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Fully-Differential Universal Frequency Filter with Dual-Parameter Control of the Pole Frequency and Quality Factor

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Abstract—The paper presents a proposal of modified 2nd-order fully-differential universal frequency filter. Both the pole frequency and quality factor (controlled without disturbing each other) of the filter can be controlled by two independent parameters (i.e. dual-parameter control) in order to increase the range of their control. It is the main improvement in comparison to recent state-of-the-art. The function of the proposed filter has been verified by PSpice simulations and also by experimental measurements.

Keywords—adjustable current amplifier; dual-parameter control; fully-differential; frequency filter; operational transconductance amplifier

I. INTRODUCTION

Although the digital signal processing is in the forefront nowadays, analogue frequency filters are still a necessity for wide range of electrical circuits. Thus, frequency filters which possess ability to adjust some of their parameters such as its pole frequency, quality factor, bandwidth, or gain in the pass-band area are beneficial. Moreover, the controllability provided in wider range is highly advantageous.

Controllability of the filter can be achieved a) electronically by adjusting one or more parameters of active element(s) [1-4], b) partly electronically by adjusting one parameter of used active element(s) and value of some passive part(s) [5], [6], c)

by variation of values of passive elements [7-9] only. Transconductance g_m [1-3], [5], [6] and current gain B [1], [4], [6] are typical used adjustable parameters of active elements. These parameters use to be controlled by DC voltage or current. Other possible parameter used for adjustability is intrinsic impedance R_X of used active element controlled by DC current [2-4] or also transresistance R_T [5]. In case of control by adjusting values of passive elements, adjusting values of resistors (conductances), capacitors [7-9] is only possible way. Table I provides a comparison of significant prior dual-parameter control filter designs.

The proposed filter provides the dual-parameter control of its pole frequency and quality factor and it has been designed in its fully-differential (F-D) form due to the advantages of F-D structures when compared to the single-ended (S-E) circuits such as greater dynamic range of the processed signals, better power supply rejection ratio, lower harmonic distortion and greater attenuation of common-mode signal [10].

II. PROPOSAL DESCRIPTION

The proposed filter is a modification of previously proposed filter from [5] in order to provide a dual-parameter control ability to electronically tune the pole frequency and quality factor of the filter without disturbing each other. The presented filtering structure is designed using multi-output

TABLE I. COMPARISON OF THE THEORETICAL, SIMULATED AND MEASURED POLE FREQUENCIES

Ref No.	[1]	[2]	[3]	[4]	[5]	[6]	Fig. 4
Dual-param. control of f_0/Q	Yes/No	Yes/No	Yes/Yes	Yes/No	No/Yes	Yes/Yes	Yes/Yes
Type of control	g_m/B	R_X/g_m	R_X/g_m	R_X/B	g_m/R_T	g_m/B	g_m/R_T
Results	measured	simulated	simulated	simulated	measured	simulated	measured
Tested range of f_0 [kHz]	38-2244	1750-7520	119-991	43-4190	-	800-2500	36-2897
$f_{0min}-f_{0max}$ ratio [-]	59.1	4.3	8.3	97.4	-	3.1	80.5
Tested range of Q [-]	-	-	1-16	-	0.47-2.83	0.6-1.9	0.44-8.3
$Q_{min}-Q_{max}$ ratio [-]	-	-	16	-	6.0	3.2	18.9

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transconductance amplifiers (MOTAs) [11], adjustable current amplifiers (ACAs) [12] and transresistance amplifier (OTRA) [13]. The MOTA element consists of two voltage input and four current output terminals and its schematic symbol is shown in Fig. 1 a). Relations between input and output terminals of the MOTA can be described as $I_{OUT+} = -I_{OUT-} = g_m(V_{IN+} - V_{IN-})$, where g_m is the transconductance of this element. Figure 1 b) shows a possible implementation of the MOTA using a universal current conveyor (UCC) [14], which has been developed in cooperation of Brno University of Technology and ON Semiconductor design center.

Fig. 1. Multi-Output Transconductance Amplifier: a) schematic symbol, b) possible implementation of the MOTA using the UCC

The ACA element consists of two current differential input and two or four current differential output terminals. Its schematic symbol and possible implementation using a UCC, universal voltage conveyor (UVC) [15] and EL2082 [16] is depicted in Fig 2 a), b), respectively. The relation between input and output terminals are given by $I_{OUT+} = -I_{OUT-} = B(I_{IN+} - I_{IN-})$, where B is the current gain of this element.

Fig. 2. Adjustable Current Amplifier: a) schematic symbol, b) possible implementation of the ACA using the UCC, UVC and EL2082

The schematic symbol of the OTRA element together with its possible implementation using two UCC chips (each chip contains one UCC and one CCII) is illustrated in Fig. 3 a), b). This element consists of two current input and two voltage output terminals. The OTRA can be described by $v_{O+} = R_T(I_P - I_N)$, $v_{O-} = -R_T(I_P - I_N)$, where R_T is the transresistance of this element.

Fig. 3. Operational Transresistance Amplifier: a) schematic symbol, b) possible implementation of the OTRA using two UCCs and two CCII

The proposed circuit structure of F-D filter is shown in Fig. 4. Since the floating capacitors are not suitable for implementation and integrability, floating capacitors in the filter structure have been replaced by grounded ones as can be seen in Fig. 4.

The denominator which is common for all obtainable transfer functions of the filter is given as:

$$D(s) = s^2 C_1 C_2 + s C_2 g_{m1} g_{m3} B_1 R_T + g_{m1} g_{m2} B_1 B_2. \quad (1)$$

The pole frequency (f_0) and quality factor (Q) of the proposed filter can be found in forms:

$$f_0 = \frac{1}{2\pi} \sqrt{\frac{g_{m1} g_{m2} B_1 B_2}{C_1 C_2}}, \quad Q = \frac{1}{g_{m3} R_T} \sqrt{\frac{g_{m2} C_1}{g_{m1} C_2}}, \quad (2)$$

where g_{m1} , g_{m2} , g_{m3} are transconductances of particular MOTA elements, R_T is the transresistance of the OTRA and B is current gain of the DACA elements. From equations (2) it is obvious that f_0 of the filter can be adjusted electronically by changing values of transconductances g_{m1} , g_{m2} when fulfilling a simple condition $g_{m1} = g_{m2}$ and/or by changing current gains B_1 , B_2 when $B_1 = B_2$. The quality factor of the filter can be adjusted electronically by changing values of g_{m3} and/or R_T . Therefore, presented topology allows two ways of the pole frequency and quality factor control. Both methods can be used independently, or they can be combined in order to extend range of control of f_0 and Q (dual-parameter control).

The obtainable transfer functions of the proposed filter are described by following equations:

$$\frac{I_{LP+}}{I_{IN}} = -\frac{I_{LP-}}{I_{IN}} = \frac{g_{m1} g_{m2} B_1 B_2}{D(s)}, \quad (3)$$

Fig. 4. Circuit structure of the proposed dual-parameter control fully-differential SITO filter

$$\frac{I_{BP+}}{I_{IN}} = -\frac{I_{BP-}}{I_{IN}} = \frac{sC_2g_{m1}g_{m3}B_1R_T}{D(s)}, \quad (4)$$

$$\frac{I_{HP+}}{I_{IN}} = -\frac{I_{HP-}}{I_{IN}} = \frac{s^2C_1C_2B_1}{D(s)}, \quad (5)$$

$$\frac{I_{BR+}}{I_{IN}} = -\frac{I_{BR-}}{I_{IN}} = \frac{s^2C_1C_2B_1 + g_{m1}g_{m2}B_1B_2}{D(s)}, \quad (6)$$

$$\begin{aligned} \frac{I_{AP+}}{I_{IN}} &= -\frac{I_{AP-}}{I_{IN}} = \\ &= \frac{s^2C_1C_2B_1 - sC_2g_{m1}g_{m3}A_1R_T + g_{m1}g_{m2}B_1B_2}{D(s)}. \end{aligned} \quad (7)$$

The equations (3) to (7) indicate availability of all common types of transfer functions (low-pass (LP), band-pass (BP), high-pass (HP), band-reject (BR) and all-pass (AP)). BR and AP transfer functions can be easily obtained by summing operation of particular outputs (LP+HP or LP+BP+HP) of the circuit because the filter is of the single-input multiple-output (SIMO) type. Thus, the proposed filter is universal. All responses were taken directly from high-impedance outputs active elements.

III. SIMULATION AND EXPERIMENTAL RESULTS

The function of the proposed filter has been verified by PSpice simulations and experimental measurement of the implemented filter in form of a printed circuit board. Simple V/I, I/V converters and network analyzer Agilent 4395A were

used for the experimental measurement. The converters, were implemented by OPA860 [17] and OPA861 [18]. According to the way of particular implementation of individual active elements using discrete integrated circuits (UCC, UVC, EL2082) as stated in section 2, the proposed filter, in case of experimental measurement, did consist of total 7 UCC chips (supply ± 1.65 V, 28 mA), 2 UVC chips (same as UCC), and 2 EL2082 chip (± 5 V, 15 mA). Thus, the resulting power consumption was approximately 1.1 W, in this particular case. We used so-called behavioral modeling using discrete chips to confirm the design correctness before rather expensive integrated circuit (IC) implementation.

Initial values of characteristic filter parameters and values of passive parts used for the simulations and experimental measurements have been selected as follows: initial values of transconductances g_{m1} , g_{m2} were set to 0.5 mS, capacitors $C_{11} = C_{12} = 440$ pF and $C_{21} = C_{22} = 880$ pF, starting values of current gains of the ACA elements were $B_1 = B_2 = 0.25$. Thus, the calculated initial value of the pole frequency is equal to 128 kHz. The value of transconductance g_{m3} has been set to 3.92 mS and transresistance $R_T = 510 \Omega$ which results in quality factor of value 0.68. Resistors have been used from E24 series.

Figure 5 compares simulation and experimental results of low-pass (green lines), band-pass (red lines) and high-pass (blue lines) transfer functions. The slope of attenuation of simulated high-pass transfer function is 38 dB per decade and the slope of attenuation of high-pass function obtained from measurement is 36 dB per decade. In case of BP it is 18.5 dB per decade for the simulation results and 17.5 dB per decade for the measurement results. The attenuation slope reaches 39.5 dB per decade for LP transfer responses in case of simulations and 37.5 dB per decade for the measurements. The largest difference between the simulation and experimental results can be seen at higher frequencies. The difference is given because of bandwidth limitations of used active elements and V/I, I/V converters. The overpeaking at approximately 10 MHz is caused by parasitic properties of the UCC. Most

likely the influence of its parasitic capacitance with resistance R_T in case of the OTRA element particular implementation.

Figure 6 illustrates f_0 tunability range. Low-pass transfer function was chosen as an example. The lowest value of f_0 reaches 35.8 kHz when g_{m1} , g_{m2} were set to 0.278 mS and B_1 , B_2 to 0.1. The transfer function does not provide unity gain in the pass-band when lower values of f_0 than indicated are set. The highest value of f_0 is 2897.1 kHz when g_{m1} , g_{m2} were set to 1 mS and B_1 , B_2 to 3. The higher values of f_0 , causes significant parasitic influence of tuning process on Q . Table II compares the theoretical values of the pole frequency with the values obtained from simulations and experimental measurements. The measured results have slightly lower values in comparison to the theoretical expectations and results of simulations. This is mainly caused by realistic characteristics of used active elements and tolerance of passive elements.

Fig. 5. Comparison of transfer functions high-pass, band-pass and low-pass of the proposed filter: simulation (solid lines) and experimental results (dashed lines)

Fig. 6. Demonstration of obtainable tunability range of the pole frequency of the proposed filter in case of the LP function: simulation (solid lines) and experimental results (dashed lines)

TABLE II. COMPARISON OF THE THEORETICAL, SIMULATED AND MEASURED POLE FREQUENCIES

B [-]/ g_m [mS]	0.1/0.278	0.25/0.5	3/1
Theoretical f_0 [kHz]	28.4	127.9	3069.3
Simulated f_0 [kHz]	26.8	122.6	2937.7
Measured f_0 [kHz]	35.8	109.8	2897.1

Available tunability range of the quality factor of the proposed filter for experimental results is depicted in Fig. 7. Band-pass transfer function was selected for this test. The lowest measured Q has value 0.44 when g_{m3} was set to 3.92 mS and R_T was set to 680 Ω . The highest obtainable Q is 8.3 when g_{m3} was set to 1 mS and R_T was set to 130 Ω . For higher values of Q , f_0 is being shifted together with Q . A comparison of the theoretical, simulation and experimental values of the quality factor can be made from Table III. The values of the quality factor obtained from the experimental measurements are slightly lower than it is for theoretical values and simulation results. This is mainly caused by the reasons mentioned above.

Fig. 7. Demonstration of obtainable tunability range of the quality factor of the proposed filter in case of the BP function: simulation (solid lines) and experimental results (dashed lines)

TABLE III. COMPARISON OF THE THEORETICAL, SIMULATED AND MEASURED QUALITY FACTORS

g_{m3} [mS]/ R_T [Ω]	3.92/680	3.92/510	1/130
Theoretical Q [-]	0.53	0.71	10.88
Simulated Q [-]	0.53	0.70	9.11
Measured Q [-]	0.44	0.66	8.33

IV. CONCLUSION

Obtained simulation and experimental results are illustrated in the paper and compared with the theoretical expectations. From the tables is evident that the simulation and experimental results are in good agreement with the theoretical results. Both simulation and experimental results prove the possibility of dual-parameter electronic control of f_0 and Q of the proposed filter. The tunability range of f_0 in case of the experimental measurement was achieved between 35.8 kHz and 2897.1 kHz that yields f_{min}/f_{max} ratio of 80.5. The tunability range of Q was tested from 0.44 to 8.33 which provides Q_{min}/Q_{max} ratio of 18.9. Thus, the tunability range has been significantly extended in comparison to standard solutions. We also obtained good results in comparison with other filters with the dual-parameter control.

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