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ESE 2025 : Prelims Exam
CLASSROOM TEST SERIES

E & T ENGINEERING

Test 8

Section A : Analog and Digital Communication Systems

Section B : Electronic Devices & Circuits-1 + Analog Circuits Topics-1

Section C : Control Systems-2 + Microprocessors and Microcontroller-2

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*Q.3 (Answer key has been Updated) (Mark to All)

DETAILED EXPLANATIONS

Section A : Analog and Digital Communication Systems

1. (c)

Carrier signal after phase modulation is,

$$S_{PM}(t) = A_c \cos[2\pi f_c t + \phi(t)]$$

Instantaneous angle,

$$\phi_i(t) = 2\pi f_c t + k_p m(t)$$

The instantaneous frequency is,

$$f_i(t) = \frac{1}{2\pi}\frac{d\phi_i(t)}{dt}$$

$$= \frac{1}{2\pi}\frac{d}{dt}\left[2\pi f_c t + k_p m(t)\right]$$

$$= \frac{1}{2\pi}\cdot 2\pi f_c + \frac{k_p}{2\pi}\frac{dm(t)}{dt}$$

Frequency deviation is given by

$$\Delta(f) = \frac{k_p}{2\pi}\frac{dm(t)}{dt}$$

$\because$ $$m(t) = A_m \sin(\omega_m t)$$

$$\Delta(f) = \frac{k_p}{2\pi} A_m \omega_m \cos \omega_m t$$

$$\Delta(f) \propto A_m f_m$$

2. (a)

The in-phase component of the noise is in phase and the quadrature noise component is in quadrature with the local oscillator signal. When multiplied by the local oscillator output in coherent detector, the quadrature component produces a signal which is attenuated by the low pass filter. Thus, the coherent detector rejects the quadrature component of noise and therefore, noise at the output has inphase component only. The inphase component of noise and output are additive at output of detector.

3. (*)

Given,

Total power, $P_t = 48$ W

We know that, $$P_t = P_c\left(1 + \frac{\mu^2}{2}\right)$$

$$48 = P_c\left[1 + \frac{(0.6)^2}{2}\right]$$

$$48 = P_c(1.18)$$

$$P_c = 40.68 \text{ W}$$

Total side band power = 48 – 40.68 = 7.32 W

Power of each sideband = $\frac{7.32}{2} = 3.66$ W

4. (c)

- QAM often has higher bandwidth efficiency than ASK since it combines both amplitude and phase modulation, allowing for more bits per symbol.
- In PSK, especially with higher modulation orders, bandwidth requirements can be influenced by the modulation index and the type of PSK (e.g. BPSK, QPSK)
- PSK is more bandwidth-efficient, requiring less bandwidth than FSK for the same data rate. Hence, PSK has a higher spectral efficiency than FSK. Thus, statement 3 is incorrect.
- MSK, a special case of FSK, has constant envelope characteristics and is more spectrally efficient than standard FSK due to its continuous phase properties.

5. (b)

$$y(t) = p + x(t)$$

The auto correlation function $R_Y(\tau)$,

$$R_Y(\tau) = E[Y(t) \cdot Y(t+\tau)]$$
$$R_Y(\tau) = E[[p + x(t)] \cdot [p + x(t+\tau)]]$$
$$R_Y(\tau) = E[p^2 + px(t) + px(t+\tau) + x(t) \cdot x(t+\tau)]$$
$$R_Y(\tau) = E[p^2] + pE[x(t)] + pE[x(t+\tau)] + E[x(t) \cdot x(t+\tau)]$$

Since, $x(t)$ is a zero mean random signal, hence

$$E[x(t)] = E[x(t+\tau)] = 0$$
$$R_Y(\tau) = p^2 + R_X(\tau)$$

6. (b)

We have,

Modulated FM signal as, $s(t) = 25\cos[(2\pi \times 10^6 t) + 20\sin(2\pi \times 1200t)]$

On comparing with standard expression of FM signal

$$s(t) = A_c \cos\{2\pi f_c t + \beta \sin(2\pi f_m t)\}$$

We get, $\beta = 20; \quad f_m = 1.2 \text{ kHz}$

and we know that, bandwidth of the FM signal is given as

$$\text{BW} = (\beta + 1)2f_m$$
$$= (20 + 1) \times 2 \times 1.2 \times 10^3 = 50.4 \text{ kHz}$$

7. (d)

- The auto correlation function $R_X[k]$ is defined as the expected value of the product $X[n]$ and $X[n+k]$, where k is the time lag between two instances of the process. This function measures how similar or correlated the values of the processes are at two different times separated by k. A higher value of $R_X[k]$ indicates a stronger correlation between $X[n]$ and $X[n+k]$, meaning that the process tends to have similar values at these time points.
- The covariance function $C_X[k]$ is related to the auto correlation function and the square of the mean μ_X^2 (where $\mu_X = E[X(n)]$). By removing μ_X^2, the covariance function focuses on the variability or spread of $X[n]$ around its mean essentially removing the contribution of the mean from the auto correlation function.
- For a wide-sense stationary (WSS) process, the autocorrelation and covariance functions depend only on the time difference (lag) k, and not on the specific time instances n and $n + k$ individually.

This is because a WSS process has time-invariant statistical properties, meaning $R_X[k]$ and $C_X[k]$ are functions of k alone rather than the absolute times.

- White noise is characterized by a lack of correlation between times. For a white noise process with zero mean, the covariance $C_X[k]$ (which is equivalent to $R_X[k]$ in this case, since $\mu_X = 0$) is zero for all non-zero lags $k \neq 0$. This means that there is no correlation between the values at different times, making the process uncorrelated at different time lags. The covariance function $C_X[k]$ is non zero only at $k = 0$, where it represents the variance of the process.

8. **(b)**

- Increasing the loop filter bandwidth can actually allow more high-frequency noise to pass through to the VCO, potentially worsening noise performance.
- The phase detector generates an output signal proportional to the phase difference, which drives the correction in the VCO.
- The loop filter is responsible for the transient dynamics of the PLL, affecting how quickly it locks and responds to changes.
- Increasing the loop filter bandwidth can make the system more responsive but may reduce stability due to higher sensitivity to noise and disturbances.

9. **(c)**

- MSK is indeed a type of continuous-phase frequency-shift keying (CPFSK). It maintains a smooth phase transition between symbols, which reduces abrupt phase changes.
- MSK has a constant envelope, making it resilient to amplitude distortions, which is beneficial in non-linear channels where amplitude changes could distort the signal.
- MSK is actually more bandwidth-efficient than BPSK. MSK achieves spectral efficiency similar to QPSK, as it requires less bandwidth than BPSK for the same bit rate.
- MSK or a variant of it (GMSK) is used in GSM systems because of its constant envelope and efficient bandwidth usage, which are well suited for mobile communication environments.

10. **(d)**

- A random process is called wide-sense stationary (WSS) if its mean and autocorrelation function do not change by shifts in time. The WSS processes are stationary in the strict sense if the probability density function (PDF) is independent of time. Since the PDF of the Gaussian process has only two parameters, mean and variance, which are independent of time shifts for WSS Gaussian process.

  Hence, if a Gaussian process is wide-sense stationary, then the process is also stationary in the strict sense.
- If a Gaussian process $X(t)$ is applied to a stable linear filter, then the random process $Y(t)$ developed at the output of the filter is also Gaussian.
- If the random variables $X(t_1)$, $X(t_2)$, ... $X(t_n)$ obtained by sampling a Gaussian process $X(t)$ at time t_1, t_2, ..., t_n, are uncorrelated, that is,

  $$E\left[\left(X(t_k - m_{X(t_k)}\right)\left(X(t_i - m_{X(t_i)}\right)\right] = 0; \; i \neq k$$

  then these random variables are also statistically independent.

11. (b)

- Huffman coding generates a prefix-free binary code (i.e. the bit string representing some particular symbol is never a prefix of the bit string representing any other symbol) that minimizes the average codeword length for a given set of symbol probabilities. "Optimal" here means that Huffman coding is the most efficient coding method (for a symbol-by-symbol scheme) in terms of minimizing the expected codeword length, given the symbol probabilities.
- Huffman coding minimizes the average code length, it does not necessarily achieve the minimum entropy of the source. Entropy is a theoretical lower bound on the average codeword, length, and Huffman coding approaches this bound but may not reach it precisely. For instance, if the probabilities do not allow an integer bit-length for codewords, Huffman coding can't match the exact entropy.
- Huffman coding is actually most efficient when source probabilities are skewed, as it takes advantage of highly probable symbols to generate shorter codewords. For sources with high entropy (where all symbols have roughly equal probabilities), Huffman coding offers less compression as codewords tend to be similar in length.
- Huffman coding produces variable-length codes based on symbol probabilities. Highly probable symbols get shorter codewords, while less probable symbols get longer ones, leading to unequal codeword lengths. Only in the rare case where all symbols have exactly the same probability, Huffman coding might produce equal-length codes.

Thus, statements 2 and 3 are not true.

12. (d)

The phase of the phase modulated signal is $\phi(t) = k_p m(t)$ and the instantaneous frequency is,

$$f_i(t) = f_c + \frac{1}{2\pi}\frac{d\phi(t)}{dt} = f_c + \frac{k_p}{2\pi}\frac{dm(t)}{dt}$$

The maximum of $f_i(t)$ is achieved for t in [0, 2] when $\frac{d}{dt}$ m(t) is maximum i.e. $\frac{d}{dt}m(t) = \frac{10^5}{2}$.

Hence, $$\max[f_i(t)] = 10^6 + \frac{3}{2\pi}\times\frac{10^5}{2} = 1.024 \text{ MHz}$$

13. (d)

$$y(t) = x(t)\ x_L(t)$$
$$= Am(t) \times 12\cos(2\pi f_c t)\cos(2\pi f_c t + 60°)$$
$$= \frac{12A}{2}m(t)\,(\cos 60° + \cos(4\pi f_c t + 60°))$$

After passing through the LPF, we get the output as,

$$z(t) = 6Am(t)\cos 60°$$

If power of $m(t)$ is P_m,

$$P_0 = (6A)^2\cos^2 60° \times P_m$$
$$= 36\,A^2\cos^2 60° \times P_m$$
$$= 36A^2 \times \frac{1}{4} \times P_m = 9A^2 P_m$$

Power of the modulated signal, $x(t) = Am(t)\cos(2\pi f_c t)$ is,

$$P_u = \frac{A^2}{2} P_m$$

$$\frac{P_u}{P_{out}} = \frac{A^2 P_m}{2 \times 9 A^2 P_m} = \frac{1}{18} = 0.055$$

14. (a)

For properly designed PLL based demodulator,

$$y(t) = \frac{k_f}{k_v} m(t) = \frac{10}{15} m(t) = \frac{2}{3} m(t)$$

$$P_y = \left(\frac{2}{3}\right)^2 P_m = \frac{4}{9} \times 25\text{ W} = \frac{100}{9}\text{ W} = 11.11\text{ W}$$

15. (a)

The noise power, P_{noise} in a bandwidth (B) at a temperature (T) is given by, $P_{noise} = kTB$, where k is the Boltzmann constant = 1.38×10^{-23} J/K. Thus,

$$\begin{aligned} P_{noise,\ AM} &= kTB_{AM} \\ &= 1.38 \times 10^{-23} \times 300 \times 10 \times 10^3 \\ &= 4.14 \times 10^{-17}\text{ W} \end{aligned}$$

$$\begin{aligned} P_{noise,\ FM} &= kTB_{FM} \\ &= 1.38 \times 10^{-23} \times 300 \times 200 \times 10^3 \\ &= 8.28 \times 10^{-16}\text{ W} \end{aligned}$$

The difference in noise power between FM and AM signals at the input of IF stage,

$$\begin{aligned} \Delta P_{noise} &= P_{noise,\ FM} - P_{noise,\ AM} \\ &= 8.28 \times 10^{-16} - 4.14 \times 10^{-17} \\ &= 7.86 \times 10^{-16}\text{ W} = 78.6 \times 10^{-17}\text{ W} \end{aligned}$$

16. (c)

Given, $R_b = 0.1$ Mb/s

and B.W = 75 kHz = 75×10^3 Hz

The bandwidth for a raised-cosine pulse is related to the data rate (R_b) and the roll-off factor (α) as

$$\text{B.W} = \frac{R_b}{2}(1+\alpha) = \frac{0.1 \times 10^6}{2}(1+\alpha)$$

$$75 \times 10^3 = \frac{0.1 \times 10^6}{2}(1+\alpha)$$

$$\frac{75 \times 2}{100} = 1 + \alpha$$

$$\alpha = \frac{3}{2} - 1 = \frac{1}{2} = 0.5$$

17. (d)

Given that

$$\Delta = 0.1 \text{ V}$$
$$f_m = 3 \text{ kHz}$$
$$\text{Sampling frequency, } f_s = 10 \times \text{N.R} = 10 \times 2 \times f_m$$
$$= 20 \times 3 \text{ kHz} = 60 \text{ kHz}$$

In delta modulation, to avoid slope overload distortion

$$\Delta f_s \geq 2\pi f'_m A_m \qquad [\, f'_m = 1 \text{ kHz for test sinusoidal signal}]$$

$$A_m \leq \frac{\Delta f_s}{2\pi f'_m}$$

$$A_m \leq \frac{0.1 \times 60 \times 10^3}{2\pi \times 10^3}$$

$$A_m \leq 0.1 \times 60 \times 0.159$$
$$A_m \leq 0.954$$

∴ Maximum amplitude of test sinusoidal signal to avoid slope overload distrotion,

$$A_{m(\max)} = 0.954$$

18. (d)

In a binary PCM system, $L = 2^n$, where n is the minimum number of binary digits.

Given, $$\left(\frac{S}{N_q}\right)_{o\,\text{dB}} = 40 \text{ dB} \Rightarrow \left(\frac{S}{N_q}\right)_o = 10^4$$

We know, the signal to noise ratio of PCM system is given by

$$\left(\frac{S}{N_q}\right)_o = \frac{3}{2} 2^n$$

$$2^n = \sqrt{\frac{2}{3}\left(\frac{S}{N_q}\right)_o} = \sqrt{\frac{2}{3} \times 10^4} = 81.6$$

The number of bits, $$n = \log_2 81.6$$
$$= 6.35 \cong 7$$

Alternate Solution:

We know, the signal to noise ratio of PCM system is given by

$$40 \text{ dB} \cong (6n + 1.76) \text{ dB}$$
$$n = 6.35 \cong 7$$

19. (c)

- In Delta modulation (DM), the step size Δ is fixed. For low-slope signals (signals that change gradually), the fixed step size can overshoot or undershoot the actual signal trajectory introducing granular noise. This is a type of quantization noise that occurs when the signal changes are too small compared to the step size.

- Adaptive Delta Modulation (ADM) adjusts the step size, Δ_{var} dynamically based on the slope of the input signal. This adaptive mechanism reduces granular noise for low-slope signals (by decreasing Δ) and slope overload distortion for high slope signals (by increasing Δ). By aligning the step size with the signal's slope, ADM minimizes quantization noise effectively.
- Delta modulation (DM) is prone to slope overload distortion for high-slope signals, as the fixed step size Δ may not be large enough to track rapid changes in the signal. In contrast, Adaptive Delta Modulation (ADM) increases the step size dynamically for high-slope signals, reducing slope overload distortion. Hence, ADM outperforms DM in high-slope signal scenarios.
- Rapidly varying signals require a larger step size to avoid slope overload distortion. ADM addresses this issue by dynamically increasing Δ for high-slope regions. This flexibility significantly reduces the error caused by a mismatch between the signal slope and the fixed step size, which is a limitation in DM systems.

Hence, statements 1, 2 and 4 are correct.

20. (a)

The Nyquist rate is twice the bandwidth of the signal. Since the signals are sampled at the Nyquist rate, hence the sampling rate of the multiplexed signal,

$$f_s = 2(1500 + 800 + 800)$$
$$= 6200 \text{ kHz}$$

Each sample is encoded using 13 bits, hence the bit rate of the multiplexed signal,

$$R_b = nf_s = 13 \times 6200 \text{ k} = 80.6 \text{ kbps}$$

Note:

- If we are using TDM + PCM and there are N-messages having same bandwidth and are sampled at same rate f_s then bit rate will be

$$R_b = N \cdot nf_s$$

- If Bandwidth of messages and sampling rates are different then overall bit rate will be

$$R_b = R_{b1} + R_{b2} + \text{.........}$$

or $R_b = n(f_{s1} + f_{s2} + \text{........})$ where, n = bits in quantizer.

21. (d)

The minimum bandwidth of BFSK signal is possible when Nyquist sampling pulses are used.

Here,

$$(\text{BW})_{\min} = f_H - f_L + R_b$$

Given

$$R_b = 2500 \text{ bps} = 2.5 \text{ kbps}$$
$$f_H - f_L = 4200 \text{ Hz} = 4.2 \text{ kHz}$$
$$(\text{BW})_{\min} = 4.2 + 2.5 = 6.7 \text{ kHz}$$

22. (c)

Average bit error probability of BPSK is given by,

$$= Q\left(\sqrt{\frac{A_c^2 T_b}{N_0}}\right)$$

Given,

$$A_c = 2 \text{ mV}, \quad \frac{N_0}{2} = 10^{-12} \text{ W/Hz}$$

$$R_b = 2.5 \text{ Mbps}$$

$$T_b = \frac{1}{R_b} = \frac{1}{2.5}\times 10^{-6}\text{ sec}$$

$$\therefore \quad P_e = Q\left(\sqrt{\frac{4\times 10^{-6}\times 10^{-6}}{2.5\times 2\times 10^{-12}}}\right) = Q(\sqrt{0.8})$$

23. (b)

- A random process is said to be stationary if its statistical properties, such as mean, variance, and higher-order moments, do not change with time. This means the probability distribution of $X(t)$ remains the same regardless of the time t. For a binary random process $X(t)$, this implies that $P[X(t)] = 1$ and $P[X(t)] = 0$ are constant over time.
- A process is ergodic if time averages over a single realization of the process equal the ensemble averages computed across multiple realizations. In a binary random process, ergodicity implies that the long-term fraction of time that $X(t)$ is in state 1 equals the probability $P(X = 1)$ and similarly for state 0.
- Cyclostationarity refers to processes whose statistical properties vary periodically with time. Binary random processes, such as those used in communications (e.g., periodic switching between 0 and 1), can exhibit cyclostationary. For example, a binary random process where the probability $P(X = 1)$ oscillates periodically with time is cyclostationary.
- For a random process to be wide-sense stationary (WSS), its mean must be constant, and its autocorrelation function $R_X(t_1, t_2)$ must depends only on the time difference $\tau = t_2 - t_1$. In a binary random process if $X(t)$ is WSS, then $R_X(\tau)$ will depend only on the time difference τ, and not on t_1 or t_2 individually.

24. (a)

- $S(t)$ is deterministic, but $N(t)$ is a random process. The randomness of $N(t)$ introduces randomness in $R(t)$. Thus, $R(t)$ is random because it depends on $N(t)$.
- The mean of $R(t)$ is given by

  $$E[R(t)] = E[S(t) + N(t)] = E[S(t)] + E[N(t)]$$

  For additive noise, $\quad E[N(t)] = 0$

  $\therefore \quad E[R(t)] = E[S(t)]$
- The variance of $R(t)$ is determined as:

  $$\text{Var}[R(t)] = \text{Var}[S(t)] + \text{Var}[N(t)] + 2\text{Cov}[S(t), N(t)]$$

  Since $S(t)$ is a deterministic signal,

  $$\text{Var}[R(t)] = \text{Var}[N(t)]$$
- The signal-to-noise ratio (SNR) is defined as the ratio of the signal power to noise power i.e.

  $$\text{SNR} = \frac{P_S}{P_N}$$

Hence, statement 4 is not correct.

25. (b)

$$U(x) = P\{2W \le x\}$$
$$= P\{W \le x/2\}$$
$$V(x) = P\{V \le x\} = P[W \le x]$$

(since W and V are identically distributed)

Since, CDF is a monotonically non-decreasing function, hence

For positive values of x,

$$U(x) - V(x) \ge 0 \quad \Rightarrow \quad x[U(x) - V(x)] \ge 0$$

For negative values of x,

$$U(x) - V(x) < 0 \quad \Rightarrow \quad x[U(x) - V(x)] \ge 0$$

Hence $[U(x) - V(x)]x/2 \ge 0$, for all values of x.

26. (c)

Probability of '0', $P(0) = \alpha$

and Probability of '1', $P(1) = \beta = 1 - \alpha$

We have,

$$\mu_X = E[X] = \sum_{i=0}^{1} x_i P(x_i) = 0 \times \alpha + 1 \times \beta = \beta = 1 - \alpha$$

$$E[X^2] = \sum_{i=0}^{1} x_i^2 P(x_i) = (0)^2\alpha + (1)^2(\beta) = \beta = 1 - \alpha$$

Variance, $$\sigma_X^2 = E[X^2] - \{E[X]\}^2$$
$$= \beta - \beta^2 = (1 - \alpha) - (1 - \alpha)^2$$
$$= (1 - \alpha) - (1 + \alpha^2 - 2\alpha)$$
$$= 1 - \alpha + 2\alpha - 1 - \alpha^2$$
$$= \alpha - \alpha^2 = \alpha(1 - \alpha) = \alpha\beta$$

27. (d)

The frequency response $H(\omega)$ of the system is

$$H(\omega) = FT.[h(t)] = \frac{5}{2}\frac{1}{j\omega + 3}$$

The mean value of $Y(t)$ is

$$\mu_Y(t) = E[Y(t)] = \mu_X H(0) = 3 \times \frac{5}{2} \times \frac{1}{3} = 2.5$$

28. (b)

- The PDF of Y is derived using the transformation rule as,

$$f_Y(y) = f_X(x)\left|\frac{dx}{dy}\right|$$

- Linearity preserves relationships but not necessarily distributions. When $g(X)$ is linear, say $Y = aX + b$, $\mu_Y = a\mu_X + b$. Thus, X and Y are not identically distributed.

- Expectation is linear, so
  $E[Y] = E[g(X)]$
- The variance of Y depends on the variance of X. For example, if $Y = aX + b$, $\text{Var}[Y] = a^2\text{Var}[X]$.

29. (d)

- When SNR is constant, the channel capacity increases with increasing B. Larger bandwidth allows for more data transmission.
- Noise power also increases with B, and capacity saturates for very large B. For an infinite bandwidth channel, $C = 1.44\dfrac{S}{N_0}$
- A higher SNR improves the $\log_2(1 + \text{SNR})$ term, thereby increasing capacity.
- For small SNR values (SNR << 1)

$$C = B\log_2(1 + \text{SNR}) = B\frac{\ln(1+\text{SNR})}{\ln 2} \approx B\frac{\text{SNR}}{\ln 2} = B(\text{SNR})\log_2 e$$

30. (b)

- The entropy $H(s)$ of the source,

$$H(s) = -\sum_{i=1}^{4} P(S_i)\log_2 P(S_i)$$
$$= -(0.4\log_2 0.4 + 0.3\log_2 0.3 + 0.2\log_2 0.2 + 0.1\log_2 0.1)$$
$$= 0.528 + 0.521 + 0.464 + 0.332$$
$$H(s) = 1.846 \text{ bits}$$

- Entropy quantifies the average bits per symbol needed for an optimal encoding scheme.
  For example, $H(s) = 1.846$ bits means approximately 1.846 bits are required per symbol when encoding this source efficiently.
- The entropy of a distribution is maximized when the probabilities are uniform (all symbols are equally likely). For a source with 4 symbols, maximum entropy occurs when

$$P(S_1) = P(S_2) = P(S_3) = P(S_4) = 0.25$$

  Since the given probabilities 0.4, 0.3, 0.2, 0.1 is not uniform, the entropy is not maximized.
- Entropy decreases when one symbol becomes more probable, because the uncertainty in the source reduces.
  For example, if $P(S_1)$ increases, the remaining probabilities $P(S_2)$, $P(S_3)$, $P(S_4)$ must decrease (as their sum must be 1). This reduction in uncertainty results in a lower entropy.

31. (a)

- Mutual information $I(X : Y)$ quantifies the amount of information that the random variables X (input) and Y (output) share. It is defined as

$I(X : Y) = H(X) - H(X/Y) = H(Y) - H(Y/X)$

where,

$H(X)$ is entropy of X.

(X/Y) is the conditional entropy of X given Y.

$H(Y)$ is the entropy of Y.

$H(Y/X)$ is the conditional entropy of Y given X.
$I(X : Y) \geq 0$; Mutual information is always non negative.
This follows from the definition and properties of entropy.
The conditional entropy $H(X/Y)$ or $H(Y/X)$ can never exceed the respective unconditional entropy $H(X)$ or $H(Y)$, so $I(X : Y) \geq 0$.

- $I(X : Y)$ is symmetric, i.e., $I(X : Y) = I(Y : X)$
  Mutual information is symmetric. The amount of information X provides about Y is equal to the amount of information Y provides about X. This is mathematically expressed as:
  $I(X : Y) = H(X) - H(X/Y) = H(Y) - H(Y/X)$
- Mutual information is maximized when X and Y are highly dependent (perfectly correlated) and not independent.
- If X and Y are independent, their mutual information is zero because knowing X provides no information about Y, and vice versa.
- Conversely, $I(X : Y)$ is maximized when $H(X/Y) = 0$ meaning Y completely determines X (or X completely determines Y).

32. (c)

The condition to detect upto 'e_d' bit errors and simultaneously correct upto 'e_c' bit errors is

$$d_{\min} \geq (e_d + e_c + 1)$$

For the given LBC, $d_{\min} = 3$

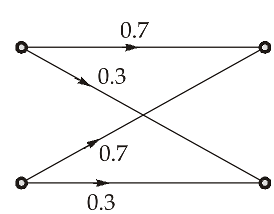
So, $$e_d + e_c + 1 \leq 3$$
$$e_d + e_c \leq 2$$

Hence, the LBC with dmin = 3 can detect upto 1 bit error and simultaneously correct upto 1 bit error.

33. (a)

- The degree of $G(x)$ determines the number of check bits that are appended to the message to generate CRC code.
- CRC is based on modulo-2 division. A valid CRC codeword $C(x)$, is designed to be exactly divisible by the generator polynomial, $G(x)$. Thus, $C(x)$ mod $G(x) = 0$ in modulo-2 arithmetic, ensures the integrity of the transmitted data.
- CRC is primarily for error detection, not correction.
- A burst error of length ≤ the degree of $G(x)$ is always detected by the CRC. Here, the degree of $G(x)$ is three, hence if a burst error of length 3 occurs, it will always be detected.

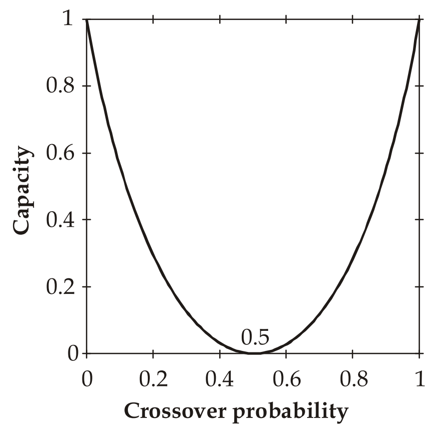
34. (d)



The capacity of a binary symmetric channel is given by

$$C = 1 - H(P) = 1 + P\log_2 P + (1-P)\log_2(1-P)$$
$$= 1 + 0.3\log_2 0.3 + 0.7\log_2 0.7 = 0.118 \text{ bits/sec}$$

The capacity of a binary symmetric channel varies with crossover probability as shown below:



- Maximum capacity is achieved when $P = 0$.
- When $P = 0.5$, then capacity, $C = 0$, i.e., the minimum value.
- For $0.3 \le P \le 0.5$, increasing P decreases C.

Hence, statements 2 and 3 are correct.

35. (b)

- Entropy is always non-negative ($H(X) \ge 0$) since probabilities are non-negative.
- Entropy is zero for deterministic variables as $P(x) = 1$ i.e., $\log_2 P(x) = 0$
- Entropy measures the uncertainty or randomness associated with a random variable. A uniform distribution where all possible N symbols are equiprobable i.e. $P(x) = \frac{1}{N}$ maximizes $H(X)$ given by $H_{max} = \log_2(N)$
- Adding an outcome increases entropy only if the distribution becomes more uniform and not necessarily for less probable outcomes. Thus, statement 4 is not correct.

36. (c)

Shannon's Source Coding Theorem states that the average number of bits required to represent a symbol from a source cannot be less than the entropy of that source, $H(s)$. Thus, the

- The theoretical minimum of average codelength, L_{avg} is $H(s)$ in the ideal case.
- In practical coding scheme, $L_{avg} \ge H(s)$, as no compression is perfectly ideal.
- $H(s)$ serves as a theoretical lower bound for lossless compression.

37. (c)

The matched filter is optimal for detecting signals in AWGN because it maximizes SNR, not because of any thresholding process. Thresholding is part of decision-making and not filtering. After the matched filter produces its output, a decision is made based on comparing the output to a threshold.

38. (a)

- According to the Shannon-Hartley theorem, the channel capacity C is given by:

$$C = B \log_2(1 + \text{SNR})$$

- For a noiseless channel, SNR $\rightarrow \infty$, So;

$$C = B \log_2(\infty) \rightarrow \infty$$

- The Shannon-Hartley theorem defines the maximum rate at which information can be transmitted over a communication channel with noise, based on the channel's bandwidth and signal-to-noise ratio. For a noiseless channel, there is no noise, and the SNR becomes infinite. Since the formula does not limit the data rate in such a case, this reason is valid.

Section B : Electronic Devices & Circuits-1 + Analog Circuits Topics-1

39. (b)

Ambipolar diffusion refers to the phenomenon where positively and negatively charged particles diffuse together with the same effective diffusion coefficient, drift mobility and life time.

40. (d)

Given,
$$n_i = 1.5 \times 10^{10}\ \text{cm}^{-3}$$
$$\sigma = 2\ (\Omega\text{-cm})^{-1}$$
$$\mu_n = 0.125\ \text{m}^2/\text{V-s} = 1250\ \text{cm}^2/\text{V-s}$$
$$\mu_p = 0.038\ \text{m}^2/\text{V-s} = 380\ \text{cm}^2/\text{V-s}$$

We know that, for p-type semiconductor,
$$\sigma \approx pq\mu_p$$
$$2 = p \times 1.6 \times 10^{-19} \times 380$$

$\therefore$
$$p = \frac{2}{1.6 \times 380 \times 10^{-19}}\ \text{cm}^{-3}$$

Using mass action law,
$$n = \frac{n_i^2}{p} = \frac{2.25 \times 10^{20} \times 1.6 \times 380 \times 10^{-19}}{2}$$

$\therefore$ Electron concentration, $n = 6.84 \times 10^3\ \text{cm}^{-3}$

41. (b)

We know that,

Electron distribution in the conduction band,
$$n(E) = N_c(E)f(E)$$
Given, $n(E) = 10^{14}\ \text{cm}^{-3}$
$$N_c(E) = 3 \times 10^{14}\ \text{cm}^{-3}$$

$\therefore$
$$f(E) = \frac{10^{14}}{3 \times 10^{14}} = \frac{1}{3}$$

The hole distribution in valence band,
$$p(E) = N_V(E)[1 - f(E)]$$

$\therefore$
$$p(E) = 3 \times 10^{14}\left[1 - \frac{1}{3}\right] = 2 \times 10^{14}$$

[For given semiconductor assume $N_C(E) = N_V(E)$]

42. (d)

The probability of state at a energy level E being occupied by an electron is given by the Fermi Dirac function as,

$$f(E) = \frac{1}{1+e^{(E-E_F)/KT}}$$

given, $$E = E_C + \frac{KT}{4};\ E_F = E_C - 0.26 \text{ eV}$$

$\therefore$ $$f\left(E_C + \frac{KT}{4}\right) = \frac{1}{1+e^{\left(E_C+\frac{KT}{4}-E_F\right)/KT}} = \frac{1}{1+e^{\left(\frac{0.26}{KT}+\frac{1}{4}\right)}}$$

$$= \frac{1}{1+e^{\left(\frac{0.26}{0.026}+0.25\right)}} \approx \frac{1}{e^{10.25}}$$

$$f\left(E_C + \frac{KT}{4}\right) = e^{-10.25}$$

43. (b)

We know that,

In an n-type semiconductor, the majority carrier concentration is

$$n_0 \approx N_d(x) = n_i \exp\left(\frac{E_F - E_t}{kT}\right)$$

Since N_d is a function of x for a non-uniformly doped semiconductor, therefore, $E_F - E_i$ should also be the function of x. Hence,

$$E_F - E_i(x) = kT \ln\left(\frac{N_d(x)}{n_i}\right)$$

The induced field is given as,

$$E(x) = \frac{1}{q}\frac{dE_i(x)}{dx}$$

The induced E-field, $$E(x) = -\left(\frac{kT}{q}\right)\frac{1}{N_d(x)}\frac{dN_d(x)}{dx}$$

We have, $$\frac{dN_d(x)}{dx} = -10^{19} \text{ cm}^{-4}$$

$\therefore$ $$E(x) = -0.026 \times \frac{1}{10^{16} - 10^{19}x}(-10^{19})$$

At $x = 0$, $$E(0) = -0.026 \times \frac{-10^{19}}{10^{16}} = 26 \text{ V/cm}$$

44. (a)

The net increase in the electron concentration per unit time is the difference between the electron flux per unit volume, minus the recombination rate. This results in the continuity equation for electrons, given as

$$\frac{\partial \delta_n}{\partial t} = \frac{1}{q}\frac{\partial J_n}{\partial x} - \frac{\delta_n}{\tau_n}$$

The electron diffusion current density is given by

$$J_n = D_n \frac{\partial \delta_n}{\partial x}.$$

Thus, we get,
$$\frac{\partial \delta_n}{\partial t} = D_n \frac{\partial^2 \delta_n}{\partial x^2} - \frac{\delta_n}{\tau_n}$$

45. (c)

Given, depletion width, $W = x_n + x_p = 0.457$

and
$$\frac{N_a}{N_d} = 200$$

Since the net charge in a depletion region is zero, hence

$$x_n N_d = x_p N_a$$

where (x_n, N_d) and (x_p, N_a) are the depletion width and doping concentration on n-side and p-side respectively

We get, depletion width on n-side,

$$x_n = \frac{W}{1+\frac{N_d}{N_a}} = \frac{0.457}{1+\frac{1}{200}} = \frac{0.457}{1+0.005}$$

$$x_n \simeq 0.455 \ \mu\text{m}$$

46. (a)

For the p-n junction, the ratio of the hole concentration at the edge of the transition region is given as,

$$\frac{p_n}{p_p} = \frac{3}{2}$$

where p_p is the hole concentration on p-side and p_n is the hole concentration on n-side. Thus,

$$p_p = N_a \quad \text{and} \quad p_n = \frac{n_i^2}{N_d}$$

We have, Contact potential,
$$V_0 = \frac{kT}{q}\ln\left(\frac{N_a N_d}{n_i^2}\right)$$

$$= \frac{kT}{q}\ln\left(\frac{p_n}{p_p}\right)$$

$$= 26\times 10^{-3}\ln\left(\frac{3}{2}\right)$$

$$= 26 \times 10^{-3}(\ln(3) - \ln(2))$$

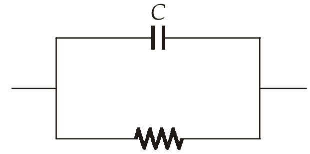
$$= 26 \times 10^{-3}(1.098 - 0.693)$$

$$= 0.01 \text{ V}$$

$$= 10 \text{ mV}$$

47. (a)

The junction capacitance is dominant when the diode is reverse biased and is the result of the charge stored in the Depletion layer whereas the Diffusion capacitance is dominant when the diode is forward biased and is the result of the stored minority carriers near the depletion region, under reverse bias condition, the equivalent circuit of pn-junction diode,



$$\therefore \quad Y = G + j\omega C_{jn}$$

where C_{jn} : Junction capacitance

48. (c)

From the given output characteristics, $V_P = -1.5$ V and $V_{GS} = 0$ V

The drain current for JFET is given by

$$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

At $V_{GS} = 0$; $I_D = I_{DSS} = 5$ mA

At $V_{GS} = -1$ V,

$$I_D = 5\text{ mA}\left(1 - \frac{-1}{-1.5}\right)^2$$

$$I_D = 1.67\text{ mA}$$

49. (b)

For the diode to be ON,

$$2 - V_{in} > 0.7 \quad \Rightarrow \quad V_{in} < 1.3\text{ V}$$

Thus, for $V_{in} < 1.3$ V, $$V_0 = V_{in} - 1k\left(\frac{V_{in} - 2 + 0.7}{2k}\right)$$

$$\therefore \quad V_0 = V_{in} - \frac{1}{2}(V_{in} - 1.3)$$

$$= V_{in} - 0.5\,V_{in} + 0.65$$

$$V_0 = 0.5V_{in} + 0.65$$

For $V_{in} > 1.3$ V, diode is OFF and acts as open circuit. Thus, $V_0 = V_{in}$. For $V_{in} = 2.3$ V, the output $V_0 = 2.3$ V.

50. (b)

Given, $$I_{DSS} = 2\text{ mA}$$

$$I_{DS} = \frac{V_{DD} - 15}{50\text{ k}\Omega} = \frac{30 - 15}{50\text{ k}\Omega} = 0.3\text{ mA}$$

For a JFET, we have, $$I_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2$$

$$\therefore \qquad V_{GS} = V_P\left[1-\sqrt{\frac{I_{DS}}{I_{DSS}}}\right] = -2\left[1-\sqrt{\frac{0.3}{2}}\right]$$

$$V_{GS} = -1.22 \text{ V}$$

From input loop; $V_{GS} + I_{DS} \times R = 0$ [Gate current, $I_G \approx 0$]

$$\therefore \qquad R = \frac{1.22}{0.3 \text{ mA}} = 4 \text{ k}\Omega$$

51. (c)

Using KVL,

$$-3 + R_C(I_C + I_B) + I_B R_B + 0.7 = 0$$

$$-3 + I_B(R_C + R_B) + \beta I_B R_C + 0.7 = 0 \qquad [\because I_c = \beta I_B]$$

$$I_B = \frac{3-0.7}{R_B + (1+\beta)R_C}$$

$$I_B = \frac{3-0.7}{(33\times 10^3)+(1+90)(1.8\times 10^3)}$$

$$\Rightarrow \qquad I_B = 11.69\ \mu\text{A}$$

$$\therefore \qquad I_C = \beta I_B$$

$$= (90) \times (11.69 \times 10^{-6})$$

$$\Rightarrow \qquad I_C \simeq 1.05 \text{ mA}$$

DC bias voltage, $V_{CE} = V_{CC} - I_C R_C$

$$= 3 - (1.05 \times 1.8)$$

$$= 1.11 \text{ V}$$

52. (c)

Given, $A_V = 200$

$\beta = 50$

$R_i = 2.5 \text{ k}\Omega;\ V_i = 5 \text{ mV}$

power gain, $A_p = A_i \cdot A_V$

where, $A_i = \beta = 50$

$\therefore \qquad A_p = 50 \times 200 = 10000 = 10^4$

A_p in dB $\Rightarrow$ $10 \log_{10} A_p = 10 \times 4 \log_{10} 10 = 40 \text{ dB}$

53. (d)

We know that,

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_T)^2 \qquad \ldots\text{(i)}$$

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \mu_n C_{ox}\frac{W}{L}(V_{GS}-V_T)$$

From equation (i), $V_{GS} - V_T = \sqrt{\dfrac{2I_D}{\mu_n C_{ox}\dfrac{W}{L}}}$

$\therefore$ $$g_m = \sqrt{2\mu_n C_{ox}\frac{W}{L}I_D} = \sqrt{2\times100\times10^{-6}\times\frac{10}{0.18}\times10^{-3}}$$

$\therefore$ $$g_m = \frac{1}{300}\text{S}$$

Small signal voltage gain, $$A_V = -g_m R_D$$

$$= \frac{-1}{300}\times1\text{ k}\Omega = -3.33$$

54. (a)

Given, $$A_{VM} = 150$$
$$A_{VL} = 100 \text{ at } f = 50 \text{ Hz}$$

We have, $$A_{VL} = \frac{A_{VM}}{\sqrt{1+\left(\frac{f_L}{f}\right)^2}}$$, where f_L is the 3-dB frequency of the amplifier

$$100 = \frac{150}{\sqrt{1+\left(\frac{f_L}{50}\right)^2}}$$

On solving, $$f_L = 55.90 \text{ Hz}$$

55. (b)

- For C.E. stage, all parasitic capacitors affect the frequency response.
- All parasitic capacitors affect the frequency response of C.B. stage.
- In C.C. stage of BJT, only C_μ will not affect frequency response since it is present between the collector and the substrates. Since, in common collector configuration, the collector terminal is grounded, therefore, C_μ gets deactivated.

56. (c)

Given, $$D_p = 12 \text{ cm}^2/\text{s}$$
$$W_b = 0.5 \text{ }\mu\text{m}$$

Transit time of holes in base,

$$\tau_t = \frac{W_b^2}{2D_p} = \frac{(0.5\times10^{-4})^2}{2\times12} = \frac{2.5\times10^{-9}}{24}$$

$$\tau_t = 1.04 \times 10^{-10} \quad \text{(or)} \quad 0.1 \text{ nsec}$$

57. (a)

In a p-njunction diode, the potential barrier, does not oppose the flow of minority carriers, but rather facilitates their movement across the junction. The drift current is due to the drift of the minority carriers by the E-field in the transition region. However, this current is small not because of the size of the barrier, but because of the few minority carriers available.

Section C : Control Systems-2 + Microprocessors and Microcontroller-2

58. (a)

For lead compensator, $\alpha = \frac{\text{Zero}}{\text{Pole}} < 1$ i.e. zero is closer to the origin than the pole. and

For lag compensator, $\alpha = \frac{\text{Zero}}{\text{Pole}} > 1$ i.e. pole is closer to the origin than the zero.

In a lead lag compensator, lead system is dominant, hence lead compensator's zero and pole are closer to the origin than the zero and pole of lag compensator. Hence, option (a) represents the pole-zero configuration of a lead lag compensator.

59. (d)

We know that, in a lead compensator, the maximum phase lead occurs at the frequency given by the square root of the product of the pole and zero locations.

$$\omega_m = \sqrt{(\text{zero's location})\,(\text{Pole's location})}$$

From the given transfer function, $G_c(s)$

$$\text{zero is at } s = \frac{-1}{4A} \text{ and}$$

$$\text{Pole is at } s = \frac{-1}{A}$$

$$\omega_m = 50 \text{ rad/sec} = \sqrt{\left(\frac{-1}{4A}\right)\left(-\frac{1}{A}\right)}$$

$$50 \text{ rad/sec} = \frac{1}{2A}$$

$$A = 0.01$$

Now, maximum phase shift, $\phi_m = \sin^{-1}\left(\frac{1-\alpha}{1+\alpha}\right)$

where, $\alpha = \frac{\text{zero}}{\text{pole}} = \frac{\left(\frac{-1}{4A}\right)}{\left(\frac{-1}{A}\right)}$

$$\alpha = \frac{1}{4}$$

$$\phi_m = \sin^{-1}\left(\frac{1-\frac{1}{4}}{1+\frac{1}{4}}\right)$$

$$\phi_m = \sin^{-1}\left(\frac{3\times 4}{4\times 5}\right)$$

$$\phi_m = \sin^{-1}\left(\frac{3}{5}\right) = 37^\circ$$

60. (a)

For controllability, we consider the controllability matrix,

$$Q_c = \begin{bmatrix} B & AB \end{bmatrix}$$

$$AB = \begin{bmatrix} a & b \\ -b & a \end{bmatrix}\begin{bmatrix} 1 \\ 1 \end{bmatrix}$$

$$AB = \begin{bmatrix} a+b \\ a-b \end{bmatrix}$$

$$Q_c = \begin{bmatrix} 1 & a+b \\ 1 & a-b \end{bmatrix}$$

Thus,
$$|Q_c| = (a - b) - (a + b)$$
$$= a - b - a - b$$
$$|Q_c| = -2b$$

Thus, we say that system's controllability is independent of 'a'.

For observability, we consider the observability matrix,

$$Q_0 = \begin{bmatrix} C^T & A^TC^T \end{bmatrix}$$

$$A^TC^T = \begin{bmatrix} a & -b \\ b & a \end{bmatrix}\begin{bmatrix} 1 \\ 1 \end{bmatrix} = \begin{bmatrix} a-b \\ a+b \end{bmatrix}$$

$$Q_0 = \begin{bmatrix} 1 & a-b \\ 1 & a+b \end{bmatrix}$$

Thus,
$$|Q_0| = (a + b) - (a - b)$$
$$|Q_0| = a + b - a + b$$
$$|Q_0| = 2b$$

Hence, observability of the system is also independent of 'a'.

61. (d)

We know that,

For open loop transfer function $G(s)H(s) = \dfrac{1}{s(sT_1+1)(sT_2+1)}$, the phase crossover frequency and the gain margin is given by

$$\omega_{pc} = \frac{1}{\sqrt{T_1T_2}} \text{ and gain margin, } GM = \frac{1}{|G(j\omega)|_{\omega=\omega_{pc}}} = \frac{T_1+T_2}{T_1T_2}$$

From the given open loop transfer function,

we get, $T_1 = 2$ and $T_2 = 3$

Thus,

$$\text{Phase crossover frequency, } \omega_{pc} = \frac{1}{\sqrt{2\times3}} = \frac{1}{\sqrt{6}} \text{ and}$$

$$\text{Gain margin, } GM = \frac{2+3}{2\times3} = \frac{5}{6}$$

62. (d)

From the given data, we have

One pole at origin, thus it implies type of the system is 1.

and as the number of branches tending towards infinity from poles is two; it implies $P - Z = 2$

where, P = total number of poles and

Z = total number of zeros

We know that,

In polar plot, for a minimum phase system i.e. with no poles and zeros in the right side of the s-plane, phase is varied from

$$(-90°\times \text{type})\big|_{\omega=0} \text{ to } (-90°\times(P-Z))\big|_{\omega=\infty}$$

Here, type = 1 and $P - Z = 2$

Hence, it is varied from, –90° × 1 = –90° at ω = 0 to –90° × 2 = –180° at $\omega = \infty$.

Therefore, option (d) is correct.

63. (a)

- The magnitude of the frequency response attains a maximum value at the resonant frequency. This maximum value is called the resonant peak.
- The resonance peak occurs for ξ less than $\frac{1}{\sqrt{2}}$
- Phase margin, PM = $\tan^{-1}\left[\frac{2\xi}{\sqrt{\sqrt{4\xi^4+1}-2\xi^2}}\right]$

Hence, option (a) is correct.

64. (c)

We have,
$$G(s) = \frac{ks}{(s+1)(s+3)}$$

$$|G(j\omega)| = \frac{j\omega k}{(j\omega+1)(j\omega+3)} = \frac{\omega k}{\sqrt{\omega^2+1}\sqrt{\omega^2+9}}$$

$$\angle G(j\omega) = 90° - \tan^{-1}(\omega) - \tan^{-1}\left(\frac{\omega}{3}\right)$$

We know that, for LTI system,

$$\text{output, } c(t) = B\sin(\omega t + \phi_2)$$

where,
$$B = 2\times |G(j\omega)|_{\text{at } \omega = 3}$$

$$B = 2\times\frac{3k}{\sqrt{10}\sqrt{18}}$$

$$B = \frac{6k}{\sqrt{180}} = \frac{k}{\sqrt{5}}$$

and
$$\phi_2 = \phi_1 + |G(j\omega)|_{\omega=3\text{ rad/sec}}$$

$$\phi_2 = 60° + 90° - 71.56° - 45°$$

$$\phi_2 = 33.44°$$

Thus, $$c(t) = \frac{k}{\sqrt{5}}\sin(3t + 33.44°)$$

65. (c)

- Transfer function based analysis is applicable for Linear time invariant (LTI) system whereas state model is applicable for any type of system. Therefore, statement 1 is incorrect.
- The transfer function is defined under the assumption of zero initial conditions whereas the state space model incorporates the initial conditions.
- The transfer function of the system is unique but the state space representation is not unique; a linear physical system can have more than one state space representation.

66. (a)

Universal Asynchronous Receiver/Transmitter or UART's main purpose is to transmit and receive serial data. Two UART's communicate directly with each other. The transmitting UART converts parallel data from controller into serial data, and transmits it to the receiving UART where data gets converted back to parallel form.

67. (c)

In 8051 microcontroller, the program status word (PSW) register is an 8-bit register. It is also referred to as the flag register. Although PSW register is 8 bits wide, only 6 bits of it are used by the 8051. The two unused bits are user-definable flags. Four of the flags are called conditional flags. These four are CY (Carry), AC (Auxiliary carry), P (Parity), and OV (overflow) used to indicate arithmetic conditions after an instruction is executed.

CY	AC	F0	RS1	RS0	OV	–	P

68. (c)

Performance is the key deciding factor in the selection of the general purpose computing system whereas application specific requirements like performance, power requirements, etc. are the key deciding factors in selecting an embedded system. Hence, statement 3 is not correct.

69. (c)

The 8253 can be operated in the following six modes:

- Mode 0 $\Rightarrow$ Interrupt on terminal count
- Mode 1 $\Rightarrow$ Programmable one shot
- Mode 2 $\Rightarrow$ Rate generator
- Mode 3 $\Rightarrow$ Square wave generator
- Mode 4 $\Rightarrow$ Software triggered strobe
- Mode 5 $\Rightarrow$ Hardware triggered strobe

70. (d)

- Preemptive scheduling allows higher-priority tasks to interrupt lower-priority ones when necessary. Hence, it is used in real-time systems to guarantee that high-priority tasks are executed within a specific time frame.

- Since cooperative scheduler relies on processes yielding the CPU voluntarily, it is more prone to CPU starvation, as a process can keep the CPU indefinitely if it doesn't voluntarily release control.

Hence, statements 1, 2 and 3 are correct.

71. (a)

Polling sequentially check the status of each device, while interrupts allow the microcontroller to continue it's operation until a specific event occurs. It helps prioritize and handle tasks as they arise without continuous monitoring.

72. (c)

- When the reset pin is activated, the 8051 jumps to address location 0000 and start executing the reset routine from 0000 H.
- 8051 microcontroller has 6 interrupt sources, out of which one is reset, two are internal (Timer interrupts: TF0, TF1), two are external (INT0, INT1) and one is a serial interrupt (RI/TI).
- In 8051 microcontroller, polling is not more efficient, it consume more CPU resources as it continuously monitor devices status, unlike interrupts which respond only when necessary.

Thus, only statement 3 is not correct. Hence, option (c) is the answer.

73. (d)

The six interrupts in the 8051 microcontroller (including the reset) are:
Reset, external interrupt 0 (INT 0); Timer 0 overflow interrupt; Timer 1 overflow interrupt; External Interrupt 1 (INT 1); and Serial Communication Interrupt (RI/TI)
Hence, INTR is not an interrupt of 8051 microcontroller.

74. (a)

Integral controller adds one pole to the system, which shifts the root locus towards right side of s-plane, thereby decreasing the stability. Both the statements are correct and statement-II is the correct explanation of statement-I.

75. (a)

Direct memory access (DMA) allows for data transfer without involving the central processing unit (CPU). This significantly reduces the CPU overhead compared to interrupt-driven I/O, where the CPU is interrupted for every byte or word transferred. Hence, DMA is more efficient for high-volume data transfers. DMA Controller requests for the data bus lines from CPU through the HOLD pin and waits for the CPU to assert the HLDA. When the HLDA pin is asserted by the CPU, the DMA controller can transfer the bulk data to the memory through these buses without an interrupt mechanism. Both the statements are correct and statement-II is correct explanation of statement-I.

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