Reconnection-Less Electronically Reconfigurable Filter with Adjustable Gain Using Voltage Differencing Current Conveyor

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Abstract—This paper presents a novel reconnection-less and electronically reconfigurable analogue filter. It employs two active elements, Voltage-Differencing Current Conveyor (VDCC) and Dual-Output Current Amplifier (DO-CA) with two independently tunable gains. Filter has the single input and single output topology (SISO). The current-mode transfer function of the filter can be tuned electronically from the low-pass (LP) response to the high-pass (HP) response, via the band-stop (BS) response, without reconnection of the input or output terminal. It is designed in order to provide tuning of both the stop-band gain and also pass-band gain of all the transfer functions, because also transfer functions with additional and intended zero in transfer function are available. The theoretical presumptions are confirmed by both simulations and lab measurements with available devices.

Index Terms—Dual-output current amplifier; DO-CA; reconfigurable filter; reconnection-less filter; voltage differencing current conveyor; VDCC.

I. INTRODUCTION

A. Reconfigurability of the Transfer Function

Possibility of modification or fine adjusting of the transfer function of an analogue circuit is always a very important requirement in the signal processing domain [1]. Reconfigurability is a very useful feature of the analogue filters since it represents the possibility of very fast change of the type of the frequency response. However, manual galvanic reconnection is usually necessary in order to change the type of the transfer response in standard multifunctional solutions. Unfortunately, manual reconnection is not possible or easily accessible in many cases (*e.g.* in case of on-chip implementation). Many recent scientific works [2]–

[7] solve the problem by the filters referred to as multifunction. These filters have several transfer functions available, but between different nodes of the network. These circuits are referred to as a single input – multiple output (SIMO) [8] or a multiple input – single output (MISO) [9]. Required change of the filtering function can be realized also by switches – this concept is similar to the SIMO and MISO circuits. However, switching of inputs or outputs is not very favourable and causes additional problems (power consumption, overshoots, interferences of clock signal to analogue path, and area consumption on chip). Additionally, some effects caused by discontinuous regime of switching are undesired. Therefore, a very interesting concept is to use two-port structure without changing internal topology or a position of input or output port, used also in microwave technique, for instance [10]. These circuits are referred to as single input – single output (SISO). Our paper focuses on the filtering solution that allows reconnection-less and switch-less control of the type of the transfer function electronically by the parameter(s) of advanced active element and also tuning of the stop-band and/or pass-band attenuation.

Many papers present the filters referred to as reconfigurable [11], [12] but in point-of-view of adjustability of bandwidth or quality factor (in the case of band-pass response). From our point of view, reconfigurable filter [13], [14] represents the type of transfer function where reconfigurability is available without any reconnection in the structure.

B. Active Elements

Many advanced active elements have been reported in recent works. A detailed summarization is available in [15]. However only active elements with the electronically adjustable parameter(s) are interesting for purposes of reconfigurable filters. We focus our attention on adjustable current amplifiers [15]–[17], operational transconductance

amplifiers (OTAs) [15], [18] and current conveyors (CC) [15], [19] because of their useful features, wide availability and easy mutual interconnection.

Several other active elements were derived from these fundamental active elements, usually by complementing two active elements into one structure with particular features. The typical examples are numerous modifications of the current differencing transconductance amplifier (CDTA) [15], [20]–[22]. Several more complex active elements, based on OTA section and CC of the second generation (CCII) have been also proposed. For example, current conveyor transconductance amplifier (CCTA) [15]–[23] utilizes independent R_x and g_m control in some of its variants [24]. Modification of CCTA in [10] employs current gain control. Controllable current gain in frame of CCII seems to be a valuable advantage [25], [26]. This active element is usually referred to as ECCII (Current Conveyor of the second generation with electronically adjustable gain between X and Z terminal). Another interesting active element is Voltage Differencing Current Conveyor (VDCC) [15], [27]–[29]. Also simple current followers (CF) and voltage buffers (VB) or their variations are useful.

C. Comparison with other SISO Solutions in Literature

Several solutions of reconnection-less and reconfigurable filters of SISO type have been published in open literature [30]–[38]. 1st-order filters are summarized in Table I, 2nd-order filters in Table II. Note that the solution presented in this paper is of the 2nd order and therefore it is included in Table II in order to provide a comparison.

TABLE I. 1ST-ORDER RECONNECTION-LESS FILTERS OF SISO TYPE WITH RECONFIGURATION.

Reference	Active elements (number of devices)	Mode of operation	Available transfer functions	Simulated (sim) and/or measured (exp)	Passive components (number)	Tunable pass-band/stop-band gain analysed
[30]	ZC-CG-VDCC (1)	VM	HP, iAP, DT	sim & exp	C (1)	no/no
[31]	OTA (2), ECCII (1)	VM	AP, LPZ, iDT, AZ	sim	C (1)	no/no
[32]	OTA (2)	VM	AP, HP, LP	sim	C (1), R (1)	no/no
[33]	MCDU (1)	CM	AP, iHP, LP	sim & exp	C (1)	no/no
[34]	MCDU (1)	CM	HP, LP, AP	sim	C (1)	no/no

TABLE II. 2ND-ORDER RECONNECTION-LESS FILTERS OF SISO TYPE WITH RECONFIGURATION.

Reference	Active elements (number of devices)	Mode of operation	Available transfer functions	Simulated (sim) and/or measured (exp)	Passive components (number)	Tunable pass-band/stop-band gain analysed
[35]	ECCII (1), VB (1)	VM	AP, BP	sim	C (2)	no/yes
[36]	OTA (5), VB (1)	VM	AP, BS	sim	C (2), R (1)	no/no
[37]	OTA (4)	VM	AP, iBP, BS, LPZ, HPZ	sim	C (2)	no/yes
[38]	OTA (4)	VM	AP, BS, HP, iBP, HPZ, LPZ	sim & exp	C (2)	no/yes
Fig. 4	VDCC (1), CA (1)	CM	LP, HP, LPZ, HPZ, BS	sim & exp	C (2), R (1)	yes/yes

Note: list of previously unexplained abbreviations used in Table I and Table II:
ZC-CCCFDITA - Z-copy Current-Controlled Current Follower Differential Input Transconductance Amplifier
ECCII - Current Conveyor of the second generation with electronically adjustable gain between X and Z terminal
MCDU - Modified Current Differencing Unit
VDCC - Voltage Differencing Current Conveyor
VM, CM - Voltage mode, Current Mode
AP, iAP - (inverting) all-pass filter
HP, iHP, HPZ - (inverting) high-pass filter (with zero)
DT, iDT - (inverting) direct transfer
LP, iLP, LPZ - (inverting) low-pass filter (with zero)
AZ - adjustable zero
BP, iBP - (inverting) band-pass filter
BS, iBS - (inverting) band-stop filter

Our solution consists of only two active elements (it is not so common), operates in the current mode (none of already published 2nd-order filters) and it provides LP, HP, LPZ, HPZ and BS functions. Its features were verified also by lab measurement. Note that its pass-band gain is tunable independently of the stop-band gain. None of the previously reported filters provides this combination of the features.

II. DEFINITION OF ACTIVE ELEMENTS

The Dual Output Current Amplifier (DO-CA) with two independent gains is the first active element with schematic symbol depicted in Fig. 1, together with a description of internal topology. Behaviour of DO-CA is as follows

$$\begin{bmatrix} v_{IN} \\ i_{B1} \\ i_{B2} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ B_1 & 0 & 0 \\ B_2 & 0 & 0 \end{bmatrix} \times \begin{bmatrix} i_{IN} \\ v_{B1} \\ v_{B2} \end{bmatrix}. \tag{1}$$

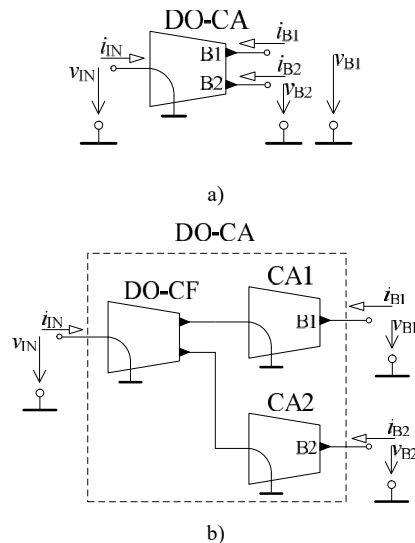Direction of all voltages and currents and meaning of variables from (1) is obvious from Fig. 1.

Fig. 1. Dual Output Current Amplifier (DO-CA) with two independent gains: (a) – schematic symbol, (b) – internal topology.
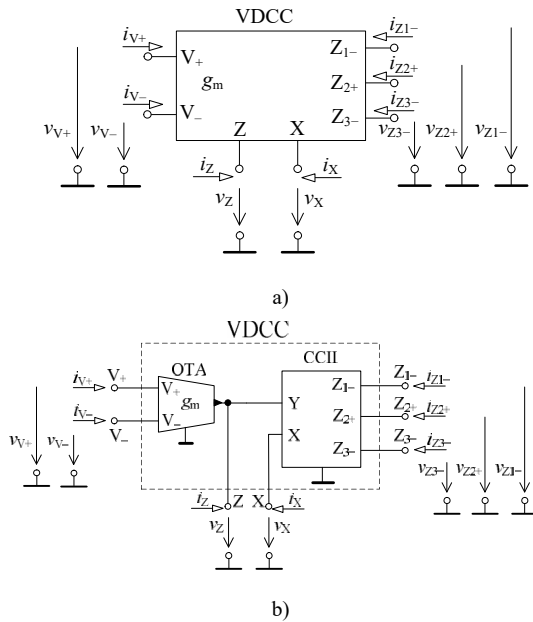
a)

b)

Fig. 2. Voltage Differencing Current Conveyor (VDCC): (a) – schematic symbol, (b) – internal topology.

The Voltage Differencing Current Conveyor (VDCC) is the second type of active element that is present in the designed filter. It is well-known active element consisting of OTA and CCII as is obvious from Fig. 2.

All voltages, currents and variables from Fig. 2 are again included in the matrix

$$\begin{bmatrix} I_{V+} \\ I_{V-} \\ I_Z \\ V_X \\ I_{Z1-} \\ I_{Z2+} \\ I_{Z3-} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ -g_m & g_m & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 & 0 & 0 & 0 \end{bmatrix} \times \begin{bmatrix} V_+ \\ V_- \\ V_Z \\ I_X \\ V_{Z1-} \\ V_{Z2+} \\ V_{Z3-} \end{bmatrix}. \quad (2)$$

III. DESIGNED RECONNECTION-LESS RECONFIGURABLE FILTER

The designed reconnection-less reconfigurable filter of the SISO type is depicted in Fig. 3. It consists only of two active elements, two capacitors and one resistor.
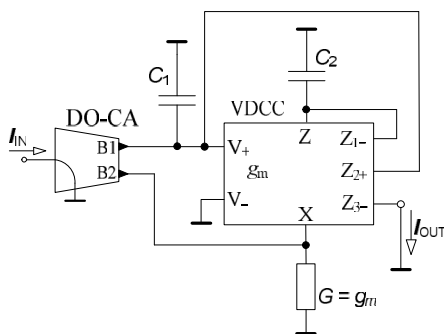

Fig. 3. Designed reconnection-less reconfigurable filter of the SISO type.

Ideal transfer function of the filter is

$$K(s) = \frac{I_{OUT}}{I_{IN}} = \frac{s^2 C_1 C_2 B_2 + G g_m B_1}{D(s)}, \quad (3)$$

where

$$D(s) = s^2 C_1 C_2 + s C_1 G + g_m G. \quad (4)$$

From (3) and (4) it is obvious that pole frequency and quality factor are described by:

$$f_P = \frac{1}{2\pi} \sqrt{\frac{g_m G}{C_1 C_2}}, \quad (5)$$

$$Q = \sqrt{\frac{g_m C_2}{C_1 G}}. \quad (6)$$

The following transfer functions are obtained by controlling of adjustable current gains (B_1, B_2):

1. LP response, if $B_2 = 0$ and $B_1 \neq 0$. Pass-band gain of the filter is adjusted by B_1. If $B_1 > 0$ ($B_1 < 0$), non-inverting (inverting) LP response is obtained and pass-band gain is directly proportional to B_1. If $|B_1| = 1$, unity gain in low-frequency pass band is achieved.

2. HP response, if $B_1 = 0$ and $B_2 \neq 0$. Pass-band gain of the filter is adjusted by B_2. If $B_2 > 0$ ($B_2 < 0$), non-inverting (inverting) HP response is obtained and pass-band gain is directly proportional to B_2. If $|B_2| = 1$, unity gain in high-frequency pass band is achieved.

3. BS response, if $B_1 = B_2 \neq 0$. Pass-band gain of the filter is adjusted by $B_1 = B_2$. If $B_1 = B_2 > 0$ ($B_1 = B_2 < 0$), non-inverting (inverting) BS response is obtained and pass-band gain is directly proportional to $B_1 = B_2$. If $|B_1 = B_2| = 1$, unity gain in both pass bands is achieved.

4. LPZ response, if $B_1 > B_2 > 0$ or $B_1 < B_2 < 0$. Also in this case, pass-band gain of the filter is adjusted by B_1, stop-band gain is dependent on B_2. Other features are similar to LP response.

5. HPZ response, if $B_2 > B_1 > 0$ or $B_2 < B_1 < 0$. Also in this case, pass-band gain of the filter is adjusted by B_2, stop-band gain is dependent on B_1. Other features are similar to HP response.

Note that $B_2 < 0$ and/or $B_1 < 0$ can be obtained with help of current mode multiplier, but this concept was not used in this paper. Moreover, if $g_m = G$ is varied, pole frequency is controlled without disturbing quality factor value (Q). Note that if only one additional current amplifier (with gain equal to B_3) is connected between Z_{1-} and node with C_2 capacitor, Q can be also controlled independently with respect to the pole frequency as obvious from the following equation. Denominator (3) changes to

$$D'(s) = g_m G + s C_1 B_3 G + s^2 C_1 C_2. \quad (7)$$

IV. NON-IDEAL AND SENSITIVITY ANALYSIS

When non-idealities of VDCC and DO-CA are taken into account, (4) changes to

$$D(s) = -g_m G n_{11} + s C_1 G n_{12} + s^2 C_1 C_2, \quad (8)$$

where n_{11} represents non-ideal current transfer from X terminal to Z_{1-} (theoretically equal to -1) and n_{12} represents

non-ideal current transfer from X terminal to Z_{2+} (theoretically equal to 1). It is an easy task to calculate non-ideal f_p and Q from (8) (therefore it is omitted in this paper). If n_{11} and/or n_{12} are not unity (in absolute value), it has observable impact only on pass-band gain of transfer functions (LP, LPZ and BS). Note that this can be compensated by tuning of B_1 gain in opposite direction, therefore it represents no problem in practice.

Another non-ideality that should be taken into account is that current and voltage gains in the terminal relationships of the active elements may not be valid at high frequencies due to the non-idealities of active components. In order to observe impact of these non-idealities, g_m, B_1, B_2, n_{11} and n_{12} were considered as frequency dependent, *i.e.* $g_m(s)$, $B_1(s)$, $B_2(s)$, $n_{11}(s)$ and $n_{12}(s)$

$$X(s) = \frac{x}{1 + s\tau_x}, \qquad (9)$$

where x should be replaced by g_m, B_1, B_2, n_{11} and n_{12} and $X(s)$ by $g_m(s)$, $B_1(s)$, $B_2(s)$, $n_{11}(s)$ and $n_{12}(s)$. Further, $\tau_x = 1/\omega_0$, where ω_0 stands for angular frequency forming the pole of the first-order function. A detailed analysis in Maple confirmed a well-known fact that in order to operate according to the theory up to frequency of, for instance, 10 MHz, $f_0 = \omega_0/2\pi$ should be at least 100 MHz (10 times higher).

Relative sensitivities of the important parameters of this filter (ω_p and Q) on its parameters can be obtained by routine calculations from (6) and are low because none of them overcome ± 0.5.

V. SIMULATION AND MEASUREMENT RESULTS

A filter from Fig. 3 was prepared to be measured. DO-CA active element was implemented by part of the UCC-N1B circuit [39] (forming DO-CF part from Fig. 1(b)) and two current-mode multipliers EL2082 (forming two CA from Fig. 1(b)). VDCC active element was implemented by another UCC-N1B circuit consisting of one CCII+/- (forming OTA part from Fig. 2(b)) and one UCC (Universal Current Conveyor), forming CCII part from Fig. 2(b). The filter parameters were: working capacitors $C_1 = 480$ pF, $C_2 = 240$ pF, starting value of quality factor $Q = 0.707$, theoretical pole frequency $f_P = \{117; 313; 919\}$ kHz obtained by $g_m = G = 1/R = \{0.25; 0.67; 1.96\}$ mS.

Implementation of current amplifier by EL2082 circuit limits possible range of current gain (B_1 or B_2) adjusting. As obvious from [40], current gain of EL2082 is adjusted by external DC voltage (V_G) and only $0 \le V_G < 5$ V is allowed. For range starting from 0 up to 2.5, current gain is approximately equal to control voltage in its numerical value ($B \approx V_G$).

In order to suppress the expected impact of parasitic capacitances in nodes of the circuit on printed-circuit board (PCB), the values of working capacitors were decreased to $C_1 = 470$ pF, $C_2 = 235$ pF.

The results for pure LP configuration ($B_1 = 1$, $B_2 = 0$) are depicted in Fig. 4 for three different pole frequencies. At frequencies above 10 MHz, the measured results are damaged by UCC-s features that have very limited operating bandwidth and are not designed to be operated at such high frequencies. However, the overall results are satisfactory in order to prove the concept. The results for pure HP configuration ($B_1 = 0$, $B_2 = 1$) and pure BS configuration ($B_1 = 1$, $B_2 = 1$) are depicted in Fig. 5 and Fig. 6 respectively. Real measured pole frequencies are only slightly different than was expected from simulation in all three cases, as summarized in Table III.
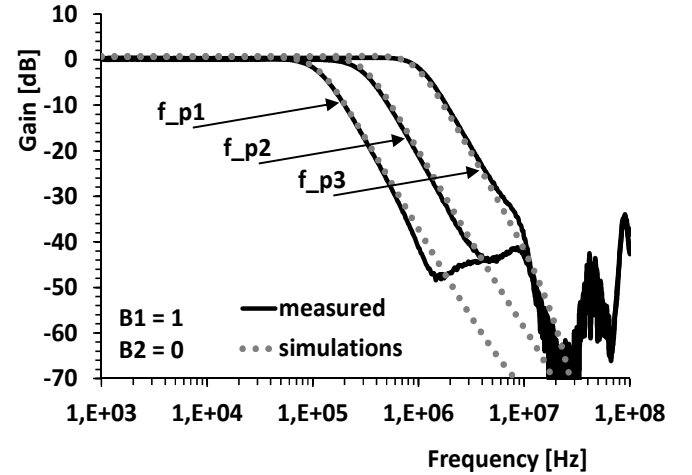

Fig. 4. Overall measurement results for LP response with unity pass-band gain while tuning pole frequency in comparison with simulations.
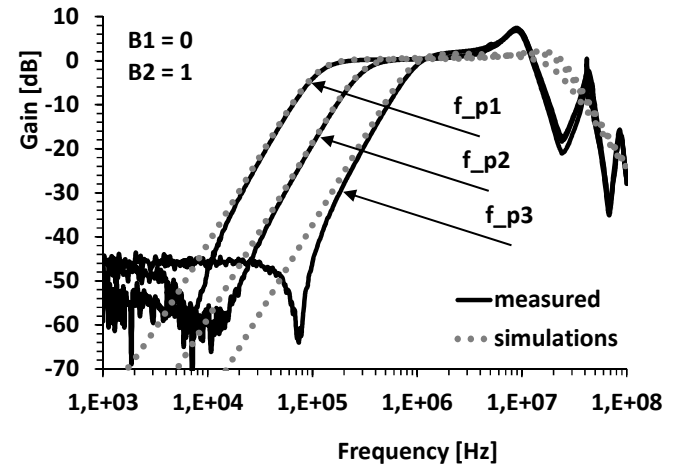

Fig. 5. Overall measurement results for HP response with unity pass-band gain while tuning pole frequency in comparison with simulations.
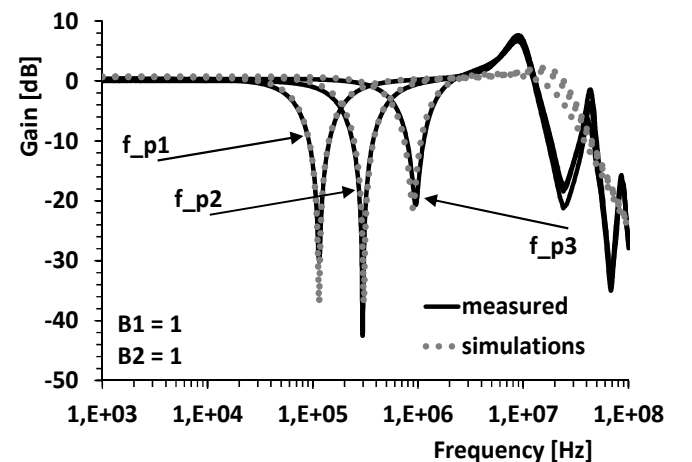

Fig. 6. Overall measurement results for BS response with unity pass-band gain while tuning pole frequency in comparison with simulations.

Figure 7 and Fig. 8 provide the results for pass-band gain tuning in case of both LP and HP filters respectively. Pass-band gain is directly proportional to the B gain as expected from (3) and obtained gains are summarized in Table IV.

TABLE III. COMPARISON OF OBTAINED VALUES IN THE CASE OF INDEPENDENT TUNING OF THE POLE FREQUENCY (FIG. 4, FIG. 5, FIG. 6).

Identification	Theoretical f_p [kHz]	Simulated f_p [kHz] of LP response	Measured f_p [kHz] of LP response	Simulated f_p [kHz] of HP response	Measured f_p [kHz] of HP response	Simulated f_p [kHz] of BS response	Measured f_p [kHz] of BS response
f_p1	117	113	113	114	114	117	115
f_p2	313	307	303	294	290	306	297
f_p3	919	987	1070	880	939	892	959

TABLE IV. COMPARISON OF OBTAINED VALUES IN THE CASE OF INDEPENDENT TUNING OF THE PASS-BAND GAIN (FIG. 7, FIG. 8).

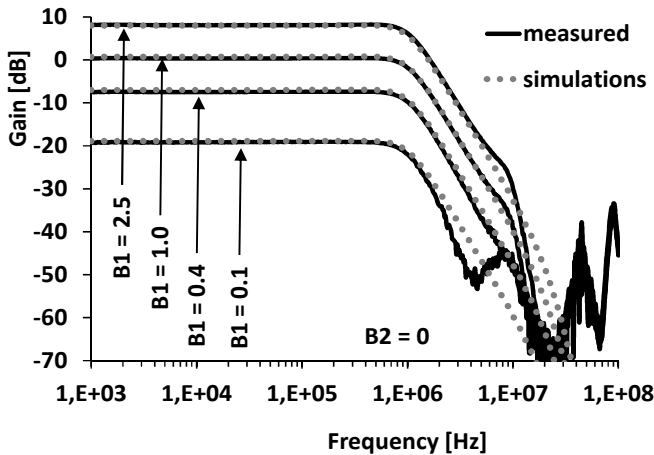| B_1 [-] or B_2 [-]	Theoretical pass-band gain [dB]	Simulated pass-band gain of LP response [dB]	Measured pass-band gain of LP response [dB]	Simulated pass-band gain of HP response [dB]	Measured pass-band gain of HP response [dB]
0.1	-20.0	-18.9	-19.2	-19.0	-18.9
0.4	-8.0	-7.0	-7.4	-7.1	-7.2
1.0	0.0	-0.7	0.3	0.6	0.5
2.5	8.0	8.0	8.1	7.9	8.1

Fig. 7. Overall measurement results for LP response with $f_{P\_THEOR} = 919$ kHz and while tuning pass-band gain by B_1 in comparison with simulations.
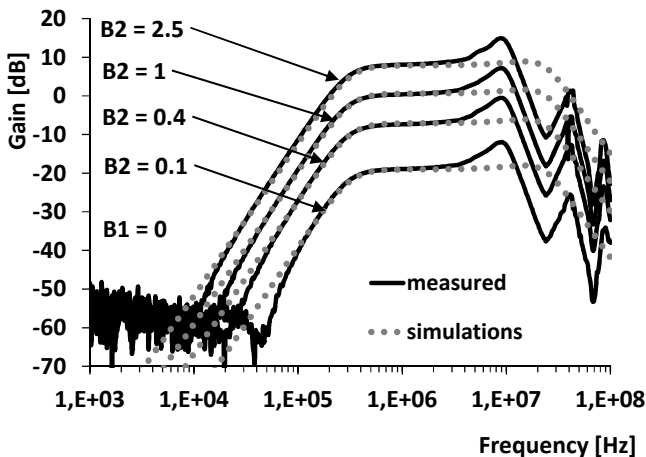

Fig. 8. Overall measurement results for HP response with $f_{P\_THEOR} = 313$ kHz and while tuning pass-band gain by B_2 in comparison with simulations.

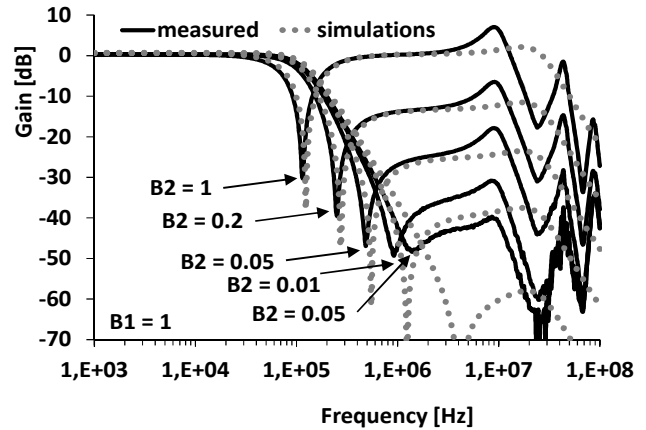Figure 9 and Fig. 10 include the measured characteristics for stop-band gain tuning.

Fig. 9. Overall measurement results for LPZ response with $f_{P\_THEOR} = 117$ kHz with unity pass-band gain while tuning stop-band gain (attenuation) by B_2 in comparison with simulations.
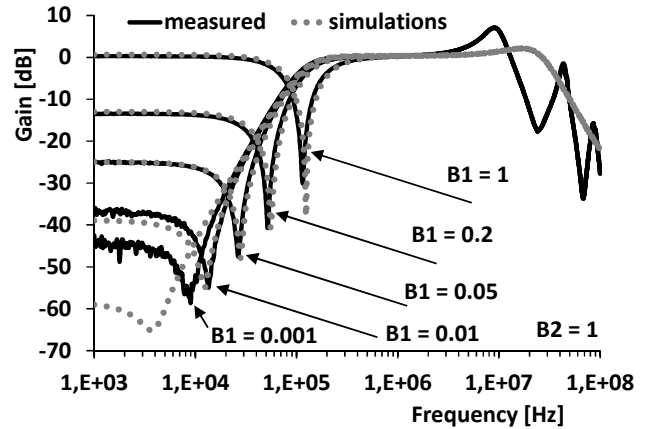

Fig. 10. Overall measurement results for HPZ response ($f_P = 117$ kHz) with unity pass-band gain while tuning stop-band gain (attenuation) by B_1 in comparison with simulations.

It is obvious from the graphs that when B is tuned from 0 to 1 (B_2 in case of LP and B_1 in case of HP), stop-band attenuation decreases (gain grows) and when $B_1 = B_2 = 1$, both functions change into BS. If $B = 0.001$ (B_2 in case of LP and B_1 in case of HP), stop-band gain should be theoretically -60 dB ($K = 20 \cdot log_{10} 0.001 = -60$ dB), however it is not possible to achieve because real implementation of filter on PCB using the above mentioned active elements is limited by -50 dB.

VI. CONCLUSIONS

The designed reconnection-less reconfigurable filter providing LP, HP, LPZ, HPZ and BS transfer function without any change or switching in the topology was presented. Its measured features are influenced by real parameters of the active elements that were used. Also real PCB and its design has direct impact on the measured characteristics. In our case, the highest attenuation was limited to 50 dB and the results are in good agreement with theory up to frequency of 10 MHz, tuning range was from 115 kHz to 959 kHz (in case of BS response). Nevertheless, we consider these results as very good. Of course, the test with real devices (UCC and EL2082) could be substituted or supplemented by the simulations with CMOS transistor

structures. Behavioural modelling (as it was used and presented in this paper) is a very useful way how to verify the design. If the filter prepared with help of universal active elements, employed as a part of the behavioural model, works well, possible non-universal transistor structure would have much better features.

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