A Simple Current-Mode Quadrature Oscillator Using Single CDTA

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Abstract. *This article presents a simple current-mode quadrature oscillator using a single Current Differencing Transconductance Amplifier (CDTA) as the active element. The oscillation condition and oscillation frequency can be electronically controlled. The circuit structure is very simple, consisting of merely one CDTA, one resistor and two capacitors. The proposed circuit is suitable for IC architecture. The PSpice simulation and experimental results are shown, and the results agree well with the theoretical assumptions.*

Keywords

Oscillator, CDTA, current-mode.

1. Introduction

Oscillator is an important basic building block, which is frequently employed in electrical engineering applications. Among several kinds of oscillator, the quadrature oscillator is widely used because it can offer sinusoidal signals with 90° phase difference, for example, in telecommunications for quadrature mixers and single-sideband modulators [1]. Presently, the current-mode technique has been more popular than the voltage-mode technique. This is due to the requirements in low-voltage environments such as portable and battery-powered equipment. The current–mode technique is ideally suited for this purpose. Today, there is a growing interest in synthesizing current-mode circuits because of their potential advantages such as larger dynamic range, higher signal bandwidth, greater linearity, simpler circuitry, and lower power consumption [2].

A reported 5-terminal active element, namely Current Differencing Transconductance Amplifier (CDTA) [3], seems to be a versatile component in the realization of a class of analog signal processing circuits, particularly analog frequency filters [3], [4]. It is actually a current-mode element whose input and output signals are currents. In addition, the output current of the CDTA can be electronically adjusted. Besides, a modified version of CDTA, whose parasitic resistances at two current input ports can be electronically controlled, has been proposed in [5]. This circuit is called Current Controlled Current Differencing Transconductance Amplifier (CCCDTA). Another CDTA modification, called ZC-CDTA (Z Copy CDTA) is proposed in [6], providing a copy of the current flowing to the *z* terminal. This copy can be used as an output signal for driving an independent load.

Several implementations of oscillator employing CDTAs or CCCDTAs have been reported in the literature [7-13]. Unfortunately, these circuits suffer from one or more of the following weaknesses: They use more than one CDTA or CCCDTA and an excessive number of passive elements, which is not convenient for IC fabrication. In addition, some reported circuits use a multiple-output CDTA or CCCDTA. Consequently, the circuits become more complicated.

The purpose of this paper is to introduce a current-mode quadrature oscillator, based on a single CDTA. The oscillation condition and oscillation frequency can be adjusted electronically. The circuit construction consists of one CDTA, one resistor, and two capacitors. Finally, this oscillator has been built by means of a CDTA chip which has been fabricated in the CMOS technology [14]. The PSpice simulation and the experimental results correspond to the theoretical analyses.

2. Principle of Operation

2.1 Current Differencing Transconductance Amplifier (CDTA)

Since the proposed circuit is based on the CDTA, a brief review of CDTA is given in this Section. The characteristics of the ideal CDTA are represented by the following hybrid matrix:

$$\begin{bmatrix} V_p \\ V_n \\ I_z \\ I_x \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 1 & -1 & 0 & 0 \\ 0 & 0 & 0 & g_m \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ V_x \\ V_z \end{bmatrix}. \tag{1}$$

In general, CDTA can contain an arbitrary number of x terminals, providing currents I_x of both directions. As an example, the symbol and the equivalent circuit of the CDTA with a pair of $x+$ and $x-$ terminals are illustrated in Fig. 1(a) and (b), respectively.

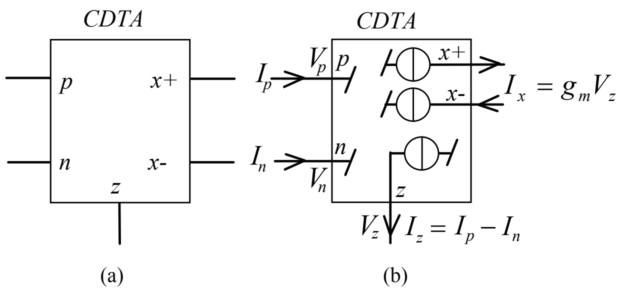

Fig. 1. CDTA (a) schematic symbol, (b) equivalent circuit.

2.2 Proposed Circuit

The proposed quadrature oscillator is designed by cascading a first-order all-pass filter and a non-inverting lossless integrator as shown in Fig. 2. Based on this block diagram, the single-CDTA quadrature oscillator can be implemented according to Fig. 3. In order to utilize the current through the capacitor C_2, an auxiliary zc terminal is used. The internal current mirror provides a copy of the current flowing out of the z terminal to the zc terminal [6].

The characteristic equation of the proposed oscillator in Fig. 3 can be expressed as follows:

$$s^2 C_1 C_2 R + s(C_2 - C_1 g_m R) + g_m = 0. \tag{2}$$

From Eq. (2), it can be seen that the proposed circuit can produce oscillations if the oscillation condition is fulfilled:

$$g_m R = \frac{C_2}{C_1}. \tag{3}$$

For example, this condition can be achieved by setting

$$C_1 = C_2 \text{ and } g_m = 1/R. \tag{4}$$

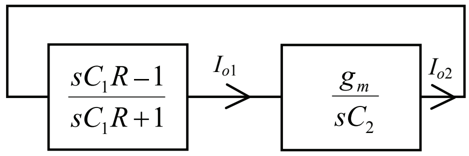

Fig. 2. Block diagram of the quadrature oscillator.

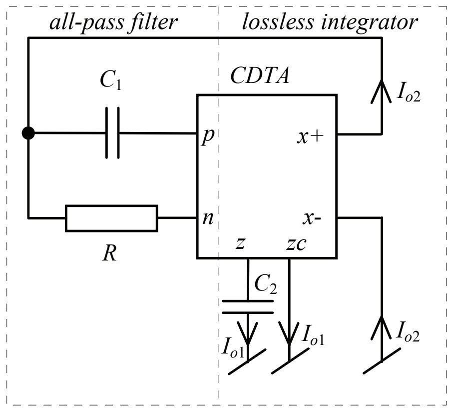

Fig. 3. Proposed CDTA-based quadrature oscillator.

Then the characteristic equation of the system becomes

$$s^2 C_1 C_2 R + g_m = 0. \tag{5}$$

From (5), the oscillation frequency is as follows:

$$\omega_{osc} = \sqrt{\frac{g_m}{C_1 C_2 R}}. \tag{6}$$

From the circuit in Fig. 3, the current transfer function from I_{o1} to I_{o2} is

$$\frac{I_{o2}(s)}{I_{o1}(s)} = \frac{g_m}{sC_2}. \tag{7}$$

For sinusoidal steady state, equation (7) becomes

$$\frac{I_{o2}(j\omega)}{I_{o1}(j\omega)} = \frac{g_m}{\omega C_2} e^{-j90^\circ}. \tag{8}$$

The phase difference ϕ between I_{o1} and I_{o2} is

$$\phi = -90^\circ, \tag{9}$$

ensuring that the currents I_{o2} and I_{o1} are in quadrature.

For the oscillation frequency, with regard to (6), equation (8) gives

$$\frac{I_{o2}(j\omega_{osc})}{I_{o1}(j\omega_{osc})} = \sqrt{g_m R \frac{C_1}{C_2}} e^{-j90^\circ}. \tag{10}$$

Taking into account oscillation condition (3), one can conclude that the oscillator will provide quadrature signals of equal magnitudes.

All the active and passive sensitivities of the oscillator are low as shown in (11):

$$S^{\omega_{osc}}_{C_1, C_2, R} = -\frac{1}{2}, S^{\omega_{osc}}_{g_m} = \frac{1}{2}. \tag{11}$$

3. Experimental Results

To prove the performance of the proposed circuit, the oscillator was constructed using the CDTA chip. The CDTA was designed and manufactured in the 0.7µm CMOS technology [14]. Its basic small-signal parameters are as follows:

- Input resistances of *p* and *n* terminals $R_p = 260\ \Omega$, $R_n = 250\ \Omega$.
- Resistance of *z* terminal $R_z = 3.48\ \mathrm{M\Omega}$.
- Resistance of *x*+, *x*- terminals $R_x = 2.2\ \mathrm{M\Omega}$.
- Current gains from *p* to *z* terminal and from *n* to *z* terminal are 1 and -1, respectively, with a 3-dB cutoff frequency of 10 MHz.
- Transconductance of the OTA amplifier is 1.22 mA/V with a 3-dB cutoff frequency of 16 MHz.

The CDTA chip was manufactured with the fixed transconductance, with *x*+ and *x*- terminals, and without the auxiliary *zc* terminal. During the experiment, the current through the *x* terminal was sensed via a I/V converter, formed by an by operational amplifier OPA 2650 with 1 kΩ feedback resistance. That is why the output voltage in millivolts corresponded to the input current in microamperes.

The quadrature oscillator was designed for an oscillation frequency of about 1MHz. The values of the external elements were designed as follows:

$$R = 1/g_m = 819\ \Omega,\ C_1 = C_2 = 150\ \text{pF}. \qquad (12)$$

According to (6), the theoretical value of the oscillation frequency is 1.294 MHz.

Actually, the measured values (12), as used in the experiment, were

$$R = 829\ \Omega,\ C_1 = 153\ \text{pF},\ C_2 = 156\ \text{pF}. \qquad (13)$$

Then the corresponding theoretical value of the oscillation frequency is 1.257 MHz.

After substituting the values from (13) to (3), we can see that the left-side value is greater than the right-side one. In other words, the current loop-gain is greater than unity, which is the condition of self-starting the oscillations.

The waveform measured, namely the output of the I/V converter measuring the current I_{o2} in Fig. 3, is shown in Fig. 4. The magnitude is approximately 75 µA. The measured frequency of 962 kHz is less than the theoretical value 1.257 MHz. In the following Section, an analysis of the real effects leads to an explanation.

Fig. 5 shows the measured spectrum components of the generated waveform. The magnitude of the second harmonics is more than 56 dB below the magnitude of the fundamental harmonics. The corresponding THD is 0.16 %. Increasing the value of *R* results in increasing the loop-gain and thus increasing the magnitude and also the distortion of generated signals.

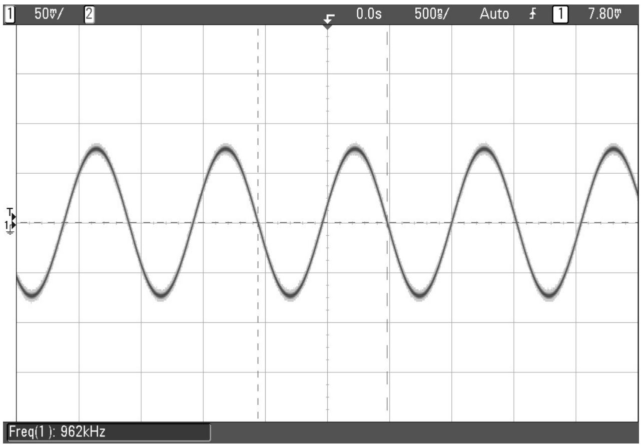

Fig. 4. Measured output waveform in millivolts, corresponding to the current I_{o2} in microamperes.

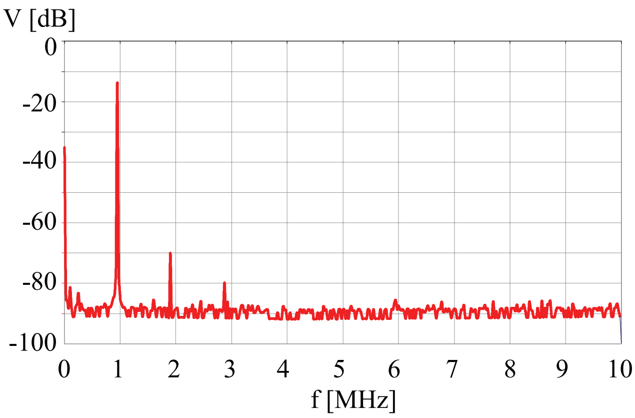

Fig. 5. Fourier analysis of generated waveform.

4. Analysis of Real Effects

The circuit operation is affected by linear and nonlinear influences. In this Section, an analysis of real effects is done in the following structure:

a) Analysis of linear effects. The oscillation frequency and the condition of oscillation are affected by parasitic impedances of CDTA terminals, by DC current transfers from *p* to *z* and from *n* to *z* terminals, which are not generally equal to one in magnitude, and by their frequency dependence. Also, the frequency characteristic of the transconductance may have an effect.

The analysis of the above linear model of the oscillator can be performed in fully symbolic form in order to obtain formulas for the oscillation condition and for the frequency of oscillation. However, more complex high-order models do not provide such results in a closed form. That is why the symbolic analysis will be evaluated only for several main influences, and the remaining one will be analyzed separately via computer simulation.

b) Analysis of nonlinear effects. There are two types of important nonlinearities of manufactured CDTA: nonlinear curves $V_p = f_p(I_p)$ and $V_n = f_n(I_n)$, and nonlinear OTA characteristic $I_x = f_x(V_z)$. These nonlinearities are responsible for properly starting - up the oscillations, for automatically achieving the oscillation condition in the steady state, and they also affect the value of the oscillating frequency.

A hand-and-paper analysis of such nonlinear effects is practically impossible. Therefore, all these characteristics were measured on the chip and transferred to PSpice as nonlinear behavioral models, built via controlled sources and the look-up table method. These models were then completed with several small-signal parameters which PSpice cannot extract from the above nonlinear curves. Such a final model has been tested and its behavior has been compared with real experiments. The coincidence was excellent. This PSpice model of the CDTA chip helped us with the analysis of nonlinear effects in the quadrature oscillator. The simulations based on this model also provide waveforms which are not easily accessible via practical measurements.

4.1 Linear Effects

The most important small-signal real influences which affect the circuit operation are modeled in Fig. 6.

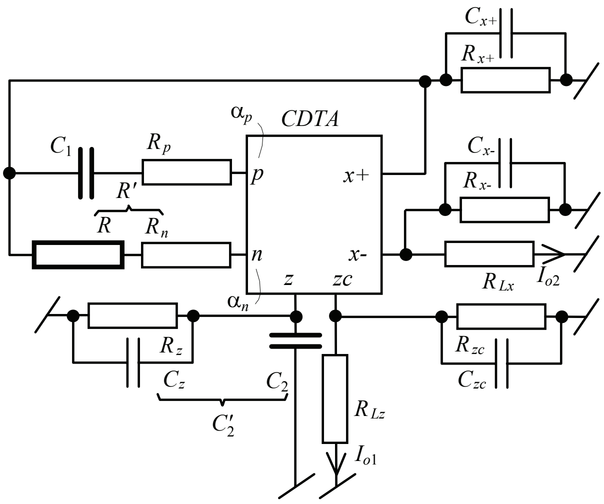

Fig. 6. Modeling linear effects in the oscillator.

In addition to the working passive elements R, C_1, and C_2 (in thick lines), several parasitic impedances show up here. The input Current Differencing Unit (CDU) is now described by the equation

$$I_z = \alpha_p I_p - \alpha_n I_n , \tag{14}$$

where the parameters α_p and α_n are current transfer values, deviating from one, depending on the internal circuit construction. R_{Lz} and R_{Lx} are loading impedances for sensing the output currents I_{o1} and I_{o2}. In conjunction with the R_{zc}, C_{zc}, or R_{x-}, C_{x-} pair they form a frequency-dependent current divider of terminal current I_{zc} or I_{x-}. Making the influence of these parasitic impedances negligible, one should choose the load resistances as low as possible. Throughout the following test, we follow this assumption. Then the remaining non-ideal factors will be as follows:

- General values of α_p and α_n. We will take them into account for the symbolic analysis in order to evaluate this influence. However, for the manufactured CDTA chip, these values are of unity-value with sufficiently high precision.
- Nonzero R_p and R_n resistances. R_n is in series to R and thus it does not change the character of this working impedance. The R_n influence can be compensated by decreasing the R value. This is not the case of R_p, which acts in series to C_1.
- Parasitic impedance of the z terminal. C_z can be considered a part of C_2. However, R_z changes the type of the impedance, which should normally be of purely capacitive character.
- Parasitic impedance of the x+ terminal. The x+ terminal is connected to low-impedance p and n terminals through C_1 and R elements. For low frequencies, the corresponding impedance level is given by R' (kilo-Ohms), which is much smaller than R_{x+}. For the high-frequency region, this impedance level is determined by R_p (hundreds of Ohms), but C_{x+} can cause a leakage of the I_{x+} current to the ground. This capacitive part of parasitic impedance increases the order of the oscillator model from the original value 2 to 3. This fact considerably complicates the evaluation of the symbolic analysis. That is why we will consider this part of the model in the second stage of the analysis.
- Frequency dependence of α_p and α_n transfer coefficients and of OTA g_m. According to the information in Section 3, the corresponding cutoff frequencies are 10 MHz and 16 MHz. For the oscillation frequencies near these values, the effect can be considerable. However, the order of circuit model is also increased. That is why we will analyze these factors separately, after the symbolic analysis.

Considering the above factors and using the notation

$$R' = R + R_n,\ C_2' = C_2 + C_z,\ G_x = 1/R_{x+},\ G_z = 1/R_z ,$$

the symbolic analysis of the model in Fig. 6 leads to the following equations:

The modified oscillation frequency

$$\omega_{osc}' = \sqrt{\frac{\alpha_n g_m + (1 + R'G_x)G_z}{C_1 C_2 (R' + R_p + R'R_p G_x)}} . \tag{15}$$

The modified condition of oscillation

$$g_m (R'\alpha_p - R_p a_n) = \frac{C_2}{C_1}(1 + R'G_x) + (R' + R_p + R'R_p G_x)G_z \tag{16}$$

Setting the numerical values of the parameters from Section 3 and a short re-arrangement lead to the conclusion that several terms in (15) and (16) can be neglected. The simplified versions of (15) and (16) are as follows:

$$\omega'_{osc} \approx \sqrt{\frac{\alpha_n g_m}{C_1 C_2 (R' + R_p)}} = \sqrt{\frac{\alpha_n g_m}{C_1 C_2 (R + R_p + R_n)}} \Rightarrow$$

$$\omega'_{osc} \approx \frac{\omega_{osc}}{a},\ a = \sqrt{1 + \frac{R_p + R_n}{R}}, \qquad (17)$$

$$g_m (R'\alpha_p - R_p a_n) \approx \frac{C_2}{C_1} \Rightarrow$$

$$g_m R\alpha_p + g_m (R_n \alpha_p - R_p a_n) \approx \frac{C_2}{C_1}. \qquad (18)$$

Equation (17) shows that the oscillation frequency decreases below its theoretical value due to nonzero parasitic resistances R_p and R_n. For $R_p = 260\ \Omega$, $R_n = 250\ \Omega$, and $R = 829\ \Omega$ (see Eq. 13), the frequency should decrease 1.172 times, i.e. from the theoretical value 1.257 MHz to 1.072 MHz. However, the real experiments yielded the value 962 kHz (see Section 3). Obviously, there are some additional influences that decrease the oscillation frequency and which were not taken into account throughout the above analysis. To find these influences in the frame of the linear models, the SNAP program [15] for symbolic analysis was used for additional modeling of the parasitic impedance of $x+$ terminal, and for modeling the cutoff frequencies of α_p, α_n, and g_m. The analysis has confirmed that all these factors decrease the oscillation frequency. For example, parasitic capacitance C_{x+} of only 1 pF decreases the frequency to 980 kHz, and the cutoff frequencies of α_p, α_n, and g_m decrease it by another 15 kHz. We can conclude that such additional modifications of oscillation frequency are approximately by one order less relevant than the changes described by (17).

Equation (18) demonstrates that for the identical values of parasitic resistances R_p and R_n and for $\alpha_p = \alpha_n = 1$, the oscillation condition is practically not affected by the above effects.

Another analysis of the model in Fig. 6 under the above simplifications leads to the conclusion that the ratio of current I_{o2} and current I_{o1} is still described by (8), thus their phase shift is 90 degrees. However, their magnitudes will not be identical any more. Utilizing (8), (17) and (18) and a short re-arrangement yield the following formula:

$$\frac{I_{o2}(j\omega'_{osc})}{I_{o1}(j\omega'_{osc})} \approx \sqrt{\frac{R' + R_p}{\alpha_n (R'\alpha_p - R_p \alpha_n)}}\, e^{-j90°}. \qquad (19)$$

For ideal values of $\alpha_p = \alpha_n = 1$, the magnitude ratio $b = abs(I_{o2}/I_{o1})$ is as follows:

$$b = \sqrt{\frac{R + R_n + R_p}{R + R_n - R_p}} \doteq 1.279. \qquad (20)$$

The above results can be summarized in a practical formula: If the oscillation frequency decreases a times due to parasitic linear effects, where a is given by (17), then the magnitudes of generated quadrature signals will no longer be equal but their ratio will be b, where b is given by (19) or (20). This formula will be confirmed in the following Subsection via PSpice simulation.

4.2 Nonlinear Effects

Nonlinear effects were studied by means of PSpice simulation, utilizing nonlinear models of the CDTA chip, briefly described in Section 4. Two CDTA models were used subsequently, first model No. 1 and then model No. 2.

Model No. 1: A simplified model of the CDU consists of linear R_p and R_n resistances, controlled source for modeling the difference of input currents, the z-terminal parasitic resistance, and single-pole frequency response with 3dB cutoff frequency. For the internal OTA, a full nonlinear model of the $I_x = f_x(V_z)$ from Fig. 7, resulting from the data measured [16], was used. In addition, the output resistances of x-terminals and the 3dB cutoff frequency were also modeled.

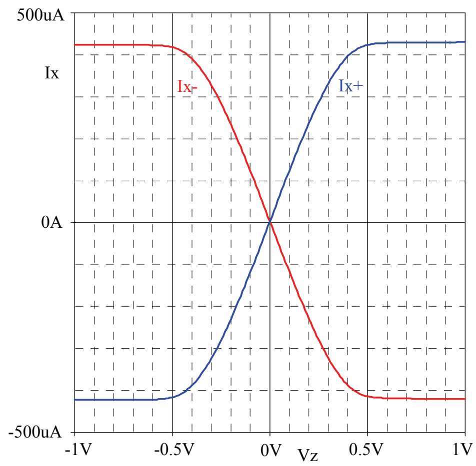

Fig. 7. Measured OTA characteristics, transferred to PSpice.

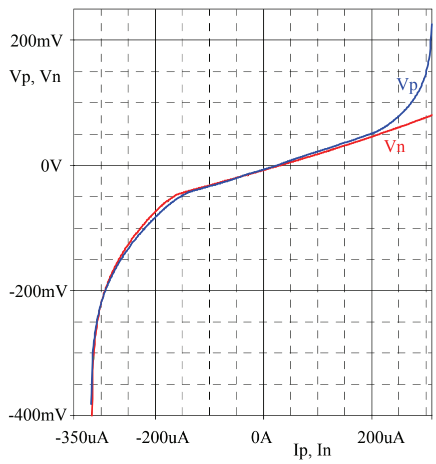

Fig. 8. Measured input characteristics of Current Differencing Unit, transferred to PSpice.

Model No. 2: A full nonlinear model of the CDU is implemented here, utilizing the measured nonlinear characteristics $V_p = f_p(I_p)$ and $V_n = f_n(I_n)$ [16], [17], see Fig. 8. The other parts of the model are identical with model No. 1.

The results of the steady-state analysis by means of model No. 1 are shown in Fig. 9. Starting up the oscillations was attained for R = 864 Ω, which is in good agreement with real measurements (829 Ω). For such a resistance value, the theoretical oscillation frequency is 1.224 MHz. The simulated value is 960.3 kHz, i.e. 1.275 times smaller. It is obvious that the generated waveforms are not of equal magnitudes. Their ratio was measured as 1.128, which roughly corresponds to the ratio 1.173, computed from (20).

As shown in Fig. 9, the peak value of the voltage at z terminal is below 200 mV. It is obvious from Fig. 7 that the transconductance amplifier works in nearly linear regime for such a voltage swing. Its equivalent transconductance is decreasing when increasing the voltage swing. The equilibrium, i.e. fulfilling the oscillation condition, is established for a concrete voltage swing. That is why the automated adjustment of the oscillation condition is due to the nonlinearity of the OTA stage.

Fig. 9 illustrates an interesting phenomenon: in spite of different magnitudes of output currents I_{o1} and I_{o2}, the magnitudes of currents I_p and I_n are equal. It can be easily proved that I_p and I_n are also orthogonal. Their magnitudes are approximately 150 µA.

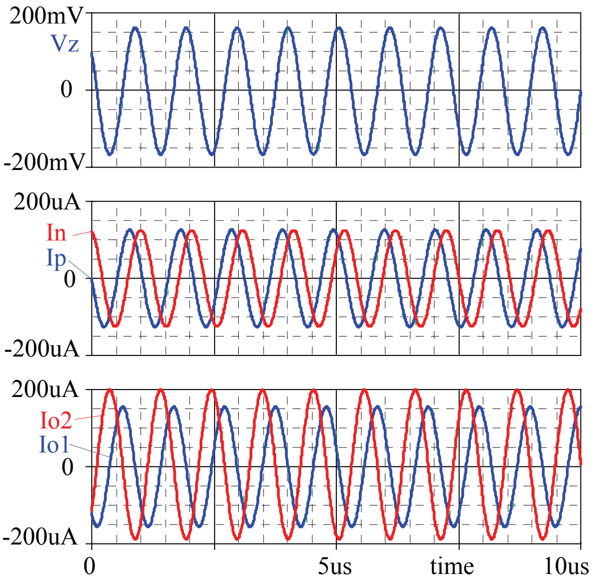

Fig. 9. Simulated steady-state oscillations, CDTA model No. 1, f_{osc} =960 kHz.

The simulation results for CDTA model No. 2 is given in Fig. 10. The parameters of the passive elements are the same as for the previous simulation. The oscillation frequency is now 964 kHz, which is in very good agreement with the real measurement (962 kHz). Also, the magnitude of I_{o2} (73 µA) now corresponds to the reality. Note that the signal levels are now lower than without detailed modeling of the CDU. This is due to the nonlinearity of p and n inputs. The R_p and R_n equivalent resistances are now dependent on signal levels, and the nonlinearity of CDU provides the automatic adjustment of the oscillation condition. It should be also noted that a DC offset appears in the waveforms, particularly in I_{o2}, which is caused by the offset of DC characteristics of the CDU (see Fig. 8).

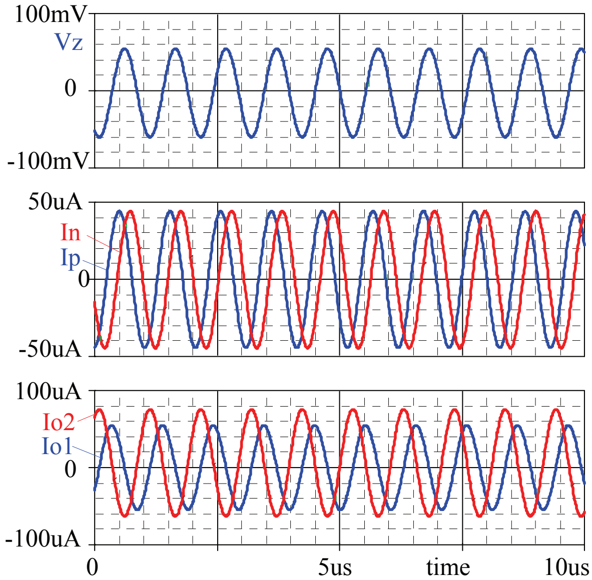

Fig. 10. Simulated steady-state oscillations, CDTA model No. 2, f_{osc} = 964 kHz.

4.3 Elimination of Real Effects

It is obvious from (17) and (20) that both real effects, i.e. decreasing the oscillation frequency and different magnitudes of generated quadrature signals, can be eliminated via increasing the working resistance R to be at least by one order higher than the parasitic resistances R_p and R_n. Let us suppose that R will be increased k times. In order to preserve the designed value of the oscillation frequency (6) and also to fulfill the condition of oscillation (3), several possible operations can accompany the R increase. Two relevant operations are summarized below:

- decreasing C_1 k times, or
- increasing g_m k times, increasing C_2 k times, decreasing C_1 k times.

The second method cannot be applied to the oscillator with CDTA chip, which has a fixed transconductance. Let us increase R ten times, to 8.64 kΩ and decrease C_1 ten times, to 15.3 pF. The results of PSpice simulation for model No. 2 is in Fig. 11. Note that the oscillation frequency is now near the theoretical value, namely 1.171 MHz, and the quadrature signals are of almost equal magnitudes.

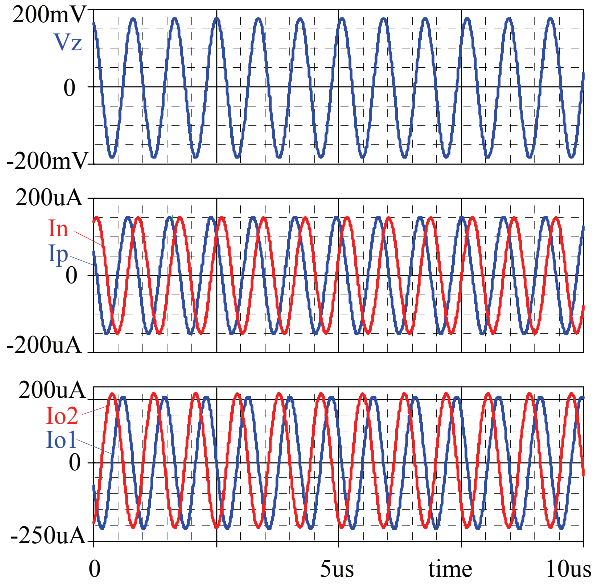

Fig. 11. Simulated steady-state oscillations, CDTA model No. 2, with elimination of real effects, f_{osc} = 1.171 MHz.

5. Conclusion

A simple quadrature oscillator based on one Current Differencing Transconductance Amplifier (CDTA), two capacitors and one resistor is described. The proposed oscillator, designed for the megahertz range, was constructed using a CDTA chip, which was fabricated in the CMOS technology. The measurements confirmed the functionality of the principle. Two problems were discussed: 1) The oscillation frequency is lower than the theoretical value. 2) The quadrature output currents are of different magnitudes although the oscillator was designed for equal magnitudes. Theoretical analysis shows the common reason for both problems: the finite values of R_p and R_n resistances of the CDTA. A quantitative description of these phenomena is given in equations (17) and (20). Based on this knowledge, a simple procedure is designed how to overcome the above difficulties.

For a practical implementation of the proposed topology of quadrature oscillator, it is useful if the CDTA has at least two *x*-terminals and one so-called *zc* terminal, which provides a copy of the *z*-terminal current. For an easy elimination of the above real influences, the current differencing unit of the CDTA should be designed with the R_p and R_n parasitic resistances as low as possible. The proposed value is at least one order lower than the value of the resistance of the working resistor connected in series to the *n* terminal.

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