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Abstract: Newly designed frequency filter of the second order with the possibility of mutually independent control of the pole frequency and quality factor is presented in this paper. Balanced-output transconductance amplifier (BOTA) and digitally adjustable current amplifier (DACA) are used as active elements in this proposal. The pole frequency and quality factor can be tuned by the adjustable current gain of three DACAs that are available in form of integrated circuit. The individual output responses of proposed frequency filter are verified using PSpice simulations and by experimental measurement in laboratory.

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Abstract *– Newly designed frequency filter of the second order with the possibility of mutually independent control of the pole frequency and quality factor is presented in this paper. Balanced-output transconductance amplifier (BOTA) and digitally adjustable current amplifier (DACA) are used as active elements in this proposal. The pole frequency and quality factor can be tuned by the adjustable current gain of three DACAs that are available in form of integrated circuit. The individual output responses of proposed frequency filter are verified using PSpice simulations and by experimental measurement in laboratory.*

1 Introduction

Analogue frequency filters are still interesting for many research teams around the world. There are substantial reasons for the continuous development of analogue filters because their application field is very wide: in analogue signal processing, telecommunications, measurement technology, etc. that evolve all the time. The analogue frequency filters can be divided in accordance to the type of input and output signals, as voltage-mode (VM) and current-mode (CM). The design of CM frequency filters is important because of possible improvement in the signal to noise ratio, increase of dynamic range, and in some cases, increasing bandwidth and linearity [1].

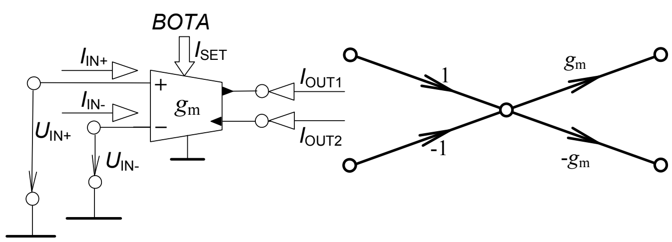
Ref. [2], [3], [4] present frequency filters, which are not able to realize several filtering functions simultaneously. Such proposed frequency filters are not able to change the individual filtering function without intervention into circuit topology of designed frequency filter. It is appropriate to propose a structure that provides suitable filtering functions simultaneously [5], [6], [7].

An electronic controllability is another important aspect in the design of the frequency filters. This is the main reason to develop active frequency filters based on active elements. Active elements useful for frequency filter design are for example Operational Transconductance Amplifier (OTA) [8], its modification (Balanced-Output OTA [9], Multiple-Output OTA [10]), current conveyor of individual generation (CCI [11], CCII [12], CCIII [13]), Current Follower (CF) [14] and its modification (Double-output CF, i.e. DO-CF) [15], Multiple-output CF (MO-CF) [16] and Current Amplifier (CA) [17], frequently used in controllable variant referred as Digitally Adjustable Current Amplifier (DACA). DACAs are also employed in the design of the filter presented in this paper. Several frequency filters employing this type of CA (DACA) have been published, see for example [18], [19], [20], [21], [22]. Some of these articles deal only with hypothetical simulation tests. However, some of them deal with the results of experimental measurement. The alternative circuit implementation of the DACA active element (behavioural model) has been used in experimental measurement in paper [20], [21]. This alternative circuit implementation utilizes features of so-called Universal Current Conveyor UCC [23], Universal Voltage Conveyor UVC [24] and variable gain control building block EL2082 [25]. In this paper are used DACA amplifiers in form of integrated circuits. This integrated circuit labelled as DACA_N [26] was produced in cooperation with ON Semiconductor in 0.35µm CMOS technology.

2 Introduction of used active elements

The frequency filter presented in this paper operates with two types of active elements. The first type of used active element is BOTA (Balanced-Output Transconductance Amplifier). BOTA is practically implemented using current conveyor with resistor R at X terminal of UCC, which allows controlling of transconductance g_m of this BOTA (Fig. 1c). This active element has differential high impedance voltage input terminals and differential high impedance current-output terminals. Note that two BOTA active elements can be established by using one UCC-N1B integrated circuit (2x UCC in single IC package). Schematic symbol, simplified signal-flow graph (SFG) and 3rd–level macro model of UCC-N1B are shown in Figure 1a, b, d. Relationships between inputs and outputs are described by this equation:

$$I_{OUT1} = -I_{OUT2} = g_m(U_{IN+} - U_{IN-}). \quad (1)$$



a) b)

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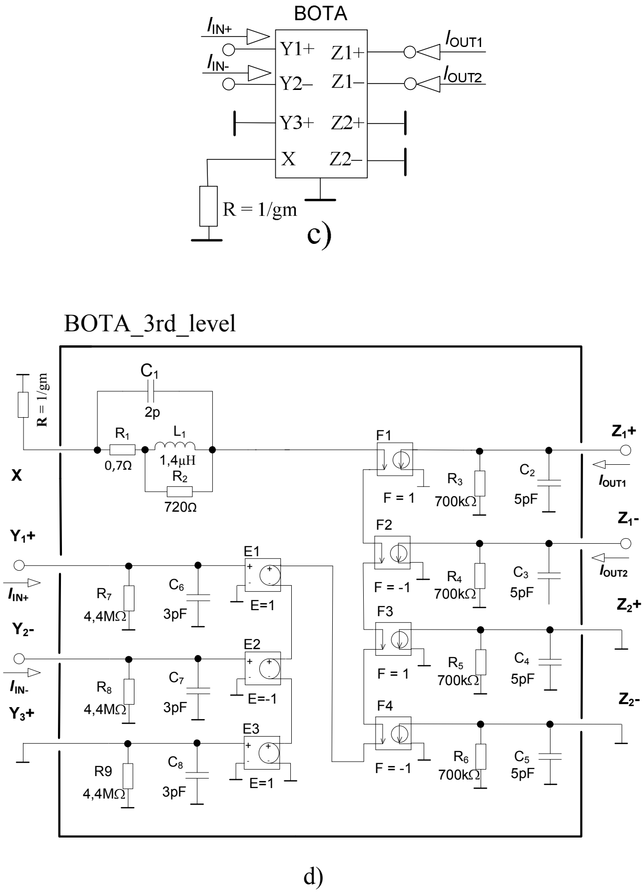

Figure 1: Balanced output transconductance amplifier (BOTA) [9] a) Schematic symbol, b) Simplified SFG, c) implementation using UCC-N1B, d) 3rd-level simulation macro-model

The second used active element is current amplifier DACA. This active element has a differential low impedance current input terminals and differential high impedance current-output terminals. DACA active element has one controllable parameter, particularly current gain referred to as A. It can be controlled via 3-bit bus with step of 1 in range from 1 to 8. DACA active elements are used in form of integrated circuits DACA_N [26] during experimental verification in this paper. Several integrated circuits of DACA_N elements were subjects to measurement of circuit characteristics and then the macro model for PSpice simulation has been proposed and presented in [27]. Design of this macro model presented in [27], was necessary for practical usability of DACA integrated circuit for computer simulations (PSpice). Its small-signal characteristics and parameters exactly correspond to real characteristics of DACA integrated circuits. The 3rd-level macro model is shown in Figure 2c (more detailed description can be found in [27]). Relationships between inputs and outputs are described in the following equations:

$$I_{OUT-} = -A\left(I_{IN+} - I_{IN-}\right), \tag{2}$$

$$I_{OUT+} = A\left(I_{IN+} - I_{IN-}\right), \tag{3}$$

$$I_{DIF\_IN} = \left(I_{IN+} - I_{IN-}\right), \tag{4}$$

$$I_{DIF\_OUT} = \left(I_{OUT+} - I_{OUT-}\right), \tag{5}$$

$$I_{DIF\_OUT} = 2AI_{DIF\_IN}, \tag{6}$$

where I_{DIF_IN}, I_{DIF_OUT} are differential input and output currents, respectively.

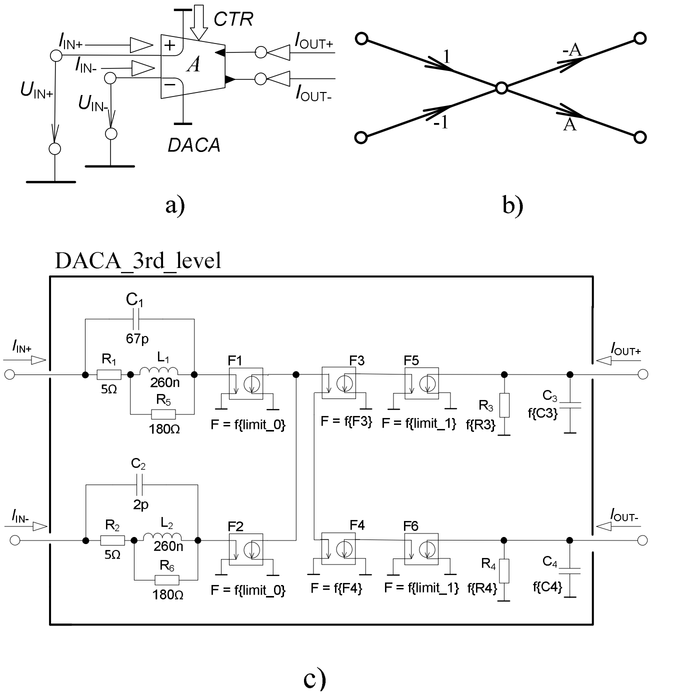

Figure 2: Digitally adjustable current amplifier (DACA) [27] a) Schematic symbol, b) Simplified SFG, c) 3rd-level simulation macro-model

3 Proposed frequency filter operating in current mode

The frequency filter was designed using signal flow graph (SFG) method based on lossless integrators in feedbacks with variable path transfer constants [28], [29]. Active elements used in this design are BOTAs (practically realized using UCC-N1B integrated circuit) and DACAs (in form of integrated circuit DACA) as described in the previous section. Simplicity and possibility of independent control of the pole frequency and the quality factor are the main goals of this work. Simplicity is represented by the lowest number of active and passive elements while maintaining the required output responses high pass (HP), low pass (LP), band pass (BP), band stop (BS) of the second order. Another benefit of this solution consists in operation in the current mode, i.e. all current responses should be taken directly from high-impedance outputs of used active elements. Current output responses (except one, HP particularly) are taken directly available at the outputs of active elements. Unfortunately, there is no possibility to obtain HP response from high-impedance output of active element (problem of topology). Thus, HP is taken from branch with grounded C_1. This disadvantage could be solved by adding of one, Multiple Output Current follower (MO-CF) as an current

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mirror to grounded C_1. However, considering the condition of the smallest number of active elements, this modification was omitted. In addition input impedance of this mirror (follower) may significantly influence transfer performances of the specific filtering transfer responses [30].

The frequency filter is designed as Single Input Multiple Output (SIMO), because it uses only one current low-impedance input. This advantage allows filtering the same input signal with different output current responses, or to get individual output current responses at the same time (if necessary).

The circuitry of the proposed frequency filter and simplified SFG can be seen in Fig. 3. The filter is designed with possibility of independent control of the pole frequency and quality factor. The pole frequency can be electronically controlled by two ways. The first case consists in control of the pole frequency by two controllable current gain parameters (A_1, A_2) of DACA elements. The second case means the possibility of control of the pole frequency by two controllable transconductances (g_{m1}, g_{m2}) of the BOTA elements. Extension of pole frequency tuning range can be achieved using appropriate combination of individual control parameters (g_{m1}, g_{m2}, A_1, A_2).Key objective of this paper is the demonstration of adjustable function of DACA element. Therefore, possibility of controlling of pole frequency using BOTAs is not presented in this paper (parameters g_{m1}, g_{m2} are constant for all types of simulations and experimental measurements). The quality factor can be electronically controlled by DACA element with controllable current gain parameter (A_3).

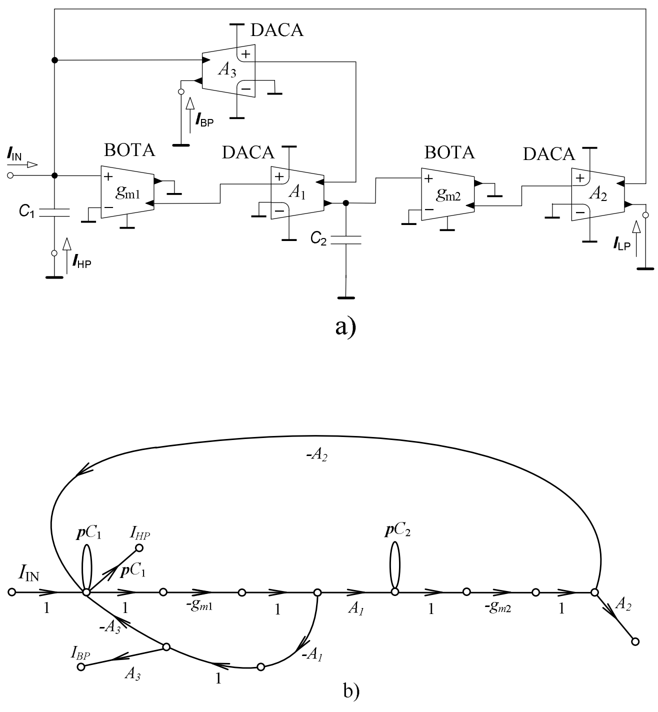

Figure 3: Proposed multifunction SIMO frequency filter of the 2nd order a) Scheme, b) Simplified SFG

The denominator of all transfer functions of the proposed filter is as follows:

$$D(s) = s^2 C_1 C_2 + sC_2 g_{m1} A_1 A_3 + g_{m1} g_{m2} A_1 A_2 , \qquad (7)$$

Following equations represent the individual filtering functions of this filter:

$$\frac{I_{LP}}{I_{IN}} = \frac{g_{m1} g_{m2} A_1 A_2}{D(s)} \qquad (8)$$

$$\frac{I_{HP}}{I_{IN}} = \frac{s^2 C_1 C_2}{D(s)} \qquad (9)$$

$$\frac{I_{BP}}{I_{IN}} = \frac{sC_2 g_{m1} A_1 A_3}{D(s)} \qquad (10)$$

$$\frac{I_{BS}}{I_{IN}} = \frac{s^2 C_1 C_2 + g_{m1} g_{m2} A_1 A_2}{D(s)} \qquad (11)$$

4 Simulation and experimental measurement results

PSpice simulation programme was used to verify functions of proposed frequency filter. In simulation, the active elements expressions are mentioned in Figure 1d for BOTA active element and in Figure 2c for DACA active element.

Experimental measurement results were obtained using network analyser 4395A. PCB (Printed Circuit Board) consists of only three integrated circuits as explained in part Introduction of active elements.

As previously noted, the pole frequency is controllable by current gain A. For independent pole frequency tuning of this filter, this condition $A = A_1 = A_2$ always must fulfilled. In the case of tuning the pole frequency of BP using parameters A_1, A_2, bandwidth of BP response is changed. This is caused by location of A_1 in numenator of (10). Tuning of pole frequency of BP without changing the bandwidth can be performed by control parameter A_2 (A_2 is not in the numerator of (10). When used only parameter A_2, the range of the pole frequency adjusting will be lower (square root dependence). The second electronically controllable parameter is quality factor (again using DACA as mentioned previously). The pole frequency f_0 tuning and quality factor Q tuning are expressed by the following equations:

$$Q = \frac{1}{A_3} \sqrt{\frac{C_1 G_2}{G_1 C_2}}, \qquad (12)$$

$$f_0 = \frac{A}{2\pi} \sqrt{\frac{G_1 G_2}{C_1 C_2}}. \qquad (13)$$

The values of passive elements are as follows: C_1=C_2=330 pF and R_1=2.2 kΩ, R_2=1.1 kΩ. These values are the same for simulation and experimental measurement. Pole frequency tuning is demonstrated in Figure 4 in case of LP response and for the following values of current gain A: 1, 2, 3, 5, 7. The value of electronically controllable parameter quality factor was set to Q = 2 in this particular case. These selected values of A are compared with the corresponding values from simulation in PSpice (dashed black curves). In Figure 4 slight deviations of simulated (dashed black curves) and experimentally measured values (colour solid curves) can be seen. These slight deviations may be caused by inaccurate values of the gains of each current amplifier within an integrated circuit DACA_N [27]. Differences at higher frequencies are mostly caused by the limited bandwidth of the DACA element which bandwidth is approximately 10 MHz. Accurate values of all individual experimentally measured pole frequencies in comparison with the simulations and calculated values are in the Table 1.

Table 1: Comparison of individual values of pole frequency of LP response for measured, simulated and calculated values (Q = 2)

	Calculated	Simulated	Measured
A [-]	f_0 [kHz]	f_0 [kHz]	f_0 [kHz]
1	310	297	279
2	620	623	553
3	930	1000	1054
4	1240	1354	1410
5	1550	1725	1730
6	1860	2094	2351
7	2170	2512	2730
8	2480	2905	3245

Possibility to control the second controllable parameter Q is demonstrated on BP response for specific values of this parameter (Table 2). Parameter serving for the pole frequency control was set as constant A = 1. Values from simulation in PSpice (dashed black curves) and from experimental measurement (colour solid curves) are compared in Figure 5. Particular values are in Table 2 for the case of calculation, simulation and also measurement.

Table 1: Comparison of individual values of quality factor of BP response for measured, simulated and calculated values (A = 1)

	A_3 [-]	1	2	3	4
Calculated	Q [-]	1.414	0.707	0.471	0.354
Simulated	Q [-]	1.920	0.924	0.593	0.476
Measured	Q [-]	1.597	0.802	0.677	0.499

Individual LP, HP, BP, BS responses from simulation in PSpice (dashed black curves) and experimental measurement results (colour solid curves) are compared in Figure 6. Controllable parameters are set to A = 1, Q = 2. Individual slopes of attenuations (LP, HP, BP) and attenuation for BS response (measured, simulated, theoretical) are compared in Table 3. Comparison of phases from individual output responses is shown in Figure 7, only for experimental measurement values for better clarity.

Table 2: Comparison of slopes of attenuations (LP, HP, BP responses) and attenuation for BS response

Response	LP	HP	BP	BS
	Slope of attenua-tion	Slope of attenua-tion	Slope of attenua-tion	Attenua tenua-tion
	[dB/dec]	[dB/dec]	[dB/dec]	[dB]
Theoreti-cal	40.00	40.00	20.00	-∞
Simulated	40.82	39.07	20.48	-39.54
Measured	38.50	35.36	18.74	-40.40

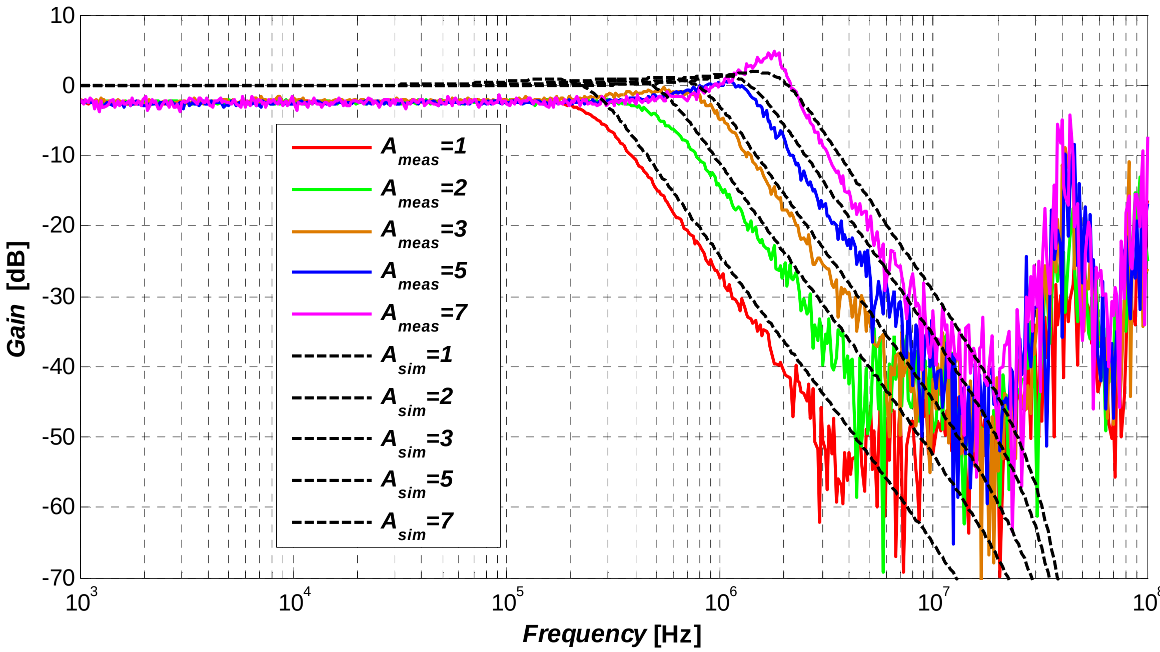

Figure 4: Tuning of the pole frequency of the LP response: comparison of measured and simulated responses (Q = 2)

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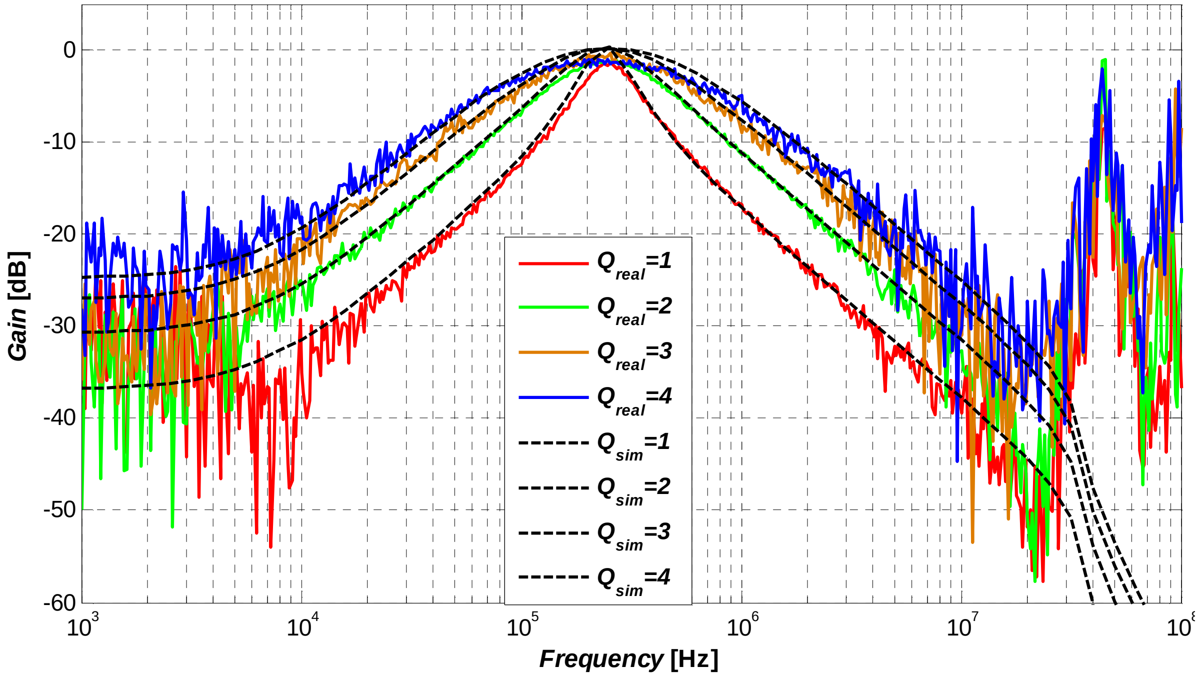

Figure 5: Tuning of the quality factor of the BP response: comparison of the measured and simulated responses (A = 1)

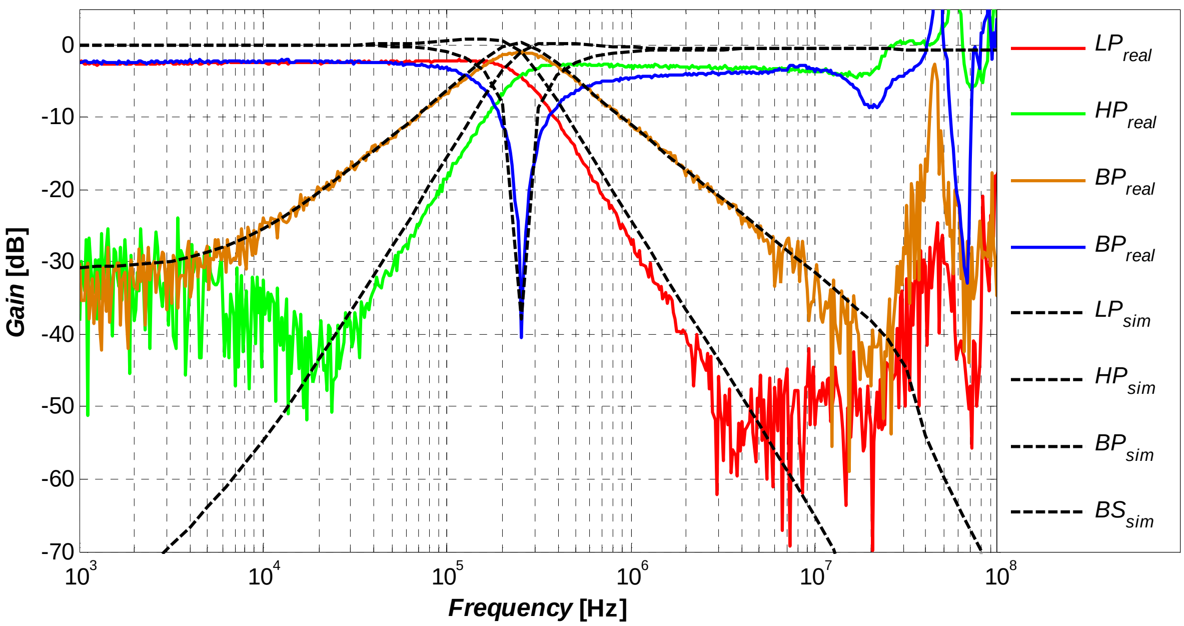

Figure 6: Comparison of measured and simulated LP, HP, BP, BS responses (A = 1, Q = 2)

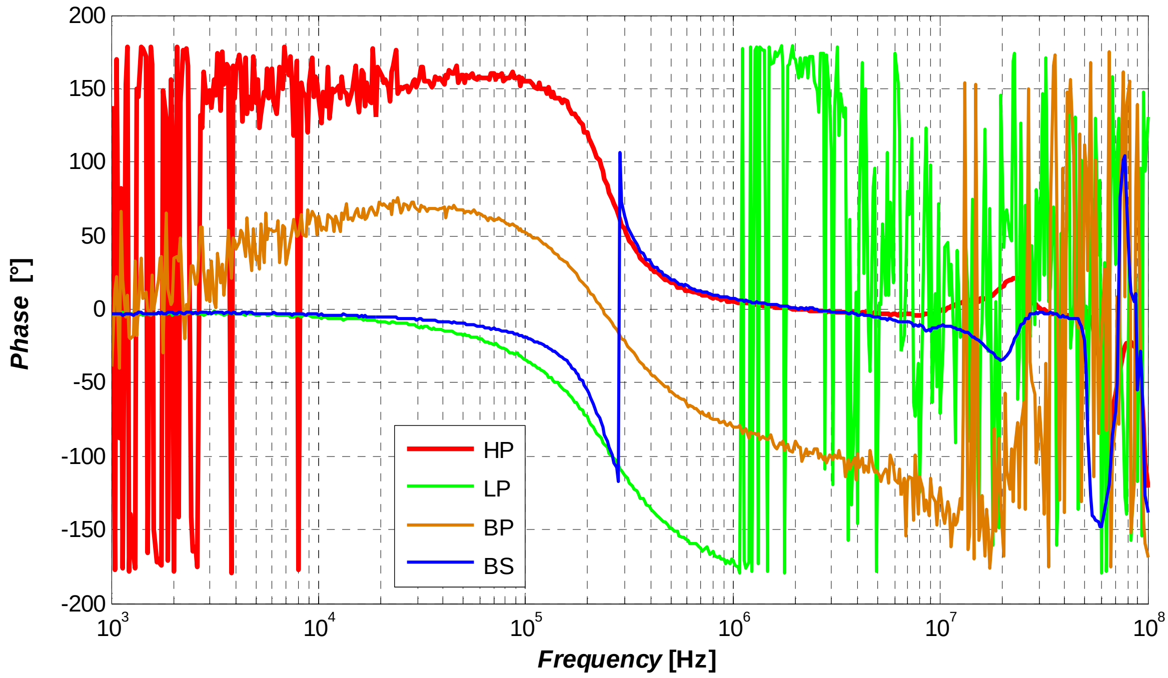

Figure 7: Measured LP, HP, BP, BS phase responses (A = 1, Q = 2)

5 Conclusion

The main topic of this paper is the newly designed frequency filter of the second order with possibility of independent electronically controlling of the pole frequency and quality factor using digitally adjustable current amplifiers. This design was focused on the minimum number of active and passive elements in filtering structure, using the digitally adjustable current amplifiers in order to control the pole frequency and quality factor. All active elements are readily available thanks to integrated circuits UCC-N1B [23] and DACA_N [26]. The proposed frequency filter contains two integrated circuits of DACA type and one integrated circuit UCC-N1B. The pole frequency is controllable using two current amplifiers (the first integrated circuit DACA_N) and quality factor can be controlled using one current amplifier (the second integrated circuit DACA_N).

Simulations and experimental measurements show that the frequency filter is able to realize all designed output responses (LP, HP, BP, BS). Distortion of the results of experimental measurements can be caused inappropriate settings of Resolution Bandwidth (RBW filter of analyzer 4395A). A difference between simulations and experimental measurements causes insignificant attenuation with a value about -2 dB. This attenuation can be caused by converters U-I and I-U used in the experimental measurement with the analyzer 4395A. All results of experimental measurements and performed simulations prove the design correctness.

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