CHAPTER 3
REALIZATION OF VOLTAGE MODE AND CURRENT
MODE FILTERS USING CCII

3.1 Introduction

Active filters are an important class of signal processing circuits widely used in communication, control and instrumentation systems. Realization of active RC-filters has been a subject of extensive research and a number of circuits operating in voltage mode (VM) have been reported [44-54]. Presently the current mode (CM) biquadratic filters [56-69] are receiving considerable attention due to the advantages of large bandwidth, better accuracy, high slew rate, low power consumption, greater linearity, simple circuitry and wider dynamic range [55,137].

In this Chapter, some cascadable voltage mode and current mode filters are realized using second generation current conveyor (CCII). The CCII is basically a current mode device, which is also suitable for high frequency applications. This has led to the CCII-based realizations gain significant popularity, in analog signal processing. The CCII can be used as an active device in the design of simple and attractive active circuits for voltage mode [44-54], as well as, current mode [56-69] operations. This has motivated circuit designer to further develop various analog voltage mode and current mode filters and circuits using CCII's.

This Chapter considers the realization and study of voltage mode and current mode first order, biquadratic and high order filters, realized using single, as well as, multiple outputs current conveyors. Some of the important and desirable features of the novel circuits given in this Chapter are:

(i) use of grounded passive components
(ii) wide frequency range of operation
(iii) high functional versatility

* Authors’ papers [P1], [P2], [P3] and [P4] are based on the material presented in chapter 3
(iv) low sensitivities to active and passive components
(v) low component count
(vi) simplicity in design
(vii) high input and low output impedance levels, for VM filters and
  high output and low input impedance levels for CM filters for
  facilitating the non-interactive cascading of the basic building
  blocks for the realization of higher order filters.

Three novel voltage mode multifunctional biquadratic filters (MBFs),
using single CCI [P1], two CCIIs and three CCIIs [P2] are respectively given
and studied in Secs. 3.2, 3.3, and 3.4. The performance features of the
presented MBFs are compared in Sec. 3.5. Section 3.6 is devoted to the
realization of VM Universal Biquadratic Filter (VM UBF) [P4], using only two
CCIIs and four passive components.

Next, the current mode circuits are considered. Section 3.7 presents
single CCI based first order current mode (CM) filters [P3], which uses only
single CCI+ having single output, along with, three grounded passive elements.
Section 3.8 is devoted to the realization of high output impedance current mode
Universal Biquadratic Filter (CM UBF), which enables easy cascading. The
proposed UBF employs only two multiple output current conveyors (MO-
CCIIs) and four grounded passive components and provides functional
versatility in the realization of multiple filter responses, such as, low pass, high
pass, band pass, band elimination and all pass. In Section 3.9, first order filter
sections are used to realize 4th order low pass filter. In Section 3.10, the
realizations of 6th order Butterworth low pass and band pass filters are
considered using the current mode UBF of Sec. 3.8. Concluding remarks are
given in Sec. 3.11. The circuits are also studied considering non-ideal behavior
of the device. Their sensitivity performance is evaluated. Simulation study,
using PSpice, are included to verify the theory.
3.2 VM Multifunctional Biquadratic Filter using Single CCII (MBF1)

Current Conveyor based realizations have gained popularity in analog signal processing, especially in low voltage signal processing applications [122-128]. Recently, active filters employing single active device have generated much interest in video signal processing and wireless communication systems [39, 41]. Such biquads have attractive features of low voltage, low power operation, low noise, small size and wide linearity.

3.2.1 Circuit description

The generalized topology of a multifunctional biquadratic filter (MBF1) [P1] is shown in Fig.3.1. In the thesis, multifunctional filter implies incorporation of LP, HP and BP responses. The realization consists of a single CMOS plus type CCII alongwith, with five passive components, out of which four are grounded and one floating.

For the CCII+, the $v-i$ relations are defined by:

$$i_y = 0, v_x = v_y \quad \text{and} \quad i_z = +i_z$$  \hspace{1cm} (3.1)

Analysis of the circuit yields the general transfer function:

$$T(s) = \frac{V_o}{V_i} = \frac{Y_2Y_4}{Y_1Y_2 + Y_3Y_4}$$  \hspace{1cm} (3.2)
Low pass realization

If the admittances are selected as: \( Y_1 = sC_1 \), \( Y_2 = 1/R_2 \), \( Y_3 = sC_3+1/R_3 \) and \( Y_4 = 1/R_4 \), as shown in Fig. 3.2 (a), a second order LP filter is realized with

\[
T_{LP}(s) = \frac{V_{LP}}{V_i} = \frac{1}{s^2 + s + \frac{1}{R_3C_3} + \frac{1}{R_4C_4} + \frac{1}{R_3R_4C_1C_3}}
\]  

(3.3)

The filter parameters, pole-\( \omega_0 \), pole-\( Q \) and gain of the low pass filter are obtained from eqn. (3.3) and given by:

\[
\omega_{0LP} = \sqrt{\frac{1}{R_3R_4C_1C_3}}, \quad Q_{LP} = \sqrt{\frac{1}{R_3C_3} + \frac{1}{R_4C_4}}, \quad H_{LP} = \frac{R_3}{R_2}
\]  

(3.4)

It can be seen from eqn. (3.4), the gain of the filter can be tuned independently through \( R_2 \), without disturbing the pole frequency and pole-\( Q \).

High pass realization

If the admittances are selected as: \( Y_1 = 1/R_1 \), \( Y_2 = sC_2 \), \( Y_3 = sC_3+1/R_3 \) and \( Y_4 = sC_4 \), HP filter shown in Fig. 3.2 (b) is realized. Its transfer function is given by:
\[
T_{\text{HP}}(s) = \frac{V_{\text{HP}}}{V_i} = \frac{s^2 C_2 / C_3}{s^2 + s \left( \frac{1}{R_3 C_3} + \frac{1}{R_4 C_4} \right) + \frac{1}{R_1 R_2 C_3 C_4}} \quad (3.5)
\]

The parameters of the HP filter are given by:

\[
\omega_{\text{HP}} = \sqrt{\frac{1}{R_1 R_2 C_3 C_4}}, \quad Q_{\text{HP}} = \sqrt{\frac{1}{R_1 R_3 C_3 C_4}}, \quad H_{\text{HP}} = \frac{C_2}{C_3} \quad (3.6)
\]

From eqn. (3.6), it can be seen that the gain of HP filter can now be tuned independently by \( C_2 \).

\[\text{Fig. 3.2 (b) High pass biquadratic filter}\]

**Bandpass realization**

For the realization of a band pass filter, the component selection is: \( Y_1 = 1/R_1, \ Y_2 = 1/R_2, \ Y_3 = sC_3 + 1/R_3 \) and \( Y_4 = sC_4 \), as shown in Fig. 3.2 (c). This gives the BP transfer function:

\[\text{Fig. 3.2 (c) Band pass biquadratic filter}\]
The pole frequency, pole-Q and gain of the BP filter are given by:

$$\omega_{BP} = \sqrt{\frac{1}{R_1 C_4 R_3 C_3}}$$
$$Q_{BP} = \sqrt{\frac{1}{\left(\frac{1}{R_1 C_4} + \frac{1}{R_3 C_3}\right)}}$$
$$H_{BP} = \frac{R_2 R_4}{R_2 (R_3 C_3 + R_4 C_4)}$$ \hspace{1cm} (3.8)

Here also, the gain can be tuned independently through $R_2$. It is observed from the expression of pole-Q that the denominator is greater than the numerator. Hence the circuit has the drawback of realizing only low Q values.

### 3.2.2 Non-ideal analysis

As discussed in Sec. 2.3.1, using the non-idealities of CCII, the modified $v-i$ relations are defined by:

$$i_y = 0, \quad v_y = \beta v_y, \quad i_z = \alpha i_z \hspace{1cm} (3.9)$$

The frequency effects of CCII, modifies the eqns. (3.3), (3.5) and (3.7) of the MBF1 to:

$$T_{lp}(s) = \frac{V_{lp}}{V_i} = \frac{\alpha \beta}{R_3 R_4 C_3 C_4} \frac{s^2}{(s + \omega_a)(s + \omega_b)} \left(\frac{1}{R_3 C_3} + \frac{1}{R_4 C_4}\right) + \frac{1}{R_3 R_4 C_3 C_4}$$ \hspace{1cm} (3.10)

$$T_{hp}(s) = \frac{V_{hp}}{V_i} = \frac{s^2 \alpha \beta C_2 \beta}{C_3} \frac{1}{(s + \omega_a)(s + \omega_b)} \left(\frac{1}{R_3 C_3} + \frac{1}{R_4 C_4}\right) + \frac{1}{R_3 R_4 C_3 C_4}$$ \hspace{1cm} (3.11)

$$T_{bp}(s) = \frac{V_{bp}}{V_i} = \frac{s \alpha \beta}{R_3 C_3} \frac{1}{(s + \omega_a)(s + \omega_b)} \left(\frac{1}{R_3 C_3} + \frac{1}{R_4 C_4}\right) + \frac{1}{R_3 R_4 C_3 C_4}$$ \hspace{1cm} (3.12)
It is clear from eqns. (3.10) to (3.12) that all the filter functions have two extra poles, due to single pole model, in addition, to the existing two complex conjugate poles. If the frequency of these two extra poles are sufficiently higher than the poles of the presented MBFl, \( \min \{ \omega_a, \omega_b \} >> (1/RC) \), (as can easily be ensured through design), then their effect on the frequency response can be minimized. It may be noted the non-idealities of CCII gives the circuit a roll off in gain at higher frequencies. Thus below 10 MHz, they have negligible effect on the circuit performance [34].

Taking the parasitics effects described in Sec. 2.3.1, into account, the analysis of the circuit of Fig. 3.1 yields the mixed LP-BP transfer functions as:

\[
T_{LP}(s) = \frac{V_{LP}}{V_i} = \frac{\alpha\beta}{s^2 + \left( \frac{1}{R_2R_4C_3C_{3P}} + \frac{1}{R_4C_1C_{3P}} \right) + \frac{1}{R_3R_4C_1C_{3P}}} + \frac{s\alpha\beta C}{R_2R_4C_3C_{3P}}
\]

(3.13)

where \( R_{3P} = R_3 / R_2, R_{4P} = R_4 / R_4, C_{1P} = C_1 + C_2, C_{3P} = C_3 + C_2 \).

For practical CCII, assuming \( C_1 >> C_2, C_3 >> C_2, C_X << C_1 \) or \( C_3 \) and \( R_3, R_4 << R_2, or R_2, \) and the current and voltage transfer ratios \( \alpha \) and \( \beta \) as unity for frequencies till tens of MHz, then the eqn. (3.13) reduces to the ideal low pass transfer function of eqn. 3.3.

Similarly the high pass and band pass transfer functions are given by:

\[
T_{HP}(s) = \frac{V_{HP}}{V_i} = \frac{s^2\alpha\beta C_2C_{3P}C_{4P}}{s^2 + \left( \frac{1}{R_{1P}C_{4P}} + \frac{1}{R_3C_2C_{3P}} \right) + \frac{1}{R_{1P}R_3C_2C_{3P}}} + \frac{s\alpha\beta C_2}{R_{1P}C_{4P}C_{3P}}
\]

(3.14)

where \( R_{1P} = R_1 / R_1, R_{3P} = R_3 / R_2, C_{2P} = C_2 + C_X, C_{3P} = C_3 + C_2, C_{4P} = C_4 + C_Y \).

Assuming, \( C_3 >> C_2, C_4 >> C_Y, C_2 >> C_X \) and \( R_3, R_4 << R_2, or R_2, \) and the current and voltage transfer ratios \( \alpha \) and \( \beta \) as unity, for frequency range of
interest, then the eqn. (3.14) simplifies to the ideal high pass transfer function of eqn. (3.5).

The band pass function is given by:

\[ T_{BP}(s) = \frac{V_{BP}}{V_i} = \frac{s\alpha\beta C_4 + s^2\alpha\beta C_4 C_X}{R_2 C_3 P C_4 P + \frac{1}{R_1 P C_4 P} + \frac{1}{R_1 P P C_3 P C_4 P}} \]  

(3.15)

where \( R_{1P} = R_1 \parallel R_1 \), \( R_{3P} = R_3 \parallel R_2 \), \( C_{3P} = C_3 + C_2 \), \( C_{4P} = C_4 + C_1 \).

Assuming \( C_3, C_4 \gg C_X, C_Y, or C_Z \), and \( R_1, R_3 \ll R_1 \) or \( R_2 \), and the current and voltage transfer ratios \( \alpha \) and \( \beta \) are nearly unity for frequencies below tens of MHz, then the eqn. (3.15) reduces to the ideal band pass transfer function of eqn. 3.7.

The filter parameters with non-idealities are given by:

\[ \omega_{OLP} = \sqrt{\frac{1}{R_{3P} R_{4P} C_{1P} C_{3P}}} \], \hspace{1cm} \omega_{QLP} = \sqrt{1 / R_{3P} R_{4P} C_{1P} C_{3P}} \] 

(3.16)

\[ \omega_{OHp} = \sqrt{\frac{1}{R_{1P} R_{2P} C_{1P} C_{3P} C_{4P}}} \], \hspace{1cm} \omega_{QHP} = \sqrt{1 / R_{1P} R_{2P} C_{3P} C_{4P}} \] 

(3.17)

\[ \omega_{OBP} = \sqrt{\frac{1}{R_{1P} R_{3P} C_{1P} C_{3P} C_{4P}}} \], \hspace{1cm} \omega_{QBP} = \sqrt{1 / R_{1P} R_{3P} C_{3P} C_{4P}} \] 

(3.18)

It is evident that the pole frequency and pole-Q get affected due to parasitics, however these parasitic effects may be minimized through proper layout and circuit design. Also, parasitic effect of CCII can be made to advantage in clever design itself.
3.2.3 Sensitivity study

Table 3.1: Sensitivity Figures of MBF1 of Fig. 3.2

<table>
<thead>
<tr>
<th>Sensitivity To / Of</th>
<th>Low pass filter</th>
<th>High pass filter</th>
<th>Band pass filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_o$</td>
<td>$Q$</td>
<td>$\omega_o$</td>
<td>$Q$</td>
</tr>
<tr>
<td>$R_1$</td>
<td>-</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
</tr>
<tr>
<td>$R_2$</td>
<td>-</td>
<td></td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td></td>
<td>$2(R_3C_3 + R_4C_4)$</td>
<td>$2(R_3C_3 + R_4C_4)$</td>
</tr>
<tr>
<td>$R_3$</td>
<td>$1/2$</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td></td>
</tr>
<tr>
<td>$R_4$</td>
<td>$1/2$</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td></td>
</tr>
<tr>
<td>$C_1$</td>
<td>$1/2$</td>
<td></td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td></td>
</tr>
<tr>
<td>$C_2$</td>
<td>$1/2$</td>
<td></td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td></td>
</tr>
<tr>
<td>$C_3$</td>
<td>$1/2$</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td></td>
</tr>
<tr>
<td>$C_4$</td>
<td>$1/2$</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td></td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td>$(R_3C_3 - R_4C_4) / (2(R_3C_3 + R_4C_4))$</td>
<td></td>
</tr>
</tbody>
</table>

The incremental sensitivities of important circuit parameters, viz., pole frequency and pole-Q of the low pass, high pass, band pass filters have been evaluated with respect to the circuit elements and are given in Table 3.1. It is evident from the Table that the nominal values of the $\omega_o$ sensitivity ($S_{\omega o}$) and the pole-Q sensitivity ($S_Q^o$) are low and attractive, being less than or equal to half in magnitude. Thus, the circuit exhibits attractive sensitivity properties.

3.2.4 Design and simulation

To verify the theory of the proposed circuits, they are designed and simulated using Level 3 PSpice parameter in 0.5 µm CMOS process at supply voltages $V_{DD} = -V_{SS} = 0.75V$. The CCII+ model of Ref. [11] is used in the simulation. Also, in all the subsequent CCII-based circuit simulations reported...
in the thesis, the same Level 3 PSpice parameters alongwith the device model will be used. To demonstrate the performance of single CCIH+ based multifunctional biquadratic filter (MBF1), the circuits of low pass, high pass and band pass filters are designed for $f_o = 200$ KHz, $Q = 0.5$ at unity gain. The design values for the filters are:

For Low pass filter: $C_1 = C_3 = 45$ pF and $R_2 = R_3 = R_4 = 17.68$ KΩ.

For High pass filter: $C_2 = C_3 = C_4 = 45$ pF and $R_1 = R_3 = 17.68$ KΩ.

For Band pass filter: $C_3 = C_4 = 45$ pF, $R_1 = R_3 = R = 17.68$ KΩ, $R_2 = 8.84$ KΩ.

The simulated LP, HP and BP responses are shown in Fig. 3.3, with the simulated results given in the Table. These are found to be close to the theoretical values.

The tunability of the BP filter has been investigated by varying the center frequency $f_o$ through passive resistor $R_3$. The BP response curves corresponding to $f_o$ of 200 KHz, 500 KHz and 1.5 MHz are shown at a constant $Q$ of 0.707. It may be noted that with the variation of frequencies, as shown in Fig. 3.4, the gain of band pass filter remains constant and equal to unity. These also exhibit close agreement with theory.

<table>
<thead>
<tr>
<th>$f_o$ (KHz)</th>
<th>$Q_{BP}$</th>
<th>$H_{LP}$</th>
<th>$H_{HP}$</th>
<th>$H_{BP}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>200.06</td>
<td>0.705</td>
<td>1.00</td>
<td>1.01</td>
<td>0.998</td>
</tr>
</tbody>
</table>

Fig. 3.3 Frequency responses of LP, HP, and BP filter
Fig. 3.4 Tunability of BPF

(a) $f_o = 200$ KHz (b) $f_o = 500$ KHz (c) $f_o = 1.5$ MHz

Also, in Table 3.2(a), the simulated values of input and output resistances are given, which show convenient cascadability, ($R_{in} \gg R_{out}$) of the proposed MBFl circuit.

<table>
<thead>
<tr>
<th>Type of MBF</th>
<th>Input Resistance</th>
<th>Output Resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>MBFl</td>
<td>$1 \times 10^{14} \text{M}\Omega$</td>
<td>21.7 K(\Omega)</td>
</tr>
</tbody>
</table>

The multifunctional biquadratic filter is realized using single CCII+ and four admittances and without the requirement of component matching. The low pass, high pass and band pass filters were obtained through proper selection of admittances. The circuit also offers several advantages, such as, high input impedance and low output impedance; which enables easy cascading, low supply voltage operation and independent tuning of the gain of filters through,
R₂ (LPF), C₂ (HPF) and R₂ (BPF), without disturbing the pole frequency and quality factor. However, the circuit has drawback of realizing only low Q values.

### 3.3 High Input Impedance Multifunctional Biquadratic Filter using Two CCIIs (MBF2)

In voltage mode signal processing, active filters with high input and low output impedances [48, 50, 52] are of great importance, because they can be directly connected in a non-interactive cascade to implement higher order filters. Several multifunctional biquadratic filters with high input impedance are available in literature [41, 43, 45, 52] which employs a large number of active and passive devices, for realizing LP, HP and BP responses. In 1997, Horng [41] proposed a circuit to realize the responses with high input impedance using only grounded passive components and four plus type CCIIs. Chang [37] proposed another circuit using four CCIIs, four resistors and two grounded capacitors. In 1999, Chang and Lee [42], proposed a circuit to realize the multiple responses using two current conveyors, three resistors and two grounded capacitors. However, this circuit lacks the advantage of high input impedance and also employs floating resistor. In 2002, Singh and Senani [44] proposed a circuit to realize KHN biquad using four CCIIs, five resistors and two grounded capacitors. The circuits of the Refs [37, 44] did not have the advantage of high input impedance; moreover, these circuits employed too many active and passive components. Horng proposed circuits [43, 52] with high input impedance using grounded passive components and three CCIIs. In 2005, Horng [45] proposed two high input impedance circuits each with one input and three outputs using three CCIIs along with four grounded capacitors and three grounded resistors. However, the circuits have the advantage of high input impedance, but employed large number of active and passive devices.

In this Section, a novel high input impedance voltage mode multifunctional biquadratic filter (MBF) is realized using two CCIIs and four
admittances only. It gives the multifunctional realization of LP, HP and BP responses through appropriate selection of admittances, as in the case of MBF1.

3.3.1 Circuit description

The proposed structure of the multifunctional biquadratic filter (MBF2) is shown in Fig. 3.5. It consists of two CCIIs and four grounded admittances.

![Fig. 3.5 Multifunctional biquadratic filter (MBF2)](image)

The generalized circuit realizes low pass, high pass and band pass responses through appropriate choice of grounded admittances without feedback requirement. Using the v-i relations of eqn. (3.1) for the CCIIs, analysis of the circuit yields the following voltage transfer function:

\[
\frac{V_v}{V_\text{in}} = \frac{Y_3 Y_5}{Y_2 Y_4}
\]  

(3.19)

**Low pass filter:** If we select, \( Y_1 = \frac{1}{R_1}, \ Y_2 = sC_2 + G_2, \) and \( Y_3 = \frac{1}{R_3}, \ Y_4 = sC_4 + G_4, \) it results in realizing low pass voltage transfer function, given by:

\[
T_{\text{LP}}(s) = \frac{V_{\text{LP}}}{V_\text{in}} = \frac{1}{R_i R_j C_2 C_4 D(s)}
\]  

(3.20)
where the characteristic polynomial, $D(s)$, is given by:

$$D(s) = s^2 + \left(\frac{1}{R_2C_2} + \frac{1}{R_4C_4}\right) + \frac{1}{R_2R_4C_2C_4}$$  \hspace{1cm} (3.21)

**High pass filter:** On selecting: $Y_1 = sC_1$, $Y_2 = sC_2 + G_2$, $Y_3 = sC_3$ and $Y_4 = sC_4 + G_4$, a high pass voltage transfer function is realized:

$$T_{hp}(s) = \frac{V_{hp}}{V_m} = \frac{s^2 C_3}{D(s)}$$  \hspace{1cm} (3.22)

with $D(s)$ as defined in eqn. (3.21).

**Band pass filter:** If we select, $Y_1 = 1/R_1$, $Y_2 = sC_2 + G_2$, $Y_3 = sC_3$ and $Y_4 = sC_4 + G_4$, then it results in the following band pass voltage transfer function:

$$T_{bp}(s) = \frac{V_{bp}}{V_m} = \frac{sC_3}{R_2C_2C_4}$$  \hspace{1cm} (3.23)

where $D(s)$ is same as defined in eqn. (3.21). From eqns. (3.20), (3.22) and (3.23), it is seen that low pass, high pass and band pass responses are realized through appropriate selection of grounded admittances, without the requirement of matching constraints. Cascading of such second order sections forms an attractive method for the realization of high order filters, due to their inherent high input impedance ($Z_m \to \infty$).

The pole frequency ($\omega_p$), bandwidth (BW) and the pole-Q of the LP, HP and BP filters obtained from $D(s)$, are given by:

$$\omega_p = \sqrt{\frac{1}{R_2R_4C_2C_4}}, \quad BW = \frac{1}{R_2C_2} + \frac{1}{R_4C_4}, \quad Q = \sqrt{\frac{1}{R_2C_2} + \frac{1}{R_4C_4}}$$  \hspace{1cm} (3.24)

The gains of the filters are given by:
From the expression of pole-Q given in eqn. (3.24), it can be seen that there is a summation term in the denominator, which makes the proposed circuit to realize low Q responses [38], as in the case of MBF1. The eqn. (3.25) shows that the gain of LP, HP and BP filters can be tuned independently through passive components \((R_1 \text{ or } R_3), \(C_1 \text{ or } C_3\) and \((R_4 \text{ or } C_4)\)) for LP, HP and BP filters, respectively.

### 3.3.2 Non-ideal analysis

Consider the non-ideal CCII characterized by the frequency independent current transfer factor \(\alpha\) and voltage transfer factor \(\beta\) for low to medium frequency applications. Using the non-idealities defined by eqn. (3.9), the analysis of the circuit of Fig. 3.5 yields the transfer functions as:

\[
T_{lp}(s) = \frac{V_{lp}}{V_{in}} = \frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 R_2 C_2 C_4} \frac{1}{D(s)}
\]

(3.26)

\[
T_{hp}(s) = \frac{V_{hp}}{V_{in}} = \frac{s^2 \alpha_1 \alpha_2 \beta_1 \beta_2 C_1 C_3}{C_2 C_4} \frac{1}{D(s)}
\]

(3.27)

\[
T_{bp}(s) = \frac{V_{bp}}{V_{in}} = \frac{s \alpha_1 \alpha_2 \beta_1 \beta_2 C_3}{R_1 C_2 C_4} \frac{1}{D(s)}
\]

(3.28)

It may be noted that the characteristic polynomial \(D(s)\) is still given by (3.21) and remains unaffected. Hence the \(\omega_c\), BW and Q are still given by eqn. (3.24).

This is one of the attractive features of the proposed filter. The gains are given by:
Slight reduction in gains is observed due to the device non-idealities.

3.3.3 Sensitivity study

The sensitivities of the filter parameters, pole-\(\omega_o\), and pole-Q, are evaluated with respect to active and passive elements and are summarized below:

\[
S_{r_1,r_2,c_1,c_4}^{\alpha_k} = -1/2, \quad S_{r_4,c_3}^{\alpha_0} = - S_{r_4,c_4}^{\alpha_0} = \frac{R_2C_4 - C_2R_4}{2(R_2C_4 + C_2R_4)} \ll |\pm 0.5|, \\
S_{r_1,r_2}^{\alpha_0} = -1, \quad S_{r_4}^{\alpha_0} = 1, \quad S_{c_3,c_4}^{\alpha_0} = -1, \quad S_{r_1}^{\alpha_0} = 1 \\
S_{c_1,c_4}^{\alpha_0} = -1, \quad S_{c_3}^{\alpha_0} = 1, \quad S_{r_1}^{\alpha_0} = -1
\] (3.30)

It is evident from eqn. (3.30) that all the sensitivities are attractive. Also, note that the filters are insensitive to the non-idealities in low to medium frequency range.

3.3.4 Design and simulation

To demonstrate the performance of CCII-based multifunctional biquadratic filter (MBF2), the low pass and band pass and high pass circuits are simulated using PSpice, as before. Their circuits are designed for \(f_o = 250\) KHz, at Q of 0.707 for unity gain. The design values for the filters are:

For low pass filter: \(C_2 = C_4 = 45\) pF and \(R_1 = R_2 = R_3 = R_4 = 14.15\) KΩ.

For high pass filter: \(C_1 = C_2 = C_3 = C_4 = 45\) pF and \(R_2 = R_4 = 14.15\) KΩ.

For band pass filter: \(C_2 = C_3 = C_4 = 45\) pF, \(R_1 = 7.07\) KΩ, and \(R_2 = R_4 = 2R_1 = 17.68\) KΩ.

The simulated responses are shown in Fig. 3.6, with simulated results given in the Table. These show good agreement with theory.
In the Table 3.2 (b), the simulated values of input and output resistances are given. These once again exhibit convenient cascadablity of MBF2, with $R_m \gg R_{out}$.

Table 3.2 (b): Input and Output Resistances of MBF2

<table>
<thead>
<tr>
<th>Type of MBF</th>
<th>Input Resistance</th>
<th>Output Resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>MBF2</td>
<td>$1 \times 10^{14}$ $\Omega$</td>
<td>36.7 $K\Omega$</td>
</tr>
</tbody>
</table>

In this Section, a novel high input impedance second order multifunctional filter (MBF2) is realized using two second generation current conveyors, alongwith, four grounded admittances without using any feedback. It realizes low pass, high pass and band pass responses with appropriate selection of all grounded admittances without any requirement for matching. High order filters can conveniently be realized by cascading the proposed second order sections. The proposed circuit permits the use of only grounded passive components, which is important for IC implementation. The circuit also
has low active and passive sensitivity and operation at low supply voltage of \( \pm 0.75V \).

### 3.4 High Input Impedance Multifunctional Biquadratic Filter Using Three CCIIIs (MBF3)

An unconditionally high input impedance multifunctional filter using only three plus type CMOS current conveyor is presented in this Section. Once again, only grounded resistors and capacitors are used for realizing second order low pass, band pass and high pass functions. The proposed circuit has the independent control of pole-\( \omega_c \), quality factor and gains of filters.

#### 3.4.1 Circuit Description

The multifunctional biquadratic filter (MBF3) [P2] is shown in Fig. 3.7.

![Fig.3.7 CCII+ Multifunctional biquadratic filter (MBF3)](image)

Using the \( v-i \) relations given in eqn. (3.1) for CCII+, the analysis of the circuit yields:

\[
T_i(s) = \frac{V_i}{V_i} = \frac{Y_3 Y_4}{Y_1 Y_5 + Y_2 Y_4}
\]

(3.31)
Proper selection of grounded admittances, \( Y_1, Y_2, Y_3, Y_4, \) and \( Y_5, \) realize standard second order LP, BP and HP filters, as given below.

**Case 1:** If the admittances are selected as: \( Y_1 = sC_1 + \frac{1}{R_1}, \) \( Y_2 = \frac{1}{R_2}, \) \( Y_3 = sC_3, \) \( Y_4 = \frac{1}{R_4}, \) \( Y_5 = \frac{1}{R_5}, \) then band pass and low pass responses can be obtained at \( V_1 \) and \( V_2, \) respectively, as given below:

\[
T_{BP}(s) = \frac{V_2}{V_i} = \frac{s}{R_2C_1D(s)}
\]

and

\[
T_{LP}(s) = \frac{V_2}{V_i} = \frac{1}{R_2R_4C_2C_3D(s)}
\]

where, denominator, \( D(s), \) is given by:

\[
D(s) = s^2 + \frac{s}{R_1C_1} + \frac{1}{R_2R_4C_1C_3}
\]

The pole frequency \( (\omega_p) \) and the quality factor \( (Q) \) obtained from the denominator polynomial, \( D(s), \) and the gain of the BP and LP filters are given by:

\[
\omega_p = \sqrt{\frac{1}{R_2R_4C_1C_3}}, \quad Q = R_1\sqrt{\frac{C_1}{R_2R_4C_3}},
\]

\[
H_{BP} = \frac{R_1}{R_5}, \quad H_{LP} = \frac{R_2}{R_5}
\]

From eqn. (3.36), it can be noted that the pole-Q can be tuned independently through \( R_1, \) without disturbing the pole-\( \omega_p. \) Also the gain of the low pass filter and band pass filter can be adjusted independently by grounded resistor \( R_5. \)
Case 2: If the admittances are selected as: 

\[ Y_1 = sC_1 + \frac{1}{R_1}, \quad Y_2 = sC_2, \quad Y_3 = \frac{1}{R_3}, \]

\[ Y_4 = sC_4, \quad Y_5 = sC_5, \]

then band pass and high pass responses are respectively obtained at \( V_1 \) and \( V_2 \). These are given by:

\[
T_{BP}(s) = \frac{V_1}{V_i} = \frac{sC_5}{C_2C_4R_3D'(s)} \tag{3.37}
\]

\[
T_{HP}(s) = \frac{V_2}{V_i} = \frac{s^2C_5}{C_4D'(s)} \tag{3.38}
\]

where, denominator, \( D'(s) \) is given by:

\[
D'(s) = s^2 + \frac{sC_1}{C_2C_4R_3} + \frac{1}{R_1R_3C_2C_4} \tag{3.39}
\]

The pole frequency \( \omega'_p \) and the quality factor \( Q' \) and the gain of the band pass and high pass filter are given by:

\[
\omega'_p = \sqrt{\frac{1}{R_1R_3C_2C_4}}, \quad Q' = \frac{1}{C_4} \sqrt{\frac{C_2C_4R_3}{R_1}}, \quad H'_{BP} = \frac{C_4}{C_1}, \quad H'_{HP} = \frac{C_5}{C_4} \tag{3.40}
\]

From equation (3.40), the pole-\( Q' \) can be tuned independently through \( C_1 \), and the gain of the band pass filter and high pass filter can be adjusted by \( C_5 \). It may be noted that multifunctional filter realizes standard low pass, band pass and high pass responses once again, without requiring matching conditions.

3.4.2 Non-ideal analysis

Considering the non-idealities of CCII of eqn. (3.9), the analysis of the circuit of Fig. 3.7 yields the transfer functions for the Case 1 as:


\[ T_{bp}(s) = \frac{V_1}{V_i} = \frac{s\alpha_1\beta_1 / R_1 C_1}{D(s)} \quad (3.41) \]

\[ T_{lp}(s) = \frac{V_2}{V_i} = \frac{\alpha_2\alpha_3\beta_2 / R_2 R_4 C_2 C_3}{D(s)} \quad (3.42) \]

where, denominator, \( D(s) \) is given by:

\[ D(s) = s^2 + \frac{s}{R_1 C_1} + \frac{\alpha_2\alpha_3\beta_2\beta_3}{R_2 R_4 C_2 C_3} \quad (3.43) \]

The filter parameters for the Case 1 are then given by:

\[ \omega_o = \sqrt{\frac{\alpha_1\alpha_2\alpha_3\beta_2\beta_3}{R_2 R_4 C_2 C_3}}, \quad Q = R_1 \sqrt{\frac{C_1\alpha_1\alpha_2\alpha_3\beta_2\beta_3}{R_2 R_4 C_2 C_3}} \quad (3.44) \]

and the low pass and band pass filter gains are:

\[ H_{LP} = \frac{R_1\alpha_1\beta_1}{R_2}, \quad H_{BP} = \frac{R_4\beta_1}{R_2\alpha_3\beta_3} \quad (3.45) \]

For the Case 2, the transfer functions are given as:

\[ T_{bp}(s) = \frac{V_1}{V_i} = \frac{sC_5\beta_1}{s^2 C_1 C_4 R_1 R_2 R_4 C_2 C_3 \alpha_1\alpha_2\alpha_3\beta_2\beta_3} \quad (3.46) \]

\[ T_{lp}(s) = \frac{V_2}{V_i} = \frac{s^2 C_5\beta_1}{D(s)} \quad (3.47) \]

where, denominator, \( D(s) \) is given by:

\[ D(s) = s^2 + \frac{s C_1}{C_2 C_4 R_1 R_4 \alpha_1\alpha_2\alpha_3\beta_2\beta_3} + \frac{1}{R_2 R_4 C_2 C_4 \alpha_1\alpha_2\alpha_3\beta_2\beta_3} \quad (3.48) \]

For the Case 1, slight reduction in pole-\( \omega_o \) and pole-Q is observed due to the device non-idealness and the non ideal gains demonstrate enhancement in \( H_{LP} \) and slight reduction in \( H_{BP} \) for \( f \leq 10 \text{ MHz} \).

The filter parameters for the Case 2 are:
\[ \omega' = \sqrt{\frac{1}{R_1 R_2 C_2 C_4 \alpha_1 \alpha_2 \alpha_3 \beta_1 \beta_2 \beta_3}}, \quad Q' = \frac{1}{C_1} \sqrt{\frac{C_2 C_4 R_1 \alpha_1 \alpha_2 \alpha_3 \beta_1 \beta_2}{R_1}} \]  

(3.49)

and the BP and HP filter gains are:

\[ H'_{BP} = \frac{C_2 \alpha_1 \beta_1}{C_1}, \quad H'_{HP} = \frac{C_2 \beta_1}{C_4 \alpha_1 \beta_2} \]  

(3.50)

For the Case 2, non-idealness causes slight enhancement in pole-\(\omega_p\) and reduction in pole-\(Q\). The filter gains demonstrate slight enhancement in \(H_{HP}\) and reduction in \(H_{BP}\).

### 3.4.3 Sensitivity study

The sensitivities of filter parameters pole-\(\omega_p\) and pole-\(Q\) are evaluated. These are summarized below:

\[ S_{R_2, R_4, C_3}^{\omega_p} = -\frac{1}{2}, \quad S_{C_3}^{Q} = \frac{1}{2}, \quad S_{R_1}^{Q} = 1, \quad S_{R_2, R_4, C_3}^{Q} = -\frac{1}{2}, \]

\[ S_{a_1, a_2, a_3, \beta_1, \beta_3}^{\omega_p} = \frac{1}{2}, \quad S_{a_1, a_2, a_3, \beta_1, \beta_3}^{Q} = -\frac{1}{2} \]  

(3.51)

Similarly, the sensitivities of \(\omega'_p\) and \(Q'\) are given by:

\[ S_{R_1, R_3, C_2, C_4}^{\omega'_p} = \frac{1}{2}, \quad S_{C_1}^{Q'} = -1, \quad S_{R_1}^{Q'} = \frac{1}{2}, \quad S_{C_3, C_4, R_3}^{Q'} = -\frac{1}{2}, \]

\[ S_{a_1, a_2, a_3, \beta_1, \beta_3}^{\omega'_p} = -\frac{1}{2}, \quad S_{a_1, a_2, a_3, \beta_1, \beta_3}^{Q'} = \frac{1}{2} \]  

(3.52)

The sensitivities for the filter gains are given by:

\[ S_{R_2, a_1, \beta_3}^{H_{HP}} = 1, \quad S_{R_1}^{H_{HP}} = -1, \quad S_{R_3, \beta_3}^{H_{HP}} = 1, \quad S_{R_1, a_1, \beta_3}^{H_{HP}} = -1, \]

\[ S_{C_1, a_1, \beta_3}^{H_{BP}} = 1, \quad S_{C_1}^{H_{BP}} = -1, \quad S_{C_3, \beta_3}^{H_{BP}} = 1, \quad S_{C_1, a_1, \beta_3}^{H_{BP}} = -1 \]  

(3.53)

Eqns. (3.51) to (3.53) show that all the sensitivities are found to be attractive, less than or equal to unity in magnitude.
3.4.4 Design and simulation

To demonstrate the performance of CCI multifunctional filter (MBF3), the low pass and band pass circuits of Case 1 are simulated using PSpice. The circuit was designed for \( f_0 = 300 \text{ KHz} \) with unity gain. The designed values are:

**For the LPF for Butterworth response (\( Q = 0.707 \)):** \( C_1 = C_3 = 45 \text{ pF, } R_2 = R_4 = R = 11.78 \text{ K\( \Omega \)} \) and \( R_4 = R_5 = 8.32 \text{ K\( \Omega \)} \).

**For BPF for a \( Q \) of 5:** \( C_1 = C_3 = 45 \text{ pF, } R_2 = R_4 = R = 11.78 \text{ K\( \Omega \)} \) and \( R_4 = R_5 = 58.9 \text{ K\( \Omega \)} \). The simulated responses are shown in Fig. 3.8 and the simulated results given in the Table. Good conformity is observed between the design and simulation.

![Simulated results](image)

<table>
<thead>
<tr>
<th>( f_0 (\text{KHz}) )</th>
<th>( Q_{\text{BP}} )</th>
<th>( H_{\text{LP}} )</th>
<th>( H_{\text{HP}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>300.06</td>
<td>4.68</td>
<td>1.00</td>
<td>1.01</td>
</tr>
</tbody>
</table>

Fig.3.8 Second order LP and BP responses at \( f_0 = 300 \text{ KHz} \)

The high pass and band pass filters of Case 2 are then designed for \( f_0 = 300 \text{ KHz, } Q = 0.707 \), at unity gain. The designed values of the BP and HP filters are:
**For band pass filter:** $C_2 = C_4 = 45 \text{ pF}$, $R_1 = R_3 = 11.78 \text{ K}\Omega$ and $C_1 = C_5 = 63.64 \text{ pF}$.

**For high pass filter:** $C_2 = C_4 = C_5 = 45 \text{ pF}$, $R_1 = R_3 = 11.78 \text{ K}\Omega$ and $C_1 = 63.64 \text{ pF}$.

The simulated high pass and band pass responses are shown in Fig. 3.9, along with the simulation results. The results exhibit good agreement with the theory.

![Simulated results table](image)

<table>
<thead>
<tr>
<th>$f_0$ (KHz)</th>
<th>$Q_{HP}$</th>
<th>$H_{HP}$</th>
<th>$H_{BP}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>300.08</td>
<td>4.69</td>
<td>1.00</td>
<td>1.02</td>
</tr>
</tbody>
</table>

Fig. 3.9 Second order HP and BP responses for $f_0 = 300 \text{ KHz}$.

The tunability of the circuit was investigated by changing the center frequency ($f_0$) of the band pass filter at a constant $Q$ of 5. The BP response curves corresponding to $f_0 = 300 \text{ KHz}$, $f_0 = 500 \text{ KHz}$, and $f_0 = 1 \text{ MHz}$ are given in Fig. 3.10. It is evident that the gain of the BPF remains nearly unity where frequency is varied. The simulated values show good agreement with the theory.
Table 3.2 (c): Input and Output Resistances of MBFs

<table>
<thead>
<tr>
<th>Type of MBF</th>
<th>Input Resistance</th>
<th>Output Resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>MBF3</td>
<td>$1 \times 10^{14} \text{M}\Omega$</td>
<td>$16.6 \text{K}\Omega$</td>
</tr>
</tbody>
</table>

A multifunctional filter is realized, using three CCII+ and grounded passive components, which is advantageous from integrated circuit implementation point of view. In addition to this, circuit offers several advantages, such as, high input impedance, which enable easy cascading, low sensitivities of filter parameters, versatility to synthesize low pass, band pass and high pass filter responses without any component matching considerations. The filter provides independent control of pole frequency, quality factor and gain of filters through separate grounded resistors and capacitors. The proposed
circuit also has low sensitivity and use of low supply voltage operation at ± 0.75V.

### 3.5 Comparative Study of Proposed MBFs

This Section presents the comparative study of the three novel multifunctional biquadratic filters (MBFs) presented in the preceding Sections. All the three MBFs give realization of standard second order LP, HP, and BP responses. Some of the parameters considered for comparison in the realization and performance of the three MBFs are:

(i) Number of CCIIs employed

(ii) Number of passive components employed

(iii) Requirement of any matching conditions

(iv) Suitability to realize high order filters (cascadablity)

(v) Independent tuning of filter parameters, such as, pole-ω₀, pole-Q

(vi) Sensitivity performance

Based on the above, the comparative performance of the MBFs is summarized in Table 3.3.

**Table 3.3 Comparison of MBFs**

<table>
<thead>
<tr>
<th>Performance parameter</th>
<th>MBF1</th>
<th>MBF2</th>
<th>MBF3</th>
</tr>
</thead>
<tbody>
<tr>
<td>(i) No. of CCIIs employed</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>(ii) No. of grounded passive elements</td>
<td>Four out of 5</td>
<td>All Six</td>
<td>All Six</td>
</tr>
<tr>
<td>(iii) Matching requirements</td>
<td>No</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>(iv) Convenient cascadablity</td>
<td>yes</td>
<td>yes</td>
<td>yes</td>
</tr>
<tr>
<td>(v) Independent tunability</td>
<td>Only of gain</td>
<td>Only of gain</td>
<td>yes</td>
</tr>
<tr>
<td>(vi) Low sensitivity</td>
<td>Yes</td>
<td>yes</td>
<td>yes</td>
</tr>
</tbody>
</table>
The MBF1 is a low component filter, realized without feedback, has independent tuning of its gain without disturbing the pole frequency and quality factor. It has the drawback of having one ungrounded component and also realizes only low Q values (Q<1). The lack of independent tuning of filter parameters, pole-\(\omega_o\) and pole-Q, is basically due to the use of low component count.

The MBF2 uses an additional active device, CCII- and also employs one more passive component. However, all passive components are grounded. In MBF1 the input resistance depends upon the L-section connected at the Y-input of CCII. However, in MBF2 signal is directly applied to Y-terminal insuring high input resistance. Its other performance features are similar to MBF1. The overall observation is that there is no appreciable improvement in MBF2 over MBF1, after using an additional passive component and a negative-type CCII.

In MBF3 the number of CCII is one more then MBF2. However all the three CCIIIs are positive type only. The circuit once again uses all six grounded elements. The MBF3 provides independent control of pole frequency, quality factor and gain of filters. Also, it does not impose restriction on Q-values as in MBF1 and MBF2. It may be seen that MBF3 presents important advantages of independent parameter tuning and realization of higher Q-values over MBF1 and MBF2. The price paid is in the use of three CCII+ devices.

3.6 VM Universal Biquadratic Filter using Two CCIIIs

Interest in the design of voltage mode universal biquadratic filter (UBF) with multi inputs and single output (MISO) [48-54] has generally the following advantages over the MBFs using single input and single output (SISO) topology: (i) realization of different filter functions from the same basic circuits, (ii) reduced number of active and passive components, (iii) greater
multifunctionality and simplicity in the design, and (iv) economical realization. Recently, many voltage mode UBFs with multi inputs and single outputs have been proposed [48-54]. The circuits of Refs [49, 52-54] have the drawback of use of excessive number of both active and passive components. Also, they are not extendable from CCII to CCCII based realizations, which have attraction of electronic tunability of filter parameters. The circuit of Ref. [51] requires complex matching constraints.

### 3.6.1 Circuit description

The proposed circuit of the MISO universal biquadratic filter [P4] is shown in Fig. 3.11. It uses low component count of only two low voltage DO-CCIIIs, along with, only two resistors and two capacitors in its realization.

![Fig. 3.11.MISO universal biquadratic filter](image)

The filter will be shown to provide six standard responses, through appropriate selection of its inputs. For an ideal DO-CCII, the v-i relations are defined by:
Routine analysis of Fig. 3.11, gives the voltage transfer function as:

\[
V_s = \frac{s^2 V_1 + s \left( \frac{1}{R_2 C_2} V_3 - \frac{1}{R_2 C_2} V_2 + \frac{1}{C_2 R_4} V_4 \right) + \frac{V_i}{R_1 R_2 C_1 C_2}}{s^2 + s \left( \frac{1}{C_2 R_1} + \frac{1}{R_1 R_2 C_1 C_2} \right)}
\]

where, the characteristic polynomial, \( D(s) \), is given by:

\[
D(s) = s^2 + s \left( \frac{1}{C_2 R_1} + \frac{1}{R_1 R_2 C_1 C_2} \right)
\]

From eqn. (3.55), various filter responses can be obtained through appropriate selection of the inputs:

(i) HP-response: \( V_4 = V_{in}, V_1 = V_2 = V_3 = 0 \),
(ii) NIBP-response: \( V_3 = V_{in}, V_1 = V_2 = V_4 = 0 \),
(iii) IBP-response: \( V_2 = V_{in}, V_1 = V_3 = V_4 = 0 \),
(iv) LP-response: \( V_1 = V_2 = V_{in}, V_3 = V_4 = 0 \), and \( R_1 = R_2 \),
(v) BE-response: \( V_1 = V_2 = V_4 = V_{in}, V_3 = 0 \), and \( R_1 = R_2 \),
(vi) AP-response: \( V_1 = V_2 = V_4 = V_{in}, V_3 = 0 \), and \( R_1 = 2R_2 \).

It may be noted that the realizations of LP, NIBP and IBP responses are completely free from matching constraints [case (i) to (iii)]. Minor constraints in the case of LP, BE and AP are present, but these are also simple to satisfy through design.

The pole frequency \( (\omega_o) \) and the pole-Q of the proposed UBF obtained from \( D(s) \) are:

\[
\omega_o = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}, \quad Q = \frac{\sqrt{R_1}}{R_2} \frac{C_2}{C_1}
\]

3.6.2 Non-ideal analysis

Taking the non-idealities of \( \alpha \) and \( \beta \) into consideration, the port relationships of DO-CCII at low and medium frequencies are given by:
Taking these non-idealities into consideration, the denominator of eqn. (3.56) of the transfer function is modified to:

\[ D'(s) = s^2 + \frac{\alpha_1 \beta_1 + \alpha_2 \beta_1 \beta_2}{C_1 R_1} + \frac{\alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2} \]  

(3.60)

Now, the filter parameters are given by:

\[
\omega_o' = \sqrt{\frac{\alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2}}, \quad Q' = \sqrt{\frac{\alpha_2 \beta_1 R_2 C_2}{\alpha_1 \beta_1 R_2 C_1}} \quad (3.61)
\]

It is evident from eqn. (3.61), that the effect of non-idealities does not affect the pole-\(Q\). However, they cause slight reduction in pole-\(\omega_o\).

### 3.6.3 Sensitivity study

The sensitivities of filter parameters \(\omega_o\) and \(Q\) with respect to active and passive components are evaluated and are summarized below:

\[
S_{R_1, R_2, C_1, C_2}^{\omega_o} = -\frac{1}{2}, \quad S_{\alpha_1, \alpha_2, \beta_1, \beta_2}^{\omega_o} = \frac{1}{2},
\]

\[
S_{R_1, C_1, \alpha_1, \beta_1}^{Q} = -\frac{1}{2}, \quad S_{R_2, C_2, \alpha_2, \beta_2}^{Q} = \frac{1}{2} \quad (3.62)
\]

From eqn. (3.62), it is clear that all the active and passive sensitivity figures are attractive, being equal to half in magnitude. This is an attractive performance feature of a filter.

### 3.6.4 Design and simulation

To demonstrate the performance of universal biquadratic filter, the circuit is simulated using PSpice. Initially the UBF was designed for \(f_o = 500\) KHz with a gain of unity at \(Q = 0.707\). For \(R_1 = R_2 = R = 15.9\) K\(\Omega\), eqn. (3.58) yields: \(C_1 = 28\) pF and \(C_2 = 14\) pF. The simulated UBF responses for LP,
HP, BP, BE, and AP are shown in Fig. 3.12, along with the simulated parameter values. This exhibits good agreement with the theory.

![Simulated results](image)

<table>
<thead>
<tr>
<th>$f_c$ (KHz)</th>
<th>$Q_{BP}$</th>
<th>$Q_{BE}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>500.07</td>
<td>0.705</td>
<td>0.703</td>
</tr>
</tbody>
</table>

Fig. 3.12 The simulated UBF response at $f_c = 500$ KHz

The UBF was then tuned by varying the resistor $R_2$. The BP responses corresponding to $f_c = 300$ KHz, $f_o = 500$ KHz and $f_c = 1$ MHz at $Q = 5$ and $H_{BP} = 1$ are given in Fig. 3.13, together with the simulated filter parameters. These are found to be in close conformity with the theory. These also exhibit convenient tunability of $f_c$ at a constant $Q$ and filter gain. Also the magnitude of BP responses remains unchanged with the variation of frequencies.
All six standard biquadratic responses, viz, low pass, high pass, non-inverting band pass, inverting band pass, band elimination and all pass are realized using a new multi inputs and single output voltage mode universal biquadratic filter. The filter enjoys attractive features, such as, low active and passive component count, low sensitivity performance, operated at low supply voltage of $\pm 0.75V$. The filter was also designed and verified using PSpice with convincing results.

### 3.7 Current Mode Realizations: First Order Filter Sections

As has been pointed out in Sec. 1.6.2, the current mode (CM) filters implemented through CCIIs have become very attractive due to their better linearity, simplicity, wider bandwidth, large dynamic range and low power consumption than their voltage mode counterparts. In this Section, a generalized approach is used to realize first order CM-LP/HP filters. The

<table>
<thead>
<tr>
<th>$f_r$ (BP)</th>
<th>$Q$ (BP)</th>
<th>$H_{BP}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>300.02 KHz</td>
<td>4.73</td>
<td>1.00</td>
</tr>
<tr>
<td>500.07 KHz</td>
<td>4.57</td>
<td>1.03</td>
</tr>
<tr>
<td>1.05 MHz</td>
<td>4.46</td>
<td>1.04</td>
</tr>
</tbody>
</table>

Fig. 3.13 Frequency tuning of BPF at constant $Q = 5$
generalized first order CM filter section is shown in Fig. 3.14. It consists of single CCII+, alongwith, three grounded admittances. Routine circuit analysis, with v/i-relationships of CCII+ defined by eqn. (3.1), yields the current transfer function as:

\[ T_i(s) = \frac{I_o}{I_i} = \frac{Y_1}{Y_2 + Y_3} \]  

(3.63)

![Fig.3.14 Generalized first order Current mode section](image)

3.7.1 *Realization of LP and HP filters*

Next the realization of low pass and high pass filters [P3] is considered as given below.

*Low pass filter:* If we select, \( Y_1 = \frac{1}{R_1} \), \( Y_2 = sC_2 \), and \( Y_3 = \frac{1}{R_3} \), then it results in the realization of low pass filter shown in Fig. 3.15, having current transfer function:

\[ T_{LP}(s) = \frac{I_o}{I_i} = \frac{1/R_1C_2}{s + 1/R_3C_2} \]  

(3.64)
Fig. 3.15 First order low pass filter

**High pass filter:** With the component selection as: \( Y_1 = sC_1, Y_2 = sC_2 \), and \( Y_3 = \frac{1}{R_3} \), a high pass filter is realized, as shown in Fig. 3.16. Its current transfer function is given by:

\[
T_{HP}(s) = \frac{I_o}{I_r} = \frac{sC_1/C_2}{s + 1/R_3C_2}
\]  

(3.65)

Fig.3.16 First order high pass filter.

In both the realizations, the pole frequency and filter gains are:

\[
\omega_o = \frac{1}{R_3C_2}, \quad H_{LP} = \frac{R_i}{R_4} \quad \text{and} \quad H_{HP} = \frac{C_1}{C_2}
\]  

(3.66)
From eqn. (3.66), it can be seen that the gain of low pass and high pass filter can be tuned independently through R1 and C1, respectively, without disturbing the pole frequency.

### 3.7.2 Non-ideal analysis

Consider the non-ideal CCII characterized by the frequency independent current transfer factor ‘α’ and voltage transfer factor ‘β’ for low to medium frequency applications. Taking these non-idealities into account, the analysis of the circuit of Fig. 3.15 and Fig. 3.16 still yields standard low pass and high pass current transfer functions, as given by:

\[
T_{lp}(s) = \frac{\alpha \beta}{s + \frac{1}{R_1 C_2}} \quad (3.67)
\]

\[
T_{hp}(s) = \frac{s \alpha \beta C_1}{s + \frac{C_2}{R_2 C_2}} \quad (3.68)
\]

The non-ideal filter parameters are given by:

\[
\omega'_o = \frac{1}{R_1 C_2}, \quad H_{lp}' = \frac{\alpha \beta R_1}{R_1} \quad \text{and} \quad H_{hp}' = \frac{\alpha \beta C_1}{C_2} \quad (3.69)
\]

It may be noted from eqn. (3.69) that only the filter gains get affected by the non-idealities and are slightly lowered. No effect is seen on pole frequency- \(\omega_o\), which is an attractive feature.

### 3.7.3 Sensitivity study

The sensitivities of filter parameters are evaluated with respect to active and passive elements. These are given as:
\[ S_{k_1}^{\text{le}} = -1, \quad S_{k_1}^{\text{ue}} = -S_{k_1}^{\text{ue}} = 1, \quad S_{\chi_1}^{\text{ue}} = -S_{\chi_2}^{\text{ue}} = 1, \]
\[ S_{\alpha, \beta}^{\text{le}} = 1, \quad S_{\alpha, \beta}^{\text{ue}} = 1 \]  

(3.70)

It is evident from eqn. (3.70) that all the sensitivity figures are found to be equal to unity in magnitude. Hence the proposed current mode circuits have reasonably low sensitivity performance.

### 3.7.4 Design and simulation

To evaluate the performance of the circuits shown in Fig. 3.15 and Fig. 3.16, PSpice simulation is used. The low pass and high pass filters are designed for \( f_0 = 500 \text{ KHz} \) and with a gain of unity. The designed values are: \( C_1 = C_2 = C = 76 \text{ pF} \) and \( R_1 = R_3 = R = 4.19 \text{ K} \Omega \). The simulated LP and HP responses are shown in Fig. 3.17. The simulated value of cutoff frequency is found to be 500.06 KHz, which exhibit close agreement with design.
The values of input and output impedances are given in Table 3.4. These have, $R_{in} \ll R_{out}$, thus exhibiting convenient cascadability of the proposed UBF circuit.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>39.7 KΩ</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>31.7 MΩ</td>
</tr>
</tbody>
</table>

The first order low pass and high pass filters with high output impedance use only one CCII+ along with all grounded components, which is attractive for IC implementation. These can be used to realize higher order low pass and high pass filters. It also has the advantages of low sensitivity, low component count and use of low supply voltage.

3.8 CM High Output Impedance UBF using MO-CCIIIs

The design of current mode universal biquadratic filters (UBF) with single input multi output (SIMO) and multi input single output (MISO) have received considerable attention due to their high performance, simple design and greater functional versatility in signal processing applications [58-67]. Several circuits for the realization of universal filters having single input multi output (SIMO) current transfer functions have been reported in the literature [58-67]. The SIMO current mode universal biquads in [60, 61, 62], require four current conveyors. In [59], SIMO current mode UBF is proposed using three multiple outputs second generation current conveyors (MO-CCIIIs), one OTA, two grounded resistors and three grounded capacitors. In [64], the current mode SIMO uses three MO-CCIIIs, two grounded resistors and two grounded capacitors. In Refs. [59 and 64], capacitors are connected at X-terminal of the MO-CCIIIs. The current conveyor based filters with X-terminal loaded by a capacitor do not exhibit good performance at high frequency due to the effect
of the parasitic resistance $R_x$ [34]. The SIMO universal biquads in [61, 65],
use four current controlled conveyors and two floating capacitors. Recently in
[66], a current mode SIMO UBF is presented employing three MO-CCIIs and
six passive components, out of which one is floating. In [67], a CM
multifunctional filter is realized using universal current conveyors with large
number of active and passive components. Thus, all the mentioned CM UBFs
suffer from large component count.

### 3.8.1 Circuit description

In this Section, a novel current mode universal biquadratic filter with
single input multi output using low voltage MO-CCII is considered. The filter
circuit is simple in structure and is shown in Fig. 3.18.

![Fig. 3.18.CM MO-CCII based UBF](image)

It consists of only two MO-CCIIs, along with, two grounded capacitors, and
two grounded resistors. The circuit presents active and passive component
minimization over the available CCII-based universal filters [56-67]. Analysis
of the circuit yields the following current transfer functions:

$$T_{LP}(s) = \frac{I_{LP}}{I_{IN}} = \frac{1}{R_1R_2C_1C_2D(s)}$$  \hspace{1cm} (3.71)
\[ T_{BP}(s) = \frac{I_{BP}}{I_{IN}} = \frac{s}{R_1 C_1 D(s)} \]  

(3.72)

\[ T_{BE}(s) = \frac{I_{BE}}{I_{IN}} = \frac{s^2 + \frac{1}{R_1 R_2 C_1 C_2}}{D(s)} \]  

(3.73)

where, the denominator is given by:

\[ D(s) = s^2 + \frac{s}{R_1 C_1} + \frac{1}{R_1 R_2 C_1 C_2} \]  

(3.74)

From eqns. (3.71), (3.72) and (3.73), it is seen that inverting low pass, inverting band pass and non-inverting band elimination responses are realized at the three outputs of the circuit. Non-inverting high pass filter is realized just by adding the high impedance outputs, \( I_{LP} \) and \( I_{BE} \). Also, an all pass filter response is realized by connecting the high impedance outputs, \( I_{BP} \) and \( I_{BE} \). The realized high pass and all pass filter responses, respectively, are given by the following equations:

\[ T_H(s) = \frac{I_{HP}}{I_{IN}} = \frac{s^2}{D(s)} \]  

(3.75)

\[ T_{AP}(s) = \frac{I_{AP}}{I_{IN}} = \frac{s^2 - \frac{s}{R_1 C_1} + \frac{1}{R_1 R_2 C_1 C_2}}{D(s)} \]  

(3.76)

The pole frequency and the quality factor of the filter, obtained from the characteristic polynomial, \( D(s) \), are given by:

\[ \omega_o = \sqrt{\frac{1}{R_1 R_2 C_1 C_2}}, \quad Q = \sqrt{\frac{R_1 C_1}{R_2 C_2}} \]  

(3.77)

### 3.8.2 Non-ideal analysis

Taking the non-idealities \( \alpha_i \) and \( \beta_i \), \( i = 1 \) and \( 2 \), into consideration, the current transfer functions of the UBF are given then by:
\[ T_{\text{LP}}(s) = \frac{I_{\text{LP}}}{I_{\text{IN}}} = -\frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2 D'(s)} \] (3.78)

\[ T_{\text{HP}}(s) = \frac{I_{\text{HP}}}{I_{\text{IN}}} = -\frac{s \alpha_1 \beta_1}{R_1 C_1 D'(s)} \] (3.79)

\[ T_{\text{BP}}(s) = \frac{I_{\text{BP}}}{I_{\text{IN}}} = -\frac{s^2 + \alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2 D'(s)} \] (3.80)

\[ T_{\text{HP}}(s) = \frac{I_{\text{HP}}}{I_{\text{IN}}} = \frac{s^2}{D'(s)} \] (3.81)

\[ T_{\text{BP}}(s) = \frac{I_{\text{BP}}}{I_{\text{IN}}} = \frac{s^2 - s \alpha_1 \beta_1 + \alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 C_1 R_1 R_2 C_1 C_2 D'(s)} \] (3.82)

and

\[ D'(s) = s^2 + \frac{s \alpha_1 \beta_1}{R_1 C_1} + \frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2} \] (3.83)

where the pole frequency \( \omega_o' \) and the quality factor \( Q' \) of the filters obtained from \( D'(s) \), are given by:

\[ \omega_o' = \sqrt{\frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2}}, \quad Q' = \sqrt{\frac{R_1 C_1 \alpha_2 \beta_2}{R_2 C_2 \alpha_1 \beta_1}} \] (3.84)

At low to medium frequencies \( (\omega \leq 10 \text{ MHz}) \), the circuit continues to provide standard second order responses. The pole-\( \omega_o \) is slightly lowered, but the pole-Q remains unaffected by the non-idealities.

### 3.8.3 Sensitivity study

The sensitivity of the filter parameters are evaluated with respect to active and passive elements and are summarized as:

\[ S_{R_1, R_2, C_1, C_2}^{\omega_o} = -\frac{1}{2}, \quad S_{R_1, C_1, \alpha_2, \beta_2}^{Q} = \frac{1}{2} \]
which are seen to be equal to half, in magnitude. This shows the attractive sensitivity feature of the UBF.

3.8.4 Design and simulation

To demonstrate the performance of CM universal biquadratic filter, the circuit is simulated using PSpice. Initially the CM UBF is designed for a pole frequency \( f_o \) = 1 MHz with unity gain and \( Q = 0.707 \) (Butterworth response). For \( C_1 = C_2 = 11 \ \mu F \), eqn. (3.77) yields \( R_1 = 10 \ \text{K}\Omega \), and \( R_2 = 20 \ \text{K}\Omega \). The simulated LP, BP, HP, BE and AP responses, of the UBF are shown in Fig. 3.19, along with, the simulated parameters. The simulated value of \( f_o \) = 1.06 MHz. These show good agreement with the designed values.

<table>
<thead>
<tr>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_o ) (MHz)</td>
</tr>
<tr>
<td>1.06</td>
</tr>
</tbody>
</table>

Fig. 3.19 The simulated UBF response at \( f_o = 1 \) MHz
The frequency tuning of the BPF response at a constant $Q$ of 5 and $H_{BP}$ of unity is next investigated by changing the center frequency ($f_0$) of the band pass realization through resistor $R_2$. The BP response curves corresponding to $f_0 = 300$ KHz, $f_0 = 500$ KHz, and $f_0 = 1$ MHz are given in Fig. 3.20. These exhibit convenient tunability of the filter.

![Fig. 3.20 Frequency tuning of BPF at Q = 5](image)

The values of low input and high output resistances are given in Table 3.5, which show $R_{in} \ll R_{out}$. This confirms convenient cascadability of the proposed CM UBF in cascade synthesis for higher order filters.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>$6.3\Omega$</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>$41.2M\Omega$</td>
</tr>
</tbody>
</table>
A novel universal CM UBF is realized using two low voltage MO-CCIIs and a minimal count of all grounded passive components. This suits IC-fabrication. The proposed filter has high output and low input impedance, which enables its direct cascading in current mode operations. It realizes high order low pass, band pass, high pass, band elimination and all pass filter responses without requirement of matching conditions. Also, the UBF has the advantages of low sensitivity, low component count and operation at low supply voltage of ±0.75V.

### 3.9 Higher Order CM Filters

There are many applications that utilize first order and second order filters as basic building blocks to synthesize various higher order active filters. Since a higher order filter requires a large number of active elements, minimizing the number of CCIIs, has its obvious advantage of low cost and low power consumption. The high order filters of any order (n) can be realized by cascading the lower order filter sections in a non-interactive cascade.

#### 3.9.1 Realization of fourth order LP and HP filters

Current mode low pass or high pass filter of any order (n) [68, 69, 70] can be obtained by cascading n-identical first order sections. Consider the realization of nth order low pass filter. If $T_1(s)$, $T_2(s)$ and $T_n(s)$ are the current transfer functions of non-interactive first, second, ... and nth section of CM low pass cascaded network, then nth order (n) low pass filter can be realized, whose transfer function is given by:

$$T(s) = \frac{I_o}{I_i} = T_1(s)T_2(s)............T_n(s)$$  \hspace{1cm} (3.86)

The transfer function for nth order LP filter is given as:

$$T_{LP}^{n}(s) = \frac{K_s^n}{s^n + a_1s^{n-1} + a_2s^{n-2} + \ldots + a_n}$$  \hspace{1cm} (3.87)
Similarly, the transfer function for nth order HP filter is given as:

\[ T_{\text{HP}^n}(s) = \frac{K_o s^n}{s^n + a_1 s^{n-1} + a_2 s^{n-2} + \ldots + a_n} \quad (3.88) \]

Now consider the realization of fourth order low pass filter [P3]. It can be obtained by cascading four identical sections of Fig.3.15, For \( C_2 = C \) and \( R_1 = R_3 = R \), the transfer function of [P3] is given by:

\[ \frac{I_o}{I_i} = T_{LP^4}(s) = \frac{K_o^4}{s^4 + \frac{4s^3}{RC} + \frac{6s^2}{R^2C^2} + \frac{4s}{R^3C^3} + \frac{1}{R^4C^4}} \quad (3.89) \]

where \( \omega_o = \frac{1}{RC} \) \quad (3.90)

It can be noted that the identical first order low pass sections are used to realize high order LP filter. Similarly the 4th order high pass filter transfer function can be obtained and is given by:

\[ \frac{I_o}{I_i} = T_{HP^4}(s) = \frac{K_o^4 s^4}{s^4 + \frac{4s^3}{RC} + \frac{6s^2}{R^2C^2} + \frac{4s}{R^3C^3} + \frac{1}{R^4C^4}} \quad (3.91) \]

The pole frequency \( (\omega_o) \) of the high pass filter is given as:

\[ \omega_o = 1/R C \quad (3.92) \]

### 3.9.2 Design and simulation

To evaluate the performance of the fourth order low pass filter, it is simulated using PSpice. The circuit of low pass filter is designed for \( f_o = 2.5 \) MHz. The designed values are \( C = 16 \) pF and \( R = 3.97 \) KΩ. The simulated fourth order LP responses are shown in Fig.3.21 with a simulated cutoff frequency of 2.501 MHz. It exhibits a good agreement with the theory.
The proposed CM fourth order low pass filter is obtained by cascading non-interactive first order low pass filters described in Sec. 3.7 without requiring additional buffers. It employs all grounded components. It also has the advantages of low sensitivity, high frequency performance, and operation at low supply voltage of ±0.75V. The theoretical and simulation results confirm the practical utility of the proposed circuits.

3.10 Sixth Order Butterworth Filters Using CM UBF

The current mode universal biquadratic filter presented in this Chapter provide non-interactive block with low input and high output impedances. These can be directly cascaded in the CM signal processing, without using additional current followers, to realize higher order filters. Thus it is possible to realize \( n^{th} \) order filter by cascading \( n/2 \)-biquadratic filter sections for even \( n \), and \( (n-1)/2 \)-biquadratic sections, along with, an additional first order section for odd values of \( n \).

3.10.1 Realization of sixth order LP and BP filters

Here we consider the realization of \( 6^{th} \) order Butterworth low pass filter using three CM UBF considered in Sec. 3.8 (Fig. 3.18). The normalized
Butterworth transfer function for the resulting 6th order CM low pass filter is given by [70]:

\[
T(s) = \frac{1}{(s^2 + 1.932s + 1)(s^2 + 1.414s + 1)(s^2 + 0.518s + 1)}
\]

(3.93)

The normalized pole frequency is at \( \omega_n = 1 \). The transfer function can be de-normalized by replacing \( S \rightarrow s/\omega_n \) to give the required sixth order filter function at given pole- \( \omega_n \) and pole-Q. The pole-Q of an individual biquadratic filter section is simply the reciprocal of the coefficients of \( s \) in eqn. (3.93) [70]. The values of \( Q \) are 0.518, 0.707, and 1.932, respectively, for the three UBF to be cascaded. The UBF can be designed using these values of \( Q \) and for a given pole frequency. It is seen that no additional buffers are employed in the realization. The filter’s pole frequency \( \omega_n \) and the quality factor \( Q \) of the current mode MO-CCII-based UBF are given by:

\[
\omega_n = \sqrt{\frac{1}{R_1 R_2 C_1 C_2}}, \quad Q = \sqrt{\frac{R_1 C_1}{R_2 C_2}}
\]

(3.94)

### 3.10.2 Design and simulation

To evaluate the performance of the sixth order Butterworth low pass filter, it is simulated using PSpice. The circuit is designed for a \( f_o = 1 \) MHz. Equal valued capacitors are selected for convenience and are assumed to be equal to 15.9 pF each. The resistors for each section are designed to satisfy eqn. (3.94). The designed values for each section are given below.

**Section-I:** \( R_1 = 5.18 \, \text{K}\Omega, \, R_2 = 19.32 \, \text{K}\Omega, \) for pole \( Q = 0.518 \)

**Section-II:** \( R_1 = 7.076 \, \text{K}\Omega, \, R_2 = 14.15, \) for pole \( Q = 0.707 \)

**Section-III:** \( R_1 = 19.31 \, \text{K}\Omega, \, R_2 = 5.18 \, \text{K}\Omega, \) for pole \( Q = 1.932 \)

The resulting theoretical and simulated frequency response curves for the sixth order low pass filter are plotted in Fig. 3.22 (a), also in DB in Fig.3.22 (b).
Simulated results

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Simulated HLP</th>
<th>Theoretical HLP</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 KHz</td>
<td>1.04</td>
<td>1.02</td>
</tr>
<tr>
<td>300 KHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td>1 MHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td>3 MHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td>10 MHz</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Fig. 3.22 (a) Frequency response of sixth order CM LPF

Gain (DB)

Fig. 3.22 (b) Frequency response of sixth order CM LPF in DB

Gain (DB)
It is seen that the simulated pole frequency of 1.04 MHz, is obtained from the simulation, which verifies the design. Figure 3.22 (b) gives the stop band attenuation of 120 DB/decade, verifying the 6\textsuperscript{th} order low pass response. Through the entire range, the simulated and theoretical responses overlap, showing close agreement with theory.

The proposed circuit can also be used to realize other higher order responses, such as, band pass, high pass and band elimination filters, through a simple electronic switching arrangement [140], for selecting the desired response of the UBF. The result of the 6\textsuperscript{th} order band pass filter is shown in Fig. 3.23 (a), with a simulated pole frequency $f_p=1.02$ MHz and a pole Q of 1.84. The same response in DB, is also shown in Fig. 3.23 (b). The slopes below $f_p$ and above $f_p$ are each 60 DB/decade, thus verifying the 6\textsuperscript{th} order band pass response. At the pole frequency $f_p=1$ MHz, the gain is equal to unity.

<table>
<thead>
<tr>
<th>Simulated results</th>
<th>$f_p$ (BP)</th>
<th>$Q_{BP}$</th>
<th>$H_{BP}$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1.02 MHz</td>
<td>1.84</td>
<td>1.01</td>
</tr>
</tbody>
</table>

Fig.3.23 (a) Frequency response of sixth order CM BPF
Fig. 3.23 (b) Frequency response of sixth order CM BPF in DB

The MO-CCII based current mode universal biquadratic filter is used to realize sixth order low pass and band pass responses by cascading biquadratic filter sections, viz., UBF of Sec. 3.8, without using any additional current followers. The proposed circuit also has low sensitivity, low component count, using low supply voltage of $\pm 0.75V$.

3.11 Conclusion

This Chapter is concerned with the realization and study of some novel good performance voltage mode and current mode multifunctional filters using second generation current conveyors (CCIIs) as the active device. The first half of the Chapter focuses on the biquadratic voltage mode filters. In all the multifunctional biquadratic filters (MBFs) considered in the Chapter, the filter realizes standard LP, HP and BP responses through component selection ($R$’s, $C$’s) for realizing the said responses. A comparative study is also included for the three MBFs. It is found that MBF2 has no significant advantage over
MBF1, in spite of using one additional active and passive device. The advantage of MBF2 is in terms of using all grounded passive components and having unconditional high input impedance. Both suffer from lack of independent tuning and realization of low Q values. MBF3 uses an additional CCII over MBF2, but enjoys better tunability and no restriction on Q values. All the three realizations are found to have low sensitivities. They are also free from matching requirements. In the filters, mostly, grounded passive components are used, which is attractive in IC fabrication.

A new VM universal biquadratic filter (UBF) is given using two DO-CCIIIs and four passive components. By UBF, it is implied in the thesis that the circuit realizes all five standard second order responses viz., LP, HP, BP, BE, and AP. The circuit enjoys attractive low sensitivity property and either the matching conditions are absent or easy to satisfy practically in the realization of various responses. Also, the circuit does not impose any limitation on the realization of high Q values.

Next, the remaining part of the thesis proposes the study of CCII-based current mode realizations of first order sections and an UBF. From a general topology, first order LP/HP filter sections are obtained through component selection. Sensitivity studies are found to be attractive. The realization scheme of current mode multioutput CCII-based UBF is suggested. It is used in the realization of five standard biquadratic responses, without requirements of matching constraints. The circuit also has attractive low sensitivities. The Chapter includes the realization of higher order CM filters using the attractive cascadability properties of the CM filters already studied in the Chapter. This is demonstrated through the realization of fourth order LP and sixth order LP and BP filters.

In conclusion, it may be mentioned that the Chapter considers VM and CM filter realizations with attractive properties. All the circuits use either completely or mostly grounded passive components. The realizations of ideal response are free from strict matching constraints. The sensitivities are found to
be reasonably low in all the circuits. Some of the circuits are also studied considering non-idealities of the active device. In general it is seen that below 10 MHz, the non-idealities do not have serious effects on the circuit performance. All the suggested realizations are put to verifications through simulation using OrCAD-10 PSpice. The results in all the cases are found to be in conformity with the theory.