DETAILED SOLUTIONS

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ESE 2026 : Prelims Exam
CLASSROOM TEST SERIES

E & T ENGINEERING

Test 10

Section A : Signals & Systems + Basic Electrical Engineering

Section B : Analog & Digital Communication Systems-1

Section C : Electronic Devices & Circuits-2 + Analog Circuits Topics-2

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Detailed Explanation
Section A : Signals & Systems + Basic Electrical Engineering

1. (d)

Given: $$x[n] = \{-2, \underset{\uparrow}{1+j}, 2, -3+2j\}$$

The conjugate antisymmetric part of the given sequence $x[n]$ is,

$$\frac{x[n]-x^*[-n]}{2}$$

At the origin, we have $x[n]\big|_{n=0} = 1+j$

∴ Conjugate antisymmetric part,

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

2. (b)

The DC value of the signal is given by the complex Fourier series coefficient C_0:

We know that, $$C_0 = \frac{1}{T}\int_0^T x(t)dt$$

Here, Time period, $T = 2$

$$C_0 = \frac{1}{2}\int_0^2 5t\,dt = \frac{5}{2}\left[\frac{t^2}{2}\right]_0^2 = \frac{5}{4}\times 4 = 5$$

∴ The DC value $C_0 = 5$

3. (c)

$$x(t) = \underbrace{4\cos\left(\frac{2\pi}{3}t+40^\circ\right)}_{x_1(t)} + \underbrace{3\sin\left(\frac{4\pi}{5}t+20^\circ\right)}_{x_2(t)}$$

For $x_1(t)$: $$\omega_1 = \frac{2\pi}{3} \Rightarrow T_1 = \frac{2\pi}{\omega_1} = \frac{2\pi}{\frac{2\pi}{3}} = 3$$

For $x_2(t)$: $$\omega_2 = \frac{4\pi}{5} \Rightarrow T_2 = \frac{2\pi}{\frac{4\pi}{5}} = \frac{5}{2}$$

Thus, the fundamental period is:

$$T = \text{LCM}(T_1, T_2)$$

$$T = \frac{\text{LCM of Numerators}}{\text{HCF of Denominators}} = \frac{3\times 5}{1} = 15\,\text{sec}$$

4. (c)

Using Parseval's theorem, the energy of the signal $x(n)$ is,

$$E = \sum_{n=-\infty}^{\infty} |x(n)|^2 = \frac{1}{N}\sum_{k=0}^{N-1} |X(k)|^2$$

Here, $N = 4$, and $X(0) = 3$; $X(1) = 2 + j$; $X(2) = j$

Using the conjugate symmetry property of the DFT:

$$X(k) = X^*(N - k)$$
$$X(3) = X^*(4 - 3)$$
$$= X^*(1)$$
$$X(3) = 2 - j$$

Thus:

$$E = \frac{1}{4}\left[|X(0)|^2 + |X(1)|^2 + |X(2)|^2 + |X(3)|^2\right]$$
$$= \frac{1}{4}\left[3^2 + (4+1) + 1 + (4+1)\right]$$
$$= \frac{1}{4}[9 + 5 + 1 + 5]$$
$$E = 5 \text{ Joules}$$

5. (a)

Let $y(n) = x^2(n) = x(n) \cdot x(n)$

Using the circular convolution property of the DFT:

$$Y(k) = \frac{1}{N}[X(k) * X(k)]$$

$$= \frac{1}{4}\begin{bmatrix} 4 & j2 & 0 & -j2 \\ -j2 & 4 & j2 & 0 \\ 0 & -j2 & 4 & j2 \\ j2 & 0 & -j2 & 4 \end{bmatrix}\begin{bmatrix} 4 \\ -2j \\ 0 \\ 2j \end{bmatrix} = \frac{1}{4}\begin{bmatrix} 24 \\ -16j \\ -8 \\ 16j \end{bmatrix}$$

$$Y(k) = \{6, -4j, -2, 4j\}$$

6. (c)

Computation of DFT requires N^2 complex multiplications and $N(N - 1)$ complex additions.

Using FFT

$$\text{Complex additions} = N \log_2 N$$

$$\text{Complex multiplications} = \frac{N}{2}\log_2 N$$

7. (d)

Given,

$$x(n) = \{-2, -3, \underset{\uparrow}{-5}, 3, 2\}$$

Using Parseval's theorem:

$$\sum_{n=-\infty}^{\infty} |x(n)|^2 = \frac{1}{2\pi}\int_{-\pi}^{\pi} |X(e^{j\omega})|^2 \, d\omega$$

$$\int_{-\pi}^{\pi} \left|X(e^{j\omega})\right|^2 d\omega = 2\pi \sum_{n=-\infty}^{\infty} |x(n)|^2$$

$$\sum_{n=-\infty}^{\infty} |x(n)|^2 = \sum_{n=-2}^{2} |x(n)|^2$$

$$= (-2)^2 + (-3)^2 + (-5)^2 + (3)^2 + (2)^2 = 51$$

$$2\int_{-\pi}^{\pi} \left|X(e^{j\omega})\right|^2 d\omega = 4\pi \times 51 = 204\pi$$

Thus, $$2\int_{-\pi}^{\pi} \left|X(e^{j\omega})\right|^2 d\omega = 204\pi$$

8. (a)

Given: $$x(t) = 5\sin(\alpha\pi t)$$

$$\text{Time period, } T = \frac{2\pi}{\omega_0} = \frac{2\pi}{\alpha\pi} = \frac{2}{\alpha}\text{sec}$$

$$5 \text{ time period} = 5 \times T = 5\times\frac{2}{\alpha} = \frac{10}{\alpha}\text{sec}$$

Since the signal is sampled at a rate of 8 samples in $\frac{10}{\alpha}$ sec.

One sample occurs every: $T_s = \frac{10}{\alpha\times 8}$

$$\text{Sampling frequency, } f_s = \frac{1}{T_s} = \frac{8\alpha}{10}$$

But $$f_s = 1.2 \times 10^3$$

$$\frac{8\alpha}{10} = 1200 \text{ Hz}$$

$$\alpha = 1500 \text{ Hz}$$

9. (b)

Using partial-fraction expansion,

$$\frac{X(z)}{z} = \frac{a}{z-1} + \frac{b}{z-2} + \frac{c}{(z-2)^2}$$

where, $$a = \left.\frac{1}{(z-2)^2}\right|_{z=1} = 1$$

$$c = \left.\frac{1}{(z-1)}\right|_{z=2} = 1$$

$$b = \left.\frac{d}{dz}\left(\frac{1}{z-1}\right)\right|_{z=2} = \left.\frac{-1}{(z-1)^2}\right|_{z=2} = -1$$

Thus, $$X(z) = \frac{z}{z-1} - \frac{z}{z-2} + \frac{z}{(z-2)^2}; |z| > 2$$

Since the ROC is $|z| > 2$, the sequence is right-sided.

Hence, $$x(n) = (1 - 2^n + n2^{n-1})u(n)$$

10. (b)

$$e^{-at^2} \xleftrightarrow{F.T} \sqrt{\frac{\pi}{a}} e^{-\frac{\omega^2}{4a}}$$

Comparing with the given expression:

$$A = \sqrt{\frac{\pi}{a}}$$

Here, $$a = 4$$

Thus $$A = \sqrt{\frac{\pi}{4}} = 0.886$$

$\Rightarrow$ $$A^2 = 0.785$$

11. (b)

$$y(t) = x\left(\sqrt{e^t}\right)$$

The given relation satisfies the principle of superposition, hence it is a linear system.

For causality: Output must not depends on the future value of the input:

At $t = 2$: $$y(2) = x\left(\sqrt{e^2}\right) = x(2.718)$$

This output depends on a future input, hence the system is non-causal.

For time variant:

Given: $$y(t) = x\left(\sqrt{e^t}\right)$$

Now let, $$y_1(t) = x_1\left(\sqrt{e^t}\right)$$

where, $$x_1\left(\sqrt{e^t}\right) = x\left(\sqrt{e^t} - t_0\right)$$

$$y_1(t) = x(\sqrt{e^t} - t_0) \quad \text{...(i)}$$

$$y(t - t_0) = x\left(\sqrt{e^{t-t_0}}\right) \quad \text{...(ii)}$$

On comparing equations (i) and (ii):

$$y_1(t) \neq y(t - t_0)$$

Hence, the system is time-variant.

12. (b)

$$y(t) = x(t) * h(t)$$

$$z(t) = \frac{d}{dt}y(t)$$

Using convolution property

$$\frac{dy(t)}{dt} = \frac{d}{dt}[x(t)*h(t)] = \left[\frac{d}{dt}x(t)*h(t)\right]$$

$$= 2\delta(t + 2) * r(t - 1)$$

$$= 2r(t + 2 - 1)$$

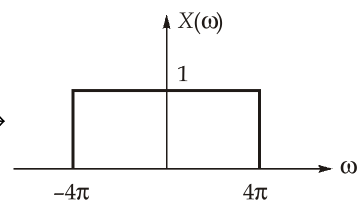
$$= 2r(t + 1)$$

13. (a)

We have, $$x(t) = \frac{\sin 4\pi t}{\pi t}$$

$$x(t) \xleftrightarrow{F.T} X(\omega)$$

$$\frac{\sin 4\pi t}{\pi t} \xleftrightarrow{F.T}$$



Energy of $x(t)$ $\Rightarrow$ $$E = \frac{1}{2\pi}\int_{-4\pi}^{4\pi} |X(\omega)|^2 \cdot d\omega$$

$$= \frac{1}{2\pi} \times 8\pi = 4 \text{ Joule}$$

We know; $$f(t) \longrightarrow E$$

$$\alpha f(\beta t + \gamma) \longrightarrow \alpha^2 \cdot \frac{E}{|\beta|}$$

Hence, Energy of $y(t)$ will be,

$$E_y = \frac{E_x}{|2|} = \frac{4}{2} = 2 \text{ Joule}$$

14. (a)

We have, $$x(n) = \sum_{k=0}^{2} (0.5)^k y(n-k)$$

$$x(n) = (0.5)^0 y(n) + 0.5y(n - 1) + (0.5)^2 y(n - 2)$$

$$x(n) = y(n) + 0.5y(n - 1) + 0.25y(n - 2)$$

Taking the z-transform

$$X(z) = Y(z) + 0.5z^{-1}\,Y(z) + 0.25\,z^{-2}Y(z)$$

$$\frac{Y(z)}{X(z)} = H(z) = \frac{1}{1 + 0.5z^{-1} + 0.25z^{-2}}$$

$$H(z) = \frac{z^2}{z^2 + 0.5z + 0.25}$$

Zeros $\Rightarrow$ 2 zeros at origin

$$\text{Poles} \Rightarrow z_1 = \frac{-1}{4} + \frac{\sqrt{3}}{4}j \Rightarrow |z_1| = 0.5$$

$$z_2 = \frac{-1}{4} - \frac{\sqrt{3}}{4}j \Rightarrow |z_2| = 0.5$$

All poles and zeros lie inside the unit circle, therefore the given system is minimum-phase system.

15. (b)

- The elliptic filter, also known as cauer filter has equirriple passband and stopband.
- For a given filter order, passband and stopband deviations, Cauer filters have less transition bandwidth as compared to Chebyshev filters. Hence, statement 2 is incorrect.
- In order to achieve linear phase, an IIR filter would require conjugate-reciprocal poles outside the unit circle, making it BIBO unstable. Hence, linear phase characteristics cannot be obtained in IIR filters.

16. (b)

An IIR (Infinite Impulse Response) is a digital filter whose impulse response extends indefinitely. The IIR filter design involves creating a prototype analog filter (such as Butterworth, Chebyshev, or Elliptic) in the analog domain. This analog filter is then transformed into a digital filter using techniques such as the Bilinear transformation and the Impulse invariance method.

17. (d)

- For a baseband signal with the highest frequency component f_m, the sampling frequency, $f_s \geq 2f_m$ to avoid aliasing effect and to recover the signal.
- A bandpass signal $m(t)$ can be recovered from its sampled signal if

$$f_s \geq \frac{2f_h}{k}; \text{ where } k = \left\lfloor \frac{f_h}{f_h - f_l} \right\rfloor$$

Here, f_h and f_l are the highest and lowest frequency components of the bandpass signal respectively.

Hence, all the given statements are correct.

18. (c)

Given, $g(t) = (2 - e^{-2t})u(t)$

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

impulse response of the system,

$$H(s) = \frac{G(s)}{U(s)} = \frac{\frac{s+4}{s(s+2)}}{\frac{1}{s}} = \frac{s+4}{s+2}$$

For input, $x(t)$ = 2 (which is D.C value) response is $H(s)\big|_{\text{at } s=0}$ $(\because \omega = 0)$

$$\therefore \quad H(0) = 2\times\frac{4}{2} = 4$$

19. (b)

To increase the ampere-hour rating of a battery, cells are connected in parallel.

20. (a)

In a lead-acid cell:

Number of negative plates = Number of positive plates + 1

21. (a)

Hydroelectric power stations are generally located in hilly areas, where dams can be conveniently built and large water reservoirs can be obtained.

22. (a)

Magnetic flux in a magnetic circuit is similar to electric current in an electric circuit. Similarly, MMF is equivalent to EMF, reluctance is equivalent to resistance, and permeability corresponds to conductivity in electric circuits.

23. (d)

Iron loss does not depend on the load and remains constant. Thus, iron loss at half load = 800 W.

Full-load copper loss = P_{cfl} = 7200 W

$$\text{Copper loss at half load} = \left(\frac{1}{2}\right)^2 P_{cfl} = \frac{7200}{4} = 1800 \text{ W}$$

Hence, option (d) is correct.

24. (b)

Voltage regulation, $VR = I(R\cos\phi - X\sin\phi)$; (at leading power factor) ...(i)

Voltage regulation = $I(R\cos\phi + X\sin\phi)$; (at lagging power factor) ...(ii)

from equation (i), zero voltage regulation is possible only at leading power factor.

25. (a)

$$\text{Full-load current, } I_{fl} = \frac{10\times10^3}{2000} = 5 \text{ A}$$

$$\text{Full-load copper loss, } P_{cfl} = 35\times\frac{(5)^2}{\left(\frac{5}{2}\right)^2}$$

$$= 35\times\frac{5^2}{5^2}\times2^2 = 35\times4$$

$$= 140 \text{ Watt}$$

$$\text{Core loss} = 40 \text{ W}$$

$$\text{Efficiency: } \eta = \frac{\text{KVA rating}\times pf}{\text{KVA rating}\times pf + \text{core loss} + \text{iron loss}}$$

$$\eta = \frac{(10 \times 10^3 \times 0.8) \times \frac{1}{2}}{\left(10 \times 10^3 \times 0.8 \times \frac{1}{2}\right) + 40 + 140 \times \left(\frac{1}{2}\right)^2}$$

$$\eta = \frac{4000}{4000 + 40 + \frac{140}{4}} \times 100 = 98.2\%$$

26. (d)

The open-circuit test is used to obtain

- Core losses (iron losses).
- Magnitude of exciting current

27. (a)

- DC series motor $\rightarrow$ high starting torque.
- DC shunt motor $\rightarrow$ constant speed.
- 3-phase induction motor $\rightarrow$ low starting torque.
- Synchronous motor $\rightarrow$ poor stability.

28. (d)

We know that,

$$P = EI = T\omega$$

$$T = \frac{EI}{2\pi\left(\frac{N}{60}\right)}$$

$$T = \frac{250 \times 12}{2\pi \times \frac{1200}{60}} = \frac{250 \times 6}{\pi \times 20} = \frac{75}{\pi}\text{N-m}$$

29. (b)

30. (b)

Distribution factor for the armature winding is given as, $D = \dfrac{\sin\left(\frac{m\gamma}{2}\right)}{m \sin\left(\frac{\gamma}{2}\right)}$.

31. (d)

We have, Supply frequency, $f = \dfrac{1000 \times 6}{120} = 50$ Hz

As, Slip frequency, s.f = 2 Hz

$$\text{slip, } s = \frac{s.f}{f} = \frac{2}{50} = 0.04$$

$\therefore$ Synchronous speed: $N_s = \dfrac{120 \times 50}{8} = 750$ rpm

$\therefore$ Rotor speed: $N_r = N_s(1 - s)$

$= 750(1 - 0.04)$

$= 720$ rpm

32. (c)

We know that,

Breadth factor for the r^{th} harmonic:

$$k_{dr} = \frac{\sin \frac{mr\alpha}{2}}{m \sin \frac{r\alpha}{2}}$$

where, α = slot angle

$$= \frac{180°}{\text{slots/pole}} = \frac{180°}{(48/4)} = 15°$$

m = slots/pole/phase

$= 48/4/3 = 4$

$\therefore$ Breadth factor for the 3rd harmonic is

$$k_{d3} = \frac{\sin\left(\frac{3 \times 4 \times 15°}{2}\right)}{4 \sin\left(\frac{15° \times 3}{2}\right)} = \frac{\sin(90°)}{4 \sin\left(\frac{45°}{2}\right)}$$

$$k_{d3} = \frac{\sin 90°}{4 \sin(22.5°)}$$

$$k_{d3} = \frac{0.25}{\sin(22.5°)}$$

33. (b)

We know that,

Synchronous speed: $N_s = \dfrac{120 \times f}{P} = \dfrac{120 \times 50}{6} = 1000$ rpm

slip, $s = \dfrac{N_s - N_r}{N_s}$ where N_r = rotational speed

$$0.04 = \frac{1000 - N_r}{1000}$$

$N_r = 1000 - 40 = 960$ rpm

34. (a)

Due to high starting torque, DC series motors are used in electric locomotives, cranes, etc. Hence, both statements are true, and statement (II) correctly explains statement (I).

35. (a)

Both Statement (I) and Statement (II) are correct and Statement (II) is correct explaination of Statement (I).

36. (c)

Ideal sampling function: $$f(t) = \sum_{n=0}^{\infty} k_n \delta[t - nT]$$

where T = sampling period

Since, $$\delta[n + n_0] \xleftrightarrow{Z.T.} z^{n_0}$$

Therefore: $$z[f(t)] = k_0 + \frac{k_1}{z} + \frac{k_2}{z^2} + ... + \frac{k_n}{z^n}$$

37. (a)

Statement (I) is true because IIR filters use feedback (they are recursive), meaning the output depends on both current/past inputs and past outputs. This feedback mechanism allows the impulse response to extend to infinity, similar to the behaviour of analog filters.

Because of the poles in their transfer function, IIR filters are much more computationally efficient and can achieve a desired magnitude response (like a sharp cutoff or narrow transition band) with a significantly lower filter order (N) than a corresponding FIR filter.

Statement (II) is true because the system function of an IIR filter is defined as a rational function:

$$H(z) = \frac{Y(z)}{X(z)} = \frac{\sum_{k=0}^{M} b_k z^{-k}}{1 + \sum_{k=1}^{N} a_k z^{-k}}$$

The presence of the denominator polynomial (terms with a_k) results in poles (zeros of the denominator) and creates the recursive structure (feedback path) when implemented in hardware or software.

Hence, the ability of an IIR filter to achieve a sharp frequency response with a lower order is directly due to its recursive nature.

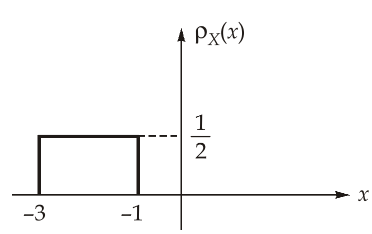
Section B : Analog and Digital Communication Systems-1

38. (b)

$F_X(x)$ is a non-decreasing function i.e.

$$F_X(x_1) \le F_X(x_2) \quad \text{for } x_1 \le x_2$$

39. (d)



For a uniform pdf, $E[X] = \frac{a+b}{2} = \frac{-3-1}{2} = -2$ $\left(\because \begin{matrix} a=-3 \\ b=-1 \end{matrix}\right)$

Now, the expectation of $Y(t) = E[Y(t)] = E[Xt + X + 5]$

$= E[Xt] + E[X] + E[5]$

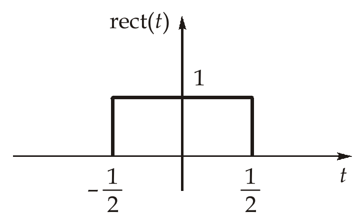
$= tE[X] + E[X] + E[5]$

$= -2t - 2 + 5$ ($\because$ E[Constant] = constant)

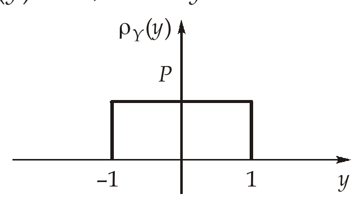
$E[Y(t)] = -2t + 3$

40. (a)

rect(t) =



$\therefore$ $\rho_Y(y) = P\,; \quad -1 \le y \le 1$



We know, $\int_{-\infty}^{\infty} \rho_Y(y)\, dy = 1$

$P \times 2 = 1$

$P = \frac{1}{2}$

Also given, $X = 2Y \Rightarrow \frac{dX}{dY} = 2$

$\therefore$ $\rho_X(x) = \left|\frac{dY}{dX}\right| \rho_Y(x)$

$\rho_X(x) = \frac{1}{2}\rho_Y(x)$ ...(i)

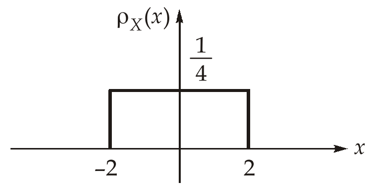
We have, $\rho_Y(x) = \frac{1}{2};\ -1 \le \frac{x}{2} \le 1$ $\left(\because Y = \frac{X}{2}\right)$

$\therefore$ $\rho_Y(x) = \frac{1}{2};\ -2 \le x \le 2$

Now from equation (i),

$\rho_X(x) = \frac{1}{2} \times \frac{1}{2};\ -2 \le x \le 2$

$\rho_X(x) = \frac{1}{4};\ -2 \le x \le 2$



$$\therefore \quad \rho_X(x) = \frac{1}{4}\text{rect}\left(\frac{x}{4}\right)$$

$$\therefore \quad Q = \frac{1}{4} = 0.25 \qquad \text{(upon comparing)}$$

41. (b)

In the case of direct (incoherent) and coherent detection, the SNR remains the same. However, in the case of non-coherent detection, the SNR decreases compared to incoherent and coherent detection.

42. (d)

We have,

Intermediate frequency, f_{If} = 455 kHz

Carrier frequency, f_c = 1155 kHz

We know that, Image frequency = $f_c + 2f_{If}$

= 1155 + 2(455)

= 2065 kHz

43. (b)

A DSB-SC signal will be

$$s(t) = m(t)\cos(\omega_c t + \theta)$$

where, $m(t)$ = envelope

So, at the output of the envelope detector = $|m(t)|$

44. (a)

In the case of an AM envelope demodulator

$$\frac{1}{f_c} < RC < W$$

$\frac{1}{f_c}$ → Carrier frequency; RC → Discharging time; W → Message Bandwidth

The forward resistance of the diode must be as low as possible so that the charging should be almost instantaneous. Normally, it is half of the carrier time period, i.e., half a cycle of the carrier.

45. (a)

In FM, high-frequency components are more immune to noise compared to lower frequencies. So at the transmitter end, we amplify the high-frequency components (i.e., pre-emphasis) to increase the S/N ratio, and at the receiver side, we reverse this using de-emphasis.

The envelope detector is used in AM demodulation. The Armstrong method is used in FM generation.

46. (d)

- Pre-emphasis and de-emphasis are used to improve SNR at receiver. But with a PLL demodulator, noise is reduced significantly at the receiver end, so to increase SNR at the receiver output, there is no requirement for pre-emphasis and de-emphasis. Also, PLL demodulators do not exhibit a threshold in S/N performance.
- PLL IC's are available with low cost.

47. (b)

- In a superheterodyne receiver, the problem of image frequency occurs.
- In FM (Frequency Modulation), the threshold effect may occur.
- In PCM (Pulse code modulation), noise occurs due to quantization error.
- In (DM) Delta modulation, granular noise occurs when the signal becomes constant.

48. (a)

We know that,

Sampling frequency, $f_s \geq 2f_m$

where, f_m = message frequency

Then, $f_s \geq 2 \times 3.6 \times 10^3$

$$f_s \geq 7.2 \text{ kHz} \quad \ldots(i)$$

And also,

$$nf_s \leq (\text{bit rate})$$

$$nf_s = \text{channel capacity}$$

$\therefore$ Channel capacity, $C = 48000$ bits/sec

$$nf_s = 48000$$

$$n \leq \frac{48000}{7200} \leq 6.67$$

$$n \leq 6.67 \quad \text{or} \quad n = 6$$

$\therefore$ Quantization level, $L = 2^6 = 64$

and sampling rate,

$$f_s = \frac{48000}{n} = 8 \text{ kHz}$$

49. (d)

The auto-correlation function is independent of phase for a sinusoidal wave.

$$R_X(\tau) = \frac{A^2}{2}\cos\omega_c\tau$$

For the random process $A\cos(\omega_c t + \theta)$

$$\therefore \quad |R_X(\tau)|_{\max} = R_X(0) = \frac{A^2}{2}$$

50. (b)

$$s(t) = e^{-at}\cos[(\omega_c + \Delta\omega)t]u(t)$$

$$\text{Complex signal} = s(t) + j\hat{s}(t)$$

$$= e^{-at}\cos[(\omega_c+\Delta\omega)t]u(t) + je^{-at}\cos[(\omega_c+\Delta\omega)t - 90°]u(t)$$

$$= e^{-at}\cos[(\omega_c+\Delta\omega)t] + j\sin[(\omega_c+\Delta\omega)t]u(t)$$

$$= \left[e^{-at}\cdot e^{j\Delta\omega t}u(t)\right]e^{j\omega_c t}$$

$$\text{Complex envelope} = e^{-at}\cdot e^{[j\Delta\omega t]}u(t)$$

51. (d)

The capture range is always less than or equal to the locking range.

52. (b)

We know that,

The standard FM signal is given by,

$$s_{FM}(t) = A_c \cos\{\omega_c t + \beta \sin\omega_m t\}$$

On comparing, we get

$$\beta = B; \quad \omega_m = 2\pi \times 100$$

$$f_m = 100 \text{ Hz}$$

Using Carson's rule, we get

$$BW = 2(\beta + 1)f_m$$

$$3.2 \times 10^3 = 2(\beta + 1)100$$

$$\beta = (1.6 \times 10) - 1$$

$$\beta = 15$$

And, we know that,

$$\beta = \frac{\Delta f}{f_m}$$

$$15 = \frac{\Delta f}{100}$$

$$\Delta f = 1.5 \text{ kHz}$$

53. (a)

We know that,

For $P_x(x)$ to be a probability density function,

$$\int_{-\infty}^{\infty} P_x(x)dx = 1$$

$$\int_{-\infty}^{\infty} ae^{-b|x|}dx = 1$$

$$2\int_{0}^{\infty} ae^{-bx}dx = 1$$

$$2a\left[\frac{e^{-bx}}{-b}\Bigg|_0^{\infty}\right] = 1$$

$$2a\left[\frac{e^{-\infty}}{-b}+\frac{1}{b}\right] = 1$$

$$\frac{2a}{b} = 1$$

$$b = 2a \quad \text{...(i)}$$

We have,

$$a + b = 3$$

$$a + 2a = 3$$

$$a = 1$$

$$2a = b \quad \Rightarrow \quad b = 2$$

54. (c)

$$\text{Mean, } E[X] = \int_{-\infty}^{\infty} x \cdot f_X(x)dx$$

$$= \int_a^b \left[\frac{1}{b-a}\right] x\,dx = 1 = \left[\frac{1}{b-a}\right]\left[\frac{x^2}{2}\bigg|_a^b\right] = 1$$

$$(b - a) = \left[\frac{b^2}{2}-\frac{a^2}{2}\right]$$

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

$$b + a = 2 \quad \text{...(i)}$$

Now,

$$\text{variance, } \sigma_x^2 = E[X^2]-(E[X])^2$$

$$1.33 = E[X^2] - (1)^2$$

$$E[X^2] = \frac{4}{3}+1$$

$$E[X^2] = \frac{7}{3} \quad \text{...(ii)}$$

Now,

$$E[X^2] = \int_{-\infty}^{\infty} x^2 f_X(x)dx = \left[\frac{1}{b-a}\right]\int_a^b x^2 dx = \frac{7}{3}$$

$$\left[\frac{1}{b-a}\right]\left[\frac{x^3}{3}\bigg|_a^b\right] = \frac{7}{3}$$

$$\frac{b^3}{3}-\frac{a^3}{3} = \frac{7(b-a)}{3}$$

$$b^3 - a^3 = 7(b - a)$$

$$(b - a)(b^2 + a^2 + ab) = 7(b - a)$$

$$b^2 + a^2 + ab = 7$$

$$(b + a)^2 - ab = 7$$

$$(2)^2 - ab = 7$$

$$ab = 4 - 7 = -3$$

$$a = \frac{-3}{b} \quad ...(iii)$$

From equation (i) and (iii), we get

$$b - \frac{3}{b} = 2$$

$$b^2 - 3 = 2b$$

$$b^2 - 2b - 3 = 0$$

$$b = 3; -1$$

As b is positive, $b = -1$ is discarded.

Thus, $b = 3$

or $$a = -1$$

55. (a)

For jointly Gaussian variables, uncorrelated (covariance zero) $\Rightarrow$ independent because the joint PDF factors appropriately. Statement (II) is correct and explains Statement (I).

56. (a)

Both the statements are correct and Statement (II) explains Statement (I).

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57. (b)

We know that

$$I_D \propto (1 + \lambda V_{DS})$$

$$\frac{I_{D1}}{I_{D2}} = \frac{1+\lambda V_{DS1}}{1+\lambda V_{DS2}}$$

$$I_{D2} = \left(\frac{1+\lambda V_{DS2}}{1+\lambda V_{DS1}}\right) I_{D1}$$

$$= \left(\frac{1+0.3\times 4}{1+0.3\times 0.5}\right) \times 1\text{ mA}$$

$$I_{D2} = 1.913\text{ mA}$$

58. (a)

Given: Emitter resistance and base resistance are

$$r_e = 500\ \Omega; \quad r_\pi = 100.5\ \text{k}\Omega$$

We know, $$r_\pi = (1 + \beta)\, r_e$$

$$1 + \beta = \frac{r_\pi}{r_e} = \frac{100.5\times 10^3}{500}$$

$$1 + \beta = 201$$

$$\beta = 200$$

Current gain of a common-collector amplifier

$$\gamma = \beta + 1$$

$$\gamma = 200 + 1$$

$$\gamma = 201$$

59. (c)

Given: Base transport factor (β) = 0.998

Emitter injection efficiency (γ) = 0.999

The current transfer ratio (α) = $\beta\gamma$

$$\alpha = (0.998) \times (0.999)$$
$$\alpha = 0.997$$

The amplification factor, $\beta = \dfrac{\alpha}{1-\alpha} = \dfrac{0.997}{1-0.997}$

$$\beta = 332.55$$
$$\beta \cong 333$$

60. (c)

For an ideal MOS capacitor, the flat band voltage V_{FB} = 0 V

and $$V_{FB} = \phi_{MS} - \frac{Q'_{ox}}{C_{ox}}$$ [for a non-ideal MOS capacitor]

61. (b)

Let the transistor be in saturation

so, $$V_{BE_{sat}} = 0.8 \text{ V}$$
$$V_{CE_{sat}} = 0.2 \text{ V}$$

$\because$ $$I_B = \frac{V_{BB} - V_{BE_{sat}}}{200k} = \frac{5-0.8}{200k} = 0.021 \text{ mA}$$
$$I_{C_{sat}} = \frac{V_{CC} - V_{CE_{sat}}}{R_C} = \frac{10-0.2}{R_C} = \frac{9.8}{R_C}$$

For the transistor to remain in saturation:

$$\beta I_B \geq I_{C_{sat}}$$
$$10^{-3} \times 100 \times 0.021 \geq \frac{9.8}{R_C}$$
$$R_C \geq \frac{9.8}{10^{-1} \times 0.021}$$

or, $$R_C \geq 4666.67$$

$\therefore$ $$R_{C_{\min}} = 4667 \ \Omega$$

62. (c)

$$\text{Power dissipation} = \frac{\text{Temperature difference}}{\text{Thermal resistance}}$$
$$= \frac{150°\text{C} - 25°\text{C}}{1°\text{C/W}} = \frac{125}{1} = 125 \text{ W}$$

63. (b)

Depletion MOSFET:

$$I_D = 4.5 \text{ mA at } V_{GS} = -2 \text{ V}$$
$$I_{DSS} = ?$$
$$V_P = -5 \text{ V}$$

Using the relation:
$$I_D = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right]^2$$
$$4.5 \text{ mA} = I_{DSS}\left[1 - \frac{-2}{-5}\right]^2$$
$$4.5 \text{ mA} = I_{DSS}\left[\frac{5-2}{5}\right]^2$$
$$I_{DSS} = \frac{4.5 \text{ mA} \times 25}{9} = 12.5 \text{ mA}$$

64. (a)

n-channel Si FET

$$a = 3 \times 10^{-4} \text{ cm}$$
$$N_D = 10^{15} \text{ electron/cm}^3$$
$$V_p = ?$$
$$\varepsilon = 12\varepsilon_0 \text{ where } \varepsilon_0 = \frac{1}{36\pi} \times 10^{-9} \text{ F/m} = 8.85 \times 10^{-14} \text{ F/cm}$$
$$V_p = \frac{a^2 q N_D}{2\epsilon} = \frac{(3\times10^{-4})^2 \times 1.6\times10^{-19} \times 10^{15}}{2\times12\times8.85\times10^{-14}}$$
$$= 6.8 \text{ V}$$

65. (c)

In a voltage series feedback configuration, the input impedance increases by a factor of $(1 + \beta A_V)$. Thus, the input impedance will be $R_i(1 + \beta A_V)$.

66. (c)

$$|\text{Voltage gain}| = \left|\text{Current gain} \times \frac{R_L}{R_i}\right|$$
$$= \left|-100 \times \frac{1}{10}\right| = 10$$
$$\text{Gain in dB} = 20 \log_{10} 10 = 20 \text{ dB}$$

67. (c)

We know, $$f = \frac{1}{2\pi RC}$$

For $C = \frac{1}{2\pi} \mu\text{F}$; $R = 1 \text{ k}\Omega$, $f = 1 \text{ kHz}$

For $C = \dfrac{1}{18\pi}\mu F$, $R = 3\,k\Omega$, $f = \dfrac{1}{2\pi \times 3 \times \dfrac{1}{18\pi}} = 3\,kHz$

Hence, both statements are correct.

68. (a)

$$R_1 = R_2 = R_3 = \bar{R}$$

$$C_1 = C_2 = C_3 = \bar{C}$$

RC phase shift oscillator

$$f = \frac{1}{2\pi RC\sqrt{6}} = \frac{1}{2 \times 3.14 \times 2.44 \times \bar{R}\bar{C}}$$

$$= \frac{1}{15.32\bar{R}\bar{C}} = \frac{0.065}{\bar{R}\bar{C}}$$

69. (b)

Using KCL at the node A:

$$\frac{10 - V_b}{1} = \frac{V_b - V_0}{1}$$

$\Rightarrow$ $$2V_b = V_0 + 10 \quad \text{...(i)}$$

Using KCL at node V_b

$$\frac{V_b}{0.2} + \frac{V_b}{1} + \frac{V_b - V_0}{1} = 0$$

$$5V_b + V_b + V_b - V_0 = 0$$

$$7V_b - V_0 = 0$$

$\Rightarrow$ $$V_0 = 7V_b \quad \text{...(ii)}$$

Substituting equation (ii) into equation (i):

$$2V_b = 7V_b + 10$$

$\Rightarrow$ $$5V_b = -10$$

$\Rightarrow$ $$V_b = -2\text{ V}$$

$$I_L = \frac{V_b}{200} = \frac{-2}{200}$$

$$I_L = -10\text{ mA}$$

70. (b)

$$V_0 = A_d V_d + A_{cm} V_{cm}$$

$$= A_d V_d \left(1 + \frac{A_{cm} V_{cm}}{A_d V_d}\right)$$

$$= A_d V_d \left(1 + \frac{1}{CMRR} \cdot \frac{V_{cm}}{V_d}\right)$$

$$\text{Error in differential output} = \frac{V_{cm}}{CMRR \times V_d} \times 100$$

where $$V_{cm} = \frac{V_1 + V_2}{2}$$

$$\Rightarrow \quad V_{cm} = \frac{1050 + 950}{2} \mu V$$

$$\Rightarrow \quad V_{cm} = 1000 \ \mu V$$

$$V_d = V_1 - V_2$$
$$= (1050 - 950) \ \mu V$$
$$= 100 \ \mu V$$

Error in differential output $= \frac{1000}{1000 \times 100} \times 100 = 1\%$

71. (b)

$$f = \frac{1}{0.69(R_A + 2R_B)C}$$

$$= \frac{1}{0.69(10 \times 10^3 + 2 \times 50 \times 10^3) \times 0.01 \times 10^{-6}} = 1.3 \text{ kHz}$$

$$\% \text{ Duty cycle} = \frac{R_A + R_B}{R_A + 2R_B} \times 100 = \frac{(10k + 50k)}{(10k + 2 \times 50k)} \times 100$$

$$= \frac{60k}{100k} \times 100 = 54.5\%$$

72. (d)

Given $$\frac{R_F}{R} = 100$$

$$R_F = 47 \text{ k}\Omega$$

If input offset voltage is considered,

$$V_0 = \left(1 + \frac{R_F}{R}\right) V_{i0} = 101 \times 6 = 606 \text{ mV}$$

If input bias current is considered

$$V_0 = I_B \times R_F = 500 \times 10^{-9} \times 47 \times 10^3$$
$$= 23.5 \text{ mV}$$

73. (b)

Statement (I) correctly defines the astable multivibrator. The astable multivibrator has no stable states. It continuously switches between its two quasi-stable states, generating a periodic square wave without requiring an external trigger.

Statement (II) describes the internal operating mechanism. The astable multivibrator circuit is designed to oscillate between the positive saturation voltage ($+V_{sat}$) or the negative saturation voltage ($-V_{sat}$). Hence, the op-amp indeed operates in the saturation region.

Hence, both statements are correct and statement (II) is not the reason of statement (I).

74. (a)

From statement (I): If the gain margin (GM) is negative, the system is unstable and will oscillate. The magnitude of a negative GM indicates how much the gain must be reduced (attenuated) to bring the system back to the brink of stability (0 dB margin), not the permissible rise. The statement incorrectly claims that the system is stable and gives a permissible rise in gain when GM is negative.

Hence, statement (I) is false.

Statement (II) refers to the Barkhausen criterion for sustained oscillations in a feedback amplifier. For oscillation to occur, two conditions must be satisfied at the frequency of oscillation.

1. Loop gain magnitude condition: The magnitude of the loop gain ($|A\beta|$) must be equal to unity.
2. Phase condition: The total phase shift around the loop must be 0° or an integer multiple of 360°.

Hence, statement (II) is true.

75. (b)

- In hard saturation, the base current is typically 2 to 10 times larger than the minimum base current $I_{B\ min}$. For fast switching, the transistor should not be driven into deep saturation, because storage time increases and switching slows down.
- In saturation, both the EB and CB junctions are under forward bias condition.

Hence both the statements are true but statements (II) is not the correct explanation of statement (I).

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