the first time. As a result, the following new information is obtained:

1. For a standard compact sedan, the distance of more than 300 m is needed to measure the accurate radiation pattern in the horizontal plane in the UHF band. Also, the pattern difference δ becomes less than 0.1 when $r > 40$ m.
2. The electric field pattern at arbitrary distance depends on polarization, frequency, antenna positions, and origin positions to define the distance.
3. The ground condition has a small influence on the distance dependence of electric field pattern.

The simulation results obtained in this study help to measure the radiation pattern of an antenna mounted on a car in the UHF band.

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A DUAL-MODE SPLIT MICROSTRIP RESONATOR AND ITS APPLICATIONS IN FREQUENCY SELECTIVE DEVICES

Boris A. Belyaev,[1,2,3] Aleksey M. Serzhantov,[1,2] and Vladimir V. Tyurnev[1,2]

[1] Kirensky Institute of Physics, Siberian Branch, Russian Academy of Sciences, Krasnoyarsk, Russia; Corresponding author: belyaev@iph.krasn.ru

[2] Institute of Engineering Physics and Radio Electronics, Siberian Federal University, Krasnoyarsk, Russia

[3] Reshetnev Siberian State Aerospace University, Krasnoyarsk, Russia

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ABSTRACT: *A dual-mode microstrip resonator whose strip conductor is partially split from its one end by a longitudinal slot has been studied. The resonator has two fundamental modes, even and odd ones. The difference of their frequencies is controlled by varying the slot length. Based on such resonators, original structures of a diplexer and passband filters including dual-band have been designed. Measured performances of their fabricated prototypes prove applicability of designed devices in microwave engineering.* © 2013 Wiley Periodicals, Inc. Microwave Opt Technol Lett 55:2186–2190, 2013; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.27806

Key words: *microstrip resonator; dual-mode resonator; bandpass filter; dual-band filter; diplexer*

1. INTRODUCTION

Use of dual-mode resonators in frequency selective devices is one of the ways to miniaturize these devices without degradation of their filtering properties. The structures of such resonators allow bringing frequencies of two oscillation modes closer. Therefore, two conventional resonators can be substituted for one dual-mode resonator enabling to decrease overall dimensions of a frequency selective device. The dual-mode resonators have proved to work well in bandpass filters including both dual-band filters [1–4] and diplexers [5,6]. Sometimes the dual-mode resonator-based devices, for example, filters, are called dual-mode ones for short [7].

Microstrip resonators are widely used in a microwave range [7]. However, some drawbacks are inherent in known dual-mode microstrip resonators. Particularly, in stepped impedance resonators [8] and in miniaturized hairpin resonators [9], it is impossible to move frequencies of two first oscillation modes closer as it does not allow using such resonators to design narrowband bandpass filters. Square-loop resonators [10] and E-shaped resonators [11] have another drawback, they have large dimensions. The mentioned imperfections are absent in the dual-mode split microstrip resonator proposed in Ref. [12] where it was applied in a narrowband bandpass in the sixth-order filter.

In this article, we study the behavior of the split microstrip resonator and its capability to be used in structures of different microwave frequency-selective devices.

2. A SPLIT MICROSTRIP RESONATOR

The layout of the split microstrip resonator is shown in Figure 1(a). There are two lower oscillation modes in the resonator. One of them is even when charges on the adjacent ends of split conductors have the same sign. Another mode is odd when charges have opposite signs. An equivalent circuit of the resonator is shown in Figure 1(b). It has a section of two coupled transmission lines connected with a section of a single transmission line. Frequencies of all even oscillation modes f_e in accordance with the shown circuit are roots of the equation [12]

$$Z_e \tan\theta_1 + 2Z_1 \tan\theta_e = 0 \tag{1}$$

and frequencies of odd oscillation modes f_o are roots of the equation

$$\cos\theta_o = 0. \tag{2}$$

Here, Z_1 and θ_1 are characteristic impedance and electrical length of the single line section and Z_e, Z_o, θ_e, θ_o are characteristic impedances and electrical lengths of the coupled lines' section for even and odd waves.

In Figure 1(c), the dependences of the resonant frequencies on the fractional slot length are plotted for the lowest even and odd oscillation modes. The computation is carried out for the case when the dielectric substrate has the permittivity $\varepsilon_r = 9.8$

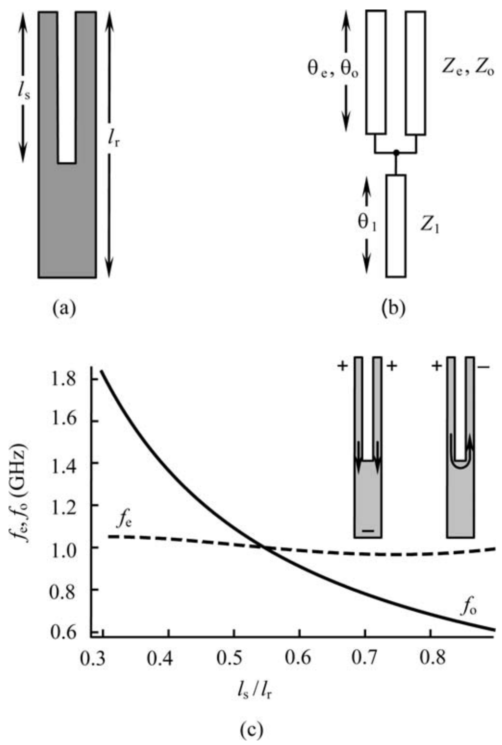


Figure 1 (a) Split microstrip resonator, (b) its 1D model, and (c) the frequencies of the even and odd modes f_e, f_o as functions of the normalized slot length

and the thickness $h = 1$ mm, the strip conductor has the total length $l_r = 55.3$ mm and the width in the nonsplit segment $w = 3$ mm, the slot has the spacing $S = 1$ mm. One can see that the frequency of the odd oscillation mode f_o rapidly decreases while the slot length l_s is increasing whereas the frequency of the even oscillation mode f_e remains practically invariable. These frequencies coincide at the point $l_s = 0.56\ l_r$ if the resonator has no couplings with a port or other resonators.

It is important to note that almost any frequency ratio of two lower oscillation modes can be realized in the split microstrip resonator. However, it is obvious that the range of practical interest is $0.5 < f_e/f_o < 2$. Really, the frequencies of high-order even modes in the considered resonator are multiple of the frequency f_e of the lowest mode like the resonant frequencies in the conventional microstrip resonator. Therefore, the frequency of the second-order even mode defines the upper range limit for f_o although the frequency f_o grows unboundedly as the slot length l_s becomes shorter. The frequency of the odd mode at maximum slot length defines the lower range limit, in this case $f_o \approx 0.5\ f_e$.

3. MICROSTRIP BANDPASS FILTERS

It is well known that a number of requirements must be met when designing conventional multisection bandpass filters. First, the resonant frequency of every resonator in presence of all its couplings must be tuned to the center frequency of the filter. Second, every pair of resonators must have a proper coupling value in order to obtain the required fractional bandwidth and equalize transmission ripple in the pass band. Third, the input and output resonators must have an optimal value of their coupling with the external transmission lines in order to obtain the required value of maximum reflection in the bandpass. In essence, these requirements must be satisfied and at designing the dual-mode bandpass filters. However, one should have in a mind that varying one of the resonant frequencies in the dual-mode resonator may be attended with the variation of coupling between two oscillation modes. So, adjusting of the dual-mode filters, as a rule, is more complicated than of the conventional ones [13].

As is known, it is practically impossible to implement dual-mode microstrip bandpass filters using half-wavelength stepped impedance resonators [14] or quarter-wavelength hairpin resonators [15,16] when the fractional bandwidth is less than 40%. However, the considered resonator, on the contrary, shows itself to advantage just in structures of narrowband dual-mode filters. We note that two different values of the ratio l_s/l_r correspond to every absolute value of the difference between f_e and f_o, which defines the passband bandwidth. The smaller value l_s/l_r corresponds to the inequality $f_o > f_e$ and the greater one corresponds to the inequality $f_o < f_e$. It is obvious that the filters built on the resonators with different ratio l_s/l_r and having the same passbands have to differ in the frequency response in stopbands. This assertion is corroborated by the frequency responses shown in Figures 2 and 3 computed for the dual-mode filters of the second and the fourth order having the tapped couplings with the ports.

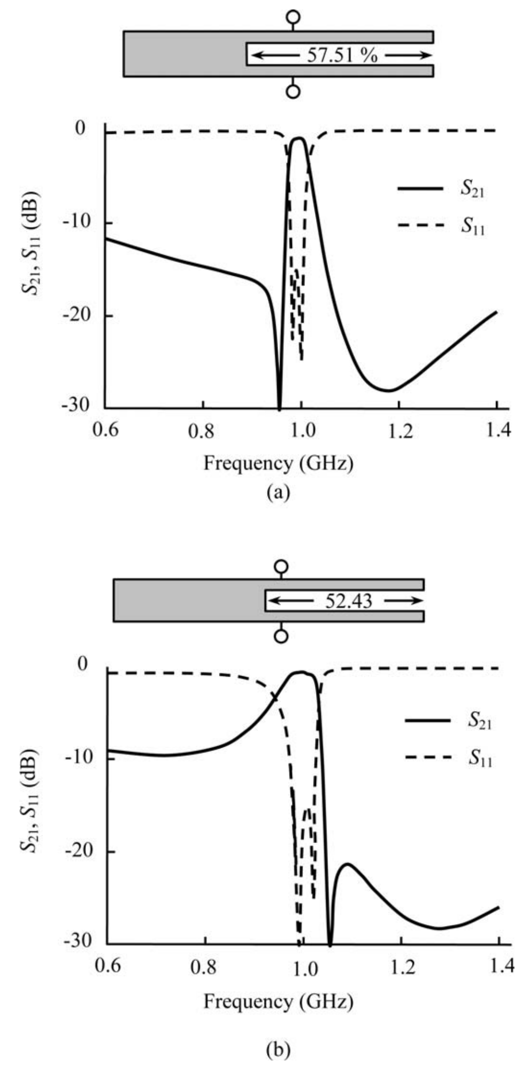
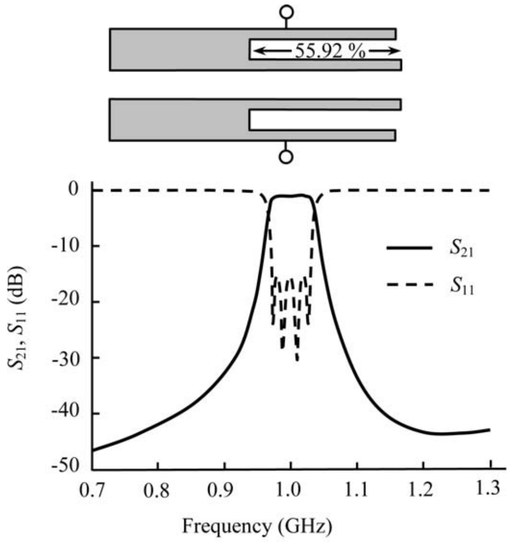


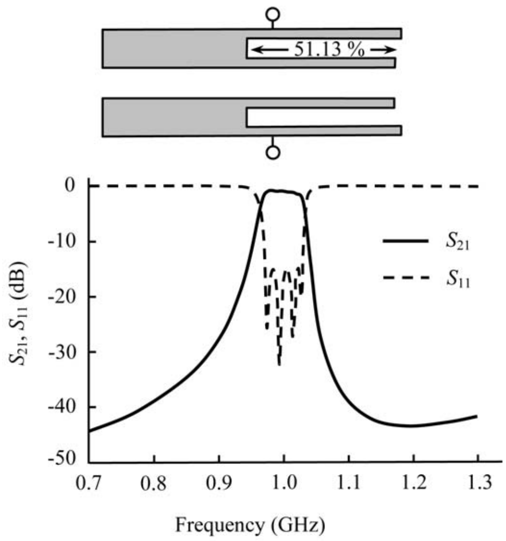
Figure 2 Computed frequency responses of the second-order filter: (a) $f_e > f_o$ and (b) $f_e < f_o$

can see that the filter with the oppositely directed resonators has sharper rejection skirts.

To evaluate accuracy of the numerical analysis used in filter design, the prototype filter of the sixth order had been fabricated. The photo and frequency response of the prototype filter are shown in Figure 5. The filter was manufactured on a ceramic substrate with dimensions of 59.2 × 36.6 × 1.0 mm^3. The substrate was fulfilled of alumina with permittivity $\varepsilon_r = 9.8$. The filter has three split microstrip resonators directed oppositely. The input and output resonators have distributed electromagnetic couplings with the ports. These couplings are ensured with the strip conductors having width of 0.25 mm. Besides, the filter structure has also a supplementary narrow nonresonant conductor intended for making an additional weak coupling between the ports. This weak coupling forms an additional transmission zero in the upper stopband near the spurious passband that enables widening the upper stopband of the filter in some degree.



(a)

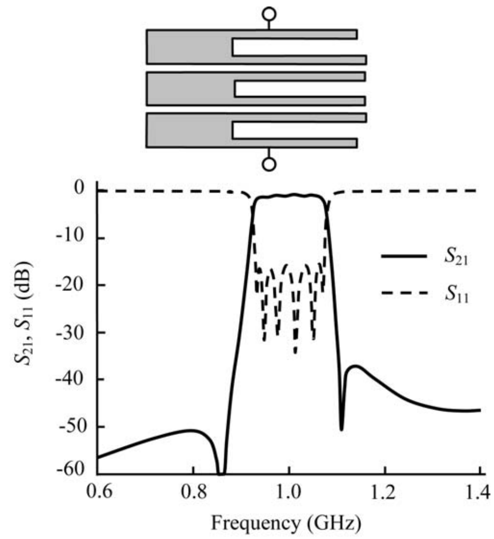


(b)

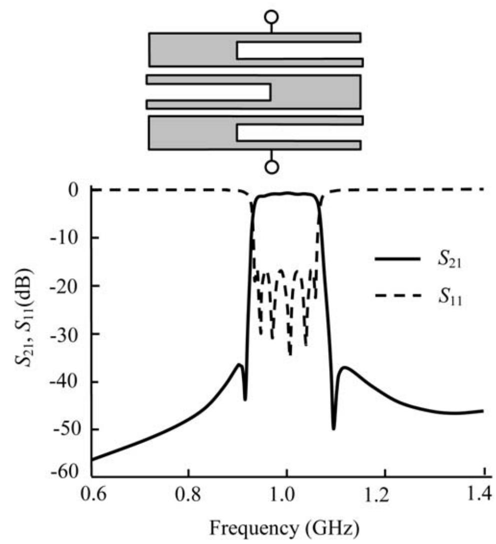
Figure 3 Computed frequency responses of the fourth-order filter: (a) $f_e > f_o$ and (b) $f_e < f_o$

It should be noted that in the fourth-order filter the lengths of the split conductors lying astride the slot in each resonators have to be different. The longer conductors may be either internal ones, as shown in Figure 3(a), or external ones, as shown in Figure 3(b), depending on the relative length of the splitting slot. This difference between lengths of internal and external split conductors is made in order to tune the resonant frequencies of coupled single-mode resonators that are equivalent to the dual-mode split resonator. The initial resonance detuning was induced by asymmetric actions of the tap coupling and the interresonator coupling.

In the sixth-order filter shown in Figure 4, the center resonator due to its symmetrical position has equal length of the conductors on the split segment. The split ends of adjacent resonators in the structure may be either same or opposite directed. These two alternatives for the case $f_e > f_o$ are shown in Figures 4(a) and 4(b). Comparing the frequency responses, one



(a)



(b)

Figure 4 Computed frequency responses of the sixth-order filter: (a) codirectional resonators and (b) oppositely directed resonators

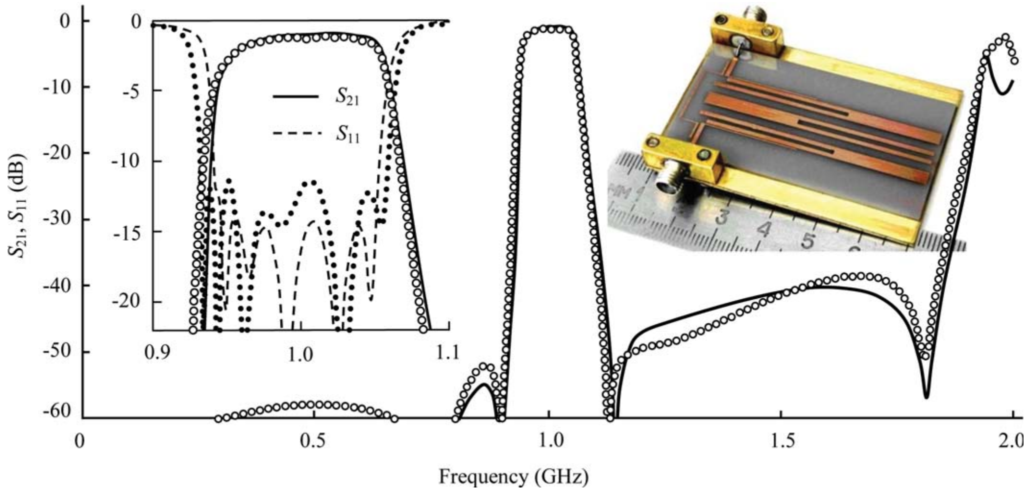


Figure 5 Computed (lines) and measured (dots) frequency responses and photograph of the prototype filter of the sixth order. Inset shows fragment of the passband. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

In Figure 5, one can see that the measured results for the fabricated filter structure are in a rather good agreement with computed frequency response of its model. The prototype filter has the center frequency $f_0 = 1.0$ GHz, the 3-dB fractional bandwidth $\Delta f/f_0 = 11.4\%$, and the minimum return loss in the passband $R = 11.4$ dB.

4. MICROSTRIP DUAL-BAND BANDPASS FILTER

The bandpass filters with two passbands where characteristics of each passband may be adjusted independently can be easily designed using the split microstrip resonators. The photograph and the frequency response of such the prototype filter are shown in Figure 6. The device structure has three oppositely directed split resonators, the external ones of which are connected to the ports through the lumped capacitors placed in certain points on the outer split conductors. The filter has been made on the substrate of alumina ($\varepsilon_r = 9.8$). It has dimensions $57 \times 17 \times 1$ mm^3.

During the process of the filter adjusting the levels of the reflection maxima in each passband are controlled by shifting the capacitors and varying their value. Increase of coupling between the external resonator and the port above an optimal value raises the reflection maxima in the passbands and lowers transmission in the stopbands.

The even oscillation mode of every resonator in this structure is involved in forming the lower passband, whereas the odd mode is involved in forming the upper passband as the splitting slot is sufficiently short. It is obvious that lengthening the slot results in lowering the upper passband and consequently in its approaching to the lower passband.

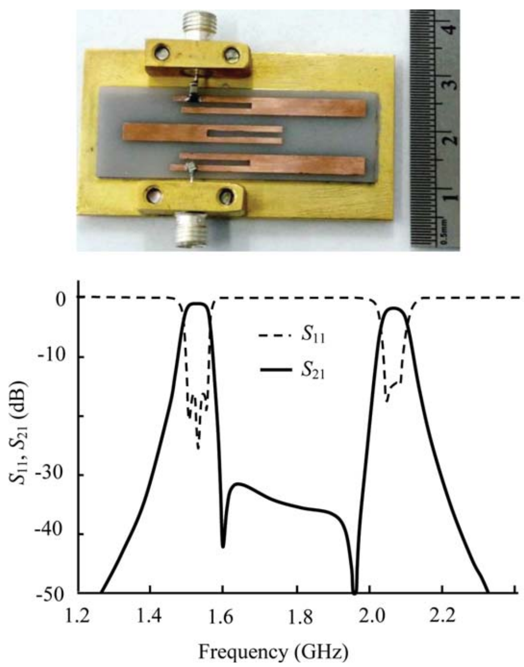


Figure 6 Dual-band microstrip filter of the third order and its measured frequency response. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

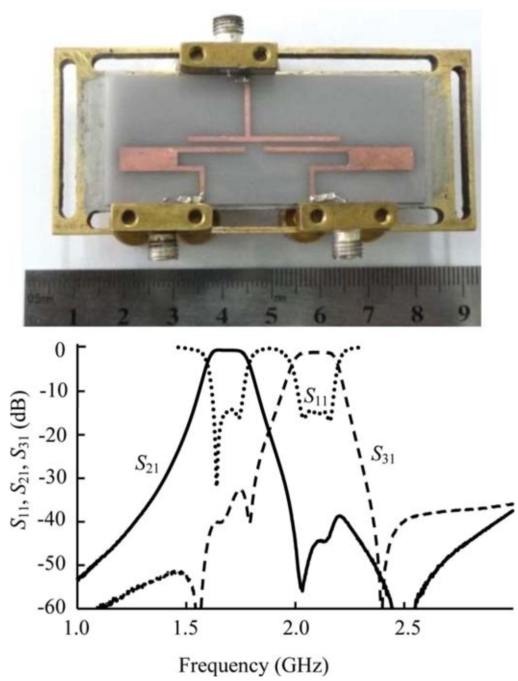


Figure 7 Microstrip diplexer of the third order using two split resonators and its measured frequency response. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

The mean width of two passbands is adjusted by varying the spacings between the resonators, where the widths of both passbands simultaneously increase as the spacings decrease and vice versa. Decrease of the coupling length between the adjacent resonators results in narrowing the lower passband and widening the upper passband.

Two passbands of the prototype filter presented in Figure 6 have the center frequencies $f_1 = 1.527$ GHz and $f_2 = 2.069$ GHz, the 3-dB bandwidths $\Delta f_1 = 71$ MHz and $\Delta f_2 = 71$ MHz, the minimum insertion losses $L_1 = 1$ dB and $L_2 = 2$ dB, and the minimum return losses $R_1 = 17$ dB and $R_2 = 14$ dB.

5. MICROSTRIP DIPLEXER

It is convenient for the split microstrip resonators to be used along with the dual-mode T-shaped microstrip resonator in order to design diplexers. The photo of the prototype device in Figure 7 shows an example of such the diplexer of the third order. Here two split resonators are electromagnetically coupled with one T-shaped resonator. The device structure was made on the alumina substrate ($\varepsilon_r = 9.8$) with dimensions $58 \times 27 \times 1$ mm^3. One end of the T-shaped resonator is connected to port 1 through the miniature capacitor $C_1 = 1$ pF. One end of each split resonator is connected to one of the rest ports (port 2, port 3) through the corresponding capacitor too. They have values $C_2 = 0.6$ pF and $C_3 = 0.4$ pF. For easy assembly of the structure, the strip conductor that is connected to the corresponding port is bent toward the substrate edge. Such bend does not hamper the normal functioning of the device.

Both oscillation modes of one of the split resonators and one of two modes of T-shaped resonator are involved in forming the passband of each channel. The center frequency of each channel is tuned by varying the total length l_r of the corresponding split resonator and the length of the corresponding stub of the T-shaped resonator. The bandwidth of each channel is adjusted by varying the slot length l_s of the corresponding split resonator and the coupling between the same split resonator and the T-shaped resonator. This coupling may be adjusted obviously either by varying the coupling length or the spacing between the resonators. The reflection maxima in the channel passbands are adjusted by varying the capacities.

The key features of the prototype device shown in Figure 7 are characterized by the following parameters. The lower channel has the center frequency $f_1 = 1.69$ GHz, the 3-dB bandwidth $\Delta f_1 = 0.19$ GHz, the minimum insertion loss $L_1 = 0.6$ dB, and the minimum return loss $R_1 = 14.2$ dB. The upper channel has the center frequency $f_2 = 2.09$ GHz, the 3-dB bandwidth $\Delta f_2 = 0.21$ GHz, the minimum insertion loss $L_2 = 1.1$ dB, and the minimum return loss $R_2 = 14.4$ dB.

6. CONCLUSION

Thus, the carried out study showed that the dual-mode microstrip split resonator possesses a number of advantages, due to which it is practical to use the resonator in all sorts of microwave devices. First, the additional lower oscillation mode arises in the resonator because of splitting its uniform strip conductor by the longitudinal slot with the same size of the resonator, so it differs in high diminutiveness. Second, the frequencies of its even and odd modes can be varied independently in a wide range. Third, in such resonator two lower resonant frequencies may be arbitrary close to each other. These features allow designing a miniature narrowband bandpass filters differing in high performance. Finally, it is easy to organize the electromagnetic coupling between dual-mode split resonators. All the mentioned advantages of the dual-mode split microstrip resonators are supported by high performance of the prototype microwave devices presented in the article. It is significant that the frequency responses obtained by numerically analyzing the structure models are in good agreement with the results of the prototype device measurements.

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