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Right-Direct (BI) Type Wide Bandpass SC Ladder Filter with Compensation for Finite Amplifier Gain and Offset Voltage

Nikolay Radev, Kantcho Ivanov, and Kalin Stanchev

Abstract: In this paper gain- and offset-compensated (GOC) modification of a sixth-order right-direct (BI) type wide bandpass switched-capacitor (SC) ladder filter is proposed. It is based on the use of simple and fast operational amplifiers (op amps) with low but precisely known and stable dc gain A. At first, the conventional integrators in the filter are replaced with GOC integrators and the unswitched capacitors in the capacitive loops are split into two capacitors. Subsequently, the nominal op amps gain value A_0 is taking into account in the capacitance sizing of some appropriately chosen capacitors.

Keywords: Filters, gain- and offset- compensation, operational amplifiers, switched-capacitor integrators

1 Introduction

SWITCHED-CAPACITOR (SC) filter structures, which simulate passive RLC ladders are often employed where low sensitivity to element value deviation is important. Among various design approaches the leapfrog ladder and coupled biquad methods are the most popular [1, 2]. One another alternative is the LUD ladder simulation method, which is based upon LU matrix decomposition technique [3]. In [4] a large family of circuit structures is revealed depending on the choice of matrix decomposition, including the existing leapfrog, coupled biquad and LUD ones as specific cases. All circuits are insensitive to parasitic capacitances. A sixth-order wide bandpass passive ladder prototype is simulated by the following four

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The authors are with Dept. Theory of Theoretical Electrical Engineering, Technical University of Sofia, 8, Kliment Ohridski Blvd., 1000 Sofia, Bulgaria (e-mails: [nradev, ivanovkp, kstan]@tu-sofia.bg).

matrix decomposition methods: left-LUD, left-direct (IA), right-LUD and right-direct (BI). The corresponding SC filter realizations are compared in term of total capacitance and capacitance spread.

SC networks with low sensitivity to operational amplifiers (op amps) characteristics such as finite dc gain A and input-referred offset voltage Vos are sometimes required, e. g., when op amp gain has to be sacrificed for bandwidth. For this reason several gain- and offset-compensated (GOC) SC building blocks (integrators, gain stages, sample-and-hold circuits) have been reported in the literature.

In this paper a GOC modification of the sixth-order right-direct (BI) type wide bandpass SC ladder filter from [4] is proposed. It is based on the use of simple and fast amplifiers with low but precisely known and stable dc gain [5].

In a first step, the conventional integrators in the filter are replaced with GOC integrators and the unswitched capacitors in the capacitive loops, around the op amps outputs and inputs, are split into two capacitors.

In a second step, the nominal op amps gain value A_0 is taking into account in the capacitance sizing of the capacitors, connected to the uncompensated outputs of the integrators.

In a third step, the precise op amp gain (POG) approach is used to minimize the gain and the phase errors of the source and load terminations. The gain errors of the remaining GOC SC integrators are reduced by modifying the values of the integrating capacitances. The variation of the op amps dc gain A from its nominal value A_0 is taken into account.

2 Right-direct (BI) type SC ladder filter with conventional integrators

The circuit schema of the sixth-order wide bandpass right-direct (BI) SC filter being considered is shown in Fig. 1 [4]. For the sampling frequency $f_s = 100$ kHz the ideal filter requirements are: lower passband edge 300 Hz; upper passband edge 3400 Hz; maximum passband ripple 0.3 dB; lower stopband edge 10 Hz; lower stopband attenuation > 30 dB; upper stopband edge 5000 Hz; upper stopband attenuation > 45 dB. The component values for the circuit are listed in Table 1

Table 1. Component values for the SC filter of Fig. 1

$C_1 = 3.147$	$C_2 = 9.392$	$C_3 = 1.0$	$C_4 = 9.821$	$C_5 = 1.615$
$C_6 = 1.266$	$C_7 = 3.435$	$C_8 = 4.044$	$C_9 = 5.433$	$C_{10} = 22.474$
$C_{11} = 3.736$	$C_{12} = 25.451$	$C_{13} = 1.0$	$C_{14} = 1.573$	$C_{15} = 1.473$
$C_{16} = 1.0$	$C_{17} = 2.137$	$C_{18} = 7.070$	$C_{19} = 1.0$	$C_{20} = 10.686$
$C_{21} = 1.252$	$C_{22} = 1.0$	$C_{23} = 1.0$		

Let us suppose that the capacitors and the switches are ideal. The op amps are assumed to have finite dc gain A and infinite bandwidth. This supposition is

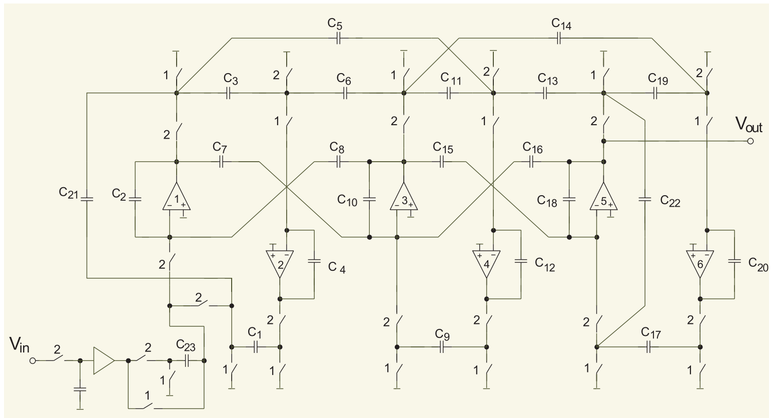
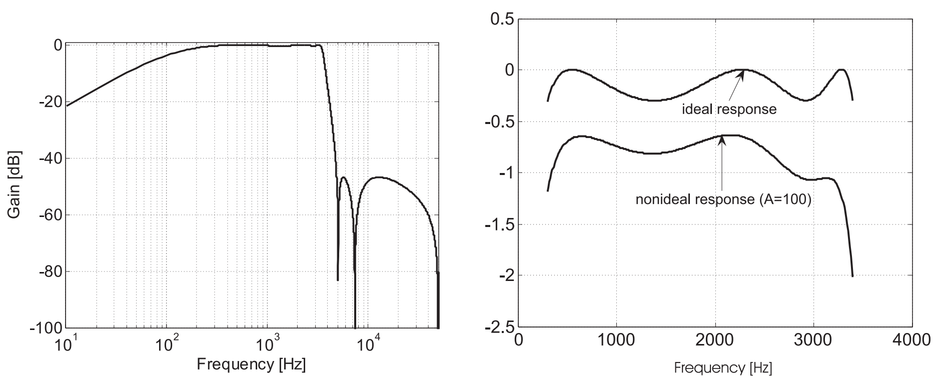


Fig. 1. Conventional sixth-order wide bandpass right-direct (BI) type SC filter.

adequate for the analysis of SC circuits containing fast and relatively low-gain amplifiers.

The filter is simulated and analyzed with MATLAB 7.1. Fig. 2a shows the ideal overall filter response. Fig. 2b gives a comparison in the passband between ideal performance and nonideal response for finite op amps dc gain $A = 100$. It is observed that the passband response is deteriorated by the op amps finite gain.



a) Overall response b) Passband response

Fig. 2. Frequency responses of the bandpass filter from Fig. 1

The input-referred op amps dc offset voltages V_{os} are modeled as voltage sources at the noninverting input terminals. The total output offset voltage $V_{out}(n)$ in steady state, for $V_{in} = 0$, is a weighted sum of the individual offset terms contributed by different paths through which an amplifier offset travels to the output. Each weighting function β_i is the transfer function from a particular offset source to the output through a particular path.

The output offset voltage of the conventional filter from Fig. 1 is given by

$$\lim_{n\to\infty} V_{out}(n) = \sum_{i=1}^{6} \frac{\alpha_i}{\Delta} V_{os_i} = \sum_{i=1}^{6} \beta_i V_{os_i} \tag{1}$$

where

$$\begin{aligned}
\alpha_1 &= \frac{1}{A} C_5 C_{14} C_{17} (C_1 + C_{21} + C_{23})(C_3 + C_6),\\
\alpha_2 &= -C_1 C_5 C_{14} C_{17} (C_3 + C_6),\\
\alpha_3 &= -\frac{1}{A} C_{14} C_{17} (C_5 + C_{11} + C_{13})\\
&\qquad \times \left\{ C_1 C_3 + \frac{1}{A}(C_3 + C_6)\left[C_{21} + \frac{1}{A}(C_1 + C_{21} + C_{23}) \right] \right\},\\
\alpha_4 &= C_{14} C_{17} (C_5 + C_{11} + C_{13})\\
&\qquad \times \left\{ C_1 C_3 + \frac{1}{A}(C_3 + C_6)\left[C_{21} + \frac{1}{A}(C_1 + C_{21} + C_{23}) \right] \right\},\\
\alpha_5 &= \frac{1}{A} C_1 (C_{14} + C_{19})(C_{17} + C_{22})(C_3 C_{11} - C_5 C_6),\\
\alpha_6 &= -C_{17}(C_{14} + C_{19}) \left\{ C_{11} \left[C_1 C_3 + \frac{1}{A} C_{21}(C_3 + C_6) \right] - C_1 C_5 C_6 \right\},\\
\Delta &= C_1 C_{17} \left[C_3 (C_{11} C_{19} - C_{13} C_{14}) - C_5 C_6 C_{19} \right]\\
&\quad + \frac{1}{A} \left[C_{17} C_{21} (C_3 + C_6)(C_{11} C_{19} - C_{13} C_{14}) + C_1 C_{22} (C_{14} + C_{19})(C_3 C_{11} - C_5 C_6) \right]\\
&\quad + \frac{1}{A^2} \left[C_{17} (C_3 + C_6)(C_1 + C_{21} + C_{23})(C_{11} C_{19} - C_{13} C_{14}) \right.\\
&\quad + C_1 (C_{14} + C_{19})(C_{17} + C_{22})(C_3 C_{11} - C_5 C_6) + C_{11} C_{21} C_{22} (C_3 + C_6)(C_{14} + C_{19})\\
&\quad \left. + C_1 C_3 C_{17} C_{19} (C_5 + C_{11} + C_{13}) \right] + \frac{1}{A^3} C_{11} C_{21} (C_3 + C_6)(C_{14} + C_{19}(C_{17} + C_{22}).
\end{aligned} \tag{2}$$

For $A = 100$ one obtains

$$\begin{aligned}
\lim_{n\to\infty} V_{out}(n) =& 0.614 V_{OS_1} - 35.80 V_{OS_2}\\
& - 0.6275 V_{OS_3} + 62.71 V_{OS_4} - 30.406 V_{OS_5} - 27.63 V_{OS_6}.
\end{aligned} \tag{3}$$

It is seen that the weighting functions β_2, β_4 and β_6 are inadmissibly large. The output offset voltage may become a significant limitation to the permissible signal swing.

3 Gain- and offset-compensated right-direct (BI) type SC ladder filter

At first, for reducing the effect of op amps imperfections (finite dc gain A and offset voltage V_{os}) the conventional integrators in the SC filter from Fig. 1 are replaced with GOC integrators. The integrators in the source and load terminations (with the op amps 1 and 5) and the third integrator are replaced with Ki-89 GOC integrators [6]. The second, the fourth and the sixth integrators are replaced with Nagaraj-86 GOC integrators [7].

The capacitors C_7 , C_8 , C_{15} and C_{16} in the unswitched capacitive loops are split into two capacitors. The resul ting filter is shown in Fig. 3, where $C_{h1} = C_2$, $C_{h2} = 1$, $C_{h3} = C_{10}$, $C_{h4} = 1$, $C_{h5} = C_{18}$ and $C_{h6} = 1$.

The relations between the output voltages of the op amps 1, 3 and 5 during the uncompensated phase 1 and during the compensated phase 2 of the clock period are frequency independent and given by the expressions

$$V_{o_i}^1(n-0.5) = \frac{1+1/A_0}{1+2/A_0} V_{o_i}^2(n-1), \qquad i = 1,3,5 \tag{4}$$

where A_0 is the nominal value of the op amp dc gain.

The uncompensated output voltages $V_{o_i}^1(n-0.5)$ are stored in the newly capacitors C_7', C_8', C_{15}' and C_{16}' during the phase 1. The charges transferred to C_7, C_8, C_{15} and C_{16} during the phase 2 and the charges transferred to C_7', C_8', C_{15}', and C_{16}' during the phase 1 can be equated by capacitance sizing of the newly capacitors.

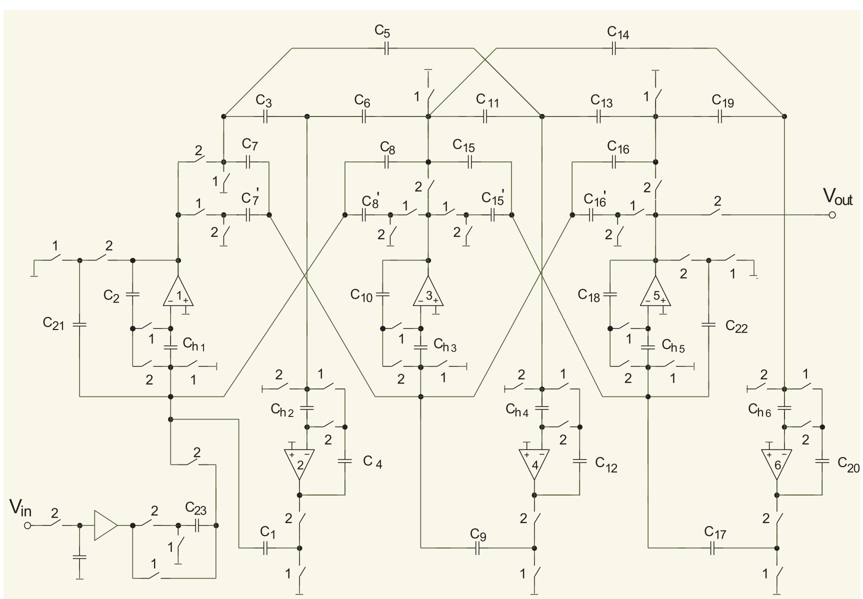


Fig. 3. GOC sixth-order wide bandpass right-direct (BI) type SC filter.

From (4) one obtains

$$C'_7 = C_7 \frac{1+2/A_0}{1+1/A_0}, \qquad C'_8 = C_8 \frac{1+2/A_0}{1+1/A_0},$$
$$C'_{15} = C_{15} \frac{1+2/A_0}{1+1/A_0}, \qquad C'_{16} = C_{16} \frac{1+2/A_0}{1+1/A_0}. \tag{5}$$

Subsequently, the POG approach is used to minimize the gain and the phase errors of the source and the load terminations during the phase 2. This is achieved by modifying the feedback capacitances C_2, C_{18} and the damping capacitances C_{21}, C_{22} on the basis of the relationships [8].

$$\begin{aligned}
C_{2p} &= C_2 \left(1+\frac{1}{A_0}\right)^{-1} - \frac{1}{A_0}(C_1+C_8+C'_8+C_{21}+C_{23}) \left(1+\frac{1}{A_0}\right)^{-2} \\
C_{18p} &= C_{18} \left(1+\frac{1}{A_0}\right)^{-1} - \frac{1}{A_0}(C_{15}+C'_{15}+C_{17}+C_{22}) \left(1+\frac{1}{A_0}\right)^{-2} \\
C_{21p} &= C_2 \left[C_{21}\left(1+\frac{2}{A_0}\right) - \frac{1}{A_0^2}(C_1+C_8+C'_8+C_{23})\right] \left(1+\frac{1}{A_0}\right)^{-2} \\
C_{22p} &= C_{22} \left[C_{22}\left(1+\frac{2}{A_0}\right) - \frac{1}{A_0^2}(C_{15}+C'_{15}+C_{17})\right] \left(1+\frac{1}{A_0}\right)^{-2}.
\end{aligned} \tag{6}$$

Finally, the gain errors of the remaining four GOC integrators during the phase 2 are reduced by modifying the integrating capacitances according to the expressions [9]

$$\begin{aligned}
C_{4p} &= \left(C_4 - \frac{C_{h2}+C_3+C_6}{A_0}\right)\left(1+\frac{1}{A_0}\right)^{-1}, \\
C_{10p} &= \left(C_{10} - \frac{C_7+C'_7+C_9+C_{16}+C'_{16}}{A_0}\right)\left(1+\frac{1}{A_0}\right)^{-1}, \\
C_{12p} &= \left(C_{12} - \frac{C_{h4}+C_5+C_{11}+C_{13}}{A_0}\right)\left(1+\frac{1}{A_0}\right)^{-1}, \\
C_{20p} &= \left(C_{20} - \frac{C_{h6}+C_{14}+C_{19}}{A_0}\right)\left(1+\frac{1}{A_0}\right)^{-1}.
\end{aligned} \tag{7}$$

From the data in Table 1 and expressions (5) and (6), for $A_0 = 100$, one finds

$$C'_7 = 3.469,\ C'_8 = 4.084,\ ,C'_{15} = 1.488,\ ,C'_{16} = 1.01,$$
$$C_{2p} = 9.166,\ C_{18p} = 6.940,\ C_{21p} = 1.251,\ C_{22p} = 0.9994022.$$

The capacitance C_{22} can be maid equal to the unit capacitance. Then, the new values of the capacitances C_{15}, C'_{15}, C_{17}, C_{18} and C_{22}, which are all incident on the "super-virtual ground" node of the load termination, are $C_{15} = 1.474$, $C'_{15} = 1.489$, $C_{17} = 2.138$, $C_{18} = 6.944$, $C_{22} = 1$.

From (7) one obtains

$$C_{4p} = 9.691,\ C_{10p} = 22.109,\ C_{12p} = 25.126,\ C_{20p} = 10.545.$$

The final capacitance values for the SC filter of Fig. 3 are listed in Table 2

Table 2. Component values for the GOC filter of Fig. 3

$C_1 = 3.147$	$C_2 = 9.166$	$C_3 = 1.0$	$C_4 = 9.691$	$C_5 = 1.615$
$C_6 = 1.266$	$C_7 = 3.435$	$C'_7 = 3.469$	$C_8 = 4.044$	$C'_8 = 4.084$
$C_9 = 5.433$	$C_{10} = 22.109$	$C_{11} = 3.736$	$C_{12} = 25.126$	$C_{13} = 1.0$
$C_{14} = 1.573$	$C_{15} = 1.474$	$C'_{15} = 1.489$	$C_{16} = 1.0$	$C'_{16} = 1.01$
$C_{17} = 2.138$	$C_{18} = 6.944$	$C_{19} = 1.0$	$C_{20} = 10.545$	$C_{21} = 1.251$
$C_{22} = 1.0$	$C_{23} = 1.0$	$C_{h1} = C_2$	$C_{h2} = 1.0$	$C_{h3} = C10$
	$C_{h4} = 1.0$	$C_{h5} = C_{18}$	$C_{h6} = 1.0$	

The passband responses of the filter from Fig. 3 for the modified capacitance values (Table 2) and op amps gain variation $A = 100 \pm 8$ [5] are shown in Fig. 4. It is obvious that these responses follow much more closely the ideal response than that of the conventional filter from Fig. 1.

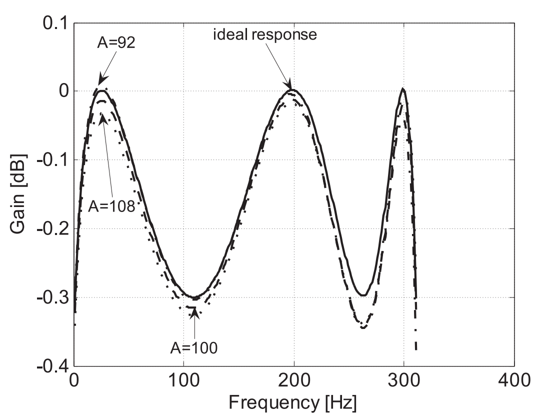


Fig. 4. Passband responses of the bandpass filter from Fig. 3 with op amps gain variation.

The output offset voltage for $A = 100$ is

$$\lim_{n\to\infty} V_{out}(n) = -0.005V_{OS_1} - 0.480V_{OS_2} + 0.008V_{OS_3} + 0.833V_{OS_4} - 0.0012V_{OS_5} - 0.363V_{OS_6}. \quad (8)$$

In this case the analytical expressions for the weighting functions α_2, α_4, α_6 and the determinant Δ from (1) are

$$
\begin{aligned}
\alpha_2 &= -\frac{C_1C_5C_{14}C_{17}(C_3+C_6)}{A(1+1/A)} \\
\alpha_4 &= \frac{C_1C_3C_{14}C_{17}(C_5+C_{11}+C_{13})}{A(1+1/A)} \\
\alpha_6 &= -\frac{C_1C_{17}(C_{14}+C_{19})(C_3C_{11}-C_5C_6)}{A(1+1/A)} \\
\Delta &= C_1C_{17}[C_3(C_{11}C_{19}-C_{13}C_{14})-C_5C_6C_{19}] \\
&\qquad + \frac{1}{A^2(1+1/A)}[C_{17}C_{21}(C_3+C_6)(C_{11}C_{19}-C_{13}C_{14}) \\
&\qquad + C_1C_{22}(C_{14}+C_{19})(C_3C_{11}-C_5C_6)].
\end{aligned}
\tag{9}
$$

By comparing the expressions for α_2, α_4 and α_6 from (9) and (2) it is seen that the values of these weighting functions for the GOC filter are approximately A times smaller than the corresponding values for the conventional filter.

4 Conclusion

An approach for reducing the effects of op amps finite gain and offset voltage in right-direct (BI) type wide bandpass SC ladder filter has been presented. It permits the substitution of narrow-bandwidth high-gain multiple-stage amplifiers by simple wide-bandwidth single-stage amplifiers with moderate gain. The passband response of the gain- and offset- compensated filter with changed topology and modified capacitances follows much more closely the ideal response than that of the conventional filter. The influence of the input-referred operational amplifiers dc offset voltages is also reduced.

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