6.1 Introduction

Basically an OFC is a current mode (CM) device and is highly suitable in the realization of CM circuits. As has been pointed out earlier, such circuits have a number of attractive features, which have made them an obvious choice of the circuit designer.

This Chapter discusses the realization and study of some novel and useful CM circuits, which find applications in ASP. A versatile OFC-based CM Basic Building Block (BBB) is given. It is used in the realization of inverting and non-inverting current amplifiers in Sec. 6.2, ideal integrator and differentiator in Sec. 6.3, first order LP and HP filter sections in Sec. 6.4, and a first order AP filter in Sec. 6.5. The BBB is also used in the realization of second order OFC-based band pass filter in Sec. 6.6. Using the non-interactive property of the building block, an MBF is obtained in Sec. 6.7, which provides LP, BP and HP responses. A low component scheme is given for the realization of insensitive CM UBF in Sect.6.8, it is shown to realize all standard second order responses. Finally, in Sec. 6.9, cascade form of design is used in the realizations of fifth order (odd n) and an eighth order (even n) CM LP Butterworth filters. The BBB used in the realization is the CM UBF derived earlier. All the circuits considered in the Chapter are critically studied for their sensitivities, matching constraints and performance with non-ideal devices. The theory is verified through PSpice simulation for all the circuits.

Authors’ paper [P12] is based on this chapter
6.2 OFC-Based Current Mode Basic Building Block

Figure 6.1 shows a versatile current mode BBB [P12], realized with an OFC as the active device, alongwith, only two grounded passive admittances. For an OFC, the v-i relations are defined by:

\[ i_y = 0, \ v_x = v_y, \ v_w = z_i i_x, \ i_z = \pm i_w \]  \quad (6.1)

Routine analysis of the circuit yields:

\[ T(s) = \frac{I_a}{I_i} = \pm \frac{Y_1}{Y_2} \]  \quad (6.2)

By convention, positive is taken for OFC+, when both \( i_x \) and \( i_w \) flow simultaneously towards or away from the operational floating conveyor. The negative sign is for OFC-, indicating the current flow in opposite direction. The proposed circuit will be shown to be useful in the realization of inverting and non-inverting amplifiers, ideal integrator, differentiator, first order and second order cascadable filter sections, without imposing matching conditions.

6.2.1 Realization of Inverting and non-inverting amplifiers

By selecting, \( Y_1 = G_1 \) and \( Y_2 = G_2 \), the basic topology of Fig. 6.1 realizes OFC-based CM inverting and non-inverting amplifiers. These are shown in Fig.6.2 and have the transfer function:
It is evident from eqn. (6.3) that inverting or non-inverting amplifiers can be obtained from the same circuit by using negative ($Z^-$) or positive ($Z^+$) output Z-terminals of OFC.

\[ A_i(s) = \frac{I_o}{I_i} = \frac{\pm R_2}{R_1} \]  
(6.3)

It is evident from eqn. (6.3) that inverting or non-inverting amplifiers can be obtained from the same circuit by using negative ($Z^-$) or positive ($Z^+$) output Z-terminals of OFC.

![Fig.6.2 OFC-based current amplifiers](image)

6.2.2 Non-ideal analysis

Here we consider the non-idealities of OFC, viz., frequency effects and parasitic effects, in the realization of CM amplifiers.

(a) Frequency effects

Taking the non-idealities, frequency dependent $\alpha(s)$, $\beta(s)$ and $\gamma(s)$ of the OFC into consideration, the current transfer function of non-inverting amplifier of Fig. 6.2 is given by:

\[ A_n(s) = \frac{I_o}{I_i} = \frac{\alpha \beta R_2}{(s + \omega_\alpha)(s + \omega_\beta)R_1} \]  
(6.4)

It is evident from eqn. (6.4) that two extra poles appear due to single pole roll off model of the device non-idealities. Their effect on the frequency response can however be minimized if the operating frequencies are selected sufficiently smaller than the current and voltage corner frequencies of OFC.
In low to medium frequency range ($f \leq 10 \text{MHz}$), the frequency effect can safely be neglected.

(b) **Parasitic effects**

Using the parasitics of the OFC into consideration, the analysis of the circuit of Fig. 6.2 yields the transfer function:

$$A_n(s) = \frac{I_o}{I_i} = \frac{\alpha \beta (G_t + s C_x - \frac{1}{Z_t})}{(G_t + s C_t + G_t)}$$

or

$$A_n(s) = \frac{I_o}{I_i} = \frac{\alpha \beta (\frac{1}{R_t} + \frac{1}{Z_t} - \frac{1}{X_{cy}})}{(\frac{1}{R_2} + \frac{1}{R_t} + \frac{1}{X_{cy}})}$$

It may be noted that in eqn. (6.6), the values of the parasitic impedance ($Z_t$), resistance ($R_t$) and reactances ($X_{cx}$ and $X_{cy}$) are many magnitude higher than the circuit resistances $R_t$ and $R_2$. Hence, over the entire range of low and medium frequency range ($f \leq 10 \text{MHz}$) they are ineffective. This simplifies eqn. (6.6) to eqn. (6.7).

$$A_n(s) = \frac{I_o}{I_i} = \frac{\alpha \beta R_2}{R_t}$$

which exhibits slight reduction in current gain.

6.2.3 **Sensitivity study**

The active and passive sensitivities of gains of non-inverting and inverting amplifiers are evaluated and are found to be reasonable. These are summarized below:

$$S_{A_{t}}^{R_t} = -1, \quad S_{A_{t}}^{A_{t}} = 1, \quad S_{A_{t},t}^{A_{t}} = 1,$$

$$S_{A_{t}}^{A_{t}} = -1, \quad S_{R_t}^{A_{t}} = 1, \quad S_{R_t,t}^{A_{t}} = 1$$

(6.8)
6.2.4 Design and simulation

To verify the theory for the proposed circuits, the amplifiers were simulated using OFC model of Ref [11] with PSpice at a supply voltages of ±0.75V. To study the transient response of the CM inverting and non-inverting amplifiers, these were designed for gains of 2 and 4, respectively. Preseting the input current, $I_{in} = 1\ \mu A$ with $R_1 = 10\ \text{K}\Omega$, the resistance $R_2$ is obtained as 20 KΩ and 40 KΩ, respectively, corresponding to gains of 2 and 4. Time domain responses of OFC-based CM inverting and non-inverting amplifiers are shown in Fig. 6.3 and Fig. 6.4, respectively.

![Fig. 6.3 Input and output current waveforms for inverting amplifier](image)

![Fig. 6.4 Input and output voltage waveforms for non-inverting amplifier](image)
It is observed that the results are in close conformity with the theory. Also, at the frequency of 1 MHz, the THD was found as 0.9% for both inverting and non-inverting amplifiers.

The frequency response of inverting amplifier circuit for gain of 2 and 4 are shown in Fig. 6.5. It is observed that the circuit has bandwidth of 59 MHz and 36 MHz for gains of 2 and 4, respectively. Also, the frequency responses of non-inverting amplifier at gains of 2 and 4 are shown in Fig. 6.6. It is now observed that the circuit has bandwidth of 85 MHz and 48 MHz for a gain of 2 and 4, respectively. This shows superior high frequency performance of non-inverting amplifier.

Fig. 6.5 Frequency response of OFC-based inverting amplifier

Fig. 6.6 Frequency response of OFC-based non-inverting amplifier
In this Section, inverting and non-inverting amplifiers are realized using single OFC, alongwith, two grounded resistors. They have the advantages of low component count, superior high frequency performance and convenient operation at low supply voltage of ± 0.75volt. In the next Section, the basic topology is used for the realization of ideal inverting and non-inverting integrator and differentiator circuits.

6.3 CM Ideal Integrator and Differentiator

Active-RC integrator and differentiator circuits are popularly used building blocks in various signal processing applications, such as, active filters, process controller, waveform generator and calibration circuits. Many active-RC integrators and differentiators have been reported in literature [112-118]. However, most of these circuits [114-117] are in dual input voltage mode operation and use an excessive number of passive and active elements, besides requiring passive elements matching to realize ideal transfer function.

In this Section, simple CM ideal integrator and differentiator circuits [P12] are realized from the generalized scheme of Fig. 6.1. Each employs only a low voltage CMOS operational floating conveyor, alongwith, two grounded passive components.

6.3.1 Ideal integrator

By selecting, \( Y_1 = G_1 \) and \( Y_2 = sC_2 \), ideal current mode integrator is realized from Fig. 6.1. The circuit is shown in Fig. 6.7 and its analysis yields:

\[
T_i(s) = \frac{I_2}{I_1} = \pm \frac{1}{sR_iC_2} = \pm \frac{1}{sr_i} \quad (6.9)
\]

where \( r_i = R_iC_2 \) is the time constant of the ideal integrator. It can be seen from eqn. (6.9) that both ideal inverting and non-inverting integrators can be realized using only a ± OFC, alongwith, grounded R, C components. Also, the integrator’s time constant \( r_i \) can be tuned linearly by a single resistor \( R_i \).
6.3.2 Ideal differentiator

For the component choice of $Y_1 = sC_1$ and $Y_2 = G_2$, an ideal differentiator, shown in Fig. 6.8, is realized with,

$$T_d(s) = \frac{I_o}{I_i} = \pm sR_2C_1 = \pm s\tau_d \quad (6.10)$$

where $\tau_d = R_2C_1$, is the time constant of the differentiator. It can once again be seen from eqn. (6.10) that both the inverting and non-inverting ideal differentiators can be obtained by using an $\pm$ OFC. Also, the time constant ($\tau_d$) can easily be tuned by single resistor ($R_2$).
6.3.3 Non-ideal analysis

Taking into consideration the non-idealities of the OFC, analysis of the circuit yields the transfer functions for non-inverting/inverting integrator as:

\[
T_i(s) = \frac{I_o}{I_i} = \pm \frac{\alpha \beta}{s R_i C_2} \frac{1}{s \tau_i}
\]  

(6.11)

where \( \tau_i = \pm \frac{R_i C_2}{\alpha \beta} \) is the time constant.

Similarly, the non-ideal analysis of differentiator of Fig.6.8 yields the transfer function:

\[
T_d(s) = \frac{I_o}{I_i} = \pm s \alpha \beta R_i C_1
\]  

(6.12)

where \( \tau_d = \pm \alpha \beta R_i C_1 \) is the time constant. In the low to medium \( f \)-range, \( f \leq 10 \text{MHz} \), only slight \( \tau_i \)-enhancement and \( \tau_d \)-reduction are observed.

6.3.4 Sensitivity study

The sensitivities of the integrator and differentiator time constants, \( \tau_i \) and \( \tau_d \), are evaluated with respect to active and passive components and are given by:

\[
S_{R_i,C_2}^{\tau_i} = 1, \quad S_{R_d,C_1}^{\tau_d} = 1, \quad S_{\alpha,\beta}^{\tau_i} = -1, \quad S_{\alpha,\beta}^{\tau_d} = 1
\]  

(6.13)

All the sensitivities are found to be unity in magnitude.

6.3.5 Design and simulation

To verify theory of the proposed circuits, these are simulated using PSpice. A rectangular waveform \( I_{in} \) of 1 \( \mu \text{A} \) (PP) at 100 KHz, shown in Fig. 6.9 (a), is applied to the integrator, having the design values of \( R_1 = 25 \text{ K}\Omega \) and \( C_2 = 75 \text{ pF} \). The output is found to be a triangular waveform of 1.8 \( \mu \text{A} \) (PP), as shown in Fig 6.9 (b). The simulated output clearly demonstrates the conversion of rectangular wave signal to a good quality triangular wave.
Similarly, a triangular waveform $I_{in}$ of 1 $\mu A$ (PP) at 100 KHz, shown in Fig. 6.10 (a), is applied to a differentiator having designed values of $C_1 = 5$ pF and $R_2 = 10$ K$\Omega$. The output is a square wave with 20 nA (PP), as shown in Fig. 6.10 (b). Once again, the simulated result for output shows good quality conversion of a triangular waveform to square waveform.
In this Section, new active CM inverting and non-inverting integrator and differentiator circuits are presented using Operational Floating Conveyor (OFC) and grounded passive R/C components. Both integrator and differentiator have linear control of time constant by varying single resistor \( R_1 \) and \( R_2 \), respectively. The proposed circuits are expected to find applications in instrumentation systems, signal processing circuits and analog filters. The realizations are free from matching constraints. The circuit also enjoys the advantages of low component count, low sensitivities, single resistor tunability, reliable high frequency performance, use of low supply voltage and suitability to IC-fabrication.

6.4 First Order LP and HP Filter Sections

In this Section, realization of novel current mode first order filters from the generalized scheme of Fig. 6.1 is given. The first order LP and HP filter sections are realized through appropriate selection of admittances in the BBB.

6.4.1 Low pass section

For \( Y_1 = G_1 \) and \( Y_2 = G_2 + sC_2 \), the basic topology of Fig. 6.1 realizes a CM low pass filter, shown in Fig. 6.11.

![Fig. 6.11 OFC-based CM low pass filter](image-url)
It consists of a single OFC+, along with two grounded resistors and single grounded capacitor. Analysis yields the following current transfer function:

\[
T_{LP}(s) = \frac{I_o}{I_i} = \frac{1}{s + \frac{1}{C_2R_2}} + \frac{1}{C_1R_1}
\]  

(6.14)

**6.4.2 High pass section**

By selecting, \( Y_1 = sC_1 \) and \( Y_2 = G_2 + sC_2 \), a first order current mode high pass filter is realized, as shown in Fig. 6.12. It consists of a single OFC+, along with, two grounded capacitors and single grounded resistor. Routine analysis yields the current transfer function as:

\[
T_{HP}(s) = \frac{I_o}{I_i} = \frac{sC_1/C_2 + 1}{s + \frac{1}{C_2R_2}}
\]  

(6.15)

The circuit parameters are:

\[
\omega_o = \frac{1}{R_2C_2}, \quad H_{LP} = \frac{R_2}{R_1} \quad \text{and} \quad H_{HP} = \frac{C_1}{C_2}
\]  

(6.16)

![Fig. 6.12 OFC-based CM high pass filter](image)

It can be seen from eqn. (6.16), the gains of LP and HP filters are independently tunable through \( R_1 \) and \( C_1 \), respectively, without disturbing the pole frequency (\( \omega_o \)).
6.4.3 Non-ideal analysis

Taking into account the non-idealities of the OFC, analysis of Fig. 6.11 yields the low pass filter transfer function as:

\[
\frac{I_o}{I_1} = \frac{\alpha \beta / R_1 C_2}{s + \frac{1}{R_2 C_2}}
\]

(6.17)

Similarly, non-ideal analysis of Fig. 6.12 gives the high pass filter transfer function:

\[
\frac{I_o}{I_1} = \frac{s \alpha \beta C_1 / C_2}{s + \frac{1}{R_2 C_2}}
\]

(6.18)

The filter parameters are now given by:

\[
\omega_p = \frac{1}{R_1 C_2}, \quad H_{LP} = \frac{s \alpha \beta R_2}{R_1} \quad \text{and} \quad H_{HP} = \frac{s \alpha \beta C_1}{C_2}
\]

(6.19)

It may be noted from eqn. (6.19) that the non-idealities have no effect on pole frequency \(\omega_p\), which is an attractive feature of the filter. However, slight reduction in gains takes place.

6.4.4 Sensitivity Study

The sensitivity parameters are evaluated with respect to active and passive elements and are given as:

\[
S_{R_2, C_2}^{\omega_p} = -1, \quad S_{R_1}^{H_{LP}} = -S_{R_1}^{H_{LP}} = 1, \quad S_{C_1}^{H_{LP}} = -S_{C_1}^{H_{LP}} = 1, \\
S_{\alpha, \beta}^{\omega_p} = 0, \quad S_{\alpha, \beta}^{H_{LP}} = 1 \quad \text{and} \quad S_{\alpha, \beta}^{H_{LP}} = 1
\]

(6.20)
It is evident from eqn. (6.20) that all the sensitivities are found to be equal to unity in magnitude. Also, note that the pole frequency \( \omega_0 \) of the filter is insensitive to the non-idealities.

### 6.4.5 Design and simulation

To evaluate the performance, the circuits are simulated using OFC+ model [11]. The LP and HP filters are designed for \( f_o = 300 \text{ KHz} \). The designed values are \( C_1 = C_2 = 21.22 \text{ pF} \) and \( R_1 = R_2 = 25 \text{ K}\Omega \). The simulated LP and HP responses are shown in Fig. 6.13, with the simulated value of cutoff frequency, \( f_o = 300.09 \text{ KHz} \). They show good agreement with the design.

![Simulated results](image)

**Fig. 6.13 Frequency response of LP and HP filters**

Table 6.1 gives the simulated values of input and output impedances. These depict convenient cascaddability, of the proposed CM circuit, as \( R_{out} >> R_{in} \).

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>25 KΩ</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>31.76 MΩ</td>
</tr>
</tbody>
</table>
These current mode OFC-based first order sections are cascaded and can be used to realize high frequency, higher order low pass and high pass filters. It enjoys all the attractive features mentioned for the CM realizations derived earlier from the generalized scheme of Fig. 6.1.

6.5 First Order All Pass Filter

First order all pass (AP) filters are widely used in analog signal processing for applications in communication systems, such as, equalizers, realization of band pass filters [130] and oscillators [129]. Several current mode first order all pass filters are available in literature. The circuits given in references [130,131,132, 133] suffer from the use of floating capacitors or resistors. In [135], several first order current mode all pass filters were realized using two CCIIs and four grounded passive components. Recently in [136], first order current mode all pass filters were realized using two CCIIs and three grounded passive components. These also have the advantage of high output impedance. However, these circuits suffer from the drawback of using large number of active and passive components.

6.5.1 Circuit description

This Section, gives a novel CM all pass filter realization using single low voltage OFC+ and two grounded admittances. It is also realized from the generalized current topology of Fig. 6.1, by applying a common input \( I_{in} \) to both the X and Y-terminals, as shown in Fig. 6.14. Routine circuit analysis yields the current transfer function:

\[
T_{AP}(s) = \frac{I_{out}}{I_{in}} = \frac{Y_2 - Y_1}{Y_1}
\]  

(6.21)

Through admittance selection, \( Y_1 = G_1 + sC_1 \), \( Y_2 = G_2 \), eqn. (6.21) becomes,

\[
T_{AP}(s) = \frac{I_{out}}{I_{in}} = \frac{G_2 - (G_1 + sC_1)}{G_1 + sC_1}
\]  

(6.22)
By using in eqn. (6.22), the condition, \( G_2 = 2G_1 \), the circuit realizes a first order all pass section with,

\[
T_{AP}(s) = \frac{I_{out}}{I_{in}} = \frac{G_1 - sC_1}{G_1 + sC_1} = \frac{s - \frac{1}{R_1C_1}}{s + \frac{1}{R_1C_1}} \quad (6.23)
\]

where

\[
\omega_0 = \frac{1}{R_1C_1} \quad (6.24)
\]

It may be seen that the design constraint, \( R_1 = 2R_2, \) is easy to satisfy practically in implementation of the filter. The proposed circuit will be shown to have high output impedance, which helps in realization of higher order filters through cascade approach. Also, the use of only one OFC and grounded capacitor and resistors is attractive from the point of view of IC fabrication.

### 6.5.2 Non-ideal analysis

Taking into account the non-idealities of the OFC, as discussed in Sec. 2.3.3, the transfer function becomes,

\[
T_{AP}(s) = \frac{I_{out}}{I_{in}} = \frac{\alpha \beta Y_2 - \alpha Y_1}{Y_1} = \frac{\alpha \beta G_2 - \alpha (G_1 + sC_1)}{G_1 + sC_1} \quad (6.25)
\]
Only minor effects of non-idealities are exhibited in eqn. (6.25), with \( \omega_o \) still given by eqn. (6.24).

### 6.5.3 Sensitivity study

The sensitivity of cutoff frequency (\( \omega_o \)) is evaluated with respect to active and passive elements and is given as

\[
S_{\eta, \epsilon_1}^{\omega_o} = -1 \quad S_{\sigma, \beta}^{\omega_o} = 0
\]

### 6.5.4 Design and simulation

To evaluate the performance of the current mode AP filter, the circuit is simulated using OFC+ model [11]. The circuit was designed at a pole frequency, \( f_o = 2.5 \) MHz, with designed values of \( R_1 = 20 \) KΩ, \( R_2 = 10 \) KΩ and \( C = 3.18 \) pF. The gain and phase plots for the circuit are shown in Fig. 6.15.

![Gain and phase plots for the proposed APF](image)

It is evident from the gain plot that the gain remains almost constant up to 100 MHz, after which it starts deviating due to the non-idealities. A phase
shift of 90° is obtained at the frequency of 2.502 MHz, which closely agrees with the design.

The time domain input and output response at 2.5MHz is shown in Fig. 6.16. As expected the output current is 90° phase shifted with respect to the input current.

![Fig. 6.16 The time domain input and output responses at 2.5MHz](image)

In Table 6.2, the simulated values of input and output impedances are given, which show convenient cascadability of the proposed circuit.

### Table 6.2

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>20 KΩ</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>31.54 MΩ</td>
</tr>
</tbody>
</table>

This Section, presents a new first order current mode all pass filter employing only a single plus type low voltage OFC, alongwith, two grounded resistors and a capacitor, which is attractive for IC implementation. The proposed circuit has high output impedance with R_{out} >> R_{in}, hence, can be cascaded to implement higher order realizations. The proposed circuit also has the advantages of low component count, low sensitivity, reliable high
frequency performance and low voltage operation at 0.75 volt, as in other circuits realized from the BBB of Fig. 6.1.

6.6 Cascadable Second order Band Pass Filter using single OFC

Active filters using single amplifier have found interest in video signal processing and wireless communication systems [39, 40]. A number of CCII and OFC-based filters using a large count of active and passive components, along with, some floating components have been reported [38, 42, 45, 106, 107]. These are not attractive for implementation in integrated form. We suggest an attractive realization.

6.6.1 Circuit description

This Section gives a CM band pass filter, shown in Fig. 6.17, using a plus type low voltage OFC, along with, four passive components, out of which three components are grounded while one is floating. Once again, the circuit is realized from the generalized scheme of Fig. 6.1.

![Fig. 6.17 OFC-based band pass filter](image)

Routine circuit analysis yields the current transfer function as:

\[
T_{BP}(s) = \frac{I_{BP}}{I_{in}} = \frac{s}{R C_1} \frac{1}{D(s)}
\]

(6.27)
where the denominator \( D(s) \) is given by:

\[
D(s) = s^2 + s \left( \frac{1}{R_1C_1} + \frac{1}{R_2C_2} \right) + \frac{1}{R_1R_2C_1C_2}
\]  

(6.28)

The circuit realizes a second-order BP filter.

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

\[
\omega_p = \sqrt{\frac{1}{R_1R_2C_1C_2}}, \quad BW = \frac{1}{R_1C_1} + \frac{1}{R_2C_2}, \quad Q = \frac{\sqrt{1/R_1R_2C_1C_2}}{1/R_1C_1} + \frac{1}{R_2C_2}
\]  

(6.29)

The gain of the filter is:

\[
H_{BP} = \frac{R_2C_1}{(R_1C_1 + R_2C_2)}
\]  

(6.30)

It is evident from eqn. (6.29), the summation term in the denominator of pole-Q limits the filter to low-Q values.

### 6.6.2 Non-ideal analysis

Taking the non-idealities of OFC into account, the analysis of the circuit of Fig. 6.17 yields the following transfer function:

\[
T_{BP}(s) = \frac{I_{BP}}{I_{in}} = \frac{s \alpha \beta}{s^2 + s \left( \frac{1}{R_1C_1} + \frac{1}{R_2C_2} \right) + \frac{1}{R_1R_2C_1C_2}}
\]  

(6.31)

Still a standard BP response is realized without affecting pole- \( \omega_p \) and pole- \( Q \) values in eqn. (6.29). The gain is now given by:

\[
H'_{BP} = \frac{\alpha \beta R_2C_1}{(R_1C_1 + R_2C_2)}
\]  

(6.32)
This depicts only a minor gain reduction in the low and medium frequency ranges.

6.6.3 Sensitivity study

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

\[ S_{R_1 R_2 C_1 C_2} = -1/2, \quad S_{\alpha \beta R_1 R_2 C_1} = 1, \]

\[ S_{R_1 C_2} = -S_{R_1 C_2} = \frac{(R_1 C_2 - C_1 R_2)}{2(R_1 C_2 + C_1 R_2)} \leq \frac{1}{2} \] (6.33)

It is evident from eqn. (6.33) that all the sensitivities are found to be reasonably low.

6.6.4 Design and simulation

To demonstrate the performance of the BP filter, it is simulated using PSpice. The circuit was designed for \( Q = 0.5, f_o = 2 \) MHz at unity gain. The designed values are, \( C_2 = C = 3.97 \) pF, \( C_1 = 2C = 7.95 \) pF, at \( R_1 = R = 10 \) KΩ and \( R_2 = 2R = 20 \) KΩ. The simulated BP response, alongwith, parameters are shown in Fig. 6.18. The simulated values are given in the Table and show close agreements with the theory.

The tunability of the BP filter was investigated by varying the center frequency \( (f_o) \) through passive resistor \( R_1 \). The BP responses corresponding to \( f_o = 500 \) KHz, 1 MHz and 3 MHz at \( R_1 \) of 40.08 KΩ, 20 KΩ and 6.7 KΩ, respectively, are shown in Fig. 6.19. These depict convenient tunability of \( f_o \).
Simulated results

<table>
<thead>
<tr>
<th>$f_o$(BP)</th>
<th>Q (BP)</th>
<th>H_{HP}</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.08 MHz</td>
<td>0.45</td>
<td>1.01</td>
</tr>
</tbody>
</table>

Fig. 6.18 Frequency response of CM OFC based band pass filter

Simulated results

<table>
<thead>
<tr>
<th>$f_o$</th>
<th>Q</th>
<th>H_{HP}</th>
</tr>
</thead>
<tbody>
<tr>
<td>500.01 KHz</td>
<td>0.48</td>
<td>1.00</td>
</tr>
<tr>
<td>1.02 MHz</td>
<td>0.43</td>
<td>1.01</td>
</tr>
<tr>
<td>3.04 MHz</td>
<td>0.42</td>
<td>1.03</td>
</tr>
</tbody>
</table>

Fig. 6.19 Frequency tuning of BPF

(a) $f_o = 500$ KHz  (b) $f_o = 1$ MHz  (c) $f_o = 3$ MHz
The simulated values of input and output impedances are given in Table 6.3. These once again, show the convenient cascadability for the proposed circuit.

### Table 6.3

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>10 KΩ</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>31.76 MΩ</td>
</tr>
</tbody>
</table>

A new CM band pass filter is realized and studied. The proposed filter employs only single plus type OFC, along with, four passive components, out of which three components are grounded. Moreover, the presented circuit is useful for cascadable realization of higher order filters. It also has low sensitivity and use of low supply voltage equal to ± 0.75V. It however, suffers from the realization of only low-Q values.

### 6.7 Realization of Cascadable Multifunctional Biquadratic Filter

In this Section, a single input multi output (SIMO), high output impedance CM multifunctional biquadratic filter (MBF) is realized by employing non-interactive cascade of two blocks of the generalized scheme of Fig. 6.1. This is conveniently possible as the building block satisfy $Z_{\text{out}} \gg Z_{\text{in}}$. The proposed MBF, shown in Fig. 6.20, uses two plus type OFCs, along with, four grounded admittances. Analysis of the circuit yields the following current transfer function:

$$T(s) = \frac{I_{\text{out}}}{I_{\text{in}}} = \frac{Y_1Y_3}{Y_2Y_4}$$  \hspace{1cm} (6.34)

Realization of various standard filter responses is considered below through appropriate choice of grounded admittances.
Fig. 6.20 Cascadable multifunctional biquadratic filter

Low pass filter: if we select, \( Y_1 = \frac{1}{R_1}, \ Y_2 = G_2 + sC_2, \ Y_3 = \frac{1}{R_3}, \) and
\( Y_4 = G_4 + sC_4, \) then this results in the low pass filter shown in Fig. 6.21 (a), having the current transfer function:

\[
T_{LP}(s) = \frac{I_{LP}}{I_m} = \frac{1}{R_2R_3C_3C_4} D(s)
\]  

where \( D(s) \) is given by

\[
D(s) = s^2 + s \left( \frac{1}{R_2C_2} + \frac{1}{R_4C_4} \right) + \frac{1}{R_2R_4C_3C_4}
\]  

Fig. 6.21 (a) Low pass biquadratic filter
**High pass filter:** With $Y_1 = sC_1$, $Y_2 = G_2 + sC_2$, $Y_3 = sC_3$ and $Y_4 = G_4 + sC_4$, the realization of high pass filter of Fig. 6.21 (b) is obtained with the current transfer function:

$$T_{HP}(s) = \frac{I}{I_{in}} = \frac{s^2 C_2 C_3}{C_4 C_5 D(s)}$$

(6.37)

![Fig. 6.21 (b) High pass biquadratic filter](image)

**Band pass filter:** On selecting, $Y_1 = sC_1$, $Y_2 = G_2 + sC_2$, $Y_3 = G_3$ and $Y_4 = G_4 + sC_4$, the band pass filter of Fig. 6.21(c) is realized, with current transfer function:

$$T_{BP}(s) = \frac{I_{BP}}{I_{in}} = \frac{s C_1}{R C_2 C_4 D(s)}$$

(6.38)

![Fig. 6.21 (c) Band pass biquadratic filter](image)
From eqns. (6.35), (6.37) and (6.38), it is seen that low pass, high pass and band pass responses are realized through appropriate selection of grounded admittances and without the requirement of matching conditions. Moreover, still higher order filters can be realized by cascading the proposed second order sections.

The pole frequency ($\omega_n$), band width (BW) and the quality factor (Q) of the filters are obtained from the characteristic polynomial $D(s)$:

$$\omega_n = \sqrt{\frac{1}{R_3 R_4 C_3 C_4}}, \quad BW = \frac{1}{R_3 C_2} + \frac{1}{R_4 C_4}, \quad Q = \frac{\sqrt{1/R_2 R_3 C_2 C_4}}{\left(\frac{1}{R_3 C_2} + \frac{1}{R_4 C_4}\right)}$$

(6.39)

Gains of the low pass, band pass and high pass filters are respectively,

$$H_{LP} = \frac{R_3 R_4}{R_1 R_2}, \quad H_{HP} = \frac{C_3}{C_2 C_4}, \quad H_{BP} = \frac{R_3 R_4 C_4}{R_3 (R_3 C_2 + R_4 C_4)}$$

(6.40)

From eqns. (6.39) it can be seen that as the denominator is greater than the numerator, the proposed circuit also realizes low Q values. From the eqn. (6.40), the gains of LP, HP and BP filters, respectively, can be tuned independently, through the passive components ($R_1$ or $R_3$), ($C_1$ or $C_3$) and ($R_3$ or $C_4$).

### 6.7.1 Non-ideal analysis

Taking the non-idealities of OFC into account, analysis of the filters of Fig. 6.21 still realizes standard low pass, high pass and band pass responses given by:

$$T_{LP}(s) = \frac{I_{LP}}{I_m} = \frac{I R_2 R_3 C_2 C_4}{D(s)}$$

(6.41)

$$T_{HP}(s) = \frac{I_{HP}}{I_m} = \frac{s^2 C_2 C_4 \alpha_2 \beta_1 \beta_2}{C_2 C_4}$$

(6.42)
It is seen that $D(s)$ remains unaltered. Hence the non-ideal filter parameters are still given by eqn. (6.39) and remain invariant. Slight deviations are noticed in the non-ideal gains of the filters. These are slightly lowered and given by:

$$
H'_{LP} = \frac{\alpha_1 \alpha_2 \beta_2 R_2 R_4}{R_1 R_3}, \quad H'_{HP} = \frac{\alpha_1 \alpha_2 \beta_1 C_1 C_3}{C_2 C_4}, \quad H'_{BP} = \frac{\alpha_1 \alpha_2 \beta_1 R_1 R_4 C_1}{R_5 (R_2 C_2 + R_4 C_4)} \tag{6.44}
$$

### 6.7.2 Sensitivity study

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

$$
S_{\alpha_1, \alpha_2, \beta_1, \beta_4}^{\omega_{0}} = -1, \quad S_{R_1, C_1}^{Q} = -1, \quad S_{R_1, C_3}^{Q} = -1, \quad S_{R_2, R_3}^{H_{LP}} = -1, \quad S_{R_4, R_5}^{H_{HP}} = -1, \quad S_{R_1, C_4}^{H_{BP}} = -1, \quad S_{R_1, R_2, R_3, R_4}^{H_{LP}} = -1
$$

It is evident from eqn. (6.45) that all the sensitivities are found to be less than or equal to unity. Also, note that the filter parameters, $(\omega_0)$ and pole-$Q$, are insensitive to the non-idealities.

### 6.7.3 Design and simulation

To demonstrate the performance of OFC-based MBF, the LP, HP and BP filters are simulated using PSpice. The circuits were designed for $f_v = 200$ KHz, $Q = 0.5$ at unity gain. The designed values are, $C_1 = C_2 = C_3 = C_4 = 45$ pF, $R_1 = R_2 = R_3 = R_4 = 17.68$ KΩ. Unity gain of the band pass filter is obtained by selecting $R_2 = R_4 = 17.68$ KΩ, and $R_3 = 8.84$ KΩ. The simulated responses are shown in Fig. 6.22 and the simulated results are given in the Table. These present conformity with theory.
The simulated values of input and output impedances are given in Table 6.4. These demonstrate convenient cascaddability of the proposed MBF.

**Table 6.4**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>23.2 KΩ</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>31.59 MΩ</td>
</tr>
</tbody>
</table>

A new single input multi output (SIMO) high output impedance MBF is realized from the BBB of Fig. 6.1. Realization of low pass, high pass and band pass responses are obtained through appropriate selection of admittances and without the requirement of matching conditions. The proposed circuits use only grounded passive components, and also have low sensitivity and operation at low supply voltage. However, it also suffers from the realization of low Q values.
6.8 Insensitive Low Component CM Universal Biquadratic Filter

Current mode universal biquadratic filters (UBF) with high output impedance are of great importance for realizing higher order CM filters for IC implementation. Several circuits for their realization have been reported in the literature [64-67]. In [64], the current mode UBF employing three MO-CCIIs, two grounded resistors and two grounded capacitors is given. The UBF in [65] uses four current controlled conveyors and two floating capacitors. Recently in [66], a 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.

This Section, presents a novel versatile current mode biquadratic filter using plus type Multi Output Operational Floating Conveyors (MO-OFC) and four grounded passive components. The circuit is simple in structure and can realize all standard second order responses, viz., low pass, high pass, band pass, band elimination and all pass, by selecting appropriate input and output terminals, without requiring matching constraints. It has high impedance outputs, which enable easy cascading in current mode operation. The proposed circuit also has the additional advantages of low component count, low sensitivity, and low voltage operation over previously reported filters [64-67]. To the best of authors' knowledge, simple universal biquadratic filter employing only two plus type multi output operational floating conveyors and four grounded components has not yet been reported in the literature.

6.8.1 Circuit description

The MO-OFC+ based UBF is shown in Fig. 6.23. It consists of only two conveyors, alongwith, two grounded capacitors and two grounded resistors. Analysis of the circuit yields the output currents $I_{o1}$, $I_{o2}$ and $I_{o3}$ as:
where the denominator $D(s)$ is given by:

$$D(s) = s^2 + \frac{s}{R_4C_3} + \frac{1}{R_1R_4C_3C_3}$$  \hspace{1cm} (6.49)
\[ \omega_n = \sqrt{\frac{1}{R_2 R_4 C_1 C_3}}, \quad Q = \sqrt{\frac{R_2 C_3}{R_2 C_1}} \]  

(6.50)

It may be seen that all standard filter responses can be realized by selecting suitable input and output terminals without matching constraints as given in Table 6.5. The gains of all the responses are found to be unity.

Table 6.5 Realization of Standard Second Order Responses

<table>
<thead>
<tr>
<th>Type of Filter</th>
<th>Inputs conditions</th>
<th>Transfer function</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low pass filter</td>
<td>( I_{o1} = I_m ) and ( I_{o2} = 0 ), in eqn. (6.47)</td>
<td>( T_{LP}(s) = \frac{I_{o1}}{I_m} = -\frac{1}{D(s)} )</td>
</tr>
<tr>
<td>Band pass filter</td>
<td>( I_{o1} = I_m ) and ( I_{o2} = 0 ), in eqn. (6.48)</td>
<td>( T_{BP}(s) = \frac{I_{o1}}{I_m} = \frac{s}{C_3 R_4 D(s)} )</td>
</tr>
<tr>
<td>High pass filter</td>
<td>( I_{o1} = I_{o2} = I_{o3} = I_m ), in eqn. (6.49)</td>
<td>( T_{HP}(s) = \frac{I_{o1}}{I_m} = \frac{s^2}{D(s)} )</td>
</tr>
<tr>
<td>Band elimination filter</td>
<td>( I_{o1} = I_{o2} = I_{o3} = I_m ) and ( I_{o4} = 0 ), in eqn. (6.49)</td>
<td>( T_{BE}(s) = \frac{I_{o1}}{I_m} = \frac{s^2 + \frac{1}{R_2 R_4 C_1 C_3}}{D(s)} )</td>
</tr>
<tr>
<td>All pass filter</td>
<td>( I_{o1} = I_{o2} = I_{o3} = I_{o4} ), in eqn. (6.49) and ( I_{o1} = I_{o2} )</td>
<td>( T_{AP}(s) = \frac{I_{o1}}{I_m} = \frac{s^2 - s \frac{1}{R_2 R_4 C_1 C_3} + \frac{1}{R_2 R_4 C_1 C_3}}{D(s)} )</td>
</tr>
</tbody>
</table>

6.8.2 Non-ideal analysis

Taking the non-idealities of MO-OFC into account, the port relationships are characterized by:

\[ i_y = 0, \quad v_x = \beta v_y, \quad v_w = z_i x, \quad i_z = \alpha i_w \]  

(6.51)

The denominator of eqn. (6.49) of the transfer functions is modified to:

\[ D'(s) = s^2 + \frac{s}{R_2 C_3} + \frac{\alpha c \beta}{R_2 R_4 C_1 C_3} \]  

(6.52)
The modified filter parameters are given by:

\[
\omega'_o = \sqrt[2]{\frac{\alpha_1 \alpha_2 \beta_1}{R_2 R_4 C_1 C_3}}, \quad Q' = \sqrt[2]{\frac{\alpha_1 \alpha_2 \beta_1 R_4 C_3}{R_2 C_1}}
\]  \hspace{1cm} (6.53)

From Table 6.5, it is evident that gains of all filters are unity. Hence, these are not affected by the non-idealities. However, in low to medium frequency range (\(\approx 10\) MHz), only slight lowering of parameters, \(\omega'_o\) and \(Q'\) is observed.

6.8.3 Sensitivity study

The sensitivity of filter parameters, pole frequency (\(\omega_o\)) and the quality factor (\(Q\)) are evaluated with respect to active and passive elements and are found to be low, being half in magnitude. This is an attractive feature of the UBF.

\[
S^{\omega_o}_{R_2, R_4, C_1, C_3} = -\frac{1}{2}, \quad S^{Q}_{R_2, C_1, \alpha_1, \alpha_2, \beta_1} = -\frac{1}{2}, \quad S^{\omega_o}_{\alpha_1, \alpha_2, \beta_1} = \frac{1}{2}, \quad S^{Q}_{R_2, C_1} = -\frac{1}{2}
\]  \hspace{1cm} (6.54)

6.8.4 Design and simulation

To demonstrate the performance of CM universal biquadratic filter, the circuit is simulated using PSpice. Initially, the UBF was designed for a \(f_o = 2.5\) MHz, \(Q = 0.707\) at unity gain. For \(R_2 = R_4 = 1.59\) K\(\Omega\), eqn. (6.50) yields \(C_1 = 56.6\) pF and \(C_3 = 28.3\) pF. The simulated responses of the UBF are shown in Fig. 6.24 and the parameter values are given in the Table. These present close agreement with the theory.

Frequency tuning of the UBF responses is next investigated by varying \(f_o\) at a constant \(Q = 0.707\), through resistor \(R_2\).
Fig. 6.24 The simulated UBF responses at $f_o = 2.5$ MHz

Fig. 6.25 (a) Frequency tuning of (i) BP, (ii) LP responses
Fig. 6.25 (b) Frequency tuning of (iii) HP, (iv) BE responses

The LP, HP, BP, BE responses for $f_o = 300$ KHz, $f_o = 700$ KHz, and $f_o = 1.5$ MHz at $R_2 = 50$ KΩ, $R_2 = 21.43$ KΩ and $R_2 = 10$ KΩ, respectively, are shown in Fig. 6.25. These exhibit convenient tunability of the filter.

The simulated values of input and output impedances, given in Table 6.6, demonstrate the convenient cascadability of the proposed UBF. This makes the realization suitable for obtaining higher order CM filters through cascade approach.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Resistance</td>
<td>63 Ω</td>
</tr>
<tr>
<td>Output Resistance</td>
<td>41.26 MΩ</td>
</tr>
</tbody>
</table>
The proposed circuit realizes novel insensitive current mode universal biquadratic filter using low voltage MO-OFC+, alongwith, only four grounded passive components. The realized UBF has low input and high output impedances and is cascadable in nature. It also has the advantages of low sensitivity, low component count, and use of low supply voltage (± 0.75V).

6.9 Design of OFC-Based CM Higher Order Filters

There are many filtering applications that utilize biquadratic 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 OFCs in the BBB has the obvious advantage of low cost and low power consumption. This Section, presents the application of the minimum component CM universal biquadratic filter of Sec.6.8, alongwith, the CM first order filters of Sec. 6.4 in the realization of higher order filters.

6.9.1 Fifth Order CM LP Butterworth filter

To show the realization of higher odd order filter functions, we consider as an example, the realization of 5th order Butterworth low pass filter by cascading two sections of second order current mode low pass filter (UBFs of Sec. 6.8), alongwith, one first order low pass filter of Sec. 6.4. The normalized Butterworth transfer function for the 5th order LP filter is given by [70]:

\[ T(s) = \frac{1}{(s+1)(s^2 + 1.618s + 1)(s^2 + 0.618s + 1)} \]  

(6.55)

where the normalized pole frequency is \( \omega_c = 1 \). The transfer function can be de-normalized by replacing \( S \rightarrow s/\omega_c \) to give the required fifth order filter function with given pole-\( \omega_c \) and pole-Q. The pole-Q of an individual biquadratic filter section is simply the reciprocal of the coefficients of s in eqn. (6.55) [70]. Then the values of Q are found to be 0.5, 0.618, and 1.618, respectively, for the one first order section and two UBFs of the cascade. The
5th order current mode low pass Butterworth filter can be designed using these values of Q for the given pole frequency. The pole frequency $\omega_o$ and the gain of the first order low pass filter of Section 6.4 are:

$$\omega_o = \frac{1}{R_2C_2}, \quad H_{LP} = \frac{R_1}{R_1}$$  \hspace{1cm} (6.56)

Also, the pole frequency and the quality factor of the second order current mode MO-OFC, based UBF, are given by:

$$\omega_o = \frac{1}{\sqrt{R_2R_4C_1C_3}}, \quad Q = \sqrt{\frac{R_4C_3}{R_2C_1}}$$  \hspace{1cm} (6.57)

### 6.9.2 Design and simulation

To evaluate the performance of the designed fifth order Butterworth low pass filter, it is simulated using novel MO-OFC model of Sec. 2.2.6. The circuit is designed for a pole frequency, $f_o = 2.5$ MHz. The values of capacitors, $C_1 = C_3 = C = 15.9$ pF are selected equal. The resistors for first order low pass section are designed to satisfy the eqn (6.56), and each individual biquadratic filter sections are designed to satisfy the eqn. (6.57). The designed values for the sections are obtained as:

- **Section-I:** $R_2 = R_4 = 4.03$ K$\Omega$, corresponding to $Q = 0.5$ (First order),
- **Section-II:** $R_2 = 2.474$ K$\Omega$, $R_4 = 6.478$ K$\Omega$, corresponding to $Q = 0.618$,
- **Section-III:** $R_2 = 6.478$ K$\Omega$, $R_4 = 2.474$ K$\Omega$, corresponding to $Q = 1.618$.

The theoretical and simulated fifth order CM low pass filter responses are shown in Fig.6.26 (a). The simulated results exhibit very close agreement with the theory in the entire frequency range of interest.
The resulting theoretical and simulated frequency response curves for the filter are also plotted in DB and given in Fig. 3.26 (b). It shows the stop band attenuation of 100DB/decade, verifying the 5th order low pass filter characteristics.
6.9.3 Eighth Order CM LP and BP Butterworth filters

Next, we show the realization of an 8th order Butterworth low pass filter by cascading four UBF sections of Sec. 6.8. The normalized Butterworth transfer function for the 8th order CM low pass filter is given by [70]:

\[ T(s) = \frac{1}{(s^2 + 1.962s + 1)(s^2 + 1.663s + 1)(s^2 + 1.111s + 1)(s^2 + 0.390s + 1)} \]  (6.58)

where the normalized pole frequency is \( \omega_n = 1 \). The transfer function can be de-normalized by replacing \( S \rightarrow s/\omega_n \) to give the required eighth order filter function with given pole-\( \omega_n \) and pole-Q. As pointed out earlier, the pole-Q of an individual biquadratic filter section is simply the reciprocal of the coefficients of s in eqn. (6.58). Hence, the values of Q are 0.51, 0.60, 0.90 and 2.56, respectively, for the four UBFs. A UBF can be designed using these values of Q for a given pole frequency. The filter parameters, i.e., pole frequency and the quality factor, of the current mode MO-OFC based UBF are given by:

\[ \omega_n = \frac{1}{\sqrt{R_2R_4C_1C_3}}, \quad Q = \frac{R_4C_3}{\sqrt{R_2C_1}} \]  (6.59)

6.9.4 Design and simulation

To evaluate the performance of the eighth order Butterworth low pass filter, it is simulated using novel MO-OFC model of Sec. 2.2.6. The circuit is designed for a pole frequency, \( f_p = 1 \) MHz. For convenience in IC fabrication, all capacitors are selected equal, i.e., \( C = 11 \) pF each. The resistors for each section are designed to satisfy the eqn (6.59). The designed values for each section are given below:

Section-I: \( R_2 = 7.37 \) K\( \Omega \), \( R_4 = 28.36 \) K\( \Omega \), corresponding to Q = 0.51,

Section-II: \( R_2 = 8.68 \) K\( \Omega \), \( R_4 = 24.11 \) K\( \Omega \), corresponding to Q = 0.60,

Section-III: \( R_2 = 13.02 \) K\( \Omega \), \( R_4 = 16.07 \) K\( \Omega \), corresponding to Q = 0.90,
Section-IV: \( R_2 = 36.73 \, \text{K}\Omega, \, R_4 = 5.65 \, \text{K}\Omega \), corresponding to \( Q = 2.56 \).

The theoretical and simulated eighth order LP responses are shown in Fig.6.27 (a) and also shown in DB in Fig.6.27 (b).

![Simulated results](image)

<table>
<thead>
<tr>
<th>( f_c ) (MHz)</th>
<th>( H_{\text{LP}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.02</td>
<td>1.01</td>
</tr>
</tbody>
</table>

Fig.6.27 (a) Frequency response of eighth order CM Butterworth LPF

![Simulated results](image)

Fig.6.27 (b) Frequency response of eighth order CM Butterworth LPF in DB
These give the simulated pole frequency of 1.02 MHz. The low pass response, shown in DB, also gives the stop band attenuation of 160DB/decade, thus verifying the 8th order low pass filter characteristics. The simulated and theoretical 8th order low pass filter characteristics are seen to be overlapping over the entire range of interest.

The MO-OFC based current mode eighth order low pass filter is realized by directly cascading the UBFs without using additional current followers. The proposed UBFs can also be used to realize higher order BPF, HPF, BEF through an electronic switching control [140]. As an example, an 8th order band pass filter is designed for $f_c = 1$ MHz and $H_{BP} = 1$ and the theoretical and simulated responses are shown in Fig. 6.28 (a) and also in DB in Fig. 6.28 (b). The simulated pole frequency comes out to be $f_c = 1.04$ MHz. It is evident from Fig. 6.28 (b) that the slopes in lower frequency and higher frequency bands are 80 DB/decade (each), thus verifying the 8th order band pass filter characteristics. At the centre frequency, the gain is equal to unity, as designed.

**Table 6.28:**

<table>
<thead>
<tr>
<th>Simulated results</th>
<th>$f_c$ (BP)</th>
<th>$Q$ (BP)</th>
<th>$H_{BP}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.04 MHz</td>
<td>2.53</td>
<td>1.02</td>
<td></td>
</tr>
</tbody>
</table>

**Fig.6.28 (a) Frequency response of eighth order CM BPF**
6.10 Conclusion

In this Chapter, the current mode OFC-based circuits are realized and critically studied. An OFC-based CM generalized building block is given. It is shown to realize current mode inverting and non-inverting amplifiers, ideal integrator and differentiator, first order LP and HP filter sections. The circuit can also give first order AP filter through minor modifications of connections at X and Y-terminals of OFC. The generalized structure is also shown to realize a second order BP filter. However, the circuit is only suitable for low Q realizations. The generalized topology can be made to ensure $R_{out}$ much greater than $R_{in}$, thus realizing a directly cascadable CM sections. On cascading two such sections an MBF is realized, which provides, LP, HP and BP responses.

An attractive realization of insensitive low component CM Universal Biquadratic Filter (UBF) is given, using only two MO-OFCs, alongwith, four grounded passive components. It is a canonic realization and provides all five standard biquadratic filter responses. The realized circuits are shown to be useful in realizing fifth and eighth order CM Butterworth LP filters through cascade form synthesis. Also, a corresponding eighth order BP filter is realized.
All the circuits considered in the Chapter are studied in detail. They are found to have low sensitivities and reliable high frequency performance. The realizations are free from matching constraints. The uses of grounded components are useful in integration. The non-interactive properties of the building blocks help in the realization of higher order filters. The realizations are verified through simulation with PSpice. Also, the simulation results show convincing performances at low operation supply voltage of ±0.75 volt.