IES 2019 MAINS TEST SERIES

ELECTRONICS ENGINEERING

ACE ACADEMY- 2019

ESE MAINS TEST SERIES

1-10 SUBJECTWISE TEST

11-14 MOCK TEST

TEST CODE: 602

ACE
Engineering Academy

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ESE – 2019 Mains Test Series

Electronics & Telecommunication Engineering

Test – 2 on Control Systems + Advanced Electronics (Paper – II)

Time Allowed: 3 Hours **Maximum Marks: 300**

INSTRUCTIONS

Please read each of the following instructions carefully before attempting questions:

There are EIGHT questions divided in TWO sections.

Candidate has to attempt FIVE questions in all.

Questions no. 1 and 5 are **compulsory** *and out of the remaining, THREE are to be attempted choosing at least ONE question from each section.*

The number of marks carried by a question / part is indicated against it.

Answers must be written in the medium authorized in the Admission Certificate which must be stated clearly on the cover of this Question-cum-Answer (QCA) Booklet in the space provided. No marks will be given for answers written in medium other than the authorized one.

Assume suitable data, if considered necessary and indicate the same clearly..

Unless otherwise mentioned, symbols and notations carry their usual standard meanings.

Attempts of questions shall be counted in sequential order. Unless struck off, attempt of a question shall be counted even if attempted partly. Any page or portion of the page left blank in the Question-cum-Answer Booklet must be clearly struck off.

SECTION-A

01.

(a) (i) Explain the terms Gain margin and Phase margin?

(ii) The open loop transfer function of a unity feedback control system is given by

$$G(s) = \frac{K}{s(1+0.1s)(1+s)} \quad (K > 1)$$

Determine the value of K so that the resonance peak M_r of the system is equal to 1.4

(6 +6)M

(b) For the open-loop transfer function $G(s)H(s) = \frac{K}{s(s+4)(s+5)}$, determine the following:

(i) Point of intersection of asymptotes with real axis

(ii) Point of intersection of root locus with imaginary axis and the value of K at this point.

(12 M)

(c) Use the properties of the unit impulse function δ(t) to evaluate the following integrals:

(i) $\int_{-\infty}^{\infty} (t^3 - \cos \pi t)\delta(t+1)dt$

(ii) $\int_{-\infty}^{\infty} [e^{-\pi t} + \sin(10\pi t)]\delta(2t+1)dt$

(6 + 6)M

(d) Calculate the 4-point DFT of $x(n) = \{0,1,2,3\}$ using

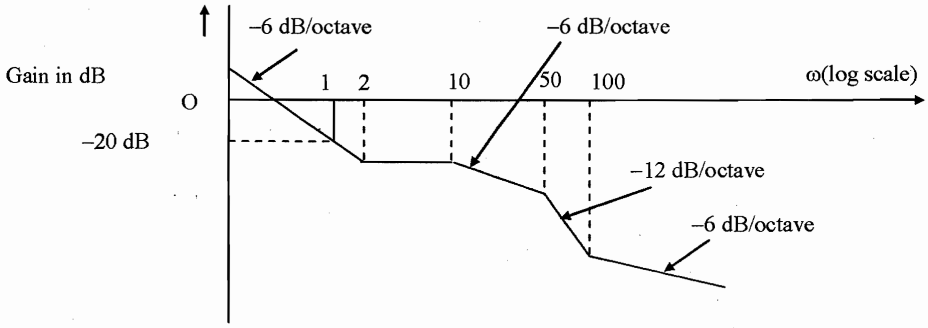
(i) DIT-FFT algorithm

(ii) DIF-FFT algorithm **(6 + 6)M**

(e) What do you mean by design flow in VLSI technology, draw the flow chart of VLSI design flow? And also draw the design iteration loop. **(12 M)**

02.

(a) The asymptotic approximation of the variation of gain with frequency of a control system (Bode plot) is shown below. Determine the transfer function of the system. **(20 M)**



(b) The eight point DFT of the sequence

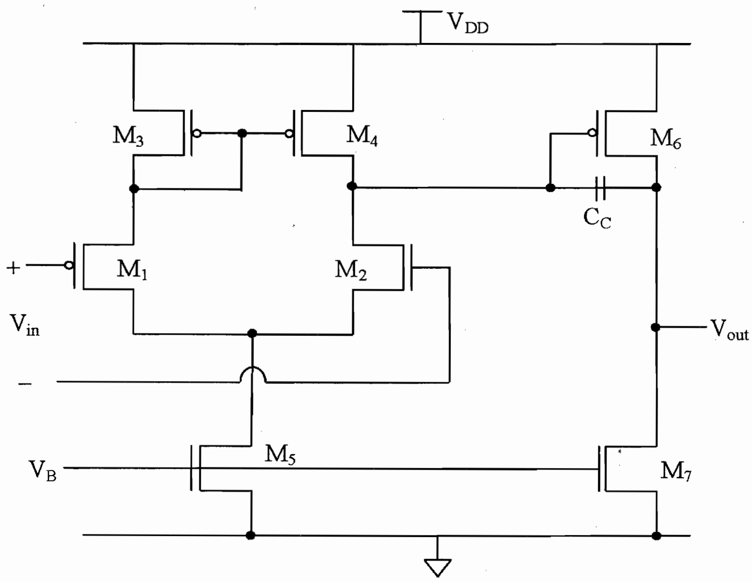
$$x(n) = \begin{cases} 1 & 0 \le n \le 3 \\ 0 & 4 \le n \le 7 \end{cases} \quad \text{is } X(K)$$

Compute the DFT of $x_1(n) = \begin{cases} 1 & n = 0 \\ 0 & 1 \le n \le 4 \\ 1 & 5 \le n \le 7 \end{cases}$ and

$$x_2(n) = \begin{cases} 0 & 0 \le n \le 1 \\ 1 & 2 \le n \le 5 \\ 0 & 6 \le n \le 7 \end{cases} \quad \text{using } X(k).$$

(10 + 10)M

(c) A simple two stage op-amp using CMOS shown in below figure need to be fabricated using VLSI technology.



Determine the aspect ratio of each MOSFET by considering the following specifications.

$K_n^1 = 120\mu A/V^2$ $V_{tn} = 0.4$ V Gain A = 80 dB SR = 20V/μs

$K_p^1 = 60\mu A/V^2$ $V_{tp} = -0.4$ V GBW = 50 MHz $\lambda = 0.05\ V^{-1}$

L = 1 μm **(20 M)**

03.

(a) A unity feedback control system has an open loop transfer function, $G(s) = \dfrac{10}{s(s+2)}$. Find the rise time, percentage overshoot, peak time and settling time for a step input of 12 units.

(15 M)

(b) The output c(t) of a control system is related to its input by

$[s^4 + 2s^3 + 2s^2 + (K+3)s + K]\ C(s) = K(s+1)R(s)$,

Where 'K' is positive gain of an amplifier.

(i) With K = 6, will the output response be stable?

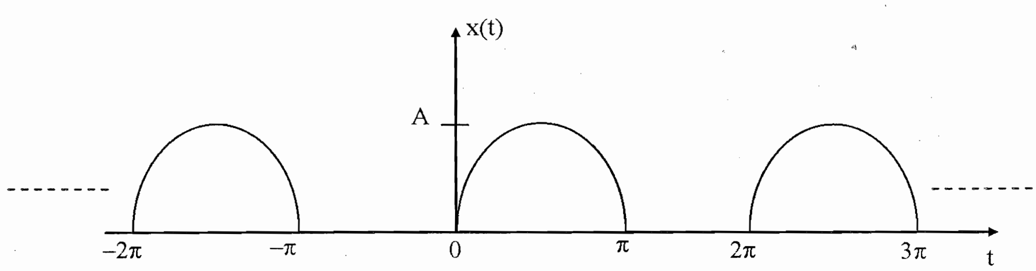
(ii) Determine the positive limiting values of K for stability

(15 M)

(c) State and Prove Initial and Final value theorems of Laplace Transform

(15 M)

(d) Find the Trigonometric Fourier series expansion of the half wave rectified sine wave shown in figure.



(15 M)

04.

(a) **(i)** What do you mean by Root-locus of a system

(ii) The open-loop transfer function of a unity feedback control system is given by $G(s) = \dfrac{K(s+4)}{s(s+1)}$

(A) Sketch the root locus diagram.

(B) Determine the range of value of the gain 'K' so that the system is Under damped, Critically damped and Over damped.

(20 M)

(b) **(i)** Consider a unity feedback system having transfer function $\dfrac{C(s)}{R(s)} = \dfrac{1}{1+s^2(s+a)(s+b)}$

Determine type of system and the steady state errors for inputs $2u(t)$, $4t^2u(t)$ and $tu(t)$

(ii) The maximum overshoot for a unity feedback control system having its forward path transfer function as $G(s) = \dfrac{K}{s(sT+1)}$ is to be reduced from 60% to 20%. The system input is a unit step function. Determine the factor by which K should be reduced to achieve afore said reduction.

(10+10)M

(c) Find the Fourier transform of the following:

(i) Gaussian signal $x(t) = e^{-at^2}$

(ii) Gaussian modulated signal $x(t) = e^{-at^2}\cos\omega_c t$

(10 + 10)M

SECTION-B

05.

(a) Design a PD controller so that the system having open loop transfer function $G(s)H(s) = \frac{1}{s(s+1)}$ will have a phase margin of 40° at 2 rad/sec.

(10 M)

(b) Obtain an expression for sensitivity of closed loop transfer function having an open loop gain 'G', with respect to changes in feedback factor 'H'. An amplifier with open loop voltage gain $A_0 = 500 \pm 10$ is available. It is required to have an amplifier whose voltage gain varies by no more than ± 0.2 percent. Find the value of the feedback factor required.

(10 M)

(c) For the open-loop transfer function $G(s)H(s) = \frac{K(s-3)(s-5)}{(s+1)(s+2)}$

Find out the break-away and break-in points, if any, for the root locus. Also specify whether the gain is maximum or minimum at these points.

(10 M)

(d) State and Prove Duality (Symmetry) Property of Fourier transform.

(10 M)

(e) Consider a stable LTI system that is characterised by the differential equation.

$$\frac{d^2y(t)}{dt^2} + 4\frac{dy(t)}{dt} + 3y(t) = \frac{dx(t)}{dt} + 2x(t)$$

Find its response for input $x(t) = e^{-t}\,u(t)$ **(10 M)**

(f) Explain in detail about the "Exposure technique" of photolithography in VLSI technology with suitable diagrams?

(10 M)

06.

(a) **(i)** Determine the impulse response of the following transfer functions and comment on their stability by drawing the poles in the s-plane

(A) $\dfrac{2}{s(3s+2)}$ **(B)** $\dfrac{6s}{(s^2+9)^2}$ **(12 M)**

(ii) Explain with proper transfer function a standard PID controller. Explain why derivative term is not employed alone.

(iii) **(A)** Enumerate the salient features of phase lead and phase lag networks

(B) Identify whether given statement is correct or incorrect. justify your answer.

"Phase lead compensation decreases the bandwidth of a system "

(4 + 4)M

(b) **(i)** Explain Nyquist stability criteria

(ii) Using the Nyquist stability criteria investigate the stability of a system whose OLTF is

$$G(s) = \frac{K(s+1)}{(s+0.5)(s-2)} \text{ for } K = 1.25$$

Also find the limiting value of K for stability

(5 + 15) M

(c) Find whether the signal

$$x(t) = \begin{cases} t-2 & ; \; -2 \le t \le 0 \\ 2-t & ; \; 0 \le t \le 2 \\ 0 & ; \; \text{otherwise} \end{cases}$$

is energy signal or power signal. Also find the energy and power of the signal.

(10 M)

(d) Determine the causal signal x(n) having the z-transform

$$X(z) = \frac{1}{(1-2z^{-1})(1-z^{-1})^2}$$

(10 M)

07.

(a) **(i)** Design a digital Butterworth filter satisfying the constraints

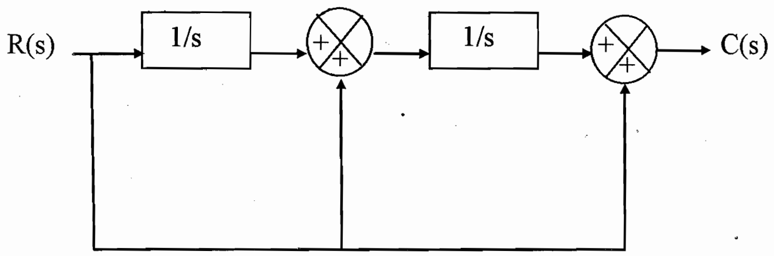
$$0.707 \le \left|H\left(e^{j\omega}\right)\right| \le 1 \quad 0 \le \omega \le \frac{\pi}{2}$$

$$\left|H\left(e^{j\omega}\right)\right| \le 0.2 \qquad \frac{3\pi}{4} \le \omega \le \pi$$

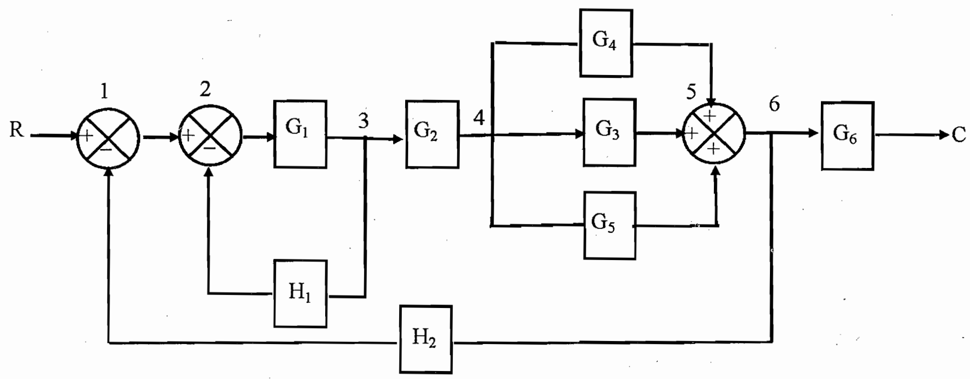
With T = 1sec, using Bilinear Transformation method **(20 M)**

(ii) Write the Comparison between butterworth and chebyshev filters **(5 M)**

(b) **(i)** For the block diagram shown in the fig. given below, obtain $\frac{C(s)}{R(s)}$ using block diagram reduction technique: **(8 M)**



(ii) Obtain signal flow graph representation for a system whose block diagram is given below and using Mason's gain formula, determine the ratio $\frac{C}{R}$ **(12 M)**



(c) **(i)** What are the differences between verification and testing in VLSI technology?

(ii) Differentiate between field programmable gate array (FPGA) and application specific integrated circuit (ASIC).

(7 + 8)M

08.

(a) The impulse response of the circuit is given as $h(t) = e^{-2t}\,u(t)$. This circuit is excited by an input of $x(t) = e^{-4t}\,[u(t) - u(t-2)]$. Determine the output of the circuit using graphical method of convolution **(20 M)**

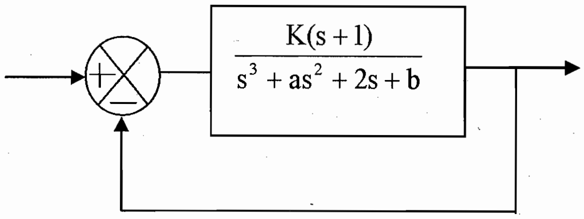
(b) (i) A closed loop control system is to be designed for an under damped response to a unit step. Sketch the desirable range of pole locations for the second order system. The specifications for the system are:

$10\% <$ percent overshoot $< 30\%$

Settling time < 0.4 sec for $\pm 2\%$ tolerance band.

(10 M)

(ii) Find the value of 'K' and obtain a relation between 'a' and 'b' for the following system to oscillate at a frequency of 3 rad/ sec



(10 M)

(c) (i) Draw the architecture of 8051 microcontroller?

(ii) Explain the addressing modes of 8051 microcontroller with examples? **(8 + 12)M**

Detailed solutions for this test will be available on our website on 08-04-2019 at 6.00 pm.
https://www.aceenggacademy.com/ese-2019-mains-offline-test-series/

This test resúlts will be displayed at on our website on 19-04-2019 at 6.00 pm.

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ESE – 2019 MAINS
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TEST – 2
SOLUTIONS

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testseries@aceenggacademy.com | Contact Us : 040 – 48539866 / 040 – 40136222

01. (a)

Sol:

(i) **Gain Margin (GM):** It is the gain which can be varied before the system becomes unstable. If the gain increases GM decreases and if the gain is doubled GM becomes half. GM is calculated at phase cross over frequency

$$GM = \frac{1}{|G(j\omega)H(j\omega)|_{\omega=\omega_{pc}}}$$

$$GM(dB) = 20 \ \log \frac{1}{|G(j\omega_{pc})H(j\omega_{pc})|}$$

Where ω_{pc} = phase cross over frequency, it is the frequency at which phase angle of G(s) H(s) is $-180°$

i.e Arg G(s) H(s) = $-180|_{\omega = \omega pc}$

Phase Margin (PM): It is the phase that can be varied before the system becomes unstable.

PM = $180 + \angle G(j\omega)H(j\omega)|_{\omega=\omega gc}$

The phase angle is measured in clock wise direction where ω_{gc} = Gain cross over frequency, it is the frequency at which magnitude of G(s) H(s) is unity (or) 0dB

$|G(j\omega)H(j\omega)|_{\omega=\omega gc} = 1$

(ii) Given that $G(s) = \frac{K}{s(1+0.1s)(1+s)}, (K > 1)$

Given that resonance peak $M_r = 1.4$

$$M_r = \frac{1}{2\xi\sqrt{1-\xi^2}} \Rightarrow 1.4 = \frac{1}{2\xi\sqrt{1-\xi^2}}$$

$\xi\sqrt{1-\xi^2} = 0.357$

$\Rightarrow \xi^2(1-\xi^2) = 0.127$

$\xi^2 - \xi^4 = 0.127$

$\xi^4 - \xi^2 + 0.127 = 0$

Let $X = \xi^2$, then $x^2 - x + 0.127 = 0$

$x_1 = 0.85 \Rightarrow \xi = 0.921$

$x_2 = 0.149 \Rightarrow \xi = 0.386$

we know that $\xi \le 0.707$

Neglecting the insignificant pole because time constant is very small. The equivalent transfer function is $G(s) = \frac{K}{s(1+s)}$. Hence characteristic equation is

$s(1 + s) + k = 0 \Rightarrow s^2 + s + k = 0$

$\omega_n = \sqrt{K}$ and $2\varepsilon\omega_n = 1$

For $\xi = 0.386$, $2\times 0.386 \times \sqrt{K} = 1$

$\Rightarrow \boxed{K = 1.67}$

01. (b)

Sol:

(i) Open Loop T.F

$$G(s)\,H(s) = \frac{K}{s(s+4)(s+5)}$$

Point of intersection of asymptotes $= \dfrac{\sum \text{poles} - \sum \text{zeros}}{P - Z} = \dfrac{-9}{3} = -3$

(ii) O.L.T.F $G(s)\,H(s) = \dfrac{K}{s(s+4)(s+5)}$

C.E = 1 + G(s) H(s) = 0

s(s + 4) (s + 5) + K = 0

$(s^2 + 4s)(s + 5) + K = 0$

$s^3 + 5s^2 + 4s^2 + 20s + K = 0$

$s^3 + 9s^2 + 20s + K = 0$

s^3	1	20
s^2	9	K
s^1	$\dfrac{9\times 20 - K}{9}$	0
s^0	K	

For marginally stable system the root locus touches to imaginary axis.

9 × 20 = K

K = 180

$9s^2 + 180 = 0$

$\therefore$ Point of intersection = $\pm j\sqrt{20}$

01. (c)

Sol:

(i) $\displaystyle\int_{-\infty}^{\infty}\left(t^3 - \cos \pi t\right)\delta(t+1)\,dt$

From sifting property of impulse signal $\displaystyle\int_{-\infty}^{\infty} f(t)\,\delta(t-a)\,dt = f(a)$

$$\int_{-\infty}^{\infty}\left(t^3 - \cos \pi t\right)\delta(t+1)\,dt = (-1)^3 - \cos\pi(-1)$$

$$= -1 + 1$$

$$= 0$$

(ii) $\displaystyle\int_{-\infty}^{\infty}\left(e^{-\pi t} + \sin 10\pi t\right)\delta(2t+1)\,dt$

From time scaling property $\delta(at+b) = \dfrac{1}{|a|}\delta\left(t + \dfrac{b}{a}\right)$

From sifting property of impulse signal $\displaystyle\int_{-\infty}^{\infty} f(t)\,\delta(t-a)\,dt = f(a)$

$$\int_{-\infty}^{\infty}\left(e^{-\pi t} + \sin 10\pi t\right)\delta(2t+1)\,dt = \frac{1}{2}\int_{-\infty}^{\infty}\left(e^{-\pi t} + \sin 10\pi t\right)\delta\left(t + \frac{1}{2}\right)dt$$

$$= \frac{1}{2}\left[e^{-\pi\left(\frac{-1}{2}\right)} + \sin 10\pi\left(\frac{-1}{2}\right)\right]$$

$$= \frac{1}{2}\left[e^{\frac{\pi}{2}} - \sin 5\pi\right]$$

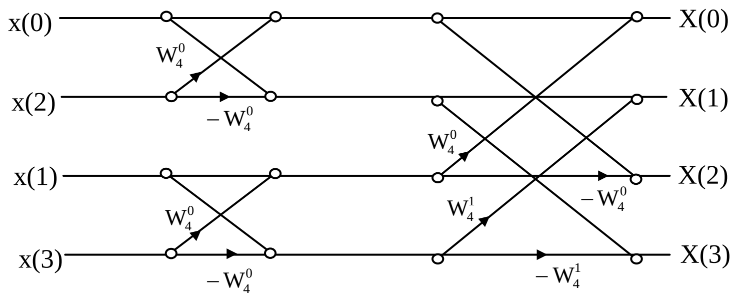
$$= \frac{1}{2}\left[e^{\frac{\pi}{2}}\right]$$

$= 2.403$

01. (d)

Sol:

(i) 4 - point DIT algorithm:-

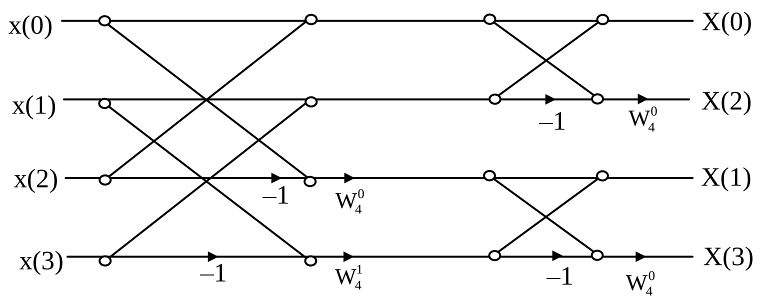


$W_4^0 = 1, W_4^1 = -j$

input	Stage 1 outputs	Stage 2 outputs
0	0 + 2 = 2	2 + 4 = 6
2	0 – 2 = –2	–2 + (–j) (–2) = –2 + 2j
1	1 + 3 = 4	2 – 4 = – 2
3	1 – 3 = – 2	– 2 – (–j) (–2) = – 2 – 2j

X(k) = {6, –2 + 2j, –2, –2 – 2j}

(ii) 4 point DIF algorithm:



inputs	Stage 1 outputs	Stage 2 outputs
0	0 + 2 = 2	2 + 4 = 6
1	1 + 3 = 4	2 – 4 = – 2
2	0 – 2 = – 2	– 2 + 2j
3	(1 – 3) (–j) = 2j	– 2 – 2j

X(k) = {6, –2 + 2j, –2, –2 – 2j}

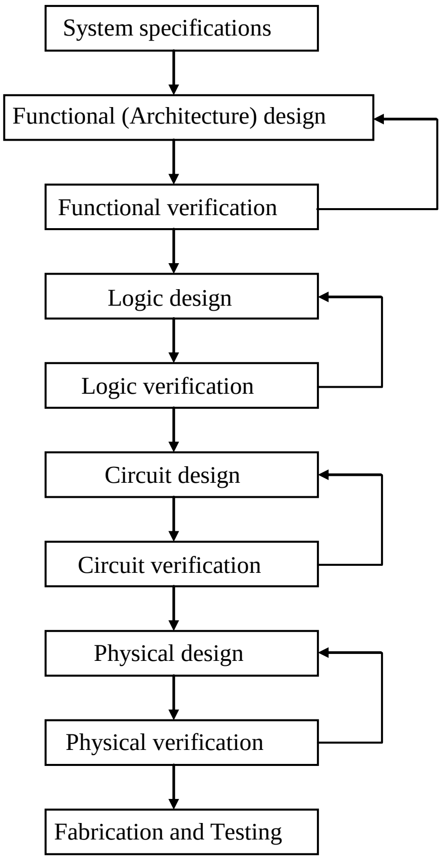
01. (e)

Sol: The VLSI design flow is a sequence of steps followed to translate the idea of a system into a chip. The flow is based on the standard design automation tools. It starts with the system specifications such as area, speed and power. Then the functional design is followed by the functional verification to check if the design is correct. In this phase, the design is described at the behavioural level.

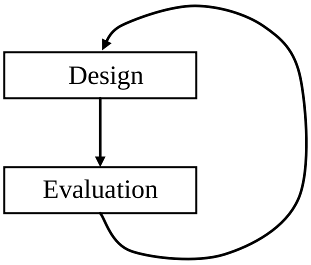
Next, the design is implemented at the logic level and verified for its correctness. This is followed by the transistor level or circuit design verification. Up to this step the flow is known as logical design.

The next phase is the physical design which actually deals with the geometry of the chip. Once the physical layout is generated, it must be verified to check if the layout really implements the actual design. The final step is fabrication and testing the chip.

Flow chart of VLSI design flow:



Design iteration loop:



02. (a)

Sol: Initial slope = – 6 dB/octave

i.e there is one pole at origin (or) one integral term portion of transfer function

$$G(s) = \frac{K}{S}$$

At ω = 2 rad/sec, slope is changed to 0 dB/oct

∴ change is slope = present slope – previous slope = 0 – (– 6) = 6dB/octave

∴ There is a real zero at corner frequency

ω_1 = 2 rad/sec

$$1 + sT_1 = \left(1+\frac{s}{\omega_1}\right) = 1+\frac{s}{2}$$

At ω = 10 rad/sec, slope is changed to – 6dB/octave

∴ Change in slope = – 6 – 0

= – 6dB/octave

∴ There is a real pole at corner frequency, ω_2 = 10 rad/s

$$\frac{1}{1+sT_2} = \frac{1}{\left(1+\frac{s}{\omega_2}\right)} = \frac{1}{\left(1+\frac{s}{10}\right)}$$

At ω = 50 rad/sec, slope is changed to – 12 dB/octave

∴ Change in slope = – 12– (–6) = – 6 dB/octave

∴ There is a real pole at corner frequency ω_3 = 50 rad/sec

$$\frac{1}{1+sT_3} = \frac{1}{1+\frac{s}{\omega_3}} = \frac{1}{1+\frac{s}{50}}$$

At ω = 100 rad/sec, the slope changed to – 6dB /octave.

∴ change in slope = – 6 – (–12) = 6dB/octave

∴ There is a real zero at corner frequency ω_4 = 100 rad/sec

$$\therefore (1 + sT_4) = 1 + \frac{s}{\omega_4} = 1+\frac{s}{100}$$

Hence, transfer function

$$= \frac{K\left(1+\frac{s}{2}\right)\left(1+\frac{s}{100}\right)}{S\left(1+\frac{s}{50}\right)\left(1+\frac{s}{10}\right)} = \frac{K(s+2)(s+100)}{s(s+50)(s+10)}\frac{\frac{1}{2}\cdot\frac{1}{100}}{\frac{1}{50}\cdot\frac{1}{10}}$$

$$= \frac{2.5K(s+2)(s+100)}{s(s+10)(s+50)}$$

In the given bode plot, at ω = 1 rad/sec

Magnitude = 20 dB

– 20 = 20 logK – 20logω

– 20 = 20log K

⇒ K = 0.1

$$\therefore \text{ Transfer function } \frac{C(s)}{R(s)} = \frac{0.25(s+2)(s+100)}{s(s+10)(s+50)}$$

02. (b)

Sol: $x(n) = \{1,1,1,1,0,0,0,0\}$

$$X(k) = \sum_{n=0}^{N-1} x(n) e^{-j\frac{2\pi}{N}nk} \qquad k = 0 \text{ to } N-1$$

$$X(k) = \sum_{n=0}^{7} x(n) e^{-j\frac{2\pi}{8}nk} \qquad k = 0 \text{ to } 7$$

$X(0) = 4$
$X(1) = 1 - j2.414$
$X(2) = 0$
$X(3) = 1 - j0.414$
$X(4) = 0$
$X(5) = 1 + j0.414$
$X(6) = 0$
$X(7) = 1 + j2.414$
$X(k) = \{4, 1 - j2.414, 0, 1 - j0.414, 0, 1 + j0.414, 0 , 1 + j2.414\}$
Given, $x_1(n) = \{1,0,0,0,0,1,1,1\}$
$x_1(n) = x((n + 3))_8$

From circular shift in time domain property $\text{DFT}[x((n-m))_N] = e^{-j\frac{2\pi}{N}mk} X(k)$

$$\text{DFT}[x((n+3))_8] = e^{j3\frac{\pi k}{4}} X(k) = X_1(k)$$

$X_1(0) = X(0) = 4$

$$X_1(1) = X(1).e^{j\frac{3\pi}{4}} = 1 + j2.413$$

$$X_1(2) = X(2).e^{j\frac{3\pi}{2}} = 0$$

$$X_1(3) = X(3).e^{j\frac{9\pi}{4}} = 1 + j0.414$$

$$X_1(4) = X(4).e^{j3\pi} = 0$$

$$X_1(5) = X(5).e^{j\frac{15\pi}{4}} = 1 - j0.414$$

$$X_1(6) = X(6).e^{j\frac{9\pi}{2}} = 0$$

$$X_1(7) = X(7).e^{j\frac{21\pi}{4}} = 1 - j2.413$$

$X_1(k) = \{4, 1 + j2.413, 0, 1 + j0.414, 0, 1 - j0.414, 0, 1-j2.413\}$
Given $x_2(n) = \{0, 0, 1, 1, 1, 1, 0, 0\}$
$x_2(n) = x((n-2))_8$

From circular shift in time domain property $\text{DFT}[x((n-m))_N] = e^{-j\frac{2\pi}{N}mk} X(k)$

$$X_2(k) = e^{-j\frac{\pi k}{2}} X(k)$$

$X_2(0) = X(0) = 4$

$$X_2(1) = X(1).e^{-j\frac{\pi}{2}} = -2.414 - j$$

$X_2(2) = 0$

$X_2(3) = X(3) . e^{-j\frac{3\pi}{2}} = 0.4141 + j$

$X_2(4) = 0$

$X_2(5) = X(5).e^{-j\frac{5\pi}{2}} = 0.414 - j$

$X_2(6) = 0$

$X_2(7) = X(7).e^{-j\frac{7\pi}{2}} = -2.414 + j$

$X_2(k) = \{4, -2.414 - j, 0, 0.414 + j, 0, 0.414 - j, 0, -2.414 + j\}$

02. (c)

Sol: Given specifications

$K_n^1 = 120\mu A/V^2$ $V_{tn} = 0.4$ V

$K_p^1 = 60\mu A/V^2$ $V_{tp} = -0.4$ V

Gain, A = 80 dB

GBW = 50 MHz

SR = 20 V/μS

$\lambda = 0.05\ V^{-1}$

L = 1 μm

We know that aspect ratio of MOSFET $= \frac{W}{L}$

let us first choose the value of compensation capacitor C_C = 2 PF

We shall bias the second stage at five times the trial current of differential stage.

Using the slew rate (SR). Let us calculate I_5

$\boxed{I_5 = SR \times C_C} = 20 \times 10^6 \times 2 \times 10^{-12}$

$\boxed{I_5 = 40\ \mu A}$

Therefore,

$I_1 = I_2 = I_3 = I_4 = \frac{I_5}{2} = 20\mu A$

The current in the output stage can be calculated as $\boxed{I_6 = I_7 = 5I_5}$

$I_6 = I_7 = 5 \times 40 = 200\mu A$

Using GBW, we can calculate g_{m1} and g_{m2} as

$\boxed{g_{m1} = g_{m2} = GBW \times C_C}$

$= 2\pi \times 50 \times 10^6 \times 2 \times 10^{-12}$

$g_{m1} = g_{m2} = 628\mu s$

$\left(\frac{W}{L}\right)_2$ can be calculated from

$$g_{m2} = \sqrt{2K_n^1\left(\frac{W}{L}\right)_2 I_2}$$

$$628 \times 10^{-6} = \sqrt{2 \times 120 \times 10^{-6}\left(\frac{W}{L}\right)_2 \times 20 \times 10^{-6}}$$

Aspect ratio of M_2 & M_1, $\boxed{\left(\frac{W}{L}\right)_2 = 82 = \left(\frac{W}{L}\right)_1}$

Let us no calculate output resistance of M_2 and M_4 as

$$r_{ds2} = r_{ds4} = \frac{1}{\lambda I_2} = \frac{1}{0.05 \times 20 \times 10^{-6}}$$

$r_{ds2} = r_{ds4} = 10^6\Omega$

$r_{ds2} \parallel r_{ds4} = 0.5 \times 10^6\Omega$

Similarly,

$$r_{ds6} = r_{ds7} = \frac{1}{\lambda I_6} = 10^5$$

$$r_{ds6} \parallel r_{ds7} = 0.5 \times 10^5 \Omega$$

Let us now calculate gm_6 as

$$A = g_{m1}(r_{ds2} \parallel r_{ds4}) \times g_{m6}(r_{ds6} \parallel r_{ds7})$$

$10{,}000 = 628\mu \times 0.5 \times 10^6 \times g_{m6} \times 0.5 \times 10^5$

$g_{m6} = 637\mu s$

$\left(\frac{W}{L}\right)_6$ can be calculated as

$$g_{m6} = \sqrt{2K_p^1 \left(\frac{W}{L}\right)_6 .I_6}$$

$$637 \times 10^{-6} = \sqrt{2 \times 60 \times 10^{-6} \times \left(\frac{W}{L}\right)_6 \times 200 \times 10^{-6}}$$

$$\boxed{\left(\frac{W}{L}\right)_6 = 17}$$

But the geometry of M_3 and M_4 has to be in the current ratio with M_6 as

$$\frac{I_3}{I_6} = \frac{\left(\frac{W}{L}\right)_3}{\left(\frac{W}{L}\right)_6}$$

$$\frac{20\mu}{200\mu} = \frac{\left(\frac{W}{L}\right)_3}{17}$$

$$\boxed{\left(\frac{W}{L}\right)_3 = 1.7}$$

Finally, let us assume that as nmos bias transistor of $\frac{W}{L} = 4$ is available with as current of $10\mu A$.

This will give the $\frac{W}{L}$ of M_5 and M_7 as 16 & 80 respectively.

Hence the aspect ratio of the MOS transistor are given by

$$\left(\frac{W}{L}\right)_1 = \frac{82\mu m}{1\mu m} \qquad \left(\frac{W}{L}\right)_2 = \frac{82\mu m}{1\mu m}$$

$$\left(\frac{W}{L}\right)_3 = \frac{1.7\mu m}{1\mu m} \qquad \left(\frac{W}{L}\right)_4 = \frac{1.7\mu m}{1\mu m}$$

$$\left(\frac{W}{L}\right)_5 = \frac{16\mu m}{1\mu m} \qquad \left(\frac{W}{L}\right)_6 = \frac{17\mu m}{1\mu m}$$

$$\left(\frac{W}{L}\right)_7 = \frac{80\mu m}{1\mu m}$$

In order to satisfy the given specifications of two stage op amp using CMOS, the above aspect ratios are maintained during fabrication process.

03. (a)

Sol: The closed loop transfer function,

$$\frac{C(s)}{R(s)} = \frac{G(s)}{1+G(s)}$$

Given that, $G(s) = \dfrac{10}{s(s+2)}$

$$\therefore \frac{C(s)}{R(s)} = \frac{\frac{10}{s(s+2)}}{1+\frac{10}{s(s+2)}} = \frac{10}{s(s+2)+10} = \frac{10}{s^2+2s+10} \quad \ldots\ldots\ldots\ldots.. (1)$$

The values of damping ratio ξ and natural frequency of oscillation ω_n are obtained by comparing the system transfer with standard form of second order transfer function.

Standard form of second order system $\dfrac{C(s)}{R(s)} = \dfrac{\omega_n^2}{s^2+2\zeta\omega_n s+\omega_n^2}$ ....(2)

On comparing equation (1) & (2) we get,

$$\omega_n^2 = 10$$

$$\therefore \omega_n = \sqrt{10} = 3.16 \text{ rad/sec}$$

$$2\zeta\omega_n = 2$$

$$\therefore \zeta = \frac{2}{2\omega_n} = \frac{1}{3.162} = 0.316$$

$$\theta = \tan^{-1}\frac{\sqrt{1-\zeta^2}}{\zeta} = \tan^{-1}\frac{\sqrt{1-0.316^2}}{0.316} = 1.249 \text{ rad}$$

$$\omega_d = \omega_n\sqrt{1-\zeta^2} = 3.16\sqrt{1-0.316^2} = 3 \text{rad/sec}$$

$$\text{Rise Time}, t_r = \frac{\pi-\theta}{\omega_d} = \frac{\pi-1.249}{3} = 0.63 \sec$$

t_r = 0.63 sec

Percentage overshoot,

$$\%M_p = e^{\frac{-\xi\pi}{\sqrt{1-\xi^2}}}\times 100 = e^{\frac{-0.316\pi}{\sqrt{1-0.316^2}}}\times 100 = 0.3512\times 100 = 35.12\%$$

$$\text{Peakovershoot} = \frac{35.12}{100}\times 12 \text{ units} = 4.2144 \text{ units}$$

Peak overshoot = 4.2144 units

$$\text{Peak time}, t_p = \frac{\pi}{\omega_d} = \frac{\pi}{3} = 1.047 \sec$$

t_p = 1.047 sec

$$\text{Time constant, } T = \frac{1}{\xi\omega_n} = \frac{1}{0.316\times 3.162} = 1\sec$$

$\therefore$ For 5% error, settling time, $t_s = 3T = 3\sec$

For 2% error, settling time, $t_s = 4T = 4\sec$

Result

Rise time, $t_r = 0.63\sec$

Percentage overshoot, $\%M_p = 35.12\%$

Peak overshoot $= 4.2144$ units, (for a input of 12 units)

Peak time, $t_p = 1.047\,\text{sec}$

Settling time, $t_s = 3\text{sec for } 5\%$ error

$= 4$sec for 2% error

03. (b)

Sol: We will first find out from Routh's array, the range of K for stability. Transfer function of system, i.e.

$$\frac{C(s)}{R(s)} = \frac{K(s+1)}{s^4 + 2s^3 + 2s^2 + (K+3)s + K}$$

The characteristic equation is given by $[s^4 + 2s^3 + 2s^2 + (K + 3)s + K] = 0$

Routh's array can be written as

s^3	1	2	K
s^4	2	K+3	0
s^2	$\frac{1-K}{2}$	K	
s^1	$\frac{(1-K)(K+3)-4K}{1-K}$	0	
s^0	K		

For stability, from s^2 row, we get $\frac{1-K}{2} > 0$ or $K < 1$

Therefore, the system will be unstable for K = 6.

Also from s^0 row, $K > 0$ for stability.

From s^1 row,

$$\frac{(1-K)(K+3)-4K}{1-K} > 0 \text{ for stability}$$

Therefore,

$$[(1 - K)(K + 3) - 4K] > 0$$

or, $3 + K - 3K - K^2 - 4K > 0$

or, $3 - 6K - K^2 > 0$

or, $K^2 + 6K - 3 < 0$

or, $(K - 0.464)(K + 6.464) < 0$

$\therefore\ K < 0.464$ or $K < -6.464$

K cannot be less than 0. Therefore, second alternative is ruled out. Range of K for stability is $0 < K < 0.464$.

03. (c)

Sol: **Initial value theorem:**

The initial value theorem enables us to calculate the initial value of a function x(t), i.e. x(0) directly from its transform X(s) without the need for finding the inverse transform of X(s). It states that

$$\text{If } x(t) \xleftrightarrow{LT} X(s)$$

$$\text{then, } \underset{t\to 0}{Lt}\, x(t) = x(0) = \underset{s\to\infty}{Lt}\, sX(s)$$

Proof: Given, L[x(t)] = X(s)

$$\text{we have, } L\left[\frac{dx(t)}{dt}\right] = \int_0^{\infty}\frac{dx(t)}{dt}e^{-st}dt = sX(s) - x(0^-)$$

Taking limit $s \to \infty$ on both sides, we get

$$\underset{s\to\infty}{Lt}\, L\left[\frac{dx(t)}{dt}\right] = \underset{s\to\infty}{Lt}\left[\int_0^{\infty}\frac{dx(t)}{dt}e^{-st}dt\right] = \underset{s\to\infty}{Lt}\left[sX(s) - x(0)\right]$$

$$\text{i.e. } 0 = \underset{s\to\infty}{Lt}\, sX(s) - x(0)$$

$$\therefore x(0) = \underset{s\to\infty}{Lt}\, sX(s)$$

$$x(0) = \underset{t\to 0}{Lt}\, x(t) = \underset{s\to\infty}{Lt}\, sX(s)$$

Final value Theorem

The final value theorem enables us to determine the final value of a function x(t), i.e. x(∞) directly from its Laplace transform X(s) without the need for finding the inverse transform of X(s). It states that

$$\text{If } x(t) \xleftrightarrow{LT} X(s)$$

$$\text{then } \underset{t\to\infty}{Lt}\, x(t) = x(\infty) = \underset{s\to 0}{Lt}\, sX(s)$$

Proof: Given, L[x(t)] = X(s)

$$\text{We have, } L\left[\frac{dx(t)}{dt}\right] = \int_0^{\infty}\frac{dx(t)}{dt}e^{-st}dt = sX(s) - x(0^-)$$

Taking the limit $s \to 0$ on both sides, we obtain

$$\underset{s\to 0}{Lt}\left[\int_0^{\infty}\frac{dx(t)}{dt}e^{-st}dt\right] = \underset{s\to 0}{Lt}\left[sX(s) - x(0^-)\right]$$

$$\therefore \int_0^{\infty}\frac{dx(t)}{dt}dt = \underset{s\to 0}{Lt}\left[sX(s) - x(0^-)\right]$$

$$\left[x(t)\right]_0^{\infty} = \underset{s\to 0}{Lt}\left[sX(s) - x(0^-)\right]$$

$$\text{i.e. } x(\infty) - x(0^-) = \underset{s\to 0}{Lt}\left[sX(s) - x(0^-)\right]$$

$$x(\infty) = \underset{s\to 0}{Lt}\, sX(s)$$

03. (d)

Sol: The periodic waveform shown in Figure with period 2π is half of a sine wave with period 2π.

$$x(t) = \begin{cases} A\sin\omega t = A\sin\frac{2\pi}{2\pi}t = A\sin t & ; \quad 0 \le t \le \pi \\ 0 & ; \quad \pi \le t \le 2\pi \end{cases}$$

Now the fundamental period $T = 2\pi$

Fundamental frequency $\omega_0 = \frac{2\pi}{T} = \frac{2\pi}{2\pi} = 1$

Let, $t_0 = 0$, $t_0 + T = T = 2\pi$

$$a_0 = \frac{1}{T}\int_0^T x(t)dt = \frac{1}{2\pi}\int_0^{2\pi} x(t)dt = \frac{1}{2\pi}\int_0^{\pi} A\sin t dt$$

$$= \frac{A}{2\pi}[-\cos t]_0^{\pi} = \frac{A}{2\pi}[-(\cos\pi - \cos 0)] = \frac{2A}{2\pi} = \frac{A}{\pi}$$

$$\therefore \quad a_0 = \frac{A}{\pi}$$

$$a_n = \frac{2}{T}\int_0^T x(t)\cos n\omega_0 t\, dt = \frac{2}{2\pi}\int_0^{2\pi} x(t)\cos nt\, dt$$

$$= \frac{1}{\pi}\int_0^{\pi} A\sin t\cos nt\, dt = \frac{A}{\pi}\int_0^{\pi}\sin t\cos nt\, dt$$

$$= \frac{A}{2\pi}\int_0^{\pi}[\sin(1+n)t + \sin(1-n)t]dt = \frac{A}{2\pi}\left[-\frac{\cos(1+n)t}{1+n} - \frac{\cos(1-n)t}{1-n}\right]_0^{\pi}$$

$$= -\frac{A}{2\pi}\left[\frac{\cos(1+n)\pi - \cos 0}{1+n} + \frac{\cos(1-n)\pi - \cos 0}{1-n}\right]$$

$$= -\frac{A}{2\pi}\left\{\left[\frac{(-1)^{n+1} - 1}{1+n}\right] + \frac{(-1)^{n-1} - 1}{1-n}\right\}$$

For odd n, $a_n = -\frac{A}{2\pi}\left[\frac{1-1}{1+n} + \frac{1-1}{1-n}\right] = 0$

For even n, $a_n = -\frac{A}{2\pi}\left[\frac{-1-1}{1+n} + \frac{-1-1}{1-n}\right] = -\frac{A}{2\pi}\left[\frac{-2}{n+1} + \frac{2}{n-1}\right] = -\frac{2A}{\pi(n^2-1)}$

$\therefore\ a_n = -\frac{2A}{\pi(n^2-1)}$ (for even n)

$$b_n = \frac{2}{T}\int_0^T x(t)\sin n\omega_0 t dt = \frac{2}{2\pi}\int_0^{2\pi} x(t)\sin nt\, dt$$

$$= \frac{1}{\pi}\int_0^{\pi} A\sin t\sin nt\, dt = \frac{A}{\pi}\int_0^{\pi}\sin t\sin nt\, dt$$

$$= \frac{A}{2\pi}\int_0^{\pi}[\cos(n-1)t - \cos(n+1)t]dt$$

$$= \frac{A}{2\pi}\left[\frac{\sin(n-1)t}{n-1} - \frac{\sin(n+1)t}{n+1}\right]_0^{\pi}$$

This is zero for all values of n except for n = 1

$$\text{For } n = 1, \quad b_1 = \frac{A}{2\pi}$$

Therefore, the Trigonometric Fourier series is:

$$x(t) = a_0 + \sum_{n=1}^{\infty} a_n \cos n\omega_0 t + b_n \sin n\omega_0 t$$

$$= a_0 + b_1 \sin t + \sum_{n=1}^{\infty} a_n \cos nt$$

$$= \frac{A}{\pi} + \frac{A}{2\pi}\sin t - \sum_{n=\text{even}}^{\infty} \frac{2A}{\pi(n^2-1)}\cos nt$$

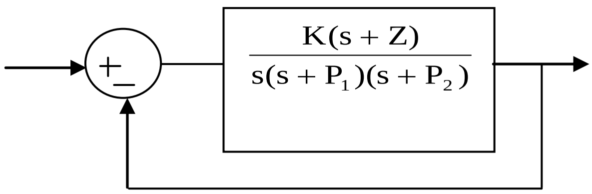
04. (a)

Sol:

(i) Root locus diagram:

Root loci diagram (RLD) is the graphical method of analyzing and designing a system

Root Locus Diagram (RLD) is a plot of loci of roots of the characteristic equation while gain 'K' is varied from 0 to ∞.



(ii)

(A) Number of root locus branches = 2(P>Z) P = 2, Z = 1

Number of Asymptotes N = P – Z= 1

$$\text{Angle of Asymptotes} = \frac{(2l+1)180^0}{P-Z} \quad l = 0$$

$$= \frac{(2(0)+1)180^0}{1} = 180^0$$

Here, only one asymptote is present, therefore centroid is not required.

Break Point:

CE is $1 + KG_1(s)H_1(s) = 0$

$$K = \frac{-1}{G_1(s)H_1(s)}$$

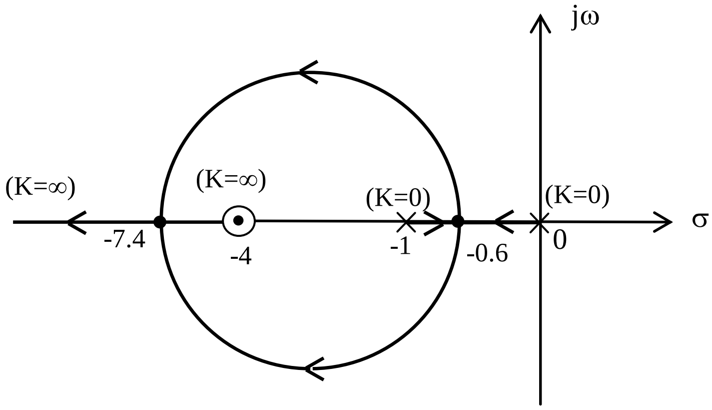
$$G_1(s)H_1(s) = \frac{s+4}{s(s+1)}$$

$$\frac{dK}{ds} = \frac{d}{ds}\left(\frac{-1}{G_1(s)H_1(s)}\right) = 0$$

$$\frac{d}{ds}\left(-\frac{s(s+1)}{s+4}\right) = 0$$

$$\frac{(s+4)(2s+1)-(s^2+s)(1)}{(s+4)^2} = 0$$

$s^2 + 8s + 4 = 0$ s = –0.6 and s = –7.4



(B) The system is critically damped when s = –0.6 and s = –7.4 (roots are real and equal)

$$K = \frac{(0.6)(0.4)}{3.4} = 0.71 \text{ (at } s = -0.6)$$

$$K = \frac{(7.4)(6.4)}{3.4} = 14 \text{ (at } s = -7.4)$$

For 0 < K< 0.71 , 14< K <∞ roots are real and distinct . Therefore the system is over damped

For K = 0.71, K = 14 , roots are real and equal .Therefore the system is critically damped.

For 0.71 < K < 14 , roots are complex and distinct . Therefore the system is under damped.

04. (b)

Sol: $\text{CLTF} = \dfrac{1}{s^2(s+a)(s+b)+1}$

Open loop transfer function G(s) = $\dfrac{\text{CLTF}}{1-\text{CLTF}}$

G(s) = $\dfrac{1}{s^2(s+a)(s+b)+1-1} \Rightarrow \dfrac{1}{s^2(s+a)(s+b)}$

G(s) has two poles at the origin, therefore type of the system is two.

Position error coefficient: $k_p = \lim_{s\to 0} G(s) = \lim_{s\to 0} \dfrac{1}{s^2(s+a)(s+b))} = \infty$

Velocity error coefficient: $k_v = \lim_{s\to 0} sG(s) = \lim_{s\to 0} \dfrac{1}{s(s+a)(s+b)} = \infty$

Acceleration error coefficient: $K_a = \lim_{s\to 0} s^2 G(s) = \lim_{s\to 0} \dfrac{1}{(s+a)(s+b)} = \dfrac{1}{ab}$

For input 2u(t): $e_{ss} = \dfrac{A}{1+k_p} = \dfrac{2}{1+\infty} = 0$

For input $4t^2$ u(t): $e_{ss} = \dfrac{A}{k_a} = \dfrac{8}{1/ab} = 8ab$

For input tu(t): $e_{ss} = \dfrac{A}{K_V} = \dfrac{1}{\infty} = 0$

(ii) The overall Transfer function is given by $\frac{C(s)}{R(s)} = \frac{\frac{K}{s(sT+1)}}{\left[1+\frac{K}{s(sT+1)}.1\right]} = \frac{K/T}{\left(s^2+\frac{1}{T}s+\frac{K}{T}\right)}$

The characteristic equation is $s^2 + \frac{1}{T}s + \frac{K}{T} = 0$

$\therefore \quad \omega_n = \sqrt{K/T} \quad \text{and} \quad 2\zeta\omega_n = 1/T$

$\therefore \quad \zeta = \frac{1}{T}.\frac{1}{2\omega_n} = \frac{1}{T}.\frac{1}{2\sqrt{K/T}} = \frac{1}{2\sqrt{KT}}$

Let K_1 be the forward path gain when $M_{p1} = 60\%$ and the corresponding damping ratio be ζ_1.

Since, $M_{p1} = e^{-\frac{\zeta_1\pi}{\sqrt{1-\zeta_1^2}}} \times 100\%$

$\therefore \; 60 = e^{-\frac{\zeta_1\pi}{\sqrt{1-\zeta_1^2}}} \times 100$

or $\quad \log_e(0.6) = -\frac{\zeta_1\pi}{\sqrt{1-\zeta_1^2}}\log_e(e)$

$\therefore \quad \zeta_1 = 0.158$

Let K_2 be the forward path gain when $M_{p2} = 20\%$ and the corresponding damping ratio be ζ_2.

Since $M_{p2} = e^{-\frac{\zeta_2\pi}{\sqrt{1-\zeta_2^2}}} \times 100\%$

$\therefore 20 = e^{-\frac{\zeta_2\pi}{\sqrt{1-\zeta_2^2}}} \times 100$

From the above relation the value of ζ_2 can be calculated as $\zeta_2 = 0.447$

Assuming time constant T to be constant

$\zeta_1 = \frac{1}{2}.\frac{1}{\sqrt{K_1T}} \quad \text{and} \quad \zeta_2 = \frac{1}{2}.\frac{1}{\sqrt{K_2T}}$

$\frac{\zeta_1}{\zeta_2} = \frac{1}{2}.\frac{1}{\sqrt{K_1T}} \times \frac{2\sqrt{K_2T}}{1}$

Hence, $\frac{K_2}{K_1} = \left(\frac{\zeta_1}{\zeta_2}\right)^2 = \left(\frac{0.158}{0.447}\right)^2 = \frac{1}{8}.$

04. (c)

Sol:

(i) Given, $x(t) = e^{-at^2}$

The Fourier transform of the given signal is:

$$X(\omega) = F\left[e^{-at^2}\right] = \int_{-\infty}^{\infty} e^{-at^2} e^{-j\omega t} dt = \int_{-\infty}^{\infty} e^{-\left(at^2 + j\omega t\right)} dt \;\; = e^{\frac{-\omega^2}{4a}} \int_{-\infty}^{\infty} e^{-\left(t\sqrt{a} + \left(j\omega/2\sqrt{a}\right)\right)^2} dt$$

Let, $p = t\sqrt{a} + \dfrac{j\omega}{2\sqrt{a}}$

$\therefore \;\; dp = \sqrt{a}dt$

$$\therefore \;\; X(\omega) = \frac{e^{-\omega^2/4a}}{\sqrt{a}} \int_{-\infty}^{\infty} e^{-p^2} dp$$

$$= \frac{e^{-\omega^2/4a}}{\sqrt{a}} \sqrt{\pi} \quad \left[\because \int_{-\infty}^{\infty} e^{-p^2} dp = \sqrt{\pi}\right]$$

$$= \sqrt{\frac{\pi}{a}} e^{-\omega^2/4a}$$

$$\therefore F\left[e^{-at^2}\right] = \sqrt{\frac{\pi}{a}} e^{-\omega^2/4a} \;\text{ or }\; e^{-at^2} \xleftrightarrow{FT} \sqrt{\frac{\pi}{a}} e^{-\omega^2/4a}$$

(ii) Given, Gaussian modulated signal $x(t) = e^{-at^2} \cos\omega_c t$

$$= e^{-at^2}\left(\frac{e^{j\omega_c t} + e^{-j\omega_c t}}{2}\right)$$

$$\therefore \;\; X(\omega) = \frac{1}{2}\left\{\left[F\left(e^{-at^2} e^{j\omega_c t}\right)\right] + \left[F\left(e^{-at^2} e^{-j\omega_c t}\right)\right]\right\}$$

By using frequency shifting property $\left[\text{i.e. } e^{-j\omega_0 t} x(t) \xleftrightarrow{FT} X(\omega + \omega_0)\right]$.

We have,

$$F\left(e^{-at^2} e^{j\omega_c t}\right) = F\left(e^{-at^2}\right)\Big|_{\omega=\omega-\omega_c}$$

$$\text{and } F\left(e^{-at^2} e^{-j\omega_c t}\right) = F\left(e^{-at^2}\right)\Big|_{\omega=\omega+\omega_c}$$

$$\therefore X(\omega) = \frac{1}{2}\left[\sqrt{\frac{\pi}{a}} e^{-(\omega-\omega_c)^2/4a} + \sqrt{\frac{\pi}{a}} e^{-(\omega+\omega_c)^2/4a}\right]$$

05. (a)

Sol: $G(s)\,H(s) = \dfrac{1}{s(s+1)}$

Let the $G_c(s) = K_p + sk_d$

Hence, $G(s)\,H(s)\,G_c(s) = \dfrac{K_p + sk_d}{s(s+1)}$

At ω = 2 rad/sec, Phase margin = 40°, hence

$G(j2)\,H(j2)\,G_c(j2) = 1\angle(180° - 40°)$

$G(j2)\,H(j2)\,G_c(j2) = \dfrac{K_p + j2k_d}{j(1+j2)}$

Hence, $-140° = -90° - \tan^{-1}2 + \tan^{-1}\dfrac{2k_d}{k_p}$

or , $-140° = -90° - 63.434° + \tan^{-1}\dfrac{2k_d}{k_p}$

So, $\tan^{-1}\dfrac{2k_d}{k_p} = 13.434$

or $\dfrac{2k_d}{k_p} = 0.2388$ --------(1)

$|G(j2)\,H(j2)\,G_C(j2)| = \dfrac{(k_p + j2k_d)}{j2(1+j2)} = 1$

Therefore, $\dfrac{\sqrt{k_p^2 + (2k_d)^2}}{2\times 2.236} = 1$

Or, $\sqrt{k_p^2 + (2k_d)^2} = 4.472$

$K_p\sqrt{1+\left(\dfrac{2k_d}{K_p}\right)^2} = 4.472$ -------- (2)

From (1) and (2), we get

$K_p\sqrt{1+0.2388^2} = 4.472$

$K_p = 4.349$

Putting this value in (1) we get

$K_d = 0.519$

Controller transfer function

$G_c(s) = k_p + k_d s \quad = 4.349 + 0.519s$

05. (b)

Sol: $M = \frac{C}{R} = \frac{G}{1+GH}$

Sensitivity of closed loop transfer function with respect to changes in H can be given by,

$$S_H^M = \frac{\partial M}{\partial H}\frac{H}{M}$$

$$= \left[\frac{\partial}{\partial H}\left(\frac{G}{1+GH}\right)\right]\frac{H}{M}$$

$$= \left[\frac{-G.G}{(1+GH)^2}\right]\frac{H}{\left(\frac{G}{1+GH}\right)}$$

$$S_H^M = \frac{-GH}{1+GH}$$

Given data

G = 500, ∂G = 10

$$\%\text{ change in } G = \frac{\partial G}{G}\times 100 = 2\%$$

$$\%\text{ change in } M = 0.2\%$$

Sensitivity of closed loop TF with respect to 'G' can be given by,

$$S_H^M = \frac{\%\text{ of change in } M}{\%\text{ of change in } G} = \frac{1}{1+GH}$$

$$\frac{0.2\%}{2\%} = \frac{1}{1+500H}$$

$$1+500H = 10$$

$$H = 18\times 10^{-3}$$

05. (c)

Sol: Open loop transfer function

$$G(s)\,H(s) = \frac{K(s-3)(s-5)}{(s+1)(s+2)}$$

To find break in (or) break away points

1 + K G(s) H(s) = 0

(s+1)(s+2) + K (s–3)(s–5) = 0

$$K = \frac{-(s+1)(s+2)}{(s-3)(s-5)}$$

$$\frac{dK}{ds} = 0$$

$$= \frac{(s^2-8s+15)(2s+3)-(s^2+3s+2)(2s-8)}{(s^2-8s+15)^2}$$

$\Rightarrow (s^2-8s+15)(2s+3) - (s^2+3s+12)(2s-8) = 0$

$\Rightarrow (2s^3 - 16s^2 + 30s + 3s^2 - 24s + 45) - (2s^3 + 6s^2 + 4s) - 8s^2 - 24s - 16$

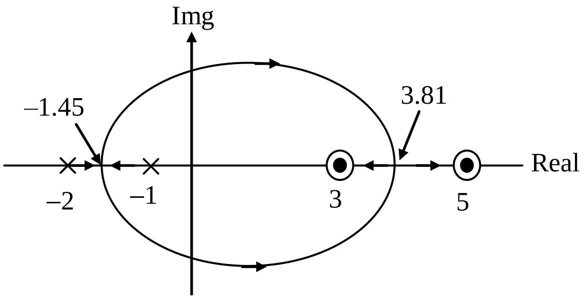
$\Rightarrow 11s^2 - 26s - 61 = 0$

By solving,

s = 3.81 and –1.45

break points are 3.81 and –1.45

Root locus



s = –1.45 lies on root locus and it is break away point.

s = 3.81 lies on root locus and as it is between two zeros and it is break in point.

As root locus start from open loop pole

(k = 0) and it will end at it's open loop zero (k = ∞).

we know that the root locus is the path traced by the roots with respect to variation of gain 'k', the gain at break away point is less than that of break in point.

05. (d)

Sol: Statement of duality property is

If $x(t) \xleftrightarrow{FT} X(\omega)$

then $X(t) \xleftrightarrow{FT} 2\pi x(-\omega)$

Proof: By definition,

Inverse Fourier transform is given by $x(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} X(\omega)e^{j\omega t}d\omega$

$$\therefore\ 2\pi x(t) = \int_{-\infty}^{\infty} X(\omega)e^{j\omega t}d\omega$$

Replace t with '-t'

$$\text{or }\ 2\pi x(-t) = \int_{-\infty}^{\infty} X(\omega)e^{-j\omega t}d\omega$$

Interchanging t and ω, we have

$$2\pi x(-\omega) = \int_{-\infty}^{\infty} X(t)e^{-j\omega t}dt = FT[X(t)]$$

$\therefore$ F[X(t)] = 2π x(–ω)

i.e, $X(t) \xleftrightarrow{FT} 2\pi x(-\omega)$

For even functions,

x(–ω) = x(ω)

$\therefore\ X(t) \xleftrightarrow{FT} 2\pi x(\omega)$

05. (e)

Sol: The given differential equation is:

$$\frac{d^2y(t)}{dt^2}+4\frac{dy(t)}{dt}+3y(t)=\frac{dx(t)}{dt}+2x(t)$$

Taking Laplace transform on both sides, we get

$[s^2Y(s) – sy(0) – y^1(0)] + 4 [sY(s) – y(0)] + 3Y(s) = [sX(s) – x(0)] + 2X(s)$

Neglecting the initial conditions, we have

$s^2Y(s) + 4sY(s) + 3Y(s) = sX(s) + 2X(s)$

i.e. $(s^2 + 4s + 3) Y(s) = (s + 2) X(s)$

$$\therefore\ Y(s)=\frac{s+2}{s^2+4s+3}X(s)=\frac{s+2}{(s+1)(s+3)}X(s)$$

Given, $x(t) = e^{-t} u(t)$

$$\therefore\ X(s)=\frac{1}{s+1}$$

$$\therefore\ Y(s)=\frac{s+2}{(s+1)(s+3)}\left(\frac{1}{s+1}\right)=\frac{s+2}{(s+1)^2(s+3)}$$

Taking partial fractions, we get

$$Y(s)=\frac{s+2}{(s+1)^2(s+3)}=\frac{A}{(s+1)^2}+\frac{B}{s+1}+\frac{C}{s+3}$$

$$Y(s)=\frac{1/2}{(s+1)^2}+\frac{1/4}{s+1}-\frac{1/4}{s+3}$$

Taking inverse Laplace transform on both sides, we get the response

$$y(t)=\left(\frac{1}{2}te^{-t}+\frac{1}{4}e^{-t}-\frac{1}{4}e^{-3t}\right)u(t)$$

05. (f)

Sol: There are mainly two types of exposure techniques:

1. Shadow printing
   a) contact printing
   b) proximity printing

2. Projection printing

In the shadow printing, the wafer and mask are either in direct contact or separated by a small gap. But in projection printing, the wafer and the mask are separated by a large distance, typically few centimetres away. Shadow printing again of two types: contact and proximity printing.

In contact printing, the mask and the wafer are in direct contact, whereas in proximity printing the mask and the wafer are separated by a small gap.

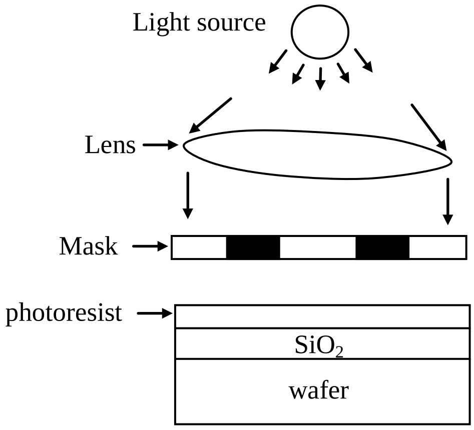


Fig: proximity printing

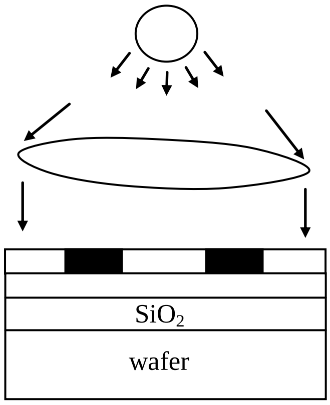


Fig: contact printing

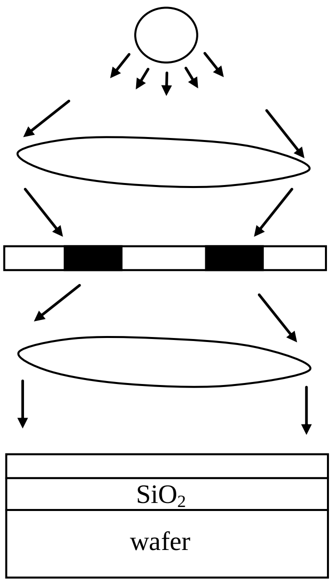


Fig: projection printing

In shadow printing, the minimum feature size that can be patterned depends on the wavelength (λ) of the light used for the exposure and the gap (g) between the mark and wafer. The minimum feature size is called critical dimension (CD)

$$CD = \sqrt{\lambda . g}$$

For projection printing, the resolution of the projection system is given by

$$\ell_m = K_1 . \frac{\lambda}{NA}$$

Where K_1 - constant

(typically 0.6)

NA - Numerical aperture.

06. (a)
Sol:

(i) (A) Given T.F $= \dfrac{2}{s(3s+2)}$

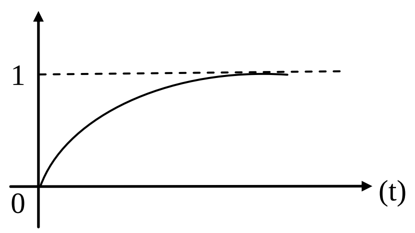
$$= \frac{1}{s} - \frac{1}{\left(s+\frac{2}{3}\right)}$$

By applying Inverse Laplace Transform on both sides, we get

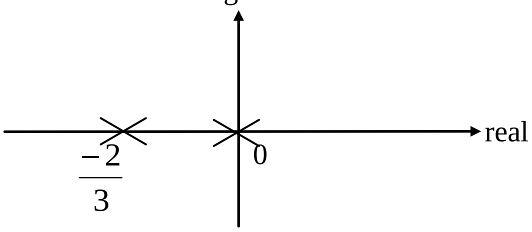
Impulse response, c(t) = L^{-1}[TF] = $L^{-1}\left[\dfrac{1}{S} - \dfrac{1}{(S+2/3)}\right]$

$$c(t) = \left(1 - e^{-\frac{2}{3}t}\right)u(t)$$

Impulse response:



Pole – Zero plot:



The given system is bounded for any bounded input. Therefore it is stable.

(B) Given T.F. $= \dfrac{6s}{\left(s^2+9\right)^2}$

The given transfer function can be written as

$$\text{T.F.} = \frac{-d}{ds}\left(\frac{3}{s^2+9}\right) \text{------(1)}$$

If x (t) $\xleftrightarrow{L.T} X(s)$ then

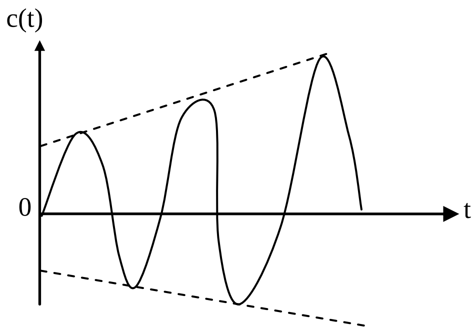
tx(t) $\xleftrightarrow{L.T} \dfrac{-d}{ds} X(s)$

By applying Inverse Laplace Transform on both sides, we get Impulse Response,

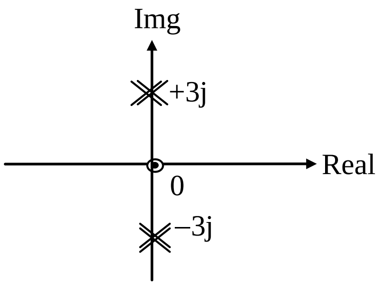
$$c(t) = L^{-1}\,[T.F] = L^{-1}\left[\frac{d}{ds}\left(\frac{3}{s^2+9}\right)\right]$$

$$c(t) = t\sin 3t\, u(t) \qquad \left[\because L^{-1}\left[\frac{3}{s^2+9}\right] = \sin 3t\; u(t)\right]$$

Impulse Response:



Pole – zero plot:



The given system is unbounded as $t \to \infty$. Therefore, it is unstable.

(ii) PID controller will have transfer function

$$G_c(s) = K_p + \frac{K_I}{s} + K_D s$$

The combination of proportional control action, integral action and derivative control action is called PID control action

Proportional controller stabilizes the gain but produces a steady state error. The integral controller reduces (or) eliminates the steady state error. The derivative controller reduces the rate of change of error.

The derivative control acts on rate of change of error and not on the actual error signal. The derivative control action is effective only during transient periods, so it doesn't produce corrective measures for only constant error. Hence derivative controller is never used alone.

(iii) **(A) Effect of phase lead network:** (Any 3 points)

- BW increases, rise time and setting time decreases, transient response is improved (faster)
- Steady state response is not affected.
- Improves the stability i.e gain and phase margins are improved
- M_r decreases
- Noise is introduced

Effect of phase lag network: (Any 3 points)

- Steady state response is improved
- ω_{ge} is reduced, BW decreases
- Rise time increases, transient response becomes sluggish /slower
- Signal to noise ratio is improved
- Improves the value of zeta, phase and gain margins are improved
- M_r decreases

(B) Phase lead compensation increases bandwidth as rise time is decreased. In phase lead controller, a zero is at right side of pole, to the forward path transfer function. The general effect is to add more damping to the closed loop system. The rise time, setting times are reduced. So, given statement is wrong

06. (b)
Sol:

(i) Nyquist stability criteria (NSC):

It states that (–1, j0) critical point should be encircled (in ACW) as many number of times as the number of poles of G(s) H(s) in the right half of s - plane by the Nyquist plot, if the Nyquist contour is defined in the clock wise direction

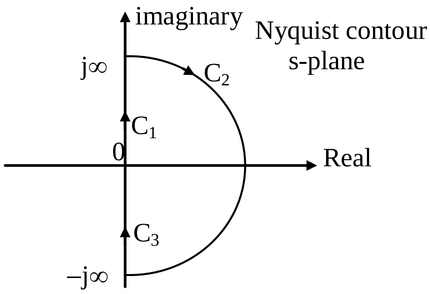
$N = P - Z$

Here,

N = Number of encirclements of (–1, j0)
critical point by the Nyquist plot
P = Number of right hand open loop poles
(or) No.of right half of s – plane poles of
F(s) and F(s) = 1 + G(s) H(s)
Z = Number of right half of s - plane poles of CLTF (or) number of right half of
s - plane zeros of F(s)

Hence, z = 0 $\Rightarrow$ stable system. i.e. N = P is called the Nyquist stability criteria

(ii) Mapping of a Nyquist contour into G(s) H(s) plane is a Nyquist plot



$\rightarrow$ Mapping of section C_1, i.e positive imaginary axis substitute $s = j\omega$, $0 \le \omega < \infty$

$$G(j\omega)\,H((j\omega) = \frac{K(j\omega+1)}{(j\omega+0.5)(j\omega-2)} = K\sqrt{\frac{\omega^2+1}{(\omega^2+0.25)(\omega^2+4)}}$$

$$\angle \tan^{-1}\omega - \left[\tan^{-1}\frac{\omega}{0.5} + 180 \quad - \quad \tan^{-1}\frac{\omega}{2}\right]$$

$\omega = 0 \rightarrow K\angle -180°$

$\omega = \omega_{pc} = 0.707 \rightarrow 0.67\,K\,\angle -180°$

$\omega = \infty \rightarrow 0\angle -90°$

Point of intersection of the plot with respect to negative real axis

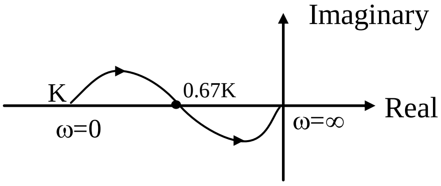
$\angle G(j\omega)\, H((j\omega) = -180°$ at $\omega = \omega_{pc}$

$$-180° + \tan^{-1}\omega_{pc} + \tan^{-1}\frac{\omega_{pc}}{2} - \tan^{-1}\frac{\omega_{pc}}{0.5} = -180°$$

$$\tan^{-1}\omega_{pc} + \tan^{-1}\frac{\omega_{pc}}{2} = \tan^{-1}\frac{\omega_{pc}}{0.5}$$

$$\Rightarrow \frac{\omega_{pc} + \frac{\omega_{pc}}{2}}{1 - \frac{\omega^2{}_{pc}}{2}} = \frac{\omega_{pc}}{0.5}$$

$\Rightarrow 1.5 = 2 - \omega_{pc}^2$



$\Rightarrow \quad \omega_{pc} = \sqrt{0.5} \Rightarrow \omega_{pc} = 0.707$

$|G\,(j\omega_{pc})\; H\,(j\omega_{pc})| = \text{POI}$

$$= \frac{K\sqrt{\omega_{pc}^2 + 1}}{(\omega_{pc}^2 + 0.25)(\omega_{pc}^2 + 4)}$$

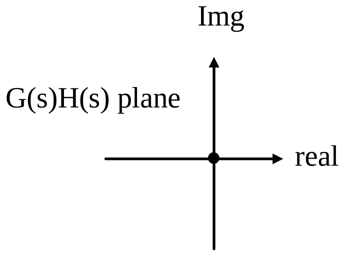
$\Rightarrow$ POI $= 0.67$ K

$\rightarrow$ Mapping of section C_2 of Nyquist contour i.e. radius 'R' semi circle.

Substitute S $= \underset{R\to\infty}{Lt}$ R.$e^{j\theta}$ $90 \geq \theta \geq -90°$

$$G(R.e^{j\theta})\; H(R.e^{j\theta}) = \frac{K(\text{Re}^{j\theta} + 1)}{(\text{Re}^{j\theta} + 0.5)(\text{Re}^{j\theta} - 2)} \cong 0$$

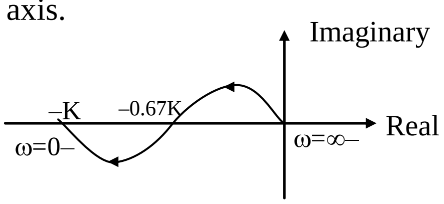
It maps to the origin



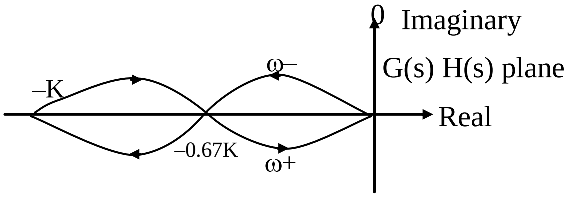
$\rightarrow$ Mapping of section C_3 of Nyquist contour i.e. negative real axis

Substitute s $= j\omega$, $-\infty \leq \omega \leq 0$

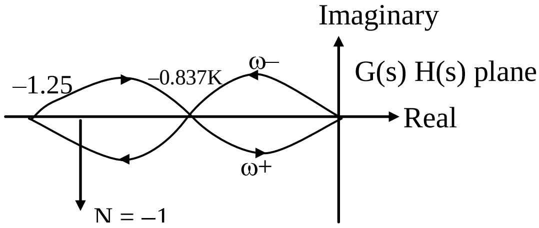
It is the image of the section C_1 and it is drawn such that the Nquist plot symmetrical w.r.t. real axis.



Combining all the above three sections the Nyquist plot is drawn below.



POI = a = 0.67K = 0.67 × 1.25 = 0.8375
N = P – Z, a = 0.8375
Z = P – N; N = – 1, P =1
Z = 1– (–1) = 2
Number of RHP = 2 ∴ unstable



∴ For stability 0.67 K > 1

$$K > \frac{1}{0.67}$$

K >1.5
∴ Stable for K >1.5

06. (c)

Sol: Energy of the signal, $E = \int_{-\infty}^{\infty} |x(t)|^2 dt = \left[\int_{-2}^{0} (t-2)^2 dt + \int_{0}^{2} (2-t)^2 dt\right]$

$$= \int_{-2}^{0} \left(t^2 - 4t + 4\right)dt + \int_{0}^{2} \left(4 + t^2 - 4t\right)dt$$

$$= \left[\frac{t^3}{3} - \frac{4t^2}{2} + 4t\right]_{-2}^{0} + \left[4t + \frac{t^3}{3} - \frac{4t^2}{2}\right]_{0}^{2}$$

$$= \frac{64}{3} \text{ joules}$$

Power of the signal, $P = \underset{T\to\infty}{\text{Lt}} \frac{1}{2T} \int_{-T}^{T} |x(t)|^2 dt$

$$= \underset{T\to\infty}{\text{Lt}} \frac{1}{2T}\left[\int_{-2}^{0} (t-2)^2 dt + \int_{0}^{2} (2-t)^2 dt\right]$$

$$= \underset{T\to\infty}{\text{Lt}} \frac{1}{2T}\left[\frac{64}{3}\right] = 0$$

Since energy is finite and power is zero, it is an energy signal.
The given signal is a non-periodic finite duration signal. So it has finite energy and zero average power. So it is an energy signal.

06. (d)

Sol: $X(z) = \dfrac{z^3}{(z-2)(z-1)^2}$

$$\frac{X(z)}{z} = \frac{z^2}{(z-2)(z-1)^2} = \frac{A}{z-2} + \frac{B}{z-1} + \frac{C}{(z-1)^2}$$

$$\frac{X(z)}{z} = \frac{4}{z-2} - \frac{3}{z-1} - \frac{1}{(z-1)^2}$$

$$X(z) = 4\frac{z}{z-2} - 3\frac{z}{z-1} - \frac{z}{(z-1)^2}$$

Apply IZT.

Then, $x(n) = 4(2)^n u(n) - 3u(n) - n\,u(n)$

07. (a)

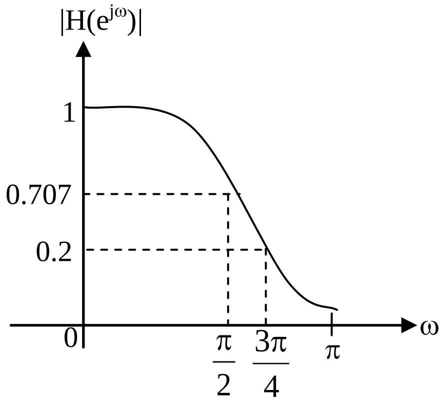
Sol:

(i) Given specifications of digital filter are

$$\frac{1}{\sqrt{1+\xi^2}} = 0.707$$

$$\frac{1}{\sqrt{1+\lambda^2}} = 0.2$$

$$\omega_p = \frac{\pi}{2}, \quad \omega_s = \frac{3\pi}{4}$$



Step 1: Convert digital filter specifications to analog filter specifications

$$\frac{1}{\sqrt{1+\xi^2}} = 0.707$$

$$\frac{1}{\sqrt{1+\lambda^2}} = 0.2$$

$$\Omega_p = \frac{2}{T}\tan\left(\frac{\omega_p}{2}\right)$$

$$\Omega_s = \frac{2}{T}\tan\left(\frac{\omega_s}{2}\right)$$

Step 2: Design normalized analog low pass filter

$$\frac{\Omega_s}{\Omega_p} = \frac{\tan(\omega_s/2)}{\tan\left(\frac{\omega_p}{2}\right)} = 2.414$$

$$N \geq \frac{\log\left[\frac{\lambda}{\xi}\right]}{\log\left[\frac{\Omega_s}{\Omega_p}\right]}$$

$\lambda = 4.898$, $\xi = 1$

$N \geq 1.803$

Round off to next higher integer. So, N = 2

$$\Omega_c = \frac{\Omega_p}{(\xi)^{1/N}} = \Omega_p = \frac{2}{T}\tan\left(\frac{\omega_p}{2}\right) = 2\text{rad/s}$$

The transfer function of N = 2 order with Ω_c = 1 rad/sec

$$H(s) = \frac{1}{s^2 + \sqrt{2}s + 1}$$

Step 3: Convert normalized analog low pass filter to un-normalized analog low pass filter using analog to analog frequency transformation.

$$H_a(s) = H(s)|_{s \to s/\Omega_c}$$

$$H_a(s) = \frac{4}{s^2 + 2.828s + 4}$$

Step 4: Convert above analog filter to digital filter using analog to digital frequency transformation i.e., bilinear transformation

Using bilinear Transformation $H(z) = H_a(s)\Big|_{S \to \frac{2}{T}\left[\frac{1-z^{-1}}{1+z^{-1}}\right]}$

$$H(z) = \frac{0.2929[1 + z^{-1}]^2}{1 + 0.1716z^{-1}}$$

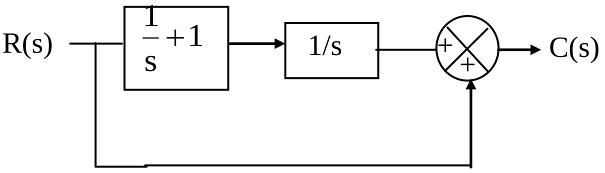
(ii)

Butterworth filter	**Chebyshev filter**
(1) The magnitude response of this filter decreases monotonically as the frequency Ω increases form 0 to ∞	(1) The magnitude response of this filter exhibits ripples in PB (or) SB according to the type
(2) The transition band is more in butterworth filter when compare to chebyshev filter.	(2) The transition band is less in chebyshev filter when compare to butterworth filter.
(3) Poles of this filter mapped onto circle in s-plane.	(3) Poles of this filter maps on to ellipse in s-plane
(4) For the same specifications the number of poles in butterworth filter are more compared to chebyshev filter i.e order of chebyshev filter is less compared to butterworth. It is the greatest advantage of chebyshev filter because less number of components will be required to construct the filter.	

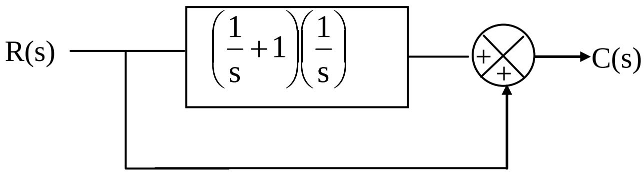
07. (b)

Sol:

(i) The given block diagram is



Step I: Cascade the two series blocks

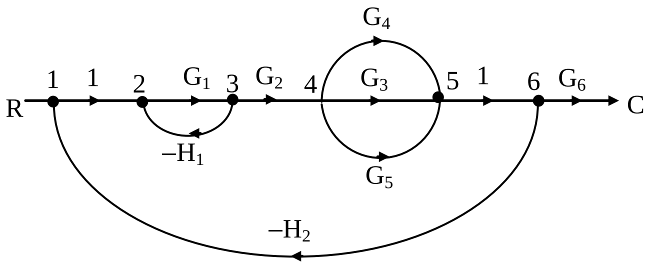


Step II: Summing the overall parallel paths

$$R(s) \xrightarrow{\quad} \frac{1}{s^2}+\frac{1}{s}+1 \xrightarrow{\quad}$$

$\therefore$ Transfer function $\dfrac{C(s)}{R(s)} = \dfrac{1}{s^2}+\dfrac{1}{s}+1$

(ii)



The No. of forward paths = 3

$$\therefore \frac{C}{R} = \frac{\sum_{K=1}^{\Delta} M_K \Delta_K}{\Delta} = \frac{M_1\Delta_1 + M_2\Delta_2 + M_3\Delta_3}{\Delta}$$

Forward path gains,

$M_1 = G_1\, G_2\, G_3\, G_6 \quad \Delta_1 = 1 – 0 = 1$

$M_2 = G_1\, G_2\, G_4\, G_6 \quad \Delta_2 = 1 – 0 = 1$

$M_3 = G_1\, G_2\, G_5\, G_6 \quad \Delta_3 = 1 – 0 = 1$

Individual loops,

$L_1 = – G_1\, H_1$

$L_2 = – G_1\, G_2\, G_3\, H_2$

$L_3 = – G_1\, G_2\, G_4\, H_2$

$L_4 = – G_1\, G_2\, G_5\, H_2$

Non – touching loops = None

$\Delta = 1 – (L_1 + L_2 + L_3 + L_4)$

$$\therefore \frac{C}{R} = \frac{G_1G_2G_3G_6 + G_1G_2G_4G_6 + G_1G_2G_5G_6}{1 + G_1H_1 + G_1G_2G_3H_2 + G_1G_2G_4H_2 + G_1G_2G_5H_2}$$

07. (c)

Sol:

(i)

Verification	Testing
1. Verifies correctness of design	1. Verifies correctness of manufactured hardware
2. Performed by simulation, hardware emulation or formal methods	2. Two - Part process i) Test generation ii) Test application
3. Performed once prior to manufacturing	3. Performed on every manufactured device
4. Responsible for quality of design	4. Responsible for quality of devices.

(ii)

FPGA	ASIC
1. It is purchased from the vendor as a standard part and then programmed by the user	1. Made to customer specification by the vendor
2. No production setup costs	2. high production setup cost
3. Fast turnaround time	3. Slow turnaround time (often at least 6 weeks)
4. Relatively high per unit cost and low capability per chip	4. Lower per unit cost but good for high volume production
5. Design requires mostly writing HDL code may be in VHDL or verilog.	5. Design often requires knowledge of physical layout of silicon inside the IC.

08. (a)

Sol: Here the impulse response and input are:

$h(t) = e^{-2t}\,u(t) = e^{-2t} \quad t > 0$

and $x(t) = e^{-4t}\,[u(t)-u(t-2)] = e^{-4t} \quad 0 < t < 2$

The output of the circuit y(t) can be obtained by convolution of x(t) and h(t).

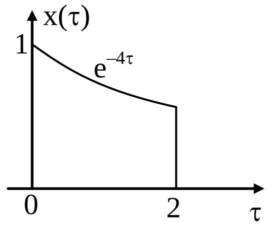
$$y(t) = x(t) * h(t) = \int_{-\infty}^{\infty} x(\tau)h(t-\tau)d\tau$$

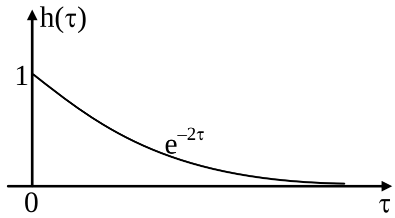
Writing x(t) and h(t) in terms of τ, we have

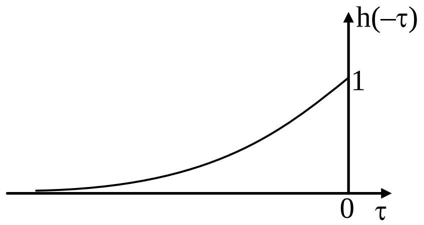
$x(\tau) = e^{-4\tau} \quad 0 < \tau < 2$

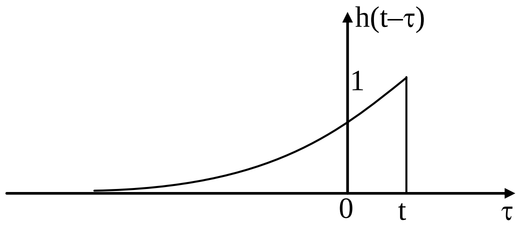
and $h(\tau) = e^{-2\tau} \quad \tau > 0$

Figure shows the plots of x(τ), h(τ) and h(–τ) w.r.t. 'τ'.









The plots of x(τ) and h(t – τ) drawn on the same time axis are shown in Figure for t < 0. The plots do not overlap.

y(t) = 0 (for t < 0)

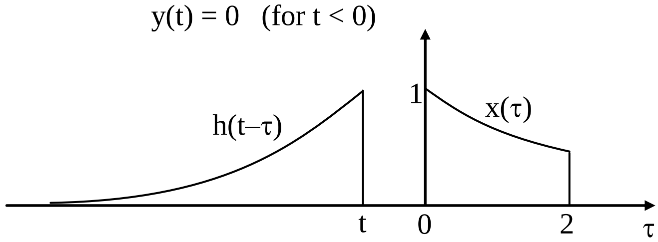


Figure. Plots of x(τ) and h(t–τ) for t < 0

For 0 < t < 2

Figure shows the plots of x(τ) and h(t – τ) for 0 < t < 2 drawn on the same time axis.

Observe that there is an overlap between x(τ) and h(t–τ) as shown by the shaded area only for 0 to t.

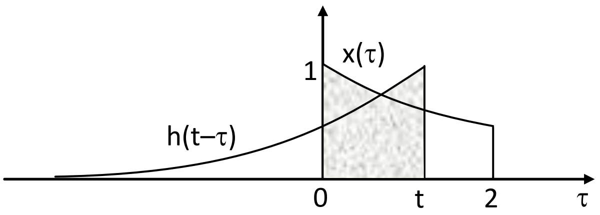


Figure. Plots of x(τ) and h(t–τ) when there is an overlap For 0 < t < 2

We can write the convolution as:

$$y(t) = \int_{-\infty}^{0} (0) h(t-\tau) d\tau + \int_{0}^{t} x(\tau) h(t-\tau) d\tau + \int_{t}^{2} x(\tau)(0) d\tau$$

$$= \int_{0}^{t} x(\tau) h(t-\tau) d\tau$$

$$= \int_{0}^{t} \left(e^{-4\tau}\right)\left(e^{-2(t-\tau)}\right) d\tau$$

$$= e^{-2t} \int_{0}^{t} e^{-2\tau} d\tau$$

$$= e^{-2t}\left[\frac{e^{-2\tau}}{-2}\right]_0^t = e^{-2t}\left(\frac{e^{-2t}-1}{-2}\right) = \frac{1}{2} e^{-2t}\left(1-e^{-2t}\right)$$

$$\therefore\ y(t) = \frac{1}{2} e^{-2t}\left(1-e^{-2t}\right) \qquad (\text{for } 0 < t < 2)$$

For t > 2

Now, consider the case t > 2. For t > 2, the plots of x(τ) and h(t – τ) drawn on the same time axis are shown in Figure. In this figure, observe that x(τ) and h(t – τ) overlap only for 0 < t < 2 as shown by the shaded area.

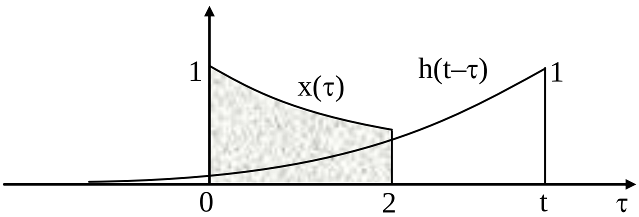


Figure. Plots of x(τ), and h(t–τ) for t > 2

Hence, we can write the convolution equation as:

$$y(t) = \int_{-\infty}^{0} (0) \times h(t-\tau) d\tau + \int_{0}^{2} x(\tau) h(t-\tau) d\tau + \int_{2}^{t} (0) \times h(t-\tau) d\tau$$

$$= \int_{0}^{2} x(\tau) h(t-\tau) d\tau = \int_{0}^{2} e^{-4\tau} e^{-2(t-\tau)} d\tau = e^{-2t} \int_{0}^{2} e^{-2\tau} d\tau$$

$$= e^{-2t}\left[\frac{e^{-2\tau}}{-2}\right]_0^2 = e^{-2t}\left(\frac{e^{-4}-1}{-2}\right)$$

$$= \frac{1}{2} e^{-2t}\left(1-e^{-4}\right) \qquad \text{for } t > 2$$

Thus, we obtained the convolution as follows:

$$y(t)=\begin{cases} 0 & ; \quad \text{for } t<0 \\ \frac{1}{2}\left(1-e^{-2t}\right)e^{-2t} & ; \quad \text{for } 0<t<2 \\ \frac{1}{2}\left(1-e^{-4}\right)e^{-2t} & ; \quad \text{for } t>2 \end{cases}$$

This function is plotted in Figure.

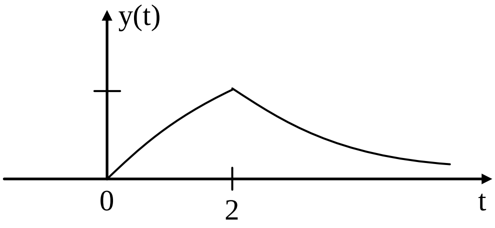


Figure. Plot of y(t)

At t = 2, the value of $y(t)=\frac{1}{2}\left(1-e^{-4}\right)e^{-4}=0.009$

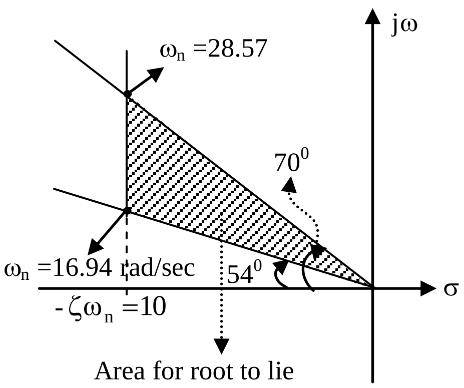
In above figure we observe that y(t) increases from t = 0 to t = 2. It has the maximum value at t = 2. The y(t) decays exponentially.

08. (b)

Sol:

(i) Given that,

$M_p = 30\% = 0.3$ and $t_s = 0.4 \sec$



We know that, $\%M_P = e^{\frac{-\zeta\pi}{\sqrt{1-\zeta^2}}} \times 100$

$$0.3 = e^{\frac{-\zeta\pi}{\sqrt{1-\zeta^2}}}$$

$$-1.2 = -\frac{\zeta\pi}{\sqrt{1-\zeta^2}}$$

$$\therefore \zeta = 0.35$$

$$\phi = \cos^{-1}(\zeta) = \cos^{-1}(0.35) = 69^0$$

If $M_p = 10\% = 0.1$, $\zeta = 0.59$

$$\phi = \cos^{-1}(\zeta) = \cos^{-1}(0.59) = 53.8^0$$

Settling time t_s (For 2% tolerance band) $= \frac{4}{\zeta\omega_n}$

$\zeta = 0.35$: $\quad \omega_n = \frac{4}{0.35 \times 0.4} = 28.57$ rad/sec

$\zeta = 0.59$: $\quad \omega_n = \frac{4}{0.59 \times 0.4} = 16.94$ rad/sec

(ii) $CE = 1 + \dfrac{K(s+1)}{s^3 + as^2 + 2s + b} = 0$

$s^3 + as^2 + (K+2)s + (K+b) = 0$

$$\begin{array}{c|cc} s^3 & 1 & K+2 \\ s^2 & a & K+b \\ s^1 & \dfrac{a(K+2)-(K+b)}{a} & 0 \end{array}$$

Given,

$\omega_n = 3$

$\Rightarrow s^1$ row = 0

s^2 row is A.E

$a(K+2) - (K+b) = 0$

$a = \dfrac{K+b}{K+2}$

$A.E = as^2 + K + b = 0$

$= \left(\dfrac{K+b}{K+2}\right)s^2 + K + b = 0$

$(K+b)\left(\dfrac{s^2}{K+2} + 1\right) = 0$

$s^2 + K + 2 = 0$

$s = \pm j\sqrt{(K+2)}$

$\omega_n = \sqrt{K+2} = 3$

$K = 7$

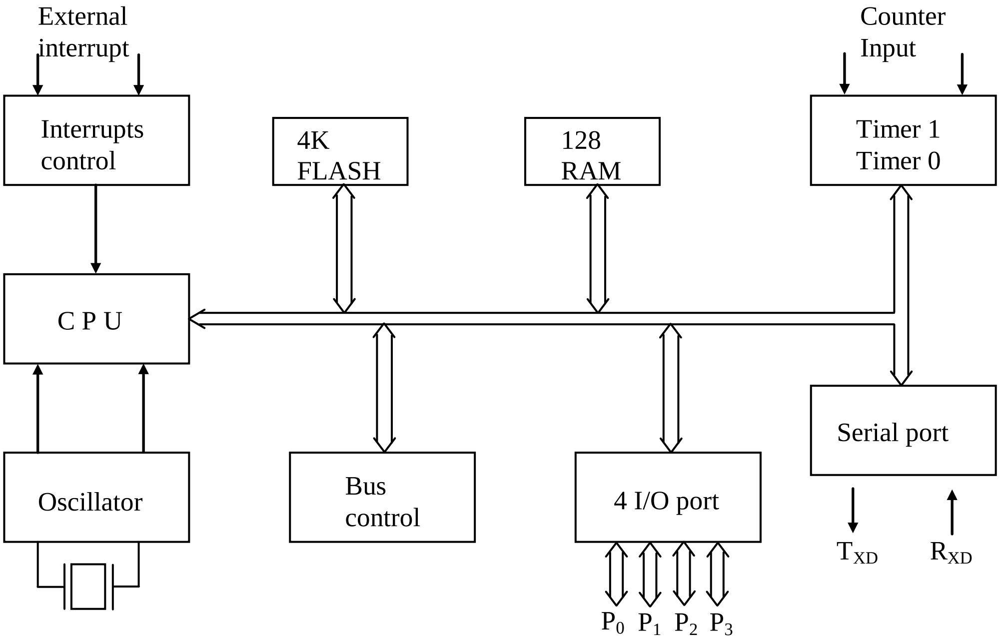
$a = \dfrac{K+b}{K+2} = \dfrac{7+b}{9}$

$\therefore 9a - b = 7$

08. (c)

Sol:

(i)



Architecture of 8051 microcontroller

(ii)

(1) **<u>Immediate addressing mode:</u>** where data is available in the instruction it self or the operand is specified in the instruction along with the op code.

Ex: ADD A, #70

A = A + 70 (decimal)

(2) **<u>Direct addressing mode:</u>** In this mode the direct address of memory location is provided in instruction to fetch the operand. Only internal RAM and SFR's address can be used in this type of instruction.

Ex: MOV A, 80H

Contents of RAM address 80H is copied into accumulator

(3) **<u>Register AM:</u>** Here the operand is contained in the specific register of microcontroller. The user must provide from where the operand/ data need to be fetched. The permitted registers are A, R0-R7 of each register bank

Ex: MOV A, R0

Content of R0 register copied into accumulator.

(4) **<u>Register indirect AM:</u>** Here the address of memory location is indirectly Provided by a register. The '@' sign indicates that the register holds the address of memory location.

Ex: MOV A, @ R0

Copy the content of memory location whose address is present in R0 register.

(5) **<u>Indexed addressing mode:</u>** the addressing mode is basically used for accessing data from look up table. Here the address of memory is indexed that is added to form the actual address of memory

Ex: MOVC A, @ A + DPTR

Here 'C' means code, content of A is added with content of DPTR and the result is address of memory location from where the data is copied to A register.