frequency of an n-stage cascaded Amplifier.</u>

Let A_L, A_m, A_h be the low, mid, high frequency gains of an individual stage.

$$A_L = \frac{A_m}{\sqrt{1+(f_L/f)^2}}$$

$$A_h = \frac{A_m}{\sqrt{1+(f/f_h)^2}}$$

f_L, f_h are lower and upper cut off frequencies of individual stage amplifier.

n-similar stages are cascaded

Let A_L, A_M, A_H are the low, mid, high frequency gains of n-stage amplifier.

Let f_L, f_H are the lower & upper cut off frequencies of n-stage amplifier.

$$A_L = \frac{A_M}{\sqrt{1+(f_L/f)^2}}$$

$$A_H = \frac{A_M}{\sqrt{1+(f/f_H)^2}}$$

$$A_L = (A_l)^n, \quad A_M = (A_m)^n, \quad A_H = (A_h)^n$$

$$A_L = (A_l)^n$$

$$\frac{A_M}{\sqrt{1+\left(\frac{f_L}{f}\right)^2}} = \left[\frac{A_m}{\sqrt{1+\left(\frac{f_l}{f}\right)^2}}\right]^n$$

$$\frac{A_M}{\sqrt{1+\left(\frac{f_L}{f}\right)^2}} = \frac{A_m^n}{\left[\sqrt{1+\left(\frac{f_L}{f}\right)^2}\right]^n}$$

$$\frac{\cancel{A_M}}{\sqrt{1+\left(\frac{f_L}{f}\right)^2}} = \frac{\cancel{A_M}}{\left(\sqrt{1+\left(\frac{f_l}{f}\right)^2}\right)^n}$$

$$\sqrt{1+\left(\frac{f_L}{f}\right)^2} = \left[\sqrt{1+\left(\frac{f_l}{f}\right)^2}\right]^n$$

$$1+\left(\frac{f_L}{f}\right)^2 = \left[1+\left(\frac{f_l}{f}\right)^2\right]^n$$

Put $f = f_L$ in above equation

$$2 = \left[1+\left(\frac{f_l}{f_L}\right)^2\right]^n$$

$$2^{1/n} = 1+\left(\frac{f_l}{f_L}\right)^2$$

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$$\left(\frac{f_l}{f_L}\right)^2 = 2^{1/n} - 1$$

$$\boxed{f_L = \frac{f_l}{\sqrt{2^{1/n} - 1}}}$$

f_L is always more than f_l

<u>Effect of cascading on upper cut off frequency (or) upper cut off frequency of n-stage cascaded Amplifier</u> :-

$$A_H = (A_h)^n.$$

$$\frac{A_M}{\sqrt{1+\left(\frac{f}{f_H}\right)^2}} = \left[\frac{A_m}{\sqrt{1+\left(\frac{f}{f_h}\right)^2}}\right]^n.$$

$$\frac{A_M}{\sqrt{1+\left(\frac{f}{f_H}\right)^2}} = \frac{A_m^n}{\left(\sqrt{1+\left(\frac{f}{f_h}\right)^2}\right)^n}$$

$$\frac{\cancel{A_M}}{\sqrt{1+\left(\frac{f}{f_H}\right)^2}} = \frac{\cancel{A_M}}{\left(\sqrt{1+\left(\frac{f}{f_h}\right)^2}\right)^n}$$

$$\sqrt{1+\left(\frac{f}{f_H}\right)^2} = \left[\sqrt{1+\left(\frac{f}{f_h}\right)^2}\right]^n$$

$$1+\left(\frac{f}{f_H}\right)^2 = \left[1+\left(\frac{f}{f_h}\right)^2\right]^n.$$

Put $f = f_H$ in the above equation

$$2 = \left[1+\left(\frac{f_H}{f_h}\right)^2\right]^n$$

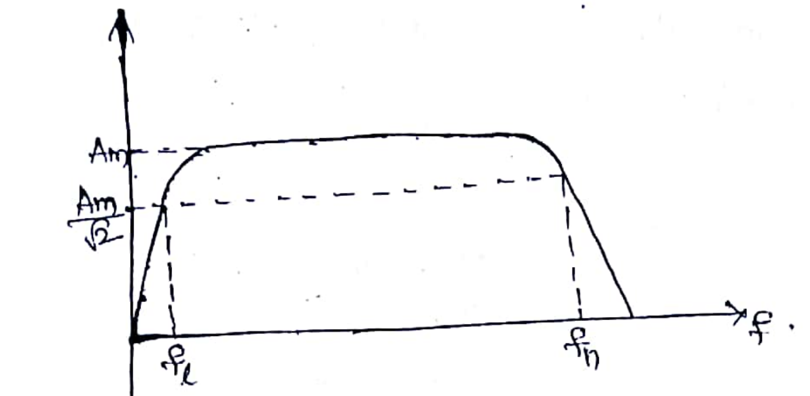
$$2^{1/n} = 1 + \left(\frac{f_H}{f_h}\right)^2$$

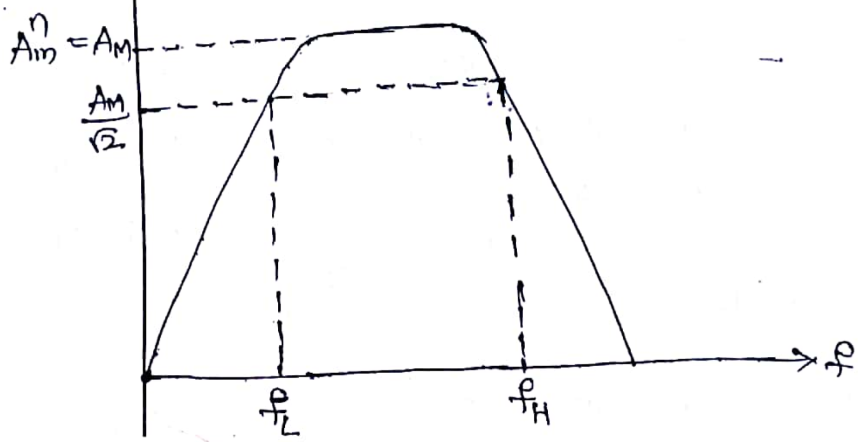
$$\left(\frac{f_H}{f_h}\right)^2 = 2^{1/n} - 1$$

$$\frac{f_H}{f_h} = \sqrt{2^{1/n}-1}$$

$$\boxed{f_H = f_h\left(\sqrt{2^{1/n}-1}\right)}$$

∴ $f_H < f_h$. (f_H is always less than f_h)





From the diagram we can observe that cascading decrease the band width.

Bandwidth of individual stage is

$BW = f_h - f_l$, normally $f_l << f_h$

$BW = f_h$

Bandwidth of n-stage amplifier

$(BW)_n = f_H - f_L$, normally $f_L << f_H$

$(BW)_n \simeq f_H$

$$(BW)_n = f_H = f_h(\sqrt{2^{1/n}-1})$$

$$\boxed{(BW)_n = (BW)(\sqrt{2^{1/n}-1})}.$$

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riday.

The band width of a 3-stage identically cascaded Amplifier extends over 20 Hz to 20 KHz determine bandwidth of individual stages and also determine the lower and upper cut off frequencies of individual stages.

· Given

$n = 3$.

$f_L = 20\,Hz$

$f_H = 20\,KHz$.

$$f_L = \frac{f_l}{\sqrt{2^{1/n}-1}}$$

$$\Rightarrow f_l = f_L(\sqrt{2^{1/n}-1})$$

$$= 20(\sqrt{2^{1/3}-1})$$

$$= 10.196\,HZ \simeq 10.2\,Hz$$

$$f_H = f_h(\sqrt{2^{1/n}-1})$$

$$f_h = \frac{f_H}{\sqrt{2^{1/n}-1}} = \frac{20\times10^3}{\sqrt{2^{1/3}-1}} = 39.2\,KHZ.$$

Band width of individual stage $= f_h - f_l$

$= 39.2K - 10.2\,HZ$

$= 39.18\,KHZ$.

2. Band width of Single stage Amplifier is 50 KHz, 4 such stage are cascaded. Determine the band width of cascaded Ampli

A. Given

$BW = 50 KHz$

$n = 4$

$(BW)_n = ?$

we know that

$(BW)_n \simeq (BW)\sqrt{2^{1/n} - 1}$

$= 50 \times 10^3 \sqrt{2^{1/4} - 1}$

$= 21.74$ K.HZ.

3. An amplifier has a gain of 20db. and lower and upper cut off frequencies of 500HZ, 50K. How many amplifiers are needed in cascade to achieve a gain of 60db. What are the lower and upper cut off frequencies. of cascaded amplifier.

A. Let A_1 db be the gain of single stage.

Let Adb be the gain of cascaded stages.

The number of stages is given by.

$$n = \frac{A\,db}{A_1\,db}.$$

In the given problem, no of required stage is

$$n = \frac{60\,db}{20\,db} = 3.$$

$f_l = 500 HZ$

$$f_L = \frac{f_l}{\sqrt{2^{1/n} - 1}} = \frac{500}{\sqrt{2^{1/3} - 1}} \rightarrow \boxed{f_L = 980.72\ HZ}$$

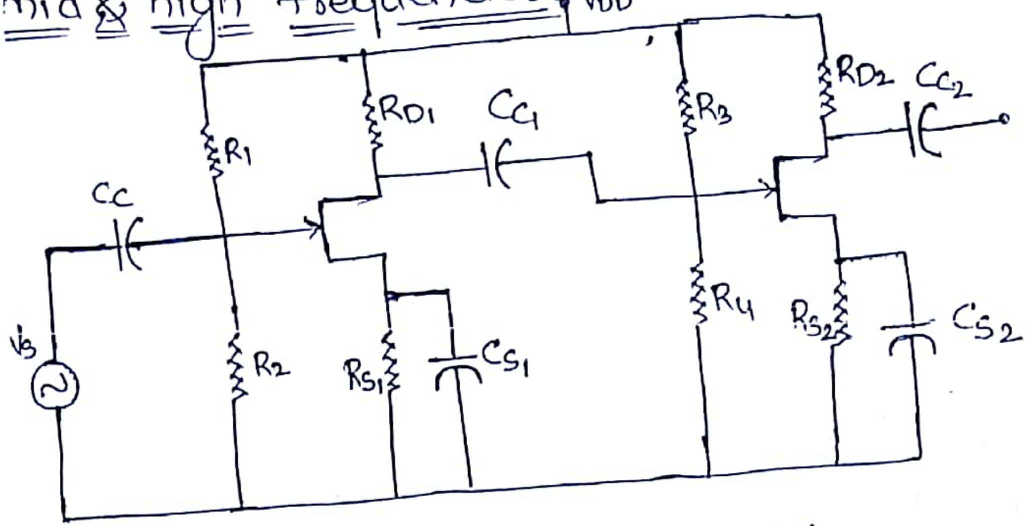
$f_h = 50 KHZ$

$f_H = f_h\left(\sqrt{2^{1/n} - 1}\right)$

$= 50 \times 10^3 \left(\sqrt{2^{1/3} - 1}\right)$

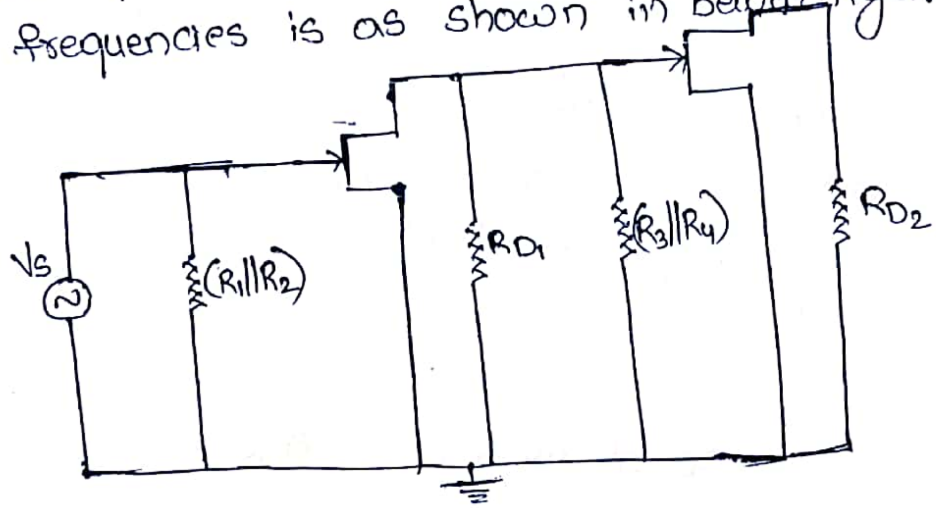
$\boxed{f_H = 25.49\ KHZ}$

Analysis of an RC-coupled FET Amplifier at low, mid & high frequencies:-

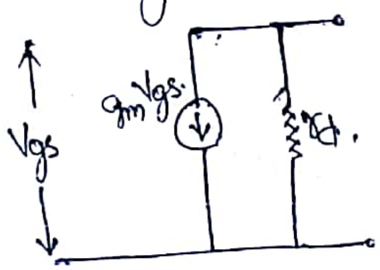


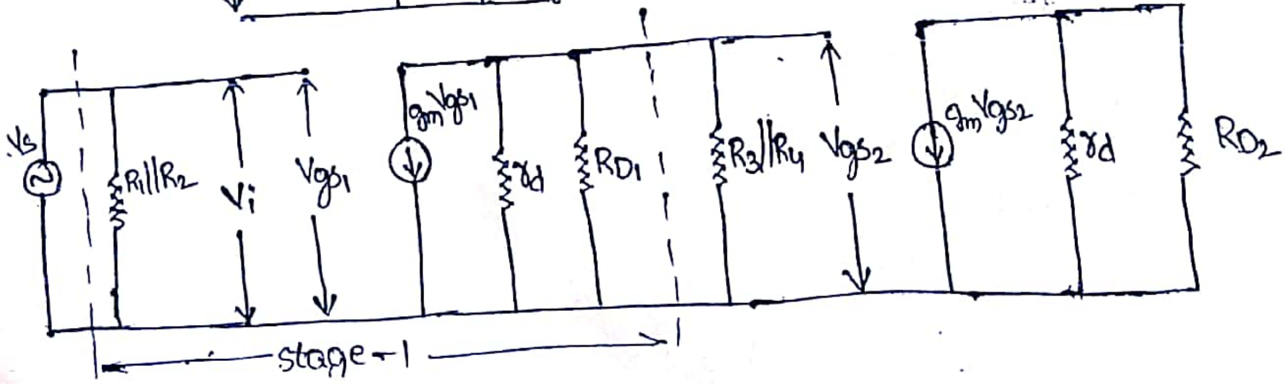
Analysis in the mid-frequency range:-

At mid frequencies coupling and bypass capacitors will be replaced with short circuit. The AC circuit at mid frequencies is as shown in below figure

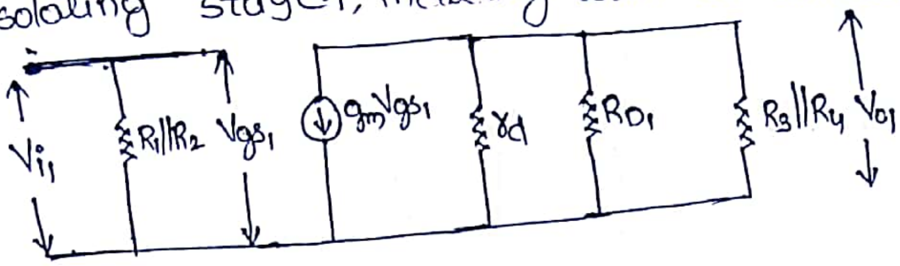


FET small signal equivalent circuit is.





Isolating stage-1, including load to calculate the gain.



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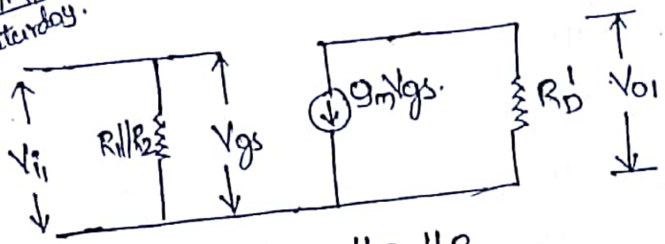
Saturday.



$$R_D' = r_d \| R_{D1} \| R_3 \| R_4 .$$

$$V_{i1} = V_{gs}$$

$$V_{o1} = -g_m V_{gs} R_D'$$

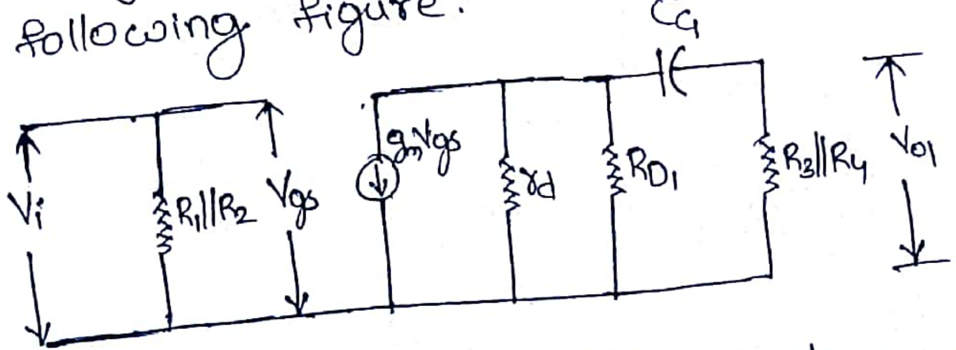
$$A_m = \frac{V_{o1}}{V_{i1}} = \frac{-g_m \cancel{V_{gs}} R_D'}{\cancel{V_{gs}}}$$

$$\boxed{A_m = -g_m R_D'}$$

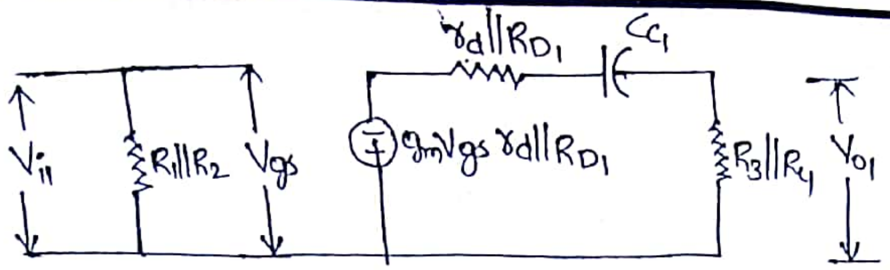
Analysis in low frequency Range:-

At low frequencies coupling capacitor will dominate the frequency response.

The equivalent circuit of a single stage FET Amplifier including loading from the next stage is shown in following figure.



By using source Transformation technique the circuit is.



$$V_{o1} = \frac{-g_m V_{gs} (r_d \| R_{D1}) \times R_3 \| R_4}{r_d \| R_{D1} + \frac{1}{j2\pi f C_{C1}} + R_3 \| R_4}$$

$$V_{o1} = \frac{\dfrac{-g_m V_{gs} (r_d \| R_{D1}) \times (R_3 \| R_4)}{1 (r_d \| R_{D1}) + (R_3 \| R_4)}}{1 + \dfrac{1}{j2\pi f C_{C1} (r_d \| R_{D1} + R_3 \| R_4)}}$$

$$V_{o1} = \frac{-g_m V_{gs} R_D'}{1 - j\left(\frac{f_L}{f}\right)}$$

where $R_D' = \dfrac{(r_d \| R_{D1}) \times (R_3 \| R_4)}{(r_d \| R_{D1}) + (R_3 \| R_4)}$

$$\therefore f_L = \frac{1}{2\pi C_{C1} (R_3 \| R_4 + r_d \| R_{D1})}$$

$V_{i1} = V_{gs}$.

$$A_L = \frac{V_{o1}}{V_{i1}} = \frac{\dfrac{-g_m \cancel{V_{gs}} R_D'}{1 - j\left(\frac{f_L}{f}\right)}}{\cancel{V_{gs}}} = \frac{-g_m R_D'}{1 - j(f_L/f)}$$

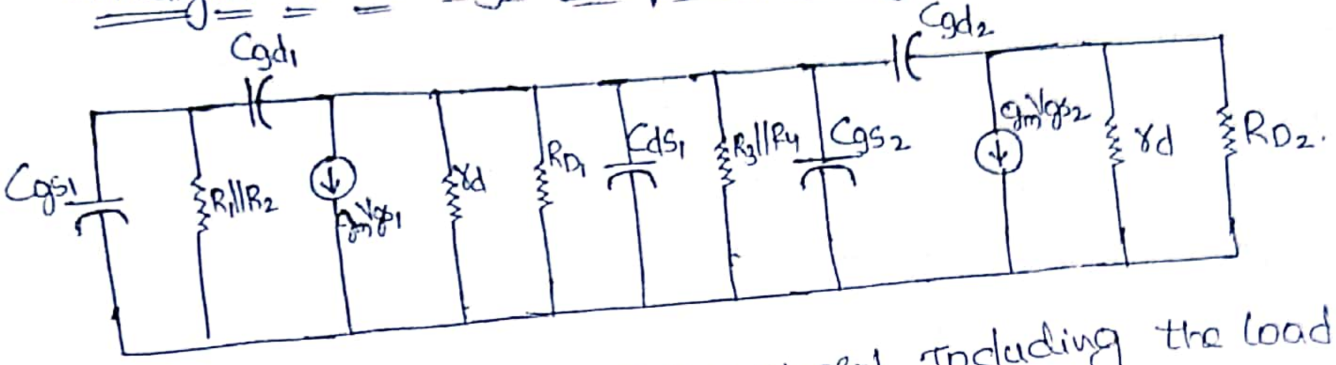
$$\boxed{A_L = \frac{A_m}{1 - j\left(\frac{f_L}{f}\right)}}$$

at $f = f_L$

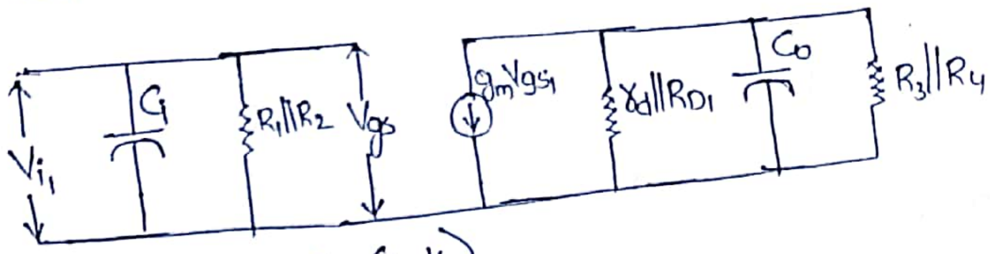
$$\boxed{A_L = \frac{A_m}{\sqrt{2}}}$$

$f_L \rightarrow$ lower cut off frequency.

* **Analysis in the high frequency Range:-**



Using miller's theorem, the stage-1 Including the load can be isolated as follows.



$$C_i = C_{gs} + C_{gd}(1-K_1)$$

$$C_o = C_{gd_1}\left(1-\frac{1}{K_1}\right) + C_{ds_1} + C_{gs_2} + C_{gd_2}(1-K_2)$$

Apply KCL at output node.

$$g_m V_{gs_1} + \frac{V_{o1}}{r_d||R_{D_1}} + \frac{V_{o1}}{\frac{1}{j2\pi f C_o}} + \frac{V_{o1}}{(R_3||R_4)} = 0.$$

$$V_{o1}\left[\frac{1}{r_d||R_{D_1}} + \frac{1}{R_3||R_4} + j2\pi f C_o\right] = -g_m V_{gs_1}$$

$$\left[\frac{1}{r_d||R_{D_1}} + \frac{1}{R_3||R_4} = \frac{1}{R_D'} \qquad V_{gs_1} = V_{i_1}\right]$$

$$V_{o1}\left[\frac{1}{R_D'} + j2\pi f C_o\right] = -g_m V_{gs_1}$$

$$V_{o1}\left[\frac{1}{R_D'} + j2\pi f C_o\right] = -g_m V_{i_1}.$$

$$\frac{V_{o1}}{V_{i_1}} = \frac{-g_m R_D'}{1+j2\pi f C_o R_D'}$$

$$\left[f_h = \frac{1}{2\pi C_o R_D'}\right]$$

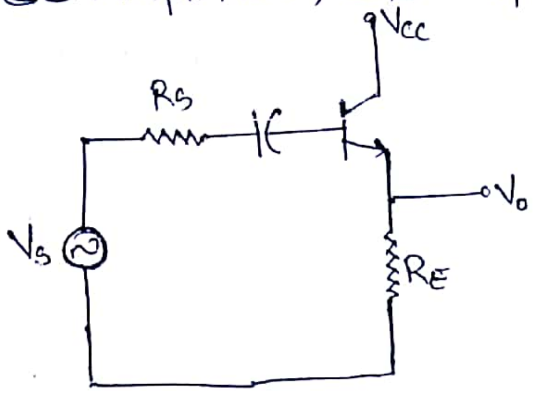
$$\therefore A_h = \frac{A_m}{1+j\left(\frac{f_h}{f}\right)}$$

1/12/18
Friday.

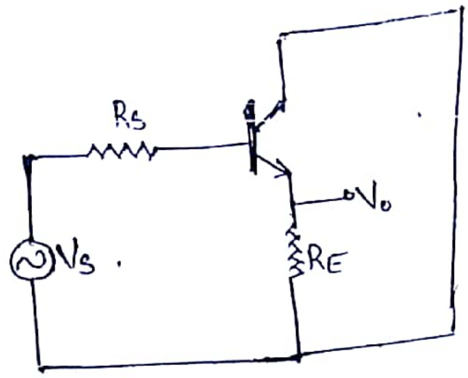
High Input Impedance circuits:-

High input impedance circuits are needed to provide impedance matching.

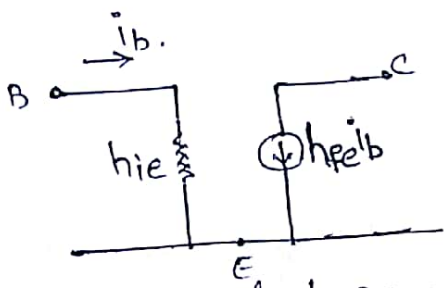
Ex:- CE Amplifier, CD Amplifier.



CC amplifier.

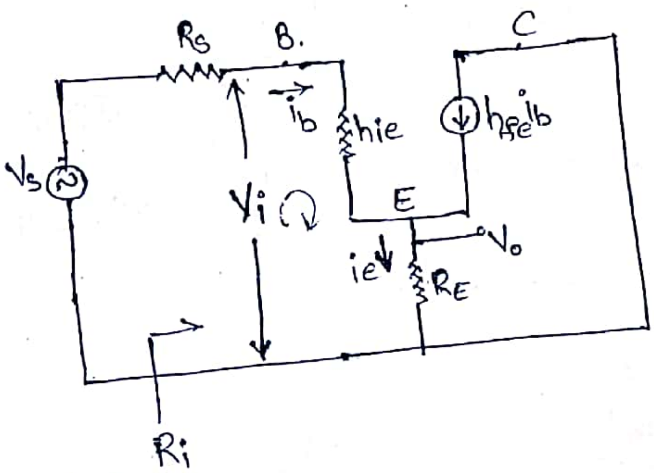


ac circuit.



Approximate h-parameter model.

Replace BJT with approximate h-parameter model.



Current Gain:-

$$A_I = \frac{i_e}{i_b}$$

Applying KCL at emitter node.

$$i_b + h_{fe} i_b = i_e$$

$(1+h_{fe}) i_b = i_e$.

$$\boxed{\therefore A_I = \frac{i_e}{i_b} = (1+h_{fe})}$$

Input Impedance :-

$$R_i = \frac{V_i}{i_b}.$$

Applying KVL to input loop:

$$-V_i + i_b h_{ie} + i_e R_E = 0.$$

$$V_i = i_b h_{ie} + (1+h_{fe}) i_b R_E$$

$$V_i = (h_{ie} + (1+h_{fe}) R_E) i_b.$$

$$\boxed{\therefore R_i = \frac{V_i}{i_b} = h_{ie} + (1+h_{fe}) R_E}$$

Voltage gain:-

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

$$= \frac{i_e R_E}{i_b R_i} = \frac{A_I R_E}{R_i}$$

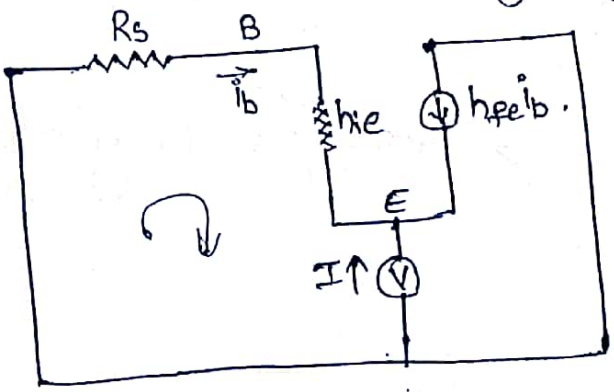
$$\left(A_I = \frac{i_e}{i_b} = (1+h_{fe})\right)$$

$$\left(R_i = h_{ie} + (1+h_{fe}) R_E\right)$$

$$\boxed{A_V = \frac{(1+h_{fe}) R_E}{h_{ie} + (1+h_{fe}) R_E}}$$

Output Impedance:-

To find output impedance Replace V_s with short circuit & disconnect R_E and connect voltage source V. find $\frac{V}{I}$ as shown in the following figure.



By applying KCL at ~~output~~ Emitter node.

$$i_b + h_{fe} i_b + I = 0$$

$$I = -(1+h_{fe}) i_b$$

Applying KCL at input

$$R_s i_b + h_{ie} i_b + V = 0$$

$$V = -(R_s + h_{ie}) i_b$$

$$R_o = \frac{V}{I} = \frac{-(R_s+h_{ie}) i_b}{-(1+h_{fe}) i_b} \Rightarrow \boxed{R_o = \frac{R_s + h_{ie}}{1+h_{fe}}}$$

Summary:-

$A_I = (1+h_{fe}) \longrightarrow$ Large.

$R_i = h_{ie} + (1+h_{fe}) R_E \longrightarrow$ Large.

$$A_V = \frac{(1+h_{fe}) R_E}{h_{ie} + (1+h_{fe}) R_E}$$

$$h_{ie} < (1+h_{fe}) R_E$$

$$A_V \simeq \frac{(1+h_{fe}) R_E}{(1+h_{fe}) R_E} \simeq 1 \longrightarrow \text{voltage gain low. approx}$$

$$R_o = \frac{R_s + h_{ie}}{1+h_{fe}} \longrightarrow \text{low.}$$

∴ CC Amplifier has high input impedance and low output impedance with unity voltage gain and large current gain.

∴ It can be used as impedance matching circuit.

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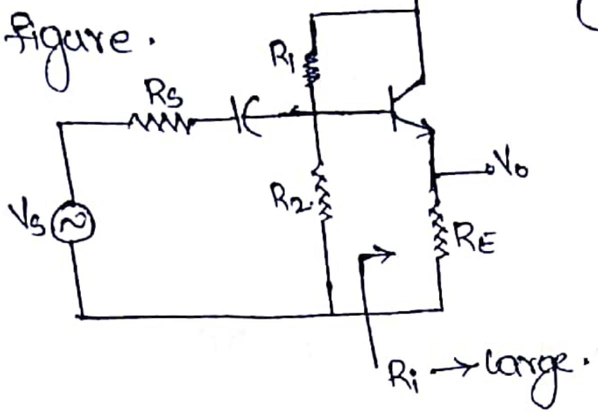
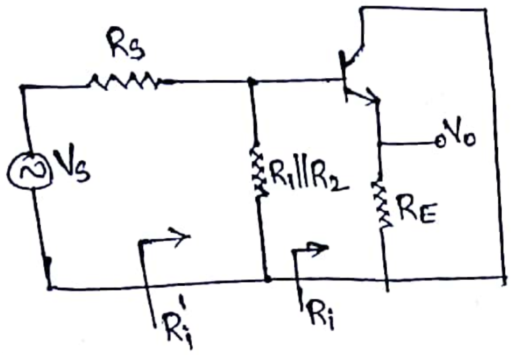
Limitation of input impedance of CC Amplifier:

CC Amplifier with biasing resistors is shown in the following figure.



The ac circuit is shown in below figure.



The input impedance R_i is.

$$R_i = h_{ie} + (1+h_{fe})R_E$$

The net input impedance R_i' is

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

But $R_1 \| R_2 \gg R_i$

$\therefore R_i' \simeq R_1 \| R_2$ which is small.

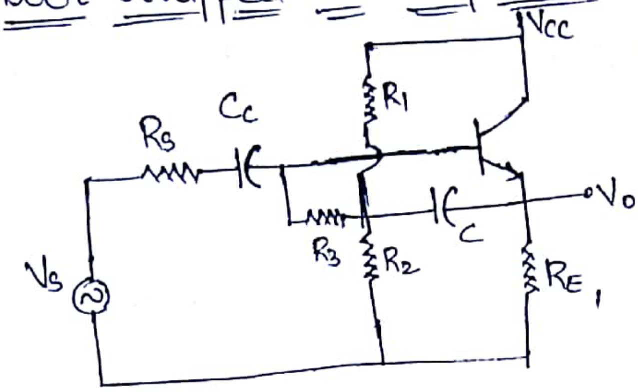
$\therefore$ Even though CC Amplifier has large input impedance.

The net input impedance R_i' is very small because of the effect of Biasing Resistors.

The above problem is overcome by using a Boot strapped CC Amplifier.

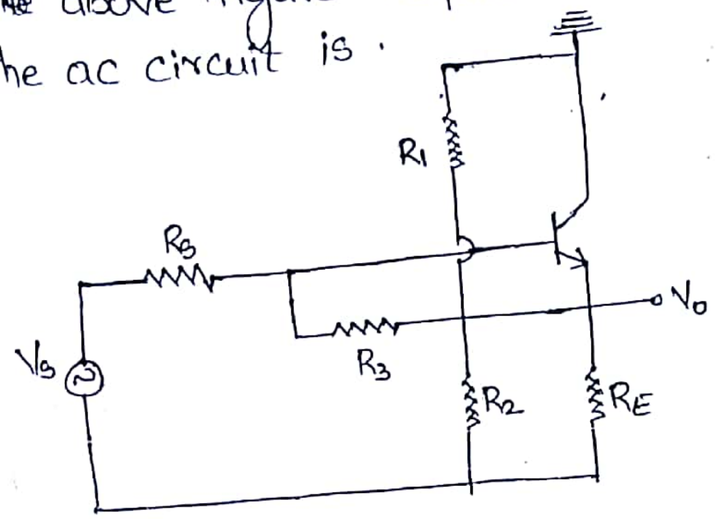
31/12/18 Monday

Boot strapped CC Amplifier :-



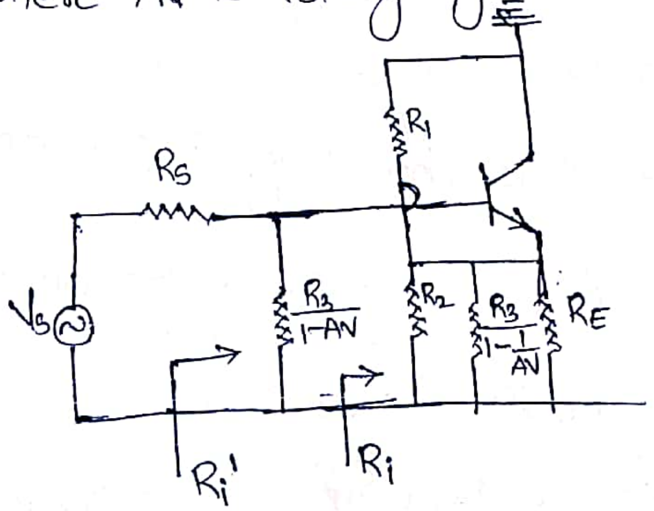
In the above figure capacitor C short circuits for ac signals.

The ac circuit is .



The resistance R_3 can be split using miller's theorem. then the equivalent value of R_3 between base and ground is $\frac{R_3}{1-A_V}$.

where A_V is voltage gain between base and emitter.



$$\therefore R_i' = \frac{R_3}{1-A_V} \parallel R_i$$

AV for a CC amplifier is approximately equal to 1.

$\therefore R_i' \simeq \infty \| R_i \simeq R_i$

$\boxed{\therefore R_i' \simeq R_i}$

Parameters.	CC amplifier. approximate h-model	CC Amplifier exact h-model.
AI	$(1+h_{fe})$	$\frac{-h_{fc}}{1+h_{oc}R_L} = \frac{(1+h_{fe})}{1+h_{oe}R_L}$
R_i	$h_{ie}+A_I R_L$	$h_{ic}+h_{rc}A_I R_L = h_{ie}+A_I R_L$
AV	$\frac{A_I R_L}{R_i}$	$\frac{A_I R_L}{R_i}$.
Y_o	$\frac{1+h_{fe}}{R_s+h_{ie}}$	$h_{oc}-\frac{h_{fc}h_{rc}}{R_s+h_{ic}} = h_{oe}+\frac{(1+h_{fe})}{R_s+h_{ie}}$

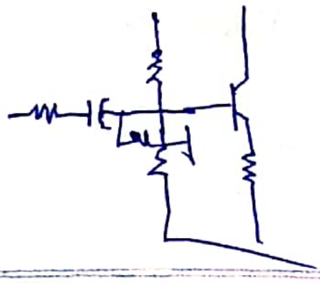
Conversion Formulas :-

$\cancel{h_{fc}}$ $h_{ic} = h_{ie}$

$h_{fc} = -(1+h_{fe})$

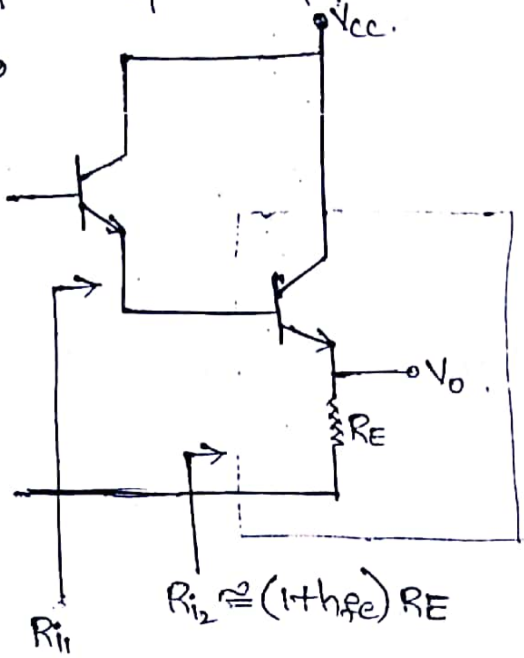
$h_{rc} = 1$

$h_{oc} = h_{oe}$.



Darlington pair :-

Darlington pair is a two stage cascaded CC Amplifier.
In applications where input impedance of single stage CC Amplifier is not sufficient, Darlington pair is used.
Darlington pair input impedance is in the order of mega ohms



$$R_{i_1} \simeq (1+h_{fe}) R_{i_2}$$
$$\simeq (1+h_{fe})^2 R_E$$

Analysis of second stage:

For second stage we can use approximate h-parameter model

Current gain $A_{I_2} = (1+h_{fe}) \longrightarrow$ large.

Input impedance $R_{i_2} = h_{ie} + A_{I_2} R_E$

$$= h_{ie} + (1+h_{fe}) R_E$$
$$\simeq (1+h_{fe}) R_E \longrightarrow \text{large}.$$

Voltage gain $A_{V_2} = \dfrac{A_{I_2} R_{L_2}}{R_{i_2}} = \dfrac{A_{I_2} R_E}{R_{i_2}} = \dfrac{R_{i_2} - h_{ie}}{R_{i_2}}$

$$= \left(1 - \frac{h_{ie}}{R_{i_2}}\right) \simeq 1$$

output Impedance. $R_{o2} = \frac{R_s + h_{ie}}{1+h_{fe}} = \frac{R_{o1} + h_{ie}}{1+h_{fe}}$

<u>Analysis of first stage</u> :-

For first stage analysis we must use exact model because $h_{oe}R_L < 0.1$ condition is not satisfied.

current gain. $A_{I_1} = \frac{(1+h_{fe})}{1+h_{oe}R_L} = \frac{1+h_{fe}}{1+h_{oe}R_{i_2}}$.

$$A_{I_1} = \frac{(1+h_{fe})}{1+h_{oe}(1+h_{fe})R_E} \Rightarrow \text{large}.$$

Input impedance. $R_{i_1} = h_{ie} + A_{I_1}R_L = h_{ie} + A_{I_1}R_{i_2}$

$$= h_{ie} + \frac{(1+h_{fe})}{1+h_{oe}(1+h_{fe})R_E}(1+h_{fe})R_E$$

$$\cong \frac{(1+h_{fe})^2 R_E}{1+h_{oe}(1+h_{fe})R_E} \rightarrow \text{very large: Typically in mega ohms}$$

voltage gain $A_{V_1} = \frac{A_{I_1}R_L}{R_{i_1}}$

$$= \frac{R_{i_1} - h_{ie}}{R_{i_1}} = \left(1 - \frac{h_{ie}}{R_{i_1}}\right)$$

$$\cong 1$$

Output Impedance. $Y_{o1} \cong h_{oe} + \frac{(1+h_{fe})}{(R_s + h_{ie})}$

overall current gain $A_I = A_{I_1} \times A_{I_2} = \text{very large}$.

over all voltage gain. $A_V = A_{V_1} \times A_{V_2}$.

$$= (<1) \times (<1)$$

$$= <1$$

Over all Input impedance $R_i = R_{i_1}$ = very large (mega ohms)

Over all output impedance $R_o = R_{o_2}$ = small

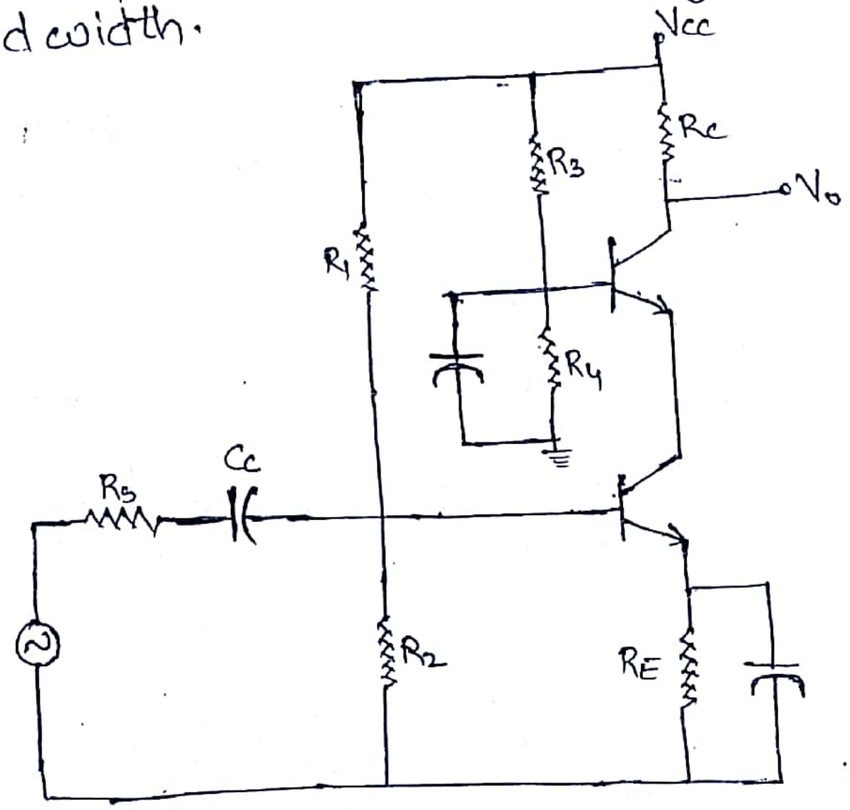
2/01/2019
Wednesday

Cascode Amplifier:-

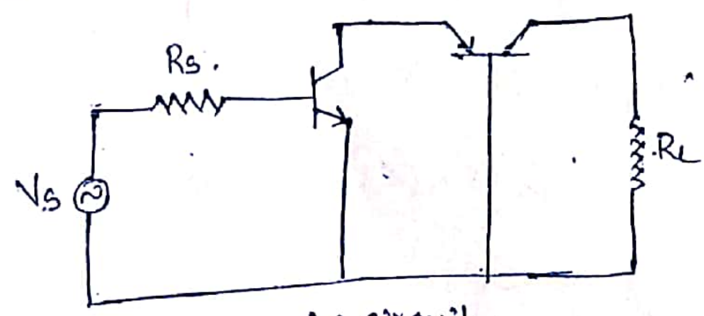
The combination of CE, CB amplifier is called cascode Amplifier. It is a high gain Amplifier

The bandwidth of cascode Amplifier is approximately same as the bandwidth of first stage CE Amplifier

Cascode Amplifier increases the gain without reducing the bandwidth.



The AC circuit can be given as.



AC circuit

over all voltage gain of the system $AV = AV_1 \times AV_2$

$AV_1 \longrightarrow$ CE voltage gain (large)

$AV_2 \longrightarrow$ CB voltage gain (large)

$\therefore AV = AV_1 \times AV_2$ = Very large.

over all current gain $AI = AI_1 \times AI_2$

$AI_1 \longrightarrow$ CE current gain (large)

$AI_2 \longrightarrow$ CB current gain ($\simeq 1$)

$\therefore AI = AI_1 \times AI_2 \simeq AI_1$

Input impedance of the cascode Amplifier is the input ~~the~~ impedance of CE Amplifier which is moderate.

The output impedance of cascode Amplifer is the output impedance of CB stage which is very large.

The disadvantage of cascode amplifier is which has very large output impedance.

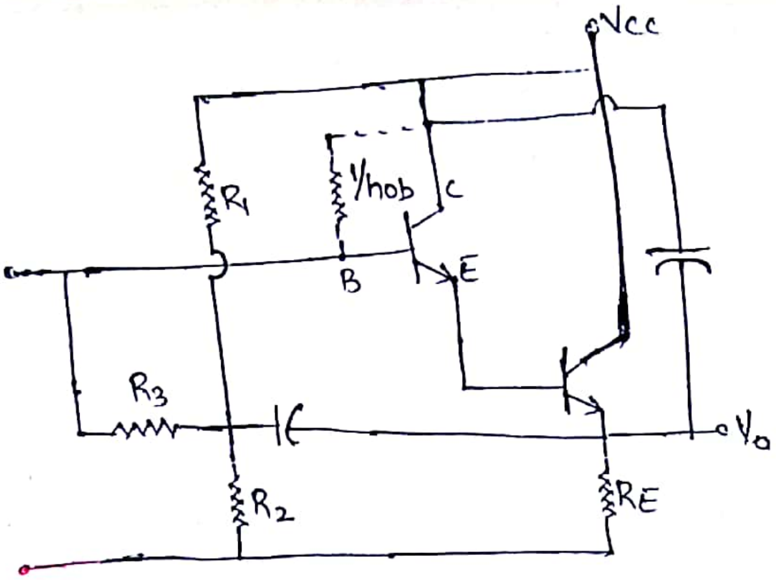
Bootstraped Darlington pair :-

Even though the input impedance of Darlington pair is very high, the net input impedance will fall drastically due to the following facts.

1) Due to Biasing resistors R_1, R_2 that appear in parallel with R_i

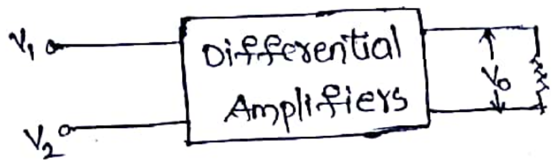
2) The miller equivalent of impedance $1/h_{ob}$ which is $\frac{1/h_{ob}}{1-K}$. this impedance will appear in parallel with R_i.

Both these effects can be avoided by using Bootstrapping technique as shown in the below figure.



Differential Amplifier :-

Differential Amplifier is a dual input amplifier. The output of differential Amplifier is proportional to the difference of the two Input signals.



$V_o \propto (V_1 - V_2)$

$V_o = A_d (V_1 - V_2)$

$V_o = A_d V_d$

A_d = Differential gain.

Practical differential Amplifiers will amplify the not only the differential signal but also the common mode signal. The output voltage for a practical differential amplifier is.

$V_o = A_d V_d + A_c V_c.$

$V_d = V_1 - V_2$ (difference signal)

$V_c = \frac{V_1 + V_2}{2}$ = common mode signal

A_d = Differential gain

A_c = common mode gain

For ideal differential Amplifier $A_d = 0$.

For practical differential Amplifier $A_d >> A_c$.

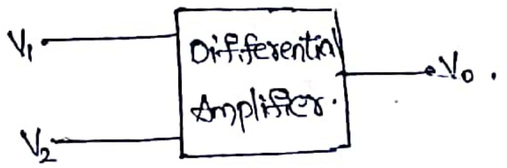
Common mode Rejection Ratio (CMRR):-

It is the ratio of differential gain to common mode gain. It is represented by ρ.

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

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

03/01/19
Thursday



since differential Amplifier is linear, it obeys superposition.

$$V_o = A_1V_1 + A_2V_2$$

But $V_o = A_dV_d + A_cV_c$

$$V_o = A_d(V_1 - V_2) + A_c\left(\frac{V_1+V_2}{2}\right)$$

$$V_o = V_1\left(A_d + \frac{A_c}{2}\right) + V_2\left(\frac{A_c}{2} - A_d\right)$$

$$V_o = A_1V_1 + A_2V_2.$$

$$\frac{A_c}{2} + A_d = A_1 \longrightarrow ①.$$

$$\frac{A_c}{2} - A_d = A_2. \longrightarrow ②.$$

$① + ② \rightarrow A_c = A_1 + A_2.$

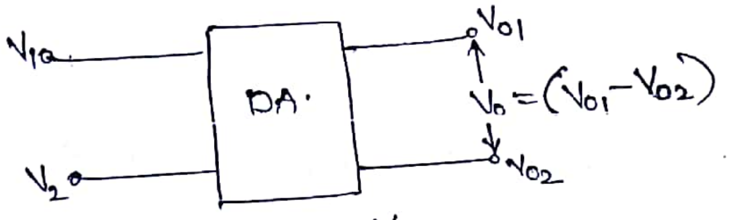
$① - ② \rightarrow 2A_d = A_1 - A_2$

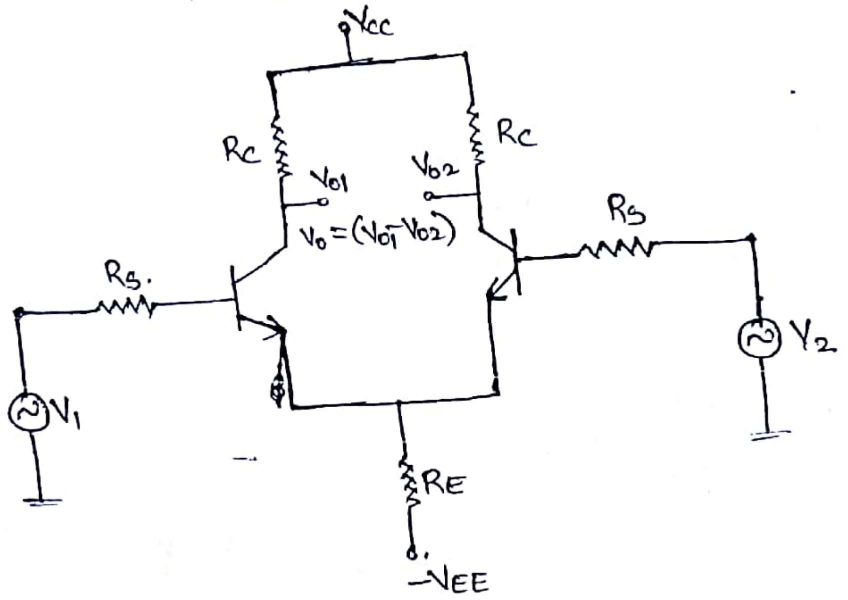
$A_d = \frac{A_1 - A_2}{2}$

Types of Differential Amplifier :-

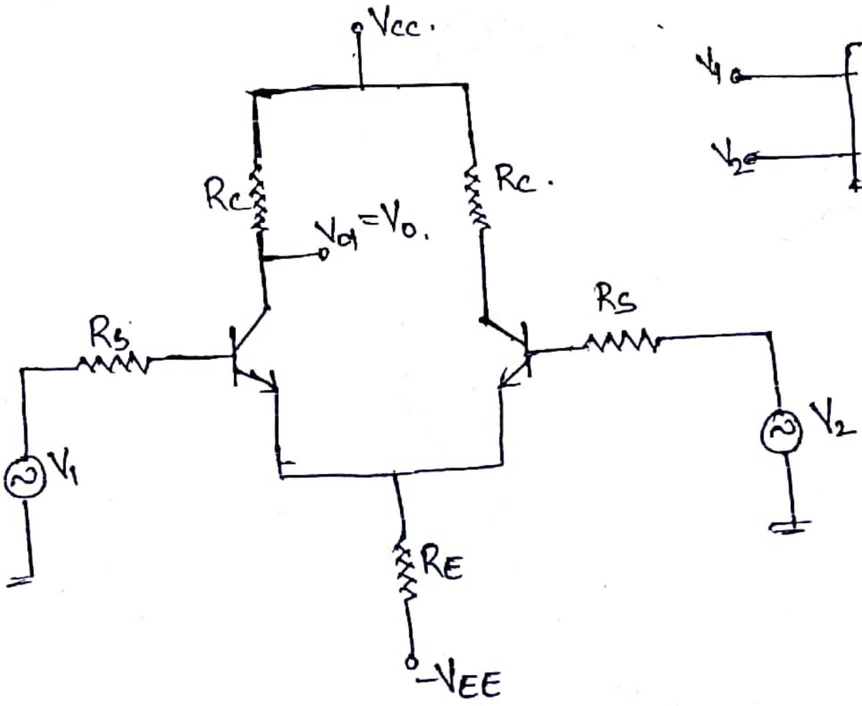
Depending upon the number of inputs & outputs differential Amplifiers can be divided in to four types.

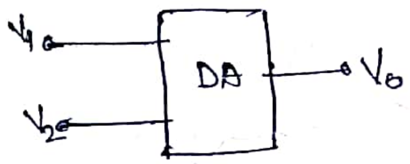
1. Dual Input balanced output differential Amplifier.



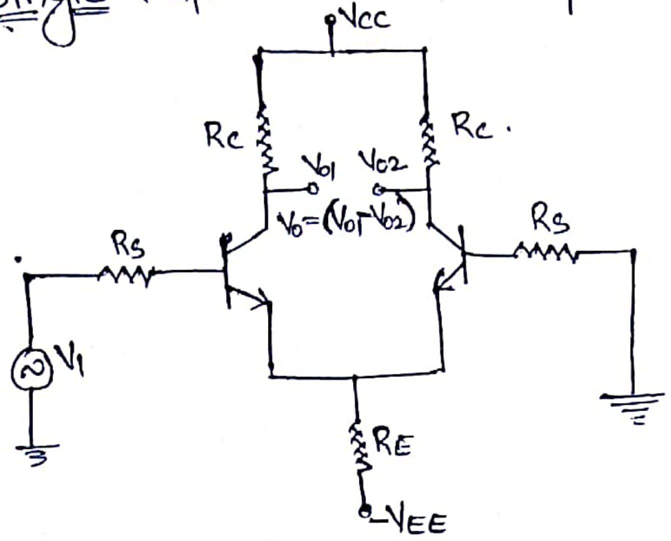


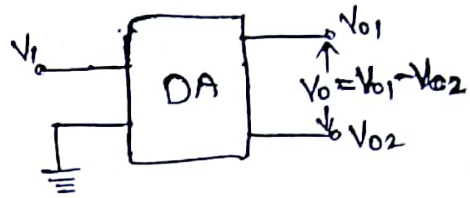
2. Dual Input unbalanced output differential Amplifier.



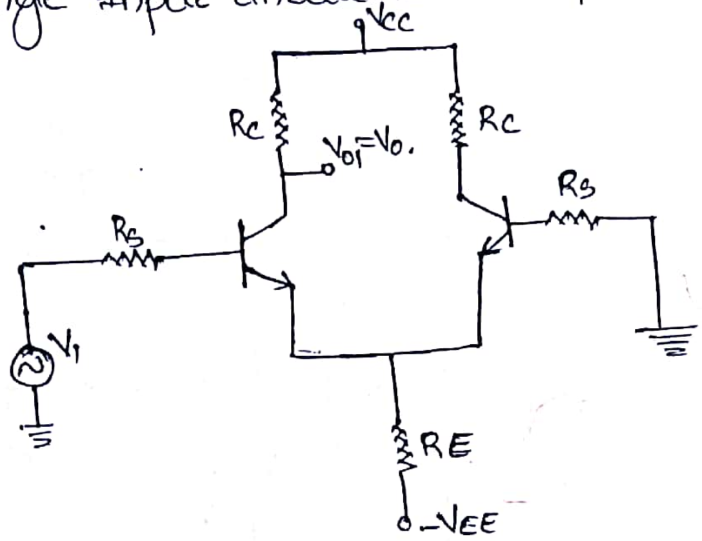


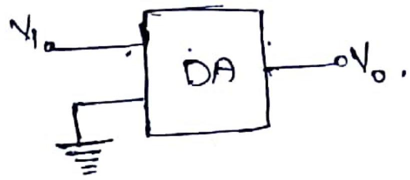
3) Single Input balanced output differential Amplifier:-





4) Single input unbalanced output differential Amplifier.





Advantages of Differential Amplifier:-

1. Differential Amplifiers will have large differential gain and small common mode gain.
2. Differential Amplifier supresses the common mode signal.
3. Differential Amplifier eliminates common noise present on the two input signal.
4. Power supply flucuations ~~are~~ ~~supressed~~. cause disturbance in the output and this can be eliminate using d.A.
5. Differential Amplifier will have high input impedance

Differential Amplifiers are direct coupled Amplifiers.
∴ Low frequency response is very good.

04/01/19
Friday.

3. Feed back Amplifiers.

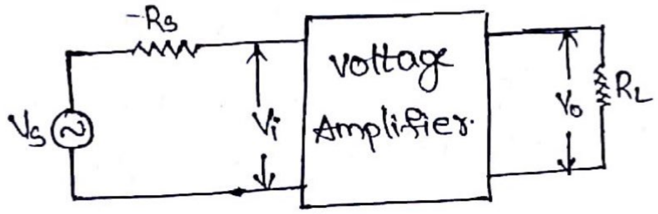
Types of Amplifiers:-

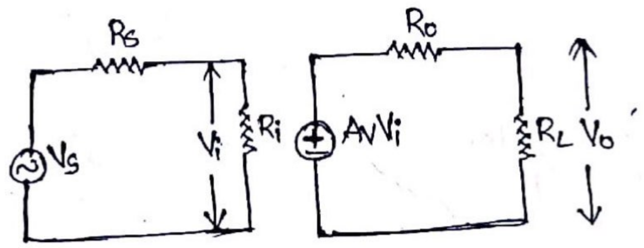
Depending upon the type of input and output signals amplifiers can be divided in to 4 types. They are

1) Voltage Amplifier
2) Current Amplifier.
3) Transconductance Amplifier.
4) Trans Resistance Amplifier.

Voltage Amplifier :-

This type of amplifier takes voltage as an input sign and generate voltage as an output signal.





$R_i, R_o \rightarrow$ Input, output impedances.

$A_V \rightarrow$ open circuit voltage gain.

$$V_i = \frac{V_s R_i}{R_i + R_s} = \frac{V_s}{1 + \frac{R_s}{R_i}}$$

If $R_i = \infty$

$V_i = V_s$.

$\therefore$ For ideal voltage amplifiers, $R_i = \infty$

For practical voltage amplifiers R_i must be large.

i.e., $R_i >> R_s$

$$V_o = \frac{A_V V_i R_L}{R_o + R_L} = \frac{A_V V_i}{1 + \frac{R_o}{R_L}}$$

If $R_o = 0$; $V_o = A_V V_i$

$\therefore$ For ideal voltage amplifiers $R_o = 0$.

For practical voltage amplifies, R_o must be small.

$$V_o = \frac{A_V V_i R_L}{R_o + R_L}$$

$$\frac{V_o}{V_i} = AV = \frac{A_V R_L}{R_o + R_L}$$

$$AV = \frac{A_V}{1 + \frac{R_o}{R_L}}$$

If $R_L = \infty$.

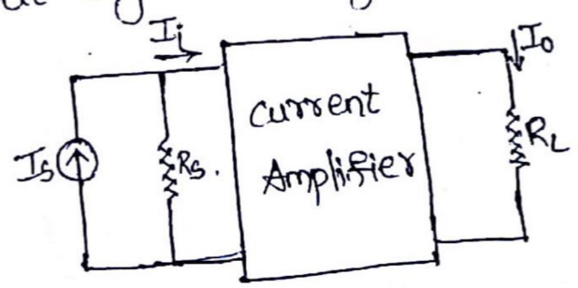
$$AV = A_V$$

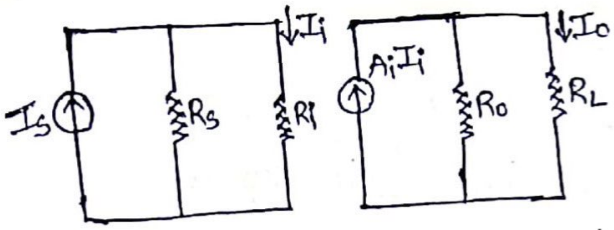
$AV \rightarrow$ voltage gain.

$A_V \rightarrow$ open circuit voltage gain.

Current Amplifier:-

This type of amplifier takes current sources as an input signal and generates current as an output signal.





$R_i, R_o \rightarrow$ input, output Impedance.
$A_i \longrightarrow$ short circuit ~~sh~~current gain

$$I_i = \frac{I_s R_s}{R_s + R_i} = \frac{I_s}{1 + \frac{R_i}{R_s}}$$

If $R_i = 0$

$\therefore I_i = I_s$.

$\therefore$ For ideal current amplifier $R_i = 0$

For practical current amplifier R_i must be very small.

$$I_o = \frac{A_i I_i R_o}{R_o + R_L} = \frac{A_i I_i}{1 + \frac{R_L}{R_o}}$$

If $R_o = \infty$,

$$I_o = A_i I_i$$

For ideal current Amplifier $R_o = \infty$

For practical current Amplifier, R_o must be very Large.

$$I_o = \frac{A_i I_i R_o}{R_o + R_L}$$

$$A_I = \frac{I_o}{I_i} = \frac{A_i R_o}{R_o + R_L}$$

$$A_I = \frac{A_i}{1 + \frac{R_L}{R_o}}$$

If $R_L = 0$.

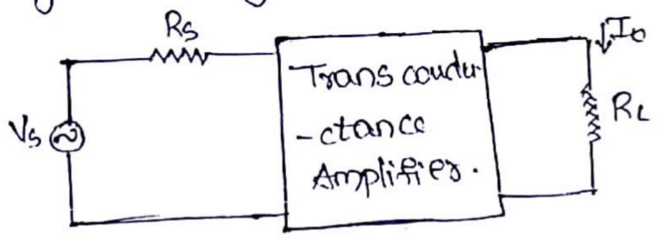
$$A_I = A_i$$

$A_I \rightarrow$ current gain

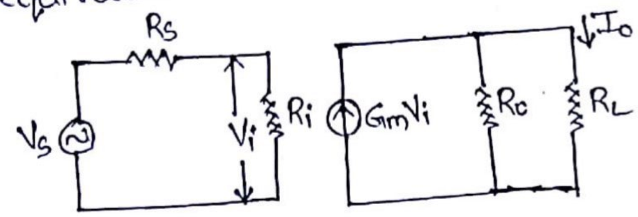
$A_i \longrightarrow$ short circuit current gain.

3) Trans conductance Amplifier:-

This type of amplifier takes voltage as an input signal and generate current as an output signal.



Equivalent circuit is.



$$V_i = \frac{V_S R_i}{R_S + R_i} = \frac{V_S}{1 + \frac{R_S}{R_i}}$$

If $R_i = \infty$,

$V_i = V_S$.

For ideal transconductance Amplifier $R_i = \infty$

For practical transconductance Amplifier R_i must be Large

$$I_o = \frac{G_m V_i R_o}{R_o + R_L}$$

$$I_o = \frac{G_m V_i}{1 + \frac{R_L}{R_o}}$$

If $R_o = \infty$,

$I_o = G_m V_i$

For ideal transconductance Amplifier $R_o = \infty$

For practical transconductance Amplifier; R_o must be as large as possible.

$$I_o = \frac{G_m V_i R_o}{R_o + R_L}$$

$$\frac{I_o}{V_i} = G_M = \frac{G_m R_o}{R_o + R_L}$$

$$G_M = \frac{G_m}{1 + \frac{R_L}{R_o}}.$$

If $R_L = 0$,

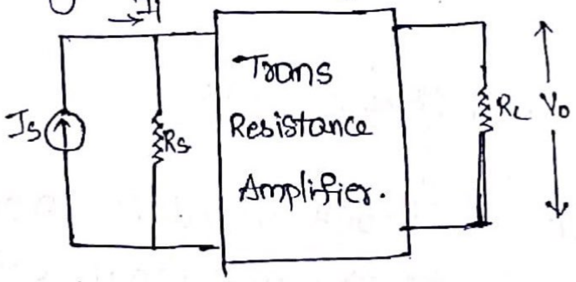
$$G_M = G_m$$

$G_M \longrightarrow$ Transconductance Gain.

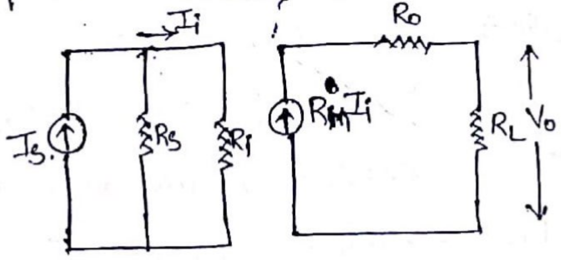
$G_m \longrightarrow$ Short circuit Transconductance Gain.

4) **Trans Resistance Amplifier :-**

This type of amplifier takes current as an input sign and generate voltage as an output signal.



equivalent circuit is.



$$I_i = \frac{I_s R_s}{R_s + R_i} = \frac{I_s}{1 + \frac{R_i}{R_s}}.$$

If $R_i = 0$; $I_i = I_s$.

For ideal Amplier $R_i = 0$

For practical transresistance Amplifier R_i must be small.

$R_i << R_s$.

$$V_o = \frac{R_m I_i R_L}{R_o + R_L} = \frac{R_m I_i}{1 + R_o}$$

If $R_o = 0$, $V_o = R_m^o I_i$

For Ideal transresistance Amplifier $R_o = 0$.
For practical transresistance Amplifier R_o should be low.

$$V_o = \frac{R_m^o I_i R_L}{R_o + R_L}$$

$$R_M = \frac{V_o}{I_i} = \frac{R_m^o R_L}{R_o + R_L} = \frac{R_m^o}{\frac{R_o}{R_L} + 1}$$

If $R_L = \infty$,

$\boxed{R_M = R_m}$

R_M → Trans resistance gain.
R_m → open circuit Trans resistance gain

05/01/19
saturday.

Definition of feed back:-

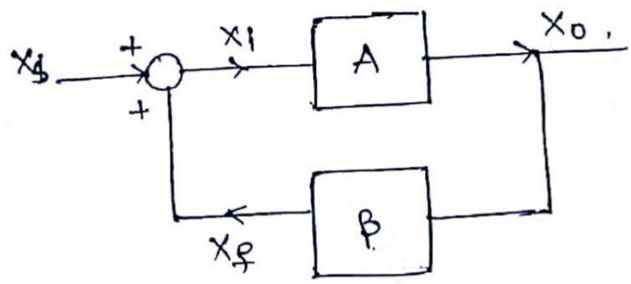
Feed back is a method (or) a process in which the output signal of an amplifier is sampled and feed back to the input.

Types of feed back:-

1. Positive feed back. 2. Negative feedback.

Positive Feedback:-

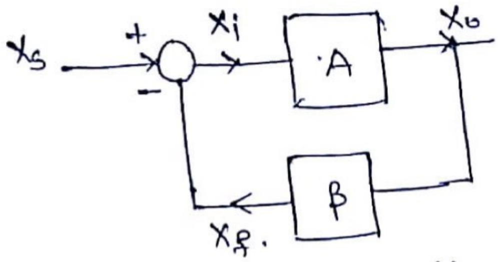
In this type of feed back, feed back signal is in phase with the input signal. This increases the net input and hence the output also increases.



Here $X_i = X_s + X_f$.

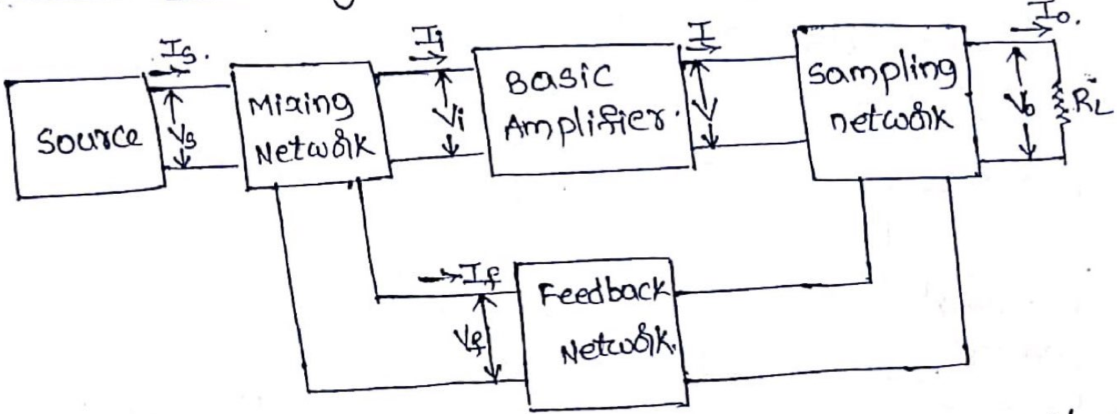
Negative Feed back.

In this type of feed back the feed back signal is out of phas with the input signal. This decreases the net input and hence the output also decreases.



Here $X_i = X_s - X_f$.

General Block diagram of feedback circuit:-



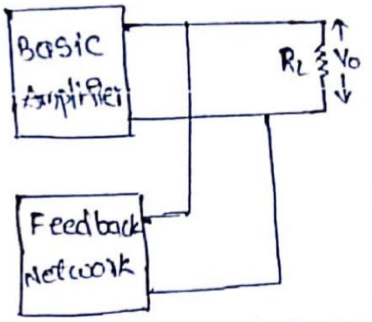
Source:- The signal source can be voltage source V_s or a current source I_s.

Basic Amplifier:- This amplifier can be a voltage Amplifier with gain A_V or a current Amplifier with gain A_I or a transresistance amplifier with gain R_M or a transconductance amplifier with gain G_M.

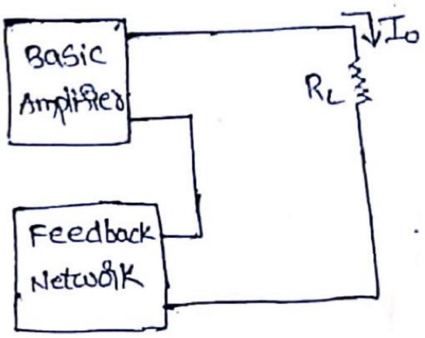
Feedback Network:- In general feed back network consists of passive elements. like resistors and capacitors in very rare applications the feed back network may also contain Active elements.

Sampling Network :-

~~There are~~ Two types of sampling is possible with the output signal is voltage then it is sampled in parallel as shown in the following figure.



If the output signal is current then the sampling is done in series as. shown in figure.

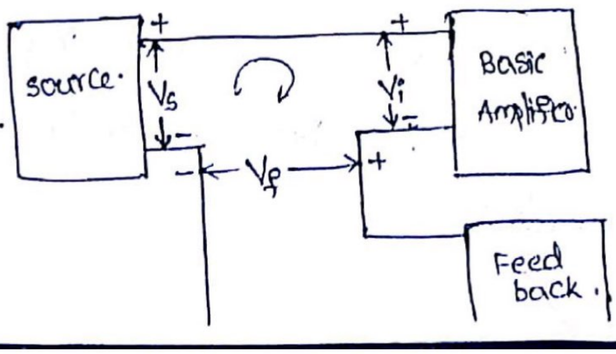


29/01/19
Wednesday

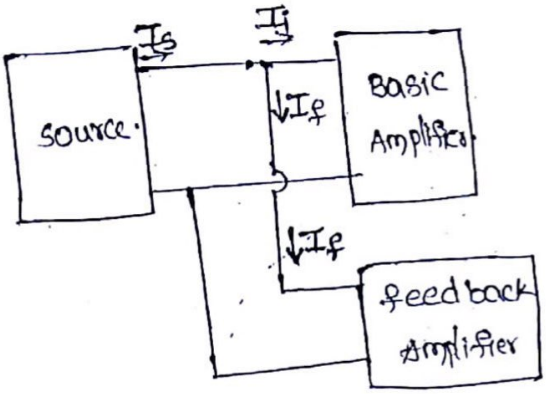
Mixing Network or Summing Network :-

Depending upon the input and feedback signal two types of mixing is possible.

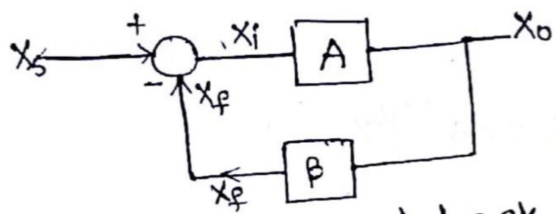
If the feedback signal is voltage signal it is added in series with the input ~~voltage~~ signal. This type of mixing is called series mixing.



If the feedback signal is current signal, then it is added in parallel (shunt) with the input signal. This type of mixing is called ~~parall~~ shunt mixing.



Analysis of Negative feedback Amplifier (or) Prove that negative feed back reduces the gain:-



when there is no feed back

$$X_s = X_i$$

The gain without feed back.

$$A = \frac{X_o}{X_s} = \frac{X_o}{X_i}$$

$$X_o = AX_i \longrightarrow ①$$

when there is feed back

$$X_i = X_s - X_f$$

$$X_f = \beta X_o$$

$$\therefore X_i = X_s - \beta X_o \longrightarrow ②$$

from ① & ②

$$X_i = X_s - \beta A X_i$$

$$X_i \quad X_i(1+\beta A)$$

Now the gain with feedback is. $\left[A = \frac{X_o}{X_i}\right]$

$$A_f = \frac{X_o}{X_s} = \frac{X_o}{X_i(1+A\beta)} = \frac{A}{1+A\beta}.$$

$$\boxed{A_f = \frac{A}{1+A\beta}}$$

where $A \rightarrow$ Gain without feedback (or) open loop gain.

$A_f \rightarrow$ Gain with feed back (or) closed loop gain

$A\beta \rightarrow$ loop gain.

$(1+A\beta). \rightarrow$ Decensitivity factor.

Normally loop gain $A\beta >> 1$

$\therefore A_f < A.$

$\therefore$ Negative feedback reduces the Gain.

0/01/19
Thursday

Advantages of Negative feed back:-

) Negative feed back stabilizes the gain.

* Negative feed back decensitizes the gain and makes the over all gain independent of A. (A - open loop gain)

2) Negative feedback increases the band width.

3) Negative feedback reduces noise.

4) Negative feedback reduces distortion.

5) Negative feedback changes input & output impedances in favour of the amplifier.

Negative feedback stabilizes the gain :-

$$A_f = \frac{A}{1+A\beta}$$

D. w.r.to A on B.S.

$$\left[\frac{d}{dx}\left(\frac{u}{v}\right) = \frac{v\frac{d}{dx}(u) - u\frac{d}{dx}v}{v^2}\right]$$

$$\frac{dA_f}{dA} = \frac{d}{dA}\left(\frac{A}{1+A\beta}\right)$$

$$\frac{dA_f}{dA} = \frac{(1+A\beta)(1) - A(\beta)}{(1+A\beta)^2}$$

$$\frac{dA_f}{dA} = \frac{1}{(1+A\beta)^2}$$

$$dA_f = \frac{dA}{(1+A\beta)^2}$$

Divide with A_f on B.S.

$$\frac{dA_f}{A_f} = \frac{dA}{A_f(1+A\beta)^2} = \frac{dA}{\left(\frac{A}{(1+A\beta)}\right)(1+A\beta)^2}$$

$$\frac{dA_f}{A_f} = \frac{dA}{A(1+A\beta)} = \frac{\frac{dA}{A}}{(1+A\beta)}$$

$$\boxed{\frac{dA_f}{A_f}\times 100 = \frac{\frac{dA}{A}\times 100}{(1+A\beta)}}$$

Negative feed back decensitizes the gain (or) makes gain independent of A :

$$A_f = \frac{A}{1+A\beta}$$

Normally loop gain is selected such that $A\beta \gg 1$

$$\therefore A_f \cong \frac{A}{A\beta} \simeq \frac{1}{\beta} \Rightarrow \boxed{A_f \simeq \frac{1}{\beta}}$$

$\therefore A_f$ is independent of A.

A_f is more stable because now it depends only on feed back factor β.
β network consists of passive elements like resistors and capacitors which are stable.

2) Negative feed back increases the band width :-

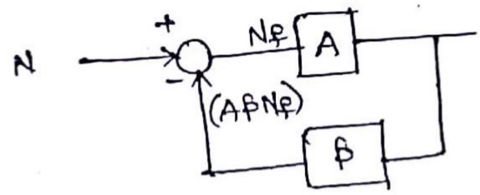
Let A, BW be the gain ; Bandwidth with out feed back.
Let A_f, BW_f be the gain, Bandwidth with feedback.
Gain bandwidth product is constant for an Amplifier.

$$\therefore A(BW) = A_f(BW_f)$$

$$A(BW) = \frac{A}{1+A\beta}(BW_f)$$

$$BW_f = (1+A\beta)(BW).$$

3) Negative feed back Reduces noise :-



In the above figure N represents noise at the input of amplifier when there is no feedback.
N_f represents noise at the input of the amplifier after introducing feedback.
From the figure we can

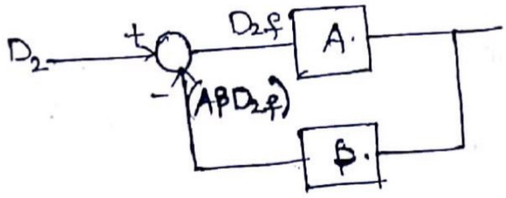
$$N - A\beta N_f = N_f$$

$$N = N_f + A\beta N_f$$

$$N = (1+A\beta) N_f$$

$$\boxed{N_f = \frac{N}{(1+A\beta)}}$$

4) Negative feedback Reduces distortion (Non-linear Distortion & Harmonic distortion):-



Let D_2 represents second harmonic distortion components without feed back.

D_{2f} represents the second harmonic component with feed back.

∴ From the figure

$$D_2 - A\beta D_{2f} = D_{2f}.$$

$$D_2 = D_{2f} + A\beta D_{2f}$$

$$D_2 = (1+A\beta) D_{2f}.$$

$$\boxed{D_{2f} = \frac{D_2}{(1+A\beta)}}$$

5). Negative feed back Changes input & output impedances in favour of the amplifier :-

If the feedback signal is returned in series, then the input impedance will increase by $(1+A\beta)$ times due to feed back.

Ex:- voltage series, current series feed back.

If the feedback signal is returned in shunt then input impedance decreases by $(1+A\beta)$ times due to feedback.
Ex:- voltage shunt, current shunt.

If the output signal is voltage negative feedback reduces the output impedance by $(1+A\beta)$ times.

Ex:- Voltage shunt, ~~current shunt~~ voltage series.

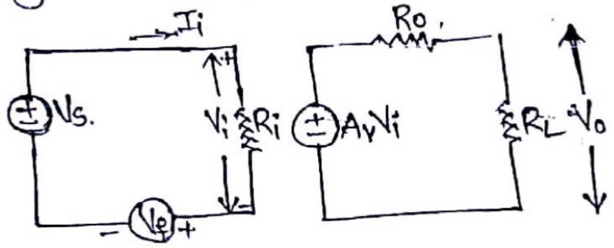
If the output signal is current feedback increases the output impedance by $(1+A\beta)$ times.

Ex:- current series, current shunt.

21/01/19
Monday

Effect of negative feedback on Input impedance :-

Voltage series feedback:-



The above figure a voltage amplifier with voltage series feedback.

when there is no feedback $V_s = V_i + V_f$ But $V_f = 0$.

$$\therefore V_s = V_i$$

Input impedance without feedback.

$$R_i = \frac{V_s}{I_i} = \frac{V_i}{I_i};$$

After feedback is introduced $V_s = V_i + V_f$.

$$\text{But } V_f = \beta V_o.$$

(V_f depends on output voltage V_o. $\therefore V_f \propto V_o$, $V_f = \beta V_o$.)

$$= \beta \frac{A_V V_i R_L}{R_o + R_L}$$

$$V_f = \beta \frac{A_V R_L}{R_o + R_L} \cdot V_i.$$

$\therefore V_f = \beta AV \cdot V_i$

$V_s = V_i + \beta AV \cdot V_i$

$V_s = V_i(1 + AV\beta)$

Input impedance with feedback is.

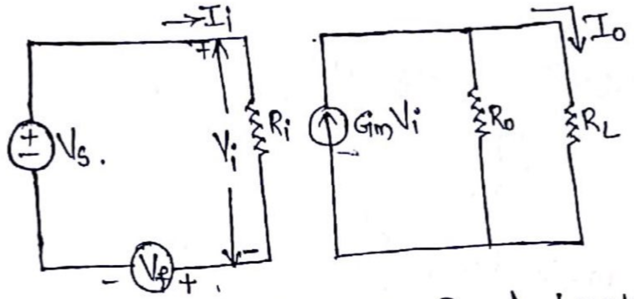
$$R_{if} = \frac{V_s}{I_i} = \frac{V_i(1+AV\beta)}{I_i} = R_i(1+AV\beta)$$

$$\boxed{\therefore R_{if} = R_i(1+AV\beta)}$$

Hence Input impedance increases by $(1+AV\beta)$ times of R_i

22/1/19
Tuesday.

Current series feedback:-



Input voltage with out feed back is $V_s = V_i$

Input impedance with out feed back.

$$R_i = \frac{V_s}{I_i} = \frac{V_i}{I_i}$$

After the feedback is introduced:

$$V_s = V_i + V_f.$$

Here $V_f = \beta I_o$.

$\therefore V_s = V_i + \beta I_o$.

$$V_s = V_i + \beta \frac{G_m V_i R_o}{R_o + R_L}.$$

$$V_s = V_i\left[1 + \beta \frac{G_m R_o}{R_o + R_L}\right]$$

$$V_s = V_i\left[1 + G_M \beta\right]$$

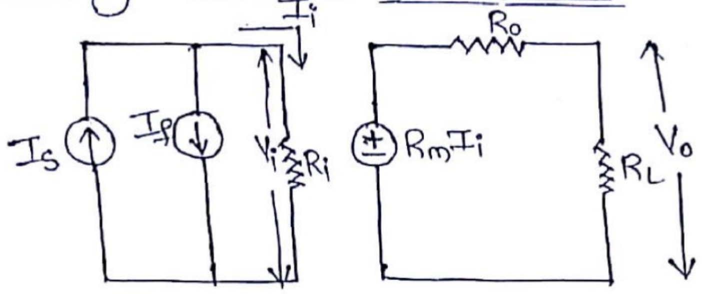
Input impedance with feedback is.

$$R_{if} = \frac{V_s}{I_i} = \frac{V_i\left[1+G_M\beta\right]}{I_i}$$

$$\boxed{R_{if} = R_i\left[1+G_M\beta\right].}$$

Hence input impedance increase by $(1+G_M\beta)$ times of R_i

Voltage shunt feedback :-



when there is no feed back $I_s = I_i$

Input impedance without feed back is

$$R_i = \frac{V_i}{I_s} = \frac{V_i}{I_i}$$

After feedback is introduced.

$$I_s = I_f + I_i$$

But $I_f = \beta V_o$.

$$I_s = \beta V_o + I_i$$

$$I_s = \beta\frac{R_m I_i R_L}{R_o+R_L} + I_i$$

$$I_s = I_i\left[\beta\frac{R_m R_L}{R_o+R_L} + 1\right]$$

$$I_s = I_i\left(1+\beta R_M\right)$$

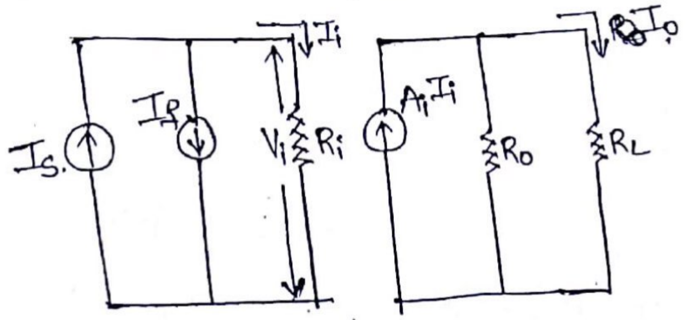
Input impedance with feedback is.

$$R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i(1+\beta R_M)} = \frac{R_i}{(1+\beta R_M)}$$

$$\therefore \boxed{R_{if} = \frac{R_i}{(1+\beta R_M)}}$$

Hence input Impedance decrease by $(1+\beta R_M)$ times of R_i

Current shunt feedback:



when there is no feed back $I_s = I_i$

Input impedance without feed back is

$$R_i = \frac{V_i}{I_s} = \frac{V_i}{I_i}.$$

After feed back is introduced.

$$I_s = I_f + I_i$$

But $I_f = \beta I_o$.

$$I_s = \beta I_o + I_i$$

$$I_s = \beta \frac{A_i I_i R_o}{R_o + R_L} + I_i$$

$$I_s = I_i \left[\beta \frac{A_i R_o}{R_o + R_L} + 1\right]$$

$$I_s = I_i \left[1 + \beta A_I\right]$$

Input impedance with feedback is.

$$R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i(1+\beta A_I)} = \frac{R_i}{(1+\beta A_I)}$$

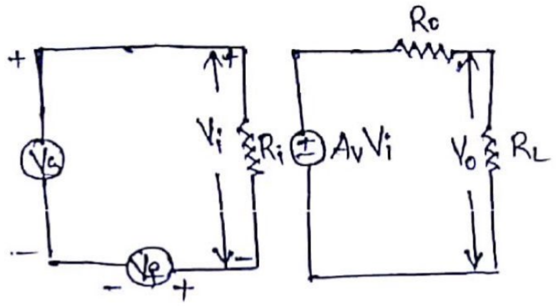
$$\therefore \boxed{R_{if} = \frac{R_i}{(1+\beta A_I)}}$$

Hence input impedance decreases by $(1+\beta A_I)$ times of R_i

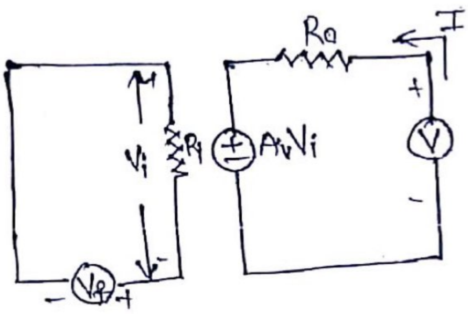
Thursday.

Effect of negative feedback on output Impedance :-

Voltage series feedback:-



To determine output impedance set $V_s = 0$ and disconnect R_L, connect voltage source V then the output impedance with feedback is given by



$$R_{of} = \frac{V}{I}$$

KVL at the input loop.

$$V_i + V_f = 0$$

$$V_i = -V_f = -\beta V_o \quad , \text{ But } V_o = V$$

$$\therefore V_i = -V_f = -\beta V$$

KVL at output loop.

$$-V + IR_o + A_V V_i = 0.$$

$$-V + IR_o + A_V(-\beta V) = 0$$

$$V[1 + A_V \beta] = IR_o$$

Output Impedance with feedback is.

$$R_{of} = \frac{V}{I} = \frac{R_o}{(1 + A_V \beta)}$$

$$\therefore R_{of} < R_o$$

output impedance with feedback including R_L

$$R'_{of} = R_{of} \| R_L$$

$$= \frac{R_{of} R_L}{R_{of} + R_L}$$

$$= \frac{\frac{R_o}{1+A_V\beta} \cdot R_L}{\frac{R_o}{1+A_V\beta} + R_L}$$

$$= \frac{R_o R_L}{R_o + R_L + A_V\beta R_L}$$

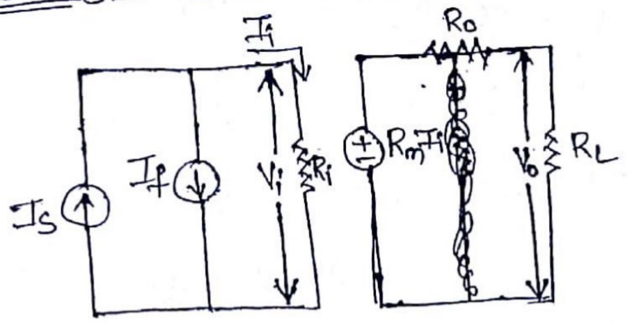
Divide both numerator & denominator with $(R_o + R_L)$

$$= \frac{\frac{R_o R_L}{R_o + R_L}}{1 + \beta\left(\frac{A_V R_L}{R_o + R_L}\right)}$$

$$R'_o = \frac{R_o R_L}{R_o + R_L}.$$

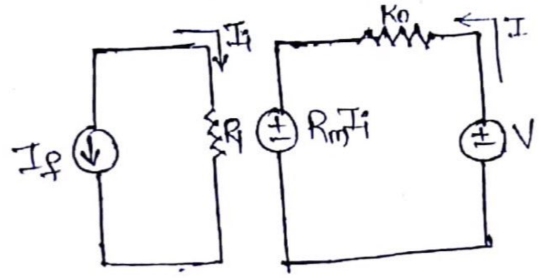
$$\boxed{\therefore R'_{of} = \frac{R'_o}{1+\beta AV}}$$

Voltage shunt feedback :-



To find output impedance with feedback set $I_S = 0$ and disconnect R_L, connect voltage source V. The output impedance with feedback is give by

$$R_{of} = \frac{V}{I}$$



Applying KCL at input loop.

$$I_i + I_f = 0 .$$

$$I_i = -I_f = -\beta V_o \; ; \; V_o = V .$$

$$\therefore I_i = -I_f = -\beta V$$

Applying KVL at Output loop.

$$-V + IR_o + R_m I_i = 0 .$$

$$-V + IR_o + R_m(-\beta V) = 0$$

$$IR_o = V(1+\beta R_m)$$

$\therefore$ output impedance with feedback is

$$R_{of} = \frac{V}{I} = \frac{R_o}{1+\beta R_m}$$

$$R_{of} < R_o .$$

Output impedance with feedback including R_L

$$R'_{of} = R_{of} \| R_L$$

$$= \frac{R_{of} R_L}{R_{of} + R_L}$$

$$= \frac{\frac{R_o}{1+\beta R_m} \cdot R_L}{\frac{R_o}{1+\beta R_m} + R_L}$$

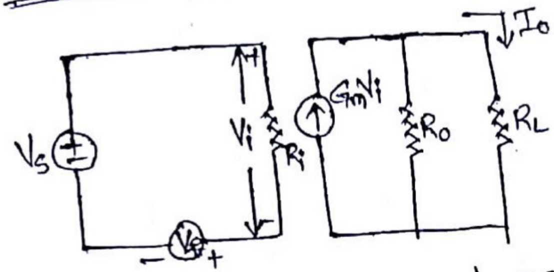
$$= \frac{R_o R_L}{R_o + R_L + \beta R_m R_L}$$

$$R'_{of} = \frac{\frac{R_o R_L}{R_o + R_L}}{1 + \beta\left(\frac{R_m R_L}{R_o + R_L}\right)}$$

$$\boxed{R'_{of} = \frac{R'_o}{1 + \beta R_M}}$$

$\therefore R'_{of} < R'_o$

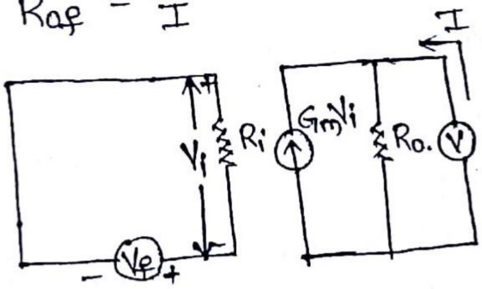
Current series feed back :-



To find output impedance set $V_s = 0$ and disconnect R_L and connect voltage source V.

The output impedance with feed back is given by

$$R_{of} = \frac{V}{I}$$



Apply KVL to the input loop.

$$V_i + V_f = 0.$$

$$V_i = -V_f = -\beta I_o. \quad (\text{But } I_o = -I)$$

$$V_i = -V_f = \beta I$$

Apply KCL at output loop.

$$G_m V_i + I = \frac{V}{R_o}.$$

$$G_m(\beta I) + I = \frac{V}{R_o}.$$

$$I(1+\beta G_m) = \frac{V}{R_o}.$$

$\therefore$ output impedance with feedback is.

$$R_{of} = \frac{V}{I} = R_o(1+\beta G_m)$$

$$R_{of} > R_o.$$

Output impedance with feedback including R_L

$$R'_{of} = R_{of} \parallel R_L$$

$$= \frac{R_{of} \cdot R_L}{R_{of}+R_L}$$

$$= \frac{R_o(1+\beta G_m)\, R_L}{R_o(1+\beta G_m) + R_L}$$

$$= \frac{R_o R_L (1+G_m\beta)}{R_o + R_L + R_o G_m \beta}$$

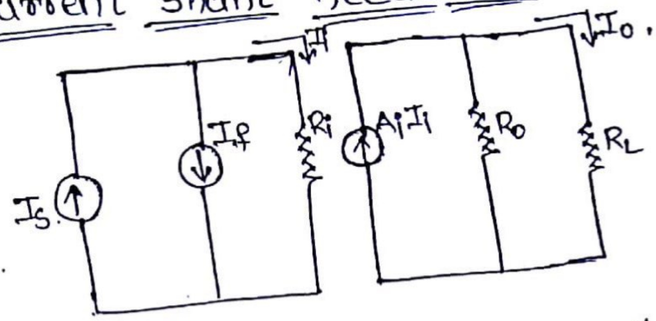
$$= \frac{\frac{R_o R_L}{R_o+R_L}(1+G_m\beta)}{1+\beta\left(\frac{G_m R_o}{R_o+R_L}\right)}$$

$$\boxed{R'_{of} = \frac{R'_o(1+G_m\beta)}{1+G_M\beta}}$$

$$G_M < G_m$$

$$\therefore R'_{of} > R'_o$$

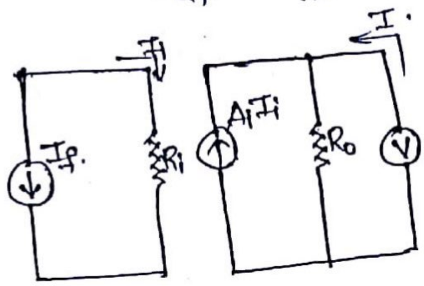
Current shunt feedback:-



To find output impedance set $I_s=0$ and disconnect R_L and connect voltage source V.

The output impedance with feedback is given by

$$R_{of} = \frac{V}{I}.$$



Apply KCL at input loop.

$$I_i + I_f = 0.$$

$$I_i = -I_f = -\beta I_o \quad ; \text{ But } (I_o = -I)$$

$$I_i = -I_f = \beta I.$$

Apply KCL at output loop.

$$A_iI_i + I = \frac{V}{R_o}.$$

$$A_i(\beta I) + I = \frac{V}{R_o}.$$

$$I(1+\beta A_i) = \frac{V}{R_o}.$$

output impedance with feedback is

$$R_{of} = \frac{V}{I} = R_o(1+\beta A_i)$$

$$R_{of} > R_o$$

Output impedance with feedback including R_L

$$R'_{of} = R_{of} \parallel R_L$$

$$= \frac{R_{of} \cdot R_L}{R_{of} + R_L}$$

$$= \frac{R_o(1+\beta A_i)R_L}{R_o(1+\beta A_i) + R_L}$$

$$= \frac{R_o R_L (1+\beta A_i)}{R_o + R_L + \beta A_i R_o}$$

$$= \frac{\frac{R_o R_L}{R_o + R_L}(1+\beta A_i)}{1+\beta A_I} \qquad A_I = \frac{A_i R_o}{R_o + R_L}$$

$$\boxed{R'_{of} = \frac{R'_o(1+\beta A_i)}{1+\beta A_I}}$$

$A_I < A_i$

$\therefore R'_{of} > R'_o$

Parameter	voltage series	Current series	voltage shunt	Current shunt
Transfer gain (A)	$AV_f = \frac{A_V}{1+A_V\beta}$	$G_{Mf} = \frac{G_M}{1+G_m\beta}$	$R_{Mf} = \frac{R_M}{1+R_m\beta}$	$A_{If} = \frac{A_I}{1+A_I\beta}$
Bandwidth	$BW_f = BW(1+A_V\beta)$ Increases	$BW_f = BW(1+G_m\beta)$ Increases	$BW_f = BW(1+R_m\beta)$ Increases	$BW_f = BW(1+A_I\beta)$ Increases
Distortion.	$D_2 = \frac{D_2}{1+A_V\beta}$ Decrease..	$D_2 = \frac{D_2}{1+G_m\beta}$ Decreases	$D_2 = \frac{D_2}{1+R_M\beta}$ Decreases	$D_2 = \frac{D_2}{1+A_I\beta}$ decreases.
Noise	$N_f = \frac{N}{1+A_V\beta}$ Decreases	$N_f = \frac{N}{1+G_M\beta}$ Decreases.	$N_f = \frac{N}{1+R_m\beta}$ Decreases	$N_f = \frac{N}{1+A_I\beta}$ Decreases
Input Impedance	$R_{if} = R_i(1+A_V\beta)$ Increase	$R_{if} = R_i(1+G_M\beta)$ Increase	$R_{if} = \frac{R_i}{1+R_M\beta}$ Decrease	$R_{if} = \frac{R_i}{1+A_I\beta}$. Decrease.

Parameter	Voltage series	Current series.	Voltage shunt	Current shunt
output Impedance.	$R_{of} = \frac{R_o}{1+A_v\beta}$ $R'_{of} = \frac{R'_o}{1+A_V\beta}$	$R_{of} = R_o(1+\beta G_m)$ $R'_{of} = \frac{R'_o(1+G_m\beta)}{(1+G_M\beta)}$	$R_{of} = \frac{R_o}{1+\beta R_m}$ $R'_{of} = \frac{R'_o}{1+\beta R_M}$	$R_{of} = R_o(1+\beta$ [illegible] $R'_{of} = \frac{R'_o(1+\beta A'_{[illegible]})}{1+\beta A_{[illegible]}}$
	Decrease	Increase.	Decrease	Increase

* Define sensitivity. & Derive an expression for the desensitivity factor in an amplifier.

A. The ratio of the percentage change in gain with feedback to the percentage change in gain without feedback.

$$\frac{dA_f}{A_f} \times 100 = \frac{\frac{dA}{A} \times 100}{1+A\beta}.$$

$$\frac{\frac{dA_f}{A_f} \times 100}{\frac{dA}{A} \times 100} = \frac{1}{1+A\beta}$$

$$\boxed{\frac{\left(\frac{dA_f}{A_f}\right)}{\left(\frac{dA}{A}\right)} = \frac{1}{1+A\beta}} = \text{sensitivity}.$$

$$\text{Desensitivity factor } D = \frac{1}{\text{sensitivity}} = \frac{1}{\left(\frac{1}{1+A\beta}\right)} \Rightarrow \boxed{D = 1 + A\beta}$$

28/01/19
Monday.

Fundamental Assumptions of feedback amplifier.

In a negative feedback amplifer $A_f = \frac{A}{1+A\beta}$ This equation is true only if 3 conditions are statisfied.

1. The input signal is transmitted to the output through Amplifier A but not through the β network

$\therefore$ β network must be unilateral. so that it does not transmit a signal from input to output

2. The feedback signal must be transmitted from output to input through the β network, But not through the amplifier.
In other words the basic amplifier must be unilateral from input too output and does not transmitt any signal form output to input.

3. The β-factor of the feed network should be independent of the source & load resistance i.e. R_L, R_s.

Method of Analysis of a Feedback Amplifier :-

steps :-

1. Identify the topology :- Is the feedback signal X_f a voltage or current, In other words, the X_f is applied in series or shunt. Is the sampled signal X_o a voltage or current
In other words is the sampled signal taken from output node or output loop.

2. Draw the basic amplifier circuit without feedback but taking the loading of the feedback into account.

3. Use thevinin's source if X_f is voltage. Use norton's source if X_f is current.

4. Replace each active device by the proper model.
(Ex: h-parameter model or hybrid π model)..

5. Evaluate feedback factor $\beta = \frac{X_f}{X_o}$.

6. Evaluate A by applying KVL, KCL to the circuit obtained in step-4

7. From the values of A & β. Determine A_f, R_{of}, R'_{of}, C

In step-2 the basic amplifier circuit without feedback but taking the effect of feedback in to account is obtain by ~~foll~~ applying the following steps.

To find the input circuit

i) set $V_o = 0$ for voltage sampling, In other words short circuit the output.

ii) set $I_o = 0$ for current ~~shunt~~ sampling, In other words open circuit the output.

To find the output circuit

i) set $V_i = 0$ for shunt ~~series~~ comparison, In other words short circuit the input node.

ii) set $I_i = 0$ for series comparison, In other words open circuit the ~~output~~ input node.

07/08/19
Thursday

4. OSCILLATORS

Oscillator:-

It is an electronic circuit which generates a periodic signal (AC Signal) at the output without an external AC input signal.

To generate the periodic signal circuit is supplied with energy from a DC source.

Classification of oscillators:-

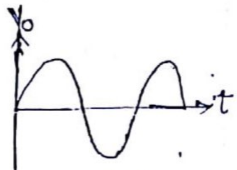
1) According to the shape of the output wave form generated, oscillators can be classified as

a) Sinusoidal oscillators.

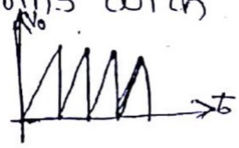
b) Non-sinusoidal & Relaxation oscillators.

c) Astable Multi vibrator.

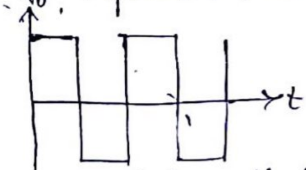
Sinusoidal oscillators produce sinusoidal signals.



Relaxation oscillators produce non-sinusoidal wave form's with abtrupt discontinities.



Astable Multi vibrator produces square or rectangular wave forms.



2) According to the frequency of the output signal, oscillators can be classified as

a) Audio frequency oscillator (upto 20KHz)

b) Radio frequency oscillators (20 KHz to 30 MHz)

c) Very high frequency oscillators (30MHz to 300MHz)

d) Ultra high frequency oscillators (300MHz to 3GHz)

e) Micro wave frequency oscillators (above 3 GHz)

3) According to the fundamental mechanism oscillators are classified into

a) Feed back oscillators.

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b) Negative resistance oscillator.

Feedback oscillators use positive feedback to produce oscillation and they satisfy the Bark hausen criteria.

Negative resistance oscillators use negative resistance device such as Tund diode, UJT, SCR.

4) According to the type of components used, oscillators are classified in to ~~RC PA~~.

a) RC phase-shift oscillator.

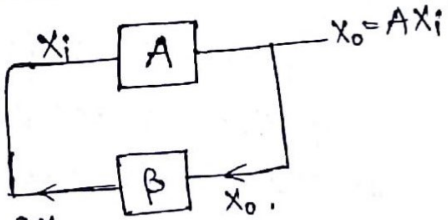
b) Wien-bridge oscillator

c) LC-oscillator.

d) Crystal oscillator.

14/02/19
Thursday.

~~Principle of~~
Working principle of oscillator, Bark-hausen criteria :-



$X_f = \beta X_o$.
$= \beta A X_i$

In the above figure the closed loop system is shown. If the loop gain $A\beta = 1$ then $X_f = X_i$ and the circuit continues to produce X_o without any external input.

Bark hausen criteria:-

The necessary and sufficient condition for a closed loop system to produce oscillations continuously is $A\beta = 1$.

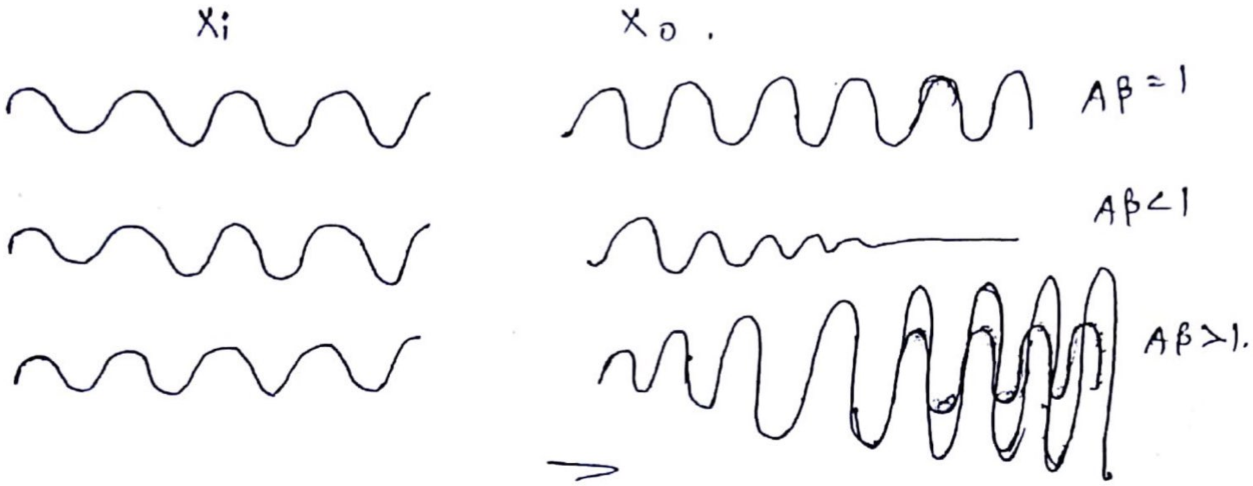
$\Rightarrow |A\beta| = 1$; $\angle A\beta = 0$ & integral multiple of 2π.

This condition is called Bark hausen Criteria.

$|A\beta| = 1$ is called magnitude criteria.

$\angle A\beta = 0$ & 2π is called Angle criteria.

If $A\beta$ is less than 1, the amplitude of the ~~output~~ input signal continuous to decrease and oscillations will stop eventually.
If ($A\beta > 1$) $A\beta$ is greater than 1, the input signal increases continuously and the hence the amplitude of the ~~output~~ output signal diverges and tends to infinity (∞)



<u>Practical considerations of Bark hausen criteria</u>:-
The required condition for oscillation is $A\beta = 1$, according to Bark hausen criteria. But due to aging & other reasons the again of the amplifier A may fall and this may lead to $A\beta < 1$ and oscillations will eventually stop.
∴ For all practical purposes $A\beta$ is always set greater than 1.
Theoretically if $A\beta > 1$ output X_o should increase continu-ously and become infinity. But practically as input signal increases, the amplifier moves into non-linear region reducing the gain of the amplifier.
∴ Output signal cannot increase beyond a certain level.
<u>Note</u>:- oscillator will have positive feed back.

RC oscillators :-

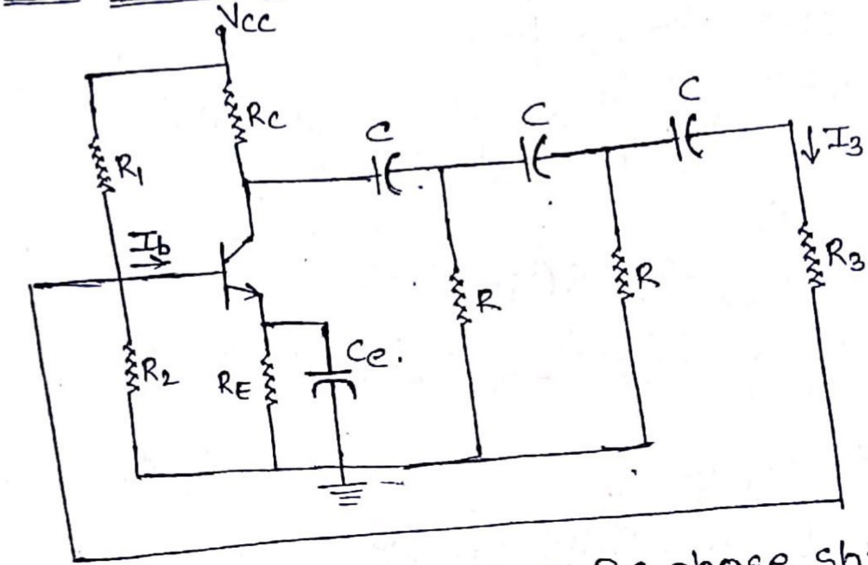
Various types of RC oscillators are.

1) BJT RC phase shift oscillator.

2) FET RC phase shift oscillator.

3) Wein bridge oscillator.

* BJT RC phase shift oscillator :-



The above figure shows a BJT RC phase shift oscillator and it produces BJT is connected in CE configuration 180° of phase shift.

The phase shift provided by each RC network is given by $\tan^{-1}\left[\frac{-1}{\omega RC}\right]$.

By properly selecting values of R & C, the phase shift provided by each RC network is set to 60°.

∴ 3-RC networks together produces 180° phase shift.

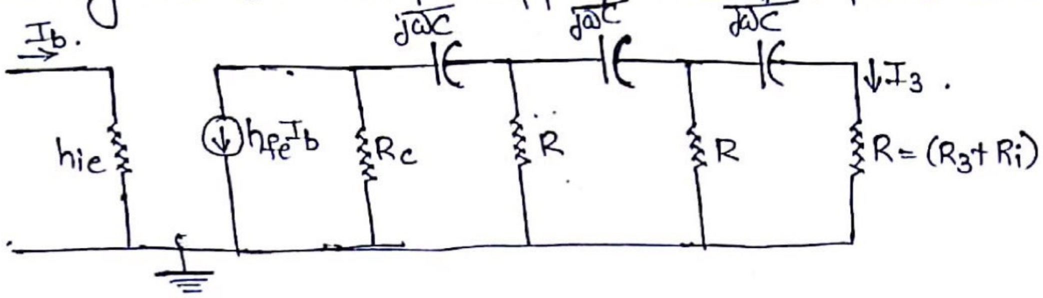
The total phase shift around the loop is 360°.

∴ The angle criteria of the oscillator is satisfied.

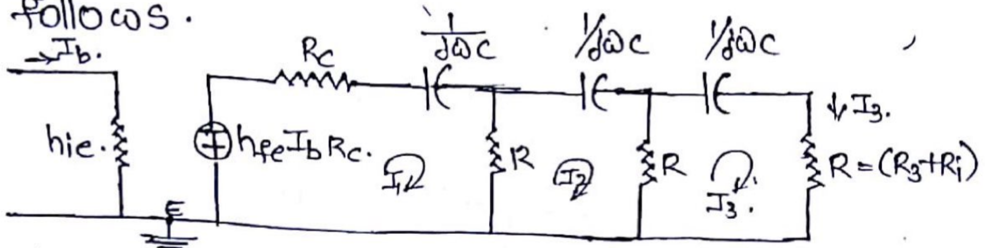
The value of R_3 selected such that $\boxed{R_3 + R_i = R}$

where R_i is the input impedance of BJT.

Replacing the BJT with approximate h-parameter model.



using source transformation, the above circuit can be modified as follows.



Apply KVL for loop ①

$$h_{fe}I_bR_c + R_cI_1 + \frac{1}{j\omega C}\cdot I_1 + R(I_1-I_2) = 0.$$

$$I_1\left(R_c+R+\frac{1}{j\omega C}\right) - I_2R = -h_{fe}I_bR_c \longrightarrow ①.$$

Apply KVL for loop ②

$$R(I_2-I_1) + \frac{1}{j\omega C}I_2 + R(I_2-I_3) = 0.$$

$$I_2\left(R+R+\frac{1}{j\omega C}\right) - I_1R - I_3R = 0.$$

$$I_2\left(2R+\frac{1}{j\omega C}\right) - R(I_1-I_3) = 0 \longrightarrow ②$$

Apply KVL for loop ③

$$R(I_3-I_2) + \frac{1}{j\omega C}\cdot I_3 + I_3R = 0.$$

$$I_3\left(2R+\frac{1}{j\omega C}\right) - I_2R = 0 \longrightarrow ③$$

$$I_1\left(R_c+R+\frac{1}{j\omega C}\right) - I_2R = -h_{fe}I_bR_c \longrightarrow ①$$

$$-I_1R + I_2\left(2R+\frac{1}{j\omega C}\right) - I_3R = 0 \longrightarrow ②$$

$$-I_2R + I_3\left(2R+\frac{1}{j\omega C}\right) = 0 \longrightarrow ③$$

Today.
The above three equations can be solved by using cramer rule.

$$I_3 = \frac{\Delta_3}{\Delta}$$

$$\Delta = \begin{vmatrix} R_c + R + \frac{1}{j\omega C} & -R & 0 \\ -R & 2R + \frac{1}{j\omega C} & -R \\ 0 & -R & 2R + \frac{1}{j\omega C} \end{vmatrix}$$

$$= \left(R_c + R + \frac{1}{j\omega C}\right)\left[\left(2R + \frac{1}{j\omega C}\right)\left(2R + \frac{1}{j\omega C}\right) - R^2\right] + R\left(-2R^2 - \frac{R}{j\omega C} - 0\right)$$
$$+ 0\,(R^2 - 0).$$

$$= \left(R_c + R + \frac{1}{j\omega C}\right)\left[3R^2 + \frac{2R}{j\omega C} + \frac{2R}{j\omega C} + \frac{1}{(j\omega C)^2}\right] - 2R^3 - \frac{R^2}{j\omega C}.$$

$$= \left(R_c + R + \frac{1}{j\omega C}\right)\left[3R^2 + \frac{4R}{j\omega C} - \frac{1}{(\omega C)^2}\right] - 2R^3 - \frac{R^2}{j\omega C}.$$

$$= R_c\left(3R^2 + \frac{4R}{j\omega C} - \frac{1}{(\omega C)^2}\right) + R\left(3R^2 + \frac{4R}{j\omega C} - \frac{1}{(\omega C)^2}\right) + \frac{3R^2}{j\omega C} + \frac{4R}{(j\omega C)^2}$$
$$- \frac{1}{j(\omega C)^3} - 2R^3 - \frac{R^2}{j\omega C}.$$

$$= 3R_cR^2 + \frac{4R_cR}{j\omega C} - \frac{R_c}{(\omega C)^2} + 3R^3 + \frac{4R^2}{j\omega C} - \frac{R}{(\omega C)^2} + \frac{3R^2}{j\omega C} - \frac{4R}{(\omega C)^2}.$$
$$- \frac{1}{j(\omega C)^3} - 2R^3 - \frac{R^2}{j\omega C}.$$

$$= R^3 + 3R_cR^2 - \frac{5R}{\omega^2 C^2} - \frac{R_c}{\omega^2 C^2} + \frac{1}{j}\left[\frac{4RR_c}{\omega C} + \frac{6R^2}{\omega C} - \frac{1}{\omega^3 C^3}\right]$$

$$= R^3\left[1 + \frac{3R_cR^2}{R^3} - \frac{5R}{R^3\omega^2 C^2} - \frac{R_c}{\omega^2 C^2 R^3} - j\left(\frac{4RR_c}{R^3\omega C} + \frac{6R^2}{R^3\omega C} - \frac{1}{R^3\omega^3 C^3}\right)\right]$$

$$\Delta = R^3\left[\left(1 + 3\left(\frac{R_c}{R}\right) - \frac{5}{(\omega RC)^2} - \frac{R_c}{R(\omega RC)^2}\right) - j\left(\frac{4R_c}{R}\cdot\frac{1}{\omega RC} + \frac{6}{\omega RC} - \frac{1}{(\omega RC)^3}\right.\right.$$

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Let $\frac{R_c}{R} = K$, $\frac{1}{\omega RC} = \alpha$.

$$\Delta = R^3\left[\left(1+3K-5\alpha^2-K\alpha^2\right) - j\left(4K\alpha+6\alpha-\alpha^3\right)\right]$$

$$\Delta = R^3\left[\left(1+3K-(5+K)\alpha^2\right) - j\left((6+4K)\alpha-\alpha^3\right)\right]$$

$$\Delta_3 = \begin{vmatrix} R_c+R+\frac{1}{j\omega C} & -R & -h_{fe}I_bR_c \\ -R & 2R+\frac{1}{j\omega C} & 0 \\ 0 & -R & 0 \end{vmatrix}$$

$$= \left(R_c+R+\frac{1}{j\omega C}\right)(0-0) + R(0-0) - h_{fe}I_bR_c\left(R^2-0\right)$$

$$= 0+0-h_{fe}I_bR_cR^2.$$

$$\Delta_3 = -h_{fe}I_bR_cR^2$$

$$I_3 = \frac{\Delta_3}{\Delta} = \frac{-h_{fe}I_bR_cR^2.}{R^3\left[\left(1+3K-(5+K)\alpha^2\right)-j\left((6+4K)\alpha-\alpha^3\right)\right]}$$

Loop gain $A\beta = \frac{I_3}{I_b} = \frac{-h_{fe}R_c.}{R\left[\left(1+3K-(5+K)\alpha^2\right)-j\left((6+4K)\alpha-\alpha^3\right)\right]}$

$$= \frac{-h_{fe}\cdot\left(\frac{R_c}{R}\right)}{\left(1+3K-(5+K)\alpha^2\right)-j\left((6+4K)\alpha-\alpha^3\right)}$$

$$\boxed{A\beta = \frac{I_3}{I_b} = \frac{-h_{fe}K.}{\left(1+3K-(5+K)\alpha^2\right)-j\left((6+4K)\alpha-\alpha^3\right)}}$$

To satisfy Bark hausen criteria $|A\beta|=1$, $\angle A\beta = 0$

$\therefore$ Imaginary part in the above equation must be equal to zero.

$$(6+4K)\alpha - \alpha^3 = 0.$$

$$(6+4K) - \alpha^2 = 0.$$

$$\alpha^2 = 6+4K$$

$$\alpha = \sqrt{6+4K}$$

$$\frac{1}{\omega RC} = \sqrt{6+4K}.$$

$$\frac{1}{2\pi f RC} = \sqrt{6+4K}.$$

$$\boxed{f = \frac{1}{2\pi RC\sqrt{6+4K}}.}$$

Condition for sustained oscillations:-

The condition for sustained oscillation is $A\beta > 1$

$$A\beta = \frac{-h_{fe}K.}{(1+3K-(5+K)\alpha^2)} > 1$$

But $\alpha^2 = 6+4K.$

$$= \frac{-h_{fe}K.}{(1+3K-(5+K)(6+4K))} > 1$$

$$= \frac{-h_{fe}K.}{1+3K-30-20K-6K-4K^2} > 1$$

$$= \frac{-h_{fe}K.}{-29-23K-4K^2} > 1$$

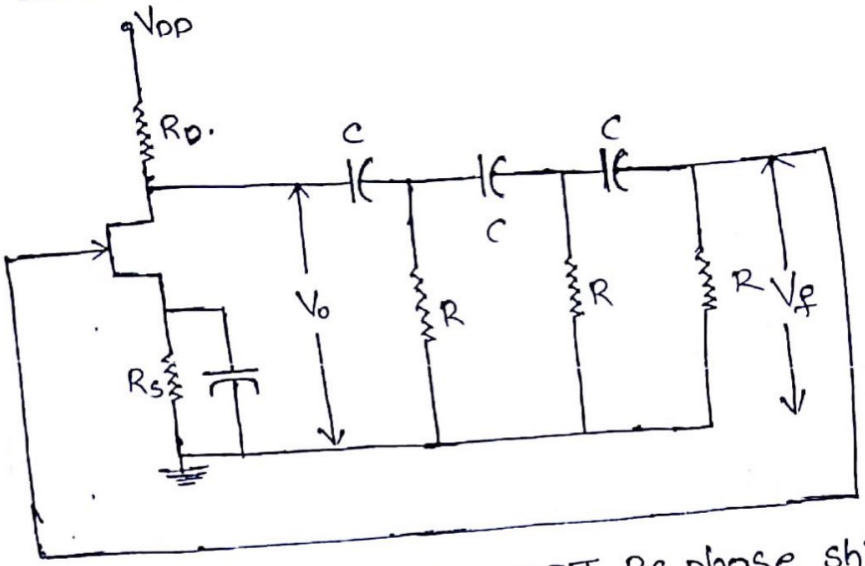
$$= \frac{h_{fe}K.}{4K^2+23K+29} > 1.$$

$$h_{fe}K > 4K^2+23K+29$$

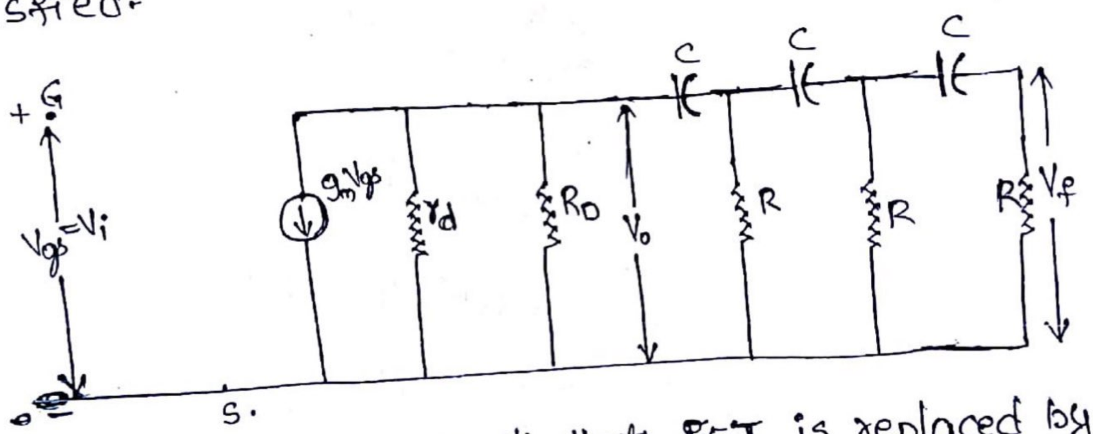
$$h_{fe} > 4K+23+\frac{29}{K}$$

$$\boxed{h_{fe} > 4K+\frac{29}{K}+23.}$$

FET RC phase shift oscillator :-



The above figure represents FET RC phase shift oscillator.
FET is in common source configuration.
$\therefore$ It produces 180° phase shift.
Each RC network produces a phase shift of $\tan^{-1}\left(\frac{-1}{\omega RC}\right)$
The values of R, C are adjusted such that the phase shift produced by each RC network is 60°.
$\therefore$ 3 RC networks together will produces 180° phase shift.
$\therefore$ The total phase shift around the loop is 360°.
The Angle condition of the Bark-hausen criteria is satisfied.

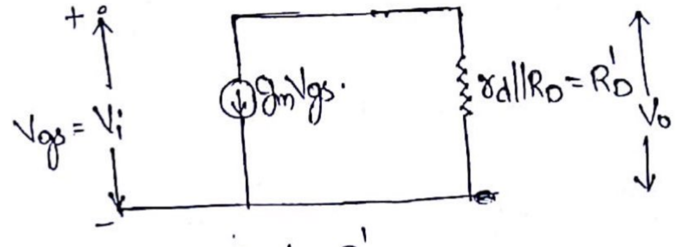


The above figure represents that FET is replaced by a approximated h-parameter model.

loop gain $= A\beta$;

$$A = \frac{V_o}{V_i} \; ; \; \beta = \frac{V_f}{V_o}$$

The gain A can be derived from the following circuit.

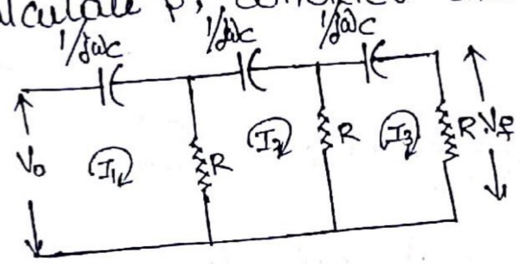


$$V_o = -g_m V_{gs} \cdot R_D'$$

$$V_o = -g_m R_D' V_i$$

$$\frac{V_o}{V_i} = A = -g_m R_D'$$

To calculate β; consider the following circuit.



Applying KVL to 1st loop.

$$\frac{I_1}{j\omega C} + R(I_1 - I_2) - V_o = 0.$$

$$I_1\left(R + \frac{1}{j\omega C}\right) - I_2 R = V_o. \longrightarrow (1)$$

Apply KVL to 2nd loop.

$$\frac{I_2}{j\omega C} + (I_2 - I_3)R + R(I_2 - I_1) = 0.$$

$$-I_1 R + I_2\left(2R + \frac{1}{j\omega C}\right) - I_3 R = 0. \longrightarrow (2)$$

Apply KVL to 3rd loop.

$$\frac{I_3}{j\omega C} + I_3 R + (I_3 - I_2)R = 0.$$

$$-I_2 R + I_3\left(2R + \frac{1}{j\omega C}\right) = 0 \longrightarrow (3)$$

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Monday.

The above 3 equations can be solved by using crammer's Rule.

$$I_3 = \frac{\Delta_3}{\Delta}.$$

$$\Delta = \begin{vmatrix} R+\frac{1}{j\omega C} & -R & 0 \\ -R & 2R+\frac{1}{j\omega C} & -R \\ 0 & -R & 2R+\frac{1}{j\omega C} \end{vmatrix}$$

$$\Delta = \left(R+\frac{1}{j\omega C}\right)\left[4R^2+\frac{2R}{j\omega C}+\frac{2R}{j\omega C}-\frac{1}{(\omega C)^2}-R^2\right] - (-R)\left(-2R^2-\frac{R}{j\omega C}\right)$$
$$+0(0-0)$$

$$=\left(R+\frac{1}{j\omega C}\right)\left[3R^2+\frac{4R}{j\omega C}-\frac{1}{(\omega C)^2}\right]+R\left(-2R^2-\frac{R}{j\omega C}\right).$$

$$=\left(R+\frac{1}{j\omega C}\right)\left[3R^2+\frac{4R}{j\omega C}-\frac{1}{\omega^2C^2}\right]-2R^3-\frac{R^2}{j\omega C}.$$

$$=3R^3+\frac{4R^2}{j\omega C}-\frac{R}{\omega^2C^2}+\frac{3R^2}{j\omega C}-\frac{4R}{\omega^2C^2}-\frac{1}{j\omega^3C^3}-2R^3-\frac{R^2}{j\omega C}.$$

$$=R^3+\frac{6R^2}{j\omega C}-\frac{5R}{\omega^2C^2}-\frac{1}{j\omega^3C^3}.$$

$$\Delta = R^3-\frac{5R}{\omega^2C^2}+\frac{1}{j}\left(\frac{6R^2}{\omega C}-\frac{1}{\omega^3C^3}\right)$$

$$\Delta = R^3\left[1-\frac{5}{R^2\omega^2C^2}+\frac{1}{j}\left(\frac{6}{\omega RC}-\frac{1}{R^3\omega^3C^3}\right)\right].$$

Let $\alpha = \frac{1}{\omega RC}$

$$\Delta = R^3\left[1-5\alpha^2+\frac{1}{j}(6\alpha-\alpha^3)\right].$$

$$\Delta_3 = \begin{vmatrix} R+\frac{1}{j\omega C} & -R & V_0 \\ -R & 2R+\frac{1}{j\omega C} & 0 \\ 0 & -R & 0 \end{vmatrix}$$

$$= \left(R+\frac{1}{j\omega C}\right)(0-0) - (-R)(0-0) + V_0\left(R^2 - \cancel{\left(2R+\frac{1}{j\omega C}\right)}\right)$$

$$\Delta_3 = 0 - 0 + V_0(R^2 \cancel{2R+\frac{1}{j\omega C}})$$

$$\Delta_3 = V_0 R^2$$

$$I_3 = \frac{\Delta_3}{\Delta} = \frac{V_0 \cancel{R^2}}{R^{\cancel{3}}\left[1-5\alpha^2+\frac{1}{j}(6\alpha-\alpha^3)\right]}$$

$$= \frac{V_0}{R\left[1-5\alpha^2-j(6\alpha-\alpha^3)\right]}$$

$$V_f = I_3 \cdot R = \frac{V_0 \cdot \cancel{R}}{\cancel{R}\left[1-5\alpha^2-j(6\alpha-\alpha^3)\right]}$$

$$V_f = \frac{V_0}{\left[1-5\alpha^2-j(6\alpha-\alpha^3)\right]}$$

$$\beta = \frac{V_f}{V_0} = \frac{1}{1-5\alpha^2-j(6\alpha-\alpha^3)}$$

$$\therefore A\beta = \frac{-g_m R_D'}{1-5\alpha^2-j(6\alpha-\alpha^3)}$$

To satisfy Bark hausen criterion $|A\beta| = 1$, $\angle A\beta = 0$

$\therefore$ Imaginary part of $A\beta$ should be zero.

$$\therefore -(6\alpha-\alpha^3) = 0$$

$$6-\alpha^2 = 0.$$

$$\alpha^2 = 6.$$

$$\alpha = \sqrt{6}$$

$$\frac{1}{\omega RC} = \sqrt{6}$$

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Scanned with CamScanner

$$\frac{1}{2\pi f RC} = \sqrt{6}.$$

$$\Rightarrow \boxed{f = \frac{1}{2\pi RC\sqrt{6}}}.$$

Condition for sustained oscillations:-

For sustained oscillations, the condition is $A\beta > 1$

$$A\beta = \frac{-g_m R_D'}{1-5\alpha^2} > 1$$

$$= \frac{-g_m R_D'}{1-5(6)} > 1$$

$$= \frac{\cancel{-} g_m R_D'}{\cancel{-}29} > 1$$

$$= \frac{g_m R_D'}{29} > 1$$

$$\therefore \boxed{g_m R_D' > 29}$$

$$|A| > 29$$

$$|A\beta| > 1$$

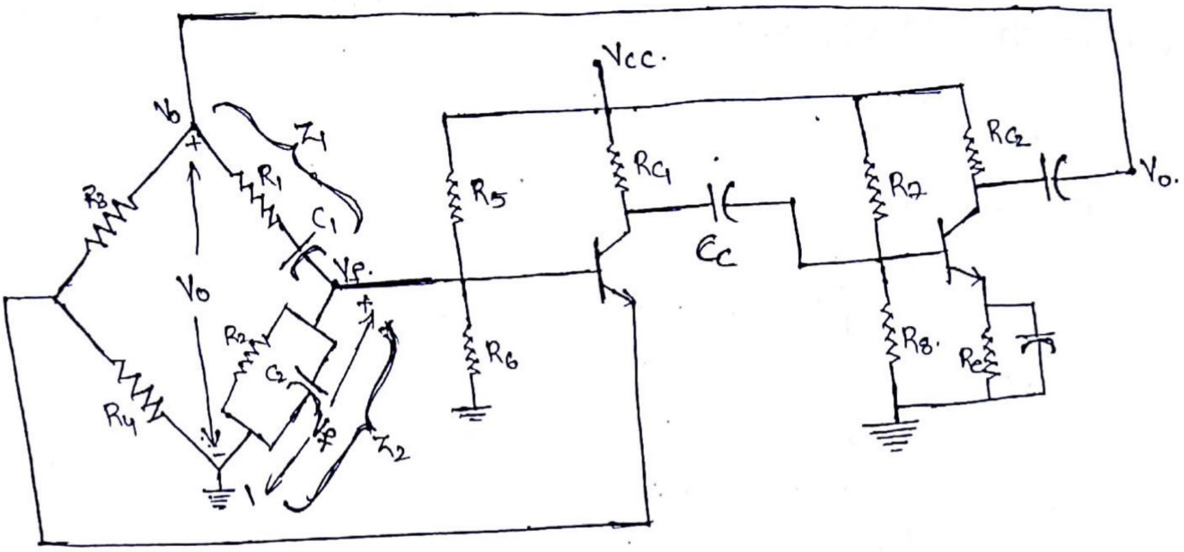
$$|\beta| > \frac{1}{A}.$$

$$|\beta| > \frac{1}{29}.$$

Wein bridge oscillator:-

The below circuit shows wien-bridge oscillator. Each CE Amplifier produces 180° phase shift.

$\therefore$ phase shift provided by both CE Amplifiers together is 360°.



Under Balanced condition the bridge will produce 0° phase shift.

∴ Total phase shift around the loop is 360°, satisfying the Bark hausen angle criteria.

Under Balance condition.

$$\frac{R_3}{R_4} = \frac{Z_1}{Z_2}.$$

$$Z_1 = R_1 + \frac{1}{j\omega C_1} = \frac{1+j\omega R_1C_1}{j\omega C_1}$$

$$Z_2 = R_2 || \frac{1}{j\omega C_2} = \frac{\frac{R_2}{j\omega C_2}}{\frac{1+j\omega R_2C_2}{j\omega C_2}} = \frac{R_2}{1+j\omega R_2C_2}.$$

$$\frac{R_3}{R_4} = \frac{\frac{1+j\omega R_1C_1}{j\omega C_1}}{\frac{R_2}{1+j\omega R_2C_2}} = \frac{(1+j\omega R_1C_1)(1+j\omega R_2C_2)}{j\omega R_2C_1}$$

$$\frac{R_3}{R_4} = \frac{1+j\omega R_2C_2+j\omega R_1C_1 - \omega^2 R_1R_2C_1C_2}{j\omega R_2C_1}.$$

$$\frac{R_3}{R_4} = \frac{1-\omega^2R_1R_2C_1C_2}{j\omega R_2C_1} + \frac{j\omega(R_2C_2+R_1C_1)}{j\omega R_2C_1}$$

$$\frac{R_3}{R_4} = \frac{-j(1-\omega^2R_1R_2C_1C_2)}{\omega R_2C_1} + \frac{R_2C_2+R_1C_1}{R_2C_1}$$

Equating the imaginary parts on bothsides of the equation.

$$0 = -j\left(1-\omega^2 R_1 R_2 C_1 C_2\right).$$

$$\Rightarrow \omega^2 R_1 R_2 C_1 C_2 = 1$$

$$\Rightarrow \omega^2 = \frac{1}{R_1 R_2 C_1 C_2}$$

$$\omega = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$2\pi f = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$f = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

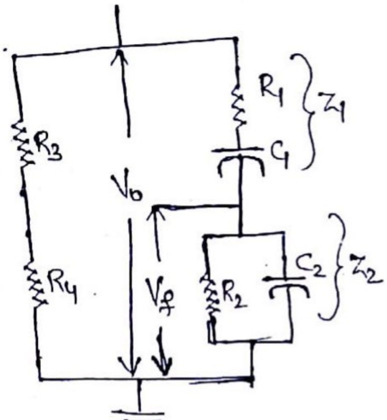
If $R_1 = R_2 = R$.

$C_1 = C_2 = C$.

$$\boxed{f = \frac{1}{2\pi R C}}$$

Condition of sustained oscillations :-

conditions for sustained oscillations is $A\beta > 1$



In above figure $\beta = \frac{V_f}{V_0}$

$$\therefore V_f = \frac{V_0 Z_2}{Z_1 + Z_2}$$

$$\beta = \frac{V_f}{V_0} = \frac{Z_2}{Z_1 + Z_2}$$

$$= \frac{\frac{R_2}{1+j\omega R_2 C_2}}{\frac{1+j\omega R_1 C_1}{j\omega C_1} + \frac{R_2}{1+j\omega R_2 C_2}}$$

$$\beta = \frac{\frac{R_2}{\cancel{(1+j\omega R_2 C_2)}}}{\frac{(1+j\omega R_1 C_1)(1+j\omega R_2 C_2) + j\omega R_2 C_1}{(j\omega C_1)\cancel{(1+j\omega R_2 C_2)}}}$$

$$\beta = \frac{j\omega R_2 C_1}{(1+j\omega R_1 C_1)(1+j\omega R_2 C_2) + j\omega R_2 C_1}$$

$$\beta = \frac{j\omega R_2 C_1}{1 + j\omega R_2 C_2 + j\omega R_1 C_1 - \omega^2 R_1 R_2 C_1 C_2 + j\omega R_2 R_1}$$

Let $R_1 = R_2 = R$

$C_1 = C_2 = C$

$$\beta = \frac{j\omega RC}{(1-\omega^2 R^2 C^2) + j3\omega RC}$$

But $\omega RC = 1$

$$\beta = \frac{j\omega RC}{j3\omega RC} = \frac{1}{3} \Rightarrow \boxed{\beta = 1/3}$$

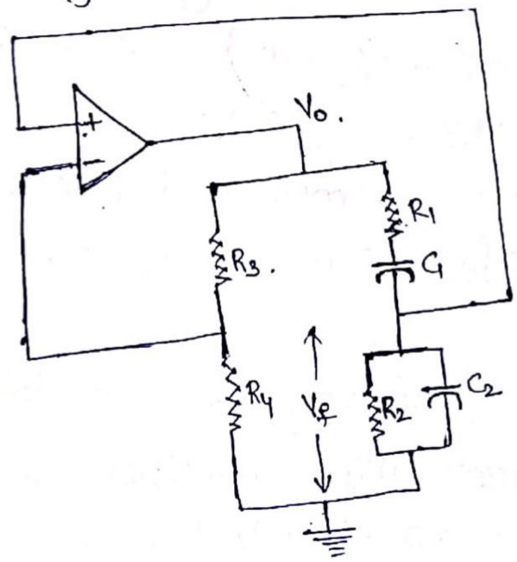
$A\beta > 1$

$A > 1/\beta$

$\boxed{A > 3}$

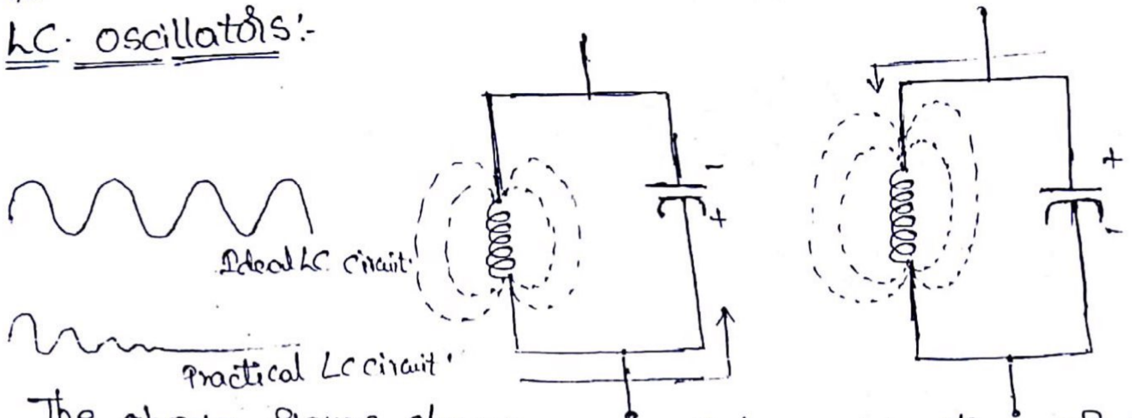
$\therefore$ Minimum gain of the amplifier in wien bridge oscillator is 3.

wein bridge oscillator using ~~OPAM~~ Op-Amp (optional Amplifier):-



3/02/19
Thursday.

LC oscillators:-



The above figure shows a parallel combination of an inductor and capacitor.

Let us assume the capacitor is initially charged and both L, C are ideal. capacitor stores energy in the form of static electric field. since it is connected to inductor. it begins to discharage through inductor.

All the energy stored in the capacitor is transferred to inductor. Inductor stores energy in the form of static magnetic field.

Once the inductor is fully charged, it begins to discharge through the capacitor.

∴ The stored magnetic energy gets converted into static electric energy.

This continues till the inductor is fully discharged.

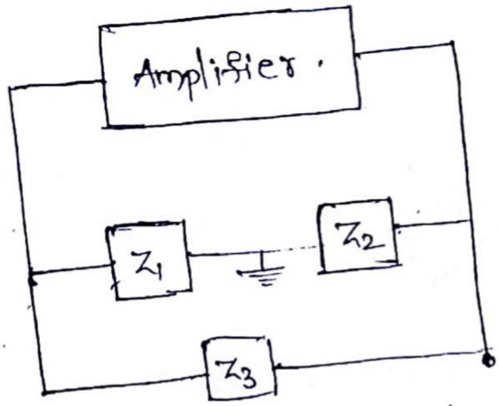
This process repeats indefinitely and produces sinusoidal oscillations across the LC circuit.

But practically there exist a small resistance in the LC circuit because ideal capacitor & ideal inductor will not exist.

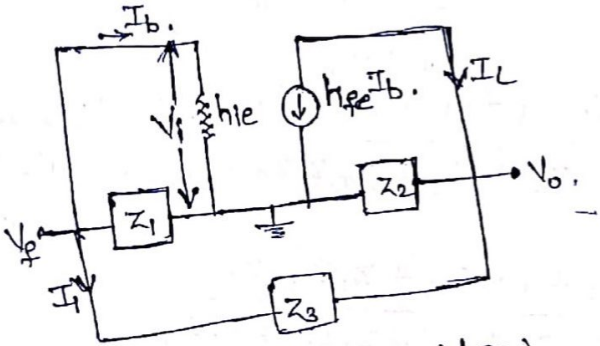
Because of this resistance energy is lost during each conversions and eventually oscillations will die.

LC circuit is also called as tank circuit.

General form of LC oscillator circuit :-



The above figure shows a general LC oscillator circuit. Impedances, Z_1, Z_2 & Z_3 will form the tank circuit. Assume the amplifier is a BJT amplifier. Replacing the amplifier with approximate h-parameter model.



To satisfy Bark Hausen criteria $A\beta$ must be 1.

$$A = \text{voltage gain} = \frac{V_o}{V_i} = \frac{I_L Z_L}{I_1 Z_1} = \left(\frac{I_L}{I_1}\right)\frac{Z_L}{Z_i}$$

$$A = \frac{AI \cdot Z_L}{Z_i}.$$

$$AI = \frac{I_L}{I_1} \quad \because I_L = -h_{fe}I_b, \; I_1 = I_b.$$

$$AI = \frac{-h_{fe}\cancel{I_b}}{\cancel{I_b}} = -h_{fe}.$$

$$Z_i = \frac{V_i}{I_i} = \frac{I_b hie}{I_b} = hie \;;$$

The load impedance

$$Z_L = Z_2 \,||\, Z_3'.$$

where $Z_3' = Z_3 + Z_1'$

$$Z_1' = Z_1 \| h_{ie}$$

$$Z_1' = \frac{h_{ie} Z_1}{Z_1 + h_{ie}}.$$

$$Z_3' = Z_3 + Z_1' = Z_3 + \frac{h_{ie} Z_1}{Z_1 + h_{ie}}.$$

$$= \frac{Z_1 Z_3 + h_{ie} Z_3 + h_{ie} Z_1}{Z_1 + h_{ie}}.$$

$$Z_L = Z_2 \| Z_3' = \frac{Z_2 \left(\frac{Z_1 Z_3 + h_{ie} Z_3 + h_{ie} Z_1}{Z_1 + h_{ie}}\right)}{Z_2 + \left(\frac{Z_1 Z_3 + h_{ie} Z_3 + h_{ie} Z_1}{Z_1 + h_{ie}}\right)}.$$

$$= \frac{Z_1 Z_2 Z_3 + h_{ie} Z_2 Z_3 + h_{ie} Z_1 Z_2}{Z_1 Z_2 + h_{ie} Z_2 + Z_1 Z_3 + h_{ie} Z_3 + h_{ie} Z_1}$$

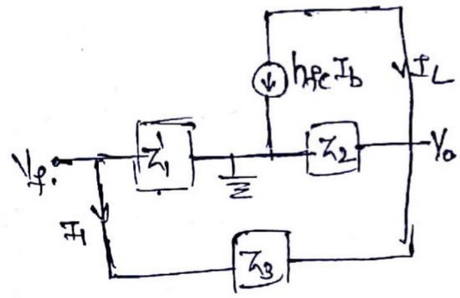
$$Z_L = \frac{(h_{ie} Z_1 + h_{ie} Z_3 + Z_1 Z_3) Z_2}{h_{ie}[Z_1 + Z_2 + Z_3] + Z_1 Z_2 + Z_1 Z_3}.$$

$$A = \frac{A_I . Z_L}{Z_i} = \frac{-h_{fe} \frac{(h_{ie} Z_1 + h_{ie} Z_3 + Z_1 Z_3) Z_2}{h_{ie}(Z_1 + Z_2 + Z_3) + Z_1 Z_2 + Z_1 Z_3}}{h_{ie}}.$$

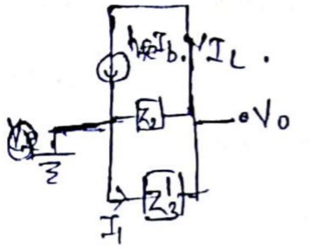
$$A = \frac{-h_{fe} (h_{ie} Z_1 + h_{ie} Z_3 + Z_1 Z_3) Z_2}{h_{ie}[h_{ie}(Z_1 + Z_2 + Z_3) + Z_1 Z_2 + Z_1 Z_3]}$$

Feedback factor $\beta = \frac{V_f}{V_o}.$

$$V_f = -I_1 Z_1' = -I_1 \frac{h_{ie} Z_1}{Z_1 + h_{ie}}.$$



$$V_0 = -I_1 Z_3' = -I_1 \times \left(\frac{Z_1Z_3 + h_{ie}Z_3 + h_{ie}Z_1}{Z_1 + h_{ie}}\right)$$



$$\beta = \frac{V_f}{V_0} = \frac{-I_1 \cdot \frac{h_{ie} Z_1}{Z_1 + h_{ie}}}{-I_1\left(\frac{Z_1Z_3 + h_{ie}Z_3 + h_{ie}Z_1}{Z_1 + h_{ie}}\right)}$$

$$\beta = \frac{h_{ie} Z_1}{h_{ie}Z_1 + h_{ie}Z_3 + Z_1Z_3}$$

$$\therefore A\beta = \left(\frac{-h_{fe}\cancel{(h_{ie}Z_1 + h_{ie}Z_3 + Z_1Z_3)}Z_2}{\cancel{h_{ie}}[h_{ie}(Z_1+Z_2+Z_3) + Z_1Z_2 + Z_1Z_3]}\right) \times \left(\frac{\cancel{h_{ie}}Z_1}{\cancel{h_{ie}Z_1 + h_{ie}Z_3 + Z_1Z_3}}\right) = 1$$

$$\Rightarrow \frac{-h_{fe}Z_1Z_2}{h_{ie}(Z_1+Z_2+Z_3) + Z_1Z_2 + Z_1Z_3} = 1$$

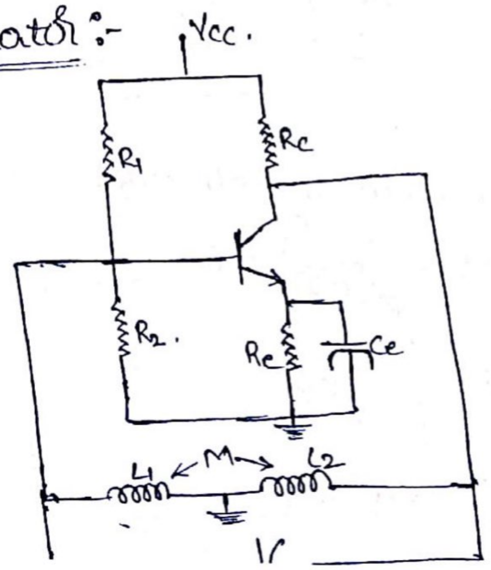
$$\Rightarrow h_{ie}(Z_1+Z_2+Z_3) + Z_1Z_2 + Z_1Z_3 = -h_{fe}Z_1Z_2$$

$$\Rightarrow \boxed{h_{ie}(Z_1+Z_2+Z_3) + Z_1Z_2(1+h_{fe}) + Z_1Z_3 = 0}$$

The above equation is called the general equation for the LC oscillator.

01/08/19
Friday.

Hartley oscillator :-



The above figure shows a hartley oscillator. The CE amplifier produces 180° phase shift & LC tank circuit produces 180° phase shift.

∴ Total phase shift is 360°, satisfying the barkhausen criteria.

The general equation for the LC oscillator is

$$h_{ie}[Z_1+Z_2+Z_3]+Z_1Z_2(1+h_{fe})+Z_1Z_3=0.$$

$$Z_1 = j\omega L_1 + j\omega M$$

$$Z_2 = j\omega L_2 + j\omega M$$

$$Z_3 = \frac{1}{j\omega C}.$$

$$h_{ie}\left[j\omega L_1+j\omega M+j\omega L_2+j\omega M+\frac{1}{j\omega C}\right]+(j\omega L_1+j\omega M)(j\omega L_2+j\omega M)(1+h_{fe})$$
$$+(j\omega L_1+j\omega M)\left(\frac{1}{j\omega C}\right)=0.$$

$$h_{ie}\left[j\omega L_1+j\omega L_2+2j\omega M+\frac{1}{j\omega C}\right]+(j\omega)[L_1+M](j\omega)[L_2+M](1+h_{fe})+(j\omega)[L_1+M]\frac{1}{(j\omega C)}=0$$

$$h_{ie}\,j\omega\left[L_1+L_2+2M+\frac{1}{(j\omega C)^2}\right]-\omega^2[L_1+M][L_2+M](1+h_{fe})+\frac{[L_1+M]}{C}=0$$

$$\underbrace{h_{ie}\,j\omega\left[L_1+L_2+2M-\frac{1}{(\omega^2 C)}\right]}_{\text{Imaginary part}}\underbrace{-\omega^2[L_1+M]\cdot[L_2+M](1+h_{fe})+\frac{[L_1+M]}{C}}_{\text{real part}}=0$$

Equating imaginary part on both sides

$$j\,h_{ie}\,\omega\left[L_1+L_2+2M-\frac{1}{\omega^2 C}\right]=0.$$

$$L_1+L_2+2M=\frac{1}{\omega^2 C}.$$

Let $L_1+L_2+2M=L_{eq}$.

$$L_{eq}=\frac{1}{\omega^2 C}.$$

$$\omega = \frac{1}{\sqrt{L_{eq} C}}$$

$$2\pi f = \frac{1}{\sqrt{L_{eq} C}}$$

$$f = \frac{1}{2\pi\sqrt{L_{eq} C}} \, ;$$

Condition for sustained oscillations :-

Equating the real part on Both sides.

$$-\omega^2 [L_1+M][L_2+M](1+h_{fe}) + \frac{[L_1+M]}{C} = 0 .$$

$$\omega^2 \cancel{(L_1+M)} (L_2+M)(1+h_{fe}) = \frac{\cancel{(L_1+M)}}{C}$$

$$(L_2+M)(1+h_{fe}) = \frac{1}{\omega^2 C} \qquad \text{But } \frac{1}{\omega^2 C} = L_{eq} = L_1+L_2+2M$$

$$(L_2+M)(1+h_{fe}) = L_1+L_2+2M .$$

$$\cancel{(L_2+M)(L_2+M)h_{fe} = L_1+L_2+2M} .$$

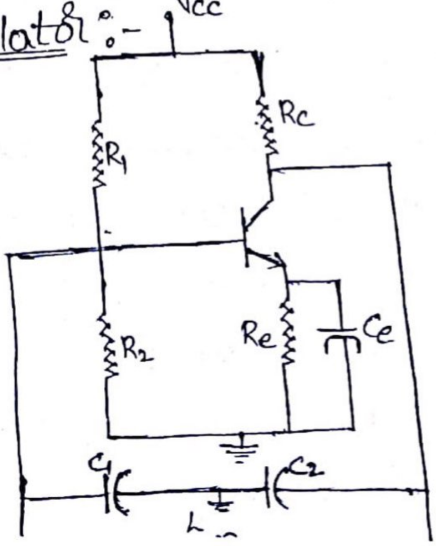
$$(L_2+M)+(L_2+M)h_{fe} = L_1+L_2+2M .$$

$$(L_2+M)h_{fe} = L_1+M .$$

$$\boxed{h_{fe} = \frac{L_1+M}{L_2+M}}$$

05/03/19
Tuesday.

Colpitt's oscillator :-



The above figure shows the colpitt's oscillator. The C.E amplifier ~~The general equation for LC oscillator~~ and the LC tank circuit produces produces 180° phase shift and the LC tank circuit produces another 180° phase shift.

$\therefore$ The total phase shift around the loop is 360°, satisfying the bark hausen criteria.

The general equation for LC oscillator is.

$$h_{ie}[Z_1+Z_2+Z_3]+Z_1Z_2(1+h_{fe})+Z_1Z_3=0.$$

$$Z_1=\frac{1}{j\omega C_1}$$

$$Z_2=\frac{1}{j\omega C_2}$$

$$Z_3=j\omega L.$$

$$h_{ie}\left[\frac{1}{j\omega C_1}+\frac{1}{j\omega C_2}+j\omega L\right]+\left(\frac{1}{j\omega C_1}\right)\left(\frac{1}{j\omega C_2}\right)(1+h_{fe})+\left(\frac{1}{j\omega C_1}\right)(j\omega L)=0$$

$$\frac{h_{ie}}{j\omega}\left[\frac{1}{C_1}+\frac{1}{C_2}+(j\omega)^2L\right]+\frac{1}{(j\omega)^2}\cdot\frac{1}{C_1C_2}(1+h_{fe})+\frac{L}{C_1}=0.$$

Let $\frac{1}{C_1}+\frac{1}{C_2}=\frac{1}{C_{eq}}.$

$$\frac{h_{ie}}{j\omega}\left[\frac{1}{C_{eq}}-\omega^2L\right]-\frac{1}{\omega^2}\frac{1}{C_1C_2}(1+h_{fe})+\frac{L}{C_1}=0.$$

Equating the imaginary part of both sides of equation.

$$\frac{h_{ie}}{\omega}\left[\frac{1}{C_{eq}}-\omega^2L\right]=0$$

$$\frac{1}{C_{eq}}-\omega^2L=0.$$

$$\frac{1}{C_{eq}}=\omega^2L.$$

$$\omega=\frac{1}{\sqrt{LC_{eq}}}.$$

$$\therefore 2\pi f = \frac{1}{\sqrt{LC_{eq}}}.$$

$$\boxed{f = \frac{1}{2\pi\sqrt{LC_{eq}}}}$$

condition for sustained oscillations:-
Equating the real part on both sides of ~~[illegible]~~ equatio,

$$\frac{-1}{\omega^2}\frac{1}{C_1C_2}(1+h_{fe}) + \frac{L}{C_1} = 0.$$

$$\frac{1}{\omega^2 \cancel{C_1} C_2}(1+h_{fe}) = \frac{L}{\cancel{C_1}}$$

$$\omega^2 L = \frac{1+h_{fe}}{C_2}.$$

$$\frac{1}{C_{eq}} = \frac{1+h_{fe}}{C_2}.$$

$$\frac{1}{C_1} + \cancel{\frac{1}{C_2}} = \cancel{\frac{1}{C_2}} + \frac{h_{fe}}{C_2}.$$

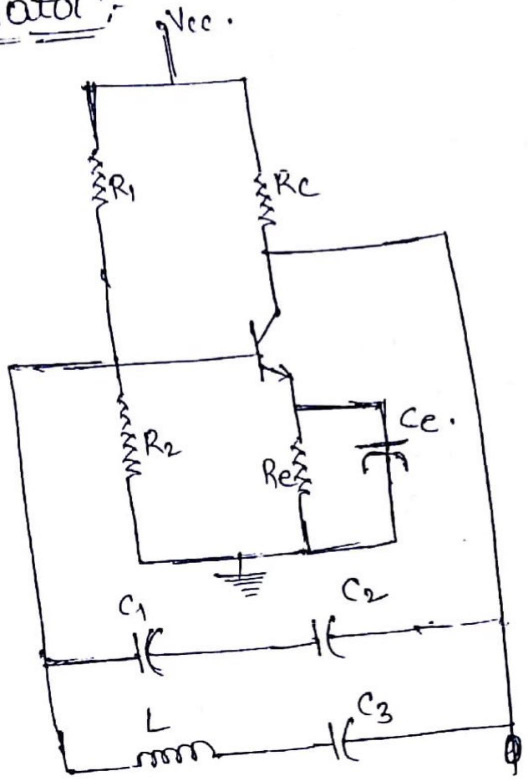
$$\frac{1}{C_1} = \frac{h_{fe}}{C_2}.$$

$$\boxed{h_{fe} = \frac{C_2}{C_1}}$$

Drawback in colpitts oscillator:
colpit's oscillator is a high frequency oscillator. The valu of C_1, C_2 will be very small. From the circuit we can observe that C_1 is between base & ground & C_2 is between collector & ground.
But there exist parasetic capacitances (junction capacitances between base & ground & collector and ground. these parasetic capacitances will be in parallel with C_1 & C_2 and hence the actual value of C_{eq} will change. Hence it is difficult to acurat the predict the frequency. This problem can be

avoided by adding a small capacitor C_3 in series with Inductor L.
C_3 is such that $C_3 << C_1, C_2$.

Clapp's oscillator:-



The above circuit is called clapp oscillator.
The CE Amplifier produces 180° phase shift and LC tank oscillator produces another 180° phase shift. The total phase shift is 360°, which satisfies the Barkhausen criteria. Thus the barkhausen angle criteria is satisfied.
The general equation is LC oscillator is

$$h_{ie}(Z_1+Z_2+Z_3) + Z_1Z_2(1+h_{fe}) + Z_1Z_3 = 0.$$

Now. $Z_1 = \frac{1}{j\omega C_1}$

$Z_2 = \frac{1}{j\omega C_2}$

$Z_3 = j\omega L + \frac{1}{j\omega C_3}$

$$h_{ie}\left[\frac{1}{j\omega C_1}+\frac{1}{j\omega C_2}+j\omega L+\frac{1}{j\omega C_3}\right]+\frac{1}{j\omega C_1}\cdot\frac{1}{j\omega C_2}(1+h_{fe})$$
$$+\frac{1}{j\omega C_1}\left(j\omega L+\frac{1}{j\omega C_3}\right)=0.$$

$$\frac{h_{ie}}{j\omega}\left[\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}+(j\omega)^2L\right]+\frac{1}{(j\omega)^2}\cdot\frac{1}{C_1C_2}(1+h_{fe})+\frac{j\omega}{j\omega C_1}\left[L+\frac{1}{(j\omega)^2C_3}\right]=0$$

Let $\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}=\frac{1}{C_{eq}}$.

$$\underbrace{\frac{h_{ie}}{j\omega}\left[\frac{1}{C_{eq}}-\omega^2L\right]}_{\text{Imaginary part}}\underbrace{-\frac{1}{\omega^2C_1C_2}(1+h_{fe})+\frac{1}{C_1}\left[L-\frac{1}{\omega^2C_3}\right]}_{\text{Real}}=0.$$

Equating the imaginary part on both sides.

$$\frac{h_{ie}}{\omega}\left[\frac{1}{C_{eq}}-\omega^2L\right]=0.$$

$$\frac{1}{C_{eq}}-\omega^2L=0$$

$$\omega_1^2=\frac{1}{LC_{eq}}.$$

$$\omega=\frac{1}{\sqrt{LC_{eq}}}.$$

$$2\pi f=\frac{1}{\sqrt{LC_{eq}}}.$$

$$\boxed{f=\frac{1}{2\pi\sqrt{LC_{eq}}}}$$

Condition for sustained oscillations:

Equating the real part on both sides.

$$\frac{-1}{\omega^2C_1C_2}(1+h_{fe})+\frac{1}{C_1}\left(L-\frac{1}{\omega^2C_3}\right)=0.$$

$$\frac{-1}{\omega^2C_2}(1+h_{fe})+\left(L-\frac{1}{\omega^2C_3}\right)=0.$$

$$\frac{-1}{\omega^2 C_2} - \frac{h_{fe}}{\omega^2 C_2} + L - \frac{1}{\omega^2 C_3} = 0.$$

Take $\frac{1}{\omega^2}$ common.

$$\frac{-1}{C_2} - \frac{1}{C_3} + \omega^2 L - \frac{h_{fe}}{C_2} = 0.$$

but $\omega^2 L = \frac{1}{C_{eq}}$.

$$\frac{-1}{C_2} - \frac{1}{C_3} + \frac{1}{C_{eq}} - \frac{h_{fe}}{C_2} = 0.$$

$$\cancel{\frac{-1}{C_2}} - \cancel{\frac{1}{C_3}} + \frac{1}{C_1} + \cancel{\frac{1}{C_2}} + \cancel{\frac{1}{C_3}} - \frac{h_{fe}}{C_2} = 0.$$

$$\frac{1}{C_1} = \frac{h_{fe}}{C_2}$$

$$\boxed{h_{fe} = \frac{C_2}{C_1}}$$

frequency $f = \frac{1}{2\pi\sqrt{L C_{eq}}}$.

$$\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

But $C_3 << C_1, C_2$.

$$\frac{1}{C_3} >> \frac{1}{C_1}, \frac{1}{C_2}$$

$$\therefore \frac{1}{C_3} \simeq \frac{1}{C_{eq}} \Rightarrow C_{eq} = C_3.$$

$$\boxed{f = \frac{1}{2\pi\sqrt{L C_{eq}}} \cong \frac{1}{2\pi\sqrt{L C_3}}}$$

$\therefore$ The frequency of the clapp oscillator depends on value of C_3 rather than on the values of C_1, C_2.

06/03/19
wednesday

Crystal oscillator:-

Piezo electric Effect:

when an electric field is applied across a piezo electric crystal mechanical vibrations are produced in a direction normal to the applied electric field.

Conversely, if mechanical stress is applied on a piezo electric crystal, electrical signal will be generated with a frequency equal to the resonant frequency of the crystal. This effect is known as piezo electric effect.

Examples of piezo electric crystals are Quartz, Rochelle salt & tourmeline.

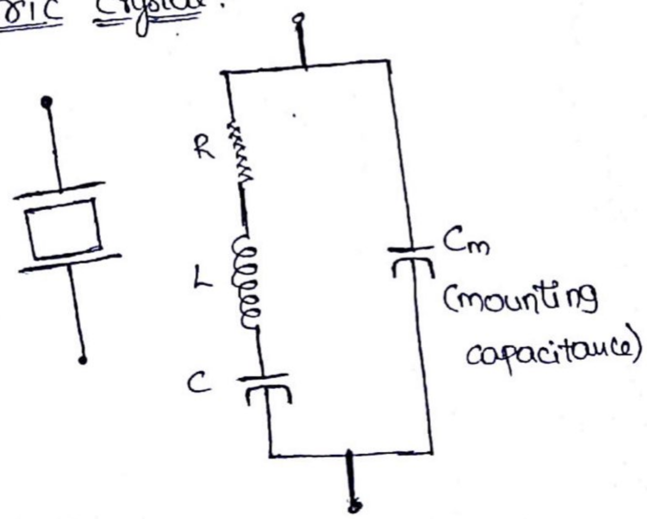
Quartz has medium frequency stability and medium Physical strength. It is easily available and is very cheap.

Rochelle salt has highest frequency stability and lowest Physical strength.

Tourmeline has lowest frequency stability, but highest Physical strength and it is more costly.

Quartz is more commonly used.

Equivalent circuit of crystal oscillator (or) electrical Equivalent circuit of Piezo Electric Crystal:-



In the above circuit, R represents the Resistance of the crystal.
which is a result of friction losses of the crystal.
L represents the inductance of the circuit, which is a result of moment of inertia of the crystal.
C represents the capacitance of the circuit, which is a result of stiffness of the crystal.

The crystal is normally placed between two conducting plates and this results in a capacitance called the mounting capacitance.

Crystal can be operated in one of the two mode.

1) Fundamental Mode:-

In this mode, the crystal will be operating at a frequency equal to the fundamental resonant frequency of the crystal.

2) Over tone mode:-

In this mode, the crystal will be operating at a frequency which is integral multiple of the fundamental frequency.

Resonance frequency of the crystal:-

Crystal will have two resonance frequencies.

Series Resonance frequency:-

Resonance is that condition where the impedance of the circuit is purely resistive.

∴ At series resonance the reactance of the series RLC circuit will be zero.

$$\therefore j\omega L + \frac{1}{j\omega C} = 0.$$

$$j\omega L = -\frac{1}{j\omega C}.$$

$$(j\omega)^2 = \frac{-1}{LC}.$$

$$\omega^2 = \frac{1}{LC}$$

$$\omega = \frac{1}{\sqrt{LC}}$$

This frequency is called series Resonance frequency & denoted by ω_s.

$$\omega_s = \frac{1}{\sqrt{LC}}$$

$$2\pi f_s = \frac{1}{\sqrt{LC}}$$

$$\boxed{f_s = \frac{1}{2\pi\sqrt{LC}}}$$

Parallel Resonance frequency :-

In parallel Resonance, the total reactance between the two electrodes of the crystal should be zero.

$$j\omega L + \frac{1}{j\omega C} + \frac{1}{j\omega C_m} = 0.$$

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

$$(j\omega)^2 L = -\left[\frac{1}{C} + \frac{1}{C_m}\right]$$

$$\omega^2 L = \left[\frac{1}{C} + \frac{1}{C_m}\right]$$

Let $\frac{1}{C} + \frac{1}{C_m} = \frac{1}{C_{eq}}$.

$$\therefore \omega^2 L = \frac{1}{C_{eq}}.$$

$$\omega = \frac{1}{\sqrt{LC_{eq}}}$$

This is called Parallel ~~series~~ Resonant frequency & denoted by ω_p.

$$\omega_p = \frac{1}{\sqrt{LC_{eq}}}.$$

$$2\pi f_p = \frac{1}{\sqrt{LC_{eq}}}$$

$$\boxed{f_p = \frac{1}{2\pi\sqrt{LC_{eq}}}}$$

$\frac{1}{C_{eq}} = \frac{1}{C} + \frac{1}{C_m}$

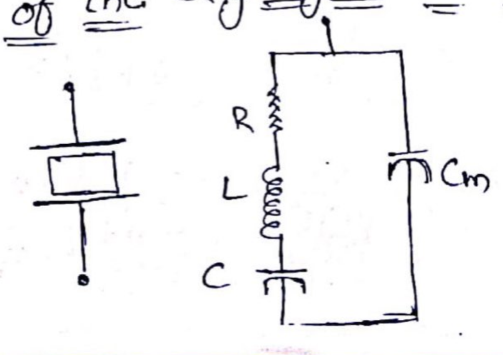
But $C_m \gg C$;

$\frac{1}{C_m} \ll \frac{1}{C}$.

$\frac{1}{C_{eq}} \simeq \frac{1}{C}$.

$C_{eq} \simeq C$. $\therefore \omega_s \simeq \omega_p$.

Impedance curve of the ~~sig~~ crystal (or) Impedance graph of the crystal :-



The impedance of the above circuit is.

$$z = \left(R + j\omega L + \frac{1}{j\omega C}\right) \Big|\Big| \frac{1}{j\omega C_m}$$

But normally R will be very small, ignore R.

$$\therefore z = \left(j\omega L + \frac{1}{j\omega C}\right) \Big|\Big| \frac{1}{j\omega C_m}$$

$$= \frac{\left(j\omega L + \frac{1}{j\omega C}\right)\left(\frac{1}{j\omega C_m}\right)}{j\omega L + \frac{1}{j\omega C} + \frac{1}{j\omega C_m}}$$

$$= \frac{\frac{j\omega L \cdot j\omega C + 1}{\cancel{j\omega C}} \cdot \frac{1}{\cancel{j\omega C_m}}}{\frac{j\omega L \cdot j\omega C \cdot j\omega C_m + j\omega C_m + j\omega C}{\cancel{j\omega C} \cdot \cancel{j\omega C_m}}}$$

$$= \frac{-\omega^2 LC + 1}{-\omega^2 LC \cdot j\omega C_m + j\omega C_m + j\omega C}$$

$$= \frac{1}{j\omega C_m}\left[\frac{-\omega^2 LC + 1}{-\omega^2 LC + 1 + \frac{j\omega C}{j\omega C_m}}\right]$$

$$= \frac{1}{j\omega C_m}\left[\frac{-\omega^2 LC + 1}{-\omega^2 LC + 1 + \frac{C}{C_m}}\right]$$

$$= \frac{1}{j\omega C_m} \frac{\cancel{LC}\left[-\omega^2 + 1/LC\right]}{\cancel{LC}\left[-\omega^2 + \frac{1}{LC} + \frac{1}{LC_m}\right]}$$

$$= \frac{1}{j\omega C_m}\left(\frac{-\omega^2 + 1/LC}{-\omega^2 + \frac{1}{L}\left(\frac{1}{C} + \frac{1}{C_m}\right)}\right)$$

$$= \frac{1}{j\omega C_m}\left(\frac{-\omega^2 + \frac{1}{LC}}{-\omega^2 + \frac{1}{LC_{eq}}}\right)$$

Th

$$= \frac{1}{j\omega C_m}\left(\frac{-\omega^2 + \omega_s^2}{-\omega^2 + \omega_p^2}\right)$$

$$Z = \frac{1}{j\omega C_m}\left(\frac{\omega^2 - \omega_s^2}{\omega^2 - \omega_p^2}\right)$$

At $\omega = \omega_s$.

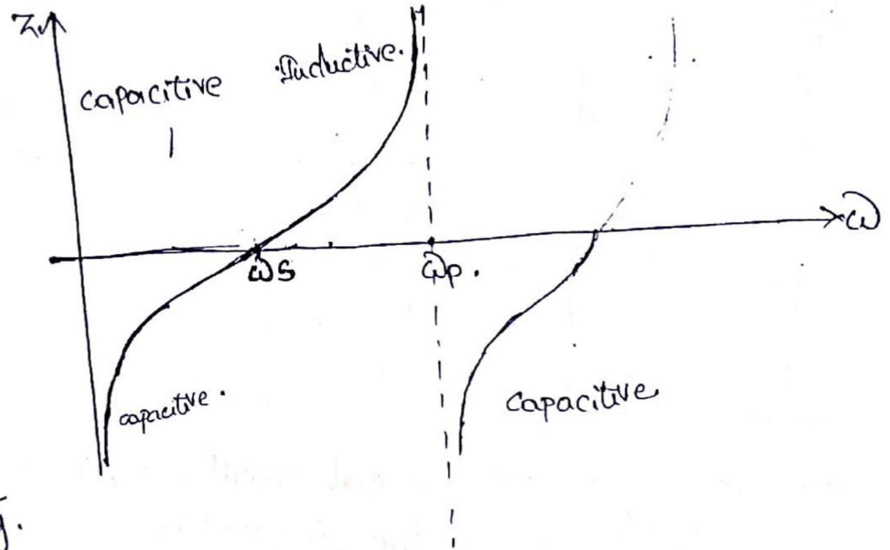
$Z = 0$.

At $\omega = \omega_p$, $Z = \infty$.

For $\omega < \omega_s$, impedance is capacitive.

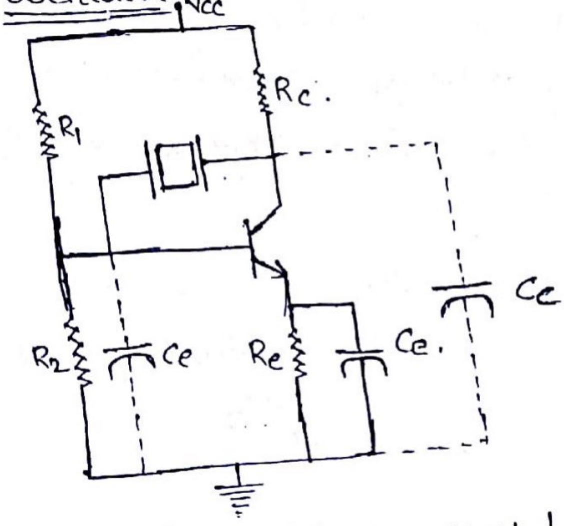
For $\omega > \omega_p$, impedance is capacitive.

For $\omega_s < \omega < \omega_p$, impedance is inductive.



07/03/19
Thursday.

Pierce crystal oscillator



The above figure shows pierce crystal. The circuit is similar to colpitt's oscillator. Crystal will behave like an

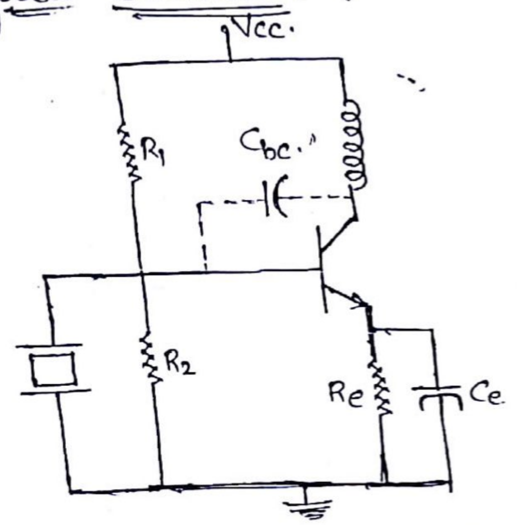
Inductor when operated at a frequency ω such that $\omega_s < \omega < \omega_p$.

The junction capacitances across the base emitter junction and collector emitter junction will provide the required capacitance for the tank circuit.

The frequency of oscillation is given by the resonant frequency of the crystal.

$$f = \frac{1}{2\pi\sqrt{LC}}.$$

Miller crystal oscillator :-



The above figure shows miller crystal oscillator, it is similar to the hartely oscillator. one of the inductors is replaced with crystal and the collector base junction capacitance provides the necessary capacitance to the tank circuit. The circuit oscillates with the resonant frequency of the crystal.

Frequency stability of oscillators:

Frequency stability of an oscillator is a measure of its ability to maintain the required frequency as precisely as possible over as long a time possible.

The main drawback in transistor oscillator is the frequency of oscillation is not stable for a long time operation.

The following are the factors that contribute to the change in frequency.

1) Due to the change in temperature, the values of the frequency determing components like resistors, inductors and capacitors will change.

2) Due to fluctations in power supply

3) Due to change in climatic conditions and aging.

4) Due to loading effects.

Loading effects can be minimized with the oscillator is connected to the load through an impedance matching circuit. The temperature effects may be minimized by using automatic temperature control mechanisms.

The frequency stability is defined as

$$S_\omega = \frac{d\theta}{d\omega}$$

where $d\theta$ is the phase shift introduced for a small frequency change in the nominal frequency ω_0.

The circuit having a larger value of $\frac{d\theta}{d\omega}$ has more frequency stability.

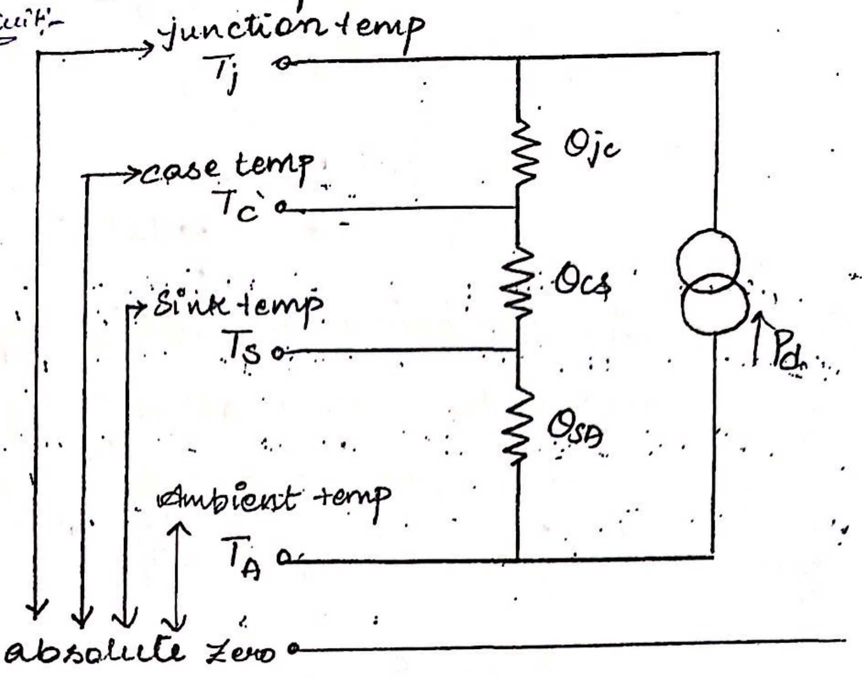
Oscillators with large Q-factor are more stable.

Amongest all the available oscillators, crystal oscillators have highest frequency stability.

Amplitude stability of oscillators:-

The amplitude of the oscillations may also drift due to temperature aging, supply flucuations and loading effects. Incase of RC oscillators, if there is any change in amplitude that can be stabilized by replacing the resistors in the ~~bridge~~ (feed back loop) RC network by sensistors which are temperature dependent resistors.

Electrical Analogous circuit:-



27/8

Pb

① Determine the junction, case & sink temperature for a power T's if it carries $I_c = 1$Amp with an average (dc) voltage of 10V. Assume $\theta_{jc} = 5°C/W$, $\theta_{SA} = 4°C/W$, $\theta_{cs} = 1°C/W$ the ambient temperature is $T_A = 25°C$

Sol

$I_c = 1$Amp, $V_{CE} = 10V$, $\theta_{jc} = 5°C/W$, $\theta_{SA} = 4°C/W$

$\theta_{cs} = 1°C/W$, $T_A = 25°C$

rise in temperature $\Rightarrow T_j - T_A = \theta_{jA} P_d$

$$P_d = V_{CE} \cdot I_c$$
$$= 10W$$

$$T_S - T_A = P_d \cdot \theta_{SA}$$

$$T_S - 25 = 10(4)$$

$$T_S = 40 + 25$$
$$= 65°C$$

$$T_C - T_S = \theta_{cs} P_d$$

$$T_C - 65 = 1 \times 10 \Rightarrow T_C = 75°C$$

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$$T_j - T_c = \theta_{jc} P_d$$

$$T_j - 75 = 5(10)$$

$$T_j = 50 + 75 = 125°C$$

② Determine the power handling capacity of a 60W power transistor rated at 25°c if derating is required above 25°c at a case temperature of 100°c. The derating factor is 0.2W/°c.

$$T_0 = 25°C$$

$$T_1 = 100°C$$

$$P(T_0) = 60W$$

$$P(T_1) = ?$$

$$D = 0.25$$

$$P(T_1) = P(T_0) - (T_1 - T_0)(D)$$

$$P(T_1) = \overset{60}{\cancel{P(T_0)}} - (75)(0.25)$$

$$\cancel{P(T_1) = 93.75}$$

$$P(T_1) = 41.25W$$

③ A power transistor with a rated power 20W and maximum junction temp of 150°C is to be operated in air with an ambient temp of 25°C. calculate its thermal resistance b/w junction & case for rated power handling capacity. If a heat sink is provided now with $\theta_{SA} = 1°C/W$ & $\theta_{CS} = 2°C/W$. Determine power handling capacity of the transistor.

$P_d = 20W$ $\theta_{SA} = 1°C/W$ } after introducing
$T_j = 150°C$ $\theta_{CS} = 2°C/W$ } heat sink.
$T_A = 25°C$
$\theta_{jC} = ?$

If the case is ideal, $T_C = T_A$, $\therefore \theta_{jC} = \frac{T_j - T_C}{P_d}$

$$\theta_{jC} = \frac{T_j - T_A}{P_d} = 6.25°C/W$$

$P_d' = ?$

$$\theta_{jA} = \theta_{jC} + \theta_{SA} + \theta_{CS}$$

$$= 6.25 + 3$$

$$= 9.25 °C/W$$

$$T_j - T_A = \theta_{jA} P_d'$$

$$\frac{150 - 25}{9.25} = P_d'$$

$$\Rightarrow P_d' = 13.51 W$$

(4) State the heat sink selection criteria for a transistor operating with $V_{CE} = 20V$, $I_C = 1.5$ Amp. Given $\theta_{jC} = 1°C/W$. junction temp should not exceed 85°c while the ambient temp is 25°c

Sol

$V_{CE} = 20V$, $P_d = 30$

$I_C = 1.5A$.

$\theta_{jC} = 1°C/W$

$T_A = 25°C$

$T_j < 85°C$

one of the specifications of heat sink is $\theta_{CA} = \theta_{CS} + \theta_{SA}$.

$$\Rightarrow \quad \theta_{CA} = \frac{85-25}{30} - 1$$

$$\Rightarrow \quad \theta_{CA} = 1°C/W$$

∴ A sink with $\theta_{CA} = 1°C/W$ should be selected.

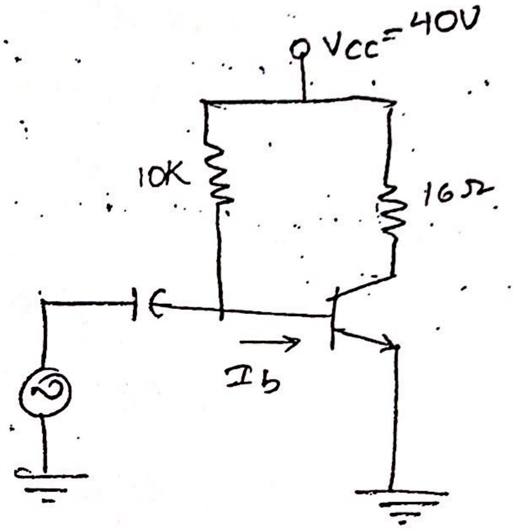
(5) A series fed class A amplifier uses a supply of 40V. The AC i/p voltage results in a base current of 6mA peak. calculate (a) Q-point ($\beta = 25$)

(b) DC i/p power

(c) AC o/p power

(d) Efficiency

(e) power dissipated



Sol

$$(I_B)_{max} = 6mA$$

$$V_{CC} = 40V$$

$$I_C = \beta I_B$$

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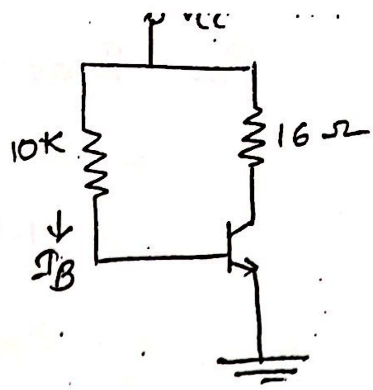
The DC equivalent ckt is



$$I_C = \beta I_B$$

$$V_{CC} = (10K) I_B + V_{BE}$$

$$I_B = \frac{V_{CC} - V_{BE}}{10K}$$

$$= \frac{40 - 0.7}{10K}$$

$(\because V_{BE} = 0.7V)$
↓
class A amp

$$= 3.93\,mA.$$

$$I_C = \beta I_B$$

$$= 25 \times 3.93\,mA$$

$$= 98.25\,mA$$

$$V_{CC} = I_C(16) + V_{CE}$$

$$V_{CE} = 40 - 98.25 \times 10^{-3} \times 16$$

$$V_{CE} = -1.572V + 40$$

$$= 38.428\,V$$

(a) Q point $= (\cancel{V_{CC}}\, V_{CE}, I_C) = (38.428, 98.25mA)$.

(b) Dc i/p power $= V_{CC} \cdot I_C = 10 \times 98.25 \times 10^{-3}$

$$= 83.93\,W$$

(c) AC o/p power $\Rightarrow P_{ac} = \frac{V_m I_m}{2} = \frac{1}{2} I_m^2 R_L$

$$(I_C)_{max} = I_{max} = \beta\,(I_B)_{max}$$

$$= 25 \times 6\,mA$$

$$= 150\,mA$$

$$I_m = I_{max} - I_c$$

$$= 150 - 98.25$$

$$= 51.75 mA.$$

$$P_{ac} = \frac{1}{2} I_m^2 R_L = \frac{1}{2} (51.75 \times 10^{-3})^2 \times 16$$

$$= 0.0214 W$$

(d) Efficiency $= \frac{P_{ac}}{P_{dc}}$

$$= \frac{0.0214}{3.93} \times 100 = 5.44 \times 10^{-3} \times 100$$

$$= 0.54$$

(e) power dissipated $= P_{dc} - P_{ac}$

$$= 3.93 - 0.0214$$

$$= 3.9 W.$$

**

⑥ A transistor amplifier with transformer coupled load produces harmonic amplitudes in the o/p as $B_0 = 1.5mA$, $B_1 = 120mA$, $B_2 = 10mA$, $B_3 = 4mA$, $B_4 = 2mA$, $B_5 = 1mA$.

(a) Determine the % of total harmonic distortion

(b) Assume an additional identical T/R is used to form a pushpull configuration. use above amplitudes to find % distortion.

(a) Given $B_0 = 1.5 mA$, $B_1 = 120mA$, $B_2 = 10mA$, $B_3 = 4mA$

$B_4 = 2mA$, $B_5 = 1mA$

Total harmonic distortion

$$D = \sqrt{D_2^2 + D_3^2 + D_4^2 + D_5^2}$$

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$$D_2 = \left|\frac{B_2}{B_1}\right| = \frac{10}{120} = 0.08$$

$$D_3 = \left|\frac{B_3}{B_1}\right| = 0.033$$

$$D_4 = \left|\frac{B_4}{B_1}\right| = \frac{2}{120} = 0.016$$

$$D_5 = \left|\frac{B_5}{B_1}\right| = 0.0083$$

$$D = \sqrt{D_2^2 + D_3^2 + D_4^2 + D_5^2}$$

$$D = \sqrt{(0.08)^2 + (0.033)^2 + (0.016)^2 + (0.008)^2}$$

$$D = \sqrt{8.31744 \times 10^{-3}}$$

$$D = 0.0912$$

$$= \cancel{8.8\%}$$

$$D = 91.2\%$$

(b) In a pushpull configuration, even harmonics are cancelled. so we take ~~only~~ $B_2 = B_4 = 0$

$$\Rightarrow D = \sqrt{D_3^2 + D_5^2}$$

$$= \sqrt{(0.033)^2 + (0.008)^2}$$

$$\% D = 3.4\%$$

(7) A sinusoidal signal with $V_s = 1.75 \sin 600t$ is fed to a power amplifier. The resulting o/p current is $I_0 = 15\sin 600t + 1.5\sin 1200t + 1.2\sin 1800t + 0.5\sin 2400t$. calculate the % Increase in power due to distortion.

$$P = (1+D^2)P_1$$

$P = \frac{1}{2} D^2 R_L$
(or)
$P = (1+D^2) P_1$.

$B_1 = 15$

$B_2 = 1.5$

$B_3 = \cancel{0.5} 1.2$

$B_4 = 0.5$

$$D_2 = \left|\frac{B_2}{B_1}\right| = \frac{1.5}{15} = 0.1$$

$$D_3 = \left|\frac{B_3}{B_1}\right| = \frac{1.2}{15} = 0.08$$

$$D_4 = \left|\frac{B_4}{B_1}\right| = \frac{0.5}{15} = 0.03$$

$$D = \sqrt{D_2^2 + D_3^2 + D_4^2}$$

$$= \sqrt{(0.1)^2 + (0.08)^2 + (0.03)^2}$$

$$= \sqrt{7.4 \times 10^{-3}}$$

$$= 0.13$$

Excess power due to distortion

$$= \frac{P - P_1}{P_1} \times 100$$

$$= D^2 \times 100$$

$$= 1.75$$

⑧ class B pushpull amplifier supplies power to a resistive load of 12Ω. The O/p T/F has a turns ratio of 3:1 & efficiency of 78.5%. obtain (i) Max. power O/p

(ii) Max. power dissipation in each transistor.

(iii) Max. base & collector current for each transistor.

Assume $h_{fe} = 25$, $V_{cc} = 20V$

Sol: $R_L = 12\Omega$

$1:n = 3:1$

$\eta = 78.5\%$

(i) $(P_c)_{max} = \frac{V_{cc}^2}{2R_L'}$

$R_L' = \frac{R_L}{n^2} = \frac{12}{1/9} = 12 \times 9 = 108$

$(P_o)_{max} = \frac{400}{2(108)} = 1.85W$

(ii) $P_C = P_i - P$

$= \frac{2V_{cc}I_m}{\pi} - \frac{V_mI_m}{2}$

$= \cancel{2V_{cc}}$

$(P_c)_{max} = \frac{2V_{cc}^2}{\pi^2 R_L'} = \frac{2 \times 400}{\pi^2 \times 108} = 0.75W$

power dissipated by each T/R $= \frac{(P_c)_{max}}{2}$

$= 0.75/2$

$= 0.375$

(iii) $[P_i = V_{cc} \cdot \frac{2I_m}{\pi} = 2.6$

$I_m = \frac{P_i \pi}{2V_{cc}} = 0.204\,Amp]$

$(P_o)_{max} = \frac{V_mI_m}{2} = \frac{V_{cc}I_m}{2} = 1.85W$

$\Rightarrow I_m = 0.185\,Amp.$

$$(I_C)_{max} = \beta (I_B)_{max}$$

$$\therefore (I_B)_{max} = \frac{(I_C)_{max}}{\beta}$$

$$= \frac{0.185}{25}$$

$$= 7.4\,mA.$$

(3) A silicon T/R operates as an ideal class B amplifier if Dc current drawn from the supply is 25mA. calculate the Ac power delivered to load. for a load resistance of 2KΩ.

Sol

$I_{DC} = 25\,mA$

$R_L = 2\,K\Omega$

$P_{ac} = ?$

$$P_{ac} = V_{rms} \cdot I_{rms}$$

$$= (I_{rms})^2 \cdot R_L \qquad \left[\because I_{rms} = \frac{I_m}{2}\right]$$

Half sinusoidal.

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

$$= \frac{1}{4} I_m^2 R_L$$

$$I_{DC} = \frac{I_m}{\pi} \Rightarrow I_m = I_{DC} \times \pi$$

$$= 25\,mA \times \pi$$

$$= 78.5\,mA$$

$$P_{ac} = \frac{1}{4}(78.5)^2 \times 2 \times 10^3$$

$$= 3.08\,W$$

$V_{cc}^2 = 80 \times \cancel{32}$

$V_{cc}^2 = \cancel{10240}\ 2560$

$V_{cc} = \cancel{101.19}$

$V_{cc} = 50.59\ V$

1/3/13

(11) A dynamic curve of a transistor amplifier is represented by $i_c = G_1 i_b + G_2 i_b^2$. The i/p signal to the transistor is $I_1 \cos w_1 t + I_2 \cos w_2 t$ show that the output will contain a Dc term sinusoidal signals with frequencies $w_1, w_2, 2w_1, 2w_2, w_1+w_2, w_1-w_2$.

Sol Given,

$$i_c = G_1 i_b + G_2 i_b^2$$

The input current, $i_b = I_1 \cos w_1 t + I_2 \cos w_2 t$

$$i_c = G_1 (I_1 \cos w_1 t + I_2 \cos w_2 t) + G_2 (I_1 \cos w_1 t + I_2 \cos w_2 t)^2$$

$$i_c = G_1 I_1 \cos w_1 t + G_1 I_2 \cos w_2 t + G_2 I_1^2 \cos^2 w_1 t + G_2 I_1 I_2 \cos w_1 \cos w_2 + G_2 I_2^2 \cos^2 w_2 t$$

$$i_c = \frac{G_2 I_1^2}{2} \; G_1 I_1 \cos\omega_1 t + G_1 I_2 \cos\omega_2 t + G_2 I_1^2 \left(\frac{1+\cos 2\omega_1 t}{2}\right)$$

$$G_2 I_2^2 \left(\frac{1+\cos 2\omega_2 t}{2}\right) + G_2 I_1 I_2 \left[\cos(\omega_1+\omega_2)t + \cos(\omega_1-\omega_2)t\right]$$

$$i_c = \frac{G_2 I_1^2}{2} + \frac{G_2 I_2^2}{2} + G_1 I_1 \cos\omega_1 t + G_1 I_2 \cos\omega_2 t +$$

$$G_2 I_1^2 \cos \frac{2\omega_1 t}{2} + G_2 I_2^2 \cos 2\omega_2 t +$$

$$G_2 I_1 I_2 \left[\cos(\omega_1+\omega_2)t + \cos(\omega_1-\omega_2)t\right]$$

$\therefore$ The o/p contains a DC term and sinusoidal signals with frequencies $\omega_1, \omega_2, 2\omega_1, 2\omega_2, \omega_1+\omega_2, \omega_1-\omega_2$

(12) The dynamic transfer curve for a given transistor is i_c (in mA) $= 50 i_b + 1000 i_b^2$ where, $i_b = 10\cos 2\pi(100t)$ mA. calculate the % of harmonic distortion.

Sol Given

$$i_c \text{ (in mA)} = 50 i_b + 1000 i_b^2 \text{ —— (1)}$$

$$i_b = 10\cos 2\pi(100t) \text{ mA}$$

$$i_c = G_1 i_b + G_2 i_b^2 \text{ —— (2)}$$

comparing (1) & (2)

$$G_1 = 50, \; G_2 = 1000$$

$$\Rightarrow i_b = I_{Bm} \cos\omega_0 t$$

$$\Rightarrow i_c = G_1 I_{Bm} \cos\omega_0 t + G_2 I_{Bm}^2 \cos^2\omega_0 t$$

$$\Rightarrow i_c = G_1 I_{Bm} \cos\omega_0 t + G_2 I_{Bm}^2 \left(\frac{1+\cos 2\omega_0 t}{2}\right)$$

$$\Rightarrow i_c = G_1 I_{Bm} \cos\omega_0 t + \frac{G_2 I_{Bm}^2}{2} + \frac{G_2 I_{Bm}^2}{2} \cos 2\omega_0 t$$

$$B_0 = \frac{G_2 I_{Bm}}{2} = \frac{1000mA \times (10mA)}{2} = 0.05 \times 10^{-}$$

$$B_1 = G_1 I_{Bm} = 50mA \times 10mA$$

$$= 500 \times 10^{-6} = 0.5 \times 10^{-3}$$

$$B_2 = \frac{G_2 I_{Bm}^2}{2} = 0.05 \times 10^{-3}$$

$$D_2 = \left|\frac{B_2}{B_1}\right| = 0.1$$

$$\Rightarrow \% D_2 = 10\%$$

(13) A silicon power transistor operated with a heat sink having thermal resistance $\theta_{cs} = 1.5°c/w$ & $\theta_{jc} = 0.5°c/w$ $\theta_{SA} = 1.5°c/w$. The T'r has to dissipate 150W of power at 25°c. Determine the dissipated power at 200°c.

Sol Given, $\theta_{jc} = 0.5°c/w$

$\theta_{cs} = 1.5°c/w$

$\theta_{SA} = 1.5°c/w$

$$\theta_{jA} = 0.5 + 1.5 + 1.5 = 3.5°c/w$$

$$T_j = 200, \quad T_A = 25$$

$$T_j - T_A = \theta_{jA} \cdot P_d$$

$$P_d = \frac{200-25}{3.5} = 50W$$

(14) calculate the second harmonic distortion if the o/p signal w/f of a pushpull amplifier has measured values of $V_{CE}(min) = 1V$, $V_{CE}(max) = 24V$ & $V_{CE}Q = 14V$

Sol Given $V_{CE}(min) = 1V$

$V_{CE}(max) = 24V$