UNIT III
ACTIVE FILTERS & OSCILLATORS

FILTER:-

A frequency selective circuit that passes electric signals of specified band of frequencies and attenuates the signals of frequencies outside the band is called an electric filter.

CLASSIFICATION OF FILTERS:-

Filters may be classified in a number of ways. some of them are

(1) Analog (or) digital

(2) Audio frequency (AF) (or) Radio frequency (RF)

(3) passive (or) Active.

(1) Analog (or) digital filters:-

Analog filters are designed to process analog signals, while digital filters process analog signals using digital techniques.

(2) Audio Frequency (or) Radio frequency filters:-

The type of element used for filtering operation dictates the operating frequency range of the filter.

For example, RC filters are commonly used for audio (or) low frequency applications operation, whereas LC (or) crystal filters are employed at RF (or) high frequencies. Especially because of their high Q value (figure of merit), the crystals provide more stable operation at higher frequencies.

(3) passive (or) Active filters:-

Depending on the type of element used in their construction, filters may be classified as passive (or) Active.

Elements used in passive filters are resistors, capacitors, and Inductors.

Active filters employ transistor (or) op-amp in addition to the resistors and capacitors.

passive filters work well for high frequencies, i.e, radio frequencies. However at audio frequencies inductors become problamatic, as the inductors become large, heavy and expensive. For low frequency application, more number of turns of wire must be used which in turn adds to the series resistance degrading inductor's performance, i.e low Q, resulting in high power dissipation. Inductors also emit magnetic fields.

The active filters overcome the above problems of the passive filters. The active filters, by enclosing a capacitor in the feedback loop, avoid using inductors. In this way inductorless active RC filters can be obtained.

Advantages of Active filters over passive filters:-

1. Gain and frequency adjustment flexibility:-

Since the op-amp is capable of providing a gain, the input signal is not attenuated as it is in a passive filter.

In addition, the active filter is easier to tune (or) adjust.

2. No loading problem:-

Because of high input resistance and low output resistance of the op-amp, the active filter does not cause loading of the source (or) load.

3. Cost:- Active filters are more economical than passive filters. This is because of the variety cheaper op-amps and absence of inductors.

4. Because of the high i/p impedance of the op-amp, a large value resistors can be used, there by reducing the value (size and cost) of the capacitors required in the design.

5. Active filters can be realized under number of class of functions such as butterworth, Thomson, chebyshev, cauer etc.

6. The frequency response (or) performance of the active filter can be improved as compared to passive filters due to ~~the~~ ready availability of high quality components.

General Applications of filters

(1). Electric filters are used in circuits which require the seperation of signals according to their frequencies.

(2) mostly used in the field of communication and signal processing and in one form (or) another in almost all ~~ele~~ sophisticated electronic systems.

(3) In Radio, television, telephone, radar, space satellites and biomedical equipment

Limitations of Active filters:-

(1) High frequency response is limited by the gain band-width (GBW) product and slew rate of the op-amp for most of the applications, a reasonbly good filter performance can be achieved approximately upto 500KHz. As against this passive filter can be used upto 500MHz.

(2). The high frequency active filters are more expensive than the passive filters. The passive filter in high frequency range is a more economic choice for applications.

(3) The active elements are much sensitive to temperature and environmental changes than the passive filter. Due to this the active filter performance may deviates from its ideal response.

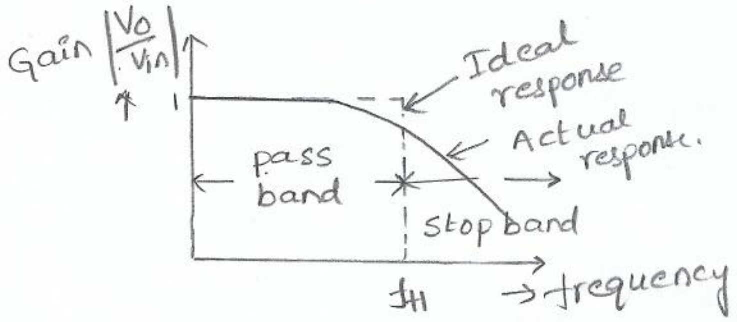
(4) The requirement of dc power supply is another limitation of Active filters. passive filters donot require the d.c supply.

The most commonly used filters are

1. Low pass filter 2. High-pass filter 3. Band-pass filter
4. Band-reject filter 5. All-pass filter.

FREQUENCY RESPONSE CHARACTERISTICS OF FILTERS

(1) Low pass filter:-



The above figure shows the frequency response characteristics of the low pass filter. The Ideal response is shown by dotted lines, while the solid lines indicate practical filter response.

A low pass filter has a constant gain 0 Hz to a high cut off frequency f_H. Therefore, the bandwidth is also f_H. In the practical response, at f_H the gain is down by 3dB, After that $(f>f_H)$ it decreases with the increase in input frequency.

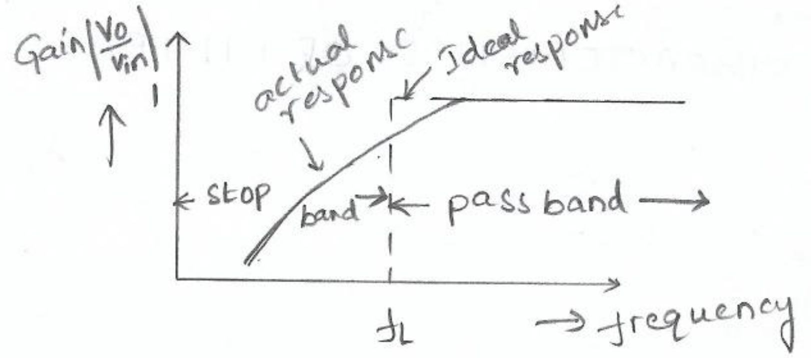
The frequencies between 0 Hz and f_H are known as the passband frequencies, whereas the range of frequencies beyond f_H, that are attenuated includes the stopband frequencies.

(2) High pass filter:-

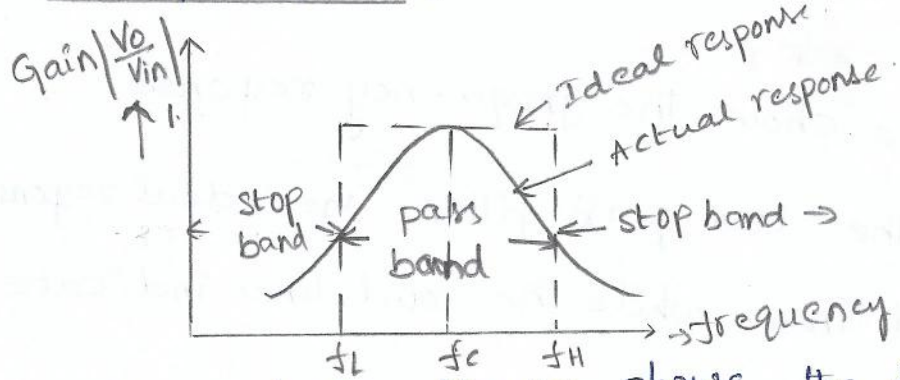
Below figure shows the frequency response characteristics of High pass filter.

For a high pass filter, f_L is the low cut off frequency.

The range of frequencies $0<f<f_L$ is the stop band where f is the operating frequency. while the range of frequencies $(f>f_L)$ are the pass band.



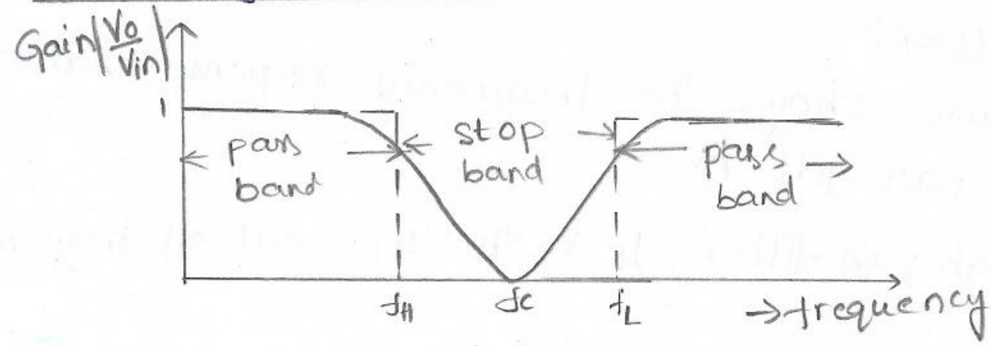
(3) Band-pass filter:-



The above figure shows the frequency response characteristics of the band-pass filter.

A band pass filter has a passband between two cut off frequencies f_H and f_L, where $f_H > f_L$. The range of frequencies $0<f<f_L$ and range of frequencies $f_H<f<\infty$ are two stop bands while the range $f_L<f<f_H$ is the pass band. The bandwidth is thus $f_H - f_L$.

(4) Band-reject filter:-



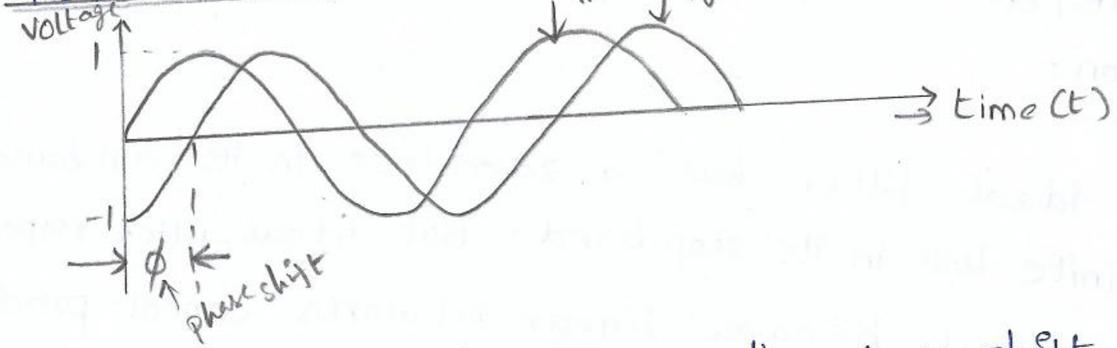
The above figure shows the frequency response characteristics of band-reject filter.

The band-reject filter performs exactly opposite to the band-pass filter; that is, it has a band stop between two cutoff frequencies f_H and f_L and two pass bands: $0 < f < f_L$ and $f > f_H$.

The band reject is also called a band-stop (or) band-elimination filter.

f_c is called center frequency since it is approximately at the center of the pass band in case of bandpass filter & stop band incase of band-stop filter.

(5) ALL PASS FILTER:-



The above figure shows the phase shift between input and output voltages of an all-pass filter. This filter passes all the frequencies equally well; That is, output and input voltages are equal in amplitude for all frequencies; with the phase shift between the two. The highest frequency up to which the input and output amplitudes remain equal is dependent on the unity gain-bandwidth of the op-amp. At this frequency the phase shift between the input and output is maximum.

In the above filters, the actual filter response in the stop band either steadily decrease (or) increase (or) both with increase in frequency.

The rate at which the gain of the filter changes in the stop band is determined by the order of the filter. For example, for the first order low pass filter the gain rolls off at the rate of 20 dB/decade in the stop band, i.e, for $f > f_H$. For the second order low-pass filter the roll-off rate is 40 dB/decade and so on. By contrast, for the first-order high pass filter the gain increases at the rate of 20 dB/decade in the stop band, that is, until $f = f_L$; the increase for the second order high-pass filter is 40 dB/decade and so on.

An ideal filter has a zero loss in its pass band and infinite loss in its stop band. But ideal filter response is not practical because linear networks cannot produce the discontinuities. However, it is possible to obtain a practical response that approximates the ideal response by using special design techniques, as well as precision component values and high-speed op-amps.

The various types of filters used in practice which approximately produce the ideal response are

(1) Butterworth filters
(2) Chebyshev filters
(3) Cauer filters

The key characteristic of the Butterworth filter is that it has a flat pass band as well as stopband. For this reason, it is sometimes called a flat-flat filter. The design of butterworth filters is very simple.

The chebyshev filter has a ripple pass band and flat stop-band.

The caur filter has a ripple passband and a ripple stop-band. Generally, the cauer filter gives the best stopband response among the three filters.

Butter worth Approximation:-

The filter in which denominator polynomial of its transfer function is butterworth polynomial is called a butter worth filter.

The butterworth polynomials of various orders are given below.

order (n)	Butterworth polynomials in factored form
1	$s+1$
2.	$s^2+\sqrt{2}s+1$
3.	$(s^2+s+1)(s+1)$
4.	$(s^2+0.76536s+1)(s^2+1.84776s+1)$
5.	$(s+1)(s^2+0.6180s+1)(s^2+1.6180s+1)$
6.	$(s^2+0.5176s+1)(s^2+\sqrt{2}s+1)(s^2+1.9318s+1)$
7.	$(s+1)(s^2+0.4450s+1)(s^2+1.2456s+1)(s^2+1.8022s+1)$
8.	$(s^2+0.3986s+1)(s^2+1.1110s+1)(s^2+1.6630s+1)(s^2+1.9622s+1)$

The coefficients of butterworth polynomials are given below

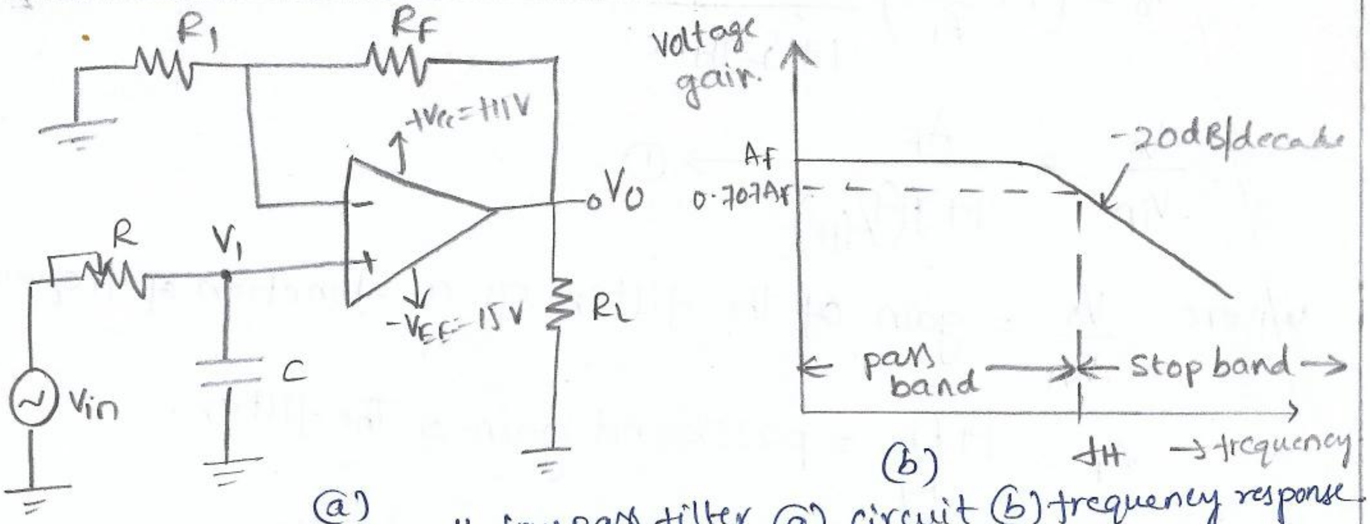
order(n)	a_1	a_2	a_3	a_4	a_5	a_6	a_7	a_8
1	1							
2.	$\sqrt{2}$	1						
3.	2	2	1					
4.	2.613	3.414	2.613	1				
5.	3.236	5.236	5.236	3.236	1			
6.	3.864	7.464	9.141	7.464	3.864	1		
7.	4.494	10.103	14.606	14.606	10.103	4.494	1	
8.	5.126	13.138	21.848	25.691	21.848	13.138	5.126	1

$a_0^* = 1$

Some important observations of butterworth polynomials :-

1. coefficient of highest power of s is always 1 for any order of filter.
2. lowest order term i.e, Constant term is always '1' for any order of filter.
3. For all odd ordered filters, one of the factors is always (s+1) while the remaining factors are quadratic in nature.
4. All the poles of butterworth polynomial are located in left half of s-plane on a circle with radius equal to one and centre at origin.
5. For all even ordered filters, all factors are quadratic in nature.

FIRST ORDER LOW PASS BUTTERWORTH FILTER :-



First-order butterworth low pass filter (a) circuit (b) frequency response.

The above figure shows first order low pass Butterworth filter. It uses an RC network for filtering along with an op-amp which is in noninverting configuration. Hence it does not load down the RC network.

The Resistors R_1 & R_F determine the gain of the filter.

According to the voltage divider rule, the voltage at the noninverting terminal (i.e, across capacitor C) is

$$V_1 = \frac{-jX_C}{R-jX_C} V_{in}$$

where $j = \sqrt{-1}$, $-jX_C = \frac{1}{j2\pi fC}$.

$$V_1 = \frac{1/j2\pi fC}{R + 1/j2\pi fC} V_{in}$$

$$V_1 = \frac{V_{in}}{1+j2\pi fCR}$$

and the output voltage $V_0 = \left(1+\frac{R_F}{R_1}\right) V_1$

$$V_o = \left(1+\frac{R_F}{R_1}\right)\frac{V_{in}}{1+j2\pi fRC}$$

$$\frac{V_o}{V_{in}} = \frac{A_F}{1+j(f/f_H)} \rightarrow ①.$$

where $\frac{V_o}{V_{in}}$ = gain of the filter as a function of frequency

$A_F = 1+\frac{R_F}{R_1}$ = passband gain of the filter.

f = frequency of the input signal.

$f_H = \frac{1}{2\pi RC}$ = high cut off frequency of the filter.

The gain magnitude and phase angle of the low pass filter can be obtained by converting eqn ① to polar form, as follows:

$$\left|\frac{V_o}{V_{in}}\right| = \frac{A_F}{\sqrt{1+(f/f_H)^2}} \rightarrow ②$$

$$\phi = \frac{\tan^{-1}0}{\tan^{-1}(f/f_H)} = 0-\tan^{-1}(f/f_H)$$

$$\phi = -\tan^{-1}(f/f_H) \rightarrow ③.$$

where ϕ is the phase angle in degrees.

The operation of the low-pass filter can be verified from the gain magnitude equation ②:

1. At very low frequencies $f < f_H$

$$\left|\frac{V_o}{V_{in}}\right| \cong A_F.$$

2. At $f = f_H$ $\left|\frac{V_o}{V_{in}}\right| = \frac{A_F}{\sqrt{2}} = 0.707\,A_F$ i.e, 3dB down to the level of A_F.

3. At $f > f_H$ $\left|\frac{V_o}{V_{in}}\right| < A_F$

Thus the low pass filter has a constant gain A_F from 0 Hz to the high cutoff frequency f_H. At f_H the gain is $0.707 A_F$. and after f_H it decreases at a constant rate with an increase in frequency. That is, when the frequency is increased tenfold (one decade), the voltage gain is divided by 10. In other words, the gain decreases 20dB ($= 20\log 10$) each time the frequency is increased by 10.

Hence the rate at which the gain rolls off after f_H is 20 dB/decade (or) 6dB/octave, where octave signifies a twofold increase in frequency.

The frequency $f = f_H$ is called cutoff frequency because the gain of the filter at this frequency is down by 3dB ($= 20 \log 0.707$) from 0 Hz. The cut off frequency is also called -3 dB frequency, break frequency (or) corner frequency.

Filter Design:-

A low-pass filter can be designed by implementing the following steps:

1. choose a value of high cutoff frequency f_H.
2. Select a value C less than (or) equal to 1μF. Mylar (or) tantalum capacitors are used for better performance.
3. Calculate the value of R using $R = \frac{1}{2\pi f_H C}$.
4. Finally, select values of R_1 & R_F dependant on the desired pass band gain 'A_F' using $A_F = 1 + \frac{R_F}{R_1}$

Frequency scaling:-

once the filter is designed, some times, it is necessary to change the value of cut off frequency f_H. The method used to change the original cut-off frequency f_H to a new cut-off frequency f_H is called a frequency scaling.

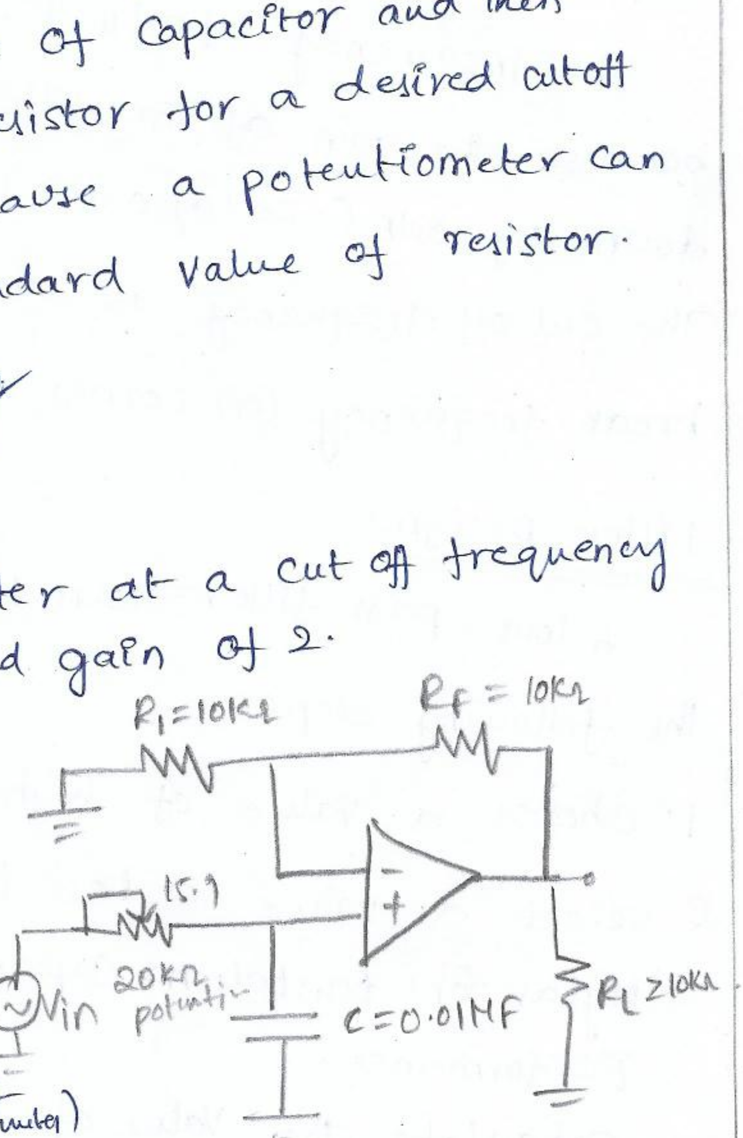
To change a high cutoff frequency, multiply R (or) C but not both by the ratio of the original cut off frequency to the new cutoff frequency.

In filter design the needed values of R and C are often not standard. Besides a variable capacitor 'C' is not commonly used. Therefore, choose a standard value of capacitor and then calculate the value of resistor for a desired cutoff frequency. This is because a potentiometer can be used for a nonstandard value of resistor.

$$R_2' = R_2 \left(\frac{f_c}{f_c'} \right)^2$$

problems:-

(1). Design a low-pass filter at a cut off frequency of 1KHz with a pass band gain of 2.



Steps

1. $f_H = 1KHz$
2. Let $C = 0.01MF$
3. $R = \frac{1}{2\pi f_H C} = \frac{1}{2\pi \times 1K \times 0.01M}$
   $= 15.9 k\Omega$ (use 20kΩ potentiometer)
4. $A_F = 2$
   $1 + \frac{R_F}{R_1} = 2 \quad R_F = R_1 = 10k\Omega$

First order High pass butterworth filter:-

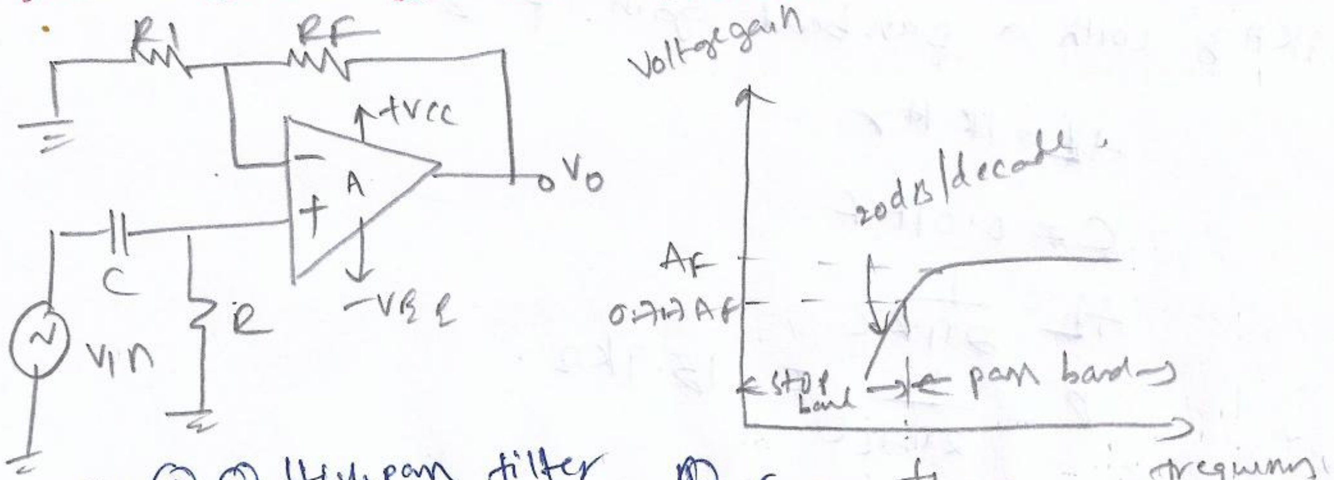


Fig ① ⓐ High pass filter ⓑ Frequency response.

High pass filters are formed simply by interchanging frequency-determining resistors & capacitor in low pass filter. It has low cutoff frequency f_L. This is the frequency at which the magnitude of the gain is 0.707 times its pass band value. All the frequencies higher than f_L are pass band frequencies, with highest frequency determined by closed bandwidth of op-amp.

For the first order high pass filter, the o/p is

$$V_o = \left(1+\frac{R_F}{R_1}\right)\frac{j2\pi fRC}{1+j2\pi fRC}\, V_{in}$$

(or) $$\frac{V_o}{V_{in}} = A_F\left[\frac{j\,(f/f_L)}{1+j(f/f_L)}\right]$$

where $A_F = 1+\frac{R_F}{R_1}$ = pass band gain of the filter.

$f_L = \frac{1}{2\pi RC}$ = low cutoff frequency (f_H)

f = frequency of the i/p signal (Hz)

The magnitude of the voltage gain is

$$\left|\frac{V_o}{V_{in}}\right| = \frac{A_F\,(f/f_L)}{\sqrt{1+(f/f_L)^2}} = \frac{A_F}{\sqrt{1+(f_L/f)^2}}$$

Scaling procedure is also applicable here as in LPF

① Design a high pass filter at a cut of frequency of 1KHz with a pass band gain of 2.

$\cancel{f_L} = 1K Hz$

$C = 0.01 \mu F$

$\cancel{f_L} = \frac{1}{2\pi RC}$

$R = \frac{1}{2\pi f_L C} = 15.9 k\Omega.$

$A_F = 2$

$1 + \frac{R_F}{R_1} = 2 \implies R_F = R_1 = 10K\Omega$

BAND-PASS FILTERS:-

A band-pass filter has a passband between two cutoff frequencies f_H and f_L such that $f_H > f_L$. Any input frequency outside this passband is attenuated.

Band pass filters are two types.

(1) wideband pass filter

(2) Narrow band pass filter.

we can define a filter as wide band pass if its figure of merit (or) quality factor $Q<10$. on the otherhand, if $Q>10$, we call the filter a narrow band-pass filter.

quality factor (Q) is a measure of selectivity, meaning the higher the value of Q, the more selective is the filter (or) the narrower its bandwidth (BW).

The relationship between Q, the 3-dB bandwidth, and the center frequency f_c is given by

$$Q = \frac{f_c}{BW} = \frac{f_c}{f_H - f_L}$$

For the wide bandpass filter, the center frequency (f_c) can be defined as

$$f_c = \sqrt{f_H f_L}$$

where f_H = High cutoff frequency (Hz)

f_L = Low cutoff frequency (Hz) of the wideband-pass filter.

In a narrow bandpass filter, the output voltage peaks at the center frequency.

WIDE BAND-PASS FILTER:-

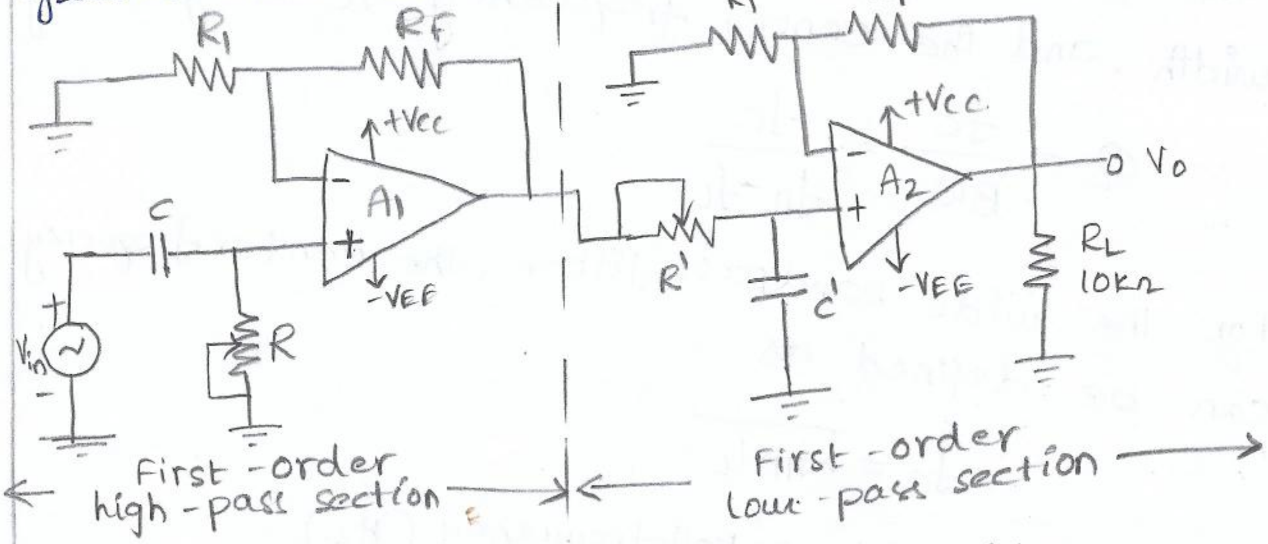
A wide band-pass filter can be formed by simply cascading high-pass and low-pass sections and is generally the choice for simplicity of design and performance.

To obtain a ±20 dB/decade band-pass, first order high-pass and first order low-pass sections are cascaded; for a ±40 dB/decade band-pass filter, second-order high pass and second-order low-pass sections are connected in series and so on.

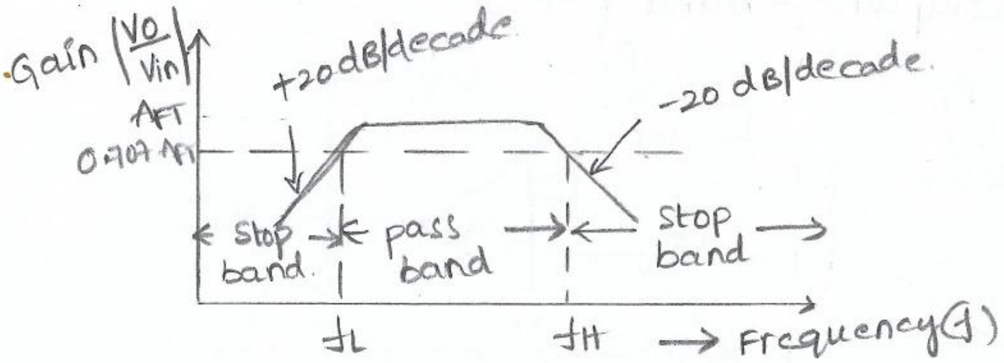
So the order of the band-pass filter depends on the order of the high pass & low-pass filter sections.

The below figure shows the ±20 dB/decade wide band-pass filter, which is composed of first-order high-pass and first-order low-pass filters.

To realize a band-pass response f_H must be greater than f_L.



(a) ±20 dB/decade wide band-pass filter



(b) Frequency response of wide band pass filter.

Problem:-

(1) (a) Design a wide band-pass filter with $f_L = 200Hz$, $f_H = 1KHz$ and a passband gain $= 4$.

(b) Draw the frequency response plot of this filter

(c) Calculate the value of Q for the filter.

Ans:- For low pass filter.

1. $f_H = 1KHz$.
2. Let $C' = 0.01\mu F$.
3. Then $R' = \frac{1}{2\pi f_H C'} = \frac{1}{2\pi \times 1\times10^3 \times 0.01\times10^{-6}} = 15.9K\Omega$

For High pass filter

1. $F_L = 200Hz$
2. Let $C = 0.05\mu F$
3. Then $R = \frac{1}{2\pi f_L C} = \frac{1}{2\pi \times 200 \times 0.05\times10^{-6}} = 15.9K\Omega$

The band-pass gain(A_{FT}) is 4

$A_{FT} = A_1 \times A_2 = 4$

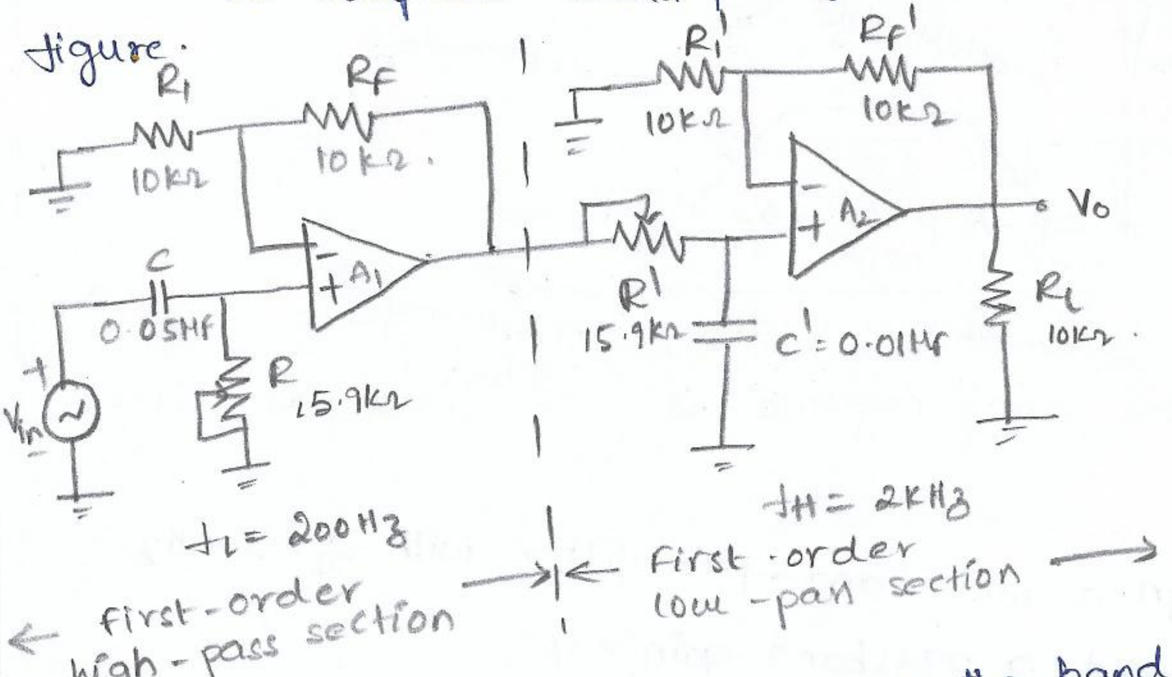
Then the gain of the high pass as well as low-pass sections could be set equal to 2.

$A_1 = A_2 = 2$.

$1 + \frac{R_F}{R_1} = 2 \quad R_F = R_1 = 10K\Omega$

$1 + \frac{R_F'}{R_1'} = 2 \quad R_F' = R_1' = 10K\Omega$

The complete -band pass filter show in below figure.



(b) The voltage gain magnitude of the band-pass filter is equal to the product of the voltage gain magnitudes of the high-pass and low-pass filters.

$$\left|\frac{V_o}{V_{in}}\right| = \frac{A_{FT}(f/f_L)}{\sqrt{\left[1+(f/f_L)^2\right]\left[1+(f/f_H)^2\right]}}$$

where A_{FT} = total pass band gain

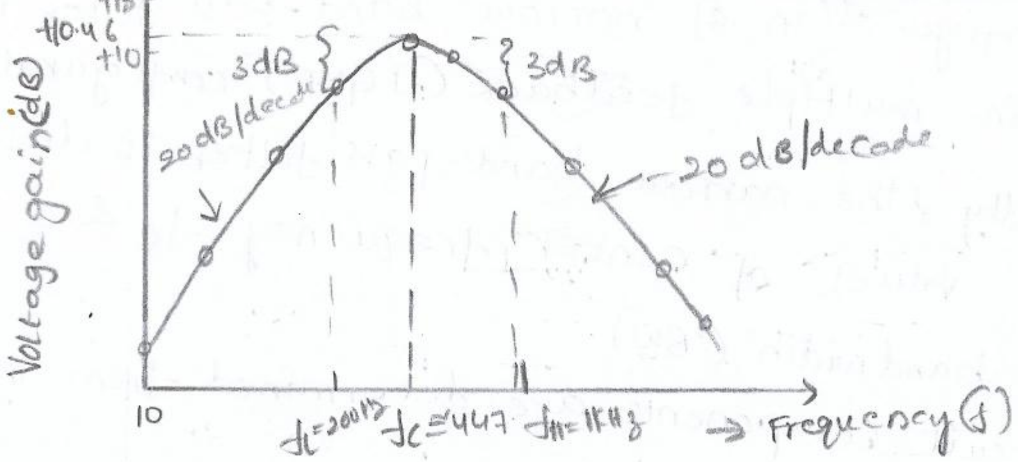
f = frequency of the input signal (Hz)

f_L = low cutoff frequency (Hz)

f_H = High cut-off frequency (Hz)

Here $A_{FT} = 4$, $f_L = 200Hz$ & $f_H = 1KHz$

Input frequency f (Hz)	Gain magnitude $\lvert V_o/V_{in}\rvert$	magnitude (dB) = $20 \log \lvert V_o/V_{in}\rvert$
10	0.1997	-13.99
30	0.5931	-4.54
100	1.780	5.01
200	2.774	8.861
447.2	3.33	10.46
700	3.151	9.969
1000	2.774	8.861
7000	0.5655	-4.951
10,000	0.3979	-8.004



Frequency response

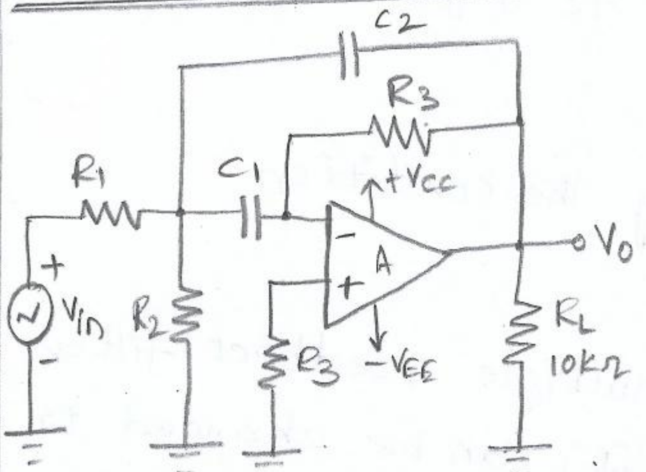
(3) $f_C = \sqrt{f_H f_L}$

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

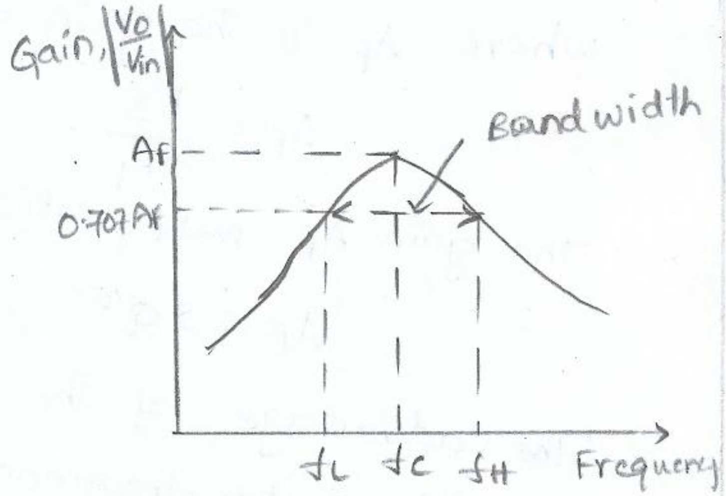
$Q = \frac{447.2}{1000 - 200} = 0.56 = \frac{f_C}{f_H - f_L}$

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

NARROW BAND-PASS FILTER:-



(a) NARROW BAND-PASS FILTER



The above figure is narrow band-pass filter. compared to all the filters, this filter is unique in the following respects:

1. It has two feedback paths, hence the name multiple-feedback filter.
2. The op-amp is used in the inverting mode.

The configuration of narrow-band-pass filter is Infinite gain multiple feedback (IGMF) configuration.

Generally, the narrow band-pass filter is designed for specific values of center frequency f_c & Q (or) f_c and bandwidth (BW).

The circuit components are determined from the following relationships:

choose $C_1 = C_2 = C$

$$R_1 = \frac{Q}{2\pi f_c C A_F}$$

$$R_2 = \frac{Q}{2\pi f_c C(2Q^2 - A_F)}$$

$$R_3 = \frac{\cancel{2}Q}{\cancel{2}\pi f_c C} \quad \frac{Q}{\pi f_c C} = \frac{1}{\pi C(BW)}$$

where A_F is the gain at f_c, given by

$$A_F = \frac{R_3}{2R_1}$$

The gain A_F must satisfy the condition

$$A_F < 2Q^2$$

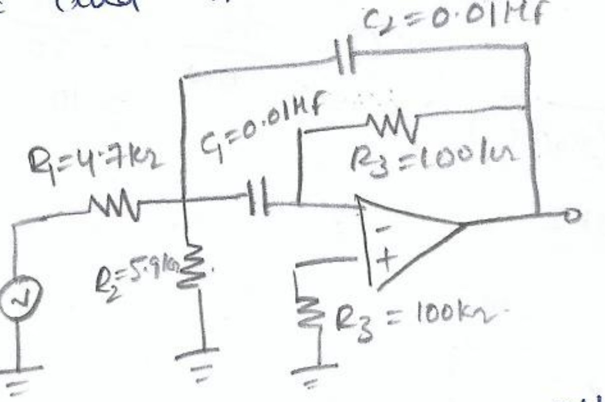
The advantage of the multiple feedback filter is that its center frequency (f_c) can be changed to a new frequency f_c' without changing the gain (or) bandwidth. This is accomplished simply by changing R_2 to R_2' so that

$$R_2' = R_2\left(\frac{f_c}{f_c'}\right)^2$$

problems:-

1) (a) Design the band pass filter show in below figure so that $f_c = 1KHz$, $Q = 3$ & $A_F = 10$.

(b) change the center frequency to $1.5KHz$, keeping A_F and the bandwidth constant.



(a) choose $C_1 = C_2 = C = 0.01\mu F$

$$R_1 = \frac{Q}{2\pi f_c C A_F} = \frac{3}{2\pi \times 1\times10^3 \times 0.01\times10^{-6} \times 10} = 4.77k\Omega$$

$$R_2 = \frac{Q}{2\pi f_c C(2Q^2 - A_F)} = \frac{3}{2\pi \times 1\times10^3 \times 0.01\times10^{-6}(2(3)^2 - 10)} = 5.97k\Omega$$

$$R_3 = \frac{Q}{\pi f_c C} = \frac{3}{\pi \times 1\times10^3 \times 0.01\times10^{-6}} = 95.5k\Omega.$$

Use $R_1 = 4.7K\Omega$ $R_2 = 5.9k\Omega$ & $R_3 = 100k\Omega$

(b) The value of R_2' required to change the center frequency from $1KHz$ to $1.5KHz$ is

$$R_2' = R_2\left(\frac{f_c}{f_c'}\right)^2 = 5.97k\left(\frac{1K}{1.5K}\right)^2 = 2.65k\Omega$$

Use $R_2' = 2.7K\Omega$

(2). Design a second order IGMF bandpass filter with the following specifications: $f_0 = 500Hz$, gain at resonance $= -5$ and band width $= 50Hz$. Use the ckt shown. Assume necessary data.

choose $C_1 = C_2 = 0.01\mu F$.

$$BW = \frac{f_c}{Q}$$

$$R_3 = \frac{Q}{\pi f_c C} = \frac{1}{\pi (BW) C} = \frac{1}{\pi \times 50 \times 0.01 \times 10^{-6}} = \underline{636.619\,k\Omega}$$

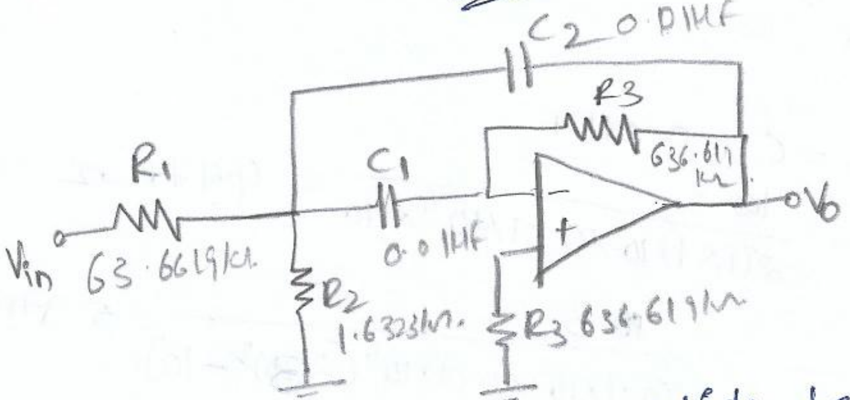
$$R_1 = \frac{Q}{2\pi f_c C A_F} = \frac{1}{2\pi (BW) C A_F} \quad \text{(or)} \quad \boxed{A_F = \frac{R_3}{2R_1}}$$

$$= \frac{1}{2\pi \times 50 \times 0.01 \times 10^{-6} \times 5}$$

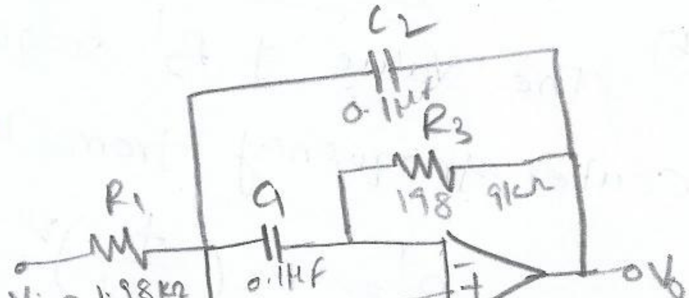
$$= \underline{63.6619\,k\Omega} \qquad Q = \frac{f_c}{BW} = \frac{500}{50} = 10$$

$$R_2 = \frac{Q}{2\pi f_c C(2Q^2 - A_F)} = \frac{1}{2\pi \times 50 \times (2\times 10^2 - 5)}$$

$$= \underline{1.6323\,k\Omega}$$



(3) Design a second order multiple feedback band pass filter using one op-amp, three resistors and two capacitors with $|A_F| = 50$. A centre frequency is 160 Hz and 3dB bandwidth of 16 Hz. Assume $C_1 = C_2 = C = 0.1\mu F$. Draw the designed circuit.



$A_F = 50$, $f_c = 160$, $BW = 16\,Hz$

$C_1 = C_2 = C = 0.1\mu F$

$$Q = \frac{f_c}{BW} = \frac{160}{16} = 10$$

$$R_1 = \frac{Q}{2\pi f_c C A_F} = \frac{10}{2\pi \times 160 \times 0.1 \times 10^{-6} \times 50}$$

$$= 1.98\,k\Omega$$

$$R_2 = \frac{Q}{2\pi f_c C(2Q^2 - A_F)} = \frac{10}{2\pi \times 160 \times 0.1 \times 10^{-6}(2\times 10^2 - 50)}$$

$$= 663.14\,\Omega$$

$$R_3 = \frac{Q}{\pi f_c C} = \frac{10}{\pi \times 160 \times 0.1 \times 10^{-6}}$$

$$= 198.94\,k\Omega.$$

BAND-REJECT FILTERS :-

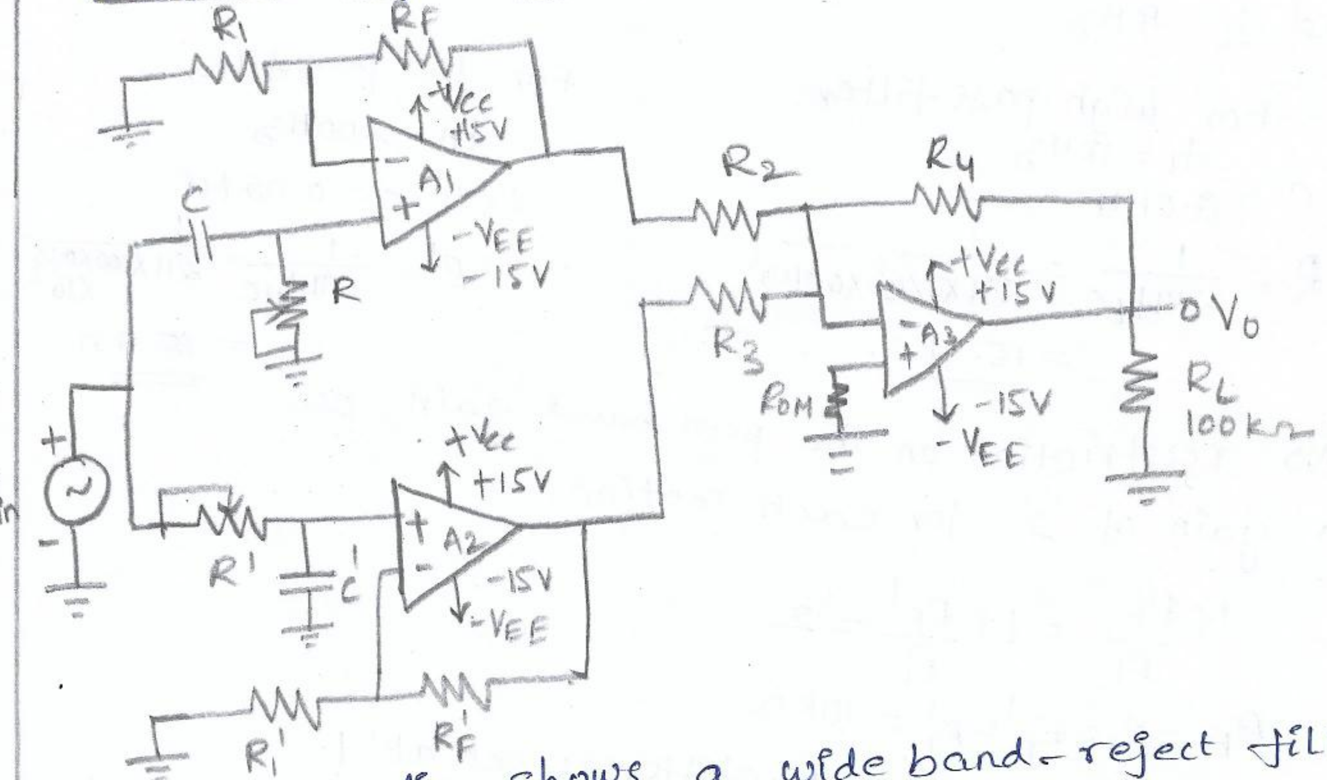
The band-reject filter is also called a band-stop (or) band-elimination filter. In this filter, frequencies are attenuated in the stopband while they are passed outside this band.

The band reject filters are two types:

(1) wide-band reject filter

(2) Narrow-band reject filter

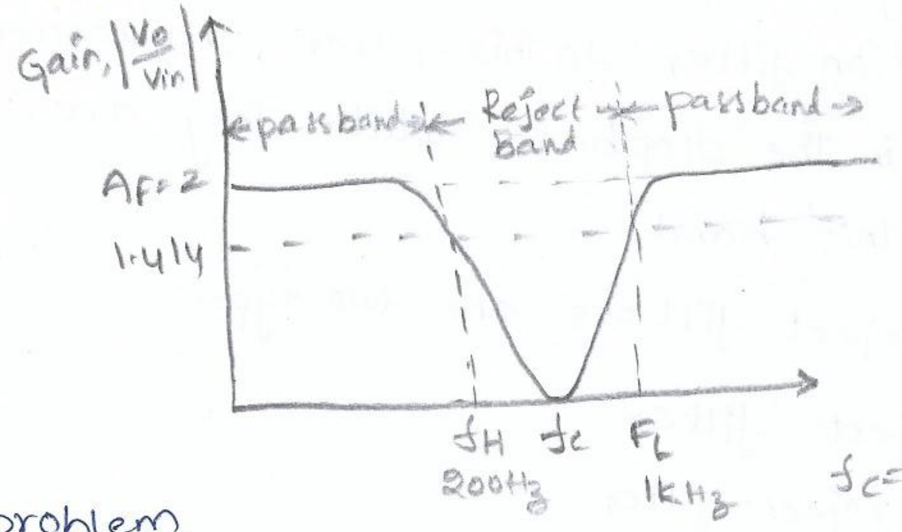
Wideband-reject filter :-



Above fig shows a wide band-reject filter using a low pass filter, a high-pass filter and a summing amplifier.

To realize a band-reject response, the low cut-off frequency f_L of the high pass filter must be larger than the high cut off frequency f_H of the low-pass filter. In addition, the passband gain of both the high-pass and low-pass sections must be equal.

The frequency response of wide band reject filter is shown below.



$f_c = \sqrt{f_H f_L} = 447.2Hz$

problem

① Design a wide band-reject filter having $f_H = 200Hz$ and $f_L = 1KHz$

For high pass filter

$f_L = 1KHz$

Let $C = 0.01\mu F$

$$R = \frac{1}{2\pi f_L C} = \frac{1}{2\pi \times 1 \times 10^3 \times 0.01 \times 10^{-6}}$$

$$= 15.9k\Omega$$

For low pass filter

$f_H = 200Hz$

Let $C' = 0.05\mu F$.

$$R' = \frac{1}{2\pi f_H C} = \frac{1}{2\pi \times 200 \times 0.05 \times 10^{-6}}$$

$$= 15.9k\Omega$$

∵ No restriction on the pass band gain, use a gain of '2' for each section.

$$1 + \frac{R_F}{R_1} = 1 + \frac{R_F'}{R_1'} = 2$$

$$R_F = R_1 = R_F' = R_1' = 10k\Omega$$

gain of the summing Amplifier is set at 1.

$R_2 = R_3 = R_4 = 10k\Omega$ & $R_{OM} = R_2 \| R_3 \| R_4 = 3.3k\Omega$

NARROWBAND-REJECT FILTER :-

The narrow band-reject filter, often called the notch filter, is commonly used for the rejection of a single frequency such as the 60 Hz power line frequency hum.

The most commonly used notch filter is the twin-T network shown in fig ⓐ. 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

$$f_N = \frac{1}{2\pi RC} \rightarrow ①$$

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 fig ⓑ. This is called active notch filter.

The frequency response of the active notch filter is shown in fig ⓒ.

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 eqn ①. For the best response, the circuit components should be very close to their indicated values

The second order Notch filter is shown in fig (a).

The transfer function of this circuit is given by

$$\frac{V_o}{V_{in}} = \frac{1-(f/f_N)^2}{1-(f/f_N)^2+\left(\frac{1}{Q}\right)j\,(f/f_N)}$$

$$f_N = \frac{1}{2\pi RC} = \text{Notch frequency}$$

$$Q = \frac{1}{2(2-A_F)}$$

$$A_F = 1+\frac{R_F}{R_1}$$

For the stability purpose, A_F must be less than 2.

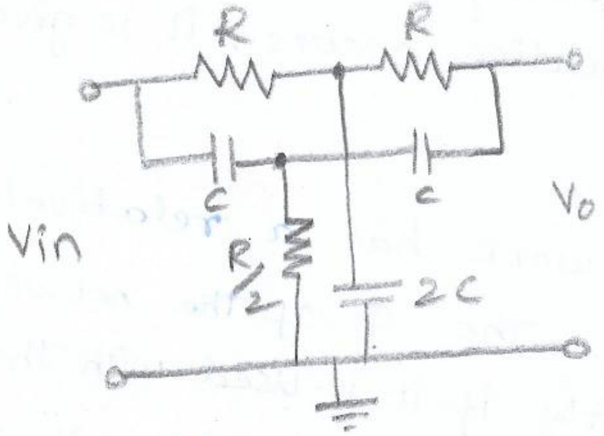
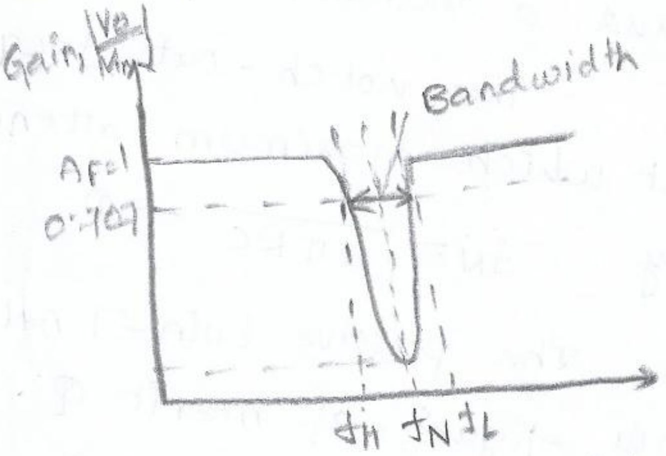


fig (a) Twin-T notch filter



fig(c) frequency Response of the active Notch Filter

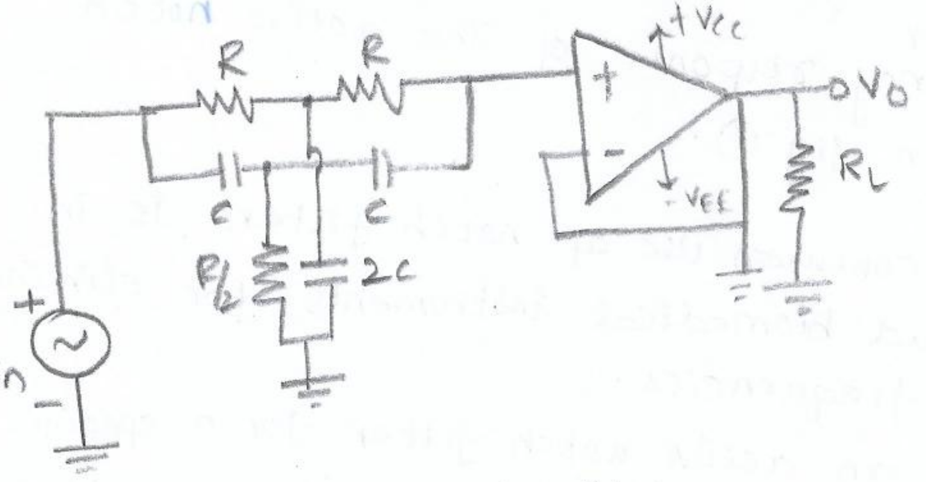


fig (b) Active Notch Filter

ALL-PASS FILTER :-

An all-pass filter passes all frequency components of the input signal without attenuation, while providing predictable phase shifts for different frequencies of the input signal.

When signals are transmitted over transmission lines, such as telephone wires, they undergo change in phase. To compensate for these phase changes, all-pass filters are required. The all-pass filters are also called delay equalizers (or) phase correctors.

Fig (a) shows an all pass filter where in $R_F = R_1$. The output voltage V_0 of filter can be obtained by using the superposition theorem:

V_{01} is the o/p when V_{in} applied to inverting i/p.

$$V_{01} = -\frac{R_F}{R_1} V_{in}$$

$R_F = R_1$

$$= -V_{in} \rightarrow (1)$$

V_{02} is the o/p when V_{in} applied to non-inverting i/p.

$$V_{02} = \left(1+\frac{R_F}{R_1}\right) V_1$$

$$= (1+1) V_1 = 2V_1 \rightarrow (2)$$

$$V_1 = \frac{-jX_C}{R-jX_C} V_{in} \rightarrow (3)$$

substitute (3) in (2).

$$V_{02} = 2\left(\frac{-jX_C}{R-jX_C}\right) V_{in} \rightarrow (4) \quad \because -j = 1/j, \; X_C = \frac{1}{2\pi f C}$$

The total output voltage is

$$V_0 = V_{01} + V_{02} = -V_{in} + 2V_{in}\left(\frac{-jX_C}{R-jX_C}\right)$$

$$= V_{in}\left(-1 + \frac{2}{j2\pi fRC+1}\right)$$

$$\frac{V_0}{V_{in}} = \frac{1-j2\pi fRC}{1+j2\pi fRC} \rightarrow (5)$$

where 'f' is the frequency of the input signal in hertz.

The magnitude of gain is

$$\left|\frac{V_o}{V_{in}}\right| = \frac{\sqrt{1+(-2\pi fRC)^2}}{\sqrt{1+(2\pi fRC)^2}} = 1 \text{ (unity)} \rightarrow (6).$$

The phase shift between V_o and V_{in} is a function of input frequency f. The phase angle ϕ is given by.

$$\phi = \tan^{-1}\left(-\frac{2\pi fRC}{1}\right) - \tan^{-1}\left(\frac{2\pi fRC}{1}\right)$$

$$= -\tan^{-1} 2\pi fRC - \tan^{-1} 2\pi fRC$$

$$= -2\tan^{-1}(2\pi fRC) \rightarrow (7)$$

where ϕ is in degrees, f in hertz, R in ohms, and 'C' in farads. Eqn (7) is used to find the phase angle ϕ if f, R, and C are known.

fig (b) shows a phase shift of 90° between the input V_{in} and output V_o. That is, V_o lags V_{in} by 90°. The phase angle ϕ changes for fixed values of R & C, the phase angle ϕ changes from 0 to −180° as the frequency 'f' is varied from 0 to ∞. In fig (a), if the positions of R & C are interchanged, the phase shift between input and output becomes positive. That is, output leads input V_{in}.

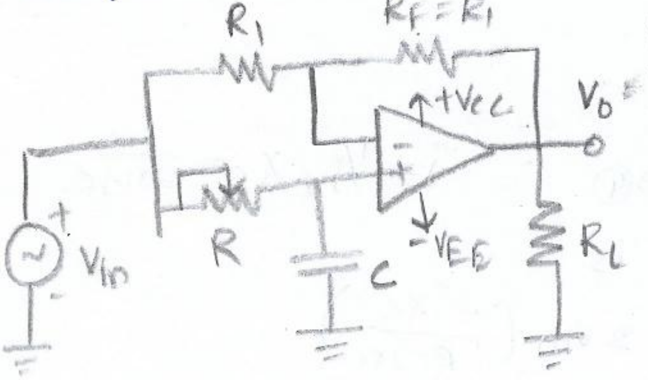
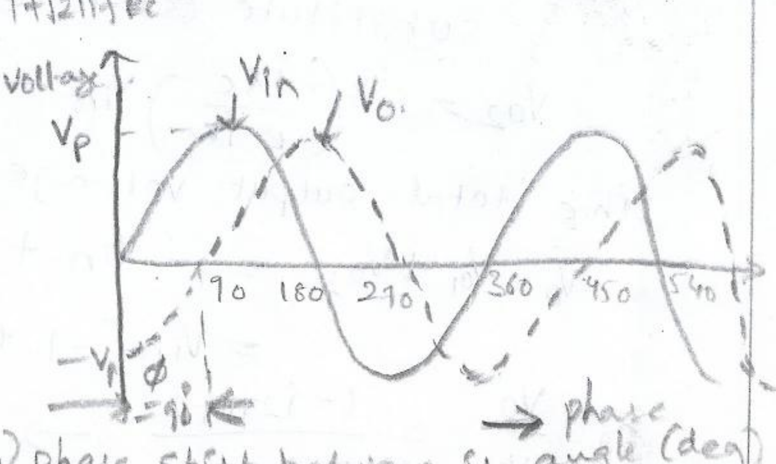


fig (a) All pass filter



(b) phase shift between i/p & o/p voltages

RC PHASE SHIFT OSCILLATOR :-

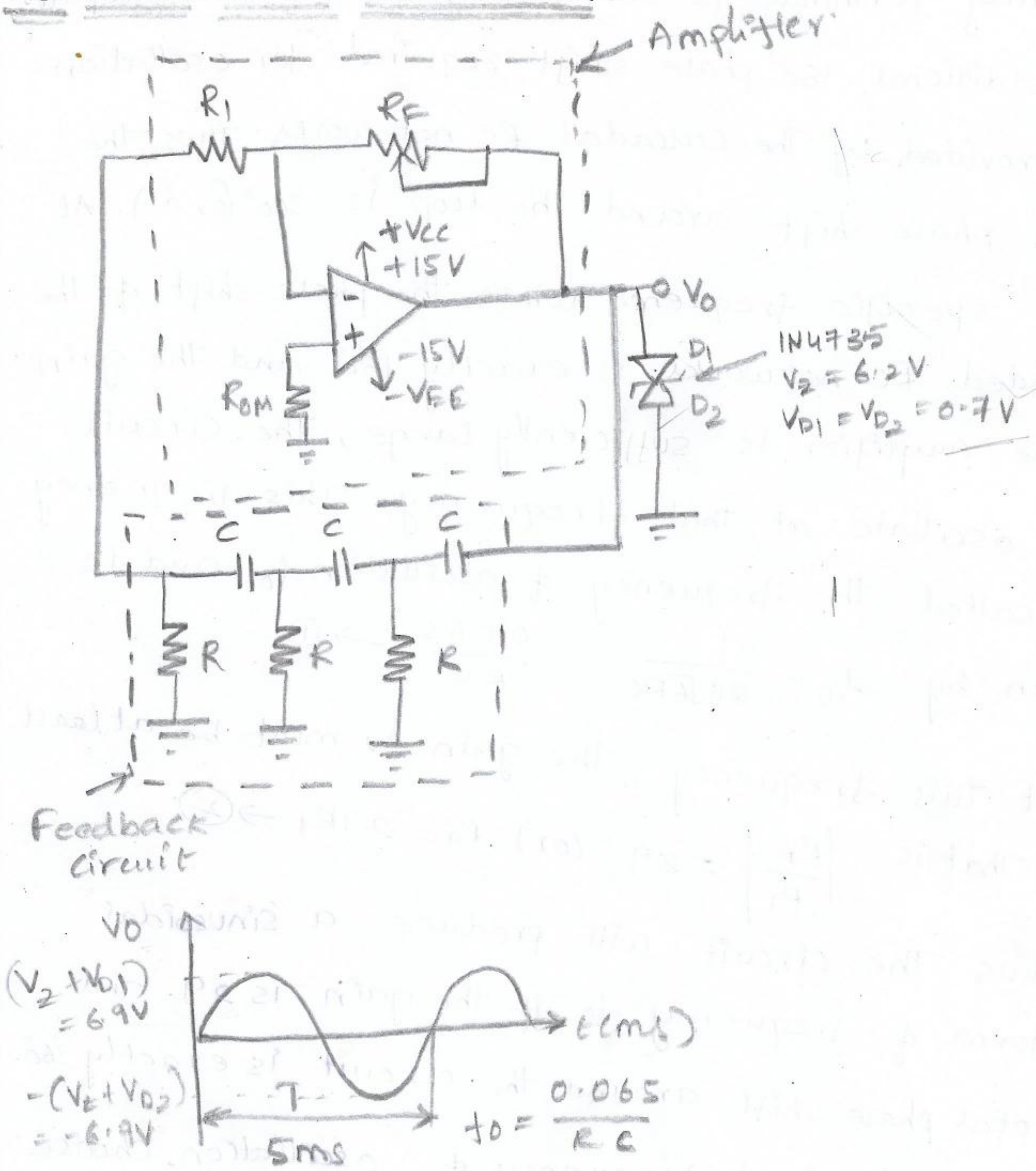


Fig ⓐ: phase shift oscillator and its o/p waveform

Fig ⓐ shows a phase shift oscillator, which consists of an op-amp as the amplifying stage and three RC cascaded networks as the feedback circuit. The feedback circuit provides feedback voltage from the output back to the input of the amplifier. The op-amp is used in the inverting mode; Therefore, any signal that appears at the

Inverting terminal is shifted by $180°$ at the output. An additional $180°$ phase shift required for oscillation is provided by the cascaded RC net-works. Thus the total phase shift around the loop is $360°$ (or $0°$). At some specific frequency when the phase shift of the cascaded RC networks is exactly $180°$ and the gain of the amplifier is sufficiently large, the circuit will oscillate at that frequency. This frequency is called the frequency of oscillation f_o and is given by $f_o = \frac{1}{2\pi\sqrt{6}RC} = \frac{0.065}{RC} \rightarrow (1)$

At this frequency, the gain A_v must be atleast 29. That is $\left|\frac{R_F}{R_1}\right| = 29$. (or) $R_F = 29R_1 \rightarrow (2)$

Thus the circuit will produce a sinusoidal waveform of frequency f_o if the gain is 29 and the total phase shift around the circuit is exactly $360°$.

For a desired frequency of oscillation, choose a capacitor C, and then calculate the value of R from eqn (1).

A desired output amplitude can be obtained with back-to-back zeners connected at the output terminal.

To show: $f_0 = \frac{1}{2\pi\sqrt{6}RC}$

$$\left|\frac{R_F}{R_1}\right| = 29$$

Proof:- First consider the feedback circuit consisting of RC combinations of the phase shift oscillator. For simplicity we use the Laplace transform. Thus, the circuit is represented in the s domain. Let us determine $V_f(s)/V_o(s)$ for the circuit.

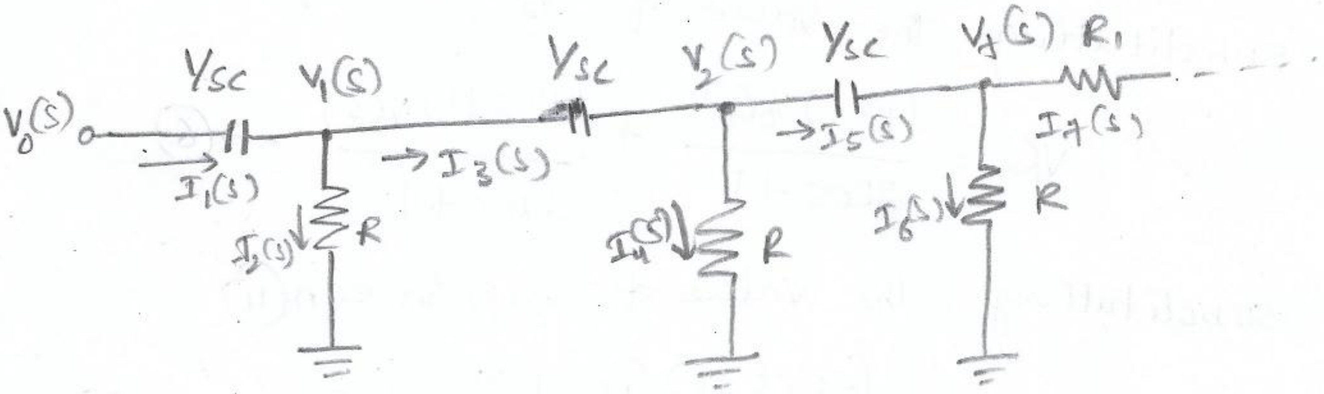


Fig (b). RC network of the phase shift oscillator transformed in to the s domain.

writing Kirchhoff's current Law (KCL) at node $V_1(s)$, we get

$$I_1(s) = I_2(s) + I_3(s)$$

$$\frac{V_o(s) - V_1(s)}{1/sC} = \frac{V_1(s)}{R} + \frac{V_1(s) - V_2(s)}{1/sC}$$

solving for $V_1(s)$, we have

$$V_1(s) = \frac{V_o(s) + V_2(s)(RCs)}{2RCs + 1} \rightarrow (3)$$

writing KCL at node $V_2(s)$,

$$I_3(s) = I_4(s) + I_5(s)$$

$$\frac{V_1(s) - V_2(s)}{1/sC} = \frac{V_2(s)}{R} + \frac{V_2(s) - V_f(s)}{1/sC}$$

solving for $V_1(s)$,

$$V_1(s) = \frac{(2RCs+1)\,V_2(s)}{RCs} - V_f(s) \rightarrow (4)$$

if $R_1 >> R$, then $I_7(s) \cong 0$ A. This means that $I_5(s) = I_6(s)$. Therefore, using the voltage-divider Rule

$$V_f(s) = \frac{R}{R + (1/sC)}\,V_2(s)$$

$$V_2(s) = \frac{(RCs+1)\,V_f(s)}{RCs} \rightarrow (5)$$

substituting the value of $V_2(s)$ in eqn (3)

$$V_1(s) = \frac{(RCs)\,V_0(s)}{2RCs+1} + \frac{(RCs+1)\,V_f(s)}{2RCs+1} \rightarrow (6)$$

substituting the value of $V_2(s)$ in eqn (4)

$$V_1(s) = \frac{(2RCs+1)(RCs+1)V_f(s)}{(RCs)(RCs)} - V_f(s) \rightarrow (7)$$

Equating eqns (6) & (7) and simplifying for $V_f(s)/V_0(s)$, we get

$$\frac{V_f(s)}{V_0(s)} = \frac{R^3C^3s^3}{(R^3C^3s^3 + 6R^2C^2s^2 + 5RCs + 1)} \rightarrow (8)$$

$$= B$$

Consider the op-amp part of the phase shift oscillator.

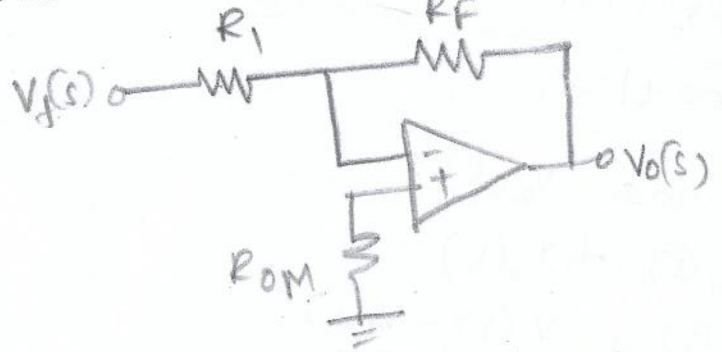


fig (c) op-amp part of the phase shift oscillator

The voltage gain of the op-amp is

$$A_v = \frac{V_o(s)}{V_f(s)}$$

$$= \frac{-R_F}{R_1} \rightarrow (9)$$

For an oscillator

$$(A_v)(B) = 1$$

using eqns (8) & (9)

$$\frac{-R_F}{R_1} \frac{R^3C^3s^3}{R^3C^3s^3 + 6R^2C^2s^2 + 5RCs + 1} = 1$$

substituting $s = j\omega$ and equating real and imaginary parts, respectively, we get

$$\left(\frac{-R_F}{R_1}\right)(-jR^3C^3\omega^3) = (-jR^3C^3s^3) - 6R^2C^2\omega^2 + j5RC\omega + 1$$

$$0 = -6R^2C^2\omega^2 + 1 \quad \text{(real part)}$$

$$\omega^2 = \frac{1}{6R^2C^2}$$

$$\boxed{f_0 = \frac{1}{2\pi\sqrt{6}RC}}$$

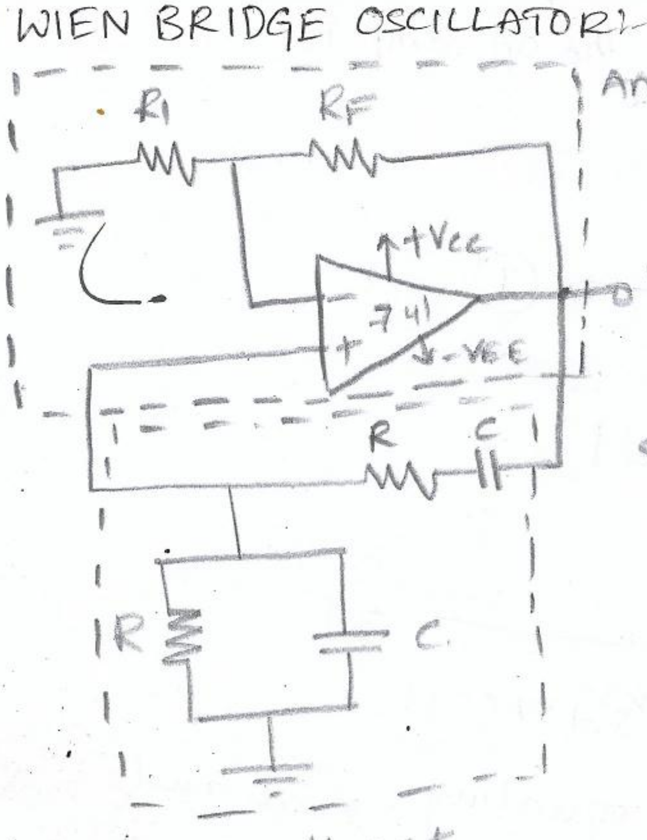
$$\left(\frac{-R_F}{R_1}\right)(-jR^3C^3s^3) = -jR^3C^3s^3 + j5RC\omega \quad \text{(imaginary part).}$$

$$-\frac{R_F}{R_1} = 1 - \frac{5}{R^2C^2\omega^2}$$

Substituting the value of ω^2 in the preceding eqn,

$$\boxed{\frac{R_F}{R_1} = 29}$$

WIEN BRIDGE OSCILLATOR:-



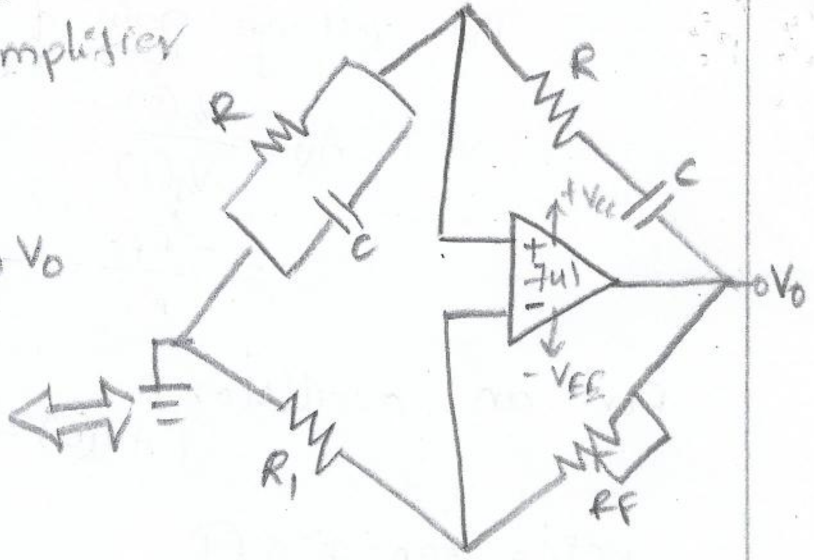


Fig ⓐ Wien bridge oscillator.

One of the most commonly used audio frequency oscillators is the wien bridge because of its simplicity and stability.

Fig ⓐ shows the wien bridge oscillator in which the wien bridge circuit is connected between The amplifier input terminals and the output terminal. The bridge has a series RC network in one arm and a parallel RC network in the adjoining arm. In the remaining two arms of the bridge, resistors R_1 and R_F are connected.

The phase angle criterion for oscillation is that the total phase shift around the circuit must be 0°. This condition occurs only when the bridge is balanced, that is, at resonance.

The frequency of oscillation f_o is exactly the resonant frequency of the balanced wien bridge and is given by

$$f_o = \frac{1}{2\pi RC} = \frac{0.159}{RC} \rightarrow (1)$$

Assuming that the resistors are equal in value, and capacitors are equal in value in the reactive leg of the wien bridge. At this frequency the gain required for sustained oscillations is given by

$$A_v = \frac{1}{B} = 3$$

That is $1 + \frac{R_F}{R_1} = 3$

(or) $R_F = 2R_1 \rightarrow (2)$

To show:

$$f_o = \frac{1}{2\pi RC}$$

$$R_F = 2R_1$$

Proof:- First consider the feedback circuit of the wien bridge oscillator. The circuit is transformed in the s domain and redrawn in fig (b)

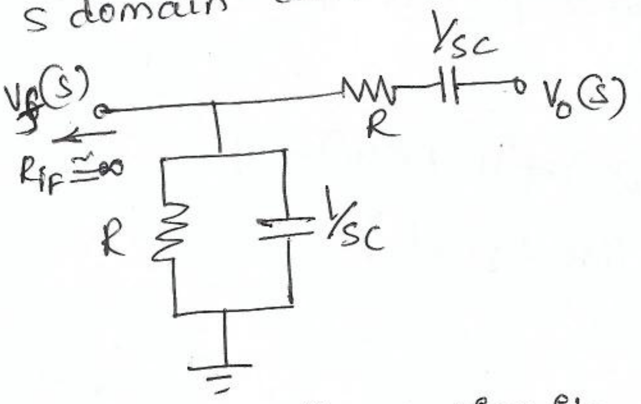


fig (b). Feedback circuit of the wien bridge oscillator represented in the time domain.

Using the voltage-divider rule.

$$V_f(S) = \frac{Z_p(S)\, V_o(S)}{Z_p(S) + Z_s(S)} \rightarrow (3)$$

where $Z_p(S) = R \| \frac{1}{sc}$

$$= \frac{R}{Rsc+1}$$

$$Z_s(S) = R + \frac{1}{sc} = \frac{Rsc+1}{sc}$$

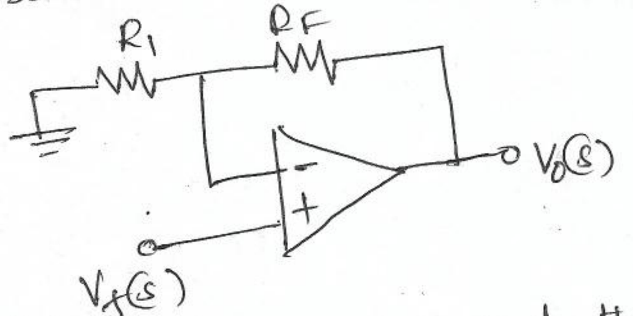
substituting $Z_p(s)$ and $Z_s(s)$ values, we get

$$V_f(s) = \frac{(RCs)\,V_o(s)}{(RCs+1)^2 + RCs}$$

$$B = \frac{V_f(s)}{V_o(s)}$$

$$= \frac{RCs}{R^2C^2s^2 + 3RCs + 1} \rightarrow (4)$$

Consider the op-amp part of the wien bridge oscillator. The circuit is redrawn in fig ©.



The voltage gain A_v of the op-amp is $A_v = \frac{V_o(s)}{V_f(s)} = 1 + \frac{R_F}{R_1} \rightarrow (5)$

Finally, the requirement for oscillation is

$$(A_v)(B) = 1$$

~~Substi~~ using equations (4) & (5), we have

$$\left(1 + \frac{R_F}{R_1}\right) \frac{RCs}{R^2C^2s^2 + 3RCs + 1} = 1$$

substituting $s = j\omega$ in this equation and then equating the real and imaginary parts, we get the frequency of oscillation f_o and the gain required for oscillation, as follows:

$$\left(1 + \frac{R_F}{R_1}\right) jRC\omega = (-R^2C^2\omega^2) + j(3RC\omega) + 1$$

$$\omega^2 = \frac{1}{R^2C^2} \text{ (real part)}$$

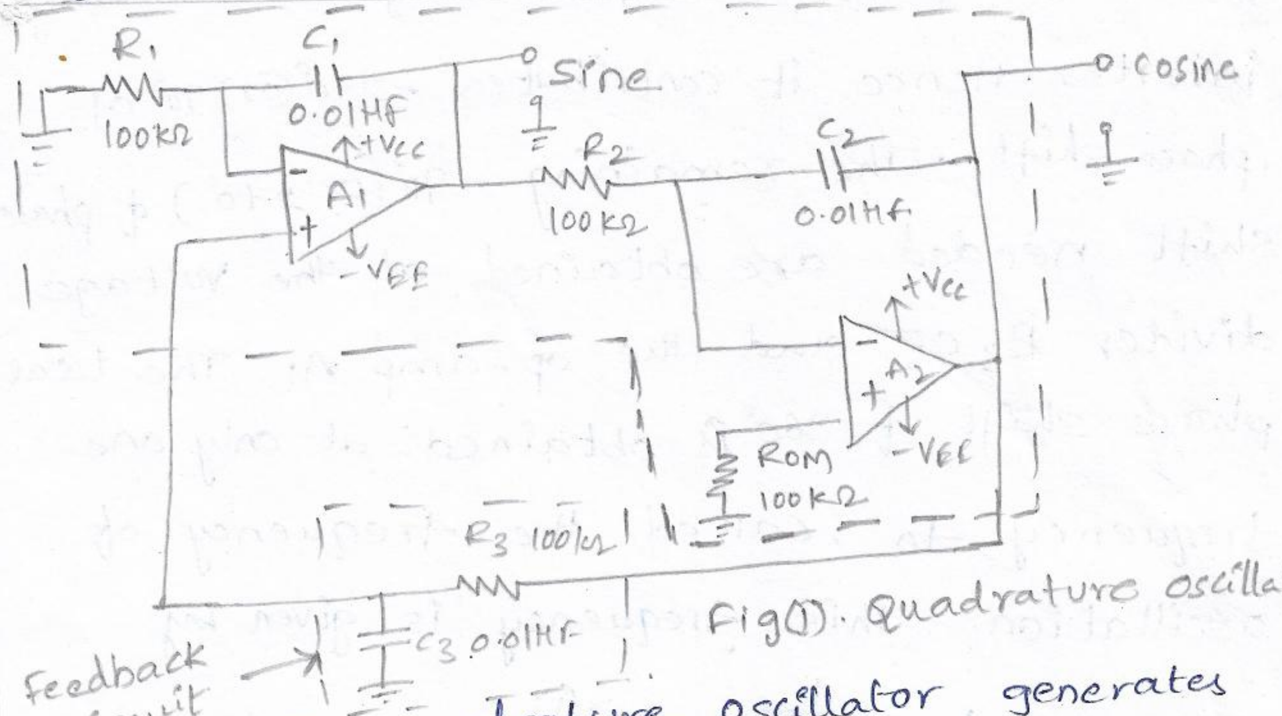
$$f_o = \frac{1}{2\pi RC}$$

and $\left(1 + \frac{R_F}{R_1}\right) jRC\omega = j(3RC\omega)$ (imaginary part)

$$1 + \frac{R_F}{R_1} = 3$$

$$R_F = 2R_1$$

QUADRATURE OSCILLATOR:-



Fig(1). Quadrature oscillator.

The quadrature oscillator generates two signals (sine and cosine) that are in quadrature, i.e, out of phase by 90°. Although the actual location of the sine and cosine is arbitrary, in the quadrature oscillator the o/p of A_1 is labeled a sine and the output of A_2 is a cosine. This oscillator requires a dual op-amp and three RC combinations. The first op-amp A_1 is operating in the noninverting mode and appears as a noninverting integrator. The second op-amp A_2 is working as a pure integrator. Furthermore, A_2 is followed by a voltage divider consisting of R_3 and C_3. The divider network forms a feedback circuit, whereas A_1 and A_2 form the amplifier stage.

The total phase shift of 360° around the loop required for oscillation is obtained in the following way:

The op-amp A_2 is a pure integrator and inverter. Hence it contributes $-270°$ (or) $90°$ of phase shift. The remaining $-90°$ (or $270°$) of phase shift needed are obtained at the voltage divider $R_3 C_3$ and the op-amp A_1. The total phase shift of $360°$ is obtained at only one frequency f_o, called the frequency of oscillation. This frequency is given by

$$f_o = \frac{1}{2\pi RC} \rightarrow ①$$

where $R_1 C_1 = R_2 C_2 = R_3 C_3 = RC$. At this frequency,

$$A_v = \frac{1}{B} = 1.414 \rightarrow ②$$ which is the second condition for oscillation.

Thus, to design a quadrature oscillator for a desired frequency f_o, choose a value of C, then from eqn ① calculate the value of R. To simplify design calculations, choose $C_1 = C_2 = C_3$ and $R_1 = R_2 = R_3$. In addition, R_1 may be a potentiometer in order to eliminate any possible distortion in the output waveforms.

(i) Design the Q.O with $f_o = 159\,Hz$

let $C = 0.01\mu F$

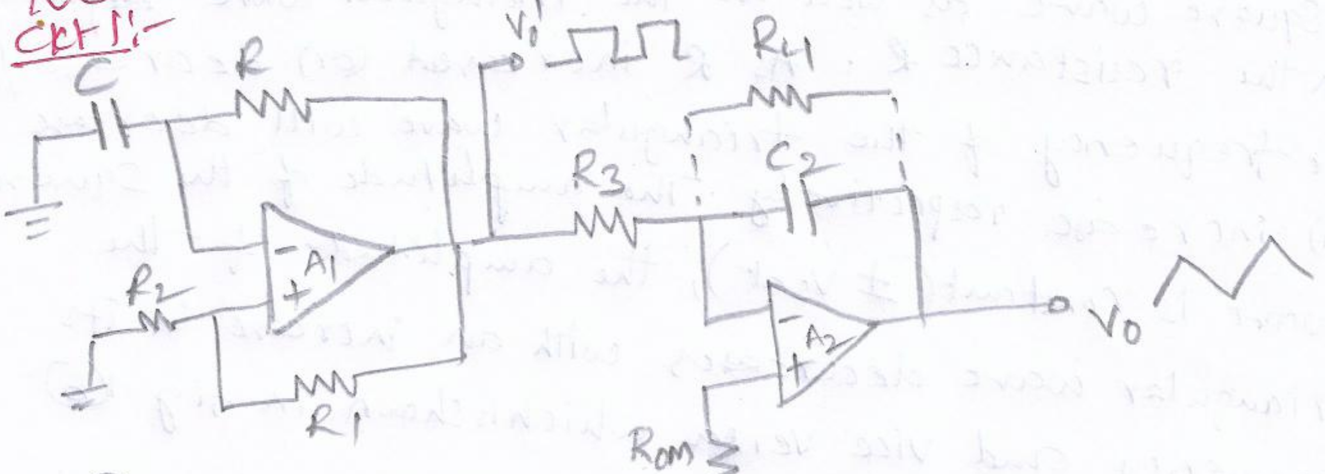
$$R = \frac{1}{2\pi f_o C} = \frac{1}{2\pi \times 159 \times 0.01 \times 10^{-6}} = 100k\Omega$$

$C_1 = C_2 = C_3 = 0.01\mu F$ & $R_1 = R_2 = R_3 = 100k\Omega$.

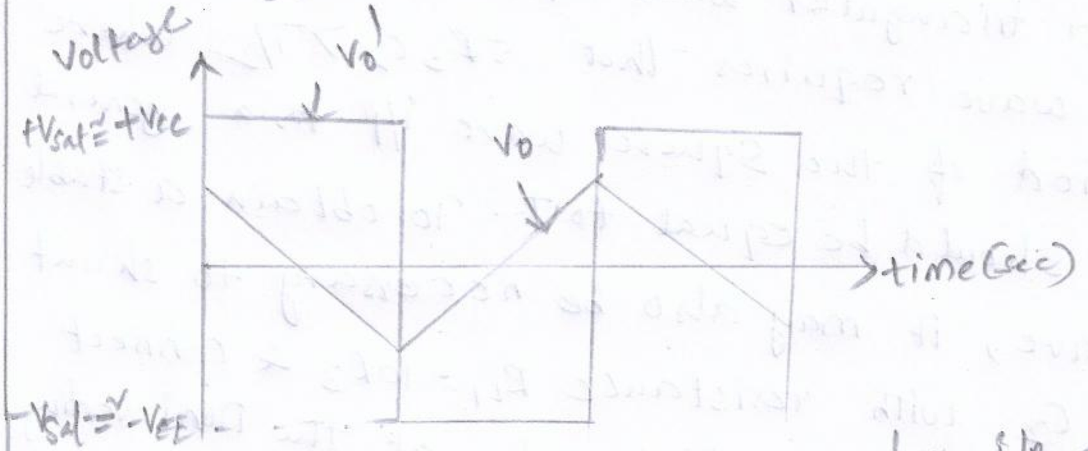
R_1 may be a $200\text{-}k\Omega$ potentiometer, which can be adjusted for undistorted o/p waveforms.

Triangular wave Generator:-

Ckt:-



1. (a) Traingular wave generator circuit



1. (b) Triangular wave generator I/p & O/p waveforms

The output waveform of integrator is triangular if its input is a square wave. The triangular wave generator can be formed by simply connecting an integrator to a square wave generator (Astable multi vibrator) which is shown in 1(a). This ckt requires a dual op-amp, two capacitors and atleast five resistors. The frequency of the square wave & triangular wave are the same.

$$f_0 = \frac{1}{2RC\ln(2R_1+R_2)/R_2}$$

if $R_2 = 1.16 R_1$

$$f_0 = \frac{1}{2RC}$$

For fixed R_1, R_2 & C values, the frequency of the square wave as well as the triangular wave depends on the resistance R. As R increased (or) decreased, the frequency of the triangular wave will decrease (or) increase respectively. The amplitude of the square wave is constant ($\pm V_{sat}$), the amplitude of the triangular wave decreases with an increase in its frequency and vice versa which is shown in fig (b)

The i/p of integrator A_2 is square wave, while its o/p is a triangular wave. For the o/p of A_2 to be a triangular wave requires that $5R_3C_2 > T/2$, where T is the period of the square wave i/p. As a general rule, R_3C_2 should be equal to T. To obtain a stable triangular wave, it may also be necessary to shunt the capacitor C_2 with resistance $R_4 = 10R_3$ & connect an offset voltage-compensating N/w at the noninverting terminal of A_2.

CKt 2:-

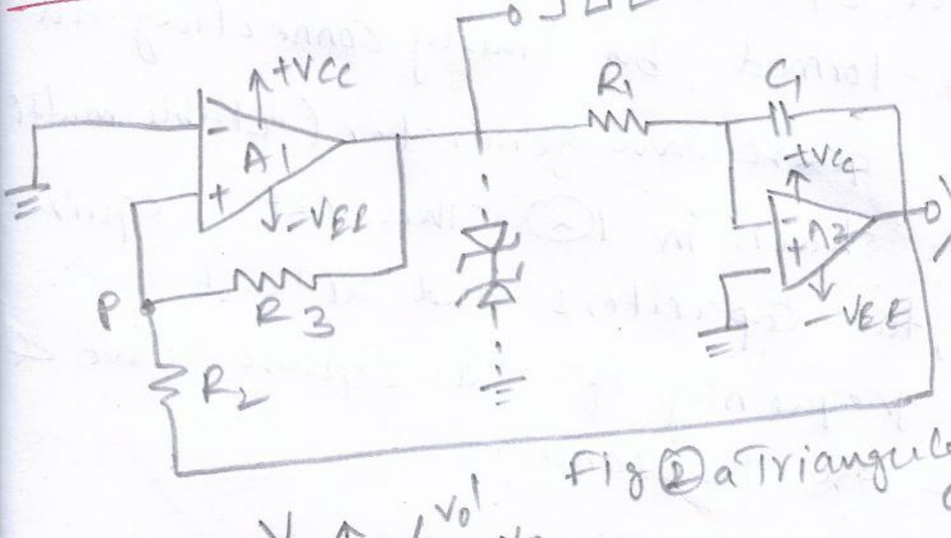


Fig (2)a Triangular wave generator circuit

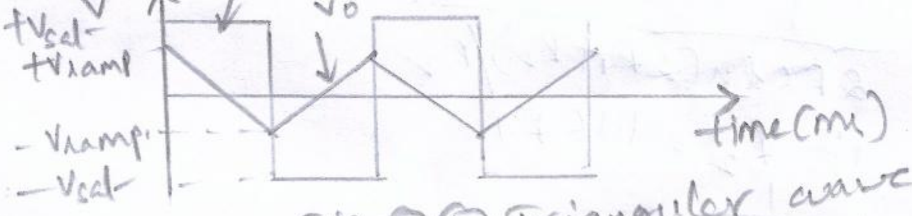


Fig (2)(b) Triangular wave generator o/p waveform

Another triangular generator, which requires fewer components is shown in fig 2(a). The generator consists of a comparator A_1 & integrator A_2. The comparator A_1 compares the voltage at point P continuously with the inverting i/p that is at 0V. when the voltage at P goes slightly below or above 0V, the o/p of A_1 is at negative (or) positive saturation level, respectively.

<u>Ckt operation</u>:- Let us set the ~~potent~~ o/p of A_1 at positive saturation $+V_{sat} \cong (+V_{cc})$. This $+V_{sat}$ is an i/p of integrator A_2. The o/p of A_2 will be a negative going ramp. Thus one end of the voltage divider $R_2 - R_3$ is the positive saturation voltage $+V_{sat}$ of A_1 & the other is negative going ramp of A_2. when a negative going ramp attains a certain value $-V_{ramp}$, point P is slightly below 0V; hence the o/p of A_1 will switch from positive saturation to negative saturation $-V_{sat} (\cong -V_{EE})$. This means that the o/p of A_2 will now stop going negatively & will begin to go positively. The o/p of A_2 will continue to increase until it reaches $+V_{ramp}$. At this time the point P is slightly above 0V. Therefore, the o/p of A_1 is switched back to the $+V_{sat}$. The sequence then repeats.

The frequency of the square wave & triangular wave are the same. The amplitude of the square wave is a fn of dc supply voltages. A desired amplitude can be obtained by using appropriate zeners at the o/p of A_1.

<u>Amplitude & frequency of triangular wave determination</u>:-

when the o/p of the comparator A_1 is $+V_{sat}$, the o/p integrator A_2 steadily decreases until it reaches $-V_{ramp}$. At this time the o/p of A_1 switches from $+V_{sat}$ to $-V_{sat}$. Just before this switching occurs, the voltage at point P is 0V.

$$+V_{sat}\frac{R_2}{R_2+R_3}+\left(-V_{ramp}\frac{R_3}{R_2+R_3}\right)=0$$

$$\frac{-V_{ramp}}{R_2}=-\frac{+V_{sat}}{R_3}$$

(or) $$-V_{ramp}=\frac{-R_2}{R_3}(+V_{sat}) \rightarrow (1)$$

Similarly $+V_{ramp}$, the o/p of A_2 at which the o/p of A_1 switches from $-V_{sat}$ to $+V_{sat}$, is given by.

$$-V_{sat}\frac{R_2}{R_2+R_3}+\left(+V_{ramp}\frac{R_3}{R_2+R_3}\right)=0.$$

$$+V_{ramp}=-\frac{R_2}{R_3}(-V_{sat}) \rightarrow (2)$$

The peak to peak o/p amplitude of the triangular wave is $V_o(pp)=+V_{ramp}-(-V_{ramp})$

$$V_o(pp)=-\frac{R_2}{R_3}(-V_{sat})-\left(-\frac{R_2}{R_3}+V_{sat}\right).$$

$$V_o(pp)=2\frac{R_2}{R_3}V_{sat} \rightarrow (3)$$

where $V_{sat}=|+V_{sat}|=|-V_{sat}|$, eqn (3) indicates that the amplitude of the triangular wave decreases with an increase in R_3.

The time it takes for the o/p waveform to swing from $-V_{ramp}$ to $+V_{ramp}$ (or from $+V_{ramp}$ to $-V_{ramp}$) is equal to half the time period $T/2$.

This time can be calculated from the integrator o/p equation by substituting $v_1=-V_{sat}$, $v_o=V_o(pp)$

$$V_o(pp)=\frac{-1}{R_1C_1}\int_0^{T/2}(-V_{sat})\,dt$$

$$=\frac{V_{sat}}{R_1C_1}\,T/2$$

Hence $\frac{T}{2} = \frac{V_o(pp)}{V_{sat}} R_1 C_1$

$$T = 2R_1C_1 \frac{V_o(pp)}{V_{sat}} \rightarrow (4)$$

Sub eqn (3) in (4)

$$T = 2R_1C_1 \frac{\left(\frac{2R_2}{R_3} V_{sat}\right)}{V_{sat}}$$

$$T = \frac{4R_1C_1R_2}{R_3} \rightarrow (5)$$

The frequency of oscillation is

$$f_o = \frac{R_3}{4R_1C_1R_2} \rightarrow (6)$$

The eqn (6) shows that the frequency of oscillations f_o increases with an increase in R_3.

The triangular wave generator is designed for a desired amplitude & frequency f_o by using equations (3) & (6).

sawtooth wave generator:-

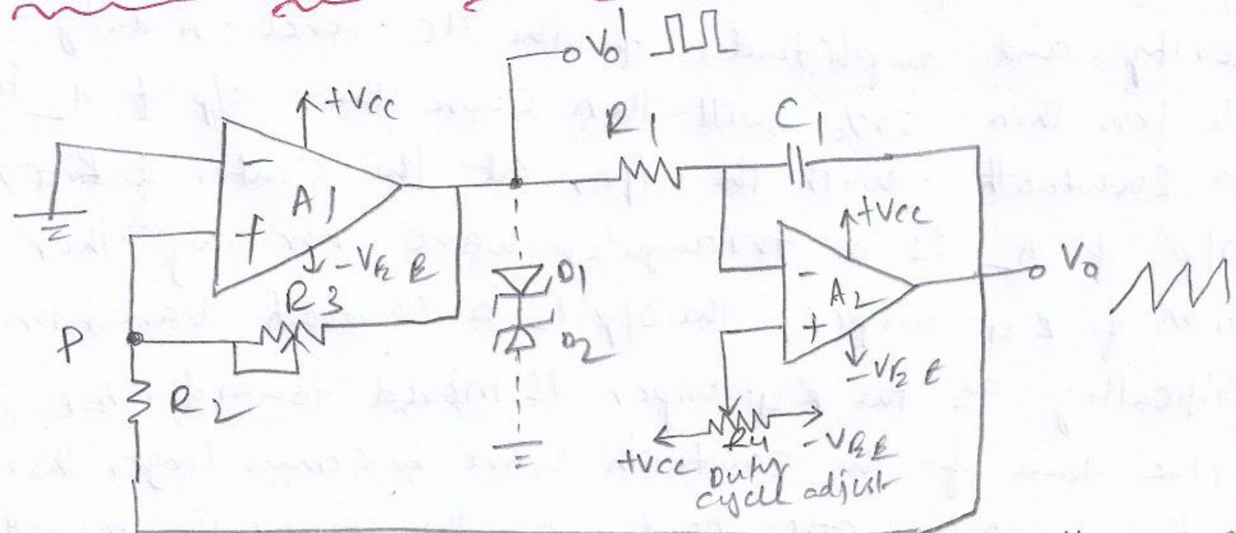


Fig 3(a) sawtooth wave generator ckt

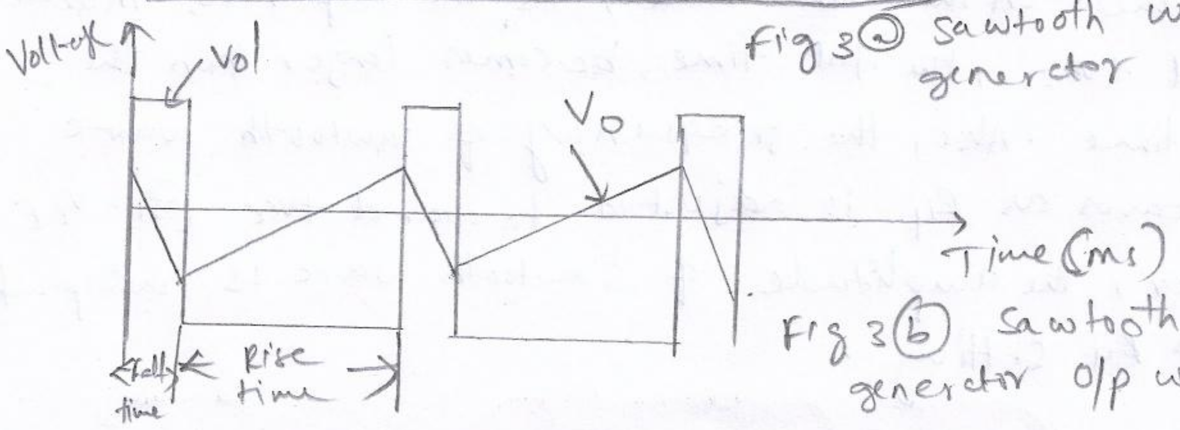


Fig 3(b) sawtooth wave generator o/p waveform

The difference between the triangular & sawtooth wave forms is that the rise time of the triangular wave is always equal to its fall time. That is, the same amount of time is required for the triangular wave to swing from $-V_{ramp}$ to $+V_{ramp}$ as from $+V_{ramp}$ to $-V_{ramp}$. On the other hand, the sawtooth has unequal rise & fall times. That is, it may rise +ve ly many times faster than it falls -ve ly, viceversa.

→ The triangular wave generator in 2(a) can be converted into a sawtooth wave generator by injecting a variable dc voltage in to the noninverting terminal of integrator A_2. This can be accomplished by using the potentiometer & connecting it to the $+V_{CC}$ & $-V_{EE}$ as shown in fig 3(a).

→ Depending on the R_4 setting, a certain dc level is inserted in the o/p of A_2. Now, suppose that the o/p of A_1 is a square wave & the potentiometer R_4 is adjusted for a certain dc level. This means that the o/p of A_2 will be a triangular wave, riding on some dc level that is a function of the R_4 setting. The duty cycle of the square wave will be determined by the polarity and amplitude of the dc level. A duty cycle less than 50% will then cause the o/p of A_2 to be a sawtooth. With the wiper at the center of R_4, the o/p of A_2 is a triangular wave. For any other position of R_4 wiper, the o/p is a sawtooth waveform. Specifically as the R_4 wiper is moved toward $-V_{EE}$, the rise time of the sawtooth wave becomes longer than fall time. On the other hand, as the wiper is moved toward $+V_{CC}$, the fall time becomes longer than the rise time. Also, the frequency of sawtooth wave decreases as R_4 is adjusted toward $+V_{CC}$ (or) $-V_{EE}$. However, the amplitude of sawtooth wave is independent of the R_4 setting.

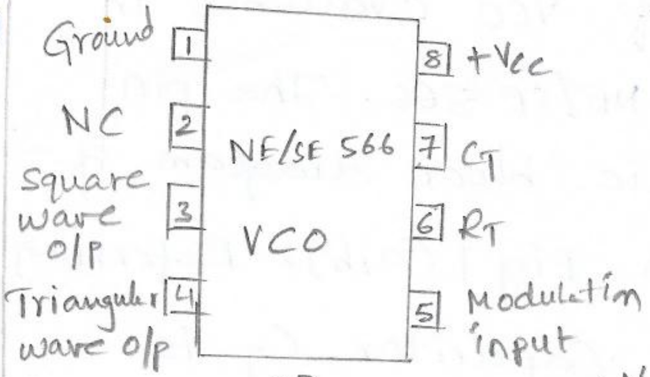
VOLTAGE CONTROLLED OSCILLATOR:-

Voltage to frequency converter

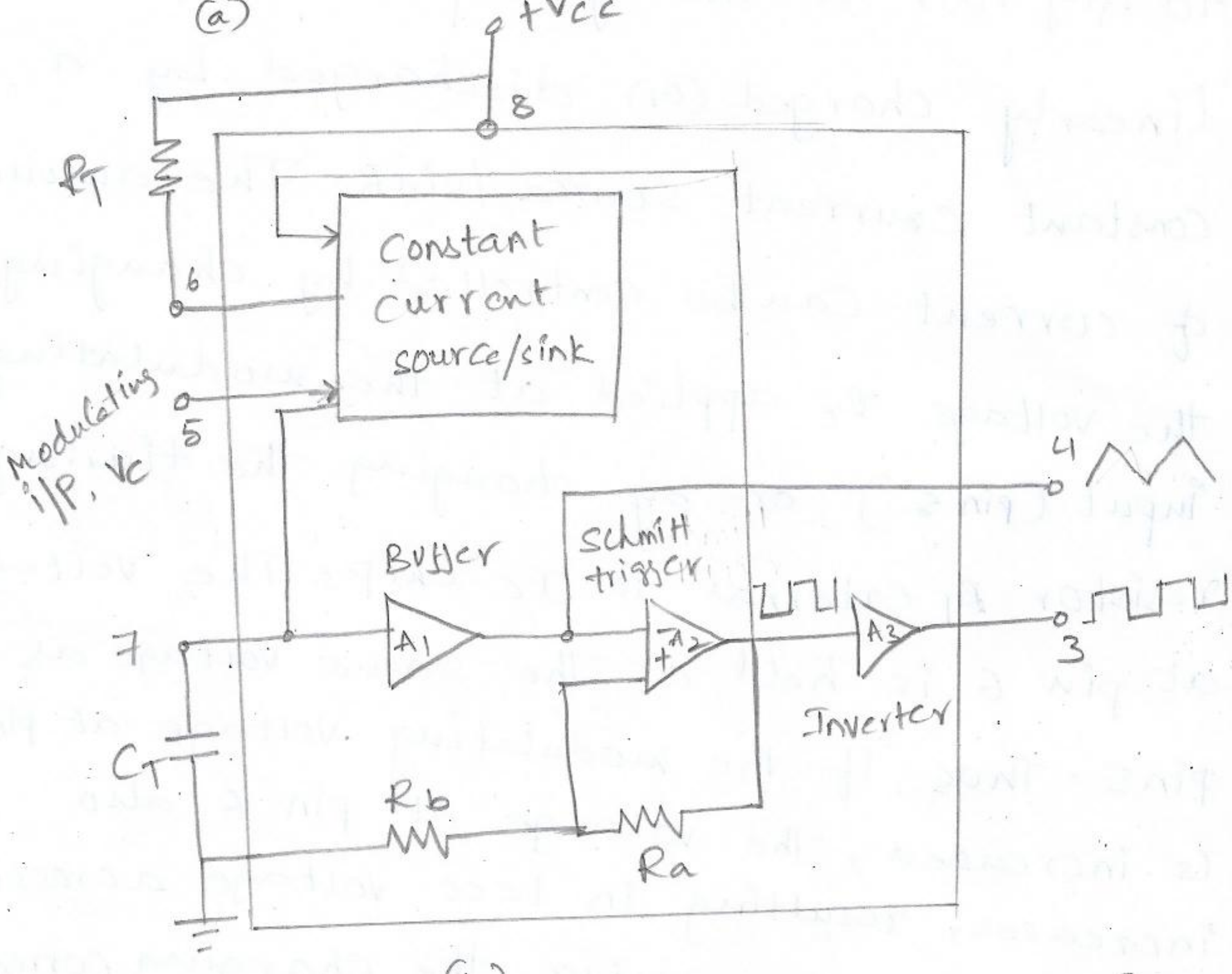
A Common type of VCO available in IC form is signetics NE/SE 566. The pin configuration and basic block diagram of 566 VCO are shown in Fig 1 (a,b). Referring to fig 1(b), a timing Capacitor C_T is linearly charged (or) discharged by a constant current source/sink. The amount of current can be controlled by changing the voltage v_c applied at the modulating input (pin 5) or by changing the timing resistor R_T external to IC chip. The voltage at pin 6 is held at the same voltage as pin5. Thus, if the modulating voltage at pin5 is increased, the voltage at pin 6 also increases, resulting in less voltage across R_T and thereby decreasing the charging current.

A small capacitor of 0.01μF should be connected between pins and 6 to eliminate possible oscillations. A VCO is commonly used in converting low frequency signals such as EEG, EKG into an audio frequency range. These audio signals can be transmitted over telephone lines (or) a two way radio communication systems for diagnostic purposes (or) can be recorded on a magnetic tape for further reference.

(14)



(a)



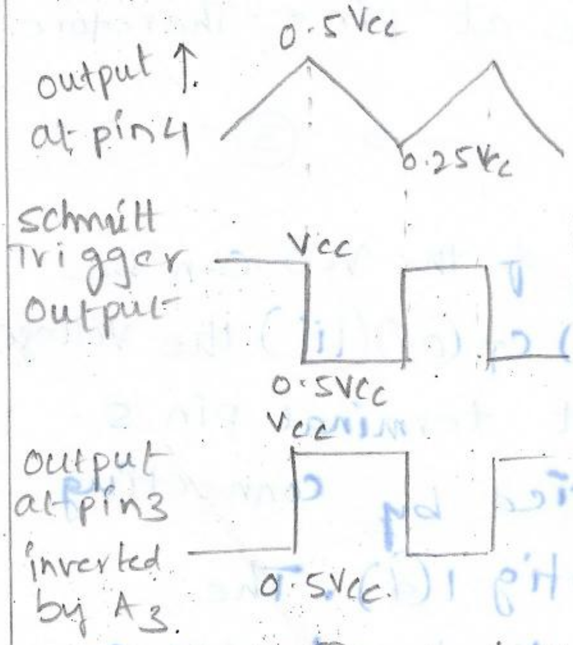
(b)

Fig 1 voltage controlled oscillator (a) pinconfiguration (b) block diagram

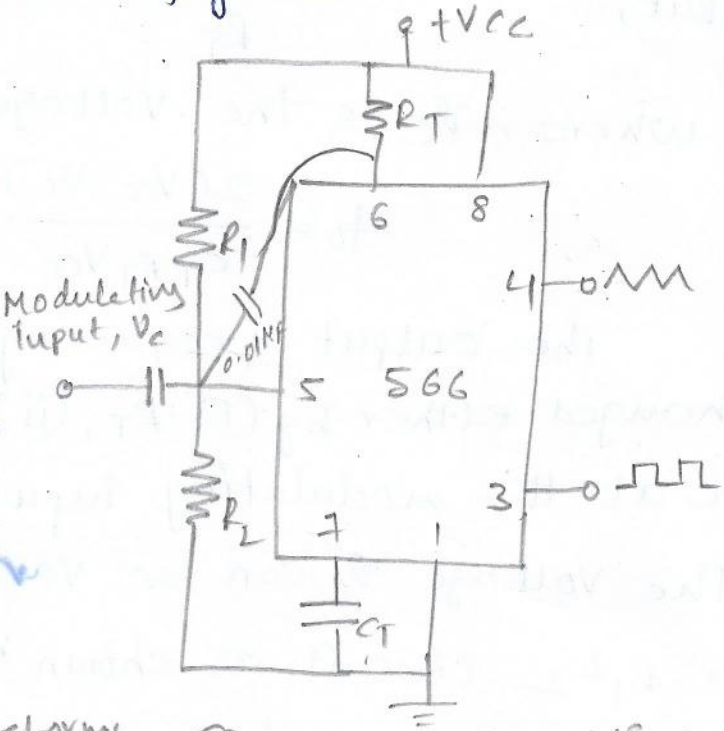
The voltage across the capacitor C_T is applied to the inverting input terminal of schmitt trigger A_2 via buffer amplifier A_1. The output voltage of the schmitt trigger is designed to swing of Vcc and 0.5 Vcc. If $R_a = R_b$ in The positive feedback loop, the voltage at the Non-inverting input terminal of A_2 swings from $0.5V_{cc}$ to 0.25 Vcc.

(46)

In fig 1(c), when the voltage on the capacitor C_T exceeds $0.5V_{cc}$ during charging, the output of the schmitt trigger goes LOW ($0.5V_{cc}$). The capacitor now discharges and when it is at $0.25V_{cc}$, the output of schmitt trigger goes HIGH (V_{cc}). Since the source and sink currents are equal, capacitor charges and discharges for the same amount of time. This gives a triangular voltage waveform across C_T which is also available at pin 4. The square wave output of the schmitt trigger is inverted by inverter A_3 and is available at pin3. The inverter A_3 is basically a current amplifier used to drive the load. The output waveforms are shown in fig 1(c).



1 (c) output waveforms



1 (d) Typical connection diagram.

(46) The output frequency of the VCO can be Calculated as follows:

The total voltage on the capacitor changes from $0.25 V_{cc}$ to $0.5 V_{cc}$. Thus $\Delta v = 0.25 V_{cc}$. The capacitor charges with a constant current source.

$$\frac{\Delta v}{\Delta t} = i / C_T$$

$$\frac{0.25 V_{cc}}{\Delta t} = i / C_T$$

$$\Delta t = \frac{0.25 V_{cc} C_T}{i} \longrightarrow (1)$$

The time period T of the triangular waveform $= 2\Delta t$. The frequency of oscillator f_o is

$$f_o = \frac{1}{T} = \frac{1}{2\Delta t} = \frac{i}{0.5 V_{cc} C_T}$$

But,

$$i = \frac{V_{cc} - v_c}{R_T} \longrightarrow (2)$$

where, v_c is the voltage at pin 5. Therefore

$$f_o = \frac{2(V_{cc} - v_c)}{C_T R_T V_{cc}} \longrightarrow (3)$$

The output frequency of the VCO can be changed either by (i) R_T, (ii) C_T (or) (iii) the voltage v_c at the modulating input terminal pin 5. The voltage v_c can be varied by connecting a $R_1 R_2$ circuit as shown fig 1(d). The components R_T and C_T are first selected so that VCO output frequency lies in the centre of the operating frequency range. Now the modulating input voltage is usually varied from

0.75 Vcc to Vcc which can produce a frequency variation of about 10 to 1. With no modulating input signal, if the voltage at pin 5 is biased at (7/8) Vcc, equation (3) gives the VCO output frequency as, (47)

$$f_o = \frac{2\left(V_{CC} - (7/8)V_{CC}\right)}{C_T R_T V_{CC}} = \frac{1}{4R_T C_T} = \frac{0.25}{R_T C_T} \rightarrow (4)$$

Voltage to Frequency conversion factor :-

A parameter of importance for VCO is voltage to frequency conversion factor k_v and is defined as

$$k_v = \frac{\Delta f_o}{\Delta v_c}$$

Here Δv_c is the modulation voltage required to produce the frequency shift Δf_o for a VCO. If we assume that the original frequency is f_o and the new frequency is f_1, then

$$\Delta f_o = f_1 - f_o = \frac{2\left(V_{CC} - v_c + \Delta v_c\right)}{C_T R_T V_{CC}} - \frac{2\left(V_{CC} - v_c\right)}{C_T R_T V_{CC}}$$

$$= \frac{2\Delta v_c}{C_T R_T V_{CC}} \rightarrow (5)$$

(or) $$\Delta v_c = \frac{\Delta f_o C_T R_T V_{CC}}{2} \rightarrow (6)$$

putting the value of $C_T R_T$ from eq (4)

$$\Delta v_c = \Delta f_o V_{CC} / 8 f_o$$

(or) $$K_V = \frac{\Delta f_o}{\Delta V_C} = \frac{8 f_o}{V_{CC}} \rightarrow (7)$$