CHAPTER – 6
MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) FILTERS

The results of the following papers have been reported in this Chapter:


The multiple-input multiple-output (MIMO) filter structures are a class in which more than one input is inserted to get simultaneously different output responses (LP, HP, BP, notch, AP etc.). These configurations are sometimes more versatile, simpler and may lead to reduction in the number of active and passive elements than SISO, SIMO and MISO filters, although the physical implementation of the circuit will require additional hardware for input selection. This concept is relatively new and the literature is limited. The author in this chapter has proposed new multiple-input multiple-output (MIMO) universal filter structures for operation in current, voltage and mixed mode.

6.1 CURRENT MODE MIMO FILTER

During the last decade and recent past a number of universal filter structures have been reported in the literature. Most of them are classified as single-input single-output (SISO), single-input multiple-output (SIMO) filters and few are multiple-input single-output (MISO). A little work has been done in the domain of multiple-input multiple-output (MIMO) current mode filters [100], [101]; hence it needs attention. Wang and Lee [100] have presented two structures – three-input single-output and single-input three-output. Both these structures use three current conveyors, namely two DOCCI and one DOCCIII and another one uses two DOCCI and one multiple output CCII, and four grounded passive components. The structure [101] realizes two-input three-output or three-input two-output filter by the use of two multiple output CCII and four grounded/ virtually grounded passive components. Although the structure [101] satisfies all the standard properties of universal filter, it lacks the orthogonality of \( \omega_0 \) and \( Q_0 \) and both the structures of [100], [101] lack electronic adjustability of filter parameters. However, the
structure [101] can be extended towards electronic adjustability by replacing current conveyor (CCII) and series resistance at port x by current controlled conveyor (CCCII) [13].

In this section, a new current mode MIMO universal filter using MOCCCIIs has been developed. The filter employs only two MOCCCIIs and three grounded passive components, which is beneficial from integrated circuit implementation point of view. The structure realizes low pass, band pass, high pass, notch and all pass responses all at high output impedance that makes it an ideal candidate for cascading in current mode (CM) operation. The filter has low passive and active sensitivities and also enjoys the attractive feature of orthogonal control of $\omega_0$ and $Q_0$ via grounded resistor or grounded resistor and bias currents of MOCCCIIs.

### 6.1.1 CIRCUIT DESCRIPTION

The proposed network (Fig. 6.1) is based on employing CCCII, which is characterized by following port relationships

$$v_X = v_Y \pm i_X | R_{xi} (i_{0i}) |, \quad i_{2z} = \pm i_X \quad \text{and} \quad i_Y = 0$$

where $R_{xi} = V_T / 2I_{0i}$, $V_T$ is the thermal voltage, $I_{0i}$ is bias current of CCCII and $i = 1, 2$. To get MOCCCIIs, current mirrors are inserted at the output of CCCII.

The analysis of the proposed circuit yields the following output functions:

$$I_{out1} = \frac{(s^2 R_{x1} R_{x2} C_1 C_2 + s G_1 R_{x1} R_{x2} C_1 + 1) I_{in3} - s R_{x1} C_1 I_{in2} + I_{in1}}{D(s)} \quad (6.1.1)$$

$$I_{out2} = \frac{-s R_{x1} C_1 I_{in2} + I_{in1}}{D(s)} \quad (6.1.2)$$
\[ I_{\text{out}3} = \frac{I_{\text{in}2} + R_{x2} (G_1 + sC_2) I_{\text{in}1}}{D(s)} \]  

(6.1.3)

where

\[ D(s) = s^2 R_{x1} R_{x2} C_1 C_2 + s G_1 R_{x1} R_{x2} C_1 + 1 \]  

(6.1.4)

Fig. 6.1 Proposed current mode multiple-input multiple-output biquad.

CASE I (Two-input and three-output):

Specializations in (6.1.1) – (6.1.3) results in the filter functions listed in Table 6.1:

**Table 6.1**

Characteristics of the filter response for two-input three-output case.

<table>
<thead>
<tr>
<th>Filter type</th>
<th>Input</th>
<th>Output</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low pass</td>
<td>( I_{\text{in}1} = 0, I_{\text{in}2} = I_{\text{in}3} = I_{\text{in}} )</td>
<td>( I_{\text{out}} = I_{\text{out}3} (I) )</td>
</tr>
<tr>
<td>Band pass</td>
<td>( I_{\text{in}1} = 0, I_{\text{in}2} = I_{\text{in}3} = I_{\text{in}} )</td>
<td>( I_{\text{out}} = I_{\text{out}2} (I) )</td>
</tr>
<tr>
<td>High pass</td>
<td>( I_{\text{in}1} = 0, I_{\text{in}2} = I_{\text{in}3} = I_{\text{in}}, R_{x2} = R_1 )</td>
<td>( I_{\text{out}} = I_{\text{out}1} + I_{\text{out}3} (N, I) )</td>
</tr>
<tr>
<td>Notch</td>
<td>( I_{\text{in}1} = 0, I_{\text{in}2} = I_{\text{in}3} = I_{\text{in}}, R_{x2} = R_1 )</td>
<td>( I_{\text{out}} = I_{\text{out}1} (N, I) )</td>
</tr>
<tr>
<td>All pass</td>
<td>( I_{\text{in}1} = 0, I_{\text{in}2} = I_{\text{in}3} = I_{\text{in}}, R_{x2} = R_1 )</td>
<td>( I_{\text{out}} = I_{\text{out}1} + I_{\text{out}2} (N, I) )</td>
</tr>
</tbody>
</table>

(Note: N. I. – noninverting, I. – inverting)

CASE II (Three-input and three-output):

In a similar manner, (6.1.1) - (6.1.3) results in the filter functions listed in Table 6.2:
Table 6.2

Characteristics of the filter response for three-input three-output case.

<table>
<thead>
<tr>
<th>Filter type</th>
<th>Input</th>
<th>Output</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low pass</td>
<td>$I_{in2} = I_{in3} = 0$, $I_{in1} = I_{in}$</td>
<td>$I_{out} = I_{out1} = I_{out2}$ (N. I.)</td>
</tr>
<tr>
<td>Band pass</td>
<td>$I_{in1} = I_{in3} = 0$, $I_{in2} = I_{in}$</td>
<td>$I_{out} = I_{out1}$ (I.)</td>
</tr>
<tr>
<td>High pass</td>
<td>$I_{in1} = 0$, $I_{in2} = I_{in3} = I_{in}$, $R_{x2} = R_1$</td>
<td>$I_{out} = I_{out1} + I_{out3}$ (N. I.)</td>
</tr>
<tr>
<td>Notch</td>
<td>$I_{in1} = 0$, $I_{in2} = I_{in3} = I_{in}$, $R_{x2} = R_1$</td>
<td>$I_{out} = I_{out1}$ (N. I.)</td>
</tr>
<tr>
<td>All pass</td>
<td>$I_{in1} = 0$, $I_{in2} = I_{in3} = I_{in}$, $R_{x2} = R_1$</td>
<td>$I_{out} = I_{out1} + I_{out2}$ (N. I.)</td>
</tr>
</tbody>
</table>

Thus the proposed structure can be viewed as two-input three-output (case I) or three-input three-output (case II) current mode universal biquad. It may be noted that very simple component matching constraint is required. Furthermore, all the current outputs are available at high impedance output ports of current controlled conveyors that enable easy cascaddability without the need of supplementary buffer circuit.

All the filters are characterized by the following filter parameters:

$$\omega_0 = \left( \frac{1}{R_{x1} R_{x2} C_1 C_2} \right)^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{R C_2}, \quad \text{and} \quad Q_0 = R \left( \frac{C_2}{R_{x1} R_{x2} C} \right)^{1/2} \quad (6.1.5)$$

It may be noted from (6.1.5) that $Q_0$ can be independently controlled keeping $\omega_0$ unchanged through the grounded resistor $R_1$ for low pass and band pass functions and through $R_1$ and then $R_{x2}$ ($I_{o2}$) and $R_{x1}$ ($I_{o1}$) in that order for high pass, notch and all pass functions. It may be noted that the resistances $R_{x1}$ and $R_{x2}$ can be easily varied to the required values by externally controlling the bias current $I_{o1}$ and $I_{o2}$ respectively of MOCCCIIs. The parameter $\omega_0$ can also be adjusted electronically by varying bias currents of MOCCCIIs without disturbing $\omega_0/Q_0$. 
6.1.2 EFFECT OF NONIDEALITIES

The frequency performance of the filter circuit may deviate from the ideal one due to nonidealities. The nonidealities effects may be categorized in two groups. The first comes from frequency dependence of internal current and voltage transfers of CCCII ($\omega(s)$, $\beta(s)$), which can be approximated by first order low pass transfer function. Their poles constitute parasitic pole of MOCCCI and thus limits the usable frequency band at higher frequencies.

The second comes from parasitics of CCC comprising of resistance and capacitances connected in terminals y and z (i.e. $R_y$, $C_y$, $R_z$ and $C_z$). The effects of these parasitics on filter response depend strongly on circuit topology. In the proposed topology parasitic capacitances are in parallel with external capacitors, so by predistorting the values of these external passive components the loading effects of the parasitics of MOCCCI may be compensated.

Considering the nonidealities outlined in section 2.3, the output functions (6.1.1) - (6.1.3) are modified to

\[
I_{out1}\bigg|_n = \frac{(s^2 R_{x1} R_{x2} C_1 C_2 + s G R_{x1} R_{x2} C_1 + \alpha_1 \alpha_2 \beta_1 \beta_2) I_{in3} - s \beta_2 R_{x1} C_1 I_{in2} + \alpha_1 \beta_1 \beta_2 I_{in1})}{D_n(s)}
\]

\[
I_{out2}\bigg|_n = \frac{-s \beta_2 R_{x1} C_1 I_{in2} + \alpha_1 \beta_1 \beta_2 I_{in1}}{D_n(s)}
\]

\[
I_{out1}\bigg|_n = \frac{-\alpha_2 \beta_1 \beta_2 I_{in2} + \beta_1 R_{x2} (G_1 + s C_2) I_{in1}}{D_n(s)}
\]
where
\[ D_n(s) = s^2 R_{x1} R_{x2} C_1 C_2 + s G_{R_{x1}} R_{x2} C_1 + \alpha_1 \alpha_2 \beta_1 \beta_2 \]  \hspace{1cm} (6.1.9)

So the output functions (6.1.6) to (6.1.8) are characterized by following filter parameters:
\[ \omega_0|_n = \left( \frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_{x1} R_{x2} C_1 C_2} \right)^{1/2}, \quad \omega_0|_n = \frac{1}{R_1 C_2}, \quad \text{and} \quad Q_0|_n = R_1 \left( \frac{\alpha_1 \alpha_2 \beta_1 \beta_2 C_2}{R_{x1} R_{x2} C_1} \right)^{1/2} \]  \hspace{1cm} (6.1.10)

It can be observed from (6.1.10) that the value of bandwidth remains unchanged in the presence of nonidealities. The values of pole \( \omega_0 \) and pole \( Q_0 \) are modified slightly which can be accommodated by tuning external bias currents (\( I_{01} \) and \( I_{02} \)).

The active and passive sensitivity analysis of pole \( \omega_0 \) and pole \( Q_0 \) are given as
\[ S_{\omega_0}^{\omega_0} = S_{\alpha_1}^{\alpha_1} = S_{\alpha_2}^{\alpha_2} = S_{\beta_1}^{\beta_1} = S_{\beta_2}^{\beta_2} = \frac{1}{2}, \quad S_{R_{x1}}^{\omega_0} = S_{R_{x2}}^{\omega_0} = S_{C_1}^{\omega_0} = S_{C_2}^{\omega_0} = -\frac{1}{2}, \quad S_{R_1}^{\omega_0} = 0 \]
\[ S_{Q_0}^{Q_0} = S_{\alpha_1}^{\alpha_1} = S_{\alpha_2}^{\alpha_2} = S_{\beta_1}^{\beta_1} = S_{\beta_2}^{\beta_2} = \frac{1}{2}, \quad S_{R_{x1}}^{Q_0} = S_{R_{x2}}^{Q_0} = S_{C_1}^{Q_0} = -S_{C_2}^{Q_0} = -\frac{1}{2}, \quad S_{R_1}^{Q_0} = 1 \]

The active and passive sensitivities of pole \( \omega_0 \) and pole \( Q_0 \) are within unity in magnitude and thus the proposed structure can be classified as insensitive.

**6.1.3 RESULTS**

To validate the theoretical predictions, the proposed circuit is simulated with SPICE using translinear current conveyor (section 2.3) and typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) of bipolar ALA arrays [112] with supply voltage of \( \pm 2.5 \) volts. To illustrate orthogonal tunability of \( Q_0 \) with \( \omega_0 \), a band pass filter is designed at a center frequency of 127 kHz for quality factors of 3.0, 5.0, 10.0 with
grounded resistor of values 0.375 kΩ, 0.625 kΩ, 1.250 kΩ respectively, C₁ = C₂ = 10 nF and I₀₁ = I₀₂ = 100 μA. The simulation results are shown in Fig 6.2. Orthogonal tunability of Q₀ with ω₀ for notch filter is also shown in Fig. 6.3. This is designed at f₀ = 127 kHz with C₁ = C₂ = 10 nF for different values of Q₀ as shown in Table 6.3.

The proposed circuit is also tested for dynamic range i.e. up to what level of input signal the output is within the permitted distortion level. Figure 6.4 shows the plot of percentage total harmonic distortion variation with input signal amplitude. It is clear from Fig. 6.4 that for large range of input signal level, output distortion is within acceptable limit [60] of the order of THD = 5%, so the proposed circuit is also useful for even large signal.

Table 6.3
Component values for various Quality factors.

<table>
<thead>
<tr>
<th>Q₀</th>
<th>R₁(kΩ)</th>
<th>Rₓ₂(kΩ) (I₀₂(μA))</th>
<th>Rₓ₁(kΩ) (I₀₁(μA))</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>0.0625</td>
<td>0.0625 (200)</td>
<td>0.25 (50)</td>
</tr>
<tr>
<td>1.0</td>
<td>0.125</td>
<td>0.125 (100)</td>
<td>0.125 (100)</td>
</tr>
<tr>
<td>2.0</td>
<td>0.25</td>
<td>0.25 (50)</td>
<td>0.0625 (200)</td>
</tr>
</tbody>
</table>

Fig. 6.2 Orthogonal tunability of band pass response.
Fig. 6.3 Orthogonal tunability of notch response.

Fig. 6.4 Dependence of total harmonic distortion (%THD) on input signal.

6.1.4 CONCLUSION

A current mode universal two-input three-output and three-input three-output filter using MOCCCI is presented. The filter employs only two MOCCCIIs and three passive components. The proposed circuit possesses all the characteristics of the universal filter as that of Wang and Lee [100] and Tu et al. [101]. Even if the first CCII of circuit [100] is replaced by CCCH, the proposed circuit requires same number of
passive components but one less number in count of current conveyors. Over and above, to realize an important feature of filter i.e. the independent adjustability of $Q_0$ keeping $\omega_0$ unchanged, the circuit [100] requires additional components i.e. at least one op amp and two resistors (one of which should be variable) [21]. Although the circuit of Tu et al. [101] requires one less number of passive components when CCIIIs are replaced by CCCIIIs, but lacks independent adjustability of $Q_0$ keeping $\omega_0$ unchanged. The proposed circuit requires a very simple matching constraint of passive elements for HP, notch and AP functions. All the passive components are grounded which is beneficial from integrated circuit implementation point of view. Besides the attractive feature of orthogonal control of $Q_0$ and $\omega_0$, the proposed circuit enjoys the electronic control of $\omega_0$, $Q_{0b}$, $\omega_0/Q_0$ and band pass gain via bias currents of MOCCCIIs, orthogonal tunability of $\omega_0$ and $\omega_0/Q_0$ and has low passive and active sensitivities. This structure may also be used as three-input two-output universal filter.

6.2 VOLTAGE MODE MIMO FILTER

A number of voltage mode (VM) multiple input single output (MISO) and multiple input multiple output (MIMO) universal filter configurations of varying complexity and features have been proposed in the literature [91] – [98], [102]. A critical investigation of available MIMO/ MISO voltage mode configurations [91] – [98], [102] reveals the following: use of excessive number of current conveyors [102], requirement of complex matching constraints [97], requirement of both normal and inverted input voltage for all pass(AP) function realization [91], [93], [96], [98], [102] and hence would require one more active device (CC or op amp etc.), requirement of capacitor injection
of excitation signals [91] - [97] or floating capacitor [98] in the circuit design, availability of all filter functions at high input impedance [98], not completely extendable from CCII to current controlled conveyor (CCCIIT) based structure [91], [92], [94], [95], [98], [102] hence not suitable for electronic adjustability without inclusion of any passive resistor.

Thus, it is observed that no MIMO/ MISO universal VM filter structure has yet been reported having simultaneously the following features:

(i) high input impedance,
(ii) use of all grounded capacitors,
(iii) non-requirement of input voltage inversion for the realization of all five generic functions of universal filter,
(iv) implementation with all CCCIIT based current controlled conveyors.

In this section, a new MIMO (three-input two-output) VM universal filter structure has been proposed that employs three DOCCCIIT and possesses all the above mentioned characteristics. Moreover the new structure employs all grounded capacitors which are ideal from integrated circuit implementation viewpoint [122]. The filter parameters are orthogonally and electronically adjustable by bias current of CCCIIT and filter possesses low component spread and hence allows realization of high $Q_0$.

6.2.1 CIRCUIT DESCRIPTION

The proposed network of Fig. 6.5 is based on employing DOCCCIITs. The DOCCCIITs have high impedance at $y$ terminal i.e. $i_y = 0$. Port relations using standard notations can be represented as
\[ V_x = V_T + i_x | R_x(I_0) |, \quad i_{z+} = i_x, \quad \text{and} \quad i_{z-} = -i_x, \]

where \( R_{xi} = V_T/2I_0 \), \( V_T \) is the thermal voltage, \( I_0 \) is bias current of DOCCII and \( i = 1, 2, 3 \).

Analysis of the circuit results in the following output voltage functions:

\[
V_{out1} = \frac{-sV_{in2}R_xC_2 + V_{in1}}{D(s)} \quad (6.2.1)
\]

\[
V_{out2} = \frac{V_{in3}D(s) + s^2R_xR_{x2}C_1C_2V_{in2} - sR_xC_1V_{in1}}{D(s)} \quad (6.2.2)
\]

where

\[
D(s) = s^2R_xR_{x2}C_1C_2 + sR_xC_1 + 1 \quad (6.2.3)
\]

**Fig. 6.5 Proposed DOCCII based MIMO voltage mode filter.**

From (6.2.1) and (6.2.2) one can see that specializations in the numerator results in the following filter responses:

(i) Low pass response: At \( V_{out1} \) with \( V_{in2} = 0 \) and \( V_{in1} = V_{in} \),

(ii) Band pass response: At \( V_{out1} \) with \( V_{in1} = 0 \) and \( V_{in2} = V_{in} \),\n
At \( V_{out2} \) with \( V_{in2} = V_{in3} = 0 \) and \( V_{in1} = V_{in} \),

(iii) High pass response: At \( V_{out2} \) with \( V_{in3} = V_{in1} = 0 \) and \( V_{in2} = V_{in} \),

(iv) Notch response: At \( V_{out2} \) with \( V_{in2} = 0, \ V_{in3} = V_{in1} = V_{in} \) and \( R_{x1} = R_{x3} \),

(v) All pass response: At \( V_{out2} \) with \( V_{in2} = 0, \ V_{in3} = V_{in1} = V_{in} \) and \( 2R_{x1} = R_{x3} \).
Thus the proposed structure can be viewed as three-input two-output voltage mode universal filter. It may be noted that very simple matching constraint is required for AP and notch responses. The filters are characterized by

\[ \omega_0 = \left( \frac{1}{R_{x1}R_{x2}C_1C_2} \right)^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{R_{x2}C_2}, \quad \text{and} \quad Q_0 = \left( \frac{R_{x2}C_2}{R_{x1}C_1} \right)^{1/2} \]  

(6.2.4)

Equation (6.2.4) reveals that \( \omega_0 \) can be adjusted by varying bias current \( I_{01} \) without disturbing \( \omega_0/Q_0 \). The \( \omega_0 \) and \( Q_0 \) are orthogonally adjustable with simultaneous adjustment of \( I_{01} \) and \( I_{02} \). Equation (6.2.4) also indicates that high values of Q-factor will be obtained from moderate values of ratios of passive components i.e. from low component spread [123]. These ratios can be chosen as \( R_{x2}/R_{x1} = C_2/C_1 = Q_0 \). Hence the spread of the component values becomes of the order of \( \sqrt{Q_0} \). This feature of the filter related to the component spread allows the realization of high \( Q_0 \) values more accurately as compared to the topologies where the spread of passive components becomes \( Q_0 \) or \( Q_0^2 \) [123]. In addition we can see that the gain of the band pass filter at \( V_{out2} = R_{x3}/R_{x1} \) can be varied independent of Q-factor of the filter and is adjustable without disturbing the \( \omega_0 \) via bias current \( I_{03} \). This also constitutes an important feature of this filter as this can be used for Q-tuning purpose [108].

6.2.2 EFFECT OF NONIDEALITIES

Taking nonidealities, as outlined in section 2.3, into consideration, (6.2.2) and (6.2.3.) are modified to
\[ V_{out1|n} = \frac{(\alpha_1 \alpha_2 + (1-\alpha_1)(1+sR_{x2}C_2))\beta_1 V_m1 - s\alpha_2 \beta_2 V_{in2} R_{x1} C_2}{D_n(s)} \]  
\[ V_{out2|n} = \frac{(\beta_3 V_{in3} D_n(s) + (1-\alpha_1 + sR_{x1} C_1) s\alpha_2 \beta_2 R_{x2} C_2 V_{in2} -)}{s\alpha_1 \alpha_2 \beta_2 R_{x3} C_2} \frac{D_n(s)}{D_n(s)} \]

where

\[ D_n(s) = s^2 R_{x1} R_{x2} C_1 C_2 + sR_{x1} C_1 + sR_{x2} C_2 (1-\alpha_1) + (1-\alpha_1) + \alpha_1 \alpha_2 \]

The filter parameters can now be expressed as

\[ \omega_0|_n = \left( \frac{\alpha_1 \alpha_2 + 1-\alpha_1}{R_{x1} R_{x2} C_1 C_2} \right)^{1/2} \quad \omega_0| = \frac{1}{R_{x2} C_2} + \frac{1-\alpha_1}{R_{x1} C_1} \quad \text{and} \]

\[ Q_0|_n = \frac{\sqrt{\alpha_1 \alpha_2 + 1-\alpha_1} \sqrt{R_{x1} R_{x2} C_1 C_2}}{R_{x1} C_1 + R_{x2} C_2 (1-\alpha_1)} \]

It may be noted from (6.2.8) that voltage nonidealities have no effect on filter parameters. The values of pole \( \omega_0, \omega_0/Q_0 \) and pole \( Q_0 \) change slightly. The effect can however be absorbed by tuning external bias currents (I_{01} and I_{02}). The active and passive sensitivity analysis of pole \( \omega_0 \) and pole \( Q_0 \) are given as

\[ S_{\omega_0} = \frac{1}{2}, S_{\beta_1} = S_{\beta_2} = S_{\beta_3} = S_{C_1} = S_{C_2} = -\frac{1}{2}, \]
\[ S_{\omega_0} = S_{\alpha_1} = S_{\alpha_2} = S_{\alpha_3} = 0, \]
\[ S_{R_{x3}} = S_{R_{x3}} = S_{C_1} = S_{C_2} = Q^2, S_{\omega_0} = S_{\beta_1} = S_{\beta_2} = S_{\beta_3} = 0, \]
\[ S_{R_{x1}} = S_{R_{x2}} = S_{C_1} = S_{C_2} = -\frac{1}{2}, S_{R_{x3}} = 0. \]
6.2.3 RESULTS

To validate the theoretical predictions, the proposed circuit is simulated with PSPICE using translinear current conveyor [13] and typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) [112] with supply voltages of ±2.5 volts. To obtain a filter with a pole frequency of \( f_0 = 127 \) kHz and quality factor of \( Q_0 = 1 \), the component values are taken as \( C_1 = C_2 = 1 \) nF, and \( I_{01} = I_{02} = 10 \) µA. Figure 6.6 shows the simulation and theoretical results for low pass and band pass responses. Orthogonal tunability of \( Q_0, \omega_0 \) and gain of band pass response is shown in Fig. 6.6 for \( C_1 = C_2 = 1 \) nF. Figure 6.7(a) shows the tuning of \( \omega_0 \) for \( Q_0 \) fixed at unity for the values of \( I_{01} = I_{02} = I_{03} \) of 10 µA, 50 µA, and 100 µA which results values of \( f_0 \) as 122 kHz, 575 kHz, and 1.18 MHz respectively.

![Graph](image)

Fig 6.6 Simulated and theoretical low pass and band pass responses of the proposed filter at \( V_{out1} \).
Fig 6.7 Orthogonal tunability of (a) $\omega_0$, (b) $Q_0$ and (c) gain of the proposed filter at $V_{\text{out2}}$.

Figure 6.7(b) shows the tuning of $Q_0$ for $f_0$ fixed at 127 kHz with: $Q_0 = 1$, $I_{02} = I_{01} = I_{03} = 10 \, \mu A$; $Q_0 = 2$, $I_{02} = 5 \, \mu A$, $I_{01} = I_{03} = 20 \, \mu A$; and $Q_0 = 4$, $I_{02} = 2.5 \, \mu A$, $I_{01} = I_{03} = 40 \, \mu A$. Figure 6.7(c) shows the tuning of gain for fixed value of $f_0 = 127$ kHz and $Q_0 = 1$ with $I_{01} = I_{02} = 10 \, \mu A$ and $I_{03} = 10 \, \mu A$, $2 \, \mu A$, $1 \, \mu A$ for gains of 1, 5 and 10 respectively. The simulations confirm workability of the proposed circuit.

6.2.4 CONCLUSION

Although a number of current conveyor based voltage mode universal MISO/ MIMO filter structures realizable with two to as many as four CCs are known in the literature,
however, no high input impedance CCII based universal voltage mode filter configuration with all grounded capacitors and without needing both types of voltage input signal has been published earlier. The work presented in this section has filled this void by presenting a novel VM universal biquad. The structure has the following attractive features:

(i) uses only three DO-CCCIIs,

(ii) realizes all the standard functions of universal filter i.e. LP, BP, HP, notch and AP using only one type of input signal,

(iii) high input impedance,

(iv) $\omega_0$, $Q_0$ and $\omega_0/Q_0$ are tunable with bias currents of DOCCCIIs,

(v) $\omega_0$ can be adjusted independently without disturbing $Q_0$ and $\omega_0/Q_0$ and vice versa.

(vi) passive sensitivities of $\omega_0$, $Q_0$ and $\omega_0/Q_0$ are low and within unity in magnitude except for $S_{\delta}^{Q_0}$.

6.3 MIXED MODE MIMO FILTER

It has already been discussed in chapter 5 that generalized mixed mode filters are a special class of filters from which one may get all four modes of filters such as current mode, voltage mode, transimpedance mode and transadmittance mode. Chapter 5 covered MISO mixed mode filters. This section is devoted for two MIMO mixed mode filters.
6.3.1 CCII +/- CCII-/ DO-CCII BASED MIMO CONFIGURATION

Generalized mixed mode filter structures can process either a voltage or current input signal and are able to produce either a voltage or current as output. The literature on generalized mixed mode filter circuits is limited and a comparison of available literature is made in section 5.3.1.

In this section, a MIMO mixed mode universal filter structure has been introduced that employs three second generation current conveyors (CCIIIs), five resistors and two capacitors. The structure realizes all five generic filter functions in all four modes (current mode, voltage mode, transimpedance and transadmittance modes) and requires a very simple matching constraint. The electronic tunability of filter parameters can be achieved by replacing two CCIIIs and their series resistance at port x by current controlled conveyors (CCCIIIs) [13]. The filter enjoys attractive features: low active and passive sensitivities and orthogonal tunability of pole quality factor $Q_0$ and pole natural frequency $\omega_0$ via grounded resistors.

6.3.1.1 CIRCUIT DESCRIPTION

Analysis of the proposed mixed mode filter of Fig. 6.8 yields the output functions as

$$V_{out} = \frac{R_4}{R_3} \frac{N_v(s) + N_i(s)}{D(s)}$$  \hspace{1cm} (6.3.1)
\[ I_{\text{out}} = \frac{1}{R_3} \frac{N_y(s) + N_i(s)}{D(s)} \quad (6.3.2) \]

where

\[ N_y(s) = V_{\text{in}3} D(s) - sC_1 R_2 V_{\text{in}1} - s^2 C_1 C_2 R_1 R_2 V_{\text{in}2} \quad (6.3.3) \]

\[ N_i(s) = R_3 I_{\text{in}3} D(s) - sC_1 R_1 R_2 I_{\text{in}2} - R_2 I_{\text{in}1} \quad (6.3.4) \]

\[ D(s) = R_1 R_2 C_1 C_2 s^2 + sG R_1 R_2 C_1 + 1 \quad (6.3.5) \]

\( V_{\text{in}1}, V_{\text{in}2}, V_{\text{in}3} \) = input voltages, \( V_{\text{out}} \) = output voltage, \( I_{\text{in}1}, I_{\text{in}2}, I_{\text{in}3} \) = input currents, and \( I_{\text{out}} \) = output current.

![Proposed mixed mode filter](image)

Fig. 6.8 Proposed mixed mode filter.

From (6.3.1) and (6.3.2) we can see that specializations in the numerator results in the following filter responses:

**Case I:** With \( I_{\text{in}1} = I_{\text{in}2} = I_{\text{in}3} = 0 \), the voltage mode (VM) and transadmittance mode responses are obtained at \( V_{\text{out}} \) and \( I_{\text{out}} \) respectively under the conditions as shown in Table 6.4.
Case II: With $V_{in1} = V_{in2} = V_{in3} = 0$, the current mode (CM) and transimpedance responses are obtained at $I_{out}$ and $V_{out}$ respectively under the conditions as shown in Table 6.4.

Thus the proposed structure can be viewed as a generalized mixed mode universal filter. It can be noted that very simple component matching constraint is required. From (6.3.1) to (6.3.5), the parameters $\omega_0, Q_0$ and $\omega_0/Q_0$ can be expressed as

$$\omega_0 = \left( \frac{1}{R_1 R_2 C_1 C_2} \right)^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{RC_2}, \quad \text{and} \quad Q_0 = R \left( \frac{C_2}{R_1 R_2 C_1} \right)^{1/2} \quad (6.3.6)$$

**Table 6.4**

Conditions for obtaining mixed mode output.

<table>
<thead>
<tr>
<th>Function type</th>
<th>Conditions Case I (VM and Transadmittance Mode)</th>
<th>Conditions Case II (CM and Transimpedance Mode)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low pass</td>
<td>$V_{in1} = V_{in2} = V_{in3} = V_{in}$, $R_1 = R$</td>
<td>$I_{in2} = I_{in3} = 0$, $I_{in1} = I_{in}$</td>
</tr>
<tr>
<td>Band pass</td>
<td>$V_{in2} = V_{in3} = 0$, $V_{in1} = V_{in}$</td>
<td>$I_{in1} = I_{in3} = 0$, $I_{in2} = I_{in}$</td>
</tr>
<tr>
<td>High pass</td>
<td>$V_{in1} = V_{in3} = 0$, $V_{in2} = V_{in}$</td>
<td>$I_{in1} = I_{in2} = I_{in3} = I_{in}$, $R_2 = R_3 = R$</td>
</tr>
<tr>
<td>Notch</td>
<td>$V_{in2} = 0$, $V_{in1} = V_{in3} = V_{in}$, $R_1 = R$</td>
<td>$I_{in1} = 0$, $I_{in2} = I_{in3} = I_{in}$, $R_2 = R_3 = R$</td>
</tr>
<tr>
<td>All pass</td>
<td>$V_{in2} = 0$, $V_{in1} = V_{in3} = V_{in}$, $R_1 = 2R$</td>
<td>$I_{in1} = 0$, $I_{in2} = I_{in3} = I_{in}$, $2R_2 = 2R_3 = R$</td>
</tr>
</tbody>
</table>

It can be seen that in case-I, $Q_0$ can be independently controlled keeping $\omega_0$ same through the grounded resistor $R$ for band pass and high pass functions and through $R$ and then $R_1$ and $R_2$ in that order for low pass, notch and all pass functions. Similarly, in case-II, for low pass and band pass functions through $R$ and for high pass, notch and all pass functions through $R$ and then $R_2$ and $R_3$ in that order. The parameter $\omega_0$ can be adjusted via $R_2$ in case-I and via $R_1$ in case-II without disturbing $\omega_0/Q_0$. 
The expression for quality factor also indicates an important feature of the implementation i.e. the achievement of high values of the Q-factor from moderate value of the passive component ratios. The ratios can be chosen as \((R/R_1) = (R/R_2) = (C_2/C_1) = Q^{23}\). In addition we can see that for current mode configuration (case II, band pass) the gain of the filter \((= R/R_3)\) is directly proportional to the Q-factor of the filter and is adjustable without disturbing \(\omega_0\). This also constitutes an important feature of this filter as the output current can be used for Q-tuning purpose [108].

6.3.1.2 EFFECT OF NONIDEALITIES

The frequency performance of the filter circuit may deviate from the ideal one due to nonidealities as discussed in section 2.2. As in [64] and [107], the loading effect of the parasitics \((R_x, R_y, R_z, C_y \text{ and } C_z)\) can be eliminated if x-port of CCIIs is terminated with suitable resistance and parallel resistance-capacitance combination of suitable value is placed at y and z ports.

In the proposed topology, the x-port of all the three CCIIs is terminated with resistances \((R_1, R_2 \text{ and } R_3)\). Hence parasitic resistances \(R_{x1}, R_{x2} \text{ and } R_{x3}\) can be fully compensated by decreasing the values of \(R_1, R_2 \text{ and } R_3\) respectively by the same amount. As the parasitic capacitances and resistances present at y and z ports of CCIIs are in parallel with external capacitors \((C_1, C_2)\), the effect of parasitics can be compensated by pre-distorting the values of \(C_1\) and \(C_2\) and selecting CCIIs with high output impedance.

Taking nonidealities outlined in section 2.2 into account, the output functions (6.3.1) and (6.3.2) modify to
\[ V_{\text{out}} \bigg|_n = \frac{\alpha_3 R_4 N_{in}(s) + N_{in}(s)}{R_3 D_n(s)} \]  
(6.3.7)

\[ I_{\text{out}} \bigg|_n = \frac{\alpha_3 N_{in}(s) + N_{in}(s)}{R_3 D_n(s)} \]  
(6.3.8)

where

\[ N_{in}(s) = V_{in3} D_n(s) - s C_1 R_2 \alpha_1 V_{in1} - s^2 C_1 C_2 R_1 R_2 V_{in2}, \]

\[ N_{in}(s) = R_3 I_{in3} D_n(s) - s C_1 R_1 R_2 I_{in2} - R_2 \alpha_1 I_{in1} \]  
(6.3.9)

and

\[ D_n(s) = R_1 R_2 C_1 C_2 s^2 + s G R_1 R_2 C_1 + \alpha_1 \alpha_2 \beta_1 \beta_2 \]  
(6.3.10)

From (6.3.7) to (6.3.10) the parameters \( \omega_0, Q_0 \) and \( \omega_0/Q_0 \) can be expressed as

\[ \omega_0 \bigg|_n = \left( \frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_1 R_2 C_1 C_2} \right)^{1/2} \equiv \left( \frac{1}{R_1 R_2 C_1 C_2} \right)^{1/2}, \quad \omega_0 \bigg|_n = \frac{1}{RC_2}, \quad \text{and} \]

\[ Q_0 \bigg|_n = R \left( \frac{\alpha_1 \alpha_2 \beta_1 \beta_2 C_2}{R_1 R_2 C_1} \right)^{1/2} \equiv R \left( \frac{C_2}{R_1 R_2 C_1} \right)^{1/2} \]  
(6.3.11)

It may be observed form (6.3.11) that bandwidth is not effected by \( \alpha \) and \( \beta \) nonidealities but \( \omega_0 \) and \( Q_0 \) are slightly affected.

The active and passive sensitivity analysis of pole \( \omega_0 \) and pole \( Q_0 \) are given as

\[ S_{a_0}^{\alpha_0} = S_{\alpha_2}^{\alpha_0} = S_{\beta_1}^{\alpha_0} = S_{\beta_2}^{\alpha_0} = \frac{1}{2}, \quad S_{R_1}^{\alpha_0} = S_{R_2}^{\alpha_0} = S_{C_1}^{\alpha_0} = S_{C_2}^{\alpha_0} = -\frac{1}{2}, \quad S_{R}^{\alpha_0} = S_{R_2}^{\alpha_0} = S_{R_4}^{\alpha_0} = 0 \]

\[ S_{Q_0}^{\alpha_0} = S_{\alpha_2}^{Q_0} = S_{\beta_1}^{Q_0} = S_{\beta_2}^{Q_0} = \frac{1}{2}, \quad S_{R_1}^{Q_0} = S_{R_2}^{Q_0} = S_{C_1}^{Q_0} = -S_{C_2}^{Q_0} = -\frac{1}{2}, \quad S_{R}^{Q_0} = 1, \quad S_{R_1}^{Q_0} = S_{R_4}^{Q_0} = 0 \]
It can further be noted that the active and passive sensitivities of $\omega_0$ and $Q_0$ are small and within unity in magnitude and thus the proposed structure can be classified as insensitive.

### 6.3.1.3 RESULTS

PSPICE simulations are performed to verify the potentialities of the circuit. The CCII/DOCCII is simulated using translinear current conveyor circuits given in section 2.3 using typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) of high performance ALA200 arrays [112] with DC supply voltage of $\pm 10$V. The simulated and theoretical responses for the transadmittance mode low pass, ban pass and high pass filters are shown in Fig. 6.9 for the component values $R = R_1 = R_2 = R_3 = R_4 = 1 \, k\Omega$, $C_1 = 1$ nF, $C_2 = 1$ nF. Orthogonal tunability of $Q_0$ and $\omega_0$ of band pass filter under transadmittance mode of operation is studied for the cases shown in Table 6.5. Figure 6.10 (a) shows the simulation results for control of $Q_0$ while keeping $\omega_0$ fixed (159.23 kHz) with $C_1 = C_2 = 1$ nF, $R_3 = R_4 = 1$ k$\Omega$. Figure 6.10(b) shows the simulation results for control of $\omega_0$ while keeping $Q_0$ (= 1) fixed for $C_1 = C_2 = 1$ nF, $R_3 = R_4 = 1$ k$\Omega$.

### Table 6.5

Components used for showing orthogonal tunability of $\omega_0$ and $Q_0$.

<table>
<thead>
<tr>
<th>Fixed $\omega_0$</th>
<th>Fixed $Q_0$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Q_0$</td>
<td>$R_1 (k\Omega)$</td>
</tr>
<tr>
<td>1.0</td>
<td>1</td>
</tr>
<tr>
<td>2.0</td>
<td>2</td>
</tr>
<tr>
<td>4.0</td>
<td>4</td>
</tr>
</tbody>
</table>
Fig. 6.9 Simulated and theoretical responses of the proposed mixed mode filter.

--- Simulated  --- Theoretical

(a) 

Fig. 6.10 (a) Orthogonal tunability of $Q_0$ and $\omega_0$ with fixed $f_0$.

(b) Orthogonal tunability of $\omega_0$ and $Q_0$ with fixed $Q_0$.

The current mode band pass filter is tested for gain and quality factor tuning while keeping pole frequency constant at 159.23 kHz, with $C_1 = C_2 = 1 \text{nF}$, $R_1 = R_2 = R_3 = 1 \text{k}\Omega$ and the value of resistance $R = 1 \text{k}\Omega$, 2 k\Omega, 3 k\Omega used for gain = quality factor = 1, 2, 3 respectively. The simulated results are shown in Fig. 6.11. The simulated results agree well with the theoretical values.
6.3.2 DOCCCI/ MOCCCI BASED MIMO CONFIGURATION

It has already been discussed in section 6.3.1 that the available mixed mode universal/ multifunctional filters [26], [103] - [106], [128] - [130] are very few in number. A comparison of the previous work [26], [103] - [106], [128] - [130] including the structure presented in section 6.3.1 is made in Table 6.6. The mixed mode structures presented in this section are developed based on refs. [64], [101]. The refs. [64], [101] are analyzed only for current mode responses, whereas the proposed circuits are developed for a generalized mixed mode operation by inserting voltage signals at suitable terminals and a little alteration of circuit.

Two new generalized mixed mode universal filters are presented in this section, one of these employs only two dual output current controlled conveyors (DOCCCIIs), one resistor and two capacitors whereas the other uses one DOCCCI, one multiple output CCCII (MOCCCI), two capacitors and two resistors. The proposed structures realize all the standard functions of generalized mixed mode universal filter i.e. LP, BP,
HP, notch and AP in all the four modes (current mode, voltage mode, transimpedance mode, transadmittance mode). Most of the responses are available in both non-inverted (N.I.) and inverted (I.) form. This feature is useful in some applications, such as in the design of inverting/ non inverting band pass filter used in hearing aid for selective amplification or attenuation of audio signal [131]. The filter, under all operations, exhibits low active and passive sensitivities.

### Table 6.6

Performance comparison of recently reported mixed mode filters.

<table>
<thead>
<tr>
<th>Related Work</th>
<th>No. Of Components Used</th>
<th>Functions Performed</th>
<th>No. of Input Currents used</th>
<th>Orthogonal tunability of $\omega_0$, $Q_0$ and $\omega/\omega_0$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current Convey or/CFO A</td>
<td>R C</td>
<td>Current Mode</td>
<td>Trans-impedance Mode</td>
<td>Trans-admittance Mode</td>
</tr>
<tr>
<td>Ref. 104</td>
<td>4 CCCIIs</td>
<td>4 2</td>
<td>—</td>
<td>LP, BP, HP</td>
</tr>
<tr>
<td>Ref. 105</td>
<td>4 CCCIIs</td>
<td>2</td>
<td>LP, BP, HP</td>
<td>LP, BP, HP</td>
</tr>
<tr>
<td>Ref. 106</td>
<td>7 CCCIIs</td>
<td>3</td>
<td>LP, BP, HP, notch, AP</td>
<td>LP, BP, HP, notch, AP</td>
</tr>
<tr>
<td>Ref. 26</td>
<td>4 CFOAs</td>
<td>3</td>
<td>LP, BP, HP, notch, AP</td>
<td>LP, BP, HP, notch, AP</td>
</tr>
<tr>
<td>Ref. 130 (structure of section 5.3)</td>
<td>3 CCCIIs</td>
<td>2</td>
<td>--do---</td>
<td>--do---</td>
</tr>
<tr>
<td>Ref. 128</td>
<td>6 CCCIIs</td>
<td>2</td>
<td>LP, BP, HP</td>
<td>LP, BP, HP</td>
</tr>
<tr>
<td>Structure of section 6.3.1</td>
<td>3 CCCIIs</td>
<td>2</td>
<td>LP, BP, HP, notch, AP</td>
<td>LP, BP, HP, notch, AP</td>
</tr>
<tr>
<td>Present work</td>
<td>2 CCCIIs</td>
<td>2</td>
<td>LP, BP, HP, notch, AP</td>
<td>LP, BP, HP, notch, AP</td>
</tr>
</tbody>
</table>

*The number of input currents can be reduced to 3 by connecting in series a resistor $R_{out}$ and a switch between $I_{out2}$ and ground in Fig. 1 [as used in Ref. 130].
6.3.2.1 CIRCUIT DESCRIPTION

The proposed network of Fig. 6.12 is based on employing DOCCCIIs. The DOCCCIIs have high impedance y terminal i.e. $i_Y = 0$. Port relationship using standard notations can be represented as

$$v_X = v_Y + i_X | R_{X_i}(I_0) |, \quad i_{Z_+} = i_X, \text{ and } i_{Z_-} = -i_X, \quad i = 1, 2.$$

where $R_{x_i} = \frac{V_T}{2I_0}, V_T$ is the thermal voltage, $I_0$ is bias current of DOCCCII.

![Diagram of proposed mixed mode circuit using DOCCCIIs.](image)

Fig. 6.12 Proposed mixed mode circuit using DOCCCIIs.

Analysis yields the following output functions:

$$I_{\text{out}1} = \frac{N_{v_1}(s) + N_{i_1}(s)}{D(s)}$$  \hspace{1cm} (6.3.12)

$$I_{\text{out}2} = \frac{N_{v_2}(s) + N_{i_2}(s)}{D(s)}$$  \hspace{1cm} (6.3.13)

$$V_{\text{out}1} = \frac{N_{v_3}(s) + N_{i_3}(s)}{D(s)}$$  \hspace{1cm} (6.3.14)

where

$$N_{i_1}(s) = I_{\text{in}2} - (G + sC_2)R_{x_2}I_{\text{in}1}$$  \hspace{1cm} (6.3.15)

$$N_{v_1}(s) = sC_2V_{\text{in}2} - (G + sC_2)V_{\text{in}3} - (V_{\text{in}1} - V_{\text{in}4})(G + sC_2)sR_{x_2}C_1$$  \hspace{1cm} (6.3.16)

$$N_{i_2}(s) = I_{\text{in}3}D(s) - sC_1R_{x_1}I_{\text{in}2} - I_{\text{in}1}$$  \hspace{1cm} (6.3.17)
\[ N_{v2}(s) = -(V_{in1} - V_{in4}) s C_1 + s C_1 R_{x1} V_{in3} + s^2 C_1 C_2 R_{x1} (V_{in3} - V_{in2}) \]  
(6.3.18)

\[ N_{i3}(s) = s C_1 R_{x1} R_{x2} I_{in2} + I_{in1} R_{x2} \]  
(6.3.19)

\[ N_{v3}(s) = V_{in3} + s (V_{in1} - V_{in4}) R_{x2} C_1 + s^2 C_1 C_2 R_{x1} R_{x2} V_{in2} \]  
(6.3.20)

\[ D(s) = s^2 R_{x1} R_{x2} C_1 C_2 + s G R_{x1} R_{x2} C_1 + 1 \]  
(6.3.21)

The functionality of the circuit of Fig. 6.12 can be enhanced by replacing second DOCCII by a MOCCII as shown in Fig. 6.13.

![Proposed mixed mode universal filter using DOCCII and MOCCII](image)

Fig. 6.13 Proposed mixed mode universal filter using DOCCII and MOCCII.

Analyzing the circuit, the output voltage \( V_{out2} \) is obtained as

\[ V_{out2} = \frac{N_{v4}(s) + N_{i4}(s)}{D(s)} \]  
(6.3.22)

where

\[ N_{i4}(s) = R_{out} (I_{in4} D(s) - s C_1 R_{x1} I_{in2} - I_{in1}) \]  
(6.3.23)

\[ N_{v4}(s) = R_{out} (-(V_{in1} - V_{in4}) s C_1 + s C_1 G R_{x1} V_{in3} + s^2 C_1 C_2 R_{x1} (V_{in3} - V_{in2})) \]  
(6.3.24)

From (6.3.12) to (6.3.14) one can see that specializations in the numerator result in filter functions as presented in Table 6.7 and Table 6.8 for the circuit of Fig. 6.13.
Case I: With $V_{in1} = V_{in2} = V_{in3} = V_{in4} = 0$, we obtain current mode and transimpedance mode responses under the conditions as shown in Table 6.7.

Case II: With $I_{in1} = I_{in2} = I_{in3} = 0$, we obtain voltage mode and transadmittance mode responses under the conditions as shown in Table 6.8.

### Table 6.7

Specialization for various CM and transimpedance mode responses.

<table>
<thead>
<tr>
<th>Function Type</th>
<th>Case I</th>
<th>Transimpedance mode</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Input</strong></td>
<td><strong>Output</strong></td>
<td><strong>Input</strong></td>
</tr>
<tr>
<td>Low pass</td>
<td>$I_{in1} = 0, I_{in2} = I_{in}$</td>
<td>$I_{out1} (N. I.)$</td>
</tr>
<tr>
<td></td>
<td>$I_{in3} = I_{in2} = 0, I_{in1} = I_{in}$</td>
<td>$I_{out2} (I.)$</td>
</tr>
<tr>
<td>Band pass</td>
<td>$I_{in3} = I_{in1} = 0, I_{in2} = I_{in}$</td>
<td>$I_{out2} (I.)$</td>
</tr>
<tr>
<td></td>
<td>$I_{in4} = I_{in1} = 0, I_{in2} = I_{in}$</td>
<td>$I_{out2} (N. I.)$</td>
</tr>
<tr>
<td>High pass</td>
<td>$I_{in3} = I_{in2} = I_{in1} = I_{in}, R_{x2} = R$</td>
<td>$I_{out2} (N. I.)$</td>
</tr>
<tr>
<td>Notch</td>
<td>$I_{in1} = 0, I_{in3} = I_{in2} = I_{in}, R_{x2} = R$</td>
<td>$I_{out2} (N. I.)$</td>
</tr>
<tr>
<td>All pass</td>
<td>$I_{in1} = 0, I_{in3} = I_{in2} = I_{in}, 2R_{x2} = R$</td>
<td>$I_{out2} (N. I.)$</td>
</tr>
</tbody>
</table>

(Note: N.I. = Non inverting, I. = Inverting)

### Table 6.8.

Specialization for various VM and transadmittance mode responses.

<table>
<thead>
<tr>
<th>Function Type</th>
<th>Voltage mode</th>
<th>Transadmittance mode</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Input</strong></td>
<td><strong>Output</strong></td>
<td><strong>Input</strong></td>
</tr>
<tr>
<td>Low pass</td>
<td>$V_{in4} = V_{in2} = V_{in1} = 0, V_{in3} = V_{in}$</td>
<td>$V_{out1} (N. I.)$</td>
</tr>
<tr>
<td>Band pass</td>
<td>$V_{in4} = V_{in3} = V_{in2} = 0, V_{in1} = V_{in}$</td>
<td>$V_{out1} (N. I.)$</td>
</tr>
<tr>
<td></td>
<td>$V_{in3} = V_{in2} = V_{in1} = 0, V_{in4} = V_{in}$</td>
<td>$V_{out1} (I.)$</td>
</tr>
<tr>
<td></td>
<td>$V_{in3} = V_{in2} = V_{in1} = 0, V_{in4} = V_{in}$</td>
<td>$V_{out2} (N. I.)$</td>
</tr>
<tr>
<td></td>
<td>$V_{in4} = V_{in3} = V_{in2} = 0$, $V_{in1} = V_{in}$</td>
<td></td>
</tr>
<tr>
<td>----------------------</td>
<td>--------------------------------------------------</td>
<td>----------------------</td>
</tr>
<tr>
<td><strong>High pass</strong></td>
<td>$V_{in4} = V_{in3} = V_{in1} = 0$, $V_{in2} = V_{in}$</td>
<td>$V_{out1}$ (N.I.)</td>
</tr>
<tr>
<td></td>
<td>$V_{in4} = V_{in3} = V_{in2} = 0$, $V_{in1} = V_{in}$</td>
<td>$V_{out2}$ (I.)</td>
</tr>
<tr>
<td><strong>Notch</strong></td>
<td>$V_{in4} = V_{in1} = 0$, $V_{in3} = V_{in2} = V_{in}$</td>
<td>$V_{out1}$ (N.I.)</td>
</tr>
<tr>
<td><strong>All pass</strong></td>
<td>$V_{in1} = 0$, $V_{in2} = V_{in3} = V_{in4} = V_{in}$, $R_{x1} = R$</td>
<td>$V_{out1}$ (N.I.)</td>
</tr>
</tbody>
</table>

It is seen from in Table 6.7 and Table 6.8 that all the universal filter responses such as LP, BP, HP, notch and AP are obtained for all the four modes (current mode, voltage mode, transimpedance mode, transadmittance mode) of generalized mixed mode operations. Tables 6.7 and 6.8 also reveal that the outputs are available both in non inverting (N.I.) and inverting (I.) modes which may be useful in typical applications such as hearing aid [131]. All the filters are characterized by

$$
\omega_0 = \left( \frac{1}{R_{x1}R_{x2}C_1C_2} \right)^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{RC_2}, \quad \text{and} \quad Q_0 = R \left( \frac{C_2}{R_{x1}R_{x2}C_1} \right)^{1/2}
$$

(6.3.25)

It may be noted that $Q_0$ can be independently controlled keeping $\omega_0$ unchanged through the grounded resistor $R$ for low pass and band pass functions in case-I and through $R$ and then $R_{x1}$ and $R_{x2}$ for high pass, notch and all pass functions. Similarly, in case-II, the parameter $Q_0$ can be adjusted for fixed $\omega_0$ for low pass, band pass and high pass functions through $R$ whereas notch and all pass functions through $R$ and then $R_{x1}$ and $R_{x2}$. It may be noted that the resistances $R_{x1}$ and $R_{x2}$ can easily be adjusted to the required values by externally controlling the bias currents ($I_{01}$ and $I_{02}$) of DOCCCIIs/
MOCCII. The parameter $\omega_0$ can also be adjusted electronically by controlling bias currents of DOCCIIIs/ MOCCIIIs without disturbing $\omega_0/Q_0$.

### 6.3.2.2 EFFECT OF NONIDEALITIES

As discussed in section 2.2, the frequency performance of the filter circuit may deviate from the ideal one due to nonidealities. The nonidealities effects may be categorized in two groups. The first comes from the frequency dependence of internal current and voltage transfers of DOCCII/ MOCCII. In the presence of the nonidealities, the port relations are modified to

$$v_X = v_T \beta_i(s) + i_X |R_{X_i}(I_{oi}|), \quad i_{Z_{0}} = i_X \alpha_{i_+(s)} \text{ and } i_{Z_{0}} = -i_X \alpha_{i_-(s)}, \quad i = 1, 2.$$  

The voltage transfer function $\beta_i(s)$ and current transfer functions $\alpha_{i_+(s)}$ and $\alpha_{i_-(s)}$ can be modeled as $\beta_i(s) = \frac{\beta_i}{s + \omega_{\beta_i}}, \alpha_{i_+(s)} = \frac{\alpha_{i_+}}{s + \omega_{\alpha_{i_+}}}$, and $\alpha_{i_-(s)} = \frac{\alpha_{i_-}}{s + \omega_{\alpha_{i_-}}}$ respectively. The terms $\beta_i, \alpha_{i_+},$ and $\alpha_{i_-}$ represent low frequency values of $\beta_i(s), \alpha_{i_+(s)},$ and $\alpha_{i_-(s)}$ respectively. The pole frequency of voltage transfer is denoted by $\omega_{\beta_i}$ whereas $\omega_{\alpha_{i_+}} (\omega_{\alpha_{i_-}})$ represent pole frequency for positive (negative) current transfers. Equations (6.3.12) to (6.3.14) and (6.3.22) are modified due to these nonidealities as

$$I_{out 1} = \frac{N_{v11}(s) + N_{m1}(s)}{D_n(s)} \quad (6.3.25)$$

$$I_{out 2} = \frac{N_{v21}(s) + N_{m2}(s)}{D_n(s)} \quad (6.3.26)$$

$$V_{out 1} = \frac{N_{v11}(s) + N_{m1}(s)}{D_n(s)} \quad (6.3.27)$$
\[ V_{\text{out2}} |_{n} = \frac{N_{\text{in4}}(s) + N_{\text{in4}}(s)}{D_n(s)} \]  

(6.3.28)

where

\[ N_{\text{in1}}(s) = \frac{\alpha_{11} \alpha_{12} \beta_1 \beta_2 I_{\text{in2}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} - \frac{\alpha_{11} \beta_1 (G_2 + sC_2) R_x I_{\text{in1}}}{(s + \omega_{\alpha_1})(s + \omega_{\beta_1})} \]  

(6.3.29)

\[ N_{\text{in2}}(s) = \frac{sC_2 \alpha_{11} \alpha_{12} \beta_1 \beta_2 V_{\text{in2}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} - \frac{\alpha_{11} \alpha_{12} \beta_1 (G_2 + sC_2)V_{\text{in3}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} \]  

(6.3.30)

\[ N_{\text{in3}}(s) = \frac{s \alpha_{11} \alpha_{12} \beta_1 \beta_2 C_1 R_s I_{\text{in2}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} - \frac{\alpha_{11} \alpha_{12} \beta_1 \beta_2 I_{\text{in1}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} \]  

(6.3.31)

\[ N_{\text{in2}}(s) = -\frac{\alpha_{11} \alpha_{12} \beta_2}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} \left( \frac{s \beta_1 C_1 V_{\text{in1}}}{(s + \omega_{\beta_1})} - sC_1 V_{\text{in4}} \right) + \]  

(6.3.32)

\[ N_{\text{in3}}(s) = sC_1 R_s R_x I_{\text{in2}} + \frac{\alpha_{11} \beta_1 R_x I_{\text{in1}}}{(s + \omega_{\alpha_1})(s + \omega_{\beta_1})} \]  

(6.3.33)

\[ N_{\text{in4}}(s) = \frac{s \alpha_{11} \beta_1 C_1 R_s V_{\text{in1}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} + \frac{\alpha_{11} \beta_1 \beta_2 I_{\text{in1}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} \]  

(6.3.34)

\[ N_{\text{in4}}(s) = R_{\text{out}} I_{\text{in3}} D_n(s) - \frac{s \alpha_{11} \beta_2 C_1 R_s I_{\text{in2}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} - \frac{\alpha_{11} \alpha_{12} \beta_1 \beta_2 I_{\text{in1}}}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})} \]  

(6.3.35)
\[
N_{vn4}(s) = -R_{out} \left( \frac{\alpha_1 \alpha_2 \beta_2}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})} \left( \frac{s \beta_1 C_{V_{in1}}}{(s + \omega_{\beta_1})} - sC_{V_{in4}} \right) \right) \\
+ \frac{s \alpha_2 \alpha_2 (G_2 + sC_2)C_1 R_{s1} V_{in3} \alpha_1 \beta_2}{(s + \omega_{\alpha_2})(s + \omega_{\beta_2})} - \frac{s^2 C_1 C_2 R_{s1} V_{in2}}{(s + \omega_{\alpha_2})(s + \omega_{\beta_2})}
\]

\[
D_a(s) = sC_1(G_2 + sC_2)R_{s1}R_{s2} + \frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{(s + \omega_{\alpha_1})(s + \omega_{\alpha_2})(s + \omega_{\beta_1})(s + \omega_{\beta_2})}
\]

Equations (6.3.25) to (6.3.37) clearly indicate that the pole frequencies (\(f_a\) and \(f_b\)) of current and voltage transfer functions (\(\alpha_i\) and \(\beta_i\)) of DOCCCIIs/ MOCCCIIs affect the overall filter response. The effect, can however be ignored if the operating frequencies are chosen sufficiently smaller than voltage and current transfer pole frequencies of DOCCCIIs/ MOCCCIIs.

The second group of nonidealities comes from parasitics of DOCCCIIs/ MOCCCIIs comprising of resistors and capacitors connected in parallel at terminals y and z (i.e. \(R_y\), \(C_y\), \(R_z\), \(C_z\)) and inductance \(L_x\) in series to \(R_x\) at terminal x [108]. The effects of these parasitics on filter response depend strongly on circuit topology. In the presence of these parasitics the circuit given in Fig. 6.13 modifies to Fig. 6.14 where \(C_{1p} = C_{y1} \parallel C_{z2}\), \(C_{2p} = C_{z1} \parallel C_{y2}\), \(G_{1p} = 1/(R_{y1} \parallel R_{z2})\), and \(G_{2p} = 1/(R_{y2} \parallel R_{z1})\). The inductance is ignored in Fig. 6.14 as it affects the frequency response only at very high frequencies.
Fig. 6.14 Proposed CCCII based mixed mode filter structure including parasitics.

Considering the parasitics outlined above, (6.3.15) to (6.3.21), (6.3.23) and (6.3.24) are modified as

\[ N_{in1}(s) = I_{in2} - (G_{2eq} + sC_{2eq})R_{x2}I_{in1} \quad (6.3.38) \]

\[ N_{vn1}(s) = sC_2V_{in2} - (G_{2eq} + sC_{2eq})V_{in3} - (sC_1V_{in1} - (G_{1p} + sC_{1eq})V_{in4})(G_{2eq} + sC_{2eq})R_{x2} \]

\[ N_{in2}(s) = I_{in3}D_n(s) - (G_{1p} + sC_{1eq})R_{x1}I_{in2} - I_{in1} \quad (6.3.40) \]

\[ N_{vn2}(s) = -(sC_1V_{in1} - (G_{1p} + sC_{1eq})V_{in4}) + (G_{2eq} + sC_{2eq})(G_{1p} + sC_{1eq})R_{x1}V_{in3} - \]

\[ sC_2(G_{1p} + sC_{1eq})R_{x1}V_{in2} \quad (6.3.41) \]

\[ N_{in3}(s) = (G_{1p} + sC_{1eq})R_{x1}R_{x2}I_{in2} + R_{x2}I_{in1} \quad (6.3.42) \]

\[ N_{vn3}(s) = V_{in3} + sC_1R_{x3}V_{in1} - (G_{1p} + sC_{1eq})R_{x2}V_{in4} + (G_{1p} + sC_{1eq})sC_2R_{x1}R_{x2}V_{in2} \quad (6.3.43) \]

\[ N_{in4}(s) = R_{vin}(I_{in3}D_n(s) - (G_{1p} + sC_{1eq})R_{x1}I_{in2} - I_{in1}) \quad (6.3.44) \]

\[ N_{vn4}(s) = R_{vin}(-(sC_1V_{in1} - (G_{1p} + sC_{1eq})V_{in4}) + (G_{2eq} + sC_{2eq})(G_{1p} + sC_{1eq})R_{x1}V_{in3} - \]

\[ sC_2(G_{1p} + sC_{1eq})R_{x1}V_{in2}) \quad (6.3.45) \]

\[ D_n(s) = (G_{1p} + sC_{1eq})(G_{2eq} + sC_{2eq})R_{x1}R_{x2} + 1 \quad (6.3.46) \]

\[ G_{2eq} = G + G_{2p}, \text{ where } G = 1/R, \]
\[ C_{1eq} = C_{1p} / C_1 \, , \, \, C_{2eq} = C_{2p} / C_2 \, , \, \text{when} \, V_{in1} = V_{in2} = 0 \] (6.3.47)

A careful investigation of (6.3.38) to (6.3.46) and Fig. 6.14 reveals that in the proposed topology parasitic conductance \( G_{2p} \) is appearing in parallel to \( G \), hence can be compensated by pre-distorting value of external passive component \( R \). Similarly the effects of parasitic capacitances \( C_{1p} \) and \( C_{2p} \) can be compensated by pre-distorting \( C_1 \) and \( C_2 \) when \( V_{in1} = V_{in2} = 0 \). To further investigate the effect of parasitics and to obtain corresponding approximate design criterion, we have taken the case of a low pass current mode filter. The current mode low pass response may be obtained as

\[ I_{out1}(s) \bigg|_n = \frac{I_{in2}}{(G_{1p} + sC_{1eq})(G_{2eq} + sC_{2eq})R_{s1}R_{s2} + 1} \] (6.3.48)

Here,

\[ \frac{\omega_0}{Q_0} \bigg|_n = \frac{G_{1p}}{C_{1eq}} + \frac{G_{2p}}{C_{2eq}} + \frac{G}{C_{2eq}} \] (6.3.49)

For \( C_1 = C_2 = C >> \) parasitic capacitances, we can write \( C_{1eq} \equiv C_{2eq} \equiv C_1 = C_2 = C \).

So, the ideal bandwidth can be written as

\[ \frac{\omega_0}{Q_0} = \frac{G}{C} \equiv \frac{G}{C_{2eq}} \] (6.3.50)

It may be observed from (6.3.49) and (6.3.50) that \( \frac{\omega_0}{Q_0} \bigg|_n \) will be approximately equal to ideal case if

\[ \frac{G}{C_{2eq}} \gg \frac{G_{1p}}{C_{1eq}} + \frac{G_{2p}}{C_{2eq}} \]

or,

\[ \frac{\omega_0}{Q_0} \gg \frac{G_{1p}}{C_{1eq}} + \frac{G_{2p}}{C_{2eq}} \]
that is, \[ C >> \left[ \frac{G_{1p} + G_{2p}}{Q_0} \right] = C_{D1} \text{(say)} \] (6.3.51)

Similarly \[ \omega_0^2 \approx \frac{1}{R_{s1} R_{s2} C_{1eq} C_{2eq}} + \frac{G_{1p} (G_{2p} + G)}{C_{1eq} C_{2eq}} \]
will be approximately equal to the ideal case if
\[ \frac{1}{R_{s1} R_{s2} C_{1eq} C_{2eq}} \gg \frac{G_{1p} (G_{2p} + G)}{C_{1eq} C_{2eq}} , \]
or,
\[ \omega_0^2 \gg \frac{G_{1p} (G_{2p} + G)}{C_{1eq} C_{2eq}} \]

or,
\[ \omega_0^2 \gg \left( \frac{G_{1p} G_{2p} + G_{1p} Q_0}{C_{1eq} C_{2eq}} \right) \]
(6.3.52)

The first term in R. H. S. of (6.3.52) can be neglected as the values of \( G_{1p} \) and \( G_{2p} \) are small and product of these terms will be a very small, so
\[ \omega_0^2 \gg \frac{G_{1p} \omega_0}{C_{1eq} Q_0} \]
or,
\[ C \gg \left[ \frac{G_{1p}}{Q_0 \omega_0} = C_{D2} \text{(say)} \right] \] (6.3.53)

Thus by choosing
\[ C = C_1 = C_2 >> \max (C_{D1}, C_{D2}) \] (6.3.54)
the effect of parasitic impedance can be practically eliminated and the filter response may approaches towards ideal response. However, the maximum frequency of operation will be limited by poles of current \( f_o \) and voltage \( f_p \) transfers which are simulated to be 61.2 MHz and 215.6 MHz respectively for the proposed circuit.
It can also be easily evaluated to show that the active and passive sensitivities of pole \( \omega_0 \) and pole \( Q_0 \) are within unity in magnitude. Thus the proposed structures can be classified as insensitive.

6.3.2.3 RESULTS

To validate the theoretical predictions, the proposed circuit of Fig. 6.13 is simulated with PSPICE using translinear current conveyor [13] and typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) [112] with supply voltages of \( \pm 2.5 \) volts. It is found that to practically eliminate the effect of parasitics of CCCII the value of \( C = C_1 = C_2 \) must satisfy (6.3.53). The values of parasitic resistances (\( R_y, R_z \)), capacitances (\( C_y, C_z \)), inductance (\( L_x \)), cut off frequencies for current and voltage transfers (\( f_u, f_\beta \)) are simulated to be \( R_y = 91 \, K\Omega, R_z = 384 \, K\Omega, C_y = 5.28 \, pF, C_z = 2.28 \, pF, L_x = 0.2 \, \mu H, f_u = 61.2 MHz \) and \( f_\beta = 215.6 MHz \) respectively for CCCII\'s bias current \( = 100 \, \mu A \). The corresponding values of \( C_{D1} \) (6.3.51) and \( C_{D2} \) (6.3.52) are calculated to be \( 21.4 \, pF \) and \( 10.71 \, pF \) respectively for bias current \( = 100 \, \mu A \). In order to show the effects of parasitics on \( f_0 \), the simulation of low pass filter under CM was carried out for \( f_0 = 1.27 \, MHz \) at \( Q_0 = 1 \). Table 6.9 shows the various values of capacitance \( C = C_1 = C_2 \) and corresponding ideal and simulated values of \( f_0 \) and plotted in Fig. 6.15. It is evident from Table 6.9 and Fig. 6.15 that when values of \( C \) are low, the simulated value of \( f_0 \) deviate strongly from ideal value and when \( C \) is about 45 times of max (\( C_{D1}, C_{D2} \)), the simulated \( f_0 \) is much closed to ideal value. Thus an output response is almost free from parasitic effect, if the value of \( C = C_1 = C_2 \) is selected about 45 times of max (\( C_{D1}, C_{D2} \)) or more.
Table 6.9

Dependence of frequency on capacitor (C) for CM filter of Fig. 6.13.

<table>
<thead>
<tr>
<th>Designed (ideal) value of $f_0$ (MHz)</th>
<th>$I_{01} = I_{02} = I_{03}$ (µA)</th>
<th>C (nF)</th>
<th>Calculated $\frac{C}{\text{Max}(C_{D1}, C_{D2})}$ (pF)</th>
<th>$\frac{C}{\text{Max}(C_{D1}, C_{D2})}$</th>
<th>$f_0$ (simulated) (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.27</td>
<td>100</td>
<td>1.0</td>
<td>21.40</td>
<td>46.70</td>
<td>1.26</td>
</tr>
<tr>
<td>1.27</td>
<td>90</td>
<td>0.9</td>
<td>20.33</td>
<td>44.26</td>
<td>1.25</td>
</tr>
<tr>
<td>1.27</td>
<td>80</td>
<td>0.8</td>
<td>19.24</td>
<td>41.58</td>
<td>1.24</td>
</tr>
<tr>
<td>1.27</td>
<td>70</td>
<td>0.7</td>
<td>18.19</td>
<td>38.48</td>
<td>1.095</td>
</tr>
<tr>
<td>1.27</td>
<td>60</td>
<td>0.6</td>
<td>16.90</td>
<td>35.50</td>
<td>0.929</td>
</tr>
<tr>
<td>1.27</td>
<td>50</td>
<td>0.5</td>
<td>15.64</td>
<td>31.96</td>
<td>0.724</td>
</tr>
<tr>
<td>1.27</td>
<td>40</td>
<td>0.4</td>
<td>14.20</td>
<td>28.16</td>
<td>0.529</td>
</tr>
<tr>
<td>1.27</td>
<td>30</td>
<td>0.3</td>
<td>12.52</td>
<td>23.96</td>
<td>0.378</td>
</tr>
<tr>
<td>1.27</td>
<td>20</td>
<td>0.2</td>
<td>9.30</td>
<td>21.50</td>
<td>0.242</td>
</tr>
<tr>
<td>1.27</td>
<td>10</td>
<td>0.1</td>
<td>7.00</td>
<td>14.28</td>
<td>0.119</td>
</tr>
<tr>
<td>1.27</td>
<td>5µA</td>
<td>0.05</td>
<td>4.50</td>
<td>11.11</td>
<td>0.0669</td>
</tr>
<tr>
<td>1.27</td>
<td>1µA</td>
<td>0.01</td>
<td>1.20</td>
<td>08.33</td>
<td>0.01448</td>
</tr>
</tbody>
</table>

Fig. 6.15 Dependence of frequency on capacitor (C) values for CM filter of Fig. 6.13.
Figure 6.16(a) shows the simulated and theoretical results for low pass, band pass and high pass responses for a pole frequency of $f_0 = 1.27$ MHz, quality factor of $Q_0 = 1$ and the component values of $C_1 = C_2 = C_3 = 1$ nF and $I_{01} = I_{02} = 100$ $\mu$A. Tunability of $Q_0$ with fixed $\omega_0$ for band pass filter is shown in Fig. 6.16(b). This is designed at $f_0 = 1.27$ MHz with $C_1 = C_2 = 1$ nF, $R_{x1} = 0.125$ k$\Omega$ (i.e. $I_{01} = 100$ $\mu$A), $R_{x2} = 0.125$ k$\Omega$ (i.e. $I_{02} = 100$ $\mu$A) and $R = 0.125$ k$\Omega$ ($Q_0 = 1$), $0.625$ k$\Omega$ ($Q_0 = 5$), $1.25$ k$\Omega$ ($Q_0 = 10$).

The simulation and theoretical results agree quite well.

![Graphs showing frequency response](image)

Fig. 6.16 (a) Frequency Response of current mode filter (— Simulated, — Theoretical)

(b) Orthogonal tunability of $Q_0$ and $\omega_0$.

The proposed circuit is also tested for dynamic range i.e. up to the level of input signal for which the output is within the permitted distortion level. The response is obtained for band pass current mode filter by applying a sinusoidal current input at $f_0 = 1.27$ MHz with $C_1 = C_2 = 1$ nF, $R = 125$ $\Omega$ and $I_{01} = I_{02} = 100$ $\mu$A. The result in Fig. 6.18 shows that for large range of input signal level, output distortion is within acceptable
limit of the order of $\text{THD} = 5\%$. It shows that the proposed circuit is also useful for even large signal.

![Graph showing variation of THD with input signal amplitude](image)

Fig. 6.18 Variation of total harmonic distortion (%THD) with input signal amplitude.

### 6.3.3 CONCLUSION

In section 6.3, we have presented MIMO mixed mode universal filter topologies that can be used in voltage mode (voltage input voltage output) or current mode (current input current output) or transimpedance mode (current input voltage output) or transadmittance mode (voltage input current output). The first proposed structure is based on second generation current conveyors (CCIs) whereas the second structure is based on multiple output current controlled conveyors (MOCCCIIs). The literature survey till date on generalized mixed mode universal filter reveals that the proposed MOCCCIII based structure uses minimum number of active components. Both the proposed structures have following attractive features: (i) realization of all generic function of universal filters i.e. LP, BP, HP, notch and all pass, (ii) orthogonal adjustability of filter parameters $\omega_0$ and $Q_0$
and $\omega_0$ and $\omega_0/Q_0$, (iii) low active and passive sensitivities of $\omega_0$ and $Q_0$, and (iv) simple resistive matching constraints. An added feature of the structure of section 6.3.2 is the availability of most of the responses in both inverted (I.) and non inverted (N.I.) forms which is useful in typical application such as hearing aid [131].