CHAPTER - 4

SINGLE-INPUT MULTIPLE-OUTPUT (SIMO) FILTERS

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


Single-input multiple-output (SIMO) configurations developed in this chapter are more versatile than SISO structures discussed in chapter 3. In SIMO topologies, more than one filter responses are simultaneously available at different nodes / branches whereas in SISO structures the desired outputs are available one at a time through a single output terminal. The author has first presented a current mode single-input three-output (SITO) filter structure and later on a voltage mode structure based on Tow Thomas approach is developed.

4.1 CURRENT MODE SIMO FILTER

Several realizations of current conveyor based current mode single-input three-output (SITO) universal filters [12], [58] – [65] are available in literature. A systematic study of these structures shows that

(i) these circuits use excessive number of active [12], [58] – [62] and passive elements [12], [58] – [61], [63], [64].

(ii) they employ ungrounded passive elements [62]– [64] which is not suitable for IC implementation [122].

(iii) the configuration [65] though uses minimum number of both active and passive components but requires additional active devices to sense currents as all the output currents (LP, HP, BP) are available on external passive components. Hence three more current conveyors will be required to implement other standard functions (notch, AP) of universal filter.
Moreover, many applications require filter tuning, in particular sophisticated techniques of signal processing demand the ability of the circuit to adapt filter characteristics dynamically. In such cases, it is desirable to vary the filter coefficients electronically. None of the above reported networks [12], [58 – [65] possess electronic tunability and thus is not adaptable in such cases. As discussed in section 2.1, the use of CCCII [13] may add electronically tunability to the filter parameters, so in recent past there has been greater emphasis on the design of current mode circuits using CCCIIs [66] – [70]. The structures [66] – [68] use 3 CCCIIs and two grounded capacitors whereas [69] uses 2 CCCIIs and two capacitors but one of the capacitors is floating. Either one or two of the outputs [66] – [69] are available on passive components. Hence one or two additional current conveyor(s) will be required to implement all the standard universal filter functions (LP, HP, BP, notch, and AP). The structures [70] enjoys high output impedance, but uses five active components and three grounded capacitors and also needs to satisfy a passive component matching for realization of notch and AP responses. Moreover all these structures [66], [67], [69], [70] except [68] use a capacitor at port x and hence are not suitable for high frequency range [64].

In this section, a new SITO universal filter structure based on MOCCCIIs is presented. The filter enjoys the following attractive features: (i) low input impedance, (ii) high output impedance; hence additional filter responses may easily be obtained, (iii) use of grounded capacitors that can absorb parasitics well, (iv) electronic and orthogonal tunability of filter parameters, (v) low active and passive sensitivities.
4.1.1 CIRCUIT DESCRIPTION

The proposed SITO filter network, illustrated in Fig. 4.1, employs three MOCCCIIs and two grounded capacitors.

![Diagram of SITO filter network](image)

Fig. 4.1 MOCCCI based current mode SITO filter.

Using the port relations of MOCCCI, as outlined in section 3.1.1, the current transfer functions may be found as

\[
\frac{I_{LP}}{I_{in}} = \frac{1}{D(s)} \quad \text{(4.1.1a)}
\]

\[
\frac{I_{BP}}{I_{in}} = \frac{sR_x C_3}{D(s)} \quad \text{(4.1.1b)}
\]

\[
\frac{I_{HP}}{I_{in}} = \frac{s^2 R_x R_y C_x C_y C_z}{D(s)} \quad \text{(4.1.1c)}
\]

where

\[
D(s) = s^2 R_x R_y C_x C_y C_z + sR_x C_z + 1 \quad \text{(4.1.2)}
\]

The coefficient of \( s \) in (4.1.2) cannot be equal to zero; hence the circuit is always stable. As seen from (4.1.1), the proposed filter produces low pass, band pass and high...
pass responses simultaneously at high impedance outputs and therefore does not require any extra hardware to get notch and all pass functions. By adding all the current outputs i.e. $I_{LP}$, $I_{BP}$ and $I_{HP}$ and treating the node thus obtained as the output node, an all pass filter can be obtained. Similarly, by adding the current outputs $I_{LP}$ and $I_{HP}$ one can obtain a notch filter. It may further be noted that no matching constraint is imposed for the realization of all pass and notch responses in contrast to refs. [64], [70]. The transfer functions represented in (4.1.1) are characterized by the following filter parameters:

\[
\omega_0 = \left( \frac{1}{R_{x2} R_{x3}} \right)^{\frac{1}{2}} \quad \omega_0 = \frac{1}{R_{x2} C_2} \quad \text{and} \quad Q_0 = \left( \frac{R_{x2} C_2}{R_{x3} C_3} \right)^{\frac{1}{2}} \quad \text{(4.1.3)}
\]

### 4.1.2 EFFECT OF NONIDEALITIES

Considering the nonidealities of the current and voltage transfers of MOCCII, as outlined in section 2.2, the transfer functions (4.1.1a) to (4.1.1c) are modified as

\[
\frac{I_{LP}}{I_{in}}_{\text{n}} = -\frac{\alpha_1 \alpha_2 \beta_2 \beta_3 R_{x3} C_3}{D_n(s)} \quad \text{(4.1.4a)}
\]

\[
\frac{I_{BP}}{I_{in}}_{\text{n}} = \frac{s \alpha_1 \beta_2 R_{x3} C_3}{D_n(s)} \quad \text{(4.1.4b)}
\]

\[
\frac{I_{HP}}{I_{in}}_{\text{n}} = \frac{s^2 R_{x2} R_{x3} C_2 C_3}{D_n(s)} \quad \text{(4.1.4c)}
\]

where

\[
D_n(s) = s^2 C_2 C_3 R_{x2} R_{x3} + s \alpha_1 \alpha_2 \beta_2 R_{x3} C_3 + \alpha_1 \alpha_2 \alpha_3 \beta_2 \beta_3 \quad \text{(4.1.5)}
\]

The various filter parameters of the realized filter transfer functions now are obtained as follows:
\[ \omega_n = \left( \frac{\alpha_2 \alpha_3 \beta_2 \beta_3}{R_{r2} R_{r3} C_2 C_3} \right)^{\frac{1}{2}}, \quad \frac{\omega_0}{Q_0} = \frac{\alpha_2 \beta_2}{R_{r2} C_2}, \quad \text{and} \quad \frac{Q_0}{\omega_0} = \left( \frac{\alpha_2 \beta_2 R_{r3} C_2}{\alpha_2 \alpha_3 \beta_2 R_{r3} C_3} \right)^{\frac{1}{2}} \] (4.1.6)

Equation (4.1.6) indicates that the filter parameters are modified due to the MOCCCII nonidealities. This variation may be accommodated by adjusting bias currents \( I_{b2} \) and \( I_{b3} \). It may be noted that voltage tracking error \( (\beta_1) \) of 1st MOCCCII has no effect on filter parameters. Equation (4.1.3) indicates that \( \omega_n \) can be adjusted by varying bias current \( I_{b3} \) without disturbing \( \omega_0/Q_0 \). The parameters \( \omega_0 \) and \( Q_0 \) are orthogonally adjustable if \( R_{r2} \) and \( R_{r3} \) are simultaneously adjusted by a common control bias current \( I_{b2} = I_{b3} = I_0 \). Equations (4.1.3) and (4.1.6) indicate that high values of Q factor can be obtained from moderate values of ratios between passive components. These ratios can be chosen as \( (R_{r2}/R_{r3}) \approx (C_2/C_3) = Q \). Hence the spread of the component values becomes of the order of \( \sqrt{Q} \). This feature of the filter related to the component spread allows the realization of high Q values more accurately as compared to the topologies where the spread of passive components becomes Q or \( Q^2 \) [123].

The active and passive sensitivity analysis of the filter parameters \( \omega_0 \) and \( Q_0 \) results in the following expressions:

\[
\begin{align*}
S_{R_{r2}}^{\omega_0} &= S_{R_{r3}}^{\omega_0} = S_{C_2}^{\omega_0} = S_{C_3}^{\omega_0} = -1/2, \quad S_{R_{r1}}^{\omega_0} = 0, \\
S_{\alpha_2}^{\omega_0} &= S_{\alpha_3}^{\omega_0} = S_{\beta_2}^{\omega_0} = S_{\beta_3}^{\omega_0} = 1/2, \quad S_{\beta_1}^{\omega_0} = 0, \\
S_{R_{r2}}^{Q_0} &= -S_{R_{r3}}^{Q_0} = S_{C_2}^{Q_0} = S_{C_3}^{Q_0} = 1/2, \quad S_{R_{r1}}^{Q_0} = 0, \\
S_{\alpha_2}^{Q_0} &= S_{\alpha_3}^{Q_0} = -S_{\beta_2}^{Q_0} = -S_{\beta_3}^{Q_0} = -1/2, \quad S_{\beta_1}^{Q_0} = 0.
\end{align*}
\]
It can be seen from the above equations that all active and passive sensitivities are low and within 1 in magnitude. Thus the circuit can be regarded as insensitive.

4.1.3 COMPARISON

The CCCII based filter structure [69] uses only two CCCIIs but realizes high pass output through the capacitor and band pass output at low impedance port x. Hence to obtain all the standard universal filter functions at high output impedance, either three or four current conveyors are required. If implemented with three conveyors then one of the conveyors has to be double output type.

Survey of literature on SITO universal current mode filter shows that four or more conveyors are necessary to implement all the standard universal filter functions (LP, HP, BP, notch, AP) except in works reported in [63], [64], and [69] where three conveyors are required as in the proposed work. Therefore, the proposed configuration is comparable with the [63], [64], and [69] in term of number of current conveyors used. A comparison with the structures [63], [64], and [69] has been summarized as follows:

(i) Requirement of passive components:

One or more passive components are ungrounded in [63], [69] which are not suitable for IC implementation [122]. All the CCIIs cannot be replaced by CCCIIs in [63], [64]. Even if one tries to replace, the minimum number of passive components cannot be reduced below three which is one more than the present work.
(ii) **Matching constraint:**

The configuration [64] needs to satisfy a matching condition to yield all pass response, in contrast to present work.

(iii) **Orthogonal and electronic control of $\omega_0$ and $Q_0$:**

Orthogonal adjustment of $\omega_0$ and $Q_0$ is not possible in [64] for all pass response. Additional active components (such as FETs) are needed for orthogonal and electronic control of $\omega_0$ and $Q_0$ in [63], [64]. In the present work, the orthogonal and electronic control do not require any additional active components but can easily be controlled by bias currents ($I_{o2}$ and $I_{o3}$) of MOCCCIIs for all the responses (LP, HP, BP, notch, AP).

(iv) **High frequency limitations:**

In contrast to the present work, all the three works [63], [64], [69] use capacitor at port which will limit the higher frequency range of operation [64].

4.1.4 **RESULTS**

The feasibility of the proposed current mode SITO universal filter structure of Fig. 4.1 has been studied with PSPICE simulations using translinear implementation of MOCCII (section 2.3) with DC supply voltage of $\pm 2.5$ V. The typical parameters of bipolar ALA array NR100N (NPN) and PR100N (PNP) [112] are used for simulation purpose. The simulated and ideal responses of low pass, band pass and high pass filters are shown in Fig. 4.2(a) for $C_1 = C_2 = 10$ nF, $I_{o1} = I_{o2} = I_{o3} = 100 \mu$A. The electronic and orthogonal control of $\omega_0$ and $Q_0$ of band pass response is shown in Fig. 4.2(b) for $Q_0 = 1$, $C_2 = C_3 = 10$ nF, $I_{o1} = I_{o2} = I_{o3}=10 \mu$A, $50 \mu$A, and $100 \mu$A.
Fig. 4.2(a) Low pass, band pass and high pass responses of the proposed filter.

--- Ideal --- Simulated

(b) Orthogonal tunability of band pass responses.

4.1.5 CONCLUSION

A new universal current mode filter has been presented. The proposed filter uses three MOCCII and two grounded capacitors. The filter has the following attractive features (i) use of only grounded capacitors makes the structure less sensitive to parasitic and easier to integrate in contrast to [63], [69], (ii) low active and passive sensitivities, (iii) independent control of $\omega_0$ without disturbing $\omega_0/Q_0$ (iv) electronic and orthogonal control of $\omega_0$ and $Q_0$, (v) realization of all standard functions of universal filter without any matching constraints (vi) a low input impedance and high output impedance thus suitable for cascading, and (vii) allows high Q with low component spread.

While writing the thesis the author has come across three more SITO universal filters [71] – [73] reported later to the present work. The work reported in [71] uses three
DOCCIIs, two resistances (one floating and one grounded) and two grounded capacitors. It uses capacitor at x port and thus has high frequency limitations; and does not provide all pass response. The structure [72] employs three MOCCIIs, four resistors (one floating and rest grounded) and two grounded capacitors. The all pass response is obtained by putting constraints on the resistance values. The configuration [73] is based on four plus type CCCIIIs and uses two floating capacitors (one out of which is at x terminal). A comparison thus reveals that the proposed SITO structure has a number of advantages over the works reported in the literature till date.

42 VOLTAGE MODE SIMO FILTER

Voltage mode biquad filters with infinite input impedance are of great interest because several such cells can be connected in cascade to implement higher order filters [74]. It is well known that one of the most attractive three op amps biquad filters is Tow Thomas biquad. Tow Thomas biquad realizations using current conveyors have been reported in the literature [74] – [76], however, none of these structures are suitable for implementation using only CCCII as an active element and hence can not be extended to electronic tunability regime in full extent.

In this section a second generation current conveyors (CCIIs) based Tow Thomas voltage mode biquad filter is developed. The circuit is completely realizable with second generation current controlled conveyors (CCCIIs) by replacing all the CCIIs and a resistance at port x by CCCIIs. The proposed circuit offers the following advantageous features: (i) realization of different filter functions from the same configuration, (ii) no requirement for component matching conditions, (iii) employment of grounded capacitors
which are ideal for integration, (iv) orthogonal and electronic control of filter parameters, (v) low sensitivity performance, (vi) electronically tunable gains via biasing current of the translinear current conveyor and (vii) employs same number of active elements but lesser number of passive elements compared to the previously reported circuits [74] – [76].

4.2.1 CIRCUIT DESCRIPTION

We have extended the op amp based Tow Thomas circuit (Fig. 4.3) to the one based on current conveyor using transformation theorem [110] discussed in section 2.4 followed by summing currents at z terminal. The various steps involved in arriving at a CCII based structure are illustrated in Fig. 4.4. The resulting circuit uses four minus type current conveyors (CCII-) and six grounded resistances and two grounded capacitors. It may be noticed from Fig. 4.4 that the purpose of BLOCK 3 is only to invert voltage $V_{LP}$, so that it can be accommodated in BLOCK 1 by reversing the polarity of 2nd CCII. Thus 4th CCII along with two resistances can be omitted without disturbing the overall functionality of the network.

![Fig. 4.3 Op amp based Tow Thomas Circuit.](image-url)
Fig. 4.4 Various steps involved in generating CCII based Tow Thomas Filter.

The resulting network is shown in Fig. 4.5 that contains only three active elements but uses fewer passive components than original op amp based Tow Thomas circuit (Fig. 4.3).

Fig. 4.5 CCII based Tow Thomas filter.

The circuit is also realized with translinear current conveyor by replacing each current conveyor and series resistance at terminal x with a translinear current conveyor (CCII). The Tow Thomas circuit so obtained is shown in Fig. 4.6 and employs only one external grounded resistance and two grounded capacitors as passive components.
Fig. 4.6 CCCII based Tow Thomas filter.

The voltage transfer functions of the circuit shown in Fig. 4.6 can be obtained as

\[
\frac{V_{BP}}{V_{in}} = \frac{1}{R_1 R_2 C_1 C_2 D(s)}
\]

(4.2.1a)

\[
\frac{V_{LP}}{V_{in}} = \frac{s}{R_1 C_1} \frac{1}{D(s)}
\]

(4.2.1b)

where

\[
D(s) = s^2 + s/(R_1 C_1) + 1/(R_2 R_3 C_1 C_2)
\]

(4.2.2)

The coefficient of \(s\) in denominator cannot be equal to zero; hence the circuit is always stable. It can be seen from (4.2.1a) and (4.2.1b) that low pass DC gain and band pass gain at \(\omega_0\) are given as

\[
G_{LP} = \frac{R_2}{R_1}
\]

(4.2.3)

\[
G_{BP} = -\frac{R_1}{R_1}
\]

(4.2.4)

The transfer functions (4.2.1a) and (4.2.1b) are characterized by following filter parameters
\[
\omega_0 = \left( \frac{1}{R_{12}R_{13}C_1C_2} \right)^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{R_1C_1}, \quad \text{and} \quad Q_0 = R_1 \left( \frac{C_1}{R_{12}R_{13}C_2} \right)^{1/2} \quad (4.2.5)
\]

Hence we see that the parameter \( \omega_0 \) can be controlled electronically by adjusting bias current \( I_{02} \) and/or \( I_{03} \) without disturbing the parameter \( \omega_0 / Q_0 \). The parameter \( \omega_0 / Q_0 \) can be controlled by adjusting \( R_1 \) without disturbing the parameter \( \omega_0 \). Furthermore, after adjusting \( \omega_0 \) by \( I_{02} \) and/or \( I_{03} \), the filter \( Q_0 \) can independently be controlled via grounded resistance \( R_1 \) while keeping \( \omega_0 \) fixed. It may also be noted that the gain of band pass and low pass filters can be adjusted by bias current \( I_{01} \) independent of \( \omega_0 \), \( Q_0 \) and \( \omega_0 / Q_0 \). This filter can also be used as channel select filter (presented in section 3.3) and has an additional feature of independent tunability of gain. Thus, the structure can be viewed as filter with embedded gain amplifier. This feature facilitates the designing of base band chain by merging gain with filtering [124].

### 4.2.2 EFFECT OF NONIDEALITIES

Considering the nonidealities in the current and voltage transfers of current conveyor outlined in section 2.2, the transfer functions represented by (4.2.1a) and (4.2.1b) are modified as

\[
\frac{V_{BP}}{V_{in}} \bigg|_n = \frac{\alpha_1 \alpha_2 \beta_1 \beta_2}{R_{12}R_{13}C_1C_2} \frac{1}{D_n(s)} \quad (4.2.6a)
\]

\[
\frac{V_{BP}}{V_{in}} \bigg|_n = -s \frac{\alpha_1 \beta_1}{R_{12}C_1} \frac{1}{D_n(s)} \quad (4.2.6b)
\]

where
\[ D_n(s) = s^2 + \frac{1}{R_1C_1} s + \frac{\alpha_1\alpha_3\beta_1\beta_2}{R_{x_2}R_{x_3}C_1C_2} \]  

(4.2.7)

The gains of the LP and BP functions are modified to

\[ G_{LP}|_n = \frac{\alpha_1R_{x_2}}{\alpha_3R_{x_1}} \]  

(4.2.8)
\[ G_{BP}|_n = -\frac{\alpha_1\beta_1R_1}{R_{x_1}} \]  

(4.2.9)

The transfer functions (4.2.6a) and (4.2.6b) are characterized by

\[ \omega_n = \left( \frac{\alpha_1\alpha_3\beta_1\beta_2}{R_{x_2}R_{x_3}C_1C_2} \right)^{1/2}, \quad \omega_n = \frac{1}{R_1C_1}, \text{ and} \]
\[ Q_n = R_1 \left( \frac{\alpha_1\alpha_3\beta_1\beta_2C_1}{R_{x_2}R_{x_3}C_2} \right)^{1/2} \]  

(4.2.10)

Equation (4.2.10) indicates that bandwidth remains unaltered in the presence of non-idealities. The values of \( \omega_n, Q_n \) and gains get changed but can be adjusted using bias currents \( I_{02} \) and \( I_{03} \) for \( \omega_n \) and \( Q_n \), and \( I_{01} \) and \( I_{02} \) for gains. The sensitivities of various filter parameters with respect to active and passive elements may be expressed as

\[ S_{\alpha_2} = S_{\beta_2} = S_{\alpha_3} = S_{\beta_3} = \frac{1}{2}, \quad S_{\alpha_1} = S_{\beta_1} = 0, \]
\[ S_{R_2}^{\alpha} = S_{R_3}^{\alpha} = S_{C_1}^{\alpha} = S_{C_2}^{\alpha} = -S_{I_{02}}^{\alpha} = -S_{I_{03}}^{\alpha} = -\frac{1}{2}, \quad S_{R_1}^{\alpha} = S_{R_{11}}^{\alpha} = -S_{I_{01}}^{\alpha} = 0, \]
\[ S_{\alpha_2} = S_{\alpha_3} = S_{\alpha_1} = S_{\beta_3} = \frac{1}{2}, \quad S_{\alpha_1} = S_{\alpha_1} = S_{\alpha_1} = S_{\beta_1} = 0, \]
\[ S_{R_2}^{Q_0} = S_{R_3}^{Q_0} = S_{C_1}^{Q_0} = S_{C_2}^{Q_0} = -S_{I_{02}}^{Q_0} = -S_{I_{03}}^{Q_0} = -\frac{1}{2}, \quad S_{R_1}^{Q_0} = S_{R_{11}}^{Q_0} = -S_{I_{01}}^{Q_0} = 0. \]
It may be noted that sensitivities of the filter parameters $\omega_0$, $Q_0$ and $\omega_0/Q_0$ are low and magnitude varies within 0 to 1.

4.2.3 RESULTS

To confirm the practical validity of the proposed biquad, the translinear current conveyor based Tow Thomas biquad is simulated with PSPICE using the implementation proposed by Abuelma'atti et. al. [125]. The transistor models PR100N (PNP) and NR100N (NPN) of bipolar ALA arrays [112] are taken for bipolar transistors. The simulated responses of the low pass and band pass filters are shown in Fig. 4.7 for $C_1 = C_2 = 1$ nF, $I_{o1} = I_{o2} = I_{o3} = 10$ $\mu$A and $R_1 = 1250$ $\Omega$. Figures 4.8 (a) and (b) show the tunability of band pass and low pass gain for constant $Q_0$ (= 1) by varying bias current $I_{o1}$ and selecting $C_1 = C_2 = 1$ nF, $I_{o2} = I_{o3} = 10$ $\mu$A, $R_1 = 1250$ $\Omega$.

![Graph showing frequency response](image)

Fig. 4.7 Frequency response of the voltage mode Tow Thomas circuit.
Fig. 4.8 (a) Simulated response for band pass filter gain tunability.
(b) Simulated response for low pass filter gain tunability.
\[ I_{\text{io1}} = 10 \mu A (\bigcirc), 25 \mu A (\ast), 50 \mu A (\checkmark), 75 \mu A (\triangle), 100 \mu A (\circ). \]

4.2.4 CONCLUSION

In this section a new Tow Thomas filter has been presented. The proposed filter utilizes three CCCIIIs, one grounded resistor and two grounded capacitors. The filter has several attractive features: (i) low active and passive sensitivities (iii) independent control of \( \omega_0 \) without disturbing \( \omega_0/Q_0 \) (iii) high input impedance (iv) independent control of \( Q_0 \) without disturbing \( \omega_0 \) and (v) electronically tunability of gains, pole \( \omega_0 \) and pole \( Q_0 \).