ISSN 1028-3358, Doklady Physics, 2017, Vol. 62, No. 6, pp. 289–293. © Pleiades Publishing, Ltd., 2017.
Original Russian Text © B.A. Belyaev, S.A. Khodenkov, An.A. Leksikov, V.F. Shabanov, 2017, published in Doklady Akademii Nauk, 2017, Vol. 474, No. 6, pp. 682–686.

PHYSICS

A Three-Mode Microstrip Resonator and a Miniature Ultra-Wideband Filter Based on It

B. A. Belyaev[a,b,*], **S. A. Khodenkov**[b], **An. A. Leksikov**[a,b],
and Academician V. F. Shabanov[c]

Received January 19, 2017

Abstract—An original microstrip resonator design with a strip conductor split by a slot at one of its ends is investigated. It is demonstrated that at the optimal slot sizes, when the eigenfrequency of the second oscillation mode hits the center between the first and third oscillation modes, the resonator can work as a third-order bandpass filter. The structure formed from only two such resonators electromagnetically coupled by split conductor sections is a miniature six-order wideband filter with high selectivity. The test prototype of the filter with a central passband frequency of ~1.2 GHz and a passband width of ~0.75 GHz fabricated on a substrate ~(45 × 11 × 1) mm^3 in size with a permittivity of 80 is characterized by minimum loss in a passband of 0.5 dB. The parametric synthesis of the filter structure was performed using electrodynamic analysis of the 3D model. The measured characteristics of the test prototype agree well with the calculated data.

DOI: 10.1134/S102833581706009X

Frequency selective microwave devices, including bandpass filters, are important elements of communication and radar systems and various special radio equipment. Filters often determine the dimensions and, most importantly, quality of microwave radio engineering devices. In recent years, filters with an ultrawide (over 50%) relative passband have been intensively developed due to the requirements for enhancing the data transmission rate in digital microwave systems, which, as is known, is directly related to the working frequency passband width. Therefore, the development of new miniature ultra-wideband filters with high selectivity, low passband loss, good manufacturability, and low cost is an urgent problem.

The widespread filter design based on regular microstrip resonators (MSRs) [1, 2] has spacings that are too small between strip conductors at a relative bandwidth of more than ~30%, which unacceptably degrades the electric strength of the device. The ultrawideband microstrip filters with a fractional bandwidth of over 50% based on quarter-wave resonators conductively coupled by line segments of a certain electric length [3] have a relatively low frequency selectivity, i.e., a narrow high-frequency stopband and low damping level in it. The MSR-based filters with direct coupling [4, 5], the passband of which is determined by a step of impedances of the forming lines are too bulky in the decimeter wavelength range. The miniature designs based on dual-mode [6] and multimode [7] microstrip resonators are too hard to tune. The filters on a suspended substrate with the double-sided strip conductor pattern [8] are fairly miniature and have high electrical performances; however, they are more difficult to produce than the microstrip filters.

In this study, we investigate a new microstrip resonator in which a conductor is split at one of its ends by a slot of certain size and demonstrate the possibility of fabrication of filters on the basis of this resonator. As is known, splitting of the strip MSR conductor by a narrow slot [9] allows reducing the distance between eigenfrequencies of its first two oscillation modes. Thus, such dual-mode resonators can be used to create not only bandpass filters with a high frequency selectivity [9], but also dual-split filters and diplexers [10]. However, the relative bandwidth of such filters is no more than 20%, whereas the resonators under study allow fabricating filters with passbands of up to ~70% with good frequency selective properties.

THREE-MODE MICROSTRIP RESONATOR

Figure 1a shows the dependences of eigenfrequencies of the first (f_1), second (f_2), and third (f_3) oscillation modes on slot width S in the resonator conductor (Fig. 1a, inset). The microstrip resonator was investi-

---

[a] *Kirenskii Institute of Physics, Federal Research Center, Krasnoyarsk Science Center, Siberian Branch, Russian Academy of Sciences, Krasnoyarsk, 660036 Russia*
[b] *Siberian State Aerospace University named after Reshetnev, Krasnoyarsk, 660037 Russia*
[c] *Krasnoyarsk Science Center, Russian Academy of Sciences, Siberian Branch, Krasnoyarsk, 660036 Russia*
e-mail: belyaev@iph.krasn.ru

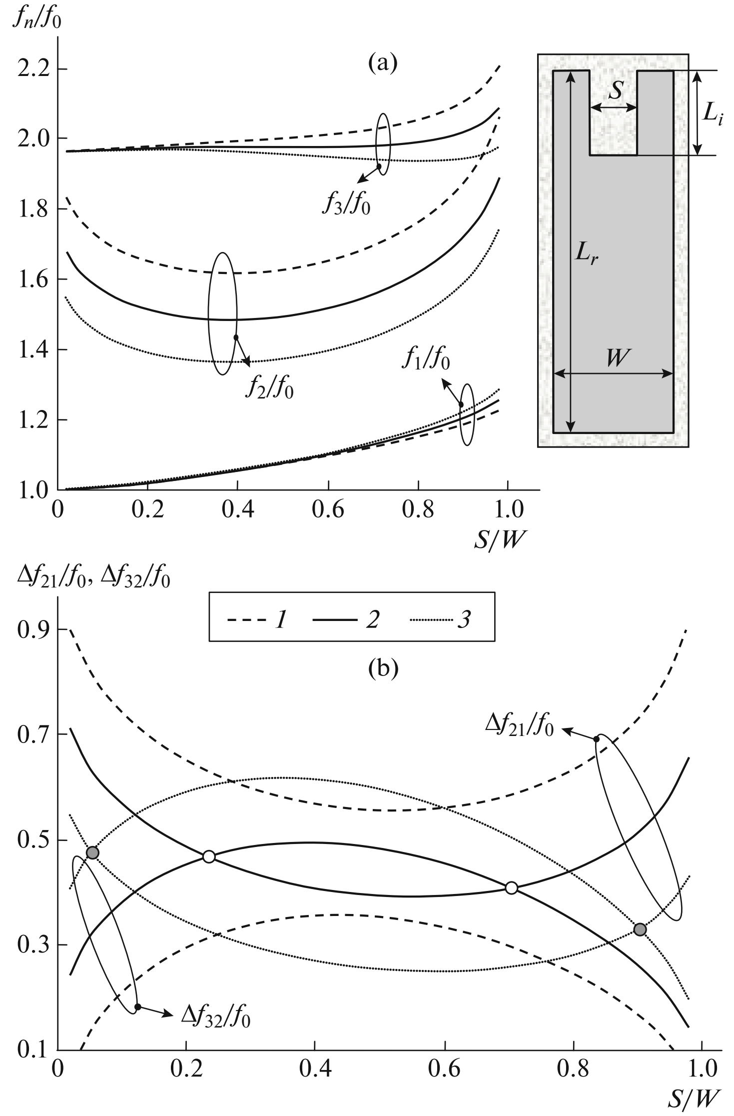

Fig. 1. Slot size dependences of (a) the first three eigenfrequencies of the resonator with a split strip conductor and (b) their differences. The dependences are normalized to frequency f_0 of the resonator with the regular conductor ($L_i = 0$) and built for three split section lengths: $L_i/L_r = 0.29$ (*1*), 0.33 (*2*), 0.38 (*3*). Inset: topology of the microstrip resonator conductor.

gated using numerical electrodynamic analysis of its 3D model at a weak capacitive coupling between the resonator and transmission lines. The eigenfrequencies are normalized to frequency f_0 of the first oscillation mode of the initial resonator with a regular strip conductor ($L_i = 0$) and the slot width is normalized to conductor width W. The resonator substrate was a 1-mm-thick plate made of high-frequency ceramics with a permittivity of $\varepsilon = 80$. The resonator conductor width and length are $W = 10$ mm and $L_r = 38$ mm, respectively ($f_0 = 0.598$ GHz). It can be seen in Fig. 1a that the frequency of the first oscillation mode (this mode is odd, since the charges at the conductor ends have opposite signs) monotonically increases with an increase in the slot width at all the slot lengths. Obviously, the growth of the frequency f_1 with increasing S is related to a decrease in the capacity of the split conductor end, where the RF electric field has an antinode for this half-wave oscillation mode [11].

The second oscillation mode is also odd, since in this mode the charges at the ends of the split conductor section have opposite signs [9, 10], and its frequency first decreases and then increases with an increasing slot width at any slot length. This oscillation mode corresponds, in fact, to a half-wave hairpin resonator formed from narrow conductors of the split section and a small segment of the adjacent unsplit conductor section. Therefore, the frequency of this oscillation mode decreases with increasing slot length in accordance with an increase in the hairpin resonator length.

The frequency of the third oscillation mode (which is even, since the charges on the conductor ends have the same sign) remains almost invariable at the small slot sizes; however, at $S/W > 0.4$, it can monotonically increase or first decrease and then increase, depending on the slot length. This behavior of f_3 upon variation in S can easily be explained by the fact that, for this oscillation mode, the full wavelength falls on the strip conductor length. As a result, the RF voltage antinodes are localized not only at the strip conductor ends, but also near its center and the RF current antinodes are localized between them [11]. As the slot width is increased, the capacitance of the split conductor section with the voltage antinode decreases, but the inductance of this section increases as well. Therefore, the observed behavior of f_3 upon variation in S depends on the length of the slot in the strip conductor of the investigated MSR.

Figure 1b shows the normalized dependences of the differences between the second and first oscillation modes $(f_2 - f_1) = \Delta f_{21}$ and between the third and second oscillation modes $(f_3 - f_2) = \Delta f_{32}$ as functions of the slot width. It can be seen that at any slot length the dependences $\Delta f_{21}/f_0(S/W)$ have minima and the dependences $\Delta f_{32}/f_0(S/W)$ have maxima. It is noteworthy that in the relatively narrow range of the relative slot length ($0.3 < L_i/L_r < 0.4$), the dependences $f_{21}/f_0(S/W)$ and $f_{32}/f_0(S/W)$ intersect at certain S/W values shown by white and black dots in Fig. 1b. The slot sizes at these points of the strip resonator are such that the frequency of its second oscillation mode is located precisely between the frequencies of the first and third oscillation modes. Only at these slot sizes is strong interaction between the first three resonances observed in the investigated resonator, if its strip conductor is conductively connected to the input and output transmission lines in the split section. This leads to the formation of an almost octave resonator passband.

To confirm the aforementioned, Fig. 2 shows the frequency responses of the resonator obtained by numerical electrodynamic analysis of the 3D model (solid lines) and measured on the resonator prototype formed on a 1-mm-thick substrate ($\varepsilon = 80$) with a conductor width of $W = 3.8$ mm, a conductor length of $L_r = 22.85$ mm, and slot sizes of $S = 2.8$ mm and $L_i = 6.95$ mm. The points of conductive connection of 50-Ω external transmission lines to the resonator are located at a distance of 2.3 mm from the ends of narrow conductors on the split section (see photograph in Fig. 2). It can be seen that the frequency response of the investigated resonator corresponds to the third-order filter, similar to the frequency response of the well-known three-mode resonator [12] based on the microstrip line segments with a step of the strip conductor width. Note that the structure formed from the direct-coupled resonators connected in series [12] is analogous to the microstrip structures from [4, 5]; however, in the former the end quarter-wave resonators are used instead of half-wave ones. This somewhat decreases the three-mode resonator sizes, but the three-mode resonator studied here has much smaller dimensions.

Note that the strip conductor size in the investigated resonator and the points of its connection to the external lines were obtained by the parametric synthesis of the structure with the use of numerical electrodynamic analysis of its 3D model. For certainty, we specified a central passband frequency of 1.25 GHz and reflection loss maxima in the passband of not higher than −20 dB. The observed minor difference between the calculated and measured characteristics of the resonator is caused by manufacturing inaccuracies.

THE TWO-RESONATOR BANDPASS FILTER

Based on the structure formed from two interacting three-mode resonators, a six-order filter can be fabricated. Figure 3 shows a photograph of the two-resonator filter prototype fabricated on a 1-mm-thick ceramic substrate ($\varepsilon = 80$). In this device, the frequency response slope near the passband grows and the spurious low-frequency passband is simultaneously suppressed (see Figs. 2 and 3), which additionally improves the frequency selectivity of the filter.

Note that in multisection filters based on multimode resonators the balance of resonator couplings at the resonant frequencies of all the oscillation modes forming the passband can be ensured using different techniques [6, 7, 9, 13]. In contrast to the filters based on two-mode resonators with the split strip conductor [9, 10], in the filter based on the investigated three-mode resonators this balance can be implemented only by connecting the resonators with narrow strip conductors in the split sections, where the RF field antinodes for all three oscillation modes are localized.

It should be emphasized that the required strong interaction of the resonators can only be obtained using counter directed MSR strip conductors (see photograph in Fig. 3). Indeed, as we mentioned above, there are RF voltage antinodes at the ends of the narrow conductors of the split section for all three oscillation modes of the resonator under study. The inductive and capacitive coupling coefficients of the counter directed resonators are summed [2, 11, 14], thereby enhancing the interaction of the resonators, which facilitates broadening of the filter passband. In the structure under study, the required interaction of the resonators and balance of their couplings are governed by the width of narrow conductors in the split section, the length of their coupling region, and spacing between them.

Note that narrowing of the connecting conductors enhances both the capacitive and inductive interaction of the resonators [11, 14]. Therefore, the optimal sizes of the narrow conductors in the split part of each res-

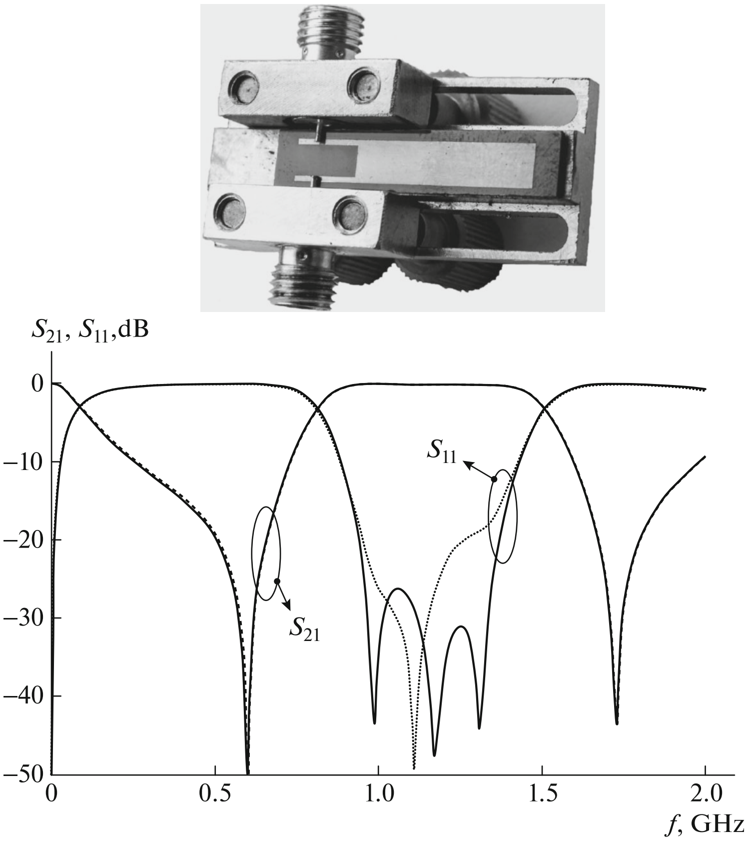

Fig. 2. Frequency responses of the three-mode microstrip resonator obtained by numerical electrodynamic analysis of the 3D model (solid line) and measured on the fabricated prototype (dots). At the top: photograph of the three-mode resonator.

onator obtained by the parametric synthesis of the filter using the electrodynamic analysis of its 3D model were found to be different. The obtained size of a couple of narrow conductors ensuring the resonator coupling balance is 8.20 × 0.15 mm at a spacing of 0.25 mm between them and their coupling region length of 4.6 mm; the obtained size of the conductors with the transmission lines (the impedance of 50 Ω) conductively connected to their ends is 7.05 × 0.80 mm (see photograph in Fig. 3). The size of the wide (unsplit) resonator conductor sections is 15.55 × 4.25 mm. Figure 3 shows that the calculated characteristics are in good agreement with the characteristics measured on the experimental sample of the two-resonator filter. The strip conductor pattern contour is 42.90 × 8.75 mm in size.

DISCUSSION

Thus, we investigated an original microstrip resonator design with a strip conductor split by a slot at one end. At the optimal slot sizes, when the eigenfrequency of the second oscillation mode hits the center between the frequencies of the first and third modes, the resonator can have a frequency response of a third-order bandpass filter. However, to obtain this, the strip conductor of the resonator should be conductively connected to the input and output transmission lines in the split section. Only then will the strong interaction between the resonances of the first three oscillation modes of the investigated resonator be implemented, which leads to the formation of its bandpass of almost octave width .

The design developed, which includes only two such three-mode resonators electromagnetically coupled by the split conductor sections, represents a miniature six-order wideband filter with a high frequency selectivity. The filter prototype with a central passband frequency of ~1.2 GHz and a passband width of ~0.75 GHz fabricated on a substrate with a permittivity of 80 has a minimum loss of merely 0.5 dB in the passband; the substrate size of the device is ~45 × 11 × 1 mm. The measured characteristics of the prototype agree well with the results of the calculation using the electrodynamic analysis of the 3D model. The results of investigations into the proposed ultra-wideband filter design demonstrate the possibility of its use in digital communication systems, broadband radar systems, and various special radio equipment.

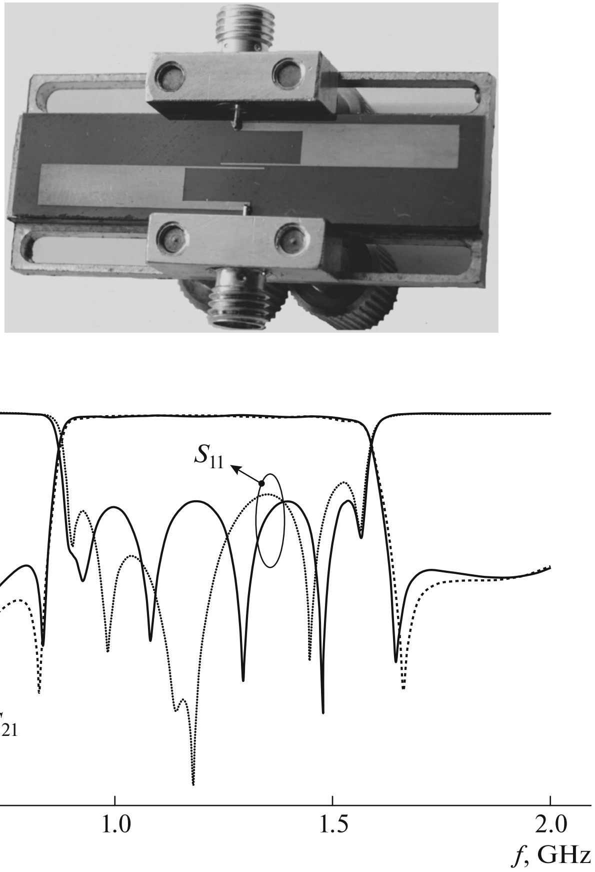

Fig. 3. Frequency response of the six-order filter based on two coupled resonators: measurements (dashes) and calculation (lines). At the top: photograph of the filter prototype.

ACKNOWLEDGMENTS

This study was supported by the Ministry of Education and Science of the Russian Federation, grant MK-9119.2016.8.

REFERENCES

1. I. C. Hunter, IET Electromagnetic Waves Ser. **48**, 353 (2006).
2. B. A. Belyaev, A. A. Leksikov, and V. V. Tyurnev, J. Comm. Technol. Electron. **49** (11), 1228 (2004).
3. J.-S. Hong and H. Shaman, Microw. Opt. Technol. Lett. **47** (3), 230 (2005).
4. B. A. Belyaev, A. S. Voloshin, and V. F. Shabanov, Dokl. Phys. **395** (6), 756 (2004).
5. B. A. Belyaev, A. S. Voloshin, and V. F. Shabanov, J. Comm. Technol. Electron. **51** (6), 653 (2006).
6. B. A. Belyaev, I. A. Dovbysh, A. A. Leksikov, and V. V. Tyurnev, J. Comm. Technol. Electron. **55** (6), 621 (2010).
7. B. A. Belyaev, S. A. Khodenkov, R. G. Galeev, and V. F. Shabanov, Dokl. Phys. **60** (3), 95 (2015).
8. B. A. Belyaev, A. S. Voloshin, A. S. Bulavchuk, and R. G. Galeev, Tech. Phys. Lett. **42** (6), 662 (2016).
9. B. A. Belyaev, A. A. Leksikov, A. M. Serzhantov, and A. M. Tyurnev, Prog. Electromagn. Res. Lett. **25**, 57 (2011).
10. B. A. Belyaev, A. M. Serzhantov, and A. M. Tyurnev, Microw. Opt. Technol. Lett. **55** (9), 2186 (2013).
11. B. A. Belyaev, A. A. Leksikov, and A. M. Serzhantov, J. Comm. Technol. Electron. **48** (1), 31 (2003).
12. L. Zhu, S. Sun, and W. Menzel, IEEE Microw. Wireless Compon. Lett. **15** (11), 796 (2005).
13. A. A. Aleksandrovskii, B. A. Belyaev, and A. A. Leksikov, J. Comm. Technol. Electron. **48** (4), 358 (2003).
14. B. A. Belyaev and A. M. Serzhantov, J. Comm. Technol. Electron. **49** (1), 20 (2004).

Translated by E. Bondareva