Current-mode Quadrature Oscillator Using CFCC

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Abstract In this communication, a Current Follower Current Conveyor (CFCC)-based quadrature oscillator circuit has been proposed which employs a resonator-negative resistor configuration and provides oscillations in quadrature from two high impedance current output terminals of the CFCCs. The workability of the quadrature oscillator has been verified using PSPICE simulations using a CMOS CFCC implementable in 0.18 micron TSMC technology.

Keywords: current follower, current conveyor, current-follower-current-conveyor, quadrature oscillator, current-mode oscillator, current-mode circuits


1. Introduction

A quadrature sinusoidal oscillator (QSO) circuit provides two sinusoids with 90° phase difference and finds several applications in communication and measurement systems. In communication systems they are used in quadrature mixers, single-sideband generators and direct-conversion receivers while in measurement systems such oscillators are used in vector generators or selective voltmeters (see [1] and references cited therein).

During the last two decades, many new active building blocks have been proposed in the domain of analog signal processing, a comprehensive review of which was presented in [2], wherein several new active building blocks (ABB) were also proposed. Many of the new ABBs proposed in [2] have been employed in the past to realize QSOs and other signal processing applications, which include Current Differencing Buffered Amplifier (CDBA) [3,4], Voltage Differencing Buffered Amplifier (VDBA) [5], Voltage Differencing Transconductance Amplifier (VDTA) [6], Voltage Differencing current conveyor (VDCC) [7], Current Differencing Transconductance Amplifier (CDTA) [8-12], Multiple-Output Current Controlled Current Differencing Transconductance Amplifier (MO-CCCDTA) [13], Modified Current Differencing Transconductance Amplifier (MCDTA) [14], Current Follower Transconductance Amplifier (CFTA) [15], Z-copy Current Follower Transconductance Amplifier (ZC-CFTA) [16], Second-Generation Current Conveyor Transconductance Amplifier (CCII-TA) [17], Differential-Input Buffered and Transconductance Amplifier (DBTA) [18], Current-Feedback Operational Amplifier (CFOA) [19], Differential Voltage Current Conveyor (DVCC) [20], Differential Voltage Current-Controlled Conventional Transconductance Amplifier (DVCCCTA) [21], Voltage Differencing Inverting Voltage Buffered Amplifier (VDIBA) [22], Voltage Differencing-Differential Input Buffered Amplifiers (VD-DIBA) [23,24] and Programmable Current Amplifier [25].

It may be mentioned that while the QSOs presented in [3,4,5,7,16-24] operate in voltage mode (VM), those presented in [6,8-17,21,25] operate in current mode (CM). Also, of the various QSO circuits quoted above, only the circuits presented in [3,4] and [12] offer fully-decoupled condition of oscillation and frequency of oscillation. Furthermore, out of [3,4] and [12], the QSOs described in [3,4] are VM oscillators while the QSO described in [12] is a CM oscillator.

It has been observed from the literature survey that a new building block named Multi-output Current Follower Current Conveyor (MO-CFCC) which was also proposed in [2] has not been utilized for the realization of QSOs so far while recently its applications in the realization of simulated impedances have been reported in [26,27]. In this letter, we propose a QSO realized with CFCC to fill this void. The workability of the proposed oscillator has been verified using PSPICE simulations in 0.18 micron TSMC technology.

2. CFCC-based Realization of the Current-mode Quadrature Oscillator

CFCC [2] is a five-terminal ABB with the following functionalities: the current at the z terminal is an inverted copy of the input current at ‘p’ terminal; the terminal ‘i’ tracks the potential at the terminal z and two complementary currents at the output terminals are available which are copies of the current at the ‘i’ terminal. To provide
additional functionality to the CFCC, a copy of the current at the ‘z’ terminal may also be provided resulting in the Z-copy CFCC (ZC-CFCC). The symbolic representation and relations between the various port variables for this ABB are shown in Figure 1, and equation (1) respectively.

\[
\begin{align*}
&V_p \quad [0 \quad 0 \quad 0 \quad 0 \quad 0] \\
&I_z \quad [1 \quad 0 \quad 0 \quad 0 \quad 0] \\
&V'_i \quad [0 \quad 1 \quad 0 \quad 0 \quad 0] \\
&I_{z^+} \quad [0 \quad 0 \quad 1 \quad 0 \quad 0] \\
&I_{z^-} \quad [0 \quad 0 \quad -1 \quad 0 \quad 0] \\
&I_{2c} \quad [1 \quad 0 \quad 0 \quad 0 \quad 0] 
\end{align*}
\]

(1)

The symbolic notation of the ZC-CFCC

![Figure 1](image1.png)

Figure 1. Symbolic notation of the ZC-CFCC

The proposed current-mode quadrature oscillator circuit using CFCCs is shown in Figure 2 which consists of a parallel RLC resonator made from the CFCC along with R4, R5, C1 and C2 and a NIC-simulated negative resistor realized by the ZC-CFCC along with the resistors R1, R2 and R3.

![Figure 2](image2.png)

Figure 2. The proposed CFCC-based current mode quadrature oscillator

A straightforward routine analysis of the circuit given in Figure 2 yields the following characteristic equation (CE):

\[
s^3 R_d R_b R_d R_f R_c C_a C_b + s^2 R_d R_c R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b = 0.
\]

(2)

From equation (2), the condition of oscillation (CO) and frequency of oscillation (FO) are found to be:

**CO:**

\[
R_d R_b = R_c R_d.
\]

(3)

**FO:**

\[
\omega_{osc} = \frac{1}{\sqrt{R_d R_b C_a C_b}}
\]

(4)

However, for quadrature oscillator design, we must look into the interrelationship between \(I_{01}\) and \(I_{02}\) which is found to be

\[
\frac{I_{01}}{I_{02}} = \frac{R_1 R_3}{s R_2 R_d R_e C_1}.
\]

(5)

Under sinusoidal steady state, equation (5) reduces to

\[
\frac{I_{01}}{I_{02}} = \frac{j}{\omega_0 R_d C_1}.
\]

(6)

The phase difference \(\phi\) is, thus, equal to 90°.

### 3. Non-ideal Considerations

For a non-ideal analysis of the proposed oscillator, we have considered the various parasitic resistances and capacitances associated with the different terminals of the CFCC as follows: \(R_p\) represents the parasitic input resistance of the p port whereas \(R_z\) and \(C_z\) represent the resistance and capacitance at the z-terminal of the CFCC; similarly, \(R_{zc}\) and \(C_{zc}\) represent resistance and capacitance at the zc terminal while \(R_i\) represents the output resistance of the voltage buffer of the MO-CCII (resistance looking into the i terminal); on the other hand, \(R_i\) and \(C_i\) represent output resistance and output capacitance at the x+ terminals of the CFCC and finally, \(R_x\) and \(C_x\) represent the output resistance and the output capacitance at the x-terminals of the CFCC. A straightforward analysis of the proposed oscillator, incorporating all the parasitic immittances described above, gives the following third order characteristic equation:

\[
s^3 R_d R_b R_d R_f R_c C_a C_b + s^2 (R_d R_d R_d R_f R_c C_a C_b) + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f C_a C_b + s (R_d R_c R_c R_d R_f C_a C_b) + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b + R_d R_b R_c R_d R_f R_c C_a C_b = 0
\]

(7)

where

\[
\begin{align*}
R_d &= R_1 + R_p, \quad R_b = R_2 || R_z, \\
R_c &= R_3 + R_c, \quad R_d = R_d + R_p, \\
R_i &= R_5 + R_1, \quad R_f = R_{z+} || R_{z-}, \\
C_a &= C_1 + C_2 \quad \text{and} \quad C_b = C_2 + C_{zc} + C_{x+} + C_{x-}.
\end{align*}
\]

We have measured the values of the various parasitic resistances and capacitances of the CFCC employed in the present work (see Figure 3) by carrying out detailed PSPICE simulations. The measured values of these non ideal parameters are found to be as summarized in Table 1.
Table 1. The SPICE-measured values of the various parasitics of the CMOS ZC-CFCC

<table>
<thead>
<tr>
<th>S.No.</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$R_p$</td>
<td>591Ω</td>
</tr>
<tr>
<td>2</td>
<td>$R_i$</td>
<td>591Ω</td>
</tr>
<tr>
<td>3</td>
<td>$R_z$</td>
<td>4.9178 MΩ</td>
</tr>
<tr>
<td>4</td>
<td>$C_z$</td>
<td>6.2988x10^{-14} F</td>
</tr>
<tr>
<td>5</td>
<td>$R_{zc}$</td>
<td>4.9178 MΩ</td>
</tr>
<tr>
<td>6</td>
<td>$C_{zc}$</td>
<td>7.3938x10^{-15} F</td>
</tr>
<tr>
<td>7</td>
<td>$R_{x+}$</td>
<td>4.9862 MΩ</td>
</tr>
<tr>
<td>8</td>
<td>$C_{x+}$</td>
<td>7.3954x10^{-15} F</td>
</tr>
<tr>
<td>9</td>
<td>$R_{x-}$</td>
<td>4.7752 MΩ</td>
</tr>
<tr>
<td>10</td>
<td>$C_{x-}$</td>
<td>7.4062x10^{-15} F</td>
</tr>
</tbody>
</table>

The following approximations (ensured by selecting appropriate values of the terminating resistances and capacitances) $R_1$, $R_2$, $R_3$, $R_4$, $R_5$, $R_f$ and $C_1$, $C_2$ lead to a second order approximation of the CE from which the non-ideal CO and FO are now given by:

\[
(CO): \quad R_1 + R_p (R_2 + R_i) = (R_2 || R_z) (R_4 + R_p) \quad (8)
\]

\[
(FO): \quad \omega_{osc} = \omega_{osc} \frac{1}{\sqrt{1 + \frac{R_p}{R_4} \left(1 + \frac{R_i}{R_2} \left(1 + \frac{C_z}{C_1} \right) \right)}} \quad (9)
\]

The interrelationship between $I_{01}$ and $I_{02}$ is found to be

\[
\frac{I_{01}}{I_{02}} = \frac{R_2 (1 + sR_iC_a)}{R_3 (1 + sR_fC_a)} = \frac{R_2 [1 + s(R_2 || R_i)C_a]}{R_3 [1 + s(R_2 || R_i)C_a]} \quad (10)
\]

The phase difference $\phi$ is given by

\[
\phi = \tan^{-1} \omega_{osc} \frac{R_c}{C_a} - \tan^{-1} \frac{\omega_{osc} R_f C_z}{R_c} \quad (11)
\]

Subject to the approximations used in the non-ideal analysis, it may be observed that the phase difference between the two output currents would be very close to 90° (as the angle corresponding to the argument of the first arctangent term will be close to ninety degree because of the very large value of $R_z$ while the angle corresponding to the argument of the second arctangent term will be very small because of the very small value of $C_z$). We have calculated the phase difference using the values of different parasitic resistances and capacitances given in Table 1 as per equation (11) and found it to be equal to 89.52° for a design frequency of 1.59 MHz.

4. PSPICE Simulation Results

We now present some SPICE simulation results to demonstrate the workability of the proposed structure. The CMOS implementation of the CFCC [27] shown in Fig. 3 using 0.18 micron TSMC process technology has been used to verify the workability of the circuit presented in this paper. The values of the DC bias currents and voltages were taken as 40 µA and ±2.5V respectively. The oscillator was designed for a frequency of 1.59 MHz by appropriately selecting the passive components as follows: $C_1 = C_2 = 10$ pF, $R_1 = 10$ kΩ, $R_2 = 10^7$ Ω, $R_3 = 10$ kΩ, $R_4 = 10$ kΩ, $R_5 = 10$ kΩ. From SPICE simulations the oscillation frequency was found to be 1.50 MHz. The output waveforms are shown in Figure 4(a). The quadrature relationship of the generated waveform is indicated by the Lissajous pattern shown in Figure 4(b). The measured phase difference was found to be 90.91°. Figure 5 shows the FFT of the generated waveforms. The output voltages and currents were found to be 3.2% and 5.3% respectively. These simulation results thus, prove the workability of the proposed circuit.

5. Concluding Remarks

In this letter, a recently proposed active building block, namely, the CFCC has been used to devise a current-mode QSO. The workability of the circuit has been substantiated by SPICE simulations based on a CMOS CFCC implementable in 0.18 µm CMOS technology. The letter has, thus, added a new application of the CFCC in the area of quadrature oscillator realization, whose applications explored and known so far were only in the realization of the simulated impedances of various kinds [26,27].

Figure 3. An exemplary CMOS implementation of the ZC-CFCC [26]
Figure 4. PSPICE Simulation Results (a) Quadrature oscillator waveforms (b) Lissajous pattern

Figure 5. FFT of the waveforms of $I_{o1}$ and $I_{o2}$ at 1.50 MHz

References


