New current-mode special function continuous-time active filters employing only OTAs and OPAMPs

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New current–mode special function continuous-time active filters employing only OTAs and OPAMPs

ERDEM SERKAN ERDOGAN†, RASIT ONUR TOPALOGLU‡, HAKAN KUNTMAN§ and OGUZHAN CICEKOGLU*†

This paper reports three current mode second order filters, each of which realizes a specific function without any external passive elements. These filters realize low-pass notch (LPN), high-pass notch (HPN) and all-pass (AP) functions. Two OPAMPs, a double output OTA and a single output OTA are employed for each circuit. The filter structures can be easily cascaded since they have high output impedances. This property is especially useful for achieving high-order filters using these LPN and HPN filters as building blocks. The presented theory is verified with macro models in SPICE simulations and, using the SPICE parameters of the layout technology, post layout simulations are carried out, with parasitics extracted from the layouts of the filter chips.

1. Introduction

Analogue continuous time active filters utilizing an operational amplifier (OPAMP) pole and the transconductance control property of the operational transconductance amplifier (OTA) have received considerable attention recently. These filters do not need to employ additional passive elements, and are therefore sometimes called active-only filters. The major advantage of these circuits is the elimination of passive elements that may result in a reduction of chip area for integrated circuit implementations. Having multiple functions in a single circuit is especially useful since the same topology can be used for different filter functions. A growing number of publications exists in the literature on OPAMP and OTA-only filters (Tsukutani et al. 1996, Abuelma’atti and Alzaher 1997, Singh and Senani 1998, Tsukutani et al. 1999, 2000a, b, c, 2001a, b).

In active-only filters, current-mode circuits are becoming more popular since they have many advantages over their voltage-mode counterparts. One of these advantages is that they have higher bandwidths. Another advantage is their higher dynamic ranges. For these types of filters, easy voltage or current control of filter parameters is important because, due to the integration process tolerances, filter characteristics deviate from the desired values. In fact, all integrated filters must be tuned after fabrication. The electronic tuning capability of these filters is achieved...
by adjusting them through the compensation capacitor (bandwidth) of the OPAMPs (Singh and Senani 1998) and/or the transconductance $g_m$ of the OTAs (Tsukutani et al. 1996, Abuelma’atti and Alzaher 1997, Tsukutani et al. 1999, 2000a, b, c. 2001a, b).

This study offers new topologies in current-mode active-only multifunction filter implementation. Three new current mode active-only filters are proposed. By cascading the proposed filters, which implement LPN and HPN functions, higher order band-pass and band-stop filter functions can be obtained. An AP function, which is implemented by the third filter circuit, is also introduced to give the chip designer more choices, depending on the application, where a phase correction may be needed.

All of the circuits are tested with SPICE simulations using both ideal and MOSFET-based OPAMP and OTA models. Layouts of the MOSFET based OPAMP and OTA cells are composed with SCnOS 2μm technology. The filter chips are implemented with these cells. Using the SPICE parameters of the layout technology, post-layout simulations are performed, with the parasitics extracted from the chip layouts. All the results are included for verification. Simulation results show that filter characteristics are in quite good agreement with theory.

2. The OTA and OPAMP models

Ideally, the OTA is assumed as an ideal voltage-controlled current source. The $g_m$ (transconductance gain), which is used to relate output current to input voltage, is a function of the bias current, $I_A$. For the case of OTAs using MOS transistors in the saturation region, the $g_m$s are proportional to $\sqrt{I_A}$; for MOS transistors operating in the weak inversion region, or for bipolar transistors, the $g_m$s are directly proportional to $I_A$. DO-OTAs (double output OTAs) are used in the circuit schematics. They differ from OTAs in that they have two outputs with separately adjustable $g_m$s.

The OTAs used in our circuits through figures 1 to 3 are designed to give out current at the same phase angle as the differential input voltage.

![Figure 1. Circuit 1, high-pass notch.](image-url)
The OPAMP, on the other hand, can be modelled by a single pole model, which can be written as $B/s$ for the operating range of frequencies, that is to say, between the first and second poles in the frequency domain. This model of the OPAMP is valid from a few kilohertz to a few megahertz. In this frequency range a bipolar monolithic OTA works as an ideal device.

3. The proposed filters

The proposed second-order filters are shown in figures 1 to 3. Circuit 1 (shown in figure 1) implements the high-pass notch function while circuit 2 (in figure 2) implements the low-pass notch function. Circuit 3 (in figure 3) realizes the all-pass function. Angular resonant frequency and quality factor, denoted by $\omega_P$ and $Q$ respectively, are independently adjustable by means of $B_1$, $B_2$, the gain–bandwidth products of both OPAMPs, assuming the open-loop gain $A(s)$ has the form of
\( A(s) = B/s \), and also by the \( g_m \) parameters belonging to the OTAs in the circuits. No component-matching constraints are imposed for the first two circuits unless one wants to have 0 dB gain for the responses at pass-bands. The filter transfer functions \( T(s) = I_{\text{out}}/I_{\text{in}} \), are given by the following equations.

**Figure 1** (high-pass notch function, circuit 1):

\[
T_{\text{HPN}}(s) = \frac{I_{\text{out}}}{I_{\text{in}}} = \frac{A_{\text{HPN}}(s^2 + \omega_0^2)}{s^2 + (\omega_p/Q)s + \omega_p^2} = -\frac{g_{m3}(s^2 + B_1B_2)}{g_{m1}s^2 + g_{m2}B_1s + B_1B_2(g_{m1} + g_{m2})}
\]

The angular resonant frequency, quality factor and pass-band gain, denoted by \( A_{\text{HPN}} \), are given by

\[
\omega_p = \sqrt{\frac{(g_{m1} + g_{m2})B_1B_2}{g_{m1}}} \quad \omega_0 = \sqrt{B_1B_2}
\]

\[
Q = \frac{1}{g_{m2}} \sqrt{\frac{g_{m1}(g_{m1} + g_{m2})B_1}{B_2}} \quad A_{\text{HPN}} = -\frac{g_{m3}}{g_{m1}}
\]

The active sensitivities of the circuit are expressed as

\[
S_{\omega_p}^{B_1} = S_{\omega_p}^{B_2} = S_{\omega_p}^{\omega_0} = S_{\omega_p}^{\omega_0} = S_{\omega_0}^{Q} = S_{\omega_0}^{Q} = -S_{\omega_0}^{A_{\text{HPN}}} = S_{\omega_0}^{A_{\text{HPN}}} = 1
\]

Thus, all sensitivities are no more than unity.

**Figure 2** (low-pass notch function, circuit 2):

\[
T_{\text{LPN}}(s) = \frac{I_{\text{out}}}{I_{\text{in}}} = \frac{A_{\text{LPN}}(s^2 + \omega_0^2)}{s^2 + (\omega_p/Q)s + \omega_p^2} = -\frac{g_{m3}(s^2 + B_1B_2)}{(g_{m1} + g_{m2})s^2 + g_{m2}B_1s + B_1B_2g_{m1}}
\]

The angular resonant frequency, quality factor and pass-band gain, denoted by \( A_{\text{LPN}} \), are given by

\[
\omega_p = \sqrt{\frac{g_{m1}B_1B_2}{(g_{m1} + g_{m2})}} \quad \omega_0 = \sqrt{B_1B_2}
\]

\[
Q = \frac{1}{g_{m2}} \sqrt{\frac{g_{m1}(g_{m1} + g_{m2})B_1}{B_2}} \quad A_{\text{LPN}} = -\frac{g_{m3}}{g_{m1} + g_{m2}}
\]
The active sensitivities of the circuit are expressed as

\[ S_{B_1}^{\alpha_p} = S_{B_2}^{\alpha_p} = S_{B_1}^{\alpha_Q} = S_{B_2}^{\alpha_Q} = -S_{B_2}^{Q} = \frac{1}{2} \quad (2d) \]

\[ S_{gm1}^{\alpha_p} = -S_{gm2}^{\alpha_p} = \frac{g_{m2}}{2(g_{m1} + g_{m2})} \quad (2e) \]

\[ S_{gm1}^{\alpha_Q} = -S_{gm2}^{\alpha_Q} = \frac{2g_{m1} + g_{m2}}{2(g_{m1} + g_{m2})} \quad (2f) \]

\[ S_{gm3}^{AP} = 1 \quad S_{gm1}^{AP} = -\frac{g_{m1}}{g_{m1} + g_{m2}} \quad S_{gm2}^{AP} = -\frac{g_{m2}}{g_{m1} + g_{m2}} \quad (2g) \]

Thus, all sensitivities are no more than unity or can be made smaller than unity.

Figure 3 (all-pass function, circuit 3):

\[ T_{AP}(s) = \frac{I_{out}}{I_{in}} = \frac{A_{AP}(s^2 - (\omega_p/Q)s + \omega_p^2)}{s^2 + (\omega_p/Q)s + \omega_p^2} = -\frac{g_{m3}(s^2 - B_1s + B_1B_2)}{g_{m2}s^2 + (g_{m1}B_1 - g_{m2}B_1)s + B_1B_2g_{m2}} \quad (3a) \]

The delay is given by

\[ \tau_{AP}(\omega) = \frac{2(\omega_p^2 + \omega^2)}{Q\omega_p((\omega_p^2 - \omega^2)\omega_p^2 + \omega^2/Q^2)} = \frac{2B_1g_{m2}(g_{m1} - g_{m2})(\omega^2 + B_1B_2)}{(g_{m2}\omega^2(\omega^2 - 2B_1B_2) + B_1^2B_2^2g_{m2}^2 + (g_{m1} - g_{m2})\omega^2))} \quad (3b) \]

The angular resonant frequency, quality factor and pass-band gain, denoted by \( A_{AP} \), are given by

\[ \omega_p = \sqrt{B_1B_2} \quad (3c) \]

\[ Q = \frac{g_{m2}}{(g_{m1} - g_{m2})} \sqrt{\frac{B_2}{B_1}} \quad A_{AP} = -\frac{g_{m3}}{g_{m2}} \quad (3d) \]

The active sensitivities of the circuit are expressed as

\[ S_{B_1}^{\alpha_p} = S_{B_2}^{\alpha_p} = S_{B_1}^{\alpha_Q} = -S_{B_2}^{Q} = \frac{1}{2} \quad (3e) \]

\[ -S_{gm1}^{\alpha_Q} = S_{gm2}^{Q} = \frac{g_{m1}}{g_{m1} - g_{m2}} \quad (3f) \]

\[ S_{gm3}^{AP} = -S_{gm2}^{AP} = 1 \quad (3g) \]

Thus, all sensitivities except the sensitivities of \( Q \) on \( g_{m1} \) and \( g_{m2} \) are no more than unity. However, this drawback can be tolerated for particular phase responses and \( g_m \) values, \( g_{m3} \) should be chosen equal to \( (g_{m1} - g_{m2}) \) for proper operation.

Low-pass notch (LPN) and high-pass notch (HPN) gain responses can be achieved by offsetting \( \omega_0 \) from \( \omega_p \). The attenuation is infinite at \( \omega_0 \). Combining circuits 1 and 2 gives a fourth-order band-pass filter. Higher order BP filters can
also be designed using LPNs and HPNs. The all-pass (AP) function can be used as a delay equalizer.

4. Simulation results, discussion and design example

To confirm the theoretical validity of the filters proposed in figure 1 (circuit 1), figure 2 (circuit 2) and figure 3 (circuit 3), a design example for each filter topology is given. Simulations are done according to the macro models obtained by CMOS implementations (Laker and Sansen 1994) of OPAMPs and OTAs with the PSPICE simulation program. In addition, the layouts of OTAs and OPAMPs are drawn to construct the filter chips. Post-layout simulations are carried out with the parasitics extracted from the chip layout. The CMOS implementations of OTAs and OPAMPs are shown in figures 4 and 5 respectively (Laker and Sansen

![CMOS OTA circuit.](image)

![CMOS OPAMP circuit.](image)
1994). The dimensions of the NMOS and PMOS transistors are given in tables 1 and 2. The model parameters used for SPICE simulations are illustrated in table 3. The circuits were supplied with symmetrical voltages of ±5 V. Although we have used DO-OTAs in some of our circuits schematically, we have, for convenience, used single-output OTAs with parallel-connected inputs to simulate DO-OTAs.

The filters are designed to realize a filter response with a resonant frequency $f_0$ of 417 kHz. To achieve this, the compensation capacitors of the OPAMPs are taken as 50 pF. A GBW of 417 kHz is obtained for both OPAMPs with these capacitances. If we were to build the filter for a higher frequency, we would need smaller capacitors and we could benefit from this property in IC implementations. The dependence of the OPAMP open-loop voltage gain on the biasing capacitor is obtained with the SPICE simulation program and is illustrated in figure 6. As can be seen from figure 6, the gain–bandwidth products are determined as 1.922 MHz, 1.026 MHz, 688 kHz, 519 kHz, 417 kHz and 348 kHz for compensation capacitor values of 10 pF, 20 pF, 30 pF, 40 pF, 50 pF and 60 pF respectively, with a 78.3 dB gain at low frequencies. Figure 7 shows the transadmittance gain of OTAs for different bias voltages. It is observed that reducing $g_m$, reduces the bandwidth. Open-loop transconductance

<table>
<thead>
<tr>
<th>Transistor</th>
<th>$L$ (µm)</th>
<th>$W$ (µm)</th>
<th>Transistor</th>
<th>$L$ (µm)</th>
<th>$W$ (µm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>M1</td>
<td>3</td>
<td>60</td>
<td>M8</td>
<td>3</td>
<td>12</td>
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<tr>
<td>M2</td>
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<td>M9</td>
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<tr>
<td>M3</td>
<td>3</td>
<td>12</td>
<td>M10</td>
<td>3</td>
<td>12</td>
</tr>
<tr>
<td>M4</td>
<td>3</td>
<td>12</td>
<td>M11</td>
<td>3</td>
<td>5</td>
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<tr>
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<td></td>
<td></td>
<td>M15</td>
<td>3</td>
<td>25</td>
</tr>
</tbody>
</table>

Table 1. Dimensions of transistors used in CMOS OTA.

<table>
<thead>
<tr>
<th>Transistor</th>
<th>$L$ (µm)</th>
<th>$W$ (µm)</th>
<th>Transistor</th>
<th>$L$ (µm)</th>
<th>$W$ (µm)</th>
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<td>180</td>
<td>M5</td>
<td>32</td>
<td>12</td>
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<tr>
<td>M2</td>
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<td>M6</td>
<td>10</td>
<td>392</td>
</tr>
<tr>
<td>M3</td>
<td>10</td>
<td>280</td>
<td>M7</td>
<td>10</td>
<td>232</td>
</tr>
<tr>
<td>M4</td>
<td>10</td>
<td>280</td>
<td>M8</td>
<td>10</td>
<td>39</td>
</tr>
</tbody>
</table>

Table 2. Dimensions of transistors used in CMOS OPAMP.

.Model PMOS (LEVEL = 2 LD=0.580687U TOX=432E-10 NSUB=1E16 VTO=--0.944048 +KP=18.5E-6 GAMMA=0.435 PHI=0.6 UO=271 UEXP=0.242315 UCRIT=20581.4 +DELTA=4.32096E-5 VMAX=33274.4 XJ=0.4U LAMBDA=0.0620011 TPG=1E12 NEFF=1.001 +NSS=1E12 TPG=1E12 RSH=9.925 CGDO=4.83117E-10 CGSO=4.83117E-10 CGBO=1.293E-9 +CJ=0.0001307 MJ=0.4247 CJSW=4.613E-10 MJSW=0.2185 PB=0.75 XQC=1
.Model NMOS (LEVEL=2 LD=0.414747U TOX=505E-10 +NSUB=1.35634E16 VTO=0.864893 KP=44.9E-6 GAMMA=0.981 PHI=0.6 UO=656 +UEXP=0.211012 UCRIT=107603 DELTA=3.53172E-5 VMAX=100000 XJ=0.4U +LAMBDA=0.00107351 NFS=1E11 NEFF=1.001 NSS=1E12 TPG=1 RSH=9.925 CGDO=2.83588E-10 +CGSO=2.83588E-10 CGBO=7.968E-10 CJ=0.0003924 MJ=0.456300 CJSW=5.284E-10 +MJSW=0.3199 XQC=1

Table 3. Model parameters of NMOS and PMOS transistors used for SPICE simulations.
gains of 200 $\mu$A/V, 270 $\mu$A/V, 370 $\mu$A/V and 420 $\mu$A/V are observed for $-3.73$ V, $-3.54$ V, $-3.26$ V and $-3.11$ V bias voltages respectively.

While implementing circuit-1, the bias voltage of $-3.26$ V used for all OTAs allowed us to achieve a transconductance gain of 370 $\mu$A/V. With these biasing voltages and compensation capacitance values, the pole quality factor of the filter is obtained as $Q = 1.41$. The zero frequency is found to be 416.87 kHz. Figures 8(a) and 8(b) show respectively the simulated frequency and phase responses of the proposed filter. In addition, post-layout simulations are performed on the filter circuit with the parasitics extracted from the chip layout. Figures 9(a) and 9(b)
Figure 8(a). Frequency response of proposed HPN circuit, simulated with CMOS macro models of OPAMPs and OTAs.

Figure 8(b). Phase response of proposed HPN circuit, simulated with CMOS macro models of OPAMPs and OTAs.

Figure 9(a). Frequency response of proposed HPN circuit, simulated with parasitics extracted from chip layout.
show respectively the post-simulation results of frequency and phase responses. Figure 9(c) shows the filter chip layout implemented with SCMOS 2 μm technology.

As stated before, the parameters can each be independently adjusted to any desired value without disturbing the others. Figure 10 shows the simulated HPN responses for $Q = 1.41$ and for $Q = 2.55$, both with the same zero frequency and pass-band gain. The $Q$ parameter is changed by tuning the $g_m$s of OTAs ($g_{m1} = g_{m3} = 420 \mu A/V$, $g_{m2} = 200 \mu A/V$), and it can be chosen as high as the dynamic range of the circuit allows. On the other hand, increasing the $Q$ factor much further brings instability and dynamic range problems to the circuit. These problems are important especially for voltage-mode cascaded circuits since the signal levels between the cascaded blocks may reach or exceed the supply rails because of the high $Q$ factor at certain frequencies. Therefore, the order of the cascaded circuits is
important for proper operation. For current mode circuits, as in the case of the proposed circuits, these problems are not encountered. The simulation results agree quite well with the theoretical analysis, as is shown by comparison with the ideal magnitude and phase responses also included in these figures.

Post-layout simulations are performed on circuit 2, with the parasitics using the same values of OTA bias voltages and OPAMP compensation capacitors that result in a pole quality factor $Q$ of 1.41. Figures 11(a) and 11(b), which belong to circuit 2, show the low-pass notch frequency and phase responses. Figure 11(c) shows the low-pass notch frequency response for various temperatures. It is observed that the circuit is capable of working well up to high temperature values.

For circuit 3, setting the compensation capacitor to 50 pF without changing the bias voltages of the OTAs and simulating the circuit with CMOS macro models,
we got an AP response as shown in figure 12. Phase response is also included in the figure as well as the ideal responses. 180° phase difference is observed at 416.97 kHz. Any delay can be obtained using circuit 3 by changing the \( g_m \) values and capacitors as needed.

The large signal behaviour of the high-pass notch filter circuit is tested with the post-layout simulation of the filter chip by applying a 1000 kHz (pass band) sinusoidal current signal to the input. The dependence of the total harmonic distortion on the input signal level is observed at the output. The results obtained are summarized in figure 13. It can be observed from figure 13 that the total harmonic distortion remains at acceptable levels below 30 \( \mu A \) input current, where THD is 4.65%, but increases rapidly for input current levels larger than 30 \( \mu A \). The current THD levels are still suitable for most analogue signal processing applications.
especially in receiver applications, because the signal levels are in the order of micro-
amperes or microvolts and are amplified for the output stages. Since the THD is
related to the linearity of the active devices employed in the filter circuit, a reduction
in THD is strongly expected with careful design of more linear OPAMPs and OTAs.
More linear devices also increase the dynamic range of the circuit and help to
increase the limits on filter parameters such as $Q$ factor. This task is left to the
VLSI designer and/or to our future studies.

The resonant frequency of the filters can be adjusted by varying the bias voltages
of the OTAs. This property is important since integrated filters must be tuned.
Current- or voltage-controlled parameters make the filter suitable for on-chip tuning
techniques. The filters have the ability to work well up to 100 MHz, a value which is
open to improvement.

The power supply noise behaviour of the circuits is simulated with $2V_{pp}$ 100 Hz
noise signal on the $V_{DD}$ line while the input signal is suppressed. Figure 14 shows
the output of circuit 1 due to power supply noise. The output current level is smaller

![Frequency response of proposed AP circuit](image)

**Figure 12.** Frequency response of proposed AP circuit, simulated with CMOS macro
models of OPAMPs and OTAs.

![THD vs Input Amplitude](image)

**Figure 13.** Dependence of total harmonic distortion at output of circuit 1 on input signal
amplitude; input is a 1000 kHz sinusoidal current signal.
than 100 nA, which is a satisfactory result for the filter circuit constructed with simple OPAMP and OTA circuits. Developing more complex OPAMP and OTA circuits with self-biasing capability would give better results, but this is not a task within the scope of this paper.

5. Conclusion

This paper reports three active-only current-mode filter structures. The filter structures are easily cascaded since they have high output impedances. This property can be easily used to achieve a circuit that can implement higher order filters. Furthermore, current mode operation is expected to overcome dynamic range limitation problems due to supply rails for cascaded sections. Simulation and post-layout simulation results (performed with parasitics extracted from the layouts) are included to verify the theory presented.

Acknowledgments

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References


Figure 14. Outputs due to power supply noise signal of 2 V_{pp} at 100 Hz.


