http://dx.doi.org/10.5755/j01.eee.21.3.10205 ELEKTRONIKA IR ELEKTROTECHNIKA, ISSN 1392-1215, VOL. 21, NO. 3, 2015

Reconnection-less OTA-based Biquad Filter with Electronically Reconfigurable Transfers

Roman Sotner[1], Jiri Petrzela[1], Jan Jerabek[2], Tomas Dostal[3]
[1]*Department of Radio Electronics, Faculty of Electrical Engineering and Communication, Brno University of Technology, Technicka 12, Brno, 616 00, Czech Republic*
[2]*Department of Telecommunications, Faculty of Electrical Engineering and Communication, Brno University of Technology, Technicka 12, Brno, 616 00, Czech Republic*
[3]*Department of Electrical Engineering and Computer Science, College of Polytechnics Jihlava, Tolsteho 16, Jihlava 586 01, Czech Republic*
sotner@feec.vutbr.cz

***Abstract*—This paper deals with operational transconductance amplifiers (OTAs) -based active voltage-mode biquad filter with electronically reconfigurable transfer functions. Due to utilization of the very favourable active devices, this design is ready for immediate CMOS design. Presented filtering solution contains four active elements where each of them is directly used for reconnection-less change of transfer function or modification including electronic control of quality factor and tuning. The filter offers availability of all-pass, high-pass, band-pass, band-reject transfer response and special transfers as high-pass with zero and low-pass with zero. Measurement results based on utilization of diamond transistors confirmed expected behaviour of the circuit.**

***Index Terms*—Active filter, biquad, circuit synthesis, electronic control, multifunctionality, operational transconductance amplifier, OTA, reconfiguration, reconnection-less.**

I. INTRODUCTION

Reconfigurability is a very useful feature of the analog filters since it represents the possibility of immediate change of type of the frequency response of two-port structure without changing internal topology or a position of input or output port. Its importance even grows in the case of full on-chip implementation of complex electronic system comprising one or several filtering stages where continuous and smooth variability of their behaviour in the frequency domain by the external dc sources became highly necessary. It is not only trend in microelectronic design but also in practical applications focused on effective high-speed signal processing. Many recent scientific publications solve simplified problem since these multifunctional second-order filters have particular transfer functions available between different nodes of the network. Many methods useful for design of “standard” multifunctional filtering functions have been published [1]–[10]. Signal flow graph synthesis [1], [2] is very useful and illuminating for these purposes. Perfect explanations of synthesis of multifunctional filters based on multiple-loop structures of integrators were given for example in [3]–[10]. Many of them are based on operational transconductance amplifiers (OTAs) [11], [12]. Unfortunately, reconnection-less reconfiguration of the filter response is not discussed in basic literature [3]–[10]. However, our aim is to avoid using the filtering structures where change of required transfer function is obtained by manual reconnection of output (so called single-input and multiple-output types - SIMO) or input (multiple-input and single-output - MISO) terminal.

Research for these structures has real importance due to complicated change of filter type in realization on chip. We can use these filters beneficially when some undesirable frequency components or distortions/noises occur in processed signal and we have to modify (tune) or change type of the transfer function (magnitude response) immediately. Physical reconnection of appropriate output/input of the filter by switches is traditional way how to provide this intentional change. However, it means additional problems (additional chip area for control logic and switches, power consumption, additional distortion from switching mechanisms – discontinuous operation – undesirable frequency components).

Several works dealing with this topic were already published in recent years. However, many of them are focused only on the first-order filters [13]–[15].

Second-order solution of the reconfigurable filtering structure was firstly reported in [16], where circuit based on two active devices providing all-pass (AP) and band-reject (BR) response was introduced. Continuous change of the AP to BR response is possible together with electronic tuning. However, quality factor of proposed solution is very low and limited and other transfer functions are not available.

Manuscript received November 3, 2014; accepted February 25, 2015.
Research described in this paper was financed by Czech Ministry of Education in frame of National Sustainability Program under grant LO1401. For research, infrastructure of the SIX Center was used. Research described in the paper was supported by Czech Science Foundation projects under No. 14-24186P. Grant No. FEKT-S-14-2281 also supported this research. The support of the project CZ.1.07/2.3.00/20.0007 WICOMT, financed from the operational program Education for competitiveness, is gratefully acknowledged.

Problems of low quality factor and other problems in tuning are solved in [17]. However, circuit is very complicated (at least five active elements) and provides again only AP and BR responses. Multiple-loop integrator structure [3]–[10] was used for synthesis in [17].

In this paper, we present structure of the reconnection-less reconfigurable biquad (second-order) filter based on four OTAs, that allows more types of transfer characteristic and more possibilities of electronic control than previously reported solutions. The paper has following structure: Section I gives detailed introduction to this area and reasons for this research, Section II deals with method of synthesis, Section III introduces behavioural model and experimental results and Section IV summarizes main findings of synthesis and main features of proposed circuit.

II. METHOD OF SYNTHESIS

We used matrix method of the unknown nodal voltages (MUNV) [18]–[21] to obtain discussed circuit. MUNV is a widely used approach dedicated to symbolical analysis of the linearized circuits. Its rules directly result from first Kirchhoff's laws [21], the individual equations represent current balance at the particular nodes. To preserve a system of the linear non-homogenous equations solvable, these nodes must be independent on each other. This property is indicated by regularity of the square admittance matrix **Y**.

The proposed design procedure starts with a defined transfer function which must be as general as possible in order to alternate transfer zeroes and poles independently on each other. Design itself starts with defined form of the voltage transfer function which should correspond to equation

$$K(s)=\frac{N(s)}{D(s)}=\frac{a_2s^2+a_1s+a_0}{b_2s^2+b_1s+b_0}=K_0\frac{s^2+\dfrac{\mu_Z}{Q_Z}s+\mu_Z^2}{s^2+\dfrac{\mu_P}{Q_P}s+\mu_P^2},\quad (1)$$

where K_0 is pass-band gain and $Q_Z > 0$ and $Q_P > 0$ are zero and pole quality factors. Note that term (1) contains four parameters (μ_Z, Q_Z, μ_P, Q_P) and this compact set covers all possible frequency responses. Let us imagine a complex plane ready to be filled by transfer zeroes and poles. To preserve some degree of stability the poles should be real or a pair of complex conjugated values but always placed in the left half-plane of the complex plane. This implies that polynomial $D(s)$ have only positive coefficients with the possibility to minimize linear part b_1 down to a reasonable value. In order to obtain all-pass filter we must be able to set a negative value of coefficient $a_1 = -b_1$. For ideal high-pass filter it is necessary to achieve equality $a_0 = a_1 = 0$ and for ideal low-pass we must be able to set $a_1 = a_2 = 0$. Typical configuration of band-pass (BP) filter is $a_0 = a_2 = 0$. Situation is a little bit confusing in the case of band-reject filter because it is bounded to the same circles of constant angular frequency for zeroes and poles, i.e. to the relations $\mu_Z = \mu_P$ and $Q_Z > Q_P$. In some specific situations, the so-called low-pass filter with zero (LPZ) and high-pass filter with zero (HPZ) is required. First of them has $\mu_Z > \mu_P$ and $Q_Z > Q_P$ and the second case $\mu_Z < \mu_P$ and $Q_Z > Q_P$.

Thus we will focus on circuits which contain just three independent nodes oriented to the overall ground reference. Marking input and output node as first and third unknown voltage respectively, Cramer rule [18] gives following formula for the voltage transfer function:

$$\mathbf{V}=\begin{pmatrix}V_1 & V_2 & V_3\end{pmatrix}^T,\quad (2)$$

$$K(s)=\frac{V_{OUT}}{V_{INP}}=\frac{V_3}{V_1}=\frac{\Delta_{1,3}}{\Delta_{1,1}},\quad (3)$$

where **V** is a column vector of the unknown voltages and $\Delta_{i,j}$ is a sub-determinant of admittance matrix after omitting i-th row and j-th column. Input impedance can be established as

$$Z(s)=\frac{V_{INP}}{I_{INP}}=\frac{1}{I_{INP}}\times\frac{I_{in}(-1)^{1+1}\Delta_{1,1}}{\Delta}=\frac{\Delta_{1,1}}{\Delta},\quad (4)$$

where I_{in} is arbitrary input current and Δ is full determinant of admittance matrix. We used described approach in system of three nodal voltages where OTAs are used. The basic OTA has differential voltage input and single current output to provide: $(V_+ - V_-).g_m = I_{out}$, where g_m ($A.V^{-1}$) is electronically adjustable transconductance [11], [12]. Our intended and specific transfer function has form

$$K(s)=\frac{s^2-\dfrac{g_{m3}}{C_1}s+\dfrac{g_{m2}g_{m4}}{C_1C_2}}{s^2+\dfrac{g_{m1}}{C_1}s+\dfrac{g_{m1}g_{m2}}{C_1C_2}}.\quad (5)$$

Supposing presence of floating capacitor between node 1 and 3 (C_1) and grounded capacitor (C_2) in node 2, we have system of linear equations in form

$$\begin{pmatrix}sC_1 & 0 & -sC_1\\ 0 & sC_2 & 0\\ -sC_1 & 0 & sC_1\end{pmatrix}\times\begin{pmatrix}V_1\\ V_2\\ V_3\end{pmatrix}=\begin{pmatrix}I_1\\ 0\\ 0\end{pmatrix}.\quad (6)$$

Fixing this degree of freedom, we have to select correct positions in admittance matrix where specific transconductances (g_{m1} to g_{m4}) are located. In accordance to (3), it leads to sub-determinants:

$$\begin{aligned}\Delta_{1,1}&=s^2C_1C_2+g_{m1}sC_2+g_{m1}g_{m2}=\\&=sC_2(sC_1+g_{m1})-g_{m1}(-g_{m2}),\end{aligned}\quad (7)$$

$$\begin{aligned}\Delta_{1,3}&=s^2C_1C_2-g_{m3}sC_2+g_{m2}g_{m4}=\\&=(-g_{m4})(-g_{m2})-(-sC_1+g_{m3})sC_2.\end{aligned}\quad (8)$$

We can really see where particular transconductances have to be located in final admittance matrix:

$$\begin{pmatrix}sC_1 & 0 & -sC_1\\ -g_{m4} & sC_2 & g_{m1}\\ -sC_1+g_{m3} & -g_{m2} & sC_1+g_{m1}\end{pmatrix}\times\begin{pmatrix}V_1\\ V_2\\ V_3\end{pmatrix}=\begin{pmatrix}I_1\\ 0\\ 0\end{pmatrix},\quad (9)$$

and prepare schematic of the solution (Fig. 1). Such biquad filter has significant degree of reconfigurability because g_{m1} influence both Q_P and μ_P while g_{m2} affects zero and pole frequency simultaneously. The relation between frequency radius for transfer zeroes and poles are uniquely controlled through g_{m4} since $\mu_P/\mu_Z = (g_{m4}/g_{m1})^{1/2}$ while Q_Z is adjusted only by g_{m3} change. This filter realizes these transfers:

1. AP response for $g_{m1} = g_{m2} = g_{m3} = g_{m4}$,
2. BR response for $g_{m1} = g_{m2} = g_{m4}$, $g_{m3} \to 0$,
3. high-pass (HP) response for $g_{m1} = g_{m2}$, $g_{m3} = g_{m4} \to 0$,
4. iBP (inverting) response for $g_{m1} = g_{m2}$, $g_{m4} \to 0$,
5. HPZ (derived from BR) and for $g_{m1} > g_{m4}$, $g_{m3} \to 0$,
6. LPZ (derived from BR) for $g_{m1} < g_{m4}$, $g_{m3} \to 0$.

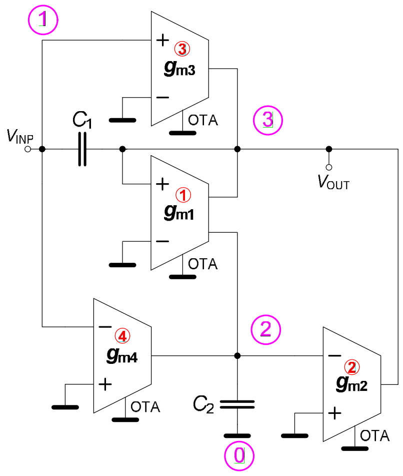


Fig. 1. Synthetized reconfigurable biquad filter based on OTAs.

Note that iBP response is not pure BP (with symmetrical sidebands – slope 20 dB/dec) because it has finite attenuation at the high-frequency stop-band (intentionally – resulting from design equations).

III. Behavioural Model and Experimental Results

Five single-output diamond transistors OPA860 [22] were used for implementation of transconductance amplifiers as evident from Fig. 2. Supply voltage became $V_{DD} = V_{SS} = 5$ V and both working capacitors have value $C_1 = C_2 = 470$ pF. Transconductances are given approximately by the so-called degradation resistors $R_{\mathrm{deg}} = 1/g_m$ which were set to get $g_{m1} = g_{m2} = 0.91$ mS. This design suppose ideal pole frequency located at $f_p = 339$ kHz and quality factor $Q = 1$. The rest of parameters (g_{m3} and g_{m4}) were changed in order to show the overall filter performances in frequency domain. Diamond transistors are beneficial due to high-frequency features and variability. Unfortunately, wide-range control of their basic parameter is possible only by changing value of the passive resistor.

Results of the measurements are given in Fig. 3 up to Fig. 10. Figure 3 shows configuration of the filter as HP response ($g_{m3} = g_{m4} = 0$ mS), see (5) for clarity. BR response is available for $g_{m3} = 0$ mS simultaneously with $g_{m1,2,4} = 0.91$ mS and is shown in Fig. 4. Attenuation control of the minimal gain of the BR response is possible as we can see in Fig. 5. Maximal measured available attenuation 40 dB was obtained for $g_{m1,2,4} = 0.91$ mS and $g_{m3} = 0.1$ mS (Fig. 5).

Example of attenuation control is shown in Fig. 6, where attenuation is set to be 30 dB were obtained for $g_{m3} = 0.16$ mS. iBP response (with gain 20 dB) for $g_{m3} = 13.7$ mS and $g_{m4} = 0.1$ mS is shown in Fig. 7. AP response is shown in Fig. 8 for $g_{m1,2,3,4} = 0.91$ mS.

Presented filter example is also easily configurable as HP or LP response both with transfer zero. Frequency responses for the first case (HPZ) is achievable for $g_{m3} = 0$ mS, $g_{m4} = 0.1$ mS and is depicted in Fig. 9. Particular results for the second case (LPZ) is reached by setting $g_{m3} = 0$ mS, $g_{m4} = 7.1$ mS and is demonstrated in Fig. 10. Of course, electronic tuning (of pole frequency) is possible, but due to limited space, we restricted our analyses to presentation of reconnection-less filter reconfiguration only.

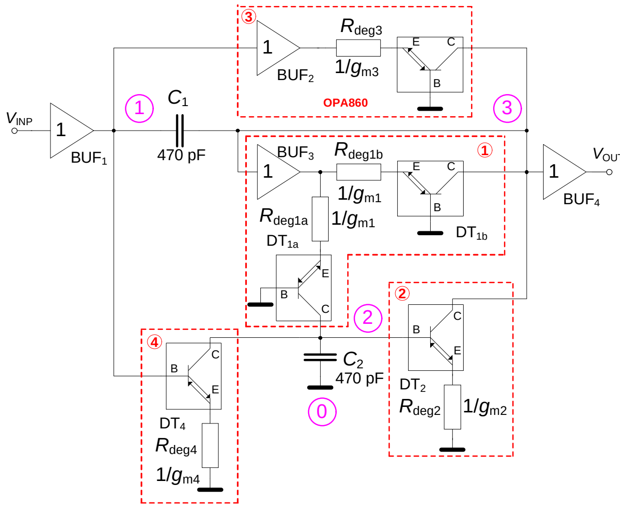
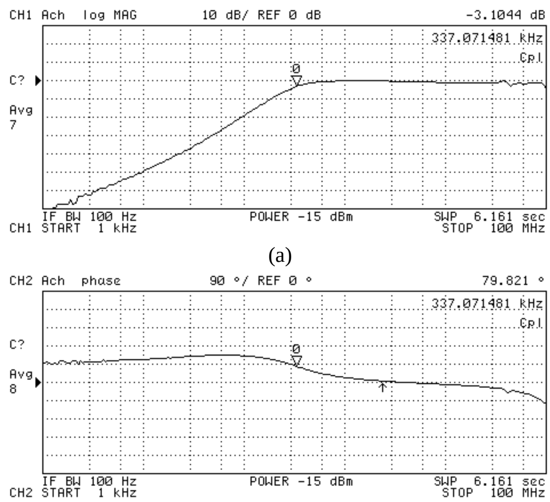


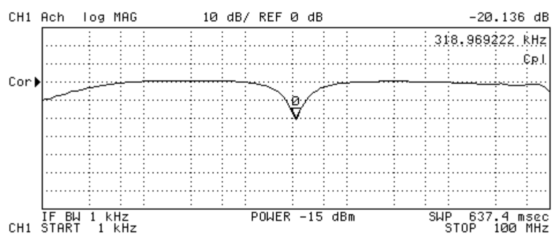
Fig. 2. Behavioural model of the biquadratic filter based on diamond transistors and voltage buffers (available in the single package).



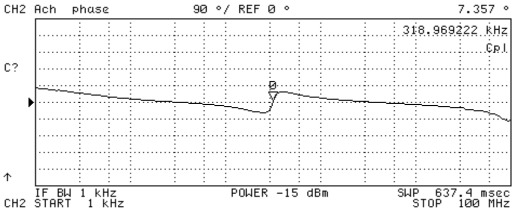
(a)

(b)

Fig. 3. High-pass configuration ($g_{m3} = g_{m4} = 0$ mS): a) magnitude response; b) phase response.

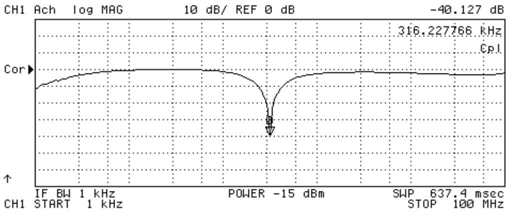


(a)

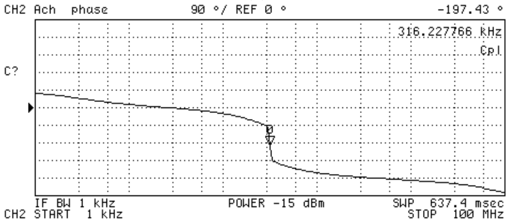


(b)

Fig. 4. Band-reject configuration (g_{m3} = 0 mS, $g_{m1,2,4}$ = 0.91 mS): a) magnitude response; b) phase response.

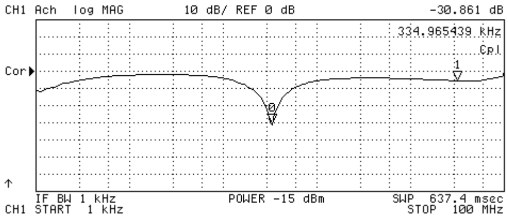


(a)

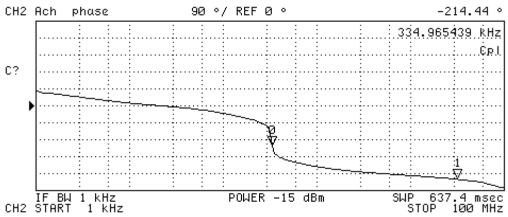


(b)

Fig. 5. Band-reject configuration (g_{m3} = 0.1mS, $g_{m1,2,4}$ = 0.91 mS) employing 40 dB attenuation: a) magnitude response; b) phase response.

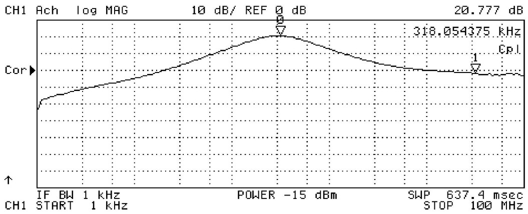


(a)

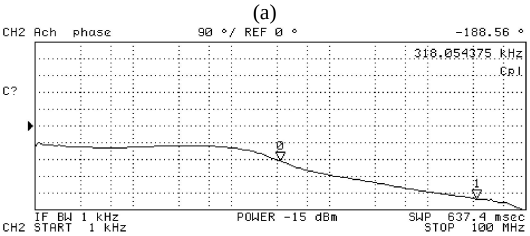


(b)

Fig. 6. Band-reject configuration (g_{m3} = 0.16 mS, $g_{m1,2,4}$ = 0.91 mS) employing 30 dB attenuation: a) magnitude response; b) phase response.

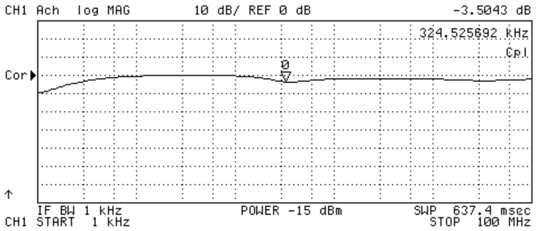


(a)

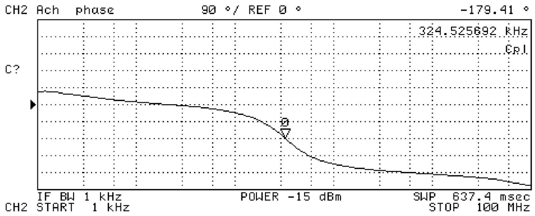


(b)

Fig. 7. Inverting band-pass configuration (g_{m3} = 13.7 mS, g_{m4} = 0.16 mS): a) magnitude response; b) phase response.

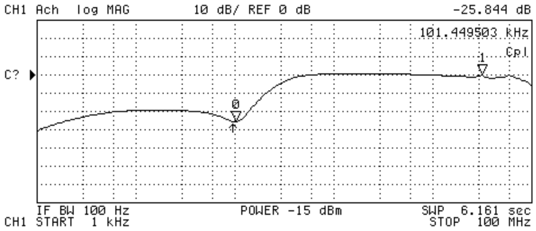


(a)

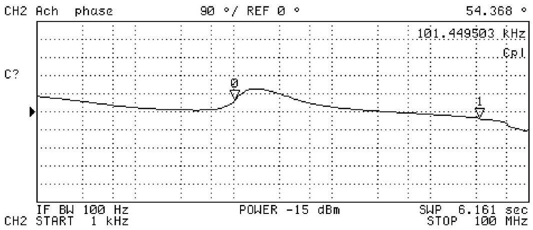


(b)

Fig. 8. All-pass configuration ($g_{m1,2,3,4}$ = 0.91 mS): a) magnitude response; b) phase response.

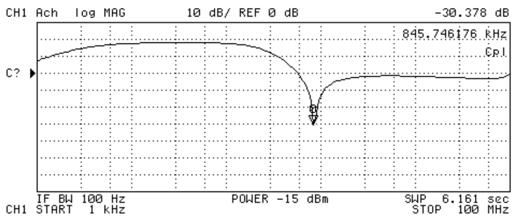


(a)

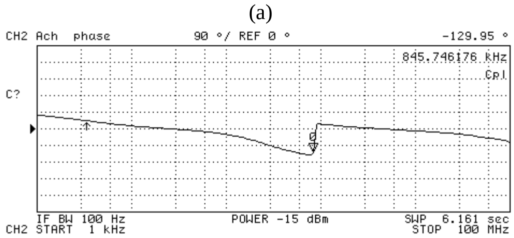


(b)

Fig. 9. High-pass with transfer zero configuration (g_{m3} = 0 mS, g_{m4} = 0.1 mS): a) magnitude response; b) phase response.



(a)



(b)

Fig. 10. Low-pass with transfer zero configuration (g_{m3} = 0 mS, g_{m4} = 7.1 mS): a) magnitude response; b) phase response.

Measurement results given above were obtained by network-vector/spectrum analyser HP4395A with input power level −15 dBm, i.e. input voltage 40 mV$_{ef}$ on 50 Ω. Magnitude drop in some results at low frequencies f_{-3dB} = 2.1 kHz is given by coupling capacitance (foil 1 μF) at the output terminal of the filter to avoid DC component dangerous for input of the analyser.

IV. CONCLUSIONS

The presented second-order filtering stage has four parameters which should be externally adjusted independently of each other. Therefore, the voltage transfer function allows setting zeroes and poles independently on each other but still being part of the group of the complex numbers. The transfer poles should be placed firstly with the possibility to achieve the complex conjugated values to get higher quality factors than 0.5; otherwise a structure becomes useless. To preserve a certain degree of stability, the migration of poles should be strictly limited in the left half-plane of the complex plane. The transfer zeros should be complex numbers as well as it is necessary to be able to move them from left to right half-plane of complex space and back in order to obtain all-pass frequency response. Several scenarios for moving zeroes can be arranged. If such movement is along a circle of the constant significant angular frequency we are able to realize continuous change of frequency response from band-reject to all-pass. If couple of the transfer zeroes are created at the origin of the imaginary axis and move towards high angular frequency we change filter nature from high-pass through modified high-pass and band-reject up to low-pass. The complexity of the directions for transfer zeros migration indicates that at least four active elements are needed for practical design of this type of the reconfigurable filter. MUNV is also very suitable for study of real parasitic influences in structures (all undesired effects – parasitic resistances and capacitances - can be simply and directly reflected to the matrix **Y**). Measurement results confirm availability of all possible transfer functions (AP, HP, iBP, BR, LPZ, HPZ) in structure based on one single input and single output from frequencies of several kHz to several tens of MHz.

REFERENCES

[1] S. J. Mason, "Feedback Theory: Further properties of Signal Flow Graphs", in *Proc. IRE*, vol. 44, no. 7, pp. 920–926, 1956. [Online]. Available: http://dx.doi.org/10.1109/jrproc.1956.275147

[2] C. J. Coates, "Flow-graph solution of linear algebraic equations", *IRE Trans. Circuit Theory*, vol. 6, no. 2, pp. 170–187, 1959. [Online]. Available: http://dx.doi.org/10.1109/TCT.1959.1086537

[3] T. Dostal, "All-pass filter in current mode", *Radioenginnering*, vol. 14, no. 3, pp. 48–53, 2005.

[4] C. M. Chang, C. L. Hou, W. Y. Chung, J. W. Horng, C. K. Tu, "Analytical synthesis of high-order single-ended-input OTA-grounded C all-pass and band-reject filter structures", *IEEE Trans. Circuits and Systems: Regular Papers*, vol. 53, no. 3, pp. 489–498, 2006. [Online]. Available: http://dx.doi.org/10.1109/TCSI.2005.859057

[5] C. M. Chang, B. M. Al-Hashimi, "Analytical synthesis of current-mode high-order OTA-C filters", *IEEE Trans. Circuits and Systems: Regular Papers*, vol. 50, no. 9, pp. 1188–1192, 2003. [Online]. Available: http://dx.doi.org/10.1109/TCSI.2003.816327

[6] T. Tsukutani, M. Higashimura, M. Ishida, S. Tsuiki, Y. Fukui, "A general class of current-mode high-order OTA-C filters", *International Journal of Electronics*, vol. 81, no. 6, pp. 663–669, 1996. [Online]. Available: http://dx.doi.org/10.1080/002072196136355

[7] Y. Sun, J. K. Fidler, "Some design methods of OTA-C and CCII-RC filters", in *Proc. IEE Colloquium on Digital and Analogue Filters and Filtering*, London, 1993, pp. 1–8.

[8] Y. Sun, J. K. Fidler, "Current-mode multiple-loop feedback filters using dual output OTAs and grounded capacitors", *International Journal of Circuit Theory and Applications*, vol. 25, no. 2, pp. 69–80, 1997. [Online]. Available: http://dx.doi.org/10.1002/(SICI)1097-007X(199703/04)25:2<69::AID-CTA950>3.0.CO;2-9

[9] Y. Sun, J. K. Fidler, "Current-mode OTA-C realization of arbitrary filter characteristics", *Electronics Letters*, vol. 32, no. 13, pp. 1181–1182, 1996. [Online]. Available: http://dx.doi.org/10.1049/el:19960807

[10] Y. Sun, J. K. Fidler, "Structure generation of current-mode two integrator dual output-OTA grounded capacitor filters", *IEEE Trans. Circuits and Systems II: Analog and Digital Signal Processing*, vol. 43, no. 9, pp. 659–663, 1996. [Online]. Available: http://dx.doi.org/10.1109/82.536762

[11] R. L. Geiger, E. Sanchez-Sinencio, "Active filter design using operational transconductance amplifiers: a tutorial", *IEEE Circuits and Devices Magazine*, vol. 1, pp. 20–32, 1985. [Online]. Available: http://dx.doi.org/10.1109/MCD.1985.6311946

[12] D. Biolek, R. Senani, V. Biolkova, Z. Kolka, "Active elements for analog signal processing: Classification, Review and New Proposals", *Radioengineering*, vol. 17, no. 4, pp. 15–32, 2008.

[13] J. Petrzela, R. Sotner, "Systematic design procedure towards reconfigurable first- order filters", in *Proc. 24th Int. Conf. Radioelektronika 2014*, Bratislava, 2014, pp. 1–4. [Online]. Available: http://dx.doi.org/10.1109/radioelek.2014.6828462

[14] R. Sotner, J. Jerabek, N. Herencsar, R. Prokop, K. Vrba, T. Dostal, "Resistor-less first-order filter design with electronical reconfiguration of its transfer function", *in Proc. 24th Int. Conf. Radioelektronika 2014*, Bratislava, 2014, pp. 63–66. [Online]. Available: http://dx.doi.org/10.1109/radioelek.2014.6828417

[15] R. Sotner, N. Herencsar, J. Jerabek, R. Prokop, A. Kartci, T. Dostal, K. Vrba, "Z-copy controlled-gain voltage differencing current conveyor: advanced possibilities in direct electronic control of first-order filter", *Elektronika Ir Elektrotechnika*, vol. 20, no. 6, pp. 77–83, 2014. [Online]. Available: http://dx.doi.org/10.5755/j01.eee.20.6.7272

[16] R. Sotner, J. Jerabek, B. Sevcik, T. Dostal, K. Vrba, "Novel solution of notch/all- pass filter with special electronic adjusting of attenuation in the stop band", *Elektronika Ir Elektrotechnika*, vol. 17, no. 7, pp. 37–42, 2011. [Online]. Available: http://dx.doi.org/10.5755/j01.eee.113.7.609

[17] R. Sotner, J. Jerabek, J. Petrzela, K. Vrba, T. Dostal, "Design of fully adjustable solution of band-reject/all-pass filter transfer function using signal flow graph approach", in *Proc. 24th Int. Conf. Radioelektronika 2014*, Bratislava, 2014, pp. 67–70. [Online]. Available: http://dx.doi.org/10.1109/radioelek.2014.6828418

[18] A. Basak, *Analogue electronic circuits and systems*. Cambridge University Press, New York, USA, 1991, p. 376. [Online]. Available: http://dx.doi.org/10.1017/CBO9781139168069

[19] P. R. Gray, P. J. Hurst, S. H. Lewis, R. G. Meyer, *Analysis and design of analog integrated circuits*. John Wiley and Sons, Inc: USA, 2009, p. 896.

[20] R. Raut, M. N. S. Swamy, *Modern Analog Filter Analysis and Design: A practical approach*. Willey-VCH Verlag GmbH and Co. KGaA, Germany, Weinheim, 2010, p. 35. [Online]. Available: http://dx.doi.org/10.1002/9783527631506

[21] W. Chen, *The Circuits and Filters Handbook*. Florida: CRC Press, Boca Raton, 2009, p. 3364.

[22] OPA860: Wide Bandwidth Operational Transconductance Amplifier and Buffer, Texas Instruments. 2005, last modified 8/2008.