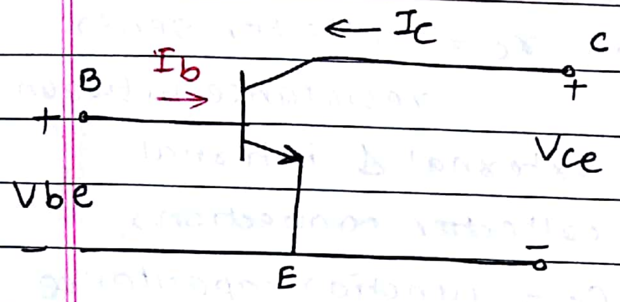
Frequency Response of amplifiers

* Extended hybrid-π equivalent circuit of BJT

Fig a) shows an npn bipolar transistor in a common emitter configuration, along with the small signal voltages & currents.



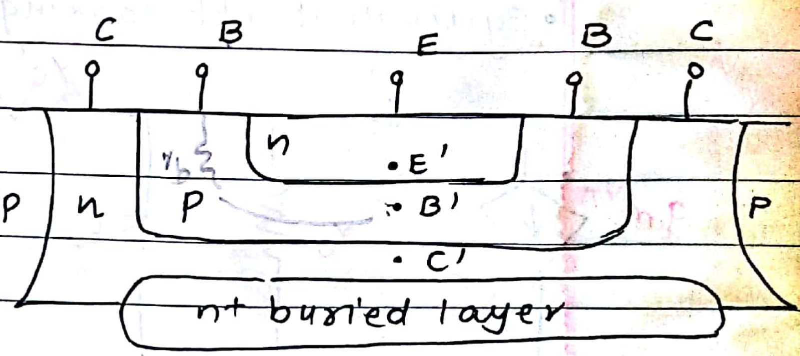
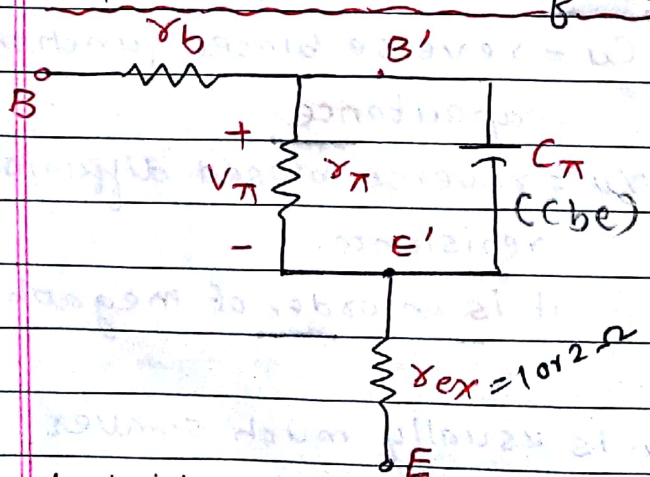


fig a) CE npn BJT

fig b) cross-section of npn BJT for the hybrid π Model.

fig b) is a cross-section of npn transistor in a (BJT amplifier ckt) classic integrated circuit configuration.
- The C, B & E terminals are the external connections to the transistor & C', B' & E' are the idealized internal collector, base and emitter regions.

• Equivalent circuit of the transistor:–



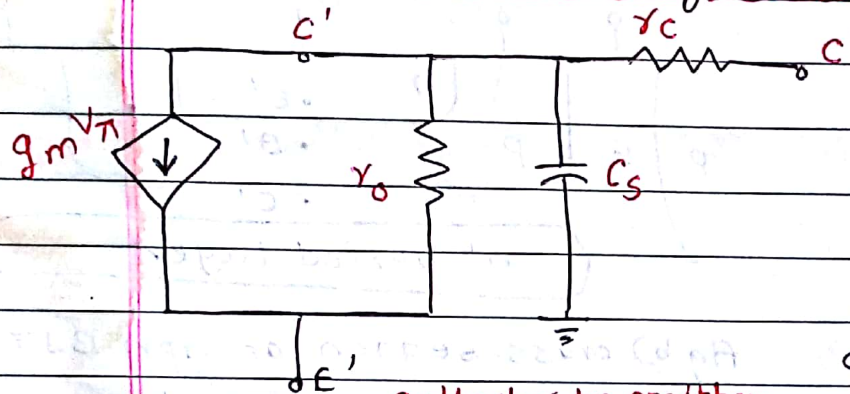
hybrid π equivalent ckt

base to emitter

- Resistance r_b is the base series resistance betn the external base terminal B & internal base region B'.
- the B'–E' junction is forward biased. therefore C_π is the forward biased junction capacitance &

r_π is the forward biased junction diffusion resistance.
C_π & r_π are the function of the junction current.
r_{ex} is the emitter series resistance between the external emitter terminal & the internal emitter region.
r_{ex} = usually very small in order of 1 to 2 Ω.

- **Equivalent ckt looking into collector terminal :-**

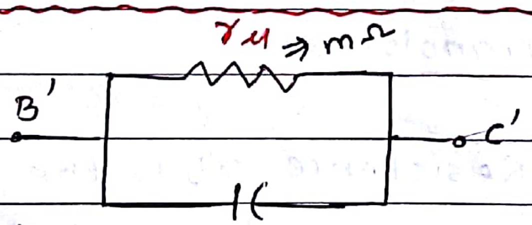


collector to emitter

r_c = collector series resistance between external & internal collector connections.
C_s = junction capacitance of reverse biased collector-substrate junction.

- The dependent current source $g_m V_\pi$ is the transistor collector current controlled by the internal base-emitter voltage.
- r_o = inverse of o/p conductance g_o & is due primarily to the Early effect.

- **Equivalent ckt of reverse biased B'-C' junction -**



$\Rightarrow C_\mu\ (C_{bc})$

Cu smaller $< C_\pi$

base to collector

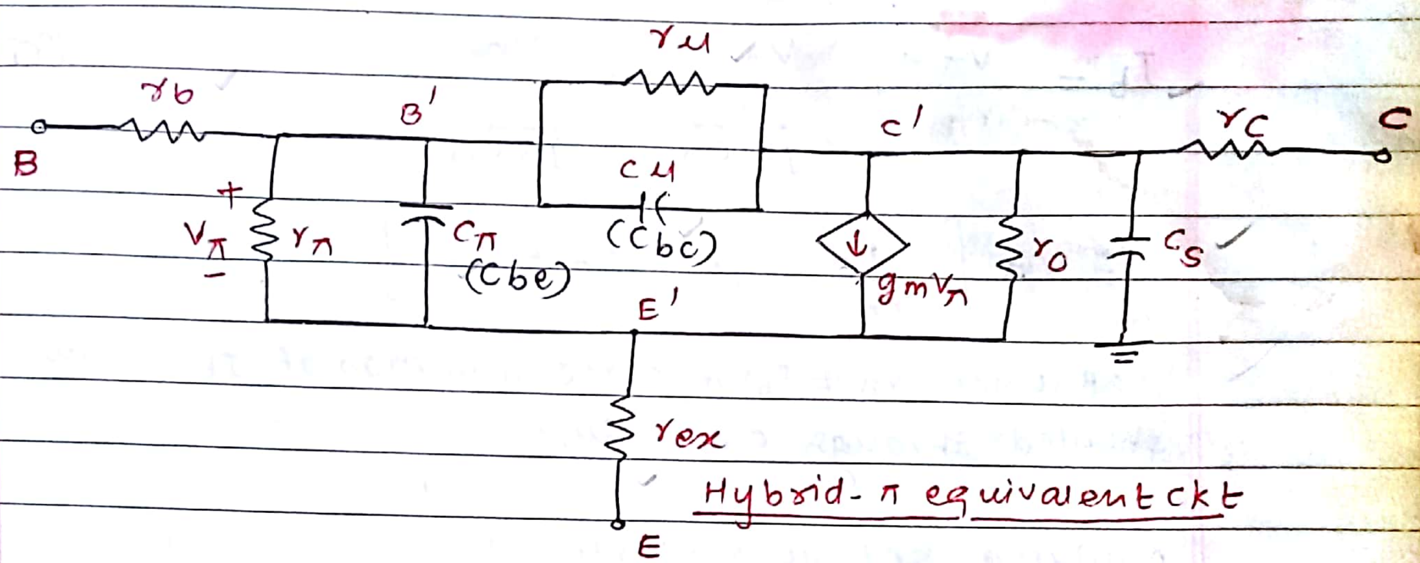
C_μ = reverse biased junction capacitance
r_μ = reverse biased diffusion resistance.
it is in order of megaohms & can be neglected.

● NOTE:- The value of C_μ is usually much smaller than C_π. however because of a phenomenon known as 'Miller effect', C_μ usually cannot be neglected.

* Hybrid π provides more accurate model for $\boxed{HF}$ effects.

$\boxed{\text{For LF to MF}}$ ⇒ r_b << small so S.C., r_c & r_{ex} << small ⇒ S.C. 3
r_μ >> large >> βr_o = O.C. so equivalent ckt = same as 'r_e' model.

• Complete hybrid-π equivalent ckt for BJT :-



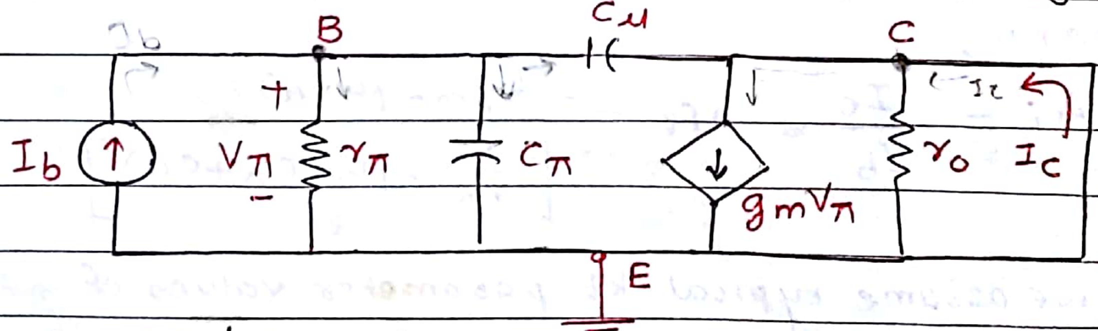
Hybrid-π equivalent ckt

- The capacitances lead to frequency effects in the transistor.
- The gain is a function of the input signal frequency.

* Short-circuit current Gain:-

We can neglect the parasitic resistances r_b, r_c & r_{ex}, the base-collector B−C diffusion resistance r_μ, and the substrate capacitance C_S. Collector is connected to signal ground.

- To understand the frequency effects of BJT we first determine the short circuit current gain



simplified hybrid π equivalent ckt for determining S.C. current gain

- We will determine small signal current gain $\boxed{A_i = \dfrac{I_c}{I_b}}$

writing a KCL equation at i/p node,

$$I_b = \frac{V_\pi}{r_\pi} + \frac{V_\pi}{\frac{1}{j\omega C_\pi}} + \frac{V_\pi}{\frac{1}{j\omega C_\mu}} \quad \text{--- eq}^n \text{ 1}$$

$$= V_\pi \left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu) \right]$$

But now $V_\pi \neq I_b r_\pi$ since a portion of I_b is now shunted through C_π & C_μ.

Applying KCL at o/p node,

$$\frac{V_\pi}{\frac{1}{j\omega C_\mu}} + I_c = g_m V_\pi \qquad I_c = V_\pi (g_m - j\omega C_\mu)$$

OR $\quad I_c = V_\pi (g_m - j\omega C_\mu)$

The i/p voltage V_π can be

$$V_\pi = \frac{I_c}{(g_m - j\omega C_\mu)} \quad \text{put this in eq}^n \text{ 1}$$

$$I_b = I_c \cdot \frac{\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]}{g_m - j\omega C_\mu}$$

The small signal current gain usually designated as h_{fe} becomes,

$$A_i = \frac{I_c}{I_b} = h_{fe} = \frac{(g_m - j\omega C_\mu)}{\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]}$$

If we assume typical ckt parameter values of

$C_\mu = 0.2\,pf$ & $f_{max} = 100\,MHz$

$g_m = 50\,mA/V$

then we see that $\omega C_\mu << g_m$

Therefore to a good approximation, the small signal

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current gain is

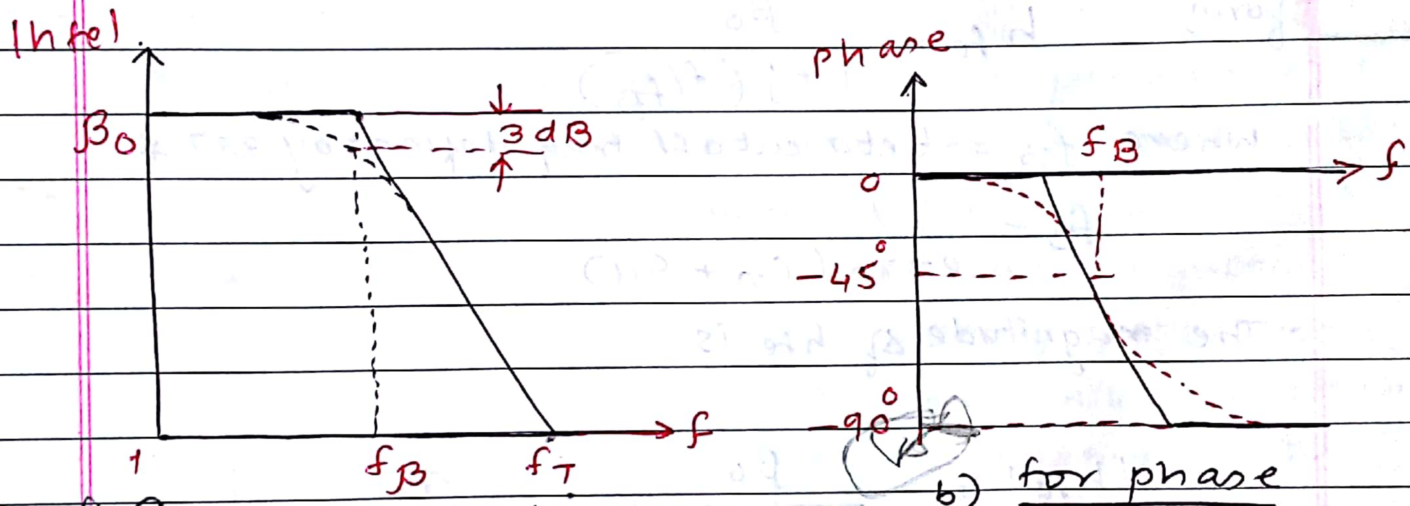
$$h_{fe} \simeq \frac{g_m}{\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]} = \frac{g_m r_\pi}{1 + j\omega r_\pi (C_\pi + C_\mu)}$$

L eqⁿ ②

since $g_m r_\pi = \beta$ then low freq current gain = is just β

Eqⁿ ② shows that in a BJT, the magnitude & phase of current gain are both functions of the frequency.

Bode plot of short ckt current gain magnitude -



fig② a) for mgnitude

b) for phase

- The corner frequency, which is also the beta cutoff frequency f_β in this case is given by

$$\boxed{f_\beta = \frac{1}{2\pi r_\pi (C_\pi + C_\mu)}} \quad — (A)$$

from fig b) above,

- As the frequency increases, the small signal collector current is no longer in phase with the small signal base current.
- At high frequencies, the collector current lags the i/p current by 90°.

* f_β is determined by a set of parameters employed in hybrid π Model.

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* **cut-off frequency :-**

Bode plot in previous fig a) shows that the magnitude of small signal current gain decreases with increasing frequency.

- At frequency f_T, which is the cutoff frequency, this gain goes to 1.
- The cut off frequency is a figure of merit of transistors.

From Eq^n (1) $h_{fe} \simeq \dfrac{g_m r_\pi}{1+j\omega r_\pi (C_\pi + C_\mu)}$

we can write the small signal current gain in the form $h_{fe} = \dfrac{\beta_0}{1+j(f/f_\beta)}$

where f_β = beta cutoff freq defined by eq^n *

$$f_\beta = \frac{1}{2\pi r_\pi (C_\pi + C_\mu)}$$

- The magnitude of h_{fe} is

$$|h_{fe}| = \frac{\beta_0}{\sqrt{1+\left(\frac{f}{f_\beta}\right)^2}} \quad (3)$$

- At cutoff freq f_T, $|h_{fe}| = 1$ & eq^n (3) can be modified as

$$|h_{fe}| = 1 = \frac{\beta_0}{\sqrt{1+\left(\frac{f_T}{f_\beta}\right)^2}}$$

Normally, $\beta_0 >> 1$, which implies that $f_T >> f_\beta$

$$\therefore \quad 1 \simeq \frac{\beta_0}{\sqrt{\left(\frac{f_T}{f_\beta}\right)^2}} = \frac{\beta_0 f_\beta}{f_T} \quad \text{or} \quad \boxed{f_T = \beta_0 f_\beta}$$

- f_β is also called bandwidth of the transistor.

{ f_T = gain BW product }

$\therefore f_T = \beta_0 f_\beta$, the cutoff freq f_T is the $\boxed{\text{gain bandwidth product}}$ of transistor, or more commonly the ~~gain~~ unity gain bandwidth.

from eq^n (A),

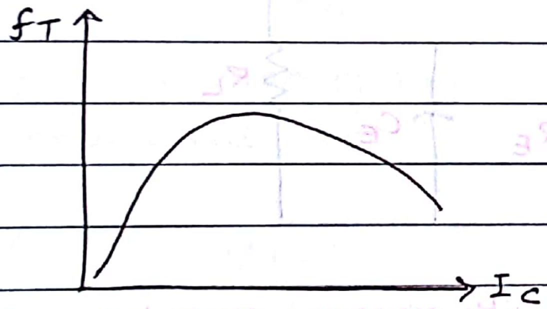
The unit gain bandwidth is

$$f_T = \beta_0 \left[\frac{1}{2\pi r_\pi (C_\pi + C_\mu)}\right] = \frac{g_m}{2\pi(C_\pi + C_\mu)}$$

(if $\beta_0 = g_m r_\pi$)

- since the capacitances are a function of the size of the transistor, we see again that high frequency transistors imply small device sizes.

- The cutoff frequency f_T is also a function of the dc collector current I_C, and the general characteristic of f_T versus I_C is shown in fig c.)



cut off frequency Vs. collector current

- The transconductance g_m is directly proportional to I_C, but only a portion of C_π is related to I_C.

- The cutoff frequency is therefore lower at low collector current levels.

However, the cut off frequency also decreases at high current levels, in the same way that β decreases at large currents.

$$\boxed{f_\beta = \frac{f_T}{\beta_0}}$$

- The cutoff frequency or unity gain bandwidth of a transistor is usually specified on the device data sheets

* Miller Effect and Miller Capacitance :-

- The Miller effect, or feedback effect, is a multiplication effect of C_μ in circuit applications.

- Next figure ① is a common-emitter circuit with a signal current source at the input.
We will determine the small-signal current gain $A_i = i_o/i_s$ of the circuit.

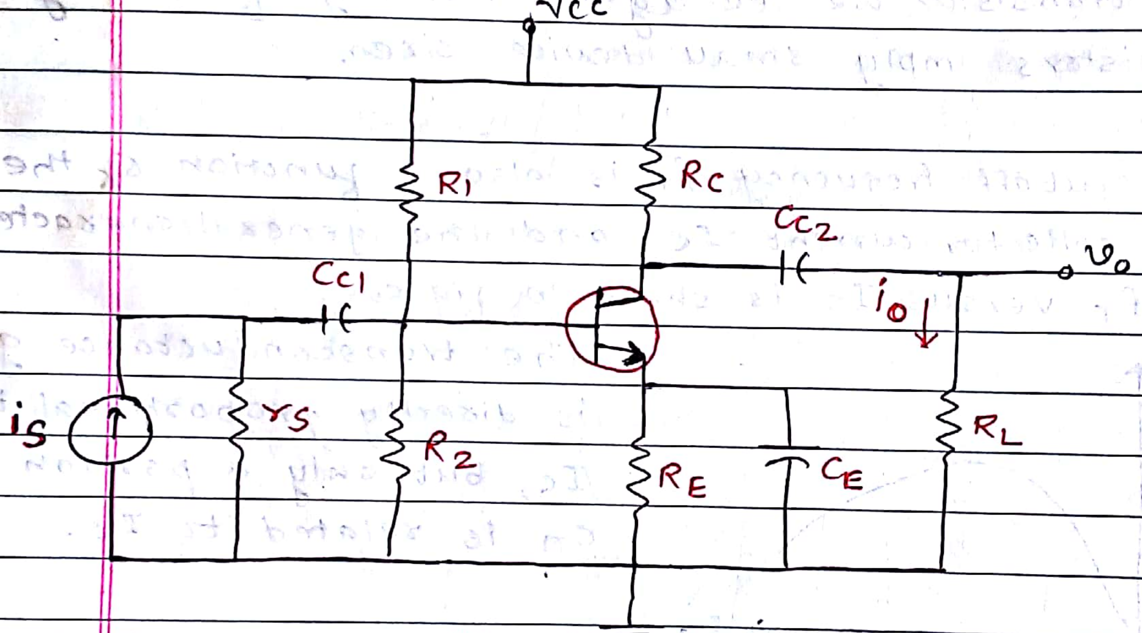


fig ① common-emitter ckt with ~~common~~ current source i/p

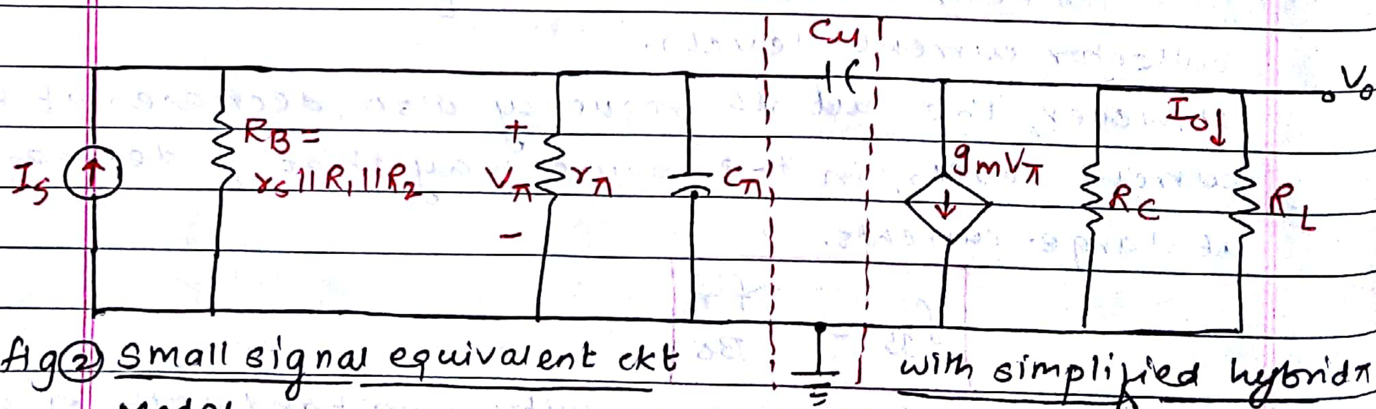


fig ② small signal equivalent ckt with simplified hybrid π Model

- fig ② is the small signal ~~current gain~~ equivalent ckt, assuming the frequency is sufficiently high for the coupling & bypass capacitors to act as short ckts.

- The capacitor C_μ is a feedback element that connects the output back to the i/p.
- The o/p vtg & current will therefore influence the input characteristics.
- We can determine the effect of C_μ on the input characteristics by finding an equivalent impedance Z_A across the plane A-A in fig ③, producing the equivalent ckt shown in fig ④.
- The current I_1 from figure ③ can be written as

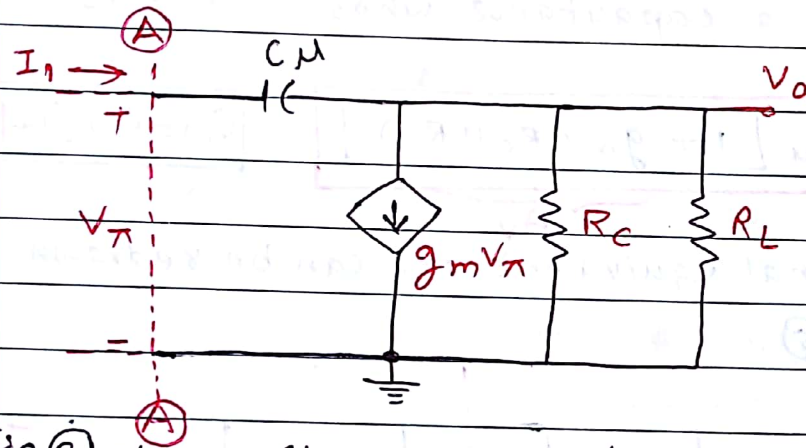


fig ③ o/p portion of small signal equivalent ckt

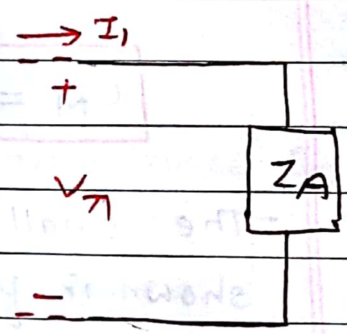


fig ④ equivalent impedance of this portion of the ckt

$$I_1 = \frac{V_\pi - V_o}{(1/j\omega C_\mu)} = (V_\pi - V_o)(j\omega C_\mu) \quad \text{—— ①}$$

summing the currents at o/p gives

$$I_1 = (V_\pi - V_o)(j\omega C_\mu) = g_m V_\pi + \frac{V_o}{R_C \| R_L}$$

$$V_\pi j\omega C_\mu - V_o j\omega C_\mu = g_m V_\pi + V_o / R_C \| R_L$$

or combining terms, we have

$$V_\pi (j\omega C_\mu - g_m) = V_o \left(\frac{1}{R_C \| R_L} + j\omega C_\mu \right)$$

But $|j\omega C_\mu| << g_m$

$$\therefore \quad -g_m V_\pi \cong V_o \left(\frac{1}{R_C \| R_L} \right)$$

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or $\boxed{V_0 = -g_m (R_C \| R_L) V_\pi}$

substituting this expression for V_0 into eqⁿ 1 yields

$$I_1 = \{V_\pi - [-g_m (R_C \| R_L) V_\pi]\} j\omega C_\mu$$

or

$$I_1 = V_\pi \cdot j\omega C_\mu [1 + g_m (R_C \| R_L)]$$

$\frac{V_\pi}{I_1} = Z_A$

$Z_A \cong C_M$

* $C_M = \frac{1}{j\omega C_M}$

- The equivalent impedance Z_A in fig 4 then corresponds to a capacitance whose value is

$$\boxed{C_M = C_\mu [1 + \underbrace{g_m (R_C \| R_L)}_{A_V}]}$$

$$\boxed{C_M = C_\mu [1 + |A_V|]}$$

- The small signal equivalent ckt can be redrawn as shown in fig 5.

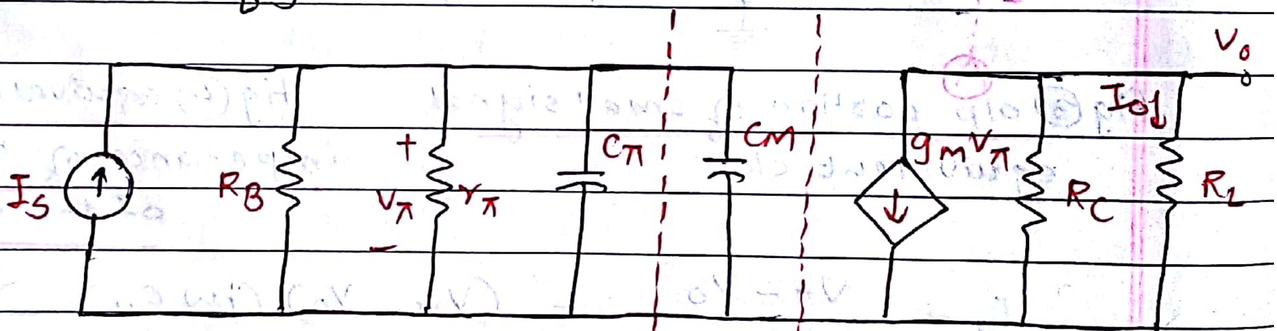


fig 5 small signal equivalent ckt including the equivalent Miller capacitance

- Capacitance C_M is called the Miller capacitance & the multiplication effect of C_μ is called the Miller effect.

- from fig 5, the i/p capacitance is now $C_\pi + C_M$ rather than just C_π if C_μ had been ignored.

Ex1) Determine the 3dB frequency of the short ckt current gain of a bipolar transistor.

consider a bipolar transistor with parameters $r_\pi = 2.6K\Omega$, $C_\pi = 2pf$ & $C_\mu = 0.1pf$

Sol: We know

$$f_\beta = \frac{1}{2\pi r_\pi (C_\pi + C_\mu)}$$

$$= \frac{1}{2\pi (2.6\times10^3)(2+0.1)(10^{-12})}$$

$$\boxed{f_\beta = 29.1MHz}$$

Ex2) calculate the bandwidth f_β & capacitance C_π of a bipolar transistor.

consider a transistor that has parameters $f_T = 500MHz$ at $I_C = 1mA$, $\beta_0 = 100$ & $C_\mu = 0.3pf$

Sol: Bandwidth is $f_\beta = \frac{f_T}{\beta_0} = \frac{500}{100} = 5MHz$

The transconductance is

$$g_m = \frac{I_C}{V_T} = \frac{1}{0.026} = 38.5mA/V$$

$$f_T = \frac{g_m}{2\pi(C_\pi + C_\mu)}$$

$$500\times10^6 = \frac{38.5\times10^{-3}}{2\pi(C_\pi + 0.3\times10^{-12})}$$

$$\boxed{C_\pi = 12.0pf}$$

Ex 3) Determine 3dB frequency of the current gain for the circuit shown in fig ⑤ (in previous page) both with & without effect of C_M.

The circuit parameters are:

$R_C = R_L = 4k\Omega$, $r_\pi = 2.6k\Omega$, $R_B = 200k\Omega$

$C_\pi = 4pf$, $C_\mu = 0.2pf$, $g_m = 38.5mA/V$

Soln

without effect of CM

$$f_{3dB} = \frac{1}{2\pi(R_B \| r_\pi)(C_\pi + C_M)}$$

neglecting effect of C_μ ($C_M = 0$)

$$f_{3dB} = \frac{1}{2\pi[(200\times10^3)\|(2.6\times10^3)](4\times10^{-12})}$$

$= 15.5\ MHz$

$$C_M = C_\mu[1 + g_m(R_C \| R_L)]$$

$= 0.2[1 + (38.5)(4\|4)]$

$C_M = 15.6pf$

with effect of CM

$$f_{3dB} = \frac{1}{2\pi(R_B \| r_\pi)(C_\pi + C_M)}$$

$$= \frac{1}{2\pi[(200\times10^3)\|(2.6\times10^3)][4+15.6](10^{-12})}$$

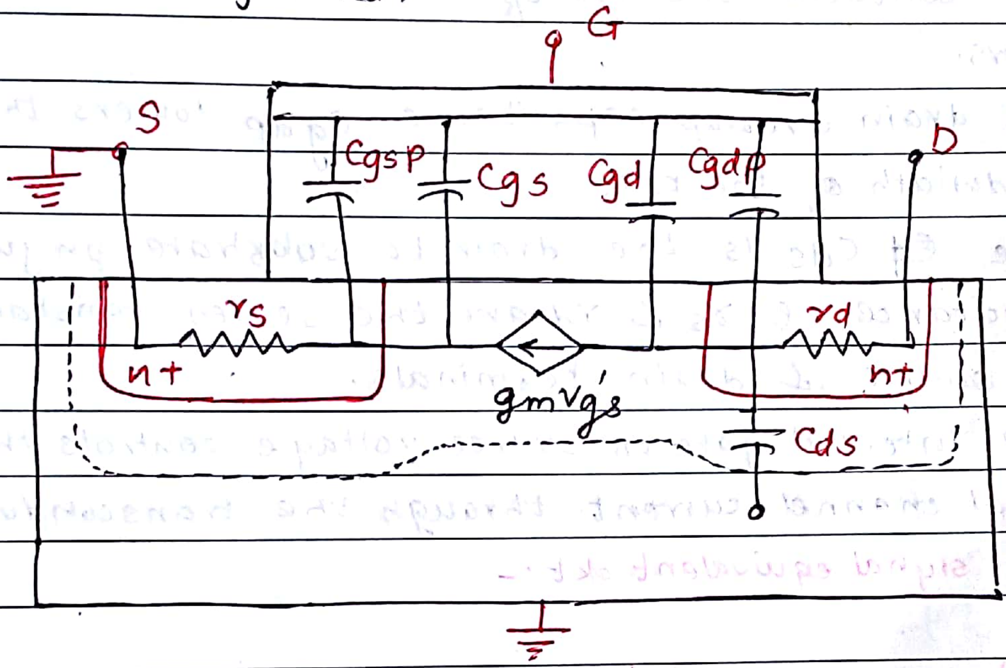
$$\boxed{f_{3dB} = 3.16MHz}$$

NOTE: If the gain is reduced then the Miller capacitance will be reduced & the bandwidth will be increased.

* High frequency Equivalent circuit of MOS FET –

Next figure shows a model based on the inherent capacitances & resistances in the n-channel MOSFET, as well as the elements representing the basic device equation.

assumption :– The source & substrate are both tied to ground.



(a) inherent resistance & capacitances in n-channel MOSFET

– Two capacitances C_{gs} & C_{gd} represent the interaction between the gate and the channel inversion charge near the source & drain elements resp.

– If the device is biased in the nonsaturation region & V_{DS} is small, the channel inversion charge is approximately uniform, which means that

$$C_{gs} \cong C_{gd} \cong \frac{1}{2} W L C_{ox}$$

where $C_{ox} (F/cm^2) = \epsilon_{ox}/t_{ox}$

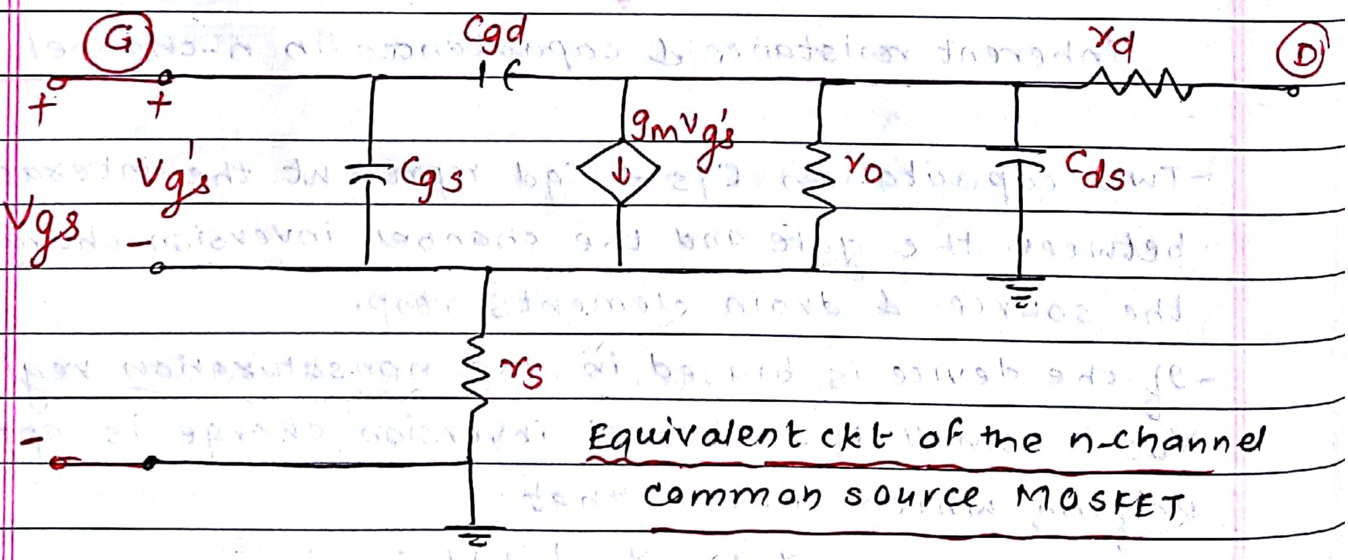
NOTE :– When the transistor is biased in the saturation region, the channel is pinched off at the drain & the inversion charge is no longer uniform. The value of C_{gd}

essentially goes to zero & C_{gs} approximately equals to $(2/3) W L C_{ox}$.

The value of C_{gs} changes as device size changes, but typical values are in the tenth of picofarad range.

- The remaining two capacitances C_{gsp} & C_{gdp} are parasitic or overlap capacitances so called because, in actual devices, the gate oxide overlaps the source & drain contacts, because of tolerances or other fabrication factors.
- The drain overlap capacitance C_{gdp} lowers the bandwidth of the FET.
- The ~~Cd~~ C_{ds} is the drain to substrate pn junction capacitance & r_s & r_d are the series resistances of the source & drain terminals.
- The internal gate to source voltage controls the small signal channel current through the transconductance.

Small signal equivalent ckt:-



Equivalent ckt of the n-channel common source MOSFET

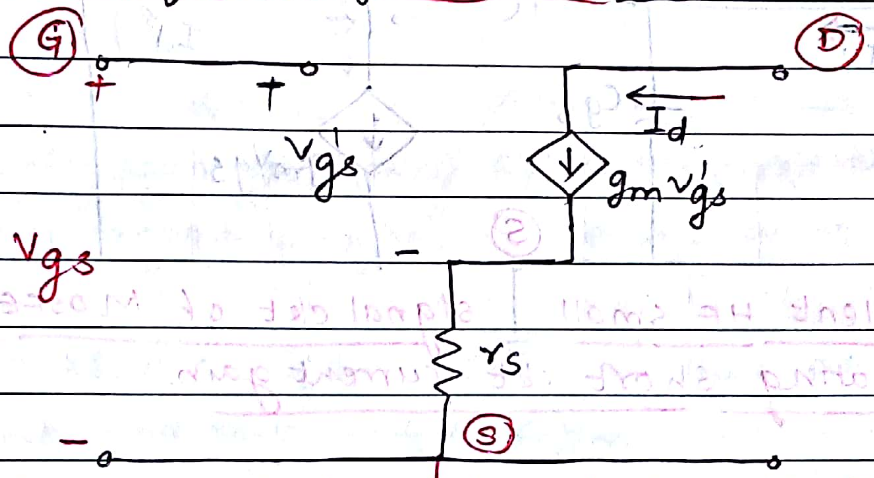
- V_{gs}' = internal gate to source vtg that controls the channel current.
- C_{gs} & C_{gd} contain the parasitic overlap capacitances.
- r_o is associated with the slope of I_D versus V_{DS}. In ideal MOSFET biased in the saturation region, I_D is

independent of V_{DS} which means r_o is infinite.

- However r_o is finite in short channel length devices, because of channel length modulation & is therefore included in the equivalent ckt.

- Source resistance r_s can have a significant effect on the transistor characteristics.

simplified low frequency equivalent ckt -



simplified LF equivalent ckt of n-channel CS MOSFET including r_s but not resistance r_o

- The drain current is $I_d = g_m V'_{gs}$

& The relationship betn V_{gs} & V'_{gs} is

$$V_{gs} = V'_{gs} + (g_m V'_{gs}) r_s$$
$$= (1 + g_m r_s) V'_{gs}$$

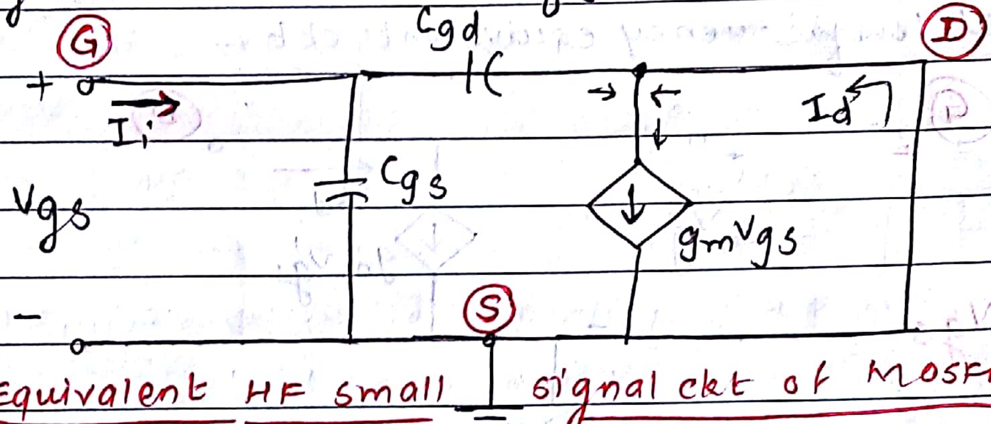
- The drain current can be written as

$$\boxed{I_d = \left(\frac{g_m}{1+g_m r_s}\right) V_{gs} = g'_m V_{gs}}$$

This eqn shows that source resistance reduces the effective transconductance or transistor gain.

* Unity Gain Bandwidth :-

- For a BJT, UGB or BW is a figure of merit for FETs.
- If we neglect r_s, r_d, r_o & C_{ds} and connect the drain to signal ground, the resulting equivalent small signal ckt is shown in figure.



Equivalent HF small signal ckt of MOSFET for calculating short ckt current gain.

- At HF, i/p gate impedance is no longer infinite, we can define short ckt current gain.
- From that we can define & calculate UGB.

writing KCL eqn at i/p node,

$$I_i = \frac{V_{gs}}{\frac{1}{j\omega C_{gs}}} + \frac{V_{gs}}{\frac{1}{j\omega C_{gd}}} \quad \text{--- (1)}$$

$$= V_{gs}[j\omega(C_{gs} + C_{gd})]$$

Apply KCL at ~~input~~ o/p node

$$\frac{V_{gs}}{\frac{1}{j\omega C_{gd}}} + I_d = g_m V_{gs}$$

OR

$$I_d = V_{gs}(g_m - j\omega C_{gd})$$

solving this eqn for V_{gs}

$$V_{gs} = \frac{I_d}{g_m - j\omega C_{gd}}$$ put this in eqn ①

$$I_i = I_d \frac{[j\omega(C_{gs}+C_{gd})]}{g_m - j\omega C_{gd}}$$

∴ small signal current gain

$$A_i = \frac{I_d}{I_i} = \frac{g_m - j\omega C_{gd}}{j\omega(C_{gs}+C_{gd})}$$

- ~~The unity gain frequency f_T is defined as frequency at which magnitude of the s.c. current~~

- If we assume $C_{gd} = 0.05\,pF$ & $g_m = 1\,mA/V$ & $f = 100\,MHz$ then we find that $\omega C_{gd} \ll g_m$.
- The small signal current gain to a good approximation is then,

$$\boxed{A_i = \frac{I_d}{I_i} \simeq \frac{g_m}{j\omega(C_{gs}+C_{gd})}}$$

- The unity gain frequency f_T is defined as the frequency at which the magnitude of the short ckt current gain goes to 1.

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

- The unity gain frequency or bandwidth is a parameter of the transistor & is independent of the ckt.

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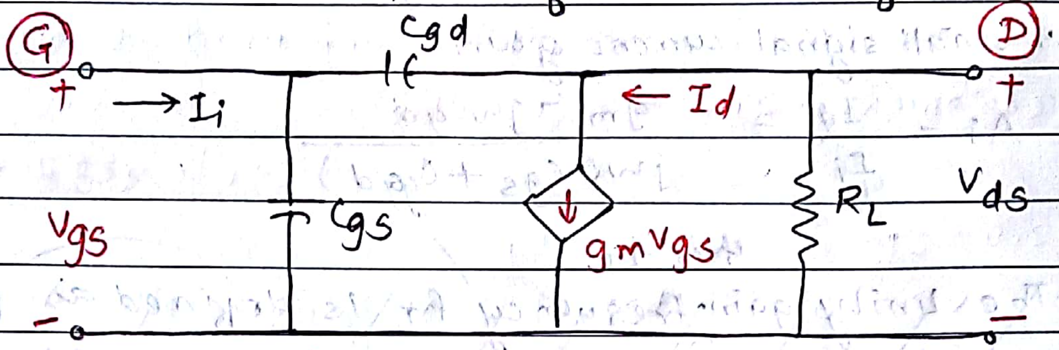
* Miller effect and Miller capacitance :-

- fig a) is a simplified high frequency transistor model, with a load resistor R_L connected to the o/p.
- We will determine the current gain in order to demonstrate the impact of the Miller effect.



fig@ Equivalent High freq small signal ckt of a MOSFET with a load resistance R_L.

- writing a KCL eqn at the i/p gate node, we have

$$I_1 = j\omega C_{gs} V_{gs} + j\omega C_{gd}(V_{gs} - V_{ds}) \quad \text{--- (1)}$$

where I_1 is the i/p current.

- summing currents at the o/p drain code we have,

$$\frac{V_{ds}}{R_L} + g_m V_{gs} + j\omega C_{gd}(V_{ds} - V_{gs}) = 0 \quad \text{--- (2)}$$

We can combine eqn (1) & (2) to eliminate vtg V_{ds}. The i/p current is

$$I_1 = j\omega \left\{ C_{gs} + C_{gd}\left[\frac{1 + g_m R_L}{1 + j\omega R_L C_{gd}}\right]\right\} V_{gs} \quad \text{--- (3)}$$

$(\omega R_L C_{gd})$ is much less than 1, therefore we can neglect $(j\omega R_L C_{gd})$ & eqn (3) becomes,

$$I_i = j\omega[C_{gs} + C_{gd}(1 + g_m R_L)] V_{gs} \quad \text{--- (4)}$$

Next fig (b) shows the equivalent ckt described by eqn (4).

The parameter C_M is the Miller capacitance & is given by

$$\boxed{C_M = C_{gd}(1+g_m R_L)} \quad \text{— (5)}$$

Eq. (5) clearly shows the effect of the parasitic drain overlap capacitance.

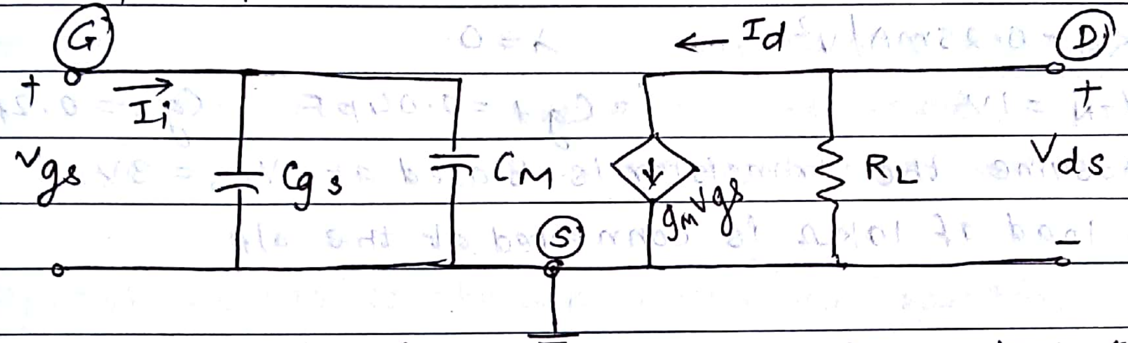


fig b) MOSFET high frequency ckt, including equivalent Miller capacitance.

NOTE: C_{gd} is the overlap capacitance (or total gate to drain capacitance). This overlap capacitance is multiplied because of the Miller effect & may become a significant factor in the bandwidth of an amplifier.

– The cut off frequency f_T of a MOSFET is defined as the frequency at which the current gain = 1, or magnitude of i/p current I_i is equal to the ideal load current I_d.

– From fig (b), we see that

$I_i = j\omega(C_{gs}+C_M)V_{gs}$ & ideal load current is

$I_d = g_m V_{gs}$

∴ magnitude of current gain is

$$|A_i| = \left|\frac{I_d}{I_i}\right| = \frac{g_m V_{gs}}{j\omega(C_{gs}+C_M)V_{gs}}$$

$$\boxed{A_i = \frac{g_m}{2\pi f(C_{gs}+C_M)}} \quad \text{— (6)}$$

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setting this eq^n (= 1) equal to 1, we find cut off freq

$$f_T = \frac{g_m}{2\pi(C_{gs}+C_M)} = \frac{g_m}{2\pi C_G}$$

where C_G = i/p gate capacitance.

Ex 1) Determine Miller capacitance & cut off freq of MOSFET ckt. & Determine unity gain bandwidth of an MOSFET. consider an n-channel MOSFET with parameters

$K_n = 0.25 mA/V^2$ $\lambda = 0$

$V_{TN} = 1V$ $C_{gd} = 0.04pF$ $C_{gs} = 0.2pF$

Assume the transistor is biased at $V_{GS} = 3V$. & load of $10k\Omega$ is connected at the o/p.

Sol^n:-

The Transconductance is

$$g_m = 2K_n(V_{GS} - V_{TN})$$
$$= 2(0.25)(3-1)$$
$$\boxed{g_m = 1 mA/V}$$

without C_M

Unity-Gain Bandwidth is (or freq) is

$$f_T = \frac{g_m}{2\pi(C_{gs}+C_{gd})} = \frac{10^{-3}}{2\pi(0.2+0.04)\times 10^{-12}}$$

$$\boxed{f_T = 6.63\times 10^8 Hz} \quad or \quad \boxed{f_T = 663 MHz}$$

Miller capacitance is

$$C_M = C_{gd}(1+g_m R_L)$$
$$= 0.04[1+(1)(10)]$$
$$\boxed{C_M = 0.44pf}$$

with C_M

cut off frequency

$$f_T = \frac{g_m}{2\pi(C_{gs}+C_M)} = \frac{10^{-3}}{2\pi(0.2+0.44)\times 10^{-12}} = 2.49\times 10^8 Hz$$

$$\boxed{f_T = 249 MHz}$$

* **NOTE**: Miller capacitance reduce the cutoff frequency of MOSFET as well as BJT

* Frequency Response : Transistor Amplifiers with circuit capacitors :-

- Three types of capacitors are considered.

① coupling capacitor ② Load capacitors ③ Bypass capacitors.

① Coupling capacitor Effects :-

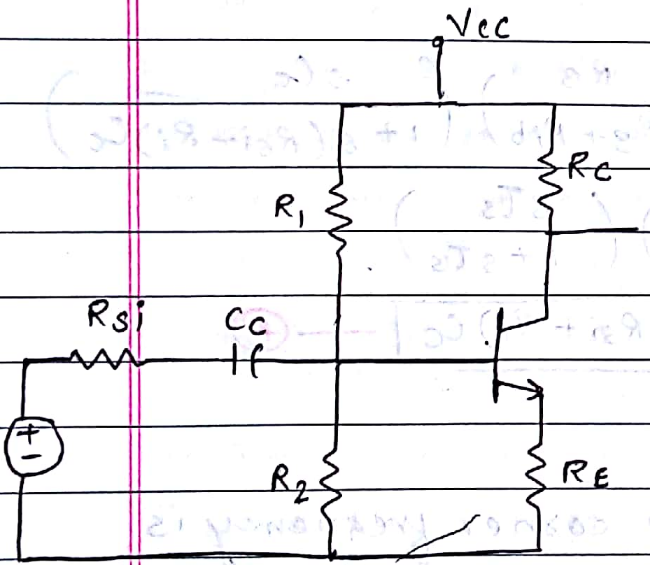
Input coupling capacitor : common emitter circuit

* **Note** :- we will use a current-voltage analysis to determine the frequency response of the circuit

Then, we will use the equivalent time constant technique.

fig a) shows BJT CE ckt with a coupling capacitor, fig b) shows corresponding small signal equivalent ckt.

r_o (small signal o/p resistance) is assumed to be infinite ($\because$ $r_o >> R_C$ & $r_o >> R_E$)



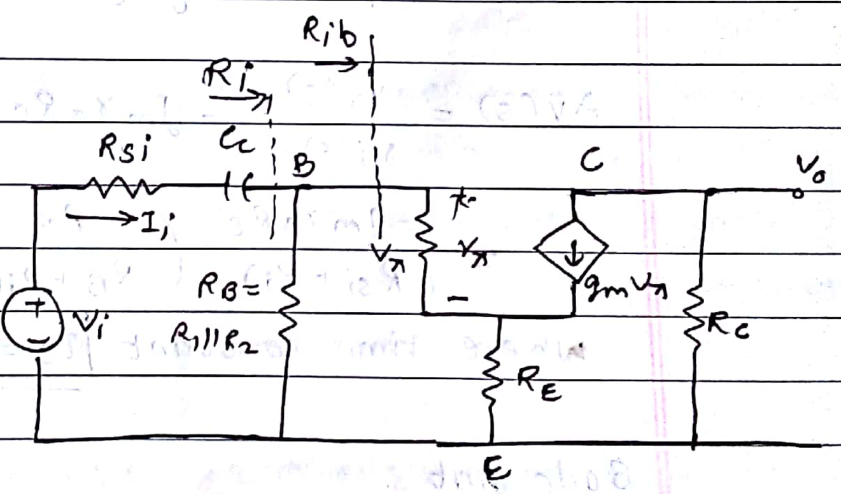


fig b) small signal equivalent ckt

fig 1.2 / fig a) CE ckt with coupling capacitor

Ⓐ Current - Voltage Analysis :-

- The I/p current can be written as

$$I_i = \frac{V_i}{R_{si} + \frac{1}{sC_s} + R_i}$$

where $R_i = R_B \| [r_\pi + (1+\beta)R_E] = R_B \| R_{ib} = R_B \| R_{ib}$

$= R_B \| R_{ib}$

$$I_i = \frac{V_i}{R_{si} + \frac{1}{sC_c} + R_i} \quad \text{--- (1)}$$

where R_i is the input resistance $= R_B \| [r_\pi + (1+\beta)R_E]$
$= R_B \| R_{ib}$

Using current divider,

$$I_b = \left(\frac{R_B}{R_B + R_{ib}}\right) I_i \quad \text{(2)} \quad \& \quad \text{then} \quad V_\pi = I_b r_\pi \quad \text{--- (3)}$$

o/p vtg is given by

$V_o = -g_m V_\pi R_C$ putting eqn (2) & (3)

$$= -g_m R_C (I_b r_\pi)$$

$$= -g_m r_\pi R_C \left(\frac{R_B}{R_B + R_{ib}}\right) I_i$$

$$V_o = -g_m r_\pi R_C \left(\frac{R_B}{R_B + R_{ib}}\right)\left(\frac{V_i}{R_{si} + \frac{1}{sC_c} + R_i}\right)$$

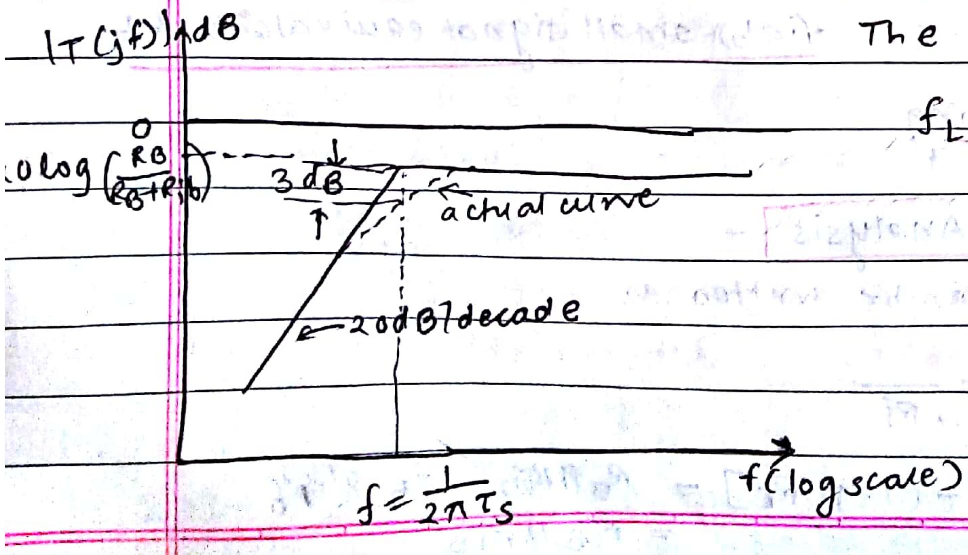
$\therefore$ small signal voltage gain is

$$A_V(s) = \frac{V_o(s)}{V_i(s)} = -g_m r_\pi R_C \left(\frac{R_B}{R_B + R_{ib}}\right)\left(\frac{sC_c}{1 + s(R_{si} + R_i)C_c}\right)$$

$$= \frac{-g_m r_\pi R_C}{(R_{si} + R_i)}\left(\frac{R_B}{R_B + R_{ib}}\right)\left(\frac{s\tau_s}{1 + s\tau_s}\right)$$

where time constant $\boxed{\tau_s = (R_{si} + R_i) C_c}$ --- (*)

Bode plot -



The corner frequency is

$$f_L = \frac{1}{2\pi\tau_s} = \frac{1}{2\pi(R_{si} + R_i)C_c}$$

The maximum magnitude in decibels is

$$|A_{v(max)}|_{dB} = 20\log_{10}\left(\frac{g_m r_\pi R_C}{R_{si}+R_i}\right)\left(\frac{R_B}{R_B+R_{ib}}\right)$$

(B) Time constant Technique –

- In general, we do not need to derive the complete circuit transfer function including capacitance effects in order to complete the bode plot & determine the frequency response.
- By looking at a circuit, we can determine if the amplifier is low-pass or high pass circuit.
We can specify the Bode plot if we know the time constant & the maximum midband gain.

- The time constant determines the corner frequency.
- The time constant technique yields good results when all poles are real.
(NOTE: This technique does not determine corner frequencies due to system zeros.)
- The major benifit of using the time constant approach is that it gives information about which circuit elements affect the −3dB frequency of the ckt.

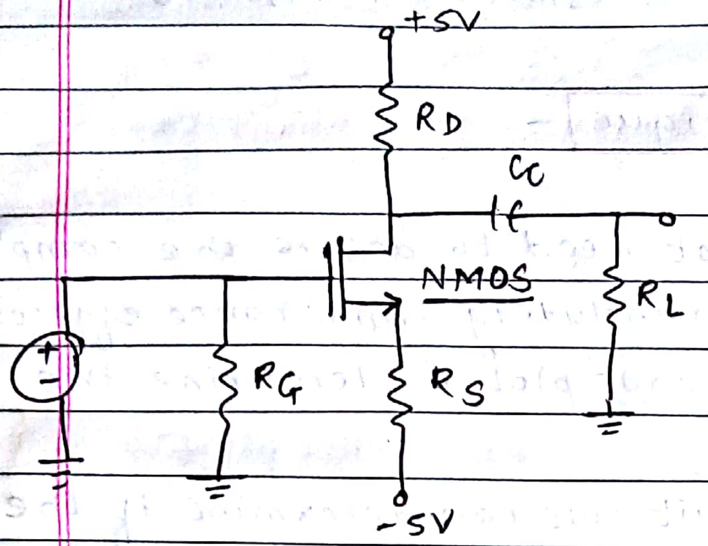
- The time constant for the ckt is a function of the equivalent resistance seen by the capacitor.
- from fig (b) (previous page), the equivalent resistance seen by the coupling capacitor C_C is $(R_{si}+R_i)$.
∴ The time constant is

$$\boxed{\tau_s = (R_{si}+R_i)\, C_C} \quad (*)$$

where $R_i = R_B \| [r_\pi + (1+\beta)R_E]$

so this is same as that determined by using current voltage analysis

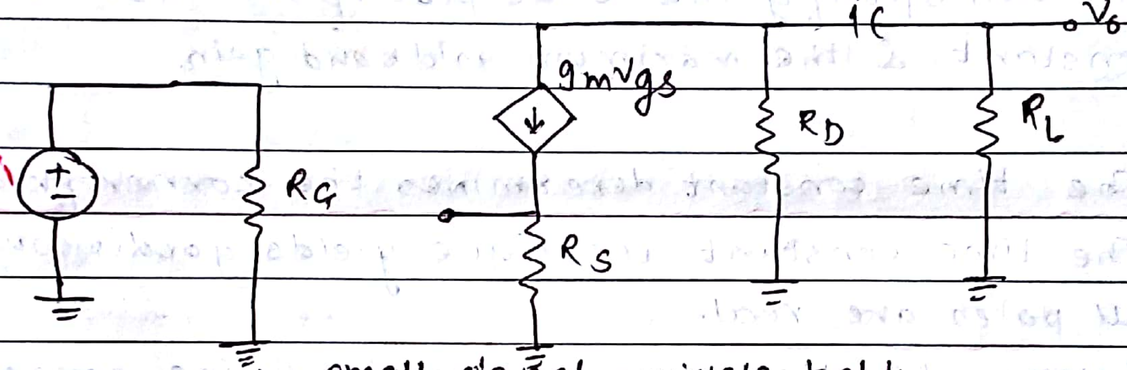
* Output coupling capacitor – Common source circuit :–



– o/p signal is connected to the load through a coupling capacitor.

Note – the resistance of the signal generator is much less than R_G. (& therefore be neglected.)

fig 1.1 CS ckt with o/p coupling capacitor.



small-signal equivalent ckt

– The Time constant is a function of the effective resistance seen by capacitor C_c, which is determined by setting all independent sources equal to zero.

, $V_i = 0$ then $V_{gs} = 0$ & $g_m V_{gs} = 0$

∴ effective resistance seen by C_c is $(R_D + R_L)$

∴ The time constant is

$$\boxed{\tau_s = (R_D + R_L) C_c}$$

& corner frequency is $\boxed{f_L = \frac{1}{2\pi \tau_s}}$

* Output coupling capacitor : Emitter follower circuit :-

- Coupling capacitor C_{C1}, is very large & it acts as a short circuit to the i/p signal.

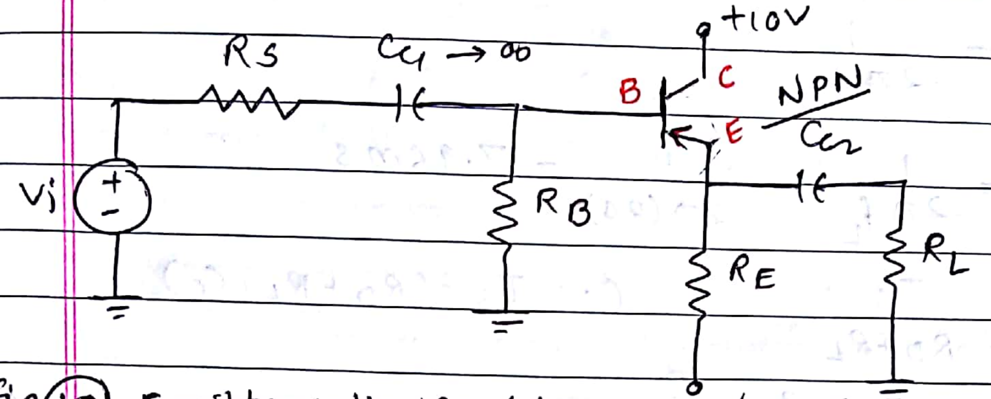
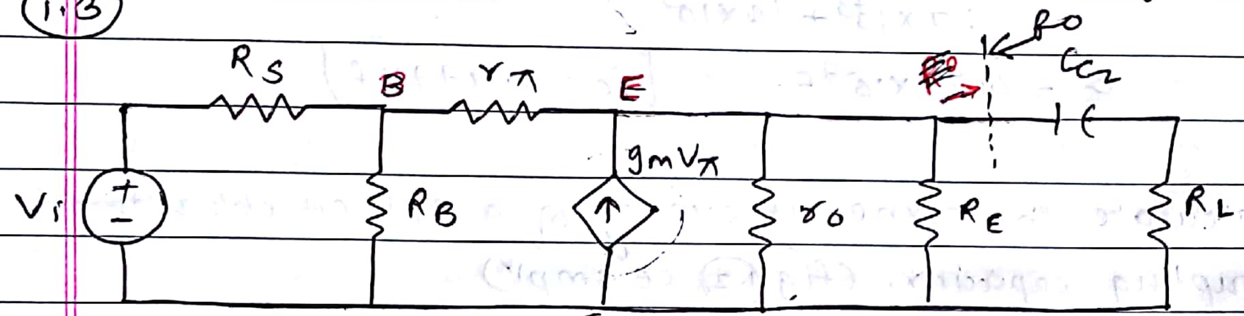


fig (1.3) Emitter follower ckt -10V with o/p coupling capacitor.



small-signal equivalent ckt

- The equivalent resistance seen by the coupling capacitor C_{C2} is $(R_o + R_L)$ & The time constant is

$$\boxed{\tau_S = [R_o + R_L] C_{C2}}$$

where R_o = o/p resistance

$$R_o = R_E \parallel r_o \parallel \left\{ \frac{[r_\pi + (R_S \parallel R_B)]}{1+\beta} \right\}$$

NOTE - To find effect of one capacitor, assume all other capacitor as short circuited.

To find B to E

Effective Resistance from Left terminal to Right ⇒ divide by $(1+\beta)$

from Right (E) to left term (B) ⇒ multiply by $(1+\beta)$

in fig (1.1) previous page (CS ampl^r)

Ex1) The ckt given is to be used as a simple audio amplifier. Design a circuit such that the lower corner frequency

$f_L = 20Hz$ $R_D = 6.7k\Omega$ $R_S = 5k\Omega$ $V^+ = 5$

$R_G = 50k\Omega$ $R_L = 10k\Omega$ $V^- = -5V$

Sol?- The corner freq in terms of time constant

$$f_L = \frac{1}{2\pi\tau_S}$$

$$\tau_S = \frac{1}{2\pi f_L} = \frac{1}{2\pi(20)} = 7.96ms$$

$\tau_S = (R_D + R_L) C_C$

$$C_C = \frac{\tau_S}{R_D + R_L} \quad (\because \tau_S = (R_D + R_L) C_C)$$

$$= \frac{7.96 \times 10^{-3}}{6.7 \times 10^3 + 10 \times 10^3}$$

$$C_C = 4.77 \times 10^{-7} F \qquad \boxed{C_C = 0.477\mu f}$$

Ex2) Calculate the corner frequency of a BJT CE ckt with a coupling capacitor. (fig (1.2) CE ampl^r)

The ckt parameters are : $R_1 = 51.2k\Omega$ $R_2 = 9.6k\Omega$

$R_C = 2k\Omega$ $R_E = 0.4k\Omega$ $R_{si} = 0.1k\Omega$ $C_C = 1\mu f$ &

$V_{CC} = 10V$, $V_{BE(on)} = 0.7V$, $\beta = 100$ & $V_A = \infty$ & $I_{CQ} = 1.81mA$

Sol?-

$$g_m = \frac{I_{CQ}}{V_T} = \frac{1.81}{0.026} = 69.6 mA/V$$

$$r_\pi = \frac{\beta V_T}{I_{CQ}} = \frac{100 \times 0.026}{1.81} = 1.44k\Omega$$

The i/p Resistance

$$R_i = R_1 \| R_2 \| [r_\pi + (1+\beta) R_E]$$

$$= 51.2 \| 9.6 \| [1.44 + (101)(0.4)]$$

$$= 6.77k\Omega$$

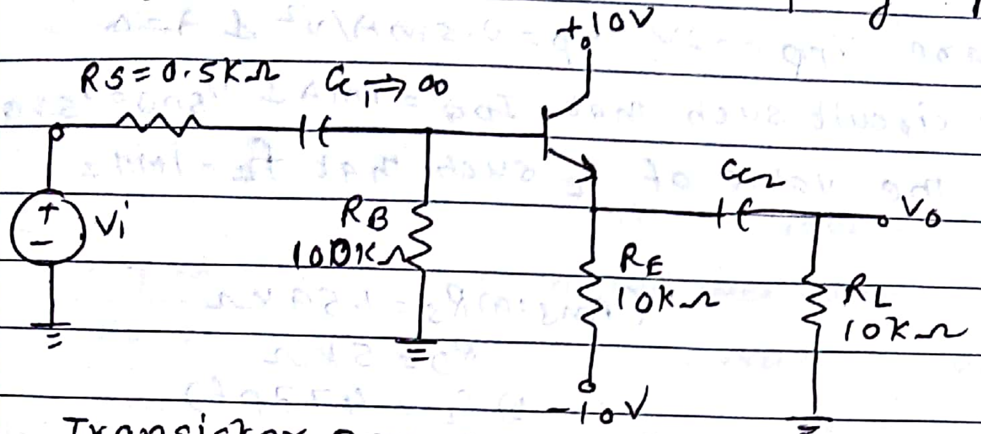
$$\tau_S = (R_{si} + R_i) C_C$$

$$= (0.1 \times 10^3 + 6.77 \times 10^3)(1 \times 10^{-6})$$

$$= 6.87 \times 10^{-3} s$$

$\boxed{\tau_S = 6.87ms}$ & $f_L = \frac{1}{2\pi\tau_S} = 23.2Hz$ $\boxed{f_L = 23.2Hz}$

3) Determine the 3dB frequency of an emitter follower amplifier circuit with an o/p coupling capacitor. fig(1.3)



Transistor parameters $\beta = 100$, $V_{BE(ON)} = 0.7V$, $V_A = 120V$

The o/p coupling capacitance is $C_{C2} = 1\mu f$, $I_{CQ} = 0.838mA$

$r_o = 143K\Omega$

$$r_\pi = \frac{\beta V_T}{I_{CQ}} = \frac{100 \times 0.026}{0.838 \times 10^{-3}} = 3.10K\Omega$$

$$g_m = \frac{I_{CQ}}{V_T} = \frac{0.838 \times 10^{-3}}{0.026} = 32.2mA/V$$

$$R_o = R_E \parallel r_o \parallel \left\{ \frac{[r_\pi + (R_S \parallel R_B)]}{1+\beta} \right\}$$

$$= 10 \parallel 143 \parallel \left\{ \frac{[3.10 + (0.5 \parallel 100)]}{101} \right\}$$

$$\boxed{R_o \cong 35.5\Omega}$$

$$\tau_s = [R_o + R_L] C_{C2}$$

$$= [35.5 + 10^4](10^{-6})$$

$$\tau_s = 1 \times 10^{-2} s$$

$$f_L = \frac{1}{2\pi \tau_s} = \frac{1}{2\pi(10^{-2})} = 15.9Hz$$

$$\boxed{f_L = 15.9Hz}$$

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Ex4) The PMOS CS ckt shown in fig ① (next page)

H.W has a load resistance $R_L = 10k\Omega$. The transistor parameters are $V_{TP} = -2V$, $K_p = 0.5mA/V^2$ & $\lambda = 0$.

ⓐ Design the circuit such that $I_{DQ} = 1mA$ & $V_{SDQ} = V_{SGQ}$

ⓑ Determine the value of C_L such that $f_H = 1MHz$

(Ans: a) $R_S = 1.59k\Omega$
$R_D = 5k\Omega$
b) $C_L = 47.7pF$)

The corner frequency

$$f_L = \frac{1}{2\pi \tau_s} = \frac{1}{2\pi(6.87\times10^{-3})} = 23.2\,Hz$$

* Load capacitor Effects :-

The model of the load circuit input impedance in parallel with a resistance. In addition there is a parasitic capacitance between ground and the line that connects the amplifier o/p to the load circuit.

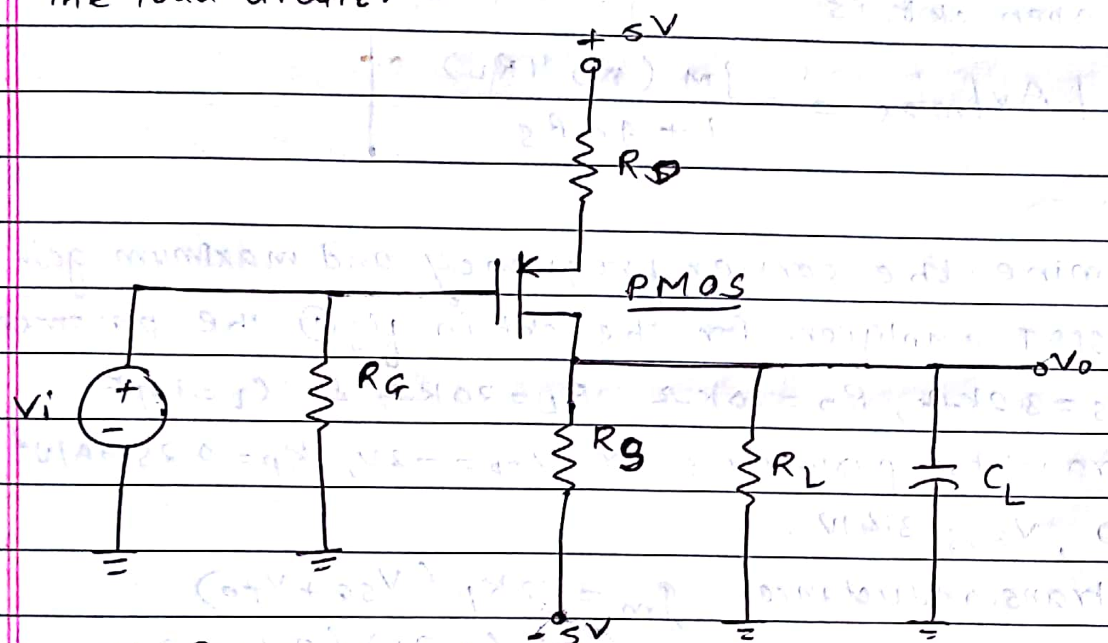

fig ① MOSFET CS ckt with a load capacitor

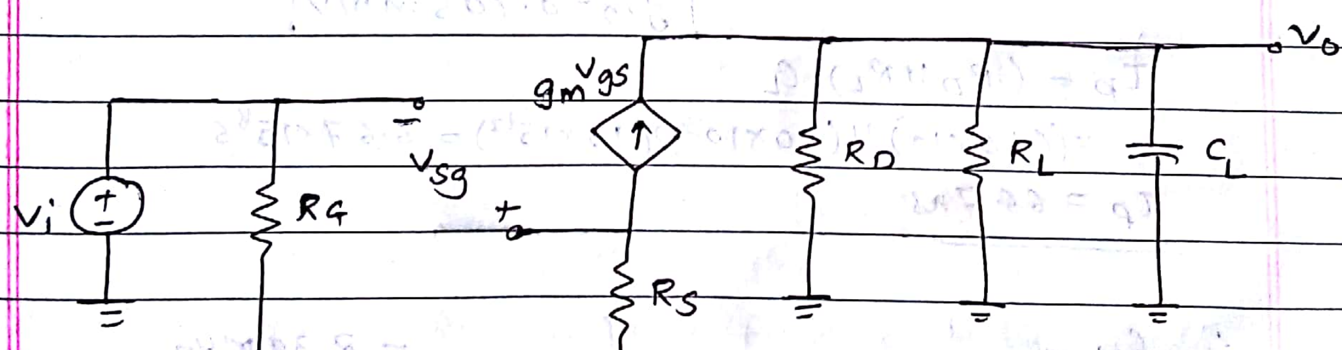

fig ② small signal equivalent ckt

- fig ① shows a MOSFET CS ampl^r with a load resistance R_L & load capacitance C_L connected to the o/p.
fig ② shows the small signal equivalent ckt.
o/p resist^r r_o is assumed to be infinite.

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- The equivalent resistance seen by the load capacitor C_L is $(R_D \| R_L)$

Since we set $V_i = 0$ then $g_m V_{gs} = 0$ which means that the dependent current source does not affect the equivalent resistance.

$\therefore$ The time constant for this ckt is

$$\boxed{\tau_p = (R_D \| R_L) C_L}$$

- The maximum gain, which is found by assuming C_L is an open ckt is

$$\boxed{|A_v|_{max} = \frac{g_m (R_D \| R_L)}{1 + g_m R_S}}$$

Ex 12) Determine the corner frequency and maximum gain of a MOSFET amplifier. for the ckt in fig ①, the parameters are $R_S = 3.2\,k\Omega$, $R_D = 10\,k\Omega$, $R_L = 20\,k\Omega$ & $C_L = 10\,pF$. The transistor parameters are $V_{TP} = -2V$, $K_p = 0.25\,mA/V^2$ & $\lambda = 0$, $V_{SGQ} = 3.41V$

Soln The transconductance $g_m = 2K_p (V_{SG} + V_{TP})$

$= 2(0.25)(3.41 - 2)$

$$\boxed{g_m = 0.705\,mA/V}$$

$\tau_p = (R_D \| R_L) \cdot C_L$

$= ((10\times10^3) \| (20\times10^3))(10\times10^{-12}) = 6.67\times10^{-8}\,s$

$\tau_p = 66.7\,ns$

$\therefore f_H = \frac{1}{2\pi \tau_p} = \frac{1}{2\pi(66.7\times10^{-9})} = 2.39\,MHz$

$|A_v|_{max} = \frac{g_m (R_D \| R_L)}{1 + g_m R_S} = \frac{0.705(10\|20)}{1 + (0.705\times3.2)} = 1.44$

$$\boxed{|A_v|_{max} = 1.44}$$

Next Example ⊛ (previous page)

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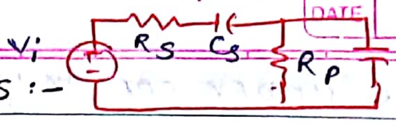
At LF, we treat load capacitor as O.C $\Rightarrow T_S \Rightarrow$ T.C. associated with C_S

At HF, we treat coupling capacitor as S.C. $\Rightarrow T_P =$ T.C associated with C_P

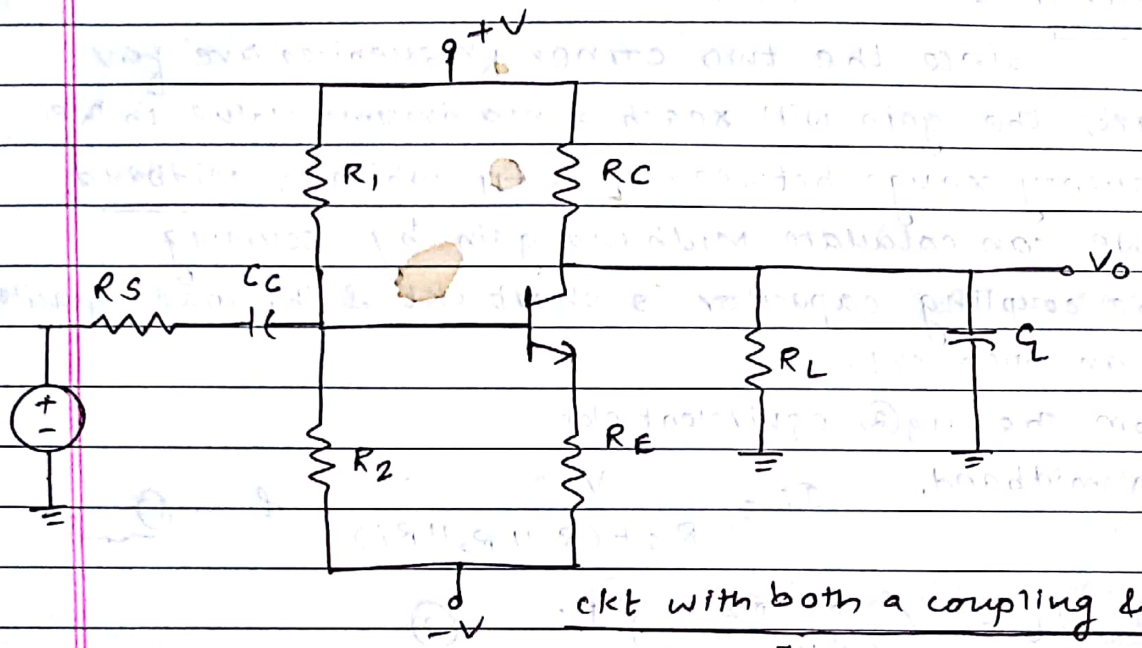


$T_S = (R_S + R_P) C_S$

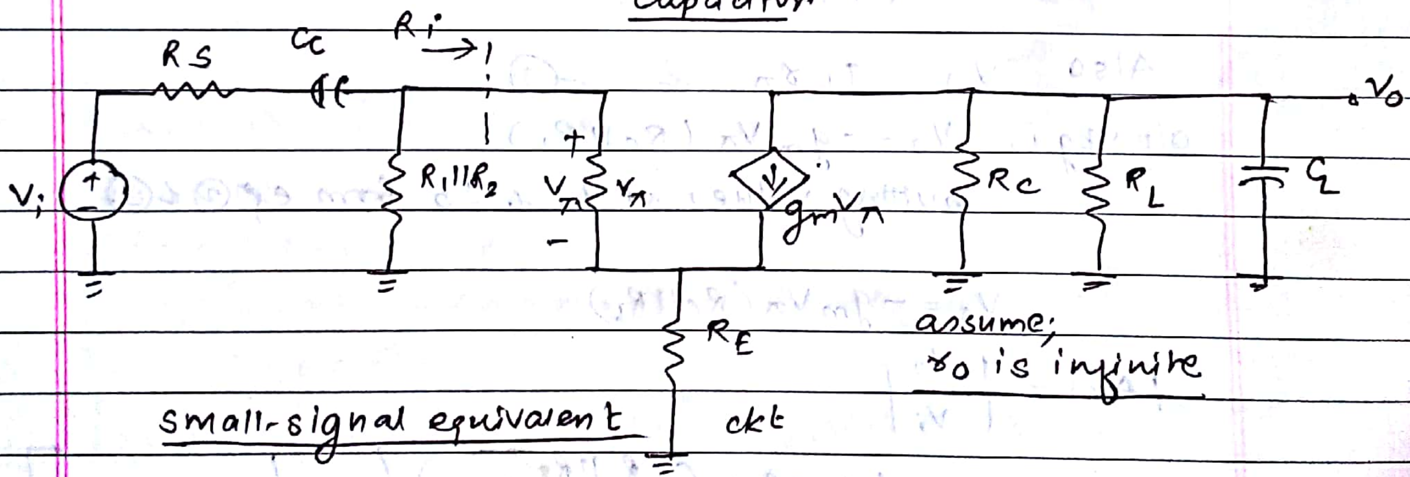
$T_P = (R_S \| R_P) C_P$

* coupling and Load capacitors :-

- since the values of the coupling capacitance & load capacitance differ by orders of magnitude, the corner frequencies are far apart & can be treated separately.



ckt with both a coupling & a load capacitor.



assume; r_o is infinite

small-signal equivalent ckt

- The lower corner frequency f_L is given by $f_L = \frac{1}{2\pi T_S}$

where T_S = time constant associated with C_C

$$T_S = [R_S + (R_1 \| R_2 \| R_i)] C_C$$

where $R_i = r_\pi + (1+\beta) R_E$

(T_S is equivalent resistance associated with the coupling capacitor by setting signal source to zero)

- The upper corner frequency f_H is given by

$$f_H = \frac{1}{2\pi \tau_p}$$

$$\tau_p = (R_C \| R_L) C_L$$ = Time constant related to C_L

Midband :-

since the two corner frequenices are far apart, the gain will reach a maximum value in the frequency range between f_L & f_H which is Midband.

- We can calculate Midband gain by assuming the coupling capacitor is short ckt & the load capacitor is an open ckt.

from the fig②, equivalent ckt in midband,

$$I_i = \frac{V_i}{R_s + (R_1 \| R_2 \| R_i)} \quad \& \quad \text{—①}$$

$$I_b = \left(\frac{R_1 \| R_2}{(R_1 \| R_2) + R_i}\right) I_i \quad \text{—②}$$

Also $V_\pi = I_b r_\pi$ & —③

o/p vtg is $V_o = -g_m V_\pi (R_C \| R_L)$

putting values of V_π & I_b from eq② & ③

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

$$|A_V| = \left|\frac{V_o}{V_i}\right|$$

$$= g_m r_\pi (R_C \| R_L) \left(\frac{R_1 \| R_2}{(R_1 \| R_2) + R_i}\right) \left(\frac{1}{[R_s + (R_1 \| R_2 \| R_i)]}\right) \quad \text{—④}$$

* Determine the midband gain, corner frequencies and bandwidth of a circuit containing both a coupling capacitor & a load capacitor.

Consider the ckt in fig a) with transistor parameters $V_{BE(ON)} = 0.7V$, $\beta = 100$, & $V_A = \infty$, $I_{CQ} = 0.99mA$

Soln:-

$$g_m = \frac{I_{CQ}}{V_T} = \frac{0.99}{0.026} = 38.1 mA/V$$

$$r_\pi = \frac{\beta V_T}{I_{CQ}} = \frac{(100)(0.026)}{0.99} = 2.63 k\Omega$$

i/p resistance R_i

$$R_i = r_\pi + (1+\beta) R_E = 2.63 + (101)(0.5) = 53.1 k\Omega$$

from Eqn (4),

$$|A_V|_{max} = \left|\frac{V_o}{V_i}\right|_{max}$$

$$= g_m r_\pi (R_C \| R_L) \left(\frac{R_1 \| R_2}{(R_1 \| R_2) + R_i}\right)\left(\frac{1}{[R_S + (R_1 \| R_2 \| R_i)]}\right]$$

$$= (38.1)(2.63)(5 \| 10)\left(\frac{40 \| 5.7}{(40 \| 5.7) + 53.1}\right)\left(\frac{1}{[0.1 + (40 \| 5.7 \| 53.1)]}\right)$$

$$\boxed{|A_V|_{max} = 6.16}$$

$$\tau_S = (R_S + R_1 \| R_2 \| R_i) C_C$$
$$= (0.1 \times 10^3 + (5.7 \times 10^3)] \| (40 \times 10^3) \| 53.1 \times 10^3))(10 \times 10^{-6})$$
$$= 4.46 \times 10^{-2} s$$

$$\boxed{\tau_S = 46.6 ms}$$

$$\tau_P = (R_C \| R_L) C_L$$
$$= ((5 \times 10^3) \| (10 \times 10^3))(15 \times 10^{-12})$$
$$= 5 \times 10^{-8} s$$

$$\boxed{\tau_P = 50 ns}$$

$$f_L = \frac{1}{2\pi \tau_S} = \frac{1}{2\pi(46.6 \times 10^{-3})} = 3.42 Hz$$

$$f_H = \frac{1}{2\pi \tau_P} = \frac{1}{2\pi(50 \times 10^{-9})} = 3.18 MHz$$

$BW = f_H - f_L$

$$\boxed{BW \simeq 3.18 MHz}$$

* Frequency Response of BJT Amplifier :-

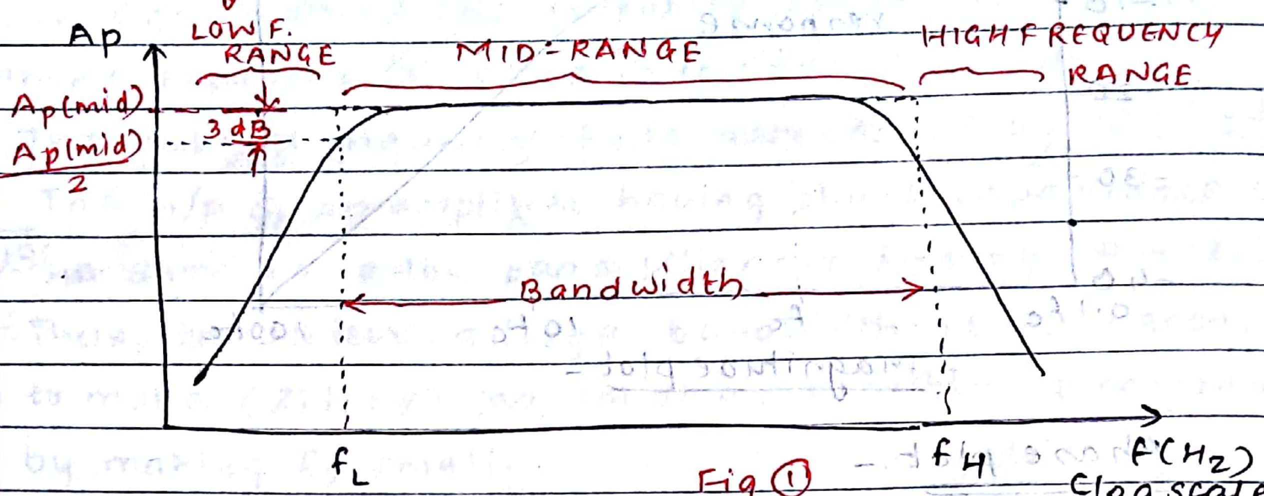


Fig ①

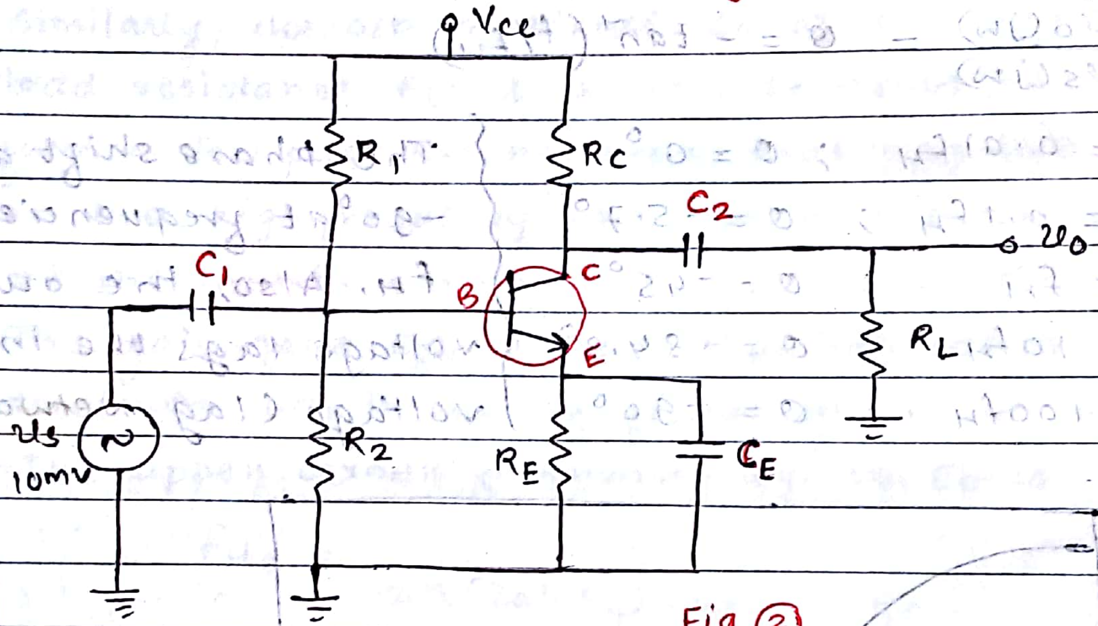


Fig ②

* In ~~HF~~ & MF region, short circuit ~~All the~~ coupling & bypass capacitors.

$$X_{C_1} = \frac{1}{2\pi F C_1} \qquad 2\pi \ \& \ C_1 \text{ are constant}$$

$$\downarrow X_{C_1} \propto \frac{1}{F\uparrow}$$

* In HF region, coupling & bypass capacitor are short circuited. & Av decreases due to junction capacitance, exist in the f_H.

— In $\boxed{\text{LF}}$ region, gain increases due to coupling & bypass capacitor. as we know $X_C \propto 1/F$

ie As $F\uparrow$, $X_C\downarrow$ & then more & more current flow

through the i/p terminal and o/p current increased &
∴ o/p vtg increases, ie A_v also increase.

- At Freq f_L (lower cut off freq), coupling & bypass capacitor will become short.
- The frequency region between lower cut off freq (f_L) & upper cut off freq (f_H) is called mid frequency region.
- At $\boxed{MF}$, these capacitors will remain short circuited. so a constant i/p current will flow through the amplifier so o/p current constant & o/p vtg remains constant. ie vtg gain remains constant in MF region.

- At a freq f_H (Upper cut off freq), the junction and parasitic capacitance will exist.
- At beyond F_H, as freq increases, in HF region, gain decreases due to this parasitic capacitance.

* Low Frequency Analysis of BJT :-

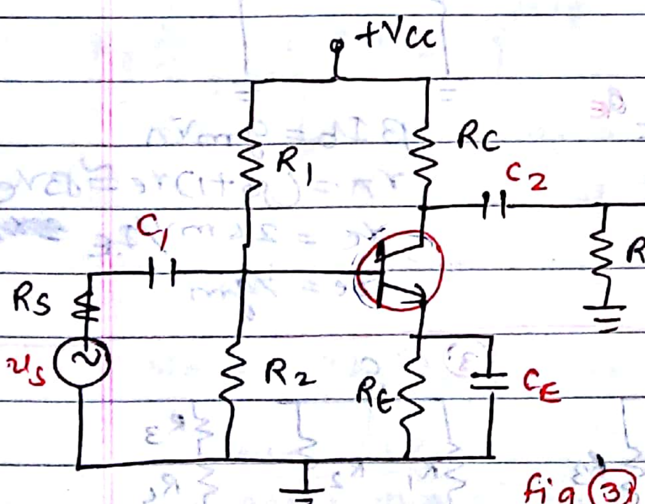


fig ③

step 1 : connect +Vcc to GND ckt.

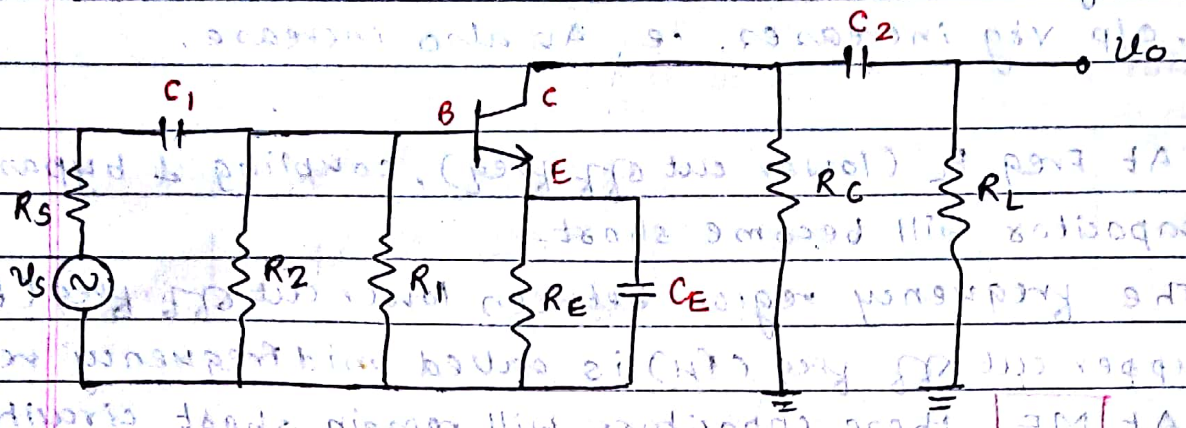
2 : Draw the small signal model for BJT (with coupling & bypass capacitors.)

3 : connect appropriate components to appropriate terminal of BJT.

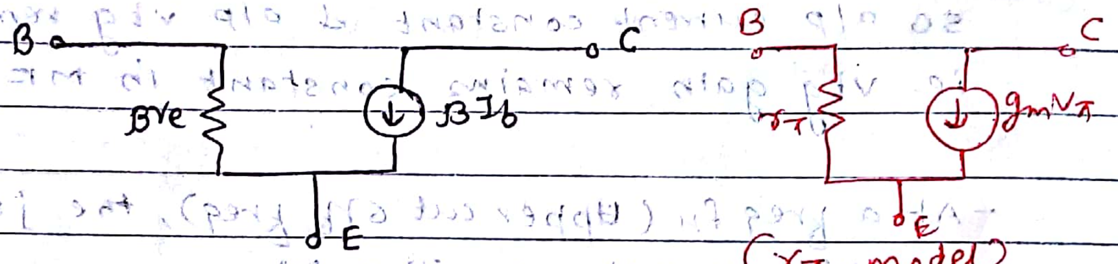
4. To find effect of one capacitor assume all other capacitor as shorted.

* Hybrid-π Model can be quite accurate for L.F ckts & can easily be adopted for higher freq ckt with addition of appropriate inter-electrode capacitance & other parasitic elements.

step ①: diagram ③ / fig ③ as:—

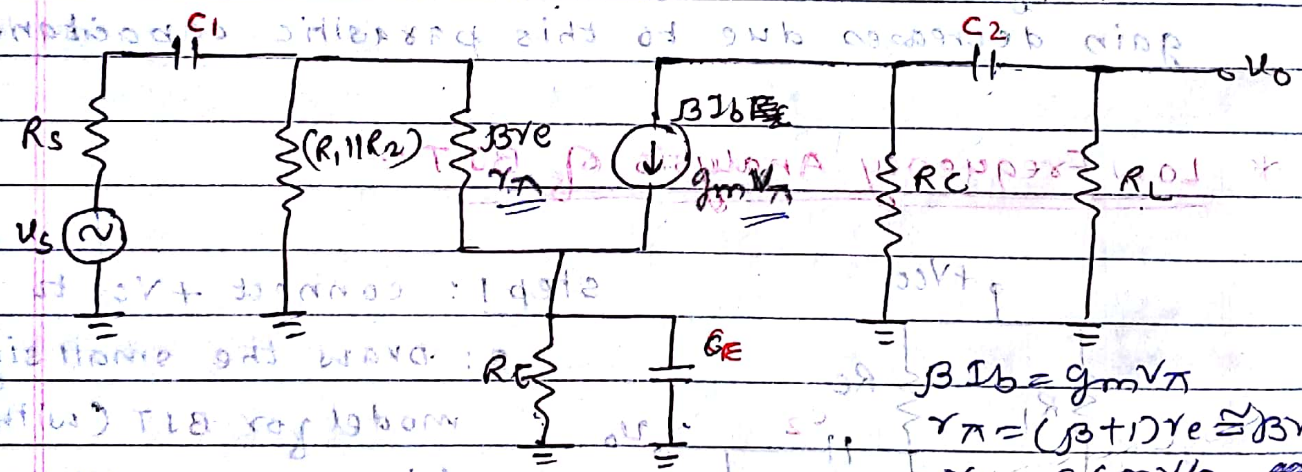


step ②: small signal model (using r_e) for CE ampl^r



(r_π model)

step ③: connect appropriate components to appropriate terminals.



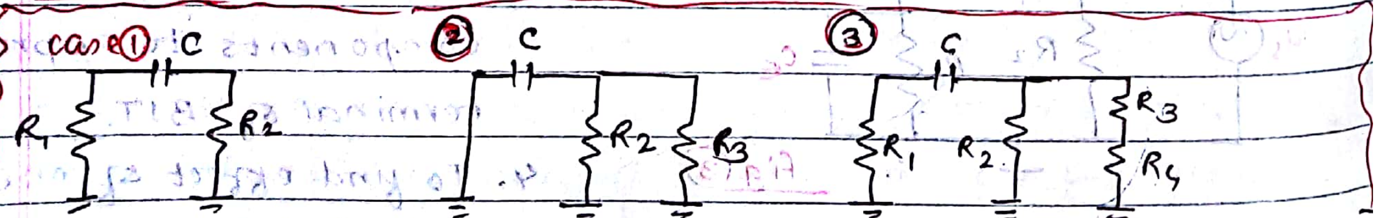
$\beta I_b = g_m V_\pi$

$r_\pi = (\beta+1) r_e \cong \beta r_e$

$r_e = 26\,mV / I_E$

$r_{be} = 1/g_m$

Note:



case ①: $X_C = R_{eq} = R_1 + R_2$

② $X_C = R_{eq} = (R_2 \| R_3)$

③ $X_C = R_{eq} = R_1 + [R_2 \| (R_3 + R_4)]$

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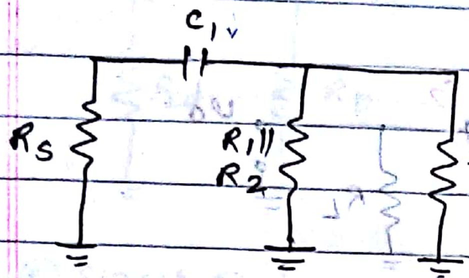
Step ④ : ⓐ Effect of C_1

- short ckt voltage source & open ckt current source.

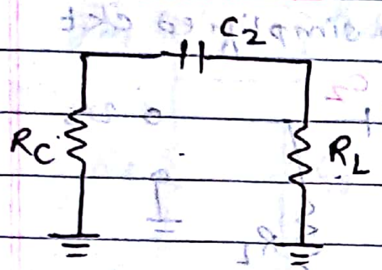


$$\therefore f_{LC_1} = \frac{1}{2\pi R_{eq} C_1}$$

$$X_{C_1} = R_{eq} = R_S + [(R_1 \| R_2 \| \beta r_e)] \qquad = \frac{1}{2\pi [R_S + (R_1 \| R_2 \| \beta r_e)] \cdot C_1}$$

$$R_{eq} = R_S + [R_B \| R_{ib}] \quad \text{where } R_{ib} = r_\pi + (1+\beta) R_E$$

ⓑ Effect of C_2

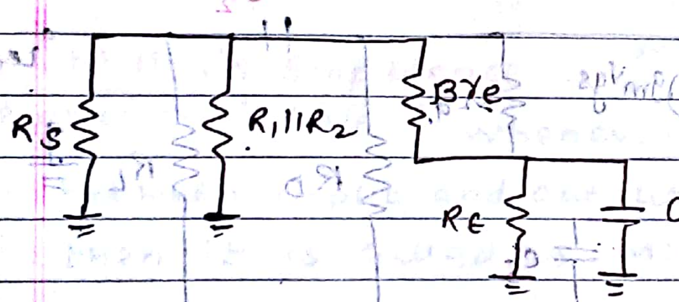


$$X_{C_2} = R_{eq} = R_C + R_L$$

$$f_{LC_2} = \frac{1}{2\pi R_{eq} C_2}$$

$$= \frac{1}{2\pi (R_C + R_L) C_2}$$

ⓒ Effect of C_E



$$X_{C_E} = R_{eq} = R_E \| \left[\frac{\beta r_e + (R_S \| R_1 \| R_2)}{(\beta + 1)} \right]$$

$$f_{LC_E} = \frac{1}{2\pi R_{eq} C_E}$$

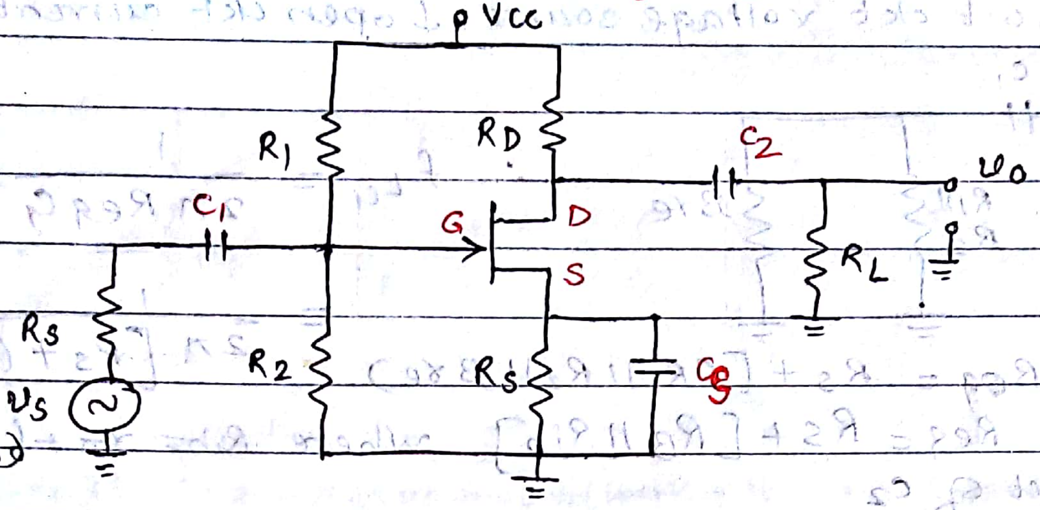
Note : The effective lower cut off frequency among f_{LC_1}, f_{LC_2} & f_{LC_E} is the highest frequency among those three.

$$\boxed{f_L = \text{Highest}(f_{LC_1}, f_{LC_2}, f_{LC_E})}$$

& $A_{V(at\,f = f_L)} = \dfrac{A_{V\,mid}}{\sqrt{2}}$ $\qquad$ $A_{V(at\,any\,LF)} = \dfrac{A_{V\,mid}}{\sqrt{1 + (f_L/f)^2}}$

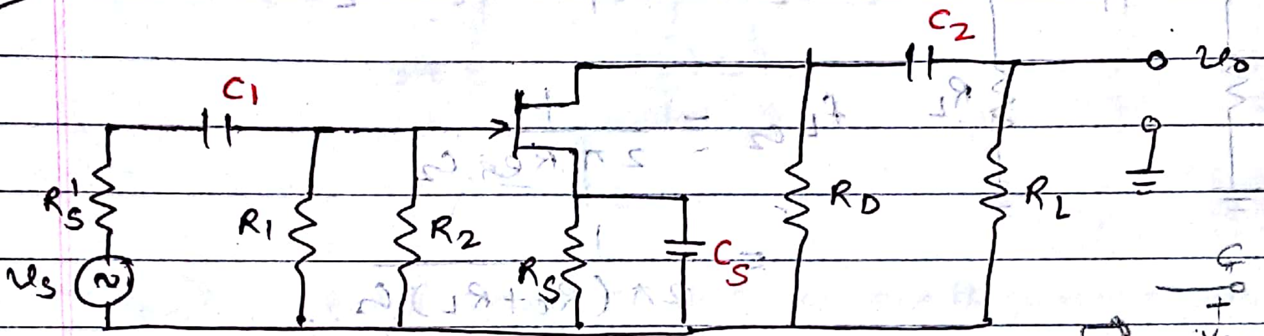
$A_{V\,mid} = \dfrac{-(R_C \| R_L)}{r_e}$ $\qquad$ $A_{V\,mid} = -g_m (R_C \| R_L)$

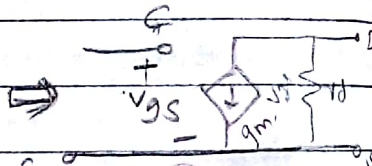
* Low Frequency Analysis of FET :-



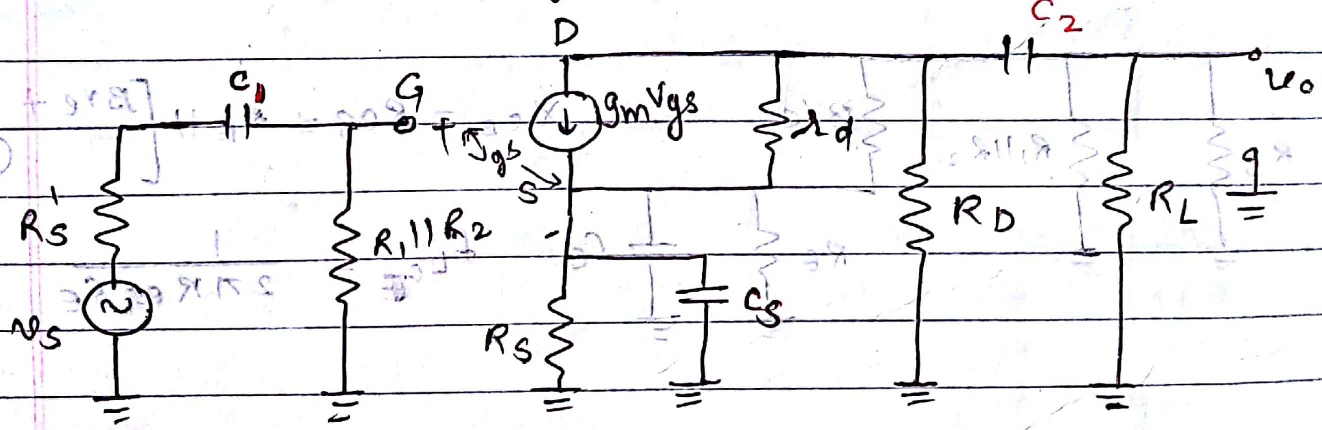
here RG can be used
RG = R1 || R2

step①: connect Vcc to ground & draw simplified ckt

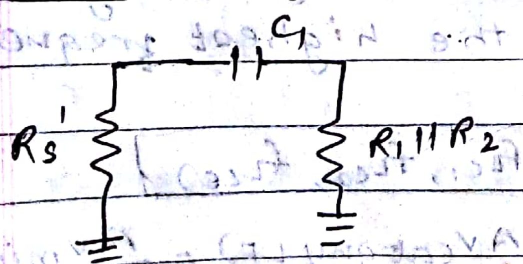




step②: draw small signal model of FET ac equivalent



step③: ⓐ effect of C_1:-

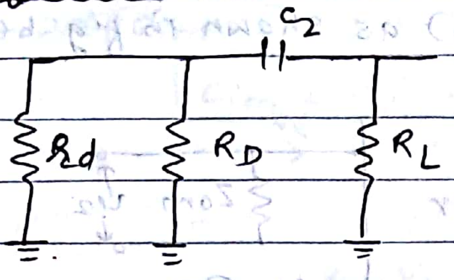


$$X_{C_1} = R_{eq} = R_s' + (R_1 \| R_2)$$

$$f_{L_{C_1}} = \frac{1}{2\pi R_{eq} C_1}$$

* In HF region, capacitance of importance are interelectrode capacitor & wiring capacitor

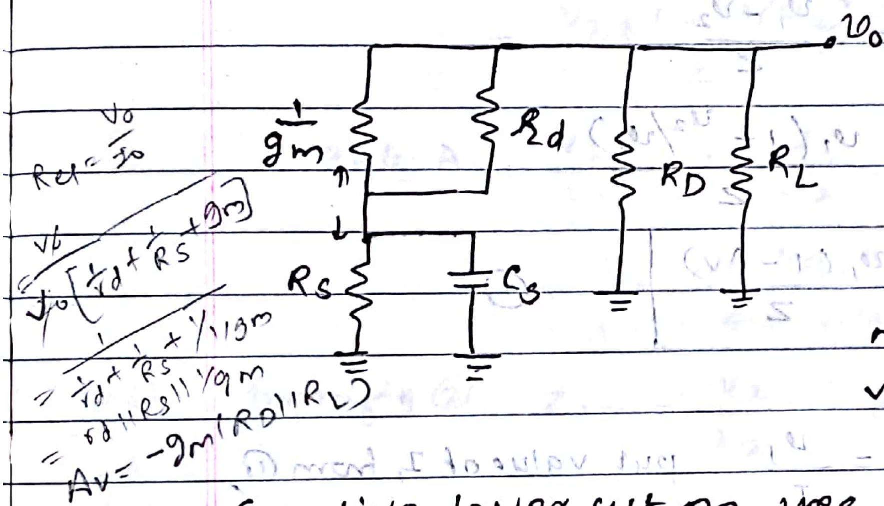
ⓑ Effect of C_2 :-



$$X_{C_2} = R_{eq} = (r_d \parallel R_D) + R_L$$

$$f_{LC_2} = \frac{1}{2\pi R_{eq} C_2}$$

ⓒ Effect of C_S :-



$$X_{C_S} = R_{eq} = R_S \parallel r_d \parallel \frac{1}{g_m}$$

$$f_{LC_S} = \frac{1}{2\pi R_{eq} C_S}$$

neglect R_D & R_L because value of both are negligible.

Effective lower cut off freq is the highest freq component amoung f_{LC_1}, f_{LC_2} and f_{LC_E}.

* **Miller's Impedance** :- Miller effect idea is used to convert an impedance across an amplr into two separate ones as at i/p & o/p. Whenever an impedance is connected between input and output terminals of an amplifier then it is called as Miller's Impedance.

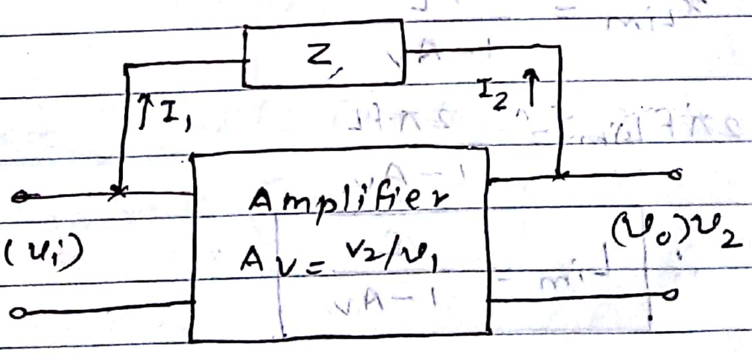


Fig ①

V_1, I_1 is i/p vtg & i/p current resp.
V_2 & I_2 is o/p vtg & o/p current resp.

Zim : Miller i/p impedance
Zom : Miller o/p impedance.

The miller(effect) impedance affect on i/p side (Zim) & o/p side (Zom) as shown in fig below.

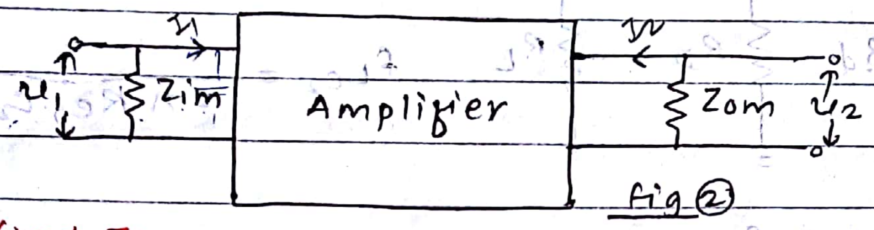


fig ②

To find Zim :-

from fig ①,
$$I_1 = \frac{v_1 - v_2}{Z}$$
$$= \frac{v_1(1 - v_2/v_1)}{Z}$$
$$\therefore \boxed{I_1 = \frac{v_1(1 - A_v)}{Z}} \quad \text{①}$$

from fig ②, $Z_{im} = \frac{v_1}{I_1}$ put value of I_1 from ①,
$$Z_{im} = \frac{v_1 Z}{v_1(1 - A_v)}$$
$$\boxed{Z_{im} = \frac{Z}{1 - A_v}}$$

For Inductive Reactance instead of 'Z' :-
$$x_{Lim} = \frac{X_L}{1 - A_v}$$
$$2\pi f L_{im} = \frac{2\pi f L}{1 - A_v}$$
$$\therefore \boxed{L_{im} = \frac{L}{1 - A_v}}$$

For capacitive Reactance :-
$$X_{cim} = \frac{X_c}{(1 - A_v)}$$

$$\frac{1}{2\pi f C_{im}} = \frac{1}{2\pi f C(1-A_v)}$$

$$\therefore \boxed{C_{im} = C(1-A_v)}$$

To find Z_{om} (from fig ①)

$$I_2 = \frac{V_2 - V_1}{Z}$$

$$= \frac{V_2(1 - v_1/v_2)}{Z}$$

But $A_v = \frac{V_2}{V_1}$ $\therefore \frac{v_1}{v_2} = \frac{1}{A_v}$

$$\therefore I_2 = \frac{V_2(1 - 1/A_v)}{Z} \quad \text{—— ②}$$

from fig ②, $Z_{om} = \frac{v_2}{I_2}$ put value of I_2 from eq^n ②,

$$Z_{om} = \frac{V_2 \cdot Z}{V_2(1 - 1/A_v)}$$

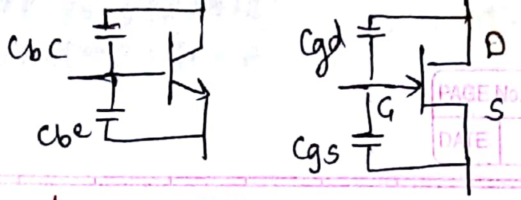
$$\boxed{Z_{om} = \frac{Z}{(1 - 1/A_v)}}$$

For Capacitive Reactance :-

$$X_{com} = \frac{X_c}{(1 - 1/A_v)}$$

$$\frac{1}{2\pi f C_{om}} = \frac{1}{2\pi f C(1 - 1/A_v)}$$

$$\therefore \boxed{C_{om} = C\left(1 - \frac{1}{A_v}\right)}$$



* High Frequency Analysis of BJT :-

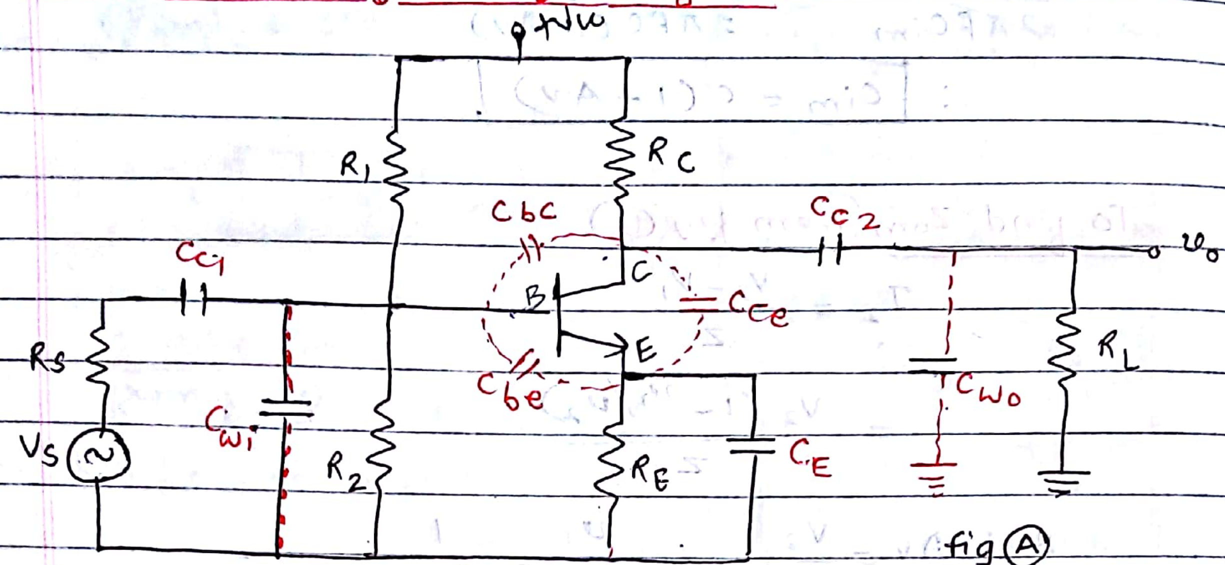


fig (A)

C_{wi} = i/p wiring capacitor

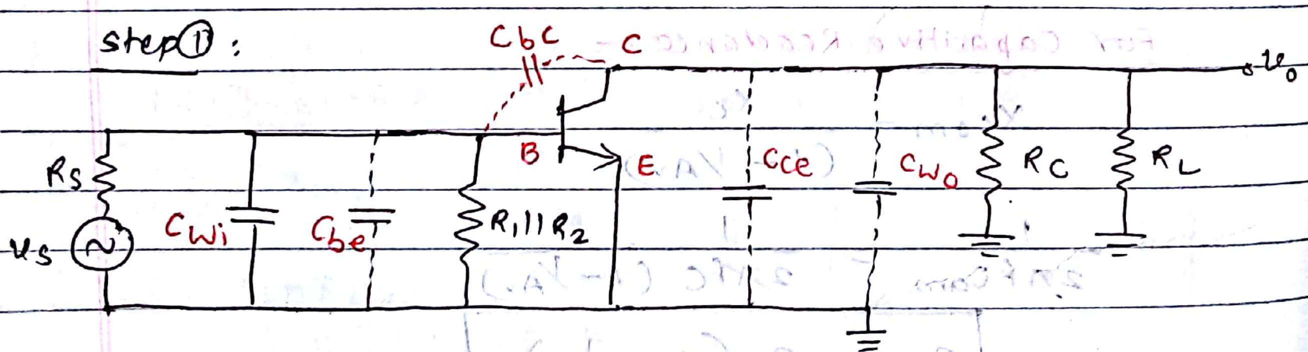
C_{wo} = o/p wiring capacitor.

Step ① : short ckt coupling & bypass capacitors & connect +Vcc to ground. and redraw simplified ckt.

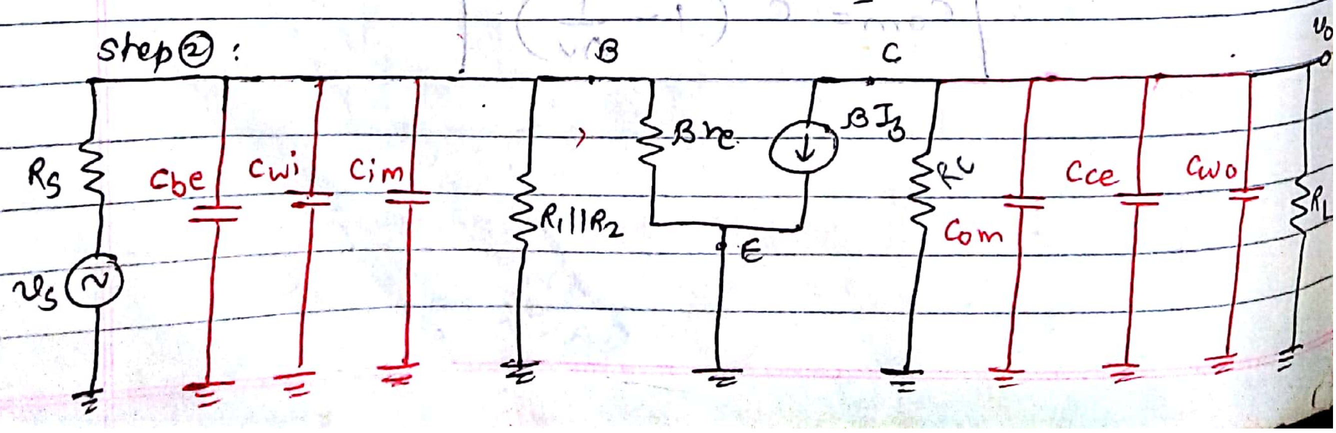
step ② : Draw the small signal model for BJT & connect appropriate components to appropriate terminals.

step ③ : Calculate F_{Hi} & F_{Ho} for i/p & o/p capacitors resp.

step ① :

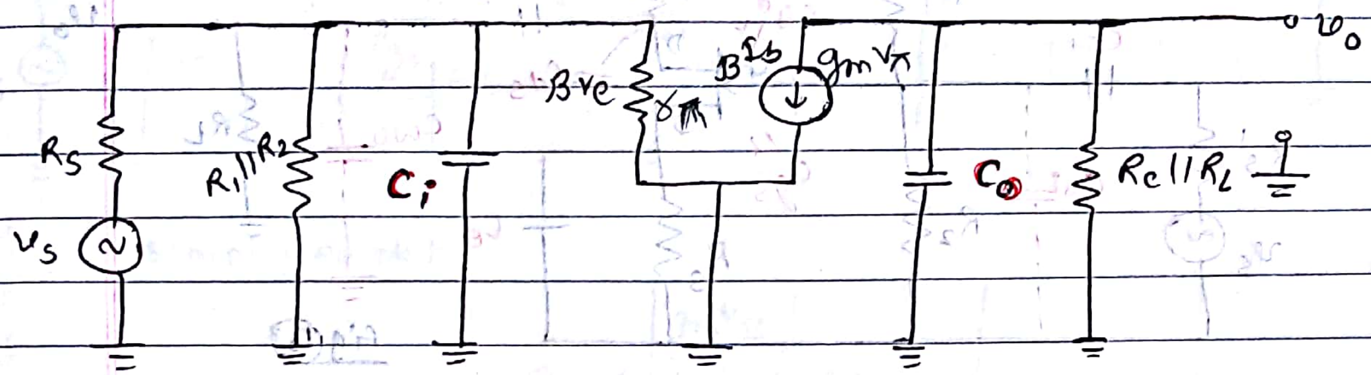


step ② :



where $C_{im} = C_{bc}(1 - A_v)$
$C_{om} = C_{bc}(1 - 1/A_v)$

simplified ckt :-



where $C_i = C_{im} + C_{be} + C_{wi}$ (parallel)
$C_o = C_{ce} + C_{wo} + C_{om}$

step ③:

a) Effect of C_i :- $f_{Hi} = \frac{1}{2\pi \cdot R_{eq} C_i}$

where $C_i = C_{im} + C_{be} + C_{wi}$ & $C_{im} = (1 - A_v) C_{bc}$

To find R_{eq}, short ckt v_s

$\therefore \boxed{R_{eq} = R_S || (R_1 || R_2) || \beta r_e}$

b) Effect of C_o :-

$f_{Ho} = \frac{1}{2\pi R_{eq} C_o}$

where $C_o = C_{ce} + C_{wo} + C_{om}$ & $C_{om} = C_{bc}(1 - 1/A_v)$

To find R_{eq}, open ckt current source.

$\therefore \boxed{R_{eq} = R_c || R_L}$

Note: Effective higher cut off freq betn f_{Hi} & f_{Ho} is the lowest freq component betn f_{Hi} & f_{Ho}.

$A_v (\text{at } f = f_H) = \frac{A_{vmid}}{\sqrt{2}}$ $A_v(\text{at any high freq}) = \frac{A_{vmid}}{\sqrt{1 + (f/f_H)^2}}$

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* High frequency Analysis of FET :-

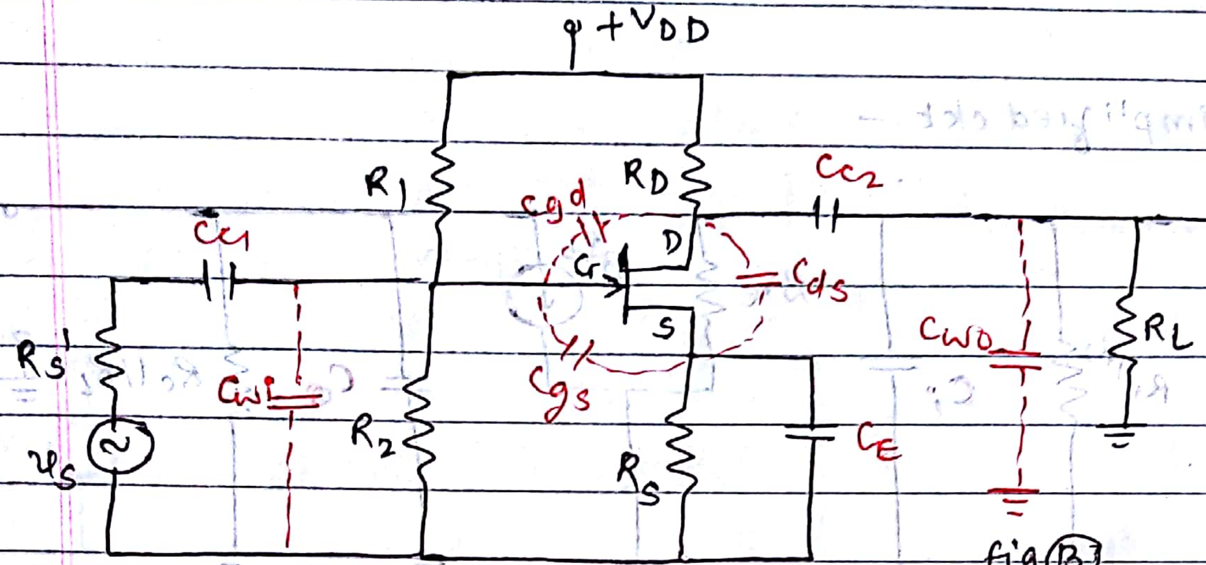


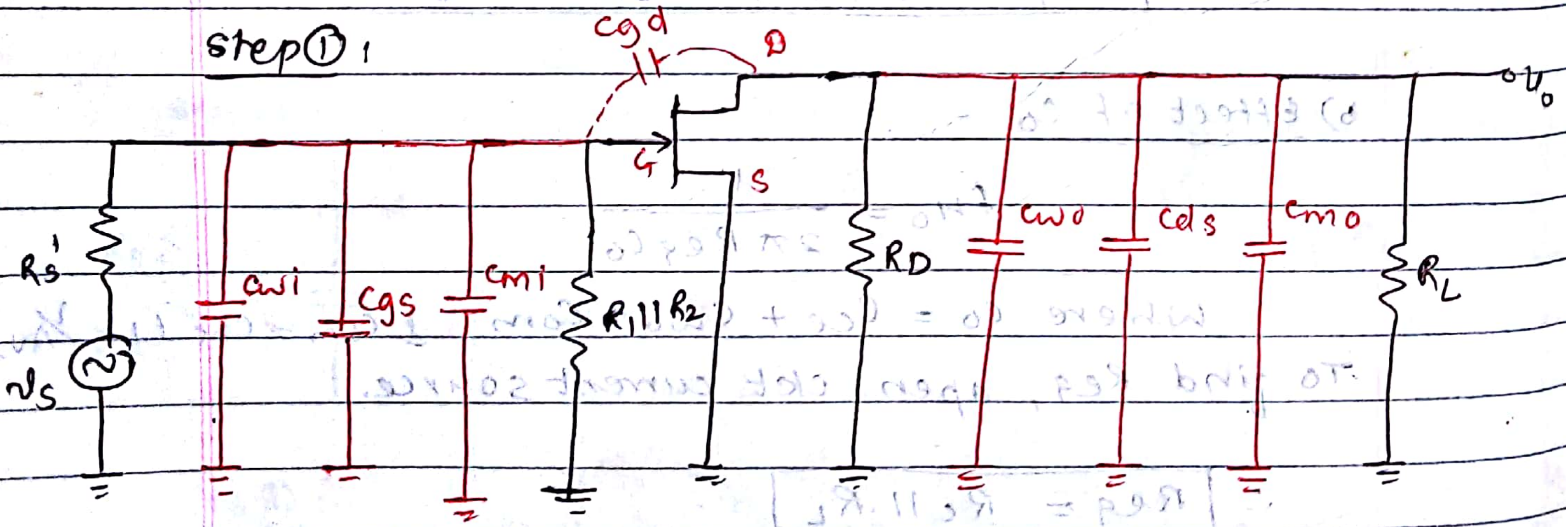
fig (B)

step ① : short ckt coupling & bypass capacitors. connect $+V_{DD}$ to ground & draw simplified ckt.

step ② : Draw small signal model for FET & connect appropriate components to appropriate terminals.

step ③ : find f_{Hi} & f_{Ho} corresponding to C_i & C_o resp.

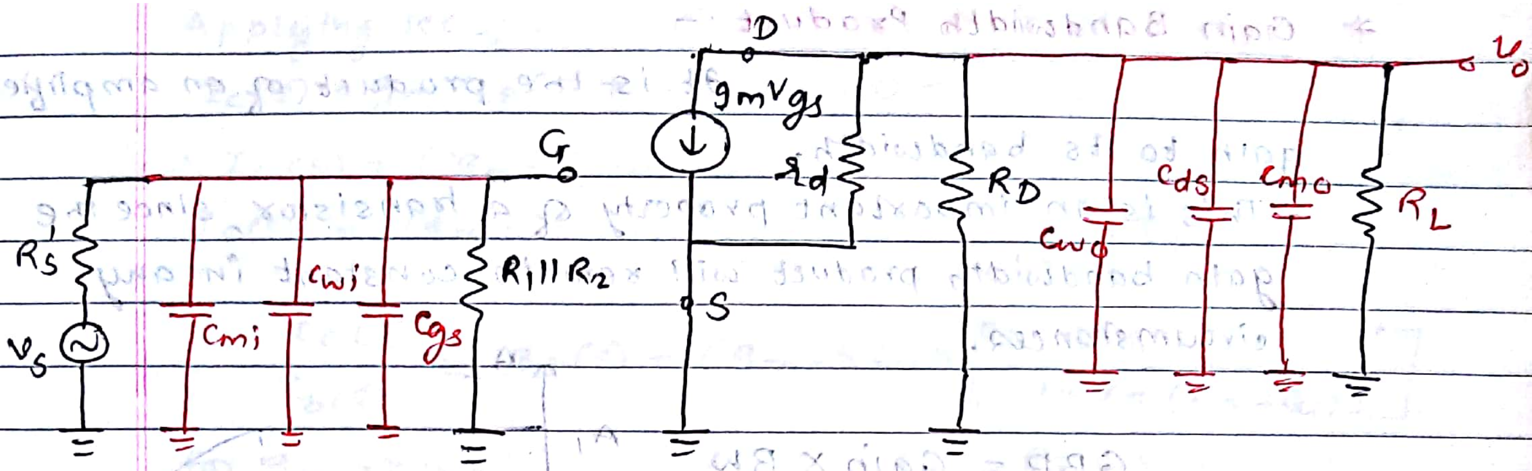
step ① :



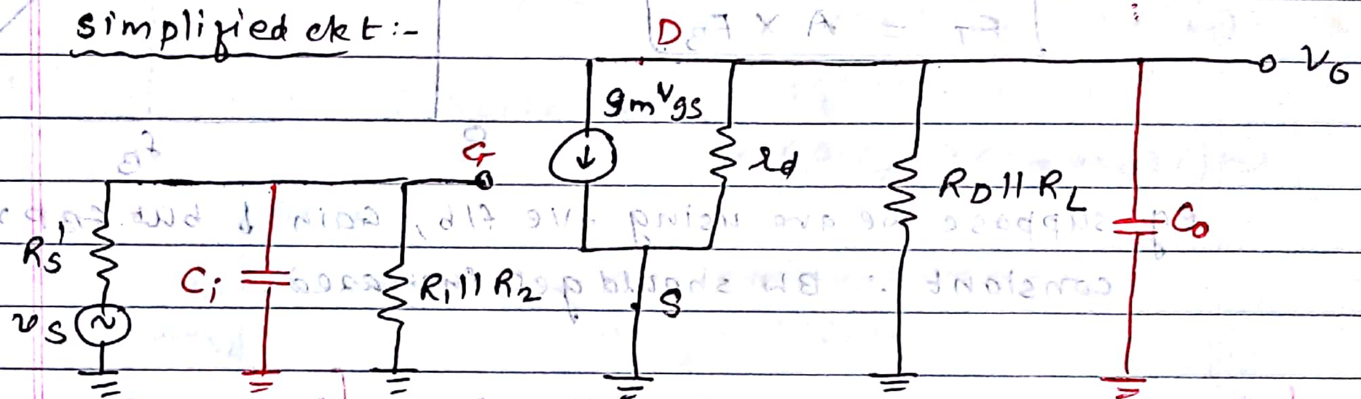
step ② : small signal model :-

$$C_{mi} = C_{gd}(1 - A_v)$$

$$C_{mo} = C_{gd}(1 - 1/A_v)$$



simplified ckt :-



where $C_i = C_{mi} + C_{wi} + C_{gs}$

$C_o = C_{mo} + C_{wo} + C_{ds}$

step (III) 1 (a) Effect of C_i :-

$$f_{Hi} = \frac{1}{2\pi R_{eq} C_i}$$ where $C_i = C_{mi} + C_{wi} + C_{gs}$

$C_{mi} = (1 - A_v) C_{gd}$

To find Req, short ckt ~~current~~ V_{TG} source (v_s).

$\therefore$ $\boxed{R_{eq} = R_s' \| R_1 \| R_2}$

(b) Effect of C_o :-

$$f_{Ho} = \frac{1}{2\pi R_{eq} C_o}$$ where $C_o = C_{mo} + C_{ds} + C_{wo}$

$C_{mo} = \left(1 - \frac{1}{A_v}\right) C_{gd}$

To find Req, open ckt current source.

$\therefore$ $\boxed{R_{eq} = r_d \| R_D \| R_L}$

$\boxed{f_H = \text{Lowest}(f_{Hi} \,\&\, f_{Ho})}$

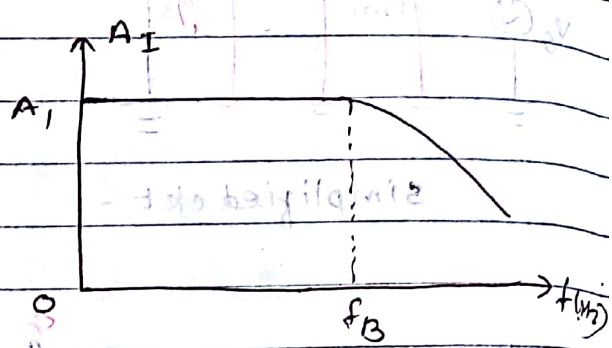
* Gain Bandwidth Product :-

It is the product of an amplifier gain to its bandwidth.

- This is an important property of a transistor, since the gain bandwidth product will remain constant in any circumstances.

$$GBP = Gain \times BW$$

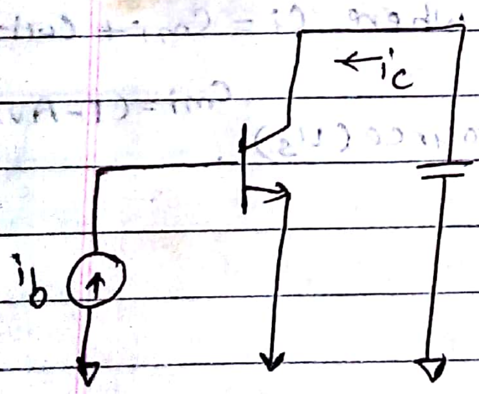
$$\boxed{F_T = A \times F_B}$$



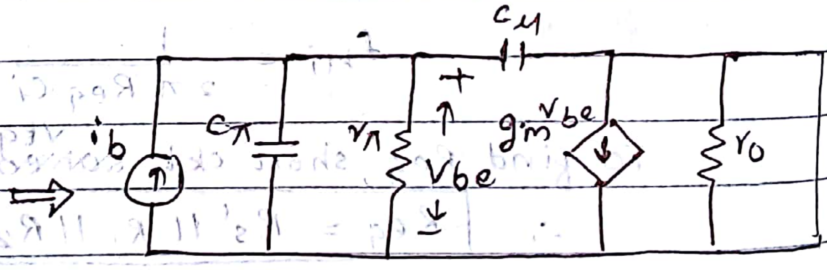
eg. suppose we are using -ve f/b, Gain ↓ but GBP remains constant ∴ BW should get increased.

Proof * GBP or UGB or Transition freq = $\boxed{f_T = \dfrac{g_m}{2\pi(C_\pi + C_\mu)}}$

Proof: applying a test current i_b to the base and short circuiting the collector for ac, for finding short ckt current gain β_f



⇒



$$V_{be}(s) = \left[r_\pi \parallel \frac{1}{(C_\pi + C_\mu)s} \right] I_b(s)$$

$$= \left[\frac{r_\pi \times \left(\frac{1}{C_\pi + C_\mu}\right)s}{r_\pi + \left(\frac{1}{C_\pi + C_\mu}\right)s} \right] I_b(s)$$

$$= \left[\frac{r_\pi}{1 + r_\pi (C_\pi + C_\mu)s} \right] I_b(s)$$

Applying KCL,

$$I_c(s) - g_m V_{be}(s) + s C_\mu V_{be}(s) = 0$$

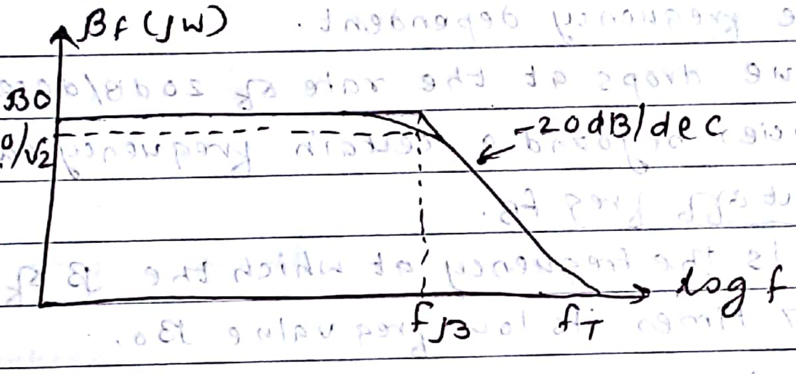
$$\therefore I_c(s) = (g_m - s C_\mu) V_{be}(s)$$

$$I_c(s) = (g_m - s C_\mu)\left[\frac{r_\pi}{1 + r_\pi (C_\pi + C_\mu) s}\right] I_b(s)$$

$$\therefore \frac{I_c(s)}{I_b(s)} = \beta_f(s) = (g_m - s C_\mu)\left[\frac{r_\pi}{1 + r_\pi (C_\pi + C_\mu) s}\right]$$

for $g_m >> \omega . C_\mu$, $\beta_f(j\omega) = \dfrac{g_m r_\pi}{1 + r_\pi (C_\pi + C_\mu)(j\omega)}$

$$\beta_f(j\omega) = \frac{\beta_0}{1 + (\beta_0 / g_m)(C_\pi + C_\mu)(j\omega)}$$



current gain β_f is a function of frequency & as freq increases, current gain decreases at a rate of -20dB per decade. (low pass characteristics)

At 3dB freq $\beta_f(j\omega_\beta) = \dfrac{\beta_0}{1+j1}$ where $\beta_0 \approx \beta_f$, low freq current gain.

$$\therefore \frac{\beta_f (C_\pi + C_\mu)\omega_\beta}{g_m} = 1$$

$$\therefore \omega_\beta = 2\pi f_\beta = \frac{g_m}{\beta_0 (C_\pi + C_\mu)}$$ ie $\boxed{f_\beta = \dfrac{g_m}{2\pi \beta_0 [C_\pi + C_\mu]}}$

L Bandwidth of BJT.

current gain will be infinity when $|\beta_f(j\omega)| = 1$

$\therefore$ UGB $= \omega_T = \dfrac{g_m}{(C_\pi + C_\mu)}$ rad/sec or $\boxed{f_T = \dfrac{g_m}{2\pi (C_\pi + C_\mu)}}$ Hz

$$f_T = \frac{1}{2\pi r_e (C_\pi + C_\mu)} \quad \because \frac{1}{r_e} = g_m$$

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for a MOSFET, $f_T = \dfrac{g_m}{2\pi[C_{gs}+C_{gd}+C_{gb})}$

f_β = Beta cutoff freq

f_T ≈ is the figure of merit of a transistor for linear application.

≈ 100 MHz to few GHz

≈ 300 MHz for BC147A, BC147B, BFW11 etc.

* High frequency performance (response) of a transistor is affected by the interelectrode capacitances that causes β to be frequency dependent.
Its value drops at the rate of 20dB/decade at frequencies beyond a certain frequency called the beta cut off freq f_β.
f_β is the frequency at which the β of a transistor is 0.707 times its low freq value β_0.
If frequency is increased beyond f_β, β continues to decrease until it eventually reaches a value of 1.

* The frequency at which β = 1, is called transition freq (f_T) or unity gain bandwidth.

For a BJT

The Beta cut off freq $f_\beta = \dfrac{g_m}{2\pi\beta_0(C_\pi+C_\mu)}$

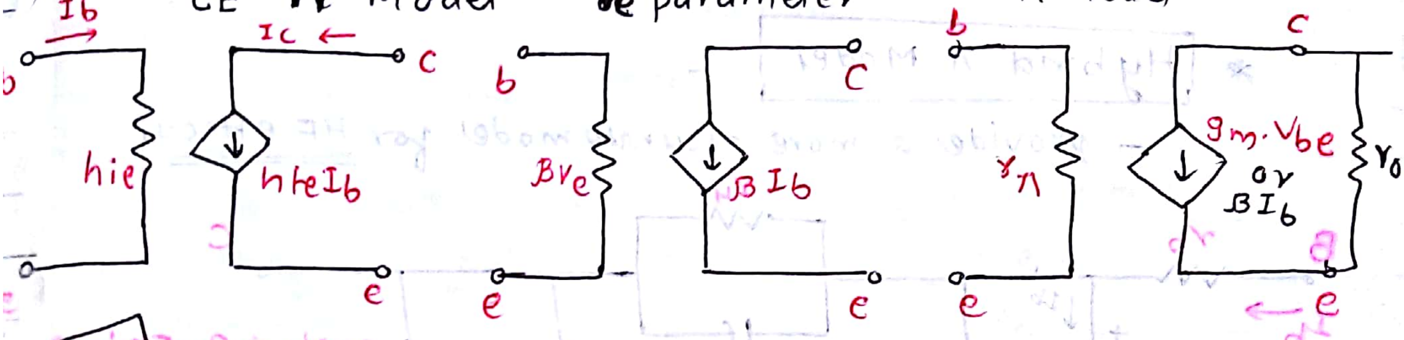
The transition frequency (GBP), $f_T = \dfrac{g_m}{2\pi(C_\pi+C_\mu)}$

For a FET

The gain bandwidth product (GBP), $f_T = \dfrac{g_m}{2\pi(C_{gs}+C_{gd})}$

Models of BJT

CE 'h' Model   're' parameter   'rπ' Model

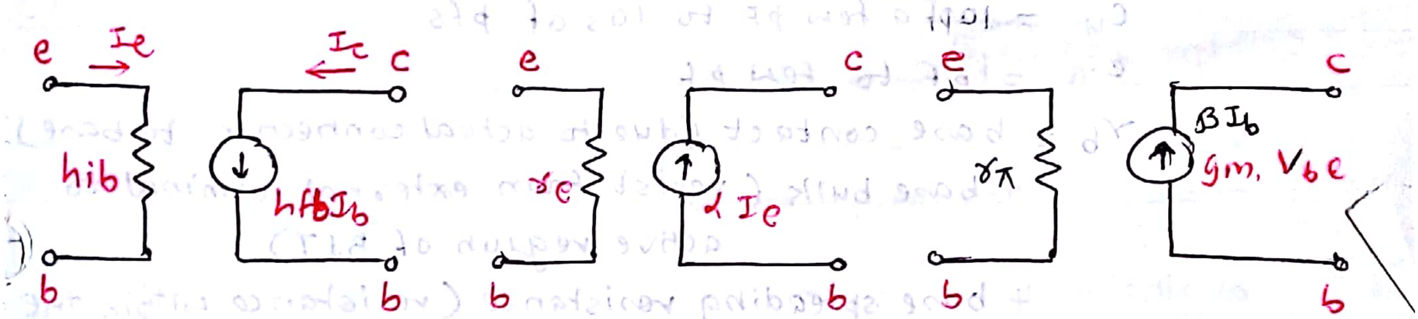


CE

$h_{ie} = \beta r_e = r_\pi$ $[\because r_\pi = \beta/g_m \; \& \; r_e = 1/g_m]$

$h_{fe} = \beta_{ac}$

$h_{oe} = 1/r_o$



CB

h parameter   re model   π Model

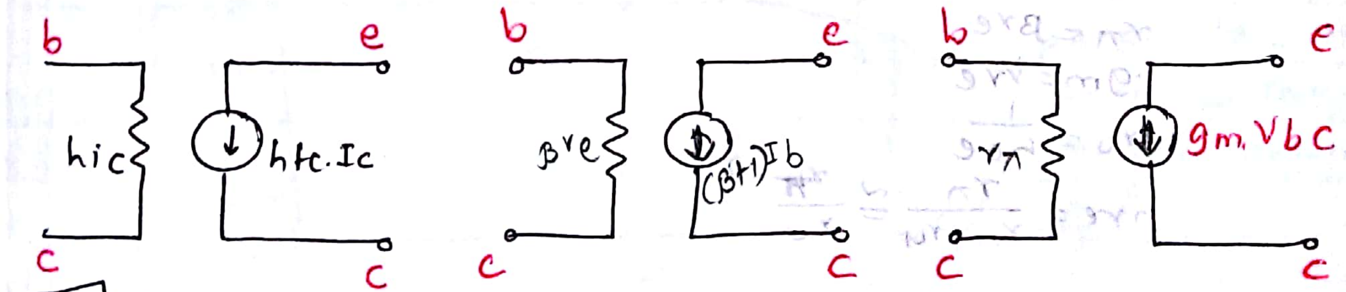
$h_{ib} = r_e \left(= \frac{h_{ie}}{1+h_{fe}}\right)$

$h_{fb} = -\alpha = -1 = \left(\frac{-h_{fe}}{1+h_{fe}}\right)$

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

$r_e = \frac{V_T}{I_E} \simeq \frac{1}{g_m}$

$I_C = \beta I_B$



CC

h parameter   re model   π Model

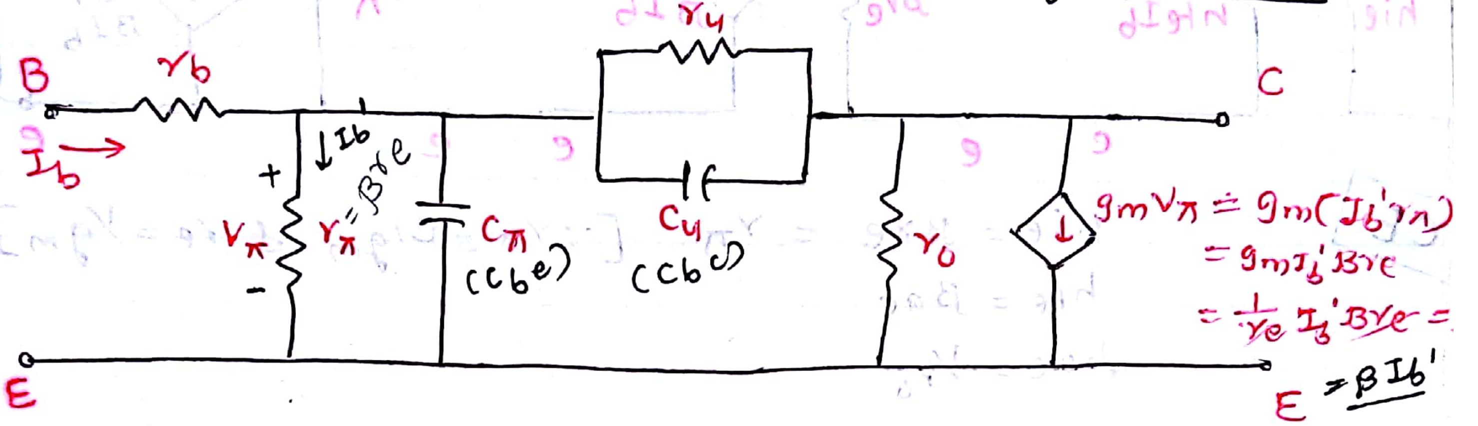
$h_{ic} = h_{ie} = \beta r_e$

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

$\left(\frac{1}{h_{oc}}\right) = \left(\frac{1}{h_{oe}}\right)$

* Hybrid π Model :-

- provide a more accurate model for HF effects.



C_u = 10pF a few pF to 10s of pFs

C_π = 1pF to few pF

r_b = base contact (due to actual connection to base
+ base bulk (resist. from external terminal to active region of BJT)
+ base spreading resistance (resistance within the active base region)

r_u = 'u' union it provides betn C & B terminal
= very large R provides f/b path from o/p to i/p in the equivalent model.

$r_\pi = \beta r_e$ = symbol π for hybrid π terminology.

r_o = o/p resistance across an applied load. = $1/h_{oe}$

$r_\pi = \beta r_e$

$g_m = 1/r_e$

$r_o = \frac{1}{h_{oe}}$

$h_{re} = \frac{r_\pi}{r_\pi + r_u} \simeq \frac{\cancel{r_\pi}}{r_u}$

For Low to MF r_b << small so S.C.
r_u >> large >> βr_o $\Rightarrow$ O.C.

so Equivalent ckt in LF & MF $\Rightarrow$ same as 'r_e' Model

* At HF, there are 2 factors that define → -3dB cutoff pt.
i) N/W capacitance (parasitic 'C')
ii) freq. depedance of h_{fe} (β)

* h_{fe} or β variation :-

we know that $A_v = \frac{1}{1+j(f/f_2)}$

similarly, $h_{fe} = \frac{h_{fe(mid)}}{1+j(f/f_\beta)}$

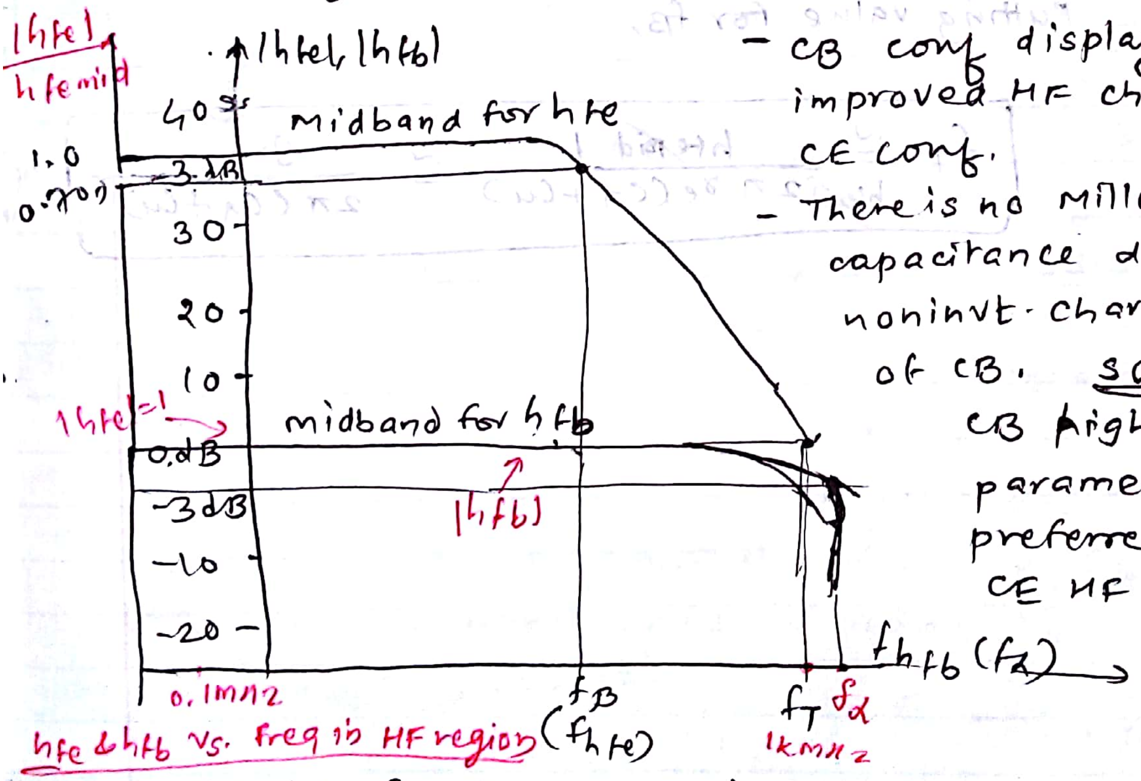
f_β is determined by a set of parameters employed in hybrid π or Giacoletto Model

$$f_\beta = f_{h_{fe}} = \frac{1}{2\pi r_\pi (C_\pi + C_\mu)}$$

But $r_\pi = \beta r_e = h_{femid} r_e$

$$\therefore f_\beta = \frac{1}{h_{femid} \cdot r_e} \frac{1}{2\pi (C_\pi + C_\mu)} = \frac{1}{2\pi \beta_{mid} r_e (C_\pi + C_\mu)}$$

r_e = is a function of N/W design
f_β = function of bias configuration.



h_{fe} & h_{fb} vs. freq in HF region

- CB conf displays improved HF char. over CE conf.
- There is no Miller effect capacitance due to noninvt. characteristics of CB. So CB high freq parameters are preferred over CE HF parameters.

Relation for direct conversion

$$\boxed{f_\beta = f_\alpha (1-\alpha)}$$

A quantity GBP is defined for transistor by condn, $\beta = 1$

$$\left| \frac{h_{fe(mid)}}{1 + j(f/f_\beta)} \right| = 1$$

so that $|h_{fe}|_{dB} = 20\log_{10} \left| \frac{h_{fe(mid)}}{1+j(f/f_\beta)} \right| = 20\log_{10} 1 = 0$

∴ The frequency at which $h_{fe}(dB) = 0\,dB$ is f_T

h_{fe} at defined condn pt $(f_T >> f_\beta)$ is given by

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

∴ $\boxed{f_T \cong h_{fe\,mid} \cdot f_\beta}$ GBP $(\because f_\beta \cong BW)$

$f_T \cong \beta_{mid} f_\beta$ & $f_\beta = f_T/\beta_{mid}$

putting value for f_β,

$$\boxed{f_T \cong \frac{\cancel{h_{fe\,mid}}\, 1}{\cancel{h_{fe\,mid}}\, 2\pi r_e (C_\pi + C_\mu)} \cong \frac{g_m}{2\pi (C_\pi + C_\mu)}}$$

Ex 1) Determine the 3dB freq & short ckt current gain of a bipolar transistor with the following parameters.

$r_\pi = 2.6k$ $C_\pi = 2pf$ $C_\mu = 0.1pf$

$A_I = \frac{\beta_0}{\sqrt{1+(f/f_\beta)^2}}$

Sol^n.- The 3dB freq $f_\beta = \frac{g_m}{2\pi \beta_f (C_\pi + C_\mu)}$

$= \frac{1}{2\pi(\beta_f/g_m)(C_\pi + C_\mu)}$

$r_\pi = \beta r_e$ $(\because r_e = \frac{1}{g_m})$

$r_\pi = \beta/g_m$

$= \frac{1}{2\pi r_\pi (C_\pi + C_\mu)}$

$= \frac{1}{2\pi \times 2.6k(2\times10^{-12} + 0.1\times10^{-12})}$

$\boxed{f_\beta = 29.16 MHz}$

2.) Calculate the bandwidth & capacitance C_π of a BJT if the unity GBW is 500MHz at $I_C = 1mA$, $\beta_f = 100$ & $C_\mu = 0.3pf$

Sol^n.- $UGB = f_T = \underbrace{f_\beta}_{BW} \times \underbrace{\beta_f}_{h_{fe}(mid)}$ $\therefore$ Bandwidth $f_\beta = \frac{500\times10^6}{100}$ (f_T) $= 5MHz$ (BW)

$g_m = \frac{I_C}{V_T} = \frac{1mA}{26mV} = 38.5mS$

$f_T = \frac{g_m}{2\pi(C_\pi + C_\mu)} \Rightarrow \therefore C_\pi + C_\mu = \frac{38.5\times10^{-3}}{2\pi\times500\times10^6}$

$C_\pi + C_\mu = 12.26pf$

But $C_\mu = 0.3pf$

$\therefore \boxed{C_\pi = 11.96pf}$

3.) Calculate the UGB of a n-channel MOSFET with the parameters $K_n = 0.25mA/V^2$, $V_{TN} = 1V$, $C_{gd} = 0.04pf$, $C_{gs} = 0.2pf$ & transistor is biased at $V_{GS} = 3V$.

Sol^n:- $I_D = K_n(V_{GS} - V_{TN})^2$

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

$= 2(0.25)(3-1)$

$= 1mS$

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

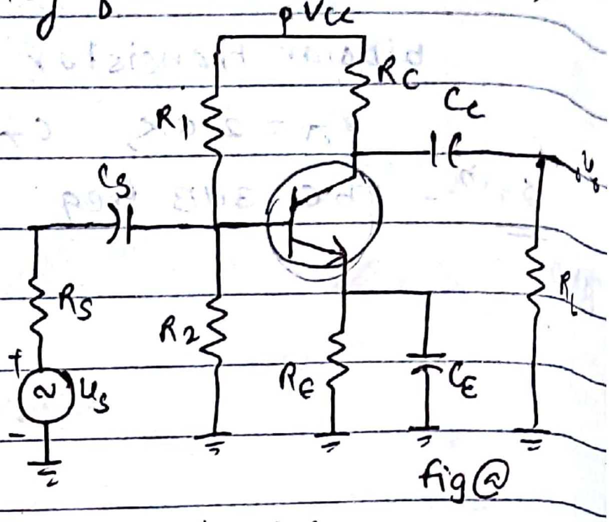
$= \frac{1\times10^{-3}}{2\pi(0.04\times10^{-12} + 0.2\times10^{-12})}$

$\boxed{f_T = 663.5MHz}$

(Ex4) a) Determine the lower cut-off frequency for the network shown using following parameters:-

LF

$C_s = 10\mu F$	$R_s = 1k\Omega$
$C_E = 20\mu F$	$R_1 = 40k\Omega$
$C_c = 1\mu F$	$R_2 = 10k\Omega$
$\beta = 100$	$R_E = 2k\Omega$
$r_o = \infty\Omega$	$R_c = 4k\Omega$
$V_{cc} = 20V$	$R_L = 2.2k\Omega$



fig(a)

Find AV mid
AVs

b) sketch the frequency response using a Bode plot.

Soln:- a) To determine re for dc conditions,

$\beta R_E = (100)(2k\Omega) = 200k\Omega \geq 10R_2 = 100k\Omega$

$\beta R_E \geq 10R_2$ cond for approximate analysis

$V_B = \frac{R_2 \cdot V_{cc}}{R_2 + R_1} = \frac{10k\Omega(20V)}{10k\Omega + 40k\Omega} = \frac{200V}{50} = 4V$

$V_E = V_B - V_{BE}$

$I_E = \frac{V_E}{R_E} = \frac{4V - 0.7}{2k\Omega} = 1.65mA$

I_{CQ}

$\rightarrow r_e = \frac{26mV}{I_E} = \frac{26mV}{1.65mA} = 15.76\Omega$

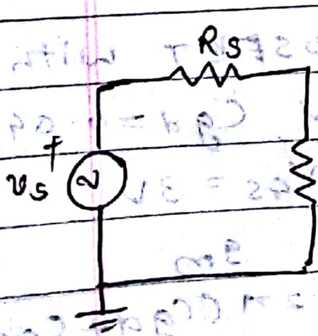
$g_m = \frac{I_E}{26mV}$ & $\beta r_e = 100(15.76) = 1576\Omega = 1.576k\Omega$

$g_m = 1/r_e$

Midbandgain :- $A_v = \frac{v_o}{v_i} = \frac{-R_c \| R_L}{r_e} = \frac{-(4k\Omega)\|(2.2k\Omega)}{15.76} \cong -90$

r_π

β / g_m



$R_{eq} = R_1 \| R_2 \| \beta r_e$
$= 40k\Omega \| 10k\Omega \| 1.576k\Omega$

$R_{eq} \cong 1.32k\Omega$

$v_i = \frac{R_{eq} \cdot v_s}{R_{eq} + R_s} = \frac{R_i \cdot v_s}{R_i + R_s}$

$\therefore \frac{v_i}{v_s} = \frac{R_{eq}}{R_{eq} + R_s} = \frac{1.32k\Omega}{1.32k\Omega + 1k\Omega} = 0.569$

effect of C_C :- $f_{Lc} = \dfrac{1}{2\pi(R_c+R_L)C_c}$

$= \dfrac{1}{6.28(4k\Omega+2.2k\Omega)1\mu F}$

$f_{Lc} \simeq 25.68 Hz$

effect of C_E :- $f_{LE} = \dfrac{1}{2\pi R_{eq} C_E}$ 27

$R_{eq} = R_E \| \left[\dfrac{\beta r_e + (R_s \| R_1 \| R_2)}{\beta}\right]$

$R_{eq} = 24.35$ $\therefore f_{LE} = 327 Hz$

$A_{v_s} = \dfrac{v_o}{v_s} = \dfrac{v_o}{v_i} \times \dfrac{v_i}{v_s} = (-90)(0.5869)$

$= -51.21$

$\therefore \boxed{A_{v_s} = -51.21}$

$f_{LS} = 1.76 Hz$
$f_{LC} = 25.6 Hz$
$f_{LE} = 327 Hz$

effect of C_s :- $f_{Ls} = \dfrac{1}{2\pi(R_{eq}+R_s)C_s} = \dfrac{1}{(6.28)(1k\Omega+1.32k\Omega)(10\mu F)}$

$= \dfrac{1}{2\pi R_{eq} C_s}$ where $R_{eq} = R_i + R_s = (R_1 \| R_2 \| \beta r_e) + R_s$

$\boxed{f_{Ls} \simeq 6.86 Hz}$

$\therefore$ $\boxed{\text{Lower cutoff freq} = 327 Hz}$ (Highest among f_{Ls}, f_{LC} & f_{LE})

Ex 5) a) Determine the lower cutoff frequency for the network shown fig@ using the following parameters.

LF

Ex 2

H.W

$C_s = 10\mu F$	$R_s = 1k\Omega$
$C_E = 20\mu F$	$R_1 = 40k\Omega$
$C_c = 1\mu F$	$R_2 = 10k\Omega$
$\beta = 100$	$R_E = 1.2k\Omega$
$r_o = \infty\Omega$	$R_c = 4k\Omega$
$V_{cc} = 10V$	$R_L = 10k\Omega$

b) sketch the frequency response using a Bode plot.

Sol? - a) To determine r_e for dc conditions,

$\beta R_E = 100(1.2k\Omega) = 120k\Omega > 10R_2 = 100k\Omega$

$V_B = \dfrac{R_2 \cdot V_{cc}}{R_1+R_2} = \dfrac{10k\Omega(10V)}{10k\Omega+40k\Omega} = \dfrac{100V}{50} = 2V$

$I_E = \dfrac{V_E}{R_E} = \dfrac{2V-0.7}{1.2k\Omega} = 1.083mA$

$r_\pi = \dfrac{\beta V_T}{I_{cq}}$ $\quad r_e = \dfrac{26mV}{I_E} = \dfrac{26mV}{1.083mA} \simeq 24.01\Omega$

$g_m = \dfrac{I_{cq}}{V_T}$ $\quad \beta r_e = 100(24.01\Omega) = 2400\Omega = 2.4k\Omega$

Midband gain :- $A_v = \dfrac{v_o}{v_i} = \dfrac{-R_c \| R_L}{r_e} = -\dfrac{(4k\Omega)\|(10k\Omega)}{24.01} = -119$

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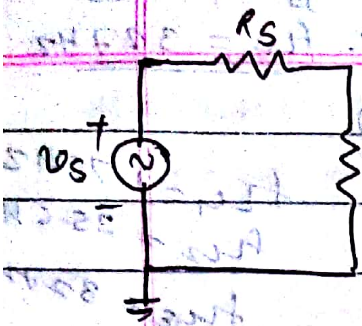


$R_i = R_1 \| R_2 \| \beta r_e$

$= 40k\Omega \| 10k\Omega \| 2.4k\Omega)$

$\cong 1.85k\Omega$

$$v_i = \frac{R_i \cdot v_s}{R_i + R_s}$$

$$\frac{v_i}{v_s} = \frac{R_i}{R_i+R_s} = \frac{1.85k\Omega}{(1.85+1)k\Omega} = 0.65$$

$$A_{vs} = \frac{v_o}{v_s} = \frac{v_o}{v_i} \cdot \frac{v_i}{v_s} = -119 \times 0.65 = -77.35$$

Effect of C_s :-

$$f_{Ls} = \frac{1}{2\pi R_{eq} C_s} = \frac{1}{2\pi(R_i+R_s)C_s} = \frac{1}{(6.28)(1k\Omega+1.85k\Omega) \times 10\mu F}$$

$f_{Ls} \cong 5.59Hz$

effect of C_c :-

$$f_{Lc} = \frac{1}{2\pi(R_c+R_L)C_c} = \frac{1}{(6.28)(4k\Omega+10k\Omega)(1\mu F)} =$$

$f_{Lc} = 11.68Hz$

Effect of C_E :-

$$f_{LCE} = \frac{1}{2\pi R_{eq} C_E} = \frac{1}{2\pi\left[R_E \| \left(\frac{\beta r_e + (R_s \| R_1 \| R_2)}{\beta}\right)\right] C_E}$$

$f_{LE} = 248.6Hz$

$\therefore$ Lower cutoff freq = Highest among f_{Ls}, f_{Lc} & $f_{LE} \Rightarrow 248.6Hz$

Next Example :- • $R_L = 2.2k\Omega$ • $V_{cc} = 20V$ } $f_{Hi} = 738.24kHz \rightarrow R_{eq} = 0.531k\Omega, C_i = 406$; $f_{Ho} = 8.6MHz \rightarrow R_{eq} = 1.418k\Omega, C_o = 13.04pF$

31

$C_{im} = [1-(-90)]4$
$C_{om} = (1 - 1/-90)4$

Ex 7.) Use the network of fig (A), with the parameters;

HF BJT (Ex3)

$R_S = 1k\Omega$	$C_S = 10\mu F$	& $C_\pi (C_{be}) = 36pF$
$R_1 = 40k\Omega$	$C_C = 1\mu F$	$C_\mu (C_{bc}) = 4pF$
$R_2 = 10k\Omega$	$C_E = 20\mu F$	$C_{ce} = 1pF$
$R_E = 1.2k\Omega$	$\beta = 100$	$C_{wi} = 6pF$
$R_C = 4k\Omega$	$r_o = \infty\Omega$	$C_{wo} = 8pF$
• $R_L = 10k\Omega$	• $V_{cc} = 10V$	

a) Determine f_{Hi} & f_{Ho}

$r_e = \frac{26mV}{I_E}$ $\quad I_E = \frac{V_E}{R_E} = \frac{V_B - V_{BE}}{R_E}$

$V_B = R_2 \cdot V_{cc}/(R_1 + R_2)$

Sol^n:-

$r_e = 15.76\Omega$ (from Ex 4)

$R_i = R_1 \| R_2 \| R_S = 1.85k\Omega$

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

$Av_{mid} = \frac{v_o}{v_i} = \frac{-(R_C \| R_L)}{r_e} = \underline{\underline{-119}}$ $\quad \underline{g_m = 0.03S}$

$Av_{mid} = -g_m(R_C \| R_L)$

$r_\pi = \frac{\beta V_T}{I_{CQ}}$

$C_i = C_{wi} + C_{be} + C_{mi}$

$= C_{wi} + C_{be} + (1 - A_v) C_{bc}$

$I_{CQ} = 1mA$

$= 6pF + 36pF + [1-(-119)4]pF$

$r_\pi = 2.6k\Omega$

$= 522pF$

$g_m = 0.038$

$f_{Hi} = \frac{1}{2\pi R_{eq} C_i}$ $\quad R_{eq} = R_S \| R_1 \| R_2 \| R_i =$

$= \frac{1}{2\pi (\quad)(522pF)}$

$\boxed{f_{Hi} = 470kHz}$

$R_{eq} = R_C \| R_L = 4k\Omega \| 10k\Omega = 2.86k\Omega$

$C_o = C_{wo} + C_{ce} + C_{mo} = 8pF + 1pF + (1 - \frac{1}{-119})4pF$

$C_o = 13.03pF$

$f_{Ho} = \frac{1}{2\pi R_{eq} C_o} = \frac{1}{2\pi(2.86k\Omega)(13.03pF)} = 8.542MHz$

$\boxed{f_{Ho} = 8.542MHz}$

Select Lowest value of f_H among f_{Hi} & f_{Ho}

$\boxed{f_H = \text{Lowest}(f_{Hi}, f_{Ho})}$

Ex 8) Determine the high-cut off frequencies for the network in fig (B) using following parameters,

HF // FET //

$C_G = 0.01\mu f$	$R_S = 1K\Omega$	$C_{gd} = 2pF$
$C_D = 0.5\mu f$	$R_L = 2.2K\Omega$	$C_{gs} = 4pF$
$C_S = 2\mu f$	$I_{DSS} = 8mA$	$C_{ds} = 0.5pf$
$R_S' = 10K\Omega$	$V_P = -4V$	$C_{wi} = 5pf$
$R_G = 1M\Omega$	$r_d = \infty\ \Omega$	$C_{wo} = 6pf$
$R_D = 4.7K\Omega$	$V_{DD} = 20V$	
&	& $V_{GSQ} = -2V$	

$g_m = g_{mo}\left(1 - \frac{V_{GSQ}}{V_P}\right)$

$g_m = \frac{2I_{DSS}}{|V_P|}\left(1 - \frac{V_{GSQ}}{V_P}\right)$

$g_m = 2mS$

Soln:- $R_{eq} = R_S' \| R_G = 10K\Omega \| 1M\Omega = 9.9K\Omega$

$A_{v\,mid} = -g_m(R_D \| R_L)$
$= -2(4.7 \| 2.2k)$
$= -3$

MOSFET

$K_n = 0.25mA/V^2$

$V_{TN} = 1V$

$V_{GS} = 3V$

$g_m = 2K_n(V_{GS} - V_{TN})$

$C_i = C_{wi} + C_{gs} + (1 - A_v) C_{gd}$
$= 5pf + 4pf + (1+3)2pf$

$C_i = 17pf$

$f_{Hi} = \frac{1}{2\pi R_{eq} C_i} = \frac{1}{2\pi (9.9K\Omega) \times 17pf} = 945.67\ kHz$

$R_{eq} = R_S' \| R_G$

$R_{eq} = R_D \| R_L$
$= 4.7K\Omega \| 2.2K\Omega = 1.5K\Omega$

$C_o = C_{wo} + C_{ds} + C_{mo} = 6pf + 0.5pf + \left(1 - \frac{1}{-3}\right)2pf = 9.17pf$

$f_{Ho} = \frac{1}{2\pi R_{eq} C_o} = \frac{1}{2\pi (1.5K\Omega)(9.17pf)} = 11.57\ MHz$

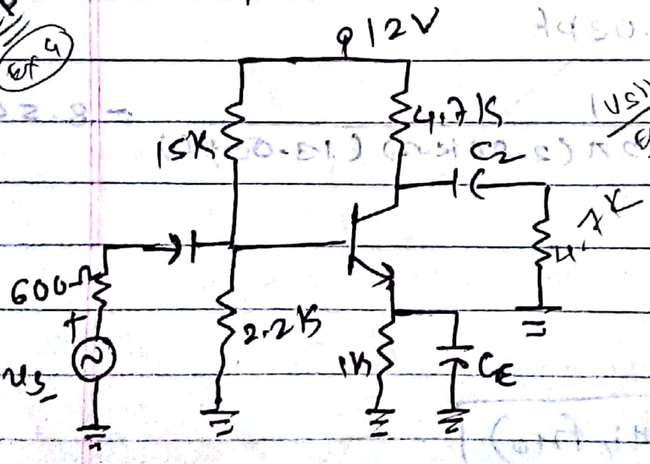
select lowest value $f_H = 945.67kHz$

Design problem

Ex 9) For the CE amplifier shown, calculate the size of bypass & coupling capacitors to provide $f_L = 10Hz$

LF

(Ex 4)



Assuming $h_{fe} = 100$

(using exact Analysis)

$V_{th} = \frac{R_2 \cdot V_{cc}}{R_1 + R_2} = 1.535$

$R_{th} = R_1 \| R_2 = 15k \| 2.2k = 1.919k$

$I_B = \frac{V_{th} - V_{BE}}{R_{th} + (1+\beta) R_E} = \frac{(1.535 - 0.7)}{1.919k + (101)1k}$

$I_B = 81.3$

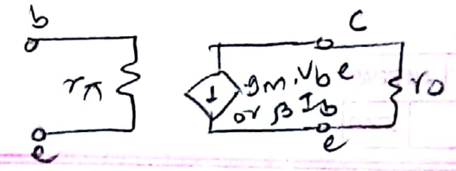
$I_C = 100 \times I_B = 0.8113mA$

Using Approx Analysis

$V_B = \frac{R_2 \cdot V_{CC}}{R_2+R_1} =$

$I_E = V_E/R_E =$

$r_e = \frac{26mV}{I_E} =$ $\quad r_\pi = \beta r_e$



$r_\pi = \frac{\beta}{g_m} = \frac{100 \times 26mV}{0.8113 \times 10^{-3}} = 3.2k$

$r_\pi = \frac{\beta V_T}{I_{CQ}}$

$R_i = R_1 || R_2 || r_\pi = 15k || 2.2k || 3.2k = 1.12k$

$g_m = \frac{I_{CQ}}{V_T}$ $\quad \cancel{Z_o = 4.7k}$

$f_{L_{C_E}} = 10Hz = \frac{1}{2\pi R_{eq} C_E}$ $\quad R_{eq} = R_E || \left[\frac{r_\pi + (R_1||R_2||R_S)}{1+\beta}\right]$

$= 1k || \left[\frac{3.2k + (15k||2.2k||0.6k)}{101}\right]$

$= 34.94\Omega$

$C_E = \frac{1}{2\pi R_{eq} f_{L_{C_E}}}$

$\therefore C_E = \frac{1}{2\pi (10) \times 34.94} = 455.74\mu f \approx 470\mu f$

$f_{L_{C_1}} = \left(\frac{f_L}{10}\right) = 1 Hz = \frac{1}{2\pi R_{eq} C_1}$ $\quad R_{eq} = R_i + R_S$

$= (R_1 || R_2 || r_\pi) + R_S$

$= (1.12k + 600)$

$=$

$\therefore C_1 = \frac{1}{2\pi R_{eq} f_{L_{C_1}}} = \frac{1}{2\pi (1.12k + 600) \times 1} = 92.58\mu f$

$= 100\mu f$

$f_{L_{C_2}} = \left(\frac{f_L}{20}\right) = \frac{10}{20} = 0.5 = \frac{1}{2\pi R_{eq} C_2}$ $\quad R_{eq} = R_C + R_L$

$= 4.7k + 4.7k$

$= 9.4k$

$\therefore C_2 = \frac{1}{2\pi R_{eq} f_{L_{C_2}}} = \frac{1}{2\pi (9.4k) \times 0.5} = 33.88\mu f \approx 39\mu f$

$\therefore$

$$\boxed{\begin{aligned} C_1 &= 100\mu f \\ C_2 &= 39\mu f \\ C_E &= 470\mu f \end{aligned}}$$

$\boxed{\text{Verification}}$:- with the above capacitor values

$f_L = f_{L_{C_E}} + f_{L_{C_1}} + f_{L_{C_2}}$

$= \frac{1}{2\pi \times 34.94 \times 470\mu f} + \frac{1}{2\pi (1.12k + 600) 100\times10^{-6}} + \frac{1}{2\pi (9.4k) \times 39\times10^{-6}}$

$= 9.696 + 0.925 + 0.434$

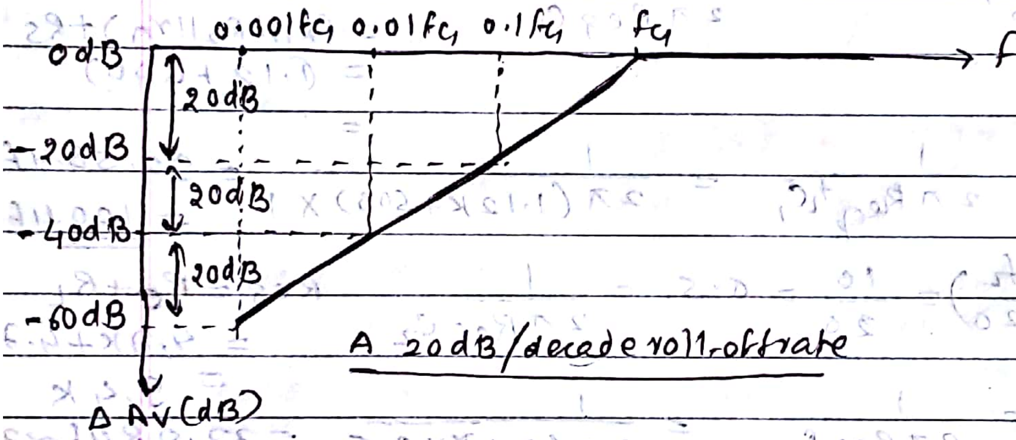
$\boxed{f_L = 11 Hz}$ & given $\boxed{f_L = 10 Hz}$ $\quad$ { If we increase $C_E \approx 560\mu f$, $f_L = 9.5 Hz < 10 Hz$ }

* Gain Roll-off :-

- The term roll-off rate is used to describe the rate at which an RC ckt causes the voltage gain of an amplifier to decrease once the f_{c_1} or f_{c_2} has been passed.

- Low-freq roll off is calculated using, $\Delta A_{V(dB)} = 20 \log \frac{1}{\sqrt{1+(\ldots}}$
  where f = freq of operation.

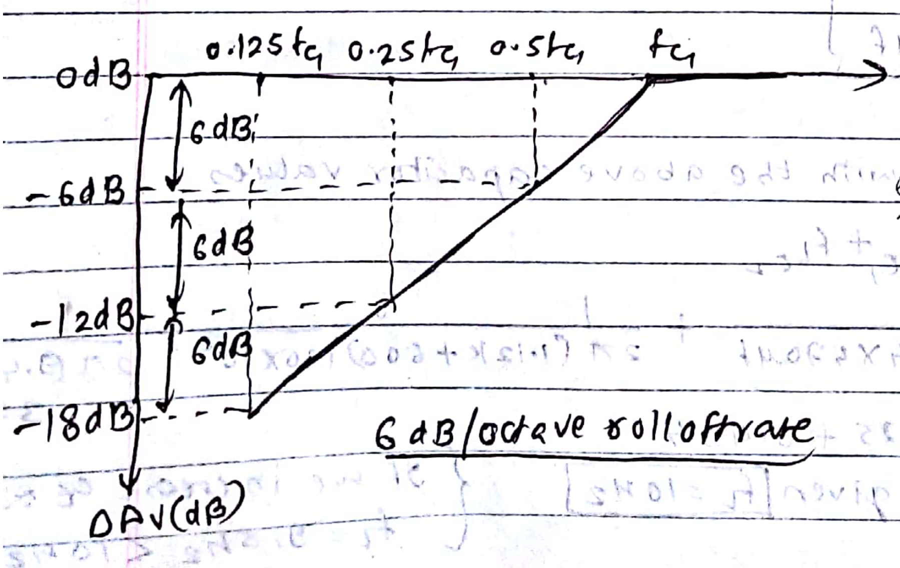
- The $\Delta A_{V(dB)}$ eqⁿ can be used to show that every RC ckt has a roll-off rate of 20 dB per decade.



A 20dB/decade roll-off rate

- The $\Delta A_{V(dB)}$ eqⁿ can also be used to show that every RC ckt has a roll-off rate of 6dB/octave.

octave corresponds to doubling or halving of frequency.

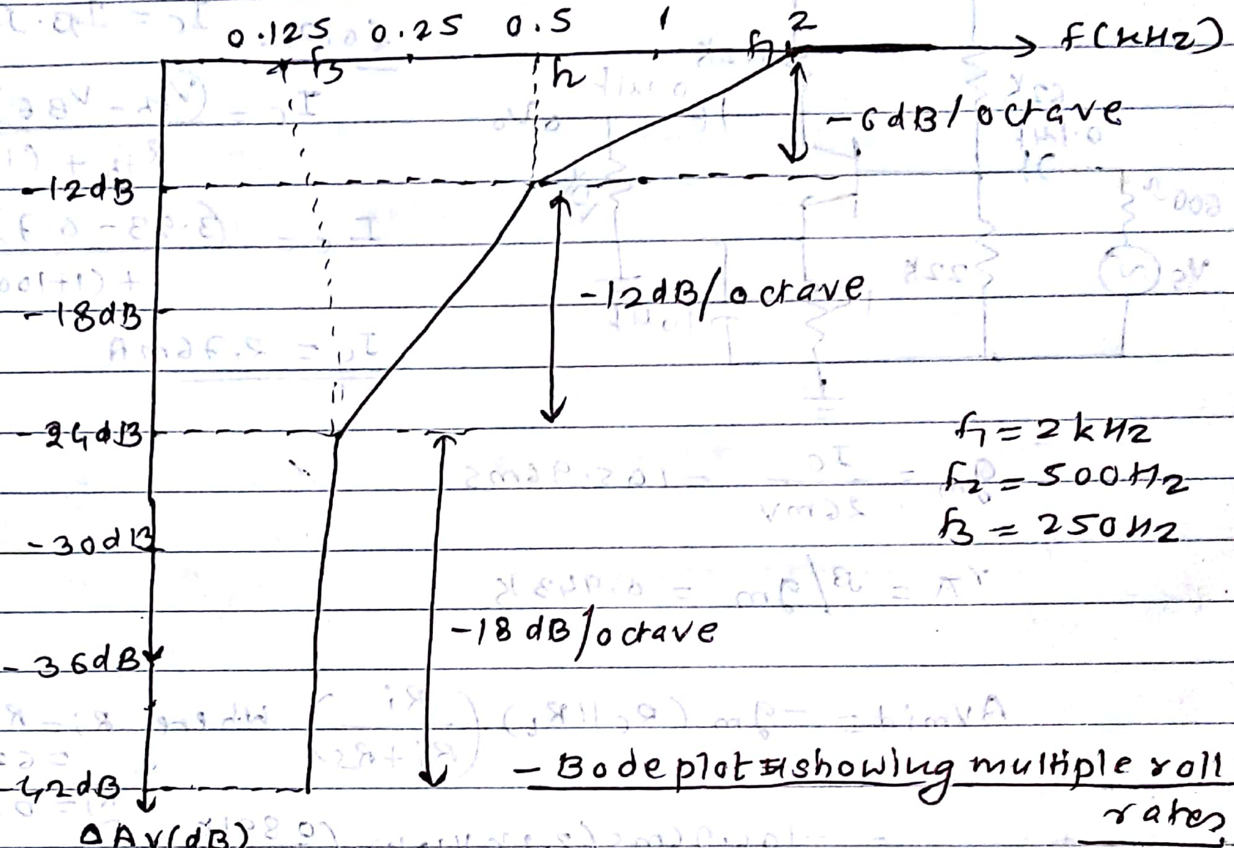


6dB/octave roll-off rate

0 20 40 80

octave1 octave2

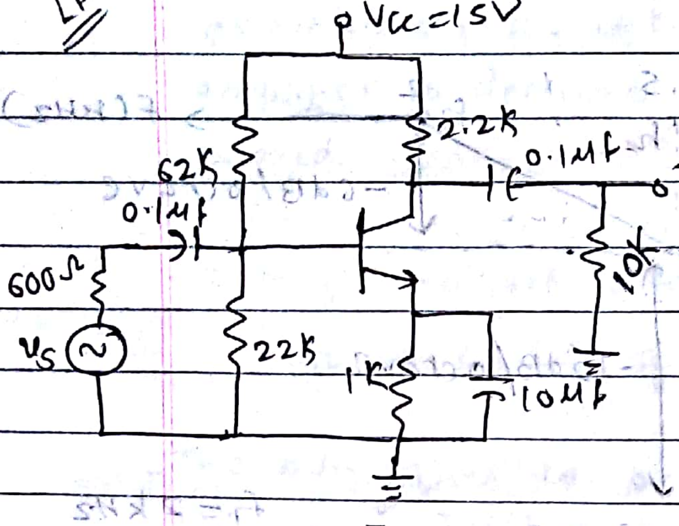
- When a circuit (such as BJT amplr) contains more than one RC ckt, the effects of $\Delta A_V(dB)$ values are additive.



- Bode plot showing multiple roll off rates -

Ex10). Determine the low frequency response of BJT amplifier shown, given $\beta = 100$.

LF



$V_{th} = \frac{R_2 V_{cc}}{R_1+R_2}$

$R_{th} = R_1 \| R_2$

Soln:- $I_C = I_B \cdot \beta$

$$I_C = \frac{(V_{th} - V_{BE})}{R_{th} + (1+\beta) R_E} \cdot \beta$$

$$I_C = \frac{(3.93 - 0.7)}{+ (1+100) \times 1k} \times 100$$

$I_C = 2.76mA$

$$g_m = \frac{I_C}{26mV} = 105.96ms$$

$r_\pi = \frac{\beta V_T}{I_{CQ}}$ $\quad r_\pi = \beta / g_m = 0.943k$

$$A_{Vmid} = -g_m (R_C \| R_L) \cdot \left(\frac{R_i}{R_i + R_s}\right)$$ where $R_i = R_1 \| R_2 \| r_\pi$

$= 62k \| 22k \| 0.943$

$R_i = 0.891k$

$$= -106.96ms \cdot (2.2k \| 10k) \cdot \left(\frac{0.891k}{0.891k + 600}\right)$$

$A_{Vmid} = -115.26$

$$A_{V_{Low}} = \frac{A_{Vmid}}{\sqrt{2}} = \frac{-115.26}{\sqrt{2}} = -81.5 \quad (\text{at } f = f_L)$$

$$f_{LC_1} = \frac{1}{2\pi (R_i + R_s) C_1} = \frac{1}{2\pi (0.891k + 0.6k)(0.1 \times 10^{-6})} = 1067.98$$

$$f_{LC_2} = \frac{1}{2\pi (R_C + R_L) C_2} = \frac{1}{2\pi (2.2k + 10k)(0.1 \times 10^{-6})} = 130.52Hz$$

$$f_{LCE} = \frac{1}{2\pi R_{eq} C_E} \qquad R_{eq} = R_E \| \left[\frac{r_\pi + (R_1 \| R_2 \| r_\pi)}{(\beta + 1)}\right]$$

$$f_{LCE} = \frac{1}{2\pi \times 1k \| \left[\frac{0.943k + (62k \| 22k \| 0.6k)}{100}\right] \times 10 \times 10^{-6}}$$

$\boxed{f_{LCE} = 1072.87Hz}$

$f_L = f_{LC_1} + f_{LC_2} + f_{LC_E} = 1067.98\,Hz + 130.52\,Hz + 1072.87\,Hz$

$$\boxed{f_L = 2.27\,kHz}$$

DC cond^n :-

or $V_B = \frac{R_2}{R_2+R_1} \cdot V_{cc} = \frac{22K}{(22+62)K} \times 15V =$

$I_E = \frac{V_E}{R_E} = \frac{V_B - 0.7}{R_E} = \frac{-0.7}{1K} =$

$r_e = \frac{26mV}{I_E} =$ $\quad r_\pi = \beta r_e =$

$g_m = \frac{1}{r_e} =$

$r_\pi = \beta / g_m = 100/.$

Mid Band Gain :- $A_{Vmid} = -g_m (R_C \| R_L) = \frac{v_o}{v_i}$

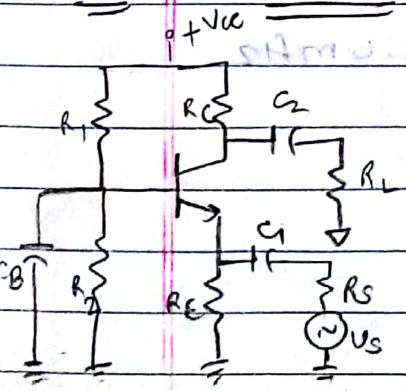
$A_V = A_{Vmid} \cdot \underbrace{\left(\frac{R_i}{R_i+R_S}\right)}_{v_i/v_s} = \frac{v_o}{v_s} = \frac{v_o}{v_i} \times \frac{v_i}{v_s}$ where $R_i = R_1 \| R_2 \| r_\pi$

Ex.11) Design a $\boxed{CB}$ amplifier to give a low 3dB freq of 150 Hz.

a) calculate values of C_1, C_2 & C_B, given $R_1 = 7K$, $R_2 = 4.3K$, $R_E = 330\Omega$, $R_C = 5K$, $R_L = 5K$, $R_S = 200\Omega$, $C_\pi = 15pF$, $C_\mu = 1pF$, $r_\pi = 1.4K$, $g_m = 57.14\,mS$, $\beta_f = 80$

b) calculate A_{mid} c) the high 3dB freq

Sol^n :- for CB, $f_L = f_{C_1} = 150 = \frac{1}{2\pi R_{eq} C_1}$ $\qquad r_\pi = \beta r_e = 1.4K$



CB Amplifier

$R_{eq} = \left[R_E \| \left(\frac{\beta r_e}{1+\beta_f}\right)\right] + R_S$

$= 16.42 + 200$

$= 216.42\,\Omega$

$\therefore C_1 = \frac{1}{2\pi \times 216.42 \times 150} = \underline{\underline{4.9\,\mu F}}$

C_B = Bypass capacitor

$$f_{L C_B} = \frac{f_L}{10} = 15Hz = \frac{1}{2\pi R_{eq} C_B}$$

$$R_{eq} = R_1 \| R_2 \| [\beta r_e + (1+\beta)(R_E \| R_S)] = 2.16k\Omega$$

$$\therefore C_B = \frac{1}{2\pi \times 2.16k \times 15} = 4.914\mu f$$

$$\underline{\underline{C_B = 4.914\mu f}}$$

$$f_{L C_2} = \frac{f_L}{20} = \frac{150}{20} = 7.5Hz = \frac{1}{2\pi R_{eq} C_2}$$

$$R_{eq} = R_C \| R_L = 5k \| 5k = \cancel{10k}\ 2.5$$

$$C_2 = \frac{1}{2\pi \times 10k \times \cancel{7.5}\ 2.5} = \cancel{2.12\mu f} = 4.24\mu f$$

$$\underline{\underline{C_2 = \cancel{2.12\mu f}}}\ 4.24\mu f$$

$$A_{mid} = g_m (R_C \| R_L)\left(\frac{Z_i}{Z_i + R_S}\right) \text{ where } Z_i = R_E \| \left(\frac{\beta r_e}{1+\beta_f}\right) = 16.42$$

$$= \cancel{(57.14 \times 10^{-3})(5k \| 5k)} \left(\frac{16.42}{16.42 + 200}\right)$$

$$\boxed{A_{vmid} = 10.84}$$

$$f_H = \frac{1}{2\pi (R_{c\pi} C_\pi + R_{cu} C_u)}$$

$$R_{c\pi} = R_S \| R_E \| \beta r_e \| (1/g_m)$$

$$= 200 \| 300 \| 1.4k \| \left(\frac{1}{57.14 \times 10^{-3}}\right)$$

$$= 15.1\Omega$$

$$R_{cu} = R_C \| R_L = 5k \| 5k = 2.5k\Omega$$

$$f_H = \frac{1}{2\pi [(15.1 \times 15pf) + (2.5k \times 1pf)]} = 58.4 MHz$$

$$\boxed{f_H = 58.4 MHz}$$

Ex 12) Design a $\boxed{CC}$ amplifier to give a lower 3dB freq $f_L = 150\,Hz$ calculate values C_b, C_1 & C_2 given $R_1 = 7K$, $R_2 = 4.3K$, $R_E = 330\Omega$, $R_S = 200\Omega$, $R_L = 5K$, $g_m = 57.14\,mS$, $B_f = 80$, $r_\pi = 1.4K$, $C_\pi = 15\,pF$, $C_\mu = 1\,pF$ calculate f_H

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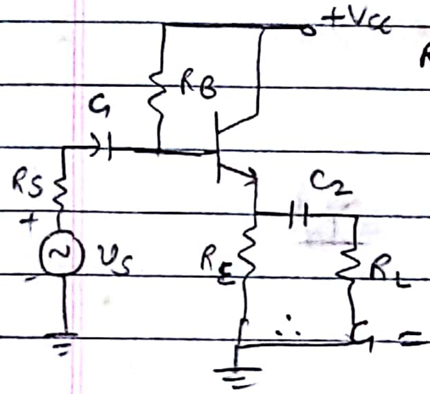
$r_\pi = \beta r_e$

$$f_L = f_{LC_1} = 150\,Hz = \frac{1}{2\pi R_{eq} C_1}$$



$$R_{eq} = Z_i + R_S = \{R_1 \| R_2 \| [\beta r_e + (1+\beta_f)(R_E \| R_L)]\} + R_S$$

$$= 7K \| 4.3K \| [1.4K + 81(330 \| 5K)]$$

$$= 2.42K + 0.2K$$

$$= 2.62K$$

$$\therefore C_1 = \frac{1}{2\pi(2.62K)(150)} = \underline{\underline{0.405\,\mu F}}$$

$$F_{LC_2} = \frac{f_L}{10} = 15\,Hz = \frac{1}{2\pi R_{eq} C_2}$$

$$R_{eq} = Z_o + R_L$$

$$= \left[R_E \| \left(\frac{\beta r_e + (R_1 \| R_2 \| R_S)}{1+\beta_f}\right)\right] + R_L$$

$$= 330 \| \left(\frac{1.4K + (7K \| 4.3K \| 200)}{81}\right)$$

$$= 18.48\,\Omega + 5K$$

$$= 5.018K$$

$$C_2 = \frac{1}{\cancel{2\pi(18.48)(15)}} \quad \frac{1}{2\pi(5.018K)(15)}$$

$$\underline{\underline{C_2 = 2.11\,\mu F}}$$

$$R_{C\pi} = \beta r_e \| \left[\frac{(R_1 \| R_2 \| R_S) + (R_E \| R_L)}{1 + g_m (R_E \| R_L)}\right]$$

$$= 1.4K \| \left[\frac{(7K \| 4.3K \| 200) + (330 \| 5K)}{1 + 57.14\times10^{-3}(330 \| 5K)}\right] = 1.4K \| 26.54$$

$$= 26.05\,\Omega$$

$$R_{CH} = (R_1 \| R_2 \| R_S) \| [\beta r_e + (1+\beta_f)(R_E \| R_L)]$$

$$= 7k \| 4.3k \| 200 \| [1.4k + 81(330 \| 5k)]$$

$$= 184.7\Omega$$

$$f_H = \frac{1}{2\pi (R_{C\pi} C_\pi + R_{C\mu} C_\mu)}$$

$$= \frac{1}{2\pi (26.05 \times 15pf + 184.7 \times 1pf)}$$

$$\boxed{f_H = 276.7 MHz}$$