Novel multiple function analog filter structures and a dual-mode multifunction filter

E. S. ERDOGAN†, R. O. TOPALOGLU†, O. CICEKOGLU†,
H. KUNTMAN*‡ and A. MORGUL†

†Department of Electrical and Electronics Engineering, Bogazici University,
Bebek-Istanbul, Turkey
‡Faculty of Electrical and Electronics Eng., Istanbul Technical University,
34469, Maslak-Istanbul, Turkey

(Received in final form 23 March 2006)

This paper reports three multi-function filters each of which realizes at least three basic functions without any external passive elements. Depending on the circuit, from one to four operational transconductance amplifiers (OTAs), and two operational amplifiers (OPAMPs) are employed. All three circuits are transimpedance-mode circuits; one of them can also operate as a current-mode filter. Therefore this filter represents a dual-mode, multifunction filter. The presented theory is verified with macro models in simulation program with integrated circuit emphasis (SPICE) simulations and post layout simulations, which are carried out with parasitics extracted from the layouts of the filter chips.

Keywords: Active only; Multifunction filter; Current-mode; Transimpedance mode

1. Introduction

Continuous time active filters find wide application in communication instrumentation. Traditionally active filters employ RC passive components and operational amplifiers. Active-R and active-C filters have also attracted attention. It is possible however to implement active filters using operational amplifiers and OTAs only. No other passive elements are required since the OTA imposes a relation between current and voltage variables, simulating the resistance. The operational amplifier can be used as an integrator (Tsuktani et al. 1996, 1999, 2000a,b,c, 2001a,b, Abuelma’atti et al. 1997, Abuelma’atti and Alzaher 1997, 1998a,b, Singh and Senani 1998, Minaei et al. 2003a,b,c, 2005, Topaloglu 2003). These filters can be used at higher frequencies compared to active RC counterparts. Very few publications exist in the literature on OPAMP and OTA active only current-mode circuits (Tsuktani et al. 1996, 1999, 2000a, Abuelma’atti and Alzaher 1997). For these types of filters the filter characteristics can be electronically tuned through the transconductance, \( g_m \) of the OTAs and/or the compensation capacitor of the OPAMPs. Most of the presented topologies in the literature are either current-mode (Tsuktani et al. 1996, 1999, 2000a, Abuelma’atti and Alzaher 1997, 1998b,
Minaei et al. 2003a,b,c, Topalog˘lu 2003) or voltage-mode (Tsuktani et al. 2000b,c, 2001a,b, Minaei et al. 2003a). It is possible, however, to design transimpedance-mode and transadmittance-mode filters. The port signals are current input and voltage output for the former and voltage input and current output for the latter. Therefore definition of four different mode filters in possible. Each type of filter has its own practical application. Employing a suitable mode filter the source and load signal requirements are easily satisfied. For example, in cases where the source is a Gilberts multiplier or DAC, the output signal is available as current where a voltage signal is usually needed. For many applications, traditionally the current output is converted to a voltage signal by some extra hardware and then processed by a voltage mode filter. However a transimpedance type of filter will perform this conversion and filtering function simultaneously saving chip area and reducing cost. One application area of the transresistive type filters is processing of signals generated by sensors. For example photodiodes convert photons to current which needs a transresistive type amplifier in order to be converted to a buffered and more easily read output voltage. Some other active-only circuit applications are also presented in the literature (Singh and Senani 1998, Abuelma’atti and Alzaher 1998, Abuelmaatti et al. 1998). A transimpedance-mode active only multifunction filter was proposed in a recent work (Minaei et al. 2005). However, the resulting topology contains excess number of active elements.

This study reports topologies in transimpedance-mode active only multifunction filters employing reduced number of active elements. Moreover one of the presented circuits allows a possibility of current-mode operation in addition to the transimpedance-mode operation, which gives to the design engineer flexibility depending on the application. Therefore this filter is called a dual-mode, multifunction filter. The circuits are simulated with macro models of OPAMPs and OTAs in SPICE. Also layouts of the OPAMPs and OTAs are drawn to construct the filter chips. Post layout simulations are done with the parasitics extracted from the chip layouts. Simulation results show that filter characteristics are in quite good agreement with theory.

2. Active components: the OTA and OPAMP models

The ideal OTA is an ideal voltage-controlled current source and the transconductance gain \( g_m \) relating the output current to the input voltage, is a function of the bias current, \( I_A \). For the case of OTAs using metal oxide semiconductor (MOS) transistors in saturation, the \( g_m \)'s are proportional to \( \sqrt{I_A} \); for MOS transistors operating in weak inversion or bipolar transistors, the \( g_m \)'s are directly proportional to \( I_A \). We have used dual output OTA (DO-OTA) in our schematics. They differ from OTAs in that they have two outputs with separately adjustable \( g_m \)'s. In fact two single output OTAs can be used to implement a DO-OTA. The current convention of DO-OTA in this paper is such that the output current \( i_o = g_m(V_+ - V_-) \) corresponding to the plus sign at the output of the OTA flows into the device while the output current corresponding to the minus sign flows out of the device. For the dual output OTA the second current flow is in the opposite direction. The OTAs used in circuit-3 are designed to give output current with 0 degrees of phase angle in the direction of the arrows used for current mode
operation. Likewise in transimpedance type of operation, positive voltage is present at the indicated output port for each transfer function unless there is a minus sign in front of it. The OPAMP on the other hand is 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 of bipolar monolithic OTA works as an ideal device.

3. The proposed filters

The proposed second order filters are shown in figure 1–3. Circuit-1 shown in figure 1 performs high-pass (HP), low-pass (LP) and band-pass (BP) functions. Circuit-2 shown in figure 2 performs band-stop (notch, BS), BP and LP functions. These two circuits work in transimpedance-mode of operation. Circuit-3 shown in figure 3 performs five functions; two (AP and BP) of them are operating in current-mode while the remaining three (BS, BP, LP) are operating in transimpedance-mode of operation. Thus circuit-3 is a dual-mode multifunction filter. 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 $g_m$ parameters of OTAs. No component matching constraints are imposed except for the AP response. The filter transfer functions are given by the following equations.

$$-B_2 + B_1 + g_m_1 - g_m_2 - I_{in} + V_1 = V_2 = V_3 = V_{LP}$$

Figure 1. Transimpedance-mode HP, BP, LP filter.
Figure 2. Transimpedance-mode BS, BP, LP filter.

Figure 3. The dual-mode multifunction filter: Transimpedance-mode BS, BP, LP and current-mode AP, BP responses.
Circuit-1 (figure 1): (HP, LP, BP functions)

\[
T_{HP}(s) = \frac{V_1}{I_{in}} = \frac{A_{HP} \cdot s^2}{s^2 + (w_p/Q)s + w_p^2} = \frac{-s^2}{g_{m2}s^2 + g_{m1}B_2s + B_1B_2g_{m1}} \tag{1a}
\]

\[
T_{LP}(s) = \frac{V_2}{I_{in}} = \frac{A_{LP} \cdot w_p^2}{s^2 + (w_p/Q)s + w_p^2} = \frac{B_1B_2}{g_{m2}s^2 + g_{m1}B_2s + B_1B_2g_{m1}} \tag{1b}
\]

\[
T_{BP}(s) = \frac{V_3}{I_{in}} = \frac{A_{BP}(w_p/Q)s}{s^2 + (w_p/Q)s + w_p^2} = \frac{-B_2s}{g_{m2}s^2 + g_{m1}B_2s + B_1B_2g_{m1}} \tag{1c}
\]

The angular resonant frequency, quality factor and the pass-band gain denoted by \( A \) are given by

\[
\omega_p \sqrt{\frac{g_{m1}B_1B_2}{g_{m2}}} \quad Q = \sqrt{\frac{g_{m2}B_1}{g_{m1}B_2}} \tag{1d}
\]

\[
A_{LP} = \frac{1}{g_{m1}} \quad A_{BP} = -\frac{1}{g_{m1}} \quad A_{HP} = -\frac{1}{g_{m2}} \tag{1e}
\]

The active sensitivities of the circuit are expressed as

\[
S_{B_1}^{\omega_p} = S_{B_2}^{\omega_p} = S_{g_{m1}}^{\omega_p} = -S_{g_{m2}}^{\omega_p} = S_{B_1}^{Q} = -S_{B_2}^{Q} = S_{g_{m2}}^{Q} = -S_{g_{m1}}^{Q} = \frac{1}{2} \tag{1f}
\]

thus all sensitivities are no more than unity.

Circuit-2 (figure 2): (LP, BP, BS functions)

\[
T_{BS}(s) = \frac{V_1}{I_{in}} = \frac{A_{BS}(s^2 + w_p^2)}{s^2 + (w_p/Q)s + w_p^2} = \frac{B_1B_2 + s^2}{g_{m1}(s^2 + B_1s + B_1B_2)} \tag{2a}
\]

\[
T_{BP}(s) = \frac{V_2}{I_{in}} = \frac{A_{BP}(w_p/Q)s}{s^2 + (w_p/Q)s + w_p^2} = -\frac{B_1s}{g_{m1}(s^2 + B_1s + B_1B_2)} \tag{2b}
\]

\[
T_{LP}(s) = \frac{V_3}{I_{in}} = \frac{A_{LP} \cdot w_p^2}{s^2 + (w_p/Q)s + w_p^2} = \frac{B_1B_2}{g_{m1}(s^2 + B_1s + B_1B_2)} \tag{2c}
\]

The angular resonant frequency, quality factor and the pass-band gain denoted by \( A \) are given by,

\[
\omega_p = \omega_z = \sqrt{B_1B_2} \quad Q = \sqrt{\frac{B_2}{B_1}} \tag{2d}
\]

\[
A_{LP} = \frac{1}{g_{m1}} \quad A_{BP} = -\frac{1}{g_{m1}} \quad A_{BS} = \frac{1}{g_{m1}} \tag{2e}
\]
The active sensitivities of the circuit are expressed as

$$s_{B_1}^{op} = s_{B_2}^{op} = s_{B_1}^{op} = s_{B_2}^{op} = -s_B^Q = -s_B^Q = \frac{1}{2}$$

(2f)

thus all sensitivities are also no more than unity.

Circuit-3 (figure 3): (BS, BP, LP, AP functions)

**Transimpedance-mode operation**

$$T_{BS}(s) = \frac{V_1}{I_{in}} = \frac{A_{BS}(s^2 + w_p^2)}{s^2 + (w_p/Q)s + w_p^2} = \frac{B_1B_2 + s}{g_{m2}s^2 + (g_{m1}B_1 - g_{m2}B_1)s + B_1B_2g_{m2}}$$

(3a)

$$T_{BP}(s) = \frac{V_2}{I_{in}} = \frac{A_{BP}(w_p/Q)s}{s^2 + (w_p/Q)s + w_p^2} = \frac{B_1s}{g_{m2}s^2 + (g_{m1}B_1 - g_{m2}B_1)s + B_1B_2g_{m2}}$$

(3b)

$$T_{LP}(s) = \frac{V_3}{I_{in}} = \frac{A_{LP} \cdot w_p^2}{s^2 + (w_p/Q)s + w_p^2} = \frac{B_1B_2}{g_{m2}s^2 + (g_{m1}B_1 - g_{m2}B_1)s + B_1B_2g_{m2}}$$

(3c)

The angular resonant frequency, quality factor and the pass-band gain denoted by \(A\) are given by

$$\omega_p = \omega_z = \sqrt{B_1B_2} \quad Q = \frac{g_{m2}}{g_{m1} - g_{m2}} \cdot \sqrt{\frac{B_2}{B_1}}$$

(3d)

$$\frac{A_{LP}}{g_{m2}} \quad \frac{A_{BP}}{g_{m1} - g_{m2}} \quad \frac{A_{BS}}{g_{m2}}$$

(3e)

The active sensitivities of the circuit are expressed as

$$s_B^{op} = s_B^{op} = s_B^{op} = s_B^{op} = s_B^Q = -s_B^Q = \frac{1}{2} \quad -s_{B_1}^Q = s_{B_1}^Q = \frac{g_{m1}}{g_{m1} - g_{m2}}$$

(3f)

thus all sensitivities except that are dependence on \(g_m\) parameters are no more than unity. However this deficiency applies only to the sensitivity of the \(Q\) parameter, which can be tolerated in particular applications. Furthermore this sensitivity can be kept at reasonable values for low \(Q\) applications.

**Current-mode operation**

$$T_{BP}(s) = \frac{I_{out1}}{I_{in}} = \frac{A_{BP}(w_p/Q)s}{s^2 + (w_p/Q)s + w_p^2} = \frac{g_{m3}B_1s}{g_{m2}s^2 + (g_{m1}B_1 - g_{m2}B_1)s + B_1B_2g_{m2}}$$

(4a)

$$T_{AP}(s) = \frac{I_{out2}}{I_{in}} = \frac{A_{AP}(s^2 - (w_p/Q)s + w_p^2)}{s^2 + (w_p/Q)s + w_p^2} = \frac{g_{m4}(B_1B_2 - B_1s + s^2)}{g_{m2}s^2 + (g_{m1}B_1 - g_{m2}B_1)s + B_1B_2g_{m2}}$$

(4b)
The condition for the all-pass response is $g_{m1} = 2g_{m2}$. The pass-band gains are given by

$$A_{AP} = \frac{g_{m4}}{g_{m2}} \quad A_{BP} = \frac{g_{m3}}{g_{m1} - g_{m2}}$$

(4c)

The pole frequency, $Q$ factor and the corresponding sensitivities are same as the ones given in equations (3d) and (3f).

Note that some of the voltage-mode signal outputs are taken from the OPAMP outputs thus they do not require any voltage buffers. For example LP and BP outputs of circuit-1, circuit-2, and circuit-3 exhibit low output impedance.

4. Simulation results, discussion and design example

To confirm the theoretical validity of the filter proposed in figure 1 (circuit-1), a design example was given. Simulations are done according to the macro models obtained by CMOS implementations of OPAMPs and Oas with PSPICE simulation program, where the related information is summarized in figures 4–9. Also the layouts of OTAs and OPAMPs are drawn to construct the filter chip. Post layout simulations are carried out with the parasitics extracted from the chip layout. For the layouts, SCMOS 2 $\mu$m technology is used. The CMOS implementations of OTAs and OPAMPs are shown in figure 8 and figure 9, respectively. The dimensions of the NMOS and PMOS transistors are given in table 1 and table 2. The model parameters used for SPICE simulations are illustrated in table 3. The circuits are supplied with symmetrical voltages of $\pm 5$ V. Although we have used DO-OTAs in some of our circuits schematically, we have used single output OTAs with paralle-connected inputs to simulate DO-OTAs for convenience.

<table>
<thead>
<tr>
<th>Transistor</th>
<th>$L$ ($\mu$m)</th>
<th>$W$ ($\mu$m)</th>
<th>Transistor</th>
<th>$L$ ($\mu$m)</th>
<th>$W$ ($\mu$m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>M1</td>
<td>3</td>
<td>60</td>
<td>M8</td>
<td>3</td>
<td>12</td>
</tr>
<tr>
<td>M2</td>
<td>3</td>
<td>60</td>
<td>M9</td>
<td>3</td>
<td>12</td>
</tr>
<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>
</tr>
<tr>
<td>M5</td>
<td>3</td>
<td>12</td>
<td>M12</td>
<td>3</td>
<td>5</td>
</tr>
<tr>
<td>M6</td>
<td>3</td>
<td>12</td>
<td>M13</td>
<td>3</td>
<td>5</td>
</tr>
<tr>
<td>M7</td>
<td>3</td>
<td>12</td>
<td>M14</td>
<td>3</td>
<td>5</td>
</tr>
<tr>
<td>M15</td>
<td>3</td>
<td>25</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Transistor</th>
<th>$L$ ($\mu$m)</th>
<th>$W$ ($\mu$m)</th>
<th>Transistor</th>
<th>$L$ ($\mu$m)</th>
<th>$W$ ($\mu$m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>M1</td>
<td>2</td>
<td>50</td>
<td>M5</td>
<td>3</td>
<td>6</td>
</tr>
<tr>
<td>M2</td>
<td>2</td>
<td>50</td>
<td>M6</td>
<td>2</td>
<td>207</td>
</tr>
<tr>
<td>M3</td>
<td>3</td>
<td>5</td>
<td>M7</td>
<td>2</td>
<td>175</td>
</tr>
<tr>
<td>M4</td>
<td>3</td>
<td>5</td>
<td>M8</td>
<td>2</td>
<td>33</td>
</tr>
</tbody>
</table>
The circuit is designed to realize a filter response with a natural frequency of $f_0 = 1054 \text{ kHz}$ and a $Q = 1$. The achieve this, the bias voltages of the CMOS OTAs are chosen as $-3.11 \text{ V}$ and the compensation capacitors of the OPAMPs are taken as $12 \text{ pF}$. A GBW of $1054 \text{ kHz}$ is obtained for both OPAMPs with these capacitances. Designing the filter for higher frequencies would require smaller capacitors and that simplifies the IC implementation. The dependence of the OPAMP open-loop voltage gain on the compensation capacitor is obtained with SPICE simulation. The gain-bandwidth products are determined as $2.472 \text{ MHz}$, $1.248 \text{ MHz}$, $1.054 \text{ MHz}$, $628 \text{ kHz}$, $422 \text{ kHz}$ and $310 \text{ kHz}$ for compensation capacitor values of $5 \text{ pF}$, $10 \text{ pF}$, $12 \text{ pF}$, $20 \text{ pF}$, $30 \text{ pF}$ and $40 \text{ pF}$ respectively with a $61.2 \text{ dB}$ gain at low frequencies. For operational transconductance amplifier of figure 8 open loop transconductance gain of $200 \mu \text{ A/V}$, $270 \mu \text{ A/V}$, $370 \mu \text{ A/V}$ and $420 \mu \text{ A/V}$ are observed for $/C_0 = 3.73 \text{ V}$, $/C_0 = 3.54 \text{ V}$, $/C_0 = 3.26 \text{ V}$ and $/C_0 = 3.11 \text{ V}$ bias voltages respectively.

While implementing circuit-1, the bias voltage of $-3.11 \text{ V}$ used for all OTAs allowed us to achieve a transadmittance gain of $420 \mu \text{ A/V}$. With these bias voltages and equal compensation capacitance values the pole quality factor of the filter is obtained as $Q = 1$. Circuit-1 is simulated with the parasitics extracted from the chip layout, which gives an idea of the real chip behaviour. Figure 4a shows the post layout simulation of LP, BP and HP frequency responses of the proposed filter as well as a comparison with ideal results. Figure 4b shows the filter chip layout implemented with SCMOS $2 \mu \text{ m}$ technology. The simulation results agree quite well with the theoretical analysis.

Figure 5 shows the LP frequency responses for various compensation capacitance values. $838 \text{ kHz}$, $555 \text{ kHz}$ and $413 \text{ kHz}$ cutoff frequencies are observed for $20 \text{ pF}$, $30 \text{ pF}$ and $40 \text{ pF}$ capacitors respectively. A pass-band transresistance gain of $2374 \Omega$ is observed which can be compared with $A_{LP}$ in (1e), which results in $1/g_{m1} = 1/42 \Omega = 2380 \Omega$. As an example, for a $21 \mu \text{ A}$ input current, an output voltage of $21 \mu \text{ A} \times 2374 \Omega = 50 \text{ mV}$ can be obtained at the output with these values of $g_{m1}$ and gain-bandwidth products.

The large signal behaviour of the circuit-1 is tested with post layout simulations by applying a $1000 \text{ kHz}$ sinusoidal current signal to the input, and observing the dependence of the total harmonic distortion on this input signal level at the BP output. The results obtained are shown in figure 6. It can be observed that the total

| Table 3. Model parameters of NMOS and PMOS transistors used for SPICE simulations. |
| .MODEL P PMOS (LEVEL = 2) LD = 0.0580687U 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 = 33724.4 |
| +XJ = 0.4U LAMBDA = 0.0620018 NFS = 1E11 NEF = 1.001 NSS = 1E12 TPG = -1 |
| +RSH = 10.25 CGDO = 4.83117E-10 CGSO = 4.83117E-10 CGBO = 1.293E-9 |
| +DJ = 0.0001307 CG = 0.4247 CJSW = 4.613E-10 MJSW = 0.2185 PB = 0.75 XQC = 1 |
| .MODEL N 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 = 656U EXP = 0.211012 UCRIT = 107603 DELTA = 3.53172E-5 |
| +VMAX = 100000 XJ = 0.4 U LAMBDA = 0.00107351 NFS = 1E11 NEF = 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 |
harmonic distortion remains at acceptable levels below 40 mV output voltage where THD is 1% and increases rapidly for output voltage levels above this value. These values are acceptable for most IC implementations. However for better responses, the OTA and OPAMP parameters should be improved, which is a task that is out of the scope of this paper. Various SPICE transient post layout simulations are run at these values of input levels to test the circuit. In these simulations, different values of resistive loads are tested and the circuit is found to be working satisfactorily.

The power supply noise behaviour of the circuit is simulated with 2 Vpp 100 Hz noise signal on the $V_{DD}$ line while the input signal is suppressed. Figure 7 shows the output functions of circuit-1 due to power supply noise. The output voltage levels are
smaller than 150 uV, which is a satisfactory result for the filter circuit constructed with simple OPAMP and OTA circuits. Developing more complex OPAMP and OTA circuits, which have self-biasing capability, would give better results but this is not a task in the scope of this paper.

Figure 5. Proposed low-pass response for circuit-1 designed with CMOS (OTA)s and (OPAMP)s. The compensation capacitors are taken as parameter: $C = 40 \text{ pF}, 30 \text{ pF}, 20 \text{ pF}, 12 \text{ pF}$ (with 12 pF largest cutoff frequency in the figure is obtained).

Figure 6. Total harmonic distortion at BP output of circuit-1 versus output signal amplitude, while input is a 1 MHz sinusoidal current signal.
Figure 7. Outputs due to Power Supply Noise Signal of 2 Vpp at 100 Hz.

Figure 8. COMS OTA circuit used in simulations.
Temperature dependency is a more common problem and analog filters in general suffer from this. However the integrated circuit filters must be tuned after fabrication. This tuning is for fabrication tolerances, for aging or for temperature variation. In some tuning techniques the filter is tuned continuously during operation. For example consecutive time slots are used for filtering and tuning. For some other techniques for example master-slave technique, two theoretically identical filters are employed. The current or voltage control possibility of angular resonant frequencies as well as $Q$ factor makes the presented filters suitable for these applications.

In the filter structures proposed the pole angular frequency and the quality factor $Q_P$ are determined basically by the gain-bandwidth products of two operational amplifiers. For the two stage amplifiers shown in figure 9 the gain-bandwidth product can be expressed as

$$B = \frac{g_{m1}}{2\pi C_C} = \frac{1}{2\pi C_C} \sqrt{\frac{I_{DS5}k_{in}'(W/L)}{L}}_{1-2}$$

(5)

Where $g_{m1}$ is the transconductance, $(W/L)_{1-2}$ is the aspect ratio, $k_{in}'$ is the process transconductance parameter and $I_{DS5}$ is the tail current of the input transistors. It can be clearly observed from (5) that it is possible to change $B$ by varying the tail current $I_{DS5}$ of the input pair. Note that the drain current $I_{DS5}$ of $M_5$ is provides the common source current therefore it is possible to tune by varying the biasing voltage $V_{B1}$. In a recent work it was also demonstrated that $B$ can be adjusted by changing the tail current (Minaei et al. 2005). Therefore no detailed description of tuning is given in this work.

Choosing adequate $C_C$ values to ensure the stability, it is possible for all of the topologies (Circuit-1, Circuit 2 and Circuit-2) presented in this work to adjust...
the filter pole frequency by keeping the ratio $B_1/B_2$ constant. For a specified compensation capacitor orthogonal fine tuning of pole frequency and quality factor can be achieved by varying the tail current. This makes the presented circuits attractive for IC implementation. One more advantage of this circuit comes from its nature, that is to say being a transimpedance-mode type. This gives the filter the ability to operate well for up to 100 MHz, a value open to improvement up to GHz levels by building high performance OTAs and OPAMPs. This task is left to the VLSI designer and/or to our future studies. It should be noted that, BP and LP responses of circuit-1 can be used at least beyond MHz region and HP response has a cut-off beyond 100 MHz with current configuration of OTAs and OPAMPs without any improvement.

Other circuit configurations are not tested with MOSFETs but with ideal models in SPICE and found to be operating well. In these models, OPAMP is modeled as an infinite bandwidth, integrating device while OTA is modelled simply as a VCCS. Finally, the gain factors $A_{LP}$, $A_{BP}$ etc. can be used to include a constant gain for all frequencies, that is to say amplifying the pass-bands as needed by an application without additional circuitry simply by adjusting the $g_m$ parameters.

Note that it is also possible to realize the transimpedance-mode filters using other components such as OTAs and capacitors (i.e. OTA-C filter) with reduced number of MOS transistors. If we compare the OTA-C filter topologies with the proposed filters employing OTAs and OAs it can be easily observed that the most important advantage of the circuit topologies proposed is the low output impedance which is necessary for transimpedance-mode operation. This low-output impedance condition is achieved simply since the transimpedance filter functions are taken generally from the output of OAs. A further advantage is the low capacitance values since the integrators are realized only with compensation capacitors of OAs. Also note that it is difficult to obtain these both properties with OTA-C structures which makes the topologies proposed more attractive for IC designers.

5. Conclusion

This paper reports three active-only multifunction transimpedance-mode filter structures one of them can simultaneously work as a current-mode filter. These types of filters find application in signal filtering of output signals of DACs and current output sensors. The presented dual-mode filter enables easy cascadability due to high output impedance if operated in current-mode. Furthermore all filters can drive loads without requiring a buffer if the corresponding signal is taken from the OPAMP outputs. Simulation and post layout simulation results (performed with parasitics extracted from the layouts) are included to verify the presented theory.

Acknowledgments

This work in part supported by Bogazici University research found with the project code 01X101.
References


