4.0 Introduction

In the cascade form synthesis, the nth-order transfer function is realized as a non-interactive cascade of (i) only second-order sections for even n and (ii) one first- or a third-order section and the remaining second-order sections, for making up the desired values of n, in case n is odd. In the case of OA-C cascade form synthesis, the basic realization philosophy remains the same, but in addition the lower order sections, acting as the basic building blocks, are required to be realized with only capacitors and operational amplifiers.

At present, the usefulness of any filter section is judged on the basis of two important considerations:

(a) Performance, and
(b) Suitability to implementation in the contemporary IC technologies, in which MOS is currently dominating the scene.

Some of the important and desirable performance aspects of the basic building blocks (BBBs) are

i) Wide frequency-range of operation;

† The material included in this chapter has led to the publication of author's papers Nos. 1, 5 and 9, listed on page vii.
ii) high functional-versatility;

iii) independent tuning of parameters, preferably electronic tunability;

iv) low sensitivities of active and passive components;

v) low component count, particularly in the active device;

vi) low spread in component values;

vii) simplicity in design procedure;

viii) high input and low output impedance levels for facilitating the non-interactive cascading of the basic blocks.

Generally, it is very rare for circuit to satisfy all the above performance criteria simultaneously; an engineering compromise is therefore required in most of the cases.

Besides the essential requirement of high performance, it is also desirable that the circuit/system should be compatible with some contemporary IC technology. The advantages of such an implementation are too well known to be repeated here. The realization of a circuit, particularly in monolithic form also adds up to the improved performance of the circuit, which can not normally be achieved in the discrete form. For a circuit to be implemented in a contemporary IC technology, it should satisfy the technological constraints of that technology, like (i) the class of components which can conveniently be fabricated and (ii) the circuit should have economical
feasibility of high degree of process standardization. In the case of the biquadratic filtering blocks, this implies the requirement of functional versatility to the realization of multiple filter responses, such as, low-pass, high-pass, band-pass, band-elimination and all-pass. Batch processing can then be employed for the manufacture of such versatile BBBs, leading to the production of reliable, high quality circuits at low cost.

This chapter on OA-C filters emphasises on the study of such BBBs, which use only OAs and capacitor-ratios in their realization and are also suitable for implementation in the popular MOS technology. The circuits will be shown to possess a number of the desirable features mentioned earlier. Unlike the previous chapter, the emphasis is not on the technique for the realization of a filter; directly the circuits are given and studied in detail for their performance and the suitability to MOS implementation. The chapter is organized as follows:

In Section 4.1, the first-order OA-C filter sections have been described. Section 4.2, basically deals with the realization of second-order filter sections. The study of the effects of non-idealities and parasites are included in Section 4.3. The higher-order filters have been discussed in Section 4.4. In Section 4.5, the experimental results on some of the filters considered in the chapter are given.
Finally, important concluding remarks are given in Section 4.6.

4.1 First-Order OA-C Sections

This section deals with the first-order low-pass, high-pass and all-pass sections, using only one op-amp and C-ratios in their realizations. As has been pointed out earlier, the first-order sections find use in the realization of filters of odd-order in the cascade form synthesis approach. In addition to this, they may also be used in various other applications, such as filter realization schemes, phase-shifter, first-order equalizer for bass and trebble audio tones, etc.

4.1.1 Low-pass section

In Fig. 4.1, four OA-C first-order sections are given for the realization of LP-characteristics. Using the first-pole roll-off characterization of an OA, the transfer functions of the sections have been derived and included in Table 4.1, where $\alpha = C_1/C_{12} = C_1/(C_1 + C_2)$.

It may be noted that the circuits of Fig. 4.1: (a), (b) and (c), realize non-inverting low-pass characteristics, while that of Fig. 4.1 (d) realizes an inverting LP-characteristics. The low-pass gains are also included in the table. These circuits exhibit electronic tunability of pole-frequency with bias-voltage.
Fig. 4.1 (a), (b), (c) Non-inverting and (d) inverting first-order low-pass sections.
Table 4.1: First-order low-pass sections

<table>
<thead>
<tr>
<th>Fig. No.</th>
<th>Transfer function T(s) = ( \frac{V_o}{V_i} )</th>
<th>Gain ( H_{lp} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.1(a)</td>
<td>( \frac{B}{s + \alpha B} )</td>
<td>( l + \frac{C_2}{C_1} )</td>
</tr>
<tr>
<td>4.1(b) and 4.1(c)</td>
<td>( \frac{\alpha B}{s + \alpha B} )</td>
<td>1</td>
</tr>
<tr>
<td>4.1(d)</td>
<td>( \frac{-(1-\alpha)B}{s + \alpha B} )</td>
<td>( \frac{C_2}{C_1} )</td>
</tr>
</tbody>
</table>

### 4.1.2 High-pass section

First-order OA-C high-pass filter section is shown in Fig. 4.2. It is characterized by the transfer function

\[
T(s) = \frac{V_o}{V_i} = \frac{(1-\alpha)s}{s + \alpha B} \tag{4.1}
\]

where, \( \alpha = \frac{C_1}{C_{12}} \) and the high-pass gain, \( H_{hp} = (1-\alpha) \).

This circuit realizes non-inverting HP-characteristics, with electronic tunability of pole frequency with bias-voltage control.

### 4.1.3 All-pass section

Consider the OA-C first-order circuit shown in Fig. 4.3. The transfer function is given by
Fig. 4.2 First-order high-pass section.

Fig. 4.3 First-order all-pass section.
\[ T(s) = \frac{V_0}{V_1} = \frac{1}{1+\alpha_2}, \quad \frac{s - \frac{B}{1+\alpha_1}}{s + \frac{B}{1+\alpha_1}} \] (4.2)

where
\[ \alpha_1 = \frac{C_1}{C_2}, \quad \text{and} \quad \alpha_2 = \frac{C_3}{C_4}. \]

**Case 1**: In case, \( \sqrt{\alpha_1 \alpha_2} - 1 = 1 \), i.e., \( \alpha_1 \alpha_2 = 2 \), the circuit realizes the function
\[ T_{AP}(s) = \frac{V_0}{V_1} = \frac{1}{1+\alpha_2}, \quad \frac{s - \frac{B}{1+\alpha_1}}{s + \frac{B}{1+\alpha_1}} \] (4.3)

and exhibits all-pass characteristic, which is in terms of \( C \)-ratios. This circuit may also be used as a phase-shifter with the magnitude given by
\[ |T_{AP}(\omega)| = \frac{1}{1 + \alpha_2} \] (4.4)

and the phase-angle by
\[ \phi = \pi - 2 \tan^{-1} \left[ \frac{w(1+\alpha_1)}{B} \right]. \] (4.5)

**Case 2**: Simple first-order equalizer for bass and treble audio tone shaping can also be realized with the circuit of Fig. 4.3. In this case, it is required that \( \alpha_1 \alpha_2 < 1 \). The responses for bass and treble equalizers may be set by choosing appropriate value for \( \alpha_2 \).

The sensitivity aspects of the first-order sections
were examined. These are summarized as follows:

\[ S_{B}^{W} = 1; \quad |S_{C_{1}}^{W}| < 1, \quad \text{where} \quad i = 1, 2. \quad (4.6) \]

It is seen that in all the cases, the sensitivity figures are reasonable. Also, the parameters of the circuits are in terms of C-ratios, which may be realized in high precision, high degree of tracking and good stability in MOS technology.

4.2 Second-Order OA-C Filter Sections

In the previous chapter, some techniques were suggested for the realization of OA-C second-order filter sections. These filters were also studied in detail. In this section, we directly give two circuits for the realization of low to medium Q filters and four circuits for medium to high Q filters. In all the configurations, only OAs and C-ratios are employed in the circuit realizations. These circuits simultaneously realize low-pass and band-pass characteristics at the output terminals of the OAs. These may, therefore, constitute as the non-interactive basic building blocks for the realization of higher-order filters, as the outputs are directly available at the outputs of the OAs and high input impedance level may be adjusted by the proper design of C-ratios. The basic circuits may also be extended to obtain a general biquadratic filter, using C-MOS inverter. Such a filter may realize all types of important standard responses.
4.2.1 **Low to medium-Q filters**

Two multifunctional OA-C filters for the simultaneous realization of low-pass and band-pass responses are shown in Figs. 4.4(a) and (b). The expressions for the voltage transfer functions are included in Table 4.2, where \( a_1 = C_1/C_{12} \) and \( a_2 = C_3/C_{34} \).

**Table 4.2**: Voltage transfer functions of filters of Fig. 4.4(a) and (b).

<table>
<thead>
<tr>
<th>Fig. No.</th>
<th>( T_{BP}(s) )</th>
<th>( T_{LP}(s) )</th>
</tr>
</thead>
</table>
| 4.4(a)  | \[
\frac{V_1}{V_i} = \frac{sB_1}{s^2 + s \alpha_2B_1 + (1 - \alpha_2)B_1B_2} = \frac{sB_1}{D(s)}
\] | \[
\frac{V_2}{V_i} = \frac{a_1B_1B_2}{D(s)}
\] |
| 4.4(b)  | \[
\frac{V_1}{V_i} = \frac{B(s + \alpha_2B_2)}{s^2 + s \alpha_2B_2 + \alpha_2 B_1B_2} = \frac{B_1(s + \alpha_2B_2)}{D(s)}
\] | \[
\frac{V_2}{V_i} = \frac{a_1B_1B_2}{D(s)}
\] |

If, \( w \gg \alpha_2B_2 \),

\[
\frac{V_1}{V_i} = \frac{sB_1}{D(s)}
\]

It may be noted that the circuit of Fig. 4.4(a) realizes standard LP and BP responses where as, that of Fig. 4.4(b) realizes an ideal LP and mixed BP responses. This circuit may, however, be designed to give a near ideal BP response, if the condition included in the table is satisfied through design.
Fig. 4.4 (a), (b): Low-to-medium-Q LP/BP filters.
**Parameters** - The important filter parameters, such as LP and BP gains, the pole-\(w_o\) and pole-\(Q\), are included in Table 4.3, where \(C_{ij} = C_i + C_j\).

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Filter of Fig. 4.4(a)</th>
<th>Filter of Fig. 4.4(b)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(H_{LP})</td>
<td>((1 + \frac{C_3}{C_4}))</td>
<td>((1 + \frac{C_4}{C_3}))</td>
</tr>
<tr>
<td>(H_{BP})</td>
<td>((1 + \frac{C_4}{C_3}))</td>
<td>(\frac{B_1}{B_2} \left(1 + \frac{C_4}{C_3}\right))</td>
</tr>
<tr>
<td>(w_o)</td>
<td>(\left[B_1B_2 \cdot \frac{C_1}{C_{12}} \cdot \frac{C_4}{C_{34}} \right]^{1/2})</td>
<td>(\left[B_1B_2 \cdot \frac{C_1}{C_{12}} \cdot \frac{C_3}{C_{34}} \right]^{1/2})</td>
</tr>
<tr>
<td>(Q)</td>
<td>(\left[\frac{C_4}{C_3} \cdot \frac{B_2}{B_1} \cdot \frac{C_1}{C_{12}} \cdot \frac{C_{34}}{C_3} \right]^{1/2})</td>
<td>(\left[\frac{B_1}{B_2} \cdot \frac{C_1}{C_{12}} \cdot \frac{C_{34}}{C_3} \right]^{1/2})</td>
</tr>
</tbody>
</table>

The expressions in the table indicate convenient realization of the filters in the MOS technology, as all the parameters are in terms of ratioed-\(C\). It may be noted that the circuits are basically, suitable for the realization of low-to-medium-\(Q\) values. For the realization of medium-range \(Q\) values, it is required that

\[C_1 > C_2, \text{ and } C_4 > C_3\]  \(\text{(4.7)}\)

by the design. The circuit enjoys independent electronic
tunability of the pole-frequency. However, the passive tunability of $w_0$ and $Q$ are found to be interactive in these circuits.

0 Sensitivity - Detailed sensitivity analysis for the two circuits was performed. The final results of active and passive sensitivities for various parameters of interest are included in Table 4.4. It may be observed that the active gain sensitivities for the filter of Fig. 4.4(a) are zero and those of Fig. 4.4(b) are low. Also, in both the cases the $w_0$ and $Q$ sensitivities, to active and passive parameters, are less than or equal to one in magnitude; with most of them being less than or equal to half in magnitude. This constitutes an attractive feature of the circuits.

4.2.2 Medium to high-$Q$ filters

In this section, four OA-C second-order filters are realized and studied, which have independent passive tuning of $Q$ with $C$-ratios and are also suitable for the realization of higher $Q$-values. The circuits are shown in Fig. 4.5 and each simultaneously realizes BP, and LP responses at the output terminals of the OAs. Conventional analysis, based on the single-pole roll-off characterization of the OAs, gives the transfer functions listed in Table 4.5. The parameters of interest for these filters are also derived and included in Table 4.6.
Table 4.4(a) : Sensitivity figures of the filter of Fig. 4.4(a)

<table>
<thead>
<tr>
<th>Sensitivity TO OF</th>
<th>$H_{LP}$</th>
<th>$H_{BP}$</th>
<th>$\omega_0$</th>
<th>$Q$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$B_1$</td>
<td>0</td>
<td>0</td>
<td>$1/2$</td>
<td>$-1/2$</td>
</tr>
<tr>
<td>$B_2$</td>
<td>0</td>
<td>0</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td>$C_1$</td>
<td>0</td>
<td>0</td>
<td>$\frac{1}{2} \cdot \frac{C_2}{C_{12}}, (\frac{1}{2})$</td>
<td>$\frac{1}{2} \cdot \frac{C_2}{C_{12}}, (\frac{1}{2})$</td>
</tr>
<tr>
<td>$C_2$</td>
<td>0</td>
<td>0</td>
<td>$-\frac{1}{2} \cdot \frac{C_2}{C_{12}}, (\frac{1}{2}) - \frac{1}{2} \cdot \frac{C_2}{C_{12}}, (\frac{1}{2})$</td>
<td></td>
</tr>
<tr>
<td>$C_3$</td>
<td>$\frac{C_3}{C_{34}}, (\frac{1}{2})$</td>
<td>$-\frac{C_3}{C_{34}}, (\frac{1}{2})$</td>
<td>$\frac{1}{2} \cdot \frac{C_3}{C_{34}}, (\frac{1}{2}) - \frac{1}{2} \cdot \frac{C_3 + 2C_4}{C_3 + C_4}, (\frac{1}{2})$</td>
<td></td>
</tr>
<tr>
<td>$C_4$</td>
<td>$-\frac{C_3}{C_{34}}, (\frac{1}{2})$</td>
<td>$\frac{C_3}{C_{34}}, (\frac{1}{2})$</td>
<td>$\frac{1}{2} \cdot \frac{C_3}{C_{34}}, (\frac{1}{2}) - \frac{1}{2} \cdot \frac{C_3 + 2C_4}{C_3 + C_4}, (\frac{1}{2})$</td>
<td></td>
</tr>
<tr>
<td>Sensitivity</td>
<td>$H_{LP}$</td>
<td>$H_{BP}$</td>
<td>$w_0$</td>
<td>$Q$</td>
</tr>
<tr>
<td>-------------</td>
<td>---------</td>
<td>---------</td>
<td>-------</td>
<td>-----</td>
</tr>
<tr>
<td>$B_1$</td>
<td>0</td>
<td>1</td>
<td>1/2</td>
<td>1/2</td>
</tr>
<tr>
<td>$B_2$</td>
<td>0</td>
<td>-1</td>
<td>1/2</td>
<td>-1/2</td>
</tr>
<tr>
<td>$C_1$</td>
<td>0</td>
<td>0</td>
<td>$\frac{1}{2} \cdot \frac{C_2}{C_{12}} \langle 1/2 \rangle$</td>
<td>$\frac{1}{2} \cdot \frac{C_2}{C_{12}} \langle 1/2 \rangle$</td>
</tr>
<tr>
<td>$C_2$</td>
<td>0</td>
<td>0</td>
<td>$-\frac{1}{2} \cdot \frac{C_2}{C_{12}} \langle 1 \rangle$</td>
<td>$-\frac{1}{2} \cdot \frac{C_2}{C_{12}} \langle 1 \rangle$</td>
</tr>
<tr>
<td>$C_3$</td>
<td>$- \frac{C_4}{C_{34}} \langle 1 \rangle$</td>
<td>$- \frac{C_4}{C_{34}} \langle 1 \rangle$</td>
<td>$\frac{1}{2} \cdot \frac{C_4}{C_{34}} \langle 1/2 \rangle$</td>
<td>$\frac{1}{2} \cdot \frac{C_4}{C_{34}} \langle 1/2 \rangle$</td>
</tr>
<tr>
<td>$C_4$</td>
<td>$\frac{C_4}{C_{34}} \langle 1 \rangle$</td>
<td>$\frac{C_4}{C_{34}} \langle 1 \rangle$</td>
<td>$- \frac{1}{2} \cdot \frac{C_4}{C_{34}} \langle 1 \rangle$</td>
<td>$\frac{1}{2} \cdot \frac{C_4}{C_{34}} \langle 1 \rangle$</td>
</tr>
</tbody>
</table>
Fig. 4.5 (a), (b), (c) and (d): Medium-to-High-Q OA-C filters.
Table 4.5: Voltage transfer functions of filters of Fig. 4.5

<table>
<thead>
<tr>
<th>Fig. No.</th>
<th>$T_{BP}(s) = V_1/V_1$</th>
<th>$T_{LP}(s) = V_2/V_1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.5(a)</td>
<td>$\frac{B_1(s + B_2C_3/C_{34})}{s^2 + s(B_2C_3/C_{34}) + B_1B_2C_1/C_{12}} \neq \frac{B_1(s + B_2C_3/C_{34})}{D(s)}$</td>
<td>$\frac{B_1B_2}{D(s)}$</td>
</tr>
<tr>
<td></td>
<td>If $w \gg B_2C_3/C_{34}$,</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$\frac{sB_1}{D(s)}$</td>
<td></td>
</tr>
<tr>
<td>4.5(b)</td>
<td>Same as for filter of Fig. 4.5(a)</td>
<td>$\frac{B_1B_2C_1/C_{12}}{D(s)}$</td>
</tr>
<tr>
<td>4.5(c)</td>
<td>$\frac{sB_1C_1/C_{12}}{s^2 + s(B_1C_3/C_{34}) + B_1B_2C_2C_5/(C_{12}C_{56})} = \frac{sB_1C_1/C_{12}}{D(s)}$</td>
<td>$\frac{-B_1B_2C_1C_5/(C_{12}C_{56})}{D(s)}$</td>
</tr>
<tr>
<td>4.5(d)</td>
<td>$\frac{sB_1C_1/C_{12}}{s^2 + sB_1 \frac{C_3C_7}{C_8C_{34} + C_7C_{34}} + \frac{B_1B_2C_2C_5}{C_{12}C_{56}}}$</td>
<td>$\frac{-B_1B_2C_1C_5/(C_{12}C_{56})}{D(s)}$</td>
</tr>
</tbody>
</table>
The filters of Fig. 4.5(a) and (b) realize the non-inverting low-pass responses at the output of OA₂ and give mixed BP responses at the output of OA₁. However, by insuring, \( w \gg \frac{C_3}{C_{34}} B_2 \), near-ideal BP-response may conveniently be realized. Only four capacitors have been used in each of the realizations and the circuit parameters are found to be in terms of C-ratios. The circuits have independent passive tunability of Q with \( C_3 \) and/or \( C_4 \), and independent electronic tunability of pole-frequency. In a practical circuit, first the pole frequency is set at its nominal value and the desired value of Q is obtained with ratioed-capacitors, \( C_4/C_3 \). The circuits are suitable for medium-Q realizations. High-Q values may also be obtained at the cost of larger spread in \( C_4/C_3 \)-ratio. The circuit of Fig. 4.5(b) can also realize low-pass gain greater than unity.

The multifunctional filter of Fig. 4.5(c) employs an additional pair of capacitors in its realization and provides ideal non-inverting BP and inverting LP characteristics at the outputs of OA₁ and OA₂, respectively. The other aspects regarding tunability, component spread, etc., are similar to those of filters of Fig. 4.5(a) and(b).

The circuit of Fig. 4.5(d) has been obtained from the filter of Fig. 4.5(c) by feeding a C-attenuator, constituted by \( C_7 \) and \( C_8 \), at node P. This circuit exhibits ideal LP and
BP responses, similar to those of filter of Fig. 4.5(c).

The additional advantage which emerges from the increase in C-count is in the convenient realization of high-Q values in low C-spread. A careful comparison of the expressions of Q and \( H_{BP} \) given in Table 4.6 for the filters of Fig. 4.5(c) and (d) shows one additional multiplication factor, \( \frac{C_{78}}{C_{7}} + \frac{C_{8}}{C_{34}} \), in the later case. This helps in the convenient design of high-Q values in low C-spread. The other features of the circuit are similar to those of filter of Fig. 4.5(c).

The incremental sensitivities of all the four filters were evaluated with respect to active and passive parameters. Only the final results are included in Table 4.7. These show all the filters to have attractive sensitivity properties with the maximum sensitivity being equal to unity in magnitude. In most of the cases sensitivities are much lesser than this value.

4.2.3 Realization of General Biquadratic Filters

It is well known in technical literature on active synthesis that a general biquadratic building block can be realized from a second-order filter having any two standard responses by using a summer. The same applies to the OA-C filters. However, in order to have MOS-compatible realization, CMOS inverter(s) are required. Any multi-
Table 4.6: Parameters of filters of Fig. 4.5

<table>
<thead>
<tr>
<th>Fig.No.</th>
<th>( H_{LP} )</th>
<th>( H_{BP} )</th>
<th>( w_o )</th>
<th>( Q )</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.5(a)</td>
<td>( \frac{C_{12}}{C_1} ) ( B_2 \cdot \frac{C_{34}}{C_3} ) ( (B_1B_2 \frac{C_1}{C_{12}})^{1/2} ) ( (1 + \frac{C_4}{C_3}) (\frac{B_1}{B_2} \cdot \frac{C_1}{C_{12}})^{1/2} )</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4.5(b)</td>
<td>( \frac{1}{B_2} \cdot \frac{C_{34}}{C_3} ) ( (B_1B_2 \frac{C_1}{C_{12}})^{1/2} ) ( (1 + \frac{C_4}{C_3}) (\frac{B_1}{B_2} \cdot \frac{C_1}{C_{12}})^{1/2} )</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4.5(c)</td>
<td>( \frac{C_1}{C_2} \cdot \frac{C_{34}}{C_{12}} \cdot \frac{C_3}{C_3} ) ( (B_1B_2 \frac{C_2}{C_{12}} \cdot \frac{C_5}{C_{56}})^{1/2} ) ( (1 + \frac{C_4}{C_3}) (\frac{B_2}{B_1} \cdot \frac{C_2}{C_{12}} \cdot \frac{C_5}{C_{56}})^{1/2} )</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4.5(d)</td>
<td>( \frac{C_1}{C_2} \cdot \frac{C_{34}^2}{C_{12}} \cdot \frac{C_7}{C_7} \cdot \frac{C_{34}}{C_{56}} ) ( (B_1B_2 \frac{C_2}{C_{12}} \cdot \frac{C_5}{C_{56}})^{1/2} ) ( (\frac{C_{34}}{C_3} \cdot \frac{C_{78}}{C_7} + \frac{C_8}{C_8}) (\frac{B_2}{B_1} \cdot \frac{C_2}{C_{12}} \cdot \frac{C_5}{C_{56}})^{1/2} )</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
Table 4.7: Sensitivity figures of filters of Fig. 4.5

<table>
<thead>
<tr>
<th>Fig. No.</th>
<th>Sensitivity TO</th>
<th>(H_{LP})</th>
<th>(H_{BP})</th>
<th>(w_0)</th>
<th>(Q)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(B_1)</td>
<td></td>
<td>0</td>
<td>1</td>
<td>1/2</td>
<td>1/2</td>
</tr>
<tr>
<td>(B_2)</td>
<td></td>
<td>0</td>
<td>-1</td>
<td>1/2</td>
<td>-1/2</td>
</tr>
<tr>
<td>(C_1)</td>
<td>(\frac{C_2}{C_{12}},(\langle 1 \rangle))</td>
<td>0</td>
<td>(\frac{1}{2}) (\frac{C_2}{C_{12}},(\langle 1/2 \rangle))</td>
<td>(\frac{1}{2}) (\frac{C_2}{C_{12}},(\langle 1/2 \rangle))</td>
<td></td>
</tr>
<tr>
<td>(4.5(a))</td>
<td>(\frac{C_2}{C_{12}},(\langle 1 \rangle))</td>
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<td>(\frac{1}{2}) (\frac{C_2}{C_{12}},(\langle 1/2 \rangle))</td>
<td>(\frac{1}{2}) (\frac{C_2}{C_{12}},(\langle 1/2 \rangle))</td>
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<tr>
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<td>(\frac{C_2}{C_{12}},(\langle 1 \rangle))</td>
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<tr>
<td>(C_3)</td>
<td>0</td>
<td>(\frac{C_4}{C_{34}},(\langle 1 \rangle))</td>
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<td>(\frac{C_4}{C_{34}},(\langle 1 \rangle))</td>
<td></td>
</tr>
<tr>
<td>(C_4)</td>
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<td>(\frac{C_4}{C_{34}},(\langle 1 \rangle))</td>
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<td>(\frac{C_4}{C_{34}},(\langle 1 \rangle))</td>
<td></td>
</tr>
<tr>
<td>Fig. No.</td>
<td>Sensitivity</td>
<td>$H_{LP}$</td>
<td>$H_{BP}$</td>
<td>$w_o$</td>
<td>$Q$</td>
</tr>
<tr>
<td>---------</td>
<td>-------------</td>
<td>---------</td>
<td>---------</td>
<td>------</td>
<td>-----</td>
</tr>
<tr>
<td>$B_1$</td>
<td></td>
<td>0</td>
<td>1</td>
<td>$1/2$</td>
<td>$1/2$</td>
</tr>
<tr>
<td>$B_2$</td>
<td></td>
<td>0</td>
<td>-1</td>
<td>$1/2$</td>
<td>-$1/2$</td>
</tr>
<tr>
<td>$C_1$</td>
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<td>0</td>
<td>0</td>
<td>$1/2 \cdot \frac{c_2}{c_{12}}, (&lt;1/2)$</td>
<td>$1/2 \cdot \frac{c_2}{c_{12}}, (&lt;1/2)$</td>
</tr>
<tr>
<td>$C_2$</td>
<td></td>
<td>0</td>
<td>0</td>
<td>$1/2 \cdot \frac{c_2}{c_{12}}, (&lt;</td>
<td>1/2</td>
</tr>
<tr>
<td>$C_3$</td>
<td></td>
<td>0</td>
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<td>1</td>
<td>)$</td>
</tr>
<tr>
<td>$C_4$</td>
<td></td>
<td>0</td>
<td>$\frac{c_4}{c_{34}}, (&lt;1)$</td>
<td>0</td>
<td>$\frac{c_4}{c_{34}}, (&lt;1)$</td>
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<tr>
<th>Fig. No.</th>
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<th>$H_{BP}$</th>
<th>$w_0$</th>
<th>$Q$</th>
</tr>
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<tbody>
<tr>
<td>$B_1$</td>
<td></td>
<td>0</td>
<td>0</td>
<td>1/2</td>
<td>1/2</td>
</tr>
<tr>
<td>$B_2$</td>
<td></td>
<td>0</td>
<td>0</td>
<td>1/2</td>
<td>-1/2</td>
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<tr>
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<td>$\frac{C_2}{C_{12}},(&lt;1)$</td>
<td>$-\frac{1}{2}\cdot\frac{C_1}{C_{12}},(&lt;</td>
<td>1/2</td>
<td>)$</td>
</tr>
<tr>
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<td>$-\frac{C_2}{C_{12}},(&lt;1)$</td>
<td>$\frac{1}{2}\cdot\frac{C_1}{C_{12}},(&lt;</td>
<td>1/2</td>
<td>)$</td>
</tr>
<tr>
<td>$C_3$</td>
<td>0</td>
<td>$-\frac{C_4}{C_{34}},(&lt;1)$</td>
<td>0</td>
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</tr>
<tr>
<td>$C_4$</td>
<td>0</td>
<td>$\frac{C_4}{C_{34}},(&lt;1)$</td>
<td>0</td>
<td>$\frac{C_4}{C_{34}},(&lt;1)$</td>
<td></td>
</tr>
<tr>
<td>$C_5$</td>
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<td>0</td>
<td>$\frac{1}{2}\cdot\frac{C_6}{C_{56}},(&lt;</td>
<td>1/2</td>
<td>)$</td>
</tr>
<tr>
<td>$C_6$</td>
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<td>0</td>
<td>$-\frac{1}{2}\cdot\frac{C_6}{C_{56}},(&lt;</td>
<td>1/2</td>
<td>)$</td>
</tr>
</tbody>
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### Table 4.7 continued

<table>
<thead>
<tr>
<th>Fig.No.</th>
<th>Sensitivity TO</th>
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<th>H_{LP}</th>
<th>H_{BP}</th>
<th>w_{0}</th>
<th>Q</th>
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<td>B_1</td>
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<td>0</td>
<td>1/2</td>
<td>1/2</td>
<td></td>
</tr>
<tr>
<td></td>
<td>B_2</td>
<td>0</td>
<td>0</td>
<td>1/2</td>
<td>-1/2</td>
<td></td>
</tr>
<tr>
<td></td>
<td>C_1</td>
<td>1</td>
<td>\frac{C_2}{C_{12}},(\langle 1/2\rangle)</td>
<td>\frac{1}{2} \cdot \frac{C_1}{C_{12}},(\langle 1/2\rangle)</td>
<td>\frac{1}{2} \cdot \frac{C_1}{C_{12}},(\langle 1/2\rangle)</td>
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</tr>
<tr>
<td></td>
<td>C_2</td>
<td>-1</td>
<td>-\frac{C_2}{C_{12}},(\langle 1/2\rangle)</td>
<td>\frac{1}{2} \cdot \frac{C_1}{C_{12}},(\langle 1/2\rangle)</td>
<td>\frac{1}{2} \cdot \frac{C_1}{C_{12}},(\langle 1/2\rangle)</td>
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</tr>
<tr>
<td></td>
<td>C_3</td>
<td>0</td>
<td>-\frac{C_7 C_{48} + C_4 C_8}{C_8 C_{347} + C_7 C_{34}},(\langle 1\rangle)</td>
<td>0</td>
<td>-\frac{C_7 C_{48} + C_4 C_8}{C_8 C_{347} + C_7 C_{34}},(\langle 1\rangle)</td>
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</tr>
<tr>
<td></td>
<td>C_4</td>
<td>0</td>
<td>\frac{C_4 C_{78}}{C_8 C_{347} + C_7 C_{34}},(\langle 1\rangle)</td>
<td>0</td>
<td>\frac{C_4 C_{78}}{C_8 C_{347} + C_7 C_{34}},(\langle 1\rangle)</td>
<td></td>
</tr>
<tr>
<td></td>
<td>C_5</td>
<td>0</td>
<td>0</td>
<td>\frac{1}{2} \cdot \frac{C_6}{C_{56}},(\langle 1/2\rangle)</td>
<td>\frac{1}{2} \cdot \frac{C_6}{C_{56}},(\langle 1/2\rangle)</td>
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Table 4.7 continued ....

<table>
<thead>
<tr>
<th>Fig. No.</th>
<th>Sensitivity</th>
<th>$H_{LP}$</th>
<th>$H_{BP}$</th>
<th>$w_0$</th>
<th>$u$</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.5(d)</td>
<td>C₆</td>
<td>0</td>
<td>0</td>
<td>$-\frac{1}{2} \cdot \frac{C_6}{C_{56}}, (\langle</td>
<td>1/2</td>
</tr>
<tr>
<td></td>
<td>C₇</td>
<td>0</td>
<td>$-\frac{C_{34}C_8}{C_{8}C_{347}+C_{34}C_{7}}, (\langle 1 \rangle)$</td>
<td>0</td>
<td>$-\frac{C_{34}C_8}{C_{8}C_{347}+C_{34}C_{7}}, (\langle</td>
</tr>
<tr>
<td></td>
<td>C₈</td>
<td>0</td>
<td>$\frac{C_{34}C_8}{C_{8}C_{347}+C_{34}C_{7}}, (\langle 1 \rangle)$</td>
<td>0</td>
<td>$\frac{C_{34}C_8}{C_{8}C_{347}+C_{34}C_{7}}, (\langle 1 \rangle)$</td>
</tr>
</tbody>
</table>
functional OA-C second-order filter with two or more standard responses may be used for the purpose. We demonstrate the principle by employing the OA-C filter of Fig. 4.5(c).

The complete biquadratic circuit realization, with filter of Fig. 4.5(c) as the basic building block (BBB), and two inverters is shown in Fig. 4.6. The technique requires the summing of three signals, from the input source and the two standard signal outputs of the BBB, through an inverting summer. In addition, the responses from the BBB are required to be of inverting nature. Since the BP response of OA-C filter of Fig. 4.5(c) is non-inverting, the additional inverter is required (only shown in the symbolic form in Fig. 4.6). It may be noted that in case the filter of Fig. 3.8(b), with available inverting BP and LP responses would have been used, the additional inverter would have been avoided. The choice of the filter only demonstrates a general case.

The analysis of the circuit of Fig. 4.6 give the overall transfer function

\[
T(s) = \frac{s^2 + s(\beta - \alpha') + (\gamma - \delta')}{s^2 + s\beta + \gamma}
\]

where

\[
\alpha = B_1 \cdot \frac{C_1}{C_1 + C_2}, \quad \beta = B_1 \cdot \frac{C_3}{C_3 + C_4},
\]

\[
\gamma = B_1B_2 \cdot \frac{C_5}{C_5 + C_6}, \quad \frac{C_2}{C_1 + C_2}, \frac{C_5}{C_5 + C_6}
\]
Fig. 4.6 General biquadratic OA–C filter.
It is noticed that the circuit realizes a general biquadratic characteristic. Here, the BP and LP responses are directly available at the outputs of OA\(_1\) and OA\(_2\), respectively, of the BBB. The other important standard responses, viz., HP, BE, and AP, can be obtained through coefficient adjustment of the inverting summer. The conditions for the realizations of such responses are included in Table 4.8. The exercise demonstrates a convenient method of realizing any standard response from the second order multifunctional filters, already studied in this thesis.

4.3 Practical Considerations

Various factors which affect the idealized behaviour of OA–C filters are considered in this section. For a reliable performance, due considerations should be given to these factors in the design and fabrication of the system. In a practical set up, it has been found that the following factors play an important role on the overall performance of the filtering systems:
Table 4.8: Realization conditions for specific responses

<table>
<thead>
<tr>
<th>Response</th>
<th>Condition</th>
</tr>
</thead>
<tbody>
<tr>
<td>HP</td>
<td>( \frac{C_{s_1}}{C_{s_2}} (1 + \frac{C_2}{C_1}) = (1 + \frac{C_4}{C_3}) ) and ( \frac{C_2}{C_1} = \frac{C_{s_3}}{C_{s_1}} ).</td>
</tr>
<tr>
<td>BE</td>
<td>( C_{s_3} = 0 ) and ( \frac{C_{s_1}}{C_{s_2}} (1 + \frac{C_2}{C_1}) = (1 + \frac{C_4}{C_3}) ).</td>
</tr>
<tr>
<td>AP</td>
<td>( C_{s_3} = 0 ) and ( \frac{C_{s_1}}{C_{s_2}} (1 + \frac{C_2}{C_1}) = \frac{1}{2} (1 + \frac{C_4}{C_3}) ).</td>
</tr>
</tbody>
</table>

i) effects of parasitic capacitors;
ii) effects of excess phase;
iii) dc bias and stability;
iv) temperature drift and voltage fluctuations; and
v) effects of input and output impedances and 3-dB frequency \((w_a)\) of OAs.

These factors are next discussed in some details.
4.3.1 Effects of parasitic capacitors

In the practical fabrication of a MOS-filter, parasitic capacitances are almost invariably present. Their effect may be reduced considerably through a proper layout of the system. At present, standard computer aided layout programmes are available for the job. In addition, in a number of filter topologies the effects of parasitic capacitors may itself be included in the design to advantage. In this section, the OA-C filter of Fig. 4.5(c) is chosen to illustrate this point.

Consider the filter redrawn in Fig. 4.7, with the parasitic capacitors between the input terminals of the OAs and the ground, shown within the dotted lines. Here \( C_{p1} \) and \( C_{n1} \) represent the parasitic values appearing between the inverting (n) and the non-inverting (p) input terminals of the ith-amplifier to ground. Analysis of the circuit gives the following expressions for the filter parameters:

\[
H_{lp} = -\frac{C_1}{C_2}, \quad H_{bp} = \frac{C_1}{C_{12p1}} \cdot \frac{C_{34n1}}{C_3},
\]

\[
\omega_0 = \left[ B_1 B_2 \cdot \frac{C_2}{C_{12p1}} \cdot \frac{C_5}{C_{56n2}} \right]^{1/2},
\]

and

\[
Q = (1 + \frac{C_{4n1}}{C_3}) \left[ \frac{B_2}{B_1} \cdot \frac{C_2}{C_{12p1}} \cdot \frac{C_5}{C_{56n2}} \right]^{1/2}
\]
Fig. 4.7 OA–C filter of Fig. 4.5 (c) with parasitic capacitors (in dotted branches).
where, $C_{ik}$ and $C_{ijk}$, respectively represent $C_{ik} = C_i + C_k$ and $C_{ijk} = C_i + C_j + C_k$. In the actual design of the system, the parasites which are appearing in parallel with the circuit capacitors may be absorbed in them. This not only removes the undesirable effect of the parasites on the performance of the filter, but also helps in the reduction of the actual values of capacitors required in the fabrication. The reduction in $C$-values is an attractive feature, as it helps in the saving of important chip area of a monolithic filter.

4.3.2 Effects of excess-phase

At higher frequencies, when $\omega > 0.1 \text{ B}$, the two-pole model of operational amplifier becomes effective. The effect of the second-pole may be represented more conveniently in terms of the excess-phase ($\tau$) exhibited by operational amplifiers. The high frequency model of an OA is thus represented by

$$A(s) = \frac{B}{s} \cdot e^{-st}. \quad (4.11)$$

The excess phase term, $\exp (-st)$, causes serious deviations in the performance of filters at high frequencies, i.e., above few hundred kilohertz for 741 type of OAs. By incorporating this term in the analysis, the realized filter parameters ($w_R$, $Q_R$, $H_{BPR}$) deviate considerably from the corresponding designed parameters based on the single-pole roll-off characterization of OAs. Therefore, filters for high frequency
applications (above a few hundred kilohertz), with general purpose OAs are required to be designed using predistortion formulae already available in technical literature. These relationships are given here for convenience:

\[ w_R = w_o \left[ 1 + \frac{w_R^2}{(2Q_R)^2} \right] = w_o, \]
\[ Q = \frac{Q}{(1 - 2Q^2 w_R^2)} \]

and

\[ H_{BPR} = \frac{H_{BP}}{1 - 2Q^2 w_R^2}. \]  (4.12)

It is evident from eqn. (4.12) that only a slight shift is exhibited in the pole-frequency, i.e., \( w_R = w_o \). However, drastic enhancements in the pole-\( Q \) and filter-gain (\( H_{BP} \)) are caused by the excess-phase of the amplifiers. This enhancement can make the filter oscillate at a frequency \( w_o \), if the designed \( Q \) is made larger than

\[ Q_{\text{max}} = \frac{1}{2 \sin \omega_0 \tau}. \]  (4.13)

It is interesting to note that sensitivities of the realized filter parameters (\( Q_R, H_{BPR} \)) with respect to \( \tau \) are:

\[ S_{Q_R} = S_{H_{BPR}} = 2Q_R w_R \tau \]  (4.14)

and with \( Q \) are

\[ S_{Q_R} = \frac{1}{1 - 2Q w_R \tau}, \quad \text{and} \quad S_{H_{BPR}} = \frac{2Q w_R \tau}{1 - 2Q w_R \tau}. \]  (4.15)

These results give a number of useful inferences. The \( Q-\)}
enhancement depicted in (4.12) is helpful in reducing the capacitors spread in the realization of high-Q filters. However, the term has an adverse effect on sensitivities of the filter. Using eqn. (4.15) with

\[ S_{Q}^{R} = S_{Q}^{R} \cdot S_{X_{1}}^{Q}, \quad \text{and} \quad S_{H_{BPR}}^{R} = S_{H_{BPR}}^{R} \cdot S_{X_{1}}^{Q} \]  \hspace{1cm} (4.16)

where, \( S_{X_{1}}^{Q} \) are parameter Q-sensitivities, the active and passive \( Q_{R} \) and \( H_{BPR} \) sensitivities may easily be evaluated. (Note that \( S_{X_{1}}^{Q} \) sensitivities have already been evaluated along with the study of a filter). It is evident from the above expressions that although the sensitivities are fairly attractive at low frequencies, the same may not be true when the excess-phase term becomes appreciable. Particularly, the \( Q \) and \( H_{BPR} \) sensitivities, at high frequencies, are increased appreciably. This makes the stabilization of \( \tau \) essential for reliable operation of OA-C filters at higher frequencies, i.e. \( \omega > 0.1 \) B. Some schemes for the stabilization of \( \tau \) have already been suggested by Schaumann in reference 48. More work may be persuaded in this direction.

4.3.3 DC bias and stability

OA-C filters are required to be provided with dc paths, for input bias-currents and for dc stability. If this path is blocked due to a capacitor, a high-value resistor (R) across the capacitor may serve the purpose, without affecting the
AC performance of the circuit, provided $R \gg Z_1$. For example, in the OA–C filter of Fig. 4.5(c), high-valued resistors, each of about 2.2 Mohm, were connected across the capacitors $C_2$, $C_4$, and $C_5$ for providing the desired biasing. Similarly, for the other filter circuits, discussed earlier, the dc paths to input terminals of OAs can be provided by placing high-valued resistors across appropriate capacitors. If the circuit is implemented in monolithic form, MOS amplifiers require very small bias currents, which may be provided directly by the leakage resistance of the circuit capacitors. Also, in MOS-circuits, compatible fabrication techniques are now available for fabricating high-valued resistors within small chip area, e.g., pinched resistors may be fabricated during the IC processing of chips, as discussed earlier.

4.3.4 Temperature drift and voltage fluctuations

It is evident from the parameter-expressions of OA–C filters that the gain bandwidth product of OAs is directly involved in these expressions. The dependence of $B$ on the pole-frequency of the filters is in direct variation. Also, $B$ is itself dependent on temperature and bias voltage. Thus, any temperature or voltage drift affects the performance of the filter. This problem may be circumvented by using extra-circuits for temperature stabilization, particularly, if the circuit is to be used under varying environmental
conditions. Some circuits are already available in literature and shall not be discussed here\textsuperscript{48,52}. Also, a stabilized power supply may be used for providing stable dc biasing.

The problem of temperature and voltage variations is not serious on the $Q$ of the filters. The expressions of $Q$ show it to be dependent on ratio of $B$s. Therefore, any variation in temperature or voltage tends to nullify the effect of each other. If the circuit is in monolithic MOS form, inherent tracking in MOS will itself make the effect insignificant on the $Q$ of the filter.

Another parameter which may be stabilized is the excess-phase term ($\tau$) of the OAs, particularly, for the filters designed in the frequency range, $w > 0.1 B$. This point has already been discussed earlier.

4.3.5 Effects of input and output impedances and $3dB$-frequency ($w_a$) of OAs\textsuperscript{73}

Besides the major amplifiers non-idealities already considered, finite values of input/output impedances affect the performance of circuits, particularly, for low gain MOS amplifiers. Also, if $w_a$ is comparable to $w$, i.e., at low frequency of operation, effect of finite-$w_a$ is exhibited on the circuit's performance.

In the case of MOS amplifiers, the input impedance is very high and its effect is negligible. Therefore, the input
impedance can be considered purely capacitive, i.e.,
\[ Z_{in}(s) = \frac{1}{s C_{in}}. \]
The effect of parasitic capacitors has already been discussed in Section 4.3.1.

The amplifiers output impedance can be considered to be dominantly resistive. This actually increases the order of the filter by one for each amplifier used in the circuit. Thus, a second-order two OAs filter becomes a fourth-order filter. For an exhaustive analysis, each filter circuit has to be analysed separately, as it is difficult to generalize the results. In practice, this is however not required, as the dominant effect of finite output resistance is only on the slight shifting of the dominant pole-pair \((w_o, Q)\). The additional two poles and a few zeros thus produced, generally do not pose danger to instability as long as the capacitive effect do not cause a serious loading of the amplifier.

In the present work, the effect of finite 3dB-frequency \((w_a)\) has not been considered, as higher frequency operation of the circuits are emphasized. However, in case \(w_a\) is comparable to the frequency of operation, its effect is exhibited on performance of the circuits. For low frequency of operation with \(\tau = 0\), the amplifiers dc gain is given by, \(A_0 = \frac{B}{w_a}\). It has been shown in reference 73 that the realized value of pole-\(Q\) and pole-frequency suffer slight deviations from the nominal values. The relationship between the realized and nominal values are given by:
\[ Q_m = Q \left( 1 - \frac{2Q w_a}{w_o} \right) \]  

(4.17)

and

\[ w_{om} = w_o \left( 1 + \frac{w_a}{w_o Q} \right)^{1/2} \]  

(4.18)

where, \( Q_m \) and \( w_{om} \) are the measured values, and the higher-order terms having negligible inference have been omitted.

In the case of band-pass response, the affect shifts the zero of the filter to \(-w_a\).

The various non-idealities considered in this section only slightly affect the pole-frequency, in case the variations in \( B_s \) of OAs are stabilized against voltage and temperature drifts. The deviation in the \( Q \)-values may be high, particularly, at frequencies which constitutes an appreciable fraction of the gain bandwidth product of the OA employed, due to the introduction of the excess-phase term. This therefore, requires a more careful design using eqn. (4.12). Also, stabilization of \( t \) is recommended at such a frequency-range.

**4.4 Realization of Higher-Order Filters**

In the op-amp based active synthesis, higher-order filters (say, nth-order) are conveniently realized through the cascade form synthesis\(^{23,59,87}\). As discussed earlier, the technique consists of realizing nth-order filter by employing a non-interactive cascade of \( n/2 \) second-order
sections, acting as the basic building blocks (BBBs), in case \( n \) is even. In the case of odd \( n \), a first or third-order section is additionally required. The \( n \)th-order transfer function \( T(s) \) can be written as

\[
T(s) = \prod_{i=1}^{n/2} T_i(s)
\]  \hspace{1cm} (4.19)

where, \( n \) is the order of overall filter and has been assumed even for convenience. The voltage transfer function \( T_i(s) \) of a general basic building blocks is of the form:

\[
T_i(s) = \frac{m_i s^2 + c_i s + d_i}{s^2 + a_i s + b_i}
\]  \hspace{1cm} (4.20)

with

\[
w_{o_i} = \left(\frac{b_i}{a_i}\right)^{1/2} \quad \text{and} \quad Q_i = \frac{\left(\frac{b_i}{a_i}\right)^{1/2}}{a_i}.
\]  \hspace{1cm} (4.21)

An important advantage of the approach lies in the convenience of tuning. As shown in (4.20), individual sections may be tuned separately for the important parameters before cascading to provide the overall desired response.

For the realization of the overall response, the individual sections are first designed. This requires each section to be characterized by a high input and low output impedances. The low output impedance is ensured if the output is taken at the output terminal of an OA; this is the case with most of the second-order active sections. In case, this does
not hold, a buffer stage will be required. The requirement of high input impedance can conveniently be obtained through a proper design of a filter section. This also helps in minimizing the inband losses. In the case of OA-C filter, the use of MOS OAs further helps in ensuring the high input impedance. Also, as the circuits employ only C-ratios as passive device, use of low C-values not only helps in economizing the chip area, but also in presenting high impedance levels.

For the design of a cascaded filter, the decomposition of a transfer function is required. The choice of optimum pole-zero pairing and gain distribution becomes very important for maximizing the dynamic range and providing the desired sensitivity properties to the filter. This aspect has been discussed in details in Reference 87. The sequencing of subsection also has a dominant effect on the signal-to-noise ratio, which is discussed in Reference 28.

4.4.1 Design example

In this section, design of a fourth-order Butterworth low-pass filter is considered. The filter is realized by cascading the second-order low-pass sections of the circuit given in Fig. 4.5 (b). The cascaded filter is shown in Fig. (4.8).

The overall transfer function of the filter is given by
Fig. 4.8 Cascaded fourth-order OA-C lowpass filter.
\[
T(s) = \frac{V_0}{V_1} = \frac{B_1 B_2 B_3 B_4 C_1 C_5}{(s^2 + s B_2 C_3 + B_1 B_2 C_1 C_7 + B_3 B_4 C_5)}
\]

(4.22)

From (4.22), the pole-frequency and pole-Q of individual sections are respectively given by:

\[
w_{o1} = (B_1 B_2 C_1 C_3) \frac{1}{2}, \quad Q_1 = (1 + C_4 C_3) \left[ \frac{B_1 C_1}{B_2 C_3} \right] \frac{1}{2}
\]

and

\[
w_{o2} = (B_3 B_4 C_3 C_5) \frac{1}{2}, \quad Q_2 = (1 + C_8 C_5) \left[ \frac{B_3 C_3}{B_4 C_5} \right] \frac{1}{2}
\]

(4.23)

The expressions clearly demonstrate the flexibility available in the design and the convenience of tuning present in the individual sections.

Next, the two cascaded sections are designed to realize a fourth-order Butterworth response for a pole frequency \(f_0 = 100\, \text{KHz}\). The standard Butterworth denominator polynomial and the corresponding Q-values for the two sections are available in Reference 87. The Q values are found to be: \(Q_1 = 1.31\), and \(Q_2 = 0.54\). Assuming the use of op-amps, CA3140, with gain bandwidth products, \(B_1 = B_2 = B_3 = B_4 = 4.5\, \text{MHz}\), equation (4.23) gives the following relationships:

\[
C_1 = 0.000494C_2 \quad (4.24)
\]

\[
C_3 = 0.017298C_4 \quad (4.25)
\]
\[ C_5 = 0.000494C_6 \quad (4.26) \]

and

\[ C_7 = 0.042845C_8. \quad (4.27) \]

On preselecting, \( C_1, C_3, C_5 \) and \( C_7 \) as 111pF, 220 pF, 111pF, and 470pF, respectively, the final C-values are obtained as:

\[
\begin{align*}
C_1 &= 111 \text{ pF}, \\
C_2 &= 224.66 \text{ nF}, \\
C_3 &= 220 \text{ pF}, \\
C_4 &= 12.7 \text{ nF}, \\
C_5 &= 111 \text{ pF}, \\
C_6 &= 224.66 \text{ nF}, \\
C_7 &= 470 \text{ pF}, \\
C_8 &= 10.95 \text{ nF}. \\
\end{align*}
\]

(4.28)

The designed circuit will also be experimentally verified in Section (4.5).

4.5 Experimental Results

In this section, experimental verifications on the discrete versions of some of the filters studied in this chapter are included. The experimental investigations can be broadly classified under three subsections. In Section 4.5.1 the experimental results on first-order low-pass and high-pass filter sections are given. The results on the second-order LP and BP filters are included in Section 4.5.2. These filters have been designed for high frequency operation using predistortion technique, discussed in Section 4.3. Finally in Section 4.5.3, experimental verification on a fourth-order filter, realized through the cascade form synthesis approach and designed in Section 4.4.1 is given.
4.5.1 First-order OA-C filters

The first-order LP and HP sections of Fig. 4.1(c) and 4.2, respectively, were designed and tested experimentally. Both the sections were designed to have the same cut-off frequency of 55.7 KHz. For the filter realization, OAs LM 741, were employed, with a measured value of gain bandwidth products (B) of 0.9 MHz at +15 V supply. On preselecting $C_2 = 0.1 \mu F$, the value of $C_1$ came out to be 6.6 nF from the relationship, $\omega_0 = BC_1/C_{12}$. In the circuit realization, polyester capacitors of 1 percent tolerance were employed. The designed circuits are included in Figs. 4.9 and 4.10. In the experimental set up, high-valued resistors of 2.2 Mohm were connected between the output and inverting input terminals of OAs for providing dc bias and stability to the circuits, without affecting the performance.

The experimental results on the circuits are shown in Fig. 4.9 and 4.10, respectively, for the LP and the HP sections. It is observed that even without resorting to tuning, the experimental results are in good conformity with the design.

Electronic tunability: To demonstrate tunability, the high-pass section of Fig. 4.10 (a) was considered. The bias voltage $V_B$ was varied from $\pm 6V$ to $\pm 18V$. The spot-values of the cut-off frequencies are included in Table 4.9, which show a
Fig. 4.9 (a) Designed first-order LP-section of Fig. 4.1(c)  
(b) Performance curve

<table>
<thead>
<tr>
<th>( \frac{V_2}{V_1} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.0</td>
</tr>
<tr>
<td>0.707</td>
</tr>
<tr>
<td>0.6</td>
</tr>
</tbody>
</table>

DESIGNED | OBSERVED
---|---
\( f_0 \) | 55.7 KHz | 55.65 KHz

Fig. 4.10 (a) Designed first-order HP-section of Fig. 4.2  
(b) Performance curve

<table>
<thead>
<tr>
<th>( \frac{V_2}{V_1} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.707</td>
</tr>
</tbody>
</table>

DESIGNED | OBSERVED
---|---
\( f_0 \) | 55.7 KHz | 55.65 KHz
corresponding variation from 39.7 KHz to 59.5 KHz. Thus, in the recommended range of the bias voltage for 741-type of OAs, a tunability of nearly 50 per cent was observed. It may be noted that corresponding variation in B was also 50 percent. The results clearly demonstrate the effect of bias-voltage control on B for providing electronic tunability over an appreciable range.

Table 4.9 : Variation of B and $f_{oh}$ with bias voltage ($V_B$)

<table>
<thead>
<tr>
<th>$V_B$</th>
<th>B</th>
<th>$f_{oh}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\pm 6$ V</td>
<td>0.64 MHz</td>
<td>39.7 KHz</td>
</tr>
<tr>
<td>$\pm 12$ V</td>
<td>0.84 MHz</td>
<td>49.6 KHz</td>
</tr>
<tr>
<td>$\pm 18$ V</td>
<td>0.96 MHz</td>
<td>59.5 KHz</td>
</tr>
</tbody>
</table>

4.5.2 Second-order OA-C filters

The second-order filter of Fig. 4.5(c) was used in the design of LP and BP responses using CA 3140 BiMOS OAs. These amplifiers were found to have a measured value of B of 4.5 MHz. The BP filter was designed for three sets of $\Omega_R (\Omega_R = 4.7, 11.6, \text{ and } 32.4)$ at a resonance frequency ($f_o$) of 653.8 KHz.

A LP-filter having Butterworth response was also designed for a cut-off frequency of 200.4 KHz. As this frequency of operation forms an appreciable fraction of B of the OAs, predistortion technique, discussed in Section 4.5, was employed,
along with the parameters given in Table 4.6. The following parameters were used from the RCA specification sheet:

\( \tau = 17 \text{ ns}, \ C_{\text{in}} = 4 \text{ pF} \). The designed values of the components are given in Table 4.10 and 4.11 for BP and LP cases, respectively. The circuits were fabricated using discrete capacitors of 1 percent tolerance. The dc feed-back and stability was provided by using 4.2 Mohm resistors across the capacitors \( C_2, C_3 \) and \( C_5 \).

The frequency responses of BP and LP filters are included in Figs. 4.11 and 4.12, respectively. The experimental results given in Tables 4.10 and 4.11 clearly demonstrate a close agreement between the theory and the experiment. It may be noted that the reported results were obtained without resorting to tuning. In the design, large-valued capacitors were deliberately chosen to minimize the effects of parasitic capacitances. In a monolithic MOS fabrication, actually low-valued capacitors, in few picofarad-range, are recommended for use to economize the chip area and also to present high-impedance levels in the circuit.

4.5.3 Higher-order OA-C filters

In this section, experimental results on a fourth-order low-pass Butterworth filter already designed for a cut-off frequency \( f_c = 100 \text{ KHz} \) in Section 4.4.1, are given. The component values are available in eqn. (4.28). The filter
Fig. 4.11 Frequency response curve of BP-filter of Fig. 4.5 (c)

Fig. 4.12 Frequency response curve of LP filter of Fig. 4.5 (c)

$Q_R = 4.7 \quad -$ 1
$Q_R = 11.6 \quad -$ 2
$Q_R = 32 \quad -$ 3
### Table 4.10: Design values and experimental results of BP filter of Fig. 4.5(c)

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Designed</th>
<th>Experimental</th>
<th>Component values</th>
</tr>
</thead>
</table>
| $f_0$      | 653.8 KHz | 652 KHz      | $C_1 = 0.72 \text{ nF}$, $C_2 = 0.116 \text{ nF}$  
|            |          |              | $C_5 = 0.112 \text{ nF}$, $C_6 = 0.62 \text{ nF}$  |
| $Q_R$      | 4.7      | 4.6          | $C_3 = 2.27 \text{ nF}$, $C_4 = 0.121 \text{ nF}$  |
| $Q_R$      | 11.6     | 11.8         | $C_3 = 0.325 \text{ nF}$, $C_4 = 9.6 \text{ nF}$  |
| $Q_R$      | 32.4     | 32.5         | $C_3 = 0.117 \text{ nF}$, $C_4 = 4.63 \text{ nF}$  |

### Table 4.11: Designed values and experimental results of LP filter of Fig. 4.5(c)

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Designed</th>
<th>Experimental</th>
<th>Component values</th>
</tr>
</thead>
</table>
| $f_0$      | 200.4 KHz | 199.5 KHz    | $C_1 = 47.7 \text{ nF}$, $C_2 = 2.23 \text{ nF}$  
| ($\omega_R = 0.707$) |          |              | $C_3 = 70 \text{ nF}$, $C_4 = 4.7 \text{ nF}$,  
|            |          |              | $C_5 = 2.2 \text{ nF}$ and $C_6 = 47.3 \text{ nF}$ |
Fig. 4.13 Frequency response curve of Butter-worth fourth-order low-pass filter of Fig. 4.8.
was fabricated in the laboratory with, CA 3140, OAs, having measured values of gain bandwidth products: $B_1 = B_2 = B_3 = B_4 = 4.5$ MHz, at a bias of $\pm 15$ V, as assumed in the design. Good quality capacitors with in 1 percent tolerance were employed in the circuit fabrication. The dc bias and stability to the circuit was provided using high valued resistors of 3.3 Mohm each, across the capacitors $C_1$, $C_3$, $C_5$ and $C_7$.

The frequency response of the filter is shown in Fig. 4.13. The observed value of the cut-off frequency was found to be 100.2 KHz, without circuit adjustments. This presents an excellent agreement with the design.

4.6 **Concluding Remarks**

In this chapter, material related to the cascade approach to OA-C synthesis has been presented. First, the basic building blocks constituted by first- and second-order filter sections were considered. First-order OA-C sections realizing important responses were realized and studied. Their important areas of applications were also mentioned. The second-order filter sections, which constitute the backbone of the cascade form synthesis were then realized and studied†. Emphasis was placed on the multifunctional capabilities of the filters to suit the monolithic fabrication. The filters were studied on the basis of their convenient

---

† The remarks include the second-order filters already realized and studied in Chapter 3.
realization of low-to-medium-Q values and medium-to-high Q values. It was also shown that these filters present sufficient design flexibility. A very attractive performance figure of the studied OA-C second-order filters is their low sensitivity properties in the frequency range, where the first-pole roll-off model is applicable.

Realization of higher-order filters, using the cascade form synthesis, was also considered. A practical design of a fourth-order filter was included. This design has also been verified experimentally with convincing results.

A section has been devoted on the effects of non-idealness on the performance of filters. The effect of parasitic capacitances has been studied. It was shown that by a proper design, these may be helpful in saving chip area of a monolithic filter. As such, the parasitic capacitances may be reduced by careful layout and routing of the circuits. The effect of excess-phase at frequencies \( w > 0.1 B \), may be serious and result in the drastic enhancement of pole-Q and gain of filters. It was recommended to use the predistortion technique in the design of filters at higher frequencies. Also, \( \tau \)-stabilization may be used. The effect of excess-phase has also been exhibited on sensitivities of the filters, which may pose serious problems at higher frequencies. The effect of temperature and voltage drifts reflects on the Bs of the OAs. This affects the pole-\( w_0 \) of the circuit.
Temperature stabilization of B and use of stabilized dc supply were recommended, particularly under changing environmental conditions.

Experimental verifications on some first-, second- and higher-order filters were performed. In all the cases, the results were found to be in good conformity with the theory. The OA-C filters were shown to provide a promising class of filters for realization in the contemporary MOS technology. The use of only OAs and ratioed-capacitors, along with the multifunctional capabilities constitute the major motivation force for their MOS implementation. These circuits, in addition, enjoy reliable high frequency performance.