GATE Electronics and Communication Engineering

Sample Paper – 6

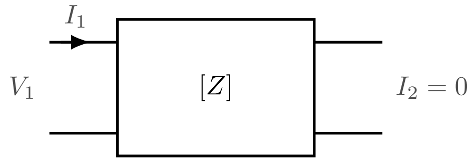
Duration: 128 Minutes **Maximum Marks: 72**

Instructions

- This paper contains **46 questions** covering the core **Electronics and Communication Engineering** subject of the GATE paper conducted by the IITs and IISc (General Aptitude and Engineering Mathematics are provided as separate papers).
- **Q1–Q20 carry 1 mark each** and **Q21–Q46 carry 2 marks each**, for a total of **72 marks**.
- Each question carries **negative marking of one-third of its allotted marks**: a wrong 1-mark question costs **0.33** and a wrong 2-mark question costs **0.67**. Unattempted questions carry no penalty.
- Every question here is a single-correct **MCQ**. The real GATE paper also has Multiple Select (MSQ) and Numerical Answer Type (NAT) questions, and **those carry no negative marking**.
- Attempt this paper in one timed sitting of about **128 minutes**, the share this core subject earns at GATE's overall pace of 180 minutes for 65 questions. Unless stated otherwise take $n_i = 1.5 \times 10^{10}$ cm^{-3} for silicon, thermal voltage $V_T = 25$ mV at room temperature, $\varepsilon_0 = 8.854 \times 10^{-12}$ F/m, $\mu_0 = 4\pi \times 10^{-7}$ H/m and $c = 3 \times 10^8$ m/s.

Networks and Circuit Analysis

Q1. A linear two-port network is described by its open-circuit impedance (Z) parameters, as sketched. With port 2 left open ($I_2 = 0$) a current $I_1 = 2$ A is driven into port 1, and the driving-point parameter is $Z_{11} = 15\ \Omega$. The resulting port-1 voltage $V_1 = Z_{11}I_1$ is: **[1 mark]**



(A) 10 V

(B) 20 V

(C) 30 V

(D) 45 V

Q2. A reciprocal two-port network described by its transmission (ABCD) parameters must satisfy $AD - BC = 1$. A particular two-port has $A = 2$, $B = 3\ \Omega$ and $C = 1$ S. For the network to be reciprocal, the parameter D must be: **[1 mark]**

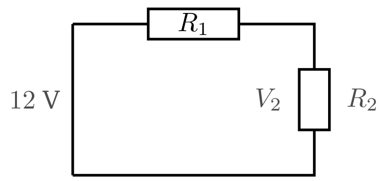
(A) 1

(B) 3

(C) 4

(D) 2

Q3. In the resistive voltage divider shown, a 12 V d.c. source drives two series resistors $R_1 = 4\ \Omega$ and $R_2 = 8\ \Omega$. The voltage across R_2, given by $V_2 = V\dfrac{R_2}{R_1 + R_2}$, is: **[1 mark]**



(A) 8 V

(B) 4 V

(C) 6 V

(D) 12 V

Q4. An ideal capacitor of $C = 2\ \mu$F is driven by a sinusoidal source of angular frequency $\omega = 5000$ rad/s. The magnitude of the capacitive reactance $X_C = \dfrac{1}{\omega C}$ is: **[1 mark]**

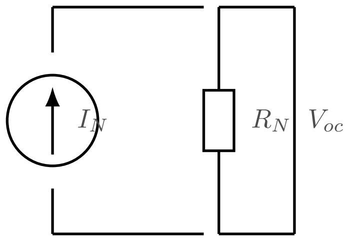
(A) $100\ \Omega$

(B) $10\ \Omega$

(C) $0.01\ \Omega$

(D) $1000\ \Omega$

Q5. A practical current source is modelled by its Norton equivalent with Norton current $I_N = 4$ A and Norton resistance $R_N = 6\ \Omega$, as shown. The open-circuit voltage at its terminals, $V_{oc} = I_N R_N$, is: **[1 mark]**



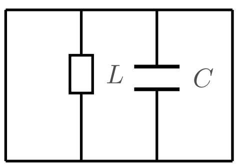
(A) 4 V

(B) 6 V

(C) 10 V

(D) 24 V

Q6. A parallel RLC circuit has $L = 4$ mH and $C = 25$ nF, as shown. The undamped natural (resonant) angular frequency $\omega_0 = \dfrac{1}{\sqrt{LC}}$ of the circuit is: **[1 mark]**



(A) 10000 rad/s

(B) 50000 rad/s

(C) 200000 rad/s

(D) 100000 rad/s

Electronic Devices

Q7. A silicon sample is uniformly doped with acceptor atoms of concentration $N_a = 5 \times 10^{16}$ cm^{-3}. Taking $n_i = 1.5 \times 10^{10}$ cm^{-3} and full ionisation, the equilibrium electron (minority-carrier) concentration $n = \dfrac{n_i^2}{N_a}$ is nearest to: **[1 mark]**

(A) 4.5×10^3 cm^{-3}

(B) 5×10^{16} cm^{-3}

(C) 1.5×10^{10} cm^{-3}

(D) 2.25×10^{20} cm^{-3}

Q8. At thermal equilibrium in an *intrinsic* silicon sample the electron and hole concentrations are equal ($n = p = n_i$). Taking the standard room-temperature value, the equilibrium electron concentration in intrinsic silicon is: **[1 mark]**

(A) 0

(B) 1.5×10^{10} cm^{-3}

(C) 2.25×10^{20} cm^{-3}

(D) 3×10^{10} cm^{-3}

Q9. In a non-degenerate silicon sample the diffusion coefficient and mobility of electrons are linked by the Einstein relation $D_n = \mu_n V_T$. With $\mu_n = 1350$ cm^2/V · s and $V_T = 0.025$ V, the electron diffusion coefficient D_n is: **[1 mark]**

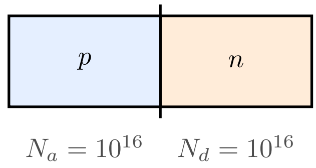
(A) 1350 cm^2/s

(B) 33.75 cm^2/s

(C) 54000 cm^2/s

(D) 0.025 cm^2/s

Q10. An abrupt silicon p–n junction is doped with $N_a = N_d = 10^{16}$ cm^{-3}, as sketched. The built-in potential is $V_{bi} = V_T \ln \frac{N_a N_d}{n_i^2}$, with $V_T = 0.025$ V and $n_i = 1.5 \times 10^{10}$ cm^{-3}. Its value is nearest to: **[1 mark]**



(A) 0.36 V

(B) 0.55 V

(C) 0.67 V

(D) 1.12 V

Q11. In an n-type silicon bar an electric field $E = 50$ V/cm is applied along its length. Taking the electron mobility $\mu_n = 1000$ cm^2/V $\cdot$ s, the electron drift velocity $v_d = \mu_n E$ is: **[1 mark]**

(A) 1000 cm/s

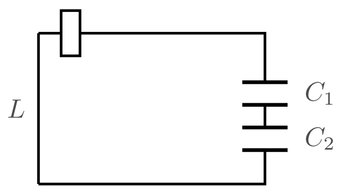
(B) 5×10^4 cm/s

(C) 50 cm/s

(D) 20 cm/s

Analog Circuits

Q12. The tank circuit of a Colpitts oscillator uses two capacitors $C_1 = 100$ pF and $C_2 = 100$ pF in series with an inductor L, as shown. The effective (series) capacitance $C = \frac{C_1 C_2}{C_1 + C_2}$ that resonates with L is: **[1 mark]**



(A) 200 pF

(B) 100 pF

(C) 50 pF

(D) 25 pF

Q13. The Colpitts tank of the previous type has effective capacitance $C = 50$ pF and inductance $L = 10\ \mu$H. The frequency of oscillation $f_0 = \dfrac{1}{2\pi\sqrt{LC}}$ is nearest to: **[1 mark]**

(A) 1.59 MHz

(B) 3.56 MHz

(C) 14.2 MHz

(D) 7.12 MHz

Q14. A Wien-bridge oscillator uses equal resistors $R = 16$ kΩ and equal capacitors $C = 1$ nF in its frequency-selective network. The frequency of oscillation $f_0 = \dfrac{1}{2\pi RC}$ is nearest to: **[1 mark]**

(A) 1 kHz

(B) 6.28 kHz

(C) 9.95 kHz

(D) 15.9 kHz

Q15. For sustained oscillation the non-inverting amplifier of a Wien-bridge oscillator must provide a voltage gain of at least 3, which requires the feedback ratio $R_f/R_1 = 2$. If $R_1 = 10$ kΩ, the feedback resistor R_f must be: **[1 mark]**

(A) 20 kΩ

(B) 10 kΩ

(C) 30 kΩ

(D) 5 kΩ

Q16. An ideal op-amp Miller integrator uses an input resistor $R = 100$ kΩ and a feedback capacitor $C = 0.1\ \mu$F. The integrator time constant $\tau = RC$ is: **[1 mark]**

(A) 0.01 s

(B) 0.1 s

(C) 1 s

(D) 0.001 s

Q17. An RC phase-shift oscillator is built with three identical cascaded RC sections in its feedback path. For the Barkhausen phase condition to be met, the total phase shift contributed by the three-section RC network must be: **[1 mark]**

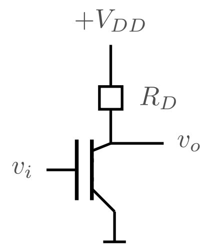
(A) $90°$

(B) $180°$

(C) $360°$

(D) $45°$

Q18. A common-source MOSFET amplifier operates with transconductance $g_m = 2$ mS and a drain resistor $R_D = 10$ kΩ, as shown. Neglecting the output resistance r_o, the magnitude of the small-signal voltage gain $|A_v| = g_m R_D$ is: **[1 mark]**



(A) 2

(B) 5

(C) 200

(D) 20

Digital Circuits

Q19. A 4-bit binary up-counter (modulus 16) starts from the state 0000. After exactly 20 clock pulses, the state held by the counter is: **[1 mark]**

(A) 0000

(B) 1010

(C) 0100

(D) 0110

Q20. A 4-bit ring counter is preset so that exactly one of its flip-flops holds a logic 1, which then circulates around the ring. The number of distinct states in its repeating sequence is: **[1 mark]**

(A) 16

(B) 4

(C) 8

(D) 2

Q21. A Johnson (twisted-ring) counter built from n flip-flops cycles through $2n$ distinct states. To obtain a counter that steps through exactly 10 distinct states, the number of flip-flops required is: **[2 marks]**

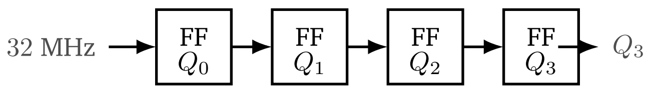
(A) 5

(B) 10

(C) 4

(D) 3

Q22. A mod-16 ripple counter is driven by a 32 MHz clock, as shown. The signal available at its most-significant output Q_3 is the clock divided by 2^4, so its frequency is: **[2 marks]**



(A) 32 MHz

(B) 16 MHz

(C) 8 MHz

(D) 2 MHz

Q23. The four-variable function $F(A, B, C, D) = \sum m(0, 2, 8, 10)$ is plotted on the Karnaugh map shown (rows AB, columns CD). The four ones fall on the corners of the map. The minimal sum-of-products expression for F is: **[2 marks]**

$AB\backslash CD$	00	01	11	10
00	1	0	0	1
01	0	0	0	0
11	0	0	0	0
10	1	0	0	1

(A) $\bar{B}\,\bar{D}$

(B) $\bar{A}\,\bar{C}$

(C) $\bar{B}D$

(D) $B\bar{D}$

Q24. A 4-bit shift register initially holds the pattern 1011 (MSB first). Its contents are shifted right *logically* by two positions, with zeros entering from the left. The final contents of the register are: **[2 marks]**

(A) 1110

(B) 1011

(C) 0101

(D) 0010

Signals and Systems

Q25. A discrete-time signal is $x[n] = (0.4)^n\,u[n]$. Its z-transform is $X(z) = \dfrac{1}{1 - 0.4z^{-1}}$ with ROC $|z| > 0.4$. The value of $X(z)$ evaluated at $z = 1$ (equivalently the DTFT at $\omega = 0$) is: **[2 marks]**

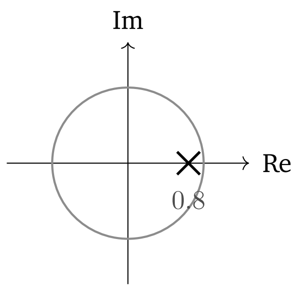
(A) 0.4

(B) 1.667

(C) 2.5

(D) 0.6

Q26. A causal discrete-time LTI system has transfer function $H(z) = \dfrac{z}{z - 0.8}$, whose pole is marked $\times$ on the z-plane below (unit circle drawn for reference). The pole of the system is located at: **[2 marks]**



(A) 0

(B) -0.8

(C) 1

(D) 0.8

Q27. The discrete-time unit impulse is $\delta[n]$. Its discrete-time Fourier transform (DTFT), $\displaystyle\sum_{n=-\infty}^{\infty} \delta[n]e^{-j\omega n}$, is: **[2 marks]**

(A) 0

(B) 1

(C) $e^{-j\omega}$

(D) $2\pi\,\delta(\omega)$

Q28. A discrete-time signal $x[n]$ has DTFT $X(e^{j\omega})$. If the signal is delayed by 3 samples to give $y[n] = x[n-3]$, then by the time-shift property the DTFT $Y(e^{j\omega})$ equals $X(e^{j\omega})$ multiplied by: **[2 marks]**

(A) $e^{-j3\omega}$

(B) $e^{+j3\omega}$

(C) $3e^{-j\omega}$

(D) $e^{-j\omega/3}$

Q29. Two finite-length discrete sequences of lengths 6 and 8 samples respectively are linearly convolved. The length (number of samples) of the resulting sequence is: **[2 marks]**

(A) 13

(B) 14

(C) 48

(D) 6

Q30. The discrete-time Fourier transform $X(e^{j\omega})$ of any discrete-time signal is a periodic function of the angular frequency ω. Its fundamental period is: **[2 marks]**

(A) π

(B) $\pi/2$

(C) 2π

(D) 4π

Control Systems

Q31. The open-loop transfer function of a unity-feedback system is $G(s) = \dfrac{100}{s}$. The gain-crossover frequency ω_{gc}, at which $|G(j\omega)| = 1$, is: **[2 marks]**

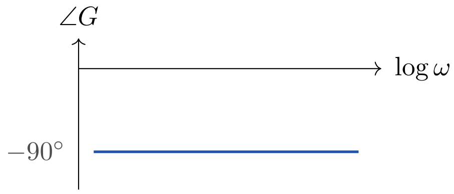
(A) 10 rad/s

(B) 100 rad/s

(C) 1 rad/s

(D) 1000 rad/s

Q32. The open-loop transfer function $G(s) = \dfrac{K}{s}$ of a system contributes a constant phase of -90° at all frequencies, as its Bode phase plot shows. The phase margin, PM $= 180^\circ + \angle G(j\omega_{gc})$, is: **[2 marks]**



(A) 0°

(B) 45°

(C) 180°

(D) 90°

Q33. On a Bode magnitude plot a dimensionless voltage gain of 100 is expressed in decibels as $20\log_{10}(100)$. This equals: **[2 marks]**

(A) 40 dB

(B) 20 dB

(C) 100 dB

(D) 10 dB

Q34. At its phase-crossover frequency (where the phase is -180°) an open-loop system has magnitude $|G(j\omega)| = 0.5$. The gain margin, GM $= -20\log_{10}|G(j\omega_{pc})|$ in decibels, is: **[2 marks]**

(A) 0.5 dB

(B) 3 dB

(C) 6.02 dB

(D) -6 dB

Q35. On a Bode plot a first-order factor $\dfrac{1}{1+s/\omega_c}$ has its corner (break) frequency at ω_c. For the factor $\dfrac{1}{1+0.02s}$ the corner frequency is: **[2 marks]**

(A) 0.02 rad/s

(B) 50 rad/s

(C) 20 rad/s

(D) 100 rad/s

Communications

Q36. A matched filter maximises the output signal-to-noise ratio at the sampling instant. For a signal of energy $E = 8 \times 10^{-6}$ J received in additive white Gaussian noise of power spectral density $\frac{N_0}{2}$ with $N_0 = 2 \times 10^{-6}$ W/Hz, the maximum attainable SNR is $\frac{2E}{N_0}$, which equals: **[2 marks]**

$$r(t) \longrightarrow \boxed{h(t) = s(T{-}t)} \longrightarrow \text{sample } t{=}T$$

(A) 2

(B) 4

(C) 8

(D) 16

Q37. For coherent binary phase-shift keying (BPSK) the bit-error probability in AWGN is $P_b = Q\left(\sqrt{2E_b/N_0}\right)$. If the received ratio is such that $\sqrt{2E_b/N_0} = 2$, and $Q(2) = 0.0228$, the bit error rate is: **[2 marks]**

(A) 0.5

(B) 0.159

(C) 0.32

(D) 0.0228

Q38. The impulse response of a matched filter for a transmitted pulse $s(t)$ of duration T is $h(t) = s(T - t)$, a time-reversed and shifted copy of the pulse. For a rectangular pulse of duration $T = 1$ ms, the matched-filter output attains its peak at the instant: **[2 marks]**

(A) 0 ms

(B) 0.5 ms

(C) 2 ms

(D) 1 ms

Q39. In a digital receiver the ratio $E_b/N_0 = 10$ (as a plain power ratio). Expressed in decibels, $10\log_{10}(E_b/N_0)$, this ratio is: **[2 marks]**

(A) 20 dB

(B) 10 dB

(C) 3 dB

(D) 100 dB

Q40. In an optimum matched-filter receiver for antipodal (BPSK) signalling, the decision statistic depends on $\sqrt{2E_b/N_0}$. With symbol energy $E_b = 10^{-5}$ J and noise density $N_0 = 2 \times 10^{-6}$ W/Hz, the value of $\sqrt{2E_b/N_0}$ is nearest to: **[2 marks]**

(A) 10

(B) 5

(C) 3.16

(D) 1.58

Q41. By Nyquist's criterion the minimum (theoretical) bandwidth required to transmit R_s symbols per second with no intersymbol interference is $R_s/2$. For a symbol rate of $R_s = 2000$ symbols/s, this minimum bandwidth is: **[2 marks]**

(A) 1000 Hz

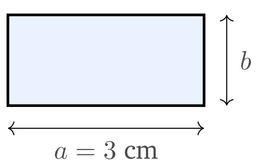
(B) 2000 Hz

(C) 4000 Hz

(D) 500 Hz

Electromagnetics

Q42. An air-filled rectangular waveguide has broad-wall dimension $a = 3$ cm, as shown. For the dominant TE_{10} mode the cut-off frequency $f_c = \dfrac{c}{2a}$ is: **[2 marks]**



(A) 2.5 GHz

(B) 5 GHz

(C) 10 GHz

(D) 1 GHz

Q43. An air-filled rectangular waveguide has dimensions $a = 2$ cm and $b = 1$ cm. The cut-off frequency of a general TE_{mn} mode is $f_{c,mn} = \frac{c}{2}\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}$. The cut-off frequency of the TE_{20} mode is: **[2 marks]**

(A) 7.5 GHz

(B) 15 GHz

(C) 30 GHz

(D) 10 GHz

Q44. In a rectangular waveguide the transverse-magnetic (TM) modes require both mode indices to be non-zero ($m \geq 1$ and $n \geq 1$). Consequently the lowest-order TM mode that can propagate is: **[2 marks]**

(A) TM_{11}

(B) TM_{10}

(C) TM_{01}

(D) TM_{00}

Q45. A rectangular waveguide is operated at a frequency $f = 1.25\, f_c$, where f_c is the cut-off frequency of the propagating mode. The phase velocity inside the guide is $v_p = \dfrac{c}{\sqrt{1-(f_c/f)^2}}$. The ratio v_p/c is: **[2 marks]**

(A) 0.6

(B) 0.8

(C) 1.667

(D) 1.25

Q46. A uniform plane wave propagates through a lossless dielectric of relative permittivity $\varepsilon_r = 4$ and relative permeability $\mu_r = 1$. Its phase velocity $v = \dfrac{c}{\sqrt{\varepsilon_r}}$ (with $c = 3 \times 10^8$ m/s) is: **[2 marks]**

(A) 3×10^8 m/s

(B) 0.75×10^8 m/s

(C) 1.5×10^8 m/s

(D) 6×10^8 m/s

Detailed Solutions

Q1.

Solution

Concept – open-circuit impedance (Z) parameters: With port 2 open, no current flows there ($I_2 = 0$), and the port-1 voltage is set entirely by the self-impedance Z_{11}: $V_1 = Z_{11}I_1 + Z_{12}I_2 = Z_{11}I_1$.

Step 1 – apply the open-circuit condition: $I_2 = 0$, so the $Z_{12}I_2$ term vanishes.

Step 2 – write the reduced relation: $V_1 = Z_{11}I_1$

Step 3 – substitute the data: $V_1 = 15 \times 2$

Step 4 – evaluate: $V_1 = 30$ V

Final Answer: $V_1 = 30 \text{ V} \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q1

Q2.

Solution

Concept – reciprocity in ABCD parameters: A reciprocal two-port satisfies the determinant condition $AD - BC = 1$.

Step 1 – write the condition: $AD - BC = 1$

Step 2 – substitute the known parameters: $2D - (3)(1) = 1$

Step 3 – simplify: $2D - 3 = 1$

Step 4 – solve for D: $2D = 1 + 3 = 4$

$$D = \frac{4}{2} = 2$$

Final Answer: $D = 2 \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q2

Q3.

Solution

Concept – resistive voltage divider: For two resistors in series across a source V, the voltage across R_2 is $V_2 = V\dfrac{R_2}{R_1 + R_2}$.

Step 1 – form the total series resistance: $R_1 + R_2 = 4 + 8 = 12\ \Omega$

Step 2 – form the divider ratio: $\dfrac{R_2}{R_1 + R_2} = \dfrac{8}{12} = \dfrac{2}{3}$

Step 3 – multiply by the source voltage: $V_2 = 12 \times \dfrac{2}{3}$

Step 4 – evaluate: $V_2 = 8$ V

Final Answer: $V_2 = 8 \text{ V} \Rightarrow \boxed{\text{A}}$

$\boxed{\text{Answer: (A)}}$ Go Back to Q3

Q4.

Solution

Concept – capacitive reactance: The reactance of a capacitor is $X_C = \dfrac{1}{\omega C}$.

Step 1 – convert the capacitance: $C = 2\ \mu\text{F} = 2 \times 10^{-6}$ F

Step 2 – form the product ωC: $\omega C = 5000 \times (2 \times 10^{-6})$

$\omega C = 1 \times 10^{-2} = 0.01$

Step 3 – take the reciprocal: $X_C = \dfrac{1}{0.01}$

Step 4 – evaluate: $X_C = 100\ \Omega$

Final Answer: $X_C = 100\ \Omega \Rightarrow \boxed{\text{A}}$

$\boxed{\text{Answer: (A)}}$ Go Back to Q4

Q5.

Solution

Concept – open-circuit voltage of a Norton source: When the terminals are open, all of the Norton current flows through the Norton resistance, so $V_{oc} = I_N R_N$.

Step 1 – list the data: $I_N = 4$ A, $R_N = 6\ \Omega$.

Step 2 – multiply: $V_{oc} = 4 \times 6$

Step 3 – evaluate: $V_{oc} = 24$ V

Step 4 – interpret: This $V_{oc} = 24$ V is exactly the Thevenin voltage of the equivalent source, confirming the source-transformation relation $V_{th} = I_N R_N$.

Final Answer: $V_{oc} = 24 \text{ V} \Rightarrow \boxed{\text{D}}$

$\boxed{\text{Answer: (D)}}$ Go Back to Q5

Q6.

Solution

Concept – resonant frequency of a parallel RLC circuit: The undamped natural frequency is $\omega_0 = \dfrac{1}{\sqrt{LC}}$, the same form as for the series circuit.

Step 1 – convert the values to base units: $L = 4 \text{ mH} = 4 \times 10^{-3}\text{ H}$

$C = 25\text{ nF} = 25 \times 10^{-9}\text{ F}$

Step 2 – form the product LC: $LC = (4 \times 10^{-3}) \times (25 \times 10^{-9})$

$LC = 100 \times 10^{-12} = 1 \times 10^{-10}$

Step 3 – take the square root: $\sqrt{LC} = \sqrt{1 \times 10^{-10}} = 1 \times 10^{-5}$

Step 4 – invert: $\omega_0 = \dfrac{1}{1 \times 10^{-5}}$

$\omega_0 = 100000\text{ rad/s}$

Final Answer: $\omega_0 = 100000\text{ rad/s} \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q6

Q7.

Solution

Concept – mass-action law for a p-type sample: At equilibrium $n\,p = n_i^2$. For a p-type sample the holes are the majority carriers, $p \approx N_a$, so the minority electrons are $n = \dfrac{n_i^2}{N_a}$.

Step 1 – square the intrinsic concentration: $n_i^2 = (1.5 \times 10^{10})^2$

$n_i^2 = 2.25 \times 10^{20}\text{ cm}^{-6}$

Step 2 – take the majority-carrier concentration: $p \approx N_a = 5 \times 10^{16}\text{ cm}^{-3}$

Step 3 – divide to get the electron concentration: $n = \dfrac{2.25 \times 10^{20}}{5 \times 10^{16}}$

Step 4 – combine the powers of ten: $n = 0.45 \times 10^{4} = 4.5 \times 10^{3}\text{ cm}^{-3}$

Final Answer: $n = 4.5 \times 10^{3}\text{ cm}^{-3} \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q7

Q8.

Solution

Concept – carrier concentration in intrinsic material: In an intrinsic (undoped) semiconductor every electron promoted to the conduction band leaves one hole, so $n = p = n_i$.

Step 1 – state the intrinsic condition: $n = p = n_i$

Step 2 – insert the standard silicon value: $n_i = 1.5 \times 10^{10}$ cm^{-3}

Step 3 – conclude: $n = 1.5 \times 10^{10}$ cm^{-3}

Why the other options are wrong:

- A, 0: a pure semiconductor is not an insulator; thermal generation gives a finite n_i.
- C, 2.25×10^{20}: is n_i^2 (units cm^{-6}), not n_i itself.
- D, 3×10^{10}: wrongly doubles n_i; the electron count equals n_i, not $2n_i$.

Final Answer: $n = 1.5 \times 10^{10}$ cm^{-3} $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q8

Q9.

Solution

Concept – Einstein relation: The diffusion coefficient and mobility are related by $D_n = \mu_n V_T$, where $V_T = kT/q$ is the thermal voltage.

Step 1 – list the data: $\mu_n = 1350$ cm^2/V · s, $V_T = 0.025$ V.

Step 2 – substitute into the Einstein relation: $D_n = 1350 \times 0.025$

Step 3 – evaluate: $D_n = 33.75$ cm^2/s

Step 4 – check the units: (cm^2/V·s) × (V) = cm^2/s, the correct unit for a diffusion coefficient.

Final Answer: $D_n = 33.75$ cm^2/s $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q9

Q10.

Solution

Concept – built-in potential of a p–n junction: The junction contact potential is $V_{bi} = V_T \ln \dfrac{N_a N_d}{n_i^2}$.

Step 1 – form the doping product: $N_a N_d = (10^{16})(10^{16}) = 10^{32}\ \text{cm}^{-6}$

Step 2 – square the intrinsic concentration: $n_i^2 = (1.5\times10^{10})^2 = 2.25\times10^{20}\ \text{cm}^{-6}$

Step 3 – form the ratio inside the logarithm: $\dfrac{N_a N_d}{n_i^2} = \dfrac{10^{32}}{2.25\times10^{20}} = 4.44\times10^{11}$

Step 4 – take the natural logarithm: $\ln(4.44\times10^{11}) \approx 26.8$

Step 5 – multiply by the thermal voltage: $V_{bi} = 0.025 \times 26.8$

Step 6 – evaluate: $V_{bi} \approx 0.67\ \text{V}$

Final Answer: $V_{bi} \approx 0.67\ \text{V} \Rightarrow \boxed{\text{C}}$

Answer: (C) **Go Back to Q10**

Q11.

Solution

Concept – drift velocity: Under an applied field the carriers acquire a drift velocity $v_d = \mu_n E$.

Step 1 – list the data: $\mu_n = 1000\ \text{cm}^2/\text{V}\cdot\text{s}$, $E = 50\ \text{V/cm}$.

Step 2 – substitute: $v_d = 1000 \times 50$

Step 3 – evaluate: $v_d = 50000 = 5\times10^4\ \text{cm/s}$

Step 4 – check the units: $(\text{cm}^2/\text{V}\cdot\text{s}) \times (\text{V/cm}) = \text{cm/s}$, a velocity.

Final Answer: $v_d = 5\times10^4\ \text{cm/s} \Rightarrow \boxed{\text{B}}$

Answer: (B) **Go Back to Q11**

Q12.

Solution

Concept – series combination of the Colpitts tank capacitors: The two feedback capacitors appear in series across the inductor, so the effective capacitance is $C = \dfrac{C_1 C_2}{C_1 + C_2}$.

Step 1 – form the product of the capacitors: $C_1 C_2 = 100 \times 100 = 10000\ \text{pF}^2$

Step 2 – form the sum of the capacitors: $C_1 + C_2 = 100 + 100 = 200\ \text{pF}$

Step 3 – divide: $C = \dfrac{10000}{200}$

Step 4 – evaluate: $C = 50$ pF

Step 5 – sanity check: For two equal series capacitors the result is half of one, and $100/2 = 50$ pF, which agrees.

Final Answer: $C = 50$ pF $\Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q12

Q13.

Solution

Concept – oscillation frequency of a Colpitts tank: $f_0 = \dfrac{1}{2\pi\sqrt{LC}}$, with C the effective series capacitance.

Step 1 – convert the values to base units: $L = 10\ \mu\text{H} = 10 \times 10^{-6}$ H

$C = 50\ \text{pF} = 50 \times 10^{-12}$ F

Step 2 – form the product LC: $LC = (10 \times 10^{-6}) \times (50 \times 10^{-12})$

$LC = 500 \times 10^{-18} = 5 \times 10^{-16}$

Step 3 – take the square root: $\sqrt{LC} = \sqrt{5 \times 10^{-16}} = 2.236 \times 10^{-8}$

Step 4 – form $2\pi\sqrt{LC}$: $2\pi\sqrt{LC} = 6.283 \times 2.236 \times 10^{-8}$

$= 1.405 \times 10^{-7}$

Step 5 – invert: $f_0 = \dfrac{1}{1.405 \times 10^{-7}}$

Step 6 – evaluate: $f_0 = 7.12 \times 10^6$ Hz ≈ 7.12 MHz

Final Answer: $f_0 \approx 7.12$ MHz $\Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q13

Q14.

Solution

Concept – oscillation frequency of a Wien-bridge oscillator: $f_0 = \dfrac{1}{2\pi RC}$.

Step 1 – form the product RC: $RC = (16 \times 10^3) \times (1 \times 10^{-9})$

$RC = 16 \times 10^{-6} = 1.6 \times 10^{-5}$ s

Step 2 – form $2\pi RC$: $2\pi RC = 6.283 \times (1.6 \times 10^{-5})$

$2\pi RC = 1.005 \times 10^{-4}$

Step 3 – invert: $f_0 = \dfrac{1}{1.005 \times 10^{-4}}$

Step 4 – evaluate: $f_0 = 9947\text{ Hz} \approx 9.95\text{ kHz}$

Final Answer: $f_0 \approx 9.95\text{ kHz} \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q14

Q15.

Solution

Concept – gain condition of a Wien-bridge oscillator: The lead-lag network attenuates by a factor of 3 at f_0, so the amplifier must supply a gain of 3. For a non-inverting stage $A = 1 + \dfrac{R_f}{R_1} = 3$, hence $\dfrac{R_f}{R_1} = 2$.

Step 1 – write the required ratio: $\dfrac{R_f}{R_1} = 2$

Step 2 – solve for R_f: $R_f = 2R_1$

Step 3 – substitute $R_1 = 10$ kΩ: $R_f = 2 \times 10\text{ k}\Omega$

Step 4 – evaluate: $R_f = 20\text{ k}\Omega$

Why the other options are wrong:

- B, 10 kΩ: gives gain 2, below the required 3; oscillation dies out.
- C, 30 kΩ: gives gain 4, above 3; the output would clip.
- D, 5 kΩ: gives gain 1.5, far too low.

Final Answer: $R_f = 20\text{ k}\Omega \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q15

Q16.

Solution

Concept – time constant of an op-amp integrator: The Miller integrator has time constant $\tau = RC$.

Step 1 – convert the values to base units: $R = 100\text{ k}\Omega = 1 \times 10^5\ \Omega$

$C = 0.1\ \mu\text{F} = 1 \times 10^{-7}\text{ F}$

Step 2 – multiply: $\tau = (1 \times 10^5) \times (1 \times 10^{-7})$

Step 3 – combine the powers of ten: $\tau = 1 \times 10^{-2}$

$\tau = 0.01\text{ s}$

Final Answer: $\tau = 0.01$ s $\Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q16

Q17.

Solution

Concept – Barkhausen phase condition: For oscillation the loop phase shift must total $360°$ (or $0°$). The inverting amplifier already provides $180°$, so the RC feedback network must supply the remaining $180°$.

Step 1 – account for the amplifier phase: Inverting amplifier phase $= 180°$.

Step 2 – find the phase the network must add: $360° - 180° = 180°$

Step 3 – check consistency with the section count: Three identical RC sections share the $180°$, about $60°$ each at f_0, which is realisable; a single section cannot reach $90°$, so three are used.

Final Answer: network phase shift $= 180° \Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q17

Q18.

Solution

Concept – voltage gain of a common-source stage: Neglecting r_o, the magnitude of the small-signal gain is $|A_v| = g_m R_D$.

Step 1 – convert the data: $g_m = 2$ mS $= 2 \times 10^{-3}$ S

$R_D = 10\ \text{k}\Omega = 1 \times 10^4\ \Omega$

Step 2 – multiply: $|A_v| = (2 \times 10^{-3}) \times (1 \times 10^4)$

Step 3 – combine the powers of ten: $|A_v| = 2 \times 10^1$

Step 4 – evaluate: $|A_v| = 20$

Final Answer: $|A_v| = 20 \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q18

Q19.

Solution

Concept – state of a modulo-16 up-counter: A 4-bit up-counter returns to its starting state every 16 pulses, so after N pulses the state is $N \bmod 16$ expressed in binary.

Step 1 – reduce the pulse count modulo 16: $20 \bmod 16 = 20 - 16 = 4$

Step 2 – convert 4 to 4-bit binary: $4 = (0100)_2$

Step 3 – state the counter output: The counter holds 0100.

Final Answer: state $= 0100 \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q19

Q20.

Solution

Concept – states of a ring counter: In an n-bit ring counter a single 1 is circulated, so the sequence repeats after n shifts, giving exactly n distinct states.

Step 1 – identify the register length: $n = 4$ flip-flops.

Step 2 – count the distinct states: The 1 occupies each of the 4 positions once per cycle, giving 4 states.

Step 3 – contrast with the maximum count: A 4-bit register can hold $2^4 = 16$ patterns, but the one-hot ring counter uses only 4 of them.

Why the other options are wrong:

- A, 16: is 2^4, the count of a full binary counter, not a ring counter.
- C, 8: is $2n$, the count of a Johnson (twisted-ring) counter.
- D, 2: is far too few for a 4-position ring.

Final Answer: 4 distinct states $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q20

Q21.

Solution

Concept – states of a Johnson counter: An n-stage Johnson (twisted-ring) counter feeds back the complement of the last stage, doubling the sequence length to $2n$ distinct states.

Step 1 – write the requirement: $2n = 10$

Step 2 – solve for n: $n = \frac{10}{2}$

Step 3 – evaluate: $n = 5$

Step 4 – verify: Five flip-flops give $2 \times 5 = 10$ states, exactly as required.

Why the other options are wrong:

- B, 10: treats the counter as a ring counter (n states), ignoring the complement feedback.
- C, 4: gives only 8 states.
- D, 3: gives only 6 states.

Final Answer: $n = 5 \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q21

Q22.

Solution

Concept – frequency division by a binary counter: Each flip-flop of a ripple counter halves the frequency, so the k-th stage output is the input clock divided by 2^k. For a mod-16 counter the MSB Q_3 divides by $2^4 = 16$.

Step 1 – identify the division factor at Q_3: Q_3 divides by $2^4 = 16$.

Step 2 – divide the clock frequency: $f_{Q_3} = \frac{32 \text{ MHz}}{16}$

Step 3 – evaluate: $f_{Q_3} = 2 \text{ MHz}$

Final Answer: $f_{Q_3} = 2 \text{ MHz} \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q22

Q23.

Solution

Concept – grouping the corner cells of a Karnaugh map: The four corner cells of a 4-variable map are logically adjacent and form one group of four; the literals constant across them give the product term.

Step 1 – locate the minterms: $m0 = 0000$, $m2 = 0010$, $m8 = 1000$, $m10 = 1010$, which are the four corners of the map.

Step 2 – find the constant bits across the group: In all four, $B = 0$ (second bit) and $D = 0$ (fourth bit); A and C both change.

Step 3 – keep only the constant literals: Constant literals are $\bar{B}$ and $\bar{D}$.

Step 4 – write the term: $F = \bar{B}\,\bar{D}$

Step 5 – verify against the minterms: $\bar{B}\,\bar{D}$ selects all cells with $B = 0$ and $D = 0$, namely $0000, 0010, 1000, 1010 = m0, m2, m8, m10$, exactly the given ones.

Final Answer: $F = \bar{B}\,\bar{D} \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q23

Q24.

Solution

Concept – logical right shift: A logical right shift moves every bit one place toward the LSB and feeds a 0 into the vacated MSB position; the LSB shifted out is discarded.

Step 1 – start with the loaded pattern: 1011

Step 2 – shift right by one (fill 0 at the left): $1011 \to 0101$

Step 3 – shift right a second time (fill 0 at the left): $0101 \to 0010$

Step 4 – read the final register contents: 0010

Final Answer: contents $= 0010 \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q24

Q25.

Solution

Concept – z-transform of a causal exponential and its value at $z = 1$: For $x[n] = a^n u[n]$, $X(z) = \dfrac{1}{1 - az^{-1}}$ for $|z| > a$. Setting $z = 1$ gives $\dfrac{1}{1-a}$, which is also the DTFT at $\omega = 0$ and the sum of all samples.

Step 1 – identify the ratio: $a = 0.4$.

Step 2 – substitute $z = 1$**:** $X(1) = \dfrac{1}{1 - 0.4(1)^{-1}} = \dfrac{1}{1 - 0.4}$

Step 3 – form the denominator: $1 - 0.4 = 0.6$

Step 4 – take the reciprocal: $X(1) = \dfrac{1}{0.6}$

Step 5 – evaluate: $X(1) = 1.667$

Final Answer: $X(1) = 1.667 \Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q25

Q26.

Solution

Concept – pole of a first-order discrete system: A pole is a value of z that makes the denominator of $H(z)$ zero.

Step 1 – set the denominator to zero: $z - 0.8 = 0$

Step 2 – solve for z**:** $z = 0.8$

Step 3 – check stability from the plot: $|z| = 0.8 < 1$, so the pole lies inside the unit circle and the causal system is stable.

Why the other options are wrong:

- A, 0: is the location of the *zero* (numerator $z = 0$), not the pole.
- B, -0.8: has the wrong sign; it would give a denominator $z + 0.8$.
- C, 1: lies on the unit circle and would make the system marginally stable, not the case here.

Final Answer: pole at $z = 0.8 \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q26

Q27.

Solution

Concept – DTFT of the unit impulse: The DTFT is $X(e^{j\omega}) = \sum_n x[n]e^{-j\omega n}$. For $x[n] = \delta[n]$ only the $n = 0$ term survives.

Step 1 – write the defining sum: $X(e^{j\omega}) = \displaystyle\sum_{n=-\infty}^{\infty} \delta[n]\, e^{-j\omega n}$

Step 2 – apply the sifting property: Only $n = 0$ contributes, giving $e^{-j\omega\cdot 0}$.

Step 3 – evaluate the surviving term: $e^{-j\omega \cdot 0} = e^0 = 1$

Step 4 – interpret: The impulse contains all frequencies with equal, constant magnitude 1; its transform is flat over ω.

Why the other options are wrong:

- C, $e^{-j\omega}$: is the DTFT of the *delayed* impulse $\delta[n-1]$.
- D, $2\pi\delta(\omega)$: is the DTFT of the constant sequence $x[n] = 1$, not of $\delta[n]$.

Final Answer: $X(e^{j\omega}) = 1 \Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q27

Q28.

Solution

Concept – time-shift property of the DTFT: Delaying a sequence by n_0 samples multiplies its DTFT by $e^{-j\omega n_0}$.

Step 1 – identify the delay: $y[n] = x[n-3]$, so $n_0 = 3$.

Step 2 – write the property: $Y(e^{j\omega}) = e^{-j\omega n_0} X(e^{j\omega})$

Step 3 – substitute $n_0 = 3$**:** $Y(e^{j\omega}) = e^{-j3\omega} X(e^{j\omega})$

Step 4 – state the multiplying factor: $e^{-j3\omega}$

Why the other options are wrong:

- B, $e^{+j3\omega}$: corresponds to an *advance* $x[n+3]$, not a delay.
- C, $3e^{-j\omega}$: wrongly scales the magnitude; a pure delay only changes phase.
- D, $e^{-j\omega/3}$: misplaces the factor of 3 into the exponent's denominator.

Final Answer: factor $= e^{-j3\omega} \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q28

Q29.

Solution

Concept – length of a linear convolution: If two sequences have lengths L_1 and L_2, their linear convolution has length $L_1 + L_2 - 1$.

Step 1 – list the lengths: $L_1 = 6$, $L_2 = 8$.

Step 2 – add the lengths: $L_1 + L_2 = 6 + 8 = 14$

Step 3 – subtract one: $14 - 1 = 13$

Final Answer: length $= 13 \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q29

Q30.

Solution

Concept – periodicity of the DTFT: The DTFT uses the complex exponential $e^{-j\omega n}$, which is periodic in ω with period 2π for every integer n; hence $X(e^{j\omega})$ repeats every 2π.

Step 1 – examine the kernel: $e^{-j(\omega+2\pi)n} = e^{-j\omega n}\, e^{-j2\pi n}$

Step 2 – simplify the extra factor: $e^{-j2\pi n} = 1$ for all integer n.

Step 3 – conclude: $X(e^{j(\omega+2\pi)}) = X(e^{j\omega})$, so the fundamental period is 2π.

Final Answer: period $= 2\pi \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q30

Q31.

Solution

Concept – gain-crossover frequency: The gain crossover ω_{gc} is where the open-loop magnitude falls to unity, $|G(j\omega_{gc})| = 1$.

Step 1 – write the magnitude of $G(j\omega)$: $|G(j\omega)| = \dfrac{100}{\omega}$

Step 2 – set the magnitude to 1: $\dfrac{100}{\omega} = 1$

Step 3 – solve for ω: $\omega = 100$

Step 4 – state the result: $\omega_{gc} = 100$ rad/s

Final Answer: $\omega_{gc} = 100$ rad/s $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q31

Q32.

Solution

Concept – phase margin: $\text{PM} = 180^\circ + \angle G(j\omega_{gc})$, the extra phase lag that would bring the system to the verge of instability.

Step 1 – read the phase of the integrator: $G(s) = K/s$ gives $\angle G(j\omega) = -90^\circ$ at every frequency, including ω_{gc}.

Step 2 – substitute into the phase-margin formula: $\text{PM} = 180^\circ + (-90^\circ)$

Step 3 – evaluate: $\text{PM} = 90^\circ$

Step 4 – interpret: A 90° margin is large and positive, so the single-integrator loop is very stable.

Final Answer: $\text{PM} = 90^\circ \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q32

Q33.

Solution

Concept – gain in decibels: A dimensionless voltage gain A in decibels is $A_{dB} = 20\log_{10} A$.

Step 1 – substitute the gain: $A_{dB} = 20\log_{10}(100)$

Step 2 – evaluate the logarithm: $\log_{10}(100) = \log_{10}(10^2) = 2$

Step 3 – multiply: $A_{dB} = 20 \times 2$

Step 4 – evaluate: $A_{dB} = 40$ dB

Why the other options are wrong:

- B, 20 dB: would correspond to a gain of 10, not 100.
- C, 100 dB: wrongly reads the number 100 as decibels directly.
- D, 10 dB: corresponds to a power ratio of 10 via $10\log_{10}$, the wrong formula for a voltage gain.

Final Answer: $A_{dB} = 40$ dB $\Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q33

Q34.

Solution

Concept – gain margin: $\text{GM} = -20\log_{10}|G(j\omega_{pc})|$, evaluated at the phase-crossover frequency where the phase is -180°.

Step 1 – read the magnitude at phase crossover: $|G(j\omega_{pc})| = 0.5$

Step 2 – take the base-10 logarithm: $\log_{10}(0.5) = -0.301$

Step 3 – multiply by -20: $\text{GM} = -20 \times (-0.301)$

Step 4 – evaluate: $\text{GM} = 6.02$ dB

Step 5 – interpret: A positive gain margin of about 6 dB means the loop gain can rise by roughly a factor of 2 before instability.

Final Answer: $\text{GM} = 6.02$ dB $\Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q34

Q35.

Solution

Concept – corner frequency of a first-order factor: For $\dfrac{1}{1 + s/\omega_c} = \dfrac{1}{1 + \tau s}$ the corner (break) frequency is $\omega_c = \dfrac{1}{\tau}$.

Step 1 – identify the time constant: Comparing $1 + 0.02s$ with $1 + \tau s$ gives $\tau = 0.02$ s.

Step 2 – take the reciprocal: $\omega_c = \dfrac{1}{0.02}$

Step 3 – evaluate: $\omega_c = 50$ rad/s

Step 4 – interpret: At $\omega_c = 50$ rad/s the magnitude has dropped by 3 dB and the slope turns to -20 dB/decade.

Final Answer: $\omega_c = 50$ rad/s $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q35

Q36.

Solution

Concept – maximum SNR from a matched filter: For a signal of energy E in white noise of two-sided PSD $N_0/2$, the matched filter yields a peak output SNR of $\dfrac{2E}{N_0}$, independent of the pulse shape.

Step 1 – list the data: $E = 8 \times 10^{-6}$ J, $N_0 = 2 \times 10^{-6}$ W/Hz.

Step 2 – form the numerator $2E$: $2E = 2 \times (8 \times 10^{-6}) = 16 \times 10^{-6}$

Step 3 – divide by N_0: $\text{SNR} = \dfrac{16 \times 10^{-6}}{2 \times 10^{-6}}$

Step 4 – evaluate: $\text{SNR} = 8$

Final Answer: $\text{SNR} = 8 \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q36

Q37.

Solution

Concept – bit-error rate of coherent BPSK: $P_b = Q\left(\sqrt{2E_b/N_0}\right)$, where $Q(\cdot)$ is the Gaussian tail probability.

Step 1 – read the argument of the Q-function: $\sqrt{2E_b/N_0} = 2$

Step 2 – substitute into the BER expression: $P_b = Q(2)$

Step 3 – use the given value: $Q(2) = 0.0228$

Step 4 – state the result: $P_b = 0.0228$

Why the other options are wrong:

- A, 0.5: is $Q(0)$, i.e. zero SNR, not this case.
- B, 0.159: is $Q(1)$, a smaller argument.
- C, 0.32: is roughly $Q(0.5)$, again a smaller argument.

Final Answer: $P_b = 0.0228 \Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q37

Q38.

Solution

Concept – peak instant of a matched-filter output: The matched filter output is the autocorrelation of the pulse, which peaks at the instant $t = T$ when the received pulse and the time-reversed template fully overlap.

Step 1 – write the matched filter response: $h(t) = s(T - t)$, matched to a pulse of duration T.

Step 2 – locate the peak of the output: Full overlap of $s(t)$ and $h(t)$ occurs at $t = T$.

Step 3 – substitute the pulse duration: $T = 1$ ms, so the output peaks at $t = 1$ ms.

Why the other options are wrong:

- A, 0 ms: is the start of overlap, where the output is still zero.
- B, 0.5 ms: is only partial overlap, a sub-maximal value.
- C, 2 ms: is $2T$, by which time the pulse has fully passed and the output has returned to zero.

Final Answer: peak at $t = 1$ ms $\Rightarrow \boxed{\text{D}}$

Answer: (D) Go Back to Q38

Q39.

Solution

Concept – converting a power ratio to decibels: A power ratio r in decibels is $10\log_{10} r$.

Step 1 – substitute the ratio: $(E_b/N_0)_{dB} = 10\log_{10}(10)$

Step 2 – evaluate the logarithm: $\log_{10}(10) = 1$

Step 3 – multiply: $(E_b/N_0)_{dB} = 10 \times 1$

Step 4 – evaluate: $(E_b/N_0)_{dB} = 10$ dB

Why the other options are wrong:

- A, 20 dB: uses the voltage formula $20\log_{10}$, wrong for a power ratio.
- C, 3 dB: corresponds to a ratio of 2, not 10.
- D, 100 dB: misreads the number 10 as an exponent.

Final Answer: 10 dB $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q39

Q40.

Solution

Concept – decision-statistic argument for BPSK: The matched-filter decision variable scales with $\sqrt{2E_b/N_0}$.

Step 1 – list the data: $E_b = 10^{-5}$ J, $N_0 = 2 \times 10^{-6}$ W/Hz.

Step 2 – form the numerator $2E_b$: $2E_b = 2 \times 10^{-5} = 2 \times 10^{-5}$

Step 3 – form the ratio $2E_b/N_0$: $\dfrac{2 \times 10^{-5}}{2 \times 10^{-6}} = 10$

Step 4 – take the square root: $\sqrt{10} = 3.16$

Final Answer: $\sqrt{2E_b/N_0} = 3.16 \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q40

Q41.

Solution

Concept – Nyquist minimum bandwidth: The theoretical minimum bandwidth for ISI-free transmission of R_s symbols/s is $B_{\min} = \dfrac{R_s}{2}$.

Step 1 – list the symbol rate: $R_s = 2000$ symbols/s.

Step 2 – halve it: $B_{\min} = \dfrac{2000}{2}$

Step 3 – evaluate: $B_{\min} = 1000$ Hz

Why the other options are wrong:

- B, 2000 Hz: equals R_s; it forgets the factor of one-half.
- C, 4000 Hz: is $2R_s$, twice too large.
- D, 500 Hz: is $R_s/4$, below the Nyquist limit and would force ISI.

Final Answer: $B_{\min} = 1000$ Hz $\Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q41

Q42.

Solution

Concept – cut-off frequency of the dominant waveguide mode: For the TE_{10} mode of an air-filled rectangular waveguide, $f_c = \dfrac{c}{2a}$.

Step 1 – convert the broad-wall dimension: $a = 3$ cm $= 0.03$ m

Step 2 – form the denominator: $2a = 2 \times 0.03 = 0.06$ m

Step 3 – divide the speed of light by $2a$**:** $f_c = \dfrac{3 \times 10^8}{0.06}$

Step 4 – evaluate: $f_c = 5 \times 10^9$ Hz $= 5$ GHz

Final Answer: $f_c = 5$ GHz $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q42

Q43.

Solution

Concept – cut-off frequency of a general TE_{mn} **mode:** $f_{c,mn} = \dfrac{c}{2}\sqrt{\left(\dfrac{m}{a}\right)^2 + \left(\dfrac{n}{b}\right)^2}$. For TE_{20}, $m = 2$ and $n = 0$.

Step 1 – insert $m = 2$, $n = 0$: $f_{c,20} = \frac{c}{2}\sqrt{\left(\frac{2}{a}\right)^2 + 0}$

Step 2 – simplify the square root: $f_{c,20} = \frac{c}{2} \cdot \frac{2}{a} = \frac{c}{a}$

Step 3 – convert the dimension: $a = 2$ cm $= 0.02$ m

Step 4 – substitute: $f_{c,20} = \frac{3 \times 10^8}{0.02}$

Step 5 – evaluate: $f_{c,20} = 1.5 \times 10^{10}$ Hz $= 15$ GHz

Final Answer: $f_{c,20} = 15$ GHz $\Rightarrow \boxed{\text{B}}$

Answer: (B) Go Back to Q43

Q44.

Solution

Concept – lowest TM mode in a rectangular waveguide: A TM_{mn} mode has a longitudinal E_z whose field pattern vanishes if either index is zero; hence a propagating TM mode needs $m \geq 1$ and $n \geq 1$.

Step 1 – apply the index requirement: Both m and n must be at least 1.

Step 2 – pick the smallest valid pair: $(m, n) = (1, 1)$, giving the TM_{11} mode.

Why the other options are wrong:

- B, TM_{10} and C, TM_{01}: have a zero index, so $E_z \equiv 0$ and the TM field cannot exist.
- D, TM_{00}: has both indices zero and carries no field at all.

Final Answer: lowest TM mode $= \text{TM}_{11} \Rightarrow \boxed{\text{A}}$

Answer: (A) Go Back to Q44

Q45.

Solution

Concept – phase velocity in a waveguide: Above cut-off the phase velocity exceeds c: $v_p = \frac{c}{\sqrt{1 - (f_c/f)^2}}$.

Step 1 – form the frequency ratio: $\frac{f_c}{f} = \frac{f_c}{1.25 f_c} = \frac{1}{1.25} = 0.8$

Step 2 – square the ratio: $(0.8)^2 = 0.64$

Step 3 – subtract from one: $1 - 0.64 = 0.36$

Step 4 – take the square root: $\sqrt{0.36} = 0.6$

Step 5 – invert to get v_p/c: $\frac{v_p}{c} = \frac{1}{0.6} = 1.667$

Step 6 – interpret: $v_p = 1.667c > c$, as expected for a waveguide; the energy (group) velocity stays below c.

Final Answer: $v_p/c = 1.667 \Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q45

Q46.

Solution

Concept – phase velocity in a lossless dielectric: For a non-magnetic dielectric ($\mu_r = 1$) the wave slows to $v = \frac{c}{\sqrt{\varepsilon_r}}$.

Step 1 – take the square root of the permittivity: $\sqrt{\varepsilon_r} = \sqrt{4} = 2$

Step 2 – divide the speed of light: $v = \frac{3 \times 10^8}{2}$

Step 3 – evaluate: $v = 1.5 \times 10^8$ m/s

Why the other options are wrong:

- A, 3×10^8: is the free-space value; it ignores the dielectric.
- B, 0.75×10^8: wrongly divides by $\varepsilon_r = 4$ instead of $\sqrt{\varepsilon_r}$.
- D, 6×10^8: exceeds c, which is unphysical for a passive dielectric.

Final Answer: $v = 1.5 \times 10^8$ m/s $\Rightarrow \boxed{\text{C}}$

Answer: (C) Go Back to Q46

Answer Key

Q	Ans	Q	Ans	Q	Ans	Q	Ans	Q	Ans
1	C	2	D	3	A	4	A	5	D
6	D	7	A	8	B	9	B	10	C
11	B	12	C	13	D	14	C	15	A
16	A	17	B	18	D	19	C	20	B
21	A	22	D	23	A	24	D	25	B
26	D	27	B	28	A	29	A	30	C
31	B	32	D	33	A	34	C	35	B
36	C	37	D	38	D	39	B	40	C
41	A	42	B	43	B	44	A	45	C
46	C								