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ELECTRODYNAMICS AND WAVE PROPAGATION

Selective Properties of Microstrip Filters Designed on Quarter-Wave Codirectional Hairpin Resonators

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Abstract—Selective properties of microstrip filters containing two to six codirectional irregular quarter-wave resonators with hairpin-shaped stripline conductors are studied. The dependence of the steepness of the slopes of the amplitude–frequency responses of such filters on the design parameters and the filter fractional bandwidth is studied in the quasi-static approximation in a wide range of operating frequencies. Possibilities for obtaining not only a symmetrical amplitude–frequency response but also a response with the maximum steepness of either the low-frequency or the high-frequency slope are demonstrated. The results of the numerical analysis of 1D models of these microstrip structures agree well with experimental data.

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INTRODUCTION

As is known, microwave frequency-selective devices occupy a substantial volume in communications, radar, and radio-navigation facilities. As a rule, these devices, designed as systems of coupled resonators, often determine overall dimensions of individual modules and, sometimes, the entire setup. Moreover, the quality and the ultimate characteristics of radio equipment depend directly on the frequency-selective properties of filtering devices. Hence, the search for new solutions for the design of miniature filters and studies aimed at miniaturization and improvement of frequency-selective properties of known designs are very topical and among the most important problems of modern radio engineering.

It is well known that the most miniature "electromagnetic" filters are microstrip filters (MSFs), which are widely used in microwave devices because of the advantages of such electronic components. These filters have small dimensions and are reliable and easy-to-manufacture. The results of their high-speed quasi-static analysis of various and complex design versions agree well with experimental data [1–3]. This fact allows the development of efficient program systems for the computer-aided design of MSFs [4, 5].

The overall dimensions of microstrip filters can be reduced using several well-known approaches, such as the use of folded stripline conductors in resonators, formation of smooth and stepwise irregularities in these conductors, and application of substrates with a high value of permittivity ε. Evidently, the maximum effect can be attained by combining several methods for reduction of the MSF dimensions. In this case, rather miniature designs can be obtained even in the metric wave band [6, 7].

In order to improve frequency-selective properties of filters, first, we should increase the steepness of the slopes of the amplitude–frequency response (AFR), increase attenuation levels of the microwave power in the filter stopbands, and expand the widths of these bands. These characteristics depend on both the number of the filter resonators and many other design parameters of the device. Therefore, the study of selective properties of particular design versions with different numbers of resonators as functions of the frequency band, fractional bandwidth, permittivity of the substrate, and other design parameters is an important practical problem. Such studies allow (1) determination of the limits of applicability of the chosen design and (2) creation of optimized MSFs that satisfy particular specifications and contain the minimum number of sections [1, 8, 9].

In this paper, we investigate selective properties of microstrip filters with number N of resonators ranging from 2 to 6. The resonators are shaped as irregular hairpins (Fig. 1) with stripline conductors consisting of segments having different widths. These designs have very small dimensions due to the application of not only folded and stepped stripline conductors but also quarter-wave resonators. Filters designed on similar but oppositely connected resonators were studied in [9]. However, the manufacture of substrates for such filters requires high accuracy of preservation of the substrate thickness and width since the conductors of the resonators are connected to the screen on the opposite sides of the substrate; hence, the width of the substrate determines the length of the resonator. Therefore, it is of interest to study selective properties of multisection filters designed on codirectional resonators, because these designs are free of this disadvantage (see Fig. 1).

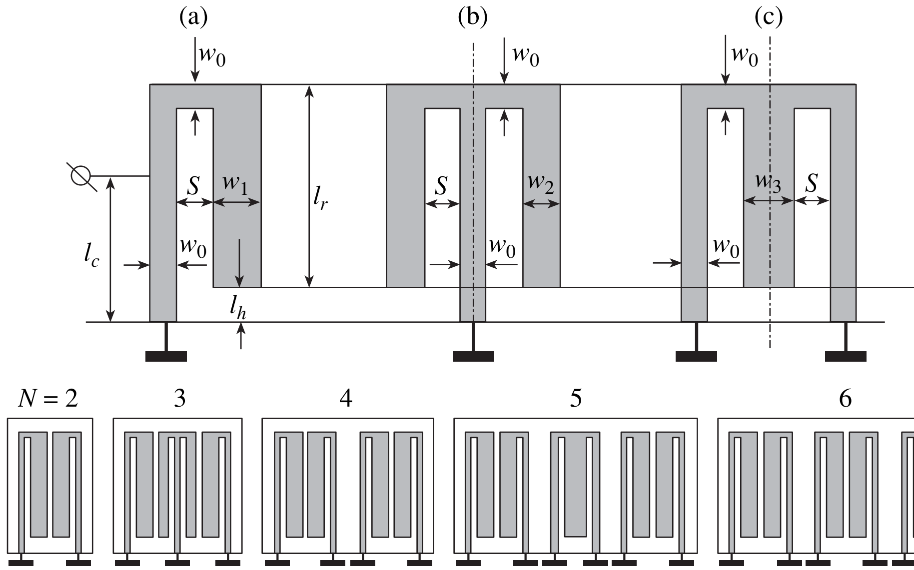

Fig. 1. Topology of conductors in irregular microstrip resonators and multisection filters designed on the basis of these resonators and having $N = 2$–6.

Note that, as in [9], if the number of sections N, is even, such designs contain identically shaped resonators (Fig. 1a) in which the free ends of the narrow parts of conductors are connected to the screen. In order to keep the mirror symmetry of the conductor topology in filters with an odd N, the center resonator has the mirror symmetry as well and consists of two hairpins connected to each other by either the segments connected to the screen (Fig. 1b) or the open-circuited segments (Fig. 1c).

It is important to note that, in these resonators, antinodes of the high-frequency current are situated at the points where the conductors are connected to the screen and antinodes of the high-frequency voltage are located at the opposite (free) ends of the stripline conductors. As a result, at frequencies of the first oscillation mode of this resonator, a narrow regular segment of the conductor of the resonator plays the role of a distributed inductance and a wide regular segment plays the role of a distributed capacitance. Therefore, by increasing the stepwise change of the width of the striplines, we not only decrease the overall dimensions of the resonators and the entire device but also increase the width of the high-frequency stopband of the filter substantially by shifting the second (parasitic) passband toward the high-frequency region [9].

1. THE MODEL AND THE CALCULATION TECHNIQUE

In the filters analyzed, the length of the coupling regions is the same for all resonators and equal to hairpin height l_r (see Fig. 1). Width w_0 of the narrow regular segments is also the same for all resonators, and the free ends of the conductors used in these sections reach the edges of the substrate and are connected to the screen exactly at these points. The wide regular segments of conductors do not reach the edges of the substrates and are separated from them by distance l_h. The width of these segments is w_1 for the outmost resonators, w_2 for the inner resonators nearest to them, and w_3 for the middle resonators used in five- and six-section filters. The width of the regular segment connecting a wide and a narrow conductor in a resonator, which are separated by spacing S, is w_0. External transmission lines with a wave impedance of 50 Ω are conductively connected to stripline conductors of the outmost resonators at a distance of l_c from the end connected to the screen.

The numerical analysis of all designs was performed using 1D models consisting of connected-in-series regular segments of isolated and multiply coupled lines whose linear parameters were calculated in the quasi-static approximation by means of a variational method [10]. The topology of the filters' stripline conductors was divided into regular segments by vertical and horizontal sections passing through the bends and ends of the conductors, as well as the points of the conductive connection of transmission lines. In the calculation, we took into account only the fundamental modes. Amplitudes of these modes and amplitudes of reflected and transmitted waves are coupled with the amplitude of the incident electromagnetic wave by a system of linear equations that can be obtained from the continuity condition for currents and voltages at the boundaries of conductors of all regular segments of the model. Dissipative losses of the microwave power were taken into account by introducing unloaded quality factor Q_0 of microstrip resonators whose value was taken from the experiment.

The analysis programs that were written for the studied filters allow calculation, in a specified frequency band, of the frequency characteristics of forward and return losses and the group delay of the transmitted wave for fixed design parameters of microstrip structures. The parameters of these structures are the geometric dimensions of conductors; thickness h and permittivity ε of the substrate; position l_c of the point of conductive connection of transmission lines to the outmost resonators; unloaded quality factor Q_0; and, if the resonator contains an upper screen, height h_a of this screen above the substrate surface. The analysis programs were connected to the Filtex32 expert system designed for the computer-aided parametric synthesis of filters with a specified characteristic in the passband. A prototype of this expert system was described in [5]. The specification for the design of such a device fixes only center frequency f_0 of the filter passband; the filter bandwidth, for example, at the half-power level (Δf_3); and the maximum level of reflections of the microwave power in the filter passband, L_R, at which all $N-1$ maxima of the return loss of the tuned filter should be placed [9].

Filters were synthesized using the optimum correction method [11]. In this method, for each filter, we select $N+1$ design parameters most strongly affecting the filter characteristics in the passband; these parameters will be tuned in the course of optimization. Such parameters are called tuning parameters and all remaining parameters are called basic parameters. For a two-section design, the tuning parameters are hairpin height l_r, which exerts the main effect on the center frequency of the filter passband, and spacing S_1 between the resonator conductors, which mainly affects the width of the filter passband. Position l_c of the point of the conductive connection of transmission lines to resonators is the last tuning parameter and determines the reflection level in the filter passband. All remaining parameters of the two-section filter are fixed in the process of optimization; i.e., they are basic parameters. A three-section filter contains an additional tuning parameter, width w_2 of the stripline conductor of the middle resonator, which affects the natural frequency of this resonator. This frequency depends also (although, to a lesser extent) on spacing S_1 between the outmost resonators. Here, resonance frequencies of all resonators must coincide with the center frequency of the filter passband. A four-section filter contains one more tuning parameter, spacing S_2 between the conductors of the middle resonators. This spacing establishes a balance between the "outer" and "inner" pairs of resonators. In a five-section filter, there is a sixth tuning parameter, width w_3 of the stripline conductor of the middle resonator, which tunes the natural frequency of this resonator. Finally, a six-section filter has an additional (seventh) tuning parameter, spacing S_3 between conductors of center resonators, which establishes a balance in the coupling between these sections and the remaining sections of the filter.

In the process of the automatic tuning of filters with the help of the Filtex32 system, optimization of tuning design parameters is assumed to be completed after fulfillment of the following three conditions. First, the deviation of the return loss maxima from –14 dB is no more than ±0.2 dB. Second, the width of the filter passband differs from the specified value by no more than ±1%. Third, the center frequency of the filter passband deviates from the specified value by no more than ±0.1% of the specified width of the passband. Owing to the latter condition, the accuracy of tuning of the center frequency increases as the filter bandwidth decreases.

Selective properties of synthesized filters were estimated from values of steepness coefficients K_l and K_h of the low-frequency and high-frequency slopes of the filter AFR [1, 6, 9],

$$K_l = \frac{\Delta f_3/2}{\Delta f_l - \Delta f_3/2} \quad K_h = \frac{\Delta f_3/2}{\Delta f_h - \Delta f_3/2}, \qquad (1)$$

where Δf_3 is the half-power width of the filter passband and $\Delta f_l = f_0 - f_l$ and $\Delta f_h = f_h - f_0$ are frequency bandwidths measured from the center of the filter passband to, respectively, the low-frequency and high-frequency slopes of the filter AFR at a level of –30 dB from the level of the minimum loss (see Fig. 2). These coefficients have better "sensitivity" to variations in selective properties of the filter as compared to traditional AFR-squareness ratios [12]. Moreover, the difference in coefficients K_l and K_h can be used to judge the degree of asymmetry of the filter's characteristic.

An experiment conducted with samples of analyzed filters with different numbers of sections has shown that theoretical results agree rather well with experimental data because we took into account end capacitances at free ends of microstrip sections in the calculations [13] and we considered frequency dispersion of permittivity in the quasi-static theory [14].

2. BASIC RESULTS OF THE INVESTIGATION

Samples of filters with fractional bandwidth $\Delta f_3/f_0 = 10\%$, which were first synthesized on V-80 ($\varepsilon = 80$) and TBNS ($\varepsilon = 40$) ceramic and alumina ($\varepsilon = 9.8$) substrates with thickness of $h = 1$ mm and than manufactured using the varnish-etching technique [15], demonstrated good agreement between theoretical and experimental results. After the manufacture, the geometry of stripline conductors of each sample was thoroughly measured by means of calculation of the averaged dimensions of all regular segments of the filter topology. These averaged values were used in the analysis program for calculation of the filter AFR and subsequent comparison of the theoretical results with experimental data.

We also tested the accuracy of calculation of filters that were designed on small-permittivity substrates and had rather "wide" passbands. Even in this case, theoret-

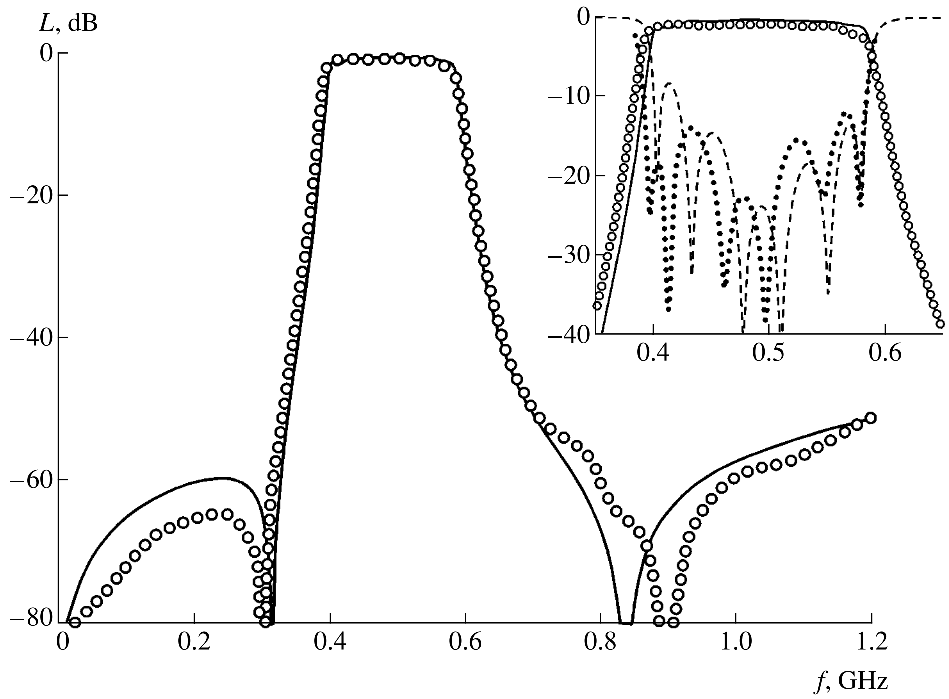

Fig. 2. Experimental (dots) and theoretical (solid and dashed lines) AFRs of a six-section filter designed on a FLAN substrate with ε = 2.8. The inset depicts a magnified fragment of the filter AFR and the return losses.

ical frequency dependences of forward and return losses agree rather well with experimental results. Figure 2 presents calculated frequency dependences of forward (solid line) and return (dashed line) losses; points in this figure are experimental results obtained for a six-section filter manufactured on a FLAN substrate with $\varepsilon = 2.8$ and a thickness of $h = 1.5$ mm. The experimental and calculated values of fractional bandwidth $\Delta f_3/f_0$ of this filter are 40.3% and 38.4%, respectively. Parameters of this design are as follows (see Fig. 1): $l_r = 44.5$, $l_h = 1.65$, $S = 1.034$, $w_0 = 0.966$, $w_1 = 3.965$, $w_2 = 2.907$, $w_3 = 3.265$, $S_1 = 0.140$, $S_2 = 0.548$, $S_3 = 0.273$, $l_c = 25.0$, and $Q_0 = 190$. All dimensions are in millimeters.

Measurements have shown that, for all tested samples, experimental values of the center frequency, bandwidth, and even the AFR steepness coefficients agree well with theoretical data. These facts allow us to perform a theoretical study of frequency-selective properties of such filters. This study can be performed using a special option of the Filtex32 expert system for computer-aided design of microstrip filters. For this purpose, we select a basic parameter of interest and specify the step and the interval of variation of the parameter's value. At each stage of the study, the filter is automatically tuned to the specified bandwidth by correcting the tuning parameters according to the technique described above. After the completion of each stage, data on the steepness coefficients of the slope of the obtained AFR, as well as on all design parameters of this filter, are recorded in a special file (table) that can be used to plot the corresponding 1D and 2D characteristics with the use of standard software. Thus, data contained in this file allow us to analyze not only variations in the selective properties of the filter under study but also the behavior of its tuning parameters as functions of basic design parameters.

It is known that the steepness of the slope of the AFR of a microstrip filter whose attenuation pole is far from the passband is solely determined by the type of coupling between the filter resonators at the passband frequencies [16]. In the case of a predominantly capacitive interaction, the steepness of the low-frequency slope is higher and, in the case of a predominantly inductive interaction, we observe the opposite behavior. It is also known that the fractional bandwidth of a filter is related to the total coupling coefficient of microstrip resonators k, which is defined as an "algebraic sum" of the inductive and capacitive coupling coefficients, k_L and k_C, respectively [17]. These coefficients can have both like and opposite signs. The value of coefficient k can be determined from the formula

$$k = \frac{k_L + k_C}{1 + k_L k_C}. \quad (2)$$

In the analyzed filters designed on codirectional hairpin resonators, coefficients k_L and k_C can have opposite signs at frequencies of the first (operating) passband. In other words, unlike filters designed on oppositely directed resonators [9], where inductive and capacitive coupling coefficients have identical signs, Values of design parameters for filters designed on codirectional and oppositely directed resonators

N	Design parameter, mm	Codirectional		Oppositely directed	
		$\Delta f_3/f_0$, %			
		5	40	5	40
2	l_r	12.860	13.000	12.860	13.320
	l_c	2.807	8.872	2.810	8.967
	S_1	1.119	0.012	1.247	0.036
3	l_r	12.870	13.030	12.870	12.930
	l_c	2.972	9.361	2.983	9.611
	S_1	1.088	0.026	1.173	0.027
	w_2	0.659	0.382	0.668	0.495
4	l_r	12.84	13.170	12.880	13.460
	l_c	2.947	9.708	3.031	9.792
	S_1	1.135	0.028	1.279	0.055
	S_2	3.410	0.372	3.400	0.473
	w_2	1.986	1.594	1.958	1.546
5	l_r	12.870	13.070	12.880	13.410
	l_c	3.093	9.773	3.082	9.904
	S_1	1.119	0.012	1.258	0.042
	S_2	2.755	0.134	2.769	0.220
	w_2	1.969	1.906	1.959	1.697
	w_3	5.051	5.217	4.535	4.057
6	l_r	12.880	13.160	12.880	13.460
	l_c	3.095	9.914	3.066	10.030
	S_1	1.113	0.021	1.251	0.048
	S_2	3.406	0.378	3.401	0.487
	S_3	1.257	0.085	1.495	0.135
	w_2	1.965	1.566	1.957	1.525
	w_3	1.969	1.764	1.960	1.667

Note: Six-section filters with f_0 = 1 GHz designed on 1-mm-thick substrates with ε = 9.8.

the total coupling coefficients between the sections of the filter are determined by the difference between the inductive and capacitive coupling coefficients. This is a fundamental difference between the filters designed on codirectional and oppositely directed resonators. As in the case of filters designed on codirectional and oppositely directed half-wave hairpin resonators [18], this difference should cause the corresponding difference between shapes of the AFRs of these devices.

Let us confirm this statement by comparing AFRs of two filters with fractional bandwidths $\Delta f_3/f_0$ = 5% (Fig. 3a) and 40% (Fig. 3b). Both filters are tuned to the center passband frequency f_0 = 1 GHz but contain different numbers N of sections. The filter substrates are made of a 1-mm-thick alumina (ε = 9.8), width w_0 of conductors connected to the screen is 1 mm, and width w_1 of free segments is 2 mm. All remaining design parameters of these filters are listed in the table.

As seen from Fig. 3, filters based on codirectional resonators have larger steepnesses of the AFR slopes than filters based on oppositely directed resonators. This difference holds for filters with both wide and narrow bandwidths and increases with the number of sections of the filter. It is of interest that six-section filters on codirectional resonators have attenuation poles situated to the left and to the right of the filter passband.

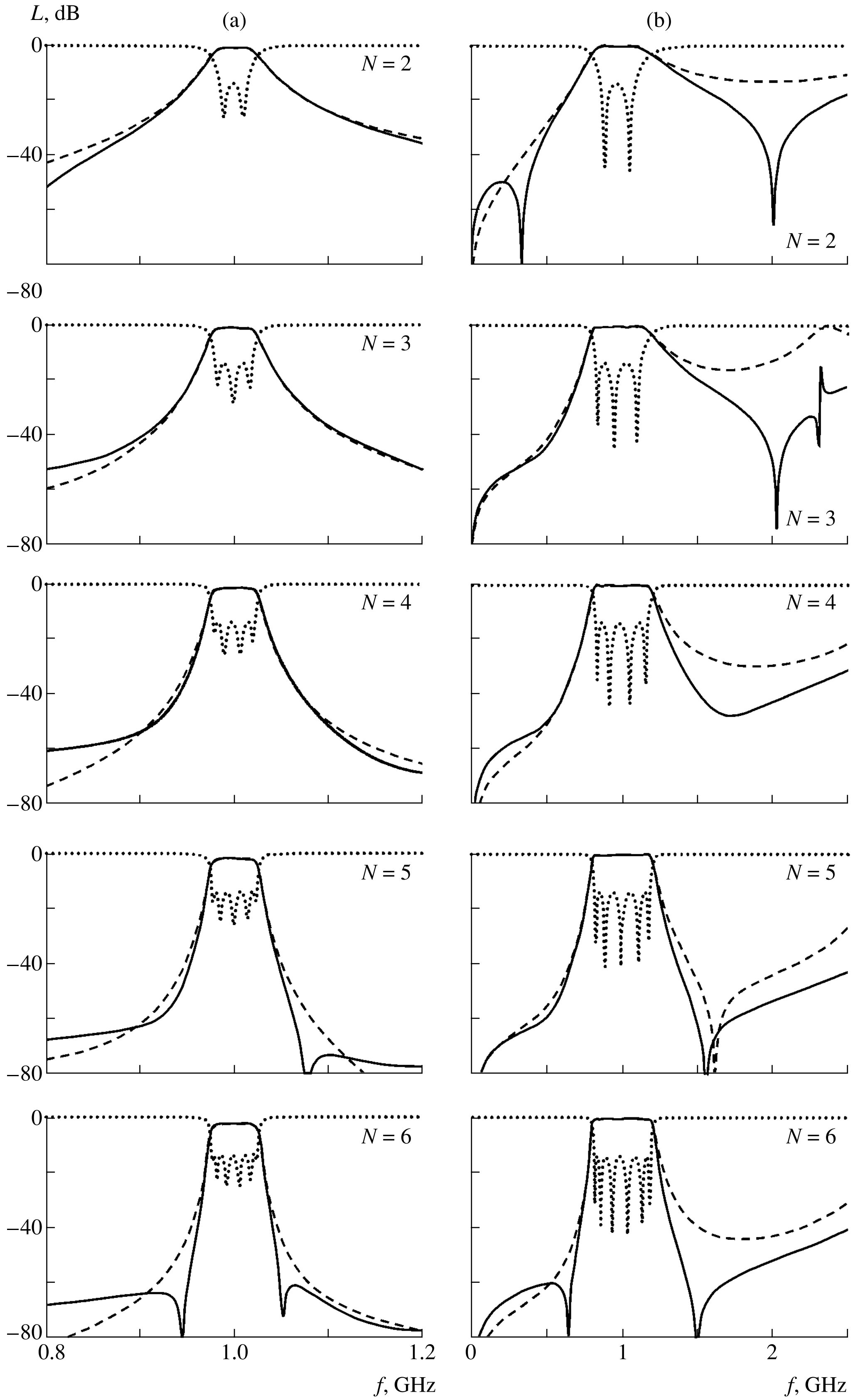

Fig. 3. Comparison between the AFRs of filters that have fractional bandwidths of (a) 5% and (b) 40%; are designed on (solid lines) codirectional and (dashed lines) oppositely directed resonators; and contain different numbers of sections, *N*. Dots indicate the frequency dependences of return losses.

This behavior is observed for any value of the filter bandwidth. The presence of such attenuation poles substantially increases the steepness of the AFR slopes and, as a result, improves frequency-selective properties of the MSF under study as compared to the properties of the filter designed on oppositely directed resonators [9].

Analyzing the data presented in the table, we should note the following. In the case of "narrowband" filters, all design parameters of the filter topology are almost the same for the filters based on both codirectional and oppositely directed resonators, irrespective of the number of the filter sections. The cause for this phenomenon is that the narrowband MSFs require a weak coupling between resonators; therefore, the spacing between the resonators exceeds the thickness of the substrate and, consequently, capacitive interaction between the sections of the filter is negligibly small as compared to inductive interaction. As a result, notwithstanding that, in a filter designed on oppositely directed resonators, inductive and capacitive coupling coefficients k_L and k_C have, as was mentioned above, identical signs, while the corresponding coefficients in a filter designed on codirectional resonators have opposite signs, the value of coefficient k_C is small and the spacing ensuring some value of the total coupling coefficient is almost that same for both designs.

Conversely, in the case of "wideband" filters, the parameters of the conductor's topology are substantially different for both considered designs (see table). This is probably related to the fact that, as the filter bandwidth increases, the spacing between the resonators should decrease accordingly, thus causing a more rapid increase of the capacitive coupling coefficient as compared to the inductive coupling coefficient [17]. Therefore, in the filters based on codirectional and oppositely directed resonators that have a fractional bandwidth of 40%, the spacings between the resonators are substantially different. As expected, owing to different signs of the inductive and capacitive coupling coefficients, filters on codirectional resonators require a smaller spacing for obtainment of the specified bandwidth. An exception is three-section filters in which spacing S_1 takes almost the same value. However, these filters have substantially different values of width w_2 of the stripline conductors of middle resonators that are not connected to the screen, thereby ensuring equal values of the total coupling coefficient in both filters. Indeed, coefficients k_L and k_C strongly depend on the width of stripline conductors in the region where this width is less than the thickness of the substrate. It is important that, in this region, the inductive coupling coefficient increases and the capacitive coupling coefficient decreases with a decrease in the width of the conductor [17].

Taking into account the character of the distribution of high-frequency currents and voltages along the stripline conductors at frequencies of the first passband, we note that the inductive interaction always prevails over capacitive interaction for pairs of resonators coupled via segments connected to the screen (see Fig. 1). However, in resonators coupled via segments not connected to the screen, capacitive coupling becomes substantial, especially for filters with a wide passband. Therefore, in the considered two- and three-section filters, the coupling between resonators is determined by capacitive interaction, whereas, for $N \geq 4$, we have an alternating type of coupling between adjacent pairs of resonators. Hence, the number of resonator pairs with inductive coupling (interacting via segments of conductors connected to the screen) is less than the number of resonator pairs with capacitive coupling (interacting via segments not connected to the screen). This rule holds for any number of sections except for $N = 5$. In the latter case, the number of aforementioned pairs is the same for both types of interaction. In this case, as expected, we observe the smallest difference between values of slope-steepness coefficients K_l and K_h.

For comparison, we obtained dependences of the AFR-slope-steepness coefficients on the fractional bandwidth of the studied filters with the number of resonators ranging from two to six (Fig. 4). In this study, passband center frequency f_0 was 500 MHz and the synthesis was performed for the following values of basic design parameters: $\varepsilon = 9.8$, $h = 1$ mm, $w_1 = 1$ mm, $w_0 = 1$ mm, $S = 1$ mm, and $Q_0 = 500$. For all designs, we synthesized a set of filters with an increment of 5 MHz to the filter bandwidth, starting from $\Delta f_3 = 10$ MHz. The synthesis procedure was conducted until the value of smallest spacing S_1 decreased to 10 μm. Note that, as the filter bandwidth increased, the height of the hairpin varied within 0.7 mm; this height first increased monotonically and then decreased monotonically. Mean height l_r was ≈26 mm. This result is due to the influence of changes in the ratio of inductive and capacitive coupling coefficients (caused by the reduction of the spacing between resonators with increasing $\Delta f_3/f_0$ and an increase in the value of parameter l_c, which is required for matching the filter at its input and output) on the resonance frequencies of the outmost resonators. Indeed, it is known that increasing capacitive coupling between resonators lowers their resonance frequencies and changing the positions of the connection points of external transmission lines causes a nonmonotonic variation of resonance frequencies as a function of parameter l_c [19].

As expected, as the number of sections of the filter sections, the steepness of both slopes of the filter AFR rapidly increases; however, for all designs, the low-frequency slope is steeper than the high-frequency slope (see Fig. 4). Note that the steepness of the slope increases monotonically with the filter bandwidth. This rule holds for all designs except for the three-section filter, for which curve $K_h(\Delta f_3/f_0)$ is nonmonotonic. The rapid increase of slope-steepness coefficients that can be observed for $\Delta f_3/f_0 < 10\%$ is a result of comparatively low values of the unloaded quality factors of the filter resonators, which strongly affect values of coefficients K_l and K_h only in narrowband filters. This fact is confirmed by the results presented in Fig. 4, where dash–dotted lines depict similar dependences for all filters in the case when $Q_0 \longrightarrow \infty$.

It is known that the permittivity of the substrate, the stepwise changes of the widths of stripline conductors, and other design parameters of MSFs with irregular resonators can substantially change the ratio of inductive and capacitive coupling between the stages and cause corresponding changes in the steepness of the AFR slopes [1, 9]. In the studied filters, almost all basic design parameters affect the steepness of the AFR slopes. This effect is caused by the aforementioned features of the filter design: (1) the alternating type of coupling between adjacent resonators and (2) the different signs of inductive and capacitive coupling at the passband frequencies.

Figure 5 shows dependences of the AFR-slope steepness on some basic parameters of six-section filters designed on FLAN (ε = 2.2), alumina (ε = 9.8), and TBNS ceramic (ε = 80) substrates.

Initial design parameters of all devices (see Fig. 1) are as follows: w_0 = 1 mm, w_1 = 1 mm, S = 1 mm, l_h = 1 mm, h_d = 1 mm, $h_a = \infty$, and Q_0 = 500. In this study, all filters have the same fractional bandwidth $\Delta f_3/f_0$ = 10% but are tuned to different center frequencies so as to obtain approximately the same length, l_r, of hairpins in filters designed on different substrates. Frequency f_0 is 3 GHz for filters designed on the FLAN substrate, 1.5 GHz for filters designed of the alumina substrate, and 0.5 GHz for filters designed on the TBNS substrate.

It can be seen that, as width w_1 of the stripline conductor or spacing S between the resonator conductors increases, the steepness of the low-frequency slope of the filter AFR rapidly decreases and the steepness of the high-frequency slope slightly increases. This behavior is observed for all devices except the filter on the FLAN substrate that has a small permittivity (ε = 2.2). In all devices, the low-frequency slope is steeper than the high-frequency slope, and the difference increases with the substrate permittivity. As substrate thickness h_d and height h_a of the shielding cap placed above the substrate increase, the steepness of the low-frequency slope of the filter AFR also increases and the steepness of the high-frequency slope decreases. This regularity is valid for all designs except the filter designed on the FLAN substrate, for which we observe a monotonic increase of the steepnesses of both slopes.

Evidently, in this design, an increase in l_h (see Fig. 1) leads to a decrease in the length of the region of interaction between the resonator pairs with dominating capacitive coupling and to an increase in the length of the region of interaction between the resonator pairs with dominating inductive coupling. Therefore, as

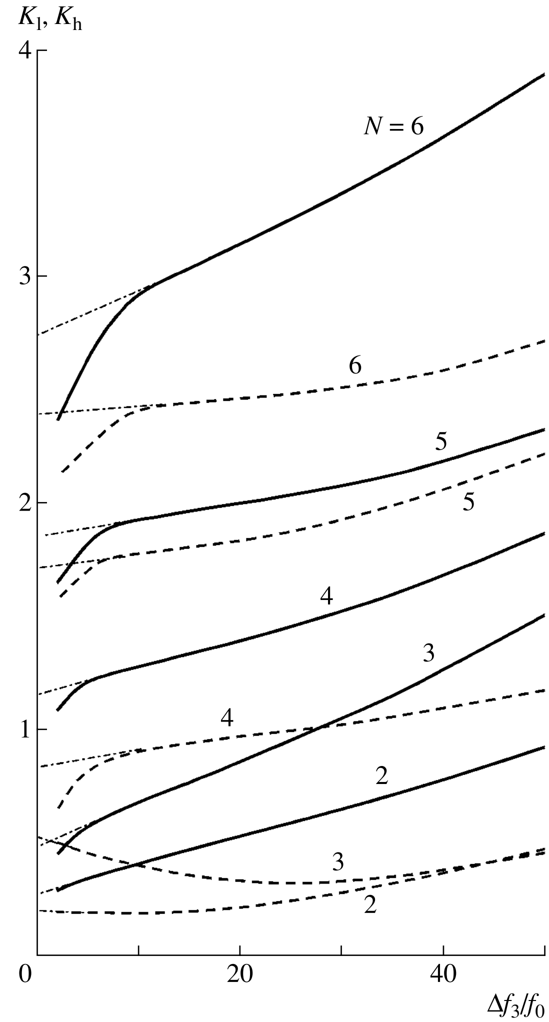

Fig. 4. Steepnesses of (solid line) the low-frequency and (dashed line) the high-frequency slopes of the filter AFR as functions of the filter bandwidth for filters with N = 2–6.

expected, as ratio l_h/l_r increases, the steepness of the low-frequency slope increases and the steepness of the high-frequency slope decreases for all filters. Note that, in this study, hairpin heights l_r vary from approximately 10.6 to 5.4 mm for the filter designed on the TBNS ceramic, from approximately 9.8 to 4.7 mm for the filter designed on the alumina substrate, and from approximately 9.6 to 4.6 mm for the filter designed on the FLAN substrate.

Investigations have revealed that, for fixed basic parameters of the considered designs, the AFR-slope-steepness coefficients depend on not only the bandwidth (see Fig. 4) but also the center frequency of the filter passband. Such dependences of the slope-steepness coefficients on normalized center frequency of the filter passband, f_0/f_1, are presented in Fig. 5 for six-section filters with the initial parameters listed above. It is seen that, as in the case of filters with oppositely directed resonators [9], coefficient K_l monotonically

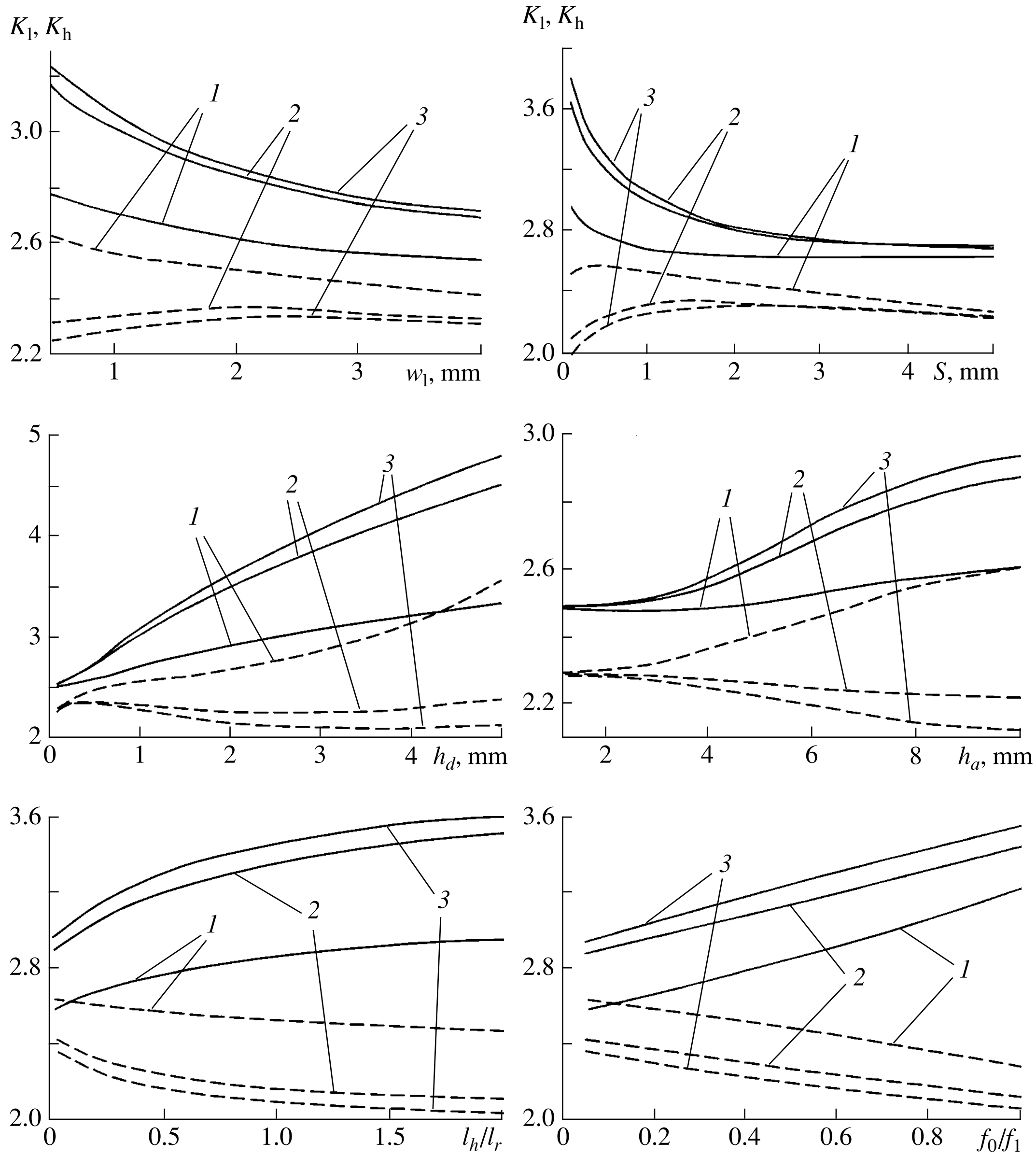

Fig. 5. Steepnesses of (solid lines) the low-frequency and (dashed lines) the high-frequency slopes of the AFRs of six-section filters that have a fractional bandwidth of $\Delta f_3/f_0$ = 10% and are designed on substrates with ε = (*1*) 2.2, (*2*) 9.8, and (*3*) 80 as functions of the filter parameters.

increases and coefficient K_h decreases as frequency increases. However, for the analyzed designs with codirectional resonators, these regularities are far more pronounced.

An advantage of filters with irregular resonators [1, 7, 9], including the designs considered above, is an increased width of the stopband. This increase is attained by shifting the second (parasitic) passband to the high-frequency region. Such a shift can be performed by increasing the difference between width w_0 of the regular segment of the stripline conductor connected to the screen (see Fig. 1), which contains, as was mentioned above, an antinode of the high-frequency current of the first (operating) mode of the resonance oscillations, and width w_1 of the regular segment not connected to the screen, which contains an antinode of the high-frequency voltage. Indeed, near the frequency of the first oscillation mode, a microstrip resonator can be treated as an oscillatory circuit in which the inductance is formed by the regular segment of the stripline conductor connected to the screen and the capacitor is formed by the open-ended regular segment. As a result, by decreasing width w_0 or increasing width w_1, we lower the natural frequency of the resonator's first oscillation mode owing to the corresponding increase of either inductance or capacitance. However, this

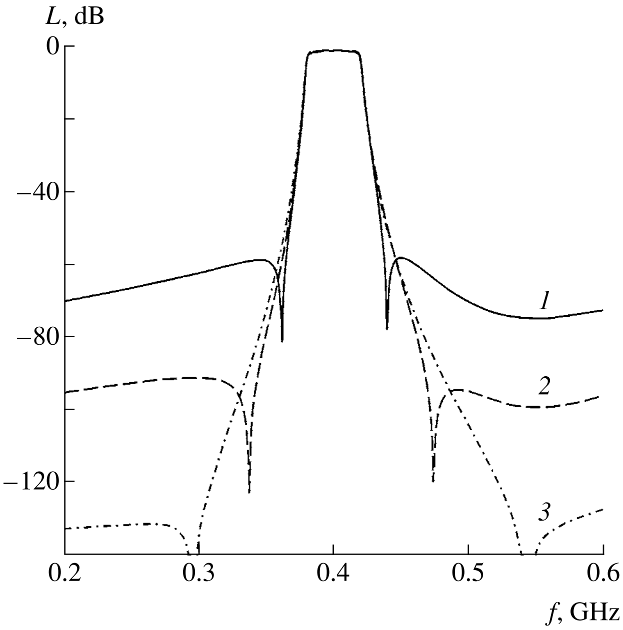

Fig. 6. Amplitude–frequency responses of a six-section filter (*1*) without a shielding cap and with a cap placed at a height of h_a = (*2*) 10 or (*3*) 5 mm.

increases the resonance frequency of the second oscillation mode because, for this mode, the second antinode of the high-frequency current in the distribution of high-frequency field is situated on the wide segment of the stripline conductor, where the linear inductance is small, and the second antinode of the high-frequency voltage is situated on the narrow segment of the conductor, where the linear capacitance is also small. As a result, in this filter, the width of the stopband may occupy several octaves [9].

It is important to note that shielding of microstrip filters substantially increases attenuation of the microwave power in the stopbands. Figure 6 shows AFRs of a six-section filter plotted for three different distances h_a between the shielding cap and the substrate. The filter under study was designed on a TBNS ceramic substrate with $\varepsilon = 80$ and $h_d = 1$ mm and had the following basic design parameters: $w_0 = 1$ mm, $w_1 = 1$ mm, $S =$ 1 mm, and $l_h = 1$ mm. It was designed for a passband center frequency of $f_0 = 0.4$ GHz and a fractional bandwidth of $\Delta f_3/f_0 = 10\%$. For the filter without a cap, height l_r of the filter resonators was 13.48 mm and the spacings between resonators were $S_1 = 0.68$ mm, $S_2 =$ 2.26 mm, and $S_3 = 0.79$ mm. Indeed, it is seen that, as the screen is moved closer to the substrate, the attenuation rapidly increases in both stopbands; however, the shape of the filter AFR remains almost unchanged up to a level of –30 dB. In other words, for the considered filter, the AFR-slope-steepness coefficients depend only slightly on the height of the cap because, even for $h_a =$ 5 mm, all spacings between the resonator conductors are substantially less than this height. As a result, at frequencies near the first filter passband, the shielding cap causes only slight perturbations of high-frequency fields of the filter resonators.

Nevertheless, we should note that, as the shielding cap approaches the substrate, capacitive coupling between filter resonators increases and inductive coupling decreases. Therefore, a corresponding increase of "capacitive" spacings S_1 and S_3 and a decrease of "inductive" spacing S_2 is required. For example, if height h_a of the upper screen is 5 mm, these parameters take the following values: $S_1 = 0.73$ mm, $S_2 = 1.94$ mm, and $S_3 = 0.88$ mm. Height l_r of the hairpin also increases slightly to 13.56 mm.

CONCLUSIONS

The investigations performed have shown that microstrip filters designed on codirectional quarter-wave hairpin resonators have not only very small dimensions but also selective properties that are better than selective properties of similar filters designed on oppositely directed resonators [9]. Such filters can be designed on substrates with permittivity ε ranging from 2 to 100 and can have a fractional bandwidth of $\Delta f_3/f_0 =$ 2–50% in a rather wide range of center frequencies f_0 from 0.1 to 5.0 GHz. However, in these designs, the steepness of the AFR slope depends strongly on basic design parameters, which opens great opportunities for obtaining a prescribed shape of the filter AFR.

An especially important advantage of the considered designs is the high manufacturability of such devices. Indeed, the production of filters designed on oppositely directed resonators requires a high accuracy of manufacture of the filter substrates, including not only the thickness of the substrate but also the width, because, in such filters, stripline conductors of the filter resonators are connected to the screen on the opposite sides of the substrate. Hence, the substrate width determines the resonator length [9]. In the designs studied above, stripline conductors of the filter resonators are connected to the screen on one side of the substrate (see Fig. 1); therefore, such devices are free of this disadvantage. Moreover, they preserve all advantages inherent in filters designed on oppositely directed resonators, including the possibility of a substantial widening of the filter stopband and rather good agreement between the experimental data and the results of numerical calculation of microstrip structures obtained in the quasi-static approximation.

It should be noted that, as in the designs studied earlier [9], filters with an odd number of sections contain resonators whose width is almost twice as large as the width of the remaining resonators. The phenomena is due to the mirror symmetry of the topology of the resonator's conductors (see Fig. 1). As a result, the overall dimensions of the substrate of a three-section filter almost coincide with the dimensions of the substrate of a four-section filter; similarly, the overall dimensions of a five-section filter are almost the same as the dimensions of a six-section filter. Therefore, at first glance, filters with odd numbers of sections are inefficient if we take into account that the steepness of the AFR slopes is substantially higher in designs with larger numbers of resonators. However, tuning of devices with smaller numbers of sections is substantially simpler and, most importantly, such devices have substantially lower losses of microwave power in the filter passband. Therefore, in some cases, when requirements for selective features of such a device are not very stringent, a three-section filter may be preferable to a four-section filter and a five-section filter may be preferable to a six-section filter.

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