Optimized Coupled Band-Pass Filters

Neven Mijat* and Dražen Jurišić*
*University of Zagreb/Faculty of Electrical Engineering and Computing
Unska 3, HR-10 000 Zagreb, Croatia
neven.mijat@fer.hr; drazen.jurisic@fer.hr

***Abstract*— In this paper we compare two versions of fourth-order band-pass (BP) filter section realized as coupled structure, having negative feedback around two Biquadratic sections in cascade, usually designated as Biquartic section (i.e. it is the fourth-order band-pass section). The frequency response of a filter is subject to vary from the nominal values due to the effects of aging, changing working conditions, the fabrication tolerances of passive and active elements, etc. In order to maintain the filter's characteristics inside given specifications at the time of manufacture, and as long as possible, the main requirement is to design filters with reduced sensitivity to component tolerances. The Biquartic section has significantly reduced sensitivities to the changes of passive element values, particularly within the pass band, in comparison to common cascade design. For that purpose this structure is very suitable for use as a building block of narrow band, high order BP filters. The sensitivities are further reduced by using three-amplifier biquadratic block instead of more sensitive single-amplifier block. We investigated the sensitivities of two versions of biquartic section with two extreme combinations of second-order sections q-factor values, using Schoeffler's measure, the dynamic range, and output thermal noise using analysis in PSpice. Biquartic sections used throughout the examples are realized as reduced amplifier version retaining the low sensitivities of biquartic design. They also possess simple tuning features and simple design; design equations are given.**

Keywords: Band-pass filters, Low-sensitivity filters, Low-noise filters, Biquartic section, Cascade realization, Schoeffler's sensitivity.

I. INTRODUCTION

Often in practice band–pass (BP) filters of high-order and narrow pass-band are difficult to be realized. They possess very high pole-Q factors, and therefore have very high sensitivity to component tolerances. This problem is becoming more pronounced if such band-pass filters are realized in integrated-circuit form, because on-chip component tolerances are much larger than those in discrete form. One of the efficient solutions of this problem is reducing the sensitivity using negative feedback: a coupled structure [1], [2].

In this paper we present the reduced-amplifier design of the coupled fourth-order BP filter with *low sensitivity* which is referred to as Biquart. We analyse the sensitivity of the obtained filter using Schoeffler's measure, on various examples and different designs. In the design we also wish to have the possibility of easy tuning the center frequency and the pass-band width without deteriorating the amplitude-frequency characteristics. It is also desirable that the new structure provides lower signal distortion and output thermal noise, as well.

II. HIGH-ORDER BP FILTERS DESIGN

To design high-order ($2N^{\text{th}}$–order) geometrically symmetrical BP filters we start from N^{th}-order low-pass (LP) prototype filter with the transfer function:

$$T_{LP}(S)=\left[\frac{k_{LP0}\cdot\gamma}{S+\gamma}\right]^{\eta}\prod_{i=1}^{(N-\eta)/2}\frac{k_{LPi}\cdot\omega_{LPi}^{2}}{S^{2}+\left(\omega_{LPi}/Q_{LPi}\right)\cdot S+\omega_{LPi}^{2}},\quad(1)$$

where η=1 if the filter order N is odd, η=0 for N even, γ is the frequency of the negative real pole for N odd, ω_{LPi} and q_{LPi} are parameters of the $(N-\eta)/2$ complex-conjugate pole pairs, and the k_{LPi} are pass-band gains. If (1) is realized in the cascade form we take care of the sequence of realizing the individual Q-factors such that $Q_{LPi}<Q_{LPi+1}<Q_{LPi+2}<\dots$. Using common LP–BP transformation:

$$S=\frac{s^{2}+\omega_{0}^{2}}{Bs},\quad(2)$$

where ω_0 is the desired center frequency, and B is the desired pass-band width of the band-pass filter, we realize the BP transfer function of the general form given by:

$$T_{CBQ}(s)=\left[T_{b0}(s)\right]^{\eta}\prod_{i=1}^{(N-\eta)/2}T_{BQi}(s),\ \ \eta=\begin{bmatrix}1 & \text{for } N \text{ odd}\\ 0 & \text{for } N \text{ even}\end{bmatrix},\quad(3)$$

where we realize $(N-\eta)/2$ BP fourth-order sections in cascade, and one second-order section if the LP prototype filter was odd. The transfer function of second-order section has the form given by (4) (without index i).

Each $T_{BQi}(s)$ in (3) is fourth order transfer function and we will continue to call it Biquartic transfer function, while the network which implements such a function is called Biquartic section or "Biquart" (hereafter BQ). $T_{b0}(s)$ is second order transfer function and is referred to as Biquadratic section or "Biquad". The BP filter of higher order having the transfer function defined by (3) is referred to as "Cascade of Biquarts" (hereafter CBQ).

III. BIQUARTIC SECTION

Biquartic section, which realizes each fourth-order BP transfer function $T_{BQi}(s)$ in (3) is the main subject of this paper. It will be realized by cascading two BP Biquadratic sections (of second-order) applying a negative feedback, as shown in Fig. 1 (see [1], [2]).

Gain β represents the feedback coefficient (β>0), A is an input gain, and $T_{b1}(s)$, $T_{b2}(s)$, are the second-order BP

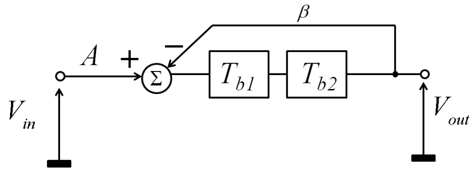

Figure 1. Biquartic BP section.

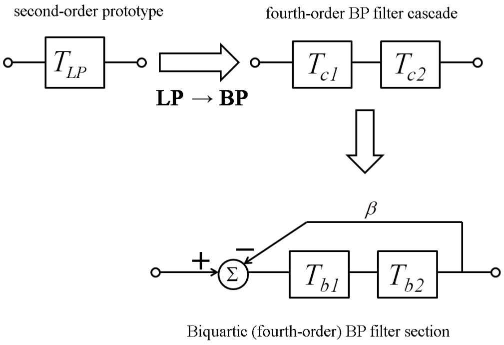

Figure 2. Complete procedure of the calculation of the fourth-order BP Biquartic section starting from second-order LP prototype via cascade.

Biquadratic sections with voltage transfer functions given by:

$$T_{bi}(s) = \frac{k_{bi} s}{s^2 + (\omega_{bi}/Q_{bi}) \cdot s + \omega_{bi}^2}; \; i = 1,2. \tag{4}$$

Voltage transfer function of the Biquartic section in Fig. 1 is given by:

$$T_{BQ}(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{A T_{b1}(s) T_{b2}(s)}{1 + \beta T_{b1}(s) T_{b2}(s)} = \frac{N_{BQ}(s)}{D_{BQ}(s)}. \tag{5}$$

The complete procedure in designing of one BP fourth-order BQ section in CBQ, from the second-order LP prototype filter over the cascaded fourth-order BP filter is shown in Fig. 2. The realization starting from the 1st-order LP producing the 2nd-order BP filters if the filter prototype order N was odd is well-known.

A. Determination of the coefficients of the cascade

In the first step in Fig. 2 it is shown the common LP–BP transformation applied on the second-order LP prototype filter and the obtained fourth-order BP filter in the cascade structure. The transfer function of a second-order LP prototype is given by (1) for N=2. Combining (2) and (1) we obtain $T_{BP}(s)=N(s)/D(s)$ in the form:

$$T_{BP}(s) = \frac{k_{LP} \cdot B^2 \omega_{LP}^2 \cdot s^2}{s^4 + B\omega_0^2 \dfrac{\omega_{LP}}{Q_{LP}} s^3 + \left(2\omega_0^2 + B^2\omega_{LP}^2\right) s^2 + B \dfrac{\omega_{LP}}{Q_{LP}} s + \omega_0^4}. \tag{6}$$

Transfer function of the fourth-order BP cascade in Fig. 2 has the form:

$$T_{CAS}(s) = \prod_{i=1}^{2} T_{ci}(s), \tag{7}$$

where $T_{c1}(s)$ and $T_{c2}(s)$ are the second-order Biquadratic sections with BP voltage transfer functions given by:

$$T_{ci}(s) = \frac{k_{ci} s}{s^2 + (\omega_{ci}/Q_{ci}) \cdot s + \omega_{ci}^2}; \; i = 1,2. \tag{8}$$

We substitute (8) into (7) and rewrite it in the form:

$$T_{CAS}(s) = \frac{k_{c1} k_{c2} \cdot s^2}{s^4 + \left(\dfrac{\omega_{c1}}{Q_{c1}} + \dfrac{\omega_{c2}}{Q_{c2}}\right) s^3 + \left(\omega_{c1}^2 + \omega_{c2}^2 + \dfrac{\omega_{c1}\omega_{c2}}{Q_{c1}Q_{c2}}\right) s^2 + \left(\dfrac{\omega_{c1}^2 \omega_{c2}}{Q_{c1}} + \dfrac{\omega_{c1}\omega_{c2}^2}{Q_{c2}}\right) s + \omega_{c1}^2 \omega_{c2}^2}. \tag{9}$$

If we equate the coefficients multiplying the potentions of complex variable "s" in the denominator of (9) and (6) we obtain the system of equations:

$$\frac{\omega_{c1}}{Q_{c1}} + \frac{\omega_{c2}}{Q_{c2}} = B \frac{\omega_{LP}}{Q_{LP}} \tag{10}$$

$$\omega_{c1}^2 + \omega_{c2}^2 + \frac{\omega_{c1}\omega_{c2}}{Q_{c1}Q_{c2}} = 2\omega_0^2 + B^2 \omega_{LP}^2 \tag{11}$$

$$\frac{\omega_{c1}^2 \omega_{c2}}{Q_{c2}} + \frac{\omega_{c1}\omega_{c2}^2}{Q_{c1}} = B \frac{\omega_{LP}}{Q_{LP}} \omega_0^2 \tag{12}$$

$$\omega_{c1}^2 \omega_{c2}^2 = \omega_0^4 \tag{13}$$

Combining (10) and (12) we obtain the condition for determining solutions:

$$(\omega_{c2} - \omega_{c1})(Q_{c2} - Q_{c1}) = 0. \tag{14}$$

Two solutions are possible depending on the poles of the LP prototype:

1. $\omega_{c1}=\omega_{c2}=\omega_0$ for real poles of the LP prototype.
2. $Q_{c1}=Q_{c2}=Q_c$ for conjugate-complex poles of the LP prototype.

In what follows we choose conjugate-complex poles of the LP prototype (case 2) and the system of equations (10)–(13) is simplifying into the system:

$$\frac{\omega_{c1} + \omega_{c2}}{Q_c} = B \frac{\omega_{LP}}{Q_{LP}} \tag{15}$$

$$\omega_{c1}^2 + \omega_{c2}^2 + \frac{\omega_0^2}{Q_c^2} = 2\omega_0^2 + B^2 \omega_{LP}^2 \tag{16}$$

$$\omega_{c1}\omega_{c2} = \omega_0^2 \tag{17}$$

Rearranging and grouping equations of the latest system, we arrive at expressions for Q_c, ω_{c1} and ω_{c2} of the cascade realization (Geffe equations) [3]:

$$Q_c = \sqrt{\frac{Q_{LP}\omega_0}{B\omega_{LP}}} \times \sqrt{Q_{LP} X + \sqrt{Q_{LP}^2 X^2 - 1}}; \quad X = 2\omega_0/(B\omega_{LP}) + B\omega_{LP}/(2\omega_0), \tag{18}$$

$$\frac{\omega_{c2}}{\omega_0} = \frac{\omega_0}{\omega_{c1}} = \frac{B\omega_{LP}}{2Q_{LP}} Q_c + \sqrt{\left(\frac{B\omega_{LP}}{2Q_{LP}} Q_c\right)^2 - 1}. \tag{19}$$

If we equate numerator in (9) and (6) we obtain:

$$k_{c1} \cdot k_{c2} = k_{LP} \cdot B^2 \omega_{LP}^2. \tag{20}$$

B. Voltage transfer function of the Biquartic section

Consider the transfer function of the Biquartic section in Fig.1 given by (5). Since the positions of the poles of T_{BQ}(s) is determined by the product $\beta k_{b1}k_{b2}$, i.e. not only by β, we will use the designation F ($F=\beta k_{b1}k_{b2}$) in order to simplify the expression. We rewrite (5) in the form:

$$T_{BQ}(s) = \frac{Ak_{b1}k_{b2}\cdot s^2}{s^4 + \left(\frac{\omega_{b1}}{Q_{b1}} + \frac{\omega_{b2}}{Q_{b2}}\right)s^3 + \left(\omega_{b1}^2 + \omega_{b2}^2 + \frac{\omega_{b1}\omega_{b2}}{Q_{b1}Q_{b2}} + F\right)s^2 + \left(\frac{\omega_{b1}^2}{Q_{b1}} + \frac{\omega_{b2}^2}{Q_{b2}}\right)s + \omega_{b1}^2\omega_{b2}^2} \quad (21)$$

Poles of the transfer function are determined by the denominator of the Biquartic transfer function:

$$D_{BQ}(s) = D(s) + Fs^2 = D(s) + \beta k_{b1}k_{b2}s^2 . \quad (22)$$

It is evident that for F=0 (β=0, k_{b1}, k_{b2} $\neq$0) the BQ consists of the two Biquads connected in the cascade, and $D(s)$ represents the denominator of (6), i.e. of the cascade.

Consider the commonly used cascade realization and compare it with the Biquartic section (both of fourth order) shown in Fig. 2. If we equate the coefficients multiplying the potentions of complex variable “s” in (21) to (9), we obtain the system of equations:

$$\frac{\omega_{c1}}{Q_{c1}} + \frac{\omega_{c2}}{Q_{c2}} = \frac{\omega_{b1}}{Q_{b1}} + \frac{\omega_{b2}}{Q_{b2}} \quad (23)$$

$$\omega_{c1}^2 + \omega_{c2}^2 + \frac{\omega_{c1}\omega_{c2}}{Q_{c1}Q_{c2}} = \omega_{b1}^2 + \omega_{b2}^2 + \frac{\omega_{b1}\omega_{b2}}{Q_{b1}Q_{b2}} + F \quad (24)$$

$$\frac{\omega_{c1}^2\omega_{c2}}{Q_{c2}} + \frac{\omega_{c1}\omega_{c2}^2}{Q_{c1}} = \frac{\omega_{b1}^2\omega_{b2}}{Q_{b2}} + \frac{\omega_{b1}\omega_{b2}^2}{Q_{b1}} \quad (25)$$

$$\omega_{c1}^2\omega_{c2}^2 = \omega_{b1}^2\omega_{b2}^2 \quad (26)$$

Because it is given above [see (18) and (19)]:

$$\omega_{c1}\omega_{c2} = \omega_0^2, \quad Q_{c1} = Q_{c2} = Q_c , \quad (27)$$

we have:

$$\omega_{b1}\omega_{b2} = \omega_0^2 , \quad (28)$$

and we choose

$$\omega_{b1} = \omega_{b2} = \omega_0 . \quad (29)$$

Now the system of equations (23)–(26) is simplifying into the system (note that (23) and (25) have become identical):

$$\frac{\omega_{c1} + \omega_{c2}}{Q_{c1}} = \omega_0\left(\frac{1}{Q_{b1}} + \frac{1}{Q_{b2}}\right), \quad (30)$$

$$\omega_{c1}^2 + \omega_{c2}^2 + \frac{\omega_0^2}{Q_c^2} = 2\omega_0^2 + \frac{\omega_0^2}{Q_{b1}Q_{b2}} + F . \quad (31)$$

Rearranging it we obtain the condition for determining solutions:

$$(\omega_{b2} - \omega_{b1})(Q_{b2} - Q_{b1}) = 0 . \quad (32)$$

We have two equations (30) and (31) and three unknowns to be determined Q_{b1}, Q_{b2} and F. One of the unknowns can be freely chosen (therefore we have one degree of freedom). We choose pole Q factors of the sections T_{b1}, T_{b2} to have two limiting values of Q-factors:

1. Sections T_{b1}, and T_{b2} have identical Q-factors.
2. One of the sections T_{b1} or T_{b2} has an infinite Q-factor.

Between these two limiting cases for Q-factors there are infinite possibilities, which turn to be more sensitive, and we do not treat them separately.

C. Equal Q-factors Q_{b1}=Q_{b2} [4]

The sections T_{b1} and T_{b2} are becoming identical, and we have:

$$Q_{b1} = Q_{b2} = \frac{2\sqrt{\omega_{c1}\omega_{c2}}}{\omega_{c1} + \omega_{c2}}Q_c . \quad (33)$$

$$\omega_{b1} = \omega_{b2} = \omega_0 = \sqrt{\omega_{c1}\omega_{c2}} . \quad (34)$$

Feedback factor F takes the value defined by an expression:

$$F = (\omega_{c2} - \omega_{c1})^2\left(1 - \frac{1}{4Q_c}\right). \quad (35)$$

D. One Q-factor is in infinity: Q_{b2}=∞

One Q-factor is becoming infinite, while other has the value:

$$Q_{b1} = \frac{\sqrt{\omega_{c1}\omega_{c2}}}{\omega_{c1} + \omega_{c2}}Q_c . \quad (36)$$

Pole frequencies are the same as in the previous case:

$$\omega_{b1} = \omega_{b2} = \omega_0 = \sqrt{\omega_{c1}\omega_{c2}} . \quad (37)$$

Feedback factor F is then given by:

$$F = (\omega_{c2} - \omega_{c1})^2 + \frac{\omega_{c1}\omega_{c2}}{Q_c^2} . \quad (38)$$

IV. EXAMPLES

In this section we realize three examples of the active-RC BP filters: one for the cascade (CAS) case in the Section III.A, and two for the Biquartic (BQ) cases in the Sections III.C and III.D. Consider two approximations: *i*) Butterworth (with normalized LP prototype parameters: ω_{LP}=1, Q_{LP}=0.707107), and *ii*) 0.5dB-ripple Chebyshev (ω_{LP}=1.231342, Q_{LP}=0.863721 [a higher pole-Q case]) that are transformed into BP characteristics with normalized center frequency ω_0=1 and bandwidth B=0.1. Filters are then denormalized to the frequency 10kHz. The active realization is shown in Fig. 3 and the element values are in Table I. Corresponding amplitude-frequency characteristics are shown in Fig. 4.

Generally, the transfer function of each Biquad shown in Fig. 3 has the form given by (4), with:

$$\omega_{bi}=\sqrt{\frac{R_{Fi}}{C_{i1}C_{i2}R_{i1}R_{i3}R_{bi}}};\ Q_{bi}=C_{i1}R_{i2}\sqrt{\frac{R_{Fi}}{C_{i1}C_{i2}R_{i1}R_{i3}R_{bi}}};$$
$$k_{bi}=\frac{R_{Fi}}{C_{i1}R_{0i}R_{i1}};\quad i=1,2. \tag{39}$$

Note that in this realization we do not use a separate operational amplifier for negative feedback. Instead, we lead the output signal to the positive input of the first operational amplifier in the first biquad. As a consequence the feedback factor F is determined by:

$$F=\beta\cdot k_{b1}\cdot k_{b2}=\beta\cdot\left(1+\frac{R_{F1}}{R_{01}}+\frac{R_{F1}}{R_{b1}}\right)\cdot\frac{1}{R_{11}C_{11}}\cdot k_{b2}, \tag{40}$$

where

$$\beta=R_{F3}/(R_{F3}+R_{03}). \tag{41}$$

It is evident from (39) that for $R_{i2}=\infty$ the pole-Q factor is infinite (e.g., for $R_{22}=\infty$ in Fig. 3 there is $Q_{b2}=\infty$).

We choose value of R_{F3} and calculate R_{03} depending on F:

$$R_{03}=\frac{R_{F3}}{F}\left[\left(1+\frac{R_{F1}}{R_{01}}+\frac{R_{F1}}{R_{b1}}\right)\cdot\frac{1}{R_{11}C_{11}}\cdot k_{b2}-F\right]. \tag{42}$$

It is evident from (40) and (41) that for $R_{03}=\infty$ and $R_{F3}=0$ the feedback factor $F=0$ and the circuit in Fig. 3 becomes cascade of Biquads.

Using (39)−(42) we calculate normalized elements of filter examples and denormalize the filter to the center frequency f_0=10kHz and bandwidth B=1kHz. We choose denormalization resistance R_0=1.591kΩ in order to obtain denormalization capacitance C_0=10nF. The obtained values are given in Table I. In order to provide maximum dynamic range it will be needed to perform gain optimization in the next step, which is in this case not possible in the simple non-iterative way. It is needed to perform numerical optimization or the optimization in an iterative way.

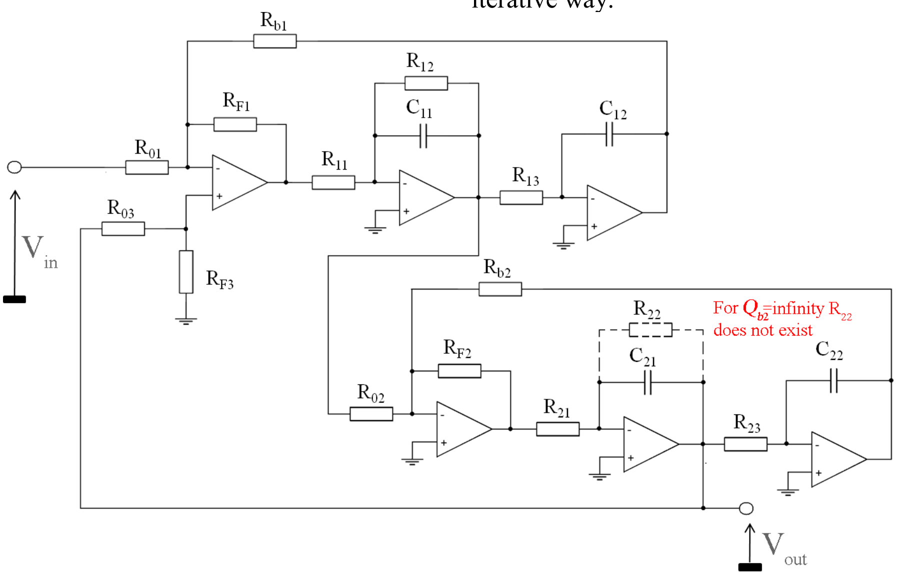

Figure 3. Realization of Biquartic section with two general purpose active-RC Biquadratic sections and feedback.

TABLE I. ELEMENT VALUES OF THE CIRCUIT IN FIG. 3 FOR THE BUTTERWORTH AND CHEBYSHEV EXAMPLES DENORMALIZED TO 10KHZ.

Elements	Butterworth			Chebyshev 0.5dB		
	CAS	BQ Q_1=Q_2	BQ Q_2=∞	CAS	BQ Q_1=Q_2	BQ Q_2=∞
R_{01}	175.923kΩ	116kΩ	95kΩ	502.16kΩ	110kΩ	100kΩ
R_{11}	1.557Ω	1.536kΩ	1.515kΩ	0.818Ω	1.475kΩ	1.412kΩ
R_{12}	22.521kΩ	22.507kΩ	11.253kΩ	31.05kΩ	22.327kΩ	11.163kΩ
R_{13}	1.895kΩ	1.633kΩ	1.617kΩ	1.681kΩ	1.746kΩ	1.612kΩ
R_{F1}	10kΩ	10kΩ	10kΩ	10kΩ	10kΩ	10kΩ
R_{b1}	9.199kΩ	10.086kΩ	10.326kΩ	19.72kΩ	9.821kΩ	11.111kΩ
C_{11}	10nF	10nF	10nF	10nF	10nF	10nF
C_{12}	10nF	10nF	10nF	10nF	10nF	10nF
R_{02}	95.784kΩ	135.714kΩ	214.364kΩ	85.749kΩ	107.843kΩ	144.144kΩ
R_{21}	1.205kΩ	1.533kΩ	1.574kΩ	1.446kΩ	1.504kΩ	1.541kΩ
R_{22}	22.521kΩ	22.507kΩ	∞	31.05kΩ	22.327kΩ	∞
R_{23}	1.972kΩ	1.608kΩ	1.591kΩ	1.75kΩ	1.629kΩ	1.591kΩ
R_{F2}	10kΩ	10kΩ	10kΩ	10kΩ	10kΩ	10kΩ
R_{b2}	9.918kΩ	10.133kΩ	10.104kΩ	9.317kΩ	10.328kΩ	10.322kΩ
C_{21}	10nF	10nF	10nF	10nF	10nF	10nF
C_{22}	10nF	10nF	10nF	10nF	10nF	10nF
R_{03}	∞	459kΩ	138.438kΩ	∞	304.758kΩ	149.566kΩ
R_{F3}	0	10kΩ	10kΩ	0	10kΩ	10kΩ

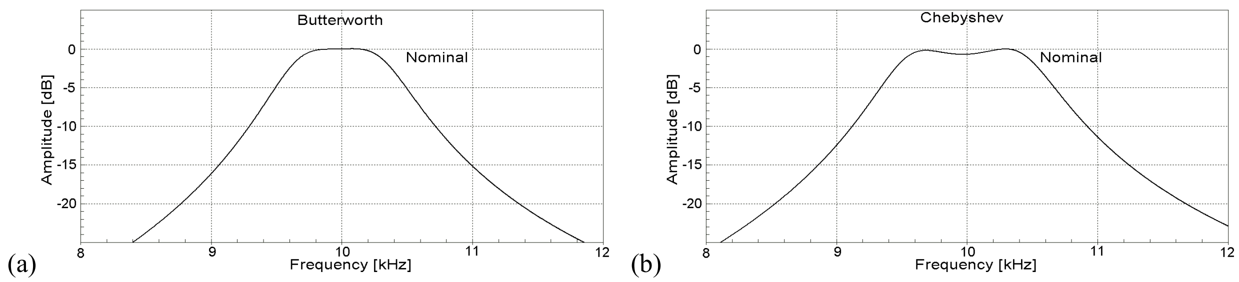

Figure 4. Simulated amplitude-frequency characteristics of filter examples in Table I using Altium Designer. (a) Butterworth. (b) Chebyshev.

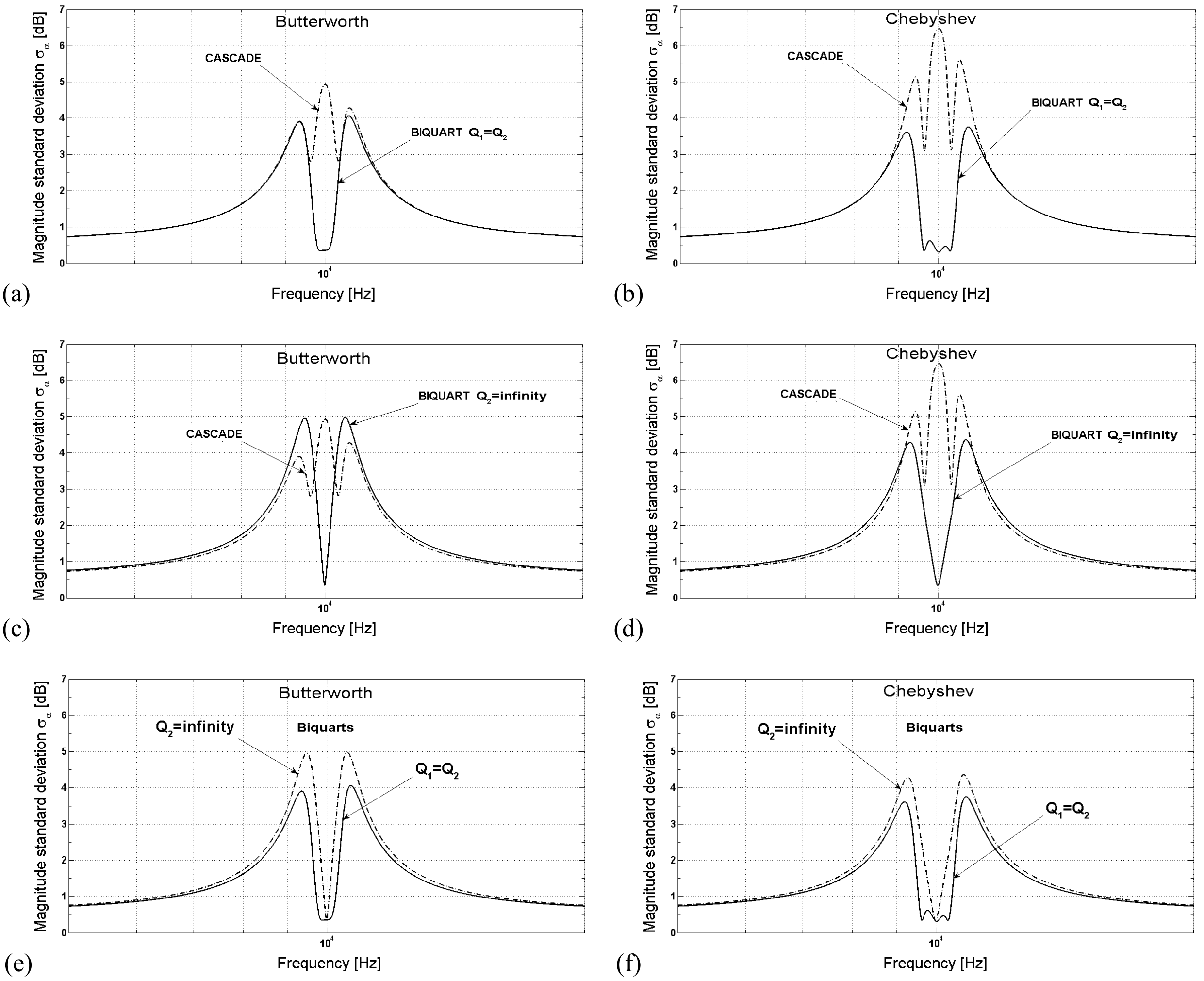

Figure 5. Simulated Schoeffler sensitivities of filter examples in Table I using Matlab. (a), (c), (e) Butterworth. (b), (d), (f) Chebyshev.

V. RESULTS OF SIMULATION AND MEASUREMENT

The sensitivity to component tolerances and output thermal noise, are compared for the cases of cascade (in Section III.A) and for the two BQ realizations: *i*) equal Q-factors: $Q_{b1}=Q_{b2}$ (in Section III.C); and *ii*) one Q-factor is in infinity: $Q_{b2}=\infty$ (in Section III.D). Sensitivity analysis is performed using MATLAB procedures given in [5], assuming relative changes of resistors and capacitors to be uncorrelated random variables, with a zero-mean Gaussian distribution and 1% standard deviation. The standard deviation (which is related to Schoeffler's sensitivities) of the variation of the logarithmic gain $\Delta\alpha=8.68588\,\Delta|T_{BP}(\omega)|/|T_{BP}(\omega)|$[dB], with respect to passive elements, is calculated for filter examples from Table I and shown in Fig. 5. We conclude that low-sensitivity BP filters can be designed using CBQ filters with equal Q-factors: $Q_{b1}=Q_{b2}$. Using the Altium Designer program output thermal noise spectral density of filter examples in Table I was generated and shown in Fig. 6. In the simulation a model of TL081/TI (Texas Instruments) FET input opamp, having typical values, as found in the data-sheets $e_{na}(f)=17\text{nV}/\sqrt{\text{Hz}}$ and $i_{na1}(f)\approx i_{na2}(f)=0.01\text{pA}/\sqrt{\text{Hz}}$ is used. From Fig. 6 one can conclude that the minimum noise possesses CBQ filters with equal Q-factors: $Q_{b1}=Q_{b2}$.

Measurements were performed on the same examples realized on separate printed circuit boards using 1% passive elements and TL081/TI opamps. For each filter, the output noise spectral density was measured using a typical university lab environment. The measurement equipment consisted of a high-quality HP 4195A Network

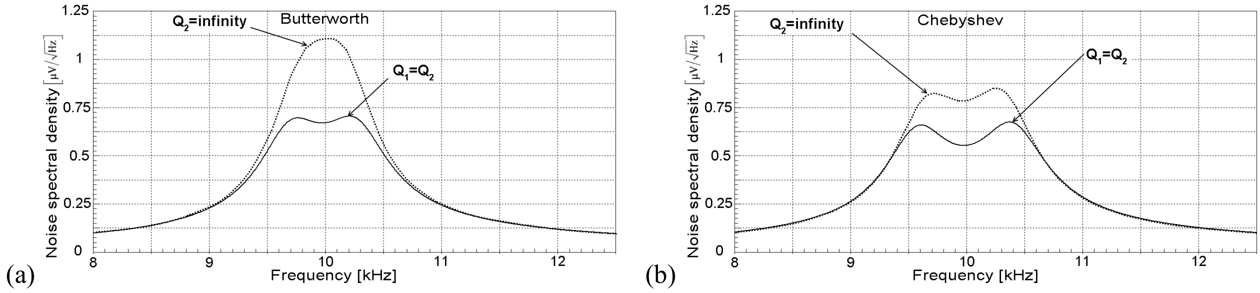

Figure 6. Simulated output thermal noise spectral densities of BQ filter examples in Table I using Altium Designer. (a) Butterworth. (b) Chebyshev.

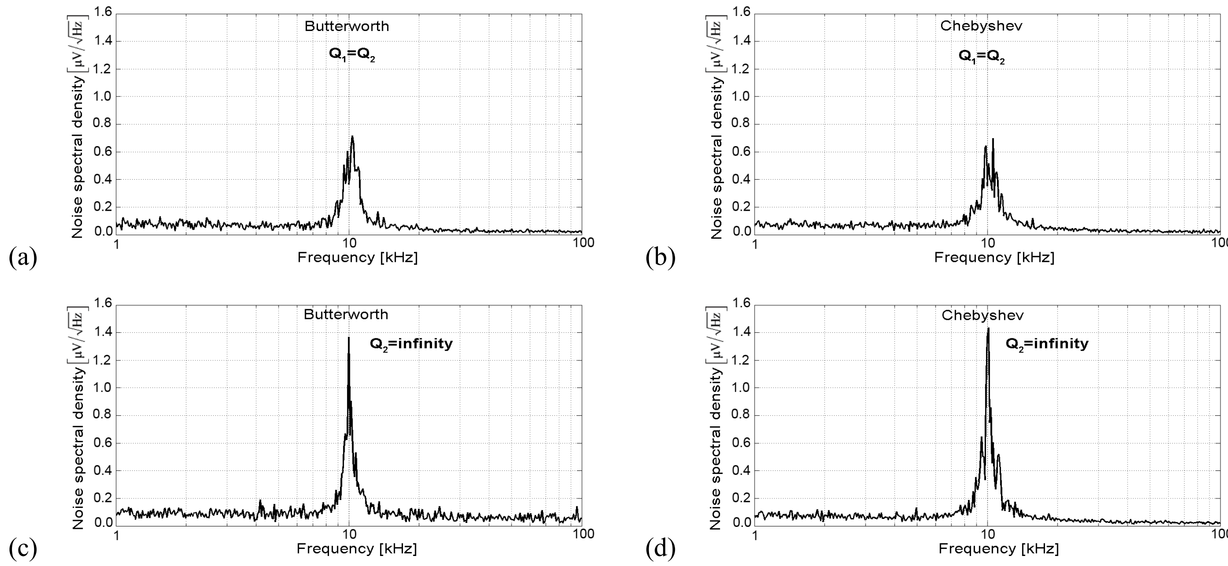

Figure 7. Measured output thermal noise spectral density of BQ filter examples in Table I. (a), (c) Butterworth. (b), (d) Chebyshev.

Analyser, which measures the spectrum of noise (Spectrum mode). Detailed description of measurement procedure and equipment is given in [6]. Measured results are shown in Fig. 7, and reconfirm the results obtained by simulation.

VI. CONCLUSION

In accordance with a simulated sensitivity by Schoeffler in realizations of amplitude-frequency characteristics large discrepancies in implementation with one infinite Q factor are expected. Thus, better results are gained using Biquartic sections with identical Q factors. Deviations in amplitude-frequency characteristics are predictable in shifting of the center frequency and amplitude distortion. Observing Schoeffler sensitivities it is visible that the sections with one infinite Q factor have increased sensitivities at the band-edges.

In order to reduce the overall sensitivities we use the three-amplifier biquads as building blocks for biquartic filter because of their considerably lower sensitivities comparing to single-amplifier structures, at the expense of the increased power consumption.

The sensitivity differences between the two CBQ cases are smaller on Chebyshev filter frequency response. Since Chebyshev transfer function has higher Q-factors than Butterworth it leads us to conclusion that the differences are smaller in the highly selective BP filters. The cascade structure has considerably bigger sensitivities in all cases.

Measured spectral density of circuit output thermal noise is very close to the simulation results. We note better results in Biquartic sections with identical Q factors. Both Biquartic realizations are preferable over the common cascade structure. In all measurements and simulations wins Biquartic section with identical Q factors and imposes as an optimal realization of low- sensitive filters.

ACKNOWLEDGEMENT

The authors wish to thank M. Sc. Kristijan Bago for measurements made on the filter examples in this paper.

REFERENCES

[1] N. Mijat and G. S. Moschytz. Multiple-critical-pole coupled active filters. *Int. J. Circ. Theor. Appl.* Vol. 12, no. 3, pp. 249−268, July 1984.

[2] S. Fotopoulos and T. Deliyannis. Active RC realization of high-order bandpass filter functions by cascading biquartic sections. *Int. J. Circ. Theor. Appl.* Vol. 12, no. 3, pp. 223−238, July 1984.

[3] P. R. Geffe. Designer's Guide to Active Bandpass Filters. *EDN* 1974; 46–52.

[4] N. Mijat and G. S. Moschytz, "Sensitivity of narrowband biquartic BP active filter block," in *Proc. of IEEE International Symposium on Network Theory and Design*, (Sarajevo, Bosnia and Herzegovina), 1984, pp. 158–163.

[5] D. Jurisic. *Applications of MATLAB in Science and Engineering. Chapter 10: Low-Noise, Low-Sensitivity Active-RC Allpole Filters Using MATLAB Optimization*. Rijeka: InTech, 2011, pp. 197−224.

[6] D. Jurisic, N. Mijat and M. Ranilovic, "Power Consumption, Noise and Bode Diagram Measurement of Active Filters," in *Proc. of the 34th International Convention* MIPRO 2011, (Opatija, Croatia), May 2013, pp. 92–95.