portable transceiver to exhibit tunable characteristics [117] with various carrier frequencies and bandwidths. Since direct conversion receivers (DCR) require only low pass filtering in the analog base band, this architecture has been applied for multi mode transceivers [118]. In the receiver, an accurate channel select filter is required for adjacent channel rejection [119]. The integrated filter for multi standard receiver should also exhibit fine and wide electronic tunability range covering different standards. While frequency operation range of switch capacitor (SC) filters is limited, operational transconductance amplifier (OTA) filters can operate on higher frequencies but their linearity and dynamic range are low. The current conveyor based filter structures [1], [43] - [45], [91] – [98], [120] which are suitable for the discussed application have the following limitations:

(i) Although the structures [1], [43] – [45] require single current conveyor (CCII) (minus type), but need floating passive components and are not adaptable to electronic tunable environment.

(ii) Use excessive number of current conveyors [91], [92], [94], [98].

(iii) Employ both types (plus and minus) of current conveyors (CCII) [91] – [93], [95], [96].

(iv) Requires complex matching constraints [97].

(v) Not completely extendible from CCII to current controlled conveyor (CCCII) based structure [91], [92], [94], [95], [98], [120] which is suitable for electronic adjustability without including any passive resistor.
Applications where power consumption and adaptation in integrated circuit environment is important, the prime concern is number of active elements. Moreover, since the use of only plus type current conveyors [46] simplify the circuit configuration, the structure that requires minimum number of only plus type current conveyors having electronic tunability is preferred. In view of above, an electronically tunable low pass filter that can be used as channel select filter and is adaptable to support multi standard receiver is presented. The proposed filter employs two plus type current controlled conveyors and two grounded capacitors and enjoys independent electronic control of filter parameters. The proposed filter exhibits both fine and wide tuning properties which are used, respectively, to provide precise frequency characteristics and for switching between different communication standards. The filter accommodates the following wireless standards: PDC, IS-54, GSM, IS-95 and WCDMA with bandwidths of 13 kHz, 15 kHz, 100 kHz, 630 kHz and 2.1 MHz respectively.

3.2.2.1 CIRCUIT DESCRIPTION

The proposed filter is shown in Fig. 3.15. Routine analysis yields the voltage transfer functions as

\[
\frac{V_{out}(s)}{V_{in}(s)} = \frac{1 + (R_{x1} - R_{x2})sC_1}{D(s)}
\]  

(3.2.19)

where

\[
D(s) = R_{x1}R_{x2}C_1C_2s^2 + sR_{x1}C_1 + 1
\]  

(3.2.20)
Equation (3.2.19) shows that low pass response can be obtained by setting $R_{x1} = R_0$. The pole $\omega_0$, band width $\omega_b/Q_0$ and quality factor $Q_0$ are given by

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

(3.2.21)

### 3.2.2 EFFECT OF NONIDEALITIES

Considering the nonidealities outlined in section 2.2, the transfer function (3.2.19) modifies to

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{\alpha_1\alpha_2\beta_1 + (R_{x1} - \alpha_1R_{x2})sC_1}{D_n(s)}$$

(3.2.22)

where

$$D_n(s) = R_{x1}R_{x2}C_1C_2s^2 + sR_{x1}C_1 + \alpha_1\alpha_2\beta_1$$

Thus the nonidealities in the current and voltage transfers of CCCII modify the various filter parameters given in (3.2.21) to

$$\omega_0|_n = \left( \frac{\alpha_1\alpha_2\beta_1}{R_{x1}R_{x2}C_1C_2} \right)^{1/2}, \quad \omega_b|_n = \frac{1}{R_{x2}C_2}, \quad \text{and} \quad Q_0|_n = \left( \frac{\alpha_1\alpha_2\beta_1R_{x2}C_2}{R_{x1}C_1} \right)^{1/2}$$

(3.2.23)
From (3.2.23) it is clear that the values of the filter parameters change in the presence of nonidealities. These changes can be easily accommodated by adjusting bias current. Further, it may be noted that voltage-tracking error of 2\textsuperscript{nd} CCCII and thus $\beta_2$ has no effect on filter parameters. Equations (3.2.21) and (3.2.23) also reveal that pole $\omega_0$ i.e. the -3 dB frequency of the proposed low pass filter can be varied without disturbing quality factor $Q_0$ by simultaneous adjustment of $R_{x1}$ (or $I_{01}$) and $R_{x2}$ (or $I_{02}$).

The active and passive sensitivities of various filter parameters are calculated as

$$S_{R_{r1}} = S_{R_{r2}} = S_{C_1} = S_{C_2} = -\frac{1}{2}, \quad S_{\alpha_1} = S_{\alpha_2} = \frac{1}{2}, \quad S_{\beta_1} = S_{\beta_2} = 0,$$

$$S_{C_{r1}} = S_{C_{r2}} = S_{\alpha_0} = S_{\beta_0} = 0, \quad S_{\alpha_{r1}} = S_{\alpha_{r2}} = S_{\alpha_{r0}} = S_{\beta_{r1}} = S_{\beta_{r2}} = 0,$$

$$S_{Q_{r1}} = S_{Q_{r2}} = S_{Q_0} = S_{C_{r1}} = S_{C_{r2}} = -\frac{1}{2}, \quad S_{\alpha_{r1}} = S_{\alpha_{r2}} = S_{\alpha_0} = S_{\beta_{r1}} = S_{\beta_{r2}} = 0.$$

It is evident that all active and passive sensitivities of pole $\omega_0$, quality factor $Q_0$ and $\omega_0/Q_0$ (band width) are low and within unity in magnitude. It may further be noted that band width does not alter in presence of nonidealities of current and voltage transfer of CCCII. Thus the proposed structure can be classified as insensitive.

### 3.2.2.3 ADVANTAGES OF THE PROPOSED STRUCTURE

It is now worthwhile to compare the proposed circuit with available plus type voltage mode low pass CC based filter structures [97], [98], and [120]. In this context we find that

(i) in comparison with circuit of [97], the proposed circuit

(a) needs one passive component less in count,
(b) is more convenient for the orthogonal control of \( \omega_0 \) and \( Q_0 \).

(i) although the input impedance in the circuit of [98] is high and requires no matching constraint of passive elements, the present circuit (a) requires less numbers of active components,

(b) is completely realized with CCCII+ and hence \( \omega_0, Q_0 \), and \( \omega_0/Q_0 \) are electronically tunable with bias currents (\( I_{o1} \) and \( I_{o2} \)) of CCCII+s. On the other hand, the circuit of [98] is not completely realizable with CCCII+ and hence electronic tunability is absent.

(iii) although the configuration of [120] utilizes two plus type CCII but needs one floating resistance and capacitance each and is not completely realizable with CCCIIIs.

322.4 RESULTS

The workability of the proposed configuration has been confirmed by PSPICE simulation. The proposed structure is simulated using typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) [112] using the circuit given in Fig. 2.8(a) in section 2.3 with supply voltage of \( \pm 2.5 \) V. Figure 3.16 shows the simulation results for the values of \( C_1 = C_2 = 1 \) nF and \( I_{o1} = I_{o2} = 1.02 \mu A, 1.17 \mu A, 7.85 \mu A, 49.45 \mu A \) and 164.85 \( \mu A \) which corresponds respectively to wireless standards PDC, IS-54, GSM, IS-95 and WCDMA. The simulation results show good agreement with theoretical calculations.

The filter is also simulated to test the linearity. Figure 3.17 shows the spectrum of the filter excited by two sinusoidal inputs at 2.1 MHz and 2 MHz for the bias currents of
18.85 μA and C1 = C2 = 1 nF. It is evident from the spectrum that the intermodulated (or mixed) signal is insignificant in comparison to the signal at 2 MHz and 2.1 MHz. This result reveals that the filter is highly linear.

Fig. 3.16 Simulation results showing wireless standard selection.

Fig. 3.17 Output spectrum of filter under two tone test.

125 CONCLUSION

A new voltage mode low pass filter using CCCII+ has been presented. The circuit uses two grounded capacitors and two CCCII+s; and does not require any resistor for the
realization. The simulation results verify the theory. The salient features of the proposed

circuit are: (i) employs only plus type of CCCIIIs and grounded capacitors which make it

ideal for integration viewpoint (ii) low active and passive sensitivities (iii) ω0 can be

varied without disturbing Q0 by bias current of CCCIIIs (iv) ω0, Q0 and ω0/Q0 are

electronically tunable via bias currents of CCCIIIs (v) low output impedance (vi) highly

linear (vii) total hardware requirement is minimum compared to reported VM filters

having features discussed above till date.

3.3 TOW THOMAS FILTER

In the last two sections we have discussed filter structures that give a single

function as output. Here realization of a biquadratic filter, that is, one in which the

numerator as well as denominator may be second order polynomial is presented. Such a

filter is capable of realizing low pass, band pass, high pass, notch and all pass responses.

Already some current conveyor based voltage mode single-input single-output (SISO)

universal biquadratic filter realizations of various complexity and features have been

reported in the literature [46] – [50]. Tomazou and Lidgey [47] proposed an active filter

configuration that can realize all second order functions namely low pass, band pass, high

pass, notch and all pass but requires seven current conveyors (both plus and minus type),

an operational amplifier buffer, two grounded capacitors and eight grounded resistors.

Tek and Anday [48] presented a configuration which can realize notch, high pass, band

pass and low pass biquadratic functions employing four current conveyors (both plus and

minus type), three grounded capacitors and five grounded resistors. The configuration

[49] implements low pass, band pass, high pass and notch biquadratic functions and
employs four plus type current conveyors, three grounded capacitors and five grounded resistors. The filter configurations [46], [47] are universal and employ same number of active and passive components. The former one [46] uses only plus type current conveyors and does not require a change in type of active component to realize all pass function [47]. The filter configuration [46] is based on well known state variable filter design namely, Tow Thomas filter employing six current conveyors (both plus and minus type), two grounded capacitors and eight grounded resistors.

In this section, a Tow Thomas biquad configuration based on CCCII- developed from conventional Tow Thomas biquad circuit [23], [121] has been presented. The circuit uses six minus type current controlled conveyors (CCCII-s), two grounded capacitors and two grounded resistors.

3.23.1 CIRCUIT DESCRIPTION

The Tow Thomas filter realizing biquadratic functions using feed forward method [23] is shown in Fig 3.18. The CCII based configuration is generated from conventional Tow Thomas biquad circuit [23], [121] using the methodology outlined in section 2.4 and the currents are summed up at z terminal instead of x terminal. The summation at z terminal leads to realization that has resistances at x terminal and facilitates CCCII based realization. The various steps involved are shown in Fig. 3.19. The complete realization of Tow Thomas filter is shown in Fig. 3.20.
Fig. 3.18  Op amp based Tow Thomas biquadratic filter.

Fig. 3.19  Various steps for generating CCII based structure.

Fig. 3.20  CCCII- based Tow Thomas filter.
The transfer function of the proposed network of Fig. 3.20 can be expressed as

$$\frac{V_{out}}{V_{in}} = \frac{R_8}{R_{x5}} \frac{N(s)}{D(s)}$$  \hspace{1cm} (3.2.24)

where

$$N(s) = s^2 + \left( \frac{1}{R_1 C_1} - \frac{R_{x5}}{R_{x1} R_{x2} C_1} \right) s + \frac{R_{x5}}{R_{x2} R_{x4} R_{x5} C_1 C_2} ,$$

$$D(s) = s^2 + \frac{1}{R_1 C_1} s + \frac{R_8}{R_{x2} R_{x3} R_{x4} C_1 C_2}$$  \hspace{1cm} (3.2.25)

The transfer function is characterized by

$$\omega_0 = \left( \frac{R_8}{R_{x2} R_{x1} R_{x4} C_1 C_2} \right)^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{R_1 C_1} , \quad \text{and} \quad Q_0 = R_1 \left( \frac{R_{x5} C_1}{R_{x2} R_{x3} R_{x4} C_2} \right)^{1/2}$$  \hspace{1cm} (3.2.26)

From (3.2.25) we can see that specializations in the numerator results in the following filter functions:

(i) low pass filter: $R_{x1}, R_{x5}$ open circuit,

(ii) band pass filter: $R_{x5}, R_{x6}$ open circuit,

(iii) high pass filter: $R_{x6}$ open circuit, $R_1 R_{x5} = R_{x1} R_{x2},$

(iv) notch filter: $R_1 R_{x5} = R_{x1} R_{x2},$ and

(v) all pass filter: $R_1 R_{x6} R_8 = 2R_{x1} R_{x2} R_{x3}.$

It may be noted that $\omega_0$ and $\omega_0/Q_0$ (band width) can be orthogonally and electronically adjusted by bias currents $I_{02}, I_{03}$ or $I_{04}$ for low pass and band pass functions, and $I_{05}$ or $I_{04}$ for high pass, notch and all pass functions. The parameter $Q_0$ can be adjusted independently with grounded resistance $R_1.$ Furthermore, the gain of low pass, band pass, high pass filters, notch and all pass filters can be adjusted independently by
urying bias currents $I_{03}$ and $I_{06}$; $I_{01}$ and $I_{02}$; and adjusting $I_{01}$, $I_{02}$ and $I_{05}$ simultaneously for remaining three filter functions.

The passive sensitivity analysis of various parameters are given as

$$S_{R_{2}}^{o} = S_{R_{1}}^{o} = S_{R_{4}}^{o} = S_{C_{2}}^{o} = S_{C_{1}}^{o} = S_{R_{3}}^{o} = S_{R_{5}}^{o} = S_{R_{6}}^{o} = -S_{R_{4}}^{o} = -S_{R_{3}}^{o} = -S_{R_{5}}^{o} = -S_{R_{6}}^{o} = -S_{R_{1}}^{o} = -1,$$

$$S_{R_{2}}^{o/Q_{0}} = S_{R_{4}}^{o/Q_{0}} = S_{R_{3}}^{o/Q_{0}} = S_{R_{5}}^{o/Q_{0}} = S_{R_{6}}^{o/Q_{0}} = S_{C_{2}}^{o/Q_{0}} = S_{C_{1}}^{o/Q_{0}} = -1,$$

Thus all the passive sensitivities are low and within 1 in magnitude.

### 3.2.3.2 EFFECT OF NONIDEALITIES

As outlined in section 2.2 and 2.3, the frequency behavior of the circuit may deviate from ideal one due to nonidealities. Considering the first group of nonidealities, the transfer function represented in (3.2.24) for the circuit of Fig. 3.23 may be modified as:

$$\frac{V_{out}}{V_{in}}\bigg|_{n} = -\frac{\alpha_{2-}(s)\beta_{2}(s)R_{6}N_{n}(s)}{R_{x3}D_{n}(s)}$$  \hspace{1cm} (3.2.27)

where

$$N_{n}(s) = s^{2} + \left(\frac{1}{R_{1}C_{1}} - \frac{\alpha_{2-}(s)\alpha_{2-}(s)\beta_{1}(s)\beta_{2}(s)R_{x5}}{\alpha_{5-}(s)\beta_{3}(s)R_{x3}R_{x2}C_{1}}\right)s$$

$$+ \frac{\alpha_{2-}(s)\alpha_{3-}(s)\beta_{3}(s)\beta_{4}(s)R_{x5}}{\alpha_{5-}(s)\beta_{3}(s)R_{x3}R_{x4}C_{1}C_{2}}$$  \hspace{1cm} (3.2.28)

and

$$D_{n}(s) = s^{2} + \frac{1}{R_{1}C_{1}} + \frac{\alpha_{2-}(s)\alpha_{3-}(s)\alpha_{4-}(s)\beta_{2}(s)\beta_{3}(s)\beta_{4}(s)R_{6}}{\alpha_{5-}(s)\beta_{3}(s)R_{x3}R_{x4}C_{1}C_{2}}$$  \hspace{1cm} (3.2.29)
Equations (3.2.27) to (3.2.29) clearly indicate that the pole frequencies of voltage and current transfer functions of CCCIIIs affect the overall filter response. The effect can, however, be ignored if the operating frequencies are chosen sufficiently smaller than the voltage and current transfer pole frequencies of CCCII-.

Taking the second group of nonidealities into account, the circuit given in Fig. 3.20 modifies to Fig. 3.21 where

\[
C_{\text{seq}} = C_1 \parallel C_{z1} \parallel C_{z2} \parallel C_{z3} \parallel C_{z4} \parallel C_{z5}, \quad C_{2\text{eq}} = C_2 \parallel C_{z2} \parallel C_{y4} \parallel C_{z6}, \quad C_{8p} = C_{z3} \parallel C_{y3} \parallel C_{z5},
\]

\[
R_{\text{seq}} = R_1 \parallel R_{z1} \parallel R_{z2} \parallel R_{z4} \parallel R_{z5}, \quad R_{2p} = R_{z3} \parallel R_{y4} \parallel R_{z6}, \quad R_{8eq} = R_8 \parallel R_{z2} \parallel R_{y3} \parallel R_{z5}.
\]

The inductance is ignored in Fig. 3.21 as it affects the frequency response only at very high frequencies.

Fig. 3.21 Proposed Tow Thomas filter structure including parasitics.

Considering the nonidealities outlined above the transfer function of circuit of Fig. 3.21 can be obtained as

\[
\left. \frac{V_{\text{out}}}{V_{\text{in}}} \right|_n = \frac{R_{\text{seq}}}{R_{x5}} \frac{as^2 + bs + d}{ms^2 + ns + p}
\]

(3.2.30)

where

\[
a = 1
\]

(3.2.31)
\[
b = \frac{1}{R_{1eq} C_{1eq}} + \frac{1}{R_{2p} C_{2eq}} - \frac{R_{5s}}{R_{x1} R_{x2} C_{1eq}} \tag{3.2.32}
\]
\[
d = \frac{R_{5s}}{R_{x2} R_{x4} R_{x6} C_{1eq} C_{2eq}} + \frac{1}{R_{l_{eq}} R_{2p} C_{1eq} C_{2eq}} - \frac{R_{5s}}{R_{l_{eq}} R_{2p} R_{x2} C_{1eq} C_{2eq}} \tag{3.2.33}
\]
\[
m = 1 + s \left( \frac{R_{8eq} C_{8p}}{R_{l_{eq}} C_{1eq}} + \frac{R_{8eq} C_{8p}}{R_{2p} C_{2eq}} \right) \tag{3.2.34}
\]
\[
n = \frac{1}{R_{l_{eq}} C_{1eq}} + \frac{1}{R_{2p} C_{2eq}} - \frac{R_{8eq} C_{8p}}{R_{l_{eq}} R_{2p} C_{1eq} C_{2eq}} \tag{3.2.35}
\]
\[
p = \frac{R_{8eq}}{R_{x2} R_{x4} R_{x6} C_{1eq} C_{2eq}} + \frac{1}{R_{l_{eq}} R_{2p} C_{1eq} C_{2eq}} \tag{3.2.36}
\]

It may be seen from (3.2.31) to (3.2.36) (except (3.2.31) for \( a \)) that the filter parameters are influenced by the parasitic elements, the ideal values of the coefficients \( d, m, n \) and \( p \) are limited to only first term; and the coefficient \( b \) is limited to first and third term. We have taken a case of low pass filter to investigate the effect of parasitics on the filter parameters and to obtain corresponding approximate design criterion. As the value of \( R_{2p} \) is very large, the low pass response may be obtained as

\[
\begin{align*}
V_{out} &= -\frac{R_{8eq}}{R_{x5}} \frac{as^2 + bs + d}{ms^2 + ns + p} = -\frac{R_{8eq}}{R_{x2} R_{x4} R_{x6} C_{1eq} C_{2eq}} \frac{1}{ms^2 + ns + p} \tag{3.2.37}
\end{align*}
\]

The coefficient \( m \) is the only coefficient that has frequency dependence. Assuming the ratios \( C_{8p} / C_{1eq} \) and \( C_{8p} / C_{2eq} \) as well as parasitic conductances as very small, we can proceed as follows:

\[
m = 1 + s R_{8eq} C_{8p} = \frac{1}{\omega_1} (s + \omega_1) \tag{3.2.38}
\]
where, corner frequency \( \omega_1 = \frac{1}{R_{seq} C_{sp}} \)  

(3.2.39)

For near ideal operation at high frequencies, the frequency of operation should be smaller than \( \omega_1 \).

Here

\[
\frac{\omega_0}{Q_0} \bigg|_n = \frac{1}{R_{1eq} C_{1eq}} + \frac{1}{R_{2p} C_{2eq}} \tag{3.2.40}
\]

For \( C_1 = C_2 = C \gg \) parasitic capacitances and \( R_1 \ll \) parasitic resistances at y and z terminals of CCCII-, we can write \( C_{1eq} \equiv C_{2eq} \equiv C_1 = C_2 = C \) and \( R_{1eq} = R_1 \). So, the ideal bandwidth of the transfer function (3.2.24) can be written as

\[
\frac{\omega_0}{Q_0} = \frac{1}{R_1 C} \equiv \frac{1}{R_{1eq} C_{1eq}} \tag{3.2.41}
\]

We find from (3.2.40) and (3.2.41) that \( \frac{\omega_0}{Q_0} \bigg|_n \) will be approximately equal to ideal case if we choose

\[
\frac{\omega_0}{Q_0} \equiv \frac{1}{R_{1eq} C_{1eq}} \gg \frac{1}{R_{2p} C_{2eq}} \]

or

\[
C \gg \left[ \frac{1}{R_{2p}} \frac{Q_0}{\omega_0} = C_{DI} \text{(say)} \right] \tag{3.2.42}
\]

Similarly

\[
\frac{\omega_0^2}{Q_0^2} \bigg|_n = \frac{R_{seq}}{R_{s1} R_{s2} R_{s4} C_{1eq} C_{2eq}} + \frac{1}{R_{1eq} R_{2p} C_{1eq} C_{2eq}}
\]

will be approximately equal to the ideal case if we choose
\[ a_0^2 \equiv \frac{R_{s_{eq}}}{R_{s_2}R_{s_3}R_{s_4}C_{1_{eq}}C_{2_{eq}}} \gg \frac{1}{R_{2_p}C_{1_{eq}}\left(\frac{1}{R_{1_{eq}}C_{2_{eq}}}\right)} \]

\[ a_0^2 \gg \frac{1}{R_{2_p}C_{2_{eq}}\left(\frac{\alpha_0}{Q_0}\right)} \]

\[ C \gg \left[\frac{1}{R_{2_p}}\frac{1}{\alpha_0 Q_0} = C_{D_2} \text{(say)}\right] \quad (3.2.43) \]

Thus by choosing

\[ C = C_1 = C_2 \gg \max (C_{D_1}, C_{D_2}) \quad (3.2.44) \]

the effect of parasitic impedances can be practically eliminated and hence filter may approach towards ideal response. However, the maximum frequency of operation will be limited by poles of current \((f_a)\) and voltage \((f_b)\) transfers.

### 3.2.3 RESULTS

To confirm the practical validity, the proposed biquad, is simulated with PSPICE using the implementation of CCCII- [125] given in Fig. 2.8(b). The transistor models PR100N and NR100N of bipolar ALA arrays [112] are used. The gain and phase responses for notch filter is shown in Fig. 3.22 for \(I_{01} = I_{02} = I_{03} = I_{04} = I_{05} = I_{06} = 10 \, \mu A\), \(R_1 = R_2 = 1250 \, \Omega\), \(C_1 = C_2 = 10 \, nF\). The circuit is also simulated for low pass, high pass, band pass and all pass filters responses. The simulated results show well agreement with theoretical analysis.