CHAPTER - IV

SWITCHED CAPACITOR NETWORKS [105][106]

4.1 Introduction:

MOS switched capacitor filters are receiving increasing attention due to recent advances in IC technology. Gregorian [83] has described first, second and third order switched capacitor filter sections. These sections can be cascaded to realize higher order sharply selective filters. The filters can be designed directly in the sampled data domain, the filters response is free from parasitic capacitance. Martin and Sedra [100] have presented the design of switched capacitor band pass filters using the bilinear Z-transform. Szentirmai and Temes [108] have described several new building blocks which can realize the general second order sampled data transfer function. These blocks can be used to realize more complex functions by cascading them. The present chapter describes the design of second order switched capacitor filters using cascaded sections. The design method does not require any approximation like $f \ll f_c$ where $f$ and $f_c$ are signal and clock frequencies. The circuits presented have less number of MOS switches as compared to Gregorian's filters. It may be predicted that the configuration presented here will require less space in monolithic fabrication without impairing the quality as compared to Gregorian's filters.
Switched-capacitor filters (s.c.f.) realized with unity-gain amplifiers are receiving increasing attention due to several standpoints. An important contribution in this direction is due to Malawka and Ghausi [104]. The circuits proposed by them are capable of realizing high-Q filters with a good dynamic range. Recently Anand Mohan et al [76] have proposed some configurations capable of realizing second-order highpass and bandpass characteristics. Although each of the networks proposed by them is canonic, it suffers from three major drawbacks: (1) the pole -Q is restricted by some constraints which make it useful only at low Q-values, (2) the spread of capacitance values is very high, particularly for Q > 10 (It is an undesirable feature for integration) and (3) the dynamic range is poor.

The purpose of this chapter is to describe some s.c.f. configurations realizing second-order transfer functions using unity-gain amplifiers. The proposed networks have the following attractive features: (a) There is no restriction on pole-Q, (b) the spread of capacitance is reasonable for integration, (c) the circuits have good dynamic range and (d) the Q-sensitivity and \( \omega_0 \) -sensitivity are of the order of typical low sensitivity active circuits, (e) the circuits are insensitive to bottom plate parasitics.

4.2  Switched capacitor filters using cascade sections

4.2.1  Basic building blocks:

The basic building-blocks which will be used here for realizing the second-order transfer functions are switched capacitor.
Fig. 4.1 shows the switched capacitor integrators. The transfer functions in Z-domain are

\[ \frac{V_{\text{out}}(z)}{V_{\text{in}}(z)} = -az \quad \text{...(4.1)} \]

and

\[ \frac{V_{\text{out}}(z)}{V_{\text{in}}(z)} = \frac{\alpha}{z-1} \quad \text{...(4.2)} \]

For inverting (Fig. 4.1(a) and noninverting (Fig. 4.1(b) integrators respectively. The integrators are insensitive to parasitic capacitances.

Using the above building blocks, a scheme has been proposed in Fig. 4.2 for realizing biquadratic transfer function in Z-domain. A straight-forward analysis of the structure of Fig. 4.2 yields the following transfer function.

\[ \frac{V_{o}(z)}{V_{i}(z)} = -\beta z \left[ z^2 + 2z \left\{ 1 - \frac{\alpha_2(\alpha_0 + \beta_0)}{2} \right\} + (1 - \alpha_2\beta_0) \right] \quad \text{...(4.3)} \]

which is biquadratic in Z-domain.

4.2.2 Design Procedure:

The indirect way of designing a sample-data (or digital) filter requires that a suitable prototype continuous (analog) filter transfer function \( G(S) \), is designed, and subsequently that is transformed via an appropriate S-plane to Z-plane mapping to give a corresponding sampled-data transfer function, \( G(Z) \).
The bilinear Z-transform

\[ s = \frac{2}{T} \left( \frac{z-1}{z+1} \right) \]  

is one suitable method of mapping the transform \( G(s) \) to \( G(Z) \).

A general continuous-time second order transfer function is of the form

\[ G(s) = \frac{a_2 s^2 + a_1 s + a_0}{s^2 + \omega_0^2 s + \omega_0^4} \]  

Using the bilinear Z-transform of (4.4) in (4.5) one gets

\[ G(z) = \frac{\left( m + \frac{z^2 - 2z \left( \frac{1}{m} (4a_2 - a_0 T^2) + \frac{1}{m} (4a_2 - 2a_1 T + a_0 T^2) \right)}{z^2 - 2 \left( \frac{2}{n} (4 - \omega_0^2 T^2)z + \frac{1}{n} (4 - \frac{2\omega_0 T}{Q} + \omega_0^2 T^2) \right) \right)}{z^2 - 2z \left( \frac{1}{m} (4a_2 - a_0 T^2) + \frac{1}{m} (4a_2 - 2a_1 T + a_0 T^2) \right)} \]  

where

\[ m = 4a_2 + 2a_1 T + a_0 T^2 \]  

\[ n = 4 + \frac{2\omega_0 T}{Q} + \omega_0^2 T^2 \]  

comprising equ. (4.6) with equ. (4.3) one gets the following identities

\[ \alpha_2 \beta_2 \beta_1 = \frac{4\omega_0 T}{nQ} \]  

\[ \alpha_1 \beta_2 ^2 = \frac{4\omega_0^2 T^2}{n} \]  

\[ \beta_2 = \frac{m}{n} \]  

\[ \alpha_3 \beta_2 = \frac{4a_1 T}{m} \]  

\[ \alpha_0 \alpha_2 = \frac{4a_0 T^2}{m} \]
One can see that the number of unknowns are more than the number of equations. Therefore, the solution is not unique. Assigning any suitable positive value to $a_2$, one can evaluate $a_0$, $a_1$, $b_0$, and $b_1$ from equ. (4.9) to equ. (4.13). The design equations for some important filter characteristics are as shown:

1. Low pass

\[
\begin{align*}
\beta_0 &= 0 \\
a_0a_2 &= 0
\end{align*}
\]  

\[\ldots \ldots (4.14)\]

2. Band-pass

\[
\begin{align*}
a_0 &= 0 \\
\alpha_0 &= 0
\end{align*}
\]  

\[\ldots \ldots (4.15)\]

3. High-pass

\[
\begin{align*}
\beta_0 &= 0 \\
a_0 &= 0
\end{align*}
\]  

\[\ldots \ldots (4.16)\]

4. North

\[
\begin{align*}
\beta_0 &= 0 \\
a_0 &= \frac{4a_2a_0T^2}{4a_2 + a_0T^2}
\end{align*}
\]  

\[\ldots \ldots (4.17)\]

The above filters are realized from the basic configuration and shown in Fig. 4.3.

4.2.3 Sensitivities

The variation of element valued affects the transfer functions
of the desired filter sections. A useful measure of these sensitivities can be obtained by transforming the denominator polynomial of \( G(Z) \) into that of an analogue transfer function using the bilinear \( S \) to \( Z \) transformation given in (4.4). The denominator polynomial of the second-order filter is given by

\[
D(s) = s^2 + \frac{\omega_0}{Q} s + \omega_0^2 \quad \ldots \ldots (4.18)
\]

\[
S_x^o = \frac{X}{Q} \frac{\partial Q}{\partial X} \quad \ldots \ldots (4.19)
\]

\[
S_x^{\omega_0} = \frac{X}{\omega_0} \frac{\partial \omega_0}{\partial X} \quad \ldots \ldots (4.20)
\]

(where \( X \) is a parameter) The above equations will provide meaningful information on the effects of element value tolerances on the response using (4.4) in (4.3) and comparing the denominator polynomial with (4.18) one gets,

\[
\zeta = \frac{1}{2 \alpha_1 \beta_2 \beta_1} \left[ \alpha_1 \beta_2 \beta_1 (4 - \alpha_1 \beta_2 - 2 \alpha_1 \beta_2 \beta_1) \right]^{1/2} \quad \ldots \ldots (4.21)
\]

\[
\omega_0 = \frac{2}{T} \left[ \frac{\alpha_1 \beta_2}{4 - \alpha_1 \beta_2 - 2 \alpha_1 \beta_2 \beta_1} \right]^{1/2} \quad \ldots \ldots (4.22)
\]

Differentiating \( \zeta \) and \( \omega_0 \), one gets the logarithmic sensitivities as

\[
\begin{align*}
S_a, \zeta^0 &= \frac{1}{8} \frac{\omega_0^2 T^2}{Q} \\
S_a, \omega_0^0 &= -\frac{1}{8} \frac{\omega_0^2 T^2}{Q} - \frac{\omega_0 T}{4Q} \\
S_p, \zeta^0 &= -\frac{1}{4Q} \\
S_p, \omega_0^0 &= -\frac{1}{8} \frac{\omega_0^2 T^2}{Q} - \frac{\omega_0 T}{4Q}
\end{align*}
\]
Let us now find the limiting values of sensitivities when poles are very high and/or \(\omega_0T\) is very low. It can be seen from (4.23) and (4.24) that

\[
\begin{align*}
S_{a1} &= \frac{1}{2} + \frac{\omega_0^2 T^2}{8} \\
S_{a2} &= \frac{1}{2} + \frac{\omega_0^2 T^2}{8} + \frac{\omega_0 T}{4Q} \\
S_{b1} &= \frac{\omega_0 T}{4Q} \\
S_{b2} &= \frac{1}{2} + \frac{\omega_0^2 T^2}{8} + \frac{\omega_0 T}{4Q}
\end{align*}
\]

Let us now find the limiting values of sensitivities when poles are very high and/or \(\omega_0T\) is very low. It can be seen from (4.23) and (4.24) that

\[
\begin{align*}
S_{a1} &\approx \frac{1}{2}, & S_{a2} &\approx -\frac{1}{2} \\
S_{b1} &\approx -1, & S_{b2} &\approx -\frac{1}{2} \\
S_{\beta1} &\approx \frac{1}{2}, & S_{\beta2} &\approx \frac{1}{2} \\
S_{\beta1} &\approx 0, & S_{\beta2} &\approx \frac{1}{2}
\end{align*}
\]

The above results show that all sensitivities of the proposed filter sections are within practically acceptable limits.

4.2.4 Design Example:

To design a switched capacitor notch filter with the following specifications:

- notch frequency = 60 Hz
- 3 dB bandwidth = 58 Hz
The Z-domain transfer function of the notch-filter is given by
\[ H(z) = \frac{0.9777(z^2 - 1.9978z + 1)}{z^2 - 1.9533z + 0.9554} \] 
\[ (4.27) \]

Comparing (4.27) with (4.3), it is possible to get the following identities
\[ \beta_0 = 0.9777 \]
\[ a_2 \beta_1 = 0.0446 \]
\[ a_1 \beta_2 = 0.0021 \]
\[ \beta_0 = 0.0000 \]
\[ a_2 a_0 = 0.0022 \] 
\[ (4.28) \]

Choosing \( \chi_2 = 0.05 \), one gets the following capacitor ratios.
\[ a_1 = 0.0410 \]
\[ \beta_1 = 0.9112 \]
\[ a_0 = 0.0440 \] 
\[ (4.29) \]

Which shows that the largest-to-smallest capacitance ratio is \( 23,8463 : 1 \).

The sensitivity functions can be obtained from (4.23) and (4.24) as
4.3 Switched-Capacitor Filters Using Unity-Gain Amplifiers

4.3.1 Realization: I

Consider the general feedback configuration of Fig. 4.4 which involves a dual input SC network $N_a$ and a single-input SC network $N_b$ in conjunction with two unity-gain amplifiers. Suppose that $N_a$ and $N_b$ are characterised by the following input output relations in $Z$-domain:

\[ V_i(z) = H_{1a}(z) V_{in}(z) + H_{2a}(z) V_o(z) \]  
\[ V_o(z) = H_{1b}(z) V_i(z) \]

then Fig. 4.4 realizes the following transfer functions:

\[ \frac{V_1(z)}{V_{in}(z)} = \frac{H_{1a}(z)}{1 - H_{1a}(z) H_{2a}(z)} \]  
\[ \frac{V_0(z)}{V_{in}(z)} = \frac{H_{1b}(z) H_{1a}(z)}{1 - H_{1a}(z) H_{2a}(z)} \]

It is possible to find different networks for $N_a$ and $N_b$ in order that Fig. 4.4 realizes second-order transfer functions. Two such realizations are shown in Fig. 4.5. The switches are operated by a 2-phase non-overlapping clock (odd phase $\theta$).
For Fig. 4.5 (a) the following relations can be obtained. Comparing (4.34) and (4.35) to (4.30) and (4.31) we obtain

\[ H_{bd}(z) = -H_{ad}(z) = \frac{1}{a(z - \frac{1}{1 + a})} \]  
\[ H_{bd}(z) = \frac{z - 1}{z - \frac{1}{1 + \beta}} \]  

Substitution of (4.36) and (4.37) in (4.32) and (4.33) yields the following bandpass transfer function \( H_{bd}(z) \):

\[ V_d(z) = \frac{V_{ad}(z)}{V_{bd}(z)} = \frac{(z - 1)}{V_{bd}(z)} \]

where

\[ D(z) = z^2 - \left(1 + \frac{1}{1 + \beta}\right)z + \frac{1}{1 + a} + \frac{\alpha}{(1 + a)(1 + \beta)} \]

For Fig. 4.5(b) assume that the input voltage \( V_{in} \) is a staircase function (coming from other switched-capacitor circuits, or from a sample/hold circuit). Then it is possible to obtain the following highpass transfer function \( H_{hd}(z) \):

\[ \frac{\bar{V}_d(z)}{V_{ad}(z)} = H_{hd}(z) = \frac{(z - 1)^2}{D(z)} \]
where $D(Z)$ is given equation. (4.39)

If the poles of the above transfer functions are located at $z = R \exp (\pm j \theta)$, then

$$R^2 = \frac{1}{1 + z} \frac{z}{(1 + 2)(1 + \beta)} \quad \cdots \quad (4.41)$$

$$\theta = \cos^{-1} \left( \frac{1}{2R} \right) \quad \cdots \quad (4.42)$$

These relations of eqn. (4.41) and (4.42) are of practical importance because they provide a convenient means for comparing the pole co-ordinates as a function of network parameters. Note that $R < 1$ for all $\omega > 0$ and $\beta > 0$. This shows that the poles lie inside the unity circle in $z$-plane which ensures stability.

### 4.3.1 Design equations and sensitivity considerations

In order to design the filters, let us make use of matched $z$-transform which yields

$$R = \exp \left( -\frac{\pi f_0}{Q \omega_c} \right) \quad \cdots \quad (4.43)$$

$$\theta = \frac{2\pi f_0}{f_c} \left( 1 - \frac{1}{4Q^2} \right)^{1/2} \quad \cdots \quad (4.44)$$

where $f_0$ is the centre frequency, $Q$ is selectivity and $f_c$ is the clock frequency.
For poles closed to unity circle (i.e. for \( f_o \ll f_c \) and \( Q \gg 1 \)) we can write

\[
\theta = \frac{2\pi f_o}{f_c} \quad \ldots (4.45)
\]

\[
R = 1 - \frac{\theta}{2Q} \quad \ldots (4.46)
\]

\[
\cos \theta = 1 - \frac{\theta^2}{2} \quad \ldots (4.47)
\]

Substitution of (4.45) to (4.47) in yields the following design equations:

\[
\beta = \frac{1 + \delta/Q}{\delta^2 - \delta/Q - 1} \quad \ldots (4.48)
\]

\[
\alpha = \delta/Q \quad \ldots (4.49)
\]

where, \( \delta = f_o/2\pi f_c \)

If we assume that no trimming of capacitor values will be performed in LSI implementation of the filters, it is necessary to investigate the sensitivity of \( Q \) and \( f_o \) to variations in \( \alpha \) and \( \beta \). Using equation (4.45) to (4.47) in eqn (4.41) and (4.42) we obtain \( Q \) and \( f_o \) in terms of \( \alpha \) and \( \beta \) as follows:

\[
Q = \frac{1}{\alpha} \left\{ \frac{(1 + \alpha)(1 + \beta)}{\beta} \right\}^{1/2} \quad \ldots (4.50)
\]

\[
f_o = \frac{f_c}{2\pi} \left\{ \frac{\beta}{(1 + \alpha)(1 + \beta)} \right\}^{1/2} \quad \ldots (4.51)
\]

Note that, for \( \alpha = \beta \) and \( \alpha < 1 \), eqn (4.50) simplified to

\[
Q = (\alpha)^{-3/2} \quad \ldots (4.52)
\]
and the spread in capacitor values in $Q$ $2/3$. Thus, the circuit is suitable for high $Q$ realization because capacitor ratios are less than $Q$. The sensitivity functions are obtained as

\[ S_Q^0 = -\frac{1}{2} + \frac{1}{2(1 + a)}, \quad S_i^0 = -1 + \frac{1}{2(1 + a)} \quad \cdots (4.53a) \]

\[ S_f^0 = \frac{1}{2} - \frac{\beta}{2(1 + \beta)}, \quad S_f^\alpha = -\frac{\alpha}{2(1 + a)} \quad \cdots (4.53b) \]

From the above equations we find that

\[ |S_Q^0| < \frac{1}{2}, \quad |S_i^0| < 1, \quad |S_f^0| < \frac{1}{2}, \quad |S_f^\alpha| < \frac{1}{2} \quad \cdots (4.54) \]

These values are typical for low-sensitivity active circuits.

4.3.1.2 Dynamic range

Referring to Fig. 4.4, assume that the output amplifier $A_2$ determines the acceptable maximum signal level and that the other amplifier has the same signal swing properties, i.e. the same slow rate and the same output-voltage and current capabilities.

Under the stated assumptions, it is clear that for maximum signal swing it is necessary to control the voltage maxima at the output of each amplifier such that for increasing signal levels no individual amplifier gets saturated sooner than its succeeding amplifier. The requirement can be put into the form

\[ |H_1(\exp j\theta)| \leq |H_2(\exp j\theta)| \quad \cdots (4.55) \]
where,

\[ H_1(z) = \frac{V_1(z)}{V_{in}(z)}, \quad H_2(z) = \frac{V_0(z)}{V_{in}(z)}, \quad \text{and} \quad z = \exp(j\theta). \]

4.3.2 Realization - II

4.3.2.1 First Order Building Blocks

First-order switched capacitor networks used as the basic building blocks for realizing second-order filter characteristics are described. It is assumed that the switches are operated from a 2-phase non-overlapping clock (even phase \( \phi_1 \) and odd phase \( \phi_2 \)).

(a) Type-A network: The network of Fig. 4.6 (a) yields the following relation in z-domain:

\[
V_0(z) = \frac{(z-1)}{z} V_1(z) + \left[ \frac{1}{1+\alpha} \right] V_2(z) \quad \ldots \ldots (4.56)
\]

Corresponding block diagram representation is shown in Fig. 4.6(b).

(b) Type-B network: The network of Fig. 4.6 (b) realizes the following relation

\[
V_0(z) = \left[ \frac{1}{1+\alpha} \right] V_1(z) - V_2(z) \quad \ldots \ldots (4.57)
\]
Using the basic building blocks of Fig. 4.6 and Fig. 4.7, schemes for realizing second order high-pass and band-pass characteristics are proposed in Fig. 4.8. The networks given realize the following transfer function in z-domain.

\[ H(z) = \frac{V_{out}(z)}{V_{in}(z)} = \frac{N(z)}{D(z)} = \frac{1 + \frac{\alpha}{1 + \beta_1}z^{-1}}{1 + \frac{\alpha_0}{1 + \beta_0}z^{-1}} \]  \hspace{1cm} \text{(4.58)}

\[ N(z) = (z-1)^2 \]

\[ D(z) = (1 - \frac{\alpha}{1 + \alpha})(z-1) \]

For designing these configurations, bilinear z-transform is used:

\[ z = \frac{1 + sT}{2} \left( \frac{1 - sT}{2} \right) \]  \hspace{1cm} \text{(4.59)}

where is the sampling frequency.

The use of bilinear z-transformation yield bilinearly pre-warped pole frequency \( \omega_0 \) and pole - Q related through the following relationships.

\[ \frac{1}{4} \delta_0 Q = \beta / (4 + 4\alpha + 2\beta + 3\alpha) \]  \hspace{1cm} \text{(4.60)}

and

\[ \frac{1}{4} \delta^2 = \alpha \beta / (4 + 4\alpha + 2\beta + 3\alpha) \]  \hspace{1cm} \text{(4.61)}

where \( \delta = (1/\omega_0 T) \).
4.3.2.3 Sensitivities:

Using (4.65) and (4.66) logarithmic sensitivities are:

\[ \begin{align*}
        s_a^Q &= \frac{(Q/Q)_{a/a}}{(\partial Q/\partial a)} = 2 \left( 1 + \frac{4 + 2\beta}{4a + 3\beta} \right)^{-1} - 1 \\
        s_a^\omega &= \frac{(\omega/\omega)_{a/a}}{(\partial \omega/\partial a)} = 2 \left( 1 - \frac{4 + 4a}{2a + 3a \beta} \right)^{-1} \\
        s_\beta^Q &= \frac{(Q/Q)_{\beta/\beta}}{(\partial Q/\partial \beta)} = 2 \left( 1 + \frac{4 + 4a}{2a + 3a \beta} \right)^{-1} - 1 \\
        s_\beta^\omega &= \frac{(\omega/\omega)_{\beta/\beta}}{(\partial \omega/\partial \beta)} = 2 \left( 1 - \frac{4 + 4a}{2a + 3a \beta} \right)^{-1}
\end{align*} \]

\( \cdots \quad (4.68) \)

Since \( \lambda \) and \( \beta \) are ratios giving capacitor spread, they are positive and we have

\[ \begin{align*}
        |s_a^Q| < 1; & \quad s_a^Q < \frac{1}{2}; \quad s_\beta^Q < \frac{1}{2}; \quad s_\beta^\omega < \frac{1}{2}; \quad s_\lambda^\omega < \frac{1}{2} \quad \cdots \quad (4.69)
\end{align*} \]

The above values are typical for low-sensitivity active filters.

4.3.2.4 Dynamic Range

As in the case of continuous time networks, the maximum magnitude occurs at the centre frequency

\[ f_0 = \frac{\Omega_0}{2\lambda} \]

The maximum magnitude of a digital band-pass filter of
From 4.60 and 4.61 we get

\[ a = \frac{1}{2} \left( \frac{4 \frac{\omega}{\omega_0} + 4}{4 \frac{\omega}{\omega_0} - 2 \frac{\omega}{\omega_0} - 3} \right) \] \quad \cdots (4.62)

\[ \alpha = \frac{1}{2} \left[ \left( 4 + 4 \alpha + 2 \beta + 3 \alpha \beta \right) \right]^{1/2} \] \quad \cdots (4.63)

and

\[ \omega_0 = \frac{2}{1} \left[ \frac{\alpha \beta}{4 + 4 \alpha + 2 \beta + 3 \alpha \beta} \right] \] \quad \cdots (4.64)

FOR \( \omega \gg 1 \), \( \beta \ll 1 \) and \( \alpha \ll 1 \) \quad \cdots (4.65)

\[ \text{If } \alpha = \frac{1}{\beta} \] \quad \cdots (4.66)

\[ Q = \alpha^{3/2} \] \quad \cdots (4.67)

and the spread in capacitor values is \( Q^{2/3} \). Thus, the circuit is suitable for high-\( Q \) realization because capacitor ratios are less than \( Q \).

The switched capacitor high-pass and band-pass filters are shown in Fig. 4.9.
the form \( \frac{K(z-1)}{z^2 - a_1 z + a_0} \) is approximately equal to \( \frac{K}{1 - a_0} \)

Let us define

\[
H_2(z) = \frac{V_o(z)}{V_1(z)}
\]

\[
H_1(z) = \frac{V_c(z)}{V_1(z)}
\]

where \( V_c \) and \( V_o \) are the output level of the amplifiers.

Then from above we get

\[
H_2(z) = \frac{K}{1 - a_0}
\]

and

\[
|H_1(z)| = |H_2(z)| \left| z - \frac{1}{1+p} \right| \left| z - 1 \right| \]

Identifying \( a_0 \) and \( K \) from (4.58) and substituting in (4.71) we get

\[
|H_2(z)| = \alpha \left( 1 + \frac{1}{1+p} \right)
\]

or

\[
H_2(z) = \frac{\zeta}{\delta} \left[ \frac{4\delta^2 + 2\gamma + 1}{4 + 4\delta} \right]
\]

Since \( \frac{1}{1+p} < 1 \) it is evident from (4.72) that

\[
|H_1(z)| > |H_2(z)|
\]
For $Q \gg 1$ and $\delta > 1$, \( H_2(z) \approx \delta Q \) (4.75)

This shows that the output level of amplifiers is high for high $Q$ values. Thus, for higher input voltages, there is a possibility of amplifiers being driven into saturation. This difficulty can be avoided by setting the input voltage level to a low value such that the amplifiers are not driven into saturation.
FIG. 4.1 SWITCHED CAPACITOR INTEGRATORS
(a) INVERTING TYPE  (b) NON INVERTING TYPE
(c) CLOCK PULSES
FIG. 4.2 BIQUADRATIC SECTION
FIG. 4.3 REALIZATION OF SECOND ORDER SWITCHED CAPACITOR FILTER (a) BAND PASS (b) HIGH PASS (c) LOW PASS WITH NOTCH CHARACTERISTICS.
FIG. 4.4 BASIC CONFIGURATION

FIG. 4.5 FILTER (a) BAND PASS (b) HIGH PASS
FIG. 4.6 (a) FIRST ORDER SWITCHED CAPACITOR NETWORK OF TYPE-A
(b) BLOCK DIAGRAM OF TYPE-A NETWORK
Fig 7 (a) First order switched capacitor network of type - B
(b) Block diagram of type - B network
FIG. 4.8 REALIZATION OF SECOND ORDER FILTERS (a) HIGH PASS (b) BAND PASS
FIG. 4.9 SWITCHED CAPACITOR IMPLEMENTATION OF THE CONFIGURATION OF FIG. 4.8
(a) HIGH PASS  (b) BAND PASS