An electronically tunable current-mode quadrature oscillator using PCAs

HERENCSÁR, N.; LAHIRI, A.; VRBA, K.; KOTON, J.

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Norbert Herencsar[a]*, Abhirup Lahiri[b], Kamil Vrba[a] and Jaroslav Koton[a]

[a]*Department of Telecommunications, Faculty of Electrical Engineering and Communication, Brno University of Technology, Purkynova 118, 612 00 Brno, Czech Republic*

[b]*36-B, J and K Pocket, Dilshad Garden, Delhi-110095, India*

The paper presents a new realization of active RC sinusoidal oscillator with electronically tunable condition and frequency of oscillation. As compared to the class of three resistors, two capacitors (3R-2C) based canonic oscillators, the proposed circuit here uses only two resistors and two capacitors as the passive components and still provides non-interactive tuning laws for the condition of oscillation (CO) and the frequency of oscillation (FO). The proposed circuit employs new bipolar programmable current amplifier (PCA) as the active building block and is capable of simultaneously providing two explicit quadrature current outputs. SPICE simulation results have been included to verify the workability of the circuit as an oscillator and the tuning range of the FO.

1. Introduction

Current amplifiers (CAs) with adjustable gains (also referred as the programmable current amplifier (PCAs)) have been presented in various forms in the past by several researchers. The first mention in the open technical literature is about the large gain current controlled current source (CCCS) amplifier from 1980 (Allen and Terry 1980). Another concept of current amplifier called translinear operational current operational amplifier (TOCA) was introduced by Fabre in 1986. In the more recent publications (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009; Alzaher 2008; Pukkalanun, Tangsrirat and Yoshida 2009), authors presented new concepts of current amplifiers. These include the use of current multiplier EL2082 and second-generation current conveyor (CCII) in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008), the use of a current follower and balanced output transconductance amplifier (BOTA) in (Koton, Vrba and Herencsar 2009) and the use of current differencing network (CDN) to implement the digitally controlled current follower (DCCF) in (Alzaher 2008; Pukkalanun, Tangsrirat and Yoshida 2009). In fact, the programmable current amplifier (PCA) realization presented in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008) is the

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*Corresponding author. Email: herencsn@feec.vutbr.cz

electronically-tunable second-generation current conveyor (ECCII) (Surakampon-torn and Thitimajshima 1988; Minaei, Sayin and Kuntman 2006) in case the voltage terminal y is grounded. The works in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009; Alzaher 2008) discusses applications of CA in biquadratic filter design and (Pukkalanun, Tangsrirat and Yoshida 2009) discusses its application in precision-rectifier design. These recent works indicate the growing interest in creating application circuits using CAs and which have advantageous features of conventional active RC counterparts using active elements from the VOA-CC family (Biolek, Senani, Biolkova and Kolka 2008). One of the primary advantages is the tunability of the circuit parameters using electrical signals like voltage or current. Also, ingenious use of CA (which can be considered as a special type of generalized current followers) can/may lead to circuit solutions with reduced/minimum passive component count. The motivation of this paper is to highlight the aforementioned advantages of the PCA presented in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009) in electronically tunable quadrature oscillator (ETQO) design.

It is well known that any active RC sinusoidal oscillator providing independent control of the condition of oscillation (CO) and the frequency of oscillation (FO), requires the use of at least three resistors and two capacitors (Gupta and Senani 2004). This class of 3R-2C "single-resistance-controlled oscillators" (SRCOs) has been extensively researched and a large catalogue of its realizations is available in the literature (see (Gupta and Senani 1998; Aggarwal 2004; Singh, Sharma, Singh, Bhaskar and Senani 2005; Tangsrirat, Prasertsom, Piyatat and Surakompontorn 2008; Bhaskar, Abdalla and Senani 2009; Kumngern and Dejhan 2009; Lahiri 2010; Tangsrirat and Surakampontorn 2009) and references cited therein). On the contrary, the proposed oscillator circuit here requires a bare minimum of two resistors and two capacitors (thereby reducing the passive component count by one). The circuit offers independently tunable condition of oscillation (CO) and the frequency of oscillation (FO) by means of the PCA current gains (unlike the circuits in (Gupta and Senani 1998; Aggarwal 2004; Singh, Sharma, Singh, Bhaskar and Senani 2005; Tangsrirat, Prasertsom, Piyatat and Surakompontorn 2008; Bhaskar, Abdalla and Senani 2009; Kumngern and Dejhan 2009; Lahiri 2010; Tangsrirat and Surakampontorn 2009) which do not offer any direct electronic tuning of the CO and the FO) and is capable of simultaneously providing two explicit quadrature current outputs. This feature of the circuit makes it suitable to be used as current-mode (CM) quadrature oscillator (QO) (a function generator producing $90°$ phase shifted sine waves) and which is desirable in communication circuits like quadrature mixers and sideband generators (Holzel 1993).

The rest of the paper is organized as follows. The conception of bipolar PCA is provided in Section 2, the proposed oscillator circuit is described in Section 3, a brief non-ideal analysis of the proposed circuit is provided in Section 4 and further discussions/comparisons of the presently reported circuit with other electronically tunable oscillators is done in Section 5. SPICE simulation results using a bipolar implementation of PCA have been included in Section 6 and which verify the functioning of the proposed circuit as an ETQO.

2.   Description of the PCA

In general, the PCA element (Figure 1) can be characterized by the following equations:

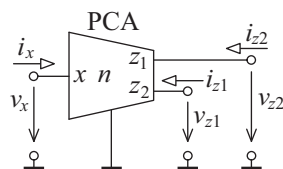$$v_x = 0, \quad i_{z1} = b_1 n i_x, \quad i_{z2} = b_2 n i_x, \tag{1}$$

Figure 1. Schematic symbol of general PCA

where $b_1, b_2 \in \{1, -1\}$ and n represents the current gain, referred to as the "mu-factor" in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009). Depending on the values of b_1 and b_2 three variants of PCA are possible, of which PCA+/− has been used in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009) for creating CM biquadratic filters. The prototype using the implementation in (Koton, Vrba and Herencsar 2009) is ready to be manufactured by AMI Semiconductor Czech, Ltd., (part of the ON Semiconductor, Ltd.) in the CMOS 0.35 μm and labeled as COAK-NAA. It is to be mentioned that the DCCF in (Alzaher 2008) and PCA in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009) have the same characterizing equations and have balanced output currents, however, as it evident from (1) that a general PCA can have three possible variants, namely PCA+/+, PCA+/−, and PCA−/−.

2.1   *Current mirror with adjustable current gain*

The cascode npn current mirror with adjustable current gain by the external bias currents is shown in Figure 2. The I_{in} and I_{out} are the input and output signal currents. Transistors Q_1–Q_4 function as a classical translinear loop, and the currents I_1 and I_2 are the external DC bias currents (Toumazou, Lidgey and Haigh 1990). In addition, the cascode stages Q_5 and Q_6 provide the high output impedance and also lead to minimize the severe peaking of the frequency responses (Fabre and Mimeche 1994). Applying the translinear principle and assuming that all the transistors are well matched with the common-emitter current gains $\beta \gg 1$, then the relationship of the collector currents can be characterized by the following equation:

$$I_{C1}I_{C3} = I_{C2}I_{C4}, \tag{2}$$

where $I_{C1} = I_1$, $I_{C2} = I_2$, $I_{C3} = I_{in}$, and $I_{C4} = I_{out}$. Therefore, the output current I_{out} of this circuit becomes:

$$I_{out} = nI_{in}, \tag{3}$$

where n is the current gain of the mirror (in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008; Koton, Vrba and Herencsar 2009) referred to as the "mu-factor")
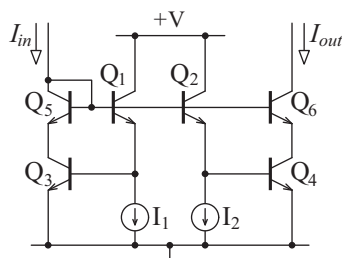

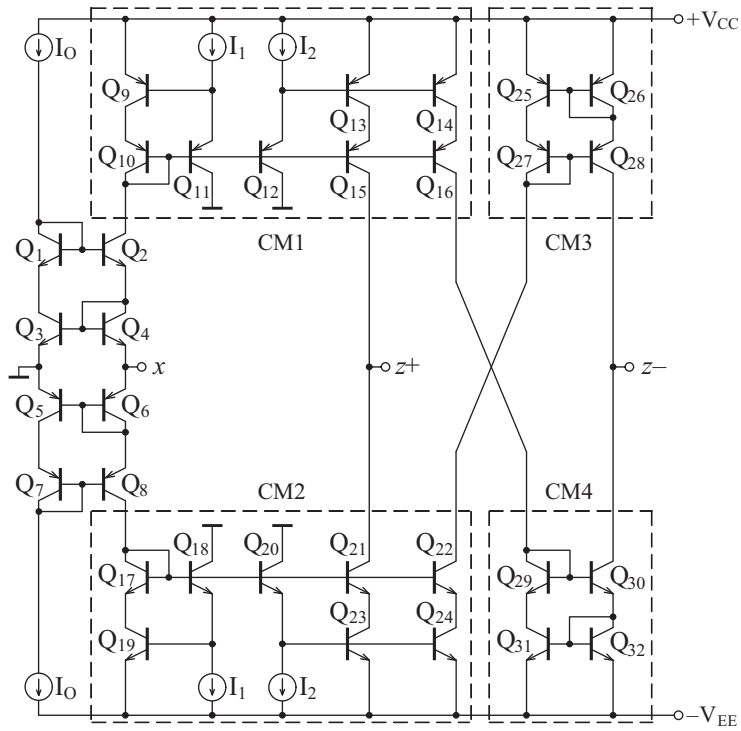Figure 2. Cascode npn current mirror with adjustable current gain

Figure 3. Bipolar implementation of the PCA+/−

and equals to the ratio of the external bias currents I_1/I_2.

2.2   *Proposed PCA*

Bipolar implementation of the plus and minus type programmable current amplifier (PCA+/−) based on the CCII with controlled current gain (Fabre and Mimeche 1994) is shown in Figure 3. Group of transistors Q_1–Q_8 form an improved translinear cell, in which Q_3–Q_6 function as a dual translinear loop. Ideally, it is required that the pair of transistors Q_3–Q_4 and Q_5–Q_6 are closely matched. The translinear cell performs a current follower, where it allows an input current i_x to source and sink at the terminal x. By two complementary variable gain current mirrors CM1 and CM2, the current i_x flowing through the port x will be reflected and inverted to the ports $z+$ and $z-$ with the current transfer ratio of n ($= i_z/i_x = I_1/I_2$). The output impedance at the port x is low, since it is looking into the emitters of translinear cell's transistors, while the output impedances of the ports $z\pm$ are high due to the effective parallel combination of output impedances of the current mirrors. Therefore, this device will provide a current transfer between ports x and $z\pm$ that the gain value is equal to n as it is defined by (1).

3.   **Proposed circuit**

The proposed circuit using only three PCAs, two resistors and two capacitors is shown in Figure 4. Using (1) and doing routine circuit analysis yields the following characteristic equation (CE):

$$s^2 C_1 C_2 R_1 R_2 + s[C_1 R_2(1 - n_1) + C_2 R_1 - n_1 n_3 C_1 R_2] + 1 + n_1 n_2 - n_1 = 0. \quad (4)$$

For the start-up of oscillation the roots of the CE should be in the right-hand
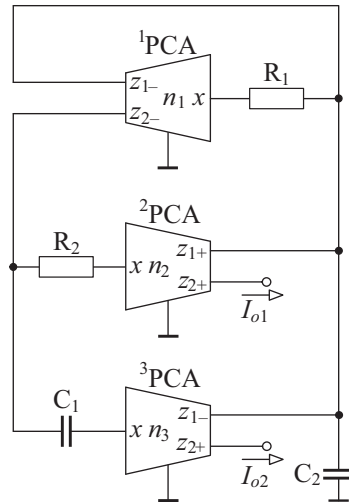
Figure 4. The proposed current-mode electronically tunable quadrature oscillator

plane and which indicates that the coefficient of 's' term in (4) should be negative. Thus the condition of oscillation (CO) and the frequency of oscillation (FO) are given as:

$$\text{CO:}\quad C_1 R_2 (1 - n_1) + C_2 R_1 \leq n_1 n_3 C_1 R_2, \tag{5}$$

$$\text{FO:}\quad f_o = \frac{1}{2\pi} \sqrt{\frac{1 + n_1(n_2 - 1)}{C_1 C_2 R_1 R_2}}. \tag{6}$$

It is evident from (5) and (6) that the CO and FO are independently tunable by means of the mu-factors n_3 and n_2, respectively. For the specific case when $n_1 = 1$ (i.e. PCA is reduced to simply an inverting current follower) the CO and FO in (5) and (6) can be rewritten as:

$$\text{CO:}\quad C_2 R_1 \leq n_3 C_1 R_2, \tag{7}$$

$$\text{FO:}\quad f_o = \frac{1}{2\pi} \sqrt{\frac{n_2}{C_1 C_2 R_1 R_2}}. \tag{8}$$

The two marked explicit current outputs (ECOs) in Figure 4 are related as:

$$|I_{o2}| = jk|I_{o1}|, \tag{9}$$

where

$$k = \frac{\omega_0 n_3 C_1 R_2}{n_2}. \tag{10}$$

It is evident from (9) and (10) that the ECOs are ideally $90°$ phase shifted and have equal amplitudes for $k = 1$.
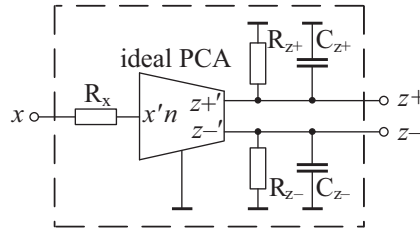
Figure 5. Model of the PCA+/− including parasitic elements

4. Non-ideal analysis

In this Section we briefly describe the important PCA non-idealities that affect the oscillator behavior, particularly the effects on the oscillator CO and FO.

- The non-ideal PCA suffers from current tracking errors between the input and output currents. Let β_{ij} represent the current transfer gains from the input terminal to j^{th} output terminal (either $z+$ or $z-$ terminal) of the ith PCA, respectively. Ideally these values are unity, but they differ slightly from unity ($\beta_{ij} = 1 - \varepsilon_{ij}$) by small current tracking errors ε_{ij} ($|\varepsilon_{ij}| \ll 1$). Considering these non-idealities the expressions in (5) and (6) modify to:

$$\text{CO:} \quad C_1 R_2 (1 - n_1 \beta_{11-}) + C_2 R_1 \leq n_1 n_3 \beta_{12-} \beta_{31-} C_1 R_2, \tag{11}$$

$$\text{FO:} \quad f_o = \frac{1}{2\pi} \sqrt{\frac{1 + n_1(n_2 \beta_{12-} \beta_{21+} - \beta_{11-})}{C_1 C_2 R_1 R_2}}. \tag{12}$$

Use of cascode mirrors (as in our scheme) reduces the current tracking errors associated with current following. The passive f_o sensitivities using (12) are:

$$|S^{f_o}_{C_1, C_2, R_1, R_2}| = \frac{1}{2}, \tag{13}$$

which indicates a satisfactory sensitivity performance. The active f_o sensitivities with respect to n_1, n_2, β_{12-}, β_{21+}, and β_{11-} are also less than unity in magnitudes.

- Another non-ideality that should be considered is the non-zero input parasitic resistance of the PCA, denoted by R_x. This parasitic resistance is absorbed in the external impedance connected at the input of the PCA, if it is of resistive nature. However, if the external impedance is a capacitor then the presence of the parasitic resistance would change the character of the impedance, which should be purely capacitive in nature. Considering Figure 4, it is evident that the parasitic resistances at [1]PCA and [2]PCA are absorbed in the external resistors R_1 and R_2, respectively. For [3]PCA, the effect discussed in (Fabre, Saaid and Barthelemy 1995) can be alleviated by considering the operating frequency $\omega_0 < 1/C_1 R_{x3}$. The parasitic resistance R_x associated with the low input impedance x terminal can be reduced by increasing the value of bias current I_O.

- The parasitic resistance $R_{z_{ij}}$ and parasitic capacitance $C_{z_{ij}}$ appear between the high output impedance z_j (where $j = 1, 2$) of the ith PCA and ground. The parasitic capacitance $C_{z_{21+}}$ and $C_{z_{31-}}$ are absorbed into external capacitor C_2 as they appear in shunt with it. Considering that the input terminals of PCA are at virtual ground (i.e. considering $R_x \to 0$), then the parasitic resistance $R_{z_{12-}}$ and capacitance $C_{z_{12-}}$ appear in parallel with external resistor R_2 and

capacitor C_1, respectively. The effects of these parasitics can be alleviated by considering external capacitors $C_1, C_2 \gg C_{z_{ij}}$, external resistor $R_2 \ll R_{z_{12-}}$ and the operating frequency $\omega_0 > 1/C_2(R_{z_{21+}}||R_{z_{31-}})$.

Considering the aforementioned non-idealities, the quadrature relation between the two marked explicit current outputs (ECOs) in Figure 4 is slightly modified to:

$$I_{o2}(s) = \frac{s\beta_{23}n_3C_1(R_2 + R_{x2})}{\beta_{22}n_2(1 + sC_1R_{x3})}I_{o1}(s). \tag{14}$$

Using reasonable assumptions that the operating frequency $\omega_0 \ll 1/C_1R_{x3}$ and the external resistor $R_2 \gg R_{x2}$, (14) approximates to the a similar equation as in (9):

$$I_{o2}(s) = \frac{s\beta_{23}n_3C_1R_2}{\beta_{22}n_2}I_{o1}(s). \tag{15}$$

5.   Brief discussion

It should be pointed that conventional unity gain current followers and current inverters (Gupta and Senani 2004) are a special class of more general PCA. PCAs simulate these unity gain cells when the $n = 1$. Thus, in a way the proposed realization here can be equivalently created using a generalized current follower (GCF) with dual/multiple outputs and which can act as either a current follower or a current conveyor. Note, however, that the resultant oscillator circuit using GCF will not have non-interactive tuning laws (a degree of freedom which is sacrificed by making $n_i = 1$). As pointed previously, a simpler realization takes place, when [1]PCA is replaced by a dual-output inverting current follower (DO-ICF). This realization still provides independent electronic control of the CO and the FO, as according to (7) and (8).

Another important point that is to be addressed is the aspect of tunability. Over the last few years, numerous oscillator realizations were reported, which provide tunable CO and FO. These are primarily based on either operational transconductance amplifier (OTA) and variants as in (Jaikla and Siripruchyanun 2007; Sotner, Jerabek, Petrzela and Vrba 2009), or second-generation current controlled conveyors (CCCII) as in (Maheshwari and Khan 2007; Maheshwari 2003; Maheshwari and Khan 2005) or OTA/CCCII hybrid elements as in (Jaikla, Siripruchyanun, Bajer and Biolek 2008; Herencsar, Koton, Vrba and Lahiri 2009; Lahiri 2009a; Lahiri 2009b; Herencsar, Koton, Vrba and Lattenberg 2010; Linares-Barranco, Rodriguez-Vazquez, Sanchez-Sinencio and Huertas 1992; Herencsar, Vrba, Koton and Lahiri 2010; Keskin and Biolek 2006). The electronic control in such circuits is via the transconductance and which is a function of the bias current ($\propto I_B/2V_T$ in BJTs, where V_T is the thermal voltage or $\propto \sqrt{2I_B\mu_0C_{ox}(W/L)}$ in MOSFETs). Considering that I_B is temperature compensated (e.g. bias derived from a first-order band-gap provides a low temperature compensated (TC) bias current), then temperature variations in g_m in BJTs is primarily due the V_T term ($\propto T$) and in MOSFETs due to the μ term ($\propto T^{-1.5}$, typically). Thus, FO controlled by g_m varies inversely with temperature. Similarly, current from beta-multiplier current source $I_B \propto \frac{1}{\mu_0C_{ox}(W/L)R^2}$ will lead to temperature compensated g_m only if low TC resistor R is used. In such a case, the control of g_m (and thus the CO and FO) can be achieved only by varying R (creating electronically tunable R, for e.g.

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using MOSFETs working in triode region, would again feature temperature dependent terms) (Talebbeydokhti, Hanumolu, Kurahashi and Moon 2006). On the contrary, the controlling mu-factor in PCA proposed here is a ratio of two currents (preferably, temperature compensated) and thus FO variations with temperature are primarily due to the variations in passive components values (due to non-zero their temperature coefficients) and non-ideal current tracking coefficients, which have non-zero TCs. Low TC resistors should be used wherever FO spread with temperature is to minimized. A detailed discussion on this topic and corresponding simulation results are beyond the scope of the present communication; the information is added just to provide a basic insight into how the presented scheme differs from OTA-C and CCCII-C oscillators.

6.   Simulation results

To prove the performances of the proposed PCA$+/-$ in Figure 3 and CM ETQO in Figure 4, SPICE simulations have been performed for examination. The PNP and NPN transistors employed in the proposed circuit were simulated by respectively using the parameters of the PR100N and NR100N bipolar transistors of ALA400-CBIC-R transistor array from AT&T (Frey 1993) and the DC supply voltages were taken as ± 3 V. Bias currents $I_O = 200$ μA and $I_1 = I_2 = 100$ μA have been chosen. The parasitic elements in Figure 5 of the proposed PCA$+/-$ in Figure 3 have been computed as $R_x = 69.816$ Ω, $R_{z+} = 14.280$ kΩ, $R_{z-} = 13.961$ kΩ, $C_{z+} = 2.829$ pF, and $C_{z-} = 1.359$ pF. The AC current transfer characteristics from the port x to the ports $z\pm$ ($i_{z\pm}/i_x$) are shown in Figure 6. The DC current gains are $\beta_{z+} \cong 0.9343$ and $\beta_{z-} \cong 0.9346$ with bandwidths $f_{\beta_{z+}} \cong 125.053$ MHz and $f_{\beta_{z-}} \cong 132.304$ MHz. The maximum values of terminal currents without producing significant distortion have been computed as $i_{z\pm,max} \cong 19.684$ mA. The total power dissipation of the PCA$+/-$ is calculated as 5.9 mW.

The possibility of adjusting the current gain n (I_1/I_2) by the external bias current I_1 of the proposed PCA$+/-$ is shown in Figure 7(a). In simulations the external bias current I_1 has been adjusted in the interval from 10 μA to 1 mA (equal to gain $n = 0.1$ to 10) whereby I_2 is set to be constant at 100 μA. Figure 7(b) shows the possibility of adjusting the current gain n by external voltage V_G of the previously presented implementation of the PCA$+/-$ in (Herencsar, Koton, Vrba, Lattenberg
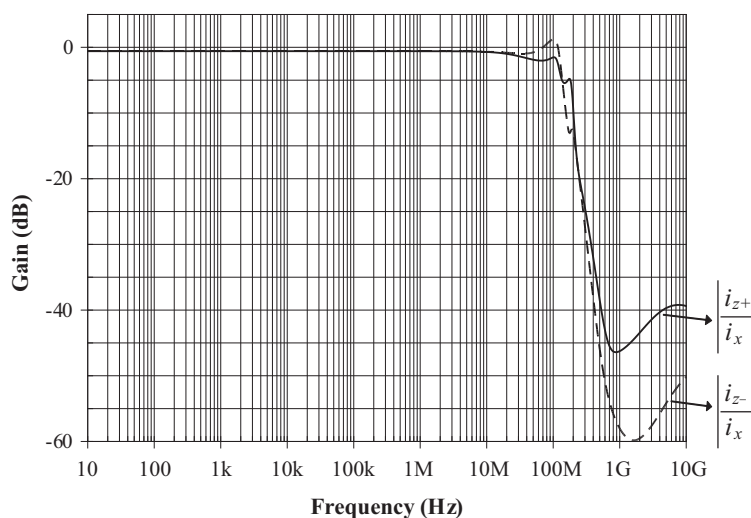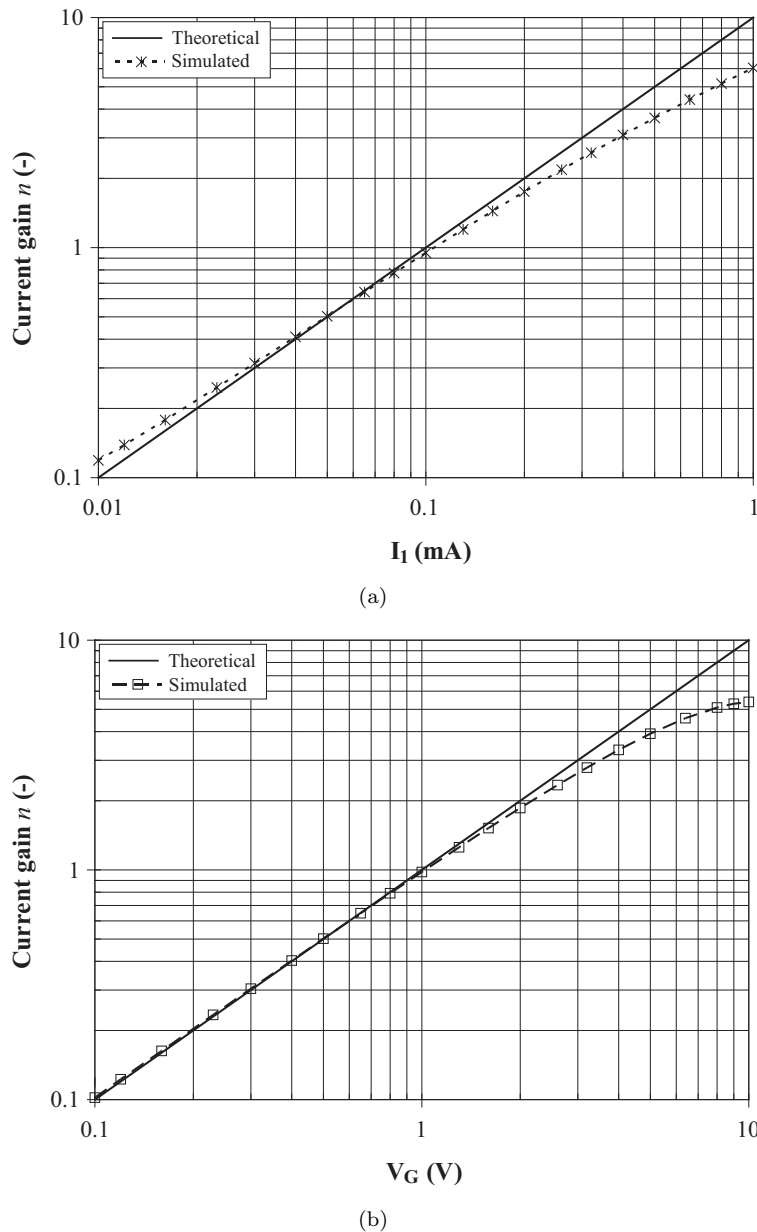

Figure 6. AC simulation results of the PCA$+/-$

(a)

(b)

Figure 7.   Possibility of adjusting the current gain n of the PCA+/–: (a) by the I_1 (when I_2=100 μA) of the bipolar implementation shown in Figure 3 and (b) by V_G of the implementation in (Herencsar, Koton, Vrba, Lattenberg and Misurec 2008)

and Misurec 2008). In this simulation the external voltage V_G has been adjusted in the same current gain interval, i.e. $n = 0.1$ to 10 ($V_G = 0.1$ V to 10 V). From simulations it is evident that in both cases the most exact result can be obtained for current gain $n = 0.4$ to 1. According to Datasheet EL2082, the gain error of the current multiplier EL2082 for $V_G = 2$ V is about –3.8 %. In the case of the proposed bipolar implementation, the gain error can be affected by using not precisely matched bipolar transistors.

To verify the theoretical study, the proposed CM ETQO in Figure 4 was designed using the following passive and active component values: $C_1 = C_2 = 1$ nF, $R_1 = R_2 = 1$ kΩ, and $n_1 = n_2 = 1$, where the value of the mu-factor n_3 of the [3]PCA was chosen as $n_3 = 1.45$ to start the oscillations. In this case the external DC bias current I_1 of [3]PCA was set as 145 μA when I_2 was set to be constant as
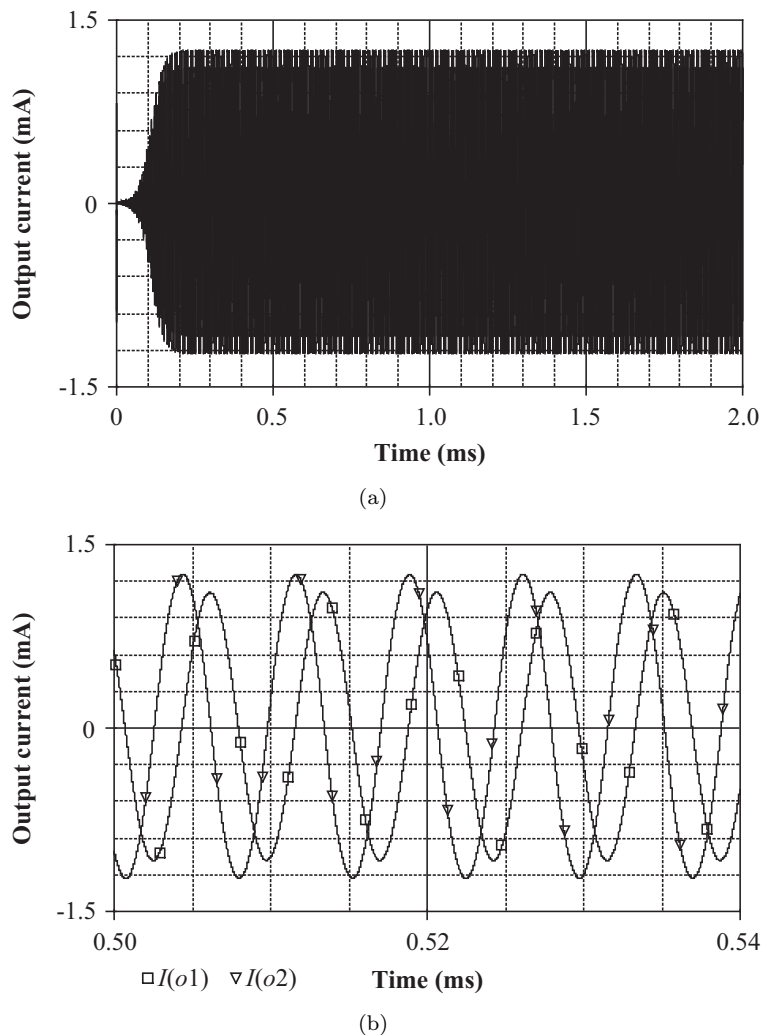
Figure 8.   Simulated output waveforms I_{o1} and I_{o2} of the proposed oscillator in Figure 4: (a) at transient stage and (b) at steady stage

100 μA. This yields oscillation frequency of 138 kHz, where the calculated value of this parameter from Eq. (8) yields 159.2 kHz. The simulated waveforms of the quadrature current outputs I_{o1} and I_{o2} in transient state and steady state are shown in Figure 8(a) and Figure 8(b), respectively. The steady state oscillations were reached within 200 μs after turning the supply sources on. No auxiliary amplitude control circuit is used for amplitude stabilization, rather the non-linearity of the active device itself limits the amplitude. If however, tight amplitude control is required, then the peak amplitude (magnitude sensing by either peak-detector or valley-detector (Park, Wilson and Ismail 2006)) can be sensed and resulting voltage can be used to reduce the bias current I_2 of [3]PCA to reduce n_3. This will reduce positive feedback action and help in achieving reduced amplitude of oscillation with low total harmonic distortion (THD). For the chosen active parameters and passive component values, theoretically, the ratio of amplitudes of the generated orthogonal signals according to (9) should be 1.45 and through simulations the values comes out to be 1.14. Figure 9 shows the frequency spectrums of the output waveforms and the value of THD at both the outputs are less than 2%. The variation of FO with mu-factor n_2 (by DC bias current I_1 of the [2]PCA when $I_2 = 100$ μA) without affecting the CO is shown in Figure 10. The total power dissipation of the proposed CM ETQO is calculated as 24.4 mW.
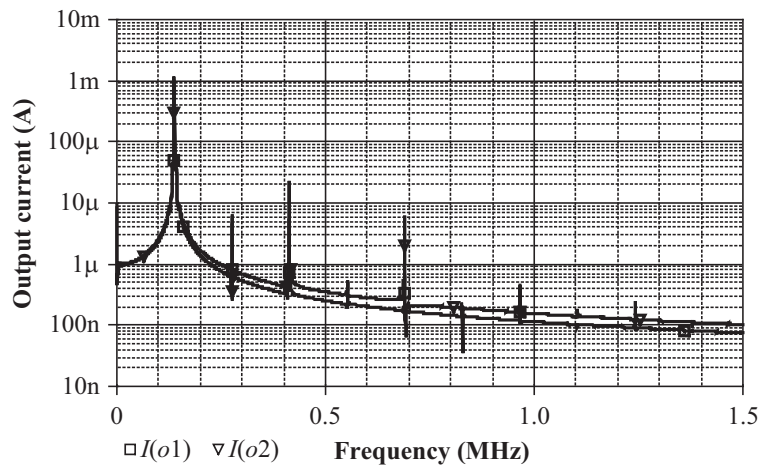
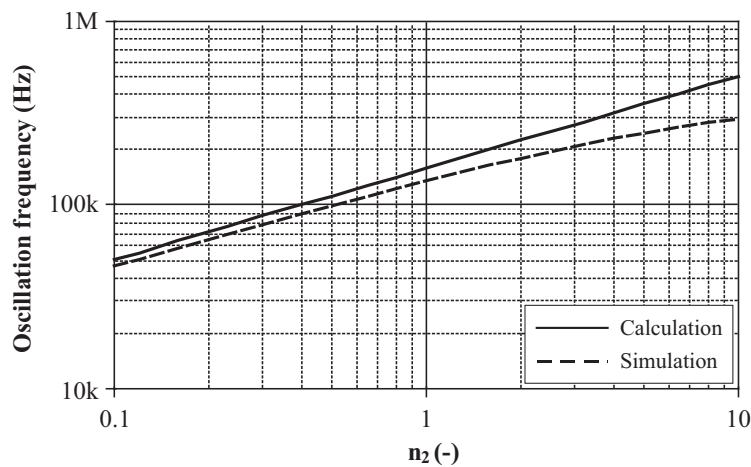Figure 9. Simulated frequency spectrums of outputs I_{o1} and I_{o2}

Figure 10. Variation of oscillation frequency with mu-factor n_2 of the ^{2}PCA

Simulation results agree quite well with the theoretical analysis. Note that the inconsistencies in output current amplitudes are due to the non-ideal effects of the PCA that is partially examined in the Section 4.

7. Conclusions

A new first of its kind programmable current amplifier (PCA) based current-mode (CM) electronically tunable quadrature oscillator (ETQO) has been reported. The advantages of PCA in oscillator design have been proved since the proposed circuit employs not only a bare minimum of four passive components, but also provides independent/non-interactive electronic control of the condition of oscillation (CO) and the frequency of oscillation (FO). The circuit also provides two explicit quadrature current outputs and simulation results have proved the workability of the circuit as a CM QO. As has been mentioned earlier that a PCA prototype using the implementation in (Koton, Vrba and Herencsar 2009) is ready to be manufactured by AMI Semiconductor Czech, Ltd., (part of the ON Semiconductor, Ltd.) in the CMOS 0.35 μm and labeled as COAK-NAA. It is expected that with the availability of such high performance PCA ICs there would be a surge of interest in the realizations of compact and tunable circuit solutions using PCA and thus PCA should prove to be a versatile active element of analog circuit design.

Acknowledgments

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