ELECTRONICALLY TUNABLE HIGH–ORDER HIGHPASS FILTERS WITH MINIMUM OF COMPONENTS

Michal Olšák — Kamil Vrba — Lukáš Matějíček *

A new method for realization of an electronically tunable high-order admittance function that can be used for the synthesis of non-cascade active frequency filters is introduced. Operational transconductance amplifiers (OTA) are employed as active elements. Electronically controlled high-order highpass filters are presented. As an example, two types of fourth-order Butterworth highpass filters are designed, simulated (using PSPICE) and measured.

K e y w o r d s: Admittance function, highpass filter, operational transconductance amplifier (OTA)

1 INTRODUCTION

Active filters based on transconductance amplifiers with a single- or balanced-current output (OTA or BOTA) [1]–[6] have recently received considerable attention. The transconductance gain g_m of the OTA and BOTA can be continuously controlled by an external voltage or current source over several decades, which lends electronic tunability to circuit parameters. It may be emphasized that electronic tunability becomes very important when the circuit is in a variety of design specifications and in the integrated form. Many authors often do not make use of this possibility. A lot of transconductance amplifiers have appeared on the market in the past few years, for example the CA3080 (Intersil), LM13600 and LM13700 (National Semiconductor), LT1228 (Linear Technology), NE5517 (Philips Semiconductors), MAX435 or MAX436 (MAXIM) circuits. The interest of technical community in these elements is evidenced by the offer of manufacturers.

It is possible to show that one-port elements with high-order immittance are suitable for non-cascade active frequency filter synthesis. Two possible realizations of such OTA based one-port elements and also their possible utilization for the tunable fourth-order highpass filter synthesis will be shown below. Results of the proposed filters simulation, on the level of SPICE models, and measurement results on function samples will be shown as well.

2 ADMITTANCE FUNCTION REALIZATION

Suppose that we have to design an analog Nth-order highpass filter using the non-cascade design procedure. An active frequency filter working in the voltage mode can be realized as a voltage divider or in the current mode as a current divider containing high-order admittance elements, as shown in Fig. 1 and Fig. 2. The transfer function in both cases is given by the following relation:

$$T(p) = \frac{V_{\rm out}}{V_{\rm in}} = \frac{I_{\rm out}}{I_{\rm in}} = \frac{Y_1(p)}{Y_1(p) + Y^{(n)}(p)}\,. \qquad (1)$$

The Nth-order admittance function $Y^{(n)}(p)$ can be expressed as

$$\begin{aligned} &Y^{(n)}(p) = Y^0(p) + Y^1(p) + \ldots + Y^n(p), \text{ where} \\ &Y^0(p) = \frac{1}{E_0} = \frac{1}{R}\,, \qquad Y^1(p) = \frac{1}{pE_1} = \frac{1}{pL}\,, \\ &Y^2(p) = \frac{1}{p^2E_2} = \frac{1}{p^2E}\,, \qquad etc. \end{aligned}$$

Note that the one-port element with 2nd-order admittance is known as a FDNR E-element [7].

However, the realization of the above individual high-order immittance elements is not economical. A general procedure for the realization of immittance function employing a relatively small number of active elements was published in [8], [9] and [10]. In the following text we show a similar design procedure for a one-port with admittance function $Y^{(n)}(p)$ on the OTA base.

Designing an nth-order highpass filter using eq. (1) we have two possible solutions: either $Y_1(p) = pC_0$ and

$$Y^{(n)}(p) = \frac{1}{E_0} + \frac{1}{pE_1} + \frac{1}{p^2E_2} + \cdots + \frac{1}{p^{N-1}E_{N-1}} \qquad (2)$$

or $Y_1(p) = G_0$ and

$$Y^{(n)}(p) = \frac{1}{pE_1} + \frac{1}{p^2E_2} + \frac{1}{p^3E_3} + \cdots + \frac{1}{p^NE_N}\,. \qquad (3)$$

The main problem is now the synthesis of admittance function $Y^{(n)}(p)$.

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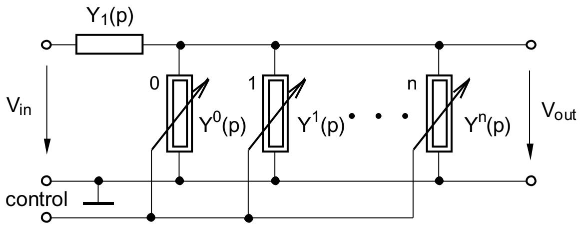

Fig. 1. Frequency filter with high-order admittance elements in the voltage mode

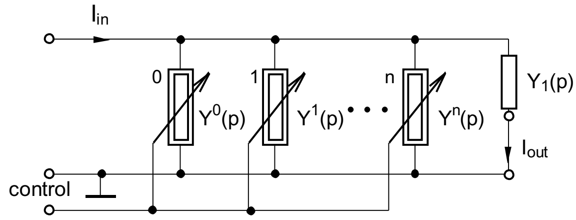

Fig. 2. Frequency filter with high-order admittance elements in the current mode

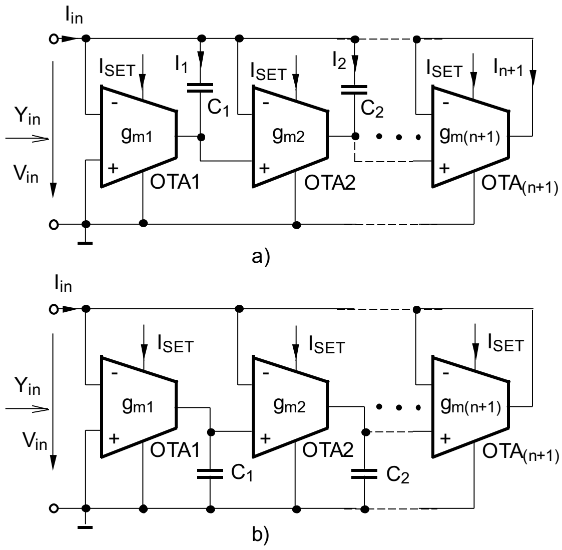

Fig. 3. One-port networks with an nth-order admittance on the OTA base

We have proposed new simple one-port networks implementing the admittance function in (2), (see Fig. 3). The input admittance of the network in Fig. 3a can be written as

$$Y^{(n)}(p)=\frac{I_{\mathrm{in}}}{V_{\mathrm{in}}}=\sum_{i=1}^{n+1}\frac{I_i}{V_{\mathrm{in}}}=g_{m1}+\frac{g_{m1}g_{m2}}{pC_1}+\frac{g_{m1}g_{m2}g_{m3}}{p^2C_1C_2}+\cdots+\frac{g_{m1}g_{m2}g_{m3}\cdots g_{m(n+1)}}{p^nC_1C_2\ldots C_n} \tag{4}$$

and the input admittance of the network in Fig. 3b can be written as

$$Y^{(n)}(p)=\frac{I_{\mathrm{in}}}{V_{\mathrm{in}}}=g_{m(n+1)}+\frac{g_{mn}g_{m(n+1)}}{pC_n}+\frac{g_{m(n-1)}g_{mn}g_{m(n+1)}}{p^2C_{(n-1)}C_n}+\cdots+\frac{g_{m1}g_{m2}g_{m3}\cdots g_{mn}g_{m(n+1)}}{p^nC_1C_2\ldots C_n}\,. \tag{5}$$

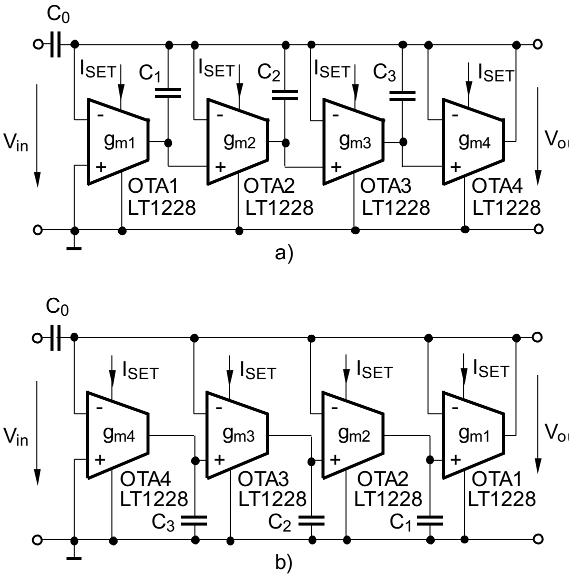

Fig. 4. Proposed electronically tunable fourth-order highpass filters in the voltage mode using OTA devices

A disadvantage of the circuit in Fig. 3a is the using of ungrounded passive elements when realizing in the integrated form. The circuit in Fig. 3b, which uses grounded passive elements, does not have this disadvantage and is thus more suitable for monolithic integration.

3 ELECTRONICALLY TUNABLE FOURTH–ORDER HIGHPASS FILTER REALIZATION

When realizing the parallel arranged admittance in Fig. 1 by one of the networks in Fig. 3 and choosing $Y_1(p)=pC_0$ and assuming $g_{m1}=g_{m2}=g_{m3}=g_{m4}=g_m$ (we consider only the fourth-order highpass filter for illustration), then according to (1) the following formula will be valid for the voltage transfer function:

$$T(p)=\frac{(p^4K_4)}{(p^4K_4+g_mp^3K_3+g_m^2p^2K_2+g_m^3pK_1+g_m^4)}\,, \tag{6}$$

where we have denoted $K_4=C_0C_1C_2C_3$, $K_3=C_1C_2C_3$, $K_2=C_2C_3$ and $K_1=C_3$,

We have obtained fourth-order highpass filters whose cut-off frequency can be set by a simultaneous change in the transconductance gains of all OTA elements by means of control current I_{SET} (see Fig. 4).

The voltage transfer function of a highpass filter in normalized form ($s=p/\omega_{3\mathrm{dBe}}$) is given by

$$T(s)=\frac{s^4}{s^4+s^3c_{41}+s^2c_{42}+sc_{43}+c_{44}} \tag{7}$$

where

$$c_{41} = \frac{g_m}{\omega_{3\mathrm{dBe}} C_0}\,, \qquad c_{42} = \frac{g_m^2}{\omega_{3\mathrm{dBe}}^2 C_0 C_1}\,,$$
$$c_{43} = \frac{g_m^3}{\omega_{3\mathrm{dBe}}^3 C_0 C_1 C_2} \qquad c_{44} = \frac{g_m^4}{\omega_{3\mathrm{dBe}}^4 C_0 C_1 C_2 C_3}\,. \tag{8}$$

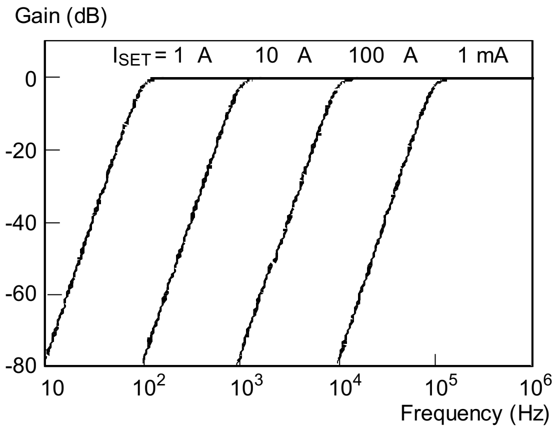

Fig. 5. Simulated magnitude frequency responses of two tunable fourth-order highpass filters from Fig. 4

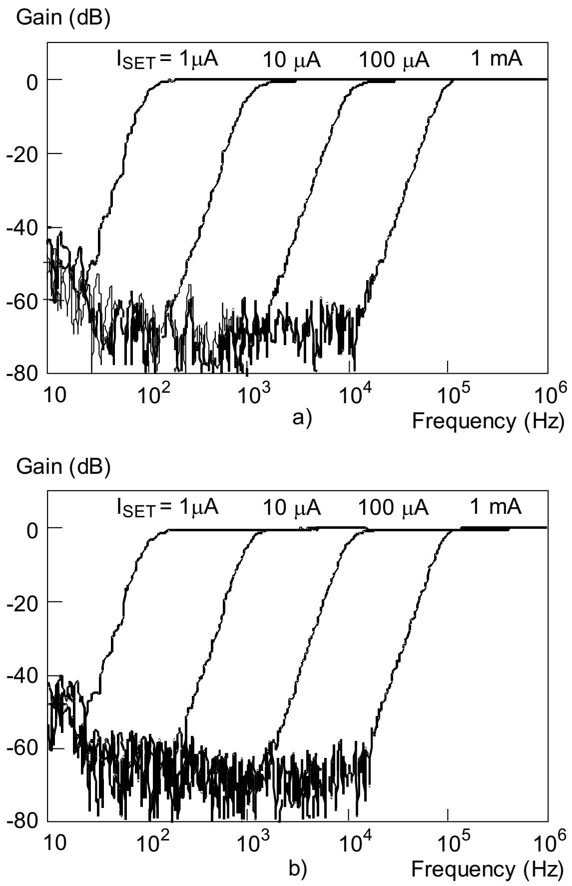

Fig. 6. Measured magnitude frequency responses of the tunable fourth-order highpass filters from Fig. 4

As an example, let us design two highpass filters from Fig. 4 with the Butterworth approximation of transfer function with cut-off frequency in the range from 100 Hz to 100 kHz. The transfer function coefficients taken from the charts for non-cascade design are: $c_{41} = 2.6131$, $c_{42} = 3.4142$, $c_{43} = 2.6131$, $c_{44} = 1.0000$. Considering LT1228 circuits [11] instead of OTA elements, it is possible to set the transconductance gain in an interval $g_m \in \langle 0.01; 10\rangle$ mS. The capacitances of capacitors can be computed from relation (8). We obtain: $C_0 = 6.09$ nF, $C_1 = 12.18$ nF, $C_2 = 20.79$ nF, $C_3 = 41.59$ nF.

3.1 Simulation results

To verify the functionality of the proposed highpass filters, PSPICE simulation was carried out. The magnitude frequency responses which are the same for both highpass filters from Fig. 4 for different values of current ISET are shown in Fig. 5.

In the simulation, a part of commercially available LT1228 circuit [11] has been used as an OTA model. Its transconductance gain is given by the formula:

$$g_m = 10 I_{\mathrm{SET}}/1V\,. \tag{9}$$

3.2 Experimental results

LT1228 circuits have also been used for experimental verification. A symmetrical supply voltage of ± 5 V has been used for these circuits. Magnitude frequency responses of the tunable highpass filters from Fig. 4 measured by the HP3589A circuit analyzer are shown in Fig. 6. These responses are quite good over a wide frequency range.

In the last LT1228 circuit the buffer which forms its part was used in order to separate the load. The buffers of the other LT1228 circuits were not made use of.

Since all OTAs have to be controlled by identical DC current, an improved multi-output Wilson current mirror with unipolar transistors in Fig. 7 [12] has been proposed. Then the filter can be controlled by a single central current source (*eg* DAC with current output). Transistor Q2 is supposed to have a fourfold channel width compared with the other transistors. The resistance of resistor R2 is therefore four times less than the resistance of resistor R1. This type of multi-output current mirror exhibits an identical unity transfer to all outputs. This feature cannot be gained with a bipolar multi-output current mirror.

The transconductance temperature coefficient has a value of -3300 ppm/°C according to the datasheet of LT 1228 circuit [11]. This temperature influence on the setting of the transconductance value can be easily compensated using, for example, an LT1004-2.5 reference diode in the source of the control current, as described in the datasheet of LT 1228 [11].

As it is necessary to change simultaneously several g_m transconductances of OTA, when the cut-off frequency of

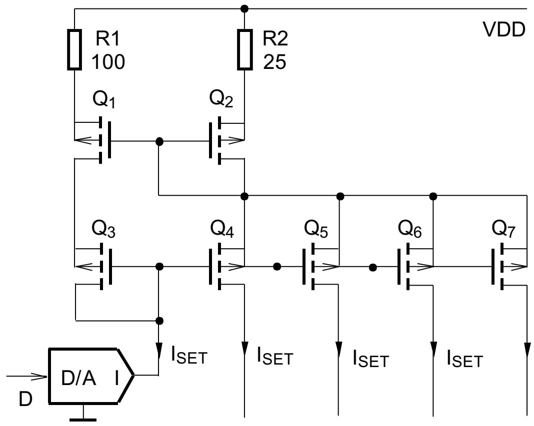

Fig. 7. Source of control currents for individual OTAs

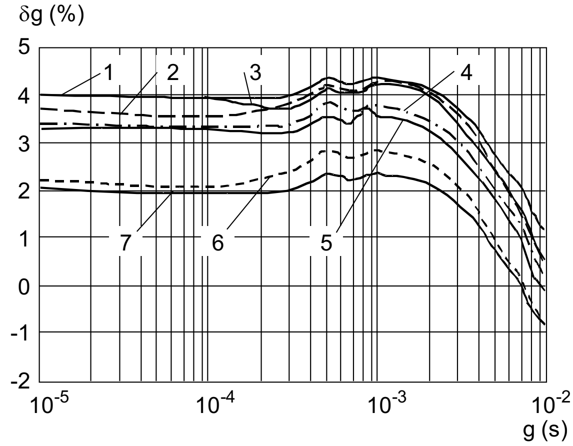

Fig. 8. Relative error of the OTA (LT 1228) transconductance setting

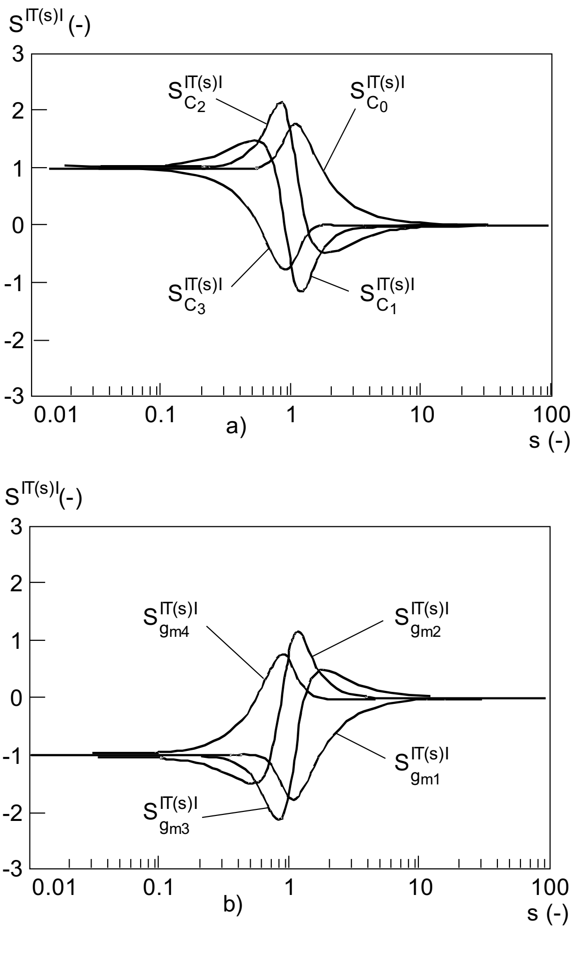

Fig. 9. Relative sensitivity of the magnitude of the transfer of the filters from Fig. 4 a) to the capacitors and b) to the OTA transconductances

the filter changes, we also examined the precision of their setting. From the results of the measurement it follows that the accuracy of all g_m settings is $\pm 1\,\%$ for each setting of g_m, see Fig. 8.

3.3 Sensitivity analysis

A detailed sensitivity analysis of the designed filters (Fig. 4 a,b) was made. As the voltage transfers of both filters have the same form, the sensitivity characteristics are also identical.

In the sensitivity computation we will start from the primary voltage transfer of the filters from Fig. 4

$$T(s) = \frac{s^4 C_0 C_1 C_2 C_3}{D}\,, \tag{10}$$

where

$$D = s^4 C_0 C_1 C_2 C_3 + s^3 C_1 C_2 C_3 g_{m1} + s^2 C_2 C_3 g_{m1} g_{m2} + s C_3 g_{m1} g_{m2} g_{m3} + g_{m1} g_{m2} g_{m3} g_{m4}\,. \tag{11}$$

The partial general sensitivity functions [13], [14] of the filter transfer function to the individual parameters of passive and active components (capacitors and transconductances g_m of OTA) are

$$\begin{aligned} S_{C_0}^{T(s)} = -S_{g_{m1}}^{T(s)} &= \frac{1}{D}\Big(s^3 C_1 C_2 C_3 g_{m1} + s^2 C_2 C_3 g_{m1} g_{m2} \\ &\quad + s C_3 g_{m1} g_{m2} g_{m3} + g_{m1} g_{m2} g_{m3} g_{m4}\Big) \\ &= \frac{s^3 c_{41} + s^2 c_{42} + s c_{43} + c_{44}}{s^4 + s^3 c_{41} + s^2 c_{42} + s c_{43} + c_{44}} \end{aligned} \tag{12}$$

$$\begin{aligned} S_{C_1}^{T(s)} = -S_{g_{m2}}^{T(s)} &= \frac{1}{D}\Big(s^2 C_2 C_3 g_{m1} g_{m2} + s C_3 g_{m1} g_{m2} g_{m3} \\ &\quad + g_{m1} g_{m2} g_{m3} g_{m4}\Big) \\ &= \frac{s^2 c_{42} + s c_{43} + c_{44}}{s^4 + s^3 c_{41} + s^2 c_{42} + s c_{43} + c_{44}} \end{aligned} \tag{13}$$

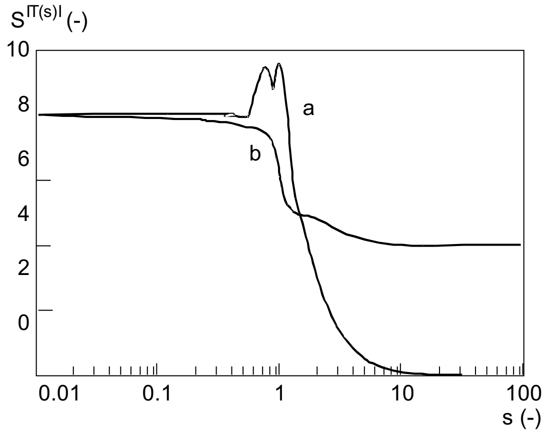

Fig. 10. Worst-case multiparameter relative sensitivity of the transfer magnitude of the filter designed with the help of the a) non-cascade synthesis and b) cascade synthesis

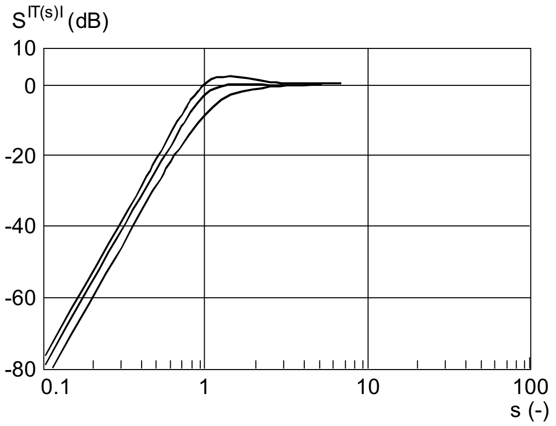

Fig. 11. Frequency magnitude response of the non-cascade variant and the limits for 5 % tolerances of the capacitors and g_m

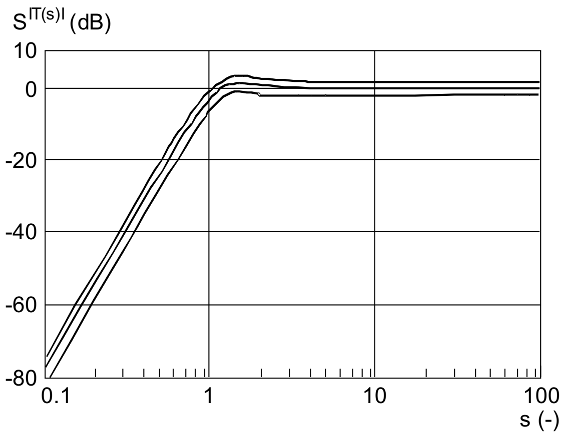

Fig. 12. Frequency magnitude response of the cascade variant and the limits for 5 % tolerances of the capacitors and resistors

$$S_{C_2}^{T(s)} = -S_{g_{m3}}^{T(s)} = \frac{sC_3 g_{m1} g_{m2} g_{m3} + g_{m1} g_{m2} g_{m3} g_{m4}}{D} = \frac{sc_{43} + c_{44}}{s^4 + s^3 c_{41} + s^2 c_{42} + sc_{43} + c_{44}} \tag{14}$$

$$S_{C_3}^{T(s)} = -S_{g_{m4}}^{T(s)} = \frac{g_{m1} g_{m2} g_{m3} g_{m4}}{D} = \frac{c_{44}}{s^4 + s^3 c_{41} + s^2 c_{42} + sc_{43} + c_{44}} \tag{15}$$

For an easy evaluation it is suitable to show these sensitivities in the graphs of the influence of individual filter parameters. It is appropriate to show in the graphs separately the real part, which expresses the relative sensitivity functions of the transfer magnitude, and the imaginary part, which express the semirelative sensitivity function of the transfer argument [14]. Figure 9 gives the partial general sensitivity function of the magnitude of the filter transfer to the capacitors and transconductances g_m for standardized frequency s. From the graphs it is evident that the sensitivities are in the pass-band area equal to zero and hence the values of the capacitors and g_m do not influence the transfer in this area and this transfer is exactly equal to one. The sensitivity of the filter transfer is mostly concentrated in the cut-off frequency area. In the stop-band area the values of the sensitivities are equal to plus or minus one. But in this area an adequate attenuation of the input signal is sufficient in the realization.

For an easy determination of the worst case of the resulting transfer magnitude influence, it is suitable to know the so-called worst-case multiparameter sensitivity of the transfer magnitude. This sensitivity expresses the maximum influence on the resulting frequency magnitude response, which is caused by the changes in the values of all filter parameters (capacitors and transconductances), as is evident from the definition equation

$$^{WS}S_{\boldsymbol{q}}^{T(s)} = \sum_{i=1}^{N} \left| \Re S_{q_i}^{T(s)} \right| , \tag{16}$$

where $\boldsymbol{q} = (q_1, q_1, q_1, \dots, q_N)$ is the vector of the filter parameters and N is the total number of parameters. The worst-case multiparameter sensitivity of the transfer magnitude for the filter from Fig. 4 is shown in graphs in Fig. 10a, again for standardized frequency. The resulting curve confirms the above facts that the sensitivity in the pass-band is equal to zero and that in the area of the cut-off frequency the sensitivity has its maximum.

To allow comparison with the cascade synthesis, the sensitivity analysis of the fourth-order high-pass filter designed with the help of cascade synthesis from two second-order high-pass filters with multifeedback [15] was also made. Figure 10b gives the worst-case multiparameter sensitivity of the transfer magnitude of the filter. From the comparison of the curves it follows that the sensitivity of the cascade filter is high in the pass-band area, in contrast to non-cascade synthesis with sensitivity equal to zero. The slightly higher sensitivity in the area of the cut-off frequency of the non-cascade synthesis usually does not matter.

The limits of the resulting frequency magnitude response are easy to determine from the worst-case multiparameter sensitivity. The frequency magnitude response of the designed filters including maximum limits for 5all filter parameters is given in Fig. 11. Figure 12 gives for comparison also the maximum limits for 5 % tolerances of all filter parameters of the cascade high-pass filter. The graphs only confirm the results that the sensitivity of the non-cascade variant is lower in particular in the pass-band area.

4 CONCLUSIONS

A method of realizing electronically tunable high-order admittance elements suitable for non-cascade synthesis of active frequency filters has been demonstrated. Two such one-port networks using modern OTA elements have been presented. One of them employs grounded passive elements. The network with grounded passive elements is more suitable for monolithic integration. It has been shown that by placing any of the one-port networks mentioned above in the frequency-dependent voltage divider we obtain an electronically tunable highpass filter in the voltage mode. We obtain an electronically tunable highpass filter in the current mode if it is appropriately located in the frequency-dependent current divider. The proposed structures are suitable only for all-pole filters (Butterworth, Chebychev). The electronically tunable highpass filters have been created, analysed for sensitivity and measured to verify theoretical results. The simulated and measured results are in excellent agreement with theoretical predictions. From the viewpoint of the sensitivity it can be said that the designed solution of the high-pass filter is less sensitive than the cascade solution, in particular in the pass-band area, where the sensitivity is equal to zero.

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