PRINCIPLES AND PRACTICE OF ENGINEERING ELECTRICAL AND COMPUTER ENGINEERING BREADTH MORNING SAMPLE TEST SOLUTIONS

Solutions (and Test) by

Kenneth L. Kaiser, P.E., Ph.D.

Professor ECE Department
Kettering University
Flint, MI

Disclaimer

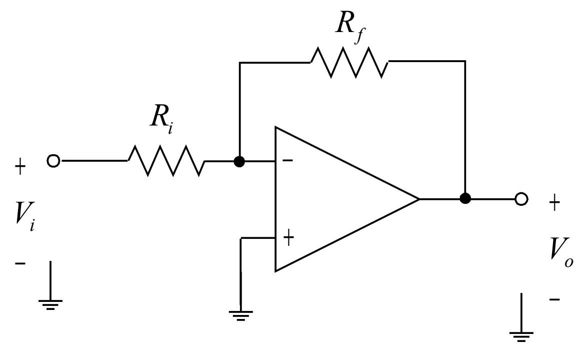
Reasonable efforts have been made to publish reliable solutions for the sample test, but the author cannot assume responsibility for the validity of all solutions and for the consequence of their use.

101. Answer is (C)

The standard gain equation for an inverting ideal amplifier is

$$A_v = -\frac{R_f}{R_i}$$

where R_f is the feedback resistance and R_i is the input resistance as shown in the following figure.

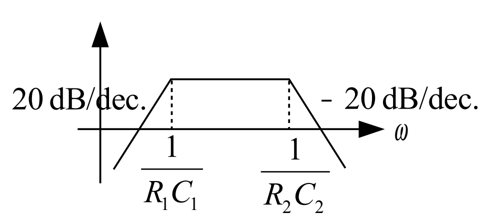


This expression can also be used for sinusoidal steady-state conditions when the feedback and input contain L's, C's, and R's

$$A_v = -\frac{\dfrac{1}{j\omega C_2}\Big\| R_2}{\dfrac{1}{j\omega C_1} + R_1} = -\frac{\dfrac{\dfrac{1}{j\omega C_2}R_2}{\dfrac{1}{j\omega C_2} + R_2}}{\dfrac{1}{j\omega C_1} + R_1} = -\frac{\dfrac{R_2}{1 + j\omega R_2 C_2}}{\dfrac{1}{j\omega C_1} + R_1} = -\frac{\dfrac{R_2}{1 + j\omega R_2 C_2}}{\dfrac{1}{j\omega C_1} + R_1}\frac{j\omega C_1}{j\omega C_1}$$

$$= -\frac{\dfrac{j\omega C_1 R_2}{1 + j\omega R_2 C_2}}{1 + j\omega C_1 R_1} = -\frac{j\omega C_1 R_2}{\left(1 + j\omega R_2 C_2\right)\left(1 + j\omega C_1 R_1\right)} = -\frac{j\omega C_1 R_2}{\left(1 + j\dfrac{\omega}{\dfrac{1}{R_2 C_2}}\right)\left(1 + j\dfrac{\omega}{\dfrac{1}{R_1 C_1}}\right)}$$

A quick partial Bode magnitude plot of this expression, which was written in standard form for Bode plotting, follows (assuming $1/\left(R_1 C_1\right) < 1/\left(R_2 C_2\right)$. This is clearly a band-pass filter with two cutoff or break frequencies. The answer is (C).



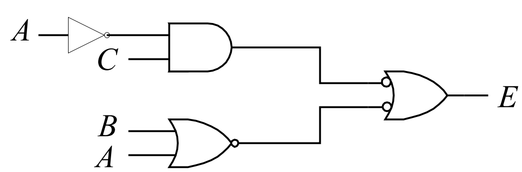
Is there a faster method of determining this solution? Yes. Recall that at low frequencies (LF) an ideal capacitor appears like an open (or nearly so), and at high frequencies (HF) an ideal capacitor appears like a short (or nearly so). Thus,

$$A_{vLF} = -\frac{R_f}{R_i} \equiv -\frac{R_2 \| \infty}{R_1 + \infty} = -\frac{R_2}{\infty} = 0, \quad A_{vHF} = -\frac{R_f}{R_i} \equiv -\frac{R_2 \| 0}{R_1 + 0} = -\frac{0}{R_1} = 0$$

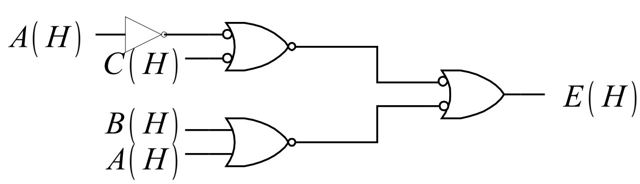
the gain at both low and high frequencies is zero. The only reasonable conclusion is that at some frequency between low and high frequencies there is some gain. Hence, this must be a band-pass filter (BPF). This quick method can be applied to many other op-amp circuits operating in sinusoidal steady state.

102. Answer is (B)

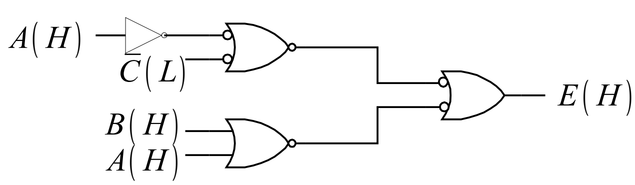
When working problems of this type, I always redraw the circuit so that the assertion levels between gates and inputs and outputs are matched. This helps eliminate the need to apply DeMorgan's law (directly). I begin with the output and work toward the input(s). Since the output is high asserting, I first draw an equivalent gate for the output-most gate, which is also high asserting:



Continuing toward the inputs, equivalent gates are substituted into the circuit when required so that "bubbles" meet "bubbles" and "nonbubbles" meet "nonbubbles:"



The high assertion levels are indicated in this last figure. In order to match the lower bubble input of the upper OR gate, rewrite $C(H) = \bar{C}(L)$:



Now that all inputs and outputs (including those of the gates) are matched, the logic expression for the output is written down by inspection, which is equal to (B):

$$E = \left(A + \bar{C}\right) + \left(B + A\right) = A + B + \bar{C}$$

103. Answer is (B)

Stability problems of this type are easy to solve as long as the Routh Test is understood. For negative feedback as shown, the closed-loop transfer function, the ratio of the output to input, is

$$G_{loop}(s) = \frac{C(s)}{R(s)} = \frac{G(s)}{1+G(s)H(s)} = \frac{\dfrac{K(s+3)}{s+5}}{1+\dfrac{K(s+3)}{s+5}\dfrac{1}{s^2}}$$

Where $G(s)$ is the transfer function of the forward path and $H(s)$ is the transfer function of the feedback path. (Carefully Note: If the feedback was positive, then the denominator of the closed-loop transfer function would be $1-G(s)H(s)$.) The Routh Test requires the use of the characteristic equation of the closed-loop transfer function. This is obtained directly from the denominator of the transfer function after some simplification:

$$G_{loop}(s) = \frac{K(s+3)s^2}{(s+5)s^2+K(s+3)} = \frac{K(s+3)s^2}{s^3+5s^2+Ks+3K}$$

The characteristic equation is $s^3+5s^2+Ks+3K$. From the coefficients of this characteristic equation, the Routh Test can be applied:

$$1K53K\frac{5K-3K}{5}03K$$

The signs of the first column elements should be the same for stability. Therefore, $5K-3K>0$ and $3K>0$, and the answer is (B).

104. Answer is (C)

(A) is not likely to be correct since capacitance is added across inductive loads (such as large HP motors) to reduce the net reactance of the plant. This is often done to reduce the current on the power lines leading to the plant. The power company is concerned about this current since it is loss of real power, due to as I^2R losses, that they are not measuring with a "real" power meter at the plant. (B) is not likely correct since capacitance is often added to reduce the overall reactance and hence reduce the reactive power in VARS. (D) is not likely correct since the parasitic capacitance of a motor is not usually considered a source of power loss (ideal capacitors do not absorb power). The answer is obviously (C). The power factor (PF) is defined as $\cos\theta$ where θ is the phase angle between the voltage across and current into the plant (or any load). By reducing the net reactance of the load, by canceling inductive reactance with capacitive reactance, the load looks more resistive and this phase angle is closer to zero degrees. Thus, the power factor is brought closer to one and, in the language of electrical engineers, the PF is improved.

105. Answer is (A)

When teaching electronics, I like to ask questions involving ideal op amps and superposition. In this circuit, imagine turning off v_2 (i.e., replacing this voltage source with a short circuit). In this case, the circuit reduces to an inverting amplifier (R_3 and R_4

should not affect the circuit much since zero current passes into the inputs of the ideal op amp) with a gain of

$$A_{v1} = -\frac{R_2}{R_1} \quad \Rightarrow \quad v_{o1} = -\frac{R_2}{R_1} v_1$$

Next, turning off v_1, the circuit reduces to a noninverting amplifier. For a standard noninverting amplifier the gain expression is

$$A_v = \left(1 + \frac{R_2}{R_1}\right)$$

However, the voltage into the + input of the amplifier is not v_2 but $v_+ = \frac{R_4}{R_3 + R_4} v_2$, which is obtained using voltage division. Therefore, the output voltage due to v_2 is

$$v_{o2} = \left(1 + \frac{R_2}{R_1}\right)\frac{R_4}{R_3 + R_4} v_2$$

Using superposition, the total output voltage is

$$v_o = v_{o1} + v_{o2} = -\frac{R_2}{R_1} v_1 + \left(1 + \frac{R_2}{R_1}\right)\frac{R_4}{R_3 + R_4} v_2$$

106. Answer is (A)

GFCI's (or just GFI's) are important devices that are often used in outlets near water and in garages, basements, and outdoors. (See the electric code for detail requirements.) Briefly, GFCI's work by detecting an imbalance in currents between the hot and neutral conductors. If these currents are not the same, then current must be returning back to the power source via some other route (e.g., the user of the product connected to the outlet). GFCI's, however, do not protect an individual in all possible electrical-hazard scenarios (see my fabulous reference book *Electromagnetic Compatibility Handbook* for more information on this and many related concepts). Circuit breakers, as stated in answer (C), are typically used to protect equipment and help prevent conductors from overheating. GFCI's (for typical residential outlets) trip around 2 mA, which is much less than the typical tripping current of 10 A or more for circuit breakers. The "let-go" current is between 6-9 mA at 60 Hz and is a function of many factors. Thus, (D) is also correct. Circuit breakers are placed in series with the hot conductor and open or break when the current in this conductor exceeds some level. Circuit breakers do not sense an imbalance in currents between the hot and neutral conductors as GFCI's do. Hence, (B) is also correct. (A) is not likely correct since circuit breakers often open or trip slowly not "too quickly."

107. Answer is (B)

Ideal resistors absorb real power and a wattmeter measures real power in watts (W). If the signal voltage is given as $v_s(t)$, then using voltage division for the two resistors in series, the voltage across the 2 Ω resistor is

$$v_2(t) = \frac{2}{2+5} v_s(t)$$

If the source voltage was dc, then the power absorbed by this resistor would be equal to

$$P_2 = \frac{\left(\frac{2}{2+5}V_{sdc}\right)^2}{2}$$

If the source voltage was sinusoidal (or ac) with zero dc offset and an amplitude of A, the time-average power absorbed by this resistor would be

$$P_2 = \frac{1}{2}\frac{\left(\frac{2}{2+5}A\right)^2}{2} = \frac{\left(\frac{2}{2+5}V_{rms}\right)^2}{2}$$

$$\text{where } V_{rms} = \frac{A}{\sqrt{2}} \text{ for sine waves with zero dc offset}$$

where V_{rms} is the rms value of the sinusoidal waveform. The rms value for other periodic waveforms with a period of T can be obtained from the general expression

$$v_{rms} = \sqrt{\frac{1}{T}\int_0^T v^2(t)\,dt}$$

Unless the waveform is relatively simple, it is probably reasonable to assume that on multiple-choice test of this type, insufficient time is available to apply this expression to obtain the rms value of a periodic waveform. The rms value of the triangular waveform of even-symmetry and dc offset given in this problem is given in my (and many other) handbooks as $12/\sqrt{3}$. Therefore, the answer to this problem is (B):

$$P_2 = \frac{\left(\frac{2}{2+5}V_{rms}\right)^2}{2} = \frac{\left(\frac{2}{2+5}\frac{12}{\sqrt{3}}\right)^2}{2} = 2\text{ W}$$

There is also another way of solving this problem that does not require the use of the exact expression for the rms value of this triangular waveform. By inspection, the average value of the waveform is 6 V (add up the total area and divide by the period). Therefore, a fast estimate for the power is

$$P_2 \approx \frac{\left(\frac{2}{2+5}6\right)^2}{2} = 1.5\text{ W}$$

The closest answer is again (B). The rms value of a waveform is always greater than or equal to its average value:

$$v_{rms} = \sqrt{\frac{1}{T}\int_0^T v^2(t)\,dt} \geq v_{avg} = \frac{1}{T}\int_0^T v(t)\,dt$$

Because the answers are well spaced, numerically, this method works. When taking multiple-choice tests, always look at all of the answers (and their signs if appropriate).

108. Answer is (D)

The original cost of \$55K and the future salvage value of \$20K should both be converted to an annual value at a 6% interest rate:

$$A = 55\text{K} \times (A/P, 6\%, 5) - 20\text{K} \times (A/F, 6\%, 5) = 55\text{K} \times 0.2374 - 20\text{K} \times 0.1774 = 9.5\text{K}$$

An interest table was used. The closest answer is (D). Note that the answer is not $(55\text{K} - 20\text{K})/5 = 7\text{K}$. A "good" multiple-choice test will have many of the "wrong-way" solutions as choices, so always carefully think through the problem and double check.

109. Answer is (C)

This problem includes both a monthly and an annual cost. The monthly interest rate can be obtained from the annual interest rate by using the equation

$$i = (1 + \phi)^k - 1$$

where i is the annual interest rate (as a fraction), ϕ is the corresponding monthly interest rate as a fraction, and k is the number of months in a year. Thus,

$$0.06 = (1 + \phi)^{12} - 1 \quad \Rightarrow \quad 1.06 = (1 + \phi)^{12} \quad \Rightarrow \quad 1.06^{\frac{1}{12}} = 1 + \phi \quad \Rightarrow \quad \phi = 0.0049$$

The monthly interest rate is 0.49%, which is close to 6%/12 = 0.5%. In some problems, this small difference may be important. The present value of the mill can now be obtained by adding the mill's initial cost to the present value of the monthly and annual maintenance costs:

$$\begin{aligned} P &= 20{,}000 + 300 \times (P/A, 0.49\%, 7 \times 12) + 1{,}500 \times (P/A, 6\%, 7) \\ &= 20{,}000 + 300 \times (P/A, 0.49\%, 7 \times 12) + 1{,}500 \times (P/A, 6\%, 7) \end{aligned}$$

Note the number of compounding periods for the monthly rate is not 12 but 84. My table did not have 0.49% interest rate or 84 compounding periods. So the actual formula for P/A will be used:

$$(P/A, i, n) = \frac{(1 + i)^n - 1}{i(1 + i)^n}$$

$$(P/A, 0.0049, 84) = \frac{(1 + 0.0049)^{84} - 1}{0.0049(1 + 0.0049)^{84}} = 68.72$$

If the 0.5% table is used with $n = 85$, the result is 69.11, which is close to that obtained. The answer is (C):

$$P = 20{,}000 + 300 \times 68.72 + 1{,}500 \times 5.582 = \$49{,}000$$

110. Answer is (B)

The synchronous speed of an induction motor with p poles operating at a frequency of f is

$$n_s = \frac{120 f}{p} = \frac{120 \times 50}{4} = 1{,}500 \text{ rpm}$$

The slip, as a fraction, is defined as

$$s = \frac{n_s - n}{n_s}$$

where n is the actual speed of the motor. Solving for n,

$$n = n_s - n_s s = n_s(1 - s) = 1{,}500(1 - 0.03) = 1{,}460 \text{ rpm}$$

The full-load rotor speed is less than the synchronous speed. Although the rotor field rotates at the synchronous speed, the rotor itself does not.

111. Answer is (D)

This is a standard K-map problem that should have been covered in all introductory courses in digital systems. The K-map is used to help organize the relationship between the inputs and the output and find a minimal expression. The K-map for the output *E* obtained directly from the given truth table is

AB\C	0	1
00	0	0
01	1	1
11	0	1
10	0	0

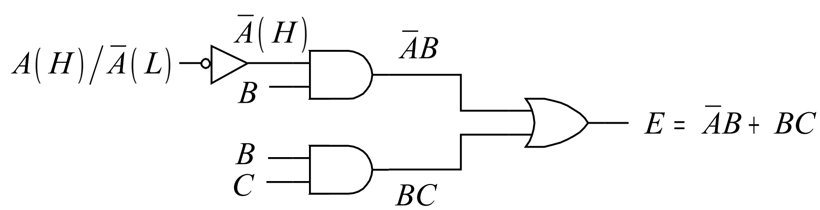
A sum-of-products (SOP) output expression is obtained by grouping together the 1's:

$$E = \overline{A}B + BC$$

A product-of-sums (POS) output expression is obtained by grouping together the 0's:

$$E = B\left(\overline{A} + C\right)$$

which, when expanded, is the same as the SOP expression. By inspection, the answer is (D). This can be easily confirmed by analyzing circuit (D):



112. Answer is (A)

The two extreme output conditions exist for a BJT when the transistor is in cutoff and saturation. When cutoff, little current passes through the transistor from the collector to emitter, and the transistor "switch" is open. Both the EB and CB junctions are reversed biased. Thus, a high impedance is seen looking into the emitter and the voltage across the load resistor is obtained using voltage division:

$$v_{L\min} = -15\frac{2}{2+10} = -2.5\text{ V}$$

(As a simple exercise, check to verify that the transistor is actually in cutoff for the given set of resistors.) When saturated, the transistor "switch" from collector to base is fully closed and the voltage from the collector to emitter is small at value of 0.2 V. The EB junction and CB junctions are forward biased. Thus, the voltage at the emitter is 0.2 V less than the voltage at the collector:

$$v_{L\max} = 15 - 0.2 = 14.8 \text{ V}$$

The answer is (A).

113. Answer is (B)

This is a series reliability problem. The probability that the transmitter will not malfunction (not fail) is 97%:

$$P_T = (1 - 0.03) = 0.97$$

The probability that the receiver will not malfunction (not fail) is 95%:

$$P_R = (1 - 0.05) = 0.95$$

Therefore, the probability that both the transmitter and receiver will not malfunction is (B):

$$P_T P_R = 0.97 \times 0.95 = 0.92$$

This is equal to $1 - (0.05 + 0.03 - 0.05 \times 0.03)$ and is less than the reliability of either the receiver or transmitter alone. For a parallel reliability problem, the reliability should be greater than either the receiver or transmitter alone. In this case, if it is acceptable to have either or both the transmitter or receiver operating properly, the reliability is 99.9%:

$$1 - (1 - P_T)(1 - P_R) = 1 - 0.03 \times 0.05 = 0.999$$

114. Answer is (D)

In this problem, the two events that can cause the amplifier to malfunction are not mutually exclusive. That is, they are not independent of each other. This can be seen by summing the two probability of failures:

$$P_{Famp} + P_{Fcap} = 0.001 + 0.002 = 0.003$$

This is not equal to the overall probability of failure given as 0.0025. Since the two events (i.e., probabilities of failure) are not mutually exclusive, there is some overlap in the Venn diagrams for these events. The overall probability of failure is then given as

$$P_{total} = P_{Famp} + P_{Fcap} - P_{Famp\&cap} = 0.001 + 0.002 - P_{Famp\&cap} = 0.0025$$

$$\Rightarrow P_{Famp\&cap} = 0.0005$$

There is 0.05% chance that the audio amplifier will fail due to a malfunction in both the final power transistor and the high-voltage capacitor. If the events were independent of each other, $P_{Famp\&cap} = 0$.

115. Answer is (A)

The intensity of the electric field between two conductors at different potentials generally increases as the applied voltage across them increases and as the distance between the conductors <u>decreases</u>. In this specific electrode configuration, the intensity is greatest directly below the cylindrical conductor where the sharpness of the conductors is the greatest (or where the radius of curvature is the smallest) and where the two conductors are the closest. The direction of the electric field at a point of interest between the conductors is determined by placing a small positive point charge at this point and qualitatively determining the direction of the force on it. (The positive charge is small so that it has little effect on the applied field from the charged conductors.) When a negative voltage is applied to the circular conductor relative to the ground plane, the ground plane is positively charged and the circular conductor is negatively charged. The small positive test charge will be attracted to the negatively charged circular conductor and repelled from the positively charged ground plane. (Recall that like charges repel and unlike charges attract.) Hence, the field at location *A* is in the direction of the circular conductor.

116. Answer is (D)

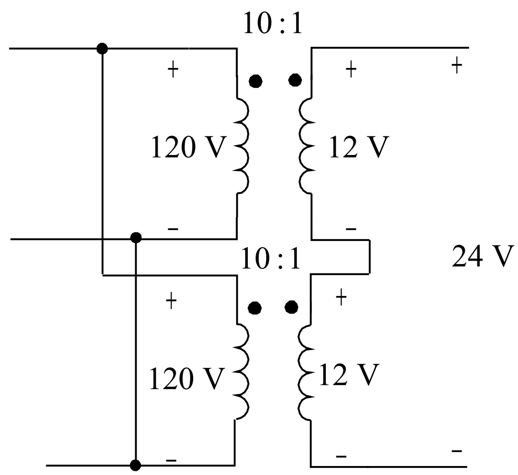
This is a favorite problem of mine that I have given (in various versions) to my students. First, it is important to remember that ideal ordinary transformers do not pass dc current. (Dc current, however, can affect the magnetic core of a real transformer.) Thus, the dc voltage on the primary or input side will not pass to the output. Second, note that the primary or input sides of the transformers are connected in parallel so the same 340 V peak-to-peak voltage is applied across both primaries. Third, notice that the secondary or output windings are connected in series adding (as determined by the polarity markers). Thus, the output voltage is

$$\frac{340}{10} + \frac{340}{10} = 2\left(\frac{340}{10}\right) \text{ V peak-to-peak}$$

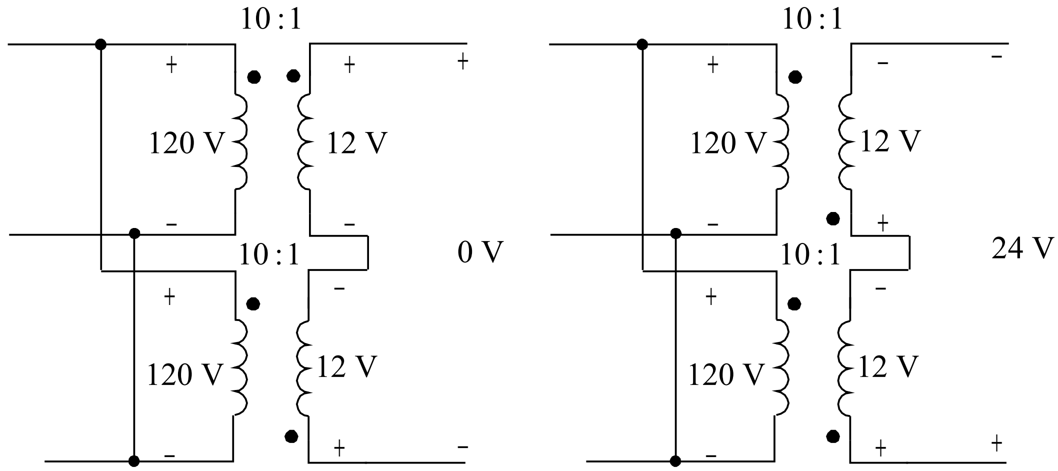
The rms value of the output is therefore (D):

$$\frac{1}{\sqrt{2}}\frac{2}{2}\left(\frac{340}{10}\right) = 24 \text{ V rms}$$

In the figure that follows, the dc offset across the primary sides is not shown and the rms values of the waveforms are shown.

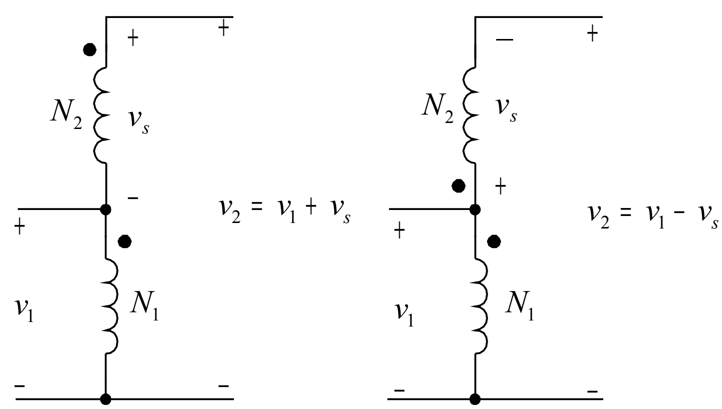


To help build your confidence with polarity markers, two other situations are shown with the same input voltage. For the 0 V output voltage situation, the secondaries are connected as series opposing or bucking.



117. Answer is (D)

An autotransformer can be constructed from an ordinary transformer by connecting one side of the primary to one side of the secondary. Hence, dc isolation is lost: (C) is correct while (D) is not correct. DC isolation implies that DC current (or voltage) cannot pass across the device. Based on how the windings are connected, the output voltage can either step up or step down the input voltage as shown in the given figures. Answer (B) is therefore correct. Finally, with autotransformers a winding is shared between the input and output, and in the given figures this corresponds to the coil with N_1 turns.



step-up autotransformer additive polarity

step-down autotransformer subtractive polarity

118. Answer is (A)

When qualitatively analyzing passive filters, four properties should be recalled:

1) an ideal capacitor has a reactance of $X_c = -1/\omega C$; thus, a capacitor appears like an open at "low" frequencies and a short at "high" frequencies
2) an ideal inductor has a reactance of $X_L = \omega L$; thus, an inductor appears like a short at "low" frequencies and an open at "high" frequencies

3) a series LC circuit appears like a short at some resonant frequency
4) a parallel LC circuit appears like an open at some resonant frequency

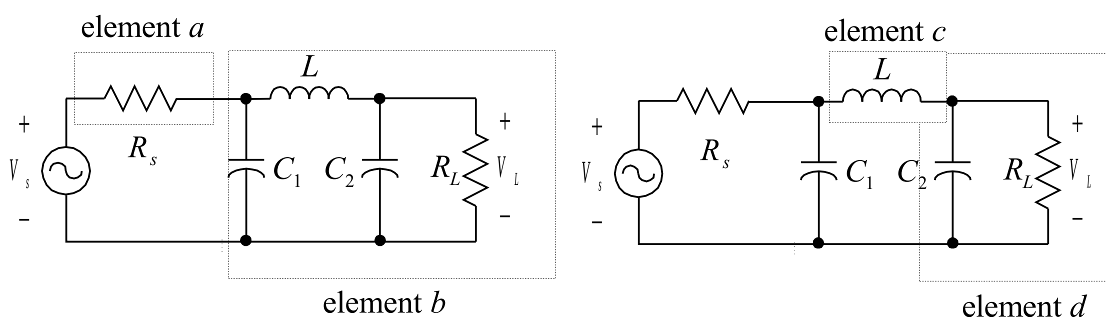
Using the first two properties, it is clear that at low frequencies the output voltage is about

$$V_L = A\frac{R_L}{R_L + R_s}$$

A "high" frequencies, on the other hand, the capacitors shunt and the inductor blocks the input signal. This is a low-pass filter.

Although mesh or nodal analysis can be used to determine the expression for the output voltage (by solving three equations with three unknowns), a one-line solution can be quickly written down by using voltage division twice:

$$\frac{V_L}{V_s} = \underbrace{\frac{\left[\left(R_L \| \frac{1}{j\omega C_2}\right) + j\omega L\right] \Big\| \frac{1}{j\omega C_1}}{\left\{\left[\left(R_L \| \frac{1}{j\omega C_2}\right) + j\omega L\right] \Big\| \frac{1}{j\omega C_1}\right\} + R_s}}_{\text{first voltage divider based on elements } a \text{ and } b} \times \underbrace{\frac{R_L \| \frac{1}{j\omega C_2}}{\left(R_L \| \frac{1}{j\omega C_2}\right) + j\omega L}}_{\text{second voltage divider based on elements } c \text{ and } d}$$



After much careful simplification, the previous equation can be reduced to answer (A).

Obviously, I designed this problem to be worked in a few minutes (not 20 minutes or more). What follows is the "trick" solution that can save you a lot of time. Of the four answers provided, only (A) and (D) have the correct low-frequency gain given previously (let $\omega = 0$ in these equations). (A) is probably more likely than (D) since this is a third-order filter (i.e., contains three separate energy storage elements), and the denominator in (A) contains a ω to the 3rd power.

119. Answer is (D)

The expression $z(t)$ is representing the standard amplitude modulated (AM) signal with a carrier:

$$z(t) = A\left[1 + km(t)\right]\cos[2\pi f_c t] = \underbrace{A\cos[2\pi f_c t]}_{\text{carrier}} + \underbrace{Akm(t)\cos[2\pi f_c t]}_{\text{sidebands}}$$

Hence, $x(t)$ is an AM signal without a carrier also referred to as double sideband (DSB) signal (or AM signal with a suppressed carrier). Therefore, the answer is either (A) or (D). There is now a 50% chance of guessing the right solution. The expression $y(t)$ is representing a phase modulated (PM) signal. The phase of the signal, often seen in textbooks as θ, is changing directly with $m(t)$:

$$y(t) = A\cos\left[2\pi f_c t + \underbrace{km(t)}_{\theta}\right]$$

The answer is therefore (D). Unless you have some background in communication theory, it may not be apparent why $s(t)$ is representing a frequency modulated (FM) signal. If the modulating signal was summed directly with the carrier frequency then it may be more obvious why the frequency is changing directly with $m(t)$: $A\cos\left\{2\pi\left[f_c + km(t)\right]t\right\}$. However, this is an uncommon way of representing an FM signal. It is more common to talk about the instantaneous frequency of the signal defined as the time derivative of the argument of the cosine:

$$\frac{d}{dt}\left[2\pi f_c t + k\int_0^t m(\lambda)\,d\lambda\right] = 2\pi f_c + km(t)$$

This instantaneous frequency is directly proportional to $m(t)$ and is thus FM.

120. Answer is (C)

A common way of representing an FM signal is

$$x(t) = A\cos\left[2\pi f_c t + k_f\int_0^t m(\lambda)\,d\lambda\right]$$

where k_f is the frequency deviation constant in rad/sec. For a sinusoidal modulating signal, $m(t) = a_m\cos(\omega_m t)$, the modulation index is defined as

$$\beta = \frac{k_f a_m}{\omega_m}$$

Carson's rule uses this modulation index to determine an (often fairly good) estimate of the bandwidth (in rad/sec) of an FM signal:

$$BW = 2(\beta + 1)\omega_m = 2\left(\frac{k_f a_m}{\omega_m} + 1\right)\omega_m = 2\left(k_f a_m + \omega_m\right)$$

As seen from this expression, the bandwidth of an FM signal is a function of the frequency deviation constant, amplitude of the modulating signal, and modulation frequency. Substituting the given information into this expression, the deviation constant is equal to

$$2\pi\left(25\times10^3\right) = 2\left[k_f a_m + 2\pi\left(10\times10^3\right)\right] \;\Rightarrow\; k_f = \frac{2\pi\left(25\times10^3\right)}{2a_m} - \frac{2\pi\left(10\times10^3\right)}{a_m}$$

(β, the phase constant, is 0.25 in this case, which implies the FM is narrow band.) Using a prime to represent the new bandwidth, amplitude, and frequency,

$$BW' = 2\left(k_f a_m' + \omega_m'\right) = 2\left\{\left[\frac{2\pi\left(25\times 10^3\right)}{2a_m} - \frac{2\pi\left(10\times 10^3\right)}{a_m}\right]2a_m + 2\pi\left(5\times 10^3\right)\right\}$$

$$= 2\left\{2\pi\left(25\times 10^3\right) - 4\pi\left(10\times 10^3\right) + 2\pi\left(5\times 10^3\right)\right\} = 2\pi\left(20\times 10^3\right)$$

The answer is (C).

121. Answer is (B)

This is a first-order circuit with a single time constant, τ, equal to $R_{eq}C$ where R_{eq} is the resistance seen by the capacitor after the switch is closed. When determining this resistance, all independent supplies are turned off: independent voltage supplies are replaced with shorts and independent current supplies are replaced with opens. The equivalent resistance is given by inspection as

$$R_{eq} = \left(R_s + R_1\right)\|R_2 = \frac{\left(R_s + R_1\right)R_2}{R_s + R_1 + R_2}$$

Based on the time constants provided, the only possible answers are (A) and (B). The general solution for any current or voltage in a single time constant circuit is

$$A + Be^{-\frac{t}{\tau}}$$

Even if some other quantity is requested, I generally will solve for the voltage across the capacitor (or current through the inductor for an *RL* circuit) since this voltage cannot change instantaneously. This capacitor voltage is then used, if necessary, to determine other quantities such as the current into the capacitor. Initially, the voltage across the capacitor was given as 0 V. This reduces the number of unknowns from two to one:

$$v_C\left(t\right) = A + Be^{-\frac{t}{\tau}} \quad\Rightarrow\quad v_C\left(0\right) = 0 = A + Be^0 \quad\Rightarrow\quad B = -A$$

$$v_C\left(t\right) = A - Ae^{-\frac{t}{\tau}} = A\left(1 - e^{-\frac{t}{\tau}}\right)$$

The variable A is determined by examining the circuit after a long time. The capacitor then appears like an open and the voltage across it is

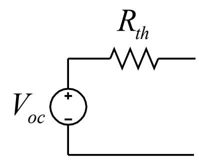
$$v_C\left(\infty\right) = V_s\frac{R_2}{R_s + R_1 + R_2}$$

The answer is (B) since $A = v_C\left(\infty\right)$:

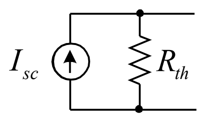
$$v_C\left(\infty\right) = A\left(1 - e^{-\frac{\infty}{\tau}}\right) = A\left(1 - 0\right) = A$$

122. Answer is (C)

The Thevenin equivalent is



while the Norton equivalent is



where $I_{sc} = V_{oc}/R_{th}$. (A) is not true since the actual circuit generally does not have the same net internal resistance as the resistance of its Thevenin and Norton equivalent circuit.

Although (B) seems very reasonable, it is generally not true. Imagine that a load resistor R_L is connected across both equivalent circuits. The powers dissipated by the two equivalent circuits are

$$P_{THEV} = \left(\frac{V_{oc}}{R_{th} + R_L} \right)^2 R_{th}, \quad P_{NORT} = \left(\frac{V_{oc}}{R_{th}} \frac{R_L}{R_{th} + R_L} \right)^2 R_{th}$$

These powers are identical only when $R_L = R_{th}$. Thus, if the temperature of two black boxes, one containing the Thevenin equivalent and one containing the Norton equivalent of the same circuit, are measured under a given load, the temperatures measured would not necessarily be the same. Imagine loading both boxes with a short

$$P_{THEV} = \left(\frac{V_{oc}}{R_{th}} \right)^2 R_{th}, \quad P_{NORT} = 0$$

or an open:

$$P_{THEV} = \left(\frac{V_{oc}}{R_{th} + \infty} \right)^2 R_{th} = 0, \quad P_{NORT} = \left(\frac{V_{oc}}{R_{th}} \frac{\infty}{R_{th} + \infty} \right)^2 R_{th} = \left(\frac{V_{oc}}{R_{th}} \right)^2 R_{th}$$

(D) is also not correct. Although in the literature individuals sometimes attempt to determine a Thevenin equivalent when a device is nonlinear and the output voltage swing is large (where small signal analysis cannot be applied), the equivalent single circuit (unless the equivalent elements are also a function of some parameter) is often of limited value and far from being an equivalent circuit.

The answer is (C). From the outside world, by definition, the load response that the equivalent circuit generates when connected to a load is identical to the actual circuit's response.

123. Answer is (D)

(A) is not correct. A battery's impedance is often low. Connecting a low-impedance load across the battery could excessively load it. The battery voltage could fall, and this load across the battery could cause an explosion.

(B) is not correct. This matching is not performed for maximum power transfer but to obtain more easily the *S* parameters.

(C) is not correct. The load impedance at the end of a lossy line is often selected to be equal to the characteristic impedance of the line to help eliminate pesky reflections.

Although some professors might disagree with my solution to this problem, (D) is the best answer.

124. Answer is (D)

This problem was included not because I think the maximum power transfer theorem is so important that it is worthy of being tested, but because many test makers seem to love problems involving it (e.g., please see the "official" sample test provided by NCEES). When dealing with an ideal transformer without a split secondary, the impedance looking into the primary with a load impedance of Z_L is given by

$$Z_{in} = \frac{Z_L}{\left(\frac{N_2}{N_1}\right)^2}$$

where N_2/N_1 is the ratio of the turns on the secondary to the primary coils of the transformer. Some sources refer to this ratio as n while others refer to it as $1/n$. With a split secondary, the input impedance of each separate load is transformed to the primary by its respective turns ratio; however, the two transformed impedance are not in series but in parallel on the primary side. Thus, the total impedance looking into the primary side of the transformer is

$$Z_{in} = \frac{Z_L}{\left(\frac{N_2}{N_1}\right)^2} \Bigg\| \frac{360}{\left(\frac{N_3}{N_1}\right)^2} = \frac{16 - j320}{\left(\frac{200}{50}\right)^2} \Bigg\| \frac{360}{\left(\frac{300}{50}\right)^2} = \frac{(1 - j20)10}{1 - j20 + 10} = \frac{(1 - j20)10}{11 - j20}$$

$$= \frac{(20.0\angle -87.1^\circ)10}{22.8\angle -61.2^\circ} = 8.77\angle -25.9^\circ = 7.9 - j3.8\,\Omega$$

If the Thévenin or equivalent impedance of the source is $Z_{th} = R_{th} + jX_{th}$, then maximum power is delivered to a load when $Z_L = R_{th} - jX_{th}$. The complex conjugate of Z_{in} is given in (D). (If the transformed impedances were incorrectly place in series, the answer would be $11 - j20$. The conjugate of this incorrect result is given in (A).)

125. Answer is (C)

This is a classical "circuit" that is usually solved in an introductory course in electromagnetics as a magnetic circuits problem. This magnetic circuit does not contain resistances but reluctances. The reluctances of the core and the air gap are in series. The reluctance decreases with increasing permeability (a magnetic material property). The corresponding reluctances are

$$\Re_c = \frac{2\pi R - g}{\mu_r \mu_o A}, \quad \Re_g = \frac{g}{\mu_o A}$$

The total reluctance is the sum of these reluctances. As the permeability of the core increases, the net reluctance decreases. Therefore, (A) is reasonable. The flux in the core and gap, which is analogous to the current in an electric circuit, is

$$\Phi = \frac{NI}{\Re_c + \Re_g} = \frac{NI}{\dfrac{2\pi R - g}{\mu_r \mu_o A} + \dfrac{g}{\mu_o A}} = \frac{NI \mu_r \mu_o A}{2\pi R + g(\mu_r - 1)}$$

where the driving force is *NI*, which is analogous to a source voltage.

The magnetic flux densities in both regions are

$$B_g = B_c = \frac{\Phi}{A} = \frac{NI \mu_r \mu_o}{2\pi R + g(\mu_r - 1)}$$

and the magnetic field intensities are

$$H_c = \frac{B_c}{\mu_r \mu_o} = \frac{NI}{2\pi R + g(\mu_r - 1)}, \quad H_g = \frac{B_g}{\mu_o} = \frac{NI \mu_r}{2\pi R + g(\mu_r - 1)}$$

The magnetic field intensities (not the magnetic flux densities) are not equal. As is clear from this expressions, the magnetic field intensities in both the core and air gap increase with *I*; hence, (D) is correct. The core losses generally increase with increasing field.

The net inductance seen by a source connected across the coil is

$$L = \frac{N\Phi}{I} = \frac{N^2 A \mu_r \mu_o}{2\pi R + g(\mu_r - 1)}$$

As is expected (at least for linear materials), this inductance is not a function of the current. As the gap length increases, the inductance decreases. This is reasonable since inductance tends to increase with the effective permeability of the overall flux path, and the effective path permeability will decrease as the air gap spacing increases. (B) is reasonable.

As the cross-sectional area of the core increases, the inductance increases (similar to standard inductors with a simple rod core). Since the impedance of the above inductance is given as $j\omega L$, the current *I* is given by

$$I = \frac{V_s}{j\omega L} = \frac{V_s}{j\omega \dfrac{N^2 A \mu_r \mu_o}{2\pi R + g(\mu_r - 1)}} = \frac{V_s \left[2\pi R + g(\mu_r - 1) \right]}{j\omega N^2 A \mu_r \mu_o}$$

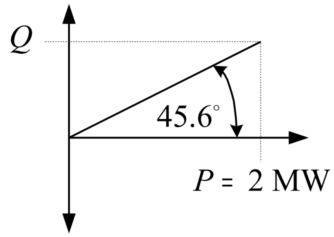
As *A* increases, the current decreases and the power factor (PF) as seen by the source decreases. (The power factor decreases with increasing inductance and increases with increasing resistance for a simple *RL* circuit.) Thus, (C) is not correct.

126. Answer is (C)

The concept of power factor (PF) is very important in power engineering. The power factor is defined as

$$\text{PF} = \cos\theta$$

where θ is the angle between the current into and the voltage across the load (input impedance looking into the mill in this problem). The power triangle is helpful in solving problems of this type:



$$\cos\theta = 0.7 \quad \Rightarrow \quad \theta = 45.6^\circ$$

where the complex power is defined as

$$S = P + jQ \text{ VA}$$

Using a little trigonometry, the imaginary power is

$$\tan\left(45.6^\circ\right) = \frac{Q}{2\times 10^6} \quad \Rightarrow \quad Q = 2 \text{ MVAR}$$

Lagging power factor implies that the paper mill, as seen by the external power connections, is inductive in nature. This is why the Q given is positive. If the power factor was zero, then the input impedance looking into the mill would be purely inductive (or capacitive). If the power factor was one, its maximum possible value, the input impedance would be purely resistive.

When a reactive load, such as a capacitor, is added in parallel with a load that requires correction, the Q's add directly. Adding a reactive load in parallel (not series) does not affect the real power delivered to the load (for a given source voltage). Thus, the new Q is

$$Q_{new} = 2 \text{ MVAR} - 1.2 \text{ MVAR}=0.8 \text{ MVAR}$$

which is still inductive, and the new phase angle is

$$\tan\theta_{new} = \frac{Q_{new}}{P} = \frac{0.8 \text{ MVAR}}{2 \text{ MW}} \quad \Rightarrow \quad \theta_{new} = 22^\circ$$

This corresponds to a PF of $\cos\left(22^\circ\right) = 0.93$. The power factor is improved (i.e., increased). The closest answer is (C).

127. Answer is (B)

For either a balanced Δ or Y load, the total real power delivered to all three loads is given by

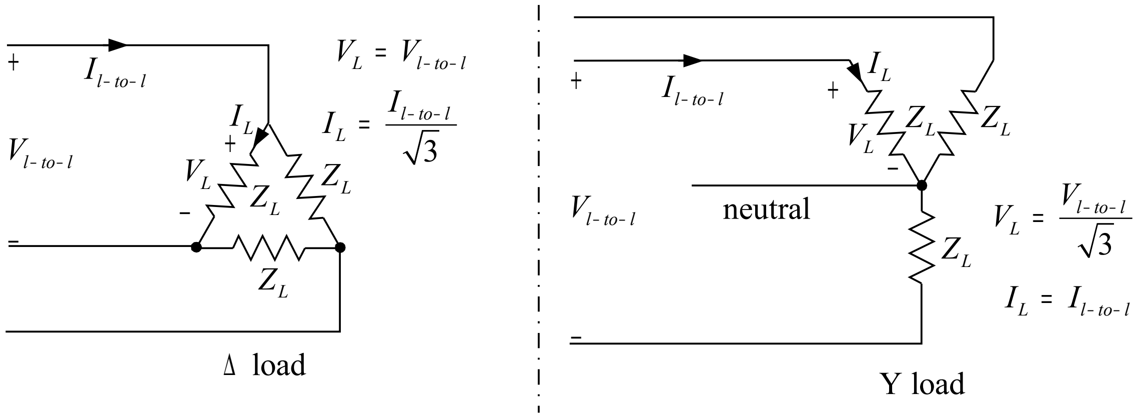
$$P_{total} = \sqrt{3}\, V_{rms} I_{rms} \cos\theta = \sqrt{3}\, V_{rms} I_{rms} PF$$

where V_{rms} is the magnitude of the rms line-to-line voltage, I_{rms} is the magnitude of the rms line current, θ is the phase angle between this voltage and current, and PF is the power factor. This is an easy problem to solve if this relationship for power is known:

$$I_{rms} = \frac{P_{total}}{\sqrt{3}\, V_{rms} PF} = \frac{150\times 10^3}{\sqrt{3}\; 230\times 0.85} = 440 \text{ A rms}$$

The leading information is not needed for this problem. The answer is (B).

Sometimes, load voltages and currents are given are requested. Whether the line-to-line voltage and line current are equal to the load voltages and currents, respectively, is dependent on the load type. It is actually quite easy to determine the magnitudes of the voltages and currents from the line-to-line voltage and line current and vice versa. These relationships are given in the following figures.



For a Δ load, the voltage across any of the loads is the same as the line-to-line voltage while the current in any of the loads is $1/\sqrt{3}$ times the line current. For the Y load, the line current is the same as the current in any of the loads while the voltage across any of the loads is $1/\sqrt{3}$ times the line-to-line voltage. Frequently, unless stated otherwise, for power-related problems, when a voltage is provided it is probably the line-to-line voltage (and an rms value).

128. Answer is (B)

For either a balanced Δ or Y load, the total real power delivered to all three loads is given by

$$P_{total} = \sqrt{3}\, V_{rms} I_{rms} \cos\theta = \sqrt{3}\, V_{rms} I_{rms} PF$$

where V_{rms} is the magnitude of the rms line-to-line voltage, I_{rms} is the magnitude of the rms line current, θ is the phase angle between this voltage and current, and PF is the power factor. Since the voltage at the motor or load is given, the impedance of each line leading to the conductor is not needed. The 20 HP rating for the motor is the deliverable power from the motor (not the input power). To convert HP to watts, multiply by 746:

$$20\ \text{HP} \times 746\ \frac{\text{W}}{\text{HP}} = 14.9\ \text{kW}$$

The actual input power to the motor is a function of the losses or efficiency of the motor. Thus, the input power is $14.9\ \text{kW}/0.88 = 17\ \text{kW}$, and the line current is (B):

$$I_{rms} = \frac{P_{total}}{\sqrt{3}\, V_{rms} PF} = \frac{17 \times 10^3}{\sqrt{3}\ 240 \times 0.78} = 52\ \text{A rms}$$

The leading information is not needed for this problem. The answer is (B).

129. Answer is (A)

Since the inductor appears like a "short" at low frequencies and an "open" at high frequencies, this is a high-pass filter. The cutoff frequency for this single-order filter can be determined two different ways. For this single-order filter, the inverse of the time constant is this cutoff frequency. The time constant is equal to

$$\tau = \frac{L}{R_{eq}} = \frac{L}{R_s \| R_L} = \frac{L}{\dfrac{R_s R_L}{R_s + R_L}}$$

Therefore, taking the inverse of this time constant, it is clear that (B) cannot be the answer. The cutoff frequency can also be determine be writing the expression for the voltage gain expression, using voltage division, and then rewriting the result in standard form for quick Bode magnitude plotting:

$$\frac{V_L}{V_s} = \frac{j\omega L \| R_L}{\left(j\omega L \| R_L \right) + R_s} = \frac{\dfrac{j\omega LR_L}{j\omega L + R_L}}{\dfrac{j\omega LR_L}{j\omega L + R_L} + R_s} = \frac{j\omega LR_L}{j\omega LR_L + j\omega LR_s + R_L R_s}$$

$$= \frac{j\omega LR_L}{j\omega \left(LR_L + LR_s \right) + R_L R_s} = \frac{j\omega LR_L}{R_L R_s \left(j\dfrac{\omega}{\dfrac{R_L R_s}{LR_L + LR_s}} + 1 \right)}$$

The cutoff frequency is seen below the variable ω, in the denominator.

Although (D) has the correct cutoff frequency, it also cannot be the correct Bode magnitude plot since its slope is 40 dB/decade. A first-order filter, containing only a single inductor, can only have a response with a slope of 20 dB/decade.

At high frequencies, the inductor appears like an “open” and the voltage gain is given by

$$\frac{R_L}{R_L + R_s}$$

Or, using the previous expression and letting the frequency approach infinity:

$$\frac{V_L}{V_s} = \frac{j\infty LR_L}{j\infty LR_L + j\infty LR_s + R_L R_s} = \frac{j\infty LR_L}{j\infty LR_L + j\infty LR_s} = \frac{R_L}{R_L + R_s}$$

The answer is therefore (A).

130. Answer is (C)

Depending on the handbook(s) that you have available, you may have the Fourier series representation for this full-rectified sinusoidal waveform with added dc offset. If you have plenty of extra time on the exam, the definition for the Fourier series may also be used:

$$x(t) = \frac{a_o}{2} + \sum_{n=1}^{\infty}\left[a_n \cos\left(\frac{2\pi\, n}{T}t\right) + b_n \sin\left(\frac{2\pi\, n}{T}t\right)\right]$$

$$\text{where } a_n = \frac{2}{T}\int_{t_o}^{t_o+T} x(t)\cos\left(\frac{2\pi\, n}{T}t\right)dt$$

$$b_n = \frac{2}{T}\int_{t_0}^{t_o+T} x(t)\sin\left(\frac{2\pi\, n}{T}t\right)dt$$

However, it is not necessary to use either a table or this integral-based definition to obtain the "best" answer for this problem. First, note that the dc or average value of the waveform is between V_{dc} and $V_{dc} + A$. This eliminates (A) and (B) since both of these dc values (those not a function of frequency) are less than V_{dc}:

$$\underbrace{V_{dc} - \frac{A}{\pi}}_{\text{dc value}} - \frac{4A}{3\pi}\cos\left(\frac{3\pi}{T}t\right) - \frac{4A}{15\pi}\cos\left(\frac{5\pi}{T}t\right) - \frac{4A}{35\pi}\cos\left(\frac{7\pi}{T}t\right) - \cdots$$

$$\underbrace{V_{dc} - \frac{A}{\pi}}_{\text{dc value}} - \frac{4A}{3\pi}\cos\left(\frac{2\pi}{T}t\right) - \frac{4A}{15\pi}\cos\left(\frac{4\pi}{T}t\right) - \frac{4A}{35\pi}\cos\left(\frac{6\pi}{T}t\right) - \cdots$$

Second, note that this waveform is even (i.e., symmetrical about the y axis). Thus, its series should only be a function of cosines (not sines) when this trigonometric form of the series is given (i.e., one without phase angles in the arguments). Recall that cosine is an even function, $x(-t) = x(t)$, while sine is an odd function, $x(-t) = -x(t)$. Thus, the most likely answer is (C). In this case, this is the series (because I pulled it directly from my book).

131. Answer is (B)

This is a simple semiconductor problem. Answers (C) and (D) are obviously incorrect (especially (D)). Since each boron atom accepts one electron, holes ("positive" charges or areas lacking an electron) are produced in the silicon. Boron is an acceptor. Thus, the material is considered a *p*-type semiconductor, (B).

132. Answer is (C)

Strain gauges are extremely important devices used in instrumentation but are only occasionally discussed in undergraduate electrical engineering courses. A change in the strain on the gauge will generate a small resistance change, which can be detected with the help of a Wheatstone Bridge. To work this problem, the definition of gauge factor should be known:

$$G = \frac{\frac{\Delta R}{R}}{\frac{\Delta L}{L}}$$

Where ΔR and ΔL are the change in resistance and length, respectively, and R and L are the manufacturers supplied nominal gauge resistance and length, respectively. Substituting the given information into the previous expression

$$3 = \frac{\frac{0.014}{350}}{\frac{\Delta L}{0.2}} \Rightarrow \Delta L = 2.7\mu\text{inches}$$

The answer is (C).

133. Answer is (D)

Although (A) seems reasonable, it is not correct. By definition, the resonant frequency is the frequency(s) where the impedance is purely real. The impedance seen by the source is equal to

$$Z = R_1 + \left[\left(R_2 + j\omega L \right) \middle\| \frac{1}{j\omega C} \right] = R_1 + \frac{\left(R_2 + j\omega L \right) \frac{1}{j\omega C}}{R_2 + j\omega L + \frac{1}{j\omega C}}$$

$$= R_1 + \frac{\left(R_2 + j\omega L \right)}{j\omega CR_2 - \omega^2 LC + 1} = R_1 + \frac{\left(R_2 + j\omega L \right)}{1 - \omega^2 LC + j\omega CR_2} \frac{1 - \omega^2 LC - j\omega CR_2}{1 - \omega^2 LC - j\omega CR_2}$$

$$= R_1 + \frac{R_2 - \omega^2 R_2 LC - j\omega CR_2^2 + j\omega L - j\omega^3 L^2 C + \omega^2 R_2 LC}{\left(1 - \omega^2 LC \right)^2 + \left(\omega CR_2 \right)^2}$$

$$= R_1 + \frac{R_2 - \omega^2 R_2 LC + \omega^2 R_2 LC}{\left(1 - \omega^2 LC \right)^2 + \left(\omega CR_2 \right)^2} + j \frac{-\omega CR_2^2 + \omega L - \omega^3 L^2 C}{\left(1 - \omega^2 LC \right)^2 + \left(\omega CR_2 \right)^2}$$

Setting the imaginary term portion of the impedance equal to zero, the resonant frequency is obtained:

$$-\omega CR_2^2 + \omega L - \omega^3 L^2 C = \omega \left(-CR_2^2 + L - \omega^2 L^2 C \right) = 0 \quad \Rightarrow \quad \omega_o = \sqrt{\frac{L - CR_2^2}{L^2 C}} = \sqrt{\frac{1}{LC} - \frac{R_2^2}{L^2}}$$

The resonant frequency, ω_o, is not equal to the commonly quoted result of $1/\sqrt{LC}$, which is the resonant frequency for simple series and parallel *RLC* circuits. Most other circuits do not have a resonant frequency of $1/\sqrt{LC}$. It was not expected that the resonant frequency be obtained in the above manner in a few minutes. Instead, I was testing whether you realized that the resonant frequency of most *RLC* circuits was not simply $1/\sqrt{LC}$.

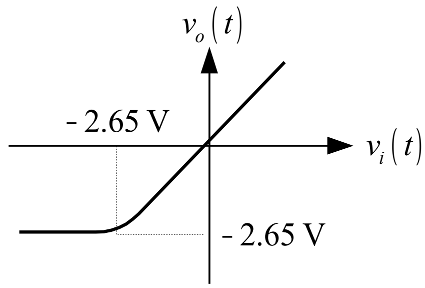
(B) is not correct. Although the maximum or minimum value in the current through or voltage across some elements can be near or at the resonant frequency, in general it is not necessarily at the resonant frequency.

(C) is not correct. If the Bode magnitude response for the current through L and R_2 is determined, it would contain several break or cutoff frequencies. Often when the phrases "low frequency" or "high frequency" are given, the frequency(s) is relative to one or more of these break frequencies. If the frequency is very high, the inductor appears like an open and the capacitor appears like a short. Thus, the current from the source mainly passes through C.

If the frequency is very low, the inductor appears like a short and the capacitor appears like an open. Thus, the current from the source mainly passes through L and R_2. Thus, (D) is the answer.

134. Answer is (B)

The $v_o(t)$ versus $v_i(t)$ plot for this limiter/clipper circuit is given below:



When the input voltage exceeds –2.65 V, the diode is reversed biased, and the output voltage is equal to the input voltage (assuming the load across the output has a very high impedance). For voltages less than –2.65 V, such as –5 V, the diode is forward biased, and the output voltage is clipped or limited to –2.65 V. When the input voltage is a sinusoid with an amplitude of 5 V, the bottom half of the sine is clipped at a value of –2.65 V. The answer is (B).

135. Answer is (D)

Careful! Is this amplifier operating in the active region? The dc voltage at the emitter is about

$$V_e = 8 - 0.7 = 7.3 \text{ V}$$

and the emitter current is

$$I_e = \frac{V_e}{2 \text{ k}\Omega} = 3.7 \text{ mA}$$

Since β is high, this is about equal to the collector current (<u>when</u> operating in the active region):

$$I_C = \frac{\beta}{\beta + 1} I_E = \frac{100}{100 + 1} 3.7 \text{ mA} = 3.6 \text{ mA}$$

The collector voltage is then

$$V_C = 12 - \left(3.6 \times 10^{-3}\right)\left(6.8 \times 10^{3}\right) = -13 \text{ V}$$

This result does not make any sense, especially since all of the applied voltages are greater than zero.

The large base voltage is driving the transistor into saturation. Both the PN junction of the base-emitter and the PN junction of the base-collector are forward biased. The voltage at the collector is low since the collector current is so great. When in saturation, the transistor's $V_{CESAT} = 0.2$ V can be used to determine the voltage at the collector:

$$V_C = V_{CESAT} + V_e = 0.2 + \left(8 - 0.7\right) = 7.5 \text{ V}$$

The power dissipated by the collector resistor is therefore

$$P_{6.8k} = \frac{V^2}{R} = \frac{(12 - 7.5)^2}{6.8 \times 10^3} = 3.0 \text{ mW}$$

The answer is (D).

136. Answer is (B)

Since β is so small, do not assume that I_E is approximately equal to I_C. For this PNP-based circuit,

$$V_E = 7.4 \text{ V}, V_B = 6.7 \text{ V}, V_C = 4 \text{ V} \quad \Rightarrow \quad V_{EB} = 0.7 \text{ V}, V_{CB} = -2.7 \text{ V}$$

Thus, the EB junction is forward biased and the CB junction is reversed biased. This transistor is operating in the active region and the standard relationships between the emitter, collector, and base currents can be used. Solving for the collector current,

$$I_C = \frac{V_C}{2 \text{ k}\Omega} = 2.0 \text{ mA}$$

The emitter current is therefore (B):

$$I_E = I_C \frac{\beta + 1}{\beta} = \frac{3 + 1}{3}(2 \text{ mA}) = 2.7 \text{ mA}$$

This current can be checked by using the base current

$$I_B = \frac{V_B}{10 \text{ k}\Omega} = 0.67 \text{ mA}$$

to determine the emitter current:

$$I_E = I_B(\beta + 1) = 0.67 \text{ mA} \times 4 = 2.7 \text{ mA}$$

137. Answer (C)

Grounding is an important topic in electrical safety (and electrical noise). A typical undergraduate (and graduate) program in electrical and computer engineering is frequently lacking in this area. Hopefully, the solution to this question will help in the understanding one aspect of safety grounding.

(A) is not correct (see the electrical code when the garage/barn is not attached to the house). By connecting both the main panel to a local earth ground and the subpanel to another local earth ground, the two panels will probably be at different potentials. Real earth has a finite conductivity and thus a potentially dangerous voltage drop will exist between these two panels. This is referred to as multiple-point grounding.

(B) is definitely not correct. This adds a third ground for this one cable further complicating the ground voltage/current issue. It is best to refer to the electric code when splicing a cable. (When is splicing into a cable for grounding purposes acceptable?)

(C) is the correct solution. There is one earth ground at the main panel. This is referred to as a single-point ground.

(D) is not correct. In most cases when an earth ground is conveniently available and has a decent conductivity, the main panel should be locally grounded. There are situations where the main panel can be grounded to a (metal) water pipe. Again, see the electric code for details. A GFCI is not a substitute for an earth ground.

138. Answer is (B)

When you worked this problem, did you fall into the trap of solving for the exact value of the input impedance (and wasting many minutes of valuable testing time) using the following expression?

$$Z_{in} = \left(\left\{ \left[\left(\frac{10}{\left(\frac{100}{1}\right)^2} \middle\| - j1.45\text{k} \right) + j20\text{k} + 14.1 \right] \middle\| j130\text{k} \middle\| 2.3\text{m} \right\} + 3\text{m} + j4.2 \right) \middle\| - j39\text{k}$$

Or, did you quickly perform the analysis in your head and determine the closest answer was (B)? No, I am not the "rain man" of circuit analysis. Recall that the impedance of two parallel elements, one much larger than the other, is approximately equal to the smaller element:

$$Z_1 \| Z_2 = \frac{Z_1 Z_2}{Z_1 + Z_2} \approx \frac{Z_1 Z_2}{Z_2} = Z_1 \text{ when } Z_1 << Z_2$$

When the two elements are in series, the net impedance is approximately equal to the larger element:

$$Z_1 + Z_2 \approx Z_2 \text{ when } Z_1 << Z_2$$

Using these two approximations, the impedance for this circuit can be analyzed in less than a minute without the use of a calculator (or slide rule):

$$\begin{aligned} Z_{in} &\approx \left(\left\{ \left[\frac{10}{(100)^2} + j20\text{k} + 14.1 \right] \middle\| j130\text{k} \middle\| 2.3\text{m} \right\} + 3\text{m} + j4.2 \right) \middle\| - j39\text{k} \\ &\approx \left(\left\{ j20\text{k} \| j130\text{k} \| 2.3\text{m} \right\} + 3\text{m} + j4.2 \right) \| - j39\text{k} \\ &\approx (2.3\text{m} + 3\text{m} + j4.2) \| - j39\text{k} = (5.3\text{m} + j4.2) \| - j39\text{k} \\ &\approx 5.3\text{m} + j4.2 \end{aligned}$$

139. Answer is (A)

This is a bridge bipolar transient protection network that uses one zener diode and four (less expensive) switching diodes.. It can handle transient voltage surges of both polarities. With a positive voltage of 14 V, diodes D_3 and D_2 will be forward biased and D_5 will be reversed biased beyond its zener breakdown voltage, thus also turned on. The answer is (A). With a negative voltage of −14 V, diodes D_1 and D_4 will be forward biased and D_5 will be again reversed biased beyond its zener breakdown voltage and turned on.

140. Answer is (B)

The total real power delivered to the electric heater and sump pump (assumed connected in parallel) should be the same after the parallel capacitance is added for power factor correction. The complex power delivered to a load is the sum of a real and an imaginary term:

$$S = P + jQ \text{ VA} \quad \text{where } \tan\theta = \frac{Q}{P}$$

The imaginary power absorbed by the heater is

$$Q_h = P_h \tan\theta_h = 1.4\times 10^3 \tan\left[\cos^{-1}(0.98)\right] = 284 \text{ VAR}$$

Since this power is lagging, it is inductive or positive. The imaginary power absorbed by the sump pump is also lagging and given as 23,000 VAR. Thus, the total VAR of the two loads is about 23.3 kVAR. The real power associated with this sump pump is

$$P_s = \frac{Q_s}{\tan\theta_s} = \frac{23\times 10^3}{\tan\left[\cos^{-1}(0.45)\right]} = 11.6 \text{ kW}$$

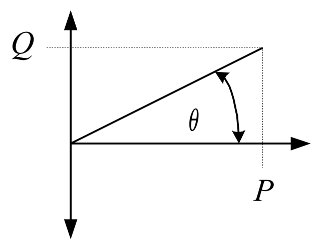
Thus, the total real power absorbed by the two loads is 11.6 + 1.4 =13 kW. Since this real power does not change with the addition of the capacitance, the required Q can be determined:

$$Q_r = P_r \tan\theta_r = 13\times 10^3 \tan\left[\cos^{-1}(0.95)\right] = 4.3 \text{ kVAR}$$

To reduce 23.3 kVAR to 4.3 kVAR, negative reactance or capacitance is needed with an associated VAR of 4.3 – 23.3 = −19 kVAR. Since the voltage across this capacitance is known, 120 V rms, the capacitance can be readily determined:

$$\Delta Q = \frac{V_{rms}^2}{\frac{-1}{\omega C}} = -\omega C V_{rms}^2 \quad \Rightarrow \quad C = \frac{\Delta Q}{2\pi f V_{rms}^2} = \frac{19\times 10^3}{2\pi(60)(120)^2} = 3.5 \text{ mF}$$

The answer is (B).



$$\cos\theta = \text{PF}$$
$$\theta = \cos^{-1}(\text{PF})$$