Second-generation current-conveyor (CCII) filters are classified into four classes based on the number of CCIIIs employed. First, several voltage-mode and current-mode single CCII filters are described. A family of CCII voltage-mode and current-mode filters based on the two-integrator loop is generated using the building blocks approach. Two universal filters realizing the three transfer functions are given: one is a voltage-mode filter which employs five CCIIIs, and the second is a current-mode filter which employs four two-output CCIIIs. Both of the universal filters employ grounded elements, and are very attractive for VLSI realization by using MOS grounded resistors. All the filters considered are evaluated based on non-ideal CCIIIs. Sensitivities and design equations for each circuit are given. © 1998 Published by Elsevier Science Ltd.

1. Introduction

Several active filters have been introduced in the literature [1-5] using the second-generation current-conveyor (CCII) as the active building block [6]. Most of these CCII filters are generated from the conventional operational amplifier (op-amp) filters using the adjoint network theorem [7], or using the transformation theorem relating a class of op-amps to CCII circuits [8], or using the nullor equivalent circuit approach [9-11]. In all these methods the generated CCII filter is equivalent to the original op-amp filter only if both active devices are ideal.

In this paper the circuits realizing voltage transfer functions are defined as 'voltage mode' circuits, and those realizing current transfer functions are defined as 'current mode' circuits.

The objectives of this paper are:

1. To classify the CCII filters based on the number of the CCIIIs employed and based on the mode of operation.
2. To identify the circuits that are related to each other by some form of transformation (thus theoretically equivalent, assuming ideal CCIIIs).
3. To compare the active sensitivities based on non-ideal CCIIIs of the circuits that are theoretically equivalent to each other.
4. To introduce some new filter circuits using the CCII as the active building block.

Two types of CCIIIs are employed in the filter circuits that are considered in this paper. The first is the single-output CCII and the second is the two-output CCII. The symbolic representations for these CCIIIs are shown in Fig. 1, and their matrix equations are described below.
The single-output CCII is a three-port active building block which is described by the following matrix equation:

\[
\begin{bmatrix}
  V_x \\
  I_y \\
  I_z
\end{bmatrix} =
\begin{bmatrix}
  0 & B & 0 \\
  0 & 0 & 0 \\
  K & 0 & 0
\end{bmatrix}
\begin{bmatrix}
  I_x \\
  V_y \\
  V_z
\end{bmatrix}
\]  

(1)

For positive \( K \) the CCII is defined as a non-inverting CCII; on the other hand, if \( K \) is negative the CCII is defined as an inverting CCII.

The two-output CCII is a four-port active building block which is defined by the following matrix equation:

\[
\begin{bmatrix}
  V_x \\
  I_y \\
  I_{Z1} \\
  I_{Z2}
\end{bmatrix} =
\begin{bmatrix}
  0 & B & 0 & 0 \\
  0 & 0 & 0 & 0 \\
  K & 0 & 0 & 0 \\
 -K & 0 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
  I_x \\
  V_y \\
  V_{Z1} \\
  V_{Z2}
\end{bmatrix}
\]  

(2)

For positive \( K \), \( Z_1 \) is defined as the non-inverting output and \( Z_2 \) as the inverting output, and vice versa for negative \( K \). This two-output CCII has been used before in other applications [4, 12].

In the ideal case \( B=K=1 \). In practice, \( B \) and \( K \) are frequency dependent and for frequencies much less than the \( f_{3dB} \) of the CCII, \( B \) and \( K \) are real quantities of magnitudes slightly less than one. For the commercially available CCII 01, both \( B \) and \( K \) have a 3 dB bandwidth \( f_{3dB}=100 \text{MHz} \) [13].

2. The single CCII filters

Several active filters are available which employ a single CCII as the active building block. In this section the most important single CCII filters are considered in detail and are classified as two classes.

2.1. Class S-1 filters

The first class of filters considered in this section is shown in Fig. 2, and is generated from the Sallen–Key (SK) filters [14]. Figure 2a represents a current-mode second-order lowpass filter [4].
which is generated from the unity gain SK lowpass filter using the adjoint network theorem [7]. The current transfer function of this filter assuming an ideal CCII is given by:

$$T_i(s) = \frac{I_o}{I_i} = \frac{1}{s^2 C_1 C_2 R_1 R_2 + s(R_1 + R_2) C_2 + 1}$$

(3)

The voltage-driven version of this circuit is shown in Fig. 2b, where the resistor $R_i$ controls the dc gain which equals $R_i/R_t$.

An alternative realization of the SK lowpass filter using the CCII is shown in Fig. 2c. This circuit employs the CCII as a voltage follower, and has a voltage transfer function as given by eq. (3). For a specified $\omega_0$ and $Q$ the design equations for each of the three circuits of Fig. 2 are given by:

$$R_1 = R_2 = R$$

(4a)

$$C_1 = \frac{2Q}{\omega_0 R} \quad \text{and} \quad C_2 = \frac{1}{2Q \omega_0 R}$$

(4b)

For this equal $R$ design, the $\omega_0$ and the $Q$ passive sensitivities are very low as in the original op-amp voltage follower version [14], and are given by:

$$S_{R_1}^{\omega_0} = S_{R_2}^{\omega_0} = S_{C_1}^{\omega_0} = S_{C_2}^{\omega_0} = -\frac{1}{2}$$

(5a)

$$S_{R_1}^{Q} = S_{R_2}^{Q} = 0, S_{C_1}^{Q} = -S_{C_2}^{Q} = \frac{1}{2}$$

(5b)

Although the circuits of Fig. 2 have the same passive sensitivities, they have different $Q$ and $T(0)$ sensitivities, when considering a non-ideal CCII as defined by eq. (1). Table 1 includes the active sensitivities of the circuits of Fig. 2. For the design given by eq. (4) it is seen that the circuits of Fig. 2a and 2b have $S_{B}^{Q} = 2KQ^2$, whereas the circuit of Fig. 2c has a $S_{B}^{Q} = 2BQ^2$. That is, the circuits of Fig. 2a and 2b are sensitive to the current error of the CCII, whereas the circuit of Fig. 2c is sensitive to the voltage error of the CCII.

The SK current-mode and voltage-mode second-order highpass filters using the CCII can be obtained from the lowpass filters of Fig. 2 by applying the $RC:CR$ transformation [14]. Designing the highpass filters with equal capacitors results in the same active sensitivities as in the SK lowpass case.

Similarly, the bandpass current-mode and voltage-mode filters using the CCII can also be generated from the corresponding conventional SK filters. For the voltage-mode SK bandpass filter using the CCII as a voltage follower with the design suggested in [15] results in a $S_{B}^{Q} = 3Q^2$ and a centre frequency gain $T(j\omega_0)=1/3$. On the other hand, the current-mode bandpass filter obtained by applying the adjoint network theorem to the conventional SK bandpass filter and using the same design as suggested in [15] results in a $S_{B}^{Q} = 3Q^2$ and a centre frequency gain of $1/3$. The voltage-driven version, however, can

<table>
<thead>
<tr>
<th>Class</th>
<th>Fig.</th>
<th>$S_{B}^{\omega_0}$</th>
<th>$S_{K}^{\omega_0}$</th>
<th>$S_{B}^{Q}$</th>
<th>$S_{K}^{Q}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>S-1</td>
<td>2a,b</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>$KQ \sqrt{\frac{C_1 R_1}{C_2 R_2}}$</td>
</tr>
<tr>
<td></td>
<td>2c</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>$BQ \sqrt{\frac{C_1 R_1}{C_2 R_2}}$</td>
</tr>
<tr>
<td>S-2</td>
<td>3b, 4a</td>
<td>1/2</td>
<td>1/2</td>
<td>1/2</td>
<td>1/2</td>
</tr>
<tr>
<td></td>
<td>3d</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>$KQ \sqrt{\frac{C_1 R_1}{C_2 R_2}}$</td>
</tr>
<tr>
<td></td>
<td>4b,c</td>
<td>0</td>
<td>0</td>
<td>$BQ \sqrt{\frac{C_1 R_1}{C_2 R_2}}$</td>
<td></td>
</tr>
</tbody>
</table>

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realize an arbitrary centre frequency gain by tuning the additional input resistor $R_i$ without affecting $\omega_0$ or $Q$ of the filter.

2.2. Class S-2 filters
The second class of the single CCII filters considered in this paper is based on the circuit shown in Fig. 3a [16], which realizes either a parallel L-R circuit or a parallel frequency-dependent negative resistance-capacitor (FDNR-C) circuit according to the proper selection of the admittances $Y_1$, $Y_2$, and $Y_3$. A second-order highpass filter can be realized if a capacitor is connected in series with the parallel L-R circuit. As stated in [16] the simulated FDNR-C circuit is very practical in realizing generalized lowpass filters; in particular, a second-order lowpass filter is realized by connecting a resistor in series with this FDNR-C circuit as shown in Fig. 3b.

The voltage transfer function of this grounded capacitor non-inverting lowpass filter is given by:

$$T_v(s) = \frac{V_0}{V_i} = \frac{1}{s^2C_1C_2R_1R_2 + s(C_1 + C_2)R_2 + 1}$$

(6)

This lowpass filter (as well as the highpass filter which is related to it by the $RC:CR$ transformation) can lead to the generation of other recently reported voltage-mode and current-mode filters [17, 18].

The capabilities of the circuit of Fig. 3a lie in the fact that the CCII employed is of inverting polarity and has infinite input impedance at node Y. Thus, the current that leaves node 3 is the same as the current entering the CCII at node 4, as shown in Fig. 3a. In other words, node 3 can be floating [16, 19]. (In fact the circuit given in [19] is exactly the same as that given in [16], and shown here in Fig. 3a after removing the ground from node 3.)

Setting $V_i$ equal to zero in the circuit of Fig. 3b and removing the ground from node 3, the circuit of Fig. 3c is obtained, which has the same characteristic equation as given by the denominator of eq. (6). This dead circuit can be excited either by a voltage source or a current source according to the desirable mode of operation.

2.3. Class S-2 voltage-mode filters
To inject a voltage signal into the circuit of Fig. 3c, the possible nodes are 3 and 1, since a voltage source cannot be applied to the X or the Z terminals of the CCII. (If $V_i$ is applied to node 4 and X is grounded, the circuit obtained is of theoretical interest only, since grounding X destroys the current conveying property.) If, however, $V_i$ is applied to node 3 and node 3' is grounded, the circuit obtained is of limited practical value since it conditionally realizes a highpass response at node 2 provided that $C_1R_1=(C_1+C_2)R_2$.

Injecting the input voltage into the Y terminal of the CCII and connecting node 1 to ground, the circuit shown in Fig. 3d is obtained [17], which has the same lowpass transfer function as given by eq. (6). It has in addition an inverting bandpass output at node 2 with a transfer function given by:

$$\frac{V_2}{V_i} = \frac{-sC_1R_1}{D(s)}$$

(7)

where $D(s)$ is the same as given by eq. (6).

Of course the circuit of Fig. 3b has also a bandpass voltage response across $R_1$ which is floating. As a non-inverting lowpass filter, however, the circuit of Fig. 3b has the advantage over the circuit of Fig. 3d in using grounded capacitors.

Another way to generate the circuit of Fig. 3d from that of Fig. 3b, and for generality, consider the circuit of Fig. 3e. The current $I_1$ is given by:
Fig. 3. The parallel L-R or FDNR-C circuit [16]. (b) The grounded capacitor second-order lowpass filter derived from (a). (c) The dead network obtained from (b). (d) The high-input impedance lowpass-bandpass filter [17]. (e) The voltage-mode generalized configuration I. (f) The voltage-mode generalized configuration II.
\[ I_1 = (V_i - V_0)Y_4 \quad (8) \]

Applying the input signal to port Y of the CCII and in order to maintain the same current magnitudes in all admittances as well as in the CCII, one obtains the circuit of Fig. 3f. It should be noted that for the same \( V_i \) the currents in the CCII and in all admittances are in opposite directions to those in the circuit of Fig. 3e, as indicated clearly on the circuit diagrams. For completeness, the generalized expressions for the voltage transfer functions are given:

\[ \frac{V_0}{V_i} = \frac{+Y_3Y_4}{Y_1Y_2 + Y_3(Y_1 + Y_2 + Y_4)} \quad (9) \]

For the circuit of Fig. 3f

\[ \frac{V_2}{V_i} = \frac{-Y_1Y_4}{Y_1Y_2 + Y_3(Y_1 + Y_2 + Y_4)} \quad (10) \]

2.4. Class S-2 current mode filters

A grounded capacitor current-mode lowpass filter can be generated from the circuit of Fig. 3c when driven by a current source at node 2 and grounding node 3. The output current of this lowpass filter can be taken from the second output of the CCII as shown in Fig. 4a. In this case the current transfer function is the same as given by eq. (6). It is worth noting that the current in \( C_1 \) represents a bandpass response, and it can be also taken as an output current using a second CCII to act as a current follower. The transfer function in this case is given by:

\[ T_i(s) = \frac{sC_1R_2}{D(s)} \quad (11) \]

From the circuit of Fig. 3c and taking node 2 as the ground node, the lowpass and the bandpass current mode filters given in [18] can be obtained if the input current is injected at node 3 or at node 1, as shown in Fig. 4b and 4c, respectively. The transfer functions are given, respectively, by:

\[ T_i(s) = \frac{1}{D(s)} \quad \text{and} \quad T_i(s) = \frac{-3C_2R_1}{D(s)} \quad (12) \]

Of course if a load is inserted between the Z output and the ground, the output current \( I_0 \) will remain unchanged.

As shown above the voltage mode circuits of Fig. 3b and 3d as well as the current-mode circuits of Fig. 4a-4c have the same \( D(s) \) (assuming ideal CCII) as given in eq. (6), and for a specified \( \omega_0 \) and \( Q \) the design equations for any of those five circuits are given by:

\[ C_1 = C_2 = C \quad (13a) \]
The $w_0$ passive sensitivities are the same as given by eq. (5a) and the $Q$ sensitivities are given by:

$$
S_{Q_1}^R = -S_{Q_2}^R = \frac{1}{2} \quad \text{and} \quad S_{Q_1}^C = -S_{Q_3}^C = -\frac{1}{2} + \frac{C_2}{C_1 + C_2}
$$

(14)

For the equal $C$ design, the $Q$ sensitivity with respect to $C_1$ or $C_2$ is equal to zero. Of course, the above results are identical to the those obtained in the conventional op-amp negative feedback topology circuit [14], since both circuits are generated by adding a series resistor to the parallel FDNR-C circuit.

The active $w_0$ and $Q$ sensitivities of the circuits of Fig. 3b, 3d and 4 assuming a non-ideal CCII are given in Table 1. For the equal $C$ design it is seen that the circuits of Fig. 3b and 4a have very low $Q$ sensitivities with respect to $K$ and $B$, whereas the circuit of Fig. 3d has a $S_{Q_2}^R = 2KQ^2$, and those of Fig. 4b and 4c have a $S_{Q_2}^B = 2BQ^2$. Thus, from all the single CCII circuits considered in this paper, the best circuits are those of Fig. 3b and 4a not only because they employ grounded capacitors but also because of the very low passive and active sensitivities (all $w_0$ and $Q$ passive and active sensitivities $\leq 0.5$).

3. The dual CCII filters

In this section several two-CCII voltage-mode and current-mode filters are considered, some of which are new. For each filter circuit the design equations, the passive and the active sensitivities assuming non-ideal CCIIIs are given.

3.1. Class D-1 filters

Figure 5a represents a voltage-mode lowpass filter which is generated from the Bach lowpass filter [20] using the adjoint network theorem, and then adding the resistor $R_1$ at the input which provides independent control on the dc gain, which is given by $K_1K_2(R_2/R_1)$. In this circuit the two CCIIIs are used as current followers and must have the same polarities. Fig. 5b represents a new highpass filter with a very high input impedance which is derived from Fig. 5a by changing the excitation port of the first CCII. This circuit is sensitive to the current errors of both CCIIIs since it results in shifting one of the zeros from the origin to $-((1 - K_1K_2)/C_1R_1)$. Of course another lowpass filter different from that of Fig. 5a can
be generated from Fig. 5b by applying the \(RC:CR\) transformation.

Figure 5c represents a new current-mode non-inverting bandpass (BP) filter which is generated from Fig. 5b by replacing the second CCII by a two-output CCII. The transfer function is given by:

\[
\frac{I_{BP}}{I_i} = \frac{sC_2R_1K_1K_2B_1}{s^2C_1C_2R_1R_2 + s[C_1R_1 + C_2R_2(1 - K_1K_2)] + 1}
\]

(15)

Assuming ideal CCIIIs, that is \(K_1K_2=1\), the \(o_0\) and Q passive sensitivities are all \(<0.5\). From Table 2, it is seen that for any of the circuits of Fig. 5 which belong to class D-1, the Q sensitivities to \(K_1\) and \(K_2\) equal \(Q^2\). For a specified \(o_0\) and Q three alternative sets of design equations for the circuits of Fig. 5 are given in Table 3.

3.2. Class D-2 filters

A new current mode bandpass filter is shown in Fig. 6a, which is generated from the circuit of Fig. 3c by injecting the input current at node 1 and grounding node 3. The current in the capacitor \(C_2\) represents a bandpass response. To utilize this current the capacitor \(C_2\) is connected to a second CCII acting as a current follower.

The transfer function is given by:

\[
T_i(s) = \frac{I_{BP}}{I_i} = \frac{sC_2K_2(K_1B_1R_1 - R_2)}{s^2C_1C_2R_1R_2 + s(C_1 + C_2)R_2 + K_1B_1}
\]

(16)

It should be noted that the case of interest, namely \(Q>0.5\), implies that \(R_1>R_2\), and thus the sign of \(T_i(s)\) is the same as the sign of \(K_2\).
### TABLE 3 Design equations for the class D, T and M circuits

<table>
<thead>
<tr>
<th>Class</th>
<th>Fig.</th>
<th>Design equations</th>
<th>Design 2</th>
<th>Design 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>D-1</td>
<td>5a-c</td>
<td>( R_1 = R_2 = R, C_1 = 1/Q \omega_0 R, )</td>
<td>( C_1 = C_2 = C, R_1 = 1/Q \omega_0 C, )</td>
<td>( C_2 = QC_1, R_1 = 1/Q \omega_0 C_1, R_2 = 1/\omega_0 C_1 )</td>
</tr>
<tr>
<td>D-2</td>
<td>6a-d</td>
<td>( C_1 = C_2 = C, R_1 = 2Q/\omega_0 C, )</td>
<td>( R_1 = 1/2Q \omega_0 C )</td>
<td><strong>Design 2</strong></td>
</tr>
<tr>
<td>D-3</td>
<td>7a, 8a</td>
<td>Design 1</td>
<td>( C_1 = C_2 = C, R_1 = 1/\omega_0 R, )</td>
<td>( C_1 = C_2 = C, R_1 = 1/2Q \omega_0 C, )</td>
</tr>
<tr>
<td></td>
<td></td>
<td>( R_1 = R_2 = R, C_1 = 1/2Q \omega_0 R, )</td>
<td><strong>Design 2</strong></td>
<td>( R_2 = Q/\omega_0 C )</td>
</tr>
<tr>
<td>T-1</td>
<td>9a-d</td>
<td>( R_1 = R_2 = R, C_1 = 1/\omega_0 R, )</td>
<td>( C_1 = C_2 = C, R_1 = 1/\omega_0 R, )</td>
<td>( C_1 = 1/Q \omega_0 R )</td>
</tr>
<tr>
<td>T-2</td>
<td>10a</td>
<td>Same as circuits of Fig. 5</td>
<td><strong>Design 2</strong></td>
<td><strong>Design 2</strong></td>
</tr>
<tr>
<td>T-2</td>
<td>10b</td>
<td>Same as circuits of Fig. 5</td>
<td><strong>Design 2</strong></td>
<td></td>
</tr>
<tr>
<td>M-1</td>
<td>11</td>
<td>( C_1 = C_2 = C, R_1 = R, )</td>
<td>( C_1 = C_2 = C, R_1 = 1/\omega_0 C, R_2 = Q/\omega_0 C )</td>
<td></td>
</tr>
<tr>
<td>M-2</td>
<td>12</td>
<td>Same as circuits of Fig. 9</td>
<td><strong>Design 2</strong></td>
<td><strong>Design 2</strong></td>
</tr>
</tbody>
</table>

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![Fig. 6](image-url) A new current-mode bandpass filter generated from Fig. 3c. (b) A grounded-capacitor current-mode lowpass filter generated from Fig. 4a. (c) The modified lowpass filter derived from Fig. 3b. (d) The high-input impedance lowpass filter [22].
Figure 6b represents a current-mode lowpass filter using two CCIIIs which is generated from Fig. 4a. Its transfer function is given by:

\[
\frac{I_{LP}}{I_i} = \frac{-K_2B_1}{D(s)}
\]

(17)

where \(D(s)\) is the same as given in eq. (16). The sign of the transfer function is opposite to the polarity of the second CCII. It is worth noting that the current in the capacitor \(C_1\) represents a bandpass response; for the equal \(C\) design, however, \(T_0=0.5\).

The circuit of Fig. 6b can also lead to the generation of the current-mode filter given in [21], which employs a non-inverting CCII and a first-generation current conveyor CCII. The latter can be replaced by a CCII for the practicality of the realization using the commercially-available dual current conveyor Catalogue No. CCLI01 [1], by disconnecting node 2 from the \(Z\) output of the first CCII and connecting it to the \(Z\) output of the second CCII (taking \(K_2=1\)) and grounding the \(Z\) output of the first CCII.

Figure 6c represents a modified version of Fig. 3b by adding a second CCII acting as a voltage follower. Although the polarity of this CCII does not affect the operation of this circuit, it is taken as a non-inverting CCII in order to demonstrate how the circuit of Fig. 6d can be generated from Fig. 6c. Assuming ideal CCIIIs it is seen that the current \(I\) which leaves \(Z_1\) to node 2 in Fig. 6c is the same as the current which leaves \(Z_2\) to ground. Thus, by disconnecting \(Z_1\) from node 2 and grounding \(Z_2\) and removing the ground from node 5 and connecting it to node 2 to supply the same current \(I\) to the passive circuit, we obtain the circuit of Fig. 6d [22]. The active sensitivities of both circuits of Fig. 6c and 6d are given in Table 2.

### 3.3. Class D-3 filters:

Figure 7a represents a non-inverting lowpass-inverting bandpass filter [23] based on the two-integrator loop, where the first CCII is a non-inverting CCII and the second is an inverting CCII. The transfer functions are given by:

\[
\frac{V_1}{V_i} = \frac{-sC_2R_2K_1}{D(s)} \quad \text{and} \quad \frac{V_2}{V_i} = \frac{K_1K_2B_2}{D(s)}
\]

(18)

where

\[
D(s) = s^2C_1C_2R_1R_2(K_1 + 1) + sC_1R_1(K_1 + 1)K_2B_1B_2 + K_1K_2B_1B_2
\]

(19)

The \(\omega_0\) and the \(Q\) active sensitivities are given in Table 2. Two alternative designs are given in Table 3, for both designs \(T_0=Q^2\).

The grounded capacitor lowpass-bandpass filter shown in Fig. 7b is generated from Fig. 7a by reflecting the floating capacitor to two grounded capacitors. If the two grounded capacitors are taken as equal and each equals twice the floating capacitor, the generated circuit will be equivalent to Fig. 7a [8]. Here, however, they are taken as different to provide an additional degree of freedom in order to have independent control of \(Q\). This is the only circuit considered in this paper with three capacitors.

The transfer functions are given by:

\[
\frac{V_1}{V_i} = \frac{-sC_2R_2K_1}{D(s)} \quad \text{and} \quad \frac{V_2}{V_i} = \frac{-K_1K_2B_2}{D(s)}
\]

(20)

where

\[
D(s) = s^2C_1C_2R_1R_2 - sC_1R_1K_1K_2B_1B_2 - K_1K_2B_1B_2
\]

(21)

It is seen that \(K_1K_2\) must be negative, which implies that two realizations are possible. In the first realization \(K_1=1\), \(K_2=-1\); in this case the
bandpass is inverting, whereas in the second realization $K_1 = -1$ and $K_2 = 1$, which realizes a non-inverting bandpass. In both cases the lowpass is non-inverting. The active sensitivities and the design equations are given in Tables 2 and 3, respectively. This design results in $T_0 = Q$.

Figure 7c represents a high-input impedance bandpass filter [22] which is also based on the two-integrator loop. The circuit uses the same components as in Fig. 7a with $V_i$ and $V_2$ interchanged and $C_1$ advantageously reflected to a single grounded capacitor only. The second integrator is the same as in the previous two circuits, whereas the first is a differential integrator [24] characterized by the following equation:

$$V_1 = \frac{K_1}{sC_1R_1} [B_1V_i - V_2]$$

Thus the transfer function obtained is:

$$V_1 = \frac{sC_2R_2K_1B_1}{V_i} \frac{sC_1C_2R_1R_2 + sC_1R_2 + K_1K_2B_2}{s^2C_1C_2R_1R_2 + sC_1R_2 + K_1K_2B_2}$$

It is worth noting that in this case $V_2$ does not represent a lowpass. The realization given in [22] employs two non-inverting CCIIs, and it is seen that it is also possible to use two inverting CCIIs, which results in an inverting bandpass response. It is also possible to use an equal C design [22] or an equal R design as given in Table 3.

Comparing the active sensitivities of the three circuits of Fig. 7, it is seen that the circuit of Fig. 7a has the lowest sensitivities to $K_1$ whereas the circuit of Fig. 7c has zero sensitivity to $B_1$.

Figure 8a represents an inverting current mode bandpass filter which is generated from Fig. 7a; its transfer function is given by:

$$I_{BP} = -\frac{sC_2R_1K_1K_2B_2}{D(s)}$$

where $D(s)$ is the same as given by eq. (19). The design equations for this circuit are given in Table 3 based on taking $R_1 = R_2$ (the equal C design is not practical in this case since it results in $T_0 = -0.5$).

Figure 8b represents a grounded capacitor current-mode bandpass filter [25], which can either by non-inverting or inverting depending on the $K_2$ polarity. The transfer function is given by:

$$I_{BP} = \frac{sC_2R_1K_1K_2B_2}{I_i}$$

where $D(s)$ is the same as given by eq. (19). The design equations for this circuit are given in Table 3 based on taking $R_1 = R_2$ (the equal C design is not practical in this case since it results in $T_0 = -0.5$).

It is seen that all the $\omega_0$ and the Q passive sensitivities equal 0.5. The active sensitivities are given in Table 2, and three alternative sets of design equations are given in Table 3.
4. The three-CCII filters

4.1. Class T-1 filters

Recently three alternative approaches have been used to realize a CCII Tow-Thomas biquad [14]. The first approach is based on the adjoint network theorem [4], the second approach is based on the nullor network criteria [9], and the third approach is based on the equivalent building blocks [8]. The realization given in [4] requires that the three current conveyors have infinite current gains, and it realizes only one type of response, that is the attractive feature of the multiple outputs biquad circuit is lost when applying the adjoint network theorem. To obtain a voltage-mode realization, the current source is replaced by a voltage source in series with a resistor, which is made such that the transfer function is the same as that of the conventional op-amp case. The realization given in [9] is of theoretical interest only, since grounding the X terminal of the CCII destroys the current conveying property. In this realization the third CCII is used as a current-controlled current source (CCCS) whose gain can be adjusted to obtain the same transfer functions as those of the op-amp circuit.

Figure 8c represents a new grounded C and grounded R inverting bandpass filter. Its transfer function is given by:

\[
\frac{I_{BP}}{I_i} = \frac{s^2 C_1 R_1 K_1 K_2 B_2}{s^2 C_1 C_2 R_1 R_2 + s C_2 R_2 \frac{R_1 R_2}{R} - K_1 K_2 B_1 B_2}
\]

It is seen that \(K_1 K_2\) must be negative, which implies that the two CCII's must have opposite polarities. The advantage of this circuit is that \(Q\) is independently controlled by varying \(R\) without affecting \(\omega_0\) of the filter. It should also be noted that the current in \(R_1\) represents a lowpass response.

4. The three-CCII filters

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\[
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\]

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\[
\frac{I_{BP}}{I_i} = \frac{s^2 C_1 R_1 K_1 K_2 B_2}{s^2 C_1 C_2 R_1 R_2 + s C_2 R_2 \frac{R_1 R_2}{R} - K_1 K_2 B_1 B_2}
\]

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Fig. 9. A voltage-mode bandpass-lowpass filter [8]. (b) A modified voltage-mode bandpass-lowpass filter. (c) A high-input impedance voltage-mode bandpass-lowpass filter [26]. (d) A grounded R, C current-mode bandpass-lowpass filter [27].

\[
\frac{V_1}{V_i} = \frac{(-K_1/C_1R_3)s}{D(s)} \quad \text{and} \quad \frac{V_2}{V_i} = \frac{(-K_1K_2B_2/C_1C_2R_1R_3)}{D(s)}
\]

where

\[
D(s) = s^2 + \frac{s}{C_1R} + \frac{K_1K_2B_2B_3}{C_1C_2R_1R_2}
\]

This circuit realizes only an inverting lowpass response, whereas the realizable bandpass can be either inverting or non-inverting. The \(\omega_0\) and the \(Q\) active sensitivities are given in Table 4.

The circuit shown in Fig. 9b is generated from that of Fig. 9a by replacing the voltage follower and the feedback resistor \(R_2\) by a voltage-controlled current source (VCCS) realized by the third CCII and the grounded resistor \(R_2\). For this circuit the transfer functions are given by eq. (27), with \(D(s)\) slightly modified, and is given by:

\[
D(s) = s^2 + \frac{s}{C_1R} + \frac{K_1K_2K_3B_2B_3}{C_1C_2R_1R_2}
\]

The advantage of this realization is its capability of realizing the bandpass and the lowpass responses with any sign combinations as given in Table 5. A more attractive modified version of this circuit can be achieved by interchanging the excitation ports as shown in Fig. 9c [26]. This circuit has the advantage of a high input impedance and all resistors and capacitors are grounded. The transfer functions are given by:

\[
\frac{V_1}{V_i} = \frac{(K_1B_1/C_1R_3)s}{D(s)} \quad \text{and} \quad \frac{V_2}{V_i} = \frac{(K_1K_2B_1B_2/C_1C_2R_1R_3)}{D(s)}
\]

where \(D(s)\) is the same as given by eq. (29). The signs for \(V_1\) and \(V_2\) are opposite to those of the circuit of Fig. 9b for the same \(K_1\), \(K_2\) and \(K_3\), as shown in Table 5.
TABLE 4  The \( \omega_0 \) and the Q active sensitivities of the three CCII (class T) filters

<table>
<thead>
<tr>
<th>Class</th>
<th>Fig</th>
<th>( S_x )</th>
<th>( \times )</th>
<th>( B_1 )</th>
<th>( B_2 )</th>
<th>( B_3 )</th>
<th>( K_1 )</th>
<th>( K_2 )</th>
<th>( K_3 )</th>
</tr>
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<tbody>
<tr>
<td>T-1</td>
<td>9a</td>
<td>( S_{20}^a S_{Q}^a )</td>
<td>0 0</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>0 0</td>
<td></td>
</tr>
<tr>
<td></td>
<td>9b–d</td>
<td>( S_{20}^b S_{Q}^b )</td>
<td>0 0</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td></td>
</tr>
<tr>
<td>T-2</td>
<td>10a</td>
<td>( S_{20}^c S_{Q}^c )</td>
<td>1/2 1/2</td>
<td>0 0</td>
<td>0 0</td>
<td>1/2 1/2</td>
<td>0 0</td>
<td>0 0</td>
<td></td>
</tr>
<tr>
<td></td>
<td>10b</td>
<td>( S_{20}^d S_{Q}^d )</td>
<td>1/2 1/2</td>
<td>1/2 1/2</td>
<td>0 0</td>
<td>1/2 1/2</td>
<td>0 0</td>
<td>0 0</td>
<td></td>
</tr>
</tbody>
</table>

TABLE 5  The signs of \( T(s) \) for the circuits of Fig. 9b–9d

<table>
<thead>
<tr>
<th>CCII polarity</th>
<th>Fig. 9b</th>
<th>Sign of ( T(s) )</th>
<th>Fig. 9c</th>
<th>Sign of ( T(s) )</th>
<th>Fig. 9d</th>
<th>Sign of ( T(s) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( K_1 )</td>
<td>( K_2 )</td>
<td>( K_3 )</td>
<td>BP</td>
<td>LP</td>
<td>( K_1 )</td>
<td>( K_2 )</td>
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<tr>
<td>+</td>
<td>+</td>
<td>+</td>
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<td>+</td>
</tr>
</tbody>
</table>

Figure 9d represents the current-mode bandpass-lowpass filter which is generated from Fig. 9b using two CCIIIs with two outputs [26]. The current transfer functions are:

\[
T_{BP} = \frac{I_{BP}}{I_i} = \frac{(K_1K_2B_2/C_1R_1)s}{D(s)}, \quad T_{LP} = \frac{I_{LP}}{I_i} = \frac{(K_1K_2K_3B_3/C_1C_2R_1R_2)}{D(s)}
\]

where

\[
D(s) = s^2 + \frac{s}{C_1R} + \frac{K_1K_2K_3B_3}{C_1C_2R_1R_2}
\]

4.2. Class T-2 filters
Two current-mode bandpass-lowpass filters are shown in Fig. 10, and are generated from the circuits of Fig. 6a and 8b, respectively.

5. Multi-CCII filters
In this section two of the most attractive universal filters are given.

5.1. Class M-1 filters
Figure 11 represents the voltage mode universal filter which was recently introduced in [3]. This universal filter has the following advantages:

1. Infinite input impedance.
2. All elements are grounded.
3. Independent control on the gain without affecting \( \omega_0 \) or Q.
4. Independent control on Q without affecting \( \omega_0 \).
5. Very low \( \omega_0 \) and Q sensitivities to all circuit components.
6. The eight possible sign combinations for the highpass, bandpass and lowpass responses are realizable by adjusting the CCII's polarities.

The transfer functions are given by:

\[
\frac{V_1}{V_i} = \frac{(K_1B_1R/R_i)s^2}{D(s)}, \quad \frac{V_2}{V_i} = \frac{(K_1K_2B_1B_3R/C_1R_1R_i)s}{D(s)} \quad \text{and} \quad \frac{V_3}{V_i} = \frac{(K_1K_2K_3B_1B_2B_3R/C_1C_2R_1R_2R_i)}{D(s)}
\]

where
Fig. 10. A current-mode bandpass-lowpass filter derived from Fig. 6a. (b) A current-mode bandpass-lowpass filter derived from Fig. 8b.

\[ D(s) = s^2 + \frac{K_1K_2K_4B_2B_4R}{C_1R_1R_4} + \frac{K_1K_2K_3K_4B_3B_2R}{C_1C_2R_1R_2R_3} \]  
\[ (34) \]

From eq. (34) it is clear that the sign products \( K_1K_2K_4 \) and \( K_1K_2K_3K_5 \) must both be positive. From eq. (33) it is seen that the highpass, bandpass and lowpass response signs are related to \( K_1, K_1K_2 \) and \( K_1K_2K_3 \), respectively. Thus, all possible eight sign combinations for the three responses are realizable by this configuration, as described in [3].

The \( Q \) sensitivities to \( K_4, B_4 \) and \( R_4 \) are given by:

\[ S_{K_4}^Q = S_{B_4}^Q = -1 \quad \text{and} \quad S_{R_4}^Q = 1 \]  
\[ (35) \]

All other \( \omega_0 \) and the \( Q \) sensitivities are either \( \pm 0.5 \) or zero.

### 5.2. Class M-2 filters

Most recently the current-mode universal filter shown in Fig. 12 has been introduced in the literature [27]. The advantages of this circuit are:

1. very low input impedance;
2. very high output impedance;
3. independent control on \( Q \) without affecting \( \omega_0 \);
4. very low \( \omega_0 \) and \( Q \) sensitivities to all circuit components.

The transfer functions are given by:

\[ I_{HP} = \frac{K_1S^2}{D(s)} \quad I_{BP} = \frac{(K_1K_2/C_1R_1)s}{D(s)} \]
\[ I_{LP} = \frac{(K_1K_2K_3B_2B_3/C_1C_2R_1R_2)}{D(s)} \]
\[ (36) \]

where \( D(s) \) is given by:

\[ D(s) = s^2 + \frac{K_1K_4B_4}{C_1R} + \frac{K_1K_2K_3B_3B_2R}{C_1C_2R_1R_2} \]
\[ (37) \]

From the above equation it is seen that the resistor \( R \) controls \( Q \) without affecting \( \omega_0 \). All the \( \omega_0 \) and the \( Q \) sensitivities are either \( \pm 0.5 \) or zero except the \( Q \) sensitivities with respect to \( K_4, B_4 \) and \( R \), which are given by \(-1, -1 \) and \( 1 \), respectively.

### 6. Conclusions

The current conveyor filters are classified into four classes based on the number of CCIIIs used; each class includes several voltage-mode and current-mode circuits. The single CCII filters (class S) are classified into two types. The first type S-1 includes three circuits that are gener-
A.M. Soliman/Current conveyor filters

Fig. 11. The high-input impedance voltage-mode universal filter [3].

Fig. 12. The current-mode universal filter [27].

ated from the SK filters. The second type S-2 includes five circuits; two of them realize voltage transfer functions and the other three realize current transfer functions. It is shown that the grounded capacitor second-order non-inverting lowpass filter based on the well-known FDNR-C circuit [16] leads to the generation of the other voltage-mode and the three current-mode filters. Although these circuits are related to each other they have different active sensitivities. It is found that the circuits of Fig. 3b and 4a have the lowest sensitivities among all the class S circuits.

The two-CCII filters (class D) are classified into three subclasses. The subclass D-1 includes three circuits: the first one is generated from the Bach lowpass filter; the second circuit, which is a highpass filter, is generated from the lowpass circuit by changing the excitation ports; and the third circuit realizes both highpass and bandpass current responses and employs a two-output CCII. It is found that subclass D-1 circuits have a high Q sensitivity of $K = Q^2$, and thus they are limited to low Q applications. The subclass D-2 includes two current-mode and two voltage-mode circuits, and they are generated from the single CCII filters of Fig. 3b. Thus they have very low sensitivities. The subclass D-3 includes six circuits, all of which have very low sensitivities. The first three are voltage mode and are all based on the two integrator loop. Two of the three current-mode circuits employ grounded capacitors. Four circuits which belong to subclass T-1 are given.
All of them realize bandpass and lowpass responses and use grounded capacitors, and have very low sensitivities. Two current-mode circuits realizing also bandpass and lowpass responses are classified as subclass T-2. Finally, two universal filters realizing the three transfer functions are given. One is a voltage-mode filter which employs five CCII and the second is a current-mode filter which employs four two-output CCIIIs. They have very low passive and active sensitivities and employ grounded resistors and capacitors.

References