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Adjustable Multiphase Sinusoidal Oscillator with Fractional–Order Elements

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Abstract—This paper presents a fractional–order multiphase oscillator with adjustable FO (Frequency of Oscillation). The oscillator utilizes the following components: CCTA (Current Conveyor Transconductance Amplifier), OTA (Operational Transconductance Amplifier), auxiliary OTRA (Operational Transresistance Amplifier), two FOC (Fractional–Order Capacitors) and auxiliary resistors. The proposed structure provides three voltage outputs and their inverted variants for three selected values of $\alpha = 0.3, 0.5$, and 0.8 of FOC. The FO is tuned by changing the values of the transconductances of the OTA and CCTA. The features of the presented circuit were verified by PSpice simulations with behavioural models of the active elements. Analysis of the dependence of the values of the transconductances on values of α and frequency is provided. A THD (Total Harmonic Distortion) and parasitic analysis of the circuit was also made.

Index Terms—fractional–order oscillator, fractional–order capacitor, THD, multiphase oscillator

I. INTRODUCTION

The fractional–order systems become popular topic of many scientific groups from many research sectors. These systems bring greater variability in the design of functional systems and in setting their parameters in comparison with conventional integer–order systems [1]. The fractional–order systems can be used in many diverse scientific fields. For example in cryptography for generating random numbers [2], in biology for modelling of biological phenomena [3], in agriculture for analysis of the fruits or vegetable properties [4], in electrical engineering for design of the precise controllers [5], frequency filters [6] and oscillators [7]–[11].

Fractional–order oscillators are characterized by arbitrary phase shift of output signals in multiples of $\alpha\pi/2$, where the parameter α is non–zero positive number in range from $0 < \alpha < 1$ [8]. Note that standard oscillators provide phase shift of output signals only in multiples of $\pi/2$. The design of the fractional–order oscillators is based on usage of the fractional–order passive element (FOE) in the circuit structure. Most often fractional–order capacitor (FOC) is used, where the slope of impedance attenuation of the FOC equals to

20α dB/decade and phase shift is $-\alpha\pi/2$ [15]. The impedance of the FOC is described by the following equation:

$$Z_{C\alpha} = \frac{1}{Y_{C\alpha}} = \frac{1}{s^\alpha C_\alpha} = \frac{1}{(j\omega)^\alpha C_\alpha}, \quad (1)$$

where the C_α is called as pseudocapacitance with unit $F/\text{sec}^{1-\alpha}$ and s is Laplacian operator [1], [7]. The value of the pseudocapacitance is calculated by $C_\alpha = C_{\text{eq}}\omega^{1-\alpha}$, where C_{eq} is value of the conventional capacitor and ω is angular frequency. The j^α can be substituted as follows:

$$j^\alpha = \cos(\alpha\pi/2) + j\sin(\alpha\pi/2). \quad (2)$$

The paper [8] presents fractional–order Colpitts oscillator based on CMOS transconductors providing two voltage outputs. The simple proposal of the fractional–order oscillator with two voltage outputs based on differential voltage current conveyor transconductance amplifier (DVCCTA) and variable gain amplifier (VGA) employing impedance converter is presented in paper [9]. The main disadvantage of the design of the circuit containing the FOEs is commercial unavailability of these passive elements. Only special prototypes are fabricated [11]. These prototypes were used in [10] and [11] to design the fractional–order oscillators. The fractional–order oscillator of orders 2.6 and 2.7 with one voltage output designed with operational amplifier AD797 is presented and verified by measurements in [10]. In paper [11], the fractional–order Hartley oscillator designed by operational amplifier AD817 is analysed and verified also by measurements.

The proposal of the fractional–order oscillator with adjustable FO (Frequency of Oscillation) is presented and analysed in this paper. The proposed oscillator provides three voltage outputs with phase shift equal to multiple of $\alpha\pi/2$ and electronically controllable FO. The analysis of the values of transconductances versus values of α and FO is also presented in this paper. The advantage of our solution is minimal number of the passive parts and active elements, control of the FO without affecting the condition of oscillation and total of six voltage outputs. Also advanced parasitic analysis of the active elements is provided, which was not performed in any of the before mentioned papers. Any before mentioned paper does not provide deep parasitic analysis of the proposed fractional–order oscillator. The FO of the presented circuit is tuned

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by changing values of the transconductances of the CCTA (Current Conveyor Transconductance Amplifier) and OTA (Operational Transconductance Amplifier). The fractional-order oscillator was proposed by the FOCs used instead both of conventional capacitors in the structure. The function of the proposed oscillator was verified using PSpice simulations with non-ideal models of the used active elements.

II. PROPOSAL OF FRACTIONAL-ORDER OSCILLATOR

The scheme of the proposed fractional-order oscillator providing three voltage outputs v_1 , v_2 , v_3 and their inverted variants v_1' , v_2' , v_3' is depicted in Fig. 1. The circuit has been designed with three active elements: CCTA [12], OTA [13], and auxiliary OTRA (Operational Transresistance Amplifier) providing required transfers and loops [14]. The function of the CCTA is described by the following equations [12]: $i_{OUT+} = -i_{OUT-} = g_m v_{Z+}$, $v_X = v_Y$ and $i_{Z+} = -i_{Z-} = i_X$, where parameter g_m is transconductance. The OTA is described by relation: $i_{OUT} = g_m(v_{IN+} - v_{IN-})$. The OTRA can be easily described as follows: $v_{Z+} = -v_{Z-} = R_m i_{IN}$.

Fig. 1. Scheme of the proposed multiphase oscillator with six voltage outputs implemented with the CCTA, OTA and OTRA active elements.

The characteristic equation of the proposed fractional-order oscillator is given by:

$$s^{\alpha+\beta} C_\alpha C_\beta - s^\alpha C_\alpha g_{m1} + \frac{g_{m2}}{R_G} = 0, \quad (3)$$

where the substitution $s = j\omega$ change the characteristic equation as follows:

$$j^{\alpha+\beta} \omega_0^{\alpha+\beta} C_\alpha C_\beta - j^\alpha \omega_0^\alpha C_\alpha g_{m1} + \frac{g_{m2}}{R_G} = 0, \quad (4)$$

The real and imaginary parts of characteristic equation obtained from (4) are as follows:

$$\Re : \omega_0^{\alpha+\beta} C_\alpha C_\beta \cos \left[\frac{\pi}{2} (\alpha + \beta) \right] - \omega_0^\alpha C_\alpha g_{m1} \cos \left(\frac{\pi}{2} \alpha \right) + \frac{g_{m2}}{R_G} = 0, \quad (5)$$

$$\Im : \omega_0^{\alpha+\beta} C_\alpha C_\beta \sin \left[\frac{\pi}{2} (\alpha + \beta) \right] - \omega_0^\alpha C_\alpha g_{m1} \sin \left(\frac{\pi}{2} \alpha \right) = 0. \quad (6)$$

The relations for calculation of values of the transconductances g_{m1} and g_{m2} are obtained by rearranging the equations (5) and (6) to:

$$g_{m1} = \frac{\omega_0^\beta C_\beta \sin \left[\frac{\pi}{2} (\alpha + \beta) \right]}{\sin \left(\frac{\pi}{2} \alpha \right)}, \quad (7)$$

$$g_{m2} = \frac{\omega_0^{\alpha+\beta} C_\alpha C_\beta \sin \left(\frac{\pi}{2} \beta \right) R_G}{\sin \left(\frac{\pi}{2} \alpha \right)}, \quad (8)$$

The equations describing relationship between the voltages v_1 , v_2 , and v_3 of the circuit from Fig. 1 are the following:

$$v_1 = \frac{v_2 g_{m2}}{\omega_0^\alpha C_\alpha} e^{-j \frac{\pi}{2} \alpha}, \quad (9)$$

$$v_3 = v_2 R_m \omega_0^\beta C_\beta e^{j \frac{\pi}{2} \beta}, \quad (10)$$

where the phase shift between v_1 and v_2 is $-\alpha\pi/2$, phase shift between v_3 and v_2 is $\beta\pi/2$ and between v_1 and v_3 is $(\alpha + \beta)\pi/2$.

For computing of the initial values of the oscillator parameters, the initial pole frequency and values of the α , β is important. In Fig. 2 (a) the Matlab simulations of dependence of values of the g_{m1} on selected frequency and $\alpha = \beta$ (we consider $\alpha = \beta$ for simplification) is depicted. The analysis is performed based on equations (7) and (1) for values of the pole frequencies in range from 0 to 100 kHz, values of $\alpha = \beta$ in range from 0 to 1 and value of equivalent capacitance is 6.8 nF. In Fig. 2 (b) is shown the dependence of the values of g_{m2} on frequency. Note that values of the g_{m2} are dependent only on frequency as is obvious from substitution the equation $C_\alpha = C_{eq} \omega^{1-\alpha}$ to (8). These analysis are very helpful in case of selecting the initial values of the oscillator parameters. For example in selecting the initial pole frequency and values of the α , β of the proposed circuit.

The tuning of the FO can be performed by changing values of the g_{m1} and g_{m2} as is seen from following equation:

$$\omega_0 = \left[\frac{g_{m1}}{2 C_\alpha \cos \left(\frac{\pi}{2} \alpha \right)} \right]^{1/\alpha} = \left[\frac{g_{m2}}{R_G C_\alpha^2} \right]^{1/2\alpha}. \quad (11)$$

Note that the condition of oscillation is fulfilled when the whole equation (11) is valid.

III. CIRCUIT SIMULATIONS

The design correctness of the proposed fractional-order multiphase oscillator from Fig. 1 was verified by simulations in OrCAD PSpice with suitable behavioural models of the used active elements. The OTA and CCTA are in this paper implemented by UCCs (Universal Current Conveyors) with resistor R , representing the transconductance with value $g_m = 1/R$, connected to X current terminal [13]. The simulation model

(a)

(b)

Fig. 2. Demonstration of the dependence of the values of transconductances on frequency and parameters $\alpha = \beta$: (a) g_{m1} , (b) g_{m2} .

of the UCC is designed based on measurement of UCC-N1B_0520 chips [16]. The OTRA is implemented by UCCs and CCII+/- with resistor R_m connected to the X terminal. The UCC chip has been developed by Brno University of Technology and by ON Semiconductor Brno.

Both of capacitors are in the structure replaced by fractional-order capacitors with variable values of the order.

For fractional-order oscillator design presented in this paper, the FOC approximation through suitable and fully sufficient RC network [15] is used. The approximation of the FOE is dependent on selected type of the RC structure, number of the RC sections, type of the calculation method, value of the α , and frequency band of the approximation validity [7], [8], [15]. The Foster I, Foster II, Cauer I, and Cauer II are types of RC structures most often used to approximate the FOC. In case of calculation of the values of passive parts in the selected RC structure, the Oustaloup's or CFE (Continuous Fraction Expansion) methods are most commonly used [7], [8], [15].

In our design, the FOC is emulated by 5th-order RC structure called Foster I, which is demonstrated in Fig. 3. The values of the passive parts are calculated by Oustaloup's algorithm in MATLAB based on selected parameter of α , number of the sections of the RC structure, value of the equivalent capacitor and frequency band of approximation validity. The

obtained values of the passive parts are summarized in Tab. I. Magnitude and phase responses of the FOC for three values of $\alpha = \beta$ at central frequency 10 kHz are depicted in Fig. 4.

Fig. 3. The 5th-order structure approximating the FOC (Fractional-Order Capacitor).

TABLE I
USED VALUES OF RESISTORS AND CAPACITORS IN THE RC STRUCTURE
FOR THREE CONSIDERED VALUES OF $\alpha = \beta$.

α [-]	0.3	0.5	0.8
C_{eq} [nF]	6.8		
$C_\alpha = C_\beta$ [μ F/sec $^{1-\alpha}$]	15.5	1.7	0.062
R_0 [Ω]	588	233.9	58.8
R_1 [Ω]	365.2	241.5	65.14
R_2 [Ω]	685.7	675.9	326.4
R_3 [Ω]	1208.2	1734.7	1470
R_4 [Ω]	2143.9	4545	7060
R_5 [Ω]	4328.1	15973	84232
C_1 [nF]	1.44	2.62	12.8
C_2 [nF]	4.85	5.91	16.2
C_3 [nF]	17.37	14.5	22.6
C_4 [nF]	61.75	35	29.7
C_5 [nF]	193	62.9	15.7

Fig. 4. Magnitude and phase responses of the FOC implemented by 5th-order RC structures for three values of $\alpha = \beta$ at central frequency 10 kHz.

The proposed circuit was tested for three values of $\alpha = \beta$ when the FO was set to 10 kHz. Using (7), (8), the calculated values of the g_{m1} and g_{m2} are summarized in Tab. II, which assume ideal active elements. For simulations were also used the following values of auxiliary passive parts: $R_G = R_{1AUX} = R_m = 1$ k Ω , $R_{2AUX} = 1/g_{m1} = \{1.315, 1.656, 3.787\}$ k Ω . The obtained waveforms of the proposed oscillator with non-ideal models of the used active elements for different values of the $\alpha = \beta$ are depicted in Fig. 5. To start up the oscillations, the values of g_{m1} had to be modified (Tab. II) due to the influence of parasitic properties of the real (non-ideal) models of the active elements. The modified

values of $g_{m1} = 1/R_{2AUX}$ for three values of $\alpha = \beta$ are: 0.787, 0.635, 0.290 mS. The impact of non-ideal properties of the active elements is further analysed in section IV.

TABLE II
CALCULATED VALUES OF THE PASSIVE COMPONENTS FOR THREE VALUES OF $\alpha = \beta$ WHEN FO WAS 10 KHz.

$\alpha = \beta$ [-]	0.3	0.5	0.8
C_{eq} [nF]	6.8		
$C_\alpha = C_\beta$ [μ F/sec $^{1-\alpha}$]	15.5	1.7	0.062
$g_{m1} = 1/R_{gm1}$ [mS] (modified g_{m1})	0.76 (0.787)	0.6 (0.635)	0.26 (0.290)
$g_{m2} = 1/R_{gm2}$ [mS]	0.1825	0.1827	0.183

The FO of the circuit in case of $\alpha = \beta = 0.3$ and $\alpha = \beta = 0.5$ equals to 10.251 kHz. For circuit with $\alpha = \beta = 0.8$ is FO 10.183 kHz. The obtained phase shifts between output voltages v_2 and v_1 for three values of $\alpha = \beta$ are: 26.27, 44.4 and 70.3°. The phase shifts between output voltages v_3 and v_2 for three values of $\alpha = \beta$ are: 26.15, 45.6 and 72.7°. Theoretical phase shifts between v_2 and v_1 , v_3 and v_2 for three values of $\alpha = \beta$ are: 27, 45 and 72°. It obvious that obtained values of phase shifts are very close to the theoretical expectation. The differences are caused by parasitic properties of the used active elements and by rounding of values of resistors and transconductances.

The proposed fractional-order oscillator provides ability to control the FO by changing the values of g_{m1} and g_{m2} . The circuit was tested for four values of the FO: 2.5, 5, 10, and 50 kHz. The values of calculated values of the g_{m1} and g_{m2} are summarized in Tab. III. The frequency range of tunability has been chosen based on validity of the FOC approximation as shown in Fig. 4. In order to extend the range of the frequency tunability of the oscillator, another FOC designed with wide range of the approximation must be used. The obtained waveforms of the circuit with non-ideal simulation models of the active elements for FO tuning are depicted in Fig. 6. To start up the oscillations, the values of the g_{m1} had to be modified. These values are stated in parentheses in the Tab. III. The values of the FO for selected output voltage v_2 are as follows: 2.468, 5.164, 10.251, and 50.01 kHz. From obtained values of FO is obvious that results correspond with theoretical expectations.

TABLE III
CALCULATED VALUES OF THE TRANSCONDUCTANCES FOR FOUR VALUES OF FO WHEN $\alpha = \beta = 0.5$

FO [kHz]	2.5	5	10	50
$g_{m1} = 1/R_{gm1}$ [mS] (modified g_{m1})	0.302 (0.326)	0.427 (0.46)	0.6 (0.635)	1.351 (1.46)
$g_{m2} = 1/R_{gm2}$ [mS]	0.046	0.091	0.183	0.913

The THD is analysed for three values of $\alpha = \beta$ of the oscillator voltage outputs v_1 , v_2 and v_3 when the FO was set to 10 kHz. The demonstration of the spectral analysis of the output voltage v_2 of the proposed circuit is demonstrated in Fig. 7. The THD was calculated for all values of $\alpha = \beta$ and

Fig. 5. Waveforms of the voltage outputs v_1 , v_2 , v_3 and their inverted variants (dashed lines) for three values of $\alpha = \beta$ when the FO (Frequency of Oscillation) was 10 kHz: (a) $\alpha = \beta = 0.3$, (b) $\alpha = \beta = 0.5$, (c) $\alpha = \beta = 0.8$.

output voltages v_1 , v_2 and v_3 . The values of THD for v_1 when $\alpha = \beta = 0.3, 0.5$, and 0.8 are following: 0.96 %, 0.71 %, and 0.78 %. The values of THD calculated for v_2 are: 1.27 %, 1.91 %, and 0.5 %. The values of THD calculated for v_3 are: 1.71 %, 1.93 %, and 1.82 %. It can be seen that THD is relatively low (below 2 %) for all values of $\alpha = \beta$ and output voltages.

IV. PARASITIC ANALYSIS

The parasitic analysis of the fractional-order oscillator was carried out. In Fig 8 are depicted most significant parasitic admittances of the output and input terminals of the used active elements. The input and output parameters of the OTA

Fig. 6. Waveforms of the voltage output v_2 for four different values of FO when $\alpha = \beta = 0.5$

Fig. 7. Spectral analysis of the results of selected voltage v_2 for three different values of $\alpha = \beta$ when the FO was 10 kHz.

element, which is implemented by UCC as is described in section II, are following: $R_{IN} \approx 1 \text{ G}\Omega$, $C_{IN} \approx 1 \text{ pF}$, $R_{OUT} \approx 200 \text{ k}\Omega$, $C_{OUT} \approx 0.6 \text{ pF}$ [13]. The input and output parameters of the OTRA are: $R_{IN} \approx 1 \text{ }\Omega$, $R_{OUT} \approx 1.1 \text{ }\Omega$ [?].

Fig. 8. Scheme of the proposed oscillator from Fig. 1 with parasitic admittances.

The parasitic admittances of the used active elements of the oscillator are given by: $Y_{P1} = G_{P1} + sC_{P1} = 1/R_{OUT-OTA} + 1/R_{IN-CCH} + sC_{OUT-OTA} + sC_{IN-CCH}$, $Y_{P2} = G_{P2} + sC_{P2} = 1/R_{OUT-BOTA} + 1/R_{IN-OTA} + 1/R_{OUT-CCH} +$

$1/R_{IN-BOTA} + sC_{OUT-BOTA} + sC_{IN-OTA} + sC_{OUT-CCH} + sC_{IN-BOTA}$, $Y_{IN-OTRA} = G_{P3} = 1/R_{IN-OTRA}$.

From before mentioned parasitic admittances, Y_{P2} has most significant effect on circuit function in comparison with other parasitic admittances. The Y_{P2} influences all output voltages of the oscillator. The G_{P2} has biggest impact on fulfilment of the condition of oscillation. The parasitic capacitance C_{P2} has most significant influence on frequency of the oscillations. The impact of parasitic resistance R_{P2} on oscillator function is depicted in Fig. 9. The analysis is performed for output voltage v_2 , three values of the $\alpha = \beta$ and six selected values of the R_{P2} (25, 50, 100, 250, 500, and 5000 k Ω) when $C_{P2} = 3.2 \text{ pF}$ and FO was 10 kHz. From obtained results of the circuit with $\alpha = \beta = 0.3$ for six selected values of R_{P2} , it is obvious that condition of oscillation is fulfilled when $R_{P2} \approx 5000 \text{ k}\Omega$. In case of oscillator with $\alpha = \beta = 0.5$ and $\alpha = \beta = 0.8$, the condition of oscillation is not fulfilled for $R_{P2} < 250 \text{ k}\Omega$. From obtained analysis is obvious that the parameter g_{m1} or g_{m2} must be modified for start up the oscillations because the calculated value of the R_{P2} , based on parallel connection of the input and output parasitic parameters of the active elements, equals to 83 k Ω .

The impact of C_{P2} on FO for three values of the $\alpha = \beta$ is shown in Fig. 10. The analysis was carried out for C_{P2} in range from 0.1 pF to 100 pF when R_{P2} was 5000 k Ω . It is obvious that higher values of the C_{P2} cause frequency shift. Most significant frequency shift is for oscillator with $\alpha = \beta = 0.3$. As mentioned above, in the simulation models, the parasitic capacitance C_{P2} result from the interconnection of active elements is approximately 3.2 pF only. This causes the error in oscillation frequency 0.58, 0.39 and 0.17 % ($\alpha = \beta = 0.3, 0.5, 0.8$) only. The results in Fig. 10 show that for higher values of C_{P2} the oscillation remains.

V. CONCLUSION

The multiphase fractional-order oscillator with controllable FO was designed and analysed in this paper. In case of $\alpha \neq \beta$, the proposed circuit provides three voltage outputs and their inverted variants with phase shift equal to multiple of $\alpha\pi/2$ and $\beta\pi/2$. Note that this oscillator with arbitrary phase shift can find the application in signal modulators for example. The fractional-order capacitors, designed by Foster I type of the RC structure, were used instead of standard capacitors in the circuit structure. The proposed oscillator was tested for three values of the $\alpha = \beta$. The circuit provides FO tuning by changing values of g_{m1} and g_{m2} and was tested for four selected values of FO in range from 2.5 to 50 kHz when $\alpha = \beta = 0.5$. The THD of the circuit, which was analysed for three values of the order, is below 2 % for all voltage outputs. The parasitic analysis was also carried out for most significant parasitic admittances of the oscillator. From this analysis was found that the Y_{P2} has greatest impact to the circuit function. The parasitic resistance R_{P2} influences the condition of oscillation. The parasitic capacitance C_{P2} causes the most significant frequency shift compared with other parasitics. The

(a)

(b)

(c)

Fig. 9. Influence of R_{p2} parasitic on waveform of the output voltage v_2 when $C_{p2} = 3.2$ pF: (a) $\alpha = \beta = 0.3$, (b) $\alpha = \beta = 0.5$, (c) $\alpha = \beta = 0.8$.

function of the circuit is supported by PSpice simulations with non-ideal models of the active elements.

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Fig. 10. Influence of C_{p2} parasitic on frequency of the output voltage v_2 when $R_{p2} = 5000$ kΩ.

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