Current-Controlled Current-Mode Quadrature Oscillator Using Translinear Current Conveyors

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***Abstract*—In this paper, a current-mode quadrature oscillator using second-generation current conveyors (CCIIs) is presented. The proposed oscillator consists of two CCIIs, two grounded capacitors and two grounded resistors. The circuit is suitable for integrated circuit implementation by using grounded capacitors. In addition, a new current-controlled current-mode quadrature oscillator using two current controlled second generation current conveyors (CCCIIs) and two grounded capacitors can be obtained by replacing CCIIs and resistors series at X terminals with CCCIIs. The condition of oscillation and frequency of oscillation can be orthogonally controlled. The frequency of oscillation can be controlled by grounded resistors and external bias currents. The proposed circuits have been simulated by SPICE simulations. The simulation results are confirmed the proposed theory.**

***Keywords*—All-pass section, quadrature oscillator, current-mode circuit, second gerenation current conveyor (CCII), CCCII.**

I. Introduction

Quadrature oscillator typically provides two sinusoids with 90° phase difference for a variety of applications such as in telecommunications for quadrature mixers, single-sideband generators, direct-conversion receivers, or in measurement purposes for vector generators and for selective voltmeters [1], [2]. Over the years, many schemes of voltage-mode and current-mode quadrature oscillators have been presented, for instance, see [3]-[18]. The structures in [3]-[8] are suitable for monolithic integrated circuit (IC) implementation by using grounded capacitors. On the other hand, current-mode oscillators with high-output current sources are of great interest because these output current sources can be directly connected to loads without using buffer circuits [12]-[15]. Several current-mode quadrature oscillators have been proposed base on different techniques. In [8]-[11], the current-mode quadrature oscillators using single active device are proposed but the circuits suffer from several disadvantages such as complexity, employment a large number of grounded or floating passive devices and low-output impedance level. In 2002 and 2003, two quadrature oscillator circuits using two current differencing buffered amplifiers (CDBAs) [12] and two differential voltage current conveyors (DVCCs) [13] with two grounded capacitors and two grounded resistors have been proposed. In 2006, two quadrature oscillator circuits using a fully-differential second-generation current conveyor (FDCCII) with two grounded capacitors and two (or three) grounded resistors have been introduced by [14]. The current-mode quadrature oscillator using two current differencing transconductance amplifiers (CDTAs), two floating capacitors and four floating resistors is proposed [15]. The circuits in [12]-[15] provide high-output impedance current sources. However, all structures lack electronic tuning capability. Moreover, the current-mode oscillator circuit in [15] employs floating passive devices.

Recently, current-controlled oscillators based on translinear current conveyors (CCCII) [18] has been interested because the parameter frequency of oscillator can be varied by external bias current. This property makes it different from current conveyor-based oscillators [13]-[14], [16]. By applying high bias current to CCCII, CCCII can be operared as conventional second-generation current conveyor (CCII) (neglecting resistance at X-terminal). The quadrature oscillators using CCCII as active device are reported in [17], [18]. However, the circuits suffer from the use of excessive number of active components (four CCCIIs in [17], three CCCIIs in [18]). Also, the realization of quadrature oscillator in [18] suffers from the use of floating capacitor.

In this paper, a current-mode quadrature oscillator using two CCIIs, two grounded capacitors and two grounded resistors is presented. The circuit provides the advantage of using grounded capacitors that are beneficial to IC implementation. Furthermore, from the proposed CCII-based quadrature oscillator, by replacing CCII and resistors connected series at X-terminal by CCCIIs, a new current-controlled quadrature oscillator using only two CCCIIs and two grounded capacitors can be obtained. The circuits provide two high-output impedance current sources with 90º phase difference. The new topology are not found in open literature. The circuits are higher suitable for IC implementation when compared with the oscillator circuits in [3]-[18].

II. Proposed Circuit

Fig. 1 and Fig. 2 show the basic building blocks of the proposed current-mode quadrature oscillator circuit. Each circuit consists of a CCII, a grounded capacitor and a grounded resistor. The multiple-output CCII can be obtained by modifying from conventional CCII [19] by adding additional current-mirrors and cross-coupled current mirrors to obtain plus- and minus-type outputs [20]. The standard notations of CCII can be characterized by $V_X = V_Y$, $I_Z = \pm I_X$ and $I_Y = 0$. Using this relation and nodal analysis, the transfer functions of circuits in Fig. 1 and Fig. 2 can be respectively expressed as

$$\frac{I_{out}}{I_{in}} = \frac{sC_1R_1 - 1}{sC_1R_1 + 1} . \quad (1)$$

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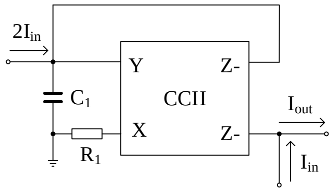


Fig. 1 Current-mode all-pass section.

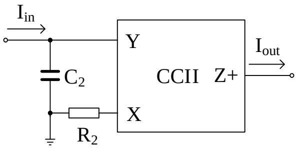


Fig. 2 Current-mode lossless integrator.

$$\frac{I_{out}}{I_{in}} = \frac{1}{sC_2R_2} \quad . \tag{2}$$

From (1) and (2), it should be noted that Fig. 1 is the all-pass section and Fig. 2 is the lossles integrator circuit. According to [21], the quadrature oscillator can be realized using an all-pass section and a lossless integrator. By using the all-pass section and lossless integrator circuits in Fig. 1 and Fig. 2, the proposed quadrature oscillator can be shown in Fig. 3. The structure is cascaded by all-pass filter circuit in Fig. 1, lossless integrator in Fig. 2 and feedback connection. The characteristic equation of Fig. 3 can be expressed by

$$s^2C_1C_2R_1R_2 + s\left(C_2R_2 - C_1R_1\right) + 1 = 0\,. \tag{3}$$

Letting R_1=R_2=R, the condition of oscillation and frequency of oscillation can be obtained respectively as

$$C_1 \geq C_2 \tag{4}$$

and

$$\omega_o = \frac{1}{R\sqrt{C_1C_2}}\,. \tag{5}$$

It is evident from (4) and (5), that the condition of oscillation can be adjusted by grounded capacitors C_1 and C_2 and the frequency of oscillation can be controlled by varying the grounded resistors R (i.e. R_1=R_2=R) without disturbing the condition of oscillation. From the circuit in Fig. 3, $CCII_2$ along with C_2 and R_2 form of the lossless integrator, the relationship between output currents I_{out1} and I_{out2} can be expressed as

$$\frac{I_{out2}}{I_{out1}} = \frac{1}{sC_2R_2} \tag{6}$$

where the phase shift is ϕ = π/2, which quarantees that proposed oscillator circuit provides the quadrature output currents I_{out1} and I_{out2}. The various passive sensitivities of the parameter ω_o with the passive devices are

$$S_{C_1}^{\omega_0} = S_{C_2}^{\omega_0} = S_{R_1}^{\omega_0} = S_{R_2}^{\omega_0} = -\frac{1}{2}\,. \tag{7}$$

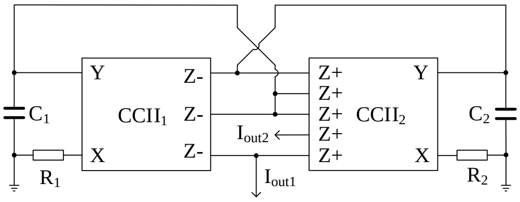


Fig. 3 Proposed CCII-based quadrature oscillator.

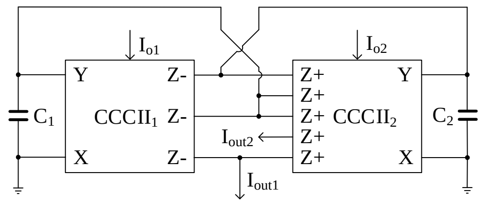


Fig. 4 Proposed CCCII-based quadrature oscillator.

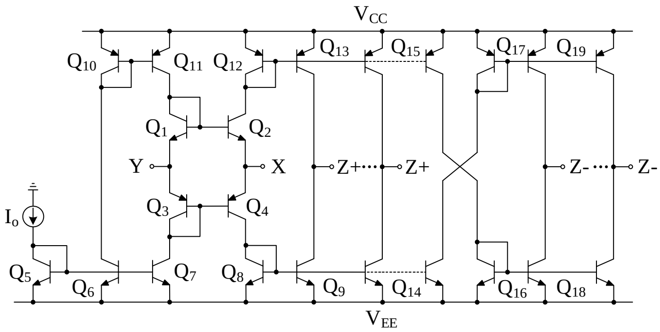


Fig. 5 Schematic implementation for CCCII [20].

Because the outputs I_{out1} and I_{out2} in Fig. 3 are the z-terminals of CCIIs which typically provide high-impedance level, hence the outputs I_{out1} and I_{out2} also possesses high-impedance level which can be directly connected to the loads.

The CCCII can be designed based on translinear elements and current mirrors [22]. The circuit has a finite resistance at X-terminal (R_X), where R_X = $V_T/2I_o$, I_o is the bias current and V_T is the thermal voltage ($V_T \cong$ 26 mV at 27 °C). Thus, CCCII will provide R_X at X-terminal that can be tuned by bias current I_o. From Fig 3, if CCIIs with resistors at the X-terminal are replaced by CCCIIs, a new CCCII-based quadrature oscillator can be obtained as shown in Fig. 4. The CCCII-based quadrature oscillator employs only two CCCIIs and two grounded capacitors. The characteristic equation of Fig. 4 can be expressed by

$$s^2C_1C_2R_{X1}R_{X2} + s\left(C_2R_{X2} - C_1R_{X1}\right) + 1 = 0 \tag{8}$$

or

$$s^2C_1C_2V_T^2 + s2V_T\left(C_2I_{o1} - C_1I_{o2}\right) + 4I_{o1}I_{o2} = 0\,. \tag{9}$$

Letting I_{o1} = I_{o2} = I_o, the condition of oscillation and frequency of oscillation can be expressed respectively as

$$C_1 \geq C_2 \tag{10}$$

and

$$\omega_o = \frac{2I_o}{V_T\sqrt{C_1C_2}}. \tag{11}$$

The condition of oscillation of Fig. 4 can be controlled by grounded capacitor C_1 and C_2 and frequency of oscillation can be controlled by varying the bias currents I_o (i.e. $I_o = I_{o1} = I_{o2}$). This implies that the circuit in Fig. 4 can be worked as current-controlled oscillator.

III. Non-Ideal Effects

Taking the non-idealities of the CCII into account, the relations of voltage and current characteristics can be expressed as

$$\begin{pmatrix} I_Y \\ V_X \\ I_Z \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 \\ \alpha & 0 & 0 \\ 0 & \pm\beta & 0 \end{pmatrix}\begin{pmatrix} V_Y \\ I_X \\ V_Z \end{pmatrix} \tag{12}$$

where α = 1-ε, $|\varepsilon|$«1 represents the voltage tracking error, β = 1-δ, $|\delta|$«1 represents the current tracking error. Using (12), the transfer function of Fig. 1 becomes

$$\frac{I_{out}}{I_{in}} = \frac{sC_1R_1 + 1 - 2\beta_1}{sC_1R_1 + 1} \tag{13}$$

where β_1 is the current gain error between z- and x-terminal of CCII1. Using (12), the transfer function of Fig. 2 becomes

$$\frac{I_{out}}{I_{in}} = \frac{\alpha_2\beta_2}{sC_2R_2}. \tag{14}$$

Therefore, the characteristic equation of Fig. 3 becomes

$$\begin{aligned} &s^2C_1C_2R_1R_2 + s\left(C_2R_2 - \beta_2^2\alpha_2C_1R_1\right) \\ &+\left(2\alpha_2\beta_1\beta_2^2 - \alpha_2\beta_2^2\right) = 0 \end{aligned}. \tag{15}$$

The modified condition of oscillation and frequency of oscillation are

$$\beta_2^2\alpha_2C_1 \geq C_2 \tag{16}$$

and

$$\omega_o = \frac{1}{R}\sqrt{\frac{2\alpha_2\beta_1\beta_2^2 - \alpha_2\beta_2^2}{C_1C_2}} \tag{17}$$

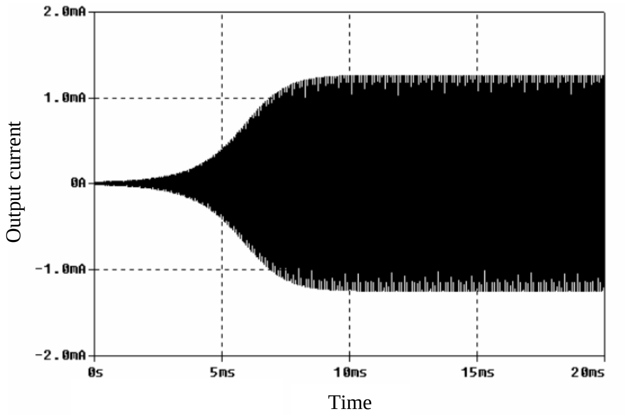
where $R = R_1 = R_2$. From (16) and (17), the voltage and current errors slightly change the condition of oscillation and frequency of oscillation. However, the condition of oscillation and frequency of oscillation can be orthogonally controlled. The active and passive sensitivities of ω_o are analysed and found within 0.5 in magnitude which express good sensitivity performance of the circuit.

IV. Simulation Results

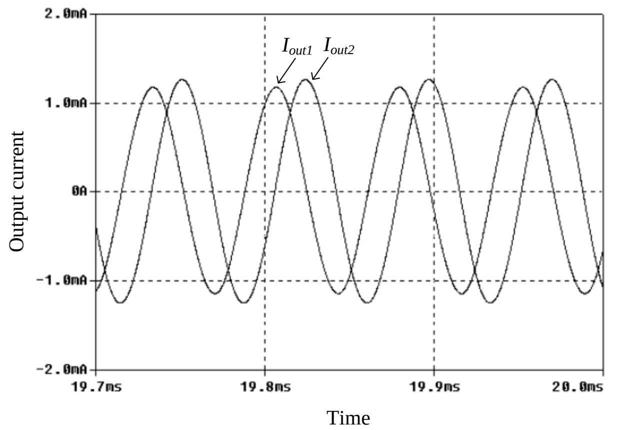
The proposed circuits were simulated using SPICE simulations. The current conveyor in Fig. 5 [20] was performed with the transistor model of NR100N and PR100N of the bipolar arrays ALA400 from AT&T [23] as listed in Table I. The voltage supply was taken as V_{CC} = 3 V, V_{EE} = -3 V. By taking high bias current I_o, CCCII in Fig. 5 can be worked as CCII (R_X is negleged). To achieve the CCII, the circuit in Fig. 5 was biased with I_o = 250 μA ($R_X \approx 50\ \Omega$). The proposed quadrature oscillator Fig. 3 was designed with R_1 = R_2 = 1 kΩ, C_1 = 12 nF, and C_2 = 10 nF, where C_1 was designed to be larger than C_2 to ensure the oscillator will start. The condition R_1, R_2 » R_X should be used to avoid the effect of parasitic resistance R_X. These passive-values were used to design the frequency of oscillation of 14.52 kHz.

Table I
Model parameters of NR200N and PR200N transistors

*NR200N-2X NPN TRANSISTOR .MODEL NX2 NPN RB=262.5 IRB=0 RBM=12.5 RC=25 RE=0.5 IS=242E-18 EG=1.206 XTI=2 XTB=1.538 BF=137.5 IKF=13.94E-3 NF=1 VAF=159.4 ISE=72E-16 NE=1.713 BR=0.7258 IKR=4.396E-3 NR=1 VAR=10.73 ISC=0 NC=2 TF=0.425E-9 TR=0.425E-8 CJE=0.428E-12 VJE=0.5 MJE=0.28 CJC=1.97E-13 VJC=0.5 MJC=0.3 XCJC=0.065 CJS=1.17E-12 VJS=0.64 MJS=0.4 FC=0.5
*PR200N-2X PNP TRANSISTOR .MODEL PX2 NPN RB=163.5 IRB=0 RBM=12.27 RC=25 RE=1.5 IS=147E-18 EG=1.206 XTI=1.7 XTB=1.866 BF=110.0 IKF=4.718E-3 NF=1 VAF=51.8 ISE=50.2E-16 NE=1.65 BR=0.4745 IKR=12.96E-3 NR=1 VAR=9.96 ISC=0 NC=2 TF=0.610E-9 TR=0.610E-8 CJE=0.36E-12 VJE=0.5 MJE=0.28 CJC=0.328E-12 VJC=0.8 MJC=0.4 XCJC=0.074 CJS=1.39E-12 VJS=0.55 MJS=0.35 FC=0.5

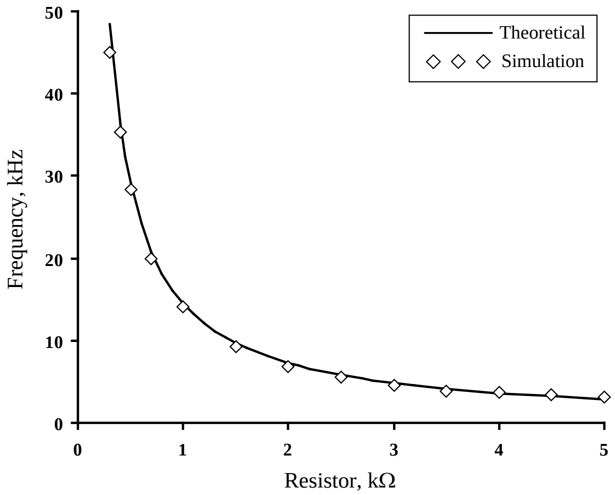


(a)

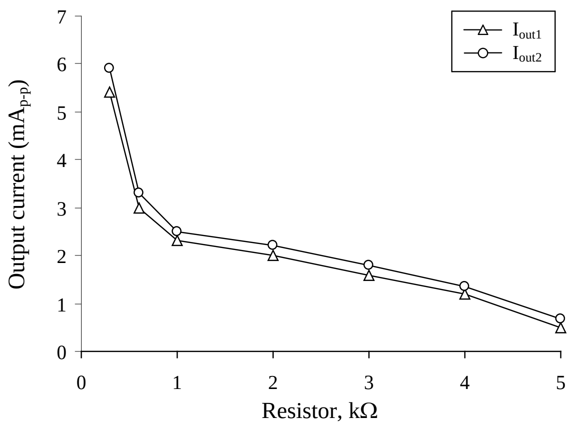


(b)

Fig. 6 The simulated output waveforms I_{out1} and I_{out2} of the proposed oscillator of Fig. 3, (a) at transient stage; (b) at steady stage.



(a)



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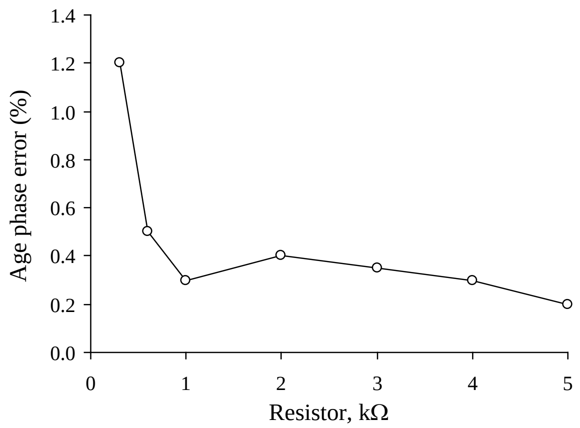

(c)

Fig. 7 Simulated results of the proposed circuit when the value of resistor R (R=R_1=R_2) is varied: (a) the frequency of oscillation, (b) the amplitude of output currents, (c) the average of phase error.

Fig. 6(a) and Fig. 6(b) show the current output waveforms I_{out1} and I_{out2} of the proposed oscillator. From Fig. 6, simulated frequency of oscillation of 13.5 kHz was expressed whereas the theoretical value was 14.52 kHz. The frequency of oscillation was 13.5 kHz instead of 14.52 kHz owing the effect of non-ideal CCII. According to (17), this drop-off would be caused by voltage and current tracking errors of CCII. In practice, the value of the voltage and current tracking errors depend on the implementation scheme that is used to realize the CCCII, for example, $\beta \approx 0.999$ and $\alpha \approx 0.991$ for bipolar technology with the ALA200 transistor arrays [23]. Fig. 6(b), the I_{out1} and I_{out2} total harmonic distortions (THD) were 3% and 0.98%, respectively. In this case, the power consumption was about 25 mW. Fig. 7(a) shows the plot of the frequency of oscillation when the value of resistor R was varied ($R = R_2 = R_2$) between 300 Ω and 5 kΩ. Note that if the resistors R_1 and R_2 are replaced by JFETs, the voltage-controlled oscillator can be obtained [23]. Under the same condition in Fig. 7(a), the amplitude of output currents I_{out1} and I_{out2} was ploted as shown in Fig. 7(b) and average phase error between currents I_{out1} and I_{out2} that deviate from 90º was ploted as shown in Fig. 7(c). The THD of currents I_{out1} and I_{out2} when the resistor R was varied ($R = R_2 = R_2$) between 300 Ω and 5 kΩ was shown in Fig. 8.

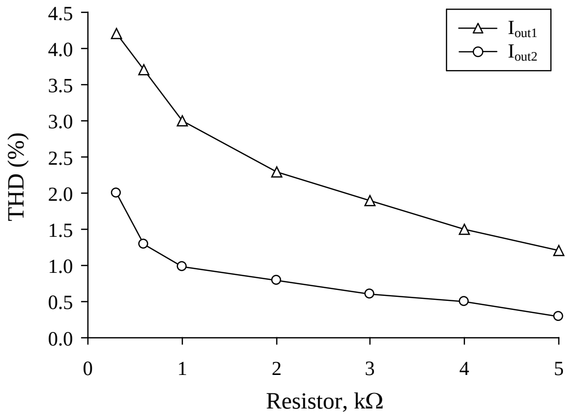

Fig. 8 Simulated total harmonic distortion of I_{out1} and I_{out2}.

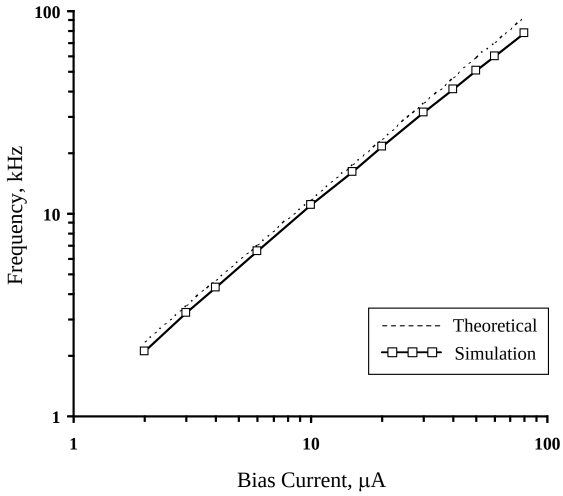

Fig. 9 Simulated the frequency of oscillation of Fig. 4 when the bias currents I_o (I_o=I_{o1}=I_{o2}) was varied.

The proposed current-controlled quadrature oscillator in Fig. 4 was simulated. The circuit was tested using V_{CC} = 3 V, V_{EE} = -3 V voltage supply, C_1 = 12 nF and C_2 = 10 nF. Because the operation of Fig. 4 is similar to Fig. 3, thus only the frequency of oscillation was investigated. Fig. 9 shows the simulated frequency of oscillation when the bias currents were varied between 2 μA and 100 μA. The theoretical value was also included to compare. This result is confirmed (11).

V. CONCLUSION

In this paper, a new current-mode quadrature oscillator based on current conveyors is presented. The proposed circuit uses two CCIIs, two grounded capacitors and two grounded resistors. The proposed CCII-based quadrature oscillator can be modified to a current-controlled CCCII-based quadrature oscillator by replacing the CCIIs and resistor series at the X terminal with CCCIIs. The new current-controlled quadrature oscillator uses only two CCCIIs and two grounded capacitors. The circuit provides two high output impedance current sources with 90º phase difference and the frequency of oscillation can be electronically controllable. The simulation results are agreeble with the theory. The proposed structure is suitable IC implementation and not found in open literature. Compared with quadrature oscillators in previous researches, the proposed oscillator employs equal active and passive elements, but the proposed oscillator provides electronic tuning capability. Compared with CCCII-based quadrature oscillators in previous researches, the proposed structure employs lesser active and passive components.

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