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SYNTHESIS AND INVESTIGATION OF THREE-SECTION MICROSTRIP FILTER ON FOLDED DUAL-MODE STEPPED-IMPEDANCE RESONATORS

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Abstract—A new structure of an ultra-wide bandpass filter is considered. An intelligence method of optimization is applied first for synthesis of three-section filter on dual-mode resonators. An area of fractional bandwidth values and values of dielectric constant of the substrate, where synthesis problem for the filter has a solution, is determined theoretically. Experimental results show good agreement with simulated values.

1. INTRODUCTION

Filters on dual-mode resonators differ from usual ones in a circumstance that two oscillation modes per resonator rather than one take part in forming the pass band. Such filters with the same selectivity require half amount of resonators. Therefore, they have smaller size and lower insertion loss in the pass band.

Structures of microstrip filters with symmetric shape of the resonators are widely considered in literature [1–3]. The resonator conductor, for example, may be square, equilateral triangle or loop of the same shape, built of rectilinear or meander-line segments. Modes of degenerate oscillations in such resonators superpose at rotation round the symmetry axis. In order to split resonant frequencies, a proper coupling element, breaking symmetry, is inserted into the resonator. Dual-mode resonators of symmetric shape, due to spatial separation of voltage antinodes for degenerate modes, may be considered as two single-mode resonators with their own resonant frequencies, couplings, and ports. Therefore, design of filters on dual-mode resonators of

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symmetric shape may be performed using the same methods as for design of filters on single-mode resonators.

More miniature microstrip filters are filters on quasi-one-dimensional dual-mode resonators, whose width may be much less than length [4, 5]. Degenerate oscillation modes do not exist in such resonators. Voltage antinode of the first mode in the resonator coincides with one of antinodes of the second mode. Therefore, design of filters on such resonators is more difficult than design of filters on single-mode resonators. Microstrip filters on tetra-mode resonators are still more miniature [6].

Systematic investigation of microstrip filters on dual-mode resonators are bounded for lack of effective optimization methods. A number of commercial microwave software is provided with utilities of standard universal optimization. These utilities are good for synthesis of simple microstrip filters on single-mode resonators. However, their efficiency sharply drops with increase of structure complexity and at poor choice of initial values for the structure parameters.

An intelligence method, using a priori physical knowledge about all types of frequency-response distortions in the pass band, resulting from mistuning of resonant frequencies and coupling of resonators, differs in high efficiency while in optimization of microstrip filters. This method is used successfully in the expert system Filtex32, built up for design and investigation of strip and microstrip filters [7].

The rules of intelligence optimization were worked out first only for filters on single-mode resonators [8, 9]. Hints how to adjust those rules for filters on dual-mode resonators were formulated later [10]. Operators for correction of frequency-response distortions of a two-section ultra-wide bandpass microstrip filter on folded stepped-impedance resonators were built there also. Software, built on their base, had allowed carrying out wide investigation of capability of the filter structure [11].

In this paper, a new structure of three-section ultra-wide bandpass microstrip filter on folded stepped-impedance resonators is considered. Application of an intelligence optimization method for synthesis of this filter is described. Properties of the filter are investigated with specially built software. Simulated amplitude-frequency response is compared with experimentally measured response.

2. FILTER STRUCTURE AND SIMULATION MODEL

Microstrip filter consists of three parallel resonators, coupled electromagnetically along their length. Shape of their strip conductors is shown in Fig. 1. A conductor of the central resonator is closed to

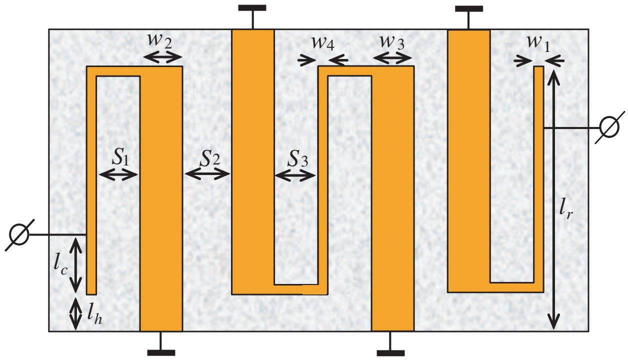

Figure 1. Microstrip filter layout.

the screen at both ends, whereas conductors of the terminal resonators are closed only at one end. Conductors of the terminal resonators have a tapping for connecting to input and output lines.

Spectrums of resonant frequencies for all resonators in the filter are not equidistant because of stepped width and folded shape of their conductors. Frequencies of two first oscillation modes are closer and both the modes are involved into forming their joint pass band.

For calculation of frequency response of the filter a one-dimensional model is used, in which phase of oscillation may change only along the strip conductor and not across it. The model contains connected sections of single and coupled microstrip lines. Only traveling fundamental waves are considered in an explicit form. Their amplitudes are found by means of solving Kirchhoff's equations for all nodes. Parameters of fundamental waves are computed in the quasi-TEM approximation. Effects of higher-order modes, localized on irregularities, are simulated by introducing effective capacities [12]. Frequency dispersion of effective dielectric constant is taken into account within an approximate analytical model [13]. Absorption of microwave power is specified with resonator unloaded Q-factor.

Figure 2 shows typical amplitude-frequency response. It answers to a 6-pole filter. On a return loss curve $R(f)$ one can see minimums of reflected power at six frequencies of coupled oscillations, between which five maximums R_1, R_2, R_3, R_4, and R_5 are disposed. In a tuned filter, these values have to coincide with the prescribed value of maximum reflection $R_{\max}$. The other parameters, specifying a pass band, are low-edge frequency f_l and high-edge frequency f_h, defined for a specified level of attenuation loss $L(f)$. Thus a current pass band of the filter we can describe with seven parameters R_1, R_2, R_3, R_4,R_5, f_l, and f_h.

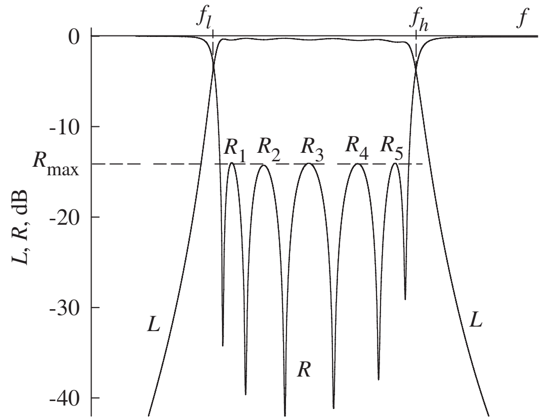

Figure 2. Typical amplitude-frequency response.

Goal of intelligence optimization of the filter is that all seven passband parameters should possess the prescribed values. It is necessary to adjust at least seven structure parameters for that. The rest of structure parameters are reserved for other possible subsequent optimizations, e.g., minimization of device dimensions. As adjustment parameters, we choose l_r, l_c, S_1, S_2, S_3, w_1, w_2, w_3, and w_4 (Fig. 1).

3. INTELLIGENCE OPTIMIZATION

In intelligence optimization of n-pole filter, as an objective function is used $n + 1$-dimensional deflection vector $\mathbf{D}$. Every its component D_i $(i = 1, 2, \ldots, n+1)$ describes its own distortion of amplitude-frequency response in the pass band. The more its absolute value is, the more distortion of ith type will be.

Filter optimization consists in by-turn elimination of ith type of distortion, corresponding to a component D_i with maximum absolute value. An own correction operator, fulfilling tuning of resonant frequencies and couplings between resonators in the filter, is used for every distortion type. Resolution of all return-loss peaks $R_1 \ldots R_5$ in the pass band is a necessary condition for feasibility of optimization.

First three components of deflection vector $\mathbf{D}$ are defined by formulas

$$D_1 = (f_c - f_0)/f_0, \tag{1}$$
$$D_2 = (\Delta f_c - \Delta f_0)/\Delta f_0, \tag{2}$$

$$D_3 = \sum_{i=1}^{n-1} (R_i - R_{\max})/(n-1), \tag{3}$$

where f_0 and f_c are required and current values of passband center frequency, Δf_0 and Δf are required and current values of passband width. Definitions of other components of vector $\mathbf{D}$ depends on n.

Other components in a six-pole filter ($n = 6$) are defined by formulas

$$D_4 = (R_1 + R_2 - R_4 - R_5)/2, \tag{4}$$
$$D_5 = (R_1 + R_4 - R_2 - R_5)/2, \tag{5}$$
$$D_6 = R_3 - (R_1 + R_5)/2, \tag{6}$$
$$D_7 = (R_1 + R_3 + R_5)/3 - (R_2 + R_4)/2. \tag{7}$$

Physical meaning of the correction operators is well known. They are operators, adjusting simultaneously resonant frequencies and couplings for certain resonators.

Rules for building of correction operators, not attached to a concrete structure, were formulated first for filters on single-mode resonators [8, 9]. However, those rules may be applied for filters on dual-mode resonators. That needs every real dual-mode resonator in the filter to be compared with proper imagine pair of coupled single-mode resonators, which frequencies of coupled oscillations coincide with resonant frequencies of real dual-mode resonator [10]. Interaction between imagine single-mode resonators has an important peculiarity. Their coupling coefficient $k(S)$ is increasing function of spacing S [10].

We suppose that structure of a filter, containing imagine pair of single-mode resonators, is symmetrical. So it is convenient to number imagine resonators and their coupling coefficients in two directions simultaneously, i.e., from terminal resonators to central resonator.

The first correction operator, conjugated to the component D_1, fulfills equal tuning of all resonant frequencies of imagine resonators, i.e., equal adjustment of f_1, f_2, f_3. Such adjustment is exercised approximately with correction of resonator length l_r.

The second operator, conjugated to D_2, performs proportional correction of the coupling coefficients of all pairs of imagine adjacent resonators, i.e., k_1, k_2, k_3. At $D_2 > 0$ such result is achieved approximately with widening of the inter-resonator spacing S_2 and simultaneous narrowing of the intra-resonator spacings S_1 and S_3.

The third operator at $D_3 > 0$ increases in first approximation a coupling of the terminal resonators with input and output transmission lines. Such result is attained with increase of the distance l_c, i.e., moving the tapping point away from the open end of the terminal resonator.

The fourth operator at $D_4 > 0$ lowers resonant frequency of the terminal imagine resonators f_1 and heightens simultaneously resonant frequencies of internal imagine resonators f_2, f_3 without changing their sum. That means increment of both resonant frequencies of the real internal resonator and changing a ratio of two first resonant frequencies of a real terminal resonator. Obviously, that needs in narrowing w_3, w_4 and simultaneous changing a ratio w_1/w_2. Results of structure simulation suggest action that is more concrete. That is narrowing w_3, w_4 to an equal degree and widening w_4 to a lesser degree.

The fifth operator at $D_5 > 0$ heightens resonant frequency of two central imagine resonators f_3 and lowers simultaneously resonant frequency of two other internal imagine resonators f_2 without changing their sum. That means heightening of both resonant frequencies of the real internal resonator and simultaneous lowering of one of two resonant frequencies of real terminal resonators. That needs in narrowing w_3, w_4 and simultaneous changing the ratio w_1/w_2. Results of structure simulation suggest again more concrete action. That is narrowing w_3, w_4 to an equal degree and widening w_2 to a lesser degree.

The sixth operator at $D_6 > 0$ decreases the coupling coefficient k_1 and simultaneously increases the coupling coefficients k_2 and k_3 to a lesser degree. That is achieved by narrowing the spacing S_1 to a greater degree and simultaneous narrowing the spacing S_2 and widening the spacing S_3 to a lesser degree.

The seventh operator at $D_7 > 0$ decreases the coupling coefficient k_3 and simultaneously increases the coupling coefficients k_2 and k_1 to a lesser degree. That is achieved by narrowing the spacing S_3 to a greater degree and simultaneous narrowing the spacing S_2 and widening the spacing S_1 to a lesser degree.

After several correction operations, one of the spacings S_1 or S_2 may be undesirably narrow, whereas another may be sufficiently wide. In this case, it is necessary to apply an additional correction operator. It simultaneously changes S_1 and $1/S_2$ to an equal degree and changes w_2 to a lesser degree without a considerable impact on the frequency response.

All correction operators, conjugated to components of the vector $\mathbf{D}$, are quasi-orthogonal. That means ith operator, removing ith distortion of frequency response, does not produce other distortions, whose values would be commensurable with value of an eliminated distortion. Exclusion is the third operator. However, its non-orthogonality may be easy eliminated with applying of other correction operators.

The rules of intelligence optimization have been applied in created software for synthesis of the considered filter, connectable to the expert system Filtex32.

4. COMPARISON OF THEORY WITH EXPERIMENT

In order to estimate an accuracy of the computational model, an unshielded microstrip filter on a substrate of microwave ceramics TBNS with dielectric constant $\varepsilon_r \approx 80$ was fabricated. The filter pass band in the performance specification was defined by the midband frequency $f_0 = 0.26$ GHz, fractional bandwidth $\Delta f_0/f_0 = 90\%$ measured at level -3 dB relative to minimum loss level, and maximum return loss level $R_{\max} = -14$ dB. The other specified parameters were the substrate thickness $h_d = 2$ mm, distance $l_h = 2$ mm, and input/output impedance $Z_0 = 50$ Ohm. For the avoidance of narrow spacings, an additional condition $S_1 \approx S_2$ was imposed.

Because of synthesis, other structure parameters have gotten the next values in millimeters: $l_r = 41.25$, $l_c = 1.72$, $w_1 = 0.54$, $w_2 = 0.80$, $w_3 = 0.49$, $w_4 = 0.31$, $S_1 = 0.30$, $S_2 = 0.28$, $S_3 = 2.56$. The substrate at these values has dimensions $45.25 \times 14.24 \times 2.00\,\text{mm}^3$.

In Fig. 3, measured curves of frequency response for the manufactured filter are compared with the computed curves. It is observed a good agreement. That allows applying the chosen simulation model for carrying out theoretical investigation of the filter.

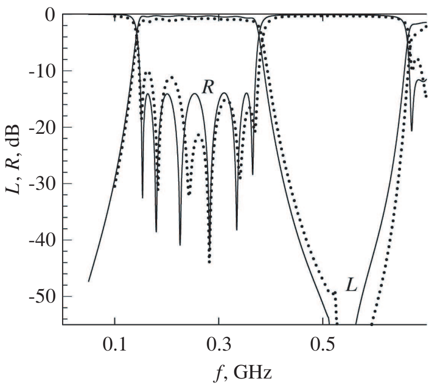

Figure 3. Simulated (solid lines) and measured (dotted lines) ampli-tude-frequency responses.

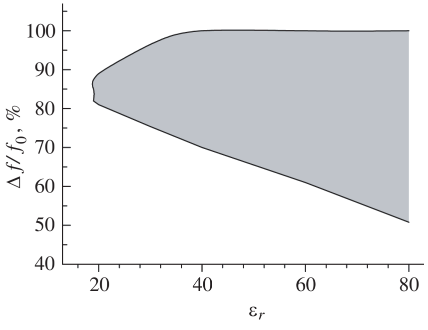

Figure 4. Domain of existence of solution for the synthesis problem.

5. INVESTIGATION OF THE FILTER

Created software allowed conducting theoretical investigation of filter properties. It was ascertained that a synthesis problem for the concerned filter structure has a solution not for any value of fractional bandwidth $\Delta f/f_0$ and not for any value of dielectric constant ε_r. Furthermore, not any solution may be realized practically. In particular, it is difficult to make a narrow spacing between strip conductors, whose value is less than metallization thickness. Therefore, we do not consider those solutions, where at least one of spacings S_1, S_2 or S_3 is less than 0.01 mm or one of widths w_1, w_2, w_3, and w_4 is less than 0.10 mm.

Figure 4 shows an area, within which the synthesis problem has an admissible solution. Outside of the area either the applied restrictions are broken, or $l_c = 0$ and an additional increment of coupling of terminal resonators with input and output lines is needed. The area is built for a case, when the center frequency $f_0 = 0.8$ GHz and the substrate thickness $h_d = 1$ mm.

It is seen that the filter may be realized on a substrate only with dielectric constant $\varepsilon_r \geq 20$. The lower boundary of a realizable pass band varies in limits from 50% to 80% depending on ε_r. The upper boundary varies from 90% to 100%. Thus, the structure under study is good for manufacture only broadband and ultra-broadband bandpass filters.

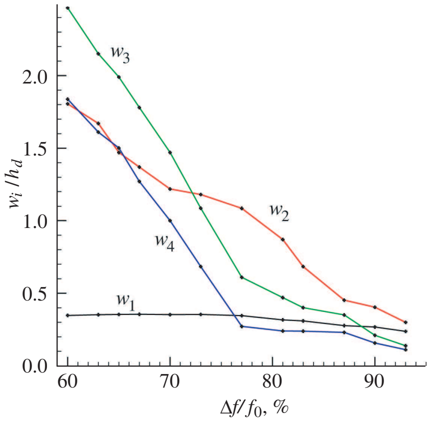

Figure 5. Strip conductor widths versus fractional bandwidth.

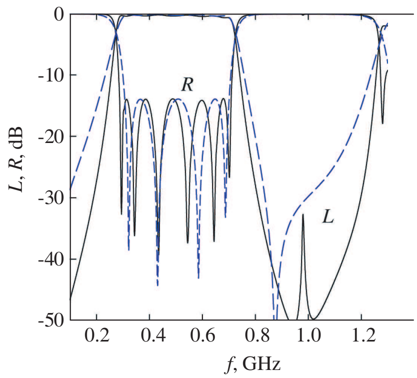

Figure 6. Amplitude-frequency responses of tree-section (solid lines) and two-section (dashed lines) filters.

The midband frequency of filter f_0 may vary in a wide range. The filters on a substrate of ceramics TBNS with standard dimensions are able to overlap a frequency band from 0.2 GHz to 2.7 GHz. Here the resonator length l_r varies in a range from 55 mm to 3.3 mm.

Figure 5 shows dependences of widths of filter strip conductors versus fractional bandwidth. Computation was carried out for $S_1 = S_2$, $\varepsilon_r = 80$, $f_0 = 0.8\,\mathrm{GHz}$, $h_d = 1\,\mathrm{mm}$. It is seen that width steps of strip conductors in irregular resonators decrease with increase of bandwidth.

Casual behavior of the curves is due to non-uniqueness of solution of the synthesis problem. Non-uniqueness arises from a circumstance, in which 9 necessary tuning parameters are responsible for keeping 8 equalities only. Therefore, the tuning parameters have one degree of freedom.

Figure 6 shows a comparison between amplitude-frequency responses of three-section and two-section filters, having the same pass band. Structure of the two-section filter on dual-mode resonators was considered in [4, 5]. It differs from structure of the three-section filter under review by absence of the central resonator. Synthesis of both filters was carried out for the substrate with $\varepsilon_r = 80$. The pass band was specified with values $f_0 = 0.5\,\mathrm{GHz}$, $\Delta f/f_0 = 90\%$, and $R_{\max} = -14\,\mathrm{dB}$. The substrate of the two-section filter took dimensions $23.31 \times 4.13 \times 1\,\mathrm{mm}^3$, and the substrate of the three-section filter took dimensions $23.77 \times 6.63 \times 1\,\mathrm{mm}^3$.

It is seen that the three-section filter has steeper skirts and larger stopband attenuation. A weakness of the three-section filter is occurrence of a weak spurious peak of transmitted power in the upper stopband. This peak arises from the third oscillation mode of the central resonator. Its resonant frequency appears always below the third resonant frequency of the terminal resonators, since both ends of the central resonator are grounded. As regards the frequency response given in Fig. 3, here casual suppression of the spurious peak occurs due to the tapping point in the terminal resonator coincides with a voltage node at the frequency of the spurious peak.

Skirt selectivity of the filter was evaluated with steepness coefficients of low and high sides of the pass band, defined by formulas

$$K_l = \frac{\Delta f/2}{\Delta f_l - \Delta f/2}, \quad K_h = \frac{\Delta f/2}{\Delta f_h - \Delta f/2}, \tag{8}$$

where Δf is a passband width at a level of $-3\,\mathrm{dB}$, $\Delta f_l = f_0 - f_l$ and $\Delta f_h = f_h - f_0$ are widths of bands, measured from the midband frequency to low and high side of the pass band accordingly at a level of $-30\,\mathrm{dB}$ relative to the minimum loss [11, 14]. The difference between the coefficients K_l and K_h allows estimating an asymmetry degree of amplitude-frequency response.

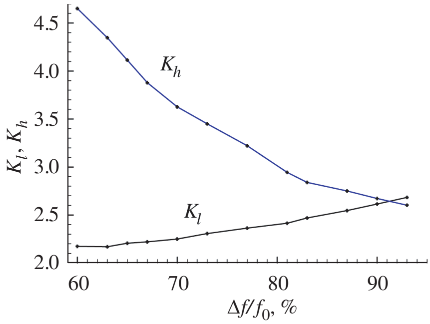

Figure 7. Steepness coefficients of passband skirts versus fractional bandwidth.

Figure 7 shows dependences of the steepness coefficients versus fractional passband width. In the most part of the area of the realizable bandwidth, the high side of the passband is steeper than the low side. An increase of the fractional bandwidth diminishes difference between the steepness coefficients of the low and high sides. The passband sides become symmetrical near the high edge of the realizable bandwidth area.

6. CONCLUSION

A new structure of a three-section mictostrip ultra-wide passband filter on folded dual-mode stepped-impedance resonators is described. A synthesis method, based on an intelligence optimization, is offered for it. Quasi-orthogonal correction operators are built for all distortion types of amplitude-frequency response of the filter.

Theoretical investigation of the filter is carried out with created software. An area of fractional bandwidth and value of substrate dielectric constant, for which practical realization of the filter is possible, is ascertained. It is proved fractional bandwidth of the filter on a substrate with thickness of 1 mm cannot be narrower than 50% and wider than 100%. Dielectric constant of the substrate must be not less than 20.

Dependences of steepness coefficients for low and high sides of the pass band versus the fractional bandwidth are investigated. A comparison of a computed amplitude-frequency response with a measured response for a fabricated structure proves adequacy of the computational filter model, used in software.

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