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NEW DOUBLE CURRENT CONTROLLED CFA (DCC–CFA) BASED VOLTAGE–MODE OSCILLATOR WITH INDEPENDENT ELECTRONIC CONTROL OF OSCILLATION CONDITION AND FREQUENCY

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In this paper, a new electronically tunable quadrature oscillator (ETQO) based on two modified versions of current feedback amplifiers (CFAs), the so called double current controlled CFA (DCC-CFAs) is presented. The frequency of oscillation (FO) of the proposed voltage-mode (VM) ETQO is electronically adjustable by current gain or by varying the intrinsic resistance of the X terminal of the active element used. The condition of oscillation (CO) is adjustable by current gain independently with respect to frequency of oscillation. Simultaneous control of current gain and intrinsic resistance allows linear control of FO and provides extension of frequency tuning range. In the proposed circuit all the capacitors are grounded. The use of only grounded capacitors makes the proposed circuit ideal for integrated circuit implementation. The presented active element realized by using BiCMOS technology and the behavior of proposed circuit are discussed in details. The theoretical results are verified by SPICE simulations based on CMOS ON-Semi C5 0.5 μm and bipolar ultra high frequency transistor arrays Intersil HFA 3096 process parameters.

K e y w o r d s: electronic control, sinusoidal oscillator, current feedback amplifier (CFA), double current controlled CFA (DCC-CFA)

1 INTRODUCTION

Recently, in [1], the fact of plenty available analog building blocks (ABBs) for analog signal processing is discussed. In [1], the review of basic, modified, and novel active elements is also given. In the last years in the literature interesting applications of components mentioned in [1], such as operational transconductance amplifier (OTA) [2, 3], current controlled second-generation current conveyor (CCCII) [4, 5], dual-X second-generation current conveyor (DXCCII) [6, 7], fully-differential second-generation current conveyer (FDCCII) [8, 9], current feedback amplifier (CFA) [10], current controlled CFA (CC-CFA) [11], modified current feedback operational amplifier (MCFOA) [12], current differencing buffered amplifier (CDBA) [13, 14], current differencing transconductance amplifier (CDTA) [15, 16], current follower transconductance amplifier (CFTA) [17, 18], programmable current amplifiers [19, 20], adjustable current followers [21], *etc*, have been published. Many of these components offer features for electronic control and great frequency response. Mentioned elements find wide range of applications in the field of sensor, control, measuring, automotive electronics, acoustic, and high-speed & radio-communication systems (amplifiers, filters, oscillators, modulators, active rectifiers, *etc*) [22–28].

The possibilities of electronic tunability of ABBs are today focused to several ways. As an example internal (intrinsic) resistance R_X control by bias current I_b [5, 11, 29], transconductance (g_m) control by bias current I_b [30–33], control by one passive element (grounded or floating resistor), so-called single resistance controlled oscillators (SRCO) [34–38] can be mentioned. Some suitable solutions allow replacement of the resistor by FET and easy voltage control of frequency of oscillation (FO) [35, 37]. Interesting approach based on synthetic negative capacitance control is published in [39]. Switching of working capacitors is also suitable method, if continuous control of FO is not necessary [40]. Another interesting way is the use of current gain control, which is not so common. Several solutions of electronically controllable current conveyors were proposed [41–45] and applied for electronic control of frequency of oscillation [44, 45].

Short review of several important previously presented solutions of oscillators with CFAs is summarized below. Siripruchyanun *et al* [11] proposed oscillator employing two CFAs and two capacitors with g_m and R_X control, but FO tunability has not been verified. Verification was provided by simulation at $f_0 = 820$ kHz with total harmonic distortion (THD) 2.7 %. In [34], Gupta *et al* presented SRCO solutions based on two CFA and five passive elements, where one of two capacitors is floating. Tunability range of the verified type was from 45 to 150 kHz with THD = 0.6%. In [35], Gupta *et al* presented similar types of SRCO oscillators employing two CFAs and 5–6 passive elements. Oscillation frequency is tunable by FET replacement of specific resistor approximately from 230 to 300 kHz with THD = 1.6 %. Lahiri *et al* proposed several

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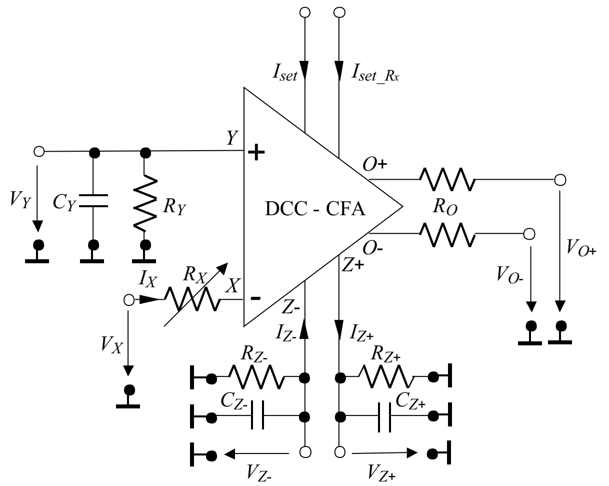

Fig. 1. Circuit symbol of the DCC-CFA including parasitic elements representing small-signal properties

Table 1. Dimensions of CMOS transistors in Figs. 3(a) and (b)

Transistor	W/L (μm)
M_1, M_2, M_{35}–M_{44}, M_{53}, M_{54}, M_{60}, M_{61}, M_{65}, M_{66}	20/1
M_3, M_4, M_7–M_{25}, M_{47}, M_{48}, M_{51}, M_{52}, M_{55}, M_{56}	40/1
M_5, M_6, M_{45}, M_{46}, M_{49}, M_{50}	10/1
M_{26}–M_{34}, M_{57}–M_{59}, M_{62}–M_{64}	5/1

Table 2. Dimensions of CMOS transistors in Fig. 3(c) (voltage buffer)

Transistor	W/L (μm)
M_1–M_3, M_8, M_9, M_{12}, M_{13}	5/1
M_4–M_7, M_{10}, M_{11}, M_{14}–M_{17}	10/1

variants of oscillators using two CFAs and 5 passive elements with grounded capacitors [36]. Furthermore, some of them have FO controllable by a single resistor. Verification was provided at $f_0 = 1.3$ MHz with THD $= 3$ %. Bhaskar *et al* [37] showed several conceptions of SRCO types employing two CFAs with 4–5 passive elements and 2 or 3 capacitors included. Tunability is based on FET replacement of resistor. The change of FO was verified approximately from 3 to 35 kHz with THD between 0.5–2 %. Herencsar *et al* [38] presented SRCO quadrature solutions based on two CFAs and 5 passive elements. Tunability was verified from 100 to 400 kHz with THD $= 0.7$ %. In [39], Lahiri *et al* introduced quadrature realizations using two CFAs and 4 passive elements, where one capacitor is replaced by negative synthetic equivalent for purposes of adjustability. In fact, the circuit contains three active and 6 passive elements. The THD values in both outputs are under 2 %. Soliman presented solutions based on 2–3 CFAs and 4–6 passive elements (all C grounded) [46]. The condition of oscillation (CO) of some of the proposed circuits is quite complicated and the independent control of the FO and CO is not allowed. Several realizations in [46] of them have quadrature outputs and the feasibility was verified on $f_0 = 159$ kHz. In [47], Singh *et al* introduced interesting solutions of SRCOs employing only one CFA and 5 passive elements. Tunability was verified from 5 to 35 kHz and 123–270 kHz with THD 1.4 % and 3 %, respectively. Senani *et al* published SRCO solution based on two CFAs and 5 passive elements with FO adjustability from 50 to 450 kHz [48]. In [49] it is introduced a SRCO, which is designed by Martinez *et al*, wherein two CFAs and 5 passive elements were used. The circuit is tunable form 40 to 400 kHz with THD < 1 %. The SRCO conception presented in [50] by Shen-Iuan *et al* contains two CFAs and 6 passive elements and it is designed for low frequencies — tunability from 10 Hz up to 550 Hz. Theoretical work [51] by Celma *et al* deals with comparison of CFA based oscillators versus classical Op-Amp solutions. The proposed SRCO solution [52] employing one CFA and 5 passive elements is tunable from 330 kHz to 1 MHz. Senani *et al* proposed several SRCO solutions published in [53], where 3 CFAs and 5 passive elements were used. Adjusting of FO is possible from 5 to 15 kHz. Some theoretical information about synthesis was given by Gupta *et al* in [54]. In [55], Abuelmaatti introduced novel oscillator solutions based on partial admittance autonomous networks with 1 and 2 CFAs and verified circuits employ 3 and 4 passive elements, respectively. Unfortunately, equations for FO and CO are quite complicated. Workability was proved at frequencies around 150 kHz. Gunes *et al* [56] presented interesting approach about several realizations of SRCO oscillators using one CFA and 5 passive elements with favorable and also with complicated design equations. Tunability was verified by R control from 5 kHz to 2 MHz by switching of the working capacitors. Very interesting solution of controllable oscillator was introduced by Bhaskar *et al* in [57], which is based on one CFA, 5 passive elements and two voltage multipliers for electronic control of FO. The tuning of FO by control voltage was verified from 15 to 73 kHz with THD in range 1.5–1.9 %. Summarization of suitable voltage controlled solutions is in [58], where two CFAs are used with two voltage multipliers and 5–6 passive elements. The tunability of the selected circuits was verified from 18 to 130 kHz and for the other variants approximately from 20 to 330 kHz. The THD is between 0.8–3.9 %. Senani *et al* [59] also proposed several conceptions employing one CFA and 6 passive elements (3C / 3R), however, tuning of FO is unfortunately not allowed (the circuit is tested at fixed FO of 11.4 kHz). In both cases the obtained THD is 0.6 and 1.2 %.

In general it can be seen that in some of introduced circuits (for example in [36, 37, 46]) the CO is very complicated, and overwhelming majority of presented solutions use R control (SRCO) for FO tuning. Hence, in case of floating resistor the adequate voltage control by FET or digital potentiometers can be difficult to implement. Therefore, to eliminate above mentioned disadvantages in the implementation of R control, we present a modified

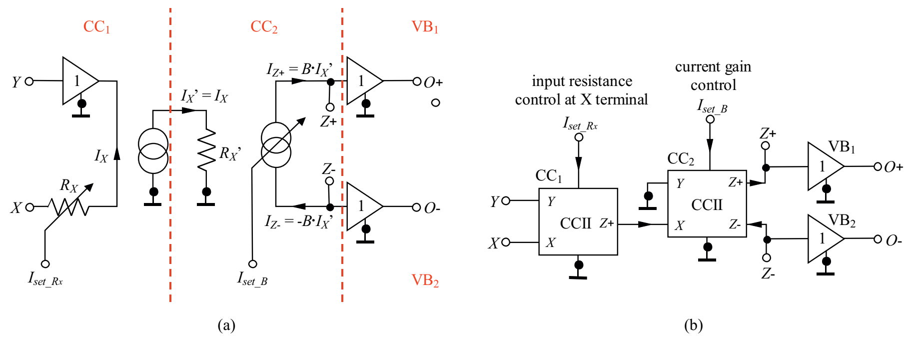

Fig. 2. (a) – behavioral model and, (b) – block diagram of DCC-CFA

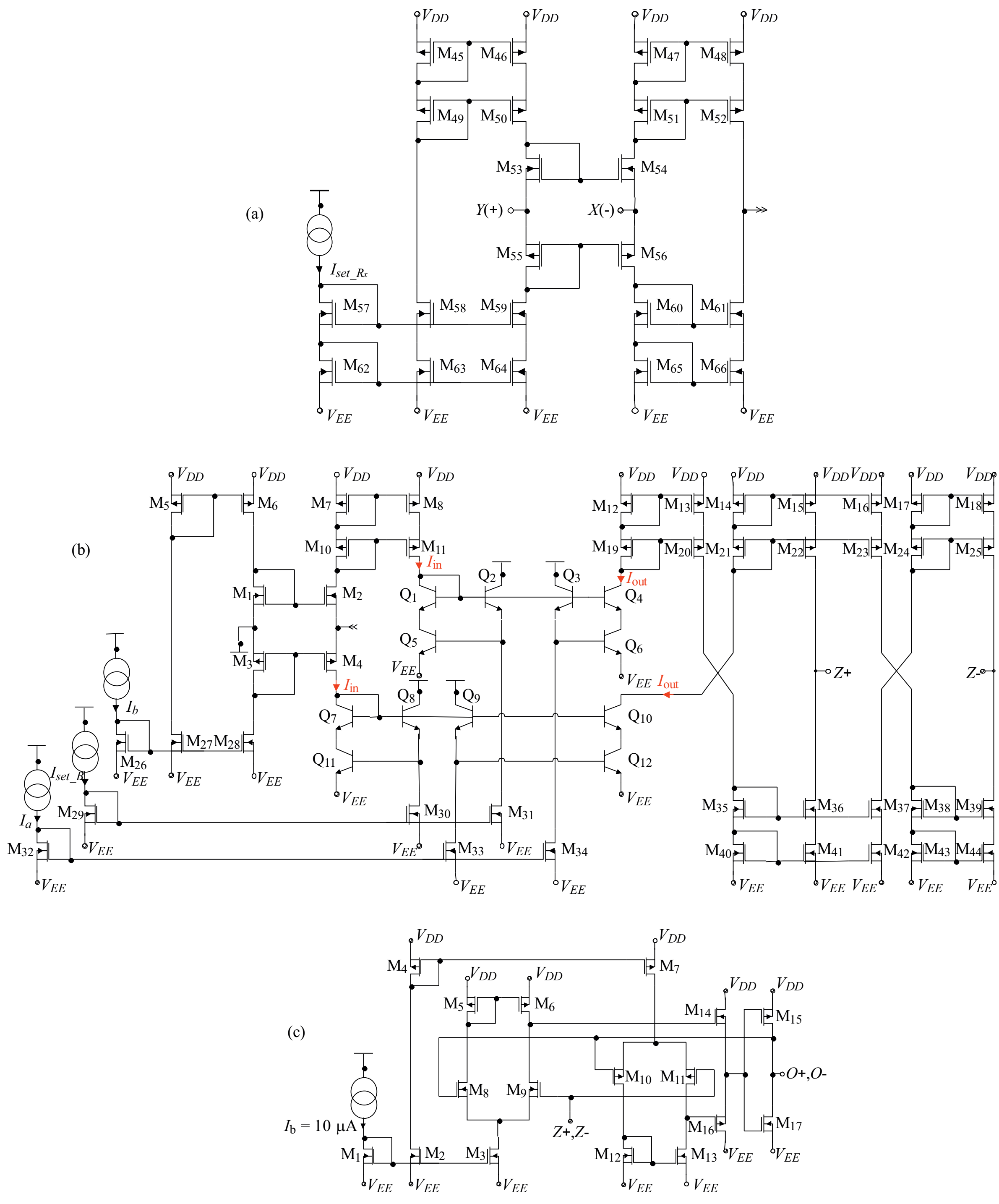

Fig. 3. Conception ofthee internal parts of DCC-CFA: (a) – CC$_1$, (b) – CC$_2$, (c) – VB

ABB of well known CFA so-called Double Current Controlled CFA (DCC-CFA) in this paper. Main advantages of presented active element is the current gain control between ports X and Z terminals and control of input (intrinsic) resistance (R_{X}) of current input X terminal without mutually affection that allow simple electronically tunable quadrature oscillator (ETQO) design with fine tuning. Moreover, this paper presents another point of view on frequency of oscillation and condition of oscillation adjusting than use oscillators in many hitherto published CFA based solutions. Hence, combination of two ways of control is possible. Here proposed ETQO provides electronic CO and except "classical" SRCO type of control allows also two another types of electronic control of FO via R_{X} ($I_{\mathrm{set\_RX}}$) and current gain B ($I_{\mathrm{set\_B}}$). In addition, simultaneous change of both parameters (their invariant ratio) allows linear FO tuning. Hence, independent control of the CO and FO without any matching conditions and without changes of any passive elements are very useful features. For wider range of FO tuning and for electronic CO control via current controlled current gain, a simple implementation of automatic amplitude gain control circuit (AGC) is presented.

2 DOUBLE CURRENT CONTROLLED CFA (DCC–CFA)

The CFA device [1, 10, 11] is equivalent circuit to second-generation current conveyor (CCII), which is followed by voltage buffer. Principle of electronic control of input (intrinsic) resistance (labeled as R_{X}) of current input terminal X is already well known [11, 29], where only one CCII and bias control (I_{b}) is sufficient [4, 29]. Implementation of current gain control is more complicated [42]. Previously, in the four terminal floating nullor (FTFN) [60–62] has been implemented such solution, which internal structure contains CCII with special type of controlled current mirror sections, two auxiliary DC currents (gain is given by ratio of these currents), and also standard bias current. Therefore, theoretically it is possible to change current gain and intrinsic resistance R_{X} in one modified CCII without their mutual affection. However, our further analyses demonstrate that R_{X} (I_{b}) bias current control causes increase of nonlinearity of input dynamical range and current gain interference ($R_{\mathrm{X}-I_{set\_RX}}$ control affects linearity of B control by I_{set_B}), if current gain (in this paper labeled by symbol B) is not unity. Therefore, in our design the section for adjusting R_{X} ($I_{\mathrm{set\_RX}}$) and gain control B ($I_{\mathrm{set\_B}}$) are separated and all interferences between R_{X} and B control are eliminated.

Circuit symbol of the DCC-CFA including parasitic elements representing small-signal properties is shown in Fig. 1, which can be described by the following hybrid matrix

$$\begin{bmatrix} I_{\mathrm{Y}} \\ V_{\mathrm{X}} \\ I_{\mathrm{Z}+} \\ I_{\mathrm{Z}-} \\ V_{\mathrm{O}+} \\ V_{\mathrm{O}-} \end{bmatrix} = \begin{bmatrix} Y_{\mathrm{Y}} & 0 & 0 & 0 & 0 & 0 \\ \alpha & R_{\mathrm{X}} & 0 & 0 & 0 & 0 \\ 0 & \beta_1 B & Y_{\mathrm{Z}+} & 0 & 0 & 0 \\ 0 & -\beta_2 B & 0 & Y_{\mathrm{Z}-} & 0 & 0 \\ 0 & 0 & \gamma_1 & 0 & Z_{\mathrm{O}+} & 0 \\ 0 & 0 & \gamma_2 & 0 & 0 & Z_{\mathrm{O}-} \end{bmatrix} \begin{bmatrix} V_{\mathrm{Y}} \\ I_{\mathrm{X}} \\ V_{\mathrm{Z}+} \\ V_{\mathrm{Z}-} \\ I_{\mathrm{O}+} \\ I_{\mathrm{O}-} \end{bmatrix}, \quad (1)$$

where $Y_{\mathrm{Y}} = sC_{\mathrm{Y}} + 1/R_{\mathrm{Y}}$, $Y_{\mathrm{Z+}} = sC_{\mathrm{Z+}} + 1/R_{\mathrm{Z+}}$, $Y_{\mathrm{Z-}} = sC_{\mathrm{Z-}} + 1/R_{\mathrm{Z-}}$, $Z_{\mathrm{O+}} = R_{\mathrm{O+}}$, $Z_{\mathrm{O-}} = R_{\mathrm{O-}}$ are parallel admittances and resistances at their relevant terminals. The α is voltage tracking error between Y and X terminals ($\alpha = 1 - \varepsilon$, $|\varepsilon| \ll 1$), $\beta_{1,2}$ are current tracking errors ($\beta = 1 - \delta$, $|\delta| \ll 1$) between X and $\mathrm{Z}_{1,2}$ terminals, and $\gamma_{1,2}$ are voltage tracking errors between Z and O terminals ($\gamma = 1 - \eta$, $|\eta| \ll 1$), respectively.

The behavioral model, the block diagram, and the proposed internal structure of DCC-CFA are shown in Figs. 2 and 3, respectively. The internal structure contains two current conveyors (CC_1 and CC_2) and two voltage buffers (VBs). In Fig. 3(a) the CC_1 section is classical CCII with cascoded mirrors [63–65]. The core of the CC_2, shown in Fig. 3(b), is based on modified CCII with additional mirrors for current gain control, which was introduced in [60–63]. In Fig. 3(c), the last part of DCC-CFA, the voltage buffer employing two classical transconductance sections is shown [65, 66]. The supply voltages of DCC-CFA were chosen as ±2.5 V.

The input intrinsic resistance of CC_1 stage is given by the following equation

$$R_{\mathrm{X}} \approx \frac{1}{g_{\mathrm{m\_M54}} + g_{\mathrm{m\_M56}}} = \frac{1}{\sqrt{2KP_N\left(\frac{W_{\mathrm{M54}}}{L_{\mathrm{M54}}}\right)I_{\mathrm{set\_RX}}} + \sqrt{2KP_P\left(\frac{W_{\mathrm{M56}}}{L_{\mathrm{M56}}}\right)I_{\mathrm{set\_RX}}}}, \quad (2)$$

where W and L are the channel width and length, respectively, and $KP = \mu_0 C_{\mathrm{OX}}$ (μ_0 is carrier mobility, C_{OX} is the gate-oxide capacitance per unit area) for the used CMOS ON-Semi C5 0.5 μm technology model (LOT: T22Y_TT, WAF: 3104) [67]. The dimensions of the MOS transistors used in the DCC-CFA implementation are listed in Tab. 1 (CC_1 and CC_2 stages) and Tab. 2 (voltage buffer), respectively.

The current gain control of DCC-CFA, which is implemented in the stage CC_2 shown in Fig. 3(b), is done by special current mirrors formed by group of transistors Q_1–Q_{12} for purposes of electronic adjustability. Moreover, in comparison with PNP types of transistors, NPN types have much better properties in particular technology, and therefore, design is done with NPN types only. In the simulations the commercially available models of ultra high frequency transistor arrays Intersil HFA 3096 were used [68]. Applying the translinear principle [20, 60, 61, 69–71] and assuming that all transistors are well matched with the common-emitter current gains

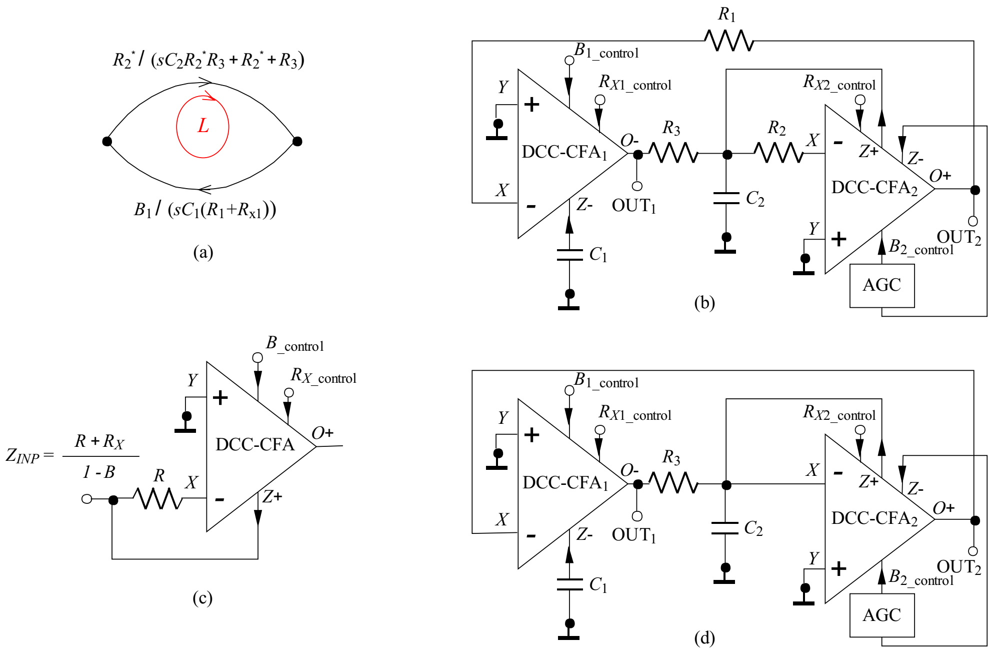

Fig. 4. (a)– signal flow graph of the proposed oscillator, (b) – the proposed oscillator, (c) – controllable negative resistance based on DCC-CFA, (d) – minimal realization of proposed oscillator using intrinsic resistances of current inputs

$h_{21e} \gg 1$, then the relationship of the collector currents can be characterized by the following equations

$$\prod_{n \in CW} I_n = \prod_{n \in CCW} I_n \,, \tag{3}$$

$$I_{C\_Q5} I_{C\_Q2} = I_{C\_Q3} I_{C\_Q6} \,, \tag{4}$$

$$I_{C\_Q11} I_{C\_Q8} = I_{C\_Q9} I_{C\_Q12} \,, \tag{5}$$

where CW means clockwise and CCW counter clockwise direction of calculation in translinear approach. In (4) and (5) I_{C_Q5} and I_{C_Q11} are input currents of the so-called "gain producing bipolar section" and I_{C_Q6} and I_{C_Q12} are the output currents. According to the previous statement, (4) and (5) can be express as

$$I_{in} I_{C\_Q2} = I_{C\_Q3} I_{out}, \ I_{in} I_{C\_Q8} = I_{C\_Q9} I_{out}. \tag{6}$$

Assuming that $I_{C\_Q2} = I_{C\_Q8} = I_{\mathrm{set\_B}}$ and $I_{C\_Q3} = I_{C\_Q9} = I_{\mathrm{a}}$, the current gain can be calculated as follows

$$B \approx \frac{I_{out}}{I_{in}} = \frac{I_{C\_Q8}}{I_{C\_Q9}} = \frac{I_{C\_Q2}}{I_{C\_Q3}} = \frac{I_{\mathrm{set\_B}}}{I_{\mathrm{b}}} \,. \tag{7}$$

From (7), it is evident that the current gain can be easily controlled by either $I_{\mathrm{set\_B}}$ and/or I_{a}.

3 OSCILLATOR DESIGN WITH TWO WAYS OF FO AND CO TUNING

The are many approaches leading to synthesis of oscillators in current technical literature[72]. As an example, loop and multi-loop integrator systems, autonomous circuit nodal analysis, or the state variable method can be mentioned [34, 37, 46]. Here proposed quadrature oscillator is based on very common approach employing integrators in the loop. The signal flow graph (SFG) of proposed oscillator and the corresponding circuit employing two DCC-CFAs and five passive elements is shown in Figs. 4(a) and (b), respectively. It is based on loop feedback system with one lossless voltage integrator (DCC-CFA$_1$, R_1, and C_1) and passive lossy voltage integrator (PLVI), which is formed by R_3 and C_2. In addition, the PLVI requires a negative resistance that is realized by another current amplifier included in frame of the DCC-CFA$_2$, as it is shown in Fig. 4(c).

The determinant of SFG [73] for the oscillator from Fig. 4(b) is shown in Fig. 4(a) and has a form

$$\Delta = 1 - \left[\frac{-B_1}{sC_1(R_1 - R_{\mathrm{X1}})} \, \frac{R_2^*}{sC_2 R_2^* R_3 + R_2^* + R_3} \right] = 0 \,, \tag{8}$$

where R_2^* represents a negative resistance given by

$$R_2^* = \frac{R_2 + R_{\mathrm{X2}}}{1 - B_2} \,. \tag{9}$$

By substituting (9) into (8) we obtain the characteristic equation (CE) of the circuit in the following simple form

$$\text{CE:} \quad s^2 + s \frac{R_3 + (R_2 + R_{\mathrm{X2}}) - R_3 B_2}{R_2 R_3 C_2} + \frac{B_1}{R_3 C_1 C_2 (R_1 + R_{\mathrm{X1}})} = 0 \,, \tag{10}$$

where the condition of oscillation and the frequency of oscillation are

$$\text{CO}: \quad B_2 \geq \frac{R_3 + (R_2 + R_{\mathrm{X2}})}{R_3} \,, \tag{11}$$

$$\text{FO}: \quad \omega_0 = \sqrt{\frac{B_1}{R_3 C_1 C_2 (R_1 + R_{\mathrm{X1}})}} \,. \tag{12}$$

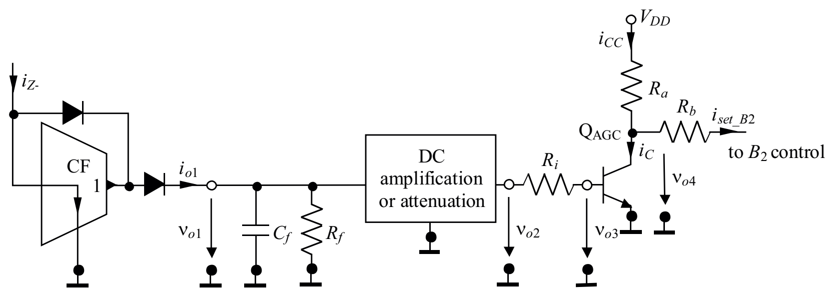

Fig. 5. Detailed AGC circuit

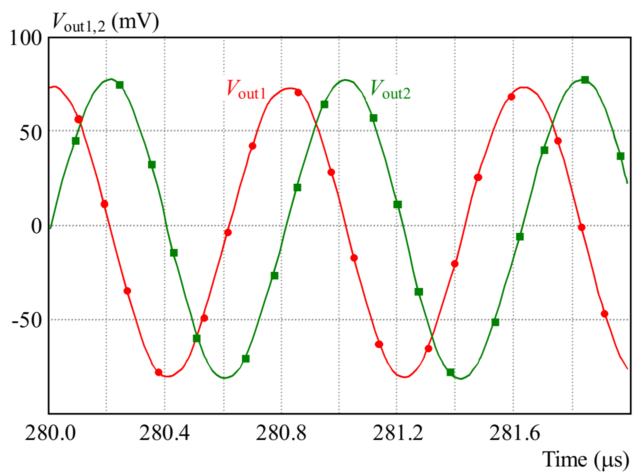

Fig. 6. Stable state output oscillations

From (11) and (12) it can be seen that in ideal case the CO and FO are independently adjustable by current gains of DCC-CFAs. Moreover, the FO can be adjustable by floating resistor R_1, therefore, this oscillator can also be added to the group of oscillators referred as SRCO types. Unfortunately, the replacement or adjusting of floating element can be complicated. Therefore, in our case we can also use intrinsic resistance control of DCC-CFA (R_X), which allows tuning by very simple way. Assuming that the current gain of DCC-CFA$_1$ is equal to 1 (fixed gain), the FO in (12) turns to

$$\text{FO}: \quad \omega_0 = \sqrt{\frac{1}{R_3 C_1 C_2 \left[R_1 + R_X(I_{\text{set\_RX}})\right]}}\,. \tag{13}$$

The equivalent minimal oscillator circuit is show in Fig. 4(d) in case of only internal intrinsic resistances $R_{X1}, R_{X2} (R_1 = R_2 = 0)$ and one external R_3 are considered.

The relative sensitivities of the frequency of oscillation (12) on circuit parameters can be calculated as

$$\begin{aligned} S_{R_1}^{\omega_0} &= -0.5\frac{R_1}{R_1 + R_{X1}}\,, \quad S_{R_{X1}}^{\omega_0} = -0.5\frac{R_{X1}}{R_1 + R_{X1}}\,, \\ S_{R_3,C_1,C_2}^{\omega_0} &= -S_{B_1}^{\omega_0} = -0.5\,, \quad S_{B_2,R_{X2},R_2}^{\omega_0} = 0\,. \end{aligned} \tag{14}$$

Equation (14) indicates that the ω_0 passive and active sensitivities are equal or not higher than 0.5 in absolute value, and hence, the circuit exhibits an attractive sensitivity performance.

4 SIMULATION RESULTS

In this Section, results concerning oscillator from Fig. 4(b), *ie* time domain, tuning properties and quality of produced signal are presented. Input resistances $R_{X1,2}$ were set to $760\,\Omega$ ($I_{\text{set_RX}} = 50\,\mu$A). External resistors have values $200\,\Omega$, therefore real values of ($R_1 + R_{X1}$) and ($R_2 + R_{X2}$) can be considered as $960\,\Omega$. The proposed oscillator is also able to operate without external resistors, as it is shown in Fig. 4(d). However, the analyses showed that the solution with external resistors of small values is better in THD. Therefore, the usage of external resistors R_1 and R_2 is matter of spectral quality of generated waveform. The current transfer $B_1 = 0.8$ is set by $I_{\text{set_B1}} = 40\,\mu$A. To fulfill the CO in (11) and to start the oscillation, the $B_2 \geq 2$ must be satisfied, which corresponds in the ideal case to $I_{\text{set_B2}} \geq 100\,\mu$A.

Since we expect wider range of electronic tuning of FO, stable output level, and lower THD, a precise automatic amplitude gain control circuit (AGC) implementation (Fig. 4(b)) is necessary. In low-voltage CMOS and BiCMOS technologies there is always problem with linearity and amplitude of output signal, which is quite small (tens - hundreds mV maximally). The implementation of AGC is quite easy, since the CO is controlled by B_2 parameter. Theoretically, the rectified output voltage is able to control B_2, but here proposed and similar types of oscillators provides output amplitude with several tens or hundreds of mV. Hence, before the rectification a sufficiently larger level amplification (several times) is necessary. Subsequently, the used amplifier before rectification must have great frequency features and also great slew rate. Such conception is better for higher output oscillation level, where pre-amplification (before rectifier) is not necessary. Therefore, we utilized the current-mode rectifier (based on current follower [74]) as it is shown in Fig. 5. We engaged unused (grounded in practice) current output Z- of DCC-CFA$_2$ in this case. The current follower in Fig. 5 was modeled by the diamond transistor OPA860 [75]. An advantage of this rectifier is in precise rectification suitable for low current amplitudes (tens of μA). In the simulations, the BAS70 [76] high speed diode with minimal threshold voltage and short recovery time was used, which provides low output amplitudes.

Figure 6 shows the output oscillation at both outputs for $B_1 = 0.8$ ($I_{\text{set_B1}} = 40\,\mu$A). Oscillator pro-

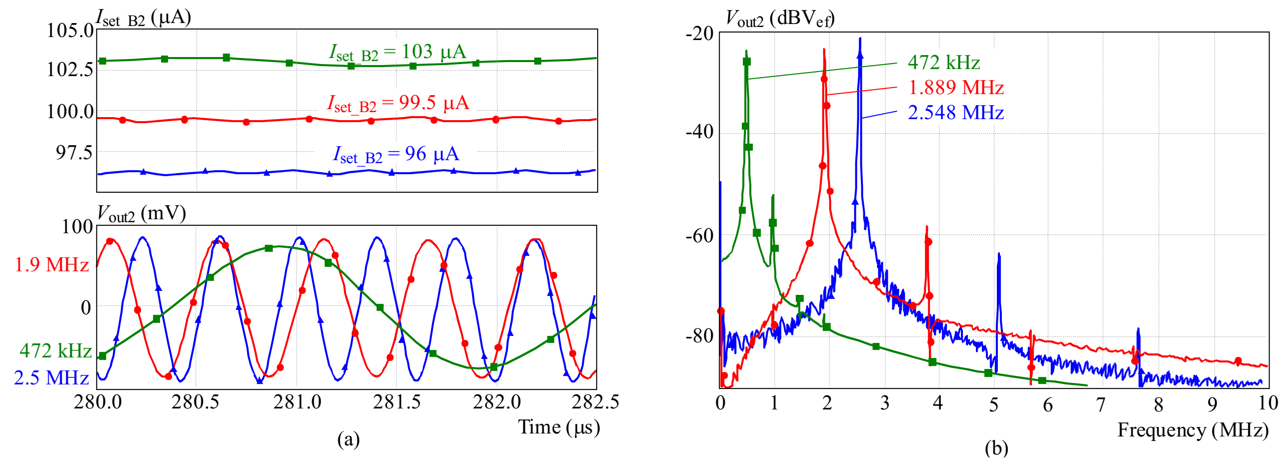

Fig. 7. (a) Tuning process observed on OUT_2 and corresponding AGC response, (b) spectrum of OUT_2 output oscillations

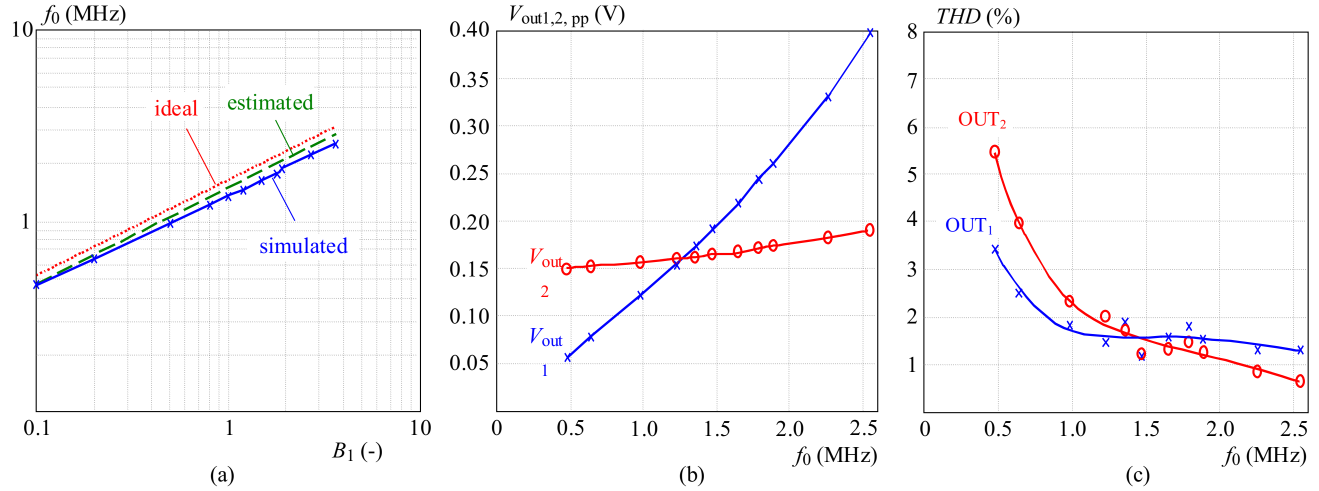

Fig. 8. Current gain control of oscillation frequency: (a) – dependence of FO on controlled current gain B_1, (b) – dependence of output amplitudes on frequency of oscillation, (c) – THD versus oscillation frequency

duces quadrature outputs (see SFG in Fig. 4(a) – relation between outputs)s in steady state operation. Simulations provided oscillation frequency 1.23 MHz. Theoretical value is 1.45 MHz. Considering the parasitic influences of the circuit, the obtained frequency of oscillation is 1.32 MHz for both outputs with amplitudes about 160 mV peak-to-peak.

4.1 FO control via current gain

The first way of control the FO of the oscillator from Fig. 4(b) is via current gain B_1 ($I_{\mathrm{set\_B1}}$). The output level of V_{OUT2} achieved values between 150-190 mV peak-to-peak, in the tuning range approximately from 470 kHz to 2.5 MHz for B_1 from 0.1 to 3.6. Figure 7 shows three output waveforms for different oscillation frequencies together with corresponding responses of AGC ($I_{\mathrm{set\_B2}}$) on change of output amplitude and spectrum of V_{OUT2}. The output V_{OUT1} changed voltage level rapidly from 56 to 400 mV, peak-to-peak (in the same frequency range of FO adjusting). The difference between the approximately determined, when small-signal parasitic influences C_{p} of DCC-CFA are considered and simulated FO is caused by non-ideal properties of active elements used and nonlinearity of B control. Figure 8 shows graphical representations of the obtained results *ie* dependence of FO on B_1, dependence of output level on FO, and THD versus FO, respectively. Obtained results are in good agreement with theory. Better results can be obtained by more careful CO and AGC settings. However, due to complex DCC-CFA transistor model, the optimization of oscillator parameters is time consuming.

4.2 FO control by intrinsic resistance

In case of the second possibility of FO control, the FO, given by (13), can be easily controlled by means of control current $I_{\mathrm{set\_RX1}}$ of $\mathrm{DCC\text{-}CFA}_1$. The dependences of FO on R_{X1} and dependences of output amplitudes on FO are shown in Fig. 9, respectively. Output level of V_{OUT2} is greater than in previous case of control, *ie* between 160–175 mV peak-to-peak. The output signal level of V_{OUT1} is changed in narrower range from 110 to 260 $\mathrm{mV_{P\text{-}P}}$. However, the tunability range is narrower than in previous case and it is approximately from

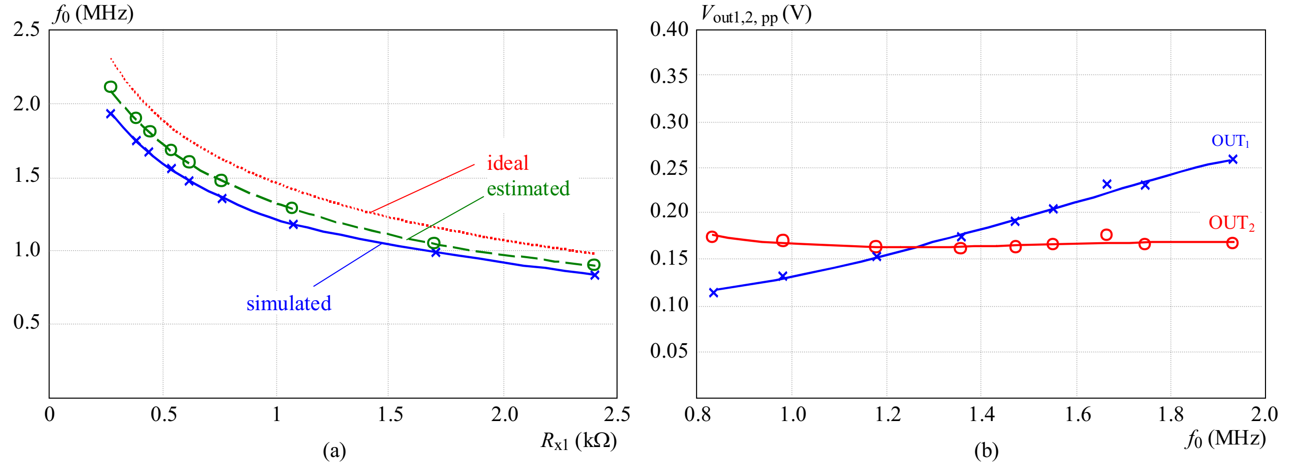

Fig. 9. Intrinsic resistance control of oscillation frequency: (a) – dependence of frequency of oscillation on intrinsic resistance, (b)– dependence of output amplitudes on frequency of oscillation

830 kHz to 1.93 MHz. The THD values are close to the THD of the first type of adjusting.

5 DISCUSSION

First type of electronic control of frequency of oscillation via B_1 is suitable for wider range of tuning. Results show that the frequency is tunable approximately from 470 kHz to 2.5 MHz. The AGC system defends oscillator against large fluctuations or outages of output amplitudes, however, in direction to higher frequencies it still brings small (40 mV) change of V_{OUT2} amplitude. The second variant of electronic control by intrinsic resistance R_{X} is suitable for shorter range of adjusting with better stability of the output level (V_{OUT2}) in the range of interest. In this case the variance of V_{OUT2} is only about 15 mV and, in comparison to B_1 control of FO, the range of oscillation frequency adjusting is approximately two times narrower, *ie* from 830 kHz to 1.93 MHz. This weakness is given by the constraining technological limits of R_{X} caused by smaller and restricted changes of R_{X} (range between saturation and linear region of transistors). The effect of R_{X} on frequency of oscillation is also determined by sensitivity of R_{X} on FO that is decreasing, if $R_1 \gg R_{\mathrm{X1}}$. In both cases the THD is comparable. If simultaneous adjusting of both parameters in Fig. 4(b) (B_1 and also R_{X}) is applied, it allows spreading of FO range from 283 kHz ($I_{\mathrm{set\_B1}} = 5\,\mu\mathrm{A}$, $I_{\mathrm{set\_RX}} = 5\,\mu\mathrm{A}$) to 3.48 MHz ($I_{\mathrm{set\_B1}} = 200\,\mu\mathrm{A}$, $I_{\mathrm{set\_RX}} = 400\,\mu\mathrm{A}$), and also linear control of FO (if B_1 and $(R_1 + R_{\mathrm{X1}})$ ratio keeps preserved).

Analyzing of (12) from Fig. 4(a) and assuming that $C_1 = C_2 = C$, the relation between V_{OUT2} and V_{OUT1} can be given as

$$V_{\mathrm{OUT1}} = V_{\mathrm{OUT2}} \sqrt{B_1 \frac{R_3}{(R_1 + R_{\mathrm{X1}})}} \,. \tag{15}$$

From (15) it is evident that the output voltage level at OUT_1 (V_{OUT1}) is dependent on current gain B_1, which affects the frequency of oscillation. Moreover, the V_{OUT1} is also dependent on $(R_1 + R_{\mathrm{X1}})$, which appears in DCC-CFA$_1$ based lossless integrator unit and both amplitudes (V_{OUT1} and V_{OUT2}) are unchangeable (or even equal) during the tuning process if invariant ratio of B_1 and $(R_1 + R_{\mathrm{X1}})$ is ensured.

6 CONCLUSION

In this paper, easy construction of oscillator with mutually independent control of frequency of oscillation and condition of oscillation was demonstrated. Used approach employs standard loop integrator system with controllable negative resistance based double current controlled CFAs. Advantage of introduced specific modification of CFA in signal processing and electronically tunable applications is clear. Proposed ABB allows current gain and internal (intrinsic) resistance control simultaneously. The proposed oscillator in minimal configuration employs three passive elements and it is tunable by current gain (B) or by internal resistance (R_{X}), both controlled by auxiliary DC control currents. The full realization of the oscillator with five passive elements belongs to the category of the so-called SRCO types, because FO control is also allowed by external floating resistor R_1. The tunability range was verified by simulations from 283 kHz to 3.48 MHz by both ways of the electronic FO control simultaneously. For stable output voltage levels and wider range of frequency tuning, a precise AGC system was designed. Amplitudes of output oscillations and total harmonic distortion are adequate to the used low voltage technology, low dynamic ranges of used active elements and quality of AGC circuit. Advantages of proposed solution are the following: sufficient range of tunability with possible extension by two types of FO control, possibility of linear control of FO, possibility of operation with equal amplitudes, grounded capacitors, easy final construction, and low sensitivities on passive components. Theoretical study was confirmed by detailed SPICE simulations.

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