BJT & JFET Frequency Response

1) Consider the following mathematical eqⁿ,

$$a = b^x$$

$$\Rightarrow \boxed{x = \log_b a}$$

2) Common logarithm $x = \log_{10} a$, base = 10

Natural logarithm $y = \log_e a = \ln a$, base = e

3) $$\ln a = 2.3 \log_{10} a$$

Proof :- $$\log_a b = \frac{\log_{10} b}{\log_{10} a}$$

$$\therefore \log_e a = \frac{\log_{10} a}{\log_{10} e} = \frac{\log_{10} a}{\log_{10} 2.718} = \frac{\log_{10} a}{0.434}$$

$$\Rightarrow \log_e a = 2.3 \log_{10} a$$

$$\Rightarrow \boxed{\ln a = 2.3 \log_{10} a}$$

4) $\log_{10} 1 = 0$, $\log_{10} 10 = 1$, $\log_{10} 100 = 2$, $\log_{10} 1000 = 3$, ---etc

Decibel :-

The background surrounding the term decibel has its origin in the established fact that power and audio levels are related on a logarithmic basis. That is, an increase in power level, say 4 to 16 watt, does not result in an audio level increase by a factor of $\frac{16}{4} = 4$. It will increase by a factor 2, i.e

$$\boxed{\log_4 16 = 2} \quad \text{or} \quad \frac{\log 16}{\log 4} = \log_4 16 = 2$$

For standardization, the bel (B) [derived from surname of Alexander Graham Bell] is defined by the following eq^n to relate power levels P_1 & P_2 :

$$\boxed{G = \log_{10} \frac{P_2}{P_1} \text{ Bel},} \quad \text{— (1)}$$

For practical purpose, deciBel (dB) is defined such that 10 deciBel = 1 Bel.

$$\therefore \boxed{G_{dB} = 10 \log_{10} \frac{P_2}{P_1} \text{ dB}} \quad \text{— (2)}$$

BPUT → 2mark

1) Difference betn dBm & dBu.

Ans : The deciBel rating is a measure of the difference between two power levels. For there must a specified terminal (output) power P_2 be a reference power level (P_1).

When the 1-mW (milli-watt) level is employed as the reference level, the deciBel symbol appears as $\boxed{\text{dBm.}}$

$$\boxed{G_{dBm} = 10 \log_{10} \frac{P_2}{1\,mW}, \text{ dBm}}$$

Similarly, If 1 - μW (Micro watt) is employed as reference level, the decibel symbol appears as $\boxed{dB\mu}$.

$$\boxed{G_{dB\mu} = 10 \log_{10} \frac{P_2}{1\,\mu W}, \ dB\mu}$$

$\rightarrow$ $G_{dB} = 10 \log \frac{P_2}{P_1}$

$$= 10 \log \left(\frac{\frac{V_2^2}{R}}{\frac{V_1^2}{R}} \right)$$

$$= 10 \log \frac{V_2^2}{V_1^2}$$

$$= 10 \log \left(\frac{V_2}{V_1}\right)^2$$

$$\boxed{G_{dB} = 20 \log \frac{V_2}{V_1}}, \ dB$$

1 watt = 30 dBm, 60 dBμ

$G_{dBm} = 10 \log \left(\frac{1}{10^{-3} W}\right) = 30\,dBm$

$d_{B\mu} = 10 \cdot \log \left(\frac{1}{10^{-6} W}\right) = 60\,dB\mu$

$\rightarrow$ If magnitude of overall voltage gain of a cascaded system is given by,

$$|A_{vT}| = |A_{v1}|\,|A_{v2}| \cdots |A_{vn}|,$$

In dB,

$$20 \log |A_{vT}| = 20 \log |A_{v1}| + 20 \log |A_{v2}| \cdots + 20 \log A_{vn}$$

$$\Rightarrow \boxed{G_{dB\,Total} = (G_{dB1} + G_{dB2} - - - + G_{dBn})\ dB}$$

General freq Considerations

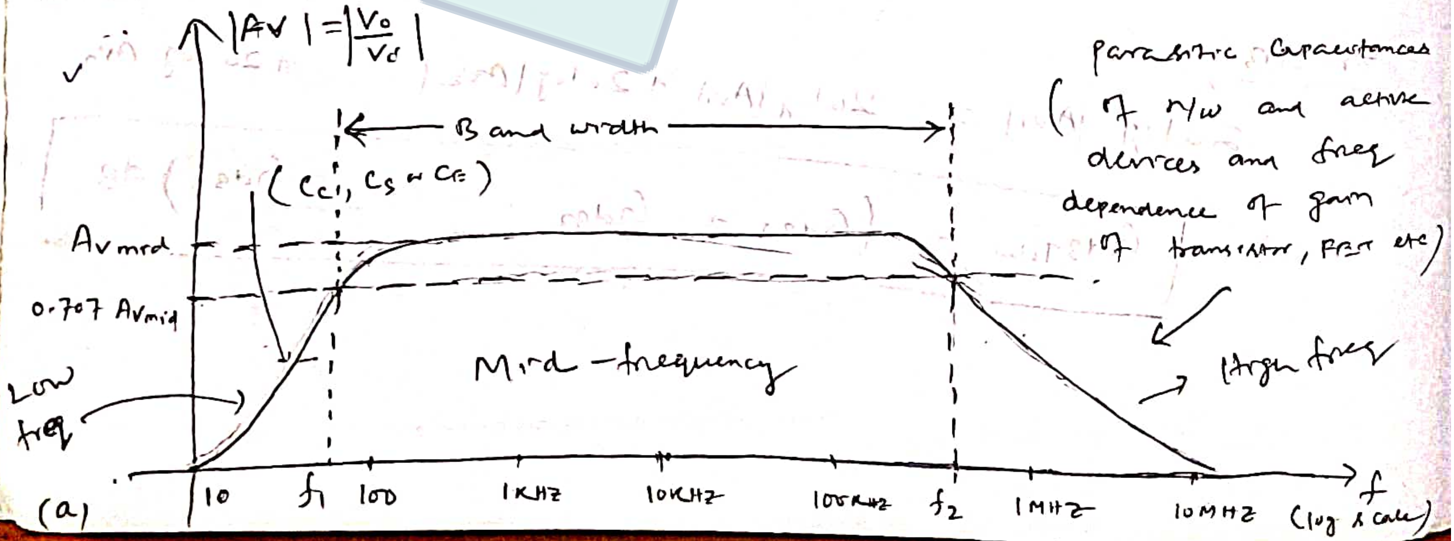
The freq of the applied signal can have a pronounced effect on the response of a single-stage or multistage n/w. All the analysis done before has been for midfreq spectrum.

At low frequency, the coupling and bypass capacitors can no longer be replaced by the short-cut approximation because of the increase in reactance of these elements.

The freq-dependent parameters of the small-signal equivalent circuits and stray capacitive elements associated with active device and the n/w will limit the high-freq response of the system.

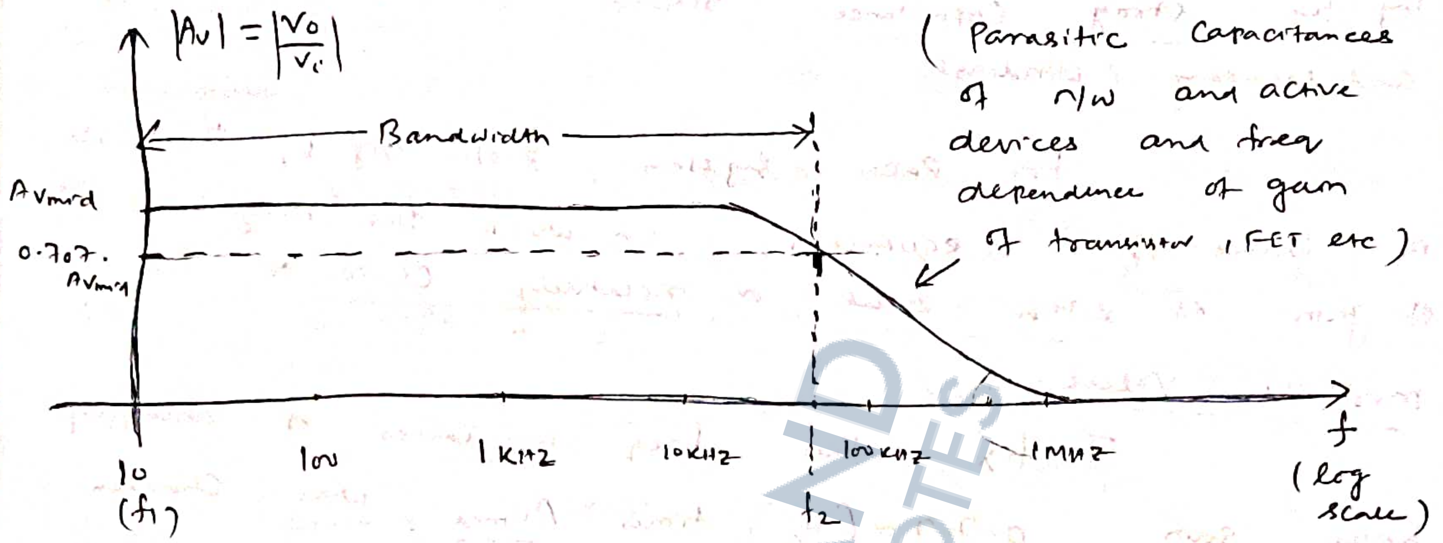
An increase in the number of stages of a cascaded system will also limit both the high and low freq response.

The magnitude of the gain response curves of an RC-coupled, direct-coupled and transformer-coupled amplifier systems are provided in fig 1, a, b, c. The primary reason for the drop in

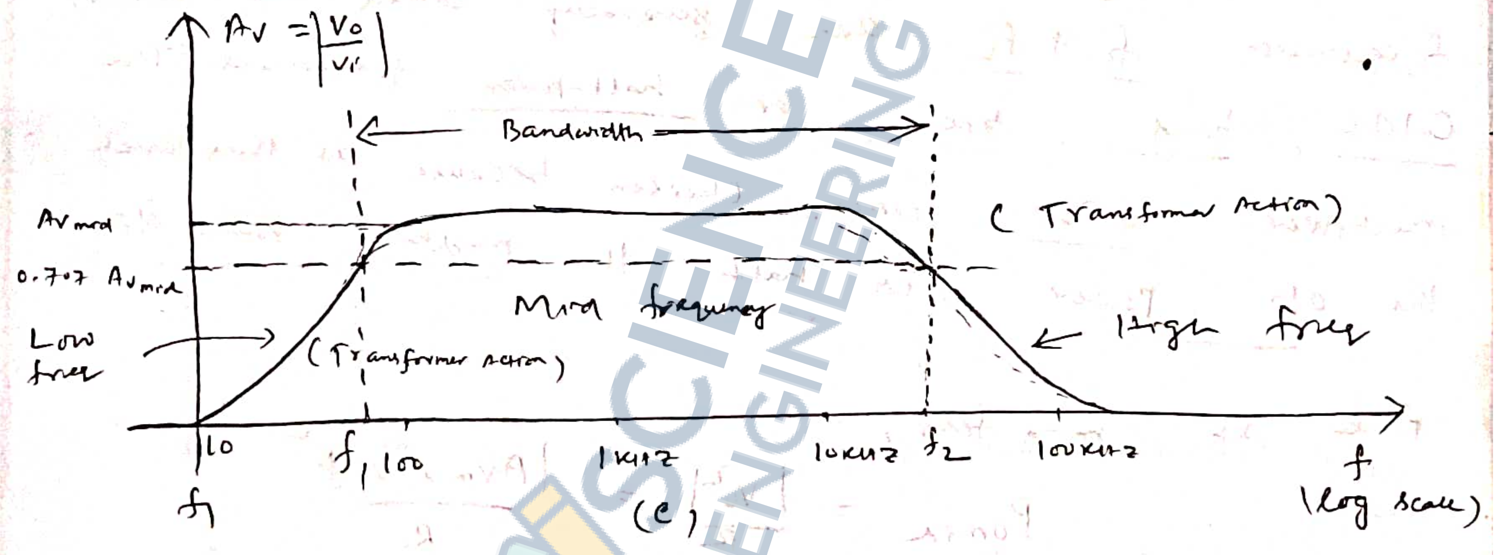


(a)

gain at low & high freq have been indicated within the parenthesis.



(b)



(c)

fig 1 :- Gain versus freq : (a) RC- Coupled amplifier (b) Direct - Coupled amplifier (c) Transformer - Coupled amplifier

<u>Note :-</u> <u>Drop in gain for transformer - Coupled System :-</u>

A low freq, ($X_L = 2\pi fL$), due to shorting effect (~~dan~~ across the i/p terminals of the transformer) of magnetizing inductive reactance. The gain must be zero at $f=0$, since at this point there is no longer change in flux established through the core to induce a secondary or o/p

Voltage.

The high freq response is controlled primarily by the stray capacitance between the turns of primary and secondary windings.

For each system of fig 1, there is a band of frequencies in which the magnitude of gain is either equal or relatively close to the midband value.

To fix the freq boundaries of relatively high gain, 0.707 or $\left(\frac{1}{\sqrt{2}}\right)$ times A_{Vmid} was chosen to be gain at the cutoff levels. The corresponding frequencies f_1 & f_2 are generally called the corner, cutoff, band, break or half-power frequencies. The multiplier 0.707 was chosen because at this level the o/p power is half the midband power o/p,

At mid freq,

$$P_{omid} = \frac{|V_o^2|}{R} = \frac{|A_{Vmid} \cdot V_i|^2}{R_o}$$

At half-power frequencies,

$$P_{oHPF} = \frac{|0.707\, A_{Vmid} \cdot V_i|^2}{R_o} = 0.5\,\frac{|A_{Vmid} \cdot V_i|^2}{R_o}$$

$$\Rightarrow \boxed{P_{oHPF} = 0.5\, P_{omid}}$$

→ The Bandwidth or pass band of each system is determined by f_1 & f_2, i.e

$$\boxed{\text{bandwidth (BW)} = f_2 - f_1}$$

Normalized Plot :-

Previously plot is A_v Vs freq

In normalized plot, plot is ~~Avmd~~ $\frac{A_v}{A_{vmid}}$ Vs freq.

When $A_v = A_{vmid}$, i.e. at mid band freq, the value of $\frac{A_v}{A_{vmid}} = 1$. At half-power freq, the resulting level is 0.707 or $\frac{1}{\sqrt{2}}$.

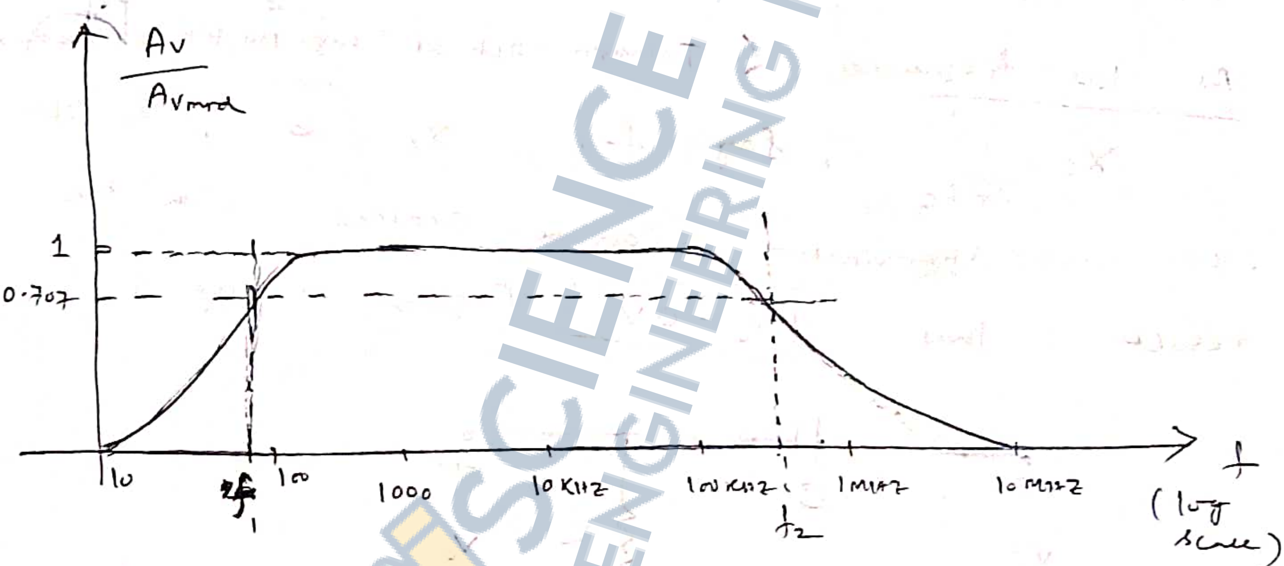


fig 2:- Normalized gain Vs freq plot

dB Plot :-

$$\left.\frac{A_v}{A_{vmid}}\right|_{dB} = 20\log_{10}\frac{A_v}{A_{vmid}}$$,

At $A_v = A_{vmid}$, $20\log 1 = 0\,dB$

At cutoff freq, $20\log\frac{1}{\sqrt{2}} = -3\,dB$.

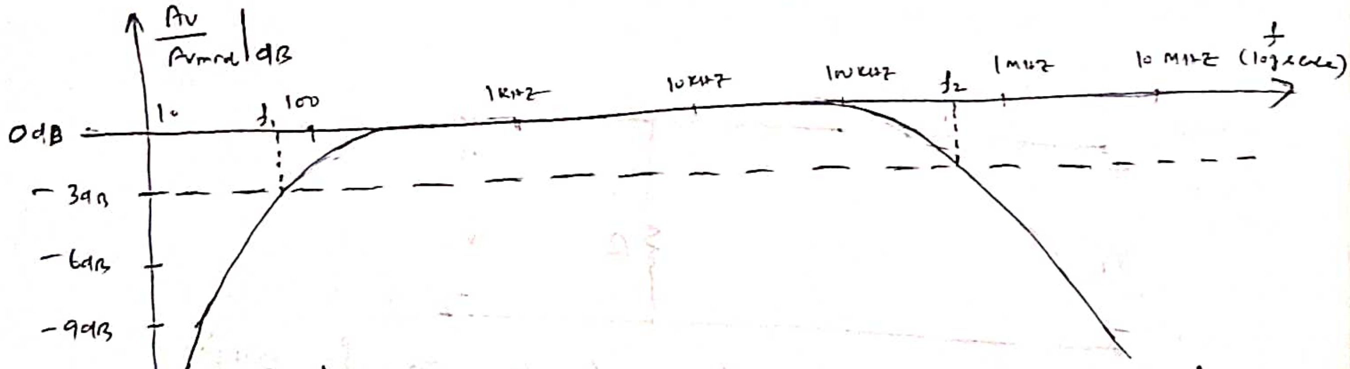


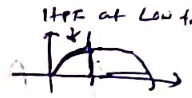
fig 3:- Decibel plot of the normalized gain Vs freq

Low-freq Analysis - Bode Plot

In the low-freq region of the single-stage BJT or FET amplifier, it is the R-C combination formed by the n/w capacitor C_C, C_E and C_s and the n/w resistive parameters that determine the cutoff frequencies.

An R-C n/w similar to fig 4 (given below) can be established for each capacitive element and the freq at which o/p voltage drops to 0.707 of its max^m value determined.

At low frequencies, < Transistor act as high pass filter



$X_C = \frac{1}{2\pi f C}$, say $f = 0$, $X_C = \infty$, an open ckt approximation can be applied, with the result that $\boxed{V_o = 0V}$. [Shown in fig 5]

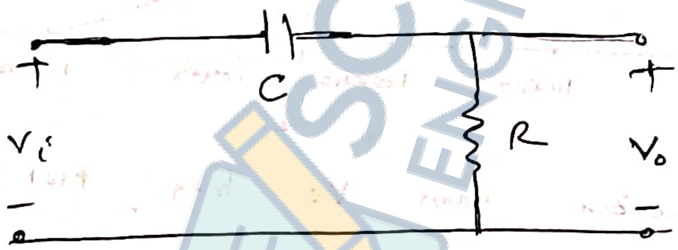


fig: 4:- RC ~~ckt of fig 3, at f=0.~~ combination that will define a low cutoff frequency.

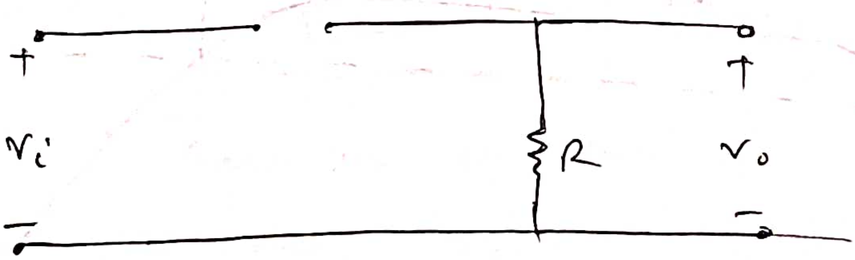


fig 5:- R-C ckt of fig 4 at f = 0Hz

At very high freq

$X_C = \frac{1}{2\pi f c} \simeq 0\,\Omega$ and the short-cut equivalent can be substituted for capacitor as shown in fig 6. The result is that $\boxed{V_o = V_i}$, at high frequencies.

fig 6:- RC equivalent of fig 4 at very high frequencies

→ Between the 2 extremes, the ratio $A_V = \frac{V_o}{V_i}$, is shown in fig 7. As the freq increases, the capacitive reactance decreases and more of the i/p voltage appears across the o/p terminals.

fig 8:- Low freq response for the R-C ckt

The o/p & i/p voltage are related by the voltage-divider rule in the following manner; [fig 4]

$$V_o = \frac{V_i}{R - jX_C} \cdot R \quad \text{—— (1)}$$

$$|V_o| = \frac{RV_i}{\sqrt{R^2 + X_C^2}}$$

$\because \frac{1}{j\omega c} = \frac{-j}{\omega c} = -jX_C$

Lets take $X_C = \frac{1}{\omega c}$

When $X_C = R$,

$$V_o = \frac{R V_i}{\sqrt{R^2 + R^2}} = \frac{\cancel{R} V_i}{\sqrt{2} \cancel{R}} = \frac{V_i}{\sqrt{2}}$$

$$\Rightarrow \quad |A_V| = \frac{V_o}{V_i} = \frac{1}{\sqrt{2}} = 0.707$$

So, the freq at which $X_C = R$, the o/p will be 70.7% of the i/p.

The freq at which this occurs is determined from,

$$X_C = R$$

$$\Rightarrow \frac{1}{2\pi f_1 C} = R$$

$$\Rightarrow \boxed{f_1 = \frac{1}{2\pi R C}} \quad \text{—— (2)}$$

<u>Interms of logs</u>,

$$G_V = 20 \log A_V = 20 \log\left(\frac{1}{\sqrt{2}}\right) = -3\,dB$$

When $V_o = V_i$, $G_V = 20 \log \frac{V_o}{V_i} = 20 \log 1 = 0\,dB$

∴ The freq at which there is a 3dB drop from midband level or the freq at which o/p is 70.7% or $\frac{1}{\sqrt{2}}$ of i/p is known as <u>-3dB freq</u> or <u>cutoff freq</u>.

$$A_v = \frac{V_o}{V_i} = \frac{R}{R - jX_C}$$ [From eqn ①, last page]

$$= \frac{1}{1 - j\left(\frac{X_C}{R}\right)}$$

$$= \frac{1}{1 - j\left(\frac{1}{2\pi f C R}\right)}$$

$$= \frac{1}{1 - j\left(\frac{\frac{1}{2\pi RC}}{f}\right)}$$

$$\boxed{A_v = \frac{1}{1 - j\left(\frac{f_1}{f}\right)}} \quad — ③$$ [$\because f_1 = \frac{1}{2\pi RC}$ eqn ②]

$$A_v = \underbrace{\frac{1}{\sqrt{1+\left(\frac{f_1}{f}\right)^2}}}_{\text{Magnitude of } A_v} \underbrace{\angle \tan^{-1}\left(\frac{f_1}{f}\right)}_{\text{Phase by which } V_o \text{ leads } V_i}$$

[$\because -\tan^{-1}\left(-\frac{f_1}{f}\right) = \tan^{-1}\left(\frac{f_1}{f}\right)$]

For the magnitude,
when $f = f_1$,

$$|A_v| = \frac{1}{\sqrt{1+1}} = \frac{1}{\sqrt{2}} = 0.707 = -3\text{ dB}.$$

In logarithmic form, the gain in dB,

$$\boxed{A_v(\text{dB}) = 20\log_{10}\frac{1}{\sqrt{1+(f_1/f)^2}}} \quad — ④$$

$$\therefore (A_v)_{dB} = -20\log_{10}\left[1+\left(\frac{f_1}{f}\right)^2\right]^{\frac{1}{2}}$$

$$= -\frac{1}{2}\times 20 \log_{10}\left[1+\left(\frac{f_1}{f}\right)^2\right]$$

$$(A_v)_{dB} = -10\log_{10}\left[1+\left(\frac{f_1}{f}\right)^2\right] \quad — ⑤$$

For frequencies, where $f << f_1$ or $\left(\frac{f_1}{f}\right) >> 1$,

Then eqⁿ ⑤, can be approximated as,

$$(A_v)_{dB} = -10 \log_{10} \left(\frac{f_1}{f}\right)^2$$

finally,

$$\boxed{(A_v)_{dB} = -20 \log_{10} \frac{f_1}{f}} \quad \text{—— ⑥} \quad \text{for } f << f_1$$

At $f = f_1$, $\frac{f_1}{f} = 1$, and $(A_v)_{dB} = -20 \log_{10} 1 = 0 \text{ dB}$

At $f = \frac{f_1}{2}$, $\frac{f_1}{f} = 2$, $(A_v)_{dB} = -20 \log 2 \simeq -6 \text{dB}$

At $f = \frac{f_1}{4}$, $\frac{f_1}{f} = 4$, $(A_v)_{dB} = -20 \log 4 \simeq -12 \text{dB}$

At $f = \frac{f_1}{10}$, $\frac{f_1}{f} = 10$, $(A_v)_{dB} = -20 \log 10 = -20 \text{ dB}$

Plot of these points is indicated in fig 9 from $0.1 f_1$ to f_1. This results in a straight line when plotted against a log scale. A straight line is also drawn for the condition of 0 dB for $f >> f_1$,

$$\left(\because \ (A_v)_{dB} = -20 \log \left[1 + \left(\frac{f_1}{f}\right)^2\right]^{\frac{1}{2}} \right.$$

$$\text{If } f >> f_1 \quad \frac{f_1}{f} \simeq 0$$

$$= -20 \log 1$$

$$\left. \simeq 0 \ \right)$$

→ ∴ For $f << f_1$, it is sloped line.
$f >> f_1$, it is 0 dB. [Asymptotes]

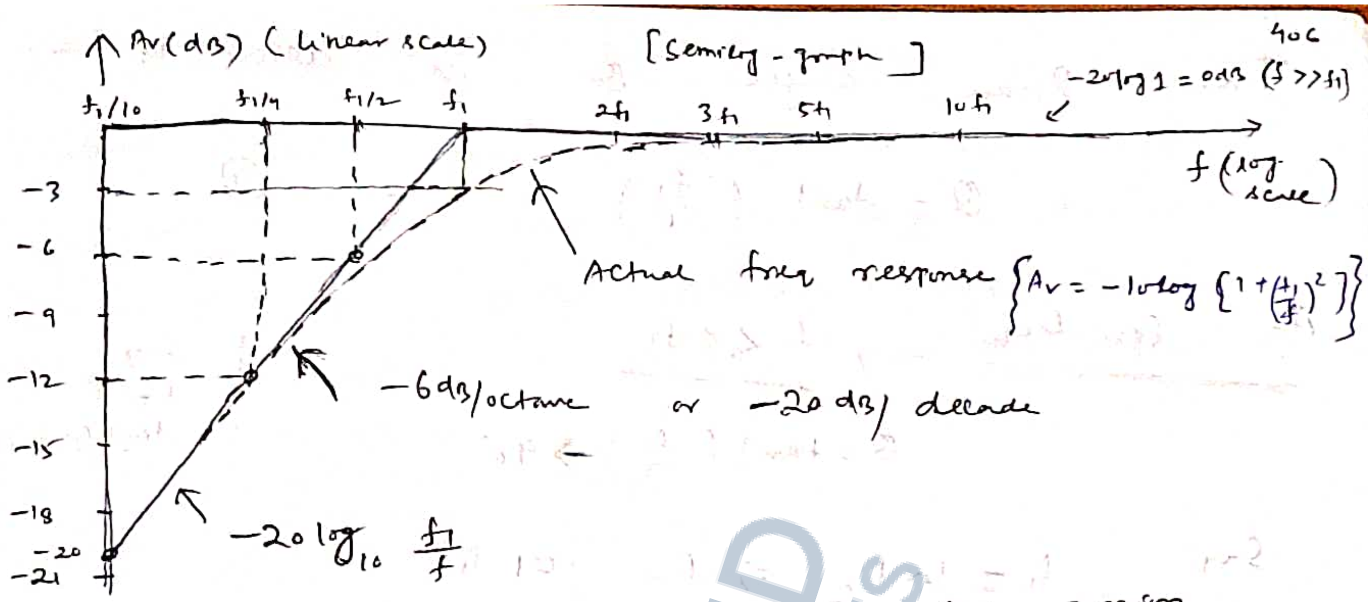


fig:-9 Bode Plot for low-freq region

($-10\log[1+1] = -3dB$)

→ When $f = f_1$, there is 3dB drop from mid-band level. The piecewise linear plot of the asymptotes and associated breakpoints is called a Bode Plot of the magnitude versus freq.

→ ~~At $f_1 = 0dB$, At $f_1/2 = -6dB$.~~

At $f = f_1$, gain $= 0dB$, At $f = f_1/2$, gain $= -6dB$

∴ A change in freq by a factor of 2, equivalent to 1 octave, results in a 6dB change in the gain.

At $f = f_1$, gain $= 0dB$, At $f = \frac{f_1}{10}$, gain $= -20dB$

∴ A change in freq by a factor of 10 equivalent to 1 decade, results in a 20dB change in the gain.

Imp:- Slope of the straight line (Asymptote) $= -6dB$/octave or $= -20dB$/decade.

The Phase angle θ is determined from,

$$\theta = \tan^{-1}\left(\frac{f_1}{f}\right) \quad \longleftarrow \quad (7)$$

At low freq, $f << f_1$

$$\theta = \tan^{-1}\left(\frac{f_1}{f}\right) \rightarrow 90^\circ$$

Say $f = 0$, $\tan^{-1}(\infty) = 90^\circ$

Say $f_1 = 100 f$, $\Rightarrow f = .01 f_1$

$$\theta = \tan^{-1}\left(\frac{f_1}{f}\right) = \tan^{-1}(100) = 89.4^\circ$$

For $f = f_1$, $\tan^{-1}(1) = 45^\circ$.

for $f >> f_1$, $\tan^{-1}\left(\frac{f_1}{f}\right) \rightarrow 0^\circ$

Ex:- $f = 100 f_1$, $\tan^{-1}(.01) = 0.573^\circ$

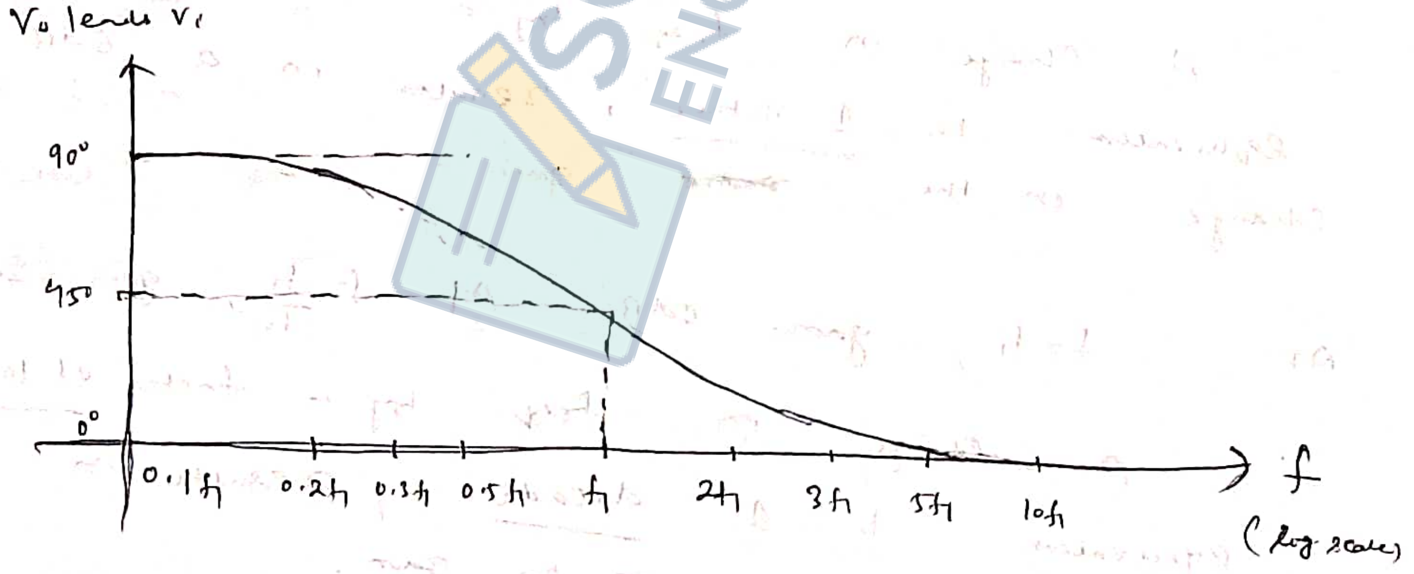


Fig: 10 - Phase Response for the R-C circuit

Low-freq response - BJT Amplifier :-

Consider a loaded Voltage-divider BJT bias configuration. As shown in the figure the capacitors C_s, C_c and C_E will determine the low-freq response. We will now examine the impact of each independently.

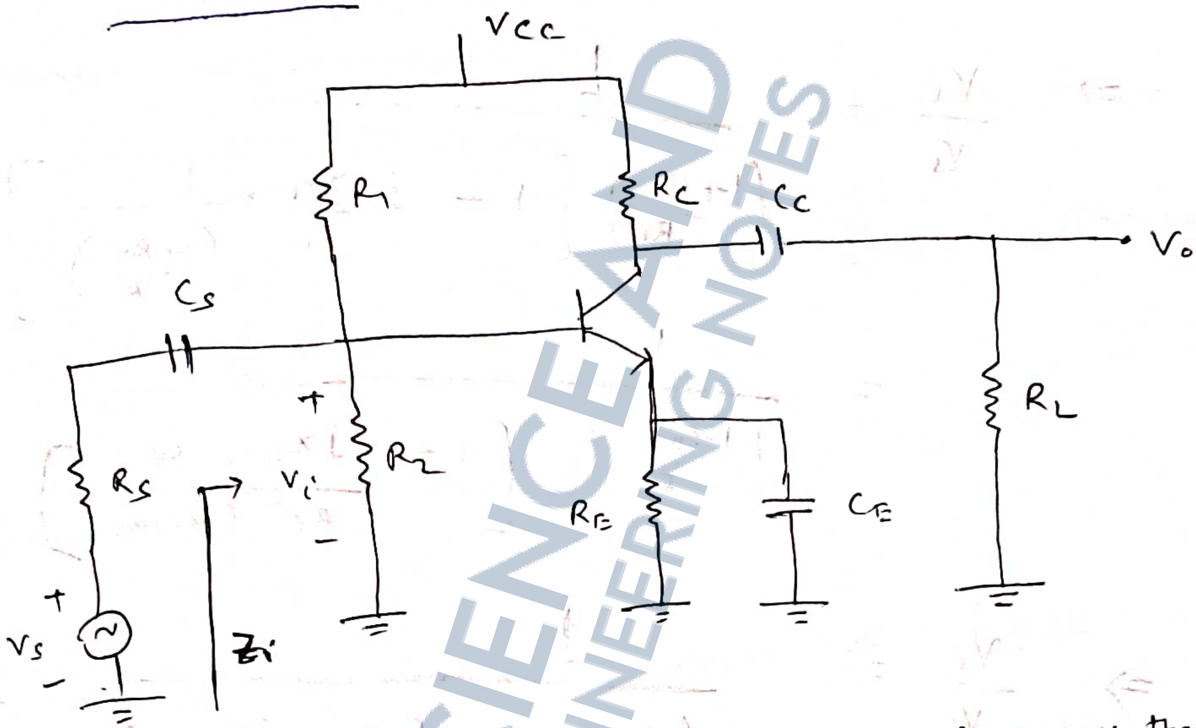


fig 11:- Loaded BJT amplifier with capacitors that affect the low-freq response.

C_s

Since C_s is normally connected between the applied source and the active device, the general form of the RC configuration is established by the n/w given below.

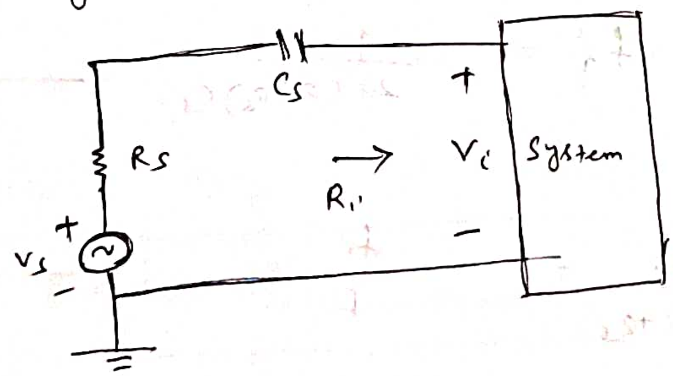


fig 12:- Determining the effect of C_s on the low freq response

Applying the voltage-divider rule;

$$V_i = \frac{V_s}{R_s + R_i - jX_{C_s}} \times R_i$$

$$\Rightarrow \frac{V_i}{V_s} = \frac{R_i}{R_s + R_i - jX_{C_s}} = \frac{1}{1 + \frac{R_s}{R_i} - j\frac{X_{C_s}}{R_i}}$$

$$\Rightarrow \frac{V_i}{V_s} = \frac{1}{\left(1 + \frac{R_s}{R_i}\right)\left[1 - j\frac{X_{C_s}}{R_i} \times \frac{1}{\left(1 + \frac{R_s}{R_i}\right)}\right]}$$

$$= \frac{1}{\left(1 + \frac{R_s}{R_i}\right)\left[1 - \frac{jX_{C_s} \times \cancel{R_i}}{\cancel{R_i}\,(R_i + R_s)}\right]}$$

$$\Rightarrow \frac{V_i}{V_s} = \frac{1}{\left(1 + \frac{R_s}{R_i}\right)\left(1 - \frac{jX_{C_s}}{R_i + R_s}\right)} \quad \text{—— (1)}$$

Let the factor,

$$\frac{X_{C_s}}{R_i + R_s} = \frac{1}{2\pi f C_s} \cdot \frac{1}{(R_i + R_s)} \quad \Big| \; \because X_C = \frac{1}{2\pi f C}$$

$$= \frac{1}{2\pi (R_i + R_s) C_s} \cdot \frac{1}{f}$$

Defining, $$f_i = \frac{1}{2\pi (R_i + R_s) C_s} \quad \text{—— (1A)}$$

we have,

$$\frac{X_{C_s}}{R_i + R_s} = \frac{f_i}{f} \quad \text{—— (2)}$$

Putting eqn ②, in eqn ①, we have

$$\frac{V_i'}{V_s} = \frac{1}{\left(1+\frac{R_s}{R_i}\right)\left(1-j\frac{f_l}{f}\right)}$$

$$\Rightarrow \boxed{A_v = \left[\frac{R_i}{R_i+R_s}\right] \cdot \left[\frac{1}{1-j\,(f_l/f)}\right]} \quad \text{——— ③}$$

~~Comparing the above~~

For midband frequencies, the n/w will appear as shown in fig below

< (Capacitor Short Ckted)

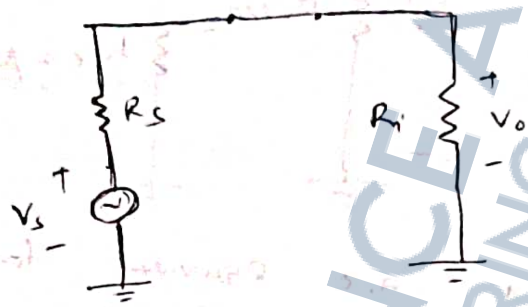


fig 13 :- High freq equivalent of fig 12.

$$A_{v\,mid} = \frac{V_o}{V_s} = \frac{R_i}{R_i+R_s} \quad \text{——— ④}$$

$\because V_o = \frac{V_s}{R_s+R_i} \times R_i$

$\Rightarrow \frac{V_o}{V_s} = \frac{R_i}{R_s+R_i}$

Dividing eqn ③ by eqn ④, we have

$$\Rightarrow \frac{A_v}{A_{v\,mid}} = \frac{1}{1-j\left(\frac{f_l}{f}\right)} \quad \text{——— ⑤}$$

Comparing eqn ⑤ with eqn $\left[A_v = \frac{1}{1-j\left(\frac{f_1}{f_a}\right)} \right.$ - eqn ③ - Page 404 $\left.\right]$

The cutoff freq is defined by 'f_l' above

and $\boxed{f_{Ls} = f_l = \frac{1}{2\pi\,(R_i+R_s)\,C_s}}$ ——— ⑥

L → Low
s → C_s

[Note:- 'f_l' is defined in eqn 1(A) Last page]

At f_{LS}, the voltage V_o will be 70.7% 411 of the midband value determined by eqn (4).

Assuming only C_S is there and C_E & C_C are short circuited, the a.c equivalent of ~~fig 11~~ n/w for i/p section of fig 11 will appear as shown below.

fig 14:- Localized a.c equivalent for C_s

The value of R_i is determined by,

$$\boxed{R_i = R_1 // R_2 // \beta r_e} \quad \text{——} \; (7)$$

C_c (Coupling Capacitor)

Since the coupling capacitor is normally connected between the o/p of the active device and the applied load, the RC configuration that determines the low-cutoff freq due to C_c appears in fig 15. From fig 15, the total series resistance is now $(R_o + R_L)$, and cutoff freq due to C_c is determined by

$$\boxed{f_{LC} = \frac{1}{2\pi (R_o + R_L) C_c}} \quad (8)$$

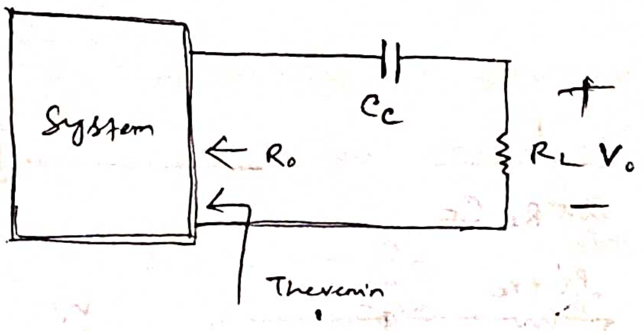


fig: 15 Determining the effect of C_c on the low-freq response.

Ignoring the effect of C_s & C_E, we have that the o/p voltage V_o will be 70.7% of its midband value at f_{Lc}.

For the n/w of fig 11, the ac equivalent n/w for the o/p section with $V_i = 0V$ appears in fig 16.

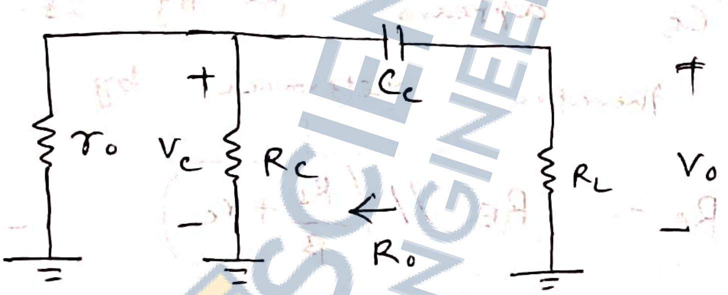


fig 16:- Localized ac equivalent for C_c with $V_i = 0V$

∴
$$\boxed{R_o = R_C \parallel r_o} \quad \text{——} \quad (9)$$

C_E (Emitter Bypass Capacitor)

To determine f_{LE}, the n/w 'seen' by C_E must be determined as shown in fig 17. Once the level of R_e is established, the cutoff freq due to C_E can be determined using the following eq^n,

$$f_{LE} = \frac{1}{2\pi R_e C_E} \qquad (10)$$

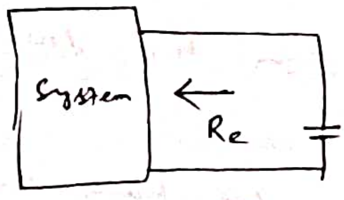


fig 17:- Determining the effect of C_E on the low-freq response

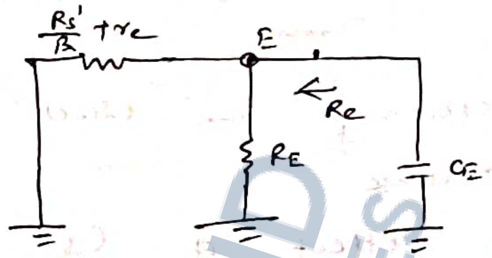


fig 18:- Localized ac equivalent of C_E

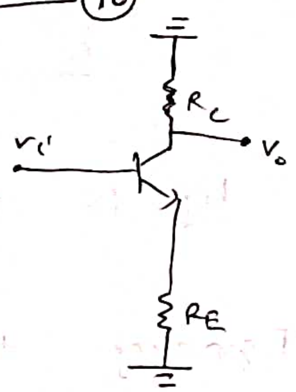


fig:19:- N/w employed to describe the effect of C_E on the amplifier gain.

→ For the n/w of fig 11, the a.c equivalent as 'seen' by C_E appears in fig 18. The value of R_e is therefore determined by

$$R_e = R_E \,//\left(\frac{R_s'}{\beta} + r_e\right) \qquad (11)$$

where $R_s' = R_s \,//\, R_1 \,//\, R_2$

Gain

$$A_v = \frac{-R_C}{R_E + r_e} \qquad (12)$$

[See voltage-divider unbypassed capacitor re model]

When $R_E = 0\Omega$, max^m gain is available.

At low freq, $X_C = \frac{1}{2\pi f c} \simeq \infty$, bypass capacitor is 'open circuited', so R_E appears in the gain eq^n above, resulting min^m gain.

At ~~large~~ freq increases, $X_C = \frac{1}{2\pi f c} \simeq 0$

'R_E' is effectively shorted by C_E. The result is a maxm or midband gain is determined,

$$A_V = \frac{-R_C}{r_e} \quad \text{--- (B)}$$

At f_{LE}, the gain will be 3 dB below midband value determined with R_E "shorted".

Ex :- 1

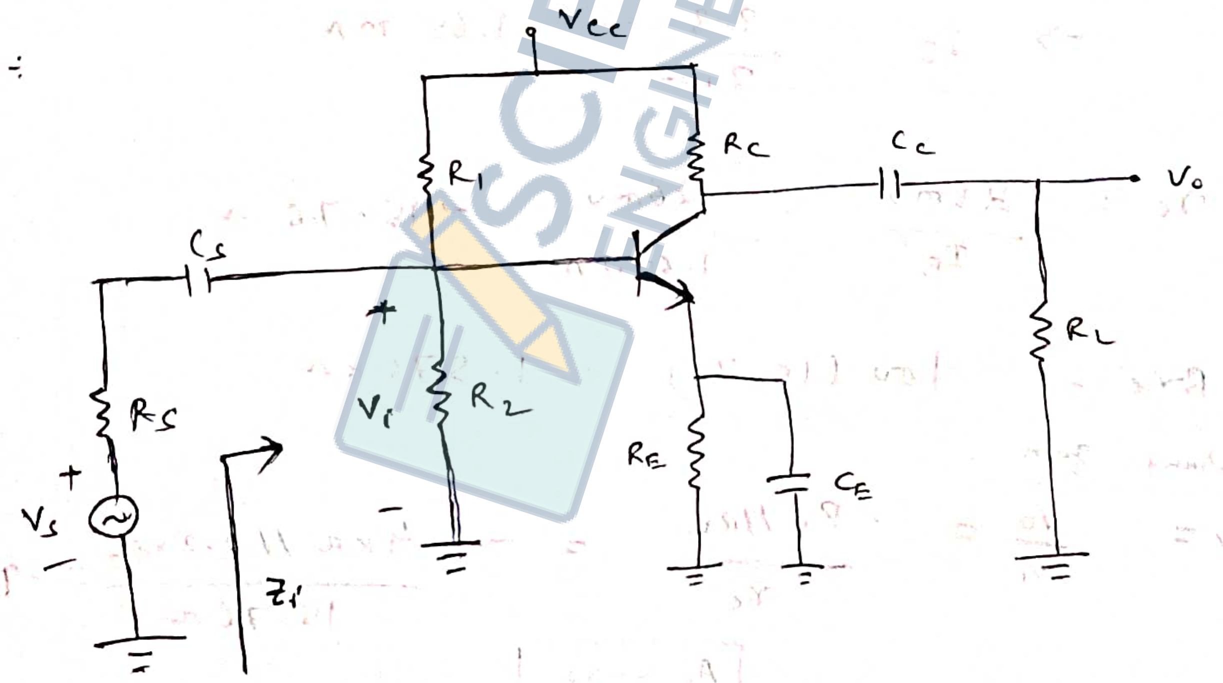
Determine the lower cutoff freq for the n/w given below, using the following parameters

$C_s = 10\mu F$, $C_E = 20\mu F$, $C_C = 1\mu F$

$R_s = 1k\Omega$, $R_1 = 40k\Omega$, $R_2 = 10k\Omega$, $R_E = 2k\Omega$, $R_C = 4k\Omega$

$R_L = 2.2k\Omega$, $\beta = 100$, $r_o = \infty$, $V_{CC} = 20V$

Ans :-



Determining 'r_e' for d.c condition

Checking the condition $\beta R_E >> 10 R_2$

$100 \cdot 2k\Omega >> 10 \times 10 k\Omega$

$200 k\Omega >> 100 k\Omega$

[For applying approximate model in voltage-divider biasing]

Since the condn is satisfied, we have

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

$$= \frac{20}{40+10} \times 10$$

$$= \frac{20}{\cancel{50}_5} \times \cancel{10}$$

$$V_B = 4 \text{ volt}$$

$$V_{BE} = V_B - V_E = 4 - V_E$$

$$\Rightarrow 0.7 = 4 - V_E$$

$$\Rightarrow V_E = 4 - 0.7 = 3.3 \text{ V}$$

$$\Rightarrow I_E R_E = 3.3 \text{ V}$$

$$\Rightarrow I_E = \frac{3.3}{2 k\Omega} = 1.65 \text{ mA}$$

$$r_e = \frac{26 mV}{I_E} = \frac{26 mV}{1.65 mA} \simeq 15.76 \, \Omega.$$

$$\beta r_e = 100(15.76) = 1.576 k\Omega.$$

Midband gain :-

$$A_V = \frac{V_o}{V_i} = \frac{-R_C \| R_L}{r_e} = -\frac{4 k\Omega \,\|\, 2.2 k\Omega}{15.76 \, \Omega} \simeq -90$$

$$\boxed{A_V = -90}$$

Input impedance

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

$$= 40 k\Omega \,\|\, 10 k\Omega \,\|\, 1.576 k\Omega$$

$$= 8 k\Omega \,\|\, 1.576 k\Omega$$

$$\therefore \boxed{Z_i = 1.32\,k\Omega} = R_i$$

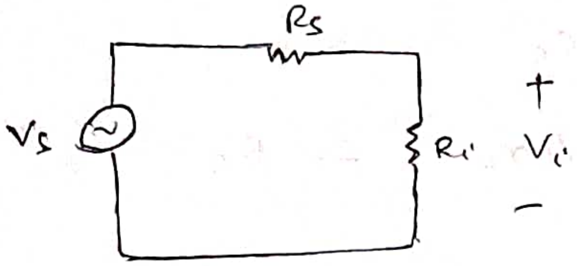


fig: 20:- Determining the effect of R_s on gain A_v.

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

or
$$\frac{V_i}{V_s} = \frac{R_i}{R_i + R_s} = \frac{1.32\,k\Omega}{1.32\,k\Omega + 1\,k\Omega} = 0.569$$

$$A_{v_s} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s} = A_v \cdot \frac{V_i}{V_s}$$

$$= (-90)\,(0.569$$

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

To determine cutoff frequencies

Due to C_s

$$f_{Ls} = \frac{1}{2\pi\,(R_i + R_s)\,C_s} = \frac{1}{2\pi\,(1.32 + 1)\,k\Omega \times 10 \times 10^{-6}}$$

$$f_{Ls} = \frac{1}{2\pi \times 2.32 \times 10^{-2}}$$

$$\boxed{f_{Ls} = 6.86\ \text{Hz}}$$

Due to C_C

$$f_{LC} = \frac{1}{2\pi (R_o + R_L) C_C}$$

where $R_o = R_C \,//\, r_o \simeq R_C$ $(\because r_o \to \infty)$

$$\therefore f_{LC} = \frac{1}{2\pi (R_C + R_L) C_C}$$

$$= \frac{1}{2\pi (4k\Omega + 2.2k\Omega) \cdot 1 \times 10^{-6}}$$

$$= \frac{1}{2\pi \times 6.2 \times 10^{-3}}$$

$$\boxed{f_{LC} \simeq 25.68 \text{ Hz}}$$

Due to C_E

$$f_{LE} = \frac{1}{2\pi R_e C_E}$$

where

$$R_e = R_E \,//\, \left(\frac{R_s'}{\beta} + r_e\right)$$

and $R_s' = R_s \,//\, R_1 \,//\, R_2$

$$\therefore R_s' = 1k\Omega \,//\, 40k\Omega \,//\, 10k\Omega$$

$$= 1k\Omega \,//\, 8k\Omega$$

$$R_s' \simeq 0.889 \, k\Omega$$

$$R_e = R_E \,//\, \left(\frac{R_s'}{\beta} + r_e\right)$$

$$\therefore \quad R_e = R_E \,//\, \left(\frac{0.889\,m}{100} + 15.76\,\Omega \right)$$

$$= 2k\Omega \,//\, 24.65$$

$$R_e \simeq 24.35$$

$$f_{LE} = \frac{1}{2 \times \pi \times 24.35 \times 20 \times 10^{-6}}$$

$$\boxed{f_{LE} \simeq 327\ Hz}$$

Low-freq response - FET amplifier

→ Analysis is similar to BJT amplifier.

→ C_s, C_c, C_E (BJT) ⟷ C_G, C_c, C_s (FET)

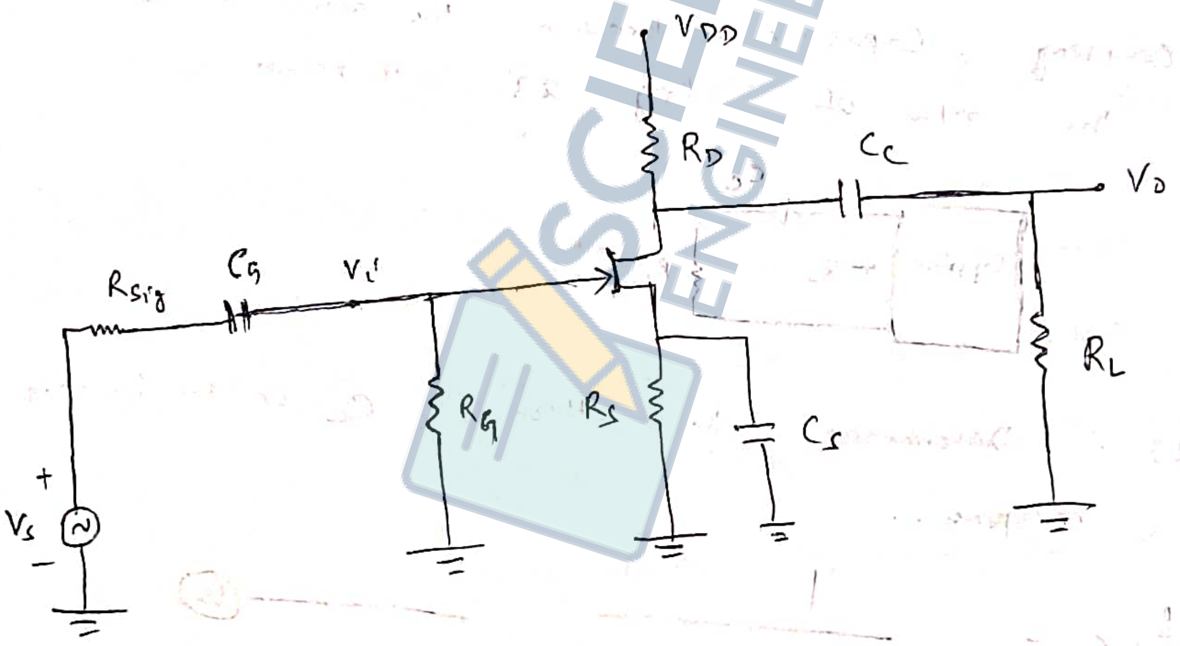


fig 21 :- Capacitive elements that affect the low-freq response of a JFET amplifier.

C_G :- For the coupling capacitor between the source & the active device, the ac equivalent n/w with appear as shown in fig 22

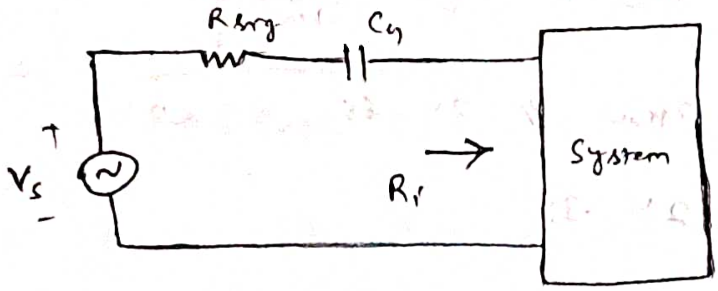


fig 22:- Determining the effect of C_G on the low-freq response.

The cutoff freq determined by C_G will then be,

$$f_{LG} = \frac{1}{2\pi (R_i + R_{sig}) C_G} \quad \text{—— (14)}$$

(Similar to eqn (6))

where $R_i = R_G$ [fig 21]

C_C :-

For coupling capacitor between the active device and the load the n/w of fig 23 with result.

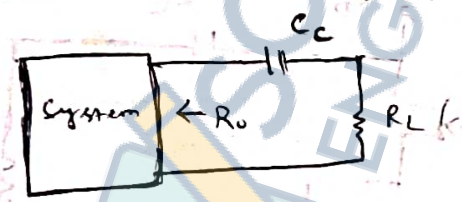


fig 23:- Determining the effect of C_C on the low-freq response.

$$f_{LC} = \frac{1}{2\pi (R_o + R_L) C_C} \quad \text{—— (15)}$$

where $R_o = R_D \,//\, r_d$

C_S :- For the source capacitor C_S, the resistance level of importance is defined by fig 24.

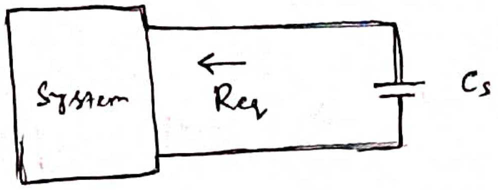


fig 24:- Determining the effect of Cs on the low-freq response

$$f_{LS} = \frac{1}{2\pi R_{eq} C_s} \quad \text{--- (16)}$$

where
$$R_{eq} = \frac{R_s}{1 + R_s\left((1+g_m r_d)/(r_d + R_D)//R_L\right)}$$

$$R_{eq} = \frac{R_s}{1 + R_s\left(\frac{1+g_m r_d}{(r_d + R_D//R_L)}\right)} \quad \text{--- (17)}$$

when $r_d \to \infty$, $R_{eq} = R_s // \frac{1}{g_m}$ --- (18)

Ex-2:- Determine the lower cutoff freq for fig 21 using the following parameters:

$C_G = 0.01\,\mu F$, $C_C = 0.5\,\mu F$, $C_S = 2\,\mu F$

$R_{sig} = 10k\Omega$, $R_G = 1M\Omega$, $R_D = 4.7k\Omega$, $R_S = 1k\Omega$

$R_L = 2.2k\Omega$, $I_{DSS} = 8mA$, $V_P = -4V$, $r_d = \infty\,\Omega$, $V_{DD} = 20V$

Ans:-
$$g_{m0} = \frac{2I_{DSS}}{|V_P|} = \frac{2\times 8mA}{|-4V|} = 4mS$$

$$g_m = g_{m0}\left(1 - \frac{V_{GSQ}}{V_P}\right) = 4\times10^{-3}\left(1 - \frac{(-2)}{-4}\right) = 2mS$$

[∴ Note V_{GSQ} is obtained either by graphical method or mathematical analysis - Discussed in FET biasing]

C_g

$$f_{Lg} = \frac{1}{2\pi (R_1 + R_{sig}) C_g}$$

$$f_{Lg} = \frac{1}{2\pi (1M\Omega + 10k\Omega) \cdot 0.01 \times 10^{-6}}$$

$\because R_1 = R_g = 1M\Omega$

$$f_{Lg} = 15.8 \text{ Hz}$$

C_c

$$f_{Lc} = \frac{1}{2\pi (R_o + R_L) C_c}$$

$$= \frac{1}{2\pi (4.7k\Omega + 2.2k\Omega) \times 0.5 \times 10^{-6}}$$

$R_o = R_D // r_d$

$= R_D // \infty$

$= R_D$

$$f_{Lc} = 46.13 \text{ Hz}$$

C_s

$$f_{Ls} = \frac{1}{2\pi R_{eq} C_s}$$

where $R_{eq} = R_s // \frac{1}{g_m}$

$= 1k\Omega // \frac{1}{2mS}$

$= 1k\Omega // 500\Omega$

$= 333.33\Omega$

$$f_{Ls} = \frac{1}{2\pi (333.33) \times 2 \times 10^{-6}} = 238.73 \text{ Hz}$$

Since f_{Ls} is the largest of 3 cutoff frequencies, it defines the low cutoff freq of the n/w.

→ Mid band gain of the system is determined by,

$$A_{vmid} = \frac{V_o}{V_i} = -g_m (R_D \| R_L)$$

$$A_{vmid} = -(2\times10^{-3})(4.7k\Omega \| 2.2k\Omega)$$

$$= -2\times10^{-3} \times 1.499\times10^{3}$$

$$\boxed{A_{vmid} \simeq -3}$$

imp

Miller Effect Capacitance :- [refer fig 30 page 431]

In high-freq region, the capacitive elements of importance are the inter-electrode (between terminals) capacitances internal to the active device and the wiring capacitance between leads of the network. The large capacitors of the n/w that controlled the low-freq response have all been replaced by their short-ckt equivalent due to their very low reactance levels. $\left[X_C = \frac{1}{2\pi f C} = 0\right]$

For inverting amplifiers (Phase shift of 180° between i/p and o/p resulting in a -ve value for Av), the i/p & o/p capacitance is increased, ~~by~~ because a capacitive level is sensitive to the interelectrode capacitance between the i/p & o/p terminals of the device and gain of the amplifier. In fig 25, this 'feedback' capacitance is defined by 'C_f'.

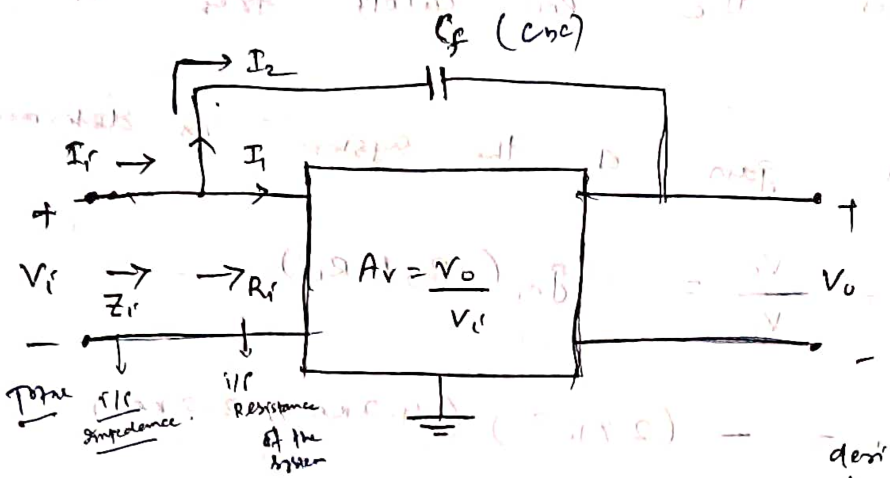


fig 25 :- N/w employed on the ~~deduction~~ derivation of an equation for the Miller i/p capacitance

Applying Kirchhoff's current law gives

$$I_i = I_1 + I_2 \quad \text{—— (19)}$$

Using Ohm's law yields

$$I_i = \frac{V_i}{Z_i}, \quad I_1 = \frac{V_i}{R_i} \quad \text{—— (20)}$$

and

$$I_2 = \frac{V_i - V_o}{X_{cf}} = \frac{V_i - A_v V_i}{X_{cf}} \qquad \left(\because A_v = \frac{V_o}{V_i} \Rightarrow V_o = A_v V_i \right)$$

$$I_2 = \frac{V_i(1 - A_v)}{X_{cf}} \quad \text{—— (21)}$$

Substituting I_i, I_1 & I_2 on eqn (19), we have

$$\frac{\cancel{V_i}}{Z_i} = \frac{\cancel{V_i}}{R_i} + \frac{\cancel{V_i}(1 - A_v)}{X_{cf}}$$

$$\Rightarrow \frac{1}{Z_i} = \frac{1}{R_i} + \frac{1}{(X_{cf}/1 - A_v)} \quad \text{—— (22)}$$

$$X_{C_f} = \frac{1}{w C_f}$$ ——— (23)

$$\therefore \frac{X_{C_f}}{1-A_V} = \frac{1}{w C_f (1-A_V)} = \frac{1}{w C_M} = X_{C_M}$$

where $C_M = C_f(1-A_V)$

Putting eqn (23), in eqn (22), we have

$$\frac{1}{Z_i} = \frac{1}{R_i} + \frac{1}{X_{C_M}}$$ ——— (24)

This equation establish a n/w

fig 26: Demonstrating the effect of the Miller effect capacitance.

Here the feedback capacitor is magnified by the gain of the amplifier. Any interelectrode capacitance at the i/p terminals to the amplifier will simply be added in parallel with elements of fig 26.

In general, the Miller effect i/p capacitance is defined by,

$$\boxed{C_{Mi} = (1-A_V) C_f}$$ ——— (25)

$\therefore$ For any inverting amplifier, the i/p capacitance will be increased by a Miller effect capacitance which is sensitive to the gain of the amplifier and the interelectrode (parasitic) capacitance between i/p & o/p terminals of the active device

→ Miller effect will also increase the level of o/p capacitance, In Fig 27, the parameters of importance to determine the o/p Miller effect are in place.

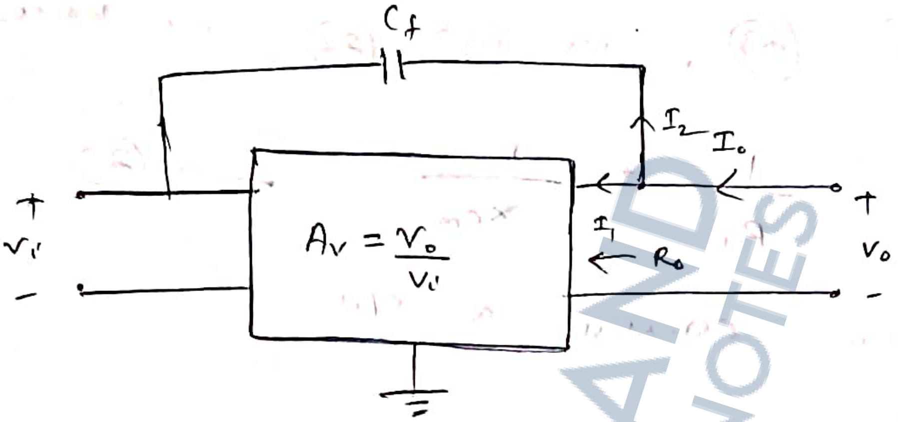


Fig 27:- N/w employed in the derivation of an equation for the Miller o/p capacitance.

Applying KLC

$$I_o = I_1 + I_2 \quad \text{———} \quad (26)$$

with $I_1 = \frac{V_o}{R_o}$ and $I_2 = \frac{V_o - V_i}{X_{Cf}}$

'R_o' is very large, $I_1 \simeq 0$

$$I_o \simeq I_2$$

$$\Rightarrow I_o = \frac{V_o - V_i}{X_{Cf}}$$

$\because A_v = \frac{V_o}{V_i}$

$\Rightarrow V_i = \frac{V_o}{A_v}$.

$$= \left(V_o - \frac{V_o}{A_v}\right) / X_{Cf}$$

$$\Rightarrow I_o = V_o\left(1 - \frac{1}{A_v}\right) / X_{Cf}$$

$$\Rightarrow I_o / V_o = \left[1 - 1/A_v\right] / X_{Cf}$$

$$\Rightarrow \frac{V_o}{I_o} = \frac{X_{cf}}{1 - 1/A_V}$$

But $X_{cf} = \frac{1}{\omega C_f}$

$$\therefore \frac{V_o}{I_o} = \frac{1}{\omega C_f (1 - 1/A_V)} = \frac{1}{\omega C_{Mo}}$$

where $C_{Mo} = C_f \left(1 - \frac{1}{A_V}\right)$

& C_{Mo} is called Miller o/p capacitance.

$$\boxed{C_{Mo} = C_f \left(1 - \frac{1}{A_V}\right)} \quad \text{—— (27)}$$

For the usual situation, $|A_V| >> 1$, eqn (27) reduce to

$$\boxed{C_{Mo} \simeq C_f} \quad \text{—— (28)}$$

$|A_V| >> 1$.

From eqn (25) & (27), both i/p & o/p ~~[illegible]~~ capacitance increased, because Av is -ve [Inverting Amplifier]

So $(1 - A_V)$ or $\left(1 - \frac{1}{A_V}\right)$ both are +ve.

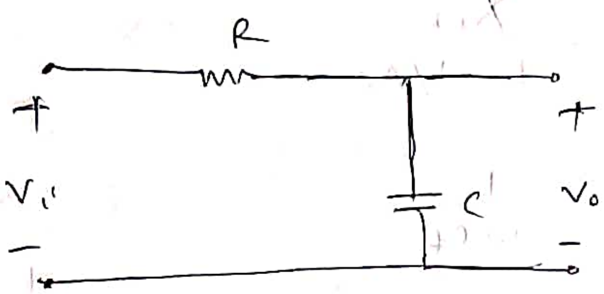
$\therefore C_{Mi}$ & C_{Mo} are always greater than or equal to C_f.

High-freq response- ~~[illegible]~~

In the high-freq region, the RC network of concern has the configuration appearing in fig 28.

$$X_C = \frac{1}{2\pi f C}$$

LPF

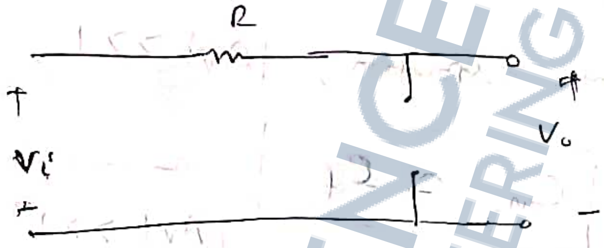


$V_o = R$

fig 28:- RC Combination that will define a high cut-off frequency.

At very low freq

$X_C = \frac{1}{2\pi f C}$, $f \to 0$, $X_C \to \infty$

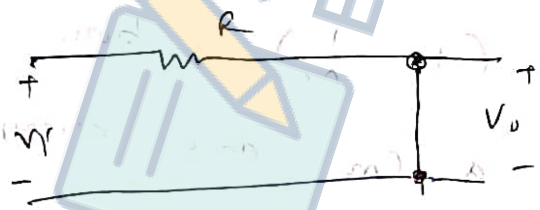


Capacitor open circuit.

$V_o = V_i$

At very high freq

$X_C = \frac{1}{2\pi f C}$, $f \to \infty$, $X_C \to 0$



Capacitor Short circuit.

$V_o = 0$

Mathematically, from fig-28, using voltage divider rule,

$$|V_o| = \frac{V_i}{R + X_C} \cdot X_C = \frac{V_i}{\frac{R}{X_C} + 1}$$

At $f \to 0$, $X_C \to \infty$, $|V_o| = \frac{V_i}{\frac{R}{\infty} + 1} = \frac{V_i}{0 + 1} = V_i$

At $f \to \infty$, $X_C \to 0$, $|V_o| = \frac{V_i}{\frac{R}{0} + 1} = \frac{V_i}{\infty + 1} = \frac{V_i}{\infty} = 0$

In general, from fig 28,

$$V_o = \frac{V_i}{R - jX_c} \cdot (-jX_c)$$

using voltage divider rule. and Reactance of capacitor is $\frac{1}{j\omega C} = -jX_c$ where $X_c = \frac{1}{\omega C}$

$$\Rightarrow \frac{V_o}{V_i} = \frac{-jX_c}{R - jX_c}$$

$$= \frac{1}{\left(\frac{R}{-jX_c}\right) + 1}$$ (Dividing n.r & d.r by $-jX_c$)

$$\Rightarrow \frac{V_o}{V_i} = \frac{1}{1 + j\frac{R}{X_c}} = \frac{1}{1 + j\frac{R}{\left(\frac{1}{2\pi f C}\right)}}$$

$\because \frac{1}{-j} = \frac{j}{-j^2} = \frac{j}{-(-1)} = j$

$$\Rightarrow \frac{V_o}{V_i} = \frac{1}{1 + j\,2\pi f RC}$$

$$\Rightarrow \frac{V_o}{V_i} = \frac{1}{1 + j(2\pi RC)f}$$

$$\Rightarrow \frac{V_o}{V_i} = \frac{1}{1 + j\frac{f}{\left(\frac{1}{2\pi RC}\right)}}$$

At, $R = X_c$, $A_v = \frac{1}{\sqrt{2}}$.

$R = \frac{1}{2\pi f_1 C} \Rightarrow f_1 = \frac{1}{2\pi RC}$

$$\Rightarrow \frac{V_o}{V_i} = \frac{1}{1 + j\frac{f}{f_H}}$$

where

$f_H = \frac{1}{2\pi RC}$ = cutoff frequency

$$\Rightarrow \boxed{A_v = \frac{1}{1 + j(f/f_H)}} \quad \text{--- (29)}$$

$$\Rightarrow |A_v| = \frac{1}{\sqrt{1 + (f/f_H)^2}} \quad \text{--- (30)}$$

In dB,

$$(A_v)_{dB} = 20 \log |A_v|$$

$$= 20 \log \frac{1}{\sqrt{1+(f/f_H)^2}}$$

$$= 20 \log \left[1+\left(\frac{f}{f_H}\right)^2\right]^{-\frac{1}{2}}$$

$$= 20 \times -\frac{1}{2} \log \left[1+\left(\frac{f}{f_H}\right)^2\right]$$

$$(A_v)_{dB} = -10 \log \left[1+\left(\frac{f}{f_H}\right)^2\right] \quad \text{——} \quad (31)$$

For large freq $f >> f_H$

So $1+\left(\frac{f}{f_H}\right)^2 \simeq \left(\frac{f}{f_H}\right)^2$

$$\therefore (A_v)_{dB} = -10 \log \left(\frac{f}{f_H}\right)^2$$

$$(A_v)_{dB} = -20 \log \left(\frac{f}{f_H}\right) \quad \text{——} \quad (32)$$

(i) At $f = f_H$, $(A_v)_{dB} = 0$ dB

(ii) At $f = 2f_H$, $(A_v)_{dB} = -20 \log (2) = -6$ dB

(iii) At $f = 4f_H$, $(A_v)_{dB} = -20 \log 4 = -12$ dB

Eqn (32) is a approximate formula, for $f >> f_H$

For Actual response eqn (31) gives, $|A_v| = -10 \log \left[1+\left(\frac{f}{f_H}\right)^2\right]$

At $f = f_H$,

$$|A_v| = -10\log(1+1) = \cancel{-\log} \; -10\log 2$$

$$|A_v| = -3\,dB.$$

i.e there is 3 dB drop from midband level (0 dB) at $f = f_H$.

$\rightarrow f << f_H$, $-10\log[1] = 0$.

$\log[1+(\frac{f}{f_H})^2] \simeq \log[1] = 0$

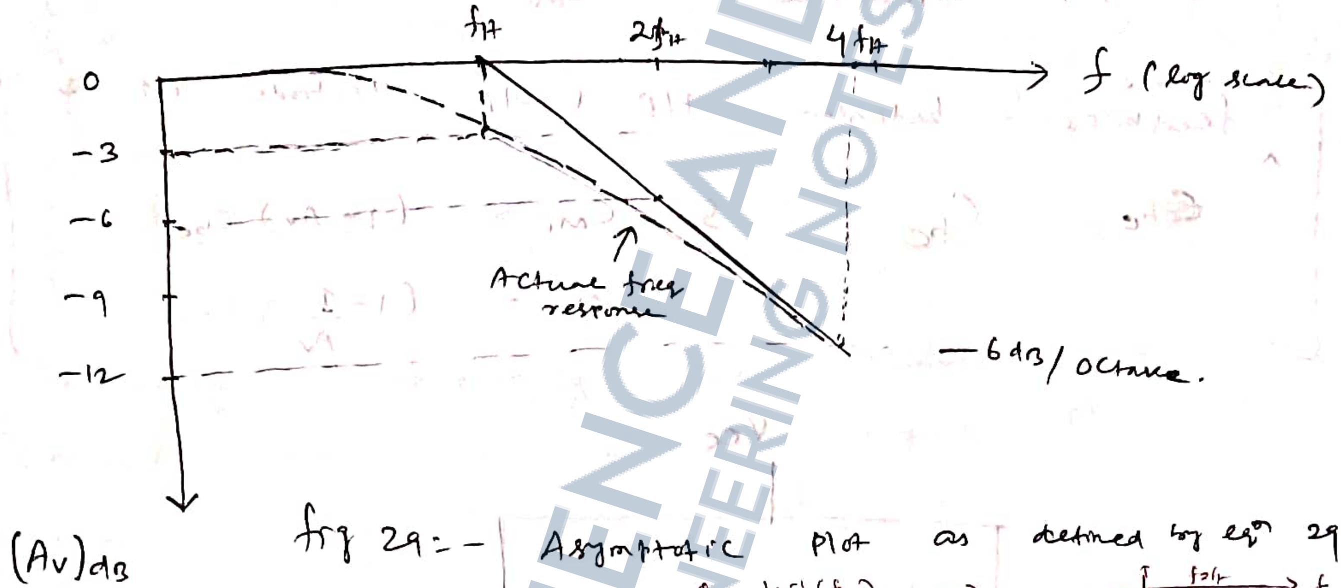
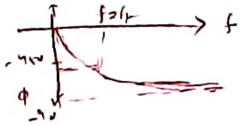


fig 29:- Asymptotic plot as defined by eqn 29

→ Phase $\phi = -\tan^{-1}(\frac{f}{f_H})$ →



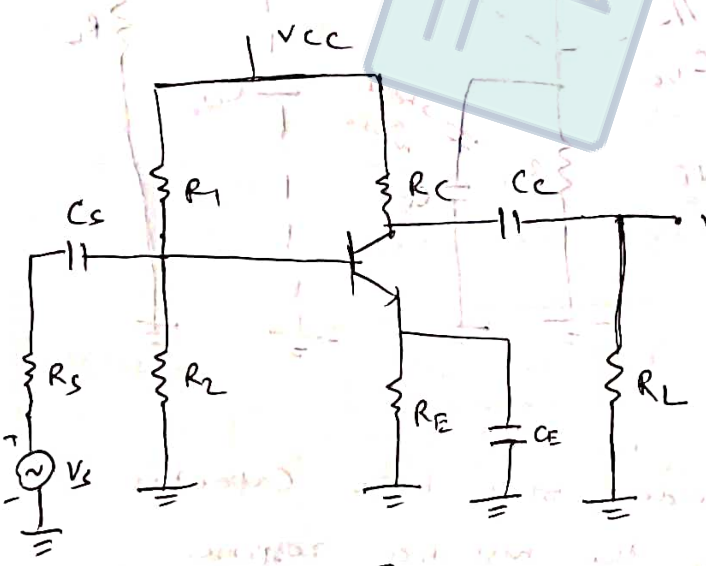
High-freq response of BJT Amplifier

At high freq operation capacitors like C_S, C_C, C_E become short ckt due to their low reactance value.

$$X_C = \frac{1}{2\pi f c}$$

At high freq,

$f \rightarrow \infty$, $X_C \rightarrow 0$.



[Normal Biasing form]

New Capacitance resulted at high freq [fig 30]

(i) Various parasitic capacitance ~ transistor Capacitance (C_{be}, C_{ce})

(ii) Wiring Capacitance [C_{wi} (at i/p), C_{wo} (at o/p)]

(c) Miller Capacitance C_{mi} & C_{mo}.

[Capacitance feedback between i/p & o/p electrode i.e. C_{bc}, So $C_{mi} = (1 - A_v)\, C_{bc}$

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

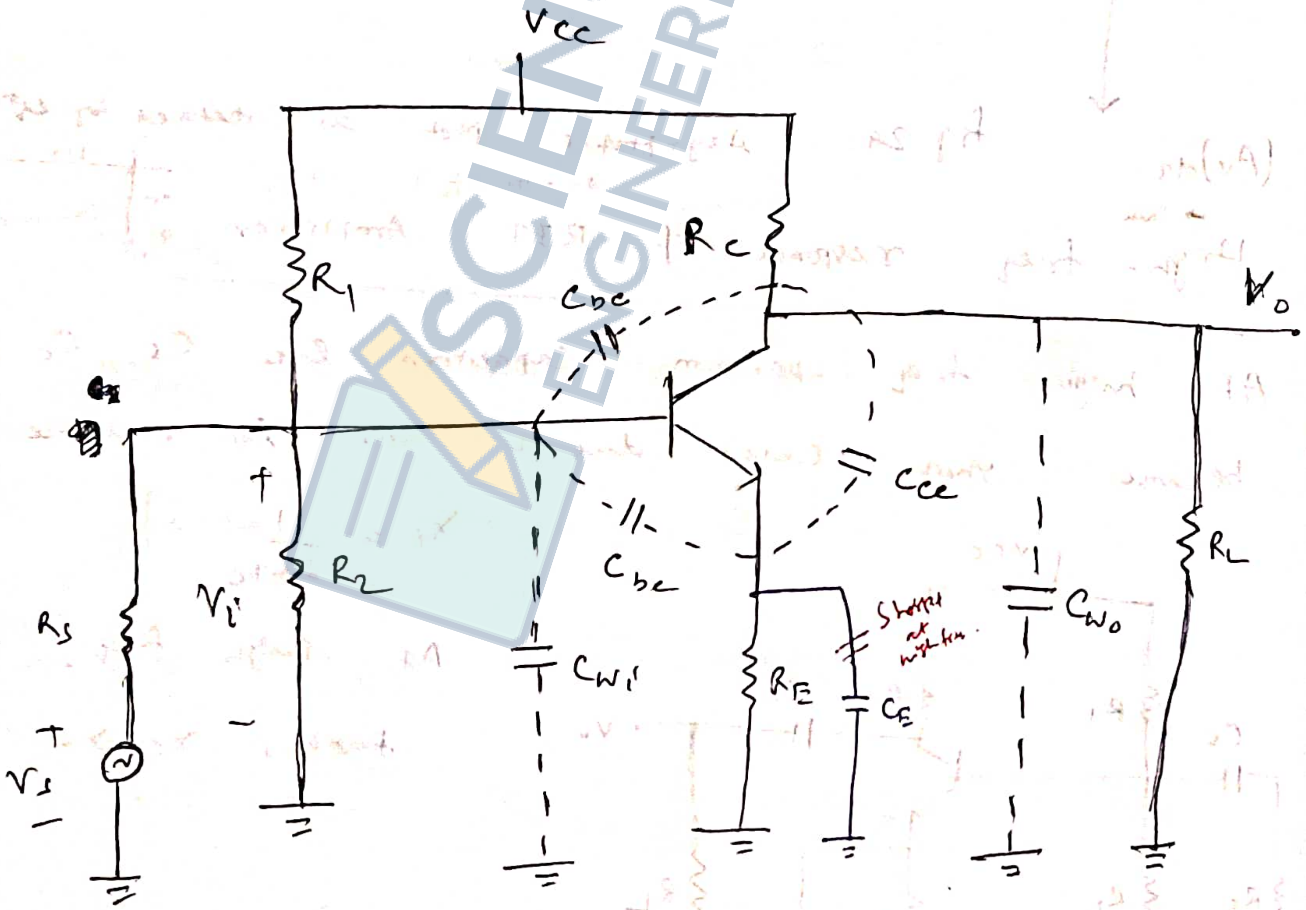


fig 30:- Voltage divider bias with Capacitors that affect the high-freq response.

In general, the capacitance C_{be} is largest parasitic capacitance & C_{ce} the smallest.

The High-freq a.c equivalent model for n/w of fig 30 is shown below.

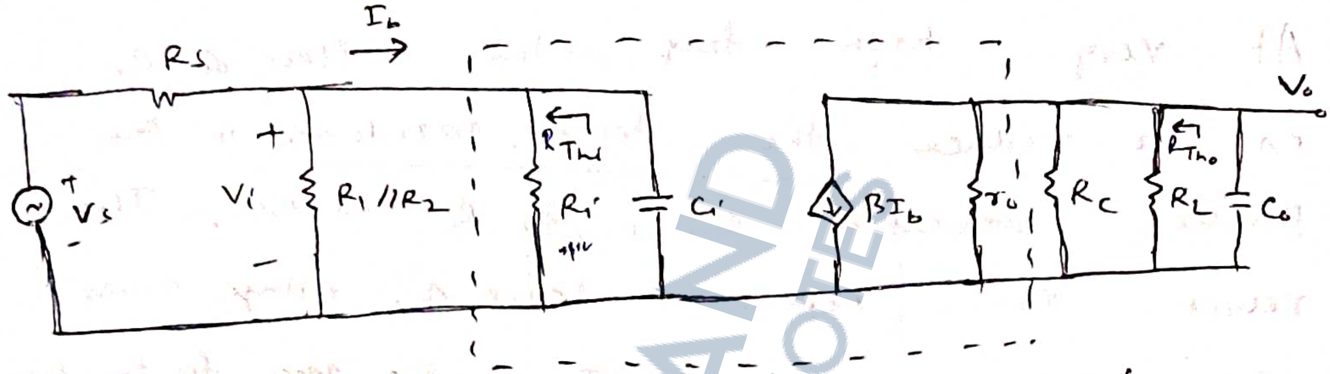


fig 31:- a.c equivalent model of fig 30.

~~Determing~~ Determining the Thevenin equivalent circuit for i/p & o/p networks of fig 31 results the configuration of fig 32.

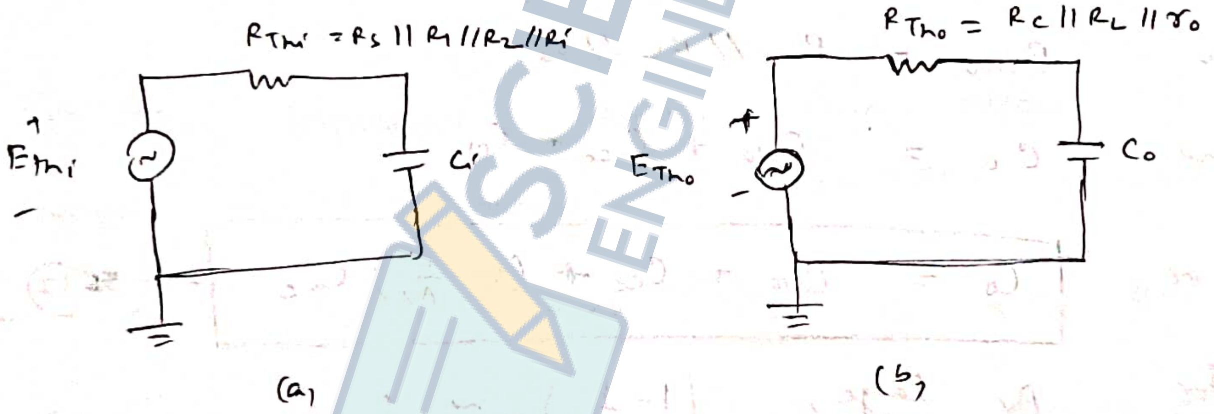


fig 32:- Thevenin ckts for i/p & o/p n/w of n/w fig 31.

Where
$$\boxed{R_{Thi} = R_s || R_1 || R_2 || R_i} \quad — (33)$$

$$\boxed{\begin{aligned} C_i &= C_{wi} + C_{be} + C_{Mi} \\ C_i &= C_{wi} + C_{be} + C_{bc}(1-A_v) \end{aligned}} \quad — (34)$$

For I/P n/w, −3dB freq is defined as

$$f_{Hi} = \frac{1}{2\pi R_{Thi} C_i} \quad \text{(35)}$$

[R_{Thi} & C_i are defined in eqn (33) & (34)]

At very high freq, the effect of C_i is to reduce the total impedance of the parallel combination of R_1, R_2, R_i and C_i. The result is reduced level of voltage across C_i, a reduction in I_b & a gain for the system.

For O/P n/w

$$f_{Ho} = \frac{1}{2\pi R_{Tho} C_o} \quad \text{(36)}$$

where $R_{Tho} = R_C \,//\, R_L \,//\, r_o$

and $C_o = C_{wo} + C_{ce} + C_{Mo}$

or

$$\boxed{C_o = C_{wo} + C_{ce} + \left(1 - \frac{1}{A_v}\right) C_{bc}} \quad \text{(37)}$$

for A_v large, $\left(1 - \frac{1}{A_v}\right) \simeq 1$

$$\therefore \quad C_o \simeq C_{wo} + C_{ce} + C_{bc}$$

At very high freq, the capacitive reactance of C_o will decrease and consequently reduce the total impedance of the O/P parallel branches. The net result is that V_o will also decline towards zero as reactance X_C becomes smaller.

The frequencies f_{Hi} & f_{Ho} each define a $-6dB/octave$ asymptote as depicted in fig 29. [* → h_{fe} or β variation – 445 page]

High freq response – FET Amplifier

→ Similar to BJT analysis.

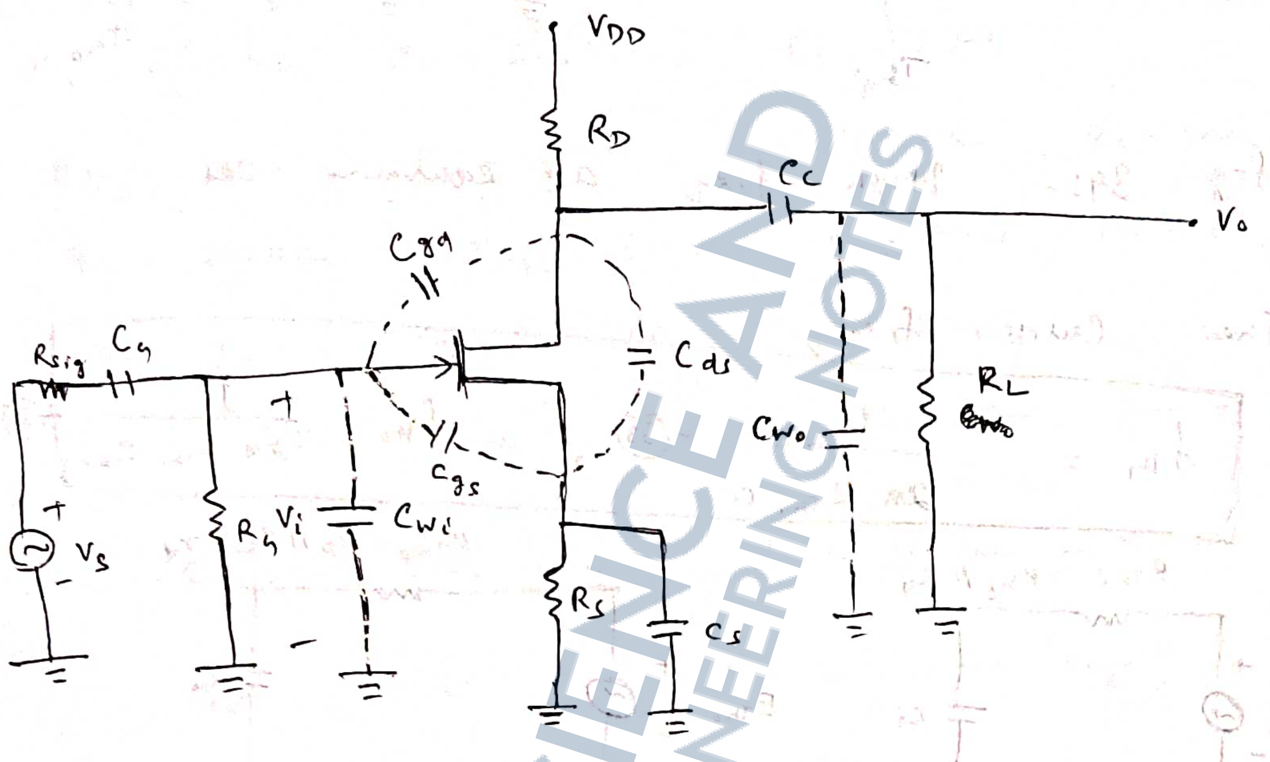


fig 33: Capacitive elements that affect the high freq response of a JFET amplifier.

→ At high freq $\underline{C_G}$, $\underline{C_S}$ & C_C are short circuited. The inter electrode and wiring capacitances determine the high-freq characteristics of the amplifier.

→ Similar to BJT, At high freq C_i will approach a short-ckt equivalent & V_{gs} will drop in value and reduce overall gain.

→ Similarly at high freq, C_o approaches its short-ckt equivalent, the parallel o/p voltage V_o will drop in magnitude.

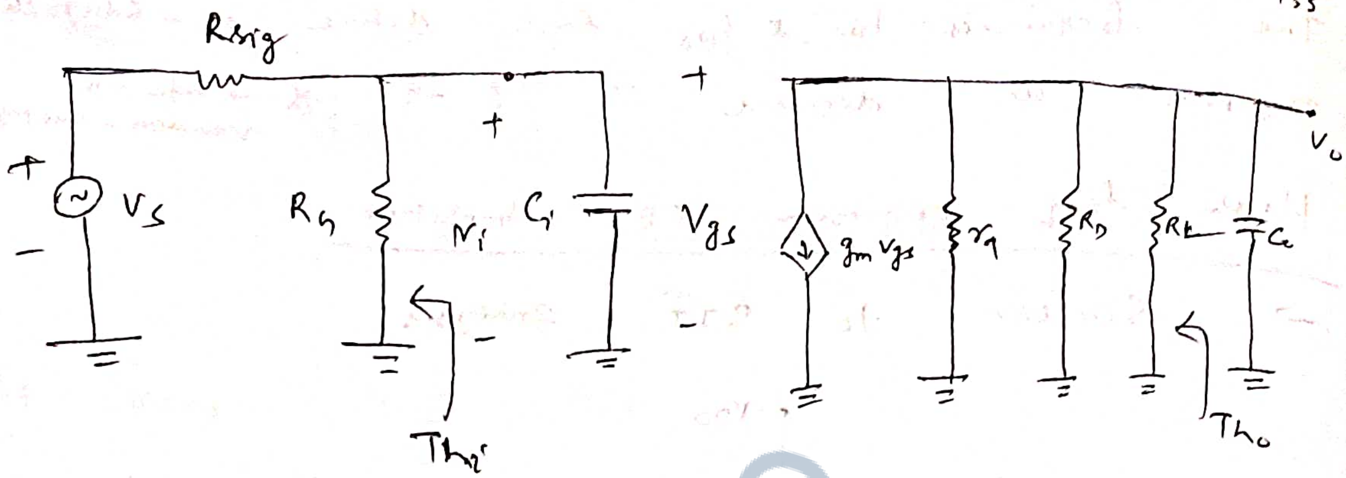


fig 34:- High freq ac equivalent ckt

The cutoff freq,

$$\boxed{f_{Hi} = \frac{1}{2\pi R_{Thi} C_i} \quad (38), \qquad f_{Ho} = \frac{1}{2\pi R_{Tho} C_o} \quad (39)}$$

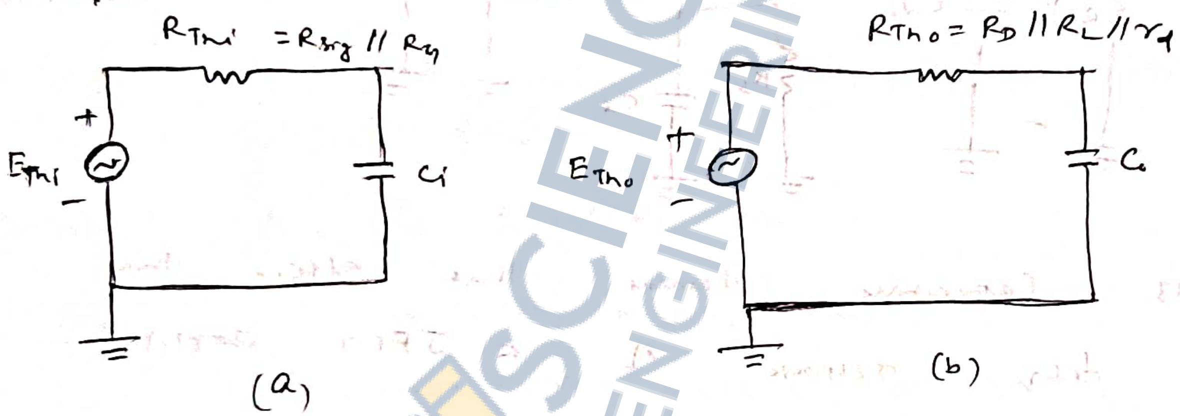


fig 35:- The Thevenin equivalent ckt for (a) the i/p ckt (b) o/p ckt

where $R_{Thi} = R_{sig} \,//\, R_g$

$C_i = C_{wi} + C_{gs} + C_{Mi}$ — (40)

where $C_{Mi} = (1 - A_v)\, C_{gd}$ = I/P Miller Capacitance

Similarly for o/p ckt,

$R_{Tho} = R_D \,//\, R_L \,//\, r_d$ — (41)

$$C_o = C_{Wo} + C_{ds} + C_{Mo} \quad \text{— (42)}$$

where $C_{Mo} = \left(1 - \frac{1}{A_v}\right) C_{gd}$ = O/P Miller Capacitance.

Ex:- 3 : Determine the high-cutoff freq for fig 33, given

$C_G = 0.01\,\mu F$, $C_C = 0.5\,\mu F$, $C_S = 2\,\mu F$

$R_{sig} = 10\,k\Omega$, $R_G = 1\,M\Omega$, $R_D = 4.7\,k\Omega$, $R_S = 1\,k\Omega$

$R_L = 2.2\,k\Omega$, $I_{DSS} = 8\,mA$, $V_P = -4V$, $r_d = \infty$,

$V_{DD} = 20V$, $C_{gd} = 2\,pF$, $C_{gs} = 4\,pF$, $C_{ds} = 0.5\,pF$,

$C_{Wi} = 5\,pF$, $C_{Wo} = 6\,pF$, Use the gain of Ex-2.

Ans: $R_{Thi} = R_{sig} \,//\, R_G = 10\,k\Omega \,//\, 1\,M\Omega = 9.9\,k\Omega$.

$A_v = -3$ [Ex-2, 422 page]

$$C_i = C_{Wi} + C_{gs} + (1 - A_v)\, C_{gd}$$

$$= 5\,pF + 4\,pF + (1-(-3))\, 2\,pF$$

$$= 9\,pF + 8\,pF$$

$$= 17\,pF$$

$$f_{Hi} = \frac{1}{2\pi R_{Thi} C_i} = \frac{1}{2\pi \times 9.9 \times 10^3 \times 17 \times 10^{-12}} = 945.67\,kHz$$

$$R_{Tho} = R_D \,//\, R_L \,//\, \infty = R_D \,//\, R_L = 4.7\,k\Omega \,//\, 2.2\,k\Omega = 1.5\,k\Omega$$

$$C_o = C_{Wo} + C_{ds} + C_{Mo} = 6\,pF + 0.5\,pF + \left(1 - \frac{1}{-3}\right) 2\,pF = 9.17\,pF$$

$$f_{HO} = \frac{1}{2\pi R_{Tho} \cdot C_o}$$

$$= \frac{1}{2\pi \cdot 1.5\times10^{3} \times 9.7\times10^{-12}}$$

$$f_{HO} = 11.57 \text{ MHz}$$

Multi-stage freq effects :-

For a second transistor stage connected directly to the O/P of a first stage, there will be a significant change in the overall freq response. In the high-freq region, the O/P capacitance C_o must now include the wiring capacitance (C_{W1}), parasitic capacitance (C_{be}) and Miller capacitance (C_{Mi}) of the following stage (i.e. the C_i of 1st stage). Further, there will be additional low-freq cutoff levels due to the second stage that will further reduce the overall gain of the system in this region. The effect of increasing number of identical stages can be clearly demonstrated by considering the situation indicated in fig 36.

Fig:-36

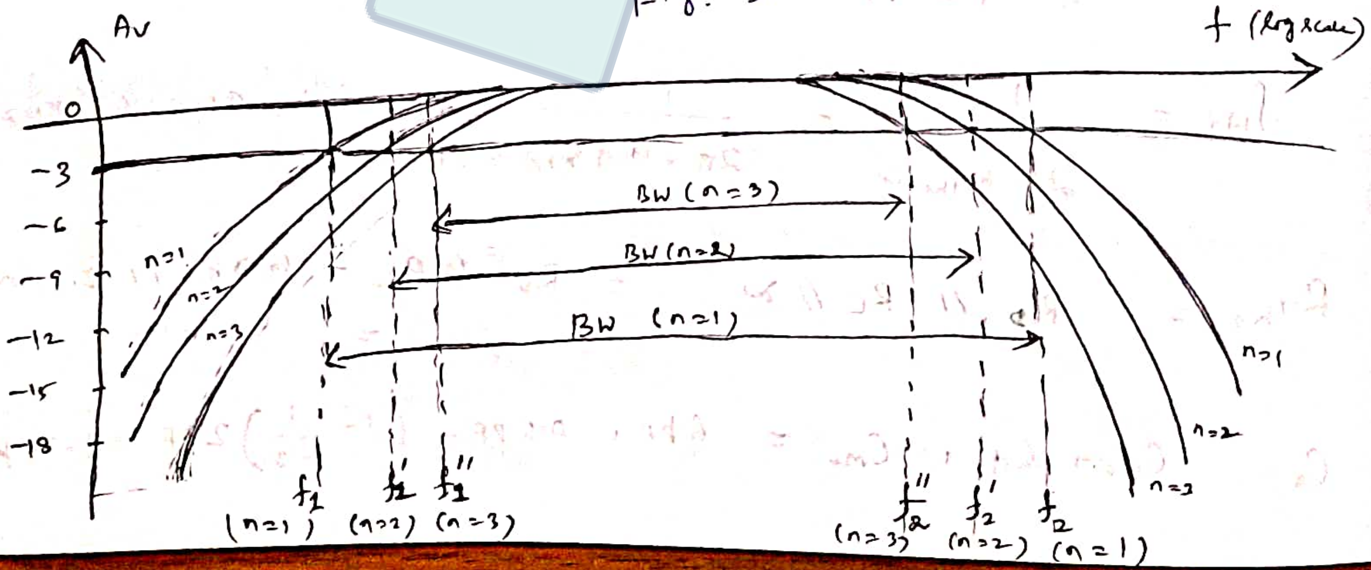


Fig 36 :- Effect of an increased number of stages on the cutoff frequencies and the BW.

In each case, the upper and lower cutoff freq of each ~~stage~~ of the cascaded stages are identical. For a single stage the cutoff freq are f_1 & f_2 as indicated. For 2 identical stages in cascade, the drop-off rate in the high freq & low freq regions has increased to -12 dB / octave or -40 dB / decade. and -3 dB point shifted to f_1' & f_2'.

A -18 dB / octave or -60 dB / decade slope will result for a 3 stage system with reduction on bandwidth (f_1'' & f_2'')

Assuming identical stages, an equation for each band freq as a function of the number of stages (n) can be determined in following manner. For low freq region

$$A_{V_{low,(overall)}} = A_{V_{1low}} \cdot A_{V_{2low}} \cdots A_{V_{n low}} \quad \text{--- (43)}$$

Since each stage is identical,

$$A_{V_{1low}} = A_{V_{2low}} = A_{V_{low}}$$

$$\therefore \quad A_{V_{low,(overall)}} = (A_{V_{low}})^n \quad \text{--- (44)}$$

But
$$A_{V_{low}} = \frac{1}{1 - j\left(\frac{f_1}{f}\right)}$$

$$|A_{V_{low}}| = \frac{1}{\sqrt{1 + \left(\frac{f_1}{f}\right)^2}} \quad \text{--- (45)}$$

$$|A_{v_{low},(overall)}| = |A_{v_{low}}|^n \quad \text{—— } 46$$

Putting eq^n 45 in eq^n 46, we have

$$|A_{v_{low}, overall}| = \left[\frac{1}{\sqrt{1+\left(\frac{f_1}{f}\right)^2}}\right]^n \quad \text{—— } 47$$

$$= \frac{1}{\left[1+\left(\frac{f_1}{f_2}\right)^2\right]^{\frac{n}{2}}}$$

At -3dB cutoff freq, $[f_T]$ $\cancel{f} = f_1$, $|A_{v_{low}}, overall| = \frac{1}{\sqrt{2}}$

$f = f_T$, $T \to$ Total

$$\therefore \quad \frac{1}{\left[1+\left(\frac{f_1}{f_T}\right)^2\right]^{\frac{n}{2}}} = \frac{1}{\sqrt{2}}$$

$$\Rightarrow \left[1+\left(\frac{f_1}{f_T}\right)^2\right]^{\frac{n}{2}} = \sqrt{2}$$

$$\Rightarrow \left[1+\left(\frac{f_1}{f_T}\right)^2\right]^{n} = 2$$

$$\Rightarrow 1+\left(\frac{f_1}{f_T}\right)^2 = 2^{\frac{1}{n}}$$

$$\Rightarrow \left(\frac{f_1}{f_T}\right)^2 = 2^{1/n} - 1$$

$$\Rightarrow \frac{f_1}{f_T} = \sqrt{2^{1/n}-1}$$

$$\Rightarrow \boxed{f_{T_{(low)}} = \frac{f_1}{\sqrt{2^{1/n}-1}}} \quad \text{—— } 48$$

~~Similar~~

n	$\sqrt{2^{1/n}-1}$
2 $\rightarrow$	0.64
3 $\rightarrow$	0.51
4 $\rightarrow$	0.43
5 $\rightarrow$	0.39

$\therefore$ ~~f_1 (low) $\rightarrow$ f_1~~

For $n=2$, $f_1' = \dfrac{f_1}{0.64} = 1.5625\, f_1$

$n=3$, $f_1'' = \dfrac{f_1}{0.51} = 1.96\, f_1$

$\therefore f_1'' > f_1' > f_1$ [fig 36]

Similarly :- For high freq region

$$|A_{v\,high,\,overall}| = \left[\frac{1}{\sqrt{1+\left(\frac{f}{f_2}\right)^2}}\right]^n = \frac{1}{\left[1+\left(\frac{f}{f_2}\right)^2\right]^{n/2}}$$

[Similar to eq^n 47]

See, for low freq $\frac{f_1}{f}$ but high freq $\frac{f}{f_2}$

At -3 dB cutoff freq, $A_{v\,high,\,overall} = \frac{1}{\sqrt{2}}$

$f = [f_T]$

$$\Rightarrow \frac{1}{\left[1+\left(\frac{f_T}{f_2}\right)^2\right]^{n/2}} = \frac{1}{\sqrt{2}}$$

$$\Rightarrow \left[1+\left(\frac{f_T}{f_2}\right)^2\right]^{\frac{n}{2}} = \sqrt{2}$$

$$\Rightarrow \left[1+\left(\frac{f_T}{f_2}\right)^2\right]^{n} = 2$$

$$\Rightarrow \quad 1 + \left(\frac{f_T}{f_2}\right)^2 = 2^{1/n}$$

$$\Rightarrow \quad \left(\frac{f_T}{f_2}\right)^2 = 2^{1/n} - 1$$

$$\Rightarrow \quad \left(\frac{f_T}{f_2}\right) = \sqrt{2^{1/n} - 1}$$

$$\Rightarrow \quad \boxed{f_{T\,(\text{Higher})} = f_2 \sqrt{2^{1/n} - 1}} \quad - (49)$$

$f_2'' < f_2' < f_2$ [fig 36]

Ex-4 :- BPUT - 2011

An amplifier consist of 3 identical stages in cascade. The BW of overall amplifier extends from 20 Hz to 20 kHz. Calculate BW of individual stages.

Ans : Given

$f_L = f_1 = 20\,Hz$, $f_H = f_2 = 20\,kHz$.

$$f_{L2} \text{ or } f_1' = \frac{f_1}{\sqrt{2^{1/n} - 1}} = \frac{20}{\sqrt{2^{1/2} - 1}} = 31.07\,Hz$$
$(n = 2)$

For $n = 3$,

$$f_{L3} \text{ or } f_1'' = \frac{f_1}{\sqrt{2^{1/3} - 1}} = \frac{20}{\sqrt{2^{1/3} - 1}} = 39.22\,Hz$$

$$\cancel{f_{H2} \text{ or } f_2' = \frac{f_2}{\sqrt{2^{1/2} - 1}} = \frac{20\,kHz}{\sqrt{2^{1/2} - 1}}}$$
$\cancel{(n = 2)}$

$$f_{H2} \approx f_2' = f_2 \sqrt{2^{1/n} - 1}$$

$[n=2]$

$$= 20\,\text{KHz} \sqrt{2^{1/2} - 1}$$

$$= 12.87\ \text{KHz}$$

$$f_{H3} \approx f_2'' = f_2 \sqrt{2^{1/3} - 1}$$

$n = 3$

$$= 20\,\text{KHz} \sqrt{2^{1/3} - 1}$$

$$f_2'' = 10.19\ \text{KHz}$$

Bandwidth of Individual Stages [fig 36]

$BW_1 = f_2 - f_1 = 20\,\text{KHz} - 20\,\text{Hz} = 19.98\ \text{KHz}$

$BW_2 = f_2' - f_1' = 12.87\,\text{KHz} - 31.07\,\text{Hz} = 12.83\,\text{KHz}$

$BW_3 = f_2'' - f_1'' = 10.19\,\text{KHz} - 39.22\,\text{Hz} = 10.15\ \text{KHz}$

$\therefore$ Bandwidth gradually reduces, with increase in number of stages.

Imp Square-wave Testing :- [BPUT - Short Note - 2010]

A sense for the freq response of an amplifier can be determined experimentally by applying a square-wave signal to the amplifier and noting the o/p response. The shape of the o/p waveform will reveal whether the high or low frequencies are being properly amplified.

Imp Why square wave is taken? (BPUT)

The use of square-wave testing

is significantly less time-consuming than applying a series of sinusoidal signals at different frequencies and magnitude to test the freq response of the amplifier.

The reason for choosing a square-wave signal for the testing process is best described by examining the Fourier Series expansion of a square wave composed of a series of sinusoidal components of different magnitudes & frequencies. The summation of the terms of the series will result in the original waveform.

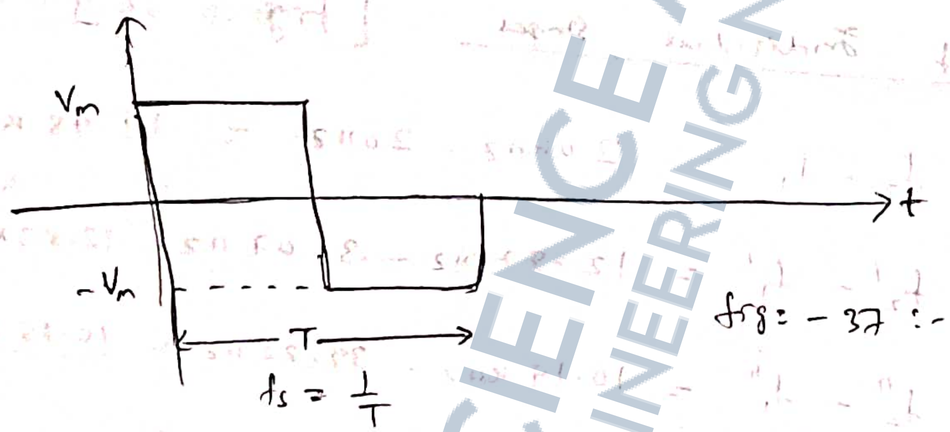


fig:- 37 :- Square wave

The fourier series expansion for the square wave given above is

$$V = \frac{4V_m}{\pi}\left(\sin 2\pi f_s t + \frac{1}{3}\sin 2\pi(3f_s)t + \frac{1}{5}\sin 2\pi(5f_s)t + \frac{1}{7}\sin 2\pi(7f_s)t + \frac{1}{9}\sin 2\pi(9f_s)t + \cdots + \frac{1}{n}\sin 2\pi(nf_s)t\right) \quad (50)$$

The first term of the series is called the fundamental term and in this case has the same freq, f_s, as the square wave.

The next term has a freq equal to three times the fundamental and is referred to as the third harmonic, Its magnitude is one third the magnitude of the fundamental term. The frequencies of the succeeding terms are odd multiples of the fundamental term and the magnitude decreases with each higher harmonic.

The generation of the square wave of fig 37 require infinite number of terms, But fig 38 shows, including fundamental, 3rd, 5th & 7th harmonics takes us a step closer to the wave form of fig 37.

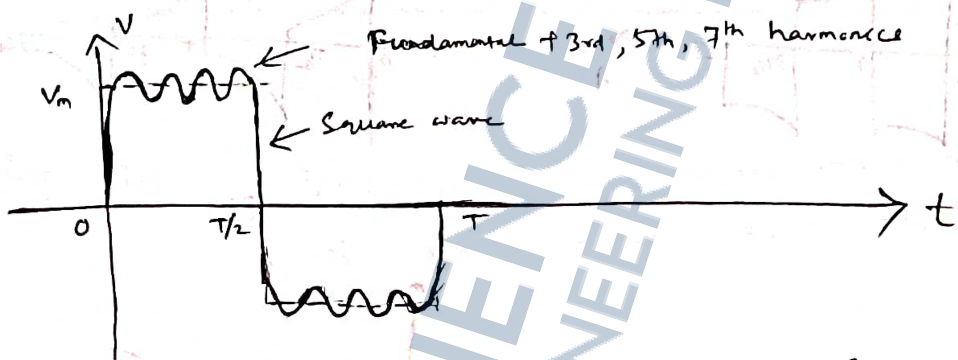


fig: 37 Harmonic content of a square wave.

Since the 9th harmonic has a magnitude greater than 10% of the fundamental term $\left[\frac{1}{9} \times 100 = 11.1\%\right]$, the fundamental term through the 9th harmonic are the major contributor to the Fourier series expansion of the square-wave function. It is therefore reasonable to assume that if the application of a square-wave of a particular freq results in a nice clean square square at the o/p, then the fundamental through the ninth harmonic are being amplified without visual distortion by the amplifier.

If the response of an amplifier to an applied square wave is an undistorted replica of the i/p, the freq response (or BW) of the amplifier is obviously sufficient for the applied frequency. If the response is as shown in fig 38 (a) and (b), the low frequencies are not amplified properly and the low cutoff freq has to be investigated

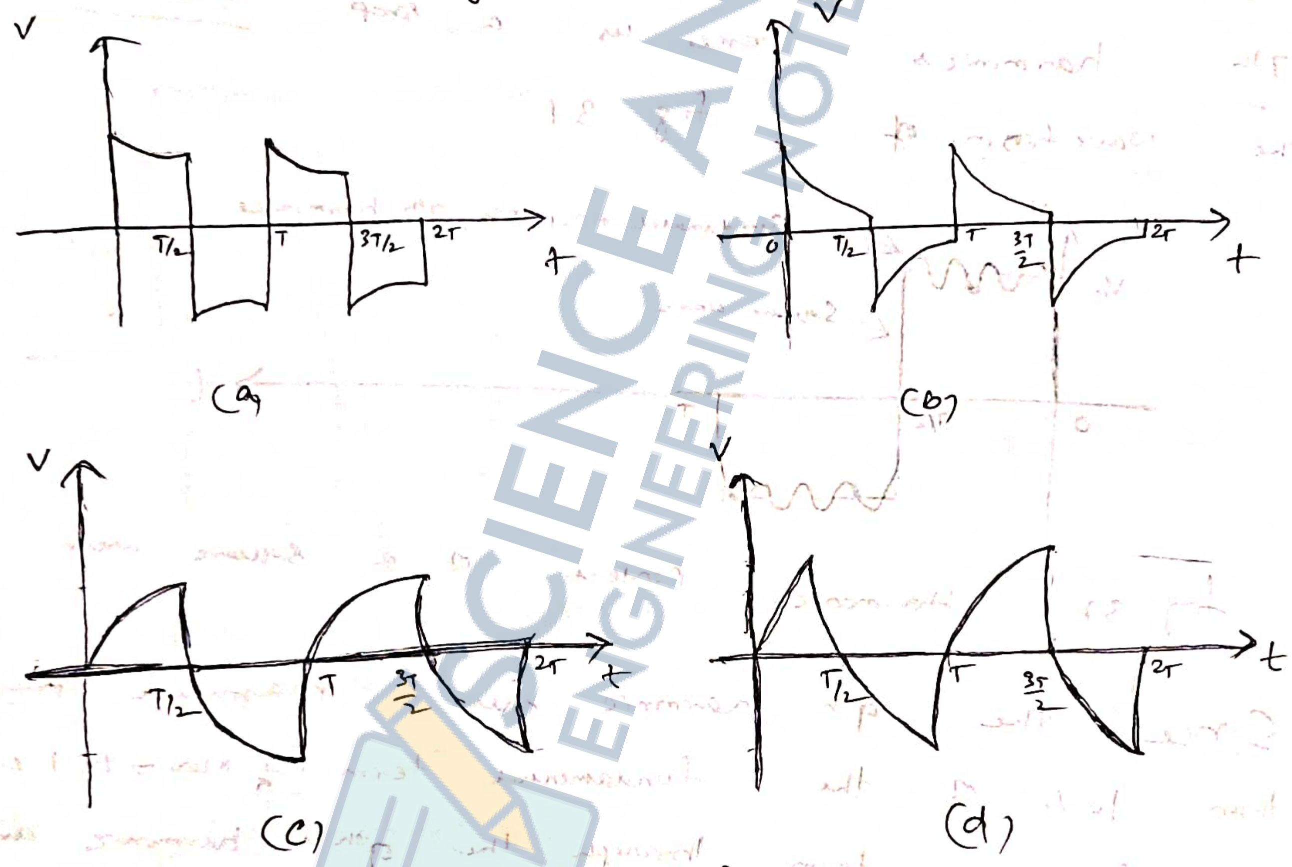


fig 38 : (a) Poor low freq response
(b) very poor low freq response
(c) Poor high-freq response
(d) Very poor high-freq response.

→ Similarly, if the waveform has appearance of fig 38 (c) & (d), the high freq components are not receiving sufficient amplification & high

Cutoff freq (or BW) has to be reviewed.

The actual high cutoff freq (or BW) can be ~~det~~ determined from the output waveform by carefully measuring the rise time defined between between 10% and 90% of the peak value, as shown in fig 39.

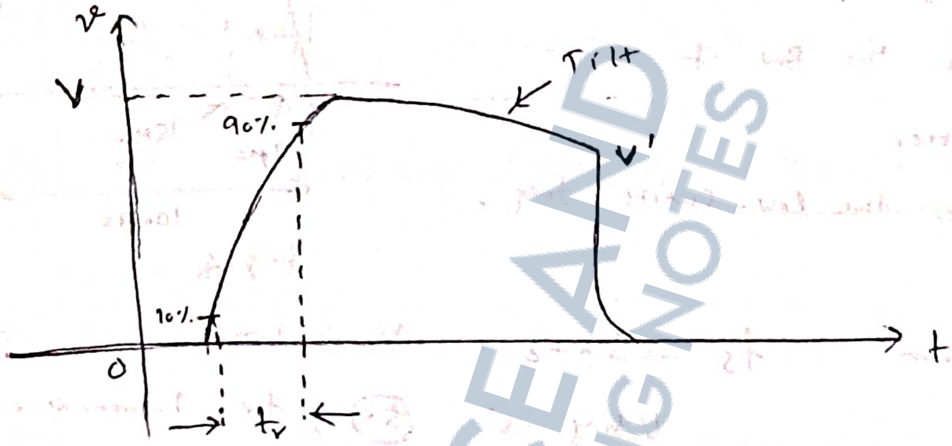


fig 39:- Defining the rise time and tilt of a square wave response.

$$BW = f_{Hi} - f_{Lo} \simeq f_{Hi}$$

$$\therefore \boxed{BW \simeq f_{Hi} = \frac{0.35}{t_r}} \qquad (51)$$

t_r = rise time

The low cutoff freq can be determined from the O/P response by carefully measuring the tilt of fig 39.

$$\% \text{ tilt} = P\% = \frac{V - V'}{V} \times 100\% \qquad (52)$$

$$\text{tilt} = P = \frac{V - V'}{V} \quad (\text{decimal form}) \qquad (53)$$

The low cutoff freq is then determined from

$$\boxed{f_{Lo} = \frac{P}{\pi} f_s} \qquad (54)$$

$f_s \rightarrow$ Fundamental freq

Ex: 5 The application of a 1 mV, 5 KHz square wave to an amplifier result on the O/P wave form [fig 40].

(a) Write the Fourier series expansion of square wave through 9th harmonic

(b) Determine the BW of the amplifier

(c) Calculate the low-cutoff freq.

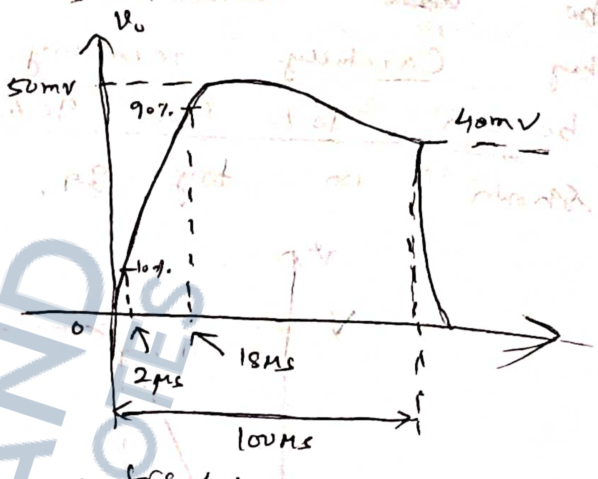


fig 40:-

Ans : (a) Given $f_s = 5\,KHz$, $V_m = 1\,mV$

Using eqn (50), for Fourier series expansion

$$V_i = \frac{4.(1\,mV)}{\pi}\Big[\sin 2\pi(5\times10^3)t + \frac{1}{3}\sin 2\pi(15\times10^3)t + \frac{1}{5}\sin 2\pi(25\times10^3)t + \frac{1}{7}\sin 2\pi(35\times10^3)t + \frac{1}{9}\sin 2\pi(45\times10^3)t\Big]$$

(b) $t_r = (18-2)\mu s = 16\,\mu s$

$$BW = \frac{0.35}{t_r} = \frac{0.35}{16\times10^{-6}} = 21,875\,Hz \simeq 4.4\,f_s$$

(c) $$P = \frac{V-V'}{V} = \frac{50\,mV - 40\,mV}{50\,mV} = \frac{10}{50} = 0.2$$

$$f_{Lo} = \frac{P}{\pi}.f_s = \frac{0.2}{\pi}\times5\times10^3 = 318.31\,Hz.$$

Ex - 6 - Internal question

Three voltage amplifiers with gain & BW ($A_1 = 100, BW_1 = 30\,KHz$), ($A_2 = 20, BW_2 = 30\,KHz$), ($A_3 = 40, BW_3 = 30\,KHz$) are connected in cascade. Calculate the overall gain in dB & overall BW.

Ans :

$$(A_v)_{overall} = 100 \times 20 \times 40 = 8 \times 10^4$$

In dB, $20 \log (8 \times 10^4) = 98.06\ dB$ (Ans)

$$\text{overall BW} = 30\,KHz \times \sqrt{2^{1/n} - 1}$$

$$= 30 \times 10^3 \sqrt{2^{1/3} - 1}$$

$$(BW)_{overall} = 15.29\ KHz \quad \text{(Ans)}$$

BPUT

High freq response - BJT Amplifier [Remaining portion]

MOT-2010

hfe or β variation

The variation of hfe or (β) with freq, is given by

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

using hybrid - π model

f_β is defined as

$$f_\beta \simeq \frac{1}{2\pi\, \beta_{mid}\, r_e\, (C_{be} + C_{bc})} \quad \text{— (2)}$$

→ Common base configuration displays improved high-freq characteristics over common-emitter configuration.

In Common base, there is absence of Miller effect Capacitance due to non-inverting ~~configuration~~ characteristics. For this reason, Common base high-freq parameters rather than CE parameters are specified for transistors - especially those designed specifically to operate in the high-freq regions.

Conversion

$$f_\beta = f_\alpha (1-\alpha) \quad \text{--- (3)}$$

Imp → A quantity called gain-BW product is defined for the transistor by the Cond^n,

$$\left| \frac{h_{fe\,mid}}{1+j\, f/f_\beta} \right| = 1 \quad \text{--- (4)}$$

From eq^n (1),

$$\therefore \quad |h_{fe}|_{dB} = 20\log \left| \frac{h_{fe\,mid}}{1+j\, f/f_\beta} \right| = 20 \log 1 = 0 \text{ dB} \quad \text{--- (5)}$$

The freq at which $|h_{fe}|_{dB} = 0$ dB is given by f_T, from eq^n (1)

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

If $f_T \gg f_\beta$, $\sqrt{1+\left(\frac{f_T}{f_\beta}\right)^2} \simeq \left(\frac{f_T}{f_\beta}\right)^2$

$$\frac{h_{fe\,mid}}{\frac{f_T}{f_\beta}} = 1$$

$$\Rightarrow \quad \frac{h_{fe\,mid} \times f_\beta}{f_T} = 1$$

$$\Rightarrow \quad \boxed{f_T = f_\beta \cdot h_{fe\,mid}}$$

$$\Rightarrow \quad \boxed{f_T = f_\beta \cdot \beta_{mid}} \quad -(5) \quad \Big| \because h_{fe} = \beta$$

Putting eqⁿ (2), in eqⁿ (5), we have

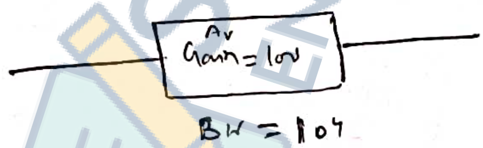
$$f_T = \frac{1}{2\pi \cdot \cancel{\beta_{mid}} \cdot r_e (C_{be} + C_{bc})} \cdot \cancel{\beta_{mid}}$$

$$\Rightarrow \quad \boxed{f_T = \frac{1}{2\pi\, r_e (C_{be} + C_{bc})}}$$

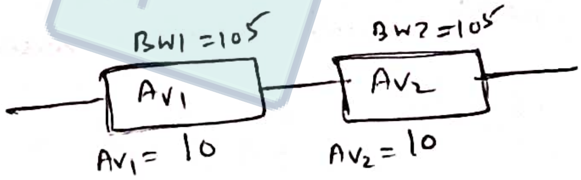
Note: In multistage, there is decrease in BW with increase in number of stages. fig - 3c, $BW_3 < BW_2 < BW_1$

→ But if midband gain can remain fixed, then BW will not decrease.

ex:



Gain × BW
$= 100 \times 10^4 = 10^6$



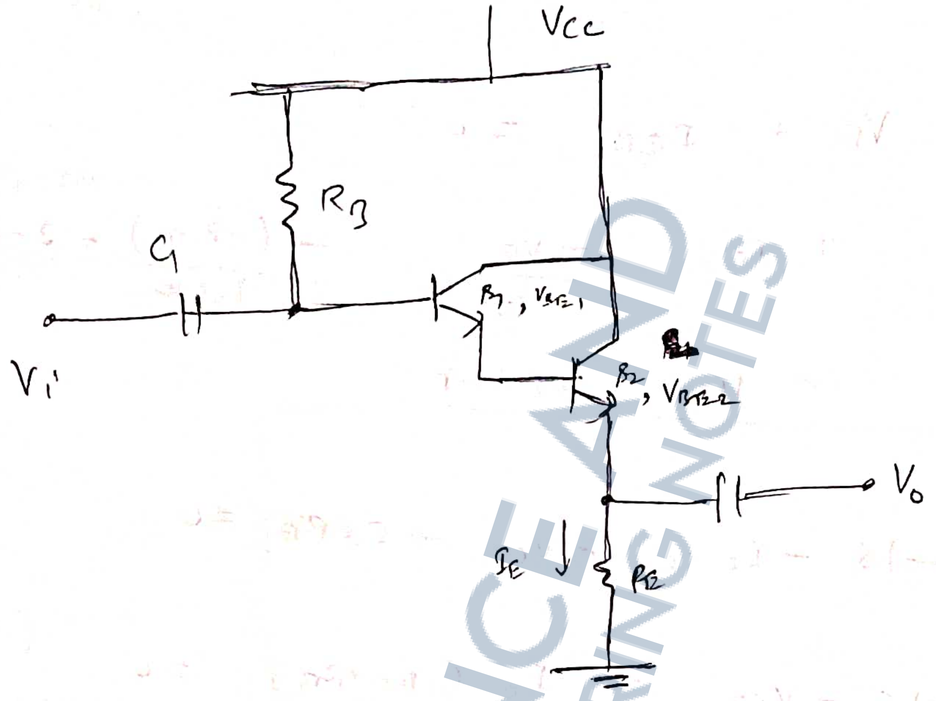
Gain = 10×10 = 100
(Same)

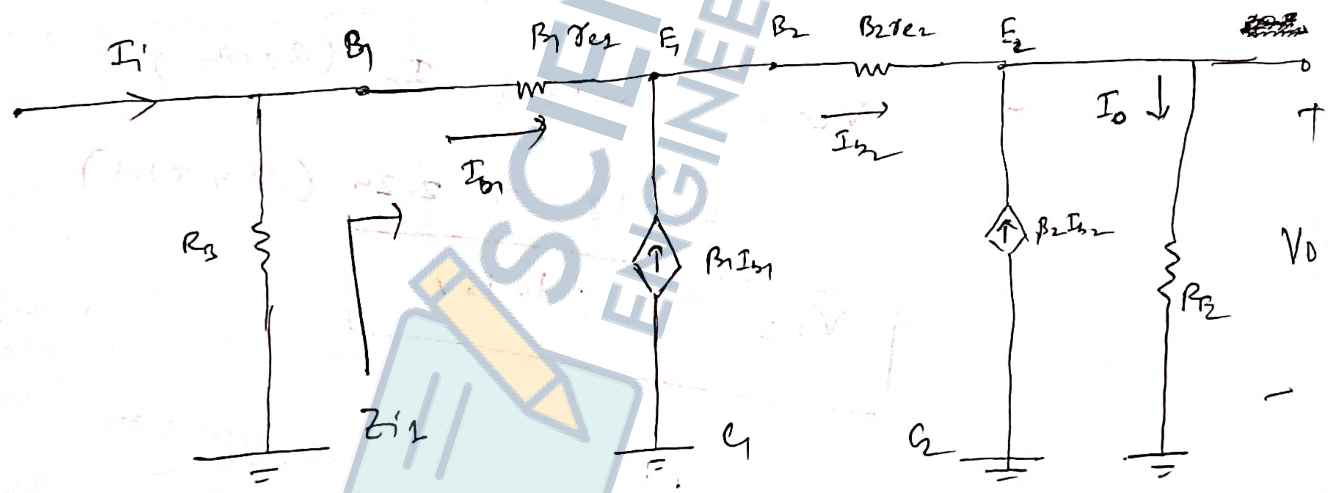
But $BW_1 = 10^5$, $10 \times 10^5 = 10^6$
$BW_2 = 10^5$, $10 \times 10^5 = 10^6$

∴ BW of each stage increased by a factor 10 to make Gain × BW = constant.

Darlington Connection :-

Current gain





Solving for O/P current

$$I_o = I_{b2} + \beta_2 I_{b2} = (\beta_2 + 1) I_{b2} \quad \text{—(1)}$$

With $I_{b2} = I_{b1} + \beta_1 I_{b1} = (\beta_1 + 1) I_{b1}$ —(2)

$$\therefore I_o = (\beta_2 + 1) I_{b2} = (\beta_2 + 1)(\beta_1 + 1) I_{b1} \quad \text{—(3)}$$

Using current division rule on the i/p circuit,

$$I_{b1} = \frac{I_i \times R_B}{R_B + Z_{i1}} = \frac{I_i R_B}{R_B + (\beta_1 \beta_2 R_E)}$$

$\therefore$ From eq^n (3),

$$I_o = (\beta_2+1)(\beta_1+1) \times \frac{I_i \times R_B}{R_B + (\beta_1 \beta_2) R_E}$$

$$\Rightarrow \frac{I_o}{I_i} = \frac{(\beta_1+1)(\beta_2+1) R_B}{R_B + \beta_1 \beta_2 R_E}$$

$$\Rightarrow A_i \simeq \frac{\beta_1 \beta_2 R_B}{R_B + \beta_1 \beta_2 R_E}$$

$$\Rightarrow \boxed{A_i \simeq \frac{\beta_D R_B}{R_B + \beta_D R_E}}$$

<u>Voltage gain</u>

$$V_o = I_o R_E \quad \text{--- (1)}$$

$$V_i = I_i \times (R_B \,//\, Z_b)$$

$$V_i = I_i \times (R_B \,//\, \beta_D R_E) \quad \text{--- (2)}$$

Dividing eqⁿ ① by eqⁿ ②,

$$\frac{V_o}{V_i} = \frac{I_o R_E}{I_i \times (R_B // \beta_D R_E)}$$

$$\Rightarrow A_v = A_i \times \frac{R_E}{R_B // \beta_D R_E}$$

$$\Rightarrow A_v = A_i \times \frac{\cancel{R_E}(R_B + \beta_D R_E)}{R_B \times \beta_D \times \cancel{R_E}}$$

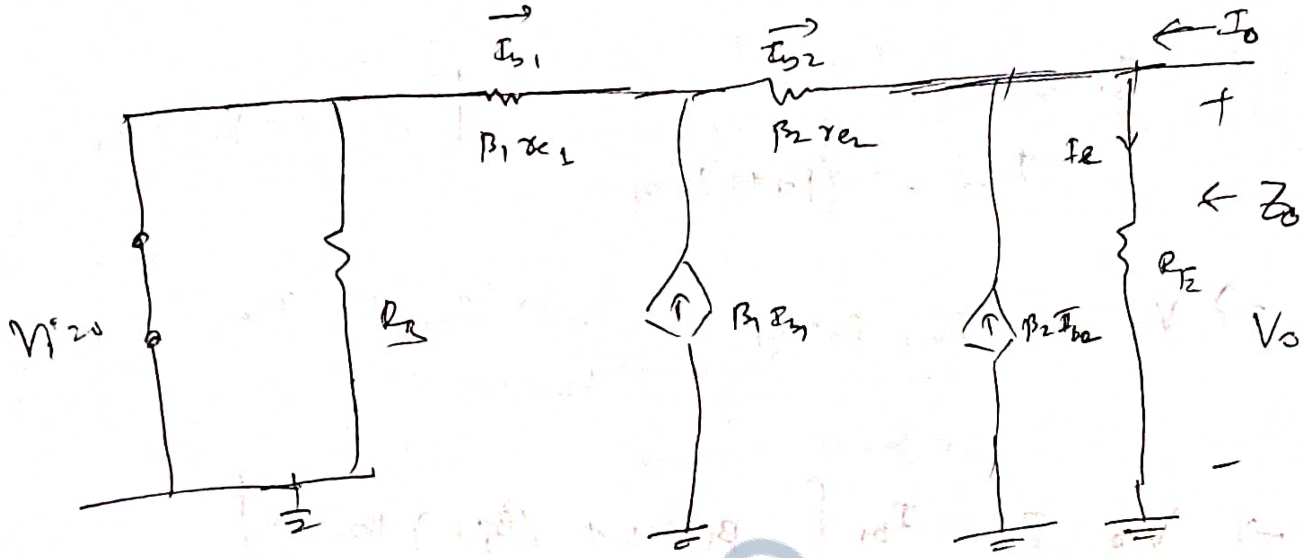
$$\Rightarrow A_v = \left(\frac{\cancel{\beta_D R_B}}{\cancel{R_B + \beta_D R_E}}\right) \times \frac{\cancel{(R_B + \beta_D R_E)}}{\cancel{R_B \times \beta_D}}$$

$$\Rightarrow \boxed{A_v = 1}$$ In Reality < 1.

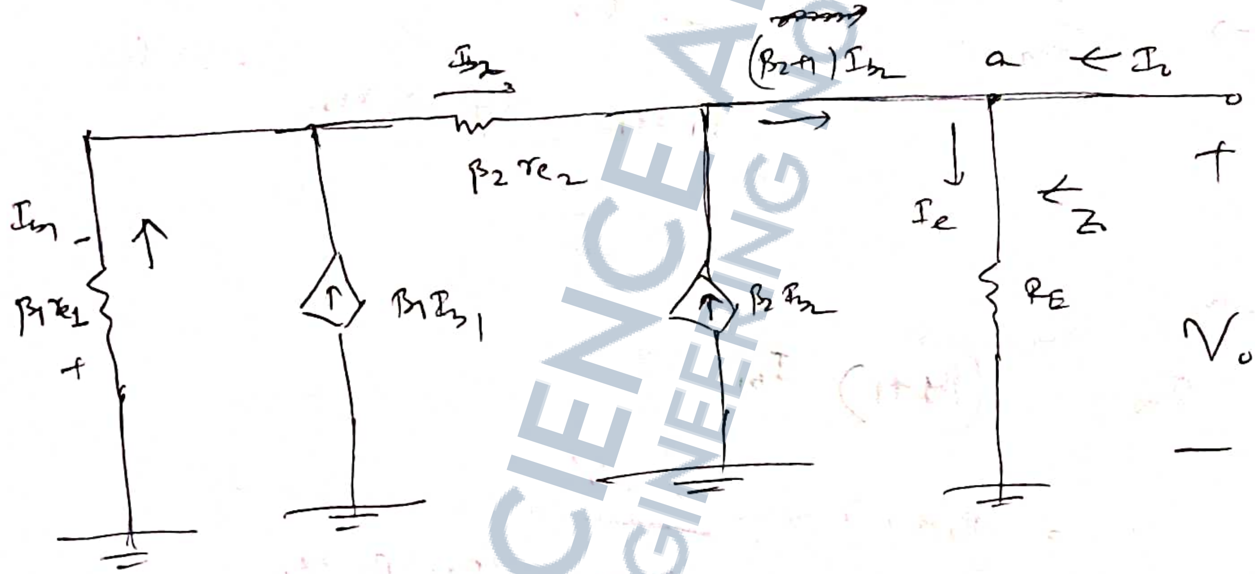
which is an expected result for the emitter-follower configuration.

O/P impedance:-

The O/P impedance will be determined by setting V_i to zero. and the resistor R_B is short circuited. The O/P current has been redefined to match the standard nomenclature and properly defined Z_o.



$\simeq$



At point 'a' KCL will result,

$$I_0 + (\beta_2+1) I_{b2} = I_e$$

$$\Rightarrow I_0 = I_e - (\beta_2+1) I_{b2} \quad \text{——— (1)}$$

Applying KVL around the entire outside loop,

$$- I_{b1} \times \beta_1 \cdot r_{e1} - I_{b2} \times \beta_2 r_{e2} - V_0 = 0$$

$$\Rightarrow V_0 = - I_{b1} \beta_1 r_{e1} - I_{b2} \beta_2 r_{e2} \quad \text{——— (2)}$$

Substituting

$$I_{b2} = (\beta_1+1) I_{b1}$$

$\because I_{b2} = \beta_1 I_{b1} + I_{b1}$

$$\Rightarrow V_0 = -I_{b1}\beta_1 r_{e1} - (\beta_1+1) I_{b1} \beta_2 r_{e2}$$

$$\Rightarrow V_0 = -I_{b1}\left[\beta_1 r_{e1} + (\beta_1+1)\beta_2 r_{e2}\right]$$

$$\Rightarrow I_{b1} = \frac{-V_0}{\beta_1 r_{e1} + (\beta_1+1)\beta_2 r_{e2}} \qquad (3)$$

WKT

$$I_{b2} = (\beta_1+1) I_{b1}$$

$$= (\beta_1+1) \times \frac{-V_0}{\beta_1 r_{e1} + (\beta_1+1)\beta_2 r_{e2}}$$

$$\Rightarrow I_{b2} = -\left[\frac{(\beta_1+1)}{\beta_1 r_{e1} + (\beta_1+1)\beta_2 r_{e2}}\right] V_0 \qquad (4)$$

From eqn (1),

$$I_0 = I_e - (\beta_2+1) I_{b2}$$

$$= I_e - (\beta_2+1) \times \left[-\frac{(\beta_1+1)}{\beta_1 r_{e1} + (\beta_1+1)\beta_2 r_{e2}}\right] V_0$$

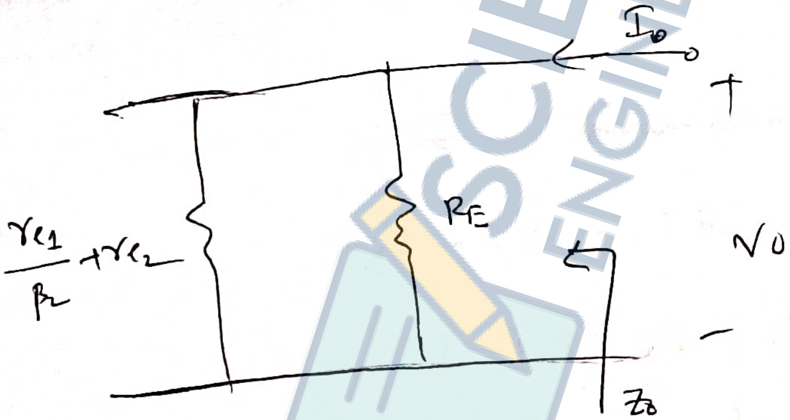
$$\Rightarrow I_o = \frac{V_o}{R_E} + \frac{(\beta_1+1)(\beta_2+1)\, V_o}{\beta_1 r_{e1} + (\beta_1+1)\beta_2 r_{e2}}$$

$$\cancel{\Rightarrow I_o \simeq \frac{V_o}{R_E} + \frac{\beta_1 \beta_2 V_o}{\beta_1 r_{e1} + (\beta_1+1)\beta_2}}$$

$$\Rightarrow I_o \simeq \frac{V_o}{R_E} + \frac{\beta_1 \beta_2 V_o}{\beta_1 r_{e1} + \beta_1 \beta_2 r_{e2}} \qquad \left(\because \begin{array}{l} \beta_1+1 \simeq \beta_1 \\ \beta_2+1 \simeq \beta_2 \end{array} \right)$$

$$\Rightarrow I_o = \frac{V_o}{R_E} + \frac{V_o}{\frac{r_{e1}}{\beta_2} + r_{e2}}$$

which defines a parallel resistance n/w.

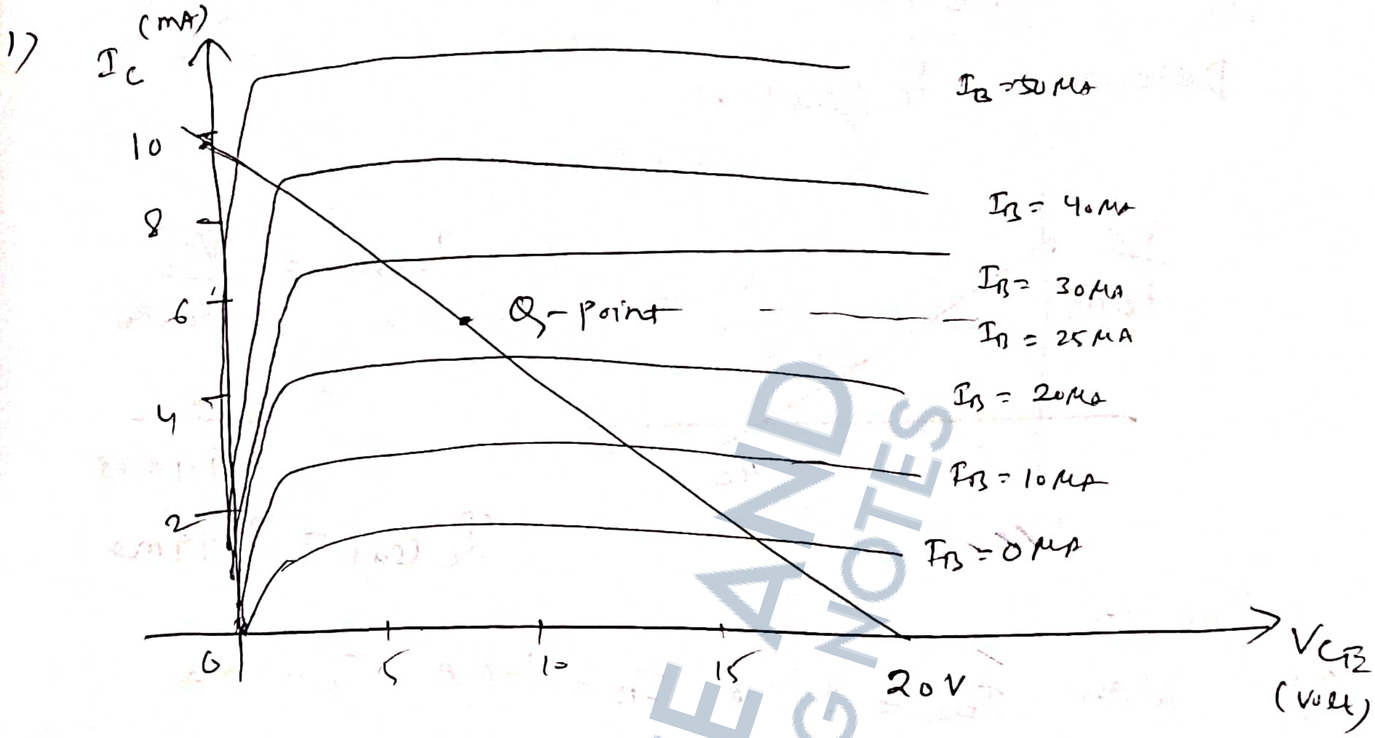


In general, $R_E >> \frac{r_{e1}}{\beta_2} + r_{e2}$

$$Z_o = R_E \parallel \left(\frac{r_{e1}}{\beta_2} + r_{e2}\right) \simeq \frac{r_{e1}}{\beta_2} + r_{e2}$$

$$\boxed{Z_o \simeq \frac{r_{e1}}{\beta_2} + r_{e2}}$$

Some Important Questions

1) 

For the given figure, determine the required Values of V_{CC}, R_C, and R_B for fixed-bias configuration.

Ans:

$$I_{C(max)} = \frac{V_{CC}}{R_C} = 10\,mA \quad \text{—} \ (1)$$

$$\boxed{V_{CC} = 20V} \quad \text{—} \ (2)$$

$$\Rightarrow \frac{20}{R_C} = 10$$

$$\Rightarrow R_C = \frac{20V}{10mA} = 2\times10^3$$

$$\Rightarrow \boxed{R_C = 2\,k\Omega}$$

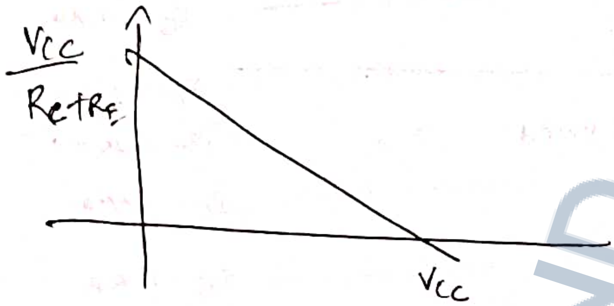
From graph, $I_B = 25\,\mu A$

$$R_B = \frac{V_{CC}-V_{BE}}{I_B} = \frac{20-0.7}{25\times10^{-6}} = 772\,k\Omega$$

2) For voltage divider bias circuit with

$V_{CC} = 16V$, $R_C = 3.9K$, $R_E = 0.68K$

Determine $I_{C(sat)}$.



$$I_{C(sat)} = \frac{V_{CC}}{R_C + R_E}$$

$$= \frac{16}{3.9 + 0.68}$$

$$I_{C(sat)} = 3.49\,mA$$

3) With $I_{CEO} = 0.5mA$, $I_B = 16\mu A$, $I_C = 2mA$

Calculate α, β, I_{CBO}.

Ans..

$$\boxed{\beta} = \frac{I_C}{I_B} = \frac{2 \times 10^{-3}}{16 \times 10^{-6}} = \frac{1}{8} \times 10^{3} = \boxed{125}$$

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

$$\Rightarrow 0.5 \times 10^{-3} = (125 + 1)\, I_{CBO}$$

$$\Rightarrow \boxed{I_{CBO}} = \frac{0.5 \times 10^{-3}}{126} = \boxed{3.96\,\mu A}$$

$$\alpha = \frac{\beta}{1 + \beta} = \frac{125}{126} = 0.99$$

4) Determine R_C, R_E, V_E, V_B, R_2 and β for V.D bias ckt with $V_{CC} = 24V$, $R_C = 3R_E$, ~~$\beta = 120$~~, $R_1 = 24K$, $I_B = 38.5\mu A$, $I_C = 5mA$

$I_{C(sat)} = 7.5 mA$

Ans:: $I_{C(sat)} = \frac{V_{CC}}{R_C + R_E}$

$\Rightarrow 7.5 mA = \frac{24}{3R_E + R_E}$

$\Rightarrow 4R_E = \frac{24}{7.5 \times 10^{-3}} = 3.2 \times 10^3$

$\Rightarrow \boxed{R_E = 0.8 K\Omega}$

$\boxed{R_C = 3R_E = 2.4 K\Omega}$

$V_E = I_E R_E = (I_C + I_B) R_E$

$\Rightarrow V_E = (5 + 0.0385) \times 10^{-3} \times 0.8 K\Omega$

$\Rightarrow \boxed{V_E = 4.0308 V}$

$V_{BE} = V_B - V_E$

$\Rightarrow 0.7 = V_B - 4.0308$

$\Rightarrow \boxed{V_B = 4.7308 V}$

$$I_B = \frac{V_{Th} - V_{BE}}{R_{Th} + (1+\beta)R_E}$$

$$= \frac{\frac{V_{CC}}{R_1+R_2} \times R_2 - V_{BE}}{\frac{R_1 R_2}{R_1+R_2} - (1+\beta)R_E}$$

$$\beta = \frac{I_C}{I_B} = \frac{5\,mA}{38.5\,\mu A} = \frac{5\times10^{-3}}{38.5\times10^{-6}} = 129.87$$

$$\Rightarrow 38.5\times10^{-6} = \frac{\frac{24}{24\times10^3 + R_2} \times R_2 - 0.7}{\frac{24\times10^3 R_2}{24\times10^3 + R_2} - (129.87+1)\times 0.8\times10^3}$$

$\Rightarrow$ $\boxed{\text{Find } R_2 =}$

or Approximate Analysis

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

$$\Rightarrow 4.7308 = \frac{24}{24\times10^3 + R_2} \times R_2$$

$$\Rightarrow 113.53\times10^3 + 4.7308\times R_2 = 24 R_2$$

$$\Rightarrow 19.26 R_2 = 113.57\times10^3$$

$$\Rightarrow R_2 = 5.89\times10^3 = 5.89\,k\Omega$$

Check

$$\beta R_E \geq 10 R_2$$

$$129.87 \times 0.8 \, k\Omega \geq 10 \times 5.89 \times 10^3$$

$$103.89 \, k\Omega \geq 50.89 \, k\Omega$$