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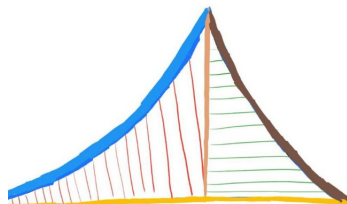
# SIGNAL PROCESSING

## Through GATE

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**EE1205-TA Group**

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# Introduction

This book provides solutions to signal processing problems in GATE.



# Chapter 1

## Harmonics

- 1.1 A damper with damping coefficient,  $c$ , is attached to a mass of 5 kg and spring of stiffness 5 kN/m as shown in figure. The system undergoes under-damped oscillations. If the ratio of the 3<sup>rd</sup> amplitude to the 4<sup>th</sup> amplitude of oscillations is 1.5, the value of  $c$  is ?

(GATE AE-62 (2022)) **Solution:**

Parameter	Value	Description
$c$	?	Damping Coefficient
$k$	5 kN/m	Stiffness
$r$	1.5	Ratio of 3 <sup>rd</sup> amplitude to 4 <sup>th</sup> amplitude of oscillations

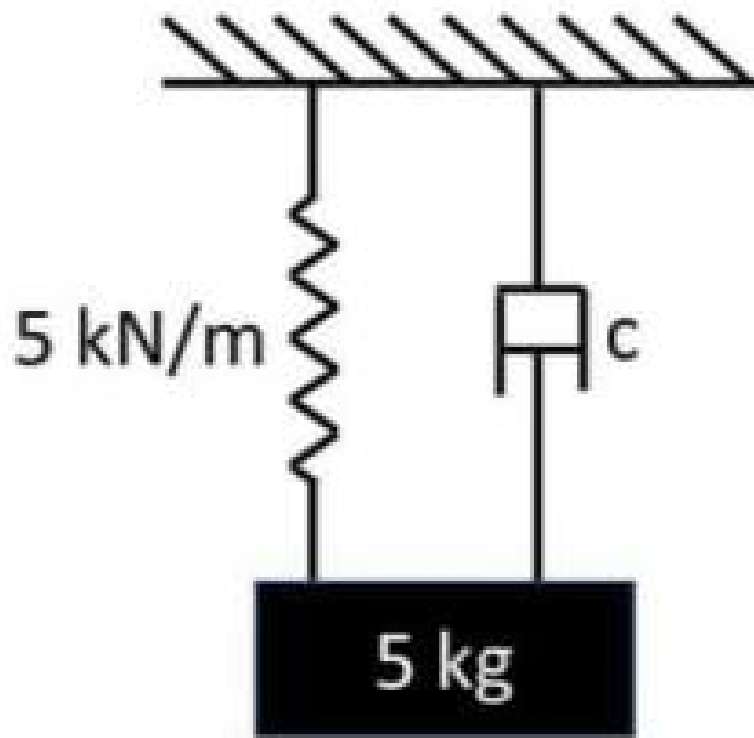
Table 1.1: Parameter Table (GATE AE-62)

Now, as the oscillation begins, from the Fig. 1.1 we write net force on the mass as,

$$F = F_1 + F_2 + mgu(t) \quad (1.1)$$

$$\Rightarrow m \frac{d^2x(t)}{dt^2} = -kx(t) - c \frac{dx(t)}{dt} + mgu(t) \quad (1.2)$$

$$\Rightarrow \frac{d^2x(t)}{dt^2} + \left(\frac{c}{m}\right) \frac{dx(t)}{dt} + \left(\frac{k}{m}\right) x(t) = gu(t) \quad (1.3)$$



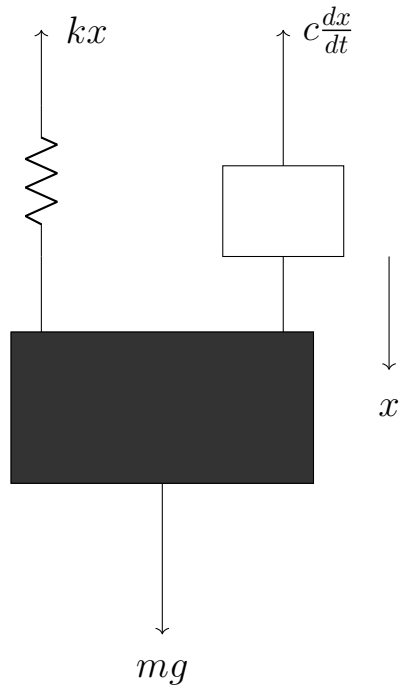
Now, taking the Laplace transform on both sides,

$$s^2 X(s) + \frac{c}{m} s X(s) + \frac{k}{m} X(s) = \frac{g}{s} \quad (1.4)$$

$$\Rightarrow X(s) = \frac{g}{s \left( s^2 + \frac{c}{m} s + \frac{k}{m} \right)} \quad (1.5)$$

$$\Rightarrow X(s) = \frac{g}{s(s - s_1)(s - s_2)} \quad (1.6)$$





Where

$$s_1 = -\frac{c}{2m} + \sqrt{\left(\frac{c}{2m}\right)^2 - \frac{k}{m}} \quad (1.7)$$

$$s_2 = -\frac{c}{2m} - \sqrt{\left(\frac{c}{2m}\right)^2 - \frac{k}{m}} \quad (1.8)$$

From (1.6) we get,

$$\begin{aligned} \Rightarrow X(s) = & \frac{g}{(s_1 - s_2)} \left[ \frac{1}{s_1(s - s_1)} - \frac{1}{s_2(s - s_2)} \right] \\ & - \frac{g}{s_1 s_2} \left( \frac{1}{s} \right) \end{aligned} \quad (1.9)$$

Now again taking the inverse Laplace transform we have,

$$x(t) = -\frac{g}{s_1 s_2} u(t) + \frac{g}{(s_1 - s_2)} \left[ \frac{1}{s_1} e^{s_1 t} - \frac{1}{s_2} e^{s_2 t} \right] u(t) \quad (1.10)$$

$$\begin{aligned} \Rightarrow x(t) = & -\sqrt{\left(\frac{mg}{k}\right)^2 + \left(\frac{gc}{2mk}\right)^2} e^{-ct/2m} u(t) \\ & \sin \left( \sqrt{\frac{k}{m} - \left(\frac{c}{2m}\right)^2} t + \tan^{-1} \left( \frac{2mg\sqrt{\frac{k}{m} - \left(\frac{c}{2m}\right)^2}}{gc} \right) \right) \\ & - \frac{mg}{k} u(t) \end{aligned} \quad (1.11)$$

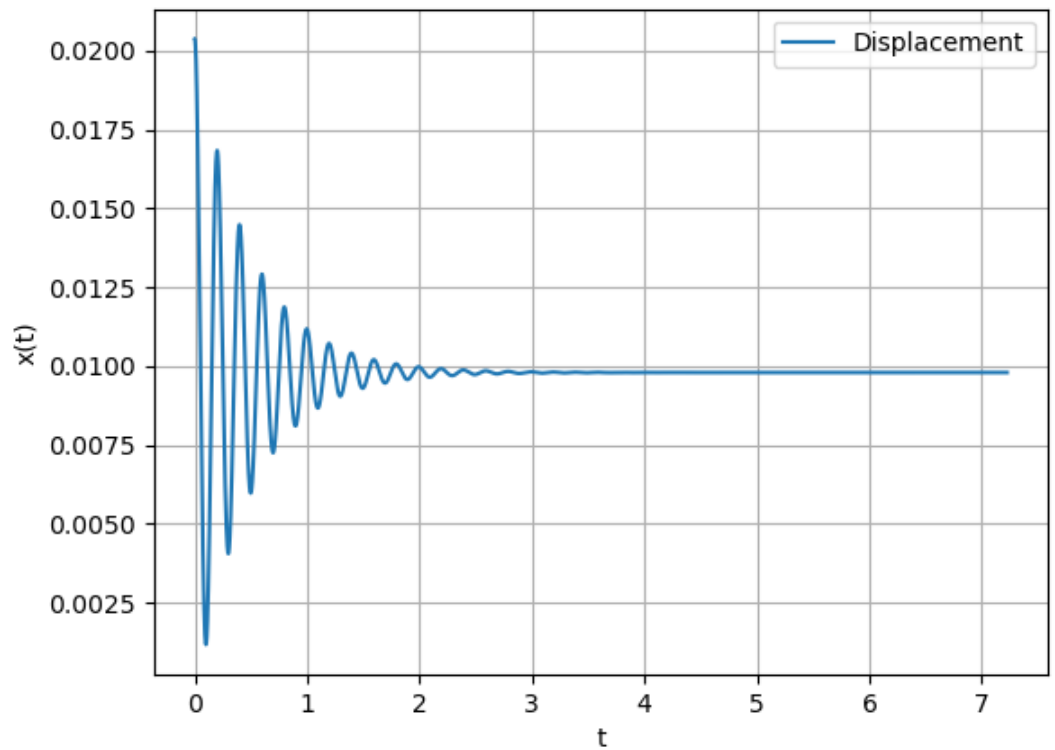
(Substituting the values of  $s_1$  and  $s_2$  from (1.7) and (1.8))

From (1.11), we have the ratio of 3<sup>rd</sup> to 4<sup>th</sup> amplitude,

$$\begin{aligned} & -\sqrt{\left(\frac{mg}{k}\right)^2 + \left(\frac{gc}{2mk}\right)^2} e^{-3cT/2m} = \\ & -\frac{3}{2}\sqrt{\left(\frac{mg}{k}\right)^2 + \left(\frac{gc}{2mk}\right)^2} e^{-4cT/2m} \end{aligned} \quad (1.12)$$

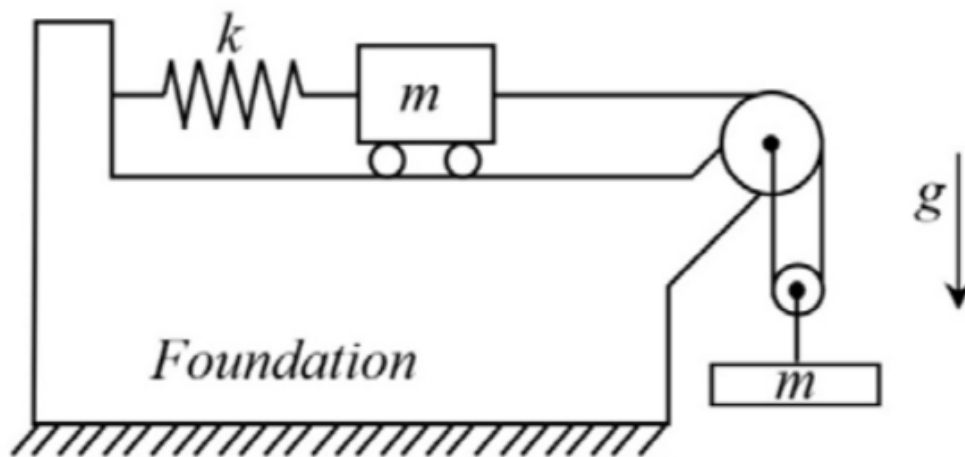
$$\Rightarrow e^{\pi c/\sqrt{mk}} = \frac{3}{2} \quad (1.13)$$

$$\Rightarrow c = \frac{\sqrt{mk} \ln \frac{3}{2}}{\pi} \quad (1.14)$$



1.2 A spring-mass system having a mass  $m$  and spring constant  $k$ , placed horizontally on a foundation, is connected to a vertically hanging mass  $m$  with the help of an inextensible string. Ignore the friction in the pulleys and also the inertia of pulleys, string and spring. Gravity is acting vertically downward as shown. The natural frequency of the system in rad/s is

- (A)  $\sqrt{\frac{4k}{3m}}$
- (B)  $\sqrt{\frac{k}{2m}}$
- (C)  $\sqrt{\frac{k}{3m}}$
- (D)  $\sqrt{\frac{4k}{5m}}$



(GATE XE 2022)

**Solution:**

Parameters	Description	Value
$x(t)$	Displacement of mass $m$ on foundation at time $t$	
$x(0)$	Displacement of mass $m$ on foundation at time $t = 0$	0
$x'(0)$	Velocity of mass $m$ on foundation at time $t = 0$	0

Table 1.2: Parameters

$$T - kx = m \frac{d^2 x}{dt^2} \quad (1.15)$$

$$mg - 2T = m \frac{d^2 \left(\frac{x}{2}\right)}{dt^2} \quad (1.16)$$

$$\implies mg - 2kx = \frac{5}{2} m \frac{d^2 x}{dt^2} \quad (1.17)$$

$$\frac{d^2 x}{dt^2} \xleftrightarrow{\mathcal{L}} s^2 X(s) - sx(0) - x'(0) \quad (1.18)$$

$$t^n \xleftrightarrow{\mathcal{L}} \frac{n!}{s^{n+1}} \quad (1.19)$$

From the Laplace transforms (1.18) and (1.19), we get

$$\frac{mg}{s} - 2kX(s) = \frac{5}{2} m (s^2 X(s) - sx(0) - x'(0)) \quad (1.20)$$

$$\implies X(s) = \frac{\frac{2g}{5}}{s \left(s^2 + \frac{4k}{5m}\right)} \quad (1.21)$$

$$= \frac{mg}{2ks} - \frac{mgs}{2k \left(s^2 + \frac{4k}{5m}\right)} \quad (1.22)$$

$$\cos at \xleftrightarrow{\mathcal{L}} \frac{s}{s^2 + a^2} \quad (1.23)$$

From the Laplace transforms (1.19) and (1.23), we get

$$x(t) = \frac{mg}{2k} \left( 1 - \cos \left( \sqrt{\frac{4k}{5m}} t \right) \right) u(t) \quad (1.24)$$

$$\implies \omega = \sqrt{\frac{4k}{5m}} \quad (1.25)$$

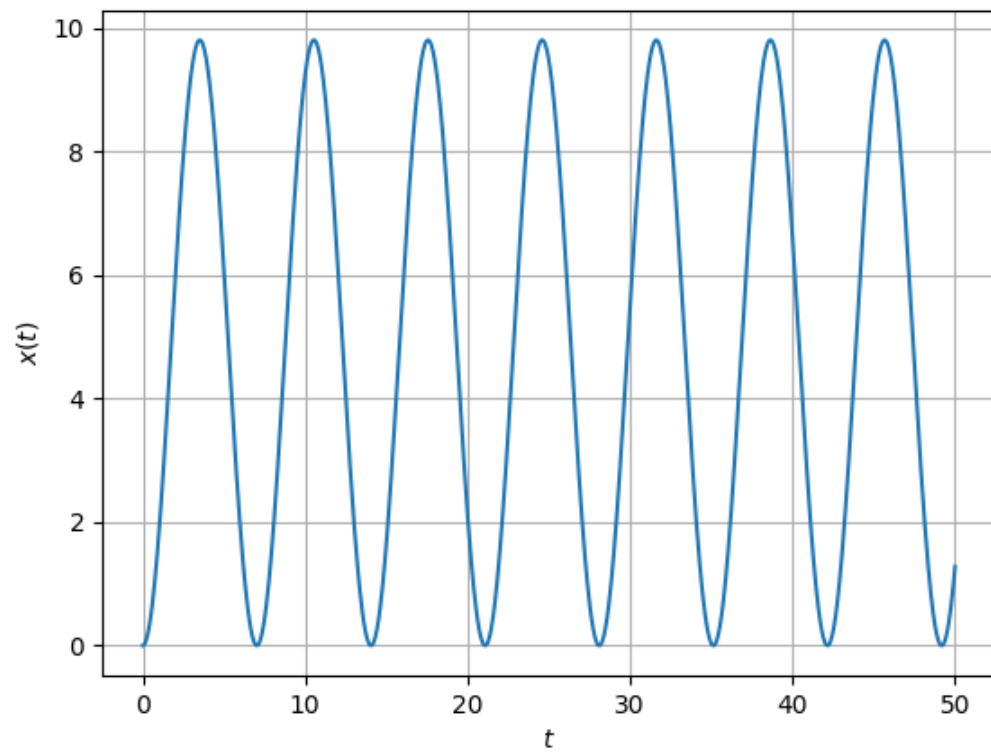


Figure 1.1: Plot of  $x(t)$  for  $m = 1kg$ ,  $k = 1N/m^2$

1.3 The time delay between the peaks of the voltage signals  $v_1(t) = \cos(6t + 60^\circ)$  and  $v_2(t) = -\sin(6t)$  is \_\_\_\_\_s

(A)  $\frac{300\pi}{360}$

(B)  $\frac{10\pi}{360}$

(C)  $\frac{50\pi}{360}$

(D)  $\frac{200\pi}{360}$

(GATE BM 2022 QUESTION 18)

**Solution:** From the values given in the Table 1.3:

Parameter	Description	Value
$v_1(t)$	Input voltage signal 1	$\cos(6t + 60^\circ)$
$v_2(t)$	Input voltage signal 2	$-\sin(6t)$
$\Delta\phi$	Phase difference between two input signals	?
$\Delta t$	Time difference between maxima of two input signals	?
$\omega$	angular frequency of input voltages	6

Table 1.3: input values

$$v_1(t) = \cos(6t + 60^\circ) \quad (1.26)$$

$$v_2(t) = -\sin(6t) \quad (1.27)$$

$$\implies v_2(t) = \cos(6t + 90^\circ) \quad (1.28)$$

From (1.27) and (1.28), phase difference between two voltage signals is  $30^\circ$ . From formula,

$$\Delta\phi = \frac{\Delta t}{\frac{2\pi}{\omega}} 360 \quad (1.29)$$

$$\therefore \Delta t = \frac{10\pi}{360} s \quad (1.30)$$

Hence, option B is correct.

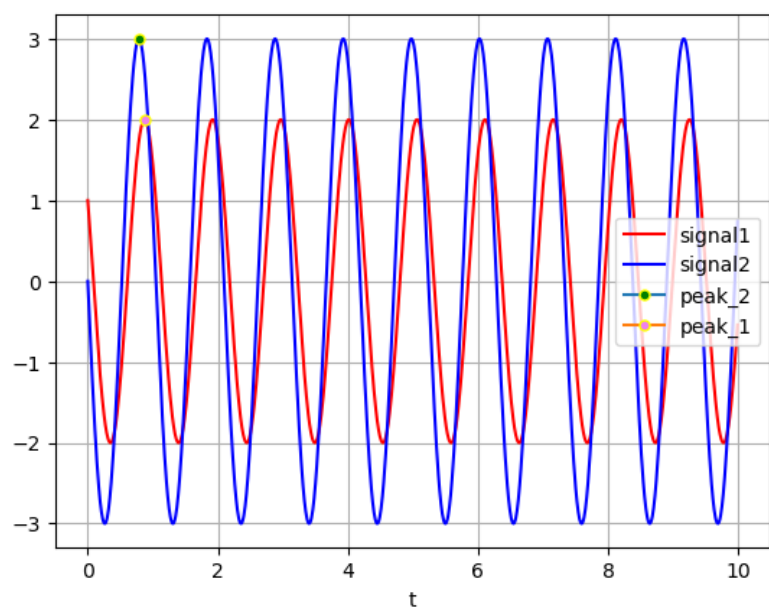


Figure 1.2: Figure of input voltage signals



1.4 A sinusoidal carrier wave with amplitude  $A_c$  and frequency  $f_c$  is amplitude modulated with a message signal  $m(t)$  having frequency  $0 < f_m \ll f_c$  to generate the modulated wave  $s(t)$  given by  $s(t) = A_c (1 + m(t)) \cos(2\pi f_c t)$  The message signal that can be retrieved completely using envelope detection is \_\_\_\_\_

(a)  $m(t) = 0.5 \cos(2\pi f_m t)$

(b)  $m(t) = 1.5 \sin(2\pi f_m t)$

(c)  $m(t) = 2 \sin(4\pi f_m t)$

(d)  $m(t) = 2 \cos(4\pi f_m t)$

(GATE IN 2022 QUESTION 16)

**Solution:**

Parameter	Description
$s(t)$	Amplitude Modulated Wave
$M(t)$	Message Signal
$c(t)$	Carrier Signal
$f_c$	Frequency of Carrier Signal
$f_m$	Frequency of Message Signal

Table 1.4: Variables and their descriptions

$$c(t) = A_c \cos(2\pi f_c t) \quad (1.31)$$

$$M(t) = A_m \cos(2\pi f_m t) \quad (1.32)$$

$$s(t) = (A_c + M(t)) \cos(2\pi f_c t) \quad (1.33)$$

$$= A_c \left( 1 + \frac{A_m}{A_c} \cos(2\pi f_m t) \right) \cos 2\pi f_c t \quad (1.34)$$

$$= A_c (1 + m(t)) \cos 2\pi f_c t \quad (1.35)$$

Modulation Index of  $s(t) = \mu = \frac{A_m}{A_c}$

- $\mu < 1$  Signal is Can be detected
- $\mu = 1$  Signal Cannot be detected
- $\mu > 1$  Over modulation

(a)  $m(t) = 0.5 \cos(2\pi f_m t)$

$$\frac{A_m}{A_c} = 0.5 \quad (1.36)$$

$$\mu < 1 \quad (1.37)$$

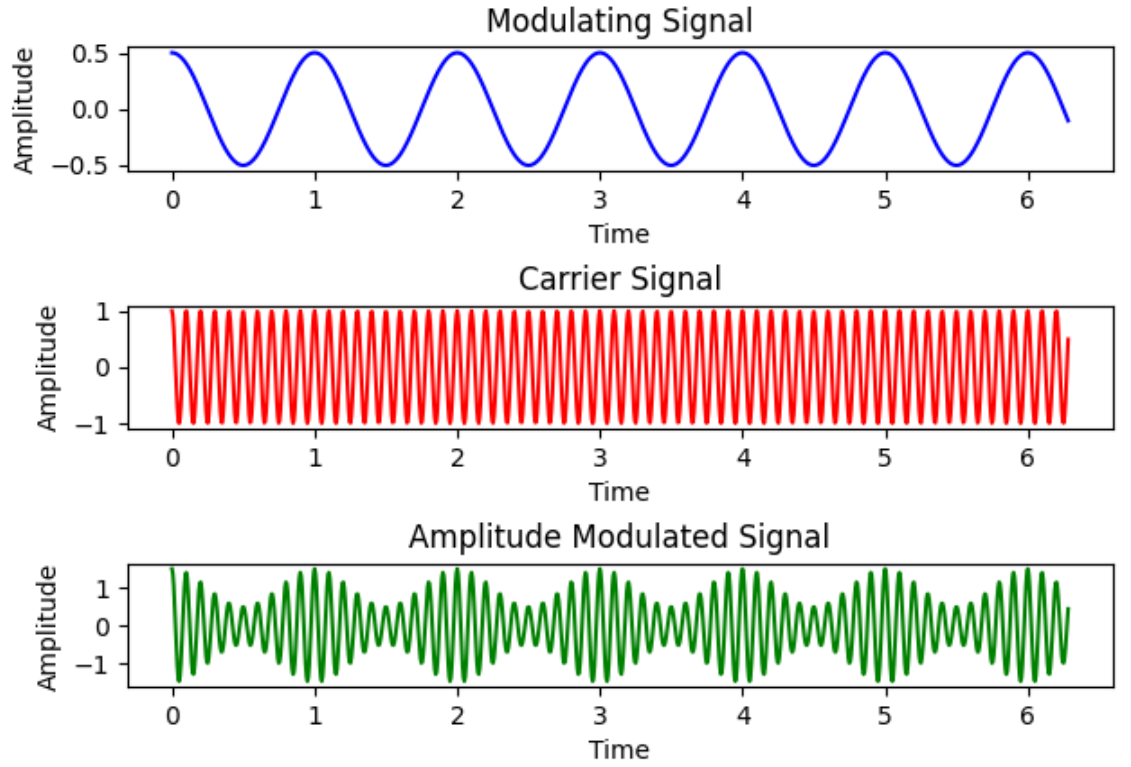
$\therefore$  Signal can be retrieved completely.

(b)  $m(t) = 1.5 \sin(2\pi f_m t)$

$$\frac{A_m}{A_c} = 1.5 \quad (1.38)$$

$$\mu > 1 \quad (1.39)$$

$\therefore$  Signal cannot be retrieved completely.



$$(c) \ m(t) = 2 \sin(4\pi f_m t)$$

$$\frac{A_m}{A_c} = 2 \quad (1.40)$$

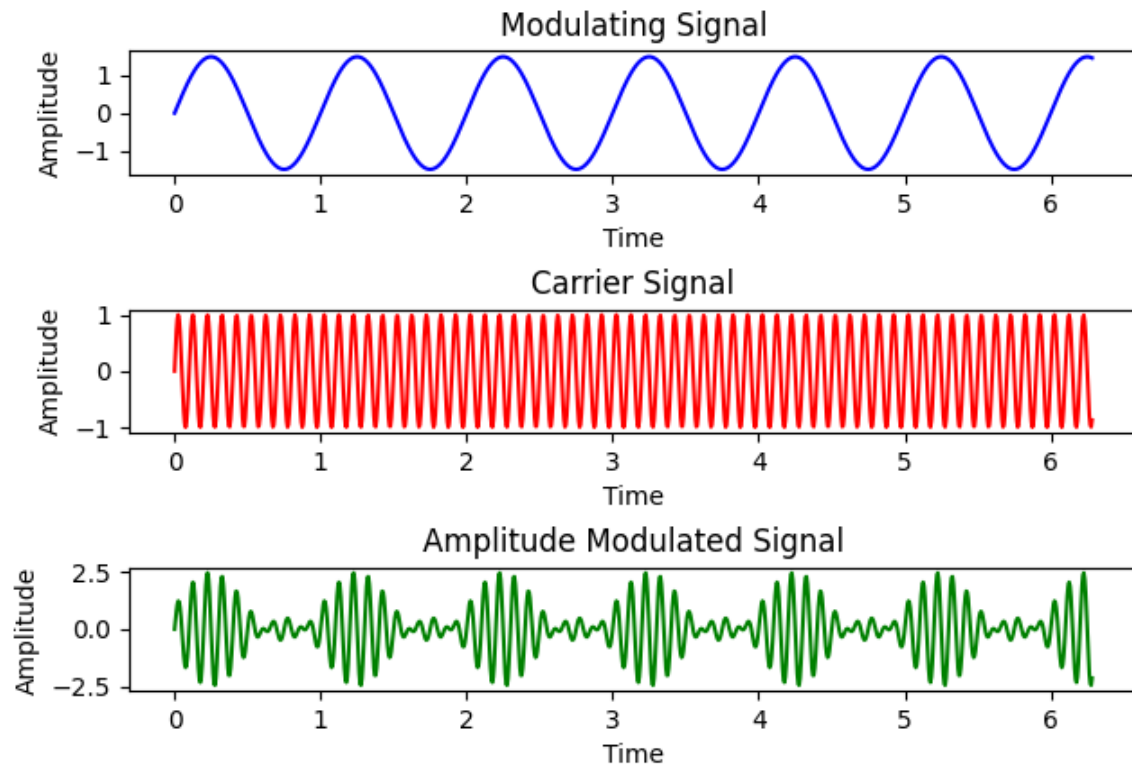
$$\mu > 1 \quad (1.41)$$

$\therefore$  Signal cannot be retrieved completely.

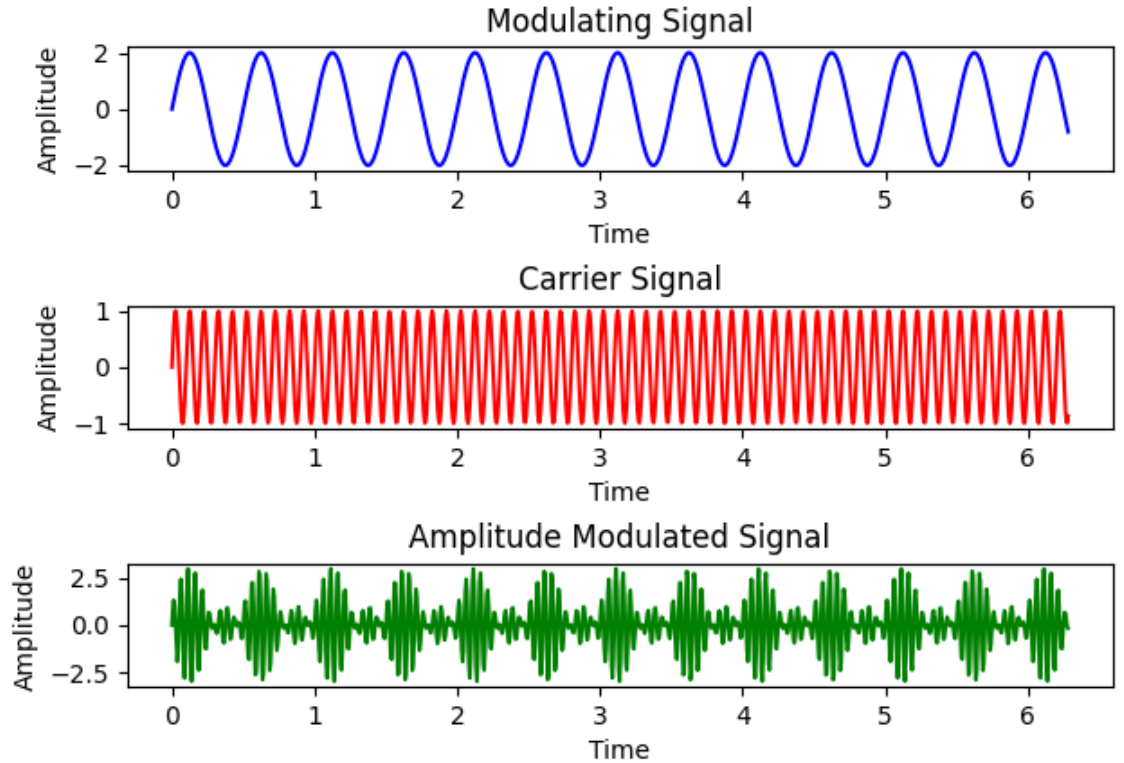
$$(d) \ m(t) = 2 \cos(4\pi f_m t)$$

$$\frac{A_m}{A_c} = 2 \quad (1.42)$$

$$\mu > 1 \quad (1.43)$$



$\therefore$  Signal cannot be retrieved completely.



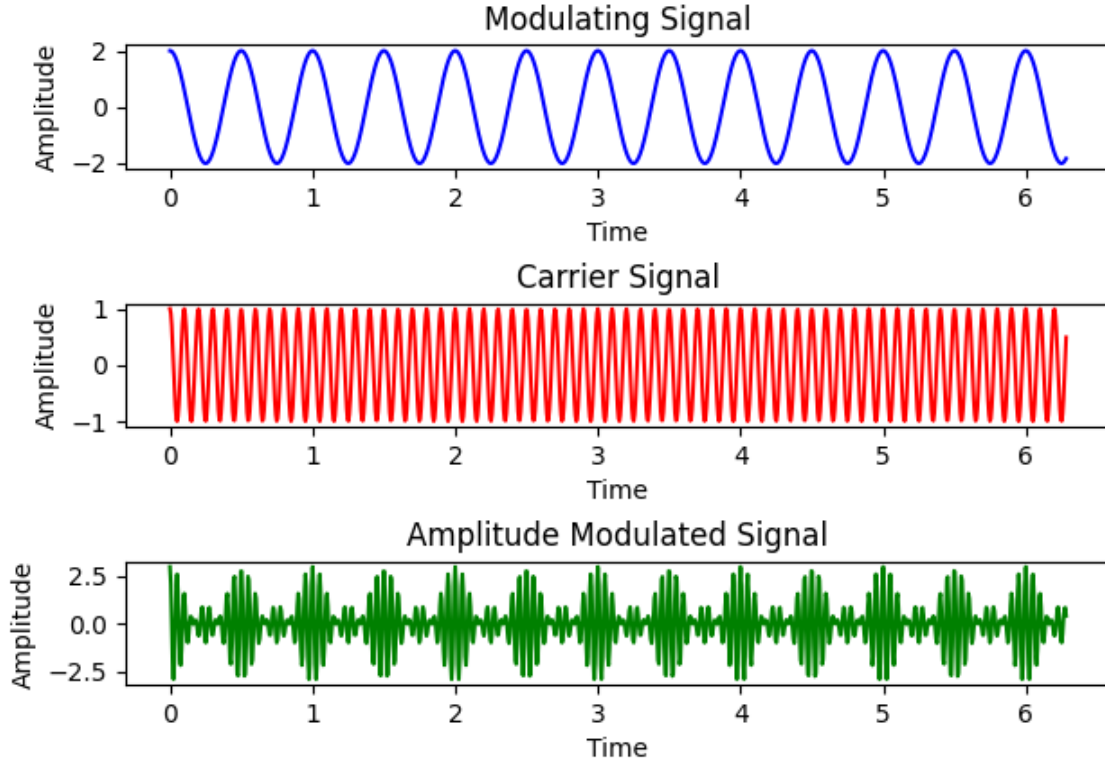
- 1.5 A uniform rigid prismatic bar of total mass  $m$  is suspended from a ceiling by two identical springs as shown in figure. Let  $\omega_1$  and  $\omega_2$  be the natural frequencies of mode I and mode II respectively ( $\omega_1 < \omega_2$ ). The value of  $\frac{\omega_2}{\omega_1}$  is \_\_\_\_\_ (rounded off to one decimal place). (GATE AE 2022 QUESTION 63)

**Solution:**

i: For vertical oscillations: from Fig. 1.4,

$$m \frac{d^2 x(t)}{dt^2} + 2kx(t) = 0 \quad (1.44)$$

Assuming the bar is at mean position and has non-zero initial velocity, we can



Parameter	Description	Value
$X(s)$	position in laplace domain	$X(s)$
$\Theta(s)$	angle rotated in laplace domain	$\Theta(s)$
$x(t)$	position of mass w.r.t time	$x(t)$
$\theta(t)$	angle rotated by mass w.r.t time	$\theta(t)$
$\alpha(t)$	angular acceleration of mass w.r.t time	$\alpha(t)$
$k$	spring constant	$k$
$m$	mass of the block	$m$
$L$	length of the mass	$L$
$\omega_o$	initial angular velocity of mass	$\omega_o$
$v(0)$	initial velocity of mass	$v(0)$

Table 1.5: input values

write it's laplace transform as:

$$s^2 m X(s) - m v(0) + 2k X(s) = 0 \quad (1.45)$$

$$16 \implies X(s) = \frac{v(0)}{s^2 + \frac{2k}{m}} \quad (1.46)$$

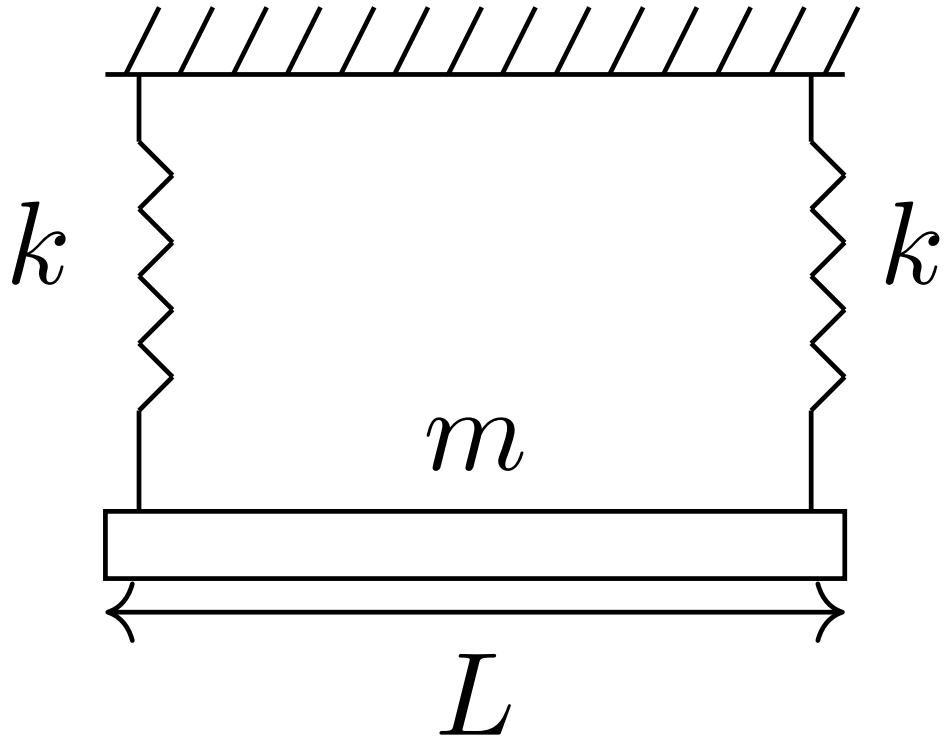


Figure 1.3: Figure given in question

On taking inverse laplace transform we get,

$$x(t) = v(0) \sqrt{\frac{m}{2k}} \sin \sqrt{\frac{2k}{m}} t \quad (1.47)$$

$$\therefore \omega_1 = \sqrt{\frac{2k}{m}} \quad (1.48)$$

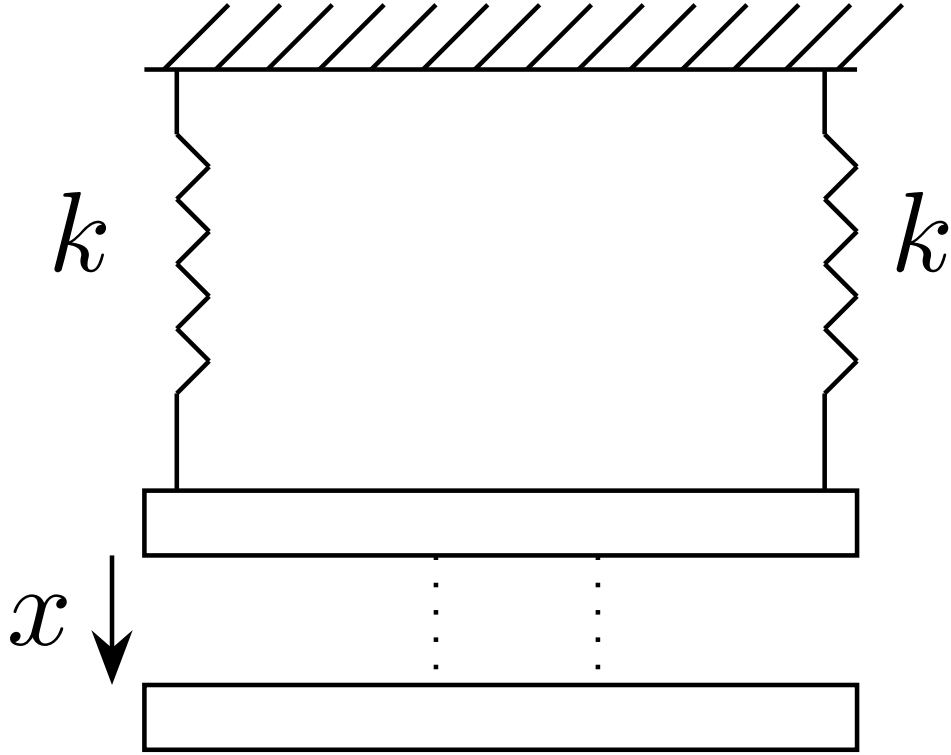


Figure 1.4: Figure for Vertical strain

ii: For torsional strain from Fig. 1.5,

$$I\alpha(t) = -\frac{kL^2\theta(t)}{2} \quad (1.49)$$

Assuming it is at mean position and having non-zero angular velocity we can write it's laplace transform as:

$$s^2 I\Theta(s) - I\omega_o + \frac{kL^2\Theta(s)}{2} = 0 \quad (1.50)$$



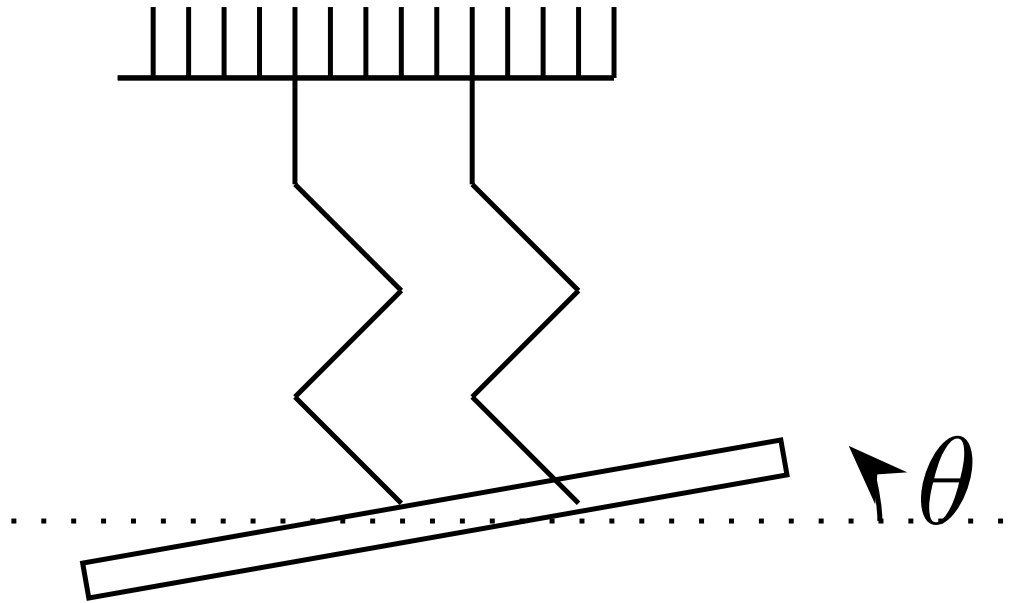


Figure 1.5: Figure for Torsional strain

substituting values from Table 1.5:

$$\Theta(s) = \frac{\omega_o}{s^2 + \frac{6k}{m}} \quad (1.51)$$

On taking inverse laplace transform we get,

$$\theta(t) = \omega_o \sqrt{\frac{m}{6k}} \sin \sqrt{\frac{6k}{m}} t \quad (1.52)$$

$$\therefore \omega_2 = \sqrt{\frac{6k}{m}} \quad (1.53)$$

From (1.48) and (1.53) we see that

$$\frac{\omega_2}{\omega_1} = \sqrt{3} \tag{1.54}$$

## Chapter 2

# Filters

- 2.1 The network shown below has a resonant frequency of 150 kHz and bandwidth of 600 Hz. The Q-factor of the network is \_\_\_\_\_ (rounded off to one decimal place).  
(GATE 2022 EC)

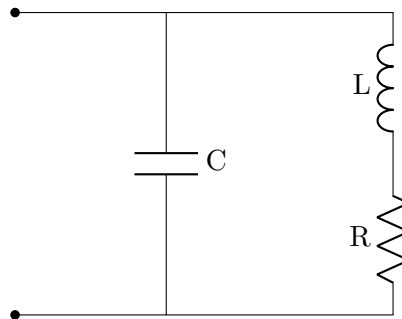


Figure 2.1: Circuit 1

**Solution:**

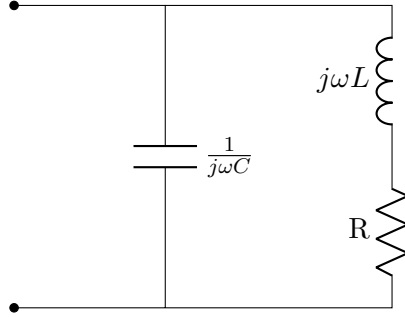


Figure 2.2: Circuit 2

Parameter	Description	Value
$f_0$	Resonant frequency	150 kHz
$B$	Bandwidth	600 Hz

Table 2.1: Parameters

At Resonance,

$$X_L = X_C \quad (2.1)$$

$$\omega_0 L = \frac{1}{\omega_0 C} \quad (2.2)$$

$$\omega_0 = \frac{1}{\sqrt{LC}} \quad (2.3)$$

$$2\pi f_0 = \frac{1}{\sqrt{LC}} \quad (2.4)$$

$$\Rightarrow f_0 = \frac{1}{2\pi\sqrt{LC}} \quad (2.5)$$

Parameter	Description	Formula
$Q$	Quality factor	$\frac{X_L}{R}$
$B$	Bandwidth	$\frac{R}{2\pi L}$
$\omega_0$	Radial resonant frequency	$2\pi f_0$
$X_L$	Inductive reactance	$\omega L$
$X_C$	Capacitive reactance	$\frac{1}{\omega C}$

Table 2.2: Formulae

Using Table 2.2,

$$Q = \frac{X_L}{R} \quad (2.6)$$

$$= \frac{\omega_0 L}{R} \quad (2.7)$$

$$= \left( \frac{1}{\sqrt{LC}} \right) \frac{L}{R} \quad (2.8)$$

$$\Rightarrow Q = \frac{1}{R} \sqrt{\frac{L}{C}} \quad (2.9)$$

From eq (2.5) and Table 2.2

$$\frac{f_0}{B} = \left( \frac{1}{2\pi\sqrt{LC}} \right) \frac{2\pi L}{R} \quad (2.10)$$

$$= \left( \frac{1}{\sqrt{LC}} \right) \frac{L}{R} \quad (2.11)$$

$$\Rightarrow \frac{f_0}{B} = \frac{1}{R} \sqrt{\frac{L}{C}} \quad (2.12)$$

From Table 2.1, eq (2.9) and eq (2.12),

$$Q = \frac{f_0}{B} \quad (2.13)$$

$$= \frac{150 \times 10^3}{600} \quad (2.14)$$

$$= 250 \quad (2.15)$$

∴ Q-factor is 250

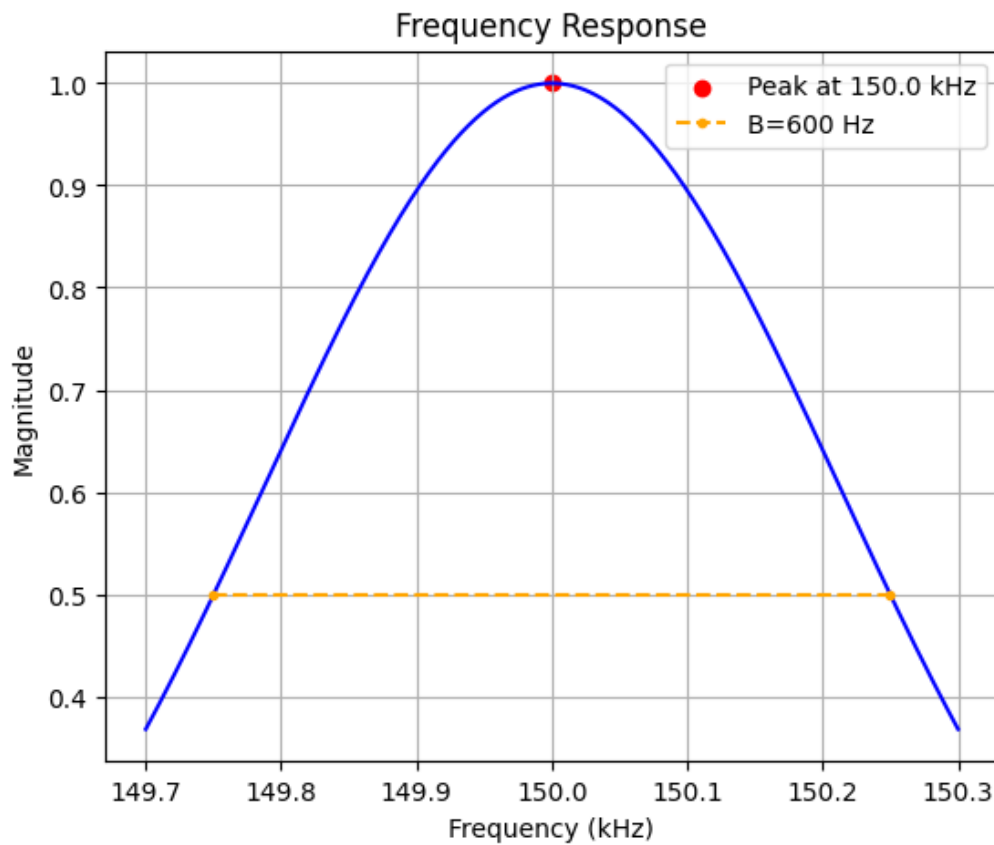
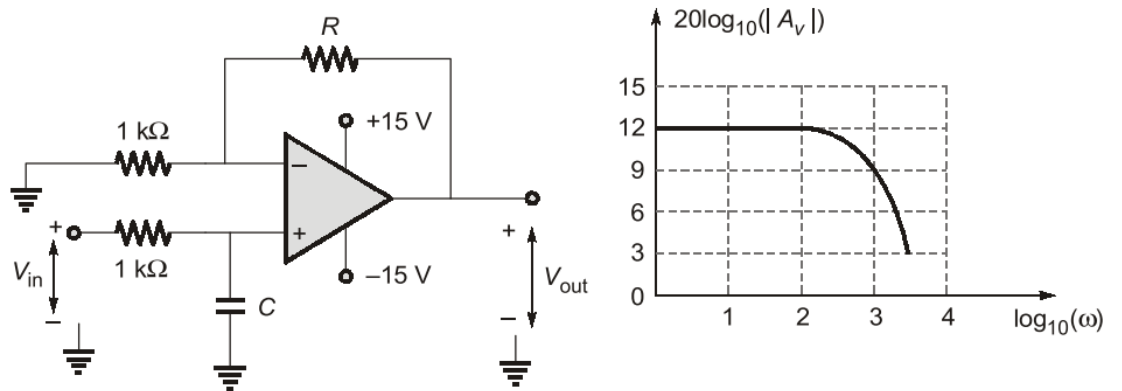


Figure 2.3: Plot of Q-factor

2.2 A circuit with an ideal OPAMP is shown. The Bode plot for the magnitude (in dB) of the gain transfer function ( $A(j\omega) = \frac{V_{out}(j\omega)}{V_{in}(j\omega)}$ ) of the circuit is also provided (here,  $\omega$  is the angular frequency in  $rad/s$ ). The values of  $R$  and  $C$  are



(A)  $R = 3\text{ k}\Omega$ ,  $C = 1\text{ }\mu\text{F}$

(B)  $R = 1\text{ k}\Omega$ ,  $C = 3\text{ }\mu\text{F}$

(C)  $R = 4\text{ k}\Omega$ ,  $C = 1\text{ }\mu\text{F}$

(D)  $R = 3\text{ k}\Omega$ ,  $C = 2\text{ }\mu\text{F}$

(GATE 2022 EC)

**Solution:**

Parameter	Description	Value
$R$	Resistance	?
$C$	Capacitance	?
$R_1$	Resistance	1000
$\omega_{dB}$	Cut-off frequency	1000
$A_V$	Gain Transfer	$\frac{V_{out}}{V_{in}}$

Table 2.3: Given Parameters

On applying KVL,

$$sR_1i_1(s) + \frac{i_1(s)}{C} = V_{IN}(s) \quad (2.16)$$

$$\frac{i_1(s)}{C} - sRi_2(s) = sV_0(s) \quad (2.17)$$

From (2.16) and (2.17),

$$\frac{sV_{IN}(s)}{sR_1C + 1} - sRi_2(s) = sV_0(s) \quad (2.18)$$

$$-\frac{i_1(s)}{C} = sR_1i_2(s) \quad (2.19)$$

From (2.16), (2.17) and (2.19) ,

$$V_{OUT}(s) = \frac{1 + 10^{-3}R}{1 + sC10^3} V_{IN}(s) \quad (2.20)$$



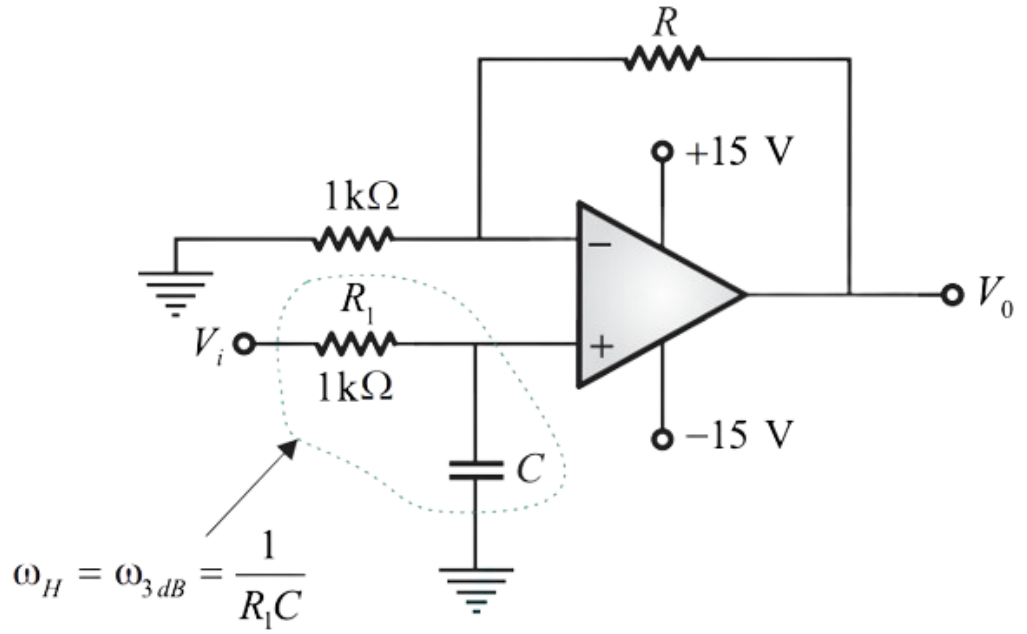


Figure 2.4: Active Low Pass Filter

The 3-dB frequency from bode magnitude plot,

$$\Rightarrow \omega_{3dB} = 1000 \text{ rad/sec} \quad (2.21)$$

$$\omega_{3dB} = \frac{1}{R_1 C} \quad (2.22)$$

$$\Rightarrow C = 1 \mu F \quad (2.23)$$

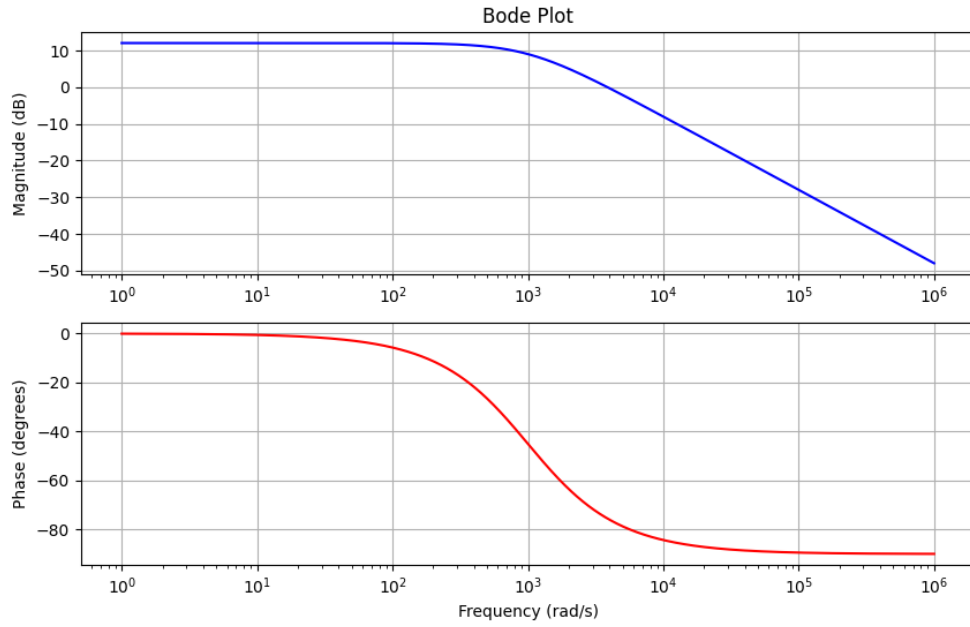


Figure 2.5: bode plot

$$\Rightarrow A(s) = \frac{V_{OUT}(s)}{V_{IN}(s)} \quad (2.24)$$

$$= \frac{1 + 10^{-3}R}{1 + sC10^3} \quad (2.25)$$

$$|A(s)| = \frac{1 + 10^{-3}R}{\sqrt{1 + \omega^2 10^{-6}}} \quad (2.26)$$

$$(2.27)$$

$A_V$  at low frequency,

$$|A_V| = 1 + 10^{-3}R \quad (2.28)$$

$$1 + 10^{-3}R = 10^{\frac{3}{5}} \quad (2.29)$$

$$R = 3k\Omega \quad (2.30)$$

Hence, The correct option is (A).

2.3 In the circuit shown, the load is driven by a sinusoidal A.C. voltage source  $V_1 = 100\angle 0^\circ V$  at  $50Hz$ . Given  $R_1 = 20\Omega$ ,  $C_1 = \left(\frac{1000}{\pi}\right)\mu F$ ,  $L_1 = \left(\frac{20}{\pi}\right)mH$  and  $R_2 = 4\Omega$ , the power factor is \_\_\_\_ (round off to one decimal place)  
(GATE 2022 IN Q52) **Solution:**

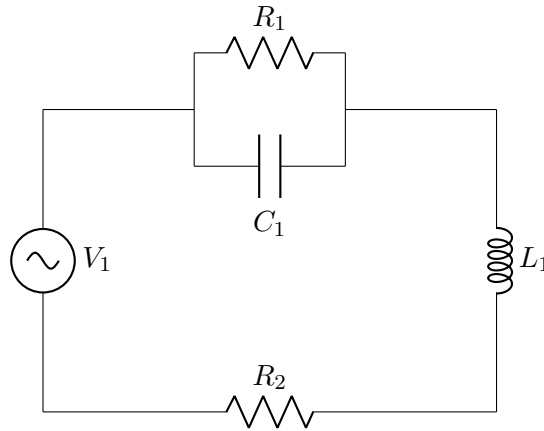


Figure 2.6:

Symbol	Value	Description
$V_1$	$100\angle 0^\circ V$	Input Voltage
$f$	$50Hz$	Frequency
$\omega$	$2\pi f$	Angular Frequency
$R_1$	$20\Omega$	Resistance
$R_2$	$4\Omega$	Resistance
$C_1$	$\left(\frac{1000}{\pi}\right)\mu F$	Capacitance
$L_1$	$\left(\frac{20}{\pi}\right)mH$	Inductance
$Z_{\text{eff}}$		Impedance
$\cos(\phi)$	$\frac{\text{Re}(Z_{\text{eff}})}{ Z_{\text{eff}} }$	Power Factor

Table 2.4: Given Parameters

$$Z_{\text{eff}} = R_2 + j\omega L_1 + \frac{\frac{R_1}{j\omega C_1}}{R_1 + \frac{1}{j\omega C_1}} \quad (2.31)$$

$$= 4 + 2j + \frac{-200j}{20 - 10j} \quad (2.32)$$

$$= 8 - 6j \quad (2.33)$$

$\therefore$  Power Factor:

$$\cos(\phi) = \frac{\text{Re}(Z_{\text{eff}})}{|Z_{\text{eff}}|} \quad (2.34)$$

$$= \frac{8}{\sqrt{8^2 + 6^2}} \quad (2.35)$$

$$= 0.8 \quad (2.36)$$

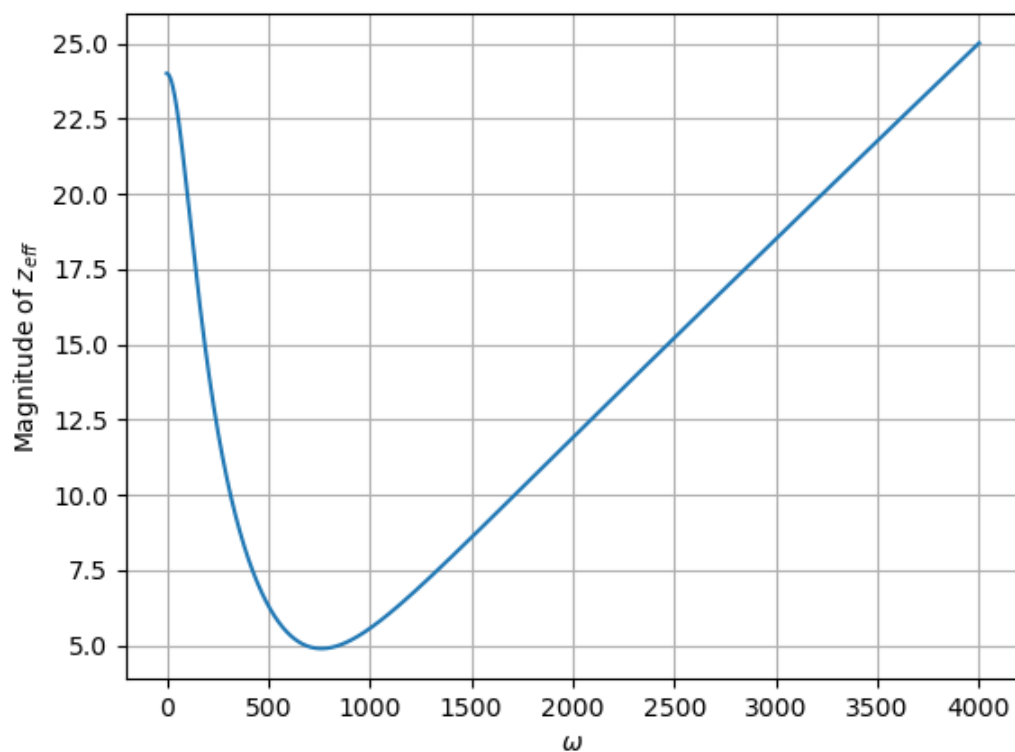


Figure 2.7: Plot of  $Z_{\text{eff}}$  vs  $\omega$

2.4 For the circuit shown, the locus of the impedance  $Z(j\omega)$  is plotted as  $\omega$  increases from zero to infinity. The values of  $R_1$  and  $R_2$  are:

- (A)  $R_1 = 2 \text{ k}\Omega, R_2 = 3 \text{ k}\Omega$
- (B)  $R_1 = 5 \text{ k}\Omega, R_2 = 2 \text{ k}\Omega$
- (C)  $R_1 = 5 \text{ k}\Omega, R_2 = 2.5 \text{ k}\Omega$
- (D)  $R_1 = 2 \text{ k}\Omega, R_2 = 5 \text{ k}\Omega$

(GATE ECE 2022 QUESTION 38)

**Solution:**

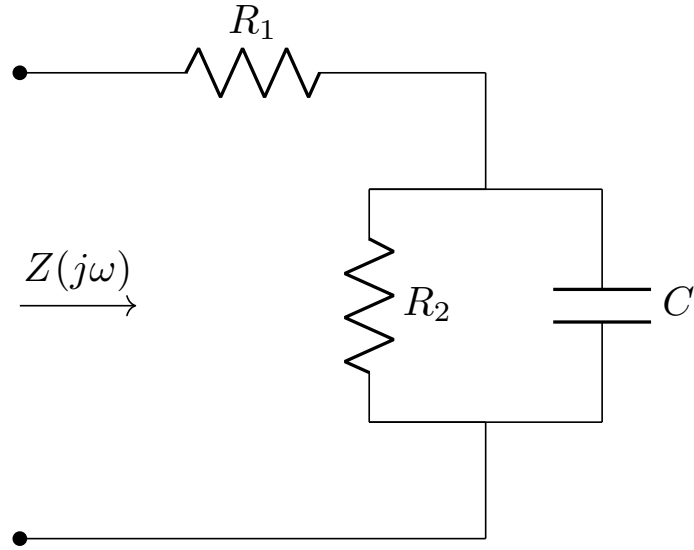


Figure 2.8: Figure of circuit

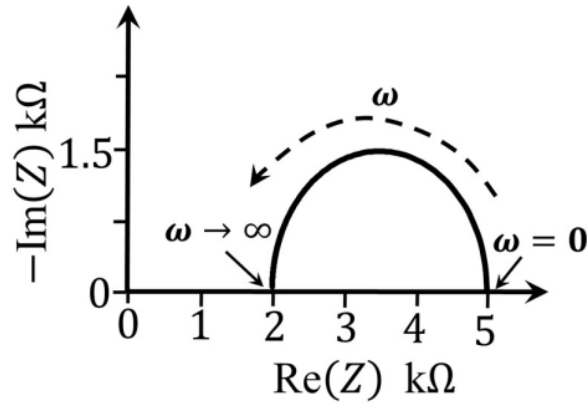


Figure 2.9:

In  $\omega$  domain (i.e. after Laplace transform) Fig. 2.8 can be represented as Fig. 2.10 So, the impedance for the circuit in  $\omega$  domain is:

$$Z(j\omega) = R_1 + \frac{1}{\frac{1}{R_2} + j\omega C} \quad (2.37)$$

Parameter	Description	Value
$Z(j\omega)$	Impedance of circuit	?
$R_1$	Resistor 1	?
$R_2$	Resistor 2	?
$C$	Capacitor	?
$\omega$	angular frequency of input voltage	$\omega$

Table 2.5: input values

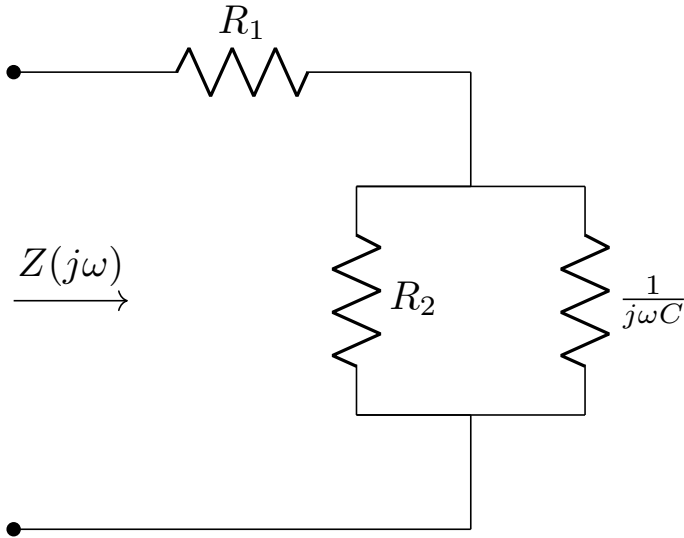


Figure 2.10:

From Fig. 2.9,  $Z(j\omega) = 2$  as  $\omega \rightarrow \infty$  and  $Z(j\omega) = 5$  as  $\omega \rightarrow 0$

$$2 = R_1 + \lim_{\omega \rightarrow \infty} \frac{1}{\frac{1}{R_2} + j\omega C} \quad (2.38)$$

$$\Rightarrow 2 = R_1 + \lim_{\omega \rightarrow \infty} \frac{\frac{1}{R_2} - j\omega C}{\left(\frac{1}{R_2}\right)^2 + (\omega C)^2} \quad (2.39)$$

$$\Rightarrow 2 = R_1 + \lim_{\omega \rightarrow \infty} \frac{\frac{1}{R_2\omega^2} - j\frac{C}{\omega}}{\left(\frac{1}{R_2\omega}\right)^2 + C^2} \quad (2.40)$$

$$\therefore 2\Omega = R_1 \quad (2.41)$$

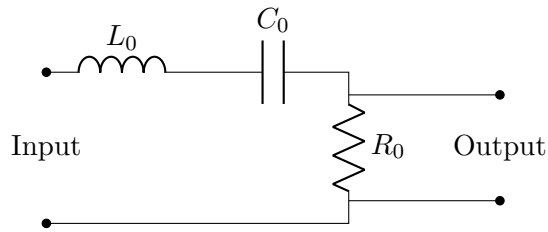
$$5 = R_1 + \frac{1}{\frac{1}{R_2} + j(0)} \quad (2.42)$$

$$\Rightarrow 5 = R_1 + R_2 \quad (2.43)$$



Hence, option (A) is correct.

2.5 In the bandpass filter circuit shown,  $R_0 = 50\Omega$ ,  $L_0 = 1mH$ ,  $C_0 = 10nF$ . The q factor of the filter is



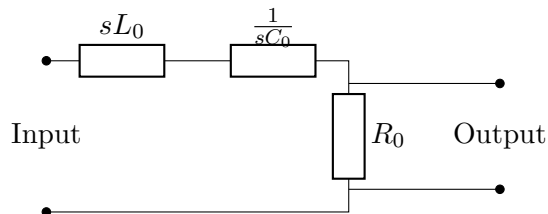
(GATE 2022 IN Q33)

**Solution:**

Variable	Description	Value
$R_0$	Resistance	$50\Omega$
$L_0$	Inductance	$1mH$
$C_0$	Capacitance	$10nF$
$\omega_0$	Resonant Angular Frequency	$\frac{1}{\sqrt{L_0 C_0}}$

Table 1: Variables and their description

The corresponding Laplace domain circuit is



Input  $X(s)$  can be written as

$$X(s) = I(s) \left( sL_0 + \frac{1}{sC_0} + R_0 \right) \quad (2.45)$$

Output  $Y(s)$  can be written as

$$Y(s) = I(s) R_0 \quad (2.46)$$

Transfer function  $H(s)$  can be written as

$$H(s) = \frac{Y(s)}{X(s)} \quad (2.47)$$

$$= \frac{sC_0R_0}{s^2C_0L_0 + C_0R_0s + 1} \quad (2.48)$$

substituting  $s = j\omega$

$$H(j, \omega) = \frac{j\omega C_0R_0}{-\omega^2 C_0L_0 + jC_0R_0\omega + 1} \quad (2.49)$$

$$\Rightarrow |H(j, \omega)| = \frac{\omega C_0R_0}{\sqrt{(1 - \omega^2 C_0L_0)^2 + (C_0R_0\omega)^2}} \quad (2.50)$$

Differentiating w.r.t  $\omega$  and equating to 0, we get

$$\begin{aligned} \frac{d|H(j, \omega)|}{d\omega} &= \frac{C_0R_0}{\sqrt{(1 - \omega^2 C_0L_0)^2 + (C_0R_0\omega)^2}} + \\ &\quad \frac{\omega C_0R_0}{2 \left( (1 - \omega^2 C_0L_0)^2 + (C_0R_0\omega)^2 \right)^{\frac{3}{2}}} \\ &\quad \left( 2\omega (C_0R_0)^2 - 2(1 - \omega^2 C_0L_0) 2\omega \right) = 0 \end{aligned} \quad (2.51)$$

$$\Rightarrow \omega_0 = \frac{1}{\sqrt{L_0C_0}} \quad (2.52)$$

from Table 1,

$$\omega_0 = 316227.76 \quad (2.53)$$

$Q$  – *factor* defined with reference to inductor

$$Q = \left| \frac{V_L}{V_R} \right|_{\omega_0} \quad (2.54)$$

$$= \frac{L_0 \omega_0}{R_0} \quad (2.55)$$

$$= \frac{1}{R_0} \sqrt{\frac{L_0}{C_0}} \quad (\text{from (2.52)}) \quad (2.56)$$

$Q$  – *factor* defined with reference to capacitor

$$Q = \left| \frac{V_C}{V_R} \right|_{\omega_0} \quad (2.57)$$

$$= \frac{1}{C_0 \omega_0 R_0} \quad (2.58)$$

$$= \frac{1}{R_0} \sqrt{\frac{L_0}{C_0}} \quad (\text{from (2.52)}) \quad (2.59)$$

Substituting the values from Table 1, we get

$$Q = 200 \quad (2.60)$$

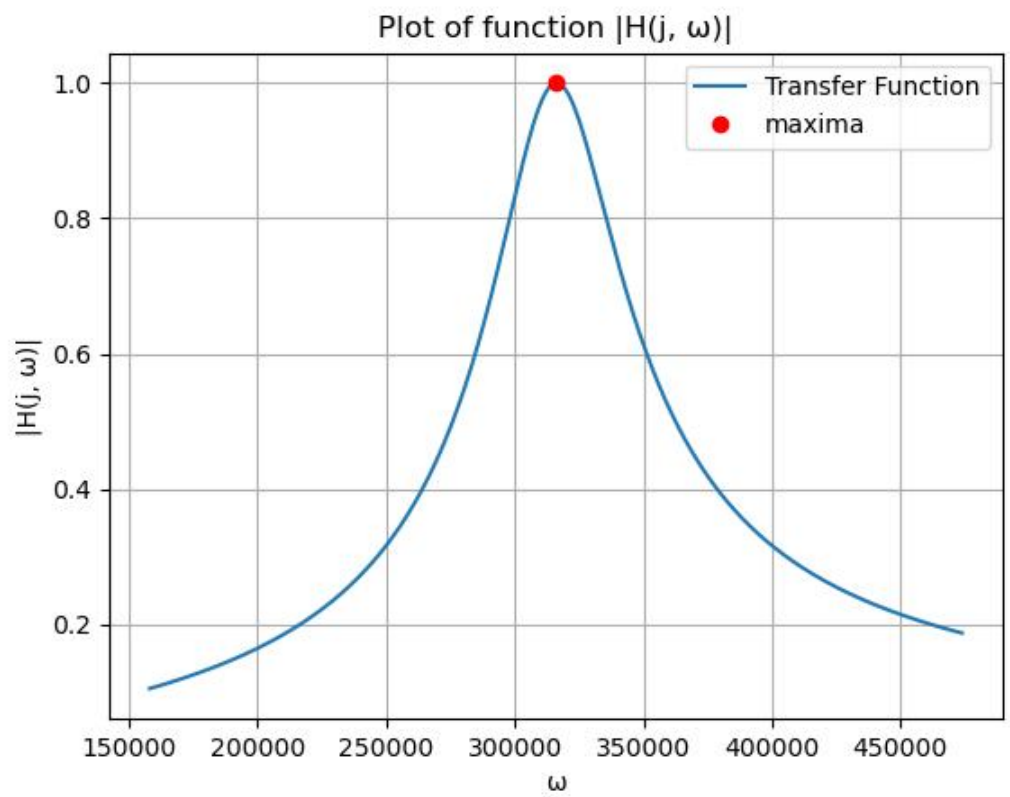


Figure 1: Transfer function  $|H(j, \omega)|$  taken from python3

- 2.6 In the circuit shown below, the switch S is closed at  $t = 0$ . The magnitude of the steady state voltage, in volts, across the  $6\Omega$  resistor is \_\_\_\_\_.(round off to two decimal places)
- (GATE 2022 EE Q31)

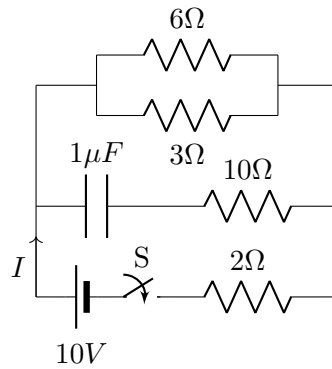


Figure 2.12:

**Solution:** Consider a sinusoidal input source of angular frequency  $\omega$ .

Symbol	Value	Description
$\omega$	0 for D.C.	Angular Frequency
$C$	$1\mu F$	Capacitance
$V_{in}(t)$	$10 \cos(\omega t)$	Input Voltage
$V_{out}(t)$		Output Voltage across $6\Omega$
$V_{out}(j\omega)$	$H(j\omega)V_{in}(j\omega)$	Output in Frequency Domain
$H(j\omega)$		Transfer Function
$I(j\omega)$		Total Current
$Z_{eff}$		Overall Impedance

Table 2.7: Given Parameters

Using KCL and KVL, we can calculate:

$$Z_{\text{eff}} = \frac{2 \left( 10 + \frac{1}{j\omega C} \right)}{12 + \frac{1}{j\omega C}} + 2 \quad (2.61)$$

$$\Rightarrow I(j\omega) = \frac{V_{in}}{\left( \frac{2 \left( 10 + \frac{1}{j\omega C} \right)}{12 + \frac{1}{j\omega C}} + 2 \right)} \quad (2.62)$$

$$\Rightarrow V_{out}(j\omega) = 2 \left[ \left( \frac{10 + \frac{1}{j\omega C}}{12 + \frac{1}{j\omega C}} \right) I(j\omega) \right] \quad (2.63)$$

$$= 2 \left[ \left( \frac{10 + \frac{1}{j\omega C}}{12 + \frac{1}{j\omega C}} \right) \frac{V_{in}(j\omega)}{\left( \frac{2 \left( 10 + \frac{1}{j\omega C} \right)}{12 + \frac{1}{j\omega C}} + 2 \right)} \right] \quad (2.64)$$

$$\Rightarrow H(j\omega) = \frac{1 + 10j\omega C}{2(1 + 11j\omega C)} \quad (2.65)$$

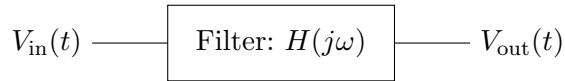


Figure 2.13: Filter Equivalent of Circuit

$$H(j\omega) = \left( \frac{\sqrt{1 + 100\omega^2 C^2}}{2\sqrt{1 + 121\omega^2 C^2}} \right) e^{j(\tan^{-1}(10\omega C) - \tan^{-1}(11\omega C))} \quad (2.66)$$

$$= \left( \frac{\sqrt{1 + 100\omega^2 C^2}}{2\sqrt{1 + 121\omega^2 C^2}} \right) e^{j \tan^{-1} \left( \frac{-\omega C}{1 + 110\omega^2 C^2} \right)} \quad (2.67)$$

$$\therefore V_{out}(t) = 10 |H(j\omega)| \cos(\omega t + \angle H(j\omega)) \quad (2.68)$$

$$= \frac{5\sqrt{1 + 100\omega^2 C^2}}{\sqrt{1 + 121\omega^2 C^2}} \cos \left( \omega t - \tan^{-1} \left( \frac{\omega C}{1 + 110\omega^2 C^2} \right) \right) \quad (2.69)$$

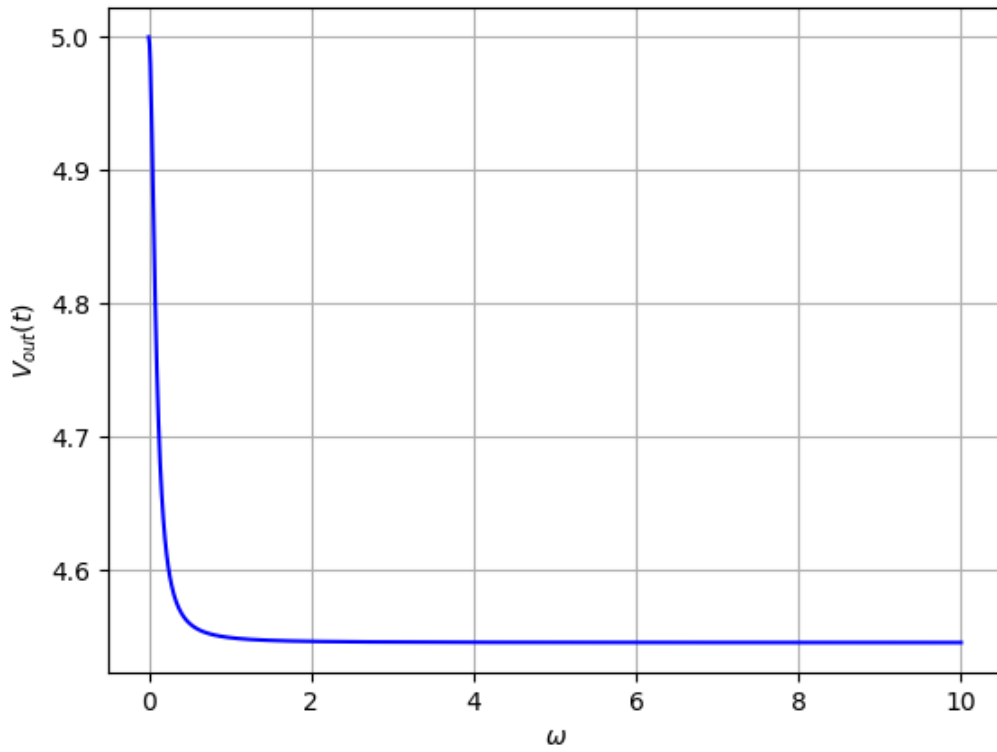


Figure 2.14: Plot of  $V_{out}(t)$  at  $t = 0$  w.r.t  $\omega$

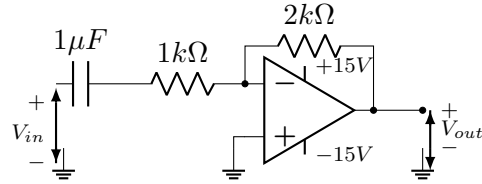
As  $\omega \rightarrow 0$ ,  $V_{in}(t)$  approaches being a D.C. input source (10V).

$\therefore$  substituting  $\omega = 0$ , we get:

$$V_{out}(t) = 5V \quad (2.70)$$



2.7 An ideal OPAMP circuit with a sinusoidal input is shown in the figure. The 3dB frequency is the frequency at which the magnitude of the voltage gain decreases by 3 dB from the maximum value. Which of the options is/are correct?



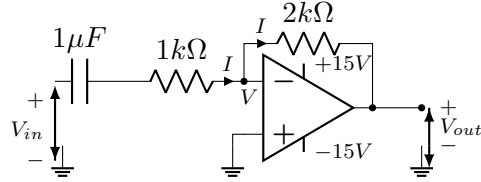
- (A) The circuit is a low pass filter.
- (B) The circuit is a high pass filter.
- (C) The 3 dB frequency is 1000rad/s.
- (D) The 3 dB frequency is  $\frac{1000}{3}$ rad/s.

(GATE EC 2022)

**Solution:**

Parameter	Description	Value
$V_{in}$	Input Voltage	—
$V_{out}$	Output Voltage	—
$C$	Capacitor	$1\mu F$
$R_1$	Resistance	$1k\Omega$
$R_2$	Feedback Resistance	$2k\Omega$
$V$	Voltage at Negative terminal	—
$V^+$	Voltage at positive terminal	0

Table 2.8: Input Parameters



$$\frac{V_{in} - V}{\frac{1}{sC} + R_1} = \frac{V - V_{out}}{R_2} \quad (2.71)$$

As Op-Amp is ideal

$$V = V^+ = 0V \quad (2.72)$$

$$\left| \frac{V_{out}}{V_{in}} \right| = \frac{sCR_2}{1 + sCR_1} \quad (2.73)$$

$$H(s) = \frac{sCR_2}{1 + sCR_1} \quad (2.74)$$

Keeping  $s = j\omega$

For determining nature of Filter

Put  $j\omega = 0$

$$H(j\omega) = 0 \quad (2.75)$$

Put  $j\omega \rightarrow \infty$

$$H(j\omega) = \frac{R_2}{R_1} = 2 \quad (\text{Finite}) \quad (2.76)$$

$\therefore$  It is high pass filter.

On simplifying (2.74) further

$$H(j\omega) = \frac{R_2}{R_1} \left( \frac{j\omega}{j\omega + \frac{1}{CR_1}} \right) \quad (2.77)$$

$$|H(j\omega)|_{max} = \frac{R_2}{R_1} \quad (2.78)$$

$$|H(j\omega)|_{\omega=\omega_c} = \frac{R_2}{R_1} \left| \frac{j\omega_c}{j\omega_c + \frac{1}{CR_1}} \right| \quad (2.79)$$

Given:

$$20 \log(|H(j\omega)|_{max}) - 20 \log(|H(j\omega)|_{\omega=\omega_c}) = 3dB \quad (2.80)$$

$$\frac{|H(j\omega)|_{max}}{|H(j\omega)|_{\omega=\omega_c}} = \sqrt{2} \quad (2.81)$$

From (2.78) and (2.79)

$$\frac{R_2}{R_1} \left| \frac{j\omega_c}{j\omega_c + \frac{1}{CR_1}} \right| = \frac{1}{\sqrt{2}} \frac{R_2}{R_1} \quad (2.82)$$

$$\left| \frac{j\omega_c}{j\omega_c + \frac{1}{CR_1}} \right| = \frac{1}{\sqrt{2}} \quad (2.83)$$

$$\implies \omega_c = \frac{1}{CR_1} \quad (2.84)$$

From Table 2.8

$$\omega_c = 1000 \text{ rad/s} \quad (2.85)$$

Where  $\omega_c$  is 3 dB frequency.

Finally, Correct options are (B) and (C).

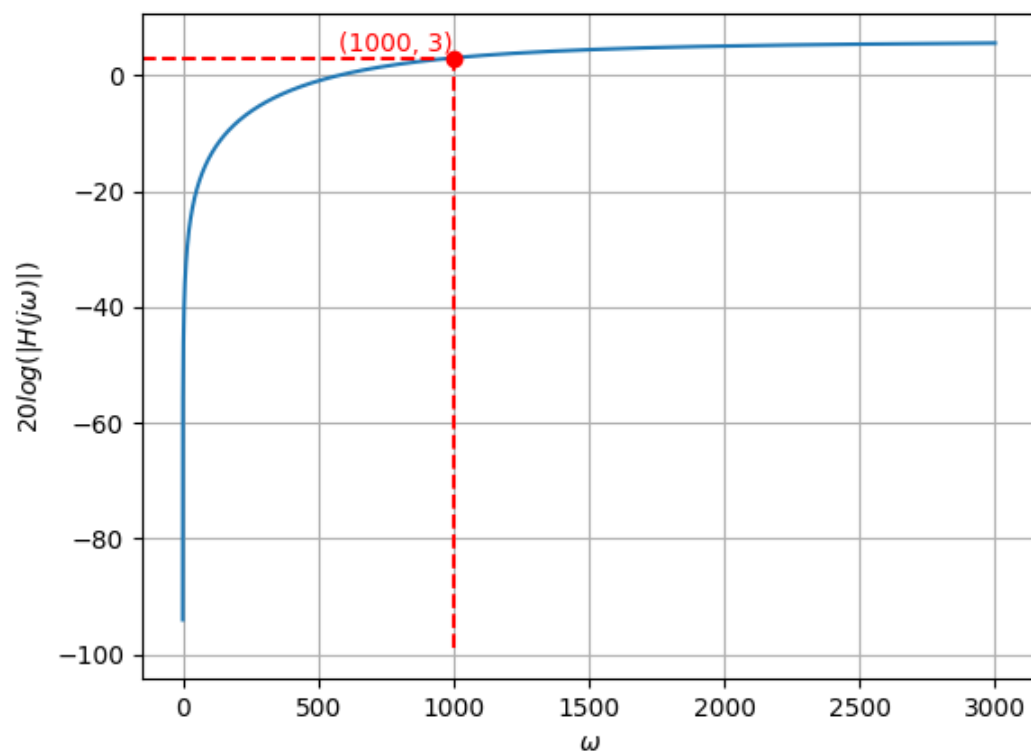


Figure 2.15: Frequency response plot

## Chapter 3

# Z-transform

3.1 Consider the following recursive iteration scheme for different values of variable  $P$  with the initial guess  $x_1 = 1$ :

$$x_{n+1} = \frac{1}{2} \left( x_n + \frac{P}{x_n} \right), \quad n = 1, 2, 3, 4, 5$$

For  $P = 2$ ,  $x_5$  is obtained to be 1.414, rounded off to 3 decimal places. For  $P = 3$ ,  $x_5$  is obtained to be 1.732, rounded off to 3 decimal places.

If  $P = 10$ , the numerical value of  $x_5$  is \_\_\_\_\_. (*round off to three decimal places*)  
(GATE CE 2022)

**Solution:**

Applying  $A.M \geq G.M$  inequality,

$$\frac{x_n + \frac{P}{x_n}}{2} \geq \sqrt{P} \tag{3.1}$$

$$\implies x_{n+1} \geq \sqrt{P} \tag{3.2}$$

Solving the equation,

$$2x_{n+1}x_n - x_n^2 - P = 0 \quad (3.3)$$

Applying  $Z$ -transform we get,

$$X(z) * X(z) = \frac{PZ^{-1}}{(1 - z^{-1})(2 - z^{-1})} \quad (3.4)$$

$$= P \left( \frac{z^{-1}}{1 - z^{-1}} - \frac{z^{-1}}{2 - z^{-1}} \right) \quad (3.5)$$

From the transformation pairs,

$$x_{n-a} \xleftrightarrow{\mathcal{Z}} z^{-a} X(z) \quad (3.6)$$

$$x_{n_1} \times x_{n_2} \xleftrightarrow{\mathcal{Z}} X_1(z) * X_2(z) \quad (3.7)$$

$$\frac{u(n-1)}{a^n} \xleftrightarrow{\mathcal{Z}} \frac{z^{-1}}{a - z^{-1}} \quad (3.8)$$

Now, applying inverse  $Z$ -transform,

$$x_n^2 = P \left( u(n-1) - \frac{u(n-1)}{2^n} \right) \quad (3.9)$$

$$\Rightarrow x_n^2 = P \left( 1 - \frac{1}{2^n} \right) \quad [\because n \geq 1] \quad (3.10)$$

Similarly,

$$x_{n+1}^2 = P \left( 1 - \frac{1}{2^{n+1}} \right) \quad (3.11)$$

$$\Rightarrow \lim_{n \rightarrow \infty} \frac{x_{n+1}}{x_n} = \lim_{n \rightarrow \infty} \sqrt{\frac{P \left( 1 - \frac{1}{2^n} \right)}{P \left( 1 - \frac{1}{2^{n+1}} \right)}} \quad (3.12)$$

$$= 1 \quad (3.13)$$

Hence, the system is convergent.

Now finding the limit of the sequence,

$$x^2 = \lim_{x \rightarrow \infty} P \left( 1 - \frac{1}{2^n} \right) \quad (3.14)$$

$$\implies x = \pm \sqrt{P} \quad (3.15)$$

From (3.2) and (3.15),

$$x_{n+1} = \sqrt{P} \quad (3.16)$$

Therefore, for  $P = 10$  the value of  $x_5$  is,

$$x_5 = \sqrt{10} \quad (3.17)$$

$$\therefore x_5 = 3.162 \quad (3.18)$$

3.2 The block diagram of a two-tap high-pass FIR filter is shown below. The filter transfer function is given by  $H(z) = Y(z)/X(z)$ .

If the ratio of maximum to minimum value of  $H(z)$  is 2 and  $|H(z)|_{max} = 1$ , the coefficients  $\beta_0$  and  $\beta_1$  are \_\_\_\_\_ and \_\_\_\_\_, respectively.

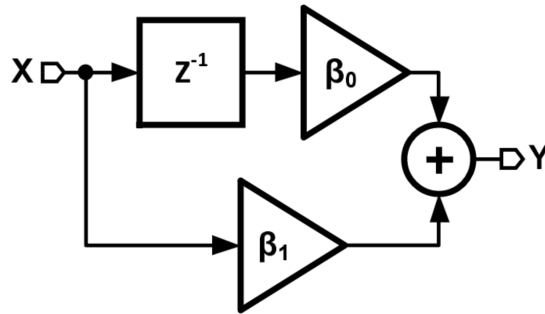


Figure 3.1: Block diagram

- (A) 0.75, -0.25
- (B) 0.67, 0.33
- (C) 0.60, -0.40
- (D) -0.64, 0.36

GATE BM 2022

**Solution:**

**Results and Proofs:**

Time Shift Property:

$$x(n) \xleftrightarrow{Z} X(z) \quad (3.19)$$

$$x(n - n_0) \xleftrightarrow{Z} z^{-n_0} X(z) \quad (3.20)$$

Proof:



Let

$$y(n) = x(n - n_0) \quad (3.21)$$

Taking z-transform

$$\mathcal{Z}(y(n)) = \mathcal{Z}(x(n - n_0)) \quad (3.22)$$

$$(3.23)$$

Simplifying LHS

$$Y(z) = \sum_{n=-\infty}^{\infty} y(n)z^{-n} \quad (3.24)$$

From (3.21)

$$Y(z) = \sum_{n=-\infty}^{\infty} x(n - n_0)z^{-n} \quad (3.25)$$

Let

$$n - n_0 = s \quad (3.26)$$

$$\implies n = s + n_0 \quad (3.27)$$

From (3.25) and (3.27)

$$Y(z) = \sum_{s=-\infty}^{\infty} x(s)z^{-(s+n_0)} \quad (3.28)$$

$$= z^{-n_0} \sum_{s=-\infty}^{\infty} x(s)z^{-s} \quad (3.29)$$

As variable in Z-transform is dummy, on replacing it, we get

$$Y(z) = z^{-n_0} \sum_{n=-\infty}^{\infty} x(n)z^{-n} \quad (3.30)$$

$$= z^{-n_0} X(z) \quad (3.31)$$

From (3.22) and (3.31)

$$\mathcal{Z}(x(n - n_0)) = z^{-n_0} X(z) \quad (3.32)$$

Hence proved

Result:

$$z^{-n_0} X(z) \xleftrightarrow{\mathcal{Z}^-} x(n - n_0) \quad (3.33)$$

**Sol:**

Variable	Description	Value
$H(z)$	Transfer Function	$\beta_0 z^{-1} + \beta_1$
$ H(z) _{max}$	Maximum value of Transfer Function	1
$ H(z) _{min}$	Minimum value of Transfer Function	$\frac{1}{2}$

Table 3.1: input parameters

In (3.33), put

$$n_0 = 1, \quad x(n) = \delta(n)$$

Since

$$1 \xleftrightarrow{\mathcal{Z}^-} \delta(n)$$

$$z^{-1} \xleftrightarrow{\mathcal{Z}^-} \delta(n-1) \quad (3.34)$$

This is a unit delay in discrete time and represents unit amplitude sinusoidal signal.  
So,

$$z^{-1} = e^{-jw} \quad (3.35)$$

$$\implies |z^{-1}| = 1 \quad (3.36)$$

Since  $H(z)$  is complex, on using Triangle Inequality, we get

$$|x + y| \leq |x| + |y| \quad (3.37)$$

And its corollary

$$||x| - |y|| \leq |x + y| \quad (3.38)$$

where x and y are complex numbers.

$$||z^{-1}\beta_0| - |\beta_1|| \leq |z^{-1}\beta_0 + \beta_1| \leq |z^{-1}\beta_0| + |\beta_1| \quad (3.39)$$

From Table 3.1

$$||z^{-1}\beta_0| - |\beta_1|| \leq |H(z)| \leq |z^{-1}\beta_0| + |\beta_1| \quad (3.40)$$

From (3.36)

$$||\beta_0| - |\beta_1|| \leq |H(z)| \leq |\beta_0| + |\beta_1| \quad (3.41)$$

So, we can conclude that

$$|H(z)|_{max} = |\beta_0| + |\beta_1| \quad (3.42)$$

Now from Table 3.1

$$1 = |\beta_0| + |\beta_1| \quad (3.43)$$

Similarly,

$$\frac{1}{2} = ||\beta_0| - |\beta_1|| \quad (3.44)$$

On solving (3.43) and (3.44), we get

$$|\beta_0| = 0.75, |\beta_1| = 0.25 \quad (3.45)$$

OR

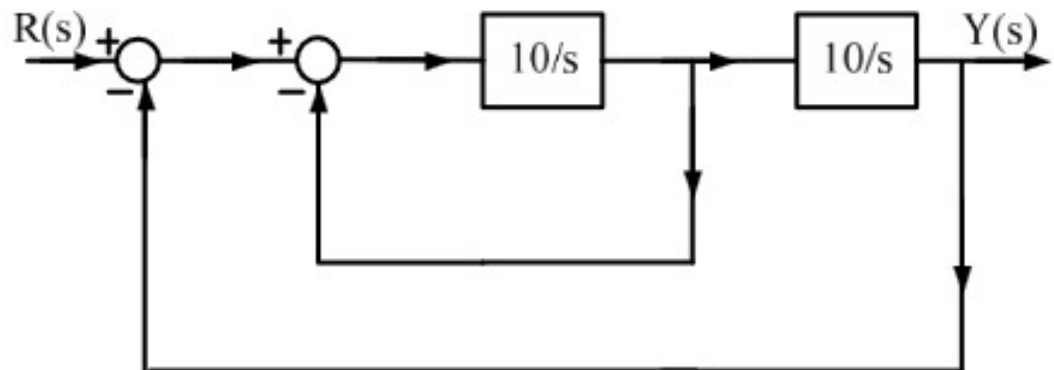
$$|\beta_0| = 0.25, |\beta_1| = 0.75 \quad (3.46)$$

Hence the correct answer is option (A)

## Chapter 4

## Systems

4.1 The damping ratio and undamped natural frequency of a closed loop system as shown in the figure, are denoted as  $\zeta$  and  $\omega_n$ , respectively. The values of  $\zeta$  and  $\omega_n$  are



- (a)  $\zeta = 0.5$  and  $\omega_n = 10$  rad/s
- (b)  $\zeta = 0.1$  and  $\omega_n = 10$  rad/s
- (c)  $\zeta = 0.707$  and  $\omega_n = 10$  rad/s
- (d)  $\zeta = 0.707$  and  $\omega_n = 100$  rad/s

(GATE EE 2022) **Solution:**

We will use Mason's Gain Formula to calculate the transfer function of this system.

Parameter	Description	Values
m	load of system	
k	stiffness of system	
$\omega_n$	Natural frequency	$\sqrt{\frac{k}{m}}$
$\zeta$	Damping ratio	$\frac{c}{2m\omega_n}$
$y(t)$	Output of system	
$x(t)$	Input to the system	
c	Damping coefficient	
$T(s)$	Transfer function of system	$\frac{Y(s)}{R(s)}$

Table 4.1: Parameter Table

First converting the given diagram to a signal flow graph :

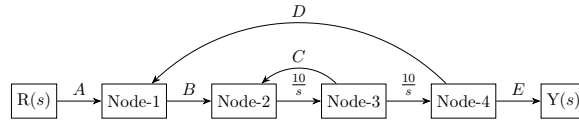


Figure 4.1: Signal Flow Diagram

Mason's Gain Formula is given by :

$$H(s) = \sum_{i=1}^N \left( \frac{P_i \Delta_i}{\Delta} \right) \quad (4.1)$$

This signal flow graph has only one forward path whose gain is given by :

$$P_1 = \frac{10}{s} \frac{10}{s} \quad (4.2)$$

$$= \frac{100}{s^2} \quad (4.3)$$

Parameter	Description
N	Number of forward paths
L	Number of loops
$P_k$	Forward path gain of $k^{th}$ path
$\Delta_k$	Associated path factor
$\Delta$	Determinant of the graph

Table 4.2: Parameter Table - Mason's Gain Law

Parameter	Formula
$\Delta$	$1 + \sum_{k=1}^L \left( (-1)^k \text{Product of gain of groups of k isolated loops} \right)$
$\Delta_k$	$\Delta$ part of graph that is not touching $k^{th}$ forward path

Table 4.3: Formula Table - Mason's Gain Law

The loop gain for loop between Node-2 and Node-3 is :

$$L_1 = \frac{10}{s} (-1) \quad (4.4)$$

$$= -\frac{10}{s} \quad (4.5)$$

The loop gain for loop between Node-1 and Node-4 is :

$$L_1 = \frac{10}{s} \frac{10}{s} (-1) \quad (4.6)$$

$$= -\frac{100}{s^2} \quad (4.7)$$

Using Table 4.3,  $\Delta$  is :

$$\Delta = 1 - \left( -\frac{10}{s} - \frac{100}{s^2} \right) \quad (4.8)$$

$$= 1 + \frac{10}{s} + \frac{100}{s^2} \quad (4.9)$$

There are no two isolated loops available. Hence all further terms will be zero.

As both the loops are in contact with the only forward path,

$$\Delta_1 = 1 \quad (4.10)$$

Using equation (4.1) :

$$H(s) = \frac{\frac{100}{s^2}}{1 + \frac{10}{s} + \frac{100}{s^2}} \quad (4.11)$$

$$= \frac{100}{s^2 + 10s + 100} \quad (4.12)$$

Referring to Table 4.1, the general equation of the damping system is second order and can be written as :

$$m\ddot{y}(t) + c\dot{y}(t) + ky(t) = x(t) \quad (4.13)$$

Take the Laplace transform and solve for  $\frac{Y(s)}{X(s)}$  :

$$\frac{Y(s)}{X(s)} = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (4.14)$$

$$\Rightarrow H(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (4.15)$$



Comparing equations (4.12) and (4.15) ,

$$\omega_n^2 = 100 \quad (4.16)$$

$$\Rightarrow \omega_n = 10 \text{ rad/s} \quad (4.17)$$

$$2\zeta\omega_n = 10 \quad (4.18)$$

$$\Rightarrow \zeta = 0.5 \quad (4.19)$$

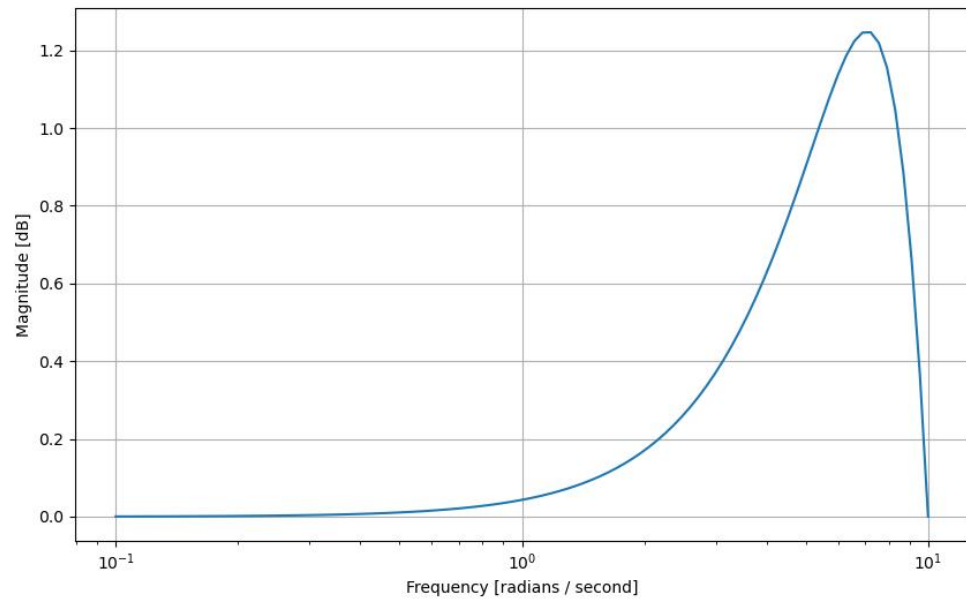


Figure 4.2: Magnitude plot

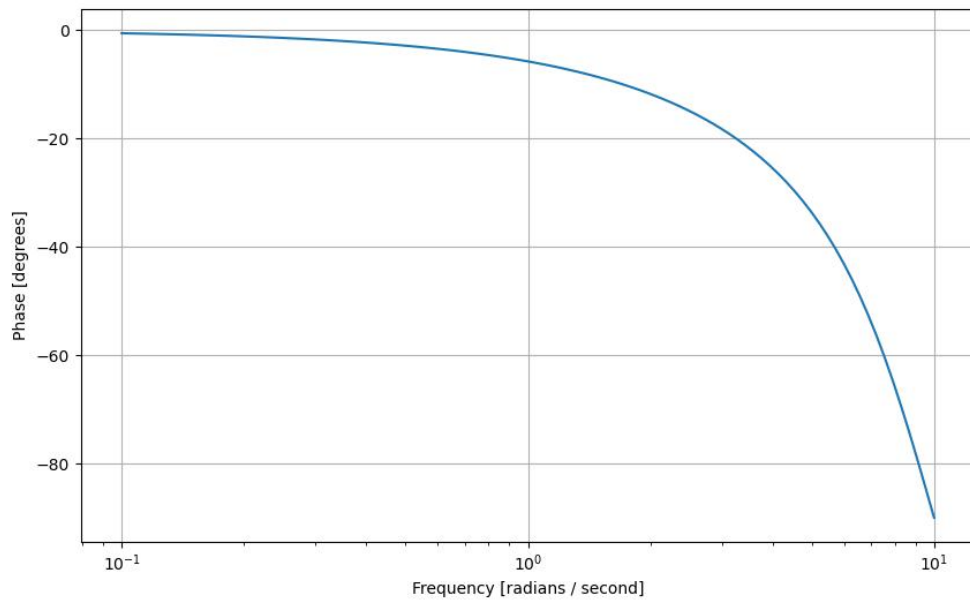
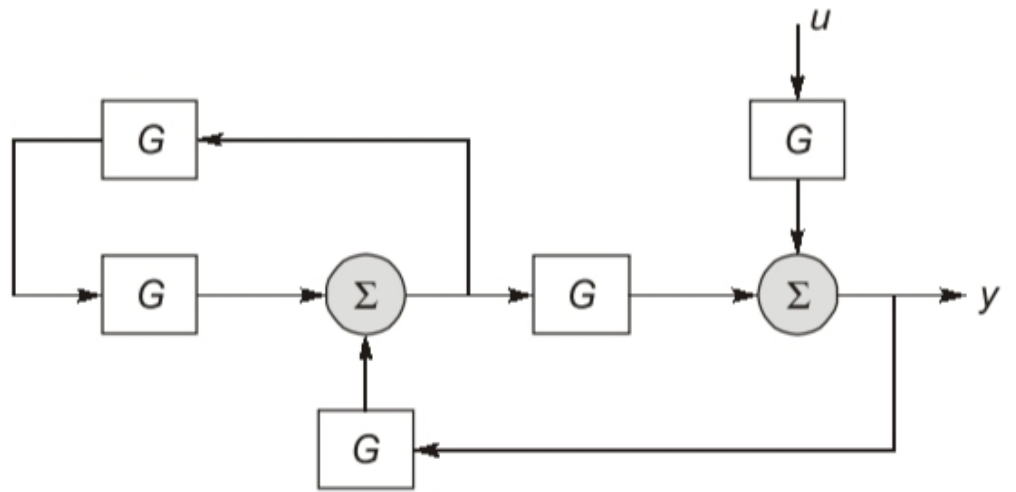


Figure 4.3: Phase plot

4.2 In the block diagram shown in the figure, the transfer function  $G = \frac{K}{\tau s + 1}$  with  $K > 0$  and  $\tau > 0$ . The maximum value of  $K$  below which the system remains stable is \_\_\_\_\_(rounded off to two decimal places) (GATE CH 2022)

**Solution:**



Parameter	Value	Description
$G$	$\frac{K}{\tau s + 1}$	Transfer function shown in blocks
$Y$		Laplace transform of $y$ (output)
$U$		Laplace transform of $u$ (input)
$X, Z$		Laplace transform of $x$ and $z$
$T$	$\frac{Y}{U}$	Transfer function of complete system

Table 4.4: Parameters

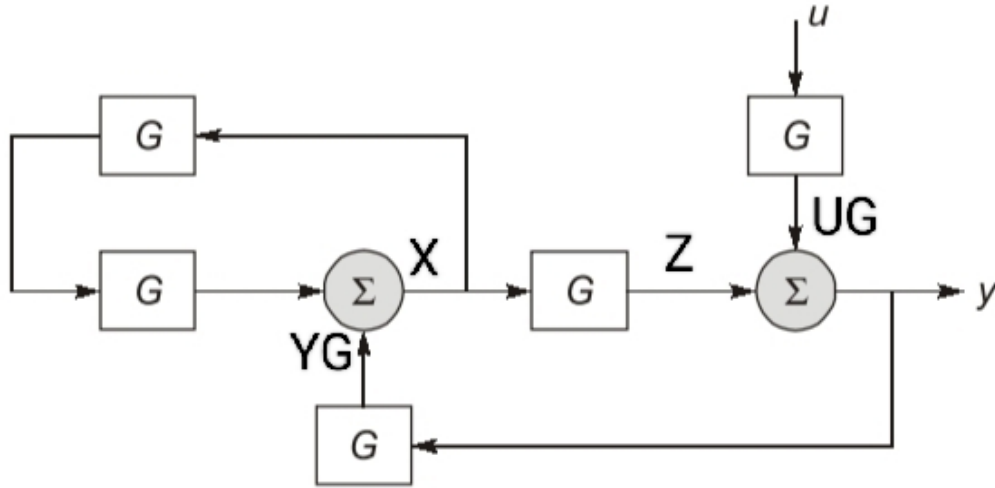


Figure 4.4: Block Diagram

$$X = XG^2 + YG \quad (4.20)$$

$$\Rightarrow X = \frac{YG}{1 - G^2} \quad (4.21)$$

$$Z = XG \quad (4.22)$$

$$Y = Z + UG \quad (4.23)$$

$$Y = XG + UG \quad (4.24)$$

$$Y = \frac{YG^2}{1 - G^2} + UG \quad (4.25)$$

$$\Rightarrow Y = \frac{UG(1 - G^2)}{1 - 2G^2} \quad (4.26)$$

From Table 4.4,

$$T = \frac{G(1 - G^2)}{1 - 2G^2} \quad (4.27)$$

$$= \frac{K \left(1 - \frac{K^2}{(\tau s + 1)^2}\right)}{\left(1 - \frac{2K^2}{(\tau s + 1)^2}\right) (\tau s + 1)} \quad (4.28)$$

$$= \frac{K(\tau^2 s^2 + 2\tau s + 1 - K^2)}{\tau^3 s^3 + 3\tau^2 s^2 + (3\tau - 2K^2\tau)s + 1 - 2K^2} \quad (4.29)$$

So, Characteristic equation :

$$\tau^3 s^3 + 3\tau^2 s^2 + (3\tau - 2K^2\tau)s + 1 - 2K^2 = 0 \quad (4.30)$$

For a characteristic equation  $a_0 s^n + a_1 s^{n-1} + a_2 s^{n-2} + \dots a_n = 0$ ,

$s^n$	$a_0$	$a_2$	$a_4$	...
$s^{n-1}$	$a_1$	$a_3$	$a_5$	...
$s^{n-2}$	$b_1 = \frac{a_1 a_2 - a_3 a_0}{a_1}$	$b_2 = \frac{a_1 a_4 - a_5 a_0}{a_1}$	...	..
$s^{n-3}$	$c_1 = \frac{b_1 a_3 - b_2 a_1}{b_1}$	$\vdots$		
$\vdots$	$\vdots$	$\vdots$		
$s^1$	$\vdots$	$\vdots$		
$s^0$	$a_n$			

Table 4.5: Routh Array

From Table 4.5:

$s^3$	$\tau^3$	$3\tau - 2K^2\tau$
$s^2$	$3\tau^2$	$1 - 2K^2$
$s^1$	$\frac{8}{3}\tau(1 - K^2)$	0
$s^0$	$1 - 2K^2$	

Table 4.6:

Given  $\tau > 0$  and  $K > 0$ , for system to be stable,

$$1 - K^2 > 0 \tag{4.31}$$

$$1 - 2K^2 > 0 \tag{4.32}$$

$$\implies 0 < K < \frac{1}{\sqrt{2}} \tag{4.33}$$

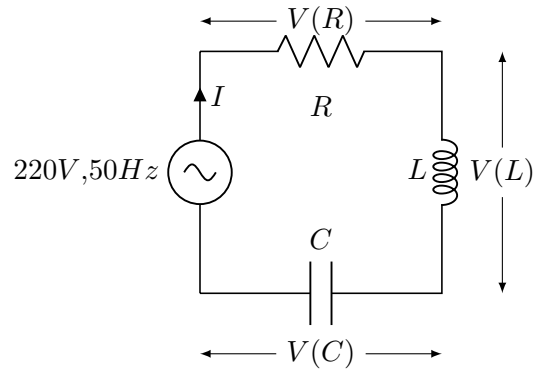
$$K_{max} \approx 0.71 \tag{4.34}$$

- 4.3 A series RLC circuit is connected to 220 V, 50 Hz supply. For a fixed a value of R and C, the inductor L is varied to deliver the maximum current. This value 0.4A and the corresponding potential drop across the capacitor is 330 V. The value of the inductor L is ? (Rounded off to two decimal places). (GATE BM 2022)

**Solution:**

Symbols	Description	Values
$V_s$	Input voltage	220 V and 50Hz
$\chi_L$	Impedance across inductor	$j\omega L$
$\chi_C$	Impedance across capacitor	$\frac{-j}{\omega C}$
$Z$	Impedance across the entire circuit	$R + j\omega L + \frac{-j}{\omega C}$

Table 4.7: Parameters, Descriptions, and Values



During maximum current  $|Z|$  is minimum .

$$I = \frac{V_s}{Z} \quad (4.35)$$

$$= \frac{V_s}{R + \chi_L + \chi_C} \quad (4.36)$$

$$= \frac{V_s}{R + j\omega L + \frac{1}{j\omega C}} \quad (4.37)$$

$$|I| = \frac{|V_s|}{\sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}} \quad (4.38)$$

Varying  $L$  for maximum value of  $I$  :

$$\omega L = \frac{1}{\omega C} \quad (4.39)$$

Putting in (4.37):

$$I_{max} = \frac{V_s}{R} \quad (4.40)$$

$I_{max}$  has same phase as  $V_s$  (Assume  $\angle\phi$ ). For impedance across the capacitor :

$$V_C|_{I=I_{max}} = I_{max}\chi_C \quad (4.41)$$

$$-330\angle(90 + \phi) = (0.4\angle\phi)\chi_C \quad (4.42)$$

$$-330\angle90 = 0.4\chi_C \quad (4.43)$$

$$\implies \chi_C = -825j\Omega \quad (4.44)$$

For value of Capacitor and inductor, using (4.39) :

$$L = \frac{825}{100\pi}H \quad (4.45)$$

$$\approx 2.63H \quad (4.46)$$

$$C = 3.858 * 10^{-6}F \quad (4.47)$$



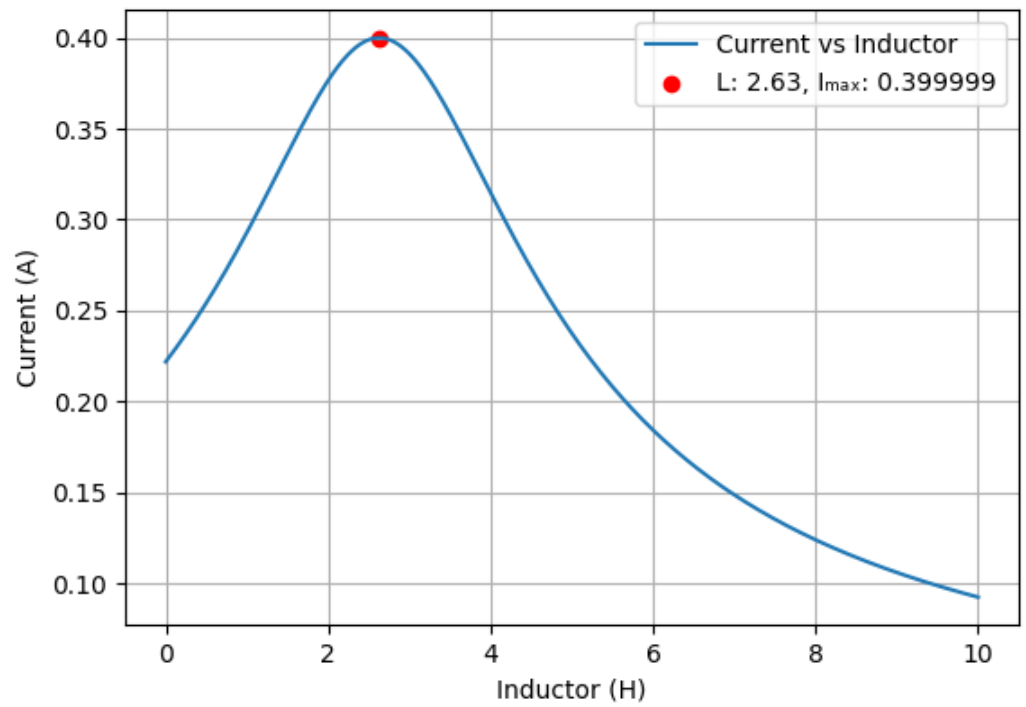


Figure 4.5:  $I$  vs  $L$

4.4 The open loop transfer function of a unity gain negative feedback system is given by

$G(s) = \frac{k}{s^2 + 4s - 5}$ . The range of  $k$  for which the system is stable, is (GATE EE 2022)

**Solution:**

Variable	Description	value
$G(s)$	Open loop transfer function	$\frac{k}{s^2 + 4s - 5}$
$1 + G(s)$	Characteristic equation	0

Table 4.8: A Table with input parameters

from Table4.8

Characteristic equation:

$$1 + G(s) = 0 \quad (4.48)$$

$$\Rightarrow 1 + \frac{k}{s^2 + 4s - 5} = 0 \quad (4.49)$$

$$\Rightarrow s^2 + 4s + (k - 5) = 0 \quad (4.50)$$

By routh table analysis, for a stable system:

$$\begin{array}{c|cc} s^2 & 1 & k-5 \\ s^1 & 4 & 0 \\ s^0 & \frac{4(k-5)-0}{4} & 0 \end{array}$$

$$\frac{4(k-5)-0}{4} > 0 \quad (4.51)$$

$$k - 5 > 0 \quad (4.52)$$

$$\Rightarrow k > 5 \quad (4.53)$$

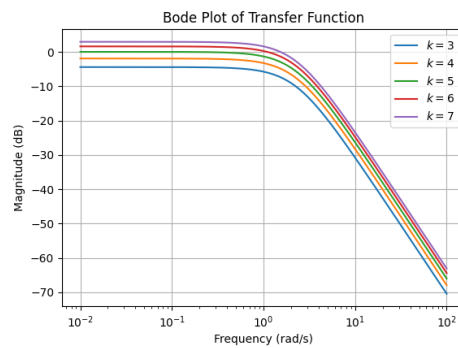


Figure 4.6: Graph showing  $k < 5, k = 5, k > 5$

For an open transfer function to be stable, its magnitude in the bode plot should be

positive for some positive frequency.

In the below graph we can observe that the above condition satisfies for  $k > 5$ .

4.5 The signal flow graph of a system is shown. The expression for  $\frac{Y(s)}{X(s)}$  is

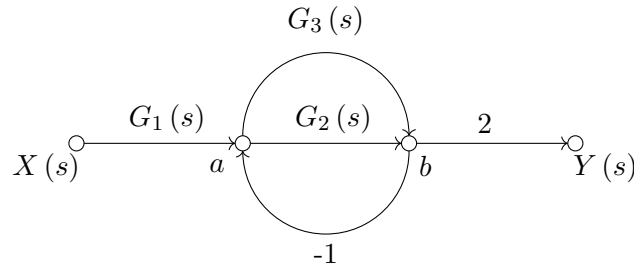


Figure 4.7: Signal Flow Graph of the System

- (a)  $\frac{2G_1(s)G_2(s)+2G_1(s)G_3(s)}{1+G_2(s)+G_3(s)}$
- (b)  $2 + G_1(s) + G_3(s) + \frac{G_2(s)}{1+G_2(s)}$
- (c)  $G_1(s) + G_3(s) - \frac{G_2(s)}{2+G_2(s)}$
- (d)  $\frac{2G_1(s)G_2(s)+2G_1(s)G_3(s)-G_1(s)}{1+G_2(s)+G_3(s)}$

(GATE 2022 IN Question 37)

**Solution:**

Parameter	Description	Value
$Y(s)$	Output node variable	
$X(s)$	Input node variable	
$\frac{Y(s)}{X(s)}$	Transfer function	?
$P_1$	Forward Path Gain a-b through $G_2(s)$	$2G_1(s) G_2(s)$
$P_2$	Forward Path Gain a-b through $G_3(s)$	$2G_1(s) G_3(s)$
$\Delta_1$	Determinant of Forward Path a-b through $G_2(s)$	1
$\Delta_2$	Determinant of Forward Path a-b through $G_3(s)$	1
$L_1$	Gain of Loop a-b through $G_2(s)$ and back	$-G_2(s)$
$L_2$	Gain of Loop a-b through $G_3(s)$ and back	$-G_3(s)$
$\Delta$	Determinant of System	$1 + G_2(s) + G_3(s)$
$n$	Number of forward paths	2

Table 4.9: Variables Used

$$P_1 = (G_1(s))(G_2(s))(2) = 2G_1(s)G_2(s) \quad (4.54)$$

$$P_2 = (G_1(s))(G_3(s))(2) = 2G_1(s)G_3(s) \quad (4.55)$$

$$\Delta_1 = 1 - (0) = 1 \quad (4.56)$$

$$\Delta_2 = 1 - (0) = 1 \quad (4.57)$$

$$L_1 = -G_2(s) \quad (4.58)$$

$$L_2 = -G_3(s) \quad (4.59)$$

$$\Delta = 1 - (L_1 + L_2) = 1 + G_1(s) + G_2(s) \quad (4.60)$$

From 4.7, Using Mason's Gain Formula,

$$\frac{Y(s)}{X(s)} = \frac{\sum_{i=1}^n P_i \Delta_i}{\Delta} \quad (4.61)$$

$$= \frac{P_1 \Delta_1 + P_2 \Delta_2}{\Delta} \quad (4.62)$$

$$= \frac{2G_1(s)G_2(s)(1) + 2G_1(s)G_3(s)(1)}{1 + G_2(s) + G_3(s)} \quad (4.63)$$

$$\Rightarrow \frac{Y(s)}{X(s)} = \frac{2G_1(s)G_2(s) + 2G_1(s)G_3(s)}{1 + G_2(s) + G_3(s)} \quad (4.64)$$

4.6 The output of the system  $y(t)$  is related to its input  $x(t)$  according to the relation

$y(t) = x(t) \sin(2\pi t)$ . This system is

- (A) Linear and time-variant
- (B) Non-Linear and time-invariant
- (C) Linear and time-invariant
- (D) Non-linear and time-variant

(GATE 2022 IN Question 14) **Solution:**

Symbol	Value	Description
$x(t)$		input signal
$y(t)$	$x(t) \sin(2\pi t)$	output signal
$\tau$		Time delay

Table 1: input parameters

From Table 1

$$y_1(t) \leftrightarrow x_1(t) \quad (4.65)$$

$$y_2(t) \leftrightarrow x_2(t) \quad (4.66)$$

$$ay_1(t) + by_2(t) \leftrightarrow ax_1(t) + bx_2(t) \quad (4.67)$$

$$ay_1(t) + by_2(t) = (ax_1(t) + bx_2(t)) \sin(2\pi t) \quad (4.68)$$

$\therefore$  satisfies principle of superposition

$$ky(t) \leftrightarrow kx(t) \quad (4.69)$$

$$ky(t) = k(x(t) \sin(2\pi t)) \quad (4.70)$$

$\therefore$  satisfies principle of homogeneity

$\therefore$  it is linear

Delay in input  $x(t)$ :

$$y_1(t) = x(t - \tau) \sin(2\pi t) \quad (4.71)$$

Delay in output  $y(t)$ :

$$y(t - \tau) = x(t - \tau) \sin(2\pi(t - \tau)) \quad (4.72)$$

$$y_2(t) = x(t - \tau) \sin(2\pi(t - \tau)) \quad (4.73)$$

$$y_1(t) \neq y_2(t) \quad (4.74)$$

$\therefore$  it is time variant

$\therefore$  (A) linear and time variant

4.7 Two linear time-invariant systems with transfer functions

$$G_1(s) = \frac{10}{s^2 + s + 1}$$

and

$$G_2(s) = \frac{10}{s^2 + s\sqrt{10} + 10}$$

have unit step responses  $y_1(t)$  and  $y_2(t)$ , respectively. Which of the following statements is/are true?

- (a)  $y_1(t)$  and  $y_2(t)$  have the same percentage peak overshoot.
- (b)  $y_1(t)$  and  $y_2(t)$  have the same steady state values.
- (c)  $y_1(t)$  and  $y_2(t)$  have the same damped frequency of oscillation.
- (d)  $y_1(t)$  and  $y_2(t)$  have the same 2% settling time.

(GATE 2022 EC Q50)

**Solution:** The general second-order transfer function is given by:

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (4.75)$$

After comparing the coefficients of  $G_1(s)$  and  $G_2(s)$ , as  $\zeta = \frac{1}{2}$  is less than 1, the



Parameter	Description	value
$X_1(s)$	input	$\frac{1}{s}$
$X_2(s)$	input	$\frac{1}{s}$
$G_1(s)$	transfer function	$\frac{10}{s^2+s+1}$
$G_2(s)$	transfer function	$\frac{10}{s^2+s\sqrt{10}+10}$
$y_1(t)$	unit step response	—
$y_2(t)$	unit step response	—
$\omega_n$	natural frequency	—
$\zeta$	damping ratio	—

Table 4.11: Given Parameters

Transfer function	$\omega_n$	$\zeta$
$G_1(s)$	1	$\frac{1}{2}$
$G_2(s)$	$\sqrt{10}$	$\frac{1}{2}$

Table 4.12: Given Parameters

system is underdamped.

$$Y(s) = X(s)G(s) \quad (4.76)$$

$$= \frac{1}{s} \left( \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \right) \quad (4.77)$$

Applying inverse laplace transform,

$$y(t) = 1 - \frac{e^{-\zeta\omega_n t}}{1 - \zeta^2} \sin(\omega_d t + \phi) \quad (4.78)$$

where  $\omega_d$  is the damped frequency of oscillation.

$$\omega_d = \omega_n \sqrt{1 - \zeta^2} \quad (4.79)$$

The percentage peak overshoot ( $PO$ ):

$$PO = \left( \frac{y_{\max} - y_{ss}}{y_{ss}} \right) \times 100\% \quad (4.80)$$

$y_{\max}$  is obtained by differentiating (4.78) with respect to time and equating it to zero, substituting the value in (4.78),

$$y_{\max} = 1 + \frac{1}{\sqrt{1 - \zeta^2}} \quad (4.81)$$

$y_{ss}$  is obtained by final value theorem,

$$y_{ss} = \lim_{s \rightarrow 0} sY(s) \quad (4.82)$$

$$= \lim_{s \rightarrow 0} s \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \frac{1}{s} \quad (4.83)$$

$$= 1 \quad (4.84)$$

Substituting the values of  $y_{\max}$  and  $y_{ss}$  in (4.80),

$$PO = \frac{1}{\sqrt{1 - \zeta^2}} \times 100\% \quad (4.85)$$

$y_1(t)$  and  $y_2(t)$  have same  $\zeta$ , they have same percentage peak overshoot. So, option (1) is correct.

The steady state value of  $y(t)$  is given by final value theorem:

$$y_{1ss} = \lim_{s \rightarrow 0} sY_1(s) \quad (4.86)$$

$$= \lim_{s \rightarrow 0} s \frac{10}{s^2 + s + 1} \frac{1}{s} \quad (4.87)$$

$$= 10 \quad (4.88)$$

$$y_{2ss} = \lim_{s \rightarrow 0} sY_2(s) \quad (4.89)$$

$$= \lim_{s \rightarrow 0} s \frac{10}{s^2 + s\sqrt{10} + 10} \frac{1}{s} \quad (4.90)$$

$$= 1 \quad (4.91)$$

as both the unit step responses have different steady state values, option (2) is incorrect.

From (4.80), as  $\omega_n$  is different for  $y_1(t)$  and  $y_2(t)$ , they have different damped frequency of oscillation. Hence option (3) is incorrect.

Settling time  $T_s$ :

$$T_s = \frac{4}{\zeta\omega_n} \quad (4.92)$$

As,  $\omega_n$  is different for  $y_1(t)$  and  $y_2(t)$ , they have different 2% settling time, Hence option (4) is incorrect.

So, only option (1) is correct.

4.8 Consider a single-input-single-output (SISO) system with the transfer function

$$G_p(s) = \frac{2(s+1)}{\left(\frac{1}{2}s+1\right)\left(\frac{1}{4}s+1\right)}$$

where the time constants are in minutes. The system is forced by a unit step input at time  $t = 0$ . The time at which the output response reaches the maximum is \_\_\_\_\_ minutes (rounded off to two decimal places). (GATE CH 2022)

**Solution:**

Parameters	Description	Value
$y(t)$	Output response	
$G_p(s)$	Transfer function	$\frac{2(s+1)}{\left(\frac{1}{2}s+1\right)\left(\frac{1}{4}s+1\right)}$
$x(t)$	Input	$u(t)$
$X(s)$	Laplace transform of x(t)	$\frac{1}{s}$
$y'(t)$	$\frac{dy}{dt}$	

Table 4.13: Parameters

$$Y(s) = G_p(s)X(s) \quad (4.93)$$

$$= \frac{16(s+1)}{s(s+2)(s+4)} \quad (4.94)$$

$$= \frac{2}{s} + \frac{4}{s+2} - \frac{6}{s+4} \quad (4.95)$$

$$u(t) \xleftrightarrow{\mathcal{L}} \frac{1}{s} \quad (4.96)$$

$$e^{-at}u(t) \xleftrightarrow{\mathcal{L}} \frac{1}{s+a} \quad (4.97)$$

From Laplace transforms (4.96) and (4.97), we get

$$y(t) = (2 + 4e^{-2t} - 6e^{-4t}) u(t) \quad (4.98)$$

For maximum value of  $y(t)$ ,

$$y'(t) = 0 \quad (4.99)$$

$$\implies -8e^{-2t} + 24e^{-4t} = 0 \quad (4.100)$$

$$e^{2t} = 3 \quad (4.101)$$

$$\implies t = \frac{\ln 3}{2} \quad (4.102)$$

$$\approx 0.55 \quad (4.103)$$

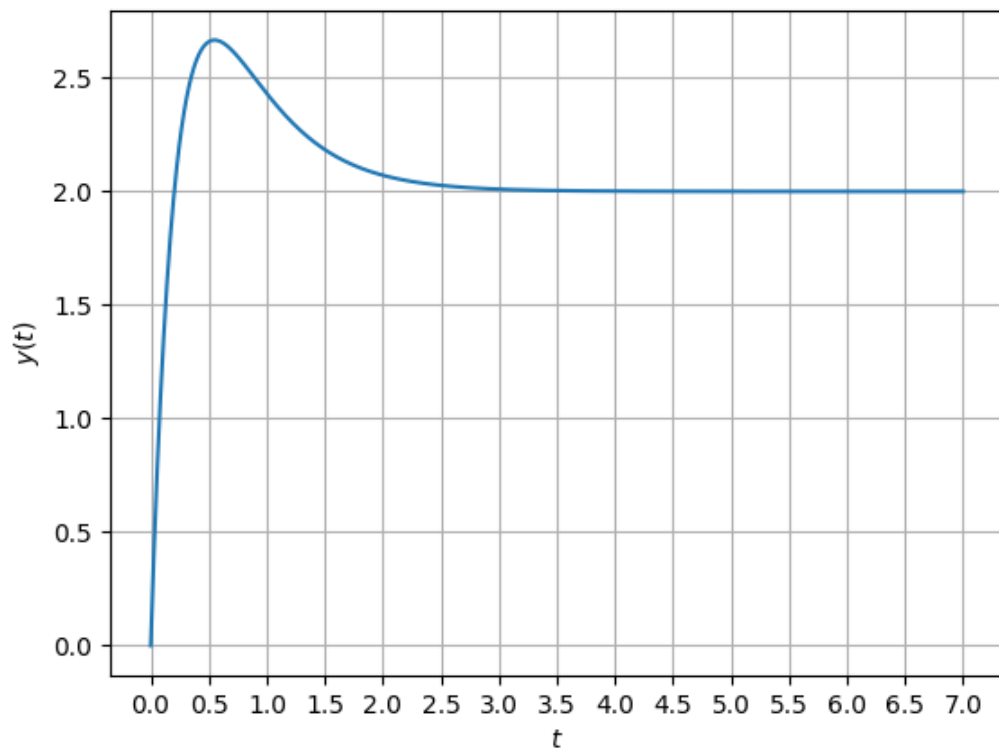


Figure 4.8: Plot of  $y(t)$

4.9 Consider the system as shown below:



The system is described by the equation

$$y(t) = x(e^{-t}).$$

The system is:

- (A) non-linear and causal.
- (B) linear and non-causal.
- (C) non-linear and non-causal.
- (D) linear and causal.

(GATE EE 2022)

**Solution:**

**Homogeneity Test:**

For input  $x_1(e^{-t})$ , the output will be  $y_1(t)$ .

$$y_1(t) = x_1(e^{-t}) \quad (4.104)$$

Multiplying both sides by a scalar quantity 'a'

$$ay_1(t) = ax_1(e^{-t}) \quad (4.105)$$

For input  $x_2(e^{-t})$ , the output will be  $y_2(t)$ .

$$y_2(t) = x_2(e^{-t}) \quad (4.106)$$

Multiplying both sides by a scalar quantity 'b'

$$by_2(t) = bx_2(e^{-t}) \quad (4.107)$$

Adding the above equations we get:

$$ay_1(t) + by_2(t) = ax_1(e^{-t}) + bx_2(e^{-t}) \quad (4.108)$$

Let us assume that, for input  $ax_1(e^{-t}) + bx_2(e^{-t})$ , the output will be  $y_3(t)$ .

$$y_3(t) = ax_1(e^{-t}) + bx_2(e^{-t}) \quad (4.109)$$

But,  $ay_1(t) + by_2(t) = ax_1(e^{-t}) + bx_2(e^{-t})$

Therefore;

$$y_3(t) = ay_1(t) + by_2(t) \quad (4.110)$$

The system satisfies homogeneity, as scaling the input scales the output.

### **Additivity Test:**

From the given system;

$$y(t) = x(e^{-t}) \quad (4.111)$$



$$y(0) = x(e^0) \tag{4.112}$$

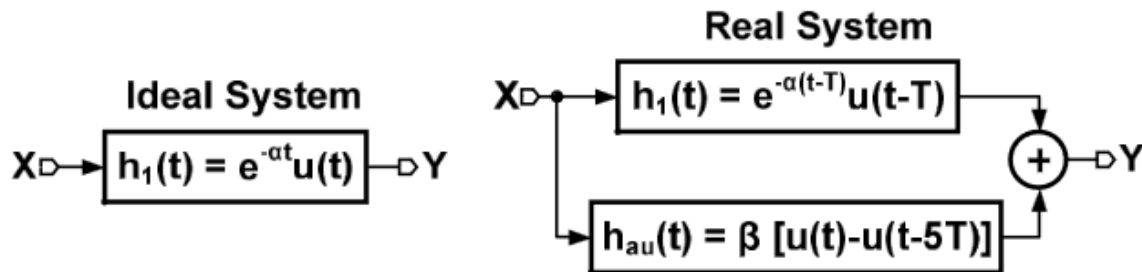
$$y(1) = x(e) = x(2.71) \tag{4.113}$$

So, the present value of output depends on the future value of input, indicating non-causality.

Therefore, the correct answer is: **(B) linear and non-causal**

4.10 The block diagrams of an ideal system and a real system with their impulse responses are shown below. An auxiliary path is added to the delayed impulse response in the real system.

For a unit impulse input  $(x(t)) = \delta(t)$  to both systems, gain  $\beta$  is chosen such that  $y(4T)$  is same for both systems. The value of  $\beta$  is:



(A)  $e^{-3\alpha T} (1 - e^{-2\alpha T})$

(B)  $-e^{-\alpha T} (1 - e^{-3\alpha T})$

(C)  $-e^{-3\alpha T} (1 - e^{-\alpha T})$

(D)  $e^{-2\alpha T} (1 - e^{-2\alpha T})$

(GATE BM 2022)

**Solution:** For both signals to be equal at  $t = 4T$ :

No.	Output	Function
1	$y_I$	$e^{-\alpha t} u(t)$
2	$y_R$	$\beta (u(t) - u(t - 5T)) + e^{-\alpha(t-T)} u(t - T)$

Table 4.14: Values

$$e^{-\alpha 4T} u(4T) = \left[ \beta (u(4T) - u(-T)) + e^{-\alpha(3T)} u(3T) \right] \quad (4.114)$$

$$e^{-\alpha 4T} = \beta + e^{-\alpha 3T} \quad (4.115)$$

$$\implies \beta = -e^{-3\alpha T} (1 - e^{-\alpha T}) \quad (4.116)$$

- 4.11 In a unity-gain feedback control system, the plant  $P(s) = \frac{0.001}{s(2s+1)(0.01s+1)}$  is controlled by a lag compensator  $C(s) = \frac{s+10}{s+0.1}$ . The slope (in dB/decade) of the asymptotic Bode magnitude plot of the loop gain at  $\omega = 3\text{rad/s}$  is \_\_\_\_\_ (in integer) (GATE 2022 IN)

**Solution:**

Parameter	Description	Value
$P(s)$	Plant Transfer Function	$\frac{0.001}{s\left(\frac{s}{0.5}+1\right)\left(\frac{s}{100}+1\right)}$
$C(s)$	Lag Compensator	$\frac{100\left(\frac{s}{10}+1\right)}{\frac{s}{0.1}+1}$
$T(s)$	Loop gain	$P(s)C(s)$
$\omega$	Angular Frequency	3rad/s

Table 4.15: Given Parameters list

$$|T(s)| = \frac{0.1 \left( \frac{s}{10} + 1 \right)}{s \left( \frac{s}{0.5} + 1 \right) \left( \frac{s}{100} + 1 \right) \left( \frac{s}{0.1} + 1 \right)} \quad (4.117)$$

Here, 10, 0.5, 100, 0.1 are corner frequencies of loop gain  $L(s)$

Corner Frequency	Description	Change in slope
10	Zero	$20\text{dB/dec}$
0.1	Pole	$-20\text{dB/dec}$
0.5	Pole	$-20\text{dB/dec}$
100	Pole	$-20\text{dB/dec}$

Table 4.16: Caption

$$\text{Gain}(K) = \lim_{s \rightarrow 0} sT(s) \quad (4.118)$$

$$K = 0.1 \quad (4.119)$$

$$|T(s)| = 20 \log_{10} K \quad (4.120)$$

$$= -20dB \quad (4.121)$$

$$T(\omega) = \begin{cases} -20 \log_{10}(w) & \omega < 0.1 \\ -20.0 (2 \log_{10}(w) - 0.1) & 0.1 \leq \omega < 0.5 \\ -20.0 (3 \log_{10}\omega - 0.1 + \log_{10}0.5) & 0.5 \leq \omega < 10 \\ -20.0 (2 \log_{10}\omega + 0.9 + \log_{10}0.5) & 10 \leq \omega < 100 \\ -20.0 (3 \log_{10}\omega - 1.9 + \log_{10}0.5) & \omega \geq 100 \end{cases}$$

Slope of Bode magnitude plot (at  $\omega = 3$ ) =  $-60$  dB/decade

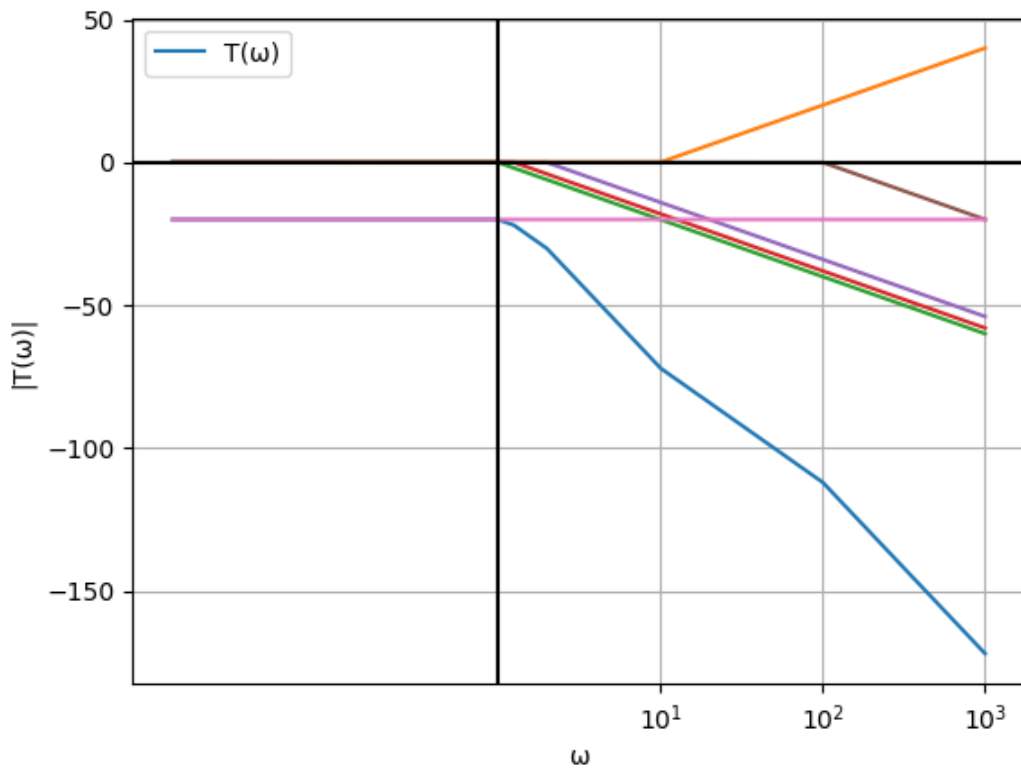


Figure 4.9: Pink Line = Bode magnitude plot of loop gain

4.12 The voltage at the input of an AC-DC rectifier is given by  $v(t) = 230\sqrt{2}\sin\omega t$ , where  $\omega = 2\pi \times 50\text{rad/s}$ . The input current drawn by the rectifier is given by

$$i(t) = 10 \sin\left(\omega t - \frac{\pi}{3}\right) + 4 \sin\left(3\omega t - \frac{\pi}{6}\right) + 3 \sin\left(5\omega t - \frac{\pi}{3}\right)$$

The power input, (rounded off to two decimal places), is\_\_\_\_\_lag. (Gate 2022 EE 33Q)

**Solution:**

$$P_{avg} = \frac{1}{T} \int_0^T V(t) I(t) dt \quad (4.122)$$

For current sources of the form  $I(t) = I_0 + I_1(t) + \dots + I_n(t)$

$$P_{avg} = \frac{1}{T} \sum_1^n \int_0^T V(t) I_n(t) dt \quad (4.123)$$

$$P_{avg} = \frac{1}{T} \sum_1^n \int_0^T V_{pk} \sin(\omega t) I_{pk(n)} \sin(\omega t + \varphi) dt \quad (4.124)$$

$$P_{avg} = \sum_0^n \frac{v_{pk} I_{(n)pk}}{2} \cos \varphi \quad (4.125)$$

For a sine wave signal  $V_{pk} = V_{rms} \sqrt{2}$

$$P_{avg} = \sum_0^n (v_{rms}) (I_{(n)rms}) \cos \varphi \quad (4.126)$$

$$\text{Power Factor} = \frac{\sum_0^n I_{(n)rms} \cos \varphi}{I_{rms}} \quad (4.127)$$

$$I_{rms} = \sqrt{\left(\frac{10}{\sqrt{2}}\right)^2 + \left(\frac{4}{\sqrt{2}}\right)^2 + \left(\frac{3}{\sqrt{2}}\right)^2} = 7.905A \quad (4.128)$$

The rms value of fundamental value of current

$$(I_{(1)rms}) = \sqrt{\left(\frac{10}{\sqrt{2}}\right)^2} \quad (4.129)$$

$$\varphi = 30^\circ \quad (4.130)$$

$$\text{Power Factor} = \frac{\frac{10}{\sqrt{2}} \cos 30}{7.905} \quad (4.131)$$

$$\Rightarrow 0.4473 \quad (4.132)$$



Symbol	Value	Description
$v(t)$	$230\sqrt{2}\sin\omega t$	input voltage
$\omega$	$100\pi rad/s$	Angular velocity
Power Factor	$\frac{P_{avg}}{V_{rms}I_{rms}}$	—
$\cos\varphi$	$\frac{1}{2}$	Fundamental displacement factor
$\varphi$	$\frac{\pi}{3}$	angle between $v(t)$ and $I_n$
$I_n$	$10\sin\left(\omega t - \frac{\pi}{3}\right)$	fundamental component of current

Table 4.17: Variable description



## Chapter 5

# Sequences

5.1 Discrete signals  $x(n)$  and  $y(n)$  are shown below. The cross-correlation  $r_{xy}(0)$  is:

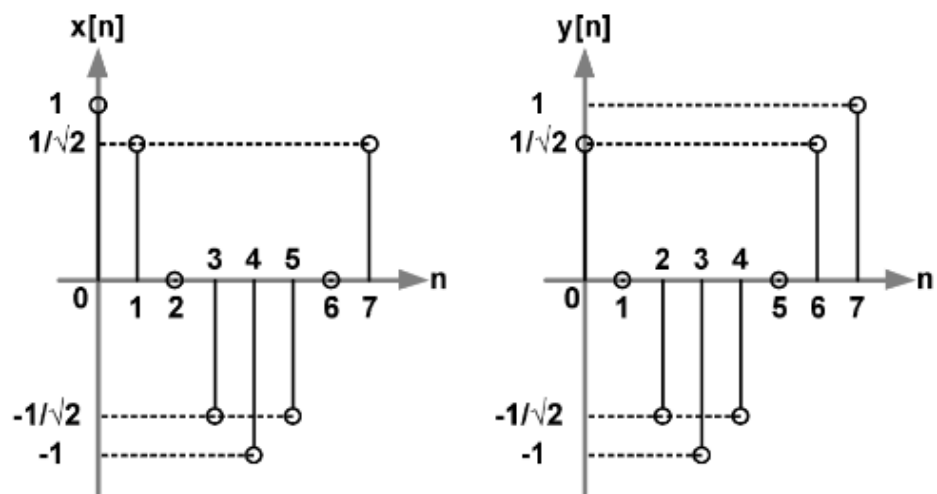


Figure 5.1: Question Figure

(GATE BM 2022)

**Solution:**

Parameter	Description	Value
$x(n)$	First Sequence	$x(n) = \begin{cases} 0 & ; n < 0 \\ \left(1, \frac{1}{\sqrt{2}}, 0, -\frac{1}{\sqrt{2}}, -1, -\frac{1}{\sqrt{2}}, 0, \frac{1}{\sqrt{2}}\right) & ; 0 \leq n \leq 7 \\ 0 & ; n > 7 \end{cases}$
$y(n)$	Second Sequence	$y(n) = \begin{cases} 0 & ; n < 0 \\ \left(\frac{1}{\sqrt{2}}, 0, -\frac{1}{\sqrt{2}}, -1, -\frac{1}{\sqrt{2}}, 0, \frac{1}{\sqrt{2}}, 1\right) & ; 0 \leq n \leq 7 \\ 0 & ; n > 7 \end{cases}$
$r_{xy}(k)$	Cross-correlation	$\sum_{m=-\infty}^{\infty} x(m) y(m-k)$

Table 1: Parameter Table

It can be seen that :

$$y(n) = x(n+1) \quad (5.1)$$

From Table 1 :

$$r_{xy}(k) = \sum_{m=-\infty}^{\infty} x(m) y(m-k) \quad (5.2)$$

$$= x(k) * y(-k) \quad (5.3)$$

From (5.1):

$$r_{xy}(k) = x(k+1) * x(-k) \quad (5.4)$$

$$= \sum_{n=-\infty}^{\infty} x(n+1) x(n+k) \quad (5.5)$$

By definition of  $x(n)$  from Table 1:

$$r_{xy}(k) = \sum_{n=0}^6 x(n+1) x(n+k) \quad (5.6)$$

$$r_{xy}(0) = \sum_{n=0}^6 x(n+1) x(n) \quad (5.7)$$

Using values from Fig. 5.1:

$$r_{xy}(0) = 2\sqrt{2} \quad (5.8)$$

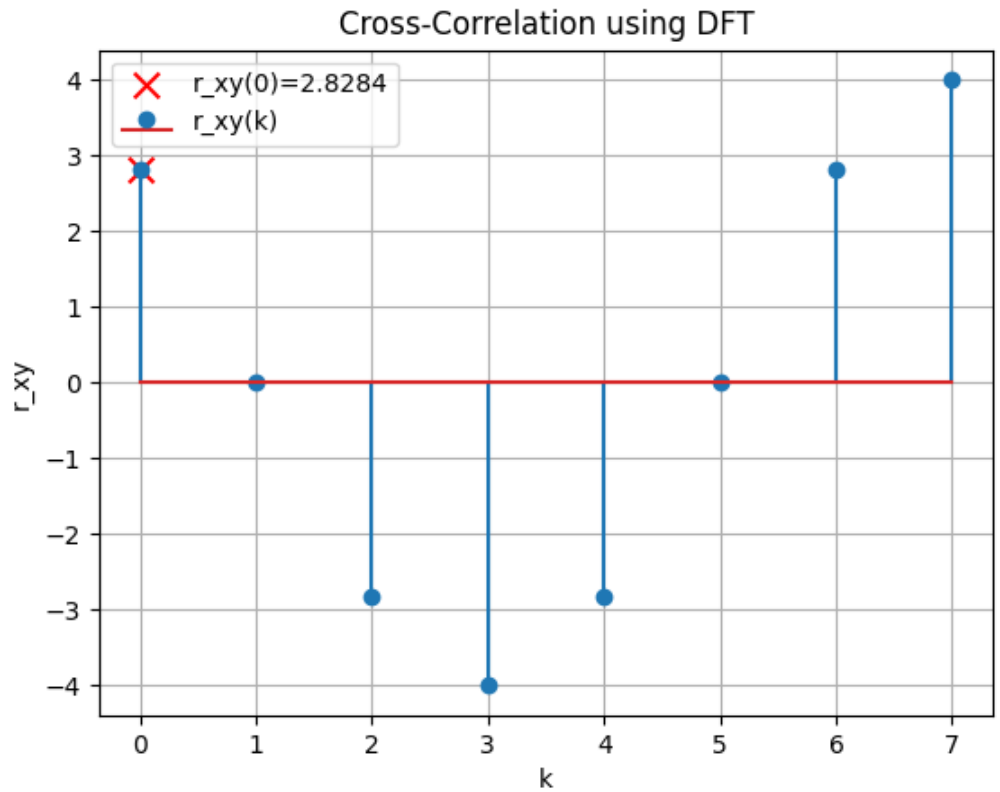


Figure 5.2: Verification of result by DFT

5.2 Which one of the following is the closed form for the generating function of the sequence  $\{a\}_{n \geq 0}$  defined below?

$$a_n = \begin{cases} n+1 & , n \text{ is odd} \\ 1 & \text{otherwise} \end{cases} \quad (5.9)$$

(A)  $\frac{x(1+x)^2}{(1-x^2)^2} + \frac{1}{1-x}$

(B)  $\frac{x(3-x^2)}{(1-x^2)^2} + \frac{1}{1-x}$

(C)  $\frac{2x}{(1-x^2)^2} + \frac{1}{1-x}$

(D)  $\frac{x}{(1-x^2)^2} + \frac{1}{1-x}$

(GATE CS 2022 QUESTION 36)

**Solution:** For the given sequence:

Parameter	Description	Value
$X(z)$	Generating function for a sequence $\{a_n\}$	?
$a_n$	$n^{\text{th}}$ term of the sequence	$(n+1)u(n)$ (when odd)
		$u(n)$ (when even)

Table 5.2: input values

$$X(z) = \sum_{k=-\infty}^{\infty} u(2k) z^{-2k} + \sum_{k=-\infty}^{\infty} ((2k+2)u(2k+1)) z^{-(2k+1)} \quad (5.10)$$

$$\Rightarrow X(z) = (1 + z^{-2} + z^{-4} + \dots) + (2z^{-1} + 4z^{-3} + 6z^{-5} + \dots) \quad (5.11)$$

$$\Rightarrow X(z) = \frac{1}{1-z^{-2}} + (2z^{-1} + 4z^{-3} + 6z^{-5} \dots) \quad |z| > 1 \quad (5.12)$$

$$\Rightarrow X(z) = \frac{1}{1-z^{-2}} + 2z^{-1} \left( \frac{1}{1-z^{-2}} + \frac{z^{-2}}{(1-z^{-2})^2} \right) \quad |z| > 1 \quad (5.13)$$

$$\therefore X(z) = \frac{1}{1-z^{-1}} + \frac{z^{-1}(1+z^{-2})}{(1-z^{-2})^2} \quad |z| > 1 \quad (5.14)$$

(5.14) is the closed form of generating function required in the question.

Hence, option (A) is correct.

$$X(z) = X_1(z) + X_2(z) \quad (5.15)$$

$$X_1(z) = \frac{1}{1-z^{-1}} \quad |z| > 1 \quad (5.16)$$

$$\implies x_1(n) = u(n) \quad (5.17)$$

$$\implies a_n = x_1(n) + x_2(n) \quad (5.18)$$

To find inverse z-transform of  $X_2(z)$  we use contour integration technique:

$$x_2(n) = \frac{1}{2\pi j} \oint_C X_2(z) z^{n-1} dz \quad (5.19)$$

$$= \frac{1}{2\pi j} \oint_C \frac{z^n (z^2 + 1)}{(z^2 - 1)^2} dz \quad (5.20)$$

We can observe that we have two poles at

$z = 1, -1$ . And poles are repeated twice, thus by applying residue theorem two times

for poles 1 and -1:

$$x_2(n) = \frac{1}{(1)!} \lim_{z \rightarrow 1} \frac{d}{dz} \left( (z-1)^2 X_2(z) \right) + \frac{1}{(1)!} \lim_{z \rightarrow -1} \frac{d}{dz} \left( (z+1)^2 X_2(z) \right) \quad (5.21)$$

$$\Rightarrow x_2(n) = \lim_{z \rightarrow 1} \frac{d}{dz} \left( (z-1)^2 \frac{z^n (z^2 + 1)}{(z^2 - 1)^2} \right) + \lim_{z \rightarrow -1} \frac{d}{dz} \left( (z+1)^2 \frac{z^n (z^2 + 1)}{(z^2 - 1)^2} \right) \quad (5.22)$$

$$\begin{aligned} \Rightarrow x_2(n) &= \lim_{z \rightarrow 1} \frac{(z+1)^2 (nz^{n-1} + (n+2)z^{n+1}) - 2z^n (1+z^2)(z+1)}{(z+1)^4} \\ &\quad + \lim_{z \rightarrow -1} \frac{(z-1)^2 (nz^{n-1} + (n+2)z^{n+1}) - 2z^n (1+z^2)(z-1)}{(z-1)^4} \end{aligned} \quad (5.23)$$

on simplification, we get

$$x_2(n) = \frac{n + n(-1)^{n-1}}{2} \quad (5.24)$$

$$\therefore a_n = u(n) + \frac{n + n(-1)^{n-1}}{2} u(n) \quad (5.25)$$

Which is the sequence given in the Question.



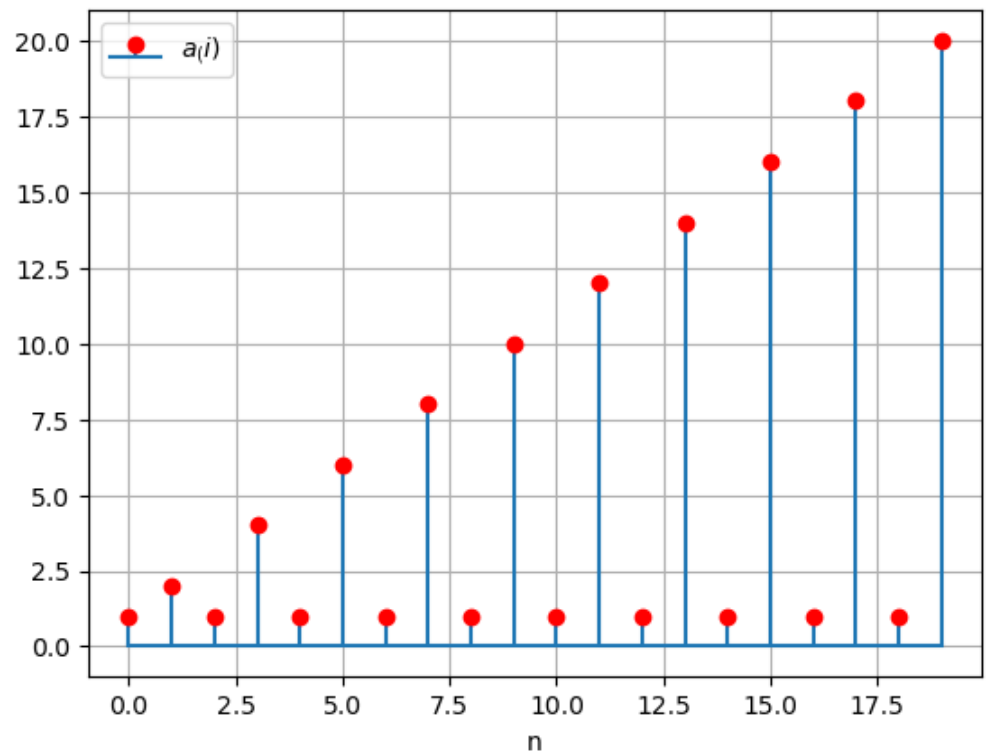


Figure 5.3: Terms of the sequence given

5.3 **Question:** Consider the following recurrence:

$$f(1) = 1; \quad (5.26)$$

$$f(2n) = 2f(n) - 1, \text{ for } n \geq 1; \quad (5.27)$$

$$f(2n+1) = 2f(n) + 1, \text{ for } n \geq 1. \quad (5.28)$$

Then, which of the following is/are **TRUE**?

(A)  $f(2^n - 1) = 2^n - 1$

(B)  $f(2^n) = 1$

$$(C) f(5 \cdot 2^n) = 2^{n+1} + 1$$

$$(D) f(2^n + 1) = 2^n + 1$$

[GATE-CS.51 2022]

**Solution:** (A)

let  $x(2^k - 1) = 2^k - 1$  for any  $k \geq 1$ ,

$$x(2^{k+1} - 1) = x(2(2^k - 1) + 1) \quad (5.29)$$

From (5.28),

$$= 2x(2^k - 1) + 1 \quad (5.30)$$

$$= 2(2^k - 1) + 1 \quad (5.31)$$

$$= 2^{k+1} - 1 \quad (5.32)$$

From (5.26),(5.32)

$$x(2 - 1) = 2 - 1 \quad (k = 0) \quad (5.33)$$

$$= 1 \quad (5.34)$$

Hence  $x(2^n - 1) = 2^n - 1$  for  $n \geq 1$

So statement A is TRUE

(B)

Let  $x(2^k) = 1$  for any  $k \geq 0$

$$x(2^{k+1}) = x(2 \cdot 2^k) \quad (5.35)$$

From (5.27)

$$= 2x(2^k) - 1 \quad (5.36)$$

$$= 1 \quad (5.37)$$

From (5.36),(5.27)

$$x(2) = 2x(1) - 1 \quad (5.38)$$

$$= 1 \quad (5.39)$$

Hence  $x(2^n) = 1$  for every  $n \geq 0$  value.

So statement B is TRUE.

(C)

Let,  $x(5 \cdot 2^k) = 2^{k+1} + 1$  be true for any  $k \geq 0$ ,

$$x(5 \cdot 2^{k+1}) = x(2(5 \cdot 2^k)) \quad (5.40)$$

From (5.27)

$$= 2x(5 \cdot 2^k) - 1 \quad (5.41)$$

$$= 2^{k+2} + 1 \quad (5.42)$$

$k = -1$  ,From (5.42)

$$x(5) = 2^1 + 1 \quad (5.43)$$

$$= 3 \quad (5.44)$$

Proof:-

$$x(5) = x(2 \cdot 2 + 1) \quad (5.45)$$

From (5.28),(5.39)

$$= 2x(2) + 1 \quad (5.46)$$

$$= 3 \quad (5.47)$$

Hence  $x(5 \cdot 2^n) = 2^{n+1} + 1$  for  $n \geq 1$

So statement C is TRUE.

(D)

$$x(2^n + 1) = x(2 \cdot 2^{n-1} + 1) \quad (5.48)$$

From (5.28),(5.39)

$$= 2x(2^{n-1}) + 1 \quad (5.49)$$

$$= 3 \quad (5.50)$$

Hence statement D is FALSE.

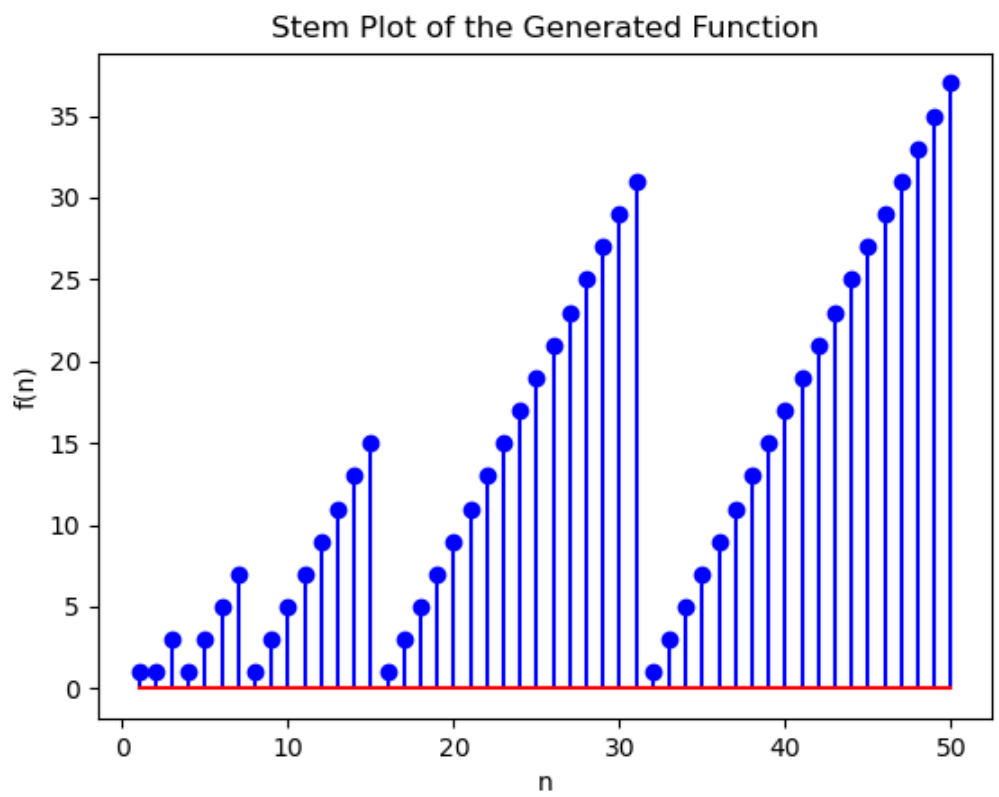


Figure 5.4: plot of  $x(n)$



## Chapter 6

# Sampling

6.1 Consider the transfer function

$$H_c(s) = \frac{1}{(s+1)(s+3)}$$

Bilinear transformation with a sampling period of  $0.1s$  is employed to obtain the discrete-time transfer function  $H_d(z)$ . Then  $H_d(z)$  is

(A)  $\frac{(1+z^{-1})^2}{(19-21z^{-1})(23-17z^{-1})}$

(B)  $\frac{(1-z^{-1})^2}{(21-19z^{-1})(17-23z^{-1})}$

(C)  $\frac{(1+z^{-1})^2}{(21-19z^{-1})(23-17z^{-1})}$

(D)  $\frac{(1+z^{-1})^2}{(21-19z^{-1})(17-23z^{-1})}$

(GATE IN 2022)

**Solution:**

$$H_c(s) \xleftrightarrow{\text{Bilinear Transform}} H_d(z) \quad (6.1)$$

Parameters	Value	Description
$H_c(s)$	$\frac{1}{(s+1)(s+3)}$	Transfer function in $s$ domain
$T_s$	$0.1s$	Sampling period
$H_d(z)$		Transfer function of sampled signal

Table 6.1: Input Parameters

To get  $H_d(z)$ , substitute  $s$  with

$$s = \frac{2}{T_s} \left( \frac{1 - z^{-1}}{1 + z^{-1}} \right) \quad (6.2)$$

where  $T_s$  is the sampling period. Then,

$$H_d(z) = \frac{1}{\left( \frac{2}{0.1} \left( \frac{1 - z^{-1}}{1 + z^{-1}} \right) + 1 \right) \left( \frac{2}{0.1} \left( \frac{1 - z^{-1}}{1 + z^{-1}} \right) + 3 \right)} \quad (6.3)$$

$$= \frac{(1 + z^{-1})^2}{(21 - 19z^{-1})(23 - 17z^{-1})} \quad (6.4)$$

ROC :  $|z| > \frac{19}{21}$

Using partial fractions,

$$H_d(z) = \frac{1}{323} + \frac{340}{323} \left( \frac{1}{21 - 19z^{-1}} \right) - \frac{380}{323} \left( \frac{1}{23 - 17z^{-1}} \right) \quad (6.5)$$

By applying inverse z-transform,

$$\delta(n) \xleftrightarrow{\mathcal{Z}} 1 \quad (6.6)$$

$$x(0)r^n u(n) \xleftrightarrow{\mathcal{Z}} \frac{x(0)}{1 - rz^{-1}} \quad (6.7)$$

$$H_d(n) = \frac{1}{323} \delta(n) + \frac{340}{6783} \left( \frac{19}{21} \right)^n - \frac{380}{7429} \left( \frac{17}{23} \right)^n \quad (6.8)$$



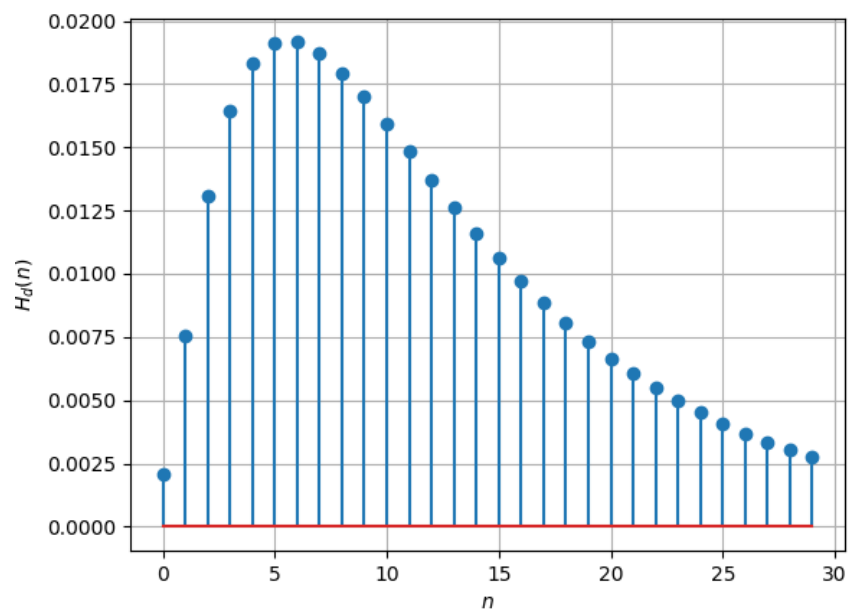


Figure 6.1: stem plot of  $H_d(n)$



## Chapter 7

# Contour Integration

7.1 In the complex  $z$ -domain, the value of integral  $\oint_C \frac{z^3-9}{3z-i} dz$  is

- (a)  $\frac{2\pi}{81} - 6i\pi$
- (b)  $\frac{2\pi}{81} + 6i\pi$
- (c)  $-\frac{2\pi}{81} + 6i\pi$
- (d)  $-\frac{2\pi}{81} - 6i\pi$

(GATE 2022 BM)

**Solution:**

Simplyfying the Contour Integral to the standard form we get,

$$\oint_C \frac{z^3-9}{3z-i} dz = \frac{1}{3} \oint_C \frac{z^3-9}{z-\frac{i}{3}} dz \quad (7.1)$$

From Cauchy's residue theorem,

$$\oint_C f(z) dz = 2\pi i \sum R_j \quad (7.2)$$

We can observe a non-repeated pole at  $z = \frac{i}{3}$  and thus  $a = \frac{i}{3}$ ,

$$R = \lim_{z \rightarrow a} (z - a) f(z) \quad (7.3)$$

$$\Rightarrow R = \frac{1}{3} \lim_{z \rightarrow \frac{i}{3}} \left( z - \frac{i}{3} \right) \frac{z^3 - 9}{z - \frac{i}{3}} \quad (7.4)$$

$$= \frac{-i}{81} - 3 \quad (7.5)$$

Therefore, from (7.2) and (7.5)

$$\oint_C \frac{z^3 - 9}{3z - i} dz = \frac{2\pi}{81} - 6i\pi \quad (7.6)$$

7.2 Consider the function

$$f(z) = \frac{1}{(z+1)(z+2)(z+3)}$$

The residue of  $f(z)$  at  $z = -1$ , is \_\_\_\_\_ (GATE 2022 IN)

**Solution:** Residue of a function  $f(z)$  at a simple pole  $c$  is

$$\text{Res}(f, c) = \lim_{z \rightarrow c} (z - c) f(z) \quad (7.7)$$

$$\Rightarrow \text{Res}(f, -1) = \lim_{z \rightarrow -1} \frac{z+1}{(z+1)(z+2)(z+3)} \quad (7.8)$$

$$= \frac{1}{2} \quad (7.9)$$

$\therefore$  residue of  $f(z)$  at  $z = -1$  is  $\frac{1}{2}$ .

7.3 The value of the integral

$$\int_C \frac{z^{100}}{z^{101} + 1} dz$$

where  $C$  is the circle of radius 2 centred at the origin taken in the anti-clockwise direction is

(A)  $-2\pi i$

(B)  $2\pi$

(C)  $0$

(D)  $2\pi i$

(GATE 2022 MA)

**Solution:**

Using B.1 and B.2 Solving the integral,

$$f(z) = \int_C \frac{z^{100}}{z^{101} + 1} dz \quad (7.10)$$

Since the pole  $z = -1$  is inside the circle, Using eq (B.2.2)

$$\text{Res } f(-1) = \lim_{z \rightarrow -1} \left( \frac{z^{100}}{z^{101} + 1} \right) (z^{101} + 1) \quad (7.11)$$

$$= 1 \quad (7.12)$$

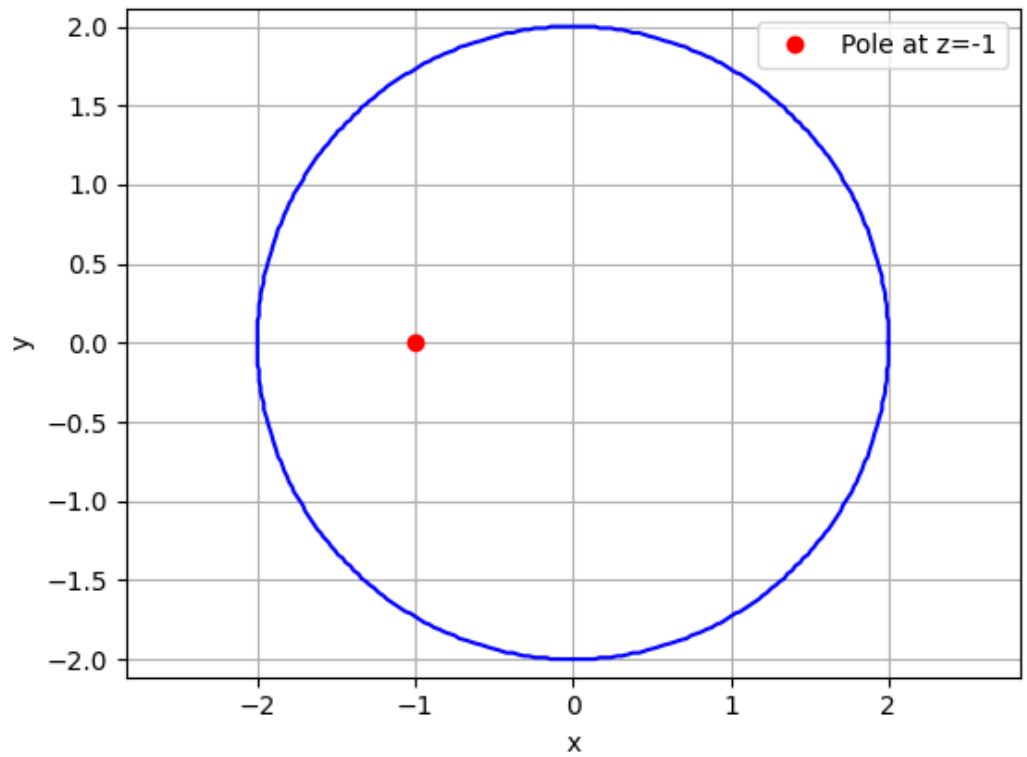


Figure 7.1: plot of  $C$  with it's pole

From eq (B.2.1), and eq (7.12)

$$\int f(z) dz = 2\pi i (1) \quad (7.13)$$

$$\Rightarrow \int_C \frac{z^{100}}{z^{101} + 1} dz = 2\pi i \quad (7.14)$$

$\therefore$  option (D) is correct.





## Chapter 8

# Laplace Transform

- 8.1 Consider the differential equation  $\frac{d^2y}{dx^2} - 2\frac{dy}{dx} + y = 0$ . The boundary conditions are  $y = 0$  and  $\frac{dy}{dx} = 1$  at  $x = 0$ . Then the value of  $y$  at  $x = \frac{1}{2}$  (GATE AE 2022)

**Solution:**

Parameters	Values	Description
$y(0)$	0	$y$ at $x = 0$
$y'(0)$	1	$\frac{dy}{dx}$ at $x = 0$

Table 8.1: Parameters

$$\frac{d^2y}{dx^2} \xleftrightarrow{\mathcal{L}} s^2Y(s) - sy(0) - y'(0) \quad (8.1)$$

$$\frac{dy}{dx} \xleftrightarrow{\mathcal{L}} sY(s) - y(0) \quad (8.2)$$

Applying Laplace Transform, using (8.1) and (8.2),

$$s^2Y(s) - sy(0) - y'(0) - 2(sY(s) - y(0)) + Y(s) = 0 \quad (8.3)$$

From Table 8.1,

$$(s^2 - 2s + 1)Y(s) - 1 = 0 \quad (8.4)$$

$$Y(s) = \frac{1}{(s-1)^2} \quad (8.5)$$

$$t^n \xleftrightarrow{\mathcal{L}} \frac{n!}{s^{n+1}} \quad (8.6)$$

$$e^{at}x(t) \xleftrightarrow{\mathcal{L}} X(s-a) \quad (8.7)$$

Taking Inverse Laplace Transform for  $Y(s)$ , using (8.6) and (8.7),

$$y(x) = xe^x \quad (8.8)$$

$$\Rightarrow y\left(\frac{1}{2}\right) = \frac{\sqrt{e}}{2} \quad (8.9)$$

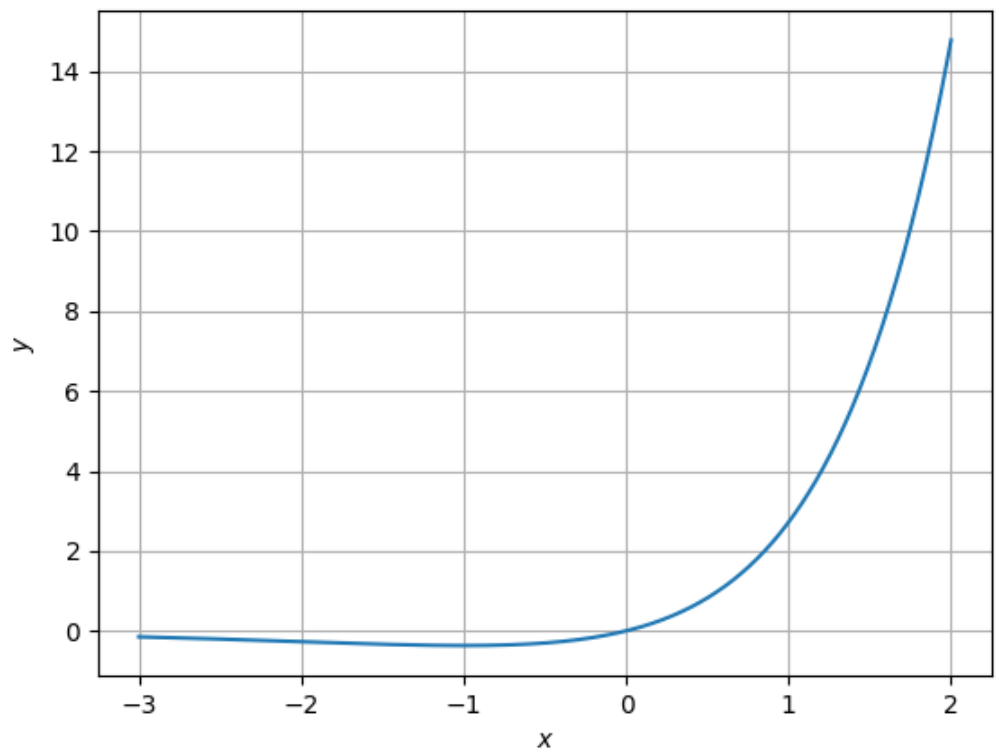


Figure 8.1: Plot of  $y(x)$

## 8.2 A process described by the transfer function

$$G_p(s) = \frac{(10s + 1)}{(5s + 1)}$$

is forced by a unit step input at time  $t = 0$ . The output value immediately after the unit step input (at  $t = 0^+$ ) is ? (Gate 2022 CH 34)

**Solution:**

Parameters	Description
$X(s)$	Laplace transform of $x(t)$
$Y(s)$	Laplace transform of $y(t)$
$G_p(s) = \frac{Y(s)}{X(s)}$	Transfer function
$x(t) = u(t)$	unit step function

Table 8.2: Given parameters

$$G_p(s) = \frac{Y(s)}{X(s)} = \frac{(10s + 1)}{(5s + 1)} \quad (8.10)$$

$$u(t) \xleftrightarrow{\mathcal{L}} \frac{1}{s} \quad (8.11)$$

From equation (8.11):

$$Y(s) = \frac{(10s + 1)}{s(5s + 1)} \quad (8.12)$$

$$= \frac{1}{s} + \frac{5}{5s + 1} \quad (8.13)$$

Taking inverse laplace transformation,

$$\frac{1}{s} \xleftrightarrow{\mathcal{L}^{-1}} u(t) \quad (8.14)$$

$$\frac{1}{s - c} \xleftrightarrow{\mathcal{L}^{-1}} e^{ct} u(t) \quad (8.15)$$

$$y(t) = \left(1 + e^{-\frac{t}{5}}\right) u(t) \quad (8.16)$$

$$y(0^+) = 2 \quad (8.17)$$

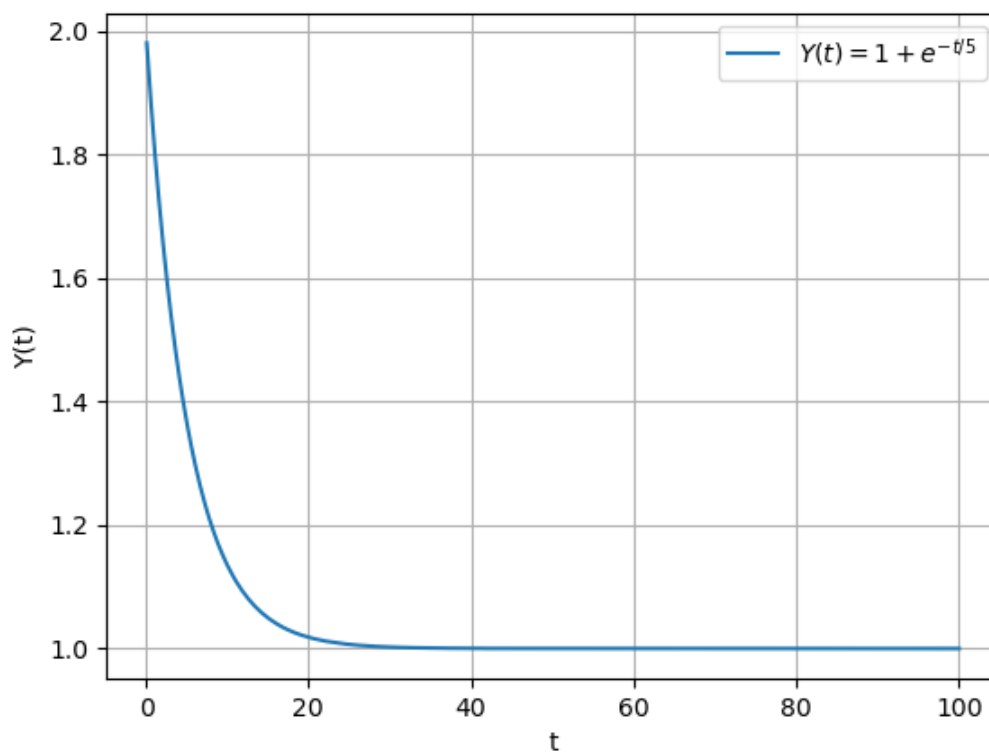


Figure 8.2: Graph of  $y(t)$

8.3 The transfer function of a real system  $H(S)$  is given as:

$$H(s) = \frac{As + B}{s^2 + Cs + D}$$

where  $A, B, C$  and  $D$  are positive constants. This system cannot operate as

- (A) Low pass filter
- (B) High pass filter
- (C) Band pass filter
- (D) An Integrator

(GATE EE 11 2022)

**Solution:** The transfer function  $H(s)$  is given by:

$$H(s) = \frac{As + B}{s^2 + Cs + D} \quad (8.18)$$

Put  $s = j\omega$  in (8.18):

$$H(j\omega) = \frac{A(j\omega) + B}{(j\omega)^2 + C(j\omega) + D} \quad (8.19)$$

$$|H(j\omega)| = \frac{\sqrt{(A\omega)^2 + B^2}}{\sqrt{(D - \omega^2)^2 + (\omega C)^2}} \quad (8.20)$$

a) Low Pass Filter:

At low frequency ( $\omega = 0$ ):

$$|H(\omega = 0)| = \frac{B}{D} \quad (8.21)$$

$\therefore H(s)$  can operate as Low pass filter.

Parameter	Description
Low Pass Filter	The gain should be finite at low frequency
High Pass Filter	The gain should be finite at high frequency
Band Pass Filter	Finite gain over frequency band
Integrator	Transfer function should have at least one pole at origin

Table 8.3: Conditions

b) High Pass Filter:

At high frequency ( $\omega = \infty$ ):

$$|H(\omega = \infty)| = 0 \quad (8.22)$$

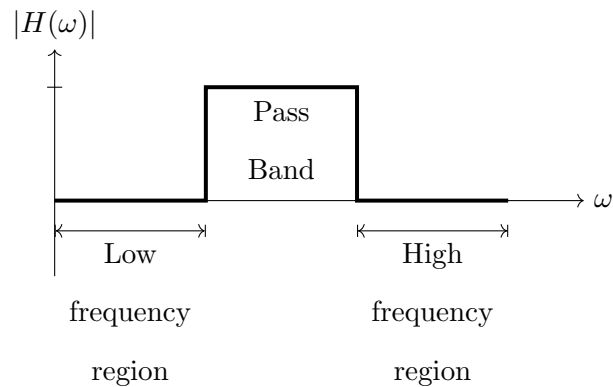
$\therefore H(s)$  cannot operate as High pass filter.

c) Band Pass Filter:

Assuming B is a very less positive valued constant as compared to others:

$$|H(j\omega)| = \frac{(A\omega)}{\sqrt{(D - \omega^2)^2 + (\omega C)^2}} \quad (8.23)$$

$$\implies |H(\omega = 0)| = 0 \text{ and } |H(\omega = \infty)| = 0 \quad (8.24)$$



$\therefore H(s)$  passes frequency be-

tween low and high frequencies.

$\therefore H(s)$  can operate as a band pass filter.

d) Integrator:

At very high value of frequency( $\omega \rightarrow \infty$ ):

$$H(s) \approx \frac{As}{s^2} \approx \frac{A}{s} \quad (8.25)$$

From Table 8.3:

$\therefore H(s)$  can operate as an Integrator.



8.4 In a circuit, there is a series connection of an ideal resistor and an ideal capacitor. The conduction current (in Amperes) through the resistor is  $2 \sin(t + \frac{\pi}{2})$ . The displacement current (in Amperes) through the capacitor is \_\_\_\_\_.

- (A)  $2 \sin(t)$
- (B)  $2 \sin(t + \pi)$
- (C)  $2 \sin(t + \frac{\pi}{2})$
- (D) 0

(GATE 2022 EC 24)

**Solution:**

Parameter	Description	Value
$I_c$	Conduction Current	$2 \sin(t + \frac{\pi}{2})$
$A$	Cross-sectional area	

Table 8.4: Parameters

Parameter	Description	Formula
$Q$	Charge	$\int I_c dt$
$D$	Electric Displacement	$\frac{Q}{A}$
$J_D$	Displacement current density	$\frac{\partial D}{\partial t}$
$I_D$	Displacement current	$J_D \times A$

Table 8.5: Formulae

S Domain	Time Domain
$\frac{1}{s}$	$u(t)$
$\frac{-s}{a^2+s^2}$	$-\cos(at)$
$\frac{a}{a^2+s^2}$	$\sin(at)$
$\frac{1}{s+a}$	$e^{-at}$

Table 8.6: Laplace transforms

$$\mathcal{L} \left[ \int f(t) dt \right] = \int_0^\infty \left[ \int f(t) dt \right] e^{-st} dt \quad (8.26)$$

$$= \int_0^\infty u dv \quad \text{where} \begin{cases} u = \int f(t) dt \\ dv = e^{-st} dt \end{cases} \quad (8.27)$$

$$= uv - v \int du \quad (8.28)$$

$$= \frac{1}{s} \int f(t) dt|_0 + \frac{1}{s} \int_0^\infty f(t) e^{-st} dt \quad (8.29)$$

$$\Rightarrow \frac{1}{s} \int f(t) dt|_0 + \frac{1}{s} F(s) \quad (8.30)$$

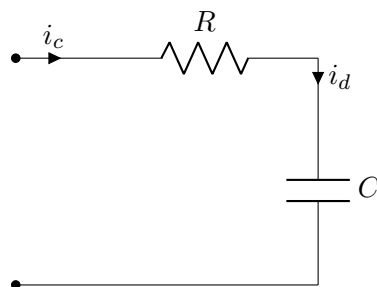


Figure 8.3: Circuit 1

From Table 8.5, Table 8.6 and eq (8.30)

$$I_c(s) = \frac{2s}{s^2 + 1} \quad (8.31)$$

$$Q_c(s) = \frac{2}{s(s^2 + 1)} \quad (8.32)$$

$$D(s) = \frac{1}{A} \left( \frac{2}{s(s^2 + 1)} \right) \quad (8.33)$$

$$J_D(s) = \frac{2}{A} \left( \frac{1}{s^2 + 1} \right) \quad (8.34)$$

$$I_D(s) = \frac{2}{s^2 + 1} \quad (8.35)$$

$$\Rightarrow I_D = 2 \sin t \quad (8.36)$$

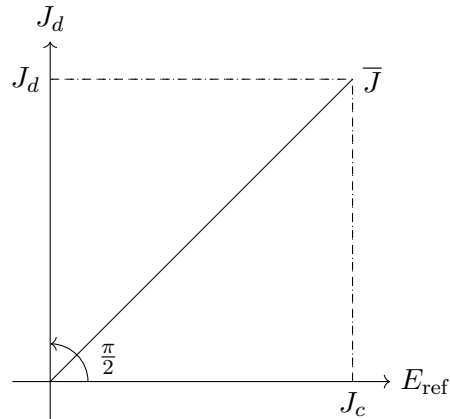


Figure 8.4: Phasor plot

From figure 8.4, phase of  $I_d$  is  $\frac{\pi}{2}$

$$\therefore I_d = 2 \sin \left( t + \frac{\pi}{2} \right) \quad (8.37)$$

$\therefore$  (C) is correct.

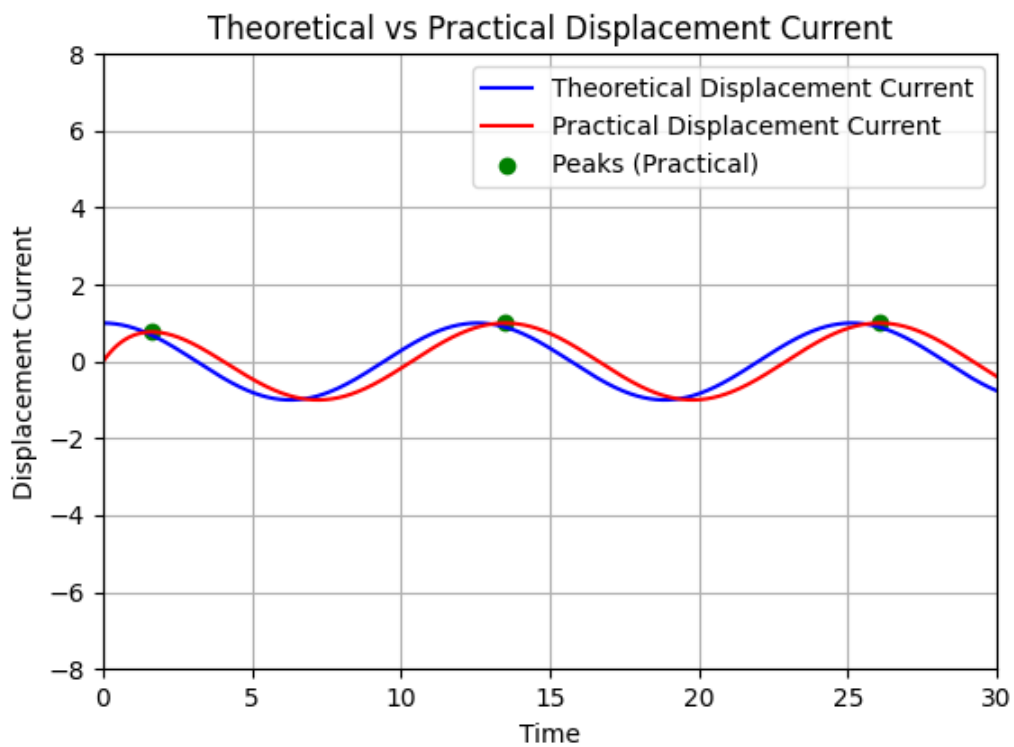


Figure 8.5: Thoritical vs Practical simulation

8.5 Given,  $y = f(x)$ ;  $\frac{d^2y}{dx^2} + 4y = 0$ ;  $y(0) = 0$ ;  $\frac{dy}{dx}(0) = 1$ . The problem is a/an

- (a) initial value problem having solution  $y = x$
- (b) boundary value problem having solution  $y = x$
- (c) initial value problem having solution  $y = \frac{1}{2} \sin 2x$
- (d) boundary value problem having solution  $y = \frac{1}{2} \sin 2x$

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**Solution:**

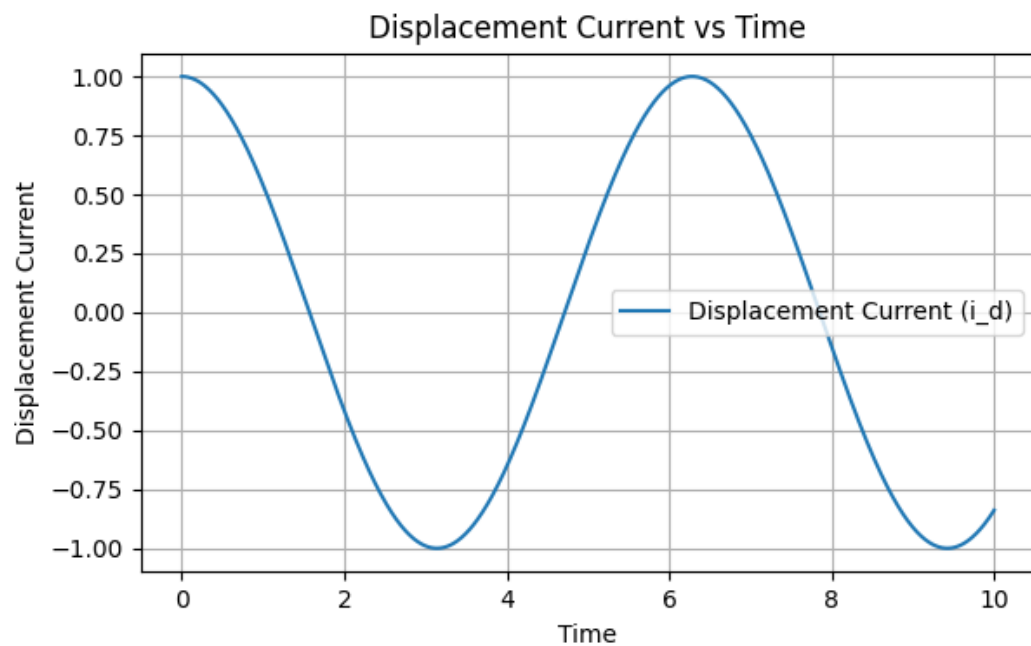


Figure 8.6: Displacement current

The above equation can be written as,

$$y''(t) + 4y(t) = 0 \quad (8.38)$$

Using the Laplace transformation pairs,

$$y''(t) \xleftrightarrow{\mathcal{L}} s^2 Y(s) - sy(0) - y'(0) \quad (8.39)$$

$$y(t) \xleftrightarrow{\mathcal{L}} Y(s) \quad (8.40)$$

$$\sin at \xleftrightarrow{\mathcal{L}} \frac{a}{a^2 + s^2} \quad (8.41)$$

Applying Laplace transform for the equation we get,

$$s^2 Y(s) - 1 + 4Y(s) = 0 \quad (8.42)$$

$$\implies Y(s) = \frac{1}{4 + s^2} \quad (8.43)$$

Now, applying inverse laplace transform we get,

$$y(t) = \frac{1}{2} \sin 2t \quad (\text{from (8.41)}) \quad (8.44)$$

Since, the conditions at the same point(0) are mentioned, it is an initial valued problem having solution  $y = \frac{1}{2} \sin 2x$ .

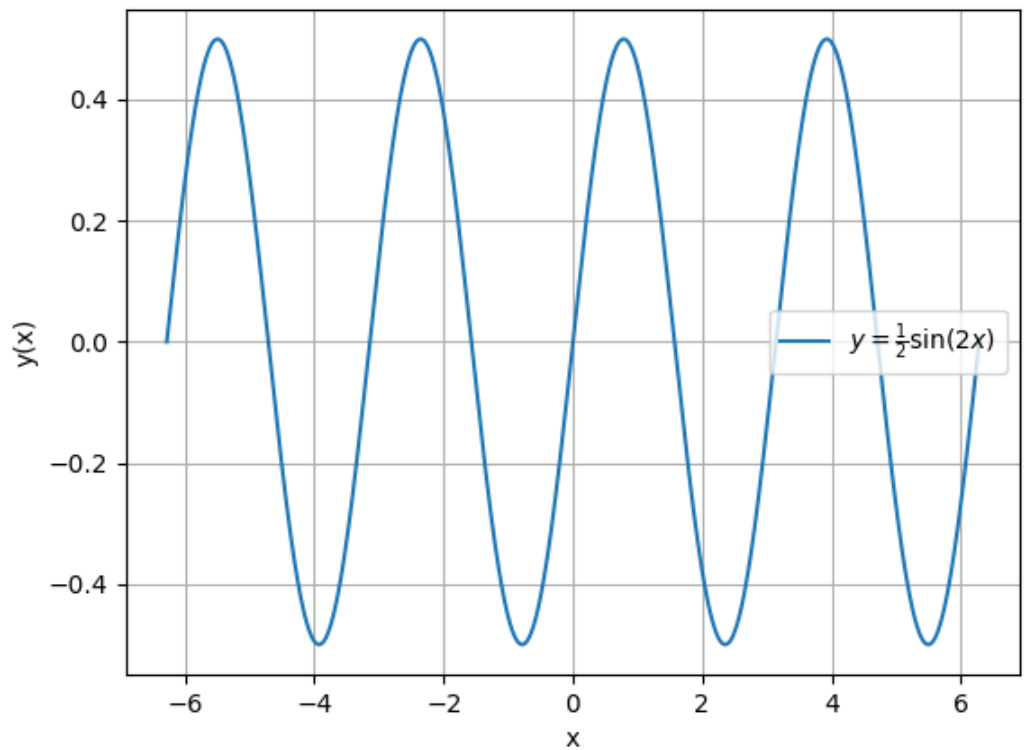


Figure 8.7:  $y(x)$  vs  $x$  graph

8.6 Let a causal LTI system be governed by the following differential equation,

$$y(t) + \frac{1}{4} \frac{dy}{dt} = 2x(t) \quad (8.45)$$

where  $x(t)$  and  $y(t)$  are the input and output respectively. It's impulse response is  
(GATE EE-2022)

**Solution: Solution:**

From (8.45), corresponding Laplace transform,

$$Y(s) + \frac{1}{4}(sY(s) - y(0)) = 2X(s) \quad (8.46)$$

Since it is causal LTI system,

$$y(0) = 0 \quad (8.47)$$

$$\Rightarrow Y(s) + \frac{1}{4}sY(s) = 2X(s) \quad (8.48)$$

$$\Rightarrow Y(s) = X(s) \frac{8}{4+s} \quad (8.49)$$

$$\Rightarrow H(s) = \frac{8}{4+s} \quad ROC : Re(s) > -4 \quad (8.50)$$

Taking inverse laplace transform and applying causality conditions

$$h(t) = 8e^{-4t}u(t) \quad (8.51)$$

8.7 Assuming  $s > 0$ ; Laplace transform for  $f(x) = \sin(ax)$  is

(A)  $\frac{a}{s^2+a^2}$

(B)  $\frac{s}{s^2+a^2}$

(C)  $\frac{a}{s^2-a^2}$

(D)  $\frac{s}{s^2-a^2}$

(GATE 2022 ES)



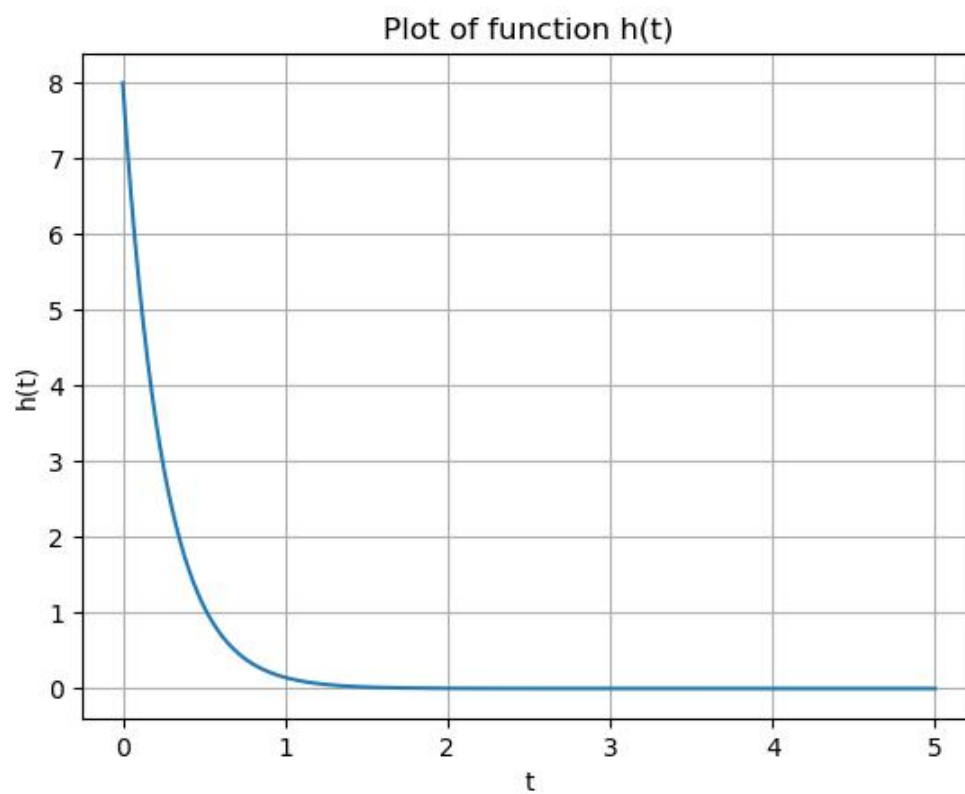


Figure 1: Plot of  $h(n)$ , taken from python3

**Solution:**

$$\mathcal{L}(f(x)) = \int_{-\infty}^{\infty} e^{-sx} f(x) dx \quad (8.52)$$

$$\text{We can write } \sin(ax) = \frac{e^{ax} - e^{-ax}}{2i} \quad (8.53)$$

From (8.53)

$$\mathcal{L}(\sin(ax)) = \int_0^\infty e^{-sx} \left( \frac{e^{iax} - e^{-iax}}{2i} \right) dx \quad (8.54)$$

$$= \frac{1}{2i} \int_0^\infty e^{-x(s-ia)} - e^{-x(s+ia)} dx \quad (8.55)$$

$$= \frac{1}{2i} \left( \frac{e^{-x(s-ia)}}{-(s-ia)} + \frac{e^{-x(s+ia)}}{-(s+ia)} \right) \Big|_0^\infty \quad (8.56)$$

$$= \frac{1}{2i} \left( \frac{1}{s-ia} - \frac{1}{s+ia} \right) \quad (8.57)$$

$$= \frac{a}{s^2 + a^2} \quad (8.58)$$

So, option (A) is correct.

8.8 The input  $x(t)$  to a system is related to its output  $y(t)$  as

$$\frac{dy(t)}{dt} + y(t) = 3x(t-3)u(t-3)$$

Here  $u(t)$  represents a unit-step function.

The transfer function of this system is

(A)  $\frac{e^{-3s}}{s+3}$

(B)  $\frac{3e^{-3s}}{s+1}$

(C)  $\frac{3e^{-(s/3)}}{s+1}$

(D)  $\frac{e^{-(s/3)}}{s+3}$

(GATE IN 2022)

**Solution:**

$$\frac{dy(t)}{dt} + y(t) = 3x(t-3)u(t-3) \quad (8.59)$$

By applying Laplace Transform on both sides

$$x(t) \xleftrightarrow{\mathcal{L}} X(s) \quad (8.60)$$

$$x(t - t_o) \xleftrightarrow{\mathcal{L}} X(s)e^{-st_o} \quad (8.61)$$

$$sY(s) + Y(s) = 3X(s)e^{-3s} \quad (8.62)$$

$$Y(s)(s + 1) = 3X(s)e^{-3s} \quad (8.63)$$

$$H(s) = \frac{Y(s)}{X(s)} = \frac{3e^{-3s}}{s + 1} \quad (Re(s) > 0) \quad (8.64)$$

$$H(j\omega) = \frac{3e^{-3j\omega}}{1 + j\omega} \quad (8.65)$$

$$= \frac{3(\cos 3\omega - j\sin 3\omega)}{1 + j\omega} \quad (8.66)$$

$$|H(j\omega)| = \frac{3}{\sqrt{1 + \omega^2}} \quad (8.67)$$

$$phase = \tan^{-1} \left( \frac{\omega \cos(3\omega) + \sin(3\omega)}{\omega \sin(3\omega) - \cos(3\omega)} \right) \quad (8.68)$$

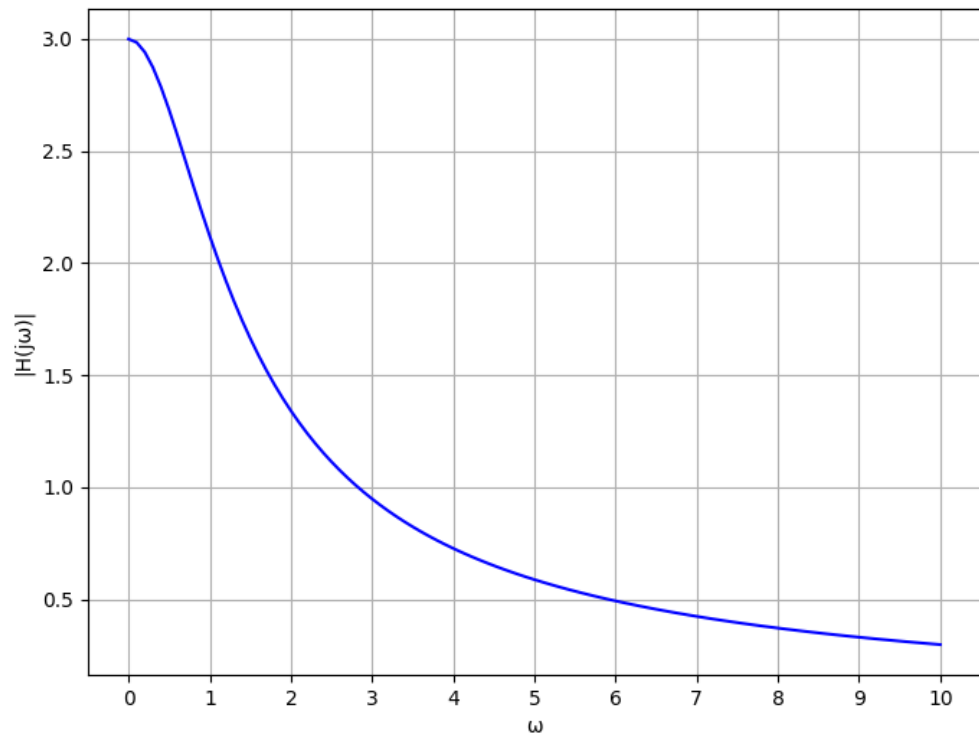


Figure 8.9: Plot for magnitude of transfer function

8.9 Let  $x_1(t) = e^{-t}u(t)$  and  $x_2(t) = u(t) - u(t - 2)$ , where  $u(\cdot)$  denotes the unit step function. If  $y(t)$  denotes the convolution of  $x_1(t)$  and  $x_2(t)$ , then  $\lim_{t \rightarrow \infty} y(t) = \underline{\hspace{2cm}}$ .  
 (Rounded off to one decimal place)  
 (GATE EC 2022 )

**Solution:**

$$y(t) = x_1(t) * x_2(t) \quad (8.69)$$

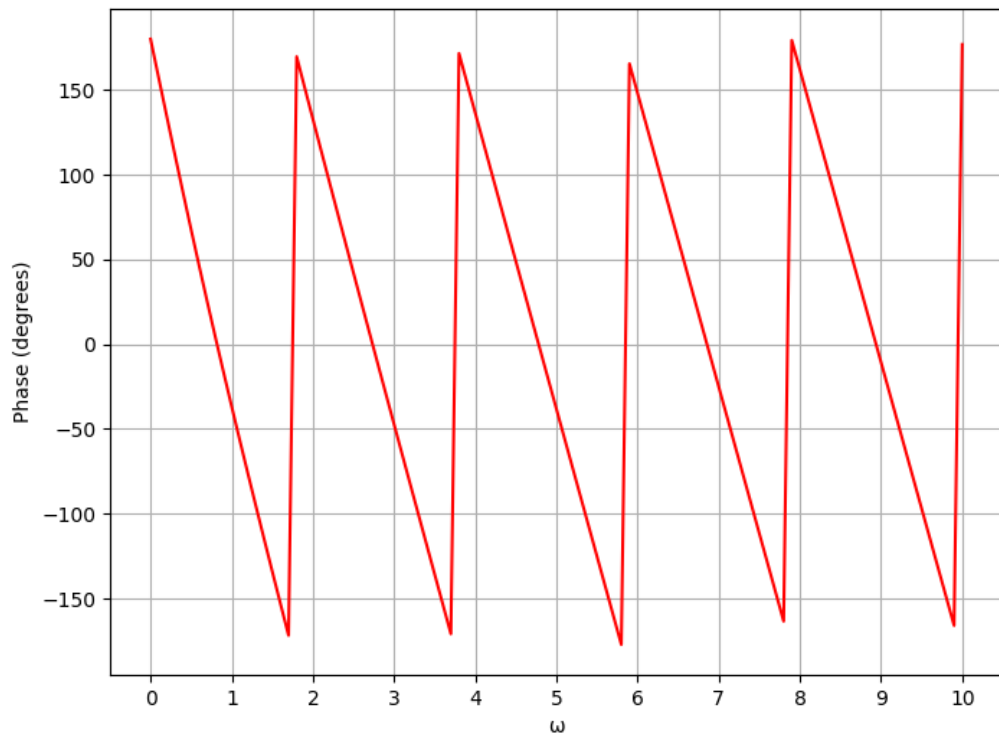


Figure 8.10: Plot for phase of transfer function

variable	value	description
$x_1(t)$	$e^{-t}u(t)$	given function 1
$x_2(t)$	$u(t) - u(t - 2)$	given function 2
$y(t)$	-	convolution of $x_1(t)$ and $x_2(t)$

Table 8.7: Table: Input Parameters

from Table 8.7

$$y(t) = e^{-t}u(t) * (u(t) - u(t - 2)) \quad (8.70)$$

By applying Laplace transform

$$Y(s) = X_1(s) \cdot X_2(s) \quad (8.71)$$

$$e^{-t}u(t) \xleftrightarrow{\mathcal{L}} \frac{1}{1+s}, \quad \operatorname{Re}(s) > -1 \quad (8.72)$$

$$u(t) - u(t-2) \xleftrightarrow{\mathcal{L}} \frac{1 - e^{-2s}}{s}, \quad \operatorname{Re}(s) > 0 \quad (8.73)$$

$$Y(s) = \left( \frac{1}{1+s} \right) \left( \frac{1 - e^{-2s}}{s} \right), \quad \operatorname{Re}(s) > 0 \quad (8.74)$$

$$= \frac{1 - e^{-2s}}{s(s+1)} \quad (8.75)$$

Final value theorem

$$\lim_{t \rightarrow \infty} x(t) = \lim_{s \rightarrow 0} sX(s) \quad (8.76)$$

$$(8.77)$$

Proof:

$$\mathcal{L}[x(t)] = X(s) = \int_0^{\infty} x(t) e^{-st} dt \quad (8.78)$$

$$\mathcal{L}\left[\frac{dx(t)}{dt}\right] = \int_0^{\infty} \frac{d}{dt} (x(t) e^{-st}) dt \quad (8.79)$$

$$= sX(s) - x(0^-) \quad (8.80)$$

$$\lim_{s \rightarrow 0} \left[ \int_0^{\infty} \frac{d}{dt} (x(t) e^{-st}) dt \right] = \lim_{s \rightarrow 0} [sX(s) - x(0^-)] \quad (8.81)$$

$$\int_0^{\infty} \frac{dx(t)}{dt} dt = \lim_{s \rightarrow 0} [sX(s) - x(0^-)] \quad (8.82)$$

$$[x(t)]_0^{\infty} = \lim_{s \rightarrow 0} [sX(s) - x(0^-)] \quad (8.83)$$

$$x(\infty) - x(0^-) = \lim_{s \rightarrow 0} [sX(s) - x(0^-)] \quad (8.84)$$

$$\implies x(\infty) = \lim_{s \rightarrow 0} sX(s) \quad (8.85)$$

$$\lim_{t \rightarrow \infty} x(t) = \lim_{s \rightarrow 0} sX(s) \quad (8.86)$$

By applying Final value theorem

$$\lim_{t \rightarrow \infty} y(t) = \lim_{s \rightarrow 0} sY(s) \quad (8.87)$$

$$= \lim_{s \rightarrow 0} s \left( \frac{1 - e^{-2s}}{s(s+1)} \right) \quad (8.88)$$

$$= \lim_{s \rightarrow 0} \left( \frac{1 - e^{-2s}}{(s+1)} \right) \quad (8.89)$$

$$= \left( \frac{1 - e^0}{0+1} \right) \quad (8.90)$$

$$\lim_{t \rightarrow \infty} y(t) = 0 \quad (8.91)$$



8.10 A unity-gain negative-feedback control system has a loop-gain  $L(s)$  given by

$$L(s) = \frac{6}{s(s-5)} \quad (8.92)$$

The closed loop system is \_\_\_\_\_

- (a) Causal and stable
- (b) Causal and unstable
- (c) Non-causal and stable
- (d) Non-causal and unstable

(GATE IN 2022)

**Solution:** From Table 8.8, the transfer function of the system is given by,

Parameter	Description	Value
$L(s)$	Forward loop transfer function	$\frac{6}{s(s-5)}$
$H(s)$	Feedback path transferfunction	1
$T(s)$	Transfer function	$\frac{L(s)}{1+L(s)H(s)}$

Table 8.8: Parameter Table

$$T(s) = \frac{\frac{6}{s(s-5)}}{1 + 1 \frac{6}{s(s-5)}} \quad (8.93)$$

$$= \frac{6}{s^2 - 5s + 6} \quad (8.94)$$

The poles of the system are given by the roots of the denominator of transfer function,

$$s^2 - 5s + 6 = 0 \quad (8.95)$$

$\therefore$  The poles of the system are  $s = 2$  and  $s = 3$ .

As the poles are positive, the output will increase without bound, causing the system to be unstable.

The transfer function of the system is ,

$$T(s) = \frac{6}{(s-2)(s-3)} \quad (8.96)$$

Clearly, it is dependent only on the past values. Hence, the system is causal.

Thus the correct option is B. The system is causal and unstable.

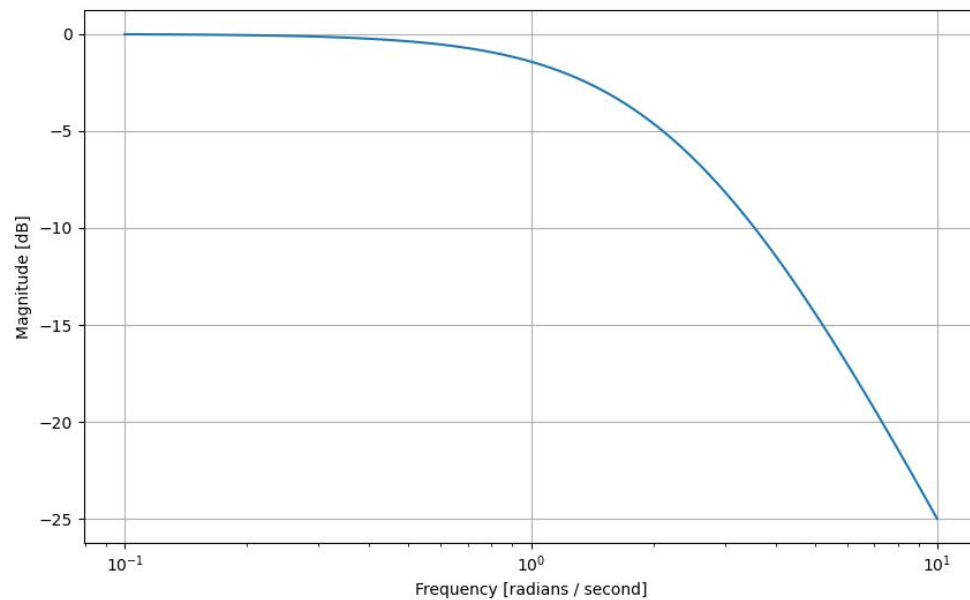


Figure 8.11: Magnitude plot for the transfer function

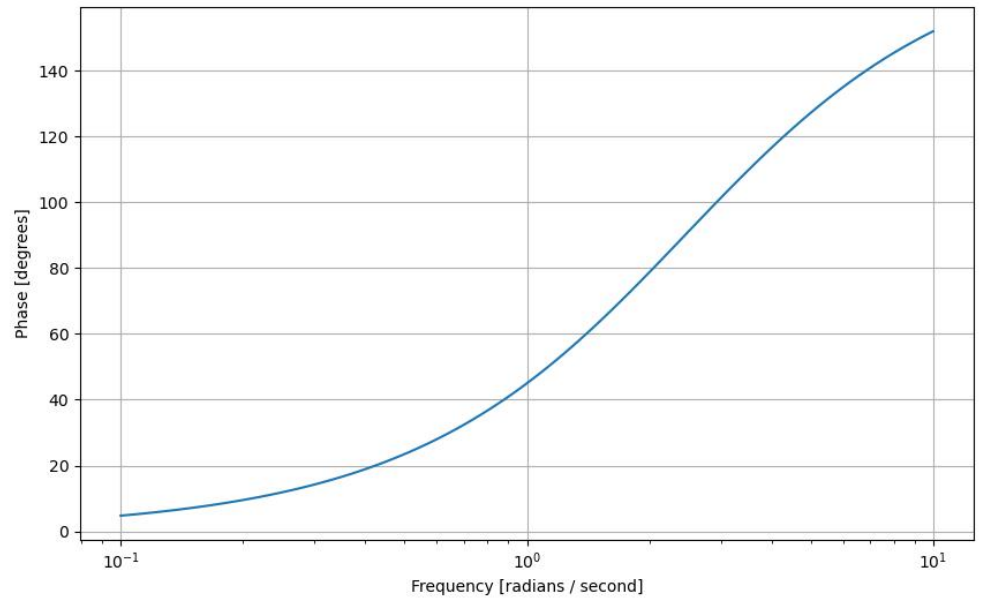


Figure 8.12: Phase plot for the transfer function

8.11 An input  $x(t)$  is applied to a system with a frequency transfer function given by  $H(j\omega)$  as shown below. The magnitude and phase response of the transfer function are shown below. If  $y(t_d) = 0$  for  $x(t) = u(t)$ , the time  $t_d(> 0)$  is.

(Gate 2022 BM.38) **Solution:**

Parameter	Description
$x(t) = u(t)$	Input signal
$y(t)$	Output signal
$X(j\omega)$	Fourier Transform of $x(t)$
$Y(j\omega)$	Fourier Transform of $y(t)$
$H(j\omega) = \frac{Y(j\omega)}{X(j\omega)}$	Transfer function

Table 8.9: Input Parameters Table

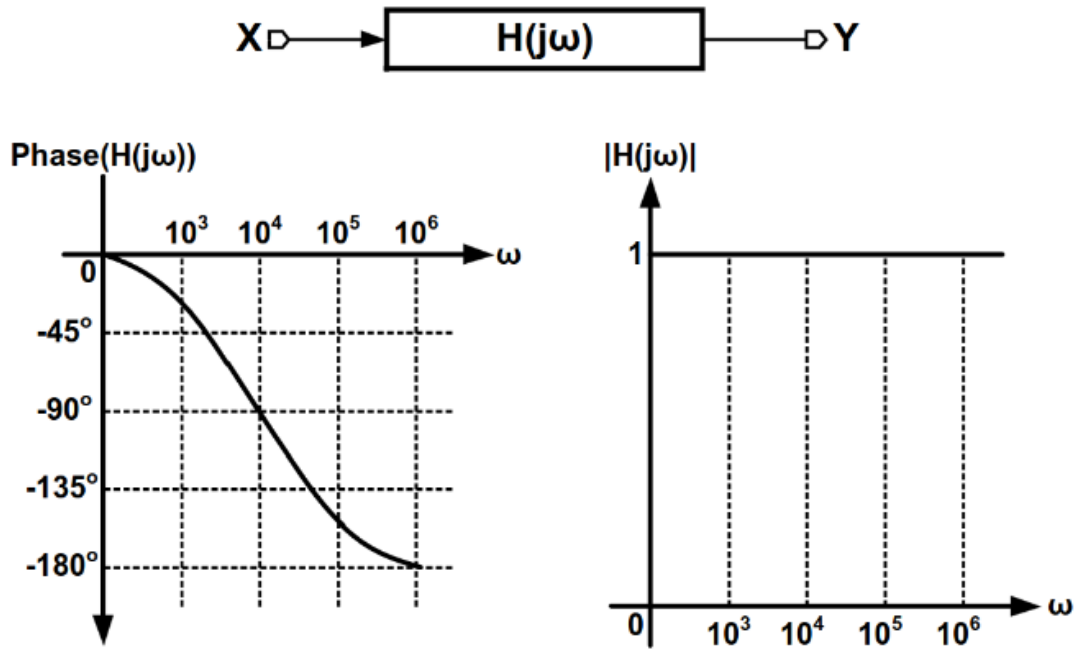


Figure 8.13: Graph of  $y(t)$

from graph 8.14

$$\angle H(j\omega) = -2 \tan^{-1} \left( \frac{\omega}{a} \right) \quad (8.97)$$

$$\text{At } \omega = 10^4, \angle H(j\omega) = -\frac{\pi}{2}$$

$$\Rightarrow a = 10^4 \quad (8.98)$$

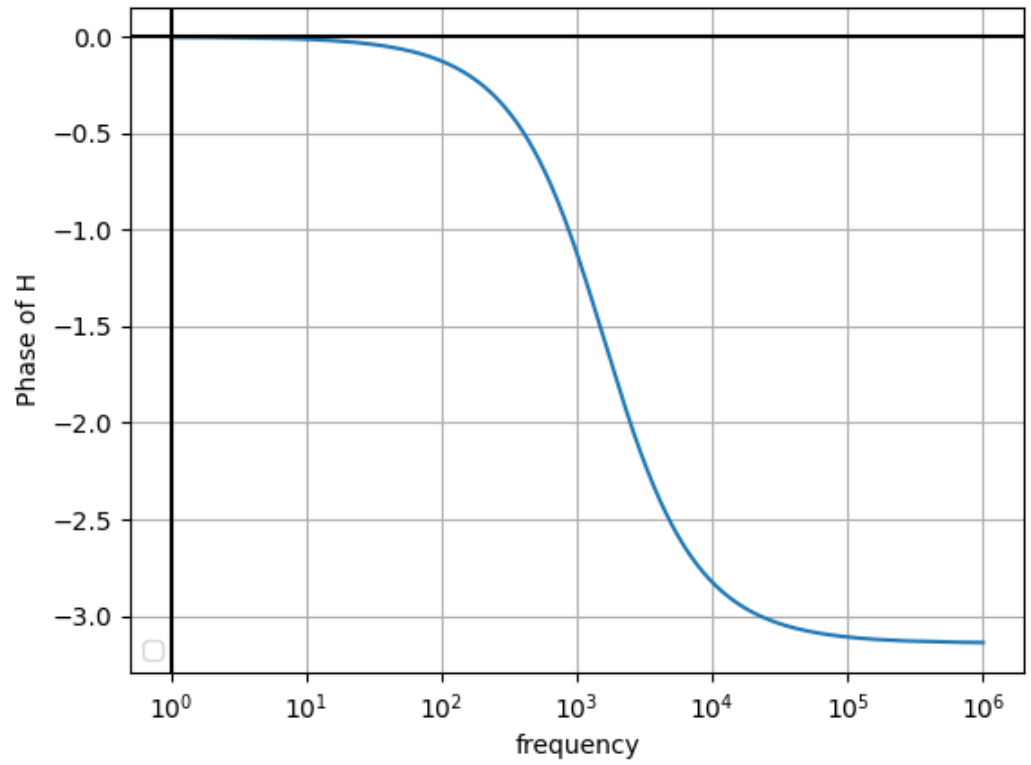


Figure 8.14: Phase of  $H(f)$

$$\angle H(j\omega) = \tan^{-1} \left( \frac{-\omega}{a} \right) - \tan^{-1} \left( \frac{\omega}{a} \right) \quad (8.99)$$

$$H(j\omega) = \frac{e^{j \tan^{-1} \left( \frac{-\omega}{a} \right)}}{e^{j \tan^{-1} \left( \frac{\omega}{a} \right)}} \quad (8.100)$$

$$= \frac{\frac{a-j\omega}{\sqrt{a^2+\omega^2}}}{\frac{a+j\omega}{\sqrt{a^2+\omega^2}}} \quad (8.101)$$

$$= \frac{a-j\omega}{a+j\omega} \quad (8.102)$$

Substitute  $s = j\omega$

$$u(t) \xleftrightarrow{\mathcal{L}} \frac{1}{s} \quad (8.103)$$

$$Y(s) = \frac{1}{s} \frac{a-s}{a+s} \quad (8.104)$$

$$= \frac{1}{s} - \frac{2}{a+s} \quad (8.105)$$

$$\frac{1}{s} \xleftrightarrow{\mathcal{L}^{-1}} u(t) \quad (8.106)$$

$$\frac{1}{a+s} \xleftrightarrow{\mathcal{L}^{-1}} e^{-at} u(t) \quad (8.107)$$

$$y(t) = (1 - 2e^{-at})u(t) \quad (8.108)$$

Verification of laplace transform:

$$\because y(t_d) = 0 \quad (8.109)$$

$$t_d = 100 \ln 2 \mu s \quad (8.110)$$

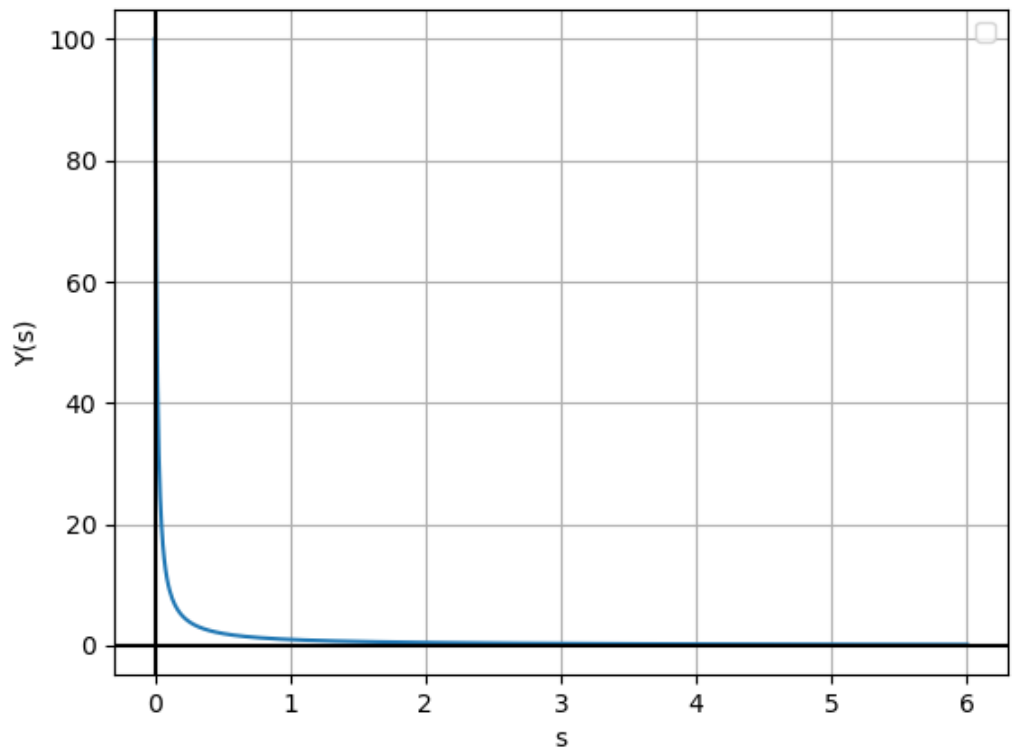


Figure 8.15: Laplace Transform of  $u(t)$

8.12 Consider the circuit shown in the figure with input  $V(t)$  in volts. The sinusoidal steady state current  $I(t)$  flowing through the circuit is shown graphically (where  $t$  is in seconds). The circuit element  $Z$  can be\_\_\_\_\_.

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- (a) a capacitor of 1 F
- (b) an inductor of 1 H
- (c) a capacitor of  $\sqrt{3}$  H
- (d) an inductor of  $\sqrt{3}$  H

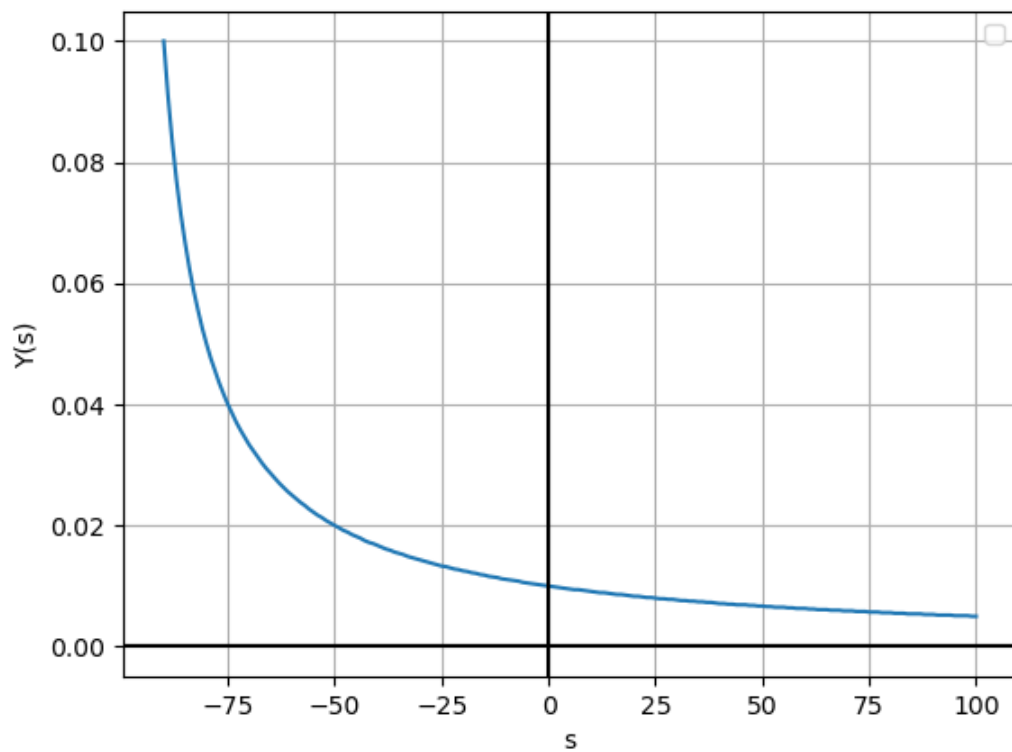
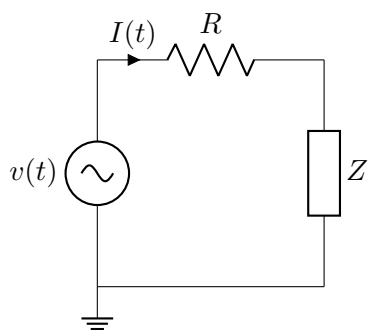
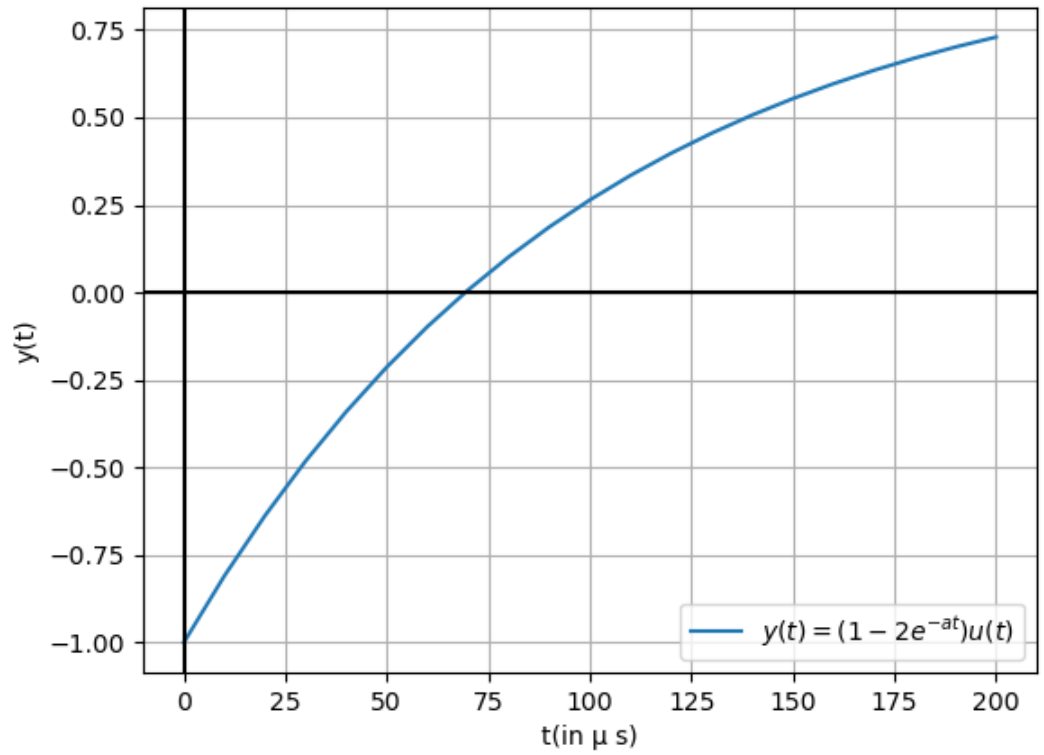


Figure 8.16: Laplace Transform of  $e^{-at}u(t)$



**Solution:**





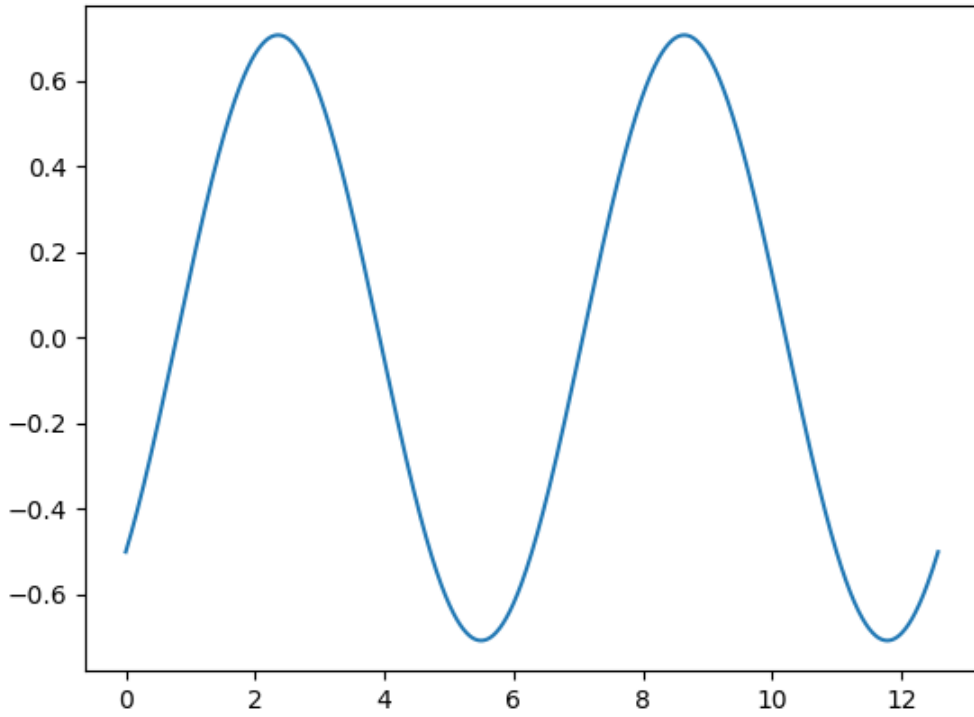
The current through the circuit can be expressed as

$$I(t) = \sin\left(t - \frac{\pi}{4}\right) \quad (8.111)$$

Since, the voltage seems to be leading the current the circuit element  $z$  is an inductor with inductance  $L$ .

Applying KVL in the circuit,

$$R.I(t) + L \frac{dI(t)}{dt} = \sin(t) \quad (8.112)$$



Applying Fourier transform to the differential equation,

$$R.I(s) + sL.I(s) - \frac{1}{s^2 + 1} = 0 \quad (8.113)$$

$$I(s)(R + sL) = \frac{1}{s^2 + 1} \quad (8.114)$$

$$\sin(at + b) \xleftrightarrow{\mathcal{L}} \frac{a \cos(b) + s \sin(b)}{a^2 + s^2} \quad (8.115)$$

$$\sin\left(t - \frac{\pi}{4}\right) \xleftrightarrow{\mathcal{L}} \frac{1 - s}{2(s^2 + 1)} \quad (8.116)$$

$$\frac{1 - s}{2(s^2 + 1)}(R + sL) = \frac{1}{s^2 + 1} \quad (8.117)$$

Symbol	Value	Description
$V(t)$	$\sin t$	Time varying voltage source
$I(t)$	$\sin t - \frac{\pi}{4}$	Current flowing in the circuit
$R$	$1\Omega$	Resistor in series to Z
$Z$	$Z$	Circuit element

Table 8.10: Variable description

Upon plugging in  $R=1\Omega$ ,

$$L = \frac{1}{s} \quad (8.118)$$

Applying inverse Laplace,

$$L = 1H \quad (8.119)$$

8.13 Consider the differential equation  $\frac{dy}{dx} = 4(x + 2) - y$  For the initial condition  $y = 3$  at  $x = 1$ , the value of  $y$  at  $x = 1.4$  obtained using Euler's method with a step-size of 0.2 is ? (round off to one decimal place) (GATE CE 2022)

**Solution:**

Symbols	Description	Values
$R$	Residue Formula	$\frac{1}{(m-1)!} \lim_{s \rightarrow a} \frac{d^{m-1}}{ds^{m-1}} ((s-a)^m f(s) e^{st})$
$\phi(x)$	Transformation of $y(x)$	$y(x+1)$
$g(x)$	Euler's Approximated function of f(x)	$g_{(n-1)}(x) + hf'(x_{n-1}, y_{n-1})$
$h$	Step-size	0.2

Table 8.11: Parameters, Descriptions, and Values

(a) Solution of the differential:

Applying the transformation from table 8.11 and laplace transform

$$s\mathcal{L}(\phi(x)) - \phi(0) = 4\left(\frac{1}{s^2} + \frac{3}{s}\right) - \mathcal{L}(\phi(x)) \quad (8.120)$$

$$\mathcal{L}(\phi(x)) = \frac{3}{s+1} + 4\left(\frac{1}{s^2(s+1)} + \frac{3}{s(s+1)}\right) \quad (8.121)$$

$$= \frac{-5}{s+1} + \frac{8}{s} + \frac{4}{s^2} \quad (8.122)$$

(b) Inverse Laplace Transform:

Using Bromwich integrals and extension of Jordans lemma :

$$\phi(t) = \frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} \mathcal{L}(\phi(x)) e^{st} dt, c > 0 \quad (8.123)$$

$$= \frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} \left(\frac{-5}{s+1} + \frac{8}{s} + \frac{4}{s^2}\right) e^{st} dt \quad (8.124)$$

Here, the poles  $s = -1$  (non repeated,  $m = 1$ ) and  $s = 0$  (repeated,  $m = 2$ ) lie

inside a semicircle for some  $c > 0$ . Using method of residues from 8.11:

$$R_1 = \lim_{s \rightarrow -1} \left( (s+1) \left( \frac{-5}{s+1} \right) e^{st} \right) \quad (8.125)$$

$$= -5e^{-t} \quad (8.126)$$

$$R_2 = \lim_{s \rightarrow 0} \left( (s) \left( \frac{8}{s} \right) e^{st} \right) \quad (8.127)$$

$$= 8 \quad (8.128)$$

$$R_3 = \frac{1}{(1)!} \lim_{s \rightarrow 0} \frac{d}{dz} \left( (s)^2 \left( \frac{4}{s^2} \right) e^{st} \right) \quad (8.129)$$

$$= 4t \quad (8.130)$$

$$\phi(t) = R_1 + R_2 + R_3 \quad (8.131)$$

$$= -5e^{-t} + 8 + 4t \quad (8.132)$$

Reverting to  $y(x)$  , we get :

$$y(x) = -5e^{-x+1} + x + 4x \quad (8.133)$$

Now, approaching to  $y(1.4)$  using euler's approximation from 8.11

$$g_{(1.2)} = 3 + (0.2) f' (1, 3) \quad (8.134)$$

$$= 4.8 \quad (8.135)$$

$$g_{(1.4)} = 4.8 + (0.2) f' (1.2, 4.8) \quad (8.136)$$

$$= 6.4 \quad (8.137)$$

$$\implies y(1.4) \approx 6.4 \quad (8.138)$$

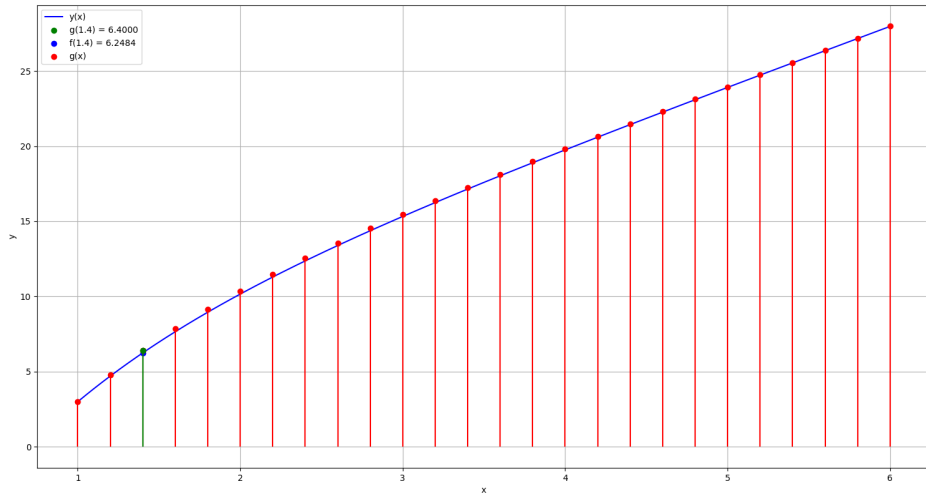


Figure 8.17: Euler's approximated function vs Original function

8.14 A proportional-integral-derivative (PID) controller is employed to stably control a plant with transfer function

$$P(s) = \frac{1}{(s+1)(s+2)} \quad (8.139)$$

Now, the proportional gain is increased by a factor of 2, the integral gain is increased by a factor of 3, and the derivative gain is left unchanged. Given that the closed-loop system continues to remain stable with the new gains, the steady-state error in tracking a ramp reference signal (GATE IN 2022)

**Solution:**

The transfer function of PID controller,

Parameter	Description
$K_P$	Proportional Gain
$K_I$	Integral Gain
$K_D$	Derivative Gain
$r(t)$	Reference Input
$G_c(t)$	Controller Output
$L(t)$	Plant Output
$e(t)$	Error Input

Table 8.12:

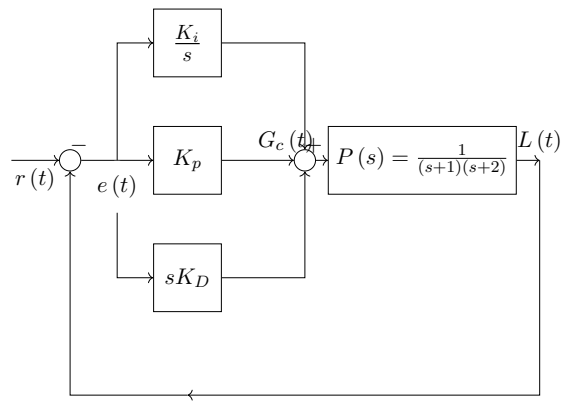


Figure 8.18: Block Diagram of System

$$G_c(s) = K_P + \frac{K_I}{s} + sK_D \quad (8.140)$$

$$= \frac{s^2K_D + sK_P + K_I}{s} \quad (8.141)$$

Overall loop-transfer function,

$$L(s) = G_c(s) \cdot P(s) \quad (8.142)$$

$$L(s) = \frac{s^2K_D + sK_P + K_I}{s(s+1)(s+2)} \quad (8.143)$$

Steady-state error due to ramp signal,

$$e_{ss} = \frac{1}{K_v} \quad (8.144)$$

where,

$$K_v = \lim_{s \rightarrow 0} sL(s) \quad (8.145)$$

$$K_v = \frac{K_I}{2} \quad (8.146)$$

$$e_{ss} = \frac{2}{K_I} \quad (8.147)$$

Now, the proportional gain is increased by a factor of 2, the integral gain is increased by a factor of 3, and the derivative gain is left unchanged.

$$K'_P = 2K_P, K'_I = 3K_I \text{ and } K'_D = K_D \quad (8.148)$$



$$K'_v = \lim_{s \rightarrow 0} sL'(s) \quad (8.149)$$

$$K'_v = \frac{3K_I}{2} \quad (8.150)$$

$$e'_{ss} = \frac{1}{K'_V} = \frac{2}{3K_I} \quad (8.151)$$

8.15 Let  $y(x)$  be the solution of the differential equation

$$y'' - 4y' - 12y = 3e^{5x} \quad (8.152)$$

satisfying  $y(0) = \frac{18}{7}$  and  $y'(0) = \frac{-1}{7}$ .

Then  $y(1)$  is \_\_\_\_\_ (rounded off to nearest integer).

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**Solution:**

Parameter	Description	Value
$y'' - 4y' - 12y = 3e^{5x}$	Differential equation	none
$y(x)$	Solution of differential equation	$y(0) = \frac{18}{7}$
$y'(x)$	First order derivative of solution of differential equation	$y'(0) = \frac{-1}{7}$

Table 8.13: Input Parameters

$$y''(t) \xleftrightarrow{\mathcal{L}} s^2Y(s) - sy(0) - y'(0) \quad (8.153)$$

$$y'(t) \xleftrightarrow{\mathcal{L}} sY(s) - y(0) \quad (8.154)$$

$$y(t) \xleftrightarrow{\mathcal{L}} Y(s) \quad (8.155)$$

$$e^{at} \xleftrightarrow{\mathcal{L}} \frac{1}{s-a} \quad (8.156)$$

Applying Laplace transform on both sides of (8.152),

$$\mathcal{L}\left(y''(t) - 4y'(t) - 12y(t)\right) = \mathcal{L}\left(3e^{5x}\right) \quad (8.157)$$

From (8.153), (8.154), (8.155), (8.156)

$$Y(s) (s^2 - 4s - 12) - y(0) (s - 4) - y'(0) = \frac{3}{s - 5} \quad (8.158)$$

$$Y(s) (s^2 - 4s - 12) - \frac{(18s - 73)}{7} = \frac{3}{(s - 5)} \quad (8.159)$$

$$Y(s) = \frac{3}{(s - 5)(s^2 - 4s - 12)} + \frac{(18s - 73)}{7(s^2 - 4s - 12)} \quad (8.160)$$

$$\Rightarrow Y(s) = \frac{1}{(s - 6)} - \frac{3}{7(s - 5)} + \frac{1}{(s + 2)} \quad (8.161)$$

$$\frac{1}{s - a} \xleftrightarrow{\mathcal{L}^{-1}} e^{at} \quad (8.162)$$

Now finding Inverse Laplace Transform on both sides of (8.161) ,

From (8.162)

$$\Rightarrow y(t) = \left( e^{6t} - \frac{3}{7}e^{5t} + 2e^{-2t} \right) u(t) \quad (8.163)$$

$$\Rightarrow y(1) = e^6 - \frac{3}{7}e^5 + 2e^{-2} \quad (8.164)$$

$$\therefore y(1) = 340 \quad (8.165)$$

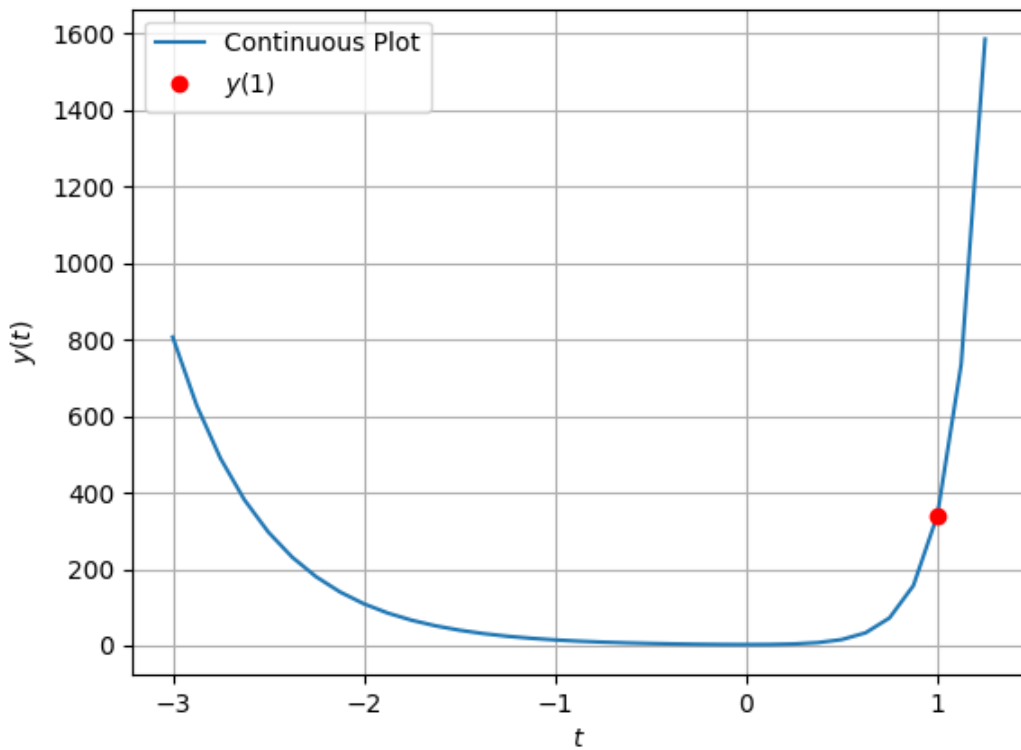


Figure 8.19:

8.16 The signal  $x(t) = (t - 1)^2 u(t - 1)$ , where  $u(t)$  is unit-step function, has the Laplace transform  $X(s)$ . The Value of  $X(1)$  is

- (a)  $\frac{1}{e}$
- (b)  $\frac{2}{e}$
- (c)  $2e$
- (d)  $e^2$

(GATE 2022 IN 40)

**Solution:**

PARAMETER	VALUE	DESCRIPTION
$x(t)$	$x(t) = (t-1)^2 u(t-1)$	Function

Table 8.14: INPUT PARAMETER TABLE

$$x(t) = (t-1)^2 u(t-1) \quad (8.166)$$

Taking Laplace-Transform:

$$t^n u(t) \leftrightarrow \frac{n!}{s^{n+1}} \quad (8.167)$$

if  $X(s)$  is Laplace transform of  $x(t)$  then,

$$x(t-t_0) = e^{-st_0} X(s) \quad (8.168)$$

using 8.167 and 8.168

$$(t-1)^2 u(t-1) \leftrightarrow \frac{2e^{-s}}{s^3} \quad (8.169)$$

$$X(s) = \frac{2e^{-s}}{s^3} \quad (8.170)$$

$$X(1) = \frac{2}{e} \quad (8.171)$$

$\therefore 2$  is Correct.

## Chapter 9

# Fourier transform

9.1 The outputs of four systems ( $S_1, S_2, S_3, S_4$ ) corresponding to the input signal  $\sin(t)$ , for all time  $t$ , are shown in the figure. Based on the given information, which of the four systems is/are definitely NOT LTI(linear and time-invariant)?

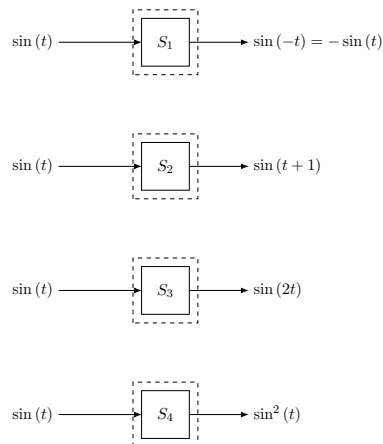


Figure 9.1: Block Diagram of Systems

(GATE22 EC Q46)

**Solution:**

Parameter	Description
$(S_1, S_2, S_3, S_4)$	Systems Given
$\sin(t)$	Input
$H(\omega)$	Transfer Function
$X(\omega)$	Fourier-Transform of input
$Y(\omega)$	Fourier-Transform of output
$\Phi(\omega)$	Phase of Transfer Function

Table 1: Parameter Table

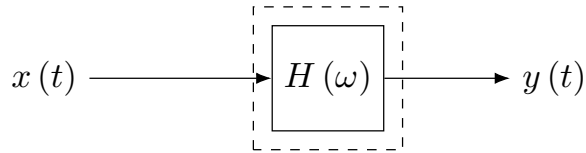


Figure 9.2: Block Diagram of LTI System

For an LTI system :

$$y(t) = h(t) * x(t) \quad (9.1)$$

$$Y(\omega) = H(\omega) X(\omega) \quad (9.2)$$

$H(\omega)$  is a complex exponential :

$$H(j\omega) = |H(j\omega)| e^{j\Phi(\omega)} \quad (9.3)$$

$x(t) = \sin(t)$ , and  $w_o = 1 \text{ rad/sec}$

$$X(\omega) = j\pi (\delta(\omega + \omega_0) - \delta(\omega - \omega_0)) \quad (9.4)$$

Now,



$$Y(\omega) = (\delta(\omega + \omega_0) - \delta(\omega - \omega_0)) \pi |H(\omega)| e^{j\Phi(\omega)} \quad (9.5)$$

$$x(t) \delta(t - t_o) = x(t_o) \delta(t - t_o) \quad (9.6)$$

Using property (9.6) in (9.5) :

$$\begin{aligned} Y(\omega) = & j\pi |H(-\omega_0)| e^{j\Phi(-\omega_0)} \delta(\omega + \omega_0) \\ & - j\pi |H(\omega_0)| e^{j\Phi(\omega_0)} \delta(\omega - \omega_0) \end{aligned} \quad (9.7)$$

By definition of the Fourier transform,

$$X(\omega) = \int_{-\infty}^{\infty} x(t) e^{-j\omega t} dt \quad (9.8)$$

$$X^*(\omega) = \int_{-\infty}^{\infty} x^*(t) e^{j\omega t} dt \quad (9.9)$$

$$X^*(-\omega) = \int_{-\infty}^{\infty} x^*(t) e^{-j\omega t} dt \quad (9.10)$$

For real-time domain signal :

$$x(t) = x^*(t) \quad (9.11)$$

Therefore , from (9.10):

$$X(\omega) = X^*(-\omega) \quad (9.12)$$

By (9.12) , Given  $h(t)$  a real-time domain signal,  $H(\omega)$  is conjugate symmetric.

$$|H(\omega)| = |H(-\omega)| \quad (9.13)$$

$$\Phi(-\omega) = -\Phi(\omega) \quad (9.14)$$

Therefore using (9.13) and (9.14) in (9.7),

$$Y(\omega) = j\pi |H(\omega_0)| \left( e^{-j\Phi(\omega_0)} \delta(\omega + \omega_0) - e^{j\Phi(\omega_0)} \delta(\omega - \omega_0) \right) \quad (9.15)$$

Taking Inverse Fourier Transform,

$$\delta(\omega - \omega_0) \xleftrightarrow{\mathcal{F}} \frac{1}{2\pi} e^{j\omega_0 t} \quad (9.16)$$

$$\delta(\omega + \omega_0) \xleftrightarrow{\mathcal{F}} \frac{1}{2\pi} e^{-j\omega_0 t} \quad (9.17)$$

$$\implies y(t) = j |H(\omega_0)| \frac{1}{2} \left( e^{-j(\omega_0 t + \Phi(\omega_0))} - e^{j(\omega_0 t + \Phi(\omega_0))} \right) \quad (9.18)$$

$$\implies y(t) = |H(\omega_0)| \sin(\omega_0 t + \Phi(\omega_0)) \quad (9.19)$$

$\omega_0 = 1$  rad/sec :

$$y(t) = |H(1)| \sin(t + \Phi(1)) \quad (9.20)$$

From (9.20) we can see output cant have scaled frequency nor a squared output. But can have a shifted output or amplitude-scaled output.

So,  $S_3$  and  $S_4$  cannot be LTI system.

9.2 The Fourier transform  $X(j\omega)$  of the signal

$$x(t) = \frac{t}{(1+t^2)^2} \text{ is } \text{_____}.$$

GATE-2022-EC-15

(A)  $\frac{\pi}{2j}\omega e^{-|\omega|}$

(B)  $\frac{\pi}{2}\omega e^{-|\omega|}$

(C)  $\frac{\pi}{2j}e^{-|\omega|}$

(D)  $\frac{\pi}{2}e^{-|\omega|}$

**Solution:**

Symbol	Value	Description
$x(t)$	$\frac{t}{(1+t^2)^2}$	Signal
$X(\omega)$	$\int_{t=-\infty}^{\infty} x(t) e^{-j\omega t} dt$	Fourier transform of $x(t)$

Table 9.2: Variable description

The Fourier transform of the form  $x(t)=e^{-a|t|}$  is

$$x(t) \xleftrightarrow{\text{F.T.}} X(\omega) \quad (9.21)$$

$$X(\omega) = \frac{2a}{a^2 + \omega^2} \quad (9.22)$$

Consider,

$$x(t) = e^{-|t|} \quad (9.23)$$

$$X(\omega) = \frac{2}{1 + \omega^2} \quad (9.24)$$

By using differentiation property from (A.1.5),

$$tx(t) \xleftrightarrow{\text{F.T.}} j \frac{d}{d\omega} X(\omega) \quad (9.25)$$

$$tx(t) \xleftrightarrow{\text{F.T.}} j \left[ \frac{d}{d\omega} \left( \frac{2}{1 + \omega^2} \right) \right] \quad (9.26)$$

$$te^{-|t|} \xleftrightarrow{\text{F.T.}} \frac{-4j\omega}{(1 + \omega^2)^2} \quad (9.27)$$

Applying duality property from (A.2.3),

$$\frac{-4jt}{(1 + t^2)^2} \xleftrightarrow{\text{F.T.}} 2\pi (-\omega) e^{-|\omega|} \quad (9.28)$$

$$\frac{t}{(1 + t^2)^2} \xleftrightarrow{\text{F.T.}} \frac{-2\pi\omega e^{-|\omega|}}{-4j} \quad (9.29)$$

$$\frac{t}{(1 + t^2)^2} \xleftrightarrow{\text{F.T.}} \frac{\pi}{2j} \omega e^{-|\omega|} \quad (9.30)$$

9.3 For a vector  $\bar{x} = [x[0], x[1], \dots, x[7]]$ , the 8-point discrete Fourier transform (DFT) is denoted by  $\bar{X} = \text{DFT}(\bar{x}) = [X[0], X[1], \dots, X[7]]$ , where

$$X[k] = \sum_{n=0}^7 x[n] \exp \left( -j \frac{2\pi}{8} nk \right).$$

Here  $j = \sqrt{-1}$ . If  $\bar{x} = [1, 0, 0, 0, 2, 0, 0, 0]$  and  $\bar{y} = \text{DFT}(\text{DFT}(\bar{x}))$ , then the value of  $y[0]$  is.

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**Solution:**

Parameter	Description	Value
$\bar{X}$	$\text{DFT}(\bar{x})$	—
$\bar{x}$	vector	$[1, 0, 0, 0, 2, 0, 0, 0]$
$\bar{y}$	$\text{DFT}(\text{DFT}(\bar{x}))$	—

Table 9.3: Given Parameters

DFT of  $\bar{x}$

$$X[k] = \sum_{n=0}^7 x[n] \exp \left( -j \frac{2\pi}{8} nk \right) \quad (9.31)$$

As the only non-zero values in  $x$  are  $x[0]$  and  $x[4]$ :

$$X[k] = x[0] + x[4] \exp(-j\pi k) \quad (9.32)$$

After substituting the values of  $k$  ranging from 0 to 7,

$$\bar{X} = \text{DFT}(\bar{x}) = [X[0], X[1], \dots, X[7]] \quad (9.33)$$

$$\bar{X} = [3, -1, 3, -1, 3, -1, 3, -1] \quad (9.34)$$

$$\bar{y} = \text{DFT}(\text{DFT}(\bar{x})) \quad (9.35)$$

$$\bar{y} = [3, -1, 3, -1, 3, -1, 3, -1] \quad (9.36)$$

$$y[0] = \sum_{n=0}^7 x[n] \quad (9.37)$$

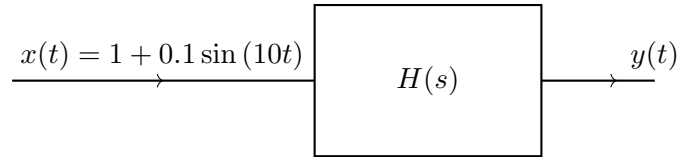
$$= x[0] + x[1] + \cdots + x[7] \quad (9.38)$$

$$= 3 - 1 + 3 - 1 + 3 - 1 + 3 - 1 = 8 \quad (9.39)$$

9.4 **Question:** An LTI system is shown in the figure where

$$H(s) = \frac{100}{s^2 + 0.1s + 10}$$

The steady state output of the system for an input  $x(t)$  is given by  $y(t) = a + b \sin(10t + \theta)$ . The values of ' $a$ ' and ' $b$ ' are



**Solution:**

Symbol	Value	Description
$x(t)$	$1 + 0.1 \sin(10t)$	Input Signal
$y(t)$	?	Output of the system
$H(s)$	$\frac{100}{s^2 + 0.1s + 10}$	Impulse Response

Table 9.4: Given Information

(a) **Theory:** If a sinusoidal input is given to a system, whose transfer function is known, the output can be calculated as follows

$$y(t) = h(t) * x(t) \quad (9.40)$$

$$Y(s) = H(s)X(s) \quad (9.41)$$

Let  $s = j\omega$

$$Y(j\omega) = H(j\omega)X(j\omega) \quad (9.42)$$

If  $\Phi$  is the phase of  $H(j\omega)$ ,

$$H(j\omega) = |H(j\omega)| e^{j\Phi(\omega)} \quad (9.43)$$

If  $x(t) = \cos(\omega_0 t)$ ,

$$X(j\omega) = \pi (\delta(\omega - \omega_0) + \delta(\omega + \omega_0)) \quad (9.44)$$

Now,

$$Y(j\omega) = (\delta(\omega - \omega_0) + \delta(\omega + \omega_0)) |H(j\omega)| e^{j\Phi(\omega)} \quad (9.45)$$

$$(9.46)$$

Since  $|H(j\omega)| \delta(\omega - \omega_0)$  is zero everywhere except at  $\omega_0$

$$Y(j\omega) = |H(j\omega_0)| e^{j\Phi(\omega_0)} \delta(\omega - \omega_0) \quad (9.47)$$

$$+ |H(-j\omega_0)| e^{j\Phi(-j\omega_0)} \delta(\omega + \omega_0) \quad (9.48)$$

As  $h(t)$  is real,

$$H(\omega) = H^*(-\omega)$$

$$\Phi(-\omega_0) = -\Phi(\omega_0)$$

Hence

$$Y(\omega) = |H(\omega_0)| \left( e^{j\Phi(\omega_0)} \delta(\omega - \omega_0) + e^{-j\Phi(\omega_0)} \delta(\omega + \omega_0) \right) \quad (9.49)$$



Taking Inverse Fourier Transform,

$$\delta(\omega - \omega_0) \xleftrightarrow{\mathcal{F}} \frac{1}{2} e^{j\omega_0 t} \quad (9.50)$$

$$\Rightarrow y(t) = |H(\omega_0)| \frac{1}{2} \left( e^{j(\omega_0 t + \Phi(\omega_0))} + e^{-j(\omega_0 t + \Phi(\omega_0))} \right) \quad (9.51)$$

$$\Rightarrow y(t) = |H(\omega_0)| \cos(\omega_0 t + \Phi(\omega_0)) \quad (9.52)$$

(b) The given input can be assumed to be a superposition of  $u(t)$  and  $0.1 \sin(\omega_0 t)u(t)$ .

$$\omega_0 = 0 \text{ and } \omega_0 = 10$$

for the constant input and the sinusoidal input respectively.

$$y(t) = |H(0)| + |H(10)| \sin(10t + \Phi(10)) \quad (9.53)$$

Here

$$H(\omega) = \frac{100}{(j\omega)^2 + 0.1(j\omega) + 10} \quad (9.54)$$

$$\Rightarrow H(\omega) = \frac{100}{10 - \omega^2 + j(0.1\omega)} \quad (9.55)$$

$$\Rightarrow |H(\omega)| = \frac{100}{\sqrt{(10 - \omega^2)^2 + (0.1\omega)^2}} \quad (9.56)$$

$$\therefore |H(0)| = 10 \text{ and } |H(10)| \approx 1 \quad (9.57)$$

The phase  $\Phi(\omega)$  is given by

$$\Phi(\omega) = \tan^{-1} \frac{0.1\omega}{\omega^2 - 10} \quad (9.58)$$

$$\Rightarrow \Phi(10) = \tan^{-1} \frac{1}{90} \quad (9.59)$$

Hence the output of the system

$$y(t) = 10 + \sin\left(10t + \tan^{-1} \frac{1}{90}\right) \quad (9.60)$$

Hence  $a = 10$  and  $b = 1$

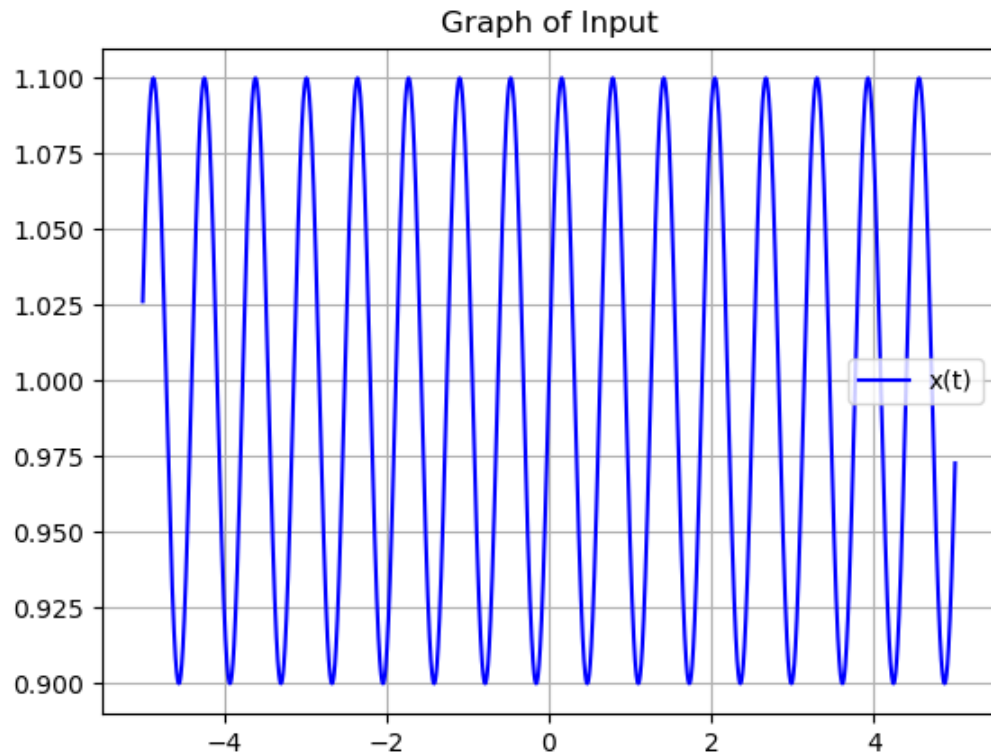


Figure 9.3: Input of the system,  $x(t)$

9.5 A periodic function  $f(x)$ , with period 2, is defined as

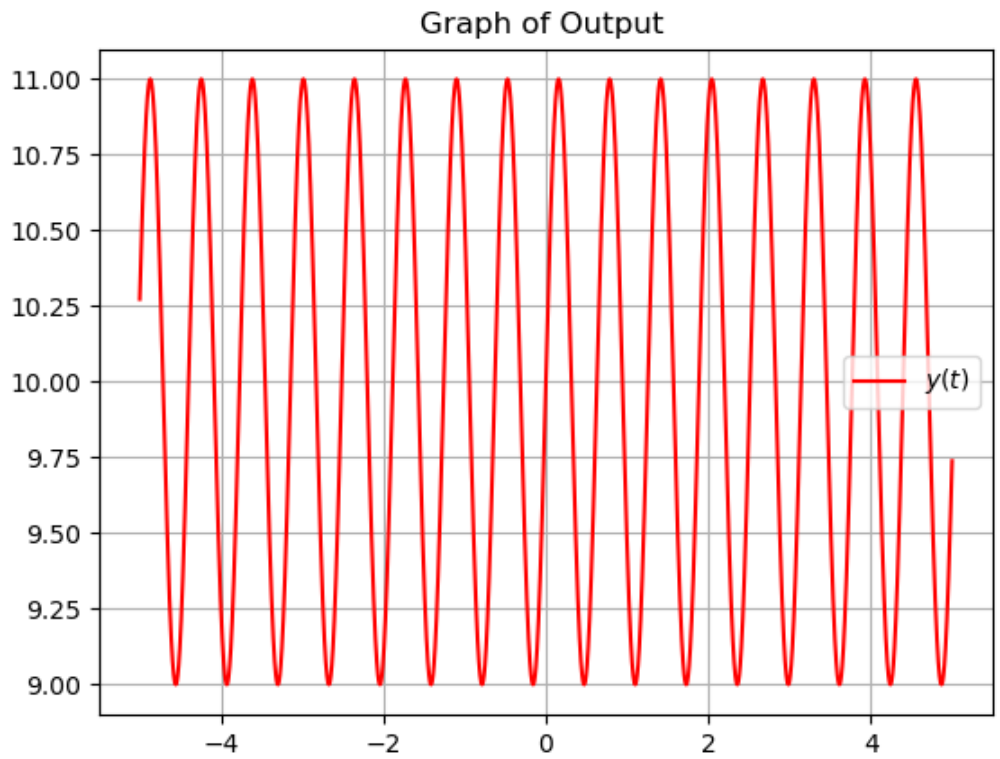


Figure 9.4: Output of the system,  $y(t)$

$$f(x) = \begin{cases} -1 - x & -1 \leq x < 0 \\ 1 - x & 0 < x \leq 1 \end{cases} \quad (9.61)$$

The Fourier series of this function contains

- A. Both  $\cos(n\pi x)$  and  $\sin(n\pi x)$  where  $n=1,2,3\dots$
- B. Only  $\sin(n\pi x)$  where  $n=1,2,3\dots$

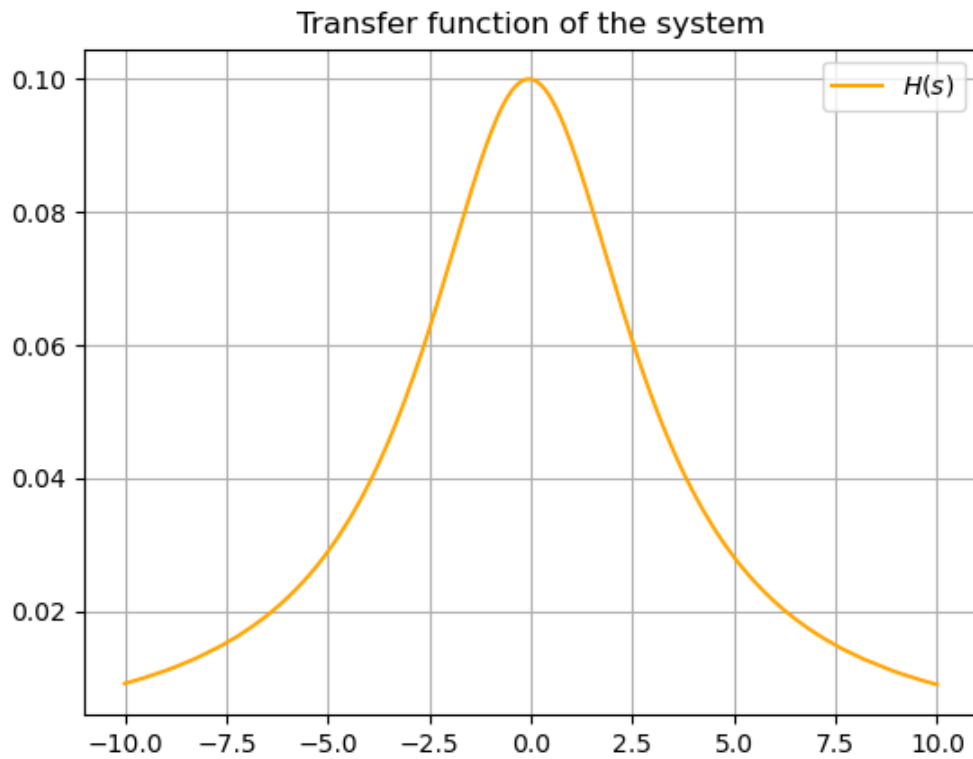


Figure 9.5: Transfer function of the system,  $H(s)$

C. Only  $\cos(n\pi x)$  where  $n=1,2,3\dots$

D. Only  $\cos(2n\pi x)$  where  $n=1,2,3\dots$

GATE IN 2022

**Solution:**

Parameter	Description
$f(x)$	Polynomial function
$2L$	Period of the Polynomial function
$c(n)$	Complex Fourier Coefficients
$a(0), a(n), b(n)$	Trigonometric Fourier Coefficients

Table 9.5: Input Parameters

The complex exponential Fourier Series of  $f(x)$  is,

$$f(x) = \sum_{n=-\infty}^{\infty} c(n) e^{jn\omega x} \quad (9.62)$$

$$\Rightarrow c(n) = \frac{1}{2L} \int_{-L}^L f(x) e^{-jn\omega x} dx \quad (9.63)$$

For  $n \neq 0$ ;

$$c(n) = \frac{1}{2} \int_{-1}^1 f(x) e^{-jn\omega x} dx \quad (9.64)$$

$$= \frac{1}{2} \left( \int_{-1}^0 (-1-x) e^{-jn\omega x} dx + \int_0^1 (+1-x) e^{-jn\omega x} dx \right) \quad (9.65)$$

$$= \frac{1}{2} \left( - \int_{-1}^0 e^{-jn\omega x} dx - \int_{-1}^1 x e^{-jn\omega x} dx + \int_0^1 e^{-jn\omega x} dx \right) \quad (9.66)$$

$$= \frac{1}{2} \left[ \frac{-1}{jn\omega} [- (1 - e^{+jn\omega}) + (e^{-jn\omega} - 1)] - \int_{-1}^1 x e^{-jn\omega x} dx \right] \quad (9.67)$$

$$= \frac{1}{2} \left[ \frac{-1}{jn\omega} [-2 + e^{+jn\omega} + e^{-jn\omega}] + \left( \frac{e^{-jn\omega x}}{jn\omega} \left[ x + \frac{1}{jn\omega} \right] \right)_{-1}^1 \right] \quad (9.68)$$

$$= \frac{-1}{jn\omega} [-1 + \cos(n\omega)] + \frac{1}{2(jn\omega)^2} [(e^{-jn\omega})(1 + jn\omega) - (e^{jn\omega})(-jn\omega + 1)] \quad (9.69)$$

$$\Rightarrow c(n) = \frac{-1}{(jn\omega)^2} [-jn\omega + j \sin(n\omega)] \quad (9.70)$$

For  $n = 0$ ;

$$c(0) = \frac{1}{2} \int_{-1}^1 f(x) dx \quad (9.71)$$

$$= \frac{1}{2} \left[ \int_{-1}^0 (-1-x) dx + \int_0^1 (1-x) dx \right] \quad (9.72)$$

$$= \frac{1}{2} \left[ \left( -x - \frac{x^2}{2} \right)_{-1}^0 + \left( x - \frac{x^2}{2} \right)_0^1 \right] \quad (9.73)$$

$$= \frac{1}{2} \left[ 0 - 1 + \frac{1}{2} + 1 - \frac{1}{2} - 0 \right] \quad (9.74)$$

$$\implies c(0) = 0 \quad (9.75)$$

The trigonometric Fourier Series of  $f(x)$  is,

$$f(x) = a(0) + \sum_{n=1}^{\infty} \{a(n) \cos(n\omega x) + b(n) \sin(n\omega x)\} \quad (9.76)$$

Finding the Fourier Coefficient  $a_0$ ,

$$a(0) = c(0) \quad (9.77)$$

$$\implies a(0) = 0 \quad (9.78)$$

Finding the Fourier Coefficients  $a(n)$ ,

$$a(n) = \frac{1}{L} \int_{-L}^L f(x) \cos(n\omega x) dx, n \geq 0 \quad (9.79)$$

$$= \frac{1}{L} \int_{-L}^L f(x) (e^{-jn\omega x} + e^{jn\omega x}) dx \quad (9.80)$$

$$\implies a(n) = c(n) + c(-n) \quad (9.81)$$

$$\implies a(n) = 0 \quad (9.82)$$

Finding the Fourier Coefficients  $b(n)$ ,

$$b_n = \frac{1}{L} \int_{-L}^L f(x) \sin(n\omega x) dx, n \geq 0 \quad (9.83)$$

$$= \frac{1}{L} \int_{-L}^L f(x) j (e^{-jn\omega x} - e^{jn\omega x}) dx \quad (9.84)$$

$$\Rightarrow b(n) = j (c(n) - c(-n)) \quad (9.85)$$

$$\Rightarrow b(n) = \frac{-2}{(n\omega)^2} [-n\omega + \sin(n\omega)] \quad (9.86)$$

$\Rightarrow$  The trigonometric Fourier Series of  $f(x)$  is,

$$f(x) = \sum_{n=1}^{\infty} \{0 + 0 + b(n) \sin(n\omega x)\} \quad (9.87)$$

$$f(x) = \sum_{n=1}^{\infty} \left\{ \frac{-2}{(n\omega)^2} [-n\omega + \sin(n\omega)] \sin(n\omega x) \right\} \quad (9.88)$$

$$f(x) = \sum_{n=1}^{\infty} \left\{ \frac{-2}{(n\pi)^2} [-n\pi + \sin(n\pi)] \sin(n\pi x) \right\} \quad (9.89)$$

$$f(x) = \sum_{n=1}^{\infty} \left\{ \frac{2}{n\pi} \sin(n\pi x) \right\} \quad (9.90)$$

$\therefore$  The Fourier series of this function contains only  $\sin(n\pi x)$  where  $n=1,2,3,\dots$

9.6 A Simple closed path C in the Complex Plane is shown in the figure.

$$\oint_C \frac{2^z}{z^2 - 1} dz = -j\pi A$$

Where  $j = \sqrt{-1}$ , Then find the value of A is \_\_\_\_\_ (Rounded of to two decimals)

(GATE 2022 EC)

**Solution:** Let

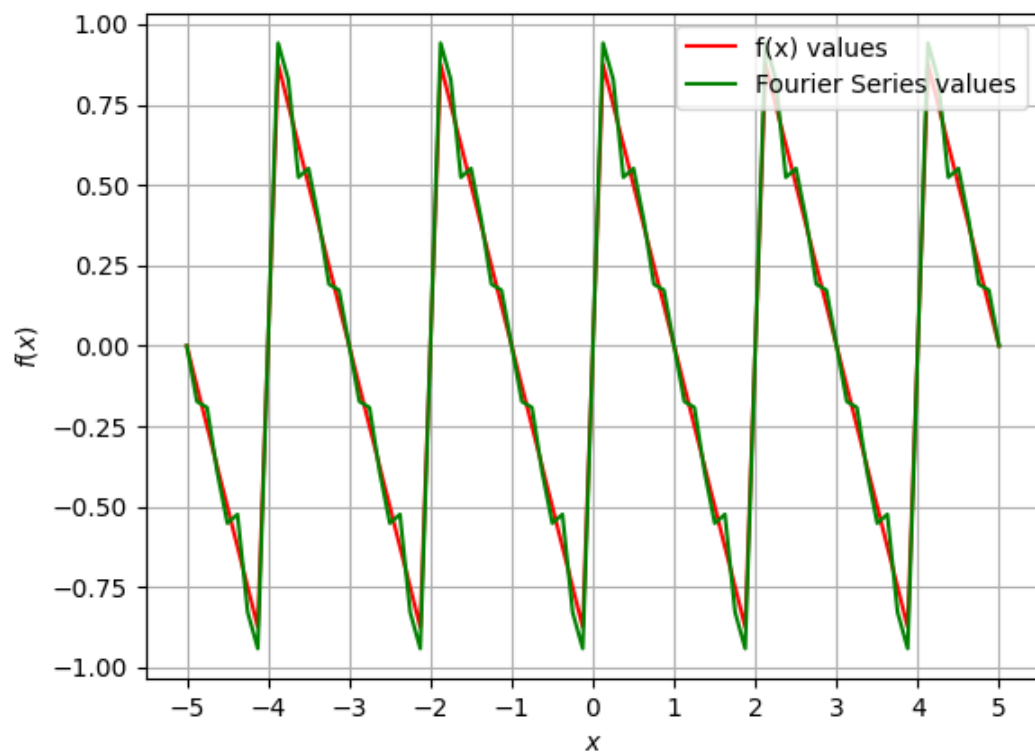
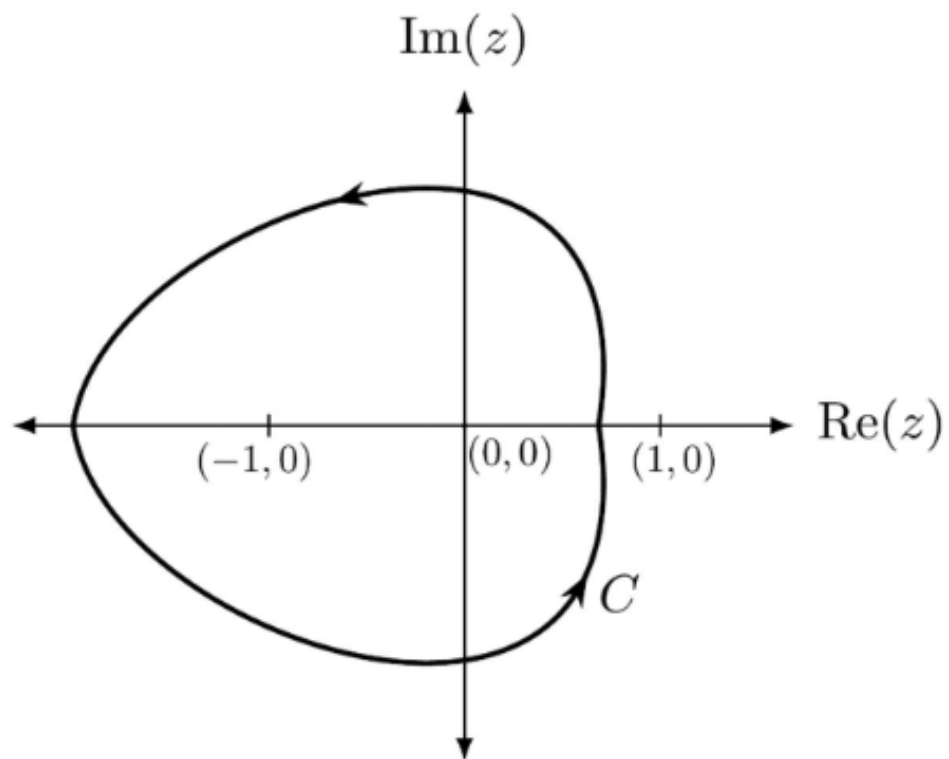


Figure 9.6:

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





For poles

$$z^2 - 1 = 0 \quad (9.91)$$

$$\implies z = \pm 1 \quad (9.92)$$

As  $Z = -1$  lies inside the  $C$  and  $z = 1$  lies outside  $C$

$$\oint_C f(z) dz = \oint_C \frac{2^z}{z+1} dz \quad (9.93)$$

$$= 2\pi j \left( \frac{2^z}{z-1} \right) \text{ At } z = -1 \quad (9.94)$$

(By Cauchy's integral formula )

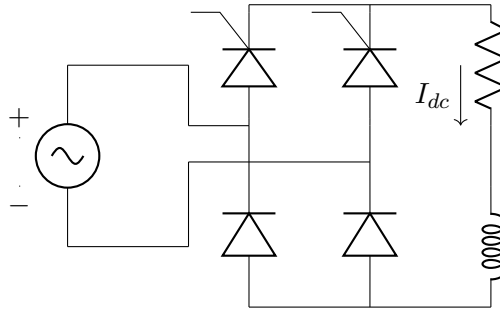
$$= 2\pi j \left( \frac{-1}{4} \right) \quad (9.95)$$

$$= -\pi j \left( \frac{1}{2} \right) \quad (9.96)$$

By comparing

$$A = \frac{1}{2} = 0.50 \quad (9.97)$$

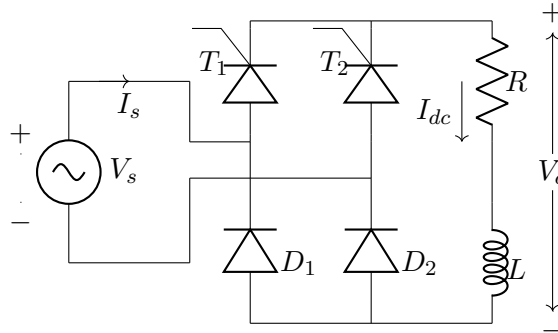
9.7 For the ideal AC-DC rectifier circuit shown in the figure below, the load current magnitude is  $I_{dc} = 15$  A and is ripple free. The thyristors are fired with a delay angle of  $45^\circ$ . The amplitude of the fundamental component of the source current, in amperes, is -----(Round off to 2 decimal places). (GATE 59 EE 2022) **Solution:**



Parameter	Description	Value
$I_{dc}$	Load current	15A
$\alpha$	Firing angle	$45^\circ$

Table 9.6:

A symmetrical single phase semi converter is shown below,



The Fourier series representation of supply current is given by:

$$i_s(t) = a_o + \sum_{n=1}^{\infty} C_n \sin(n\omega t + \theta_n) \quad (9.98)$$

where,

$$a_o = \frac{1}{2\pi} \int_0^{2\pi} i_s(t) d\omega t \quad (9.99)$$

$$C_n = \sqrt{a_n^2 + b_n^2} \quad (9.100)$$

$$\theta_n = \tan^{-1} \left( \frac{a_n}{b_n} \right) \quad (9.101)$$

$$\Rightarrow a_o = \frac{1}{2\pi} \int_{\alpha}^{\pi} I_o d\omega t - \int_{\pi+\alpha}^{2\pi} I_o d\omega t = 0 \quad (9.102)$$

$$\Rightarrow a_n = \frac{1}{\pi} \int_{\alpha}^{\pi} I_o \cos n\omega t d\omega t - \int_{\pi+\alpha}^{2\pi} I_o \cos n\omega t d\omega t \quad (9.103)$$

$$a_n = \begin{cases} \frac{-2I_o}{n\pi} \sin n\alpha & \text{for } n = 1, 3, 5... \\ 0 & \text{for } n = 2, 4, 6... \end{cases} \quad (9.104)$$

$$\Rightarrow b_n = \frac{1}{\pi} \int_{\alpha}^{\pi} I_o \sin n\omega t d\omega t - \int_{\pi+\alpha}^{2\pi} I_o \sin n\omega t d\omega t \quad (9.105)$$

$$b_n = \begin{cases} \frac{2I_o}{n\pi} (1 + \cos n\alpha) & \text{for } n = 1, 3, 5... \\ 0 & \text{for } n = 2, 4, 6... \end{cases} \quad (9.106)$$

From (9.100):

$$\therefore C_n = \frac{2\sqrt{2}I_o}{n\pi} (\sqrt{1 + \cos n\alpha}) \quad (9.107)$$

$$\Rightarrow C_n = \frac{4I_o}{n\pi} \cos \frac{n\alpha}{2} \quad (9.108)$$

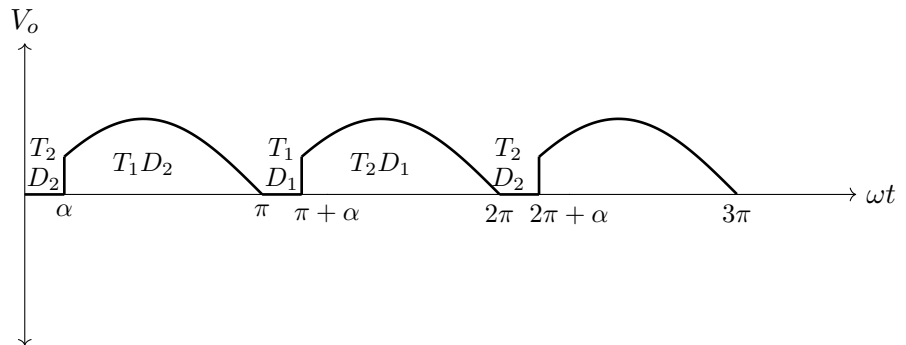
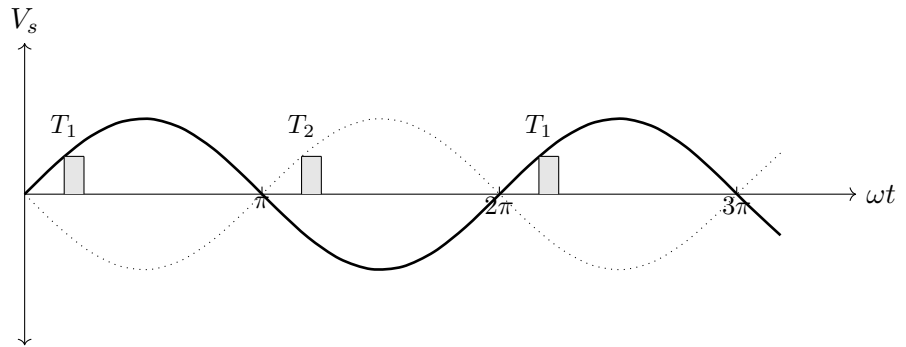
From (9.101):

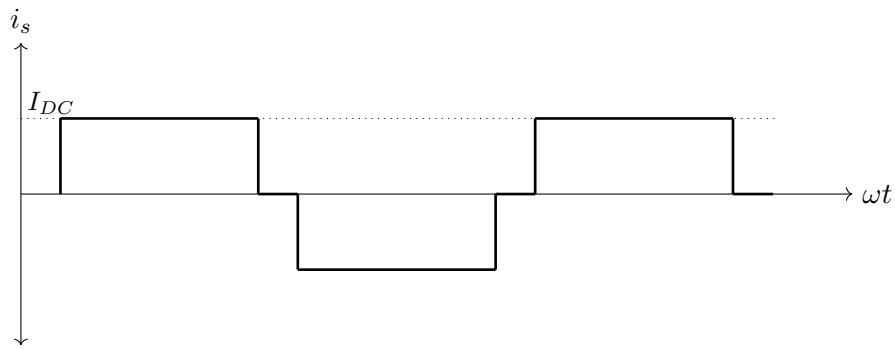
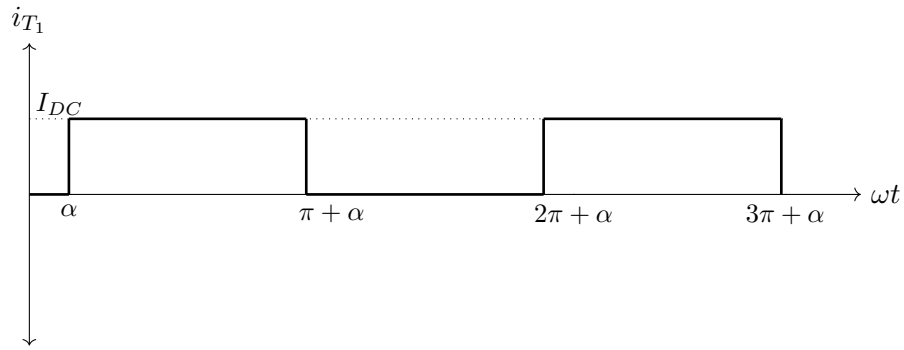
$$\theta_n = \tan^{-1} \left( \frac{-\sin n\alpha}{1 + \cos n\alpha} \right) \quad (9.109)$$

$$\Rightarrow \theta_n = \frac{-n\alpha}{2} \quad (9.110)$$

From (9.98), (9.108) and (9.110):

$$I_s(t) = \sum_{n=1,3,5,\dots}^{\infty} \frac{4I_o}{n\pi} \cos \frac{n\alpha}{2} \sin \left( n\omega t - \frac{n\alpha}{2} \right) \quad (9.111)$$





From Table 9.6:

$$(I_{s1})_{peak} = \frac{4I_{dc}}{\pi} \cos\left(\frac{\alpha}{2}\right) \quad (9.112)$$

$$= \frac{4 \times 15}{\pi} \times \cos \frac{45^\circ}{2} \quad (9.113)$$

$$= 17.64A \quad (9.114)$$

9.8 If

$$\frac{a_0}{2} + \sum_{n=1}^{\infty} a_n \cos(nx)$$

is the Fourier cosine series of the function

$$f(x) = \sin(x), 0 < x < \pi$$

then which of the following are TRUE?

(a)  $a_0 + a_1 = \frac{4}{\pi}$

(b)  $a_0 = \frac{4}{\pi}$

(c)  $a_0 + a_1 = \frac{2}{\pi}$

(d)  $a_1 = \frac{2}{\pi}$

(GATE 2022 NM Q24)

**Solution:**

Symbol	Value	Description
$a_0, a_n, b_n$		Fourier Series Coefficients
$T$	$\pi$	Time Period
$n$		Positive Integer

Table 9.7: Input Parameters

Fourier series of a function  $f(x)$ :

$$f(x) = a_0 + \sum_{n=1}^{\infty} a_n \cos(n\omega x) + \sum_{n=1}^{\infty} b_n \sin(n\omega x) \quad (9.115)$$

where,

$$a_0 = \frac{1}{T} \int_T f(x) dx \quad (9.116)$$

$$a_n = \frac{2}{T} \int_T f(x) \cos(n\omega x) dx \quad (9.117)$$

$$b_n = \frac{2}{T} \int_T f(x) \sin(n\omega x) dx \quad (9.118)$$

Calculating for given function:

$$\frac{a_0}{2} = \frac{1}{\pi} \int_0^\pi \sin(x) dx \quad (9.119)$$

$$\implies a_0 = \frac{4}{\pi} \quad (9.120)$$

$$a_n = \frac{2}{\pi} \int_0^\pi \sin(x) \cos(nx) dx \quad (9.121)$$

$$\implies a_1 = \frac{2}{\pi} \int_0^\pi \sin(x) \cos(x) dx \quad (9.122)$$

$$= 0 \quad (9.123)$$

Calculating general  $a_n$ :

$$a_n = \frac{2}{\pi} \int_0^\pi \sin(x) \cos(nx) dx \quad (9.124)$$

$$= \frac{1}{\pi} \int_0^\pi (\sin(x + nx) + \sin(x - nx)) dx \quad (9.125)$$

$$= \frac{1}{\pi} \left[ \frac{-\cos((n+1)x)}{n+1} + \frac{-\cos((1-n)x)}{1-n} \right]_0^\pi \quad (9.126)$$

$$= \frac{2(1 + \cos(n\pi))}{\pi(1 - n^2)} \quad (9.127)$$



From (9.120) and (9.123),

$$a_0 + a_1 = \frac{4}{\pi} \quad (9.128)$$

$$a_0 = 0 \quad (9.129)$$

$\therefore$  correct options are (a) and (b).

9.9 The fourier series expansion of  $x^3$  in the interval  $-1 \leq x \leq 1$  with periodic continuation has

- (a) only sine terms
- (b) only cosine terms
- (c) both sine and cosine terms
- (d) only sine terms and a non-zero constant

(GATE 2022 ME)

**Solution:**

Fourier series expansion of the function  $x(t)$  in the interval  $[-L, L]$  can be given by:

$$x(t) = a_0 + \sum_{n=1}^{\infty} a_n \cos\left(\frac{n\pi t}{L}\right) + \sum_{n=1}^{\infty} b_n \sin\left(\frac{n\pi t}{L}\right) \quad (9.130)$$

where,

$$a_0 = \frac{1}{2L} \int_{-L}^L f(t) dt \quad (9.131)$$

$$a_n = \frac{1}{2L} \int_{-L}^L f(t) \cos\left(\frac{n\pi t}{L}\right) dt \quad (9.132)$$

$$b_n = \frac{1}{2L} \int_{-L}^L f(t) \sin\left(\frac{n\pi t}{L}\right) dt \quad (9.133)$$

$$(9.134)$$

Therefore, the expansion can be given by:

$$t^3 = a_0 + \sum_{n=1}^{\infty} a_n \cos(n\pi t) + \sum_{n=1}^{\infty} b_n \sin(n\pi t) \quad (9.135)$$

Since  $t^3$  is an odd function,

$$a_0 = a_n = 0 \quad (9.136)$$

$$b_n = \frac{1}{2} \int_{-1}^1 t^3 \sin(n\pi t) dt \quad (9.137)$$

$$= (-1)^{n+1} \left( \frac{2}{n\pi} - \frac{12}{(n\pi)^3} \right) \quad (9.138)$$

$$\Rightarrow t^3 = \sum_{n=1}^{\infty} b_n \sin\left(\frac{n\pi t}{L}\right) \quad (9.139)$$

∴ It contains only sine terms.

9.10 The discrete time Fourier series representation of a signal  $x[n]$  with period  $N$  is written as  $x[n] = \sum_{k=0}^{N-1} a_k e^{j(2kn\pi/N)}$ . A discrete time periodic signal with period  $N = 3$ , has the non-zero Fourier series coefficients:  $a_{-3} = 2$  and  $a_4 = 1$ . The signal is

(A)  $2 + 2e^{-j\frac{2\pi}{6}n} \cos\left(\frac{2\pi}{6}n\right)$

(B)  $1 + 2e^{j\frac{2\pi}{6}n} \cos\left(\frac{2\pi}{6}n\right)$

(C)  $1 + 2e^{j\frac{2\pi}{3}n} \cos\left(\frac{2\pi}{6}n\right)$

(D)  $2 + 2e^{j\frac{2\pi}{6}n} \cos\left(\frac{2\pi}{6}n\right)$

(GATE EE 2022)

**Solution:**

Parameters	Description	Value
$x[n]$	Signal	
$N$	Period	3
$a_k$	Fourier series coefficient	
$a_{-3}$	$a_k$ at $k = -3$	2
$a_4$	$a_k$ at $k = 4$	1

Table 9.8: Parameters

$$x[n] = \sum_{k=-\infty}^{\infty} a_k e^{j\left(\frac{2k\pi}{3}n\right)} \quad (9.140)$$

$$= a_{-3} e^{j\frac{-6\pi}{3}n} + a_4 e^{j\frac{8\pi}{3}n} \quad (9.141)$$

$$= 2 + e^{j\frac{2\pi}{3}n} \quad (9.142)$$

$$= 1 + 1 + e^{j\frac{2\pi}{3}n} \quad (9.143)$$

$$= 1 + e^{j\frac{2\pi}{6}n} e^{-j\frac{2\pi}{6}n} + e^{j\frac{2\pi}{6}n} e^{j\frac{2\pi}{6}n} \quad (9.144)$$

$$= 1 + 2e^{j\frac{2\pi}{6}n} \left( \frac{e^{j\frac{2\pi}{6}n} + e^{-j\frac{2\pi}{6}n}}{2} \right) \quad (9.145)$$

$$= 1 + 2e^{j\frac{2\pi}{6}n} \cos\left(\frac{2\pi}{6}n\right) \quad (9.146)$$

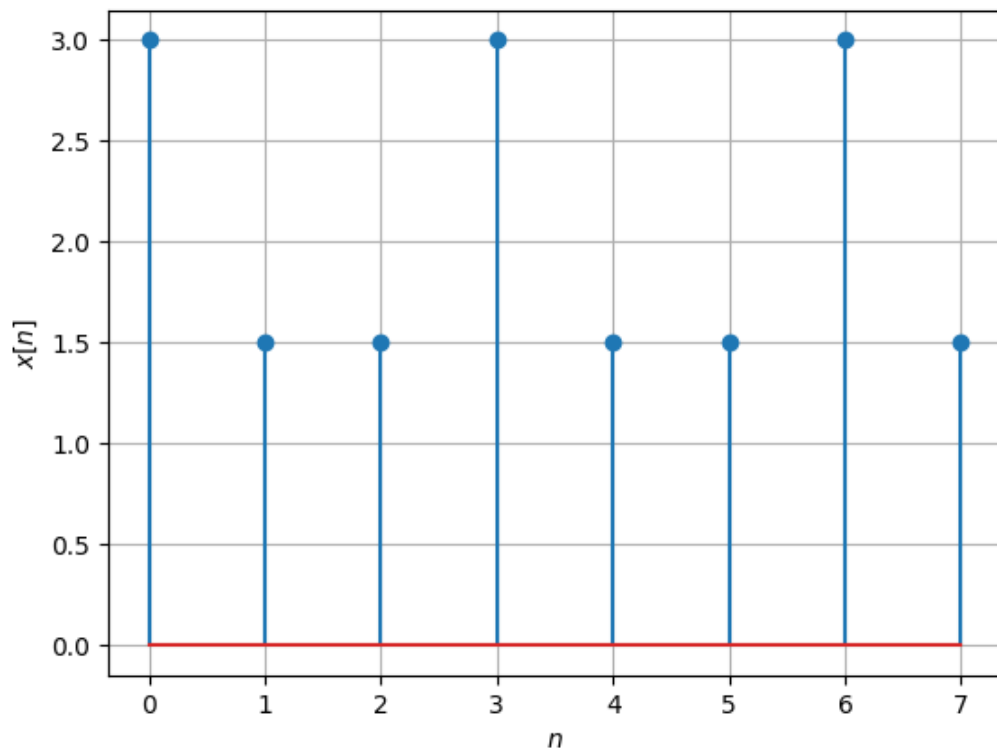


Figure 9.7: Stem Plot of  $x[n]$

## Appendix A

# Fourier transform

A.1 The Differentiation in frequency domain is as follows

Let  $x(t)$  be a signal such that,

$$x(t) \xleftrightarrow{\text{F.T.}} X(\omega) \quad (\text{A.1.1})$$

$$X(\omega) = \int_{t=-\infty}^{\infty} x(t) e^{-j\omega t} dt \quad (\text{A.1.2})$$

$$\frac{d}{d\omega} X(\omega) = \int_{t=-\infty}^{\infty} x(t) (-jt) e^{-j\omega t} dt \quad (\text{A.1.3})$$

$$j \frac{d}{d\omega} X(\omega) = \int_{t=-\infty}^{\infty} tx(t) e^{-j\omega t} dt \quad (\text{A.1.4})$$

$$tx(t) \xleftrightarrow{\text{F.T.}} j \frac{d}{d\omega} X(\omega) \quad (\text{A.1.5})$$

A.2 The duality property is as follows

From inverse Fourier transform we get,

$$x(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} X(\omega) e^{j\omega t} d\omega \quad (\text{A.2.1})$$

Replacing  $t$  by  $-t$  and multiplying  $2\pi$  on both sides we get,

$$2\pi x(-t) = \int_{-\infty}^{\infty} X(\omega) e^{-j\omega t} d\omega \quad (\text{A.2.2})$$

$$X(t) \xleftrightarrow{\text{F.T.}} 2\pi x(-\omega) \quad (\text{A.2.3})$$



## Appendix B

# Contour Integration

B.1 Cauchy's Theorem:

From Figure B.1.1

$$\int_C f(z) dz = \int_{C_r} f(z) dz \quad (\text{B.1.1})$$

Since  $g(z)$  is continuous we know that  $|g(z)|$  is bounded inside  $C_r$ . Say,  $|g(z)| < M$ .

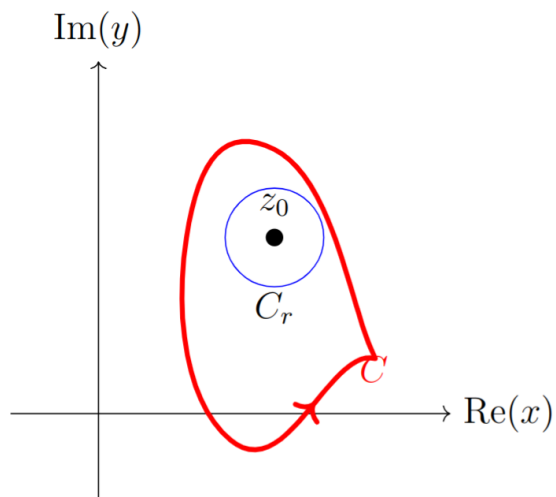


Figure B.1.1: Figure1

The corollary to the triangle inequality says that

$$\left| \int_{C_r} f(z) dz \right| \leq M2\pi r. \quad (\text{B.1.2})$$

Since  $r$  can be as small as we want, this implies that

$$\int_{C_r} f(z) dz = 0 \quad (\text{B.1.3})$$

let

$$g(z) = \frac{f(z) - f(z_0)}{z - z_0} \quad (\text{B.1.4})$$

$$\lim_{z \rightarrow z_0} g(z) = f'(z_0) \quad (\text{B.1.5})$$

$$\int_C g(z) dz = 0, \implies \int_C \frac{f(z) - f(z_0)}{z - z_0} dz = 0 \quad (\text{B.1.6})$$

Thus,

$$\int_C \frac{f(z)}{z - z_0} dz = \int_C \frac{f(z_0)}{z - z_0} dz = 2\pi i f(z_0) \quad (\text{B.1.7})$$

Using Cauchy's Theorem,

$$\int_C f(z) dz = 0 \quad (\text{B.1.8})$$

$$f(a) = \frac{1}{2\pi i} \oint_C \frac{f(z)}{z - a} dz \quad (\text{B.1.9})$$

$$f^n(a) = \frac{n!}{2\pi i} \oint_C \frac{f(z)}{(z - a)^{n+1}} dz \quad (\text{B.1.10})$$

## B.2 Residue Theorem:

From eq (B.1.7)

$$\int_C f(z) dz = 2\pi i \sum \text{Res } f(a) \quad (\text{B.2.1})$$

where, for n repeated poles,

$$\text{Res } f(a) = \lim_{z \rightarrow a} \frac{1}{(n-1)!} \left( \frac{d^{n-1}}{dz^{n-1}} [(z-a)^n f(z)] \right) \quad (\text{B.2.2})$$

