CHAPTER - 3

SINGLE-INPUT SINGLE-OUTPUT (SISO) FILTERS

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


This chapter deals with the development of various second generation current conveyor/second generation current controlled conveyor based filter structures. The background for current and voltage mode single-input single-output (SISO) filters have been specifically discussed. A generalized multiple loop feedback current mode single-input single-output biquadratic filter structure has been introduced that can generates many biquad structures. Voltage mode filter structures have been proposed for all pass filters, channel select filter and Tow Thomas biquad.

3.1 CURRENT MODE SISO FILTER

A wide variety of single-input single-output CM current conveyor based topologies are available in the literature [26] – [34]. An extensive study of these biquadratic CM current conveyor based single-input single-output (SISO) structures reported till date, exhibits the following features:

(i) use of large number of current conveyors typically six to ten [26] – [29],

(ii) use of floating passive components that are more affected by parasitics and poses restriction in integration [26], [27], [30] – [34],

(iii) less versatile as modification in the biquad topology is required for realization of multiple filter function [26], [30] – [34],

(iv) not electronically tunable [26], [29] – [34]. The electronic tunability may be required in many applications, in particular sophisticated techniques of signal processing demand the ability of the circuit to adjust filter characteristics dynamically,
(v) use of different types of current conveyors for achieving electronic tunability, such as variable gain CCIIIs and simple CCIIIs [27], [28]. This is not considered good for integrated circuit layout viewpoint (regularity and modularity).

It thus reveals that possibly there is no current mode SISO biquadratic structure that can simultaneously possess the following advantageous features: (i) use of less current conveyors (ii) uses of all grounded passive components (iii) versatility (iv) electronic tunability of filter parameters and (v) use of same types of current conveyors. The present work is an attempt to fill this void.

The work presented in this section is an extension of the approach for MO-OTA based current mode biquad generation [111] to CCCII based current mode biquads. The approach is based on multiple loop feedback and facilitates in producing new filter structures. Therefore, the work basically focuses on presenting a generalized method to synthesize current mode MOCCII – C filters. The structure uses three MOCCIIIs and two grounded capacitors. The performance of the proposed structure has been confirmed by PSPICE simulation.

3.1.1 CIRCUIT DESCRIPTION

The generalized multiple loop feedback MOCCII based biquadratic filter model with all capacitors grounded is shown in Fig. 3.1. The MOCCII (three output structure) is characterized by following port relationships:

\[ v_x = v_y + i_x | R_a (j \omega) |, \quad i_y = 0, \quad i_z = -i_x, \]  

(3.1.1a)
\[ i_{i,2} = k_{i1}i_x, \quad i_{i,12} = k_{i2}i_x \]  
\[ (3.1.1b) \]

where \( k_{i1} \) and \( k_{i2} = \pm 1 \) for \( i = 1, 2, 3 \); \( R_{X1} = V_T/2I_{01} \) is the bias current dependent resistance; \( V_T \) is the thermal voltage and \( I_{01} \) is bias current of \( i^{th} \) CCCII [13]. To get MOCCCII, current mirrors are inserted at the output of CCCII as discussed in section 2.3.

![Feedback Network Diagram](image)

Fig. 3.1 Proposed multiple loop feedback current mode biquadratic filter structure.

The proposed filter uses notation \( p \) and \( q \) instead of \( x \) and \( y \) as \( p(q) \) can represent \( x(y) \) or \( y(x) \). The current flowing through \( k_{i0}Z, k_{i1}Z, \) and \( k_{i2}Z \) are used for feedback and feed-forward and can flow into or out of CCCII depending upon required filtering function respectively.

3.1.1 SYSTEM EQUATIONS AND TRANSFER FUNCTION

The multiple loop feedback system shown in Fig. 3.1 can be divided into two parts: feedback network and feed-forward network. The feedback network can be described as

\[ I_p = \sum_{j=1}^{i} f_{ij}I_{ij} \quad i = 1, 2, 3 \]
\[ (3.1.2) \]
where $f_{ij}$ is current feedback coefficient represents a connection from output of $j^{th}$ CCCII to the input of $i^{th}$ CCCII. The coefficient $f_{ij}$ can have zero or unity value depending upon whether there is an open circuit or direct feedback connection between $i^{th}$ and $j^{th}$ CCCII. The coefficient $f_{ij} = 0$ specifies nonexistence of feedback whereas $f_{ij}=1$ represents feedback. Also, $I_{in}$ is the feedback current at input of $i^{th}$ CCCII and $I_{sxj}$ is the output current from $j^{th}$ CCCII.

Equation (3.1.2) can also be written as

$$I_f = FI_x$$  \hspace{1cm} (3.1.3)$$

where the feedback current matrix, the output current matrix and the feedback coefficient matrix are represented by $I_f = [i_{f1} i_{f2} i_{f3}]^T$, $I_x = [i_{x1} i_{x2} i_{x3}]^T$ and $F = [f_{ij}]_{3x3}$ respectively; and the superscript $T$ represents transpose.

The feed-forward network equations can be written as

$$I_i + i_{f1} = a_i i_{x1}$$ \hspace{1cm} (3.1.4)

$$k_{11} i_{x1} + sC_2 V_{p2} + a_2 i_{x2} = i_{f2}$$ \hspace{1cm} (3.1.5)

$$k_{21} i_{x2} + sC_3 V_{p3} + a_3 i_{x3} = i_{f3}$$ \hspace{1cm} (3.1.6)

where $a_i = 0(1)$ if connection is made via $y(x)$ terminal. Using this notation the voltage, $V_{pi}$ can be expressed as

$$V_{pi} = b_i i_{si} R_{si}, b_i = 1(-1) \text{ if } a_i = 1(0)$$ \hspace{1cm} (3.1.7)

Using (3.1.7), (3.1.5) and (3.1.6) modify to

$$k_{11} i_{x1} + sC_2 b_2 R_{x2} i_{x2} + a_2 i_{x2} = i_{f2}$$ \hspace{1cm} (3.1.8)

$$k_{21} i_{x2} + sC_3 b_3 R_{x3} i_{x3} + a_3 i_{x3} = i_{f3}$$ \hspace{1cm} (3.1.9)

In matrix form (3.1.4), (3.1.8) and (3.1.9) can be represented as
\[
\begin{bmatrix}
i_{s1} \\
i_{s2} \\
i_{s3}
\end{bmatrix} =
\begin{bmatrix}
a_1 & 0 & 0 \\
a_2 + b_2 s C_2 R_{x2} & a_3 + b_3 s C_3 R_{x3} & 0 \\
0 & k_{21} & 0
\end{bmatrix}
\begin{bmatrix}
i_{s1} \\
i_{s2} \\
i_{s3}
\end{bmatrix} - 0 I_i
\]  
(3.1.10)

Subtraction of (3.1.10) from (3.1.3) gives

\[
\begin{bmatrix}
f_{11} - a_1 & f_{12} & f_{13} \\
-f_{22} - (a_2 + b_2 s C_2 R_{x2}) & f_{23} & [i_{s1}] \\
0 & -k_{21} & f_{33} - (a_3 + b_3 s C_3 R_{x3}) [i_{s3}]
\end{bmatrix}
\begin{bmatrix}
i_{s1} \\
i_{s2} \\
i_{s3}
\end{bmatrix} =
\begin{bmatrix}
1 \\
0 I_i \\
0
\end{bmatrix}
\]

or

\[
\begin{bmatrix}
i_{s1} \\
i_{s2} \\
i_{s3}
\end{bmatrix} = -D^{-1}(s) . I . I_i
\]  
(3.1.11)

Total output current of the whole system from Fig. 3.1 is

\[
I_{out} = k_{12} i_{s1} + k_{22} i_{s2} + k_{31} i_{s3} + k_{32} i_{s3}
\]

Hence, the general transfer function of the whole system is

\[
\frac{I_{out}}{I_i} = \frac{-1}{D(s)} (k_{12} D_{11} + k_{22} D_{12} + k_{31} D_{13} + k_{32} D_{13}) 
\]

\[
= \frac{-N(s)}{D(s)}
\]  
(3.1.12)

where

\[
N(s) = D_{11} k_{12} + D_{12} k_{22} + D_{13} k_{31} + D_{13} k_{32}
\]

\[
D_{11} = (f_{22} - (a_2 + b_2 s C R_{x2}))(f_{33} - (a_3 + b_3 s C R_{x3})) + f_{23} k_{21}
\]
\[ D_{12} = k_{11} (f_{33} - (a_3 + b_3 s CR_{x3})) \]
\[ D_{13} = k_{11} k_{21} \]

and

\[ D(s) = (f_{11} - a_1) s^2 C^2 R_{x2} R_{x3} b_2 b_3 - f_{12} k_{11} b_3 s CR_{x3} \]
\[ + (f_{11} - a_1) s(-b_2 CR_{x3} f_{22} + a_2 b_3 CR_{x3} - b_2 CR_{x2} f_{33} + a_3 b_2 CR_{x2}) \]
\[ + (f_{11} - a_1) (f_{22} f_{33} - a_3 f_{22} - a_2 f_{33} + a_2 a_3 + f_{23} k_{21}) \]
\[ + f_{13} k_{11} k_{21} + f_{12} k_{11} (f_{33} - a_3) \] (assuming \( C_2 = C_3 = C \))

The coefficients \( k_{12}, k_{22}, k_{31} \) and \( k_{32} \) can take values 1, 0, -1, indicating current flowing into, no current flow in the output and current flowing out of MOCCCIIs respectively.

### 3.1.1.2 CHARACTERISTICS OF FEEDBACK COEFFICIENT MATRIX

The feedback coefficient matrix \( F \) is defined by (3.1.3) has property that

\[ f_{ij} = \begin{cases} 
1, & \text{for feedback from } j^{th} \text{ output to } i^{th} \text{ input} \\
0, & \text{otherwise} 
\end{cases} \]

To maintain unity feedback, the feedback coefficient matrix should have only one non-zero (unity) element in each column. The unity feedback coefficient can be constructed via pure wire connection. It is apparent that there is one to one correspondence between feedback matrix \( F \) and the circuit configuration. The different \( F \) leads to different circuit structures. The feedback matrix is upper triangular according to the features of \( F \) and has only one unit element in each column leading to either \( f_{11} = 1 \) or \( a_i = 1 \). Hence there are \( 2^{*}3! \) different combinations, that is, 12 different filter structures.

Moreover, for 2\(^{nd}\) and 3\(^{rd}\) MOCCCIIs, feedback connections can be established via \( x \) or \( y \)
terminal which further gives $2^2$ possible topologies for each of $2 \times 3!$ connections. Thus
the total number of structures will be 48. Again if one takes all possible values of $k_{11}$, $k_{21}$,
$k_{12}$, $k_{22}$, $k_{31}$ and $k_{32}$, then the number will go to a very high value. The next section
explores the feasibility of these structures to function as biquad filters.

3.1.2 FILTER SYNTHESIS PROCEDURE

The transfer function for the proposed multiple loop feedback current mode
biquad filter model of Fig 3.1 can be obtained as

$$\frac{I_{out}}{I_i} = -\frac{N(s)}{D(s)}$$

The 12 basic structures for various valid combinations of feedback coefficients
($f_i$) and $a_i$ are tabulated in Table 3.1. For each of 12 structures, four different biquad
topologies can be obtained for various combinations of $a_2$ and $a_1$. The first structure and
various topologies obtained from it are discussed in detail now. Similar analysis can be
done for other structure as well.

Structure 1:

Figure 3.2 represents structure 1, where $a_1 = f_{12} = f_{23} = 1$, all other $f_i = 0$ and $C_2 = C_3$

$$= C, \text{ then}$$

$$\frac{I_{out}}{I_i} = \frac{-k_{12}((a_2 + b_2sR_{x_2}C)(a_1 + b_3sR_{x_3}C) + k_{21}) - k_{22}k_{11}(a_1 + b_3sR_{x_3}C) + k_{31}k_{11}k_{21} + k_{32}k_{11}k_{21}}{-b_2b_3s^2R_{x_2}R_{x_3}C^2 - (a_2b_3sR_{x_3}C + a_2b_2sR_{x_2}C + k_{11}b_3sR_{x_3}C) - a_2a_3 - k_{21} - k_{11}a_3}$$

(3.1.14)
Fig. 3.2 Structure-1 generated from proposed topology.

The transfer function of (3.1.14) gives four different topologies depending upon the values of \( a_2 \) and \( a_3 \) as follows:

**TOPOLOGY - A**

Cascading through \( y \) terminal of 2\(^{nd} \) MOCCII and \( y \) terminal of 3\(^{rd} \) MOCCII i.e. \( a_2 = 0, b_2 = -1, a_3 = 0, b_3 = -1 \) and considering \( k_{11} = -1 \) and \( k_{21} = 1 \) (Fig. 3.3), the transfer function (3.1.14) can be written as

\[
\frac{I_{out}}{I_i} = \frac{k_{12}(s^2R_{x2}R_{x3}C^2 + 1) - k_{22}(sR_{x3}C) - k_{31} - k_{32}}{s^2R_{x2}R_{x3}C^2 + sR_{x3}C + 1}
\]  

(3.1.15)

Fig. 3.3 Structure-1 topology A.

Different filter functions can be derived using the following specializations in the numerator of (3.1.15):
Low pass realization: \( k_{12} = 0, \ k_{22} = 0, \ k_{31} = 1, \ k_{32} = 0; \) Band pass realization: \( k_{12} = 0, \ k_{22} = 1, \ k_{31} = 0, \ k_{32} = 0; \) High pass realization: \( k_{12} = 1, \ k_{22} = 0, \ k_{31} = 1, \ k_{32} = 0; \) Notch realization: \( k_{12} = 1, \ k_{22} = 0, \ k_{31} = 0, \ k_{32} = 0; \) All pass realization: \( k_{12} = 1, \ k_{22} = 1, \ k_{31} = 0, \ k_{32} = 0. \)

**TOPOLOGY - B**

Cascading through \( x \) terminal of 2\(^{nd} \) MOCCCII and \( y \) terminal of 3\(^{rd} \) MOCCCIIs i.e. \( a_1 = b_2 = 1, \ a_3 = 0, \ b_3 = -1 \) and considering \( k_{11} = 1, \ k_{21} = -1, \) the transfer function (3.1.14) can be written as

\[
\frac{I_{out}}{I_{i}} = \frac{k_{12} ((1 + sR_{x2}C)(-sR_{x3}C) - 1) + k_{22} (sR_{x3}C) - k_{31} - k_{32}}{s^2 R_{x2} R_{x3} C^2 + 2sR_{x3}C + 1}
\]  

(3.1.16)

The various filter functions can be obtained from (3.1.16) using the following specializations:

Low pass realization: \( k_{12} = 0, \ k_{22} = 0, \ k_{31} = 1, \ k_{32} = 0; \) Band pass realization: \( k_{12} = 0, \ k_{22} = 1, \ k_{31} = 0, \ k_{32} = 0; \) High pass realization: \( k_{12} = 1, \ k_{22} = 1, \ k_{31} = -1, \ k_{32} = 0; \) Notch realization: \( k_{12} = 1, \ k_{22} = 1, \ k_{31} = 0, \ k_{32} = 0; \) All pass realization is also possible if number of outputs in 2\(^{nd} \) MOCCCII is increased by 1.

**TOPOLOGY - C**

Cascading through \( y \) terminal of 2\(^{nd} \) MOCCCII and \( x \) terminal of 3\(^{rd} \) MOCCCIIs i.e. \( a_1 = 0, \ b_2 = -1, \ a_3 = b_3 = 1 \) and considering \( k_{11} = -1 \) and \( k_{21} = -1, \) the transfer function (3.1.14) can be written as
\[
\frac{I_{\text{out}}}{I_i} = \frac{k_{12}((-s R_{x2} C)(1 + s R_{x3} C) - 1) + k_{22}(1 + s R_{x3} C) + k_{31} + k_{32}}{s^2 R_{x2} R_{x3} C^2 + s(R_{x2} + R_{x3}) C + 2}
\]  
(3.1.17)

The following filter functions can be obtained using the specializations in the numerator of (3.1.17):

- **Low pass realization:** \(k_{12} = 0, k_{22} = 0, k_{31} = 1, k_{32} = 0\);
- **Band pass realization:** \(k_{12} = 0, k_{22} = 1, k_{31} = 1, k_{32} = -1\);
- **High pass realization:** \(k_{12} = 1, k_{22} = 1, k_{31} = 0, k_{32} = 0, R_{x2} = R_{x3}\);
- **Notch realization:** \(k_{12} = 1, k_{22} = 1, k_{31} = -1, k_{32} = -1, R_{x2} = R_{x3}\).

**TOPOLOGY – D**

Cascading through x terminal of 2\(^{\text{nd}}\) MOCCII and x terminal of 3\(^{\text{rd}}\) MOCCIIIs i.e. \(a_2 = b_2 = 1, a_3 = b_3 = 1\) and considering \(k_{11} = -1\) and \(k_{21} = 1\), the transfer function (3.1.14) can be written as

\[
\frac{I_{\text{out}}}{I_i} = \frac{k_{12}((1 + s R_{x2} C)(1 + s R_{x3} C) + 1) + k_{22}(1 + s R_{x3} C) - k_{31} - k_{32}}{s^2 R_{x2} R_{x3} C^2 + s R_{x2} C + 1}
\]  
(3.1.18)

We can derive different filter functions using the following specializations in the numerator of (3.1.18):

- **Low pass realization:** \(k_{12} = 0, k_{22} = 0, k_{31} = 1, k_{32} = 0\);
- **Band pass realization:** \(k_{12} = 0, k_{22} = 1, k_{31} = 1, k_{32} = 0\).

It can be seen from (3.1.15) through (3.1.18) that topologies A, B and D enjoy independent electronic tunability and low sensitivity of filter parameters. It is also evident that one may get 48 biquad configurations form 12 structures of Table 3.1 for various combinations of \(a_2\) and \(a_3\). The possible numbers of biquads may enhance to very high value if one takes all possible combinations of \(k_{11}, k_{21}, k_{12}, k_{22}, k_{31}\) and \(k_{32}\).
Table 3.1

Different realizations of SISO filters generated from circuit of Fig. 3.1.

<table>
<thead>
<tr>
<th>Structure</th>
<th>Condition on $f_{ij}$ and $a_i$</th>
<th>Schematic Configuration</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>$a_i = f_{12} = f_{23} = 1$, other $f_{ij} = 0$</td>
<td>![Schematic Configuration 1]</td>
</tr>
<tr>
<td>2.</td>
<td>$f_{11} = f_{12} = f_{13} = 1$, $a_i = 0$, other $f_{ij} = 0$</td>
<td>![Schematic Configuration 2]</td>
</tr>
<tr>
<td>3.</td>
<td>$a_i = f_{12} = f_{13} = 1$, other $f_{ij} = 0$</td>
<td>![Schematic Configuration 3]</td>
</tr>
<tr>
<td>4.</td>
<td>$f_{11} = f_{12} = f_{23} = 1$, $a_i = 0$, other $f_{ij} = 0$</td>
<td>![Schematic Configuration 4]</td>
</tr>
<tr>
<td>5.</td>
<td>$f_{11} = f_{12} = f_{33} = 1$, $a_i = 0$, other $f_{ij} = 0$</td>
<td>![Schematic Configuration 5]</td>
</tr>
</tbody>
</table>
6. \( a_i = f_{12} = f_{33} = 1, \ f_{11} = 0, \ \text{other} \ f_{ij} = 0 \)

7. \( f_{11} = f_{22} = f_{13} = 1, \ a_i = 0, \ \text{other} \ f_{ij} = 0 \)

8. \( a_i = f_{22} = f_{13} = 1, \ f_{11} = 0, \ \text{other} \ f_{ij} = 0 \)

9. \( f_{11} = f_{22} = f_{23} = 1, \ a_i = 0, \ \text{other} \ f_{ij} = 0 \)

10. \( a_i = f_{22} = f_{23} = 1, \ f_{11} = 0, \ \text{other} \ f_{ij} = 0 \)

11. \( f_{11} = f_{22} = f_{33} = 1, \ a_i = 0, \ \text{other} \ f_{ij} = 0 \)

12. \( a_i = f_{22} = f_{33} = 1, \ f_{11} = 0, \ \text{other} \ f_{ij} = 0 \)
3.13 INFLUENCE OF CCCII NONIDEALITIES ON FILTER CHARACTERISTICS AND DESIGN CRITERION

The frequency performance of the filter circuit may deviate from the ideal one due to nonidealities of MOCCCIIs. The nonidealities effects may be categorized in two groups. The first comes from frequency dependence of internal current and voltage transfers of MOCCCIIs. Due to this non-ideality, the port relations of MOCCCIIs may be expressed as

\[ v_x = v_i \beta_i(s) + i_x | R_x | (I_{o_i}) \], \ i_f = 0, \ i_{Z_i^+} = i_x \alpha_{i^+}(s) \text{ and } i_{Z_i^-} = -i_x \alpha_{i^-}(s); \ i = 1, 2, 3. \]

The voltage transfer function \( \beta_i(s) \) and current transfer functions \( \alpha_{i^+}(s) \) and \( \alpha_{i^-}(s) \) can be modeled as:

\[ \beta_i(s) = \frac{\beta_i}{s + \omega_{\beta_i}}, \quad \alpha_{i^+}(s) = \frac{\alpha_{i^+}}{s + \omega_{\alpha_{i^+}}}, \quad \text{and} \quad \alpha_{i^-}(s) = \frac{\alpha_{i^-}}{s + \omega_{\alpha_{i^-}}}. \]

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. The transfer function represented in (3.1.15) for the circuit of Fig. 3.3 may be modified as

\[ \frac{I_{out}}{I_i} \bigg|_{n} = \frac{as^2 + bs + d}{D_n(s)} \]  \hspace{1cm} (3.1.19)

where

\[ a = \frac{\alpha_{i^+}}{(s + \omega_{\alpha_{i^+}})}, \quad b = -\frac{\alpha_{i^-} \alpha_{i^+} \beta_{i^+}}{(s + \omega_{\alpha_{i^-}})(s + \omega_{\alpha_{i^+}})(s + \omega_{\beta_i})}, \quad k_{22} \]  \hspace{1cm} (3.1.20a)
\[
\begin{align*}
    d = & \frac{\alpha_1 \alpha_2, \alpha_3, \beta_2, \beta_3}{(s + \omega_{\alpha_2})(s + \omega_{\alpha_3})(s + \omega_{\beta_2})(s + \omega_{\beta_3})} \frac{(k_{31} + k_{32})}{R_{x2}R_{x3}C^2} \\
    & + \frac{\alpha_1 \alpha_2, \alpha_3, \beta_2, \beta_3}{(s + \omega_{\alpha_2})(s + \omega_{\alpha_3})(s + \omega_{\beta_2})(s + \omega_{\beta_3})} \frac{k_{12}}{R_{x2}R_{x3}C^2} \\

D_x(s) = & s^2 + \frac{\alpha_1 \alpha_2, \beta_2, \beta_3}{(s + \omega_{\alpha_2})(s + \omega_{\alpha_3})(s + \omega_{\beta_2})(s + \omega_{\beta_3})} \frac{1}{R_{x2}C} \\
    & + \frac{\alpha_2, \alpha_1, \beta_2, \beta_3}{(s + \omega_{\alpha_2})(s + \omega_{\alpha_3})(s + \omega_{\beta_2})(s + \omega_{\beta_3})} \frac{1}{R_{x2}R_{x3}C^2} 
\end{align*}
\]

Equations (3.1.20a) to (3.1.20c) clearly indicate that the pole frequencies of voltage and current transfer functions of MOCCCIIs affect the overall filter response. The effect, however, can be ignored if the operating frequencies are chosen sufficiently smaller than the voltage and current transfer pole frequencies of MOCCCI.

The second group of nonidealities comes from parasitics of MOCCCI comprising of resistances and capacitances connected in parallel at terminals y and z (i.e. R_y, C_y, R_z, and 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. 3.3 modifies to Fig. 3.4 where

\[
\begin{align*}
    C_{2eq} &= C \parallel C_{y2} \parallel C_{z2} \parallel C_{z1}, C_{3eq} = C \parallel C_{z2} \parallel C_{y3}, C_1 = C_{z2}, \\
    G_{1p} &= \frac{1}{R_{z2}}, G_{2p} = \frac{1}{(R_{z1} \parallel R_{y2} \parallel R_{z3})}, G_{3p} = \frac{1}{(R_{z2} \parallel R_{y3})}, \\
    C_2 &= C_3 = C.
\end{align*}
\]

The inductance is ignored in Fig. 3.4 as it affects the frequency response only at very high frequencies.
Fig. 3.4  Proposed filter structure including parasitics.

Considering the nonidealities outlined above the transfer function of the circuit shown in Fig. 3.4 can be given as

\[ \frac{I_{\text{out}}}{I_{i}} = \frac{as^2 + bs + d}{ms^2 + ns + p} \]  \hspace{1cm} (3.1.21)

where

\[ a = k_{12} \]  \hspace{1cm} (3.1.22)

\[ b = -k_{22} \frac{1}{R_x C_{2eq}} + k_{12} \left( \frac{G_{2p}}{C_{2eq}} + \frac{G_{3p}}{C_{3eq}} \right) \]  \hspace{1cm} (3.1.23)

\[ d = \frac{(k_{12} - k_{31} - k_{32})}{R_{x2} R_x C_{2eq} C_{3eq}} + \frac{G_{2p} G_{3p} k_{12}}{C_{2eq} C_{3eq}} - \frac{k_{22} G_{3p}}{R_x C_{2eq} C_{3eq}} \]  \hspace{1cm} (3.1.24)

\[ m = 1 + s C_{1p} R_{x1} + R_{x1} G_{2p} + \frac{C_{1p} R_{x1} G_{2p}}{C_{2eq}} + \frac{C_{1p} R_{x1} G_{3p}}{C_{3eq}} \]  \hspace{1cm} (3.1.25)

\[ n = \frac{1}{R_{x2} C_{2eq}} + \left( \frac{G_{3p}}{C_{3eq}} + \frac{G_{2p}}{C_{2eq}} \right) + \frac{C_{1p} R_{x1}}{R_{x2} R_x C_{2eq} C_{3eq}} + \frac{C_{1p} R_{x1} G_{2p} G_{3p}}{C_{2eq} C_{3eq}} \]  \hspace{1cm} (3.1.26)

\[ p = \frac{1}{R_{x2} R_x C_{2eq} C_{3eq}} + \frac{G_{3p}}{R_x C_{2eq} C_{3eq}} + \frac{G_{2p} G_{3p}}{C_{2eq} C_{3eq}} + \frac{R_{x1} G_{1p}}{R_{x2} R_x C_{2eq} C_{3eq}} + \frac{R_{x1} G_{1p} G_{2p} G_{3p}}{C_{2eq} C_{3eq}} \]  \hspace{1cm} (3.1.27)
It may be seen from (3.1.22) to (3.1.27) (except (3.1.22)) that the filter parameters are influenced by the parasitic elements, and the ideal values of the coefficients b, d, m, n and p are limited to only first term. The design criterion may be derived by approximating the coefficient values as discussed now.

The coefficient m is the only coefficient that has frequency dependence. Assuming the ratios $C_{1p}/C_{3eq}$ and $C_{1p}/C_{2eq}$ and parasitic conductances as very small, we can ignore last three terms in (3.1.25) and (3.1.26) and last two terms in (3.1.27). Therefore

$$m = 1 + sC_{1p}R_{x1} = \frac{1}{\omega_1} (s + \omega_1)$$

(3.1.28)

where corner frequency $\omega_1 = 1/C_{1p}R_{x1}$

For near ideal operation at high frequencies, the frequency of operation should be smaller than $\omega_1$. It may be noted from (3.1.22) to (3.1.27) that the values of $\omega_0$ and $\omega_0/Q_0$ for zeros and poles are equal to those obtained from (3.1.15) if the parasitics are ignored. So the values of $\omega_0$ and $\omega_0/Q_0$ are considered same for zeros and poles in the further calculations.

For $C_2 = C_3 = C >>$ parasitic capacitances, we can write $C_{2eq} = C_{3eq} = C_2 = C_3 = C$.

So, the ideal bandwidth of the transfer function given in (3.1.14) can be written as

$$\frac{\omega_0}{Q_0} = \frac{1}{R_{x2}C} \approx \frac{1}{R_{x2}C_{2eq}}$$
We, therefore, find from (3.1.23) that b will be approximately equal to the ideal case, if we set

\[ k_{22} \frac{a_0}{Q_0} \gg k_{12} \left( \frac{G_{2p}}{C_{eq}} + \frac{G_{3p}}{C_{eq}} \right) \]  

or,

\[ C \gg [k_{12}(G_{2p} + G_{3p})] \cdot \frac{1}{k_{22} \frac{Q_0}{a_0}} = C_4 \text{(say)} \] \tag{3.1.30}

Equation (3.1.30) clearly indicates that if external capacitance \( C = C_{2eq} = C_{3eq} \) is selected with a value higher than \( C_4 \), the influence of the parasitic elements to the coefficient \( b \) will be insignificant. Similar analysis for parameter \( d \) yields that for

\[ \frac{a_0^2}{Q_0} \gg \frac{G_{3p}}{C_{eq}} \left( \frac{G_{2p} k_{12}}{C_{eq}} - \frac{k_{22}}{R_{22} C_{eq}} \right) \frac{1}{(k_{12} - k_{31} - k_{32})} \]

or,

\[ \frac{a_0^2}{Q_0} \gg \frac{G_{3p}}{C_{eq}} \left( \frac{G_{2p} k_{12}}{C_{eq}} - k_{22} \frac{a_0}{Q_0} \right) \frac{1}{(k_{12} - k_{31} - k_{32})} \] \tag{3.1.31}

and applying (3.1.30), we can write (3.1.31) as

\[ \frac{a_0^2}{Q_0} \gg \frac{G_{3p}}{C_{eq}} \frac{a_0}{Q_0} \frac{k_{22}}{Q_0} \frac{1}{(k_{12} - k_{31} - k_{32})} \]

or,

\[ C \gg \frac{G_{3p}}{a_0 Q_0} \frac{k_{22}}{(k_{12} - k_{31} - k_{32})} = C_5 \text{(say)} \] \tag{3.1.32}

Similar analysis of (3.1.26) and (3.1.27) gives respectively,

\[ C \gg \frac{([G_{2p} + G_{3p}] Q_0}{a_0} = C_6 \text{(say)} \] \tag{3.1.33}
\[ C \gg \left[ \frac{G_{3p}}{\omega_0 Q_0} = C_7 \text{(say)} \right] \quad (3.1.34) \]

Thus by choosing

\[ C \gg \max(C_4, C_5, C_6, C_7) \quad (3.1.35) \]

the effect of parasitic impedances can be practically eliminated and thus the filter may approach towards its ideal response. However the maximum frequency of operation will be limited by parasitic poles of CCCII and corner frequency \( \omega_0 \).

3.1.4 RESULTS

To validate the theoretical predictions, the proposed circuit is simulated with PSPICE circuit simulation program using MOCCCI schematic presented in section 2.3 with typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) [112] and supply voltages of ±2.5 V. It is found that to practically eliminate the effect of parasitics of CCCII the value of \( C = C_2 = C_3 \) has to satisfy (3.1.35). The values of parasitic resistances, capacitances, \( \alpha \) and \( \beta \) are found to be \( R_y = 91 \, \text{k}\Omega \), \( C_y = 5.28 \, \text{pF} \), \( R_e = 384 \, \text{k}\Omega \), \( C_e = 5.28 \, \text{pF} \), \( \alpha = 0.0985 \), \( \beta = 0.999 \). The values of \( C_4, C_5, C_6 \) and \( C_7 \) are calculated respectively as 23.8 pF, 10.88 pF, 24.20 pF and 10.71 pF. In order to show the effect of parasitics, the simulation of band pass filter was carried out for \( f_0 = 1.27 \, \text{MHz} \) at \( Q_0 = 1 \). Table-3.2 shows the various values of capacitance \( C (= C_2 = C_3) \) and corresponding ideal and simulated values of \( f_0 \). Figure 3.5 shows the variation of frequency with capacitance. It is evident form Table 3.2 and Fig. 3.5 that when value of \( C \) is much lower than \( \max(C_4, C_5, C_6, C_7) \) then the simulated value of \( f_0 \) deviates strongly
form the ideal value. It is observed that when C is about 50 times of max (C₄, C₅, C₆, C₇), the simulated f₀ is much closer to ideal value. Hence to get an output almost free from parasitic effects, the value of C = C₂ = C₃ should be about 50 times of max (C₄, C₅, C₆, C₇) or more.

Table 3.2
Designed and simulated values of natural frequency with bias current and capacitance values.

<table>
<thead>
<tr>
<th>Iᵢ₀₁ = Iᵢ₀₂ = Iᵢ₀₃ (µA)</th>
<th>C (nF)</th>
<th>Designed (ideal) value of f₀ (MHz)</th>
<th>f₀(simulated) (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100.0</td>
<td>1.0</td>
<td>1.27</td>
<td>1.24</td>
</tr>
<tr>
<td>10.0</td>
<td>0.1</td>
<td>1.27</td>
<td>1.10</td>
</tr>
<tr>
<td>1.0</td>
<td>0.01</td>
<td>1.27</td>
<td>0.515</td>
</tr>
<tr>
<td>0.1</td>
<td>0.001</td>
<td>1.27</td>
<td>0.068</td>
</tr>
</tbody>
</table>

Fig. 3.5 Dependence of frequency on capacitor (C) for band pass filter.

Figure 3.6 (a) shows the simulation results for low pass, band pass and high pass responses for a pole frequency f₀ = 1.27 MHz, quality factor Q₀ = 1, capacitor C = C₂ = C₃ = 1 nF and Iᵢ₀₁ = Iᵢ₀₂ = Iᵢ₀₃ = 100 µA. The orthogonal tunability of ω₀ and Q₀ of notch response is shown in Fig. 3.6 (b) for Q₀ = 1, C₂ = C₃ = 1nF, Iᵢ₀₁ = Iᵢ₀₂ = Iᵢ₀₃ = 20 µA, 40 µA, 60 µA, 80 µA, and 100 µA. To test the large signal behavior of the