ANALYTICAL SYNTHESIS AND IMPLEMENTATION OF LOW-FREQUENCY PRECISION ELECTRONICALLY CONTROLLED CFOA – BASED QUADRATURE OSCILLATOR

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Abstract. This paper describes the structure and operational principle of the proposed low-frequency digitally controlled quadrature oscillator, using monolithic Current-Feedback Operational Amplifiers (CFOAs). The created oscillator is based on an all-pass filter that is synthesized by using state variable approach and includes cascade structure of two digitally controlled integrators and two adders. The oscillator employs linear digital control of the oscillation frequency within $\pm 0.1\%$ accuracy by controlling the time constants of the two integrators through two 1024 – position digital potentiometers. The efficiency of the circuit is demonstrated by comparing the simulation results for the electrical parameters with experimental results of sample circuits, working up to 130 kHz with Total Harmonic Distortion $< 1\%$. The obtained results show good agreement with the theoretical hypotheses and analytical formulae.

Keywords: analog circuits, all-pass filter, current-feedback operational amplifiers, digital potentiometer, quadrature oscillator.

1. INTRODUCTION

The quadrature oscillators are termed oscillators that simultaneously generate two sinusoidal voltages with the same frequency which differ in phase with 90° . In both outputs of the quadrature oscillator a sinusoidal and co-sinusoidal signal is obtained. In general, suitable for this purpose is to use one integrating or one differentiating stage. The main drawback of this solution is that the output amplitude of the stages depends on the frequency. For the classical integrators the output signal amplitudes decrease with increasing the

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frequency (low-pass filter is obtained) and for the differentiators the amplitude increases proportionally to the frequency (high-pass filter is obtained).

The integrators, implemented with Switched Capacitors (SC), do not suffer of such disadvantage. Suitable for synthesizing of low-frequency quadrature generators are the electronic circuits of Flicker–Laker SC filters [1]. The Flicker–Laker SC filters are synthesized by using “follow-the-leader”-feedback (FLF) topology, proposed by Laker, Schaumann and Ghausi in 1979. The basic advantages of the active-SC filters are that they are very high precision with no tuning, realize some specific transfer functions, small-die area and low power dissipation at $f < 20$ kHz. The drawbacks of the SC filters are the sample-data effects (increased noise, reduced PSRR and aliasing), necessity of antialiasing reconstruction output filters and limited bandwidth (i.e., oversampling) [2].

The progress of the microelectronic technologies leads to continuous development of new and improved types of op amps, such as monolithic Current-Feedback Operational Amplifiers (CFOAs). Furthermore, the CFOAs can be represented as a cascade structure of positive second-generation current conveyors (CCII+) and additional output stage (voltage follower). In comparison with the conventional op amp the main advantage of CFOAs is wider dynamic range, higher linearity and the possibility to operate at higher frequencies. As a result, the CFOAs have found a broad application in the analysis and design of various analog and mixed-signal circuits and systems [3–9]. The analysis of papers in the last ten years revealed a wide variety of quadrature generators, employing different types of active building blocks, based on CFOAs [10–16]. Furthermore, it has been achieved an operating frequency up to several megahertz at Total Harmonic Distortion (THD) for some circuits not greater than 1%. However, the majority of the known circuits of CFOA – based quadrature generators [10–14] do not provide an analog or digital control of the frequency of the output signals.

In 2013, Biolek, Biolková, Kolka, and Bajer represent a quadrature generator using Z-copy Controlled-gain Current Differencing Buffered Amplifier (ZC-CG-CDBA) [15]. The proposed oscillator contains two ZC-CG-CDBAs, two grounded capacitors and three resistors. For the implementation of digital control 12-bit multiplying R-2R D/A converters are used. The frequency control of the output signals is performed in the range from 50 kHz to 100 kHz. The error is approximately 10% and the THD is not higher than 0.9%. In [16], Çiçekli and Gökçen proposed a circuit CMOS CCII – based quadrature oscillator with analog control of the oscillation frequency. The circuit contains two cascaded composite CCIIs. The first amplifier is composite CCII +

or CCCII+ and the second one is composite CCII- or CCCII-. The output signals are obtained on the grounded capacitors. To control the frequency of the output signals variable resistors are used, implemented with two parallel connected NMOS transistors, operated in a triode region.

The development of digitally programmable quadrature oscillators is essential for the application of analog circuits in digital systems. One method of creating electronically controlled quadrature oscillators with linear frequency adjustment is by using programmable Current Followers (CFs) and/or Voltage Followers (VFs) as active building blocks. In [17], CMOS digitally programmable quadrature oscillators based digitally CFs (DCCFs) and VFs are presented. The current gain of the proposed DCCF is implemented by using a Current Division Network (CDN) [18]. The CMOS digitally programmable oscillators include five followers and six passive elements. The frequency of oscillation can be tuned from 300 kHz to 1.5 MHz by a digital word from 0 up to 55 (in decimal form). The THD of the output signals were found to be better than -60 dB. In [19], electronically controlled oscillator using Dual-output Controlled Gain Current Buffered Amplifiers (DO-CG-CFBAs) is reported. The electronic circuit includes three active elements, three resistors and two grounded capacitors. The current gain of the DO-CG-CFBAs is controlled by external DC voltage or current. The oscillation frequency adjustment range is from 50 kHz to 6.76 MHz. The THD is between 0.4 and 0.9%. Recently, in [9], a circuit CMOS voltage controlled quadrature oscillator is reported. The circuit contains one MVDVTA and three grounded passive components. The proposed MVDVTA is composed of one OTA and one DVCC. The oscillation frequency adjustment range is from 1.7 MHz up to 1.9 MHz at $100 \text{ mV} \leq V_c \leq 220 \text{ mV}$ (where V_c is the control voltage of the proposed oscillator) with THD greater than 1%.

In this paper, a high precision electronically controlled CFOA-based all-pass filter and quadrature oscillator are developed. The digitally programmable filter is generated using well-known state variable approach and the CFOA-based universal active filter, reported in [20]. For the electronic circuit of the created oscillator, obtained by the all-pass filter, the frequency control is implemented by using two CFOA-based non-inverting programmable integrators with dual CMOS digital potentiometers (or digital pots).

2. ACTIVE BUILDING BLOCKS – MONOLITHIC CFOAs

The most common CFOAs are equivalent to a CCII+ plus an output voltage buffer and are known as three-terminal op amp. These op amps have a

Fig. 1. A simple linear model for the most common CFOA

high impedance non-inverting y input, a low-impedance inverting x input and voltage output o . In some of the CFOAs, the z terminal between the first stage (CCII+) and the second stage (voltage buffer) is defined as an external pin, Fig. 1. The port o is the output of the voltage buffer, where the resistance is very low (magnitude of several ohms). As a result, four-terminal op amps are obtained, known as 4-terminal CFOAs. For an ideal CFOA the parameters α_{yx} , α_{xz} and α_{zo} are equal to 1. Also, the parasitic series resistance r_x is equal to zero ohms and output resistance r_z is approximately equal to infinity.

The ideal relation between input and output voltages and currents of an ideal CFOA can be given by the following hybrid matrix:

$$\begin{bmatrix} i_y \\ u_x \\ i_z \\ u_o \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} u_y \\ i_x \\ u_z \end{bmatrix}. \quad (1)$$

For example, the commercially available integrated circuits AD844, OPA615, OPA860 and OPA681 can be used as four-terminal CFOAs. Moreover, in some of the op amps, such as OPA615 and OPA860 the CCII+ is not internally connected to the output buffer, which allows those op amps to be used directly for the implementation of current conveyors.

3. STRUCTURE AND BASIC PARAMETERS OF THE DEVELOPED CFOA – BASED PROGRAMMABLE ALL-PASS FILTER

Based on the electronic circuit of Flicker–Laker SC filter universal active CFOA-based filter [20], shown in Fig. 2, providing realization of all-pass functions and possibility for independent digital control of the pole frequency

Fig. 2. Digitally controlled CFOA-based all-pass filter [20]

is obtained. As can be seen, the capacitors $C_1 = C_2 = C$ and the wipers of the RDACs are connected to the ground, which can lead to significantly reduce the noise level, and this – to ensure ease tuning the time constants of the integrators.

Considering the CFOAs to be characterized by the hybrid matrix, given in matrix (1) and the values of the variable resistances of the digital potentiometers, presented with formula (2) for the ideal transfer function of the all-pass filter, at $R_2 = R_4$ and $R_{IN} = 2R_{BP}$ ($R_{IN} = 2R$ and $R_{BP} = R$) it is obtained [20]:

1) the quality factor (or Q-factor)

$$Q_p = R_4 / \sqrt{R_1 R_3} \quad (2)$$

and

2) the pole frequency (self-oscillating frequency or undamped natural frequency)

$$\omega_p = \frac{1}{\sqrt{R_{WB1} R_{WB2} C_1 C_2}} \sqrt{\frac{R_3}{R_1}}. \quad (3)$$

As can be seen, from formulae (2) and (3), the pole frequency can be varied by the value of the control word D_{10} of the digital potentiometers, without

changing the Q-factor. The Q-factor can be modified independently through the resistor R_4 .

4. STRUCTURE AND OSCILLATION CONDITIONS OF THE SYNTHESIZED OSCILLATOR

The proposed circuit of low-frequency quadrature oscillator is shown in Fig. 3. This electronic circuit is based on the digitally controlled CFOA-based all-pass filter, Fig. 2, presented briefly in the previous section. For this purpose, in the circuit in Fig. 2, the output signal v_{AP} is applied to the input port in . Thus, it is obtained an electronic circuit with positive feedback network.

Fig. 3. Electrical circuit of the synthesized digitally controlled precision CFOA-based quadrature oscillator

The output signals v_{o1} and v_{o2} are obtained at the output ports of the two programmable integrators. The wiper positions of the digital potentiometers are controlled dynamically by microcontroller through SPI or I²C interface.

4.1. Theoretical analysis of the circuit for ideal active components

By routine analysis of the oscillator (at $R_5 = 2R$, $R_{IN} = 2R$ and $R_{BP} = R$) for the ideal characteristic equation in steady-state oscillation (i.e. oscillations that continue without change of the amplitude and frequency of the output signals) is found:

$$s^2 + \left(\frac{R_3}{R_4} - 2\frac{R_3}{R_2} \right) \frac{1}{R_{WB1}C_1} s + \frac{R_3}{R_1} \frac{1}{R_{WB1}R_{WB2}C_1C_2} = 0. \quad (4)$$

Based on the equation (5) for the oscillation condition it is obtained:

$$\frac{R_3}{R_4} = 2\frac{R_3}{R_2} \quad \text{or} \quad R_2 = 2R_4. \quad (5)$$

The oscillation frequency can be found by formula (3), i.e. $\omega_0 = \omega_p$.

Thus, R_4 appears only in the expression for the oscillation condition and R_{WB1} (R_{WB2}) appears only in the formula for the self-oscillating frequency. Therefore, each one of them can be independently controlled through the grounded resistor R_4 and the grounded variable resistors R_{WB1} (R_{WB2}), respectively.

The oscillator is started by tuning the resistor R_4 . The limit cycle stability is guaranteed by the nonlinear internal mechanism of the used CFOAs. Furthermore, to obtain stable oscillations the resistors R_{WB1} and R_{WB2} have to be with equal values for each frequency in operating range, i.e., the digital potentiometer has to work as a dual potentiometer.

In general, according to formulae (1) and (3) the minimal and maximal values of the ideal oscillation frequency are given by:

$$\omega_{0,min} = \frac{1}{\sqrt{(R_{AB1} + R_{W1})(R_{AB2} + R_{W2})C_1C_2}} \sqrt{\frac{R_3}{R_1}} \quad (6)$$

and

$$\omega_{0,max} = \frac{1}{\sqrt{R_{W1}R_{W2}C_1C_2}} \sqrt{\frac{R_3}{R_1}}. \quad (7)$$

These values of the frequency can be obtained by variation of the control word D_{10} (specified in decimal form) from 0 to $2^n - 1$.

4.2. Theoretical analysis of the circuit for non-ideal (real) active components

Also, the proposed oscillator is analyzed for the following non-idealities of the used active elements:

– The non-ideal linear model for CFOA (Fig. 1) contains three parameters to be determined by measurements, i.e. α_{yx} , the y to x voltage gain α_{xz} – the x to z current gain and α_{zo} – the z to o voltage gain. As a result for a non-ideal model the general relation between input and output voltages and the currents can be given as $i_y = 0$, $v_x = \alpha_{yx}v_y + i_x r_x$, $i_z = \alpha_{xz}i_x$, $v_z = i_z r_z$ and $u_o = \alpha_{zo}u_z$.

Fig. 4. A simple non-ideal linear model for monolithic digital potentiometers

The parasitic capacitances and inductances to the terminals of the monolithic CFOAs in comparison with the parasitic capacitances of the digital potentiometers are with much smaller values. Therefore, for operating oscillation frequencies up to 10 MHz the parasitic effects of the CFOAs can be neglected.

– Using the structures for the digital potentiometers [21] a simplified equivalent electrical circuit with basic elements is given in Fig. 4. The parasitic capacitances C_A , C_B and C_W , at the three terminals, can vary under change of the position of the wiper (middle point of the potentiometer), as their values can vary from several pF up to several tenths of pF . Moreover, the parasitic large capacitances value limits the frequency bandwidth. In general, the values of parasitic capacitances are function on the microelectronic technology, the structure of the electrical circuit, the value of the nominal resistance R_{AB} and the number of the steps (or resolution) of the wiper.

In steady-state oscillation for the non-ideal characteristic equation (using the condition $r_x \ll r_z$) the following approximate expression is found:

$$s^2 + \left(\frac{R_3}{R_4} + \alpha_{xz2}\alpha_{zo2} \frac{R_3}{R_1} \frac{C_{B1}R_{WB1} + C_{B2}R_{WB2}}{(R_{WB1} + r_{x1})C_2} - 2\frac{R_3}{R_2} \right) \frac{\alpha_{xz1}\alpha_{zo1}}{(R_{WB1} + r_{x1})C_1} s + \frac{R_3}{R_1} \frac{\alpha_{xz1}\alpha_{zo1}\alpha_{xz2}\alpha_{zo2}}{(R_{WB1} + r_{x1})(R_{WB2} + r_{x2})C_1C_2} \approx 0. \quad (8)$$

Based on equation (8) the oscillation condition and oscillation frequency, given in (3) and (5), are modified to:

$$\omega_0 = \sqrt{\frac{\alpha_{xz1}\alpha_{zo1}\alpha_{xz2}\alpha_{zo2}}{(R_{WB1} + r_{x1})(R_{WB2} + r_{x2})C_1C_2}} \sqrt{\frac{R_3}{R_1}} \quad (9)$$

and

$$\frac{R_3}{R_4} + \alpha_{xz2}\alpha_{zo2} \frac{R_3}{R_1} \frac{C_{B1}R_{WB1} + C_{B2}R_{WB2}}{(R_{WB1} + r_{x1})C_2} = 2\frac{R_3}{R_2}. \quad (10)$$

For independent control of the amplitude of the output signal it is required $R_{WB1} = R_{WB2}$ or to use dual monolithic digital potentiometer and $R_{WB1} \gg r_{x1}$. In this case, the formula for the oscillation condition is simplified and yield change simultaneously the resistances:

$$\frac{R_3}{R_4} + \alpha_{xz2}\alpha_{zo2} \frac{R_3}{R_1} \frac{C_{B1} + C_{B2}}{C_2} = 2\frac{R_3}{R_2}. \quad (11)$$

For the majority of the commercially available CFOAs the parameters α_{yx} , α_{xz} and α_{zo} are approximately equal to unity (the deviation of those parameters from unity is less than 0.1% [22]). As a result, they will have very small influence over the oscillation frequency and oscillation condition.

5. SIMULATION AND EXPERIMENTAL TESTING

To illustrate the workability of the proposed programmable quadrature oscillator (Fig. 3) the calculated and simulation results are compared with experimental testing of sample electronic circuit. For this purpose, an electronic circuit of oscillator is designed, providing operation within frequency range from several hundred hertz up to approximately 130 kHz. Moreover, the passive elements in the circuit diagram are the following values: $R_1 = R_2 = R_3 = 1.5 \text{ k}\Omega$, $R_5 = R_{IN} = 2 \text{ k}\Omega$, $R_{BP} = 1 \text{ k}\Omega$ and $C_1 = C_2 = 10 \text{ nF}$. For the implementation of the circuit a dual potentiometer AD5235 with nominal resistance $R_{AB} = 25 \text{ k}\Omega$ and 1024 wiper steps is used. The AD5235 is biased with

Fig. 5. Output waveforms ($v_{01} - \text{CH1}$ and $v_{02} - \text{CH2}$) of the oscillator at $D_1 = D_2 = 832$

Fig. 6. Output waveforms ($v_{01} - \text{CH1}$ and $v_{02} - \text{CH2}$) of the oscillator at $D_1 = D_2 = 512$

Fig. 7. Output waveforms ($v_{01} - \text{CH1}$ and $v_{02} - \text{CH2}$) of the oscillator at $D_1 = D_2 = 0$

± 2.5 V. For stable operation it is required to change simultaneously the resistances R_{WB1} and R_{WB2} , i.e. the chosen digital potentiometer has to work as a twin potentiometer. Also, the oscillator is designed using CFOA AD844AN, biased with ± 5 V supplies. The computer simulation was performed within OrCAD PSpice A/D, using AD844AN/AD standard SPICE macro-model [23] and AD5235 a SPICE macro-model, presented in [24].

The output waveforms and the measured oscillation frequency are shown in Figs 5, 6 and 7 at calculated frequency 729 Hz (at $D_1 = D_2 = 832$), 1180 Hz (at $D_1 = D_2 = 512$) and 135 kHz (at $D_1 = D_2 = 0$), respectively. The measured oscillation frequencies were found to be 711 Hz, 1149 Hz and 128.387 kHz, respectively. The oscillator spectrum at $D_1 = D_2 = 512$ is shown in Fig. 8.

Fig. 8. Oscillator spectrum at $D_1 = D_2 = 512$ ($V_{SS}^+ = -V_{SS}^- = 5V$)

Fig. 9. Output signals in a XY – format ($v_{01} - \text{CH1}$ and $v_{02} - \text{CH2}$) of the oscillator at $D_1 = D_2 = 512$

The THD of the output signals were found to be less than 1%.

In many practical cases, such as various types of synchronous demodulators, the built-in oscillator needs to produce two quadrature control voltages. Furthermore, if the frequency of the oscillator changes within certain limits, the synchronous detectors can be used to build spectral analyzers. For example, the spectral characteristics of a number of low-frequency biomedical signals, such as electrocardiographic, electromyographic and electroencephalographic signals are very important from a scientific point of view. To obtain accurate signal format information, the phase shift between the output signals from the oscillator has to be exactly equal to 90° .

At $D_1 = D_2 = 512$, it has been tested whether a phase shift between the two output signals has been established 90° . For this purpose, the output signals of the oscillator are connected to the inputs X and Y of the oscilloscope, Fig. 9. As can be seen, on the oscilloscope a circle pattern is displayed. The parasitic phase shift at the center position of the wipers for the digital potentiometers is less than 0.1° . Mainly, the value of the error is due to the influence of parasitic capacitances C_A , C_B and C_W to terminals of the potentiometers. Moreover, the capacitances can vary under change of the position of the wiper, as their values can vary from several pF up to several tenths of pF . The values of parasitic capacitances are function on the microelectronic technology, the structure of the electrical circuit and the number of the steps of the wiper. In the manufacture datasheets for the majority of the monolithic potentiometers the values of the parasitic capacitances are usually defined only at the center position (i.e. code = half-scale) of the wiper. At low frequencies (up to 1 MHz), for analysis and design of programmable circuits with digital potentiometers, the values of the parasitic capacitances at the half-scale of the wiper can be used. Actually, from a practical point of view, it is better to use the largest values of the parasitic capacitances, which can most accurately ensure the specified bandwidth and minimal value of the parasitic phase shift for all combinations of the binary code.

The deviation between the measured and simulated frequencies versus D_{10} in the operational range of oscillation (Fig. 10) is approximately equal to 5%, which confirms the workability of the proposed circuit. In the operating frequency range, the amplitude of the output signals $V_{o,m}$ is approximately equal to 2 V (peak to peak) at $R_L = 2 \text{ k}\Omega$. Moreover, the change in the amplitude is within 20 mV. The DC voltage of terminal z of each of the CFOAs for the entire frequency range does not vary more than $\pm 10 \text{ mV}$ and the output offset is less than 1 mV.

Fig. 10. Oscillation frequency versus digital code D_{10} using dual digital potentiometer AD5235 with resistance 25 k Ω

To compare the proposed programmable oscillator with some other circuits reported in the literature parameter Figure of Merit (FoM) is defined [10]:

$$FoM = \frac{\text{operation frequency}}{\text{power dissipation} \times \text{number of active devices}} \left[\frac{\text{Hz}}{\text{W}} \right]. \quad (12)$$

For the created oscillator it is found approximately 540 kHz/W at ± 5 V supplies. This value is less than the values of the digital controllable generators presented in [9, 15, 16, 17, 19], but with the exception of [17], for the proposed electronic circuit the power dissipation has lower value, due to the smaller number of active elements. Increasing the FoM for the proposed circuit can be achieved by using digital potentiometers with a higher cutoff frequency. For example, using the IC AD5143 (IC from Analog Dev.) the FoM can reach value equal to few megahertz per watt. However, the number of wiper steps and hence the number of oscillation frequencies is reduced to 256. An additional increase in the number of steps without reducing the oscillation frequency can be achieved by using some of the techniques proposed in [25]. As it is given in [25], the number of adjustable steps can be increased from 1024 up to 4096 by using a cascade structure with four digital potentiometers and one additional 4×1 output analog multiplexer in the integrator stages.

Table 1. Technical specification summary for the created oscillator

Oscillation frequency f_0	Programmable from 500 Hz to 130 kHz
Relative error between measured and simulated frequencies	$\pm 5\%$ at 25°C
Resolution	$\approx 0.1\%$ when using 10 Bit CMOS digital potentiometers
Maximum output voltage	2 V (peak to peak) at $R_L = 2\text{ k}\Omega$
Output offset voltage U_{oo} (25°C)	$\pm 1\text{ mV}$ at operating range for $D_{10} = 0$ to 1023
Input resistance r_{iA}	up to $25\text{ k}\Omega$
Output resistance r_{oA}	$< 1\ \Omega$ at 1 V amplitude
Total Harmonic Distortion – THD	$< 1\%$ at 130 kHz
Parasitic phase shift	$< 0.1^\circ$ at the center position of the wipers
Figure of Merit – FoM	540 kHz/W at $\pm 5\text{ V}$ supplies
Power dissipation	$\approx 40\text{ mW}$
Power supply voltages $V_{CC} = -V_{EE}$	$5\text{ V} \pm 0.5\%$

The presented hand calculations and the simulation results are on the basis of creating a summary of the technical specification for the proposed circuit, Table 1.

It was found by comparing the hand calculations and the simulation results with experimental results that the maximum relative error is lower than 6%. The achieved results confirm the accuracy of the above theoretical analysis and guarantee the efficiency of the proposed electronic circuits.

6. CONCLUSION

In this paper a precision digitally controlled CFOA-based quadrature oscillator with small number of active elements is proposed. The created oscillator is with good stability and linearity of the output signals by changing the electrical parameters in a broad frequency range. To illustrate the workability of the proposed low-frequency oscillator experimental results are achieved by using commercially available monolithic CFOA AD844AN and one dual CMOS potentiometer AD5235 with 1024 wiper steps (10 Bit). The frequency of oscillation is obtained from several tens of hertz up to 130 kHz. Moreover, the tuning resolution depends on the number of wiper steps of the chosen digital potentiometers. For the experimental testing with 10 Bit CMOS digital po-

tentiometers, the oscillation frequency of the oscillator can be adjusted with accuracy of 0.1%. The parasitic phase shift at 130 kHz is less than 0.1° and the THD of the output signals were found to be less than 1%.

The created electronic circuit can be useful for design of various digitally controlled analog electronic systems, such as high-resolution programmable filters in dynamometers for measuring forces, vector generators and selective voltmeters.

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