FBAR Filter with Asymmetric Frequency Response and Improved Selectivity and Passband Width

Ivan Uzunov, Dobromir Gajdajiev, and Ventsislav Yantchev

***Abstract*—The paper is dedicated on the improving of the frequency response of FBAR filters by replacement some of the single FBARs with two parallel connected resonators. The method is applied for the basic lattice filter architecture and its modification with twice less number of FBARs. The conditions, which must satisfy the resonator parameters, are derived by theoretical considerations and computer simulations.**

***Index Terms*—Film Bulk Acoustic-wave Resonators (FBAR), radio-frequency filters, FBAR filters, piezo-resonator filters.**

I. Introduction

The film bulk acoustic-wave resonators (FBAR) found in the last years fast growing application for realizing of high selective filters for the analog front-ends in the radio transmitters and receivers. Basic FBAR advantages, which led to their wide application, are their very low losses, better temperature stability, the appropriate frequency range of operation, ability to handle relatively high powers, good compatibility with the existing CMOS technology, etc. [1-6].

There are two basic architectures of FBAR filters using separated resonators: ladder and lattice, shown in Fig. 1 [1], [5]. Double mode filters, using decoupled stacked bulk acoustic resonators (DSBAR) [5], are also proposed, however they are outside of the scope of this paper and they will not be commented here.

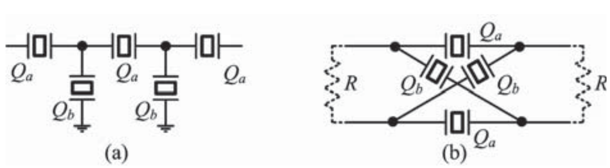

Fig. 1. (a) Ladder FBAR filter with 5 resonators; (b) lattice filter.

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The series resonance frequency f_{sa} of the series resonators Q_a in the ladder filters must be equal to the parallel resonance frequency f_{pb} of the shunt resonators Q_b. This requirement comes from the basic principle of operation of the ladder filters: the series FBARs have zero impedance at f_{sa} and realize short circuit between input and output, while the shunt resonators are equivalent to open circuits at the same frequency ($f_{pb} = f_{sa}$) and do no shunt the signal (power losses in the resonators are neglected). The resonators produce transmission zeros at their two other resonance frequencies: the shunt resonators Q_b at their series resonances f_{sb} and series resonators Q_a at their parallel resonances f_{pa}. The transmission zeros at f_{sb} and f_{pa} limit the passband bandwidth and cause returning of the filter frequency response to relatively low attenuation at frequencies below f_{sb} and above f_{pa}. The increasing of the attenuation in this region is achieved usually by cascading of few Γ-type sections, each one consisting of one series and one shunt FBAR (e.g. the filter in Fig 1(a) consists of 2½ sections). Side effects of this approach are further shrinking of the filter passband and undesirable increasing of the passband losses.

Therefore the passband width in the ladder filters is defined basically by the distance between series and parallel resonances of the resonators. Computer simulation shows that the width of the passband of single Γ-type section, measured at -0.3dB level, is about 2.3% from the passband central frequency for FBARs with ratio $f_p/f_s \approx 1.027$ (effective coupling factor $k^2_{t\,eff} = 6.2\%$). The proposed methods for extending of the filter passband try to increase this distance. One way is to increase the effective coupling factor $k^2_{t\,eff}$ of the resonators by technological means, since it determines the ratio between their series and parallel resonance frequencies [6],[7]. Other methods increase the distance between f_s and f_p by connecting small inductors in series to the resonators [5],[8]. Their effect can be explained if consider the frequency behavior of the imaginary parts of the FBAR impedance (its reactance), simulated by using of the modified Butterworth - Van Dyke (mBVD) FBAR model. The model and the frequency response are shown in Fig. 2. The impedance is negative with relatively small magnitude below the FBAR series resonance and adding small inductance (few nH) in series moves downward the series resonance frequency of the

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combination. An inductor in parallel to the resonator also increases the distance between the series and the parallel resonances, however the necessary inductor value is too large. A drawback of this method is the necessity of extra inductors, which consume large chip area.

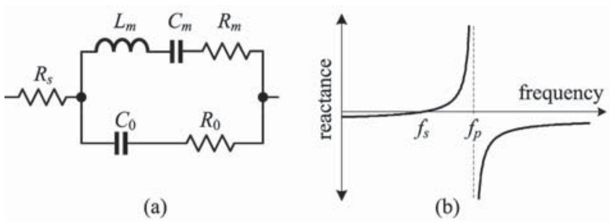

Fig. 2. (a) Modified Butterworth - Van Dyke (mBVD) model of FBAR; (b) FBAR impedance vs. frequency when the resistances are neglected.

The lattice filters employ two different pairs with identical FBARs Q_a and Q_b in each pair (Fig. 1(b)) and a requirement, similar to $f_{pb} = f_{sa}$, is valid also. The reason for this condition is different and it comes from the theory of LC lattice filters: the passband of these filters is formed by the frequency bands, in which the reactances of Q_a and Q_b have opposite signs [9]. Thus passband without gaps can be formed only if the series resonance of one from the FBAR pairs is the same as the parallel resonance of the other pair. To be more specific we will assume in the following that

$$f_{pa} = f_{sb} \,. \tag{1}$$

The passband in the lattice filters extends theoretically in the whole region where the FBAR reactances have opposite signs, i.e. between f_{sa} and f_{pb}. This statement follows from the theory of image parameters and assumes frequency dependent terminating impedances. The passband is narrower when the terminating resistors R are constant and then the filter has two frequencies of maximum gain. They are the frequency $f_{pa} = f_{sb}$ and the frequency, at which is satisfied the following condition

$$R^2 = Z_a Z_b \,, \tag{2}$$

where Z_a and Z_b are the impedances of Q_a and Q_b (losses are neglected). The filter frequency response is approximately flat with a small fall between these two frequencies. Outside of them it decreases rapidly and there is significant attenuation at f_{sa} and f_{pb}. The condition (2) can be satisfied only in the region between f_{sa} and f_{pb}, where Z_a and Z_b have opposite signs.

Thus the distance between FBAR series and parallel resonances limits the passband width in the lattice filters too. The passband extension of the lattice filters could be done in the same way as in the ladder filters: technologically by increasing of FBAR effective coupling factor or by connecting small inductors in series with the resonators.

Another way for improving the filter characteristic is to use series or parallel connected resonators with different parameters instead of the single resonators in the arms of the circuits in Fig. 1. Then the degree of the impedances of the combined resonators increases, which increases the order of the filter transfer function. The higher order usually gives steeper slopes between the passband and the stopbands and possibly wider passband. This method is not new: it is considered for the quartz-crystal filters in [9]; another application for ladder FBAR filters is given in [10]. Nevertheless there are still questions, which do not have complete answers yet. Some of them are: more accurate estimation of the parameters of the frequency response of connected resonators; more complete consideration of the effect of added resonators on the filter frequency response; the choice of resonance frequencies and other parameters of all resonators in the filter, etc.

This paper tries to answer to some of the above questions related to the lattice filter architecture and to a new circuit of FBAR filter [11] (shown in Fig. 6(a)), which is modification of the lattice filter. Section II considers the change of the resonance frequencies of two parallel or series connected FBARs; then in Section III is discussed the simplest case, when one of Q_a or Q_b in Fig. 1(b) is replaced by two parallel FBARs; and in Section IV are given comparisons with other FBAR filters based on computer simulations.

II. Parallel and Series Connected Resonators

Fig. 3(a) shows two resonators, connected in parallel and their equivalent circuit, based on mBVD model. The loss resistances in mBVD model are neglected, since they are very small and their neglecting permits applying the theory of the passive LC circuit in the deriving of the properties of the connected resonators. Then the impedances and admittances of the resonators are pure imaginary, which allows to drop out the imaginary unit in the considerations. The resonators are assumed with different resonance frequencies.

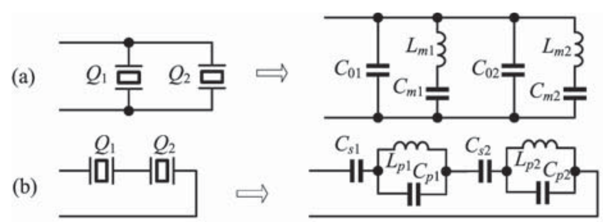

Fig. 3. (a) Two FBAR connected in parallel and their equivalent circuit; (b) two FBAR connected in series and their equivalent circuit. The loss resistances in mBVD model are neglected.

The first conclusion, which follows immediately from Fig. 3(a) is that the series resonances remain the same as they are in Q_1 and Q_2. These frequencies are defined by the groups $L_{m1}C_{m1}$ and $L_{m2}C_{m2}$, which do not change when other elements are connected in parallel. The capacitors C_{01} and C_{02} can be united in one capacitor and then the equivalent circuit in Fig 3(a) corresponds to Foster 2 form of a 5[th] order LC admittance [12] having zero in the origin. It has two series and two parallel resonances, which alternate on the frequency axis and the lowest is the series resonance.

The change of the parallel resonance frequencies can be determined if plot together the frequency responses of the admittances of both resonators in Fig. 3(a), which is done in Fig. 4. The total admittance Y is sum of both admittances Y_1 and Y_2 and the parallel resonances of Y appear at frequencies, at which Y_1 and Y_2 have equal magnitudes and opposite signs. This means that they are in the frequency bands, where Y_1 or Y_2 behave like inductances, i.e. in the intervals (f_{s1}, f_{p1}) and (f_{s2}, f_{p2}). Thus the new locations f_{p1}' and f_{p2}' of the parallel resonances are closer to the series resonances than the locations of the parallel resonances in the individual resonators. This conclusion is valid also in the case when the areas, where the FBARs are equivalent to inductances, partly overlap (i.e. when $f_{s2} < f_{p1}$). It is so, because f_{s1} and f_{s2} do not change and there is always a parallel resonance between two consecutive series resonances.

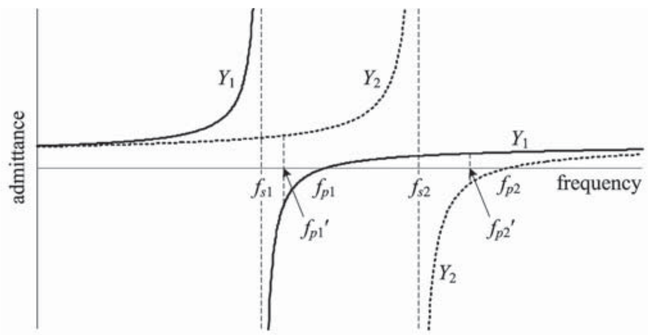

Fig. 4. Plots of the frequency responses of the admittances of two parallel connected FBARs.

The expressions for the admittances Y_1 and Y_2 are:

$$Y_1 = \frac{j\omega C_{01}\left(\omega_{p1}^2 - \omega^2\right)}{\left(\omega_{s1}^2 - \omega^2\right)}; \quad Y_2 = \frac{j\omega C_{02}\left(\omega_{p2}^2 - \omega^2\right)}{\left(\omega_{s2}^2 - \omega^2\right)}, \tag{3}$$

where ω is the angular frequency. The zeroing of $Y_1 + Y_2$ gives the following equation about the new positions of the parallel resonance frequencies:

$$C_{01}\left(\omega_{p1}^2 - \omega^2\right)\left(\omega_{s2}^2 - \omega^2\right) + C_{02}\left(\omega_{p2}^2 - \omega^2\right)\left(\omega_{s1}^2 - \omega^2\right) = 0. \tag{4}$$

This is quite general equation with many parameters, which makes difficult its investigation. It can be simplified if the following condition is satisfied:

$$\frac{\omega_{p1}}{\omega_{s1}} = \frac{\omega_{p2}}{\omega_{s2}}, \tag{5}$$

which means that both resonators have equal effective coupling factors. The resonators have the same maximum coupling factor due to technological reasons: they are located physically close to each other and they are manufactured by the same technological process. The coupling factor of individual resonators can be tuned by placing of insulating layer between the piezoelectric and one of FBAR electrodes [13]. However this is equivalent to connecting of a capacitor in series to the resonator, which reduces the coupling factor. Thus the above assumption is reasonable if one of the goals in the filter design is extending of the passband.

If the ratio in (5) is marked by α and the ratio C_{02}/C_{01} by γ, then (4) can be transformed to

$$\omega^4 - \left(\frac{\gamma + \alpha^2}{1+\gamma}\omega_{s1}^2 + \frac{1+\gamma\alpha^2}{1+\gamma}\omega_{s2}^2\right)\omega^2 + \alpha^2\omega_{s1}^2\omega_{s2}^2 = 0. \tag{6}$$

The solutions $\omega_{p1}'^2$ and $\omega_{p2}'^2$ and the series resonances are connected by the relationship $\omega_{p1}'^2\omega_{p2}'^2 = \alpha^2\omega_{s1}^2\omega_{s2}^2$ as it follows from the Viète formulas for the polynomial roots. This relationship together with (5) gives $\omega_{p1}^2\omega_{p2}^2 = \alpha^2\omega_{p1}'^2\omega_{p2}'^2$. Both relationships give the following approximate estimation about the new parallel resonance frequencies:

$$\omega_{p1}' \approx \sqrt{\alpha}\,\omega_{s1};\ \omega_{p2}' \approx \sqrt{\alpha}\,\omega_{s2} \text{ or } f_{p1}' \approx \sqrt{\alpha}\,f_{s1};\ f_{p2}' \approx \sqrt{\alpha}\,f_{s2}. \tag{7}$$

This estimation gives an idea about the decreasing of the distances between series resonances and corresponding parallel resonances. Its accuracy is illustrated in Fig. 5 for the case when α =1.03. The general conclusion from Fig. 5 is that formulas (7) give the upper boundary for f_{p1}' and the lower boundary for f_{p2}' in most of the cases. This conclusion is not valid only when both resonators have significantly different parallel capacitances (γ = 0.5 and less) and when the ratio between the series resonances is relatively large.

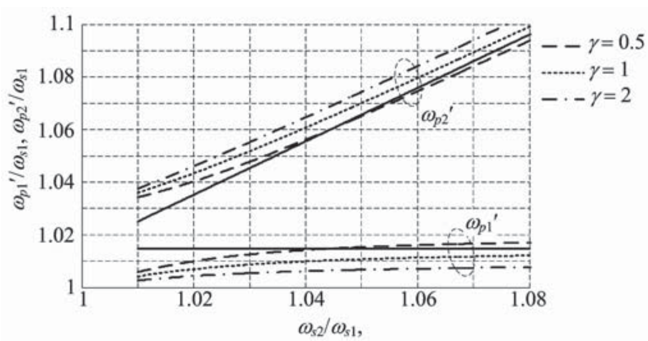

Fig. 5. Dependence of the new parallel resonances from the ratio of the series resonances. The continuous lines give the estimation according (7). The parameter α = 1.03 (effective coupling factor of 7%).

It is more convenient to consider the series connected resonators by using of their other equivalent circuit, consisting of parallel LC tank in series with a capacitor (Fig. 3(b)). It is based on Foster 1 form of LC impedances [12]. Fig. 3(b) shows immediately that the parallel resonances of the combination are the same as the parallel resonances of the individual FBARs. The series resonances change and move closer to the parallel resonances. The second resonator is equivalent to a capacitor C_{eq2} in the frequency band, where the first resonator behaves like inductor. The capacitances C_{s1} (in Fig. 3(b)) and C_{eq2} are in series, which reduces the total series capacitance and gives higher first series resonance frequency, closer to f_{p1}. The other series resonance moves closer to f_{p2} due to similar reason.

The equation for the new series resonance frequencies is derived by zeroing of the sum $Z_1 + Z_2$, where Z_1 and Z_2 are the

FBAR impedances, i.e. $Z_1 = 1/Y_1$ and $Z_2 = 1/Y_2$. This equation is the same as the equation (4) for the parallel resonances of parallel connected FBARs. The explanation of this interesting peculiarity is: eq. (4) gives two solutions ω_1' between ω_{s1} and ω_{p1} and ω_2' between ω_{s2} and ω_{p2}. When the resonators are in parallel, then the series resonances ω_{s1} and ω_{s2} are fixed and ω_1' and ω_2' give the new positions of the parallel resonances ω_{p1}' and ω_{p2}'. When the FBARs are in series, then ω_{p1} and ω_{p2} do not change and ω_1' and ω_2' are the new series resonances ω_{s1}' and ω_{s2}'. Then it follows from (7), that series and parallel connection of two FBARs give approximately the same ratio between the higher parallel resonance and the lower series resonance.

These considerations can be continued for parallel or series connections of few resonators. The general conclusions are:

- the parallel connection keeps the series resonances;
- the series connection keeps the parallel resonance;
- the distances between corresponding pairs of series and parallel resonances get smaller.

III. Filter With Parallel Connected FBARs

The effect of the parallel connected resonators will be considered for the filter, proposed in [11], which basic circuit is shown in Fig. 6(a). The first amplifier is assumed as transconductance amplifier; however other options are also possible. The resistors R_a and R_b represent the total resistances at the amplifier output and could be external resistors, amplifier output resistances or combination of both. The second amplifier has zero or very low input impedance and it will be considered as transresistance amplifier.

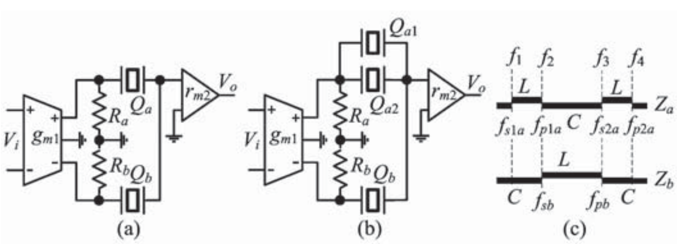

Fig. 6. (a) The basic circuit of the considered filter; (b) its modification with two FBARs on one of the arms; (c) location of the resonance frequencies of the impedances in the arms of the circuit in (b).

If Z_a and Z_b are the impedances of Q_a and Q_b respectively, then the filter transfer function is

$$T_v = \frac{V_o}{V_i} = g_{m1} r_{m2} R_a R_b \frac{Z_b/R_b - Z_a/R_a}{(Z_a + R_a)(Z_b + R_b)} . \quad (8)$$

The formula is valid also for the circuit in Fig. 6(b), which will be investigated below in more details. Then Z_a is the total impedance of the parallel connected Q_{a1} and Q_{a2}. The transfer function of the lattice filter in Fig. 1(b) is given also by (8) if accept $R_a = R_b = R$ and $g_{m1}r_{m2} = 1$. Thus the most of the results derived below for the circuit in Fig. 6(b) are valid also for the lattice filter when Q_a consists of two FBARs in parallel.

The requirement for opposite signs of the reactances of both arms in Fig. 6(b) sets the series resonance of Z_b to be equal to the first parallel resonance of Z_a and the parallel resonance of Z_b to be the same as the second series resonance of Z_a (Fig. 6(c)). The designations of the frequencies in Fig. 6(c) will be used in the following considerations. The resonances of Q_b should be chosen first and then f_{sa2} is defined also by f_3. The series resonance of Q_{a1} can be determined by iterative numerical technique in a way, which equalizes f_{pa1} to f_{sb}. If all FBARs have equal effective coupling coefficients, then (7) gives that the ratio f_4/f_1 is approximately α^2. The same is the ratio between the boundaries of the frequency band, in which Q_a and Q_b in Fig. 6(a) have reactances with opposite signs. Thus the extra resonator in Fig. 6(b) does not extend the theoretical passband bandwidth of the filter and one of the goals in the next considerations is to determine how to use this bandwidth more effectively.

The circuit in Fig 6(b) has three frequencies of maximum gain, equal to $g_{m1}r_{m2}$. These frequencies are located between f_1 and f_4 and they are f_2, f_3 and the frequency f_0, at which is satisfied

$$R_a R_b = Z_a Z_b . \quad (9)$$

The impedance Z_a tends to $j\infty$ at f_2; the other impedance Z_b tends to $j\infty$ at f_3. Then the fractional part together with R_aR_b in (8) is equal to 1 and the filter gain is equal to $g_{m1}r_{m2}$. The frequencies f_2 and f_3 are fixed by Q_b and only the frequency f_0 is necessary to be determined properly. The expressions for Z_a and Z_b are

$$Z_a = \frac{(\omega_1^2 - \omega^2)(\omega_3^2 - \omega^2)}{j\omega C_a(\omega_2^2 - \omega^2)(\omega_4^2 - \omega^2)}; \; Z_b = \frac{(\omega_2^2 - \omega^2)}{j\omega C_b(\omega_3^2 - \omega^2)}, \quad (10)$$

where C_a is the sum of the parallel capacitances in the mBVD models of Q_{a1} and Q_{a2} and C_b is the same capacitance of Q_b. It follows from (10) that the product Z_aZ_b starts from 0 at f_1 and monotonically increases to infinity when the frequency approaches f_4 (Fig. 7(a)). Thus the product R_aR_b could be any positive value and its choice determines f_0.

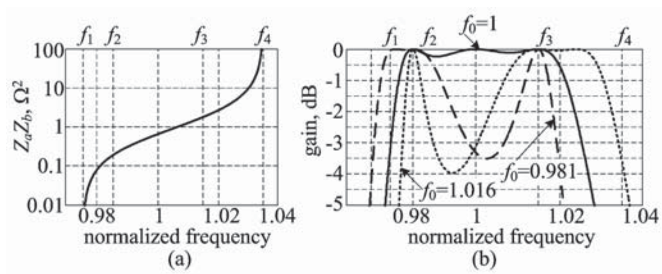

Fig. 7. (a) The product Z_aZ_b vs. frequency; (b) filter passband at different values of f_0. The frequencies are normalized to the geometric mean of f_2 and f_3 and the normalization resistance is chosen so that C_a = C_b = 1F. The resonators have effective coupling factor of 7% (α = 1.03), which gives f_1 = 0.97567, f_2 = 0.98533, f_3 = 1.0149 and f_4 = 1.0351.

The filter frequency response has failures between the f_2, f_3 and f_0. They depend on the distances between these frequencies and the choice of f_0 can minimize them. The intuitive guess about the proper place of f_0 is between f_2 and f_3 and Fig. 7(b) confirms it. Fig. 7(b) is plotted for the case when α = 1.03, however similar plots at different values of α confirm the basic conclusion: the drop of the frequency response between f_2 and f_3 is large (few dB) when f_0 is outside of the interval (f_2, f_3). The optimal position of f_0 is that, which equalizes the minima of the filter gain between f_2 and f_0 and f_0 and f_3. The nonsymmetrical filter frequency response makes difficult to find exact formula for positioning of f_0 and the best position could be done by numerical optimization. The variation of Z_aZ_b between f_2 and f_3 is relatively large (8-9 times) and does not require high accuracy for R_a and R_b to fix the desired value of f_0.

The stopband filter behavior depends also from the zeros of the filter transfer function. If express the impedances in (10) in Laplace transform domain ($-\omega^2 = s^2$) then the following equation for the zeros can be derived from (8):

$$sc\left(s^2+\omega_2^2\right)^2\left(s^2+\omega_4^2\right)-s\left(s^2+\omega_1^2\right)\left(s^2+\omega_3^2\right)^2=0\,, \qquad (11)$$

where

$$c=\tau_a/\tau_b\,;\ \tau_a=C_aR_a;\ \tau_b=C_bR_b\,. \qquad (12)$$

Equation (11) is of 7th degree and contains only odd degrees of s – thus it has a zero in the origin. The denominator of the filter transfer function (8) is of 8th degree. Thus the filter has a zero in the origin and a zero in the infinity. Equation (11) depends on two parameters: the ratio c from (12); and the ratio α of the FBAR resonance frequencies, which defines ω_1, ω_2, ω_3 and ω_4. The numerical investigation for values of α between 1.02 and 1.035 ($k_{t\,eff}^2=4.75-8\%$) and variation of c between 0.1 and 10 (in logarithmic scale) show that (11) has two conjugated imaginary zeros and four complex zeros for most of the values of c. The complex zeros have quadrant symmetry and they are located very close to the imaginary axis when $c < 0.8$ or $c > 1.2$ (quality factors of the zeros vary between 25 and 200). The complex zeros move far from the imaginary axis when c approaches unity from below or from above. The coefficient at highest degree of s in (11) is 0, when c = 1, and it can be proved, that then the lowest degree coefficient is also 0. Thus the filter transfer function has three zeros in the origin and three zeros in the infinity when c = 1. There is a narrow interesting region $1 < c < 1.0235$ (approximately not depending on the α when α is in the considered interval 1.02-1.035), in which the complex zeros transform into two different pairs of pure imaginary zeros. This region is interesting because the imaginary zeros can increase the stopband filter attenuation and dependence of their frequencies from c is shown in Fig. 8(a). Fig. 8(b) shows the dependence of the frequency of the pure imaginary zero, which exists always, from the parameter c.

The effect of the different positions of the zeros, respectively of different values of c, is illustrated in Fig. 9. Fig. 9(a) compares the filter frequency response at three significantly different values of c and similar comparison is done in Fig. 9(b) for values of c close to 1. Evidently, the cases when c is far from 1 must be discarded due to the low stopband attenuation. Fig. 9(b) shows that variation of c by ±5% has small effect in the region, where the attenuation does not exceed 25 dB. However, if higher stopband attenuation is required (in the range of 40 dB), then the limits are significantly tighter: between 1 and 1.0235. The parameter c is product of the ratios of the capacitances C_a and C_b and of the resistances R_a and R_b. A 5% tolerance is not difficult to achieve and more problematic is the 1% tolerance needed for higher attenuation. The smaller tolerance for c can be possible if the resistors R_a and R_b are tuned. The tuning could be done in different ways: by realizing of R_a and R_b as digitally controlled resistor banks; by controlling of the output impedances of the first amplifier; etc.

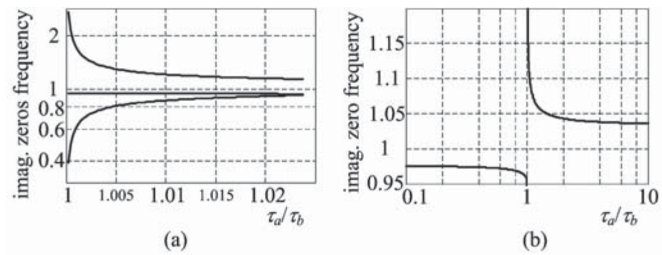

Fig. 8. (a) The three imaginary zeros vs. $c = \tau_a/\tau_b$ in the region, where they exist; (b) the frequency of the always existing imaginary zero from c. All plots are for α = 1.03 (effective coupling coefficient of 7%).

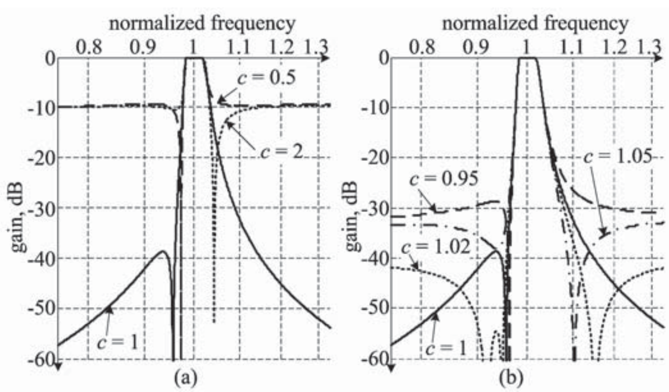

Fig. 9. Comparison of stopband frequency response of the filter at different values of c: (a) large differences of the values; (b) values of c close to 1. The normalizing frequency is the center of the passband.

The most of the conclusions above are valid also for the lattice filter in Fig. 1(b). There are differences only about the resistors R_a and R_b. They are the terminating resistors in the lattice filter and they must be equal. Thus, they can not be used for tuning the parameter c in the desired limits. Also, their resistance is usually prescribed and the satisfying of (9) should be achieved by proper choice of C_a and C_b.

IV. Simulation Based Comparisons with Other FBAR Filters

The effect of the extra resonator can be seen and estimated better by comparison of the frequency response of the new filter in Fig. 6(b) with the frequency response of the filter in Fig. 6(a), having one FBAR in each arm. The necessary frequency responses can be simulated by PSpice. The resonators are replaced by their mBVD models and the amplifiers are replaced by corresponding controlled sources. The values of the model elements, used in the simulations, are determined by using of the measured data given in [14]. These values are used for resonator Q_b in Fig. 6(b). The resonance frequencies of Q_{a1} and Q_{a2} are determined by applying of the derived recommendations (assuming c = 1). The element values of Q_{a1} and Q_{a2} are determined by proportional change of the values of Q_b in order to receive the desired resonance frequencies and to keep the same Q-factors of the resonators. This approach is applied also when determining the model elements of the resonators in the filters, used for comparison with the filter in Fig. 6(b).

TABLE I
PARAMETERS OF THE RESONATORS USED IN THE SIMULATIONS.

		L_m, nH	C_m, fF	C_0, pF	R_m, Ω	R_s, Ω	R_0, Ω	f_s, GHz	f_p, GHz
Fig. 6(b)	Q_{a1}	141.6	39.35	0.74	2.072	1.614	0.4	2.132	2.188
	Q_{a2}	132.15	39.35	0.74	2	1.56	0.4	2.207	2.265
	Q_b	69.59	78.7	1.48	1.027	0.8	0.2	2.151	2.207
Fig. 6 (a)	Q_a	70.5	79.73	1.5	1.026	0.8	0.197	2.123	2.179
	Q_b	68.69	77.69	1.46	1.026	0.8	0.203	2.179	2.236
Fig. 1 (a)	Q_a	68.69	77.69	1.46	1.027	0.8	0.203	2.179	2.236
	Q_b	70.5	79.73	1.5	1.027	0.8	0.197	2.123	2.179

This filter is compared with the filter in Fig. 6(a) and also with a ladder filter with the same number of resonators: two series and one shunt resonator. Both filters are designed to have similar passband central frequency as the filter in Fig. 6(b). The Q-factors of all resonators are about 500 – the same as in [14]. The values of the resistors R_a and R_b vary between 25Ω and 40Ω after their adjustment for every circuit separately in order to achieve wider passband. The parameters and the model elements for all resonators used in the simulations are summarized in Table 1.

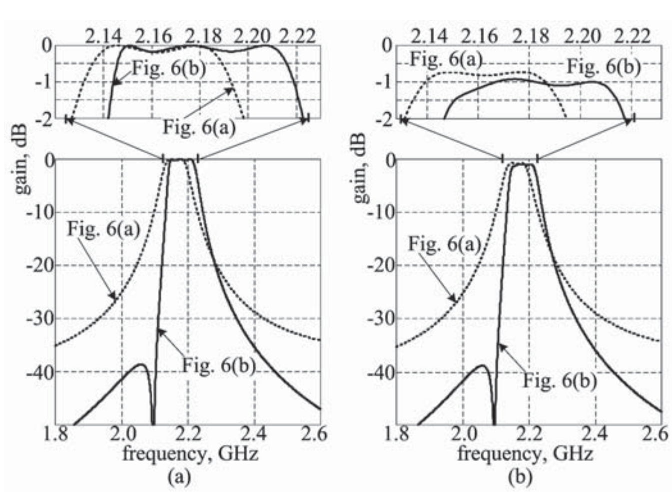

Fig. 10. Frequency responses of filter with three resonators (Fig. 6(b)) and filter with two resonators (Fig. 6(a)): (a) without including the FBAR losses; (b) FBAR losses are included.

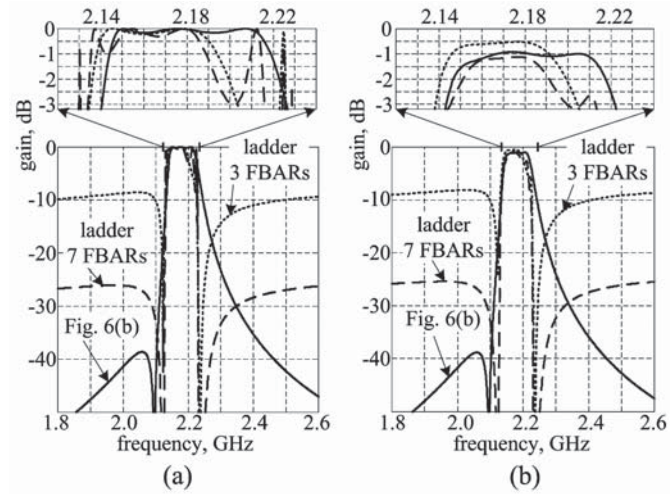

Fig. 11. Comparison of the frequency responses of the filter in Fig. 6(b) with the frequency responses of ladder filters with 3 FBARs and with 7 FBARs: (a) FBAR losses are not included; (b) FBAR losses are included.

Fig. 10 and Fig. 11 illustrate well that the extra resonator in Fig. 6(b) gives steeper slopes and higher attenuations in the stopband. The width of the passband (measured at -0.3dB level) for this circuit is 67.6MHz at 2.1801GHz passband central frequency (3.1% relative bandwidth) when the FBAR losses are not taken into account (Fig. 10(a)). The same data for Fig. 6(a) are: 49.6MHz bandwidth at 2.1622GHz central frequency (2.3% relative bandwidth). Thus the extra resonator gives a 35% extension of the relative bandwidth. However, this is the upper limit for the passband extension since the losses deteriorate more the frequency response of the filter in Fig. 6(b). Fig. 10(b) compares the frequency responses when the losses are included and the same data are: 53MHz absolute and 2.4% relative bandwidth at 2.1849GHz central frequency for the circuit in Fig. 6(b); 47.6MHz absolute and 2.2% relative bandwidth at 2.1632GHz central frequency for the circuit in Fig. 6(a). The increasing of the bandwidth is 10% only. This is a rather pessimistic estimation because Q-factor of 500 is around the lower boundary for Q [6]. The simulation at Q = 1000 (all resistors values in the mBVD model are divided by 2) gives relative bandwidths of 2.84% for Fig. 6(b) and 2.24% for the circuit in Fig. 6(a) – now the relative difference is 26%. Another advantage of the circuit with an extra resonator is the existence of a zero in the lower stopband which significantly improves the attenuation in the stopband right next to the passband. This can be very useful in cases where a transmit channel is situated right below the receive channel with little spacing.

The frequency response of the circuit in Fig. 6(b) is compared in Fig. 11 with the frequency responses of two ladder filters – one with 3 resonators and one with 7 resonators. The plots confirm the basic disadvantage of the ladder filters: more resonators are necessary for achieving reasonable stopband attenuation, but an increase in the number of resonators leads to narrower passband. The attenuation and passband data from the simulations is summarized in Table 2. The comparison of the considered filter with lattice filter with 7 resonators shows, that the circuit in Fig. 6(b) has 80% wider passband and gives more than 10 dB better stopband attenuation.

TABLE II
NUMERICAL DATA EXTRACTED FROM THE SIMULATED FREQUENCY RESPONSES.

	Min. attenuation in the lower stopband, dB	Min. attenuation in the upper stopband, dB	Min. passband, attenuation, dB	Passband width at 0.3dB, MHz
Fig 6 (a), w/o losses	29.14*	27.87*	0	49.6
Fig 6 (a), w/ losses	29.43*	27.67*	0.82	47.6
Fig 6 (b), w/o losses	38.6	35.2*	0	67.6
Fig 6 (b), w/ losses	38.79	35.94*	0.916	53.1
Fig 1 (a), 3 resonators, w/o losses	8.58	6.16	0	46.7
Fig 1 (a), 3 resonators, w/ losses	8.14	6.52	0.54	40.5
Fig 1 (a), 7 resonators, w/o losses	26.1	22.6	0	34.6
Fig 1 (a), 7 resonators, w/ losses	25.4	23.03	1.13	29.5

* The stopband attenuation is measured at frequencies which are at ±10% from the passband center frequency.

V. CONCLUSIONS

The considerations in the paper show that the replacement of the single FBARs in the arms of the lattice filters by two parallel or series connected FBARs can give better frequency response of the filter. The improvement is twofold: significantly better selectivity due to higher order of the transfer function and a moderate extending of the passband. It can be achieved when the parallel or series connected resonators have different parameters, satisfying some conditions. These conditions are investigated in details for the modification of the lattice circuit, proposed in [11], which basic version has two resonators instead of four. In order to avoid large increasing of the circuit complexity, two parallel FBARs are placed only in one of the arms. The asymmetry between the arms of the circuit reflects in an asymmetric frequency response. The considerations in the paper show that the modified lattice circuit from [11] proposes two basic advantages in the considered case: only one resonator is necessary to be added instead of two as it is in the basic lattice filter; and the tuning mechanism of modified circuit allows easier satisfying of the derived design conditions.

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