Tunable voltage-mode universal filter based on single DDCCTA

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Abstract
This paper presents a novel voltage-mode universal biquadratic filter performing completely five standard functions: lowpass, highpass, bandpass, bandreject and allpass functions, based on a single differential difference current conveyor transconductance amplifier (DDCCTA). The proposed configuration uses one DDCCTA, two grounded capacitors and two resistors. The filter can be used as either a high-input impedance single-input four-output multifunction filter or two-input four-output universal filter. The proposed circuit enjoys orthogonal controllability of the parameters resonance angular frequency (ω₀) and quality factor (Q), the use of only grounded capacitors, and low active and passive sensitivities. The HSPICE simulation results are included to verify the workability of the proposed filter. The total power consumption is 0.387 mW at ± 0.9 V supply voltages.

Keywords: active filter, universal biquad, voltage-mode circuits, DDCCTA

1. Introduction
Recently, a new active building block for analogue signal processing, namely, differential difference current conveyor transconductance amplifier (DDCCTA), was introduced in the literature (Pandey and Paul 2011). A DDCCTA is produced by cascading the differential difference current conveyor (DDCC) (Chiu and Horng 2007) with the operational transconductance amplifier (OTA) (Tsukutani, Sumi and Fukui 2007) in a monolithic chip for compact implementation of analogue function circuits. DDCCTA provides electronic controllability and the ability to generate various circuits. From the point of view of the advantages of low cost, space saving and power dissipation, the filters using single active element receive more attention at present. Subsequently, two high-input impedance DDCCTA-based voltage-mode multifunction biquadratic filters with a single input and three outputs, able to simultaneously realise voltage-mode highpass (HP), bandpass (BP) and lowpass (LP) filtering functions without passive component-matching conditions, were proposed in Pandey and Paul (2011), Tangsrirat
The resonance angular frequencies ($\omega_o$) and quality factors ($Q$) of these two circuits cannot be orthogonally controllable. In 2012, Channumsin et al. described a novel voltage-mode universal biquadratic filter using two DDCCTAs, two grounded capacitors and two grounded resistors that needs a component-matching condition to realise allpass (AP) filtering function. In 2013, an electronically controllable voltage-mode DDCCTA-based universal biquadratic filter was first proposed in the literature (Tangsrirat, Channumsin and Pukkalanun 2013). This circuit requires three DDCCTA active components. In 2014, Chen presented a voltage-mode DDCCTA-based universal biquadratic filter. The circuit was composed of two DDCCTAs combined with four all grounded passive components; however, the characteristic parameters $\omega_o$ and $Q$ cannot be orthogonally controllable. A circuit topology operated in voltage-mode universal biquadratic filter was created using two DDCCTAs combined with two grounded capacitors and two resistors (Chen, Wang, Huang and Hsieh 2014). This circuit joins one more important advantage of orthogonal controllability of the parameters $\omega_o$ and $Q$, but the resistors used are not all grounded. Another study revealed an improved topology that reduced the number of active components (Channumsin and Tangsrirat 2013). This circuit was composed of a single DDCCTA, two grounded capacitors and two resistors, and all the five standard biquadratic filtering functions: LP, BP, HP, bandreject (BR) and AP, can be obtained from the circuit configuration. However, this circuit needs to impose component-matching conditions for realizing BR and AP filtering functions.

From the wealth of knowledge on RC active filters, it is known that it is possible to design filter biquadratic using a single active element, two capacitors and two resistors, which are the minimum components necessary for realizing a biquadratic filtering function from the same topology. Minimal realizations are mainly desirable for three reasons: (i) minimum power consumption, (ii) reduced likelihood of circuit failure owing to element failure and (iii) reduced element cost, space saving, and parasitic effects. This paper is to present a new tunable voltage-mode DDCCTA-based universal biquadratic filter. The proposed configuration only uses a single DDCCTA, two grounded capacitors and two resistors. The filter can be used as either a high-input impedance single-input four-output multifunction filter or two-input four-output universal filter. In the operation of high-input impedance voltage-mode multifunction filter, the circuit can realise LP, HP and two BP filtering functions simultaneously and all passive components are grounded connections. In the operation of voltage-mode universal filter, the circuit can realise LP, BP, HP, BR and AP filtering functions without any passive component-matching conditions. The proposed circuit enjoys the following advantages: (i) the employment of single DDCCTA, (ii) the employment of only
grounded capacitors, (iii) orthogonal controllability of $\omega_o$ and $Q$, (iv) realization of filter transfer functions, (v) no need to impose components choice conditions, (vi) high-input impedance good for cascadeability the voltage-mode circuits, and (vii) low active and passive sensitivity performances. Table 1 compares the proposed circuit and those of previously published studies with DDCCTA-based voltage-mode multifunction/universal biquadratic filters. With respect to the filter structure in Tangsrirat and Channumsin (2011), the proposed circuit permits orthogonal controllability of the parameters $\omega_o$ and $Q$. With respect to the filter structures in Channumsin, Pukkalanun and Tangsrirat (2012), Tangsrirat, Channumsin and Pukkalanun (2013), Chen (2014), Chen, Wang, Huang and Hsieh (2014), the proposed circuit employs only one DDCCTA active component and permits orthogonal controllability of the parameters $\omega_o$ and $Q$. With respect to the filter structure in Channumsin and Tangsrirat (2013), the proposed circuit permits the implementation of all basic filtering functions LP, BP, HP, BR and AP responses without any impose component-matching conditions and permits orthogonal controllability of the parameters $\omega_o$ and $Q$.

Table 1. Comparison of the proposed circuit with previously published DDCCTA-based voltage-mode multifunction/universal filters.

<table>
<thead>
<tr>
<th>Circuits</th>
<th>Criteria</th>
<th>(i)</th>
<th>(ii)</th>
<th>(iii)</th>
<th>(iv)</th>
<th>(v)</th>
<th>(vi)</th>
<th>(vii)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tangsrirat and Channumsin (2011)</td>
<td></td>
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<td>yes</td>
<td>no</td>
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<tr>
<td>Channumsin et al. (2012)</td>
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<td>no</td>
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<td>AP</td>
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<tr>
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<td>yes</td>
<td>no</td>
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<td>nil</td>
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<tr>
<td>Chen (2014)</td>
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<td>no</td>
<td>all five</td>
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<tr>
<td>Chen et al. (2014)</td>
<td></td>
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<td>all five</td>
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<tr>
<td>Channumsin and Tangsrirat (2013)</td>
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<td>BR, AP</td>
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<td>Proposed filter type A</td>
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<tr>
<td>Proposed filter type B</td>
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<td>all five</td>
<td>nil</td>
<td>no</td>
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</tr>
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</table>
2. Circuit description

2.1. Basic concept of DDCCTA

As a current-mode active device, the DDCC has the advantages of both second-generation current conveyor (such as large signal bandwidth, great linearity, wide dynamic range) and the differential difference amplifier (such as high-input impedance and arithmetic operation capability). The DDCC has three voltage input terminals: \( Y_1 \), \( Y_2 \) and \( Y_3 \), which have high-input impedance. Terminal \( X \) is a low-impedance current input terminal. Terminal \( Z \) is a high-impedance current output terminal. The DDCCTA is a useful active building block that can be simply the circuit implementation. The DDCCTA is based on DDCC and consists of differential amplifier, current mirrors, and transconductance amplifier (Tangsrirat and Channumsin 2011, Channumsin, Pukkalanun and Tangsrirat 2012, Tangsrirat, Channumsin and Pukkalanun 2013, Chen 2014, Chen, Wang, Huang and Hsieh 2014, Channumsin and Tangsrirat 2013). This device possesses the advantages of the availability of addition and subtraction signal at the input port and electronic controllability of the transconductance gain appearing at the rear end. Figure 1 shows the electrical symbol of the DDCCTA. The port relations can be characterized by the following matrix equation (Tangsrirat, Channumsin and Pukkalanun 2013, Channumsin and Tangsrirat 2013):

\[
\begin{bmatrix}
I_{Y_1} \\
I_{Y_2} \\
I_{Y_3} \\
V_X \\
I_Z \\
I_O \\
I_{-O}
\end{bmatrix} =
\begin{bmatrix}
0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 \\
1 & -1 & 1 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 1 & 0 & 0 & 0 \\
0 & 0 & 0 & g_m & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & -g_m & 0 & 0
\end{bmatrix}
\begin{bmatrix}
V_{Y_1} \\
V_{Y_2} \\
V_{Y_3} \\
I_X \\
V_Z \\
V_O \\
V_{-O}
\end{bmatrix}
\tag{1}
\]

Figure 1. The symbol representation of the DDCCTA.
where \( g_m \) is the transconductance parameter of DDCCTA. The \( g_m \)-value is electrically controllable by an external bias current, which lends electronic controllability to design circuit parameters. It may be emphasized that electronic controllability becomes important when the circuit is in a variety of design specifications and in the integrated circuit form.

2.2. Proposed tunable voltage-mode universal biquadratic filter

The proposed DDCCTA-based voltage-mode biquadratic universal circuit is shown in Figure 2. Derived by each nodal equation of the proposed circuit, the following four output voltages can be derived as:

\[
V_{o1} = \frac{sC_2(V_{i1} + V_{i2})}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m}
\]  
(2)

\[
V_{o2} = \frac{g_m(V_{i1} + V_{i2})}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m}
\]  
(3)

\[
V_{o3} = \frac{-sC_2g_mR_1V_{i1} + (s^2C_1C_2R_1 + g_m)V_{i2}}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m}
\]  
(4)

\[
V_{o4} = \frac{s^2C_1C_2R_1(V_{i1} + V_{i2})}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m}
\]  
(5)

A. High-input impedance voltage-mode multifunction biquadratic filter

If \( V_{i2} = 0 \) (namely, the resistor \( R_2 \) is grounded) and \( V_{i1} = V_{in} \) (the input voltage signal with high-input impedance), then the following four voltage transfer functions are obtained as:

![Figure 2. Proposed voltage-mode universal biquadratic filter with single DDCCTA.](image-url)
As indicated by (6)–(9), a non-inverting BP filtering function was obtained from \( V_{o1} \), a non-inverting LP filtering function was obtained from \( V_{o2} \), an inverting BP filtering function was obtained from \( V_{o3} \) and a non-inverting HP filtering function was obtained from \( V_{o4} \). It may be noted that the input signal, \( V_{i1} = V_{in} \), is connected to the high-input impedance input node of the DDCCTA (the \( Y_3 \) terminal of the DDCCTA). So the proposed first type of filter provides high-input impedance and uses all grounded passive elements.

**B. Voltage-mode universal biquadratic filter**

If \( V_{i2} = V_{in} \) (the input voltage signal) and \( V_{i1} = 0 \) (grounded), then the following four voltage transfer functions are obtained as:

\[
\begin{align*}
V_{o1} &= \frac{sC_2}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m} \quad (6) \\
V_{o2} &= \frac{g_m}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m} \quad (7) \\
V_{o3} &= \frac{-sC_2g_mR_2}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m} \quad (8) \\
V_{o4} &= \frac{s^2C_1C_2R_1}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m} \quad (9)
\end{align*}
\]

As indicated by (10)–(13), a non-inverting BP filtering function was obtained from \( V_{o1} \), a non-inverting LP filtering function was obtained from \( V_{o2} \), a non-inverting BR filtering function was obtained from \( V_{o3} \) and a non-inverting HP filtering function was obtained from \( V_{o4} \). If \( V_{i1} = V_{i2} = V_{in} \), then the AP transfer function is easily obtained from the node of \( V_{o3} \) as

\[
\begin{align*}
V_{o3} &= \frac{s^2C_1C_2R_1 - sC_2g_mR_2 + g_m}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m} \quad (14)
\end{align*}
\]

As indicated by (10)–(14), the second type of filter is a voltage-mode universal biquadratic filter, too.

Based on the denominator polynomial of the transfer functions, the parameters \( \omega_o \) and \( Q \) of the filter are expressed as:
\[ \omega_0 = \sqrt{\frac{g_m}{R_{1}C_{2}}}, \quad Q = \frac{1}{R_{2}} \sqrt{\frac{R_{1}C_{1}}{g_mC_{2}}} \quad (15) \]

Equation (15) shows that the parameter \( Q \) can be independently tuned by using one resistor \( R_2 \) without disturbing \( \omega_0 \). In other words, parameters \( \omega_0 \) and \( Q \) are orthogonal controllable through the grounded \( R_1 \) and/or \( g_m \) and then the \( R_2 \) in that order. This is a desired property in biquadratic filters due to the offered design and tuning flexibility. It should be also mentioned at this point that the parameter \( \omega_0 \) can also be adjusted without influencing \( Q \), e.g. by simultaneously changing \( R_1 \) and \( g_m \) and keeping the product \( g_mR_1 \) constant, for equal-valued resistor and capacitor design. Thus, parameters \( \omega_0 \) and \( Q \) are restrictively orthogonal.

### 2.3. Nonideality analysis and design considerations

By taking into account the non-idealities of DDCCTA, the relationship of the terminal voltages and currents can be rewritten as

\[ V_X = \beta_1 V_{Y1} - \beta_2 V_{Y2} + \beta_3 V_{Y3}, \quad I_Z = \alpha I_X, \quad I_O = \gamma g_m V_Z \]

and \( I_O = -\eta g_m V_Z \), where \( \beta_k = 1 - \varepsilon_{vk} \) for \( k = 1, 2, 3 \), \( \alpha = 1 - \varepsilon_{ai} \), \( \gamma = 1 - \varepsilon_{gi} \) and \( \eta = 1 - \varepsilon_{qi} \). Here, \( \varepsilon_{vk} \) \((|\varepsilon_{vk}|<<1)\), \( \varepsilon_{ai} \) \((|\varepsilon_{ai}|<<1)\), \( \varepsilon_{gi} \) \((|\varepsilon_{gi}|<<1)\) and \( \varepsilon_{qi} \) \((|\varepsilon_{qi}|<<1)\) represent the voltage and current tracking errors of DDCCTA, respectively. If non-ideal current and voltage gains are considered, the parameters \( \omega_0 \) and \( Q \) of the proposed circuit in Figure 2 were obtained as

\[ \omega_0 = \sqrt{\frac{\alpha \beta_2 g_m}{R_{1}C_{2}}}, \quad Q = \frac{1}{\beta_1 \eta R_{2}} \sqrt{\frac{\beta_2 \gamma R_{1}C_{1}}{\alpha g_m C_{2}}} \quad (16) \]

The active and passive sensitivities of the proposed circuit were given as

\[ S^a_{\omega_0} = S^p_{\beta_2} = S^p_{\gamma} = S^p_{\alpha} = -S^a_{R_1} = -S^a_{\omega_0} = -S^a_{C_1} = -S^a_{C_2} = \frac{1}{2} \quad (17) \]

\[ S^Q_{\beta_2} = S^Q_{\gamma} = S^Q_{R_1} = S^Q_{C_1} = -S^Q_{\alpha} = -S^Q_{\omega_0} = -S^Q_{C_2} = \frac{1}{2} \quad (18) \]

\[ S^Q_{\beta_1} = S^Q_{\eta} = S^Q_{R_2} = -1 \quad (19) \]

These results indicated that all the sensitivities were low and not larger than unity in absolute value. The proposed circuit thus exhibited low-sensitivity performance.

A study is next carried out on the effects of various parasitic of the DDCCTA used in the proposed circuit. The equations (2)–(5) were obtained using an ideal DDCCTA. The three \( Y \) terminals exhibit an infinite input resistance, the \( X \) port exhibits zero input resistance, the two \( O \) ports exhibit an infinite output resistance and port \( Z \) exhibits an infinite output resistance. In practical applications implementing the active element by
using transistors, these resistances must be assumed to have finite values, depending upon the device parameters. Similarly, high frequency effects must be accounted for by assuming capacitances at these ports. Applied DDCCTA has parasitic ports: port \(Y\) takes the form \(R_Y \parallel C_Y\), port \(Z\) takes the form of \(R_Z \parallel C_Z\), port \(O\) takes the form \(R_O \parallel C_O\) and port \(X\) takes the form of \(R_X\) (Channumsin and Tangsrirat 2013). In the presence of these parasitic elements, the circuit presented in Figure 2 is modified to become the circuit presented in Figure 3, in which \(C_{1p} = C_O \parallel C_Y\), \(C_{2p} = C_O \parallel C_Z\), \(R_{1p} = R_O \parallel R_Y\) and \(R_{2p} = R_O \parallel R_Z\). The proposed circuit of Figure 2 employs external capacitors \(C_2\) and \(C_1\) connected in parallel at the terminals \(O\) and \(Z\), respectively. Consequently, the effects of the parasitic capacitances \(C_O\), \(C_Z\) and \(C_Y\) can be absorbed, because \(C_1 \gg C_Z\) and \(C_2 \gg (C_O + C_Z\). One grounded resistor is connected at the \(X\) terminal to absorb series parasitic resistance at the \(X\) terminal of the DDCCTA. Accounting for the parasitic elements in Figure 3, the input-output relationship matrix form of Figure 3 can be expressed as

\[
\begin{bmatrix}
    s & n_1 & -n_1 & 0 & V_{o1} \\
    -n_2 & s & 0 & 0 & V_{o2} \\
    g_m R_2 & 0 & m & 0 & V_{o3} \\
    0 & H_1 & -H_1 & 1 & V_{o4}
\end{bmatrix} = \begin{bmatrix}
    n_1 V_{i1} \\
    0 \\
    H_2 V_{i2} \\
    H_1 V_{i1}
\end{bmatrix}
\]

where

\[
n_1 = \left(\frac{1}{R_1}\right) \frac{s R_Z}{1 + s R_Z C_1} = \left(\frac{1}{R_1 C_1}\right) \frac{s}{s + \omega_1}, \quad \omega_1 = \frac{1}{R_1 C_1}
\]

\[
n_2 = \frac{s g_m R_{2p}}{1 + s R_{2p} C_2} = \left(\frac{g_m}{C_2}\right) \frac{s}{s + \omega_2}, \quad \omega_2 = \frac{1}{R_{2p} C_2}
\]

Figure 3. Proposed voltage-mode universal filter including the parasitic elements of DDCCTA.
\[ m = 1 + sR_1C_{\text{ip}} = \frac{1}{\omega_3} (s + \omega_3) \quad ; \quad \omega_3 = \frac{1}{R_2C_{\text{ip}}} \]  

\[ C'_1 = C_1 + C_Z, \quad C'_2 = C_2 + C_{2p}, \quad R'_1 = R_1 + R_x, \quad R_2 = R_2//R_{1p}, \quad H_1 = \frac{R_1}{R'_1}, \quad H_2 = \frac{R'_2}{R_2} \]

From (20), (2)–(5) can be rewritten as follows:

\[ V_{o1} = \frac{(mV_{i1} + H_2V_{i2})n_1s}{ms^2 + n_1g_mR'_2s + mn_n_2} \]  

\[ V_{o2} = \frac{(mV_{i1} + H_2V_{i2})n_2s}{ms^2 + n_1g_mR'_2s + mn_n_2} \]  

\[ V_{o3} = \frac{-n_1g_mR_2sV_{i1} + H_2(s^2 + n_1n_2)V_{i2}}{ms^2 + n_1g_mR'_2s + mn_n_2} \]  

\[ V_{o4} = \frac{(mV_{i1} + H_2V_{i2})H_1s^2}{ms^2 + n_1g_mR'_2s + mn_n_2} \]

As (21)–(23) illustrate, the effects of the parasitic elements are dependent on three parasitic poles yielded by the non-idealities of DDCCTA. For close to ideal operation at high frequencies, the frequency of operation should be larger than \( \omega_1 \) and \( \omega_2 \), and smaller than \( \omega_3 \). Therefore, the useful frequency range of the proposed filter is limited by the following conditions:

\[ 10 \times \max \{ \omega_1, \omega_2 \} < < \omega < < 0.1 \omega_3 \]  

It is not difficult to satisfy this condition, because the external capacitance can be set much higher than parasitic capacitance. If the conditions of \( \frac{1}{s(C_1 + C_Z)} < < R_Z \),  

\( \frac{1}{s(C_2 + C_{o} + C_{y}^2)} < < R_0//R_{y_2} \) and \( R_2 < < \frac{1}{s(C_{-o} + C_{y_1})} // R_{-o}//R_{y_1} \) are satisfied, the influence of the DDCCTA parasitic elements on the proposed filter in Figure 3 can be ignored.

According to (28), the influence of parasitic elements on the coefficients \( n_1, n_2 \) and \( m \) is diminished when the conditions of \( |s| > > \omega_1, \quad |s| > > \omega_2 \) and \( |s| < < \omega_3 \) are satisfied, and hence \( n_1 \approx \frac{1}{RC_1}, \quad n_2 \approx \frac{g_m}{C_2} \) and \( m \approx 1 \). Therefore, the four output voltages in (24)–(27) can be rewritten as

\[ V_{o1} = \frac{(V_{i1} + H_2V_{i2})C_2s}{s^2C_2R_1 + sC_2g_mR_2 + g_m} \]  

\[ V_{o2} = \frac{(V_{i1} + H_2V_{i2})g_m}{s^2C_2R_1 + sC_2g_mR_2 + g_m} \]
\[ V_{o3} = -sC_2g_mR_1V_1 + H_2(s^2C_1C_2R_1 + g_m)V_{i2} \]
\[ V_{o4} = \frac{(V_{i1} + H_2V_{i2})s^2C_1C_2R_1}{s^2C_1C_2R_1 + sC_2g_mR_2 + g_m} \]

In this case, \( \omega_o \) and \( Q \) are changed to
\[ \omega_o = \sqrt{\frac{g_m}{R_2C_2}} \quad Q = \frac{1}{R_2} \sqrt{\frac{R_1C_1}{g_mC_2}} \]

Thus the influence of DVCCTA parasitic elements to the proposed filter of Figure 2 can be ignored.

3. Simulation results
3.1. HSPICE simulation results

To verify the theoretical study, HSPICE simulations were carried out to demonstrate the feasibility of the proposed circuit. The CMOS implementation of the DDCCTA is shown in Figure 4 (Tangsrirat, Channumsin and Pukkalanun 2013, Channumsin and Tangsrirat 2013). The dimensions of MOS transistors used in implementation of the DDCCTA are given in Table 2. The supply voltages were \( V_{DD} = -V_{SS} = 0.9 \) V, and the biasing voltage was \( V_{BB} = -0.5 \) V. Figure 5 shows the simulated gain responses for the BP \( (V_{o1}) \), LP \( (V_{o2}) \) and HP \( (V_{o4}) \) filters of Figure 2 with \( V_{i1} = 0 \) (grounded) and \( V_{i2} = \) input voltage signal. Figure 6 shows the simulated gain and phase responses for the BR \( (V_{o3}) \) filter of Figure 2 with \( V_{i1} = 0 \) (grounded) and \( V_{i2} = \) input voltage signal. Figure 7 shows the simulated gain and phase responses for the AP \( (V_{o3}) \) filter of Figure 2 with \( V_{i1} = V_{i2} = \) input voltage signal. The component values of Figures 5–7 are chosen as \( g_m = 200 \) \( \mu \)A/V \( (I_B = 96.5 \) \( \mu \)A), \( R_1 = R_2 = 5 \) k\( \Omega \), \( C_1 = 3 \) pF and \( C_2 = 6 \) pF, which is designed to obtain voltage-mode filter with a pole frequency \( f_0 = 7.503 \) MHz and the quality factor of \( Q = 0.707 \). Figure 8 represents the controllability of \( f_o \) with \( Q \) unchanged for BP filter of Figure 2 with \( V_{i2} = 0 \) (grounded) and \( V_{i1} = \) input voltage signal. This is designed for a constant value of \( Q = 1 \) with \( C_1 = 3 \) pF and \( C_2 = 6 \) pF for different values of \( f_o \) as given in Table 3. Figure 9 represents the controllability of \( Q \) without disturbing \( f_o \) for BP filter of Figure 2 with \( V_{i2} = 0 \) (grounded) and \( V_{i1} = \) input voltage signal. This is designed for a constant pole frequency of \( f_0 = 1.326 \) MHz with \( g_m = 50 \) \( \mu \)A/V \( (I_B = 6.04 \) \( \mu \)A), \( R_1 = 20 \) k\( \Omega \), \( C_1 = 12 \) pF and \( C_2 = 3 \) pF. The values of \( R_2 \) is selected as 5 k\( \Omega \), 10 k\( \Omega \), 20 k\( \Omega \), 40 k\( \Omega \) and 80 k\( \Omega \), respectively, which result in \( Q = 8, 4, 2, 1 \) and 0.5, respectively. The results demonstrate that tuning of \( Q \)-value without affecting the \( f_o \)-value can be performed via different values of \( R_2 \). To test the input dynamic range of Figure 2, the simulation has been repeated for a sinusoidal input signal at \( f_o = 7.5 \) MHz. Figure 10
shows that the input dynamic range for the BP response at $V_{o3}$ output terminal with $V_{i1}$ = input voltage signal, $V_{i2} = 0$, $g_m = 200 \mu$A/V, $R_1 = R_2 = 5$ kΩ, $C_1 = 3$ pF and $C_2 = 6$ pF, which extends up to amplitude of 0.5 V (peak to peak) without significant distortion.

In Figure 10, the percentage of total harmonic distortion was 2.56 %.

![Figure 4. CMOS implementation of DDCCTA.](image)

Table 2. The aspect ratios of the CMOS transistors in DDCCTA implementation.

<table>
<thead>
<tr>
<th>Transistors</th>
<th>$L$ (μm)</th>
<th>$W$ (μm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>M1–M4</td>
<td>0.35</td>
<td>8.75</td>
</tr>
<tr>
<td>M5–M8</td>
<td>0.18</td>
<td>17.5</td>
</tr>
<tr>
<td>M9–M12</td>
<td>0.18</td>
<td>8.75</td>
</tr>
<tr>
<td>M13–M14</td>
<td>0.5</td>
<td>10</td>
</tr>
<tr>
<td>M15–M20</td>
<td>0.8</td>
<td>25</td>
</tr>
<tr>
<td>M21–M24</td>
<td>0.8</td>
<td>8</td>
</tr>
</tbody>
</table>

Table 3. Component values for orthogonal adjustment of $f_o$ with $Q$.

<table>
<thead>
<tr>
<th>Transconductance $g_m$, $\mu$A/V</th>
<th>Bias current $I_B$, $\mu$A</th>
<th>Resistances $R_1 = R_2$, kΩ</th>
<th>Calculated value of $f_o$, MHz</th>
<th>Simulated value of $f_o$, MHz</th>
<th>Frequency error, %</th>
</tr>
</thead>
<tbody>
<tr>
<td>250</td>
<td>150.84</td>
<td>4</td>
<td>9.378</td>
<td>9.441</td>
<td>0.67</td>
</tr>
<tr>
<td>200</td>
<td>96.5</td>
<td>5</td>
<td>7.503</td>
<td>7.447</td>
<td>-0.75</td>
</tr>
<tr>
<td>100</td>
<td>24.14</td>
<td>10</td>
<td>3.751</td>
<td>3.767</td>
<td>0.43</td>
</tr>
<tr>
<td>66.67</td>
<td>10.73</td>
<td>15</td>
<td>2.501</td>
<td>2.506</td>
<td>0.2</td>
</tr>
<tr>
<td>50</td>
<td>6.04</td>
<td>20</td>
<td>1.876</td>
<td>1.884</td>
<td>0.43</td>
</tr>
</tbody>
</table>
Figure 5. Gain responses of BP ($V_{o1}$), LP ($V_{o2}$) and HP ($V_{o3}$) filters with $V_{i1} = 0$ and $V_{i2} =$ input voltage signal.

Figure 6. Gain and phase responses of BR ($V_{o3}$) filter with $V_{i1} = 0$ and $V_{i2} =$ input voltage signal.
Figure 7. Gain and phase responses of AP \((V_{o3})\) filter with \(V_{i1} = V_{i2} =\) input voltage signal.

Figure 8. Gain responses of BP \((V_{o1})\) filter for difference component values \(f_o = 1.844\) MHz: blue; \(f_o = 2.506\) MHz: red; \(f_o = 3.767\) MHz: green; \(f_o = 7.477\) MHz: pink; and \(f_o = 9.441\) MHz: purple).
Figure 9. Gain responses of BP ($V_{o1}$) filter when $R_2$ is varied ($Q = 0.5$: blue; $Q = 1$: red; $Q = 2$: green; $Q = 4$: pink; and $Q = 8$: purple).

Figure 10. The input (blue line) and output (red line) waveforms of inverting BP filter at $V_{o1}$ designed with $g_m = 200$ $\mu$A/V, $R_1 = R_2 = 5$ k$\Omega$, $C_1 = 3$ pF and $C_2 = 6$ pF for a 7.5 MHz sinusoidal input voltage of 0.5 V (peak to peak).
3.2. Critical temperature and Monte Carlo simulations

The proposed filter was also simulated to test against temperature variations. Figure 11 shows the results of temperature simulations for BP gain-frequency responses at $V_{o3}$ output terminal of Figure 2 with $V_{i1} = V_{in}$, $V_{i2} = 0$, $g_m = 200 \, \mu A/V$, $R_1 = R_2 = 5 \, k\Omega$, $C_1 = 3 \, \text{pF}$ and $C_2 = 6 \, \text{pF}$. The simulated temperature variations at $-20 \, ^\circ\text{C}$, $0 \, ^\circ\text{C}$, $20 \, ^\circ\text{C}$, $40 \, ^\circ\text{C}$ and $60 \, ^\circ\text{C}$ had the centre frequencies of $f_o = 7.9983 \, \text{MHz}$, $f_o = 7.7804 \, \text{MHz}$, $f_o = 7.2778 \, \text{MHz}$, $f_o = 7.1945 \, \text{MHz}$ and $f_o = 7.1068 \, \text{MHz}$ with errors of $6.607 \, \%$, $3.703 \, \%$, $-2.996 \, \%$, $-4.107 \, \%$ and $-5.276 \, \%$, respectively. The results are summarized in Table 4. At operating temperature between $-20 \, ^\circ\text{C}$ and $60 \, ^\circ\text{C}$, the $f_o$-value of the BP filter was affected from $+6.607 \, \%$ to $-5.276 \, \%$. To collect statistical data regarding mismatch and the variation effect, Monte Carlo simulations were conducted. The device mismatch was modelled as a set of randomly generated samples that represented the probability distributions of the device parameters. The circuit was then repeatedly simulated with the random device samples, and performance data were collected. Figure 12 shows the Monte Carlo results for 200 simulation, regarding the BP frequency responses at $V_{o3}$ output terminal of Figure 2 with $V_{i1} = V_{in}$, $V_{i2} = 0$, $g_m = 200 \, \mu A/V$, $R_1 = R_2 = 5 \, k\Omega$, $C_1 = 3 \, \text{pF}$ and $C_2 = 6 \, \text{pF}$, in which the capacitors $C_1$ and $C_2$ had a variation of $5 \, \%$ Gaussian deviation and the resistors $R_1$ and $R_2$ had a variation of $20 \, \%$ Gaussian deviation. According to the simulation, with a variation of centre frequency between $6.823 \, \text{MHz}$ to $8.222 \, \text{MHz}$, the $f_o$-value of the BP filter was affected in the range of $-9.03 \, \%$ to $+9.63 \, \%$. Figure 13 compares the histogram of the centre frequency obtained from the Monte Carlo analysis. As the Monte Carlo analysis results indicated, the proposed filter exhibited reasonable sensitivity performance.

Table 4. Temperature variations of BP response of filter in Figure 11 with $g_m = 200 \, \mu A/V$, $R_1 = R_2 = 5 \, k\Omega$, $C_1 = 3 \, \text{pF}$ and $C_2 = 6 \, \text{pF}$. 

<table>
<thead>
<tr>
<th>Temperature, °C</th>
<th>Calculated value of $f_o$, MHz</th>
<th>Simulated value of $f_o$, MHz</th>
<th>Frequency error, %</th>
</tr>
</thead>
<tbody>
<tr>
<td>-20</td>
<td>7.5026</td>
<td>7.9983</td>
<td>6.607</td>
</tr>
<tr>
<td>0</td>
<td>7.5026</td>
<td>7.7804</td>
<td>3.703</td>
</tr>
<tr>
<td>20</td>
<td>7.5026</td>
<td>7.2778</td>
<td>-2.996</td>
</tr>
<tr>
<td>40</td>
<td>7.5026</td>
<td>7.1945</td>
<td>-4.107</td>
</tr>
<tr>
<td>60</td>
<td>7.5026</td>
<td>7.1068</td>
<td>-5.276</td>
</tr>
</tbody>
</table>
Figure 11. Simulated characteristics of BP gain-frequency responses for different temperatures (-20 °C: blue; 0 °C: red; 20 °C: green; 40 °C: pink; and 60 °C: purple).

Figure 12. Monte Carlo analysis results for 200 simulations of the BP (V_{o3}) frequency responses with V_{i1} = V_{in} and V_{i2} = 0.
Figure 13. Monte Carlo analysis results with ± 5% deviation on the capacitor values and ± 20% deviation on the resistor values.

Figure 14. The chip layout of proposed voltage-mode universal biquadratic filter.

3.3. Post-layout simulations

The overall chip layout and the detail layout of the filter core were shown in Figures 14 and 15, respectively. The component values of Figure 14 were given by $g_m =$
100 \mu A/V, \ R_1 = R_2 = 10 \ k\Omega, \ C_1 = 1 \ pF \ and \ C_2 = 2 \ pF, \ leading \ to \ a \ centre \ frequency \ f_0 = 11.25 \ MHz. \ The \ layout \ floorplan \ is \ shown \ in \ Figure \ 16 \ which \ explains \ elements \ placement. \ All \ these \ processes \ are \ carried \ out \ using \ Cadence \ Virtuoso \ Schematic \ and \ Layout \ editor \ tool \ for \ 0.18 \ \mu m \ process \ technology. \ The \ post-layout \ simulations \ were \ carried \ out \ to \ check \ the \ functionality \ of \ the \ design. \ Figure \ 17 \ shows \ the \ post-simulated \ results \ of \ BP, \ HP, \ BR \ and \ AP \ gain-frequency \ responses \ with \ V_{i1} = 0 \ and \ V_{i2} = V_{in}. \ The \ total \ power \ dissipation \ is \ found \ to \ be \ 0.387 \ mW. \ The \ chip \ area \ without \ pads \ is \ only \ 74.699 \times 83.075 \ \mu m^2.

![Diagram](image)

**Figure 15.** The core of the proposed voltage-mode universal biquadratic filter.

![Diagram](image)

**Figure 16.** The layout floorplan.
4. Conclusions

In this paper, a new tunable DDCCTA-based voltage-mode universal biquadratic filter was proposed. The proposed filter structure only uses a single DDCCTA, two grounded capacitors and two resistors, which are the minimum components necessary for realizing a biquadratic filtering function from the same topology. Either high-input impedance multifunction biquadratic filter or universal biquadratic filter can be realized in the same configuration. The proposed circuit has the following features: (i) providing five standard filters functions, that is, LP, BP, HP, BR and AP filters from the same configuration, (ii) imposing component choices is unnecessary, (iii) using only grounded capacitors, (iv) using only one active component, (v) orthogonal controllability of the parameters $\omega_0$ and $Q$ and (vi) low active and passive sensitivity performances. HSPICE simulations with TSMC 0.18 $\mu$m process technology and $\pm$ 0.9 V supply voltages confirm the theoretical predictions.

References


