Realization of CMOS Current Controlled Current Conveyor Transconductance Amplifier (CCCCTA) and Its Applications

MONTREE SIRIPRUCHYANUN¹, PHAMORN SILAPAN² AND WINAI JAIKLA³

¹Department of Teacher Training in Electrical Engineering, Faculty of Technical Education, King Mongkut’s University of Technology North Bangkok, THAILAND. E-mail: mts@kmutnb.ac.th
²Electric and Industrial Program, Faculty of Industrial Technology, Uttaradit Rajabhat University, THAILAND
³Electric and Electronic Program, Faculty of Industrial Technology, Suan Sunandha Rajabhat University, THAILAND. E-mail: jnai2004@yahoo.com

This article presents a basic building block for analog signal processing, namely current controlled current conveyor transconductance amplifier (CCCCTA). Its parasitic resistance at current input port can be controlled by an input bias current. It is very suitable to use in either current-mode or voltage-mode signal processing systems. The proposed element was realized in a CMOS technology and was examined the performances through PSPICE simulations. They display usabilities of the new active element, where the maximum bandwidth is 107.34 MHz. The supply voltages can be as low as ±1.5 V and maximum power consumption is 899 µW. The CCCCTA performs tuning over a wide current range. In addition, some examples as a voltage and current-mode universal biquad filter and a grounded inductance simulator are included. They occupy only single CCCCTA.

Keywords: CMOS, CCCCTA, Current-mode, Voltage-mode, Filter, Oscillator, Inductance simulator.

1 INTRODUCTION

There has been much effort to reduce the supply voltage of electronic circuits in last decade. This is due to the command for portable and battery-powered equipment. Since a low-voltage operating circuit becomes necessary, the current-mode technique is ideally suited for this purpose more than the voltage-
mode one. Consequently, there is a growing interest in synthesizing the current-mode circuits because of more their potential advantages such as larger dynamic range, higher signal bandwidth, greater linearity, simpler circuitry and lower power consumption [1–2]. Many active elements able to function in current-mode such as OTA, current conveyor and current differencing buffered amplifier (CDBA), have been introduced to response these demands.

Recently, a reported 5-terminals active element, namely current conveyor transconductance amplifier (CCTA) [3], seems to be a versatile component in the realization of a class of analog signal processing circuits, especially analog frequency filters [3–5]. It is supposed for usage mostly in current-mode circuits, but it is also choice in case of voltage mode and/or hybrid (voltage-current) circuits (e.g. V/I converters). In addition, it can also be adjusted the output current gain. However, from our investigations, there are seen that the CCTA can not be controlled by the parasitic resistance at input port so when it is used in some circuits, it must unavoidably require some external passive components, especially the resistors. This makes it not appropriate for IC implementation due to occupying more chip area. In addition, the mentioned CCTA has the third-generation current conveyor (CCIII) as an input stage which has flexibility for applications less than a second-generation current conveyor (CCII).

The purpose of this paper is to design and synthesize a modified-version CCTA, which is newly named current controlled current conveyor transconductance amplifier (CCCCTA). The parasitic resistance at current input port can be controlled by an input bias current, and then it does not need a resistor in practical applications. Even it can be implemented by employing second-generation current controlled current conveyor (CCCII) and OTA, it would be more very convenient and useful if the CCCCTA is realized in monolithic chip. Some example applications as voltage and current-mode universal filters, grounded inductance simulator and oscillator are comprised. They confirm that only single CCCCTA is employed for each application.

2 CIRCUIT CONFIGURATION

2.1 Basic Concept of CCCCTA

CCCCTA properties are similar to the conventional CCTA, except that the CCCCTA has finite input resistance $R_x$ at the x input terminal. This parasitic resistance can be controlled by the bias current $I_{bi}$ as shown in the following equation

\[
\begin{bmatrix}
I_y \\
V_x \\
I_z \\
I_o
\end{bmatrix} =
\begin{bmatrix}
0 & 0 & 0 & 0 \\
R_x & 1 & 0 & 0 \\
1 & 0 & 0 & 0 \\
0 & 0 & \pm g_m & 0
\end{bmatrix}
\begin{bmatrix}
I_x \\
V_y \\
V_x \\
V_o
\end{bmatrix},
\] (1)
where

\[ R_x = \frac{1}{\sqrt{8\beta_n I_{B1}}}, \]  

(2)

and

\[ g_m = \sqrt{\beta_n I_{B2}}. \]  

(3)

g_m is the transconductance of the CCCCTA, \( \beta_n = \mu_n C_{ox} (W/L) \) is the physical parameter of MOS transistor and \( V_T \) is the thermal voltage. The symbol and the equivalent circuit of the CCCCTA are illustrated in Figs. 1(a) and (b), respectively.

2.2 A Second Generation Current Controlled Current Conveyor (CCCII)

Fig. 2 displays a class AB translinear loop, which is used as input section. By straightforward analysis, we will obtain the parasitic resistance at input terminal as

\[ R_{in} = \frac{1}{I_{B1}} \left( \frac{1}{g_{m2} + g_{m4}} \right) - \frac{I_{B1}}{I_{in}} \left( \frac{g_{m2}}{g_{m1}} - \frac{g_{m4}}{g_{m3}} \right). \]  

(4)

From Eq. (4) if \( g_{m1} = g_{m3} \) and \( g_{m2} = g_{m4} = g_m \), we will get

\[ R_{in} = \frac{1}{2g_m} = \frac{1}{\sqrt{8\beta_n I_{B1}}}. \]  

(5)

Based on the use of the finite input resistance input stage of Fig. 2, the second generation current controlled current conveyor can be shown in Fig. 3. When \( V_y \) is grounded, the output current \( I_z \) can be expressed as

\[ I_z = \alpha I_x + \varepsilon, \]  

(6)
where $\alpha$ and $\varepsilon$ are current gain and error terms, respectively. By straightforward analysis of circuit in Fig. 3, we will obtain the $\alpha$ and $\varepsilon$ as

$$
\alpha = \frac{g_{m4}g_{m13}}{g_{m12} (g_{m2} + g_{m4})} + \frac{g_{m2}g_{m9}}{g_{m8} (g_{m2} + g_{m4})},
$$

(7)

and

$$
\varepsilon = \left\{ \begin{array}{c}
\frac{I_{B1} g_{m4} g_{m13}}{g_{m2} + g_{m4}} - \frac{I_{B1} g_{m2} g_{m9}}{g_{m8} (g_{m2} + g_{m4})} + \\
\frac{I_{B1} g_{m2} g_{m8}}{g_{m1} g_{m3}} \left( \frac{g_{m4} g_{m13}}{g_{m12}} + \frac{g_{m2} g_{m9}}{g_{m8}} \right) \end{array} \right\}.
$$

(8)
If \( g_{m1} = g_{m4}, g_{m2} = g_{m4}, g_{m8} = g_{m9} \) and \( g_{m12} = g_{m13} \), then the output current can be found to be

\[
I_z = I_z. \tag{9}
\]

The output resistance looking into the z terminal \( (r_z) \) can be respectively expressed as

\[
r_z \cong \frac{r_{o9}r_{o13}}{r_{o9} + r_{o13}}, \tag{10}
\]

where \( r_o \) is the drain-source resistance seen at the mentioned output terminal.

### 2.3 Transconductance Amplifier

In this section, a simple differential pair amplifier [6] is employed to achieve simpler circuit description of the proposed CCCCTA as shown in Fig. 4. From Fig. 4, transistors \( M_{14} \) and \( M_{15} \) function as a differential amplifier to convert an input voltage to an output current. \( M_{16} \) and \( M_{17} \) work as a simple current mirror when \( I_B^2 \) is an input bias current. When \( V_{in} \) is applied, this makes \( I_{D14} \) and \( I_{D15} \) flowing in \( M_{14} \) and \( M_{15} \), respectively. The relationship of \( I_o \) and \( V_{in} \) of the transconductance amplifier is given by

\[
I_O = \beta_1 g_{m14} V_1 - \beta_2 g_{m15} V_2 + \epsilon, \tag{11}
\]

where \( \beta_1 \) and \( \beta_2 \) are transconductance ratios. By straightforward analysis of circuit in Fig. 4, we will obtain the \( \beta_1 \) and \( \beta_2 \) as

\[
\beta_1 = \frac{g_{m17} g_{m16} - g_{m14} g_{m15}}{g_{m16} (g_{m14} + g_{m15})}, \tag{12}
\]

![FIGURE 4](image)
The simple transconductance amplifier.
\[
\beta_2 = 1 - \frac{(g_{m15}g_{m16} - g_{m14}g_{m17})}{g_{m16}(g_{m14} + g_{m15})}, \quad (13)
\]

and

\[
\varepsilon = -\frac{I_B (g_{m15}g_{m16} - g_{m14}g_{m17})}{g_{m16}(g_{m14} + g_{m15})}. \quad (14)
\]

If \( g_{m14} = g_{m15} = g_m \) and \( g_{m16} = g_{m17} \), Eq. (11) can be expressed as

\[
I_O = g_m (V_1 - V_2), \quad (15)
\]

where \( g_m = \sqrt{\beta_n I_{B2}} \).

The output resistance looking into the x terminal \( (r_o) \) can be respectively expressed as

\[
r_O \cong \frac{r_{o15} r_{o17}}{r_{o15} + r_{o17}}. \quad (16)
\]

### 2.4 Proposed Current Controlled Current Conveyor Transconductance Amplifier

The proposed CCCCTA consists of two principal blocks: a current controlled second generation current conveyor (CCCII) circuit and an operational transconductance amplifier (OTA) circuit. The proposed realization of the CCCCTA in a CMOS technology to achieve a wide-range of frequency responses is shown in Fig. 5. The circuit implementation consists of mixed translinear loop (M1-M4). The mixed loop is DC biased by using current mirrors (M5-M7 and M10-M11). The \( R_x \) can be obtained by Eq. (2). The output z-terminal that generates the current from x terminal is realized using transistors (M8-M9 and M12-M13). The simple-version transconductance amplifier is
Realized using transistors (M₁₄-M₁₇), whose transconductance gain can be adjusted by $I_{B₂}$.

3 SIMULATION RESULTS

To prove the performances of the proposed CCCCTA, the PSPICE simulation program was used. The PMOS and NMOS transistors employed in the proposed element in Fig. 2 were simulated by respectively using the parameters of a 0.35 µm TSMC CMOS technology [7] with ±1.5 V supply voltages. The aspect transistor ratios of PMOS and NMOS are listed in Table 1. Fig. 6 depicts the parasitic resistance $R_x$ at input terminal when $I_{B₁}$ has varied. Fig. 7 shows the transconductance value when $I_{B₂}$ was varied from $1\mu A - 200\mu A$.

Fig. 8 displays DC transfer characteristic of the proposed CCCCTA, when $I_{B₁} = 10\mu A, 50\mu A$ and $100\mu A$. So it is seen that it is linear in range of $−7mA ≤ I_x ≤ 7mA$. Moreover, the bandwidths of output terminals are shown in Fig. 9. We found that the $−3dB$ bandwidth of $I_z / I_x$, $V_z / V_x$, $I_a / I_x$ and $I_v / V_z$ are respectively located at 333.48 MHz, 4.12 GHz, 107.34 MHz and 115.22 MHz. The summarized properties of the CMOS CCCCTA are shown in Table 2.

<table>
<thead>
<tr>
<th>CMOS Transistors</th>
<th>$W(\mu m) / L(\mu m)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>M1-M2</td>
<td>12/1</td>
</tr>
<tr>
<td>M3-M4</td>
<td>4/1</td>
</tr>
<tr>
<td>M5, M7</td>
<td>10/1</td>
</tr>
<tr>
<td>M6</td>
<td>8/1</td>
</tr>
<tr>
<td>M8-M9</td>
<td>10/5</td>
</tr>
<tr>
<td>M10-M11</td>
<td>30/1</td>
</tr>
<tr>
<td>M12-13</td>
<td>30/5</td>
</tr>
<tr>
<td>M14-M17</td>
<td>30/5</td>
</tr>
</tbody>
</table>

![Parasitic resistance at x terminal relative to $I_{B₁}$.](image.png)
FIGURE 7
Transconductance value relative to $I_{B2}$.

FIGURE 8
DC transfer characteristic of the CCCCTA.

FIGURE 9
Frequency responses at output terminals.
TABLE 2
Conclusions of the CMOS CCCCTA parameters.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power supply voltages</td>
<td>±1.5 V</td>
</tr>
<tr>
<td>Power consumption</td>
<td>899 µW</td>
</tr>
<tr>
<td>3dB Bandwidth</td>
<td>333.48 MHz (Iz/Ix), 4.12 GHz (Vx/Vy), 107.34 MHz (Io/Ix), 115.22 MHz (Io/Vy)</td>
</tr>
<tr>
<td>Input current linear range</td>
<td>-0.7 mA to 0.7 mA</td>
</tr>
<tr>
<td>Irrange</td>
<td>491.09 Ω-15.76 MΩ</td>
</tr>
<tr>
<td>Input bias current range for controlling Irr</td>
<td>1 nA-95.29 µA</td>
</tr>
<tr>
<td>Transconductance range</td>
<td>0.28 mS-1 mS</td>
</tr>
<tr>
<td>Input bias range for controlling transconductance amplifier</td>
<td>1 µA-180 µA</td>
</tr>
<tr>
<td>Irr</td>
<td>140.16 kΩ</td>
</tr>
<tr>
<td>Rz</td>
<td>207.87 kΩ</td>
</tr>
<tr>
<td>Ry</td>
<td>9.32 MΩ</td>
</tr>
</tbody>
</table>

4 APPLICATION EXAMPLES

4.1 Digitally-programmable current-mode universal biquad filter

The first application of the proposed CCCCTA is a current-mode biquad filter shown in Fig. 10. It employs only one active element and 2 grounded capacitors, which is easy to fabricate, differing from previous circuits [8–9]. The CCCCTA in Fig. 10 is slightly modified from the proposed CCCCTA in Fig. 5 by using dual-output CCCCTA which can be easily achieved by using the current mirrors to copy current from z1 to z2 terminal to extend the usability of the CCCCTA. Straightforwardly analyzing the circuit in Fig. 10 and using CCCCTA properties in section 2, the transfer functions of the network can be obtained as 2nd

\[
I_o = \frac{I_{in1} s C_2 + g_m I_{in2} - I_{in3} (s^2 C_1 C_2 R_1 + s C_2 + g_m)}{s^2 C_1 C_2 R_1 + s C_2 + g_m}. \tag{17}
\]

From Eq. (17), the magnitudes of current inputs \( I_{in1}, I_{in2}, \) and \( I_{in3} \) are chosen as in Table 3 to obtain a standard function of the 2nd order network. The circuit for selection can be seen in [10]. From Eq. (17), the pole frequency \( (\omega_0) \) and quality factor \( (Q_0) \) of each filter response can be expressed as
\[ \omega_0 = \frac{g_m}{\sqrt{C_1 C_2 R_x}}, \quad (18) \]

and

\[ Q_0 = \frac{C_1 R_x g_m}{\sqrt{C_2}}. \quad (19) \]

Substituting the intrinsic resistance and transconductance as depicted in Eqs. (2) and (3) into Eqs. (18) and (19), it yields

\[ \omega_0 = \left( 8 \beta_n \frac{I_{B1} I_{B2}}{C_1 C_2} \right)^{1/2}, \quad (20) \]

**TABLE 3**

The \( I_{in1}, I_{in2}, \) and \( I_{in3} \) value selections for each filter function response.

<table>
<thead>
<tr>
<th>Filter Responses</th>
<th>( I_{in1} )</th>
<th>( I_{in2} )</th>
<th>( I_{in3} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>BP</td>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>HP</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>LP</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>BR</td>
<td>1</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>AP</td>
<td>2</td>
<td>0</td>
<td>1</td>
</tr>
</tbody>
</table>
From Eqs. (20) and (21), by maintaining the ratio of $I_{B1}$ and $I_{B2}$ constant, the pole frequency can be adjusted by $I_{B1}$ and $I_{B2}$ without affecting the quality factor.

To prove the performances of the proposed circuit, the PSPICE simulation program was used. $C_1 = C_2 = 1nF$, $I_{B1} = 50 \mu A$, $I_{B2} = 150 \mu A$ are chosen to obtain resistance and transconductance values of 2.59 kΩ and 0.236 mS, respectively. This yields the pole frequency of 53.70 kHz. Calculated value of this parameter from Eq. (18) is 48.06 kHz. The results shown in Fig. 11 are the gain and phase responses of the proposed biquad filter obtained from Fig. 11. There are clearly seen that the proposed biquad circuit can provide low-pass, high-pass, band-pass, band-reject and all-pass functions dependent on selections as shown in Table 3, without modifying circuit topology.

Fig. 12 displays magnitude responses of band-pass function for different $I_B$ values. It is shown that the pole frequency of the responses can be adjusted by the input bias current $I_{B1}$.

### 4.2 Grounded Inductance Simulator

The grounded inductance simulator based on the CCCCTA is shown in Fig. 13. It employs only single CCCCTA, which contrasts to ordinarily proposed circuits [11–12]. From routine analysis and using the CCCCTA properties, an input impedance of the circuit can be written as

$$Z_{in} = \frac{V_{in}}{I_{in}} = \frac{sCR_s}{g_m}. \quad (22)$$

From Eq. (22), it is obvious that the circuit shown in Fig. 13 performs a grounded inductance with a value

$$Z_{in} = \frac{V_{in}}{I_{in}} = \frac{sCR_s}{g_m}. \quad (23)$$

From Eq. (23), the inductance value $L_{eq}$ can be adjusted electronically by either $I_{B1}$ or $I_{B2}$.

The impedance and phase of the simulator relative to frequency, which are compared to ideal inductor are also shown in Fig. 14 where $C = 1nF$, $I_{B1} = 30 \mu A$ and $I_{B2} = 100 \mu A$. Fig. 15 shows impedance values relative to frequency of the simulator for different $I_{B2}$.
FIGURE 11
Gain and phase responses of the biquad filter in Fig. 10 (a) LP (b) HP (c) BP (d) BR (e) AP.
FIGURE 12
Band-pass responses for different values of $I_b$.

FIGURE 13
Grounded inductance simulator based on the CCCCTA.

FIGURE 14
The impedance and phase relative to frequency of the grounded inductance simulator.

FIGURE 15
The impedance values relative to frequency of the simulators with different $I_{b2}$. 
4.3 Sinusoidal Oscillator

The sinusoidal oscillator based on the CCCCTA is shown in Fig. 16. It consists of single output CCCCTA and 2 grounded capacitors. Considering the circuit in Fig. 16 and using the CCCCTA properties, it yields characteristic equation of this circuit as

$$s^2 C_1 C_2 R_x + s (C_1 - C_2) + g_m = 0.$$  \hfill (24)

From Eq. (24), it can obviously seen that the proposed circuit can be set to be a sinusoidal oscillator if

$$C_1 = C_2.$$  \hfill (25)

Eq. (25) is called as the condition of oscillation. Then the characteristic equation of the system becomes

$$s^2 + \frac{g_m}{C_1 C_2 R_x} = 0.$$  \hfill (26)

From Eq. (26), the oscillation frequency of this system can be obtained as

$$\omega_0 = \sqrt{\frac{g_m}{C_1 C_2 R_x}} = \left(8 \beta_n \frac{\sqrt{I_{B1} I_{B2}}}{C_1 C_2}\right)^{1/2}.$$  \hfill (27)

It can be seen that, from Eq. (27), the oscillation frequency ($\omega_0$) can be controlled by bias current. The confirmed result of the oscillator can be seen in Fig. 17. They show the responses of oscillator with bias currents: $I_{B1}$ and $I_{B2}$ are respectively set to 50 $\mu$A and 100 $\mu$A. Fig. 18 shows the simulated output spectrum, where the total harmonic distortion (THD) is about 1.07%.

![Figure 16: Oscillator based on the CCCCTA.](image-url)
The final application of the proposed CCCCTA depicted in this paper is a voltage-mode universal biquad filter shown in Fig. 19. For straightforward analysis of the circuit in Fig. 19 and using CCCCTA properties in section 2, the transfer functions of the network can be obtained as

\[
V_o = \frac{V_{in1}s^2C_1C_2 + V_{in2}sC_2G_1 + V_{in3}G_xg_m}{s^2C_1C_2 + sC_2G_1 + G_xg_m}.
\]

(28)

![FIGURE 17](image)
The output waveform obtained from the oscillator in Fig. 16.

![FIGURE 18](image)
The simulation result of output spectrum.

![FIGURE 19](image)
Voltage-mode universal biquad filter based on the CCCCTA.
From Eq. (28), the magnitudes of input voltages: $V_{i_{n1}}, V_{i_{n2}}$ and $V_{i_{n3}}$ are chosen as in Table 4 to obtain a standard function of the 2nd order network. From Eq. (28), the pole frequency ($\omega_0$) and quality factor ($Q_0$) of each filter response can be expressed as

$$\omega_0 = \sqrt{n \frac{g_m}{C_1 C_2 R_x}} = \left( \frac{8 \beta_n \sqrt{I_{R1} I_{R2}}}{C_1 C_2} \right)^{1/2}.$$  \hspace{1cm} (29)

and

$$Q_0 = R_1 \sqrt{\frac{C_1 R_1 g_m}{C_2}} = \left( \frac{C_1 \sqrt{I_{R2}}}{C_2 \sqrt{I_{R1}}} \right)^{1/2}.$$  \hspace{1cm} (30)

It is either found that, from Eqs. (29) and (30), the quality factor can be adjusted by $R_1$ without affecting the pole frequency. Reversely, the pole frequency can be controlled via $R_x$ or $g_m$.

To prove the performances of the proposed circuit, the PSPICE simulation program was also used. $C_1 = C_2 = \ln F$, $I_{R1} = 50 \mu A$, $I_{R2} = 150 \mu A$ and $R_1 = 4.7 k\Omega$ are chosen. The results shown in Fig. 20 are the gain and phase responses of the proposed biquad filter obtained from Fig. 19. It is seen that the proposed biquad circuit can provide low-pass, high-pass, band-pass, band-reject and all-pass functions dependent on selection as shown in Table 4, without modifying circuit topology. Fig. 21 displays gain responses of band-pass function for different $R_1$ values. It is shown that the quality factor can be controlled by $R_1$ without affecting the pole frequency.

<table>
<thead>
<tr>
<th>Filter Responses</th>
<th>Input</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_0$</td>
<td>$V_{i_{n1}}$</td>
</tr>
<tr>
<td>BP</td>
<td>0</td>
</tr>
<tr>
<td>HP</td>
<td>1</td>
</tr>
<tr>
<td>LP</td>
<td>0</td>
</tr>
<tr>
<td>BR</td>
<td>1</td>
</tr>
<tr>
<td>AP</td>
<td>1</td>
</tr>
</tbody>
</table>

TABLE 4
The $V_{i_{n1}}, V_{i_{n2}}$ and $V_{i_{n3}}$ value selections for each filter function response.
FIGURE 20
Gain and phase responses of the biquad filter in Fig. 19 (a) LP (b) HP (c) BP (d) BR (e) AP.
FIGURE 21
Band-pass responses for different values of $R_1$.

5 CONCLUSION

The new basic building block, called CCCCTA, has been introduced via this paper. The usabilities have been proven by the simulation and application examples. They consume few numbers of components while electronic controllability is still available, which differs from the recently proposed elements. This modified element is very appropriate to realize in commercially-purposed integrated circuit. Our future work is to find more applications of the CCCCTA, emphasizing on analog signal processing circuits such as multiplier/divider, rectifier, etc.

REFERENCES

