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Internal Assesment Test - II

Sub:	OPERATIONAL AMPLIFIERS & LINEAR ICS						Code:	17EE46	
Date:	20/04/2019	Duration:	90 mins	Max Marks:	50	Sem:	4th(A & B)	Branch:	EEE

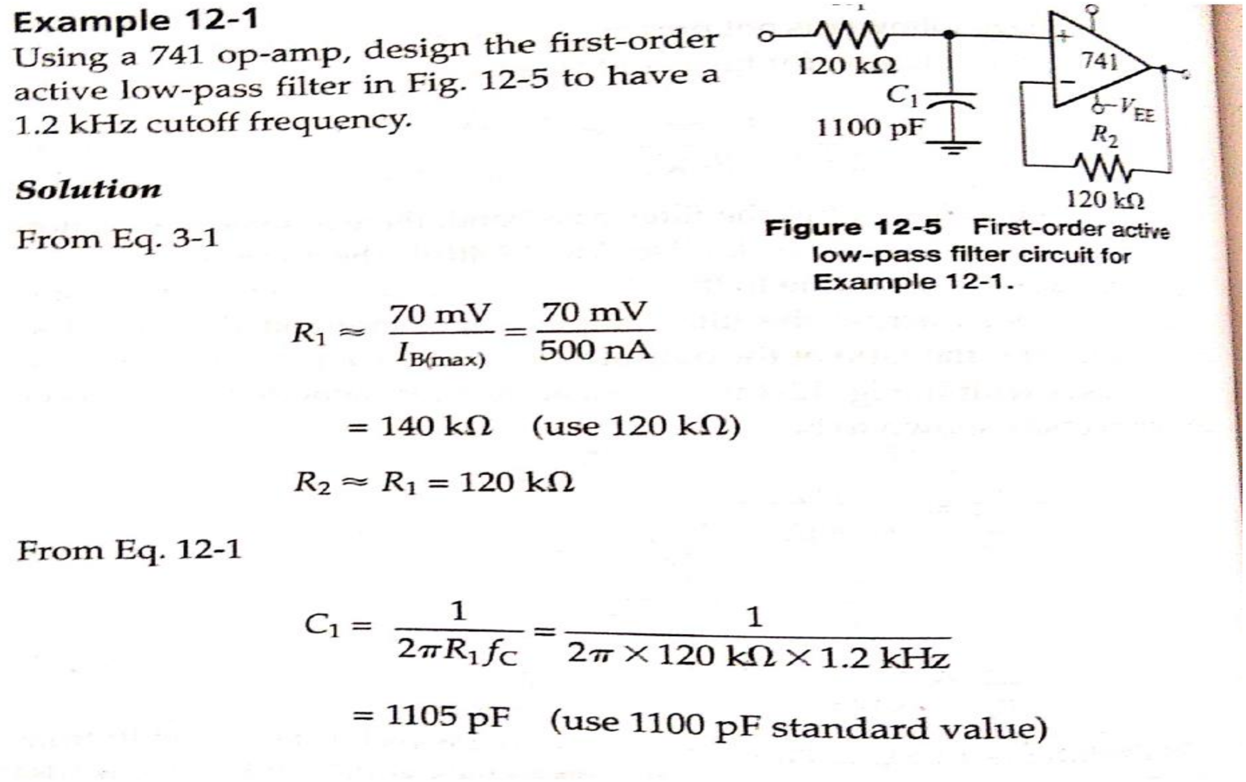
Answer any five questions.

		Marks	OBE	
			CO	RBT
1	Design a first order Low pass Butterworth filter with cut off frequency 1.2kHz and second order Low pass Butterworth filter with cut off frequency12 kHz.	10	CO6	L3
2	Explain the working of notch filter. Draw its frequency response.	10	CO6	L2
3	What are the advantages of active filters over passive filters? Design a two stage wide band pass filter and band reject filter having f_L=200Hz and f_H=1KHz and pass band gain of 4.Calculate Q value and center frequency. Assume capacitor values for high pass section=0.05uF and for the low pass section=0.01uF.Draw the circuit diagram.	10	CO6	L3
4	Explain the performance parameters of voltage regulators. An unregulated dc power supply output changes from 20v to 19.7v when the load increases from zero to maximum. The voltage also increases to 20.2v when the ac supply increases by 10%.Calculate the load and source effects and the load and line regulation.	10	CO6	L3
5	Design an adjustable voltage regulator to satisfy the following specifications: Output voltage Vo=5 to 12 v, output current Io=1 A. Voltage regulator is LM317.	10	CO6	L3
6	Explain the circuit of zero crossing detector and basic comparator.	10	CO6	L2
7	Demonstrate the operation of inverting Schmitt trigger with a neat circuit diagram, output waveforms and input output characteristics. For inverting Schmitt trigger circuit R1=100Ω,R2=56kΩ,Vin=+/-15v Determine the threshold voltages Vut and Vlt .	10	CO6	L3

******All the Best******

Solution for IAT-II

Q.1



Example 12-1

Using a 741 op-amp, design the first-order active low-pass filter in Fig. 12-5 to have a 1.2 kHz cutoff frequency.

Figure 12-5 First-order active low-pass filter circuit for Example 12-1.

Solution

From Eq. 3-1

$$R_1 \approx \frac{70\text{ mV}}{I_{B(max)}} = \frac{70\text{ mV}}{500\text{ nA}}$$

$$= 140\text{ k}\Omega \quad (\text{use } 120\text{ k}\Omega)$$

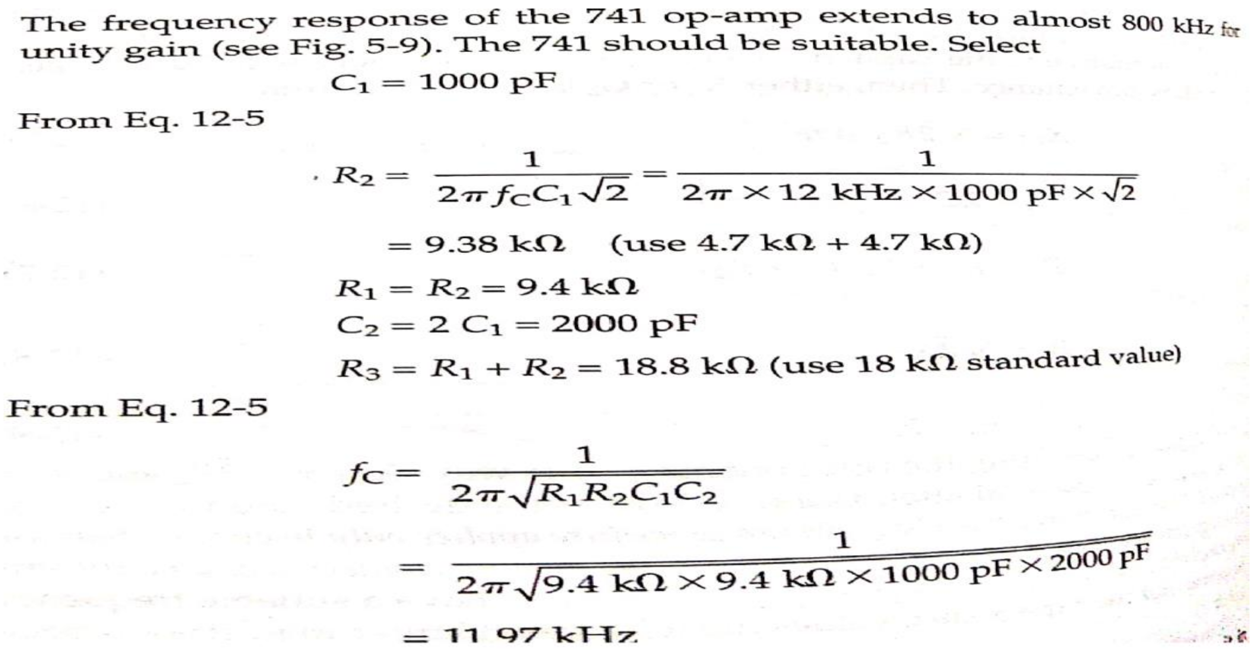
$$R_2 \approx R_1 = 120\text{ k}\Omega$$

From Eq. 12-1

$$C_1 = \frac{1}{2\pi R_1 f_C} = \frac{1}{2\pi \times 120\text{ k}\Omega \times 1.2\text{ kHz}}$$

$$= 1105\text{ pF} \quad (\text{use } 1100\text{ pF standard value})$$

Second order low pass butterworth filter:



The frequency response of the 741 op-amp extends to almost 800 kHz for unity gain (see Fig. 5-9). The 741 should be suitable. Select

$$C_1 = 1000 \text{ pF}$$

From Eq. 12-5

$$R_2 = \frac{1}{2\pi f_C C_1 \sqrt{2}} = \frac{1}{2\pi \times 12 \text{ kHz} \times 1000 \text{ pF} \times \sqrt{2}}$$

$$= 9.38 \text{ k}\Omega \quad (\text{use } 4.7 \text{ k}\Omega + 4.7 \text{ k}\Omega)$$

$$R_1 = R_2 = 9.4 \text{ k}\Omega$$

$$C_2 = 2\ C_1 = 2000 \text{ pF}$$

$$R_3 = R_1 + R_2 = 18.8 \text{ k}\Omega \ (\text{use } 18 \text{ k}\Omega \text{ standard value})$$

From Eq. 12-5

$$f_C = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

$$= \frac{1}{2\pi\sqrt{9.4 \text{ k}\Omega \times 9.4 \text{ k}\Omega \times 1000 \text{ pF} \times 2000 \text{ pF}}}$$

$$= 11.97 \text{ kHz}$$

Q.2

The narrow band-reject filter, often called the *notch filter*, is commonly used for the rejection of a single frequency such as 60-Hz power line frequency hum. The most commonly used notch filter is the *twin-T* network shown in Figure 7–15(a). This is a *passive filter* composed of two T-shaped networks. One T network is made up of two resistors and a capacitor, while the other uses two capacitors and a resistor. The *notch-out* frequency is the frequency at which maximum attenuation occurs; it is given by

$$f_N = \frac{1}{2\pi RC} \tag{7–16}$$

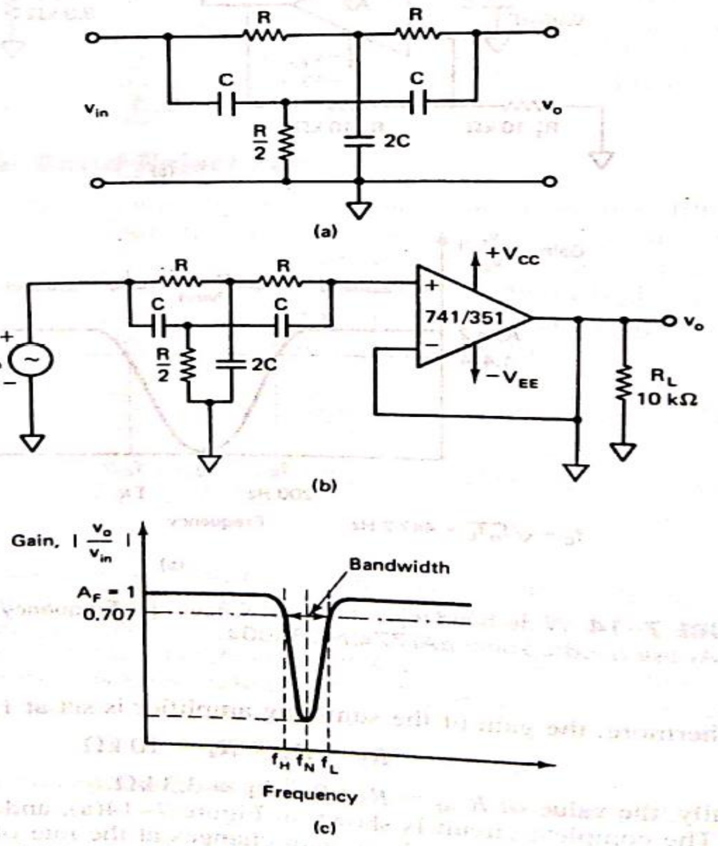
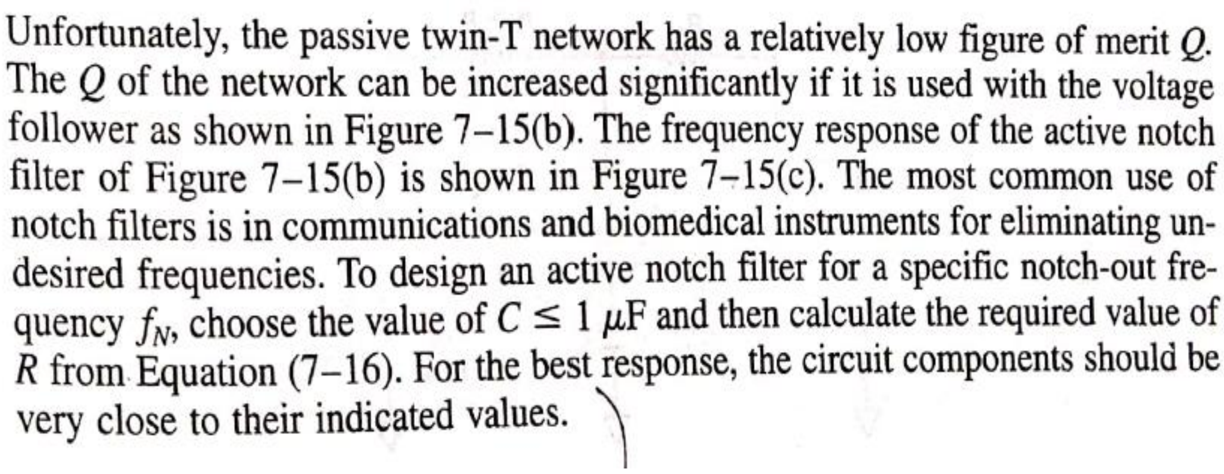
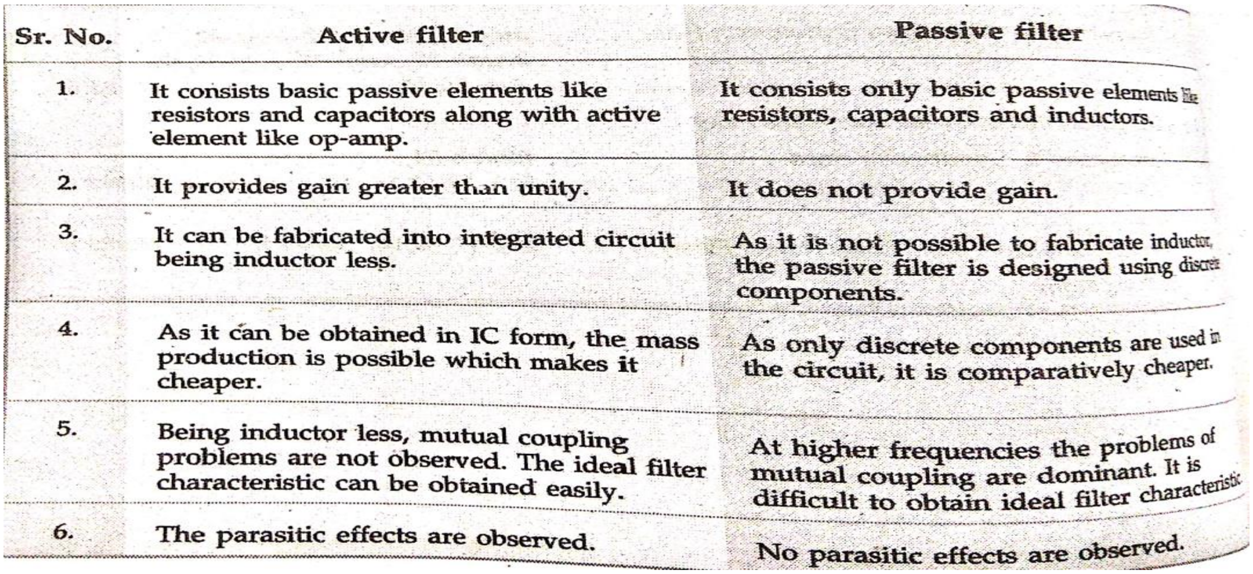


FIGURE 7–15 (a) Twin-T notch filter. (b) Active notch filter. (c) Frequency response of the active notch filter.

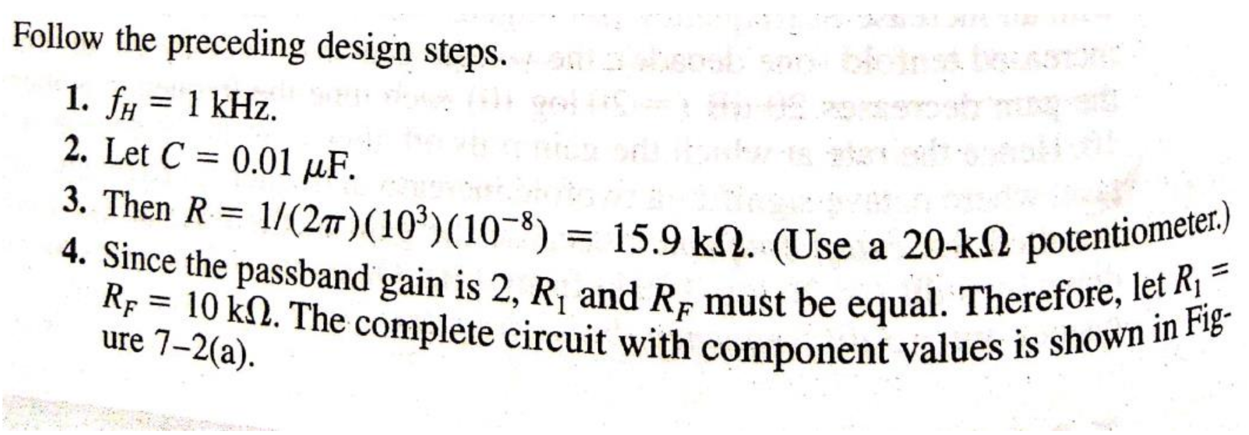


Unfortunately, the passive twin-T network has a relatively low figure of merit Q. The Q of the network can be increased significantly if it is used with the voltage follower as shown in Figure 7–15(b). The frequency response of the active notch filter of Figure 7–15(b) is shown in Figure 7–15(c). The most common use of notch filters is in communications and biomedical instruments for eliminating undesired frequencies. To design an active notch filter for a specific notch-out frequency f_N, choose the value of $C \leq 1\ \mu F$ and then calculate the required value of R from Equation (7–16). For the best response, the circuit components should be very close to their indicated values.

Q.3



Sr. No.	Active filter	Passive filter
1.	It consists basic passive elements like resistors and capacitors along with active element like op-amp.	It consists only basic passive elements like resistors, capacitors and inductors.
2.	It provides gain greater than unity.	It does not provide gain.
3.	It can be fabricated into integrated circuit being inductor less.	As it is not possible to fabricate inductor, the passive filter is designed using discrete components.
4.	As it can be obtained in IC form, the mass production is possible which makes it cheaper.	As only discrete components are used in the circuit, it is comparatively cheaper.
5.	Being inductor less, mutual coupling problems are not observed. The ideal filter characteristic can be obtained easily.	At higher frequencies the problems of mutual coupling are dominant. It is difficult to obtain ideal filter characteristic
6.	The parasitic effects are observed.	No parasitic effects are observed.



Follow the preceding design steps.

1. $f_H = 1$ kHz.
2. Let $C = 0.01\ \mu F$.
3. Then $R = 1/(2\pi)(10^3)(10^{-8}) = 15.9\ k\Omega$. (Use a 20-k$\Omega$ potentiometer.)
4. Since the passband gain is 2, R_1 and R_F must be equal. Therefore, let $R_1 = R_F = 10\ k\Omega$. The complete circuit with component values is shown in Figure 7–2(a).

(a) A low-pass filter with $f_H = 1$ kHz was designed in Example 7–1; therefore, the same values of resistors and capacitors can be used here, that is, $R' = 15.9\ \text{k}\Omega$ and $C' = 0.01\ \mu\text{F}$. As in the case of the high-pass filter, it can be designed by following the steps of section 7–3–1:

1. $f_L = 200$ Hz.
2. Let $C = 0.05\ \mu\text{F}$.
3. Then

$$R = \frac{1}{2\pi f_L C} = \frac{1}{(2\pi)(200)(5)(10^{-8})}$$
$$= 15.9\ \text{k}\Omega$$

$$f_C = \sqrt{(1000)(200)} = 447.2\ \text{Hz}$$

$$Q = \frac{447.2}{100 - 200} = 0.56$$

Thus Q is less than 10, as expected for the wide band-pass filter.

Q.4

Ripple Rejection

The *ripple rejection* is a measure of how much a voltage regulator attenuates the supply voltage ripple from the unregulated power supply. It is usually expressed in decibels. With a supply ripple of V_{rs} and an output ripple of V_{ro}

$$\text{Ripple rejection} = 20\log\left(\frac{V_{rs}}{V_{ro}}\right) \qquad (13\text{-}5)$$

Source Effect

The ac supply to the input of a transformer in a dc power supply does not always remain constant. A ±10% variation in the ac source voltage is not unusual, and this causes some variation in the dc output voltage from a regulated power supply. This output voltage change (ΔV_o) due to a supply voltage change is termed the *source effect*. If the output varies by 100 mV when the source voltage changes by ±10%, the source effect is 100 mV. An alternative way of stating this output change is to express ΔV_o as a percentage of the dc output voltage (V_o). In this case, the term *line regulation* is used.

$$\text{Source effect} = \Delta V_o \text{ for a 10\% change in supply} \tag{13-1}$$

$$\text{Line regulation} = \frac{(\Delta V_o \text{ for a 10\% change in } V_S) \times 100\%}{V_o} \tag{13-2}$$

Load Effect

Power supply output voltage is also affected by changes in load current (I_L). The output voltage decreases when I_L is increased and rises when I_L is reduced. The *load effect* defines how the output voltage changes when the load current is increased from zero to its specified maximum level ($I_{L(max)}$). If the load current change (ΔI_L) produces a voltage change (ΔV_o) of 100 mV, the *load effect* is 100 mV. As for the source effect, the load effect can also be expressed as a percentage of the output voltage. This is termed the *load regulation*.

$$\text{Load effect} = \Delta V_o \text{ for } \Delta I_{L(max)} \tag{13-3}$$

$$\text{Load regulation} = \frac{(\Delta V_o \text{ for } \Delta I_{L(max)}) \times 100\%}{V_o} \tag{13-4}$$

$$\text{load effect} = \Delta V_o \text{ for } \Delta I_{L(max)} = 20\text{ V} - 19.7\text{ V}$$

$$= 300\text{ mV}$$

From Eq. 13-4

$$\text{load regulation} = \frac{(\Delta V_o \text{ for } \Delta I_{L(max)}) \times 100\%}{V_o}$$

$$= \frac{300\text{ mV} \times 100\%}{20\text{ V}}$$

$$= 1.5\%$$

From Eq. 13-1

$$\text{source effect} = \Delta V_o \text{ for a 10\% change in supply} = 20.2\text{ V} - 20\text{ V}$$

$$= 200\text{ mV}$$

From Eq. 13-2

$$\text{line regulation} = \frac{(\Delta V_o \text{ for a 10\% change in } V_S) \times 100\%}{V_o}$$

$$= \frac{200\text{ mV} \times 100\%}{20\text{ V}}$$

$$= 1\%$$

Q.5

For the LM317, $I_{ADJ} = 100\ \mu A$ maximum. If we use $R_1 = 240\ \Omega$, then for V_o of 5 V the value of R_2 from Equation (9–17b) is

$$5 = 1.25\left(1 + \frac{R_2}{240}\right) + (10^{-4})R_2$$

$$R_2 = \frac{3.75}{(5.3)(10^{-3})}$$

$$= 0.71\ k\Omega$$

Similarly, for $V_o = 12$ V, the value of R_2 is

$$12 = 1.25\left(1 + \frac{R_2}{240}\right) + (10^{-4})R_2$$

or

$$R_2 = \frac{10.75}{(5.3)(10^{-3})}$$

$$= 2.01\ k\Omega$$

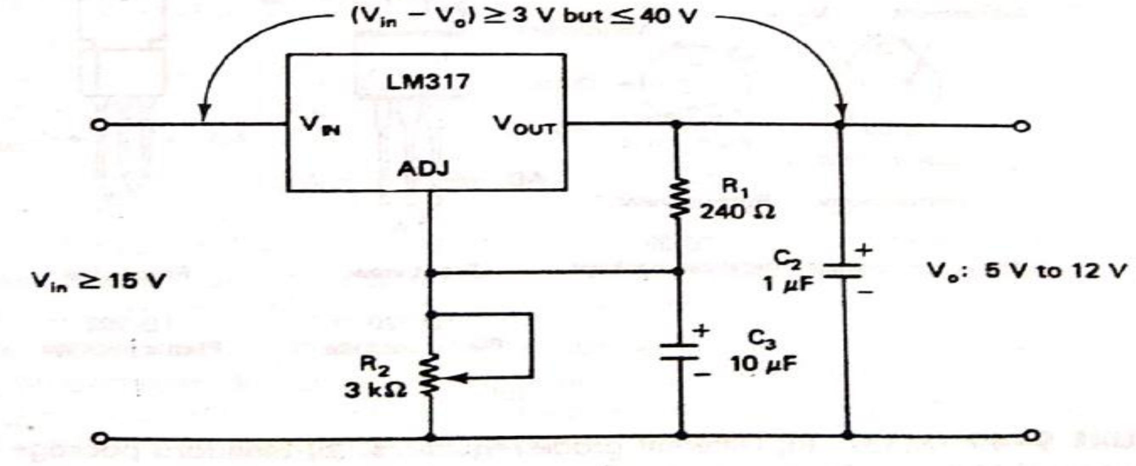
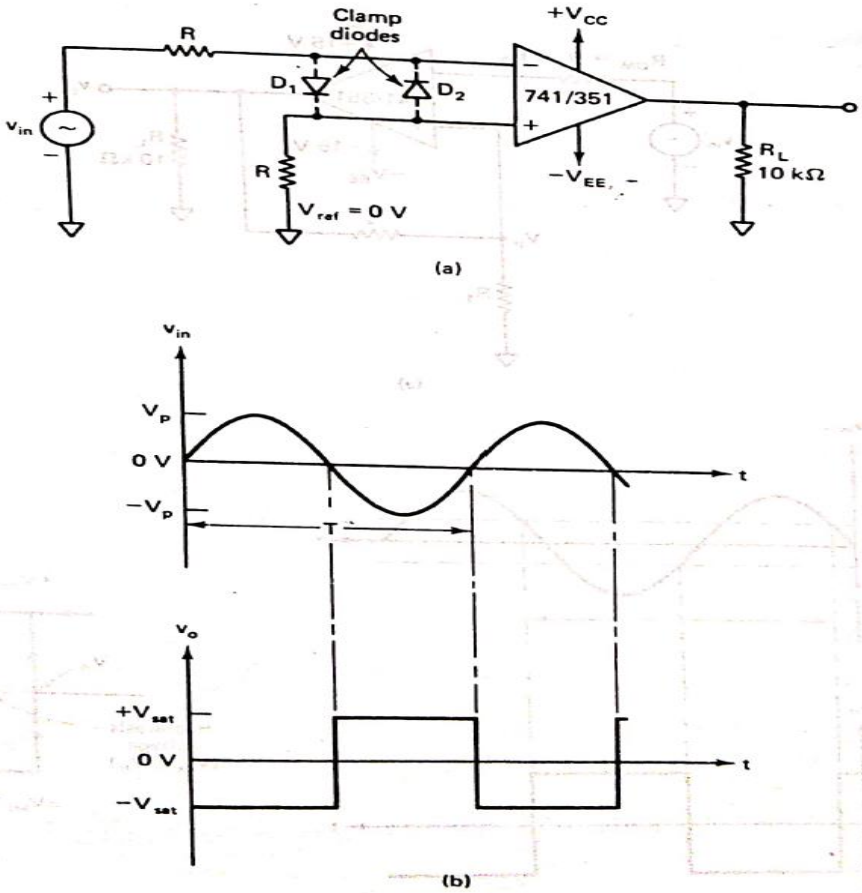


FIGURE 9–48 Adjustable voltage regulator for Example 9–11.

Q.6

8-3 ZERO-CROSSING DETECTOR

An immediate application of the comparator is the *zero-crossing detector* or *sine wave-to-square wave converter.* The basic comparator of Figure 8–1(a) or Figure 8–2(a) can be used as the zero-crossing detector provided that V_{ref} is set to zero ($V_{\text{ref}} = 0$ V). Figure 8–3(a) shows the inverting comparator used as a zero-crossing detector. The output voltage v_o waveform in Figure 8–3(b) shows when and in what direction an input signal v_{in} crosses zero volts. That is, the output v_o is driven into negative saturation when the input signal v_{in} passes through zero in the positive direction. Conversely, when v_{in} passes through zero in the negative direction, the output v_o switches and saturates positively.



(a)

(b)

IGURE 8–3 (a) Zero-crossing detector. (b) Its typical input and output waveforms.

8-2 BASIC COMPARATOR

Figure 8–1(a) shows an op-amp used as a comparator. A fixed reference voltage V_{ref} of 1 V is applied to the $(-)$ input, and the other time-varying signal voltage v_{in} is applied to the $(+)$ input. Because of this arrangement, the circuit is called the *noninverting comparator.* When v_{in} is less than V_{ref}, the output voltage v_o is at $-V_{\text{sat}}$ $(\cong -V_{EE})$ because the voltage at the $(-)$ input is higher than that at the $(+)$ input. On the other hand, when v_{in} is greater than V_{ref}, the $(+)$ input becomes positive with respect to the $(-)$ input, the v_o goes to $+V_{\text{sat}}$ $(\cong +V_{cc})$. Thus v_o changes from one saturation level to another whenever $v_{\text{in}} \cong V_{\text{ref}}$, as shown in Figure 8–1(b). In short, the comparator is a type of analog-to-digital converter. At any given time the v_o waveform shows whether v_{in} is greater or less than V_{ref}. The comparator is sometimes also called a *voltage-level detector* because, for a desired value of V_{ref}, the voltage level of the input v_{in} can be detected.

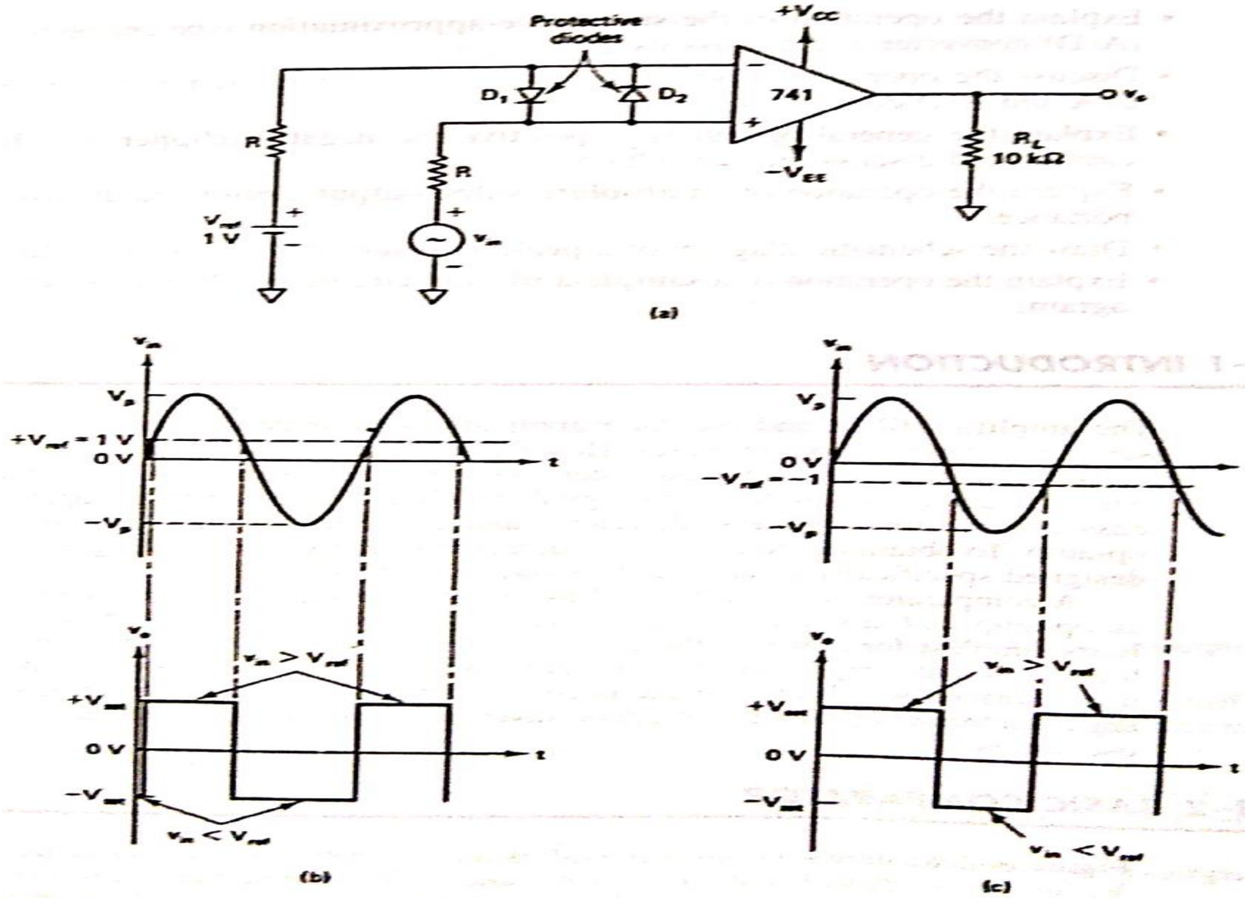


FIGURE 8–1 (a) Noninverting comparator and its input and output waveforms. (b) If V_{ref} is positive. (c) If V_{ref} is negative.

Q.7

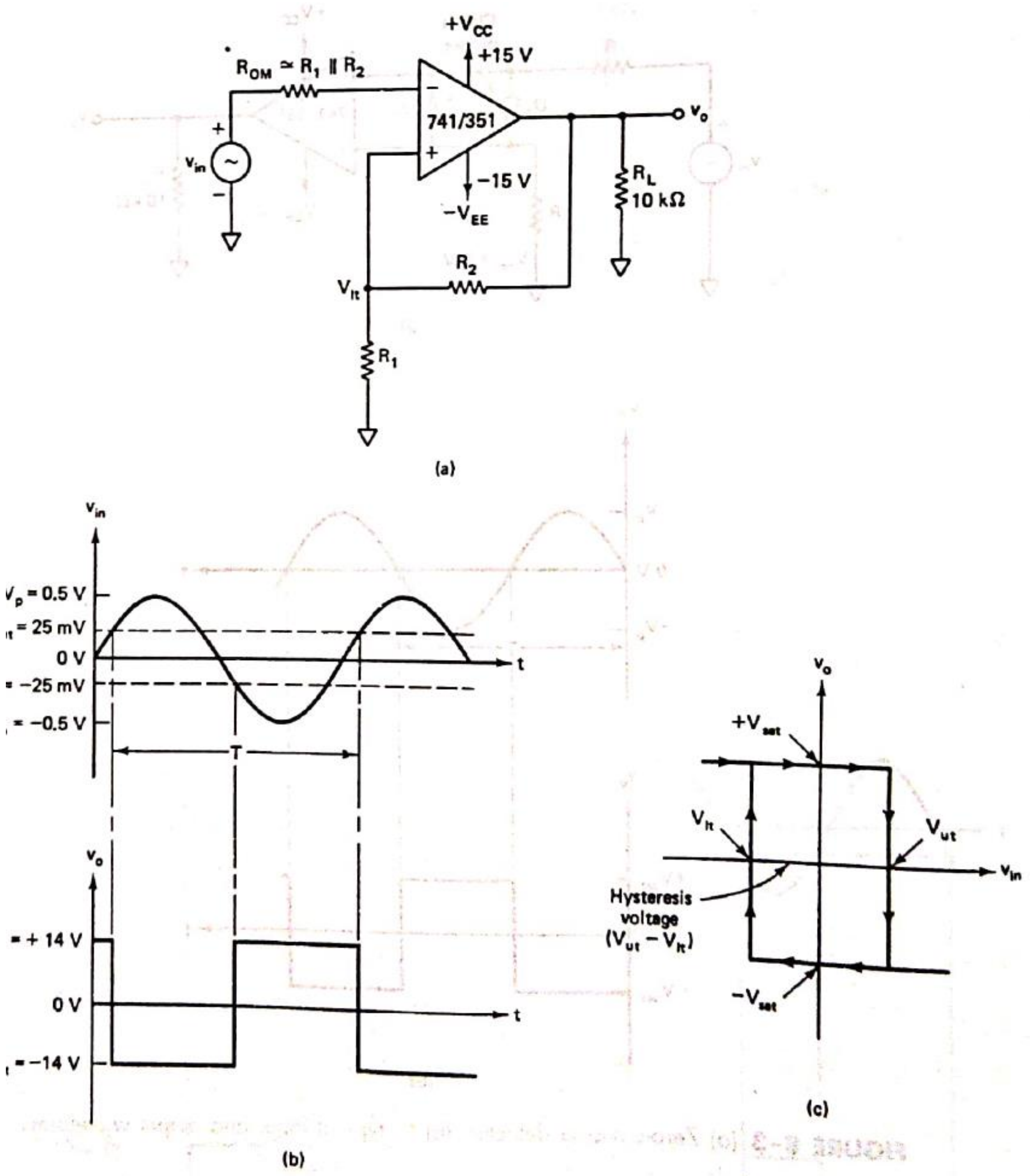


FIGURE 8–4 (a) Inverting comparator as Schmitt trigger. (b) Input and output waveforms of Schmitt trigger. (c) v_o versus v_{in} plot of the hysteresis voltage.

In Figure 8–4(a), these threshold voltages are obtained by using the voltage divider $R_1 - R_2$, where the voltage across R_1 is fed back to the (+) input. The voltage across R_1 is a variable reference threshold voltage that depends on the value and polarity of the output voltage v_o. When $v_o = +V_{sat}$, the voltage across R_1 is called the *upper threshold voltage*, V_{ut}. The input voltage v_{in} must be slightly

more positive than V_{ut} in order to cause the output v_o to switch from $+V_{sat}$ to $-V_{sat}$. As long as $v_{in} < V_{ut}$, v_o is at $+V_{sat}$. Using the voltage-divider rule,

$$V_{ut} = \frac{R_1}{R_1 + R_2}(+V_{sat}) \tag{8–1a}$$

On the other hand, when $v_o = -V_{sat}$, the voltage across R_1 is referred to as the *lower threshold voltage*, V_{lt}. v_{in} must be slightly more negative than V_{lt} in order to cause v_o to switch from $-V_{sat}$ to $+V_{sat}$. In other words, for v_{in} values greater than V_{lt}, v_o is at $-V_{sat}$. V_{lt} is given by the following equation:

$$V_{lt} = \frac{R_1}{R_1 + R_2}(-V_{sat}) \tag{8–1b}$$

Thus, if the threshold voltages V_{ut} and V_{lt} are made larger than the input noise voltages, the positive feedback will eliminate the false output transitions. Also, the positive feedback, because of its regenerative action, will make v_o switch faster between $+V_{sat}$ and $-V_{sat}$. In Figure 8–4(a), resistance $R_{OM} \cong R_1 \| R_2$ is used to minimize the offset problems.

SOLUTION

For 741 the maximum output voltage swing is ± 14 V, that is, $+V_{sat} = 14$ V and $-V_{sat} = -14$ V. From Equations (8–1a) and 8–1b),

$$V_{ut} = \frac{100}{56{,}100}(14) = 25 \text{ mV}$$

$$V_{lt} = \frac{100}{56{,}100}(-14) = -25 \text{ mV}$$

The output v_o waveform is shown in Figure 8–4(b). From Equation (8–2), the hysteresis voltage $V_{hy} = 50$ mV.