Current-mode MISO filter using CCCDTAs and grounded capacitors

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This paper presents the current-mode multi-input single-output (MISO) biquadratic filter using current-controlled current differencing transconductance amplifier (CCCDTA) as an active building block. The proposed circuit comprises three CCCDTAs and two grounded capacitors performing completely standard functions: low-pass, high-pass, band-pass, band-reject and all-pass functions and does not require double input current signals as well as inverting input currents. The pole frequency and quality factor of the proposed circuit can be adjusted electronically/independently via dc bias currents. In addition, the filter circuit has low-input and high-output impedance which facilitates easy connecting for current-mode circuit. The proposed circuit uses all grounded capacitors which is very suitable to further develop into an integrated circuit and without requiring any matching condition. Moreover, the active and passive sensitivities are low. The PSPICE simulation results are included to show the workability of the proposed circuit.

Keywords: Current-mode, MISO filter, CCCDTA, Grounded capacitor

1 Introduction

An analog filters are extremely useful circuits in many applications such as in communication system, signal processing, instrument and measurement system¹-⁶ etc. Now a days, current-mode active building block is useful more than voltage-mode devices wherewith it has high performance such as wide bandwidth, higher slew-rate, greater linearity, wider dynamic range, simple circuitry¹-⁶ etc. Especially, it can be used at low voltage and low power which is suitable for integrated circuit (IC) implementation for applications in mobile or wireless communication systems⁷-⁹. From the advantages mentioned above, the active filter networks are mainly referred as current-mode circuits. Principally, the modern current-mode circuits must have the following features (i) low-input and high-output impedances that are easier to use cascading or directing load¹,⁵,¹³,¹⁹,²¹,²²,²⁴, (ii) avoid the use of an external resistor due to the loss of power consumption and chip area¹⁰,¹⁴,²²-²⁵, (iii) use grounded capacitor to compensate the latent capacity at the node of circuit or output terminals and to reduce the chip area of IC¹⁰,¹⁴,²²-²⁶,²⁸-²⁹. The active filters which provide five standard filter responses including high-pass filter (HPF), low-pass filter (LPF), band-pass filter (BPF), band-rejected filter (BRF) or eliminate frequency and all-pass filter or phase-shifted signal (APF) in the same circuit topology are the main research topic now a days. Especially, various multi-input single-output (MISO) filter configurations have been paid attention in literature¹⁰-³². The review of previous MISO filter topology is in the following text. The circuits in¹¹-¹²,¹⁵,¹⁷,¹⁹,²¹,²⁷,³⁰ have high-output impedance that make them suitable for cascading or driving to load without additional current buffers. However, they use more than two external resistors which are not ideal for IC implementation⁷. Moreover, they are absent of electronic controllability for tuning the pole frequency (ω₀) and quality factor (Q). Although it can be used as digital potentiometer for instead of an external resistors but the circuits become complicated⁸. Some circuits in¹⁰,²⁰,²³ are benefited to electronic tuning of ω₀ and Q by the parasitic resistance Rₓ or the transconductance gain gₘ but they afflict from the requirement of double input current signals. In Refs.18, 25-26, 28-29 the ω₀ and Q can be electronically adjusted. Also they are not needed the
external resistors as well as matching conditions which make them easier for tuning and fabricating in IC. However, the input nodes do not exhibit low impedance which is not ideal for current-mode configuration. The current-mode MISO filter in \cite{31} can be electronically/independently controlled for \( \omega_0 \) and \( Q \). But, it uses an excessive number of active elements (five MCFTAs). The current-mode MISO filter presented in Ref. 32 has low-input and high-output impedance and can be electronically adjusted for \( \omega_0 \) and \( Q \). Unfortunately, the \( \omega_0 \) and \( Q \) are inconvenient for independently or orthogonally tuning. The proposed current-mode MISO filters are compared with previously published\cite{10-32} current-mode MISO filters and the results are presented in Table 1.

The current-mode MISO filter is presented and the proposed circuit provides the following advantageous features:

1. The output response of proposed circuit performs low-pass, high-pass, band-pass, band-reject and all-pass functions from the same circuit configuration without the requirement of double input current signals.
2. The proposed circuit has low-input and high-output impedance which is easy to cascade in high-order filter or drive load without using a buffering device.
3. The pole frequency and quality factor can be electronically/independently adjusted via dc bias current of CCCDTAs.
4. The proposed circuit uses only grounded capacitors, which is advantageous from the point of view of integrated circuit implementation and compensation of the latent capacity.
5. The active and passive sensitivities are low.

To verify the workability of the proposed current-mode MISO filter, the PSPICE simulation results of a BJT technology implementation have been included.

2 Principle of Proposed Circuit

2.1 Descriptions of CCCDTA

This section is a description of CCCDTA\cite{33}. It was proposed by Siripruchyanun and Jaikla\cite{34} which is similar to the conventional CDTA, except that CCCDTA has finite input resistances (\( R_p \) and \( R_n \)) at the \( p \) and \( n \) input terminal. The characteristic of voltages and currents of CCCDTA are shown in Eq. (1):

<table>
<thead>
<tr>
<th>Table 1 — Comparison between various MISO filter</th>
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Proposed MISO CCCDTA 3 0+2 Yes Yes No No Yes Yes
\[
\begin{bmatrix}
V_p \\
V_n \\
I_z \\
I_r
\end{bmatrix} = \begin{bmatrix}
R_p & 0 & 0 & 0 \\
0 & R_n & 0 & 0 \\
1 & -1 & 0 & 0 \\
0 & 0 & 0 & \pm g_m
\end{bmatrix} \begin{bmatrix}
I_p \\
I_n \\
V_z \\
V_r
\end{bmatrix}
\]  \hspace{1cm} (1)

For the CCCDTA implemented by a BJT technology, the parasitic resistances \( R_p \) and \( R_n \) and transconductance \( g_m \) can be expressed to be:

\[
R_p = R_n = \frac{V_T}{2I_{b1}} \quad \ldots(2)
\]

and

\[
g_m = \frac{I_{b2}}{2V_T} \quad \ldots(3)
\]

\( V_T \) is the thermal voltage. From Eqs (2) and (3), it can be seen that the parasitic resistances can be tuned by adjusting the \( dc \) bias current \( I_{b1} \) and the transconductance can be tuned by adjusting the \( dc \) bias current \( I_{b2} \). The symbol and the equivalent circuit of the CCCDTA are shown in Figs 1(a) and (b), respectively.

### 2.2 Proposed Current-mode MISO Filter

The proposed current-mode MISO filter is shown in Fig. 2. It consists of 3 CCCDTAs and 2 grounded capacitors, which is advantageous from the point of view of integrated circuit implementation since grounded capacitors can be reduced the area for ICs implementation and compensated the stay of capacitance at node or output terminals of CCCDTAs. From the CCCDTA properties, the output current can be given as:

\[
I_{out} = k \left\{ \frac{-I_{in1} \frac{C_2}{g_{m3}} s + I_{in2} \left[ \frac{C_1 C_2}{g_{m1} g_{m3}} s^2 + 1 \right] - I_{in3}}{g_{m1} \frac{C_2}{g_{m3}} s^2 + \frac{C_2 k}{g_{m3}} s + 1} \right\} \ldots(4)
\]

where \( k = \frac{g_{m2} R_{in2}}{2} \). According to Eq. (4), the selection of three input currents \( I_{in1}, I_{in2} \) and \( I_{in3} \) to obtain five standard second order filters, inverting LPF, HPF, inverting BPF, BRF and APF functions are achieved in Table 2. It is clear that the output-current filter responses obtain a standard biquadratic function without requirement of double input current signals as well as inverting input currents, that is easy to choose the output filter response without requirement of modifying circuit and additional double or inverting amplifier. Moreover, the selection can be digitally controlled by using the digital selection circuit in Ref. 35.

Furthermore, the pole frequency \( (\omega_0) \) and quality factor \( (Q) \) are written as:

\[
\omega_0 = \sqrt{\frac{g_{m1} g_{m3}}{C_1 C_2}} \quad \ldots(5)
\]

and

\[
Q = \frac{2}{g_{m2} R_{in2}} \sqrt{\frac{C_1 g_{m3}}{C_2 g_{m1}}} \quad \ldots(6)
\]
Substituting the parasitic resistance \( R_{s2} = (V_T / 2I_{g2}) \) and transconductance \( g_{m1} = (I_{g2} / 2V_T) \), \( g_{m2} = (I_{g4} / 2V_T) \) and \( g_{m3} = (I_{g6} / 2V_T) \) into Eqs (5) and (6), the \( \alpha_b \) and \( Q \) can be expressed as:

\[
\alpha_b = \frac{I_{g3}I_{g6}}{2V_T} \frac{C_1}{C_2} \quad \ldots (7)
\]

and

\[
Q = \frac{8I_{g3}}{I_{g4}} \sqrt{\frac{C_1}{C_2}} \quad \ldots (8)
\]

It is evident from Eqs (7) and (8) that the \( \alpha_b \) and \( Q \) can be electronically controlled by adjusting the \( dc \) bias currents of CCCDTAs. Moreover, if \( I_{g2} = I_{g6} = I_g \) (in practically, this can be achieved by using current mirror to copy current \( I_b \) to \( I_{g2} \) and \( I_{g6} \)), the \( \alpha_b \) can be adjusted independently and linearly without effecting the \( Q \) values. Also, the \( Q \) can be adjusted independently by changing \( I_{g3} \) and \( I_{g4} \) without disturbing \( \alpha_b \). Then, \( Q \) is higher by varying \( I_{g3} \) more than \( I_{g4} \) as described in Eqs (8) and (9).

\[
\alpha_b = \frac{I_g}{2V_T} \frac{1}{C_1C_2} \quad \ldots (9)
\]

and

\[
Q = \frac{8I_{g3}}{I_{g4}} \sqrt{\frac{C_1}{C_2}} \quad \ldots (10)
\]

Additionally, the current input nodes \( I_{in1} \) and \( I_{in3} \) exhibit low impedance by setting \( I_{B1} \) and \( I_{B3} \) as high as possible. Furthermore, the output current node \( I_{out} \) has high-impedance which is easy to directly connecting to load or cascading to high order of filters without using the current buffer devices that make it ideal for using in current-mode configuration. As a drawback, the impedance of input current node \( I_{in2} \) is not of low impedance character. It depends on the adjusting of \( R_{g2} \) and \( R_{g2} \) (ideally, they are equal). Thus, if the tuning of \( Q \) is done by adjusting \( I_{B3} \), the impedance at current input node \( I_{in2} \) is also changed. To alleviate this effect, the tuning of \( Q \) should be done by adjusting \( I_{B4} \).

2.3 Sensitivities of Proposed MISO Filter

The sensitivities of the active and passive elements of MISO filter are low as given in Eqs (11 and 12):

\[
S_{V_{in}}^{\alpha_b} = -\frac{1}{2}, S_{I_{in1},I_{in2}}^{\alpha_b} = \frac{1}{2}, S_{C_1,C_2}^{\alpha_b} = -\frac{1}{2} \quad \ldots (11)
\]

and

\[
S_{I_{in1},I_{in2}}^{Q} = -1, S_{C_1,C_2}^{Q} = \frac{1}{2}, S_{C_1,C_2}^{Q} = -\frac{1}{2} \quad \ldots (12)
\]

It is to be noted that the relative sensitivities of the proposed circuit are equal or less than unity in magnitude.

2.4 Non-ideal Effects

It is necessary to take into account the non-ideal effects of CCCDTA on proposed MISO filter, which are voltage/current tracking errors and parasitic elements. The first effect comes from tracking errors of current differencing amplifiers and transconductance amplifiers\(^{23,33}\). These tracking errors are caused by the mismatch of transistors in internal construction of CCCDTA. The characteristic of voltages and currents of CCCDTA are modified as:

\[
\begin{bmatrix}
V_P \\
V_n \\
I_z
\end{bmatrix}
= \begin{bmatrix}
R_p & 0 & 0 & 0 & 0 & I_p \\
0 & R_n & 0 & 0 & 0 & I_n \\
0 & \alpha_p - \alpha_n & 0 & 0 & V_s \\
0 & 0 & 0 & 0 & \pm \beta g_m & V_z
\end{bmatrix}
\begin{bmatrix}
I_p \\
I_n \\
V_s \\
V_z
\end{bmatrix}
\quad \ldots (13)
\]

where \( \alpha_p \) and \( \alpha_n \) are tracking error of the current transfer gains from \( p, n \) terminals to \( z \) terminal, respectively. \( \beta \) is the tracking error of the transconductance gains. The output current of the proposed circuit is shown in Fig. 2, get modified to:

\[
I_{out} = \frac{\beta_2 \alpha_p \alpha_{m2} R_{m2}}{2} \frac{C_1 C_2}{\beta \beta_3 \alpha_{p1} \alpha_{p2} g_{m1} g_{m3} s^2 + 1}
\]

\[
\begin{aligned}
&+ \frac{C_1 C_2}{\beta \beta_3 \alpha_{p1} \alpha_{p2} g_{m1} g_{m3} s^2 + 1} + \frac{1}{2} \beta_2 g_{m1} R_{m2} \frac{\alpha_{m3}}{\alpha_{p2}} \frac{C_1 C_2}{\beta \beta_3 \alpha_{p1} \alpha_{p2} g_{m1} g_{m3} s^2 + 1} + \frac{1}{\beta_2 g_{m1}} \frac{C_1 C_2}{\beta \beta_3 \alpha_{p1} \alpha_{p2} g_{m1} g_{m3} s^2 + 1} + 1
\end{aligned}
\]

\[
\ldots (14)
\]
Then, the $\omega_0$ and $Q$ of the proposed circuit become:

$$\omega_0 = \frac{1}{2V_T} \sqrt{\frac{B_0 B_2}{C_1 C_2}}, \quad \ldots (15)$$

and

$$Q = \frac{8 I_{B3}}{B_1 B_4} \sqrt{\frac{B_0 B_2}{B_1 B_2}}. \quad \ldots (16)$$

The second effect comes from parasitic elements of CCCDTA. The parasitic resistances and capacitances at high-impedance terminals affect the $\omega_0$ and $Q$ of the proposed circuit. There are $R_z, C_z$ at z terminal and $R_x, C_x$ at x terminal. The proposed circuit with parasitic elements is shown in Fig. 3. Then, the output current is obtained as:

$$\begin{align*}
I_{m1} &= k \left[ \frac{-Y_2}{g_{m3}} (1 + Y_5 R_p) \right] I_{n1} \\
&+ \frac{Y_2}{g_{m1} g_{m3}} (1 + Y_5 R_p) \left( 1 + Y_1 R_p \right) + 1 \right] I_{n2} \\
&- (1 + Y_1 R_p) I_{n3} \\
&+ k \frac{Y_2}{g_{m3}} (1 + Y_1 R_p) + 1
\end{align*}$$

... (17)

where

$$k = \frac{g_{m2} R_z}{(2 + R_z R_p) (1 + Y_5 R_p)}, \quad Y_1 = s(C_1 + C_{z1}) + G_{z1},$$

$$Y_2 = s(C_2 + C_{z2}) + G_{z2}, \quad Y_3 = Y_4 = sC_{z1} + G_{z1},$$

$$Y_5 = sC_{z2} + G_{z2} \quad \text{and} \quad Y_6 = s(C_{z2} + C_{z3}) + G_{z2} + G_{z3}.$$ 

From Eq. (16), if the operational frequency ($f_o$) is more lower than $G_{z1} R_p + 1$, $G_{z2} R_p + 1$, $2\pi C_{z1} R_p$, $2\pi C_{z2} R_p$, $G_{z1} R_p + 1$ and $R_{p2}\equiv 0$, $R_{p3}\equiv 0$ (in practically, $2\pi (C_{z2} + C_{z3}) R_p$), this can be achieved by setting $I_{B1}$ and $I_{B5}$ as high as possible or carefully designing the internal construction of CCCDTA to:

$$\begin{align*}
I_{m1} &= -I_{m2} \left( \frac{s C_{z1} + G_{z1}}{g_{m3}} + I_{n2} \right) \left( \frac{s C_{z2} + G_{z2}}{g_{m3}} + I_{n3} \right) \\
&- I_{m3} \left( \frac{s C_{z1} + G_{z1}}{g_{m3}} + I_{n2} \right) \left( \frac{s C_{z2} + G_{z2}}{g_{m3}} + I_{n3} \right)
\end{align*}$$

... (18)

Fig. 3 — Model of circuit for high frequency analysis.
where \( k = \frac{g_{m2}R_2}{2} \), \( C_1' = C_1 + C_{13} \) and \( C_2' = C_2 + C_{34} \).

The \( \omega_0 \) and \( Q \) of the proposed circuit changed to:

\[
\omega_0 = \sqrt{\frac{G_{12}G_{34} + kG_{m1}G_{34} + g_{m3}g_{m3}}{C_1'C_1}} \quad \ldots(19)
\]

and

\[
Q = \frac{g_{m3}g_{m3}}{kG_{m1}C_2' + C_1'G_{34} + C_2'G_{34}} \times \sqrt{\frac{G_{34}G_{34} + kG_{m3}G_{34} + g_{m3}g_{m3}}{C_1'C_2'}} \quad \ldots(20)
\]

3 PSPICE Simulations Results

To verify the theoretical analysis of the proposed circuit as shown in Fig. 2, it was simulated using the \textit{pnp} and \textit{nnp} transistors by the PR200N and NR200N bipolar transistors of ALA400 transistor array from AT&T. Internal construction of CCCDTA used in simulation as shown in Fig. 4. The power supply voltage was taken as ±1.5 V and the capacitors of the configuration values were chosen as \( C_1 = C_2 = 0.3\text{nF} \). The simulation set for \( Q = 1 \) and \( f_0 = 510.111 \text{kHz} \) with \( I_{B1} = 100 \mu A \), \( I_{B3} = 25 \mu A \), \( I_{B4} = 200 \mu A \), \( I_{B5} = 100 \mu A \) and \( I_{B2} = I_{B6} = I_B = 50 \mu A \). Figure 5(a) shows the frequency response of output-current gain of the BPF, LPF, HPF and BRF, dependent on selection as presented in Table 2, without modifying circuit topology. Furthermore, the simulation results of gain and phase response of APF are shown in Fig. 5 (b). The simulated pole frequency of the proposed circuit was obtained as 483.509 kHz, while the calculated value from Eq. (7) is about 510.111 kHz (deviated about 5.21%). The deviation has been affected from error of voltage, current transfer gains and the parasitic elements of the CCCDTAs. The results in Fig. 6 show the tuning of \( Q \) for the BPF response by electronics adjusting at \( Q = 1, 2.5, 4, 8 \) and 16, when kept to be \( I_{B4} = 200 \mu A \) and varied \( I_{B3} \), as 25, 50, 100, 200 and 400 \( \mu A \), respectively. Figure 7 shows the \( Q \) of the BPF response for difference values of \( I_{B4} \) as 25, 50, 100, 200 and 400 \( \mu A \), respectively, while \( I_{B3} = 25 \mu A \). The results of the \( Q \) are varied from 8, 4, 2, 1 and 0.5, respectively. These results are confirmed that \( Q \) can be electronically adjusted without affecting of pole frequency, as shown in Figs (7) and (8). Figure 8 shows the pole frequency of the BPF response varied as 118.577, 238.781, 483.059 and 990.832 kHz for...
value of $I_B$ as 12.5, 25, 50 and 100 µA, respectively. It can be seen that the pole frequency can be electronically adjusted without influence of the quality factor by dc bias current $I_B$.

The tolerance of capacitors may result in the pole frequency deviation from the requirement. In this case, a Monte Carlo analysis can be used for analyzing the error from the tolerance. The statistical results of a Monte Carlo analysis within 5% tolerances to all the capacitors by using the Gaussian probability distribution and 200 trials are shown in Fig. 9. The histograms of the possible spread of the pole frequency are shown in Fig. 9. The standard deviation of the pole frequency of BPF response is 19.310 kHz. The possibility for the maximum and minimum of the pole frequency are 573.679 kHz and 453.807 kHz, respectively.

4 Conclusions

A current-mode MISO filter using CCCDTAs and grounded capacitors has been presented. The proposed current-mode MISO filter consists of three CCCDTAs and two grounded capacitors and offers the following advantages: (i) it performs low-pass, high-pass, band-pass, band-rejected and all-pass functions from the same configuration, without requirement of double input current signals and inverting input currents (ii) the pole frequency and quality factor can be independently adjusted with bias currents, (ii) the employment of two grounded capacitors, which is attractive for either IC implementation (iv) the active and passive sensitivities are low (v) availability of explicit low-input and high-output impedances, enables easy cascading in high-order filter or driving load without external current buffers (vi) the quality factor is high possibility by adjusting the dc bias currents. The PSPICE simulation results agree well with the theoretical anticipation.

References