Model Question Paper-2 with effect from 2019-20 (CBCS Scheme)

USN | | | | | | | | | |

Fourth Semester B.E. Degree Examination

Subject Title: Analog Circuits

TIME: 03 Hours **Max. Marks: 100**

Note: Answer any **FIVE** full questions, choosing at least **ONE** question from each **MODULE.**

		Module -1	***Bloom's Taxonomy Level**	**Marks**
Q.01	a	Derive the following relations with respect to small signal operation of BJT: i)Transconductance ii)Voltage gain	L2	6
	b	A BJT having β=100 is biased at a DC collector current of 1mA. Find the value of g_m, r_e and r_π at the bias point.	L3	6
	c	With the small signal equivalent model of MOSFET, derive an expression for voltage gain and transconductance.	L2	8
		OR		
Q.02	a	Derive the following relations with respect to small signal operation of BJT: i)Input resistance ii) Emitter resistance Also derive the relation between emitter and base resistance.	L2	8
	b	A MOSFET is to operate at I_D=0.1mA and is to have g_m=1mA/V. If k_n' = 50µA/V^2. Find the required W/L ratio and the overdrive voltage.	L3	6
	c	State the disadvantage of fixed V_{GS} biasing technique and explain how stability of operating point is achieved in drain to gate feedback resistor biasing technique in a MOSFET amplifier	L1 , L2	6
		Module-2		
Q. 03	a	With a neat circuit diagram and ac equivalent circuit, derive the expressions for R_{in}, A_{vo}, A_v and R_o for a Source follower.	L2	8
	b	A CS amplifier utilizes a MOSFET biased at I_D=0.25mA with V_{OV}=0.25V and R_D=20kΩ. The device has V_A=50V. The amplifier is fed with a source having R_{sig}=100kΩ, and a 20-kΩ load is connected to the output. Find R_{in}, A_{vo}, A_v and R_o and G_V. If to maintain reasonable linearity, the peak of the input sine-wave signal is limited to 10% of ($2V_{OV}$) what is the peak of the sinewave voltage at the output?	L3	8
	c	In an RC Phase shift oscillator, R=200kΩ and C=200pF. Find the frequency of the BJT based oscillator.	L3	4
		OR		
Q.04	a	Draw and explain the complete frequency response of a common source amplifier. Derive the expression for its lower cutoff frequency	L1,L2	10
	b	Find the midband gain A_M , and the upper 3-dB frequency f_H of a CS amplifier fed with a signal source having an internal resistance R_{sig} = 100 kΩ. The amplifier has R_G = 4.7MΩ, R_D = R_L = 15 kΩ, , g_m=1mA/V, r_o=150kΩ , C_{gs}=1pF and C_{gd}=0.4pF	L3	6
	c	Explain the working of a Colpitts oscillator.	L1	4
		Module-3		

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Scheme & Solution
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Q. 05	a	With a neat block diagram explain the working of a Voltage series feedback amplifier. How are the overall gain, input and output impedances affected in these amplifiers?	L1,L2	8
	b	Show how Gain can be desensitized and bandwidth increased with the application of negative feedback.	L3	8
	c	Draw the circuit of a practical Voltage Shunt (or transresistance) feedback amplifier and explain its working.	L2	4
		OR		
Q. 06	a	Explain a Class B Output stage. Prove that the maximum conversion efficiency of a Class B transformer coupled amplifier is 78.5%.	L1,L2	8
	b	A transformer coupled class A power amplifier supplies to an 80Ω load connected across the secondary of a step down transformer having a turns ratio 5:1. Determine the maximum power output for a zero signal collector current of 120mA.	L3	6
	c	What is cross over distortion? How can it be eliminated?	L2	6
		Module-4		
Q. 07	a	Explain with a neat diagram and relevant expressions, an opamp voltage series feedback amplifier	L1,L2	8
	b	Explain the following: 1) Virtual ground 2) Opamp AC amplifier	L1	6
	c	For an opamp non-inverting amplifier using 741 IC with R_L=1 KΩ and R_F=10KΩ, A= 200,000; Ri =2MΩ, Ro = 75Ω, fo =5 Hz; supply voltages ±15V, output voltage swing = ±13V, Compute A_F, R_{if}, R_{of}, f_F.	L3	6
		OR		
Q. 08	a	Explain an Instrumentation amplifier using transducer bridge with relevant equations.	L1	8
	b	Explain the basic comparator circuit using an opamp. How can this circuit be used in an application as a zero crossing detector?	L1	6
	c	For a Schmitt trigger circuit; R_1= 150Ω and R_2= 68kΩ, v_{in} =500mVp-p sine wave and saturation voltages are = ±14 V. Determine the threshold voltages V_{ut} and V_{lt}. Draw the output waveforms.	L3	6
		Module-5		
Q. 09	a	Explain the operation of 4-bit R-2R DAC with neat circuit. For the R-2R DAC, with R=10kΩ and R_F=20kΩ and V_{REF}=5V, determine the output voltage when the inputs b0=b1=5V and b2=b3=0V	L2,L3	8
	b	Explain the operation of a Successive -approximation ADC with neat circuit diagram.	L2	6
	c	Draw the circuit and frequency response of a first order low pass filter. Design a first order low pass filter to have a cutoff frequency of 1kHz with a passband gain of 2.	L1,L3	6
		OR		
Q. 10	a	Draw and Explain the circuit and frequency response of a wide band-pass filter.	L1	6
	b	Explain the operation of a monostable multivibrator with relevant diagrams and waveforms.	L1,L2	8

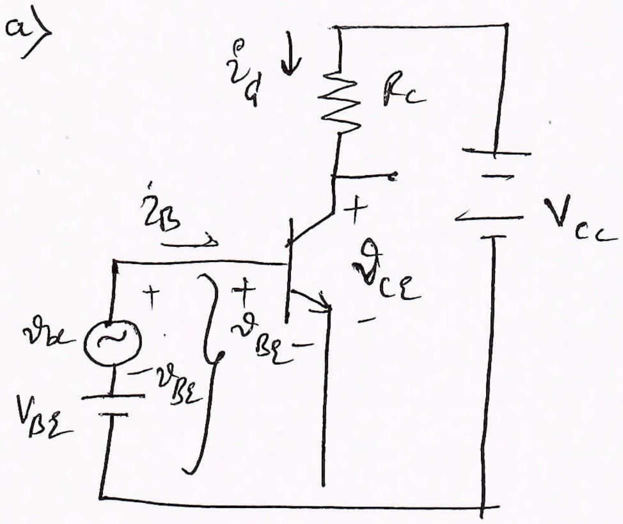
	c	In the astable multivibrator R_A=2.2kΩ, R_B=3.9kΩ and C=0.1μF. Determine the positive pulse width t_c, negative pulse width t_d and free-running frequency.	L3	6

*Bloom's Taxonomy Level: Indicate as L1, L2, L3, L4, etc. It is also desirable to indicate the COs and POs to be attained by every bit of questions.

Analog Circuits - 18EC42

Set - II

1a)



Total Marks 6M

→ 1M.

Instantaneous base emitter voltage is given by

$$v_{BE} = V_{BE} + v_{be} \rightarrow (1)$$

WKT,

$$i_C = I_S e^{v_{BE}/V_T} \rightarrow (2)$$

$$\therefore i_C = I_S e^{(V_{BE}+v_{be})/V_T}$$

$$i_C = I_S e^{V_{BE}/V_T} e^{v_{be}/V_T}$$

$$i_C = I_C e^{v_{be}/V_T}$$

If $v_{be} \ll V_T$ we can write

$$i_C = I_C \left(1 + \frac{v_{be}}{V_T}\right) \rightarrow (3)$$

$$\therefore i_C = \underbrace{I_C}_{= \text{DC comp}} + \underbrace{\frac{I_C}{V_T} v_{be}}_{\text{small s/g component}} \rightarrow (4)$$

$$i_c = \frac{I_C}{V_T} v_{be} \rightarrow (5)$$

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$$\boxed{i_c = g_m v_{be}} \rightarrow (6)$$

$$\therefore \boxed{g_m = \frac{I_C}{V_T}} \quad \text{or} \quad \boxed{g_m = \frac{i_c}{v_{be}}} \rightarrow (7)$$

1M

Applying KCL at o/p side

$$-V_{CC} + i_C R_C + v_C = 0 \rightarrow (8)$$

$$v_C = V_{CC} - i_C R_C \rightarrow$$

$$= V_{CC} - (I_C + i_c) R_C$$

$$= (V_{CC} - I_C R_C) - i_c R_C$$

$$v_C = \underbrace{V_C}_{\text{DC comp}} - \underbrace{i_c R_c}_{\text{AC comp}} \rightarrow (9)$$

$$\therefore v_c = -i_C R_C$$

$$v_c = -(g_m v_{be}) R_C$$

$$v_c = (-g_m R_C) v_{be}$$

$$\boxed{A_v = \frac{v_c}{v_{be}} = -g_m R_C} \rightarrow (10)$$

3M

Rahul.C.R

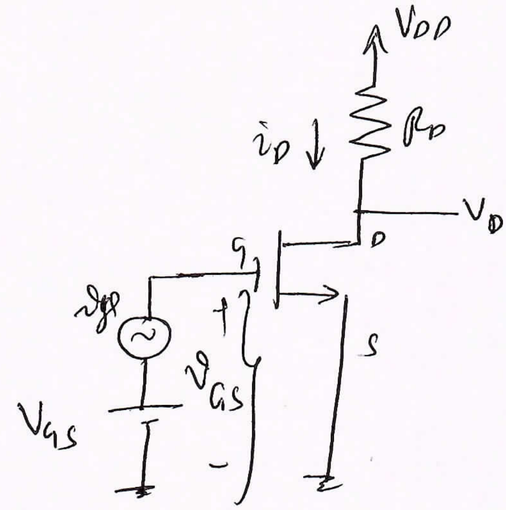
1b) $\beta = 100$ $I_C = 1mA$ $\boxed{\text{Total Marks :- 6M}}$ (3)

i) $g_m = \frac{I_C}{V_T} = \frac{1mA}{25mV} = 40mA/V$ $\rightarrow \boxed{2M}$

ii) $r_e = \frac{1}{g_m} = \frac{1}{40mA/V} = 25\Omega$ $\rightarrow \boxed{2M}$

iii) $r_\pi = (1+\beta) r_e = 2.5K\Omega$ $\rightarrow \boxed{2M}$

1c)



$\boxed{\text{Total Marks 8M}}$

$\rightarrow$ (1M)

1) from above figure

$$v_{GS} = v_{gs} + V_{GS} \rightarrow (1)$$

2) The instantaneous drain current i_D is given by

$$i_D = \frac{1}{2} K_n' \frac{W}{L} (v_{GS} - V_t)^2 \rightarrow (2)$$

$$i_D = \frac{1}{2} K_n' \frac{W}{L} \cdot [(v_{gs} + V_{GS}) - V_t]^2$$

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$$= \frac{1}{2} K_n' \frac{W}{L} [(V_{GS} - V_t) + v_{gs}]^2$$

$$i_D = \frac{1}{2} K_n' \frac{W}{L} [(V_{GS} - V_t)^2 + 2(V_{GS} - V_t) v_{gs} + v_{gs}^2]$$

$$i_D = \frac{1}{2} k_n' \frac{W}{L}(V_{GS}-V_t)^2 + k_n' \frac{W}{L}(V_{GS}-V_t)\, v_{gs} + \frac{1}{2} k_n' \frac{W}{L} v_{gs}^2 \rightarrow (4)$$

1) First term → dc bias current I_D

2) Second term → current component directly proportional to v_{gs}

3) Third term → current component proportional to square of i/p signal v_{gs}

Third term is undesirable because it represents non-linear distortion → [2M]

$$\frac{1}{2} k_n' \frac{W}{L} v_{gs}^2 \ll k_n' \frac{W}{L}(V_{GS}-V_t)\, v_{gs} \rightarrow (5)$$

$$v_{gs} \ll 2(V_{ov}) \rightarrow (6)$$

WKT,

$$i_D = I_D + i_d \rightarrow (7)$$

Where

$$i_d = k_n' \frac{W}{L}(V_{GS}-V_t)\, v_{gs} \rightarrow (8)$$

$$\boxed{g_m = \frac{i_d}{v_{gs}} = k_n' \frac{W}{L}(V_{GS}-V_t)} \rightarrow (9)$$

→ [1M]

$$\boxed{g_m = k_n' \frac{W}{L}(V_{ov})} \rightarrow (10)$$

Rahul.c.r

Voltage Gain

5

The instantaneous drain voltage is given by.

$$-V_{DD} + i_D R_D + v_D = 0$$

$$v_D = V_{DD} - i_D R_D \rightarrow (1)$$

Under small signal condition we have,

$$i_D = I_D + i_d \rightarrow (2)$$

Substituting (2) in (1)

$$v_D = V_{DD} - (i_d + I_D) R_D$$

$$= \underbrace{V_{DD} - I_D R_D}_{DC} \underbrace{- R_D i_d}_{AC}$$

$$v_D = V_D - i_d R_D \rightarrow (3)$$

WKT $$v_D = V_D + v_d \rightarrow (4)$$

Comparing (3) & (4)

$$\boxed{v_d = -i_d R_D} \rightarrow (5)$$

WKT. $$g_m = \frac{i_d}{v_{gs}} \rightarrow (6)$$

$$v_d = -g_m v_{gs} R_D$$

$$v_d = (-g_m R_D) v_{gs}$$

→ 4M

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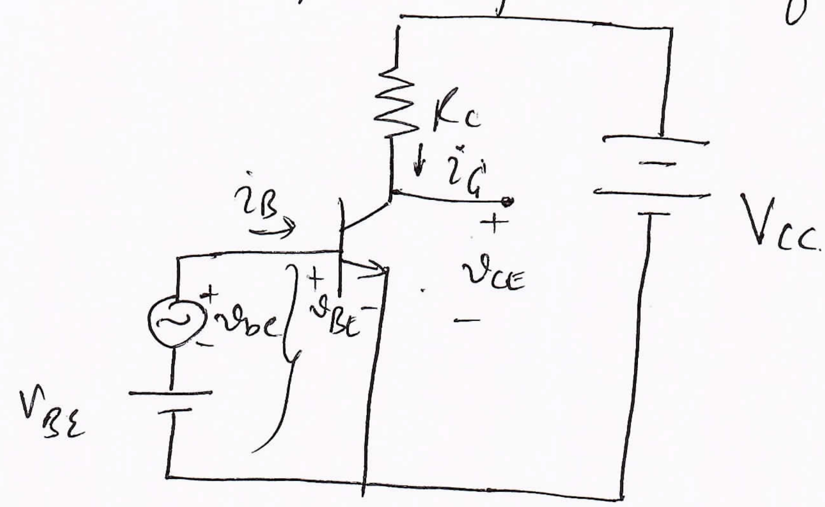
$$\therefore \boxed{A_v = \frac{v_d}{v_{gs}} = -g_m R_D} \rightarrow (7)$$

2a) Small Signal operation of BJT.

Total Marks 8M



→ 1M

The total base current i_B is given by:

$$i_B = \frac{i_C}{\beta} \rightarrow (1)$$

$$i_C = I_C + \frac{I_C}{V_T} v_{be}$$

Hence we get,

$$i_B = \frac{I_C}{\beta} + \frac{I_C}{\beta V_T} v_{be} \rightarrow (2)$$

Since $i_B = I_B + i_b \rightarrow (3)$

Comparing (3) & (2)

$$I_B = I_C/\beta$$

$$\boxed{i_b = \frac{1}{\beta} \frac{I_C}{V_T} v_{be}} \rightarrow (4) \rightarrow \frac{1}{2}M$$

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

$$\boxed{i_b = \frac{g_m}{\beta} v_{be}} \rightarrow 5$$

$\boxed{\frac{1}{2}M}$

Hence small signal input resistance b/w base and emitter looking from base is denoted by

$$\boxed{r_\pi = \frac{v_{be}}{i_b} = \frac{\beta}{g_m}} \rightarrow 6$$

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

$$r_\pi = \frac{\beta}{I_C/V_T}$$

$$\therefore \boxed{r_\pi = \beta \frac{V_T}{I_C}} \rightarrow 7$$

$$\therefore \boxed{r_\pi = \frac{V_T}{I_B}} \rightarrow 8$$

$\boxed{2M}$

2) Emitter Resistance

The total emitter current is given by

$$i_E = \frac{i_C}{\alpha} = \frac{I_C}{\alpha} + \frac{i_c}{\alpha} \rightarrow 1$$

$$i_E = I_E + i_e \rightarrow 2$$

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Comparing (1) & (2) (8)

$$i_E = I_E + i_e$$

$$\boxed{I_E = \frac{I_C}{\alpha}} \rightarrow (3)$$

$$\boxed{i_e = \frac{i_c}{\alpha} = \frac{I_C}{\alpha V_T} v_{be} = \frac{I_E}{V_T} v_{be}} \rightarrow (4)$$

$$\therefore \boxed{r_e = \frac{v_{be}}{i_e} = \frac{V_T}{I_E}} \rightarrow$$

2M

Relationship between r_π & r_e

WKT,

$$r_\pi = \frac{v_{be}}{i_b} \rightarrow (1)$$

$$r_e = \frac{v_{be}}{i_e} \rightarrow (2)$$

from (1) & (2)

$$v_{be} = r_\pi i_b = r_e i_e \rightarrow (3)$$

Hence we get

$$r_\pi = \frac{i_e}{i_b} r_e$$

$$\therefore \boxed{r_\pi = (1+\beta) r_e} \rightarrow (4)$$

2M

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2b) $g_m = 1\,mA/V$ $\quad K_n' = 50\,\mu A/V^2$ $\quad I_D = 0.1\,mA$ (4)

$\boxed{\text{Total Marks :- 6M}}$

$I_D = \frac{1}{2} K_n' \frac{W}{L} (V_{GS} - V_t)^2 = \frac{1}{2} K_n' \frac{W}{L} (V_{ov})^2 \rightarrow (1)$

$g_m = K_n' \frac{W}{L} (V_{ov}) \rightarrow (2)$

$\rightarrow \boxed{1M}$

$0.1\,mA = \frac{1}{2} \times 50\mu \times \frac{W}{L} \times (V_{ov})^2 \rightarrow (3)$

$1\,m = 50\mu \times \frac{W}{L} \times (V_{ov}) \rightarrow (4)$

$\left(\frac{W}{L}\right) \times V_{ov} = 20 \rightarrow (5)$ $\quad \rightarrow \boxed{2M}$

Substituting (5) in (3)

$0.1\,mA = 25\mu \times 20 \times V_{ov}$

$\boxed{V_{ov} = 0.2V} \rightarrow (6)$ $\quad \rightarrow \boxed{2M}$

Substituting (6) in (5)

$W/L = \frac{20}{V_{ov}} = 100 \rightarrow \boxed{1M}$

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2c) Disadvantage of biasing by fixing V_{GS} (10)

Total Marks 6M

$$I_D = \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_{GS} - V_t)^2$$

C_{ox} = oxide capacitance.

μ_n = mobility of electrons

W/L = transistor aspect ratio

V_t = threshold voltage

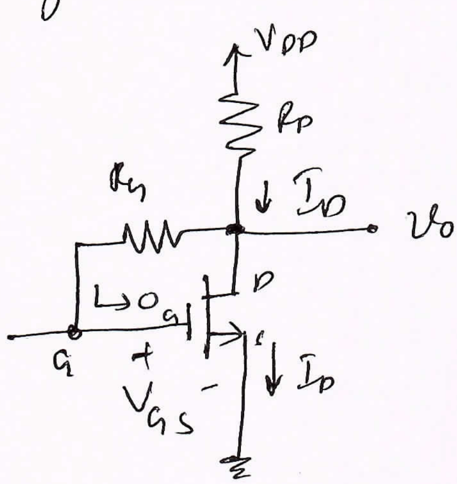
1) The transistor aspect ratio (W/L) varies widely from MOSFET to MOSFET. This causes change in value of I_D

2) Furthermore μ_n & V_t depends on temperature.

Hence by fixing V_{GS}, the drain current I_D becomes more temperature dependent.

→ 2M

Biasing using Drain to Gate feedback resistor



→ 1M

1) In above figure we can see that R_G provides negative feedback between drain & gate.

2) R_G is usually selected very large (MΩ) so that dc voltage at gate is same as at drain.

Applying KVL at o/p side (11)

$$-V_{DD} + I_D R_D + V_{DS} = 0$$

$$\therefore \boxed{V_{DD} = V_{DS} + I_D R_D} \rightarrow (1) \qquad \rightarrow \boxed{1M}$$

Because R_G is very large we can write

$$\boxed{V_{DS} = V_{GS}} \rightarrow (2)$$

Substituting (2) in (1)

$$\boxed{V_{DD} = V_{GS} + I_D R_D} \rightarrow (3) \qquad \rightarrow \boxed{1M}$$

from eqn (3) we can see that if I_D increases, V_{GS} decreases, If V_{GS} decreases, I_D will decrease ($\because V_{DD}$ is constant)

Here R_D - provides negative feedback & makes I_D stable. $\rightarrow \boxed{1M}$

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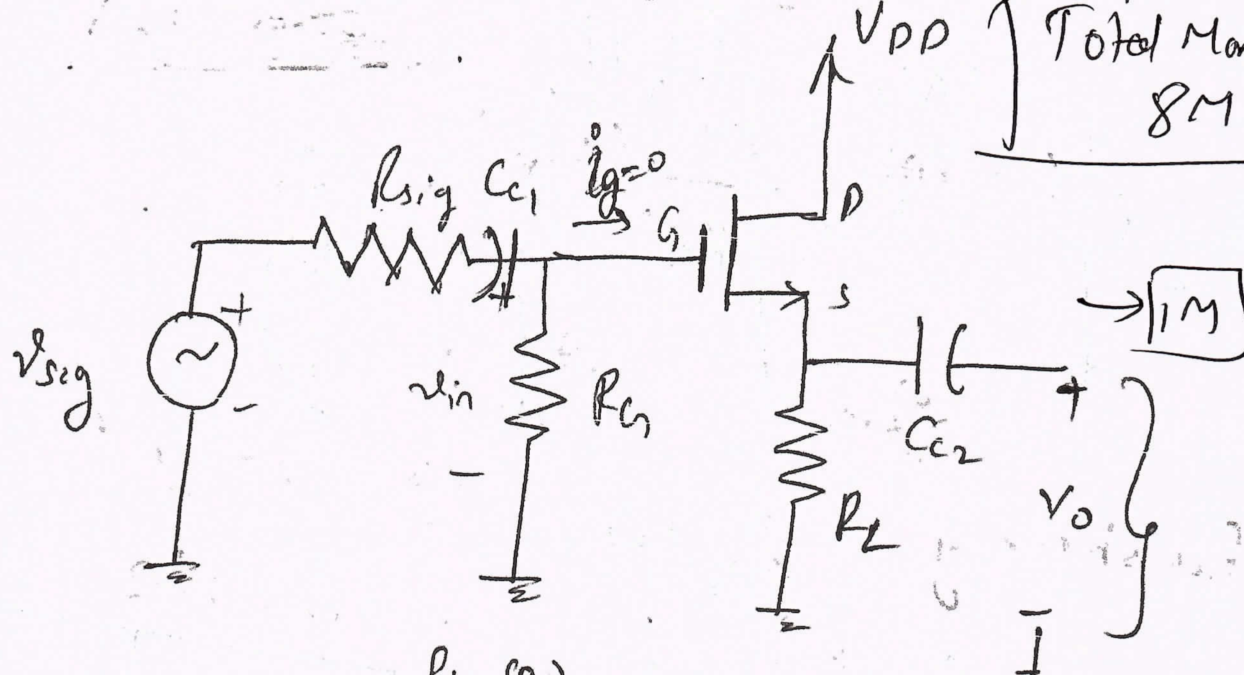
3a) Common-Drain Amplifier or Source follower

Total Marks 8M



→ 1M

fig (a)

1) fig (a) shows Source follower configuration.

2) Here input signal v_{sig} is applied to gate via R_{sig} & o/p is taken at Source. Hence the name "source follower".

3) Capacitor C_{c1}, C_{c2} acts as ac short circuit & DC open circuit.

4) O/p voltage is measured across R_L.

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AC Equivalent circuit

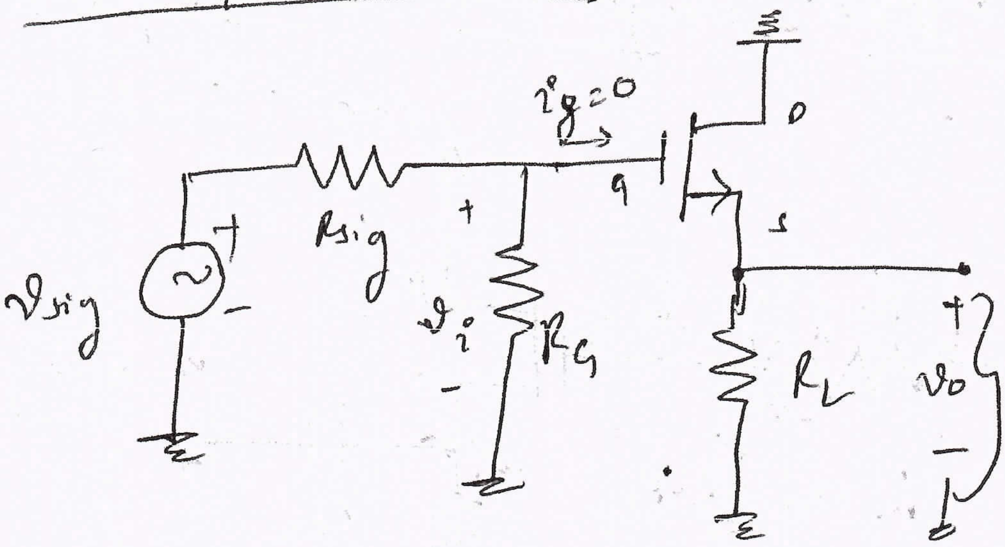

fig (b)

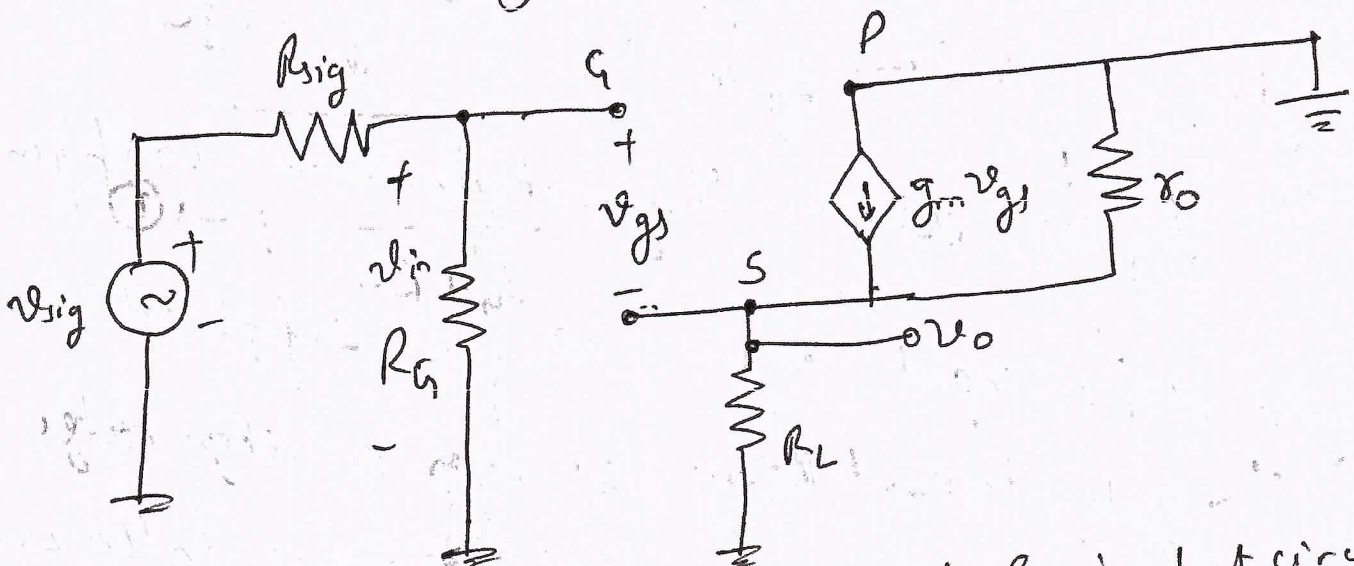

fig (c) :- Small signal Equivalent circuit

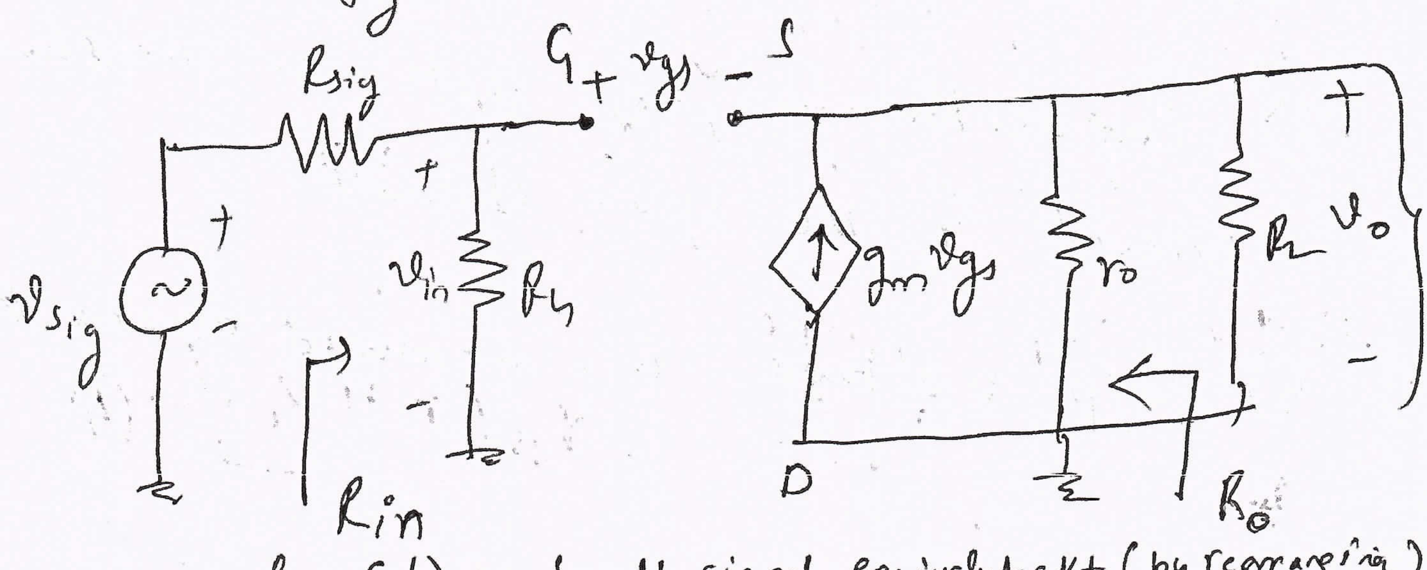

fig (d) : Small signal equivalent ckt (by rearranging)

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5) fig (d) can be used to determine input resistance, output resistance & voltage gain.

6) Input Resistance R_{in} is given by

$$\boxed{R_{in} = R_G} \rightarrow (1) \rightarrow \boxed{1M}$$

7) The input voltage V_{in} is determined by applying voltage divider bias formula.

$$\boxed{V_{in} = V_{sig} \times \frac{R_G}{R_G + R_{sig}}} \rightarrow (2)$$

8) The input gate Resistance R_G is very large (ranges from $1M\Omega - 10M\Omega$).

Hence $R_G >> R_{sig}$.

$$\therefore \boxed{V_{in} = V_{sig}} \rightarrow (3)$$

9) The o/p voltage is given by

$$\boxed{V_o = g_m V_{gs}(r_o \| R_L)} \rightarrow (4)$$

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10) Applying KVL b/w i/p & o/p (15)

$$-v_{in} + v_{gs} + v_o = 0$$

$$\therefore \quad v_{in} = v_o + v_{gs} \rightarrow (5)$$

Substituting (4) in (5)

$$\therefore \quad v_{in} = v_{gs} + g_m v_{gs} (r_o \| R_L)$$

Hence

$$A_v = \frac{v_o}{v_{in}} = \frac{g_m \cancel{v_{gs}} (r_o \| R_L)}{\cancel{v_{gs}} [1 + g_m (r_o \| R_L)]}$$

$$A_v = \frac{g_m (r_o \| R_L)}{1 + g_m (r_o \| R_L)}$$

$$\boxed{A_v = \frac{(r_o \| R_L)}{\frac{1}{g_m} + (r_o \| R_L)}} \rightarrow (6)$$

$\rightarrow$ $\boxed{2M}$

if $r_o = \infty$

$$\boxed{A_v = \frac{R_L}{1/g_m + R_L}} \rightarrow (7)$$

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To determine R_o

$\rightarrow \boxed{2M}$

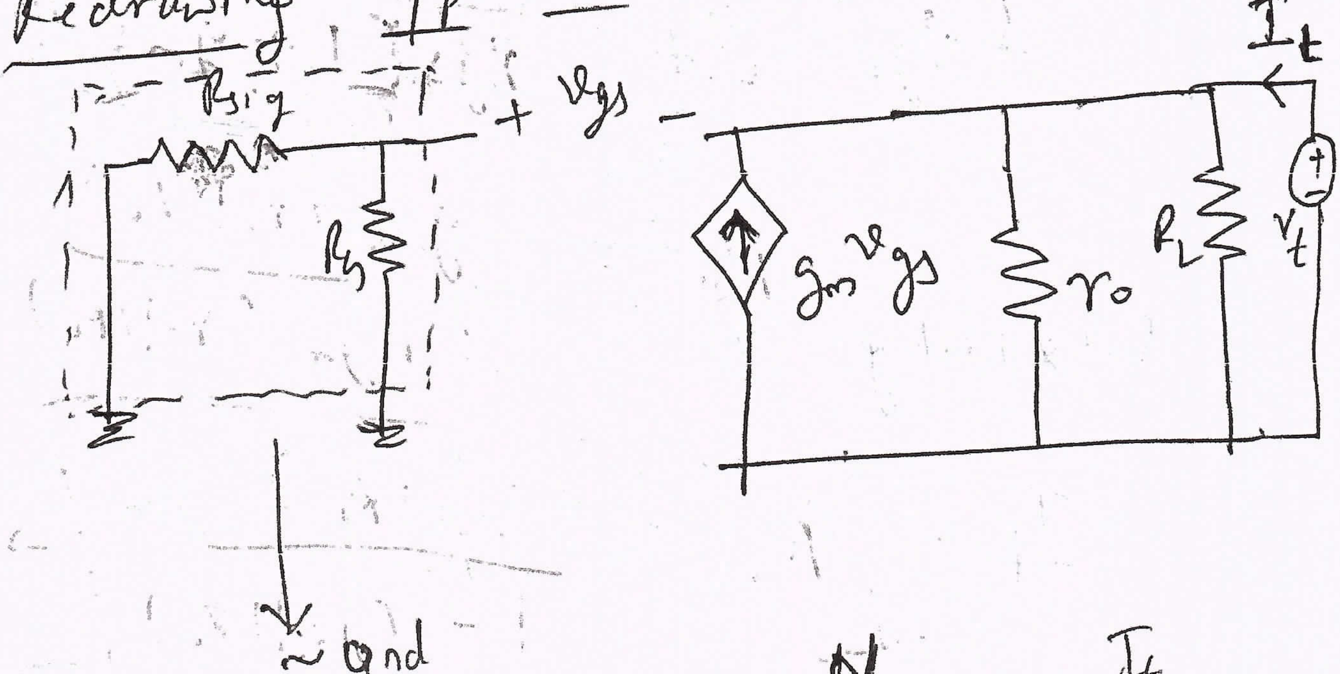
1) Make $v_{sig} = 0$

2) Apply test voltage v_t at o/p terminal

Since $v_{sig} = 0$ by eqn (3) we can write

$$v_{sig} = v_i = 0$$

Redrawing o/p circuit we have:



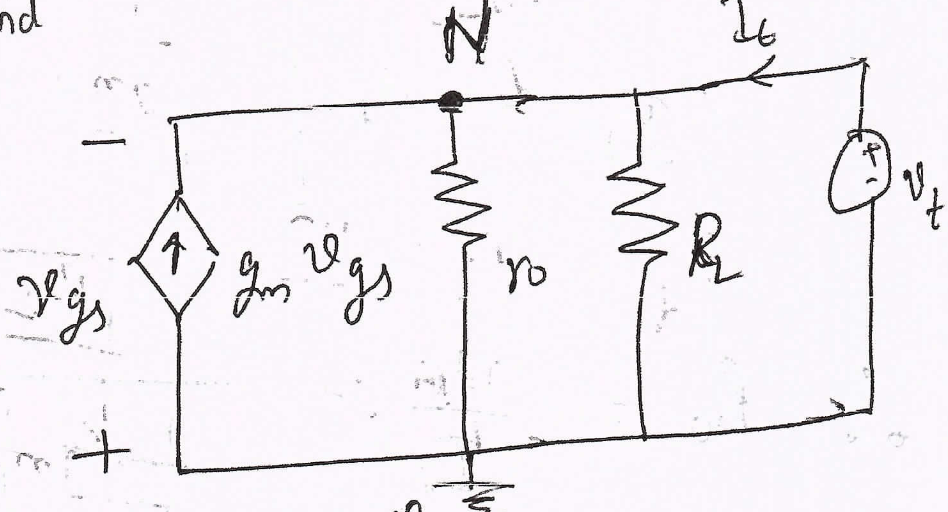


Fig (e)

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Applying KCL at node N.

$$I_t + g_m v_{gs} = \frac{v_t}{r_0} + \frac{v_t}{R_L}$$

$$I_t = v_t\left(\frac{1}{r_0} + \frac{1}{R_L}\right) - g_m v_{gs} \rightarrow (8)$$

Applying KVL at o/p side in fig (c)

$$+v_{gs} + v_t = 0$$

$$\therefore \boxed{v_{gs} = -v_t} \Rightarrow (9)$$

Substituting (9) in (8)

$$I_t = v_t\left(\frac{1}{r_0} + \frac{1}{R_L}\right) - g_m(-v_t)$$

$$I_t = v_t\left[\frac{1}{r_0} + \frac{1}{R_L} + g_m\right]$$

$$\therefore R_0 = \frac{v_t}{I_t} = \frac{1}{\frac{1}{r_0} + \frac{1}{R_L} + g_m}$$

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$$\therefore R_o = \frac{1}{\frac{1}{r_o} + \frac{1}{R_L} + \frac{1}{1/g_m}}$$

$$\therefore \boxed{R_o = r_o \parallel R_L \parallel 1/g_m} \rightarrow (10)$$

2M

Usually $r_o \gg R_L$

$$\therefore \boxed{R_o = R_L \parallel \frac{1}{g_m}} \rightarrow (11)$$

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5b)

1) $R_{in} = \infty$

2) $A_{vo} = -g_m(R_D \| r_o)$

Total Marks 8M

WKT

$$g_m = k_n' \frac{W}{L}(V_{ov})$$

But, $I_D = \frac{k_n'}{2}\frac{W}{L}(V_{GS}-V_t)^2$

$$I_D = \frac{k_n'}{2}\frac{W}{L}(V_{ov})^2$$

$$I_D = \frac{g_m}{2} V_{ov}.$$

→ 1M

$$g_m = \frac{2I_D}{V_{ov}} = 2mS$$

→ 2M

$$r_o = \frac{V_A}{I_D} = \frac{50}{0.25mA} = 200k\Omega.$$

→ 2M.

$$A_{vo} = -2mS(R_D \| r_o).$$

$$= -36.4$$

3) $R_o = R_D \| r_o = 18.2k\Omega$

→ 1M

4) $A_v = G_v = -g_m(R_D \| r_o \| R_L) = -19.$

→ 1M

5) $v_{gs} = (10\%)\, 2V_{ov}$

$= 0.05V$

$v_o = A_v v_{gs} = 19 \times 0.05 = 0.95V.$ → 1M.

3b) $R = 200k\Omega$ $C = 200pF$.

Total Marks 4M

$$f = \frac{1}{2\pi RC\sqrt{6}}$$ → 1M

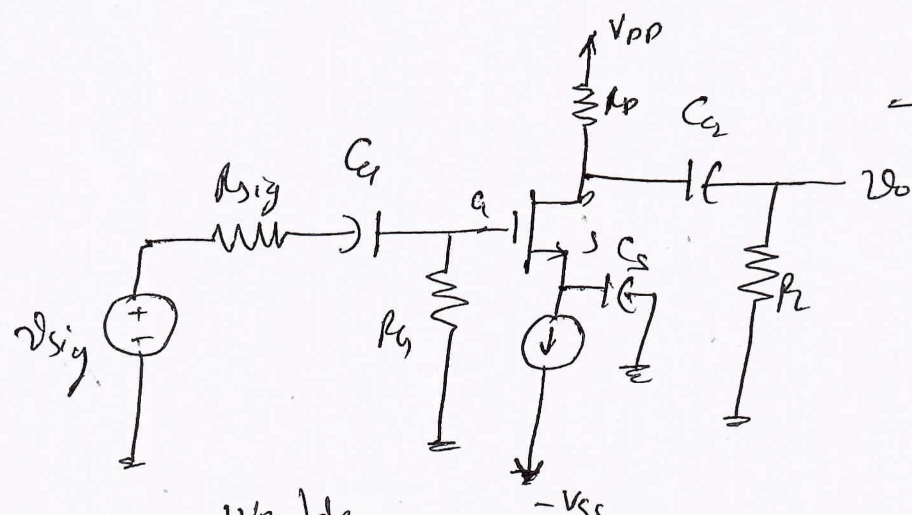
$$= \frac{1}{2\pi \times 200\times10^{3} \times 200\times10^{-12} \times \sqrt{6}}$$ → 1M

$$= \frac{1}{0.00061{,}311}$$ → 1M

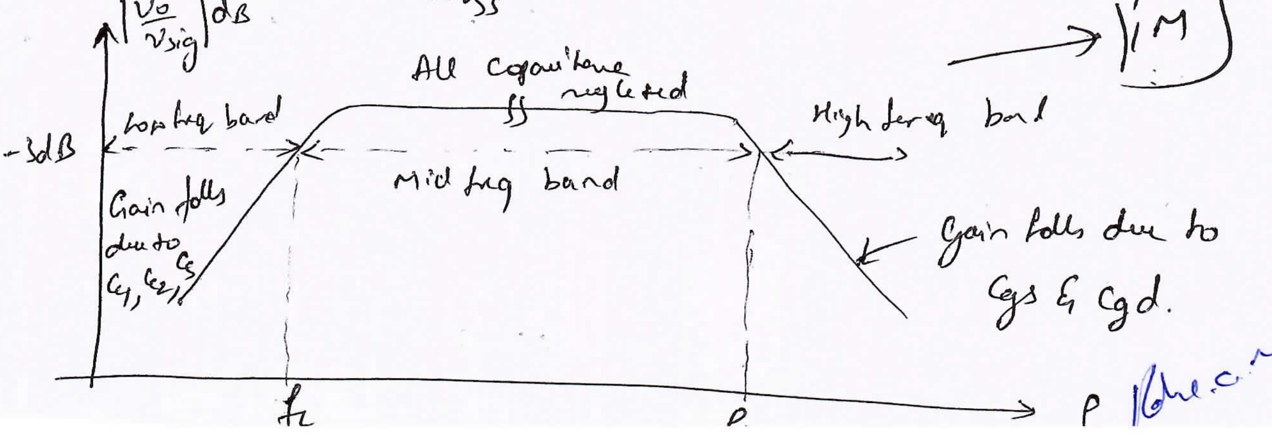
$= 1625\ Hz$ or $1.625\ KHz$. → 1M

4a) Common Source Amplifier

Total Marks 10M



→ 1M



→ 1M

1) from the gain frequency graph we can see that at lower frequency the gain falls due to effect of C_{c1}, C_s, C_{c2}. (21)

$(f < f_L)$

2) At mid freq. the capacitor C_{c1}, C_{c2}, C_s act like short circuit & we get maximum gain at mid-frequency band

3) As frequency is increased further in high frequency range $(f > f_H)$ gain falls off due to effect of parasitic capacitance C_{gs} & C_{gd}. $\rightarrow \boxed{2M}$

The gain of CS amplifier is given by.

$$A_M = \frac{v_o}{v_{sig}} = -\frac{R_G}{R_{sig}+R_G} g_m (r_o || R_D || R_L)$$

At $-3dB$ we get bandwidth given by.

$$B.W = f_H - f_L.$$

The figure of merit for C.S Amplifier is gain-bandwidth product

$$GB = |A_M| B.W. \rightarrow \boxed{1M}$$

Dhal.cm

Low frequency Response

To determine low frequency response we have to make $V_{DD} = 0$ & ignore r_o, Now when v_{sig} is applied

$$v_g = v_{sig} \cdot \frac{R_G}{R_G + \frac{1}{sC_{C1}} + R_{sig}}$$

$$v_g = v_{sig} \times \frac{R_G}{R_G + R_{sig}} \times \frac{s}{s + \frac{1}{C_{C1}(R_G + R_{sig})}}$$

The break frequency ω_{p1} is given by $\hookrightarrow$ 2M

$$\omega_{p1} = \omega_o = \frac{1}{C_{C1}(R_G + R_{sig})} \rightarrow 1M$$

We determine I_d by dividing V_g by total impedance

$$I_d = \frac{V_g}{\frac{1}{g_m} + \frac{1}{sC_S}}$$

$$I_d = g_m V_g \cdot \frac{s}{s + \frac{g_m}{C_S}}$$

Kohl. C M.

$$I_d = g_m V_g \left(\frac{S}{S + \frac{g_m}{C_S}} \right)$$

C_S introduces frequency dependent factor. Hence another break frequency is given by

$$\therefore \omega_{p2} = \frac{g_m}{C_S}$$

$$\therefore I_o = -I_d \frac{R_D}{R_D + \frac{1}{S C_{C2}} + R_L}$$

$$V_o = I_o R_L = -I_D \frac{R_D R_L}{R_D + R_L} \frac{S}{S + \frac{1}{C_{C2}(R_D + R_L)}}$$

$\rightarrow \boxed{RM}$

Finally C_{C2} will introduce another break frequency

$$\omega_{p3} = \frac{1}{C_{C2}(R_D + R_L)}$$

$$\therefore \frac{V_o}{V_{sig}} = -\left(\frac{R_{in}}{R_{in} + R_{sig}}\right) (g_m (R_D \| R_L)) \left(\frac{S}{S + \omega_{p1}}\right)\left(\frac{S}{S + \omega_{p2}}\right)\left(\frac{S}{S + \omega_{p3}}\right)$$

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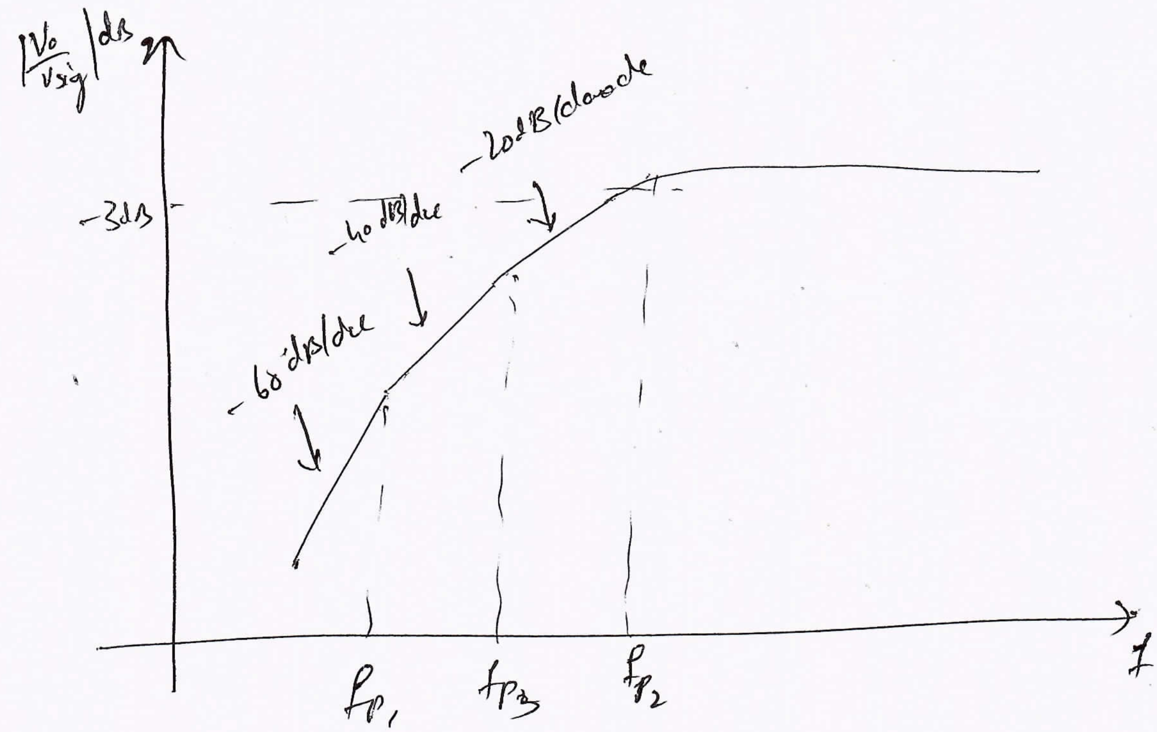


4b) $A_m = -\frac{R_G}{R_G + R_{sig}} g_m R_L'$

Total Marks 6M

$R_L' = r_o \| R_D \| R_L = 150 \| 15 \| 15 = 7.14\,k\Omega$ → 1M

$g_m R_L' = 1 \times 7.14 = 7.14\,V/V.$ → 1M

Thus $A_M = -\frac{4.7}{4.7+0.1} \times 7.14 = -7\,V/V.$ → 1M

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

$= (1 + 7.14) \times 0.4 = 3.26\,pF$

[illegible]

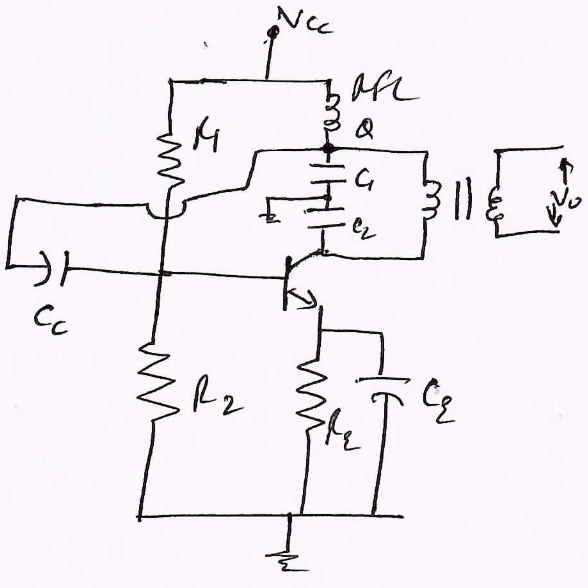
$C_{in} = C_{gs} + C_{eq} = 1 + 3.26 = 4.26\ pF$ → $\boxed{1M}$ 25

$$f_H = \frac{1}{2\pi C_{in}(R_{sig} \| R_G)}$$

$$= \frac{1}{2\pi \times 4.26 \times 10^{-12} (0.1 \| 4.7) \times 10^{6}}$$

$= 382\ KHz$ → $\boxed{2M}$

4c) Colpitts Oscillator



→ $\boxed{2M}$

) The oscillator consists of two stages. i.e Common Emitter stage. which introduces phase shift of 180°, where R_1 & R_2 are used to establish desired operating point.

(26)

2) The voltage across C_1 is feedback to amplifier input through coupling capacitor.

3) Due to split capacitor arrangement, the tank ckt introduces 180° phase shift.

4) Overall phase shift is 360°. → 2M

frequency of oscillation is given by

$$f = \frac{1}{2\pi\sqrt{L\,C_{eq}}}$$

when $C_{eq} = \frac{C_1 C_2}{C_1 + C_2}$.

Condition for sustained oscillation is given by

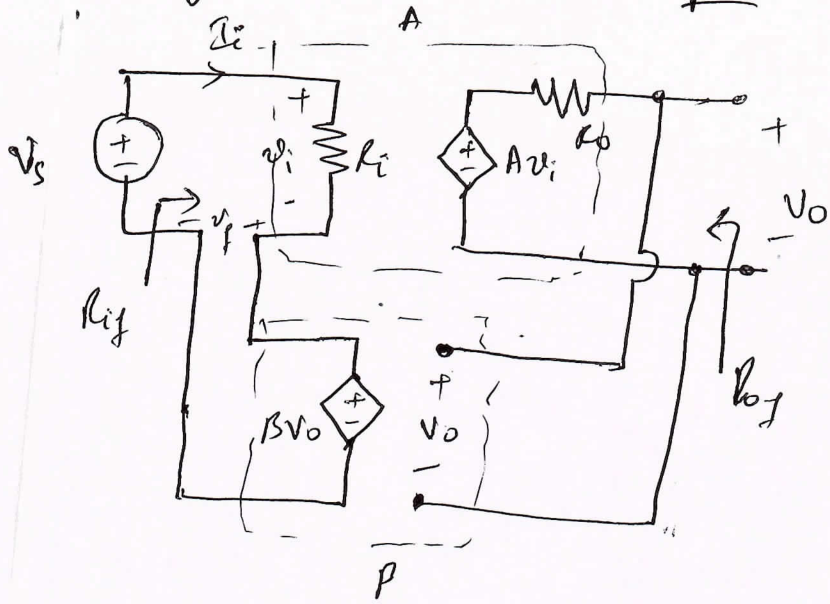
$$h_{fe} \geq \frac{C_2}{C_1}.$$

→ 2M

Rahul. CR

Voltage Series Feedback Amplifier

Total Marks 8M



→ 1M

Voltage Gain

$$v_o = A v_i$$

$$= A(v_s - v_f)$$

$$v_o = A(v_s - \beta v_o)$$

$$v_o(1 + A\beta) = A v_s$$

$$\therefore \boxed{A_f = \frac{v_o}{v_s} = \frac{A}{1 + A\beta}} \rightarrow (1) \rightarrow 1M$$

With feedback the closed loop voltage gain is reduced by factor $(1+A\beta)$

ii) $R_{if} = \frac{V_s}{I_i} = \frac{V_s}{V_i / R_i} = R_i \frac{V_s}{V_i} \rightarrow$

Applying KVL at i/p side

2M

$$-V_s + V_i + V_f = 0$$

$$V_s = V_i + V_f$$

$$V_s = V_i + \beta V_o$$

$$V_s = V_i + \beta(A V_i)$$

$$R_{if} = \frac{R_i (V_i + A\beta V_i)}{V_i}$$

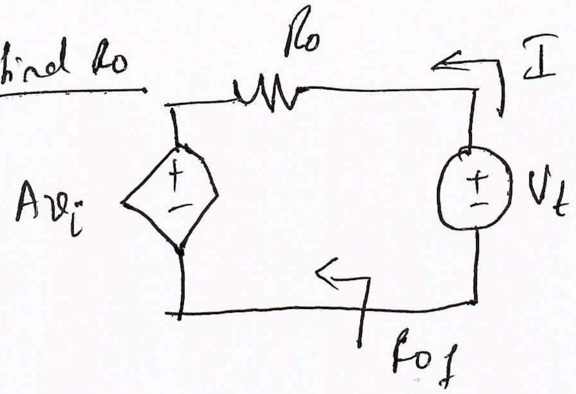
$$\boxed{R_{if} = R_i(1 + A\beta)} \rightarrow (2)$$

fdd.cn

From (2) input impedance with negative feedback is increased by factor $(1+A\beta)$

(28)

<u>To find Ro</u>



$\rightarrow$ 1M

$$R_{of} = V_t/I$$

where $I = \dfrac{V_t - Av_i}{R_o}$

Since $v_s = 0$, $v_i = -v_f = -\beta v_o = -\beta V_t$.

$$I = \frac{V_t + A\beta V_t}{R_o}$$

$$\therefore \boxed{R_{of} = \frac{R_o}{(1+A\beta)}} \rightarrow (3) \qquad \rightarrow 2M$$

from above eqn we can see that o/p impedance with feedback is reduced by factor $(1+A\beta)$ $\rightarrow$ 1M

Total Marks = 8M

5b) <u>Gain desensitivity</u>

With negative feedback closed loop voltage gain will be less sensitive to variation in value of circuit components

W.K.T. $A_f = A/1+A\beta \rightarrow (1)$ $\rightarrow$ 1M

Differentiating (1) wrt A.

Rahul C.R.

(99)

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

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

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

$$\therefore \frac{dA_f}{A_f} = \frac{dA\,/\,(1+A\beta)^2}{A/\cancel{(1+A\beta)}}$$

$$\boxed{\frac{dA_f}{A_f} = \frac{1}{(1+A\beta)}\frac{dA}{A}} \rightarrow (3)$$

$\rightarrow$ 2M

from above eqn we can see that % change in A_f is smaller than % change in A by factor $(1+A\beta)$ $\rightarrow$ 1M

Hence $(1+A\beta)$ is also known as desensitivity factor.

2) Bandwidth Extension

B.W. with feedback is much larger than bandwidth without feedback.

The high frequency response can be approximated to a single pole s/m. in midfreq & high freq.

This can be expressed as $\rightarrow$ 1M

$$A(S) = \frac{A_M}{1+S/\omega_H} \rightarrow (1)$$

$A_m \rightarrow$ mid band gain $\omega_H \rightarrow$ upper 3dB cutoff.

With -ve feedback, eqn (1) can be written as

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$$A_f(s) = \frac{A(s)}{1+\beta A(s)} \rightarrow (2)$$

Substituting (1) in (2)

$$A_f(s) = \frac{A_M/(1+s/\omega_H)}{1+\beta\left(\frac{A_M}{1+s/\omega_H}\right)}$$

$$A_f(s) = A_M \Big/ \left(1+\frac{s}{\omega_H}\right)+\beta A_M$$

$$\therefore A_f(s) = A_M \Big/ 1+A_M\beta+\frac{s}{\omega_H} \rightarrow (3)$$

Dividing (3) by $(1+A_M\beta)$ at numerator & denominator.

$$A_f(s) = \frac{\frac{A_M}{(1+A_M\beta)}}{\frac{1+A_M\beta+s/\omega_H}{(1+A_M\beta)}} = \frac{A_M/(1+A_M\beta)}{1+\frac{s}{\omega_H(1+A_M\beta)}} \rightarrow (4)$$

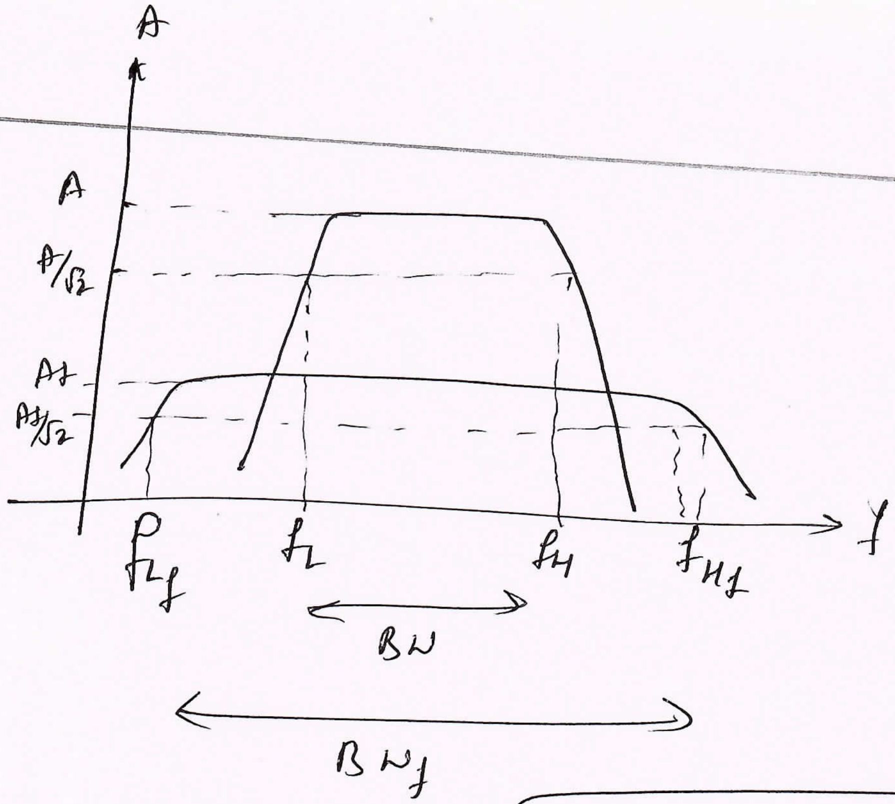
Comparing (4) & (1)

$$\boxed{A_{Mf} = A_M/(1+A_M\beta)} \rightarrow (5) \longrightarrow \boxed{1M}$$

$$\boxed{\omega_{Hf} = \omega_H(1+A_M\beta)} \rightarrow (6) \longrightarrow \boxed{1M}$$

ω_{Hf} is increased by factor $(1+A_M\beta)$ III^ly $\omega_{Lf} = \frac{\omega_L}{(1+A_M\beta)}$

Mohd. am.



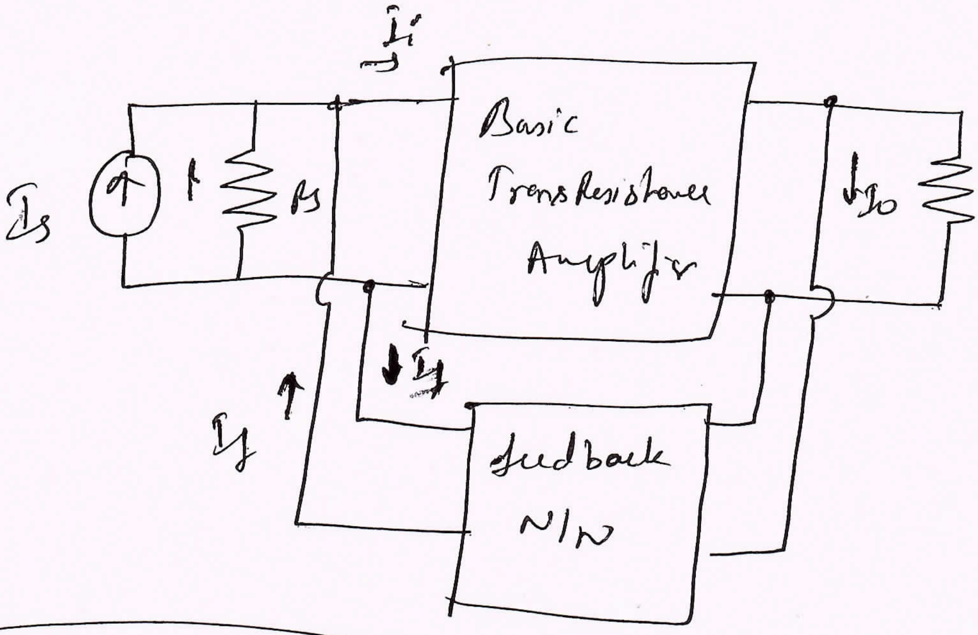
→ $\boxed{1M}$

5c) Trans Resistance Amplifier — $\boxed{\text{Total Marks:- 4M}}$

1): Here input signal is current & O/p signal is voltage

2> This topology is also called shunt shunt feedback

3) It has Low input & Low O/p. impedance.



→ $\boxed{2M}$

$$\boxed{R_{if} = \frac{R_i}{(1+A\beta)}} \rightarrow \boxed{1M}$$

$$\boxed{R_{of} = \frac{R_o}{(1+A\beta)}} \rightarrow \boxed{1M}$$

6a) Class B o/p stage

Total Marks :- 8M

i) Here transistor conducts for only one half cycle of input signal

ii) Here conduction angle is 180°

iii) Here I_c is zero

iv) Here Q point is located in cutoff region.

v) It has good power conversion efficiency.

→ 4M

Power conversion efficiency

$$\% \eta = \frac{P_{o(ac)}}{P_{i(dc)}} \times 100\% \rightarrow (1)$$

→ 1M

$$P_{o(ac)} = V_{rms} I_{rms}$$

$$= \frac{V_m}{\sqrt{2}} \times \frac{I_m}{\sqrt{2}} = \frac{V_m I_m}{2} \rightarrow (2)$$

→ 1M

where V_m = peak value of o/p vol = V_{CC}

I_m = " " " " current = $I_{C(p)}$

$$P_{i(dc)} = I_{dc} V_{CC} = \frac{2 I_m}{\pi} V_m \rightarrow (3)$$

Substituting (2) & (3) in (1)

$$\% \eta = \frac{V_m I_m / 2}{2 I_m / \pi \; V_m} \times 100\%$$

→ 1M

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$\therefore \quad \% \eta = \frac{\pi}{4} \times 100\%$

$$\boxed{\% \eta = 78.5\%} \rightarrow \boxed{1M}$$

6b) $R_L = 80\,\Omega$, $I_{CQ} = 120\,mA$, $\alpha = \frac{N_1}{N_2} = 5$

$\boxed{\text{Total Marks} = 6M}$

$P_{o(ac)} = I_{rms} V_{rms}$

$P_{o(ac)} = I_{rms}^2 \times R_L' \rightarrow \boxed{1M}$

$I_{rms} = \frac{I_m}{\sqrt{2}} = \frac{1}{\sqrt{2}}\left(\frac{I_{C(max)} - I_{C(min)}}{2}\right)$

$= \frac{1}{\sqrt{2}}\left(\frac{2I_{CQ} - 0}{2}\right)$

$I_{rms} = \frac{I_{CQ}}{\sqrt{2}} = \frac{120\,mA}{\sqrt{2}} \rightarrow \boxed{2M}$

$\therefore \quad P_{o(ac)} = \left(\frac{120\,mA}{\sqrt{2}}\right)^2 \times R_L' \rightarrow \boxed{1M}$

$R_L' = \left(\frac{N_1}{N_2}\right)^2 \times R_L = \left(\frac{5}{1}\right)^2 \times 80 = 2000\,\Omega$

$$\therefore \boxed{P_{o(ac)} = 14.4\,W} \rightarrow \boxed{2M}$$

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6c) 1) In class B power Amplifier circuit there will be (34) cross over distortion i.e. whenever V_i goes positive & exceeds above 0.5V. the transistor conducts & operates as emitter follower.

$$V_o = V_I - V_{BE_N} \rightarrow (1)$$

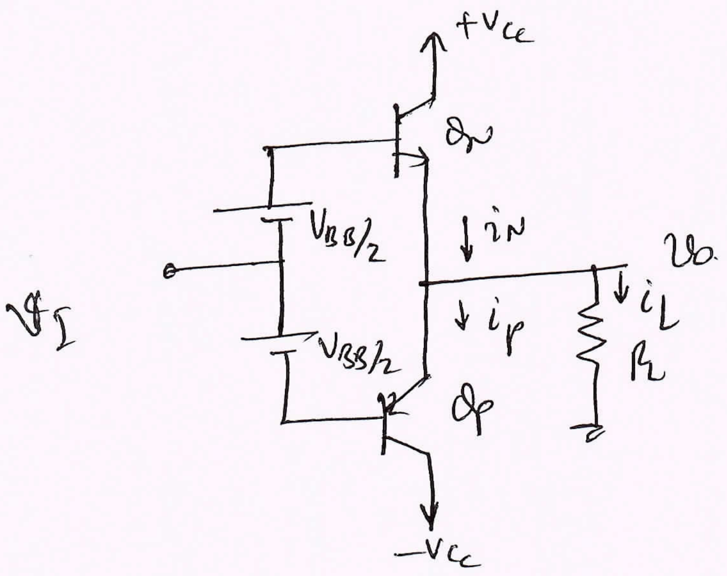
(16)
Total Marks
6M

2) When i/p voltage goes negative & exceeds above -0.5V. Q_p conducts & operates as emitter follower.

$$V_o = -V_I + V_{BEP} \rightarrow (2)$$

→ 2M

2) Here in order to overcome this we go for class AB power Amplifier.



→ 2.M

3) Here small dc voltage $V_{BB}/2$ is applied to base of transistor Q_N & Q_B.

5) $V_{BB}/2$ is selected such that it nullifies V_{BE} drop.

6) Here we bias transistor in cut-in region//.

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$-V_I - V_{BB}/2 + V_{BEN} + V_0 = 0 \rightarrow ①$ (38)

$$V_0 = V_I + V_{BB}/2 - V_{BEN}$$

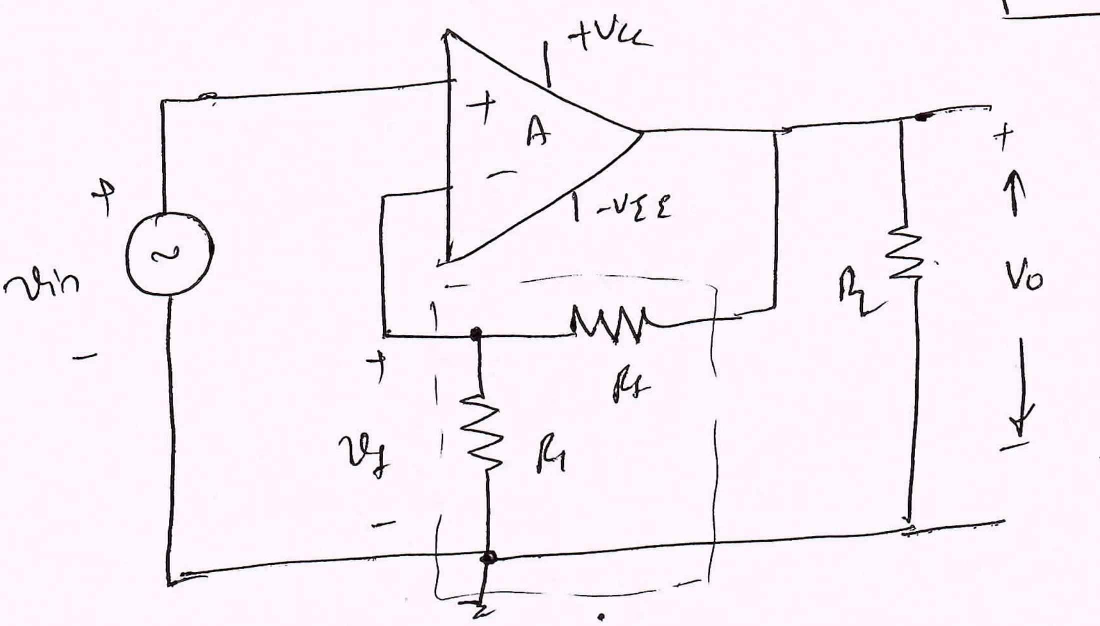
Since $V_{BB}/2 = V_{BEN}$

$$\boxed{V_0 = V_I} \rightarrow ② \qquad \rightarrow \boxed{2M}$$

from eqn ② we can see that crossover distortion is overcome by applying bias voltage.

7a) Voltage Series feedback Amplifier

Total Marks 8M



$\rightarrow \boxed{1M}$

1) The above figure shows voltage series feedback Amplr with open loop gain A & feedback circuit consisting of 2 resistors R_1 & R_f. It is also known as Non-Inverting Amplr.

2) $A = V_0/V_{id}$, $A_f = \frac{V_0}{V_{in}}$, $\beta = V_f/V_0$.

$\hookrightarrow \boxed{2M}$

i) <u>Negative feedback</u>

Applying KVL we get

$$v_{id} = v_{in} - v_f \rightarrow (1) \qquad \rightarrow \boxed{1M}$$

v_{in} = input voltage

v_f = feedback voltage.

v_{id} = difference input voltage.

from eqn (1) we can see that op-Amp amplifies difference b/w input voltages (v_{id})

ii) <u>Closed loop voltage gain</u>

$$A_f = \frac{v_o}{v_{in}} \longrightarrow (1)$$

But $\quad v_o = A(v_1 - v_2) \rightarrow (2)$

$$v_1 = v_{in} \rightarrow (3)$$

$$v_2 = v_f = \frac{R_1 v_o}{R_1 + R_f} \qquad R_i >> R_1$$

$$\therefore \quad v_o = A\left(v_{in} - \frac{R_1 v_o}{R_1 + R_f}\right)$$

we get

$$\boxed{A_f = \frac{v_o}{v_{in}} = \frac{A(R_1 + R_f)}{(R_1 + R_f) + AR_1}} \rightarrow (2)$$

$\rightarrow \boxed{2M}$

Rahul.c.M.

Since A is very large,

$AR_1 \gg (R_1 + R_f)$

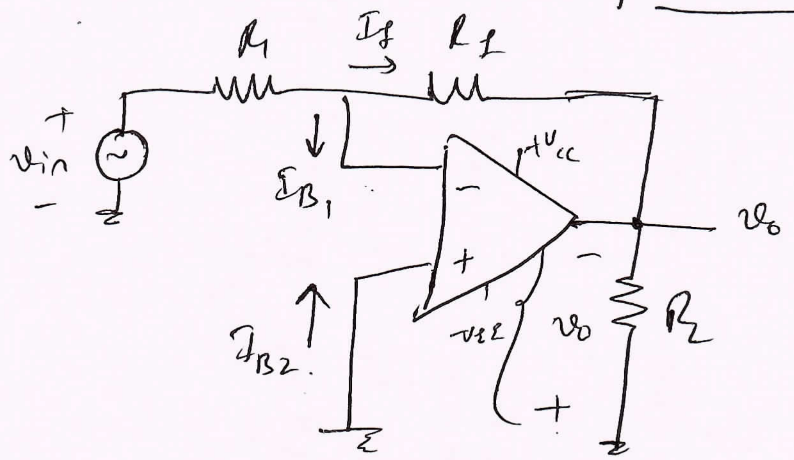
$$\therefore \boxed{A_f = \frac{v_o}{v_{in}} = 1 + \frac{R_f}{R_1}} \rightarrow (3) \rightarrow \boxed{1M}$$

$$\boxed{\beta = \frac{v_f}{v_o} = \frac{R_1}{R_1 + R_f} = \frac{1}{A_f}} \rightarrow (4) \rightarrow \boxed{1M}$$

from eqn (3) we can see that closed loop gain A_f depends only on R_f & R_1. & hence gain is constant.

7b) i) <u>Virtual Ground</u>

Total Marks :- 6M



$\rightarrow \boxed{1M}$

1) The above figure shows inverting Amplifier

2) Since non-inverting terminal is grounded & input signal is connected to inverting input terminal via R_1

3) The differential input voltage is ideally zero i.e. the voltage at inverting terminal v_2 is approximately equal to voltage at non-inverting terminal (v_1)

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line v_1 is connected to ground v_2 will be a virtual ground.

(38)

$$\therefore \quad i_{in} = i_f$$

$$\frac{v_{in} - v_2}{R_1} = \frac{v_2 - v_0}{R_f}$$

However, $v_1 = v_2 = 0$

$$\frac{v_{in}}{R_1} = -\frac{v_0}{R_f}$$

$$\therefore \quad \boxed{A_f = \frac{v_0}{v_{in}} = -R_f/R_1}$$

→ 2M

AC Amplifier

1) Here opamp responds wrt changes in ac input.

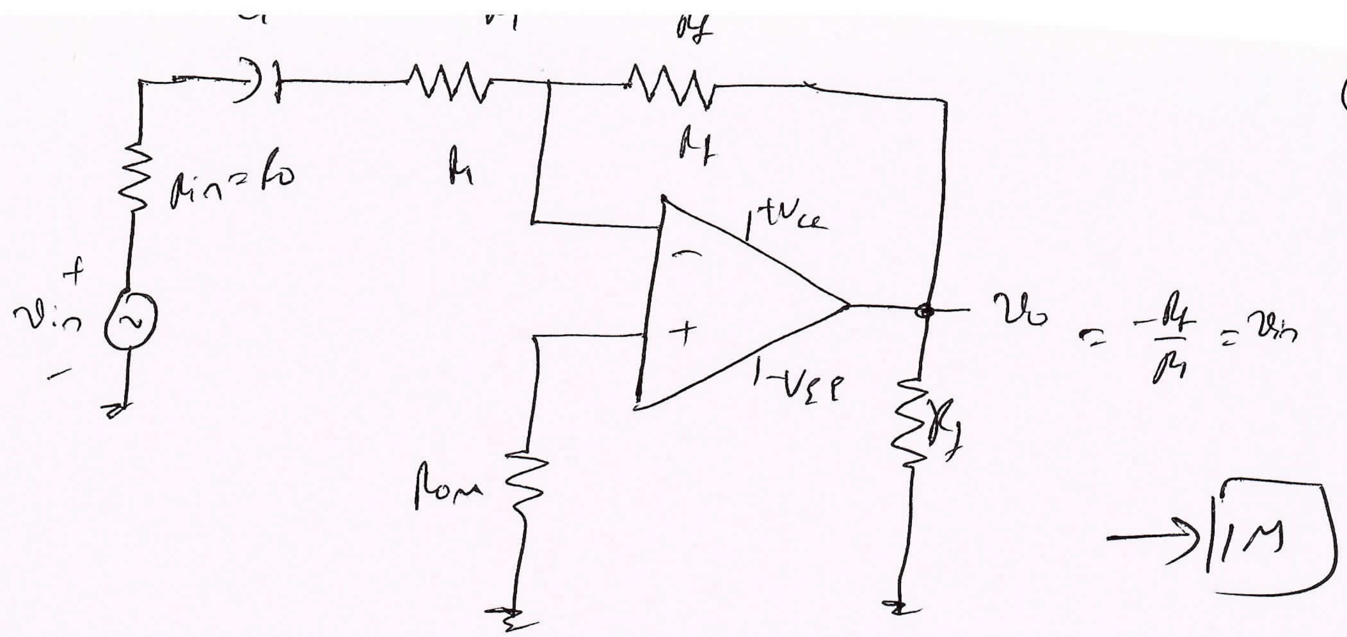
2) If ac input is having some dc level, then it is necessary to use a coupling capacitor.

3) The coupling capacitor not only blocks dc voltage but also pass AC & sets lower cutoff frequency

$$\boxed{f_L = \frac{1}{2\pi C_i (R_{if} + R_o)}}$$

→ 1M

Bhw. M.



→ 1M

4) The higher cutoff frequency f_H depends on closed loop voltage gain of amplr.

5) The B.W is given by $\boxed{BW = f_H - f_L}$

6) To minimize effect of o/p offset voltage ~~produced~~ produced by resistor R_{OM} or o/p coupling capacitor "C_o" may ~~be~~ used.

→ 1M

7c) $R_1 = 1K\Omega$, $R_f = 10K\Omega$, $A = 200000$, $R_i = 2M\Omega$, $R_o = 75\Omega$

$f_o = 5Hz$.

$\boxed{\text{Total Marks} = 6M}$

$$A_f = 1 + \frac{R_f}{R_1} = 11$$ → 2M

$$R_{if} = R_i(1 + A\beta) = 2M\left(1 + 200000 \times \frac{1}{11}\right) = 3.63 \times 10^{10}\Omega.$$ → 2M

$$R_{of} = R_o/(1 + A\beta) = 4.12 m\Omega$$ → 1M

$f_f =$

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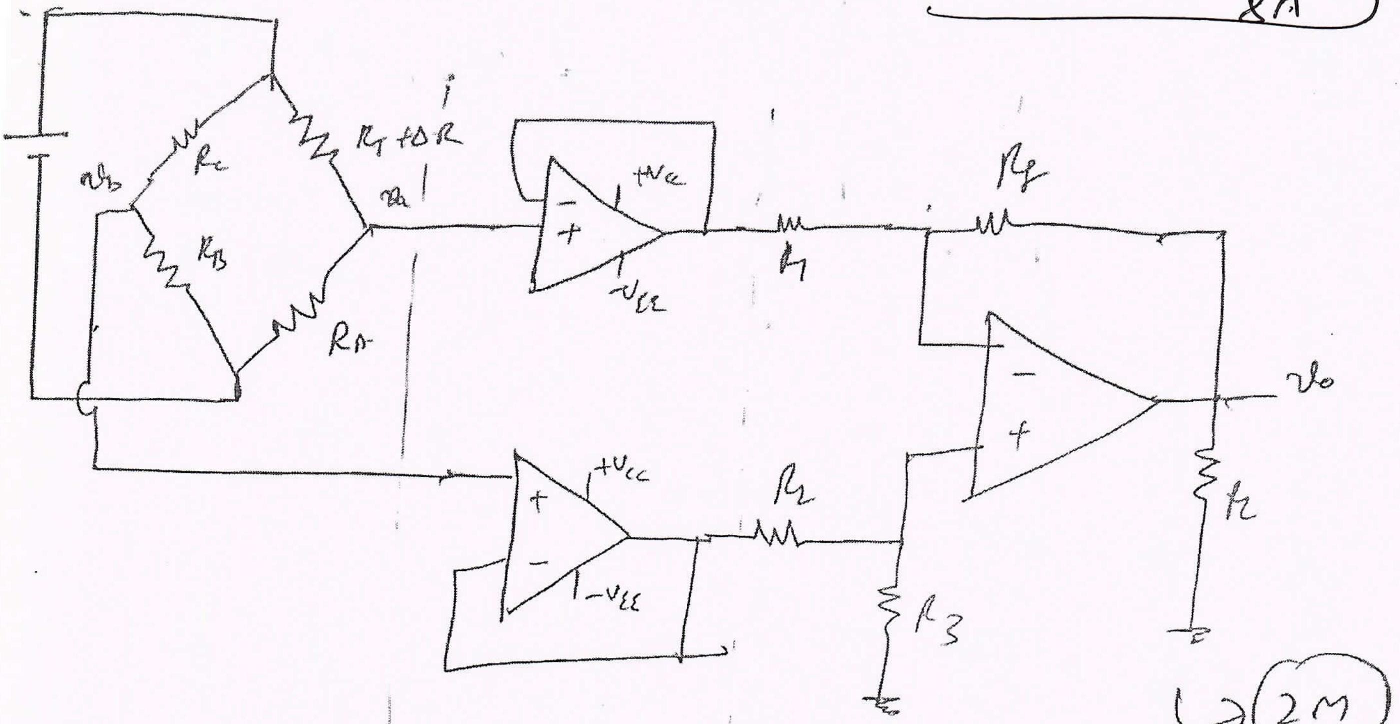
→ 1M

8a) IA

Instrumentation Amplifier

Total Marks (3) 3 / 8A

(40)



↳ (2M)

⇐ Differential i/p, o/p ampls ⇒ ⇐ Difference Amplifier ⇒

1) The above figure shows instrumentation amplifier using transducer bridge.

2) Here a resistive transducer (strain gauge) is used whose resistance changes as function of physical energy applied

3) It is denoted by $R_T + \Delta R$, where R_T is resistance of transducer & ΔR is change in resistance of R_T.

4) At bridge balance

Expln → (1M)

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(38) (41)

$$\left.\begin{aligned} V_a &= V_b \\ \frac{V_{dc} \times R_B}{R_C + R_B} &= \frac{V_{dc} \times R_A}{R_A + R_T} \end{aligned}\right\} \rightarrow (2M)$$

5) Generally R_A, R_B, R_C & R_T chosen to be same at some reference temperature.

6) The bridge is balanced at reference condn., as the physical qty to be measured changes, the resistance of transducer also be changed, causing bridge imbalance

$V_a \neq V_b$

7) Let ΔR be change in resistance of transducer, Since R_B & R_C are fixed resistor, the voltage V_b is constant However V_a varies as the function of transducer resistor

$$\left.\begin{aligned} V_a &= \frac{V_{dc} \times R_A}{R_A + (R_T + \Delta R)} \rightarrow (2) \\ V_b &= \frac{V_{dc} \times R_B}{R_C + R_B} \rightarrow (3) \\ V_{ab} &= V_a - V_b \\ &= \frac{V_{dc} \times R_A}{R_A + (R_T + \Delta R)} - \frac{V_{dc} \times R_B}{R_C + R_B} \end{aligned}\right\} (2M)$$

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$$V_{ab} = \frac{-V_{dc} \times \Delta R}{2(2R + \Delta R)} \rightarrow (5)$$

39 37

Here -ve sign indicates Va is less than Vb.

The o/p voltage Vab is applied to differential i/p, o/p amplifier consisting of 3 opamp.

$$\therefore \boxed{V_o = \frac{R_f}{R_1}\left(\frac{V_{dc} \times \Delta R}{2(2R + \Delta R)}\right)} \rightarrow (6)$$

Since $2R >> \Delta R \Rightarrow 2R + \Delta R \approx 2R$

$$\boxed{V_o = \frac{R_f}{R_1} \frac{\Delta R}{4R} V_{dc}}$$

1M

$V_o \propto \Delta R$

Rahul .cm

8b)

Basic Comparator

Total Marks 6M

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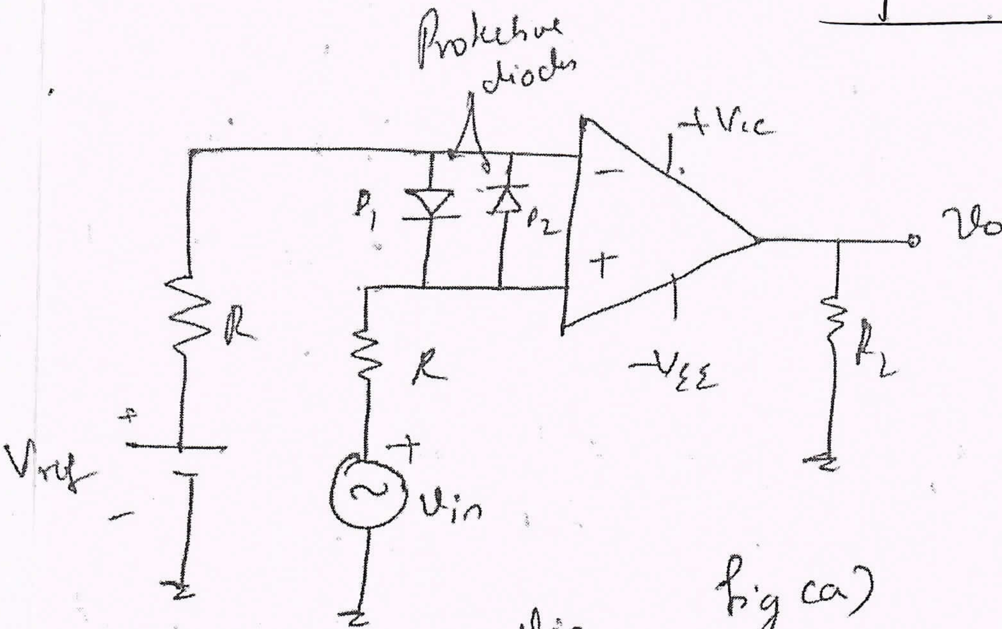
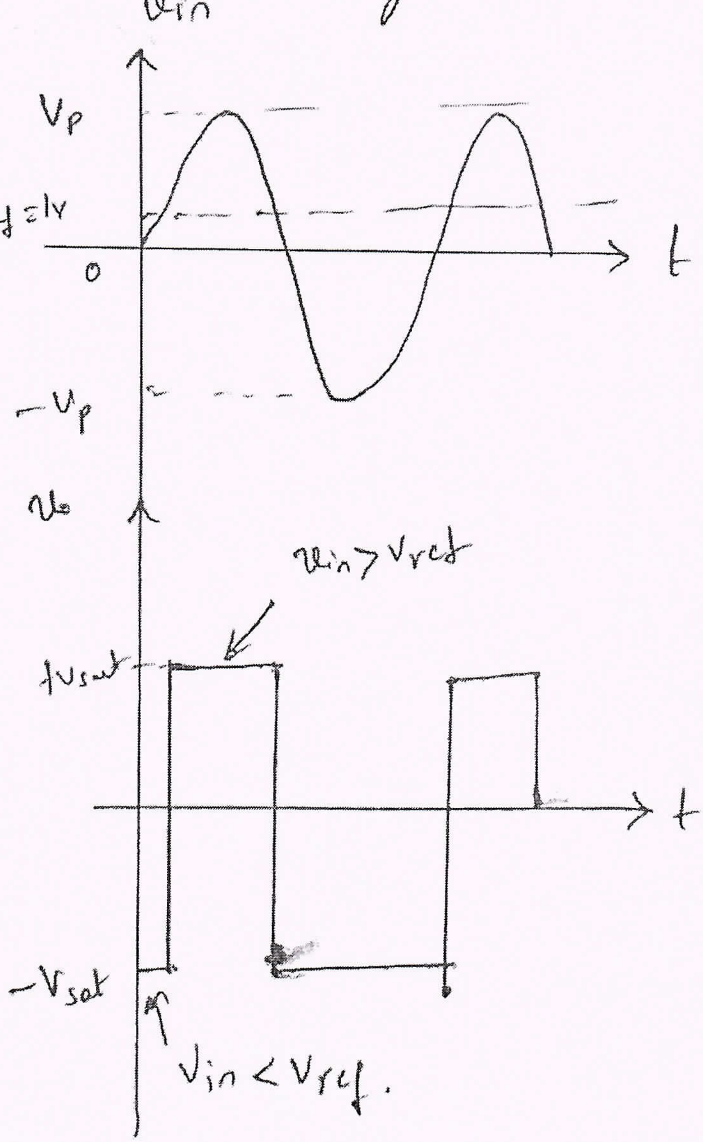


fig (a)



→ 2M

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1) The figure shows op-Amp used as a Comparator.

2) A fixed reference voltage $V_{ref} = 1V$ is applied to (−) input ; & other input is applied with V_{in}

3) Whenever $V_{in} < V_{ref}$ inverting input terminal will be at higher potential & hence O/P of the op-Amp goes to negative Satn. $-V_{sat}$

4) Whenever $V_{in} > V_{ref}$ non-inverting input will be at higher +ve potential & hence O/P of the op-Amp will be at +ve Satn. $+V_{sat}$.

5) The ckt is also known as ~~sigre~~ Sine wave to square wave converter.

Expln :- $\boxed{2M}$

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Zero Crossing detector

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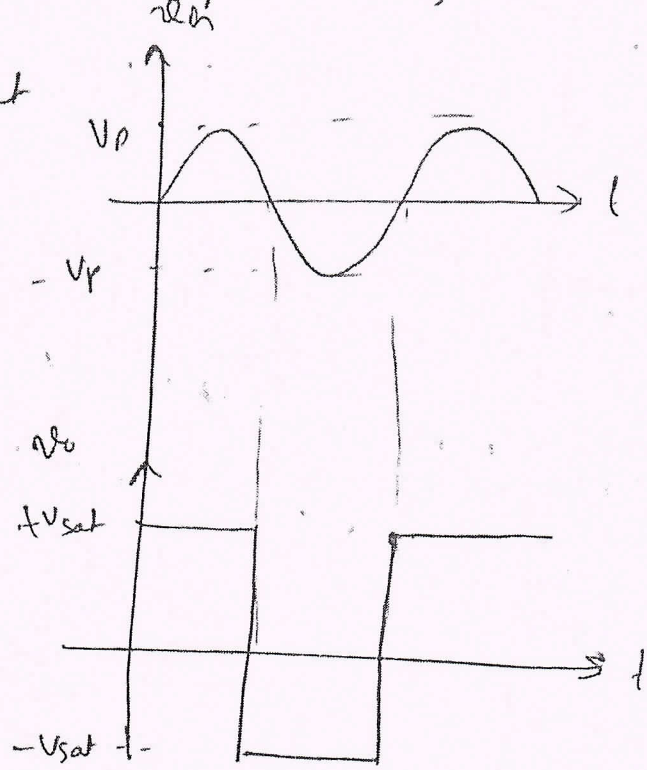
1) In basic Comparator {fig a} if $V_{ref} = 0$ then it will act as zero crossing detector

2) When $V_{in} < 0$, inverting terminal will be at higher +ve potential, hence o/p will be $-V_{sat}$

2) When $V_{in} > 0$, non-inverting terminal will be at higher +ve potential, hence o/p will be $+V_{sat}$

→ [2M]



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8c) $R_1 = 150\Omega$, $R_2 = 68K\Omega$, $V_{in} = 500mV_{p-p}$. $\boxed{\text{Total Marks } 6M}$

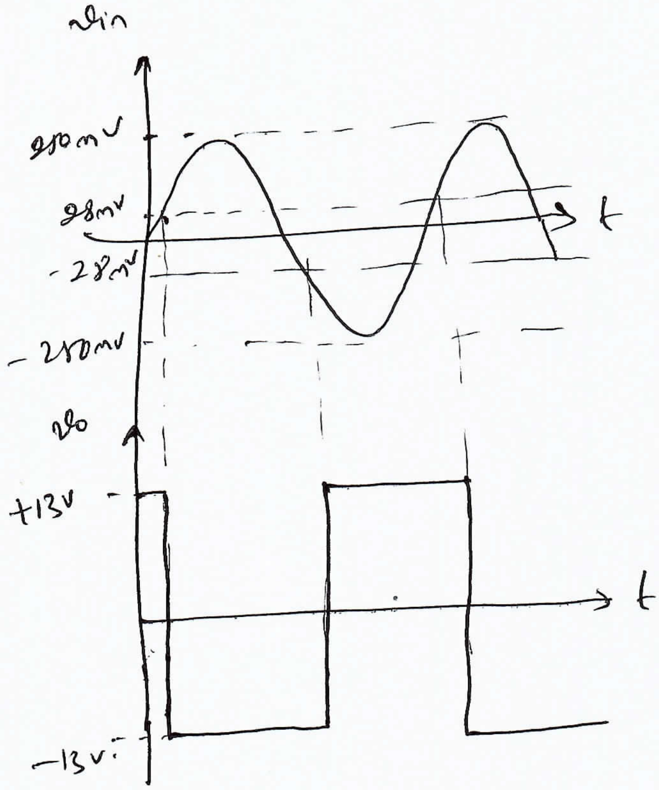
46

$$V_{ut} = V_{sat}\left(\frac{R_1}{R_1+R_2}\right) = 13\left(\frac{150}{68K+150}\right) = 28mV.$$

$\hookrightarrow \boxed{2M}$

$$V_{lt} = -V_{sat}\left(\frac{R_1}{R_1+R_2}\right) = -13\left(\frac{150}{68K+150}\right) = -28mV.$$

$\hookrightarrow \boxed{2M}$



$\rightarrow \boxed{2M}$

$\boxed{\text{Total Marks : 8M}}$

9a) R-2R DAC

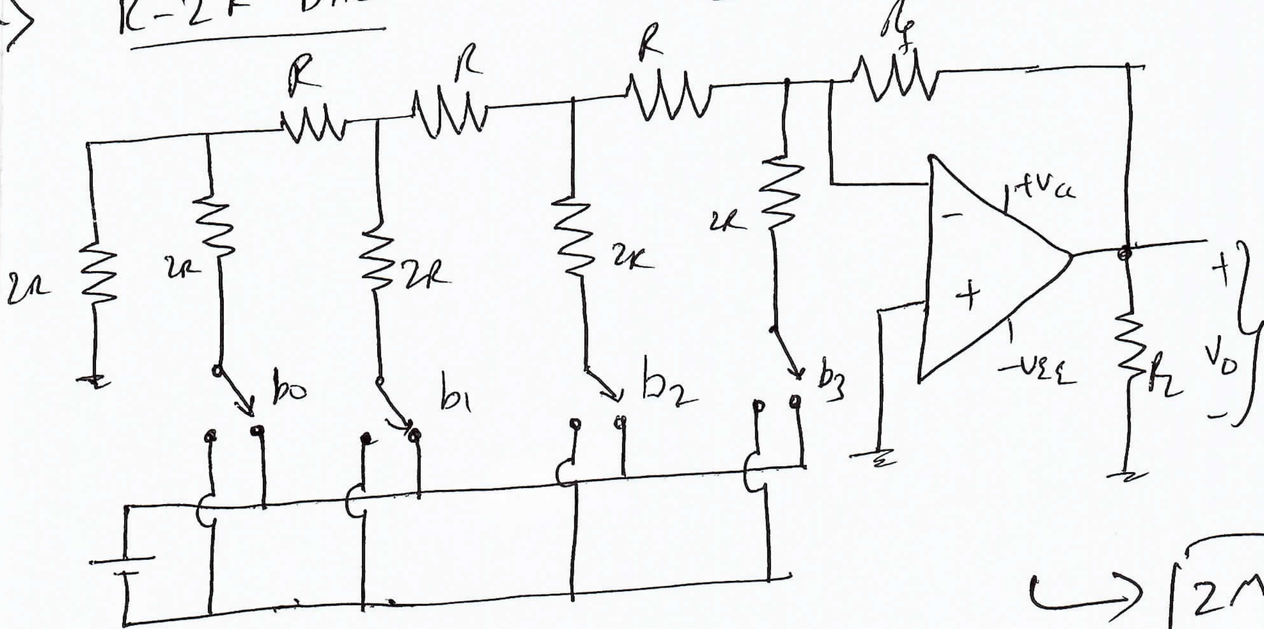


Fig (a)

$\hookrightarrow \boxed{2M}$

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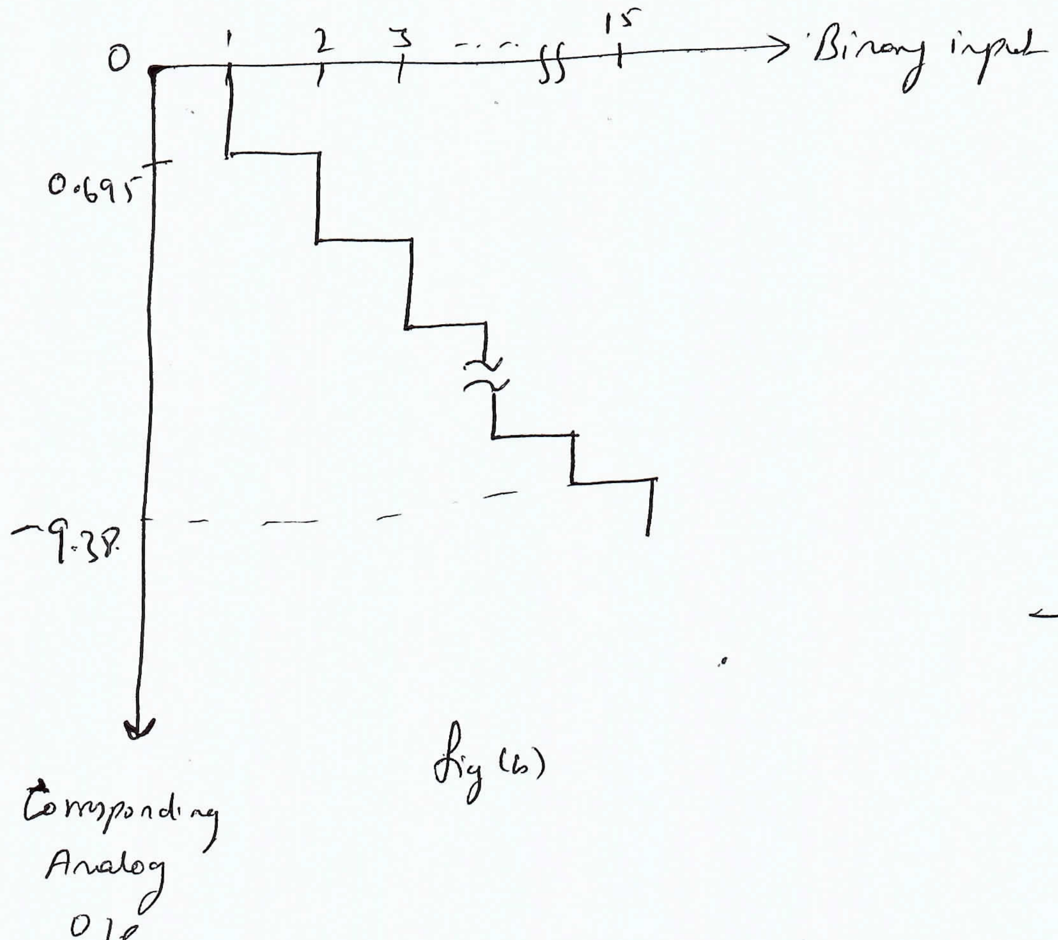

fig (b)

→ 1M

1) figure (a) shows Digital to Analog converter with R-2R register.

2) Binary inputs are given via switches b_0 to b_3.

3) output voltage is directly proportional to binary input

4) The problem with binary weighted DAC is overcome by using R-2R DAC, since it uses only 2 resistors.

5) If MSB bit is connected to 5V & other switch are connected to ground, we get thevenin equivalent resistor given by

$$R_{th} = (((2R \parallel 2R) + R) \parallel 2R) + R) \parallel 2R = 2R$$

48

$$\cancel{V_0 = -V_{ref}\frac{R_f}{R}\left[\frac{b_3}{R} + \frac{b_2}{4}\right.}$$

$$\boxed{V_0 = -\frac{R_f}{R} V_{ref}\left[\frac{B_0}{16} + \frac{B_1}{8} + \frac{B_2}{4} + \frac{B_3}{2}\right]} \rightarrow \boxed{2M}$$

Given $R = 10K\Omega$, $R_f = 20K\Omega$, $V_{ref} = 5V$.

$$V_0 = -\frac{\overset{2}{\cancel{20K}}}{\cancel{10K}} \times 5\left[\frac{5}{16} + \frac{5}{8} + 0 + 0\right]$$

$$= -10\left[\frac{40+80}{128}\right] \rightarrow \boxed{1M}$$

$$V_0 = -0.976V$$

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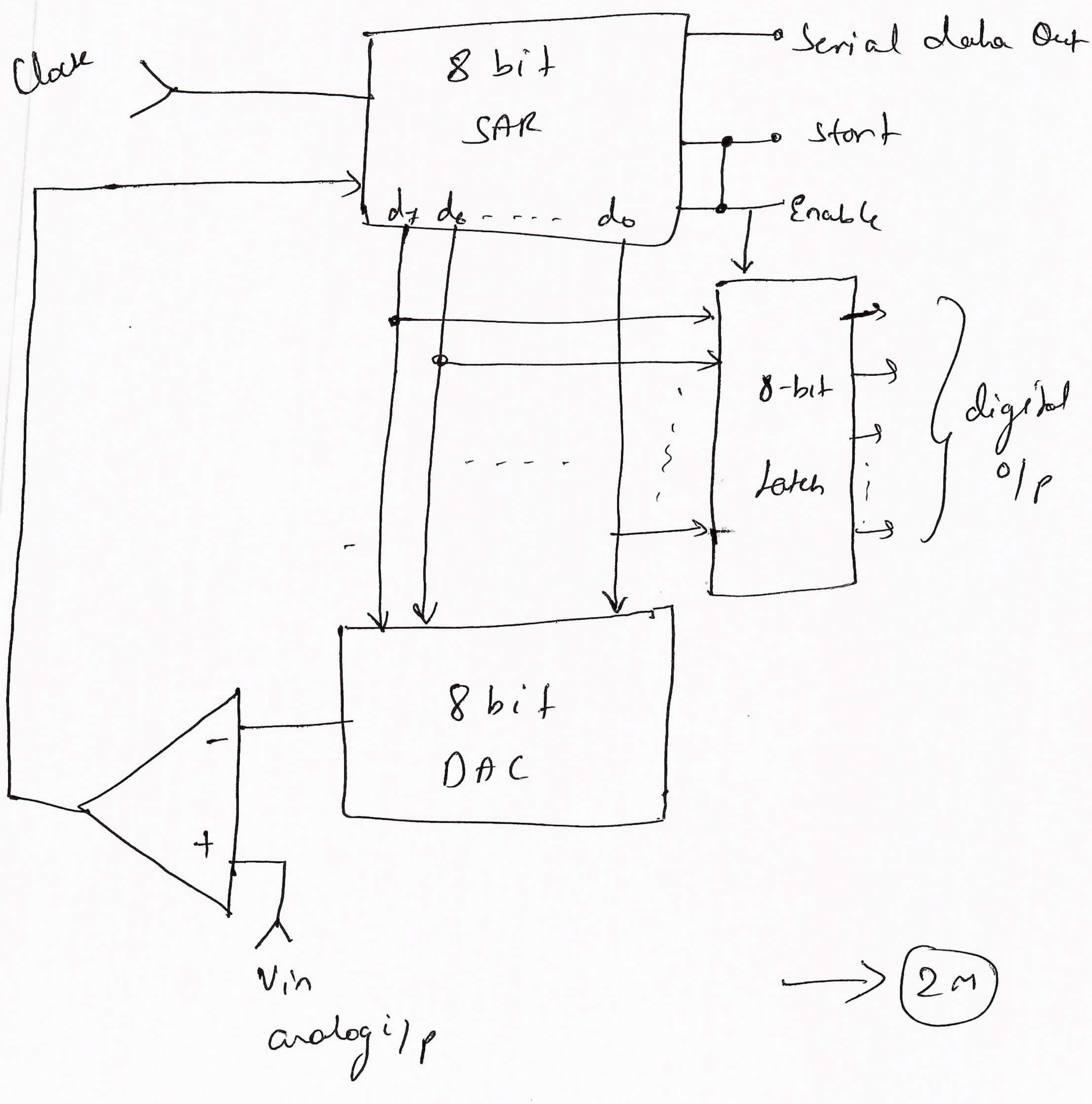
9b) Successive Approximation type ADC (4)

[6M]



→ (2M)

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V_{in}	SAR (o/p) Q_7 Q_6 Q_5 Q_4 Q_3 Q_2 Q_1 Q_0	Comparator o/p
4.156	1 0 0 0 0 0 0 0	1
(212)	1 1 0 0 0 0 0 0	1
	1 1 1 0 0 0 0 0	0
	1 1 0 1 0 0 0 0	1
	1 1 0 1 1 0 0 0	0
	1 1 0 1 0 1 0 0	1
	1 1 0 1 0 1 1 0	0 → 2M
	1 1 0 1 0 1 0 1	0
	1 1 0 1 0 1 0 0	0

1) The heart of circuit is 8 bit SAR
2) The o/p of SAR is given to 8-bit DAC.
3) The o/p of 8-bit DAC is compared with analog i/p V_{in}
4) Initially Start = 1 & SAR is reset
On first clock pulse, MSB is set & all other bits are reset

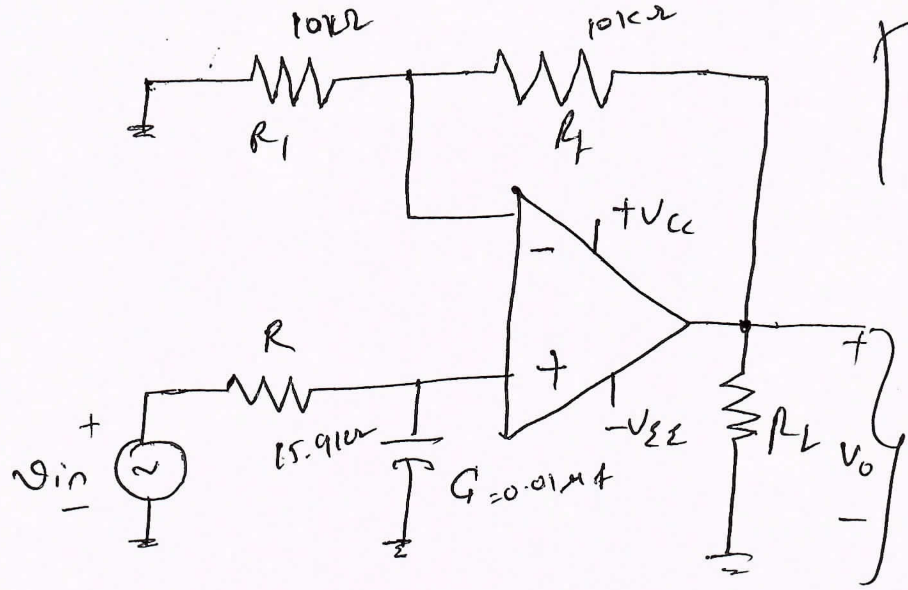
$\boxed{\text{Expln} - 2M.}$

i) If $V_{in} < V_a$. O/P of comparator is zero
SAR resets current bit & sets next bit.
ii) If $V_{in} > V_a$ O/P of comparator is 1
SAR retains current bit & sets next bit.

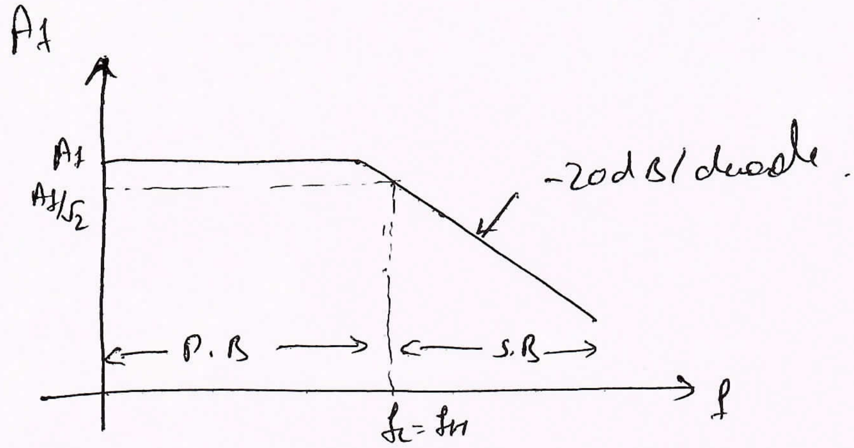
2M

9c) First Order low pass Butterworth filter :- (57)

Total Marks 6M



→ 2M



→ 1M

① $f_H = 1KHz$

② Let $C_1 = 0.01\mu F$

$$R = \frac{1}{2\pi f_H C_1} = 15.91k\Omega$$ → 1M

$$A_f = 1 + \frac{R_f}{R_1} = 2$$

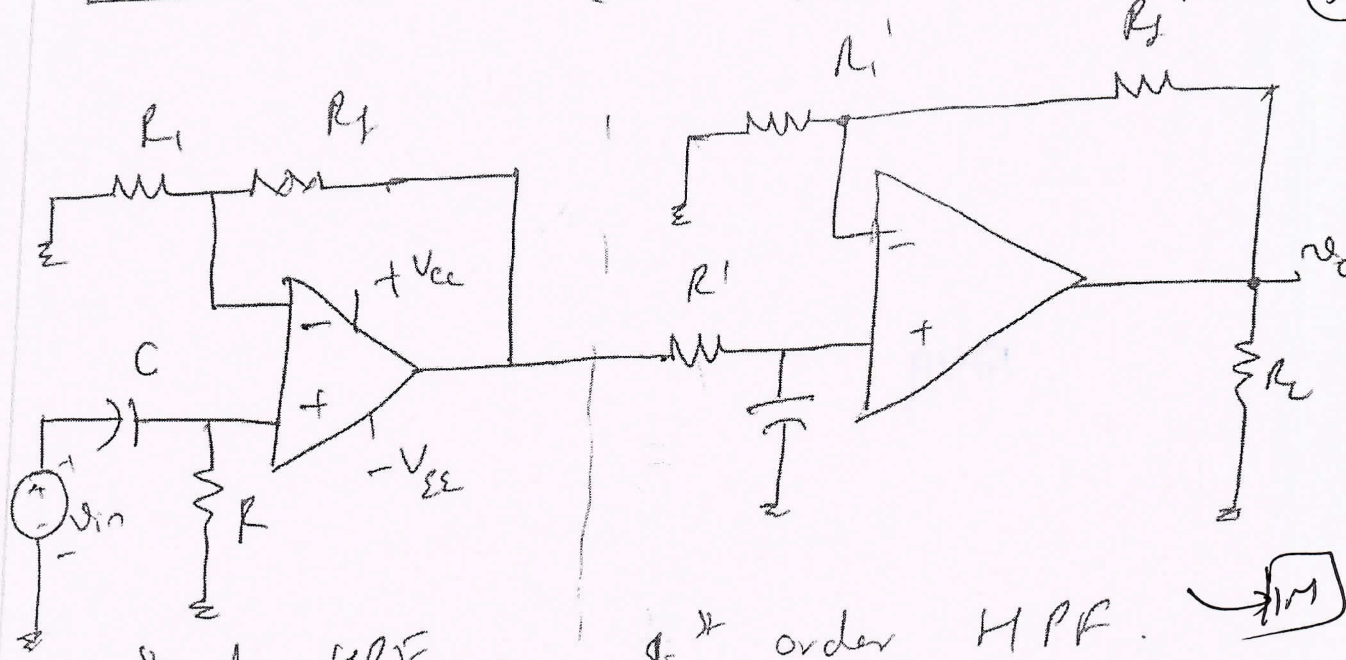
$\therefore R_f = R_1$

Let $R_f = R_1 = 10k\Omega$ → 2M

10a) Wide Band Pass filter

Total Marks :- 6M

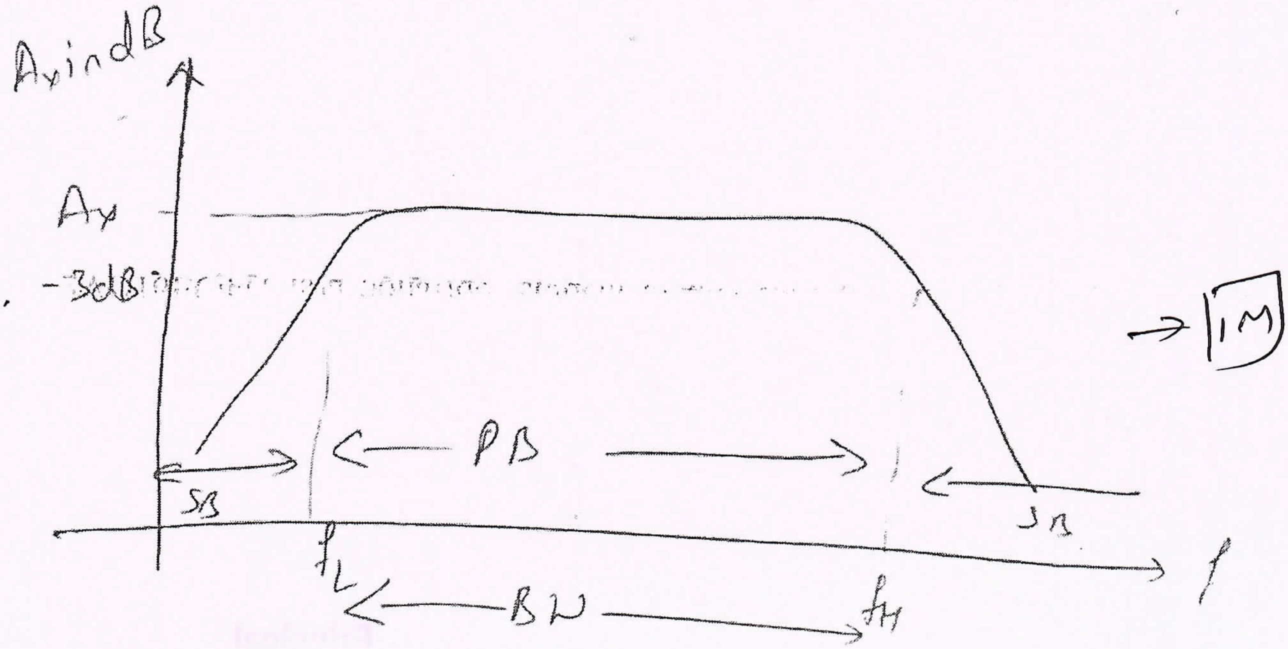
52



1st order HPF

1st order HPF

1M



1M

1) Wide band pass filter is obtained by cascading HPF & LPF

2) To obtain ± 20dB/decade band pass filter,

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first order HPF & first order LPF are cascaded.

3) To obtain ± 40 dB/decade BPF second order HPF & LPF are cascaded

4) Here f_H should be larger than f_L ($f_H > f_L$)

$$\left|\frac{V_o}{V_{in}}\right| = \frac{A_f\,(f/f_L)}{\sqrt{\left(1+(f/f_L)^2\right)\left(1+(f/f_H)^2\right)}} \rightarrow \boxed{1M}$$

5) Since pass band gain of HBPF is 4. The pass band gain of LPF & HPF will be '2'.

Expln :- 2M

Q - factor

1) It is called "figure of merit" for filter ckt.

2) It defines selectivity of filter is passing. Centre freq & rejecting other frequency.

$$\boxed{Q = \frac{f_c}{BW}} \longrightarrow \boxed{1M}$$

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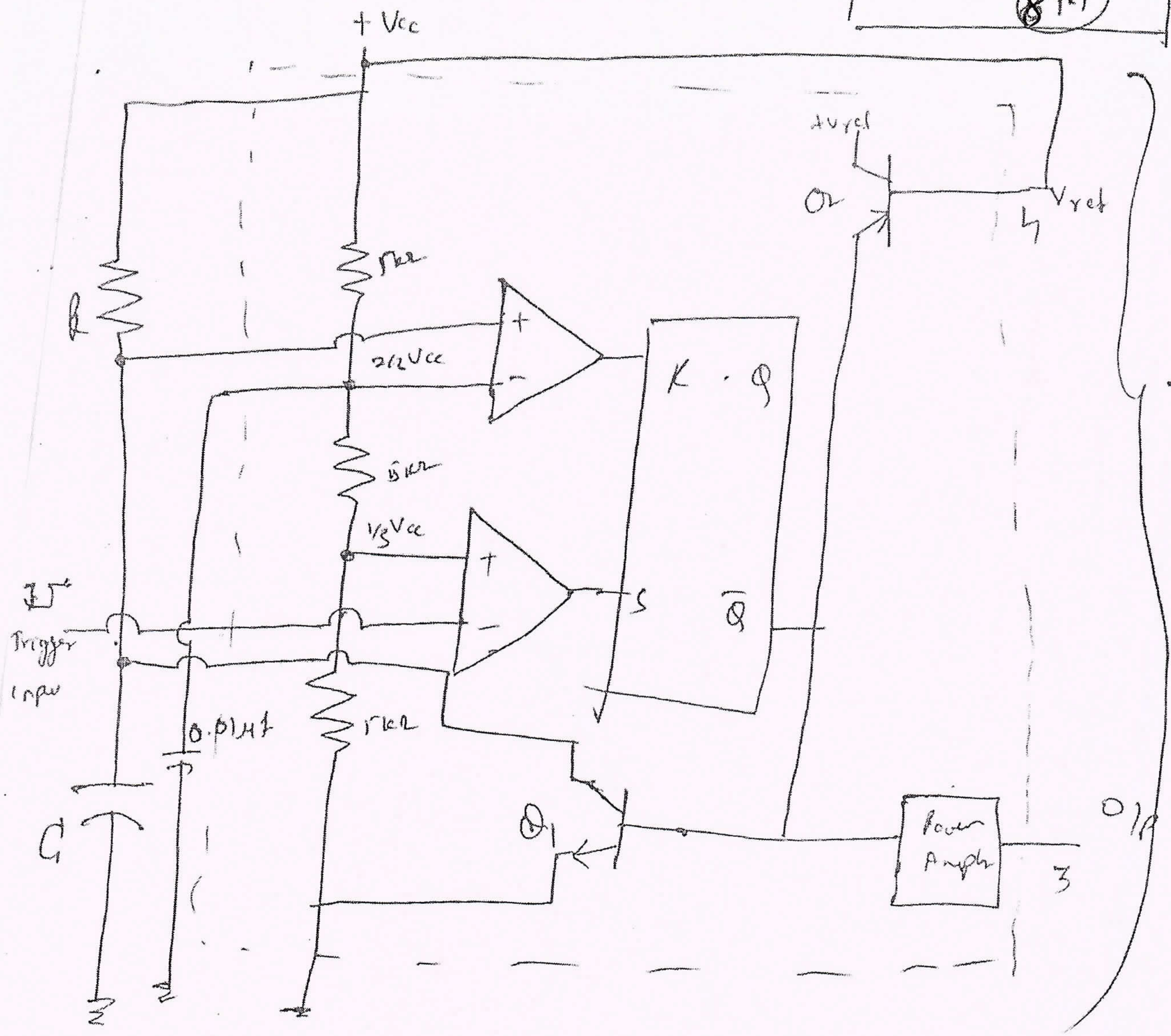
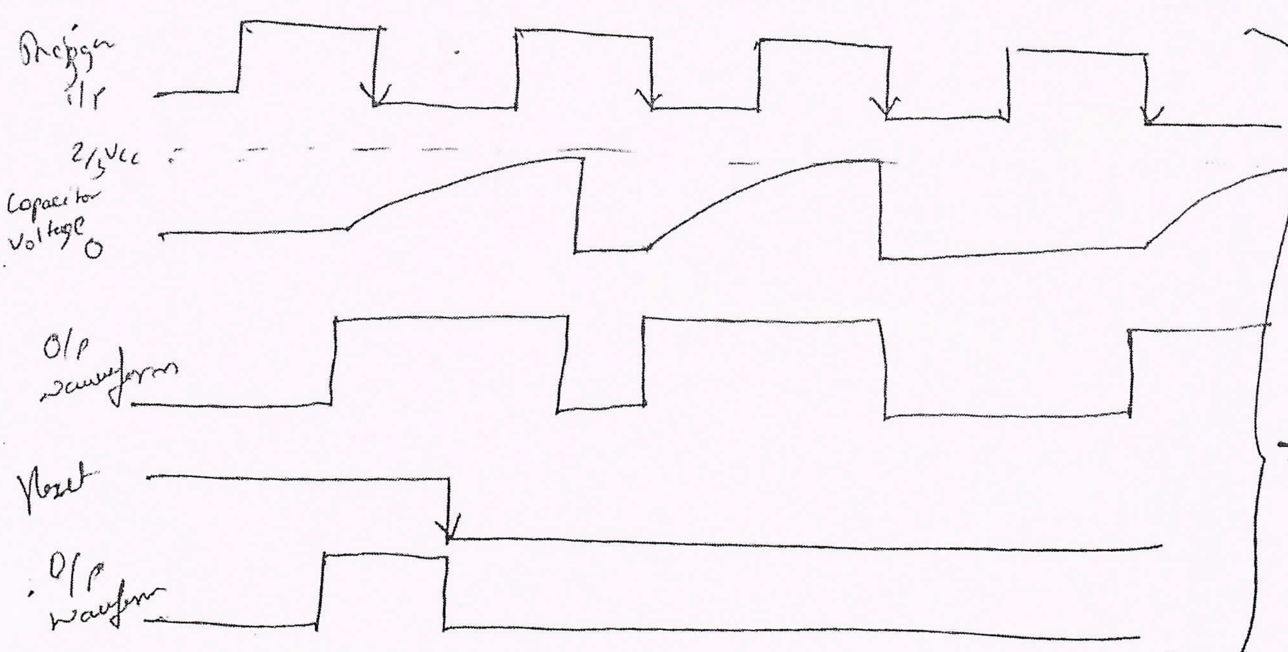
10 b) <u>Monostable</u> <u>Multivibrator</u>

Total Marks (8M)



→ 3M



→ 2M

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1) In standby state $\bar{Q}$ o/p of SR flip-flop = 1. (50)

2) Hence o/p of 555 timer = 0, discharge transistor Q_1 is ON, clamping capacitor to ground.

3) When trigger voltage is applied & Trigger vol. goes below $\frac{1}{3}V_{cc}$, the o/p of L.C. is S=1, Q=1. $\therefore \bar{Q} = 0$. The o/p of timer is (high).

4) The discharge transistor Q_1 is off, Now capacitor starts charging towards V_{cc} via R.

5) The capacitor continues to charge until voltage of capacitor exceeds $\frac{2}{3}V_{cc}$, When vol. across capacitor goes above $\frac{2}{3}V_{cc}$, R=1, $\bar{Q}$=1 & o/p of 555 timer goes low.

6) The discharge transistor Q_1 is ON, clamping external timing capacitor C to gnd.

7) From the waveform we can see that once triggered the o/p remain in high state until time T_p

8) When reset goes low, o/p is forced to zero since ~~discharge~~ reset transistor is turned ON.

$\boxed{\text{Explaination} \rightarrow 3M}$

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10 c)

$R_A = 2.2k\Omega$, $R_B = 3.9k\Omega$, $C = 0.1\mu f$.

Total Marks 6M

$t_c = 0.693(R_A + R_B)C$

$= 0.42\ msec$ // → 2M

$t_d = 0.693\ R_B C$

$= 0.693 \times 3.9k\Omega \times 0.1\mu f.$

$= 0.27\ msec$ // → 2M

$f = \dfrac{1}{t_c + t_d} = 1.44\ kHz$ // → 2M

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