Microstrip Slow-Wave Open-Loop Resonator Filters with Reduced Size and Improved Stopband Characteristics

Prayoot Akkaraekthalin and Jaruek Jantree

This paper presents a new class of microstrip slow-wave open-loop resonator filters with reduced size and improved stopband characteristics. A comprehensive treatment of both ends loaded with triangular and rectangular ends is described, leading to the invention of a microstrip slow-wave open-loop resonator. Two-resonator and four-resonator bandpass filters are designed at the operating frequency of about 2 GHz, and a bandwidth of 60 MHz. The size of the slow-wave open-loop resonator is optimized from the standpoint of the unloaded Q-factor. The filters are not only compact in size due to the slow-wave effect, but also have a wider upper stopband resulting from the dispersion effect. The filter designs of this type are described in details. The experimental results are demonstrated and discussed.

Keywords: Microstrip slow-wave open-loop resonator, bandpass filter.

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I. Introduction

Radio frequency and microwave planar bandpass filters are presently required in a wide variety of applications of wireless communication systems [1]. Recent advances in high-temperature superconducting (HTS) circuits and microwave monolithic integrated circuits (MMIC) have additionally stimulated the development of various planar filters, especially narrow-band bandpass filters which play an important role in modern communication systems [2]-[6]. Currently, filters with compact size which suppress spurious sidebands and have wider upper stopbands are necessarily required for several wireless communication systems. However, most of the planar bandpass filters built on microstrip structures are large in size and their first spurious resonance frequencies appear at $2f_0$ and $3f_0$, where f_0 is the center frequency, which maybe closed to the desired frequencies. The half-wavelength resonators inherently have a spurious passband at $2f_0$, while quarter-wavelength resonator filters have the first spurious passband at $3f_0$, but they require short-circuit connections with via holes, which are not quite compatible with planar fabrication techniques. Previously, most studies proposed techniques for the suppression of spurious passbands of the filters using microstrip parallel-coupled lines. Parallel-coupled microstrip filters with over-coupled end stages have been proposed to extend the electrical length of the odd mode to compensate the difference in the phase velocities [7]. The wiggly-line microstrip filter using a continuous perturbation of the width of the coupled lines following a sinusoidal law has been studied [8]. Parallel-coupled microstrip filters with the width of slots in the ground

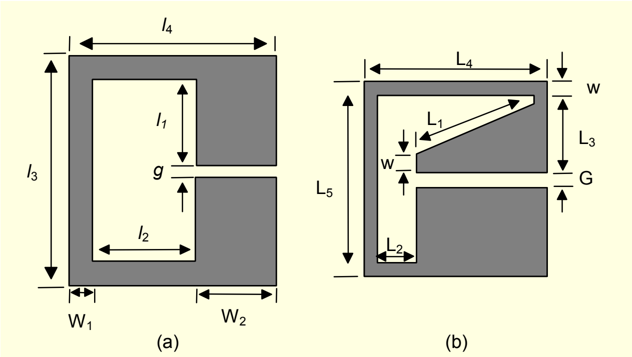

Fig. 1. (a) A conventional resonator and (b) the proposed microstrip slow-wave open-loop resonator.

plane adjusted to compensate unequal modal electrical lengths to obtain spurious suppression have been proposed [9], [10].

The substrate suspension structure has been designed to substantially speed up the even-mode phase velocity to make the modal phase velocities equal for the suppression of spurious bands [11]. These published microstrip-coupled line filters have some key drawbacks. All of these filters are very long due to their straight structures, especially when the filter order becomes high. The filters in [9] and [10] need ground-plane apertures and that in [11] needs a suspended substrate, which increases the complexity and cost. Besides, many other proposed filters use stepped impedance resonators (SIR) to shift away the first higher order resonance frequency [12]-[14]. However, the step impedance method is to move, not to suppress, the first spurious band and sometime a large impedance stepping ratio is required and makes the layout of the filter difficult. Capacitively-loaded transmission lines as slow-wave open-loop resonators and SIR have been found advantageous for controlling the spurious bands, however, their spurious responses are still large [12]-[16].

Cross-coupled resonator filters are traditionally realized using waveguide cavities or dielectric resonators, which are bulkier than planar structures. Previously, planar cross-coupled filters have been proposed, which are mostly based on open-loop microstrip resonators [17]-[19]. The cross-coupling between nonadjacent resonators creates transmission zeros that improve the skirt rejection. In order to obtain the transmission zeros, this filter structure needs at least four resonators, however, it has been found recently that by using a 0^{o} feed structure, two transmission zeros near the passband can be created and the stopband rejection is significantly increased [14]. Therefore, we introduce a new class of microstrip slow-wave open-loop resonator bandpass filters based on cross-coupled and 0^{o} feed structures. The proposed filters include maximally flat and elliptic or quasi-elliptic function responses, resulting in not only more compact size but also a wide upper stopband.

In section II, a brief comparison of the proposed microstrip slow-wave open-loop resonators and the conventional one is made in view of their sizes, coupling characteristics, and fundamental and first spurious resonance frequencies. In section III, we describe the size optimization of the proposed resonator from the standpoint of the unloaded Q-factor. The designs and measured results of both a two-resonator filter and a four-resonator cross-coupling filter using the proposed resonators are described in detail in section IV. Finally, our conclusions are given in section V.

II. Comparison of the Conventional and Proposed Microstrip Slow-Wave Open-Loop Resonators

Figure 1(a) shows a conventional microstrip hairpin resonator [14]. The proposed microstrip slow-wave open-loop resonator composed of a microstrip line loaded with triangular and rectangular ends is shown in Fig. 1(b). The slow-wave behavior is caused by the capacitive loading effect from both ends. The resonance responses of the conventional and proposed microstrip slow-wave open-loop resonators are computed by using IE3D [20], a commercial electromagnetic simulator. An RT/Duroid 3003 substrate, which has a given dielectric constant of 3 and a thickness of 1.524 mm is used, resulting in curves as given in Fig. 2. The resonance responses are obtained from the microstrip slow-wave open-loop resonators with the same size. It can be noticed that the proposed resonator has a resonance frequency of 2 GHz, compared to 2.45 GHz of the conventional one. This means that the proposed microstrip resonator can be made smaller than the conventional structure when they resonate at the same frequency. The coupling coefficients are also computed by using IE3D simulation software. Figure 3 shows a magnetic coupling. The variation of coupling coefficient K versus the distance S between two resonators is drawn by dashed and solid lines for the conventional resonators and the proposed resonators, respectively. It is seen that the proposed resonators have smaller K than conventional resonators. This can be explained by the fact that both the conventional and proposed resonators have electric and magnetic coupling, but the magnetic couplings are dominant. The magnetic fringing fields of the conventional resonators are stronger near the center of the resonators, but the magnetic fringing fields of the proposed microstrip resonators distribute asymmetrically at the resonator line. Therefore, both the size and the coupling property of a resonator, as well as the specifications of a filter, need to be considered in choosing resonators. Roughly speaking, if a narrowband filter is designed, smaller coupling coefficients between resonators will be required [2], therefore the proposed

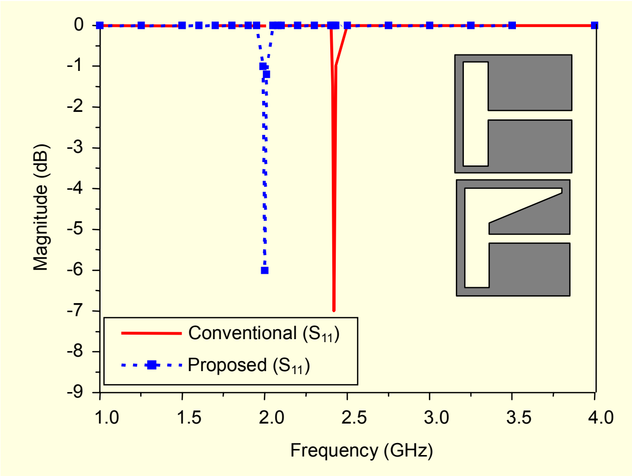

Fig. 2. Performance of the conventional resonator with l_1 = 4.2 mm, l_2 = 0.5 mm, l_3 = 9 mm, l_4 = 9 mm, W_1 = 0.5 mm, W_2 = 8 mm, and g = 0.5 mm; and the proposed microstrip slow-wave open-loop resonator with L_1 = 7 mm, L_2 = 0.5 mm, L_3 = 4.25 mm, L_4 = 9 mm, L_5 = 9 mm, W = 0.5 mm, and G = 0.5 mm.

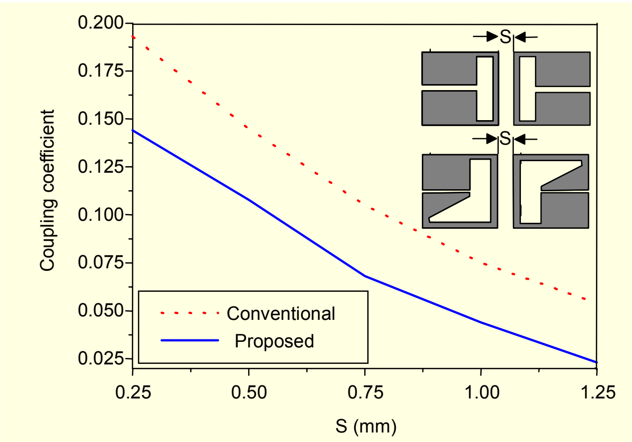

Fig. 3. Variation of the coupling coefficient K versus the distance S between two coupled resonators of the conventional resonator with l_1 = 4.87 mm, l_2 = 0.5 mm, l_3 = 10.75 mm, l_4 = 10.75 mm, W_1 = 0.5 mm, W_2 = 9.75 mm, and g = 0.5 mm; and the proposed resonator with L_1 = 7 mm, L_2 = 0.5 mm, L_3 =4.25 mm, L_4 = 9 mm, L_5 = 9 mm, W = 0.5 mm, and G = 0.5 mm.

microstrip slow-wave open-loop resonators are preferred; if a wideband response is required, it may be preferable to choose conventional resonators.

III. An Optimized Resonator and Its Bandstop Characteristic

1. An Optimized Resonator

The size of the proposed microstrip slow-wave open-loop resonator has been optimized from the standpoint of the unload

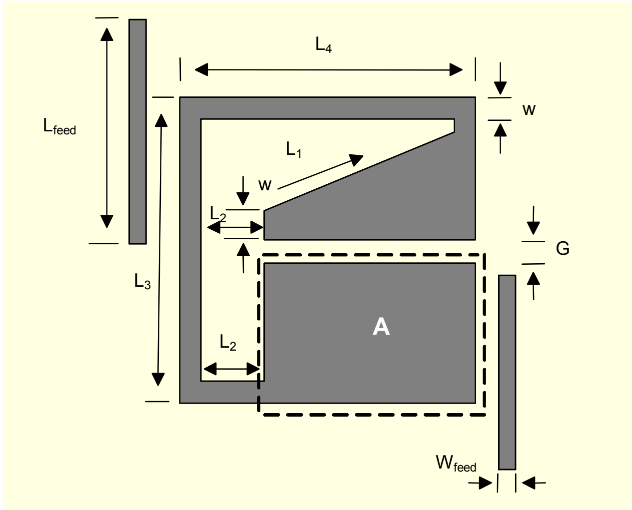

Fig. 4. Layout for the simulation of Q_u with L_{feed} = 11 mm, W_{feed} = 0.5 mm, and $L = L_3+L_4$ mm when $L_3 = L_4$.

Q-factor of the resonator with an IE3D simulator. We assume that the conductivity of the metal is $\sigma = 5.8 \times 10^7$ [S/m], the dielectric loss tangent is tan δ = 0.0013, the dielectric constant is 3, and the thickness is 1.524 mm. In Fig. 4, the line width W and gap G are varied, while the lengths L and L_2 have been adjusted so that the resonance frequency of the microstrip slow-wave open-loop resonator f_0 is equal to 2 GHz. The unloaded *Q*-factor Q_u can be calculated by the following equation [21]:

$$Q_u = \frac{Q_L}{1-10^{-\frac{L_0}{20}}}, \tag{1}$$

where Q_L is the loaded *Q*-factor and L_0 is the insertion loss in decibels of the resonance frequency that is obtained from the simulated result of the resonator. The resonator layout used for our simulation is shown in Fig. 4. Figure 5(a) shows the variation of the unloaded *Q*-factor Q_u against the line width *W* and the gap G. It is seen that as the width of gap G increases, the unloaded *Q*-factor Q_u will increase and saturate when the width W is a large number, however, the area of the resonator increases. The key area of the resonator is assumed to be A, since when adjusting width W and gap G, area A of the resonator will be affected more than other parts. We obtain Q_u /A by dividing the unloaded *Q* -factor Q_u by occupation area A, as shown in Fig. 5(b). The higher the Q_u /A value is, the higher the *Q* -factor Q_u for each unit area is. In other words, the area of the resonator that has the same *Q*-factor Q_u can be reduced, so we can reduce the area of the filter without increasing its insertion loss. The point is, G = 0.5 mm and W = 0.5 mm, seem to be optimum because the Q_u /A value is very high as we can clearly see in Fig. 5(b).

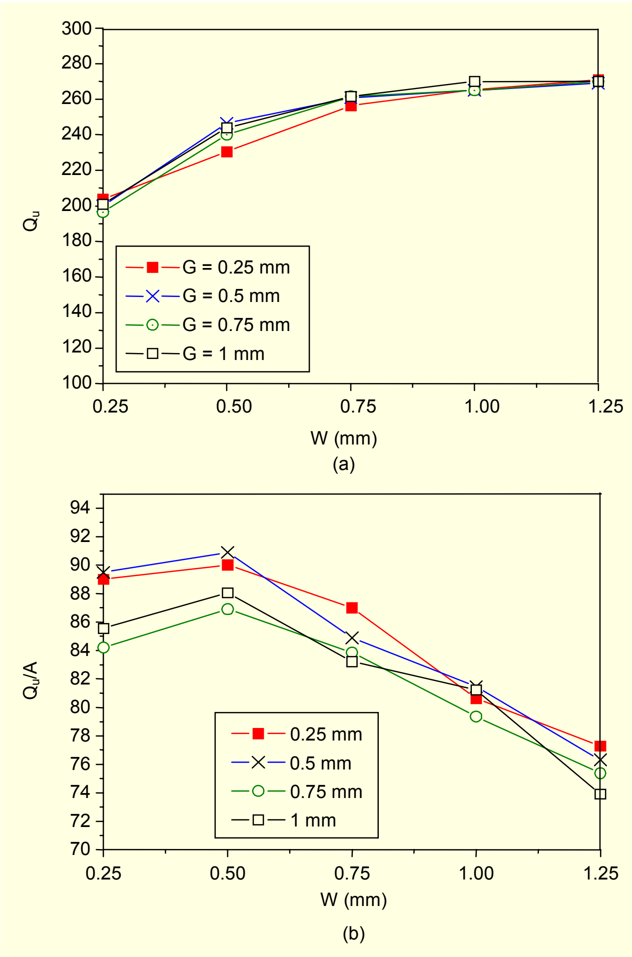

Fig. 5. Variation of Q_u and Q_u/A with the W and G of the proposed slow-wave open-loop resonator.

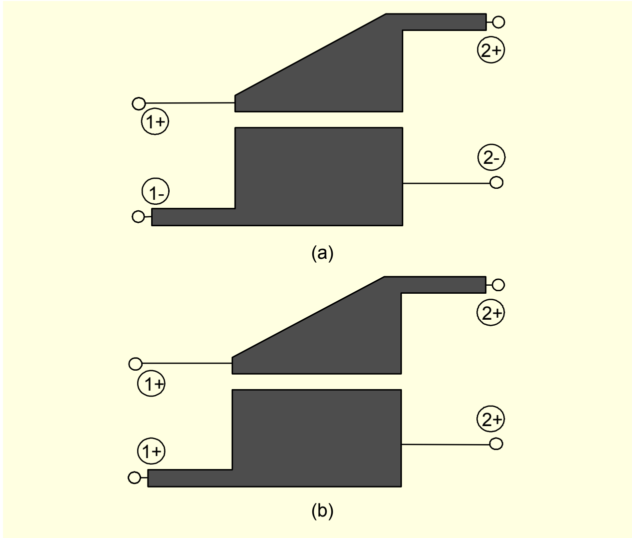

Fig. 6. The loading capacitance part of the proposed resonator forming an asymmetrical parallel-coupled line and modeled as an IE3D differential two-port for (a) odd- and (b) even-mode excitations.

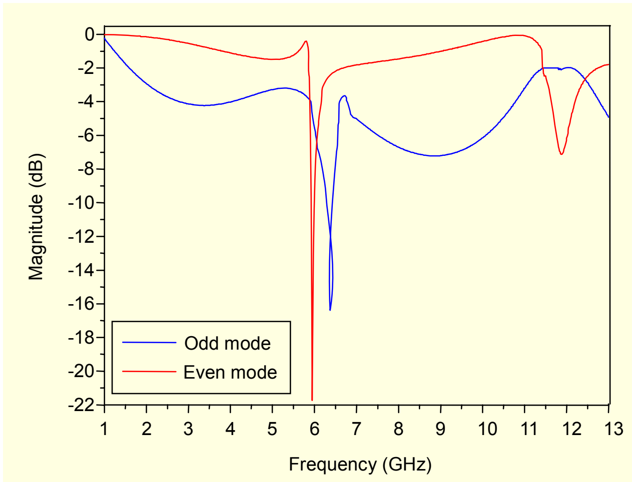

Fig. 7. Frequency responses (S_{21}) of the asymmetrical parallel-coupled line with odd- and even-mode excitations.

2. Bandstop Characteristic for Spurious Suppression

The proposed microstrip slow-wave open-loop resonator has been optimized, resulting in the final dimensions as shown in the previous subsection. The resonator has an inherent bandpass characteristic with the fundamental resonance frequency of 2 GHz and the first spurious resonance frequency of about 6 GHz. The loading capacitances at the open-ends of the proposed resonator have been further studied in detail. These loading capacitance parts are formed in an asymmetrical parallel-coupled line structure as shown in Fig. 6. The IE3D has been employed to evaluate the characteristics of the asymmetrical parallel-coupled line section by using differential two-port models for two modes of excitations, known as odd- and even-modes, as shown in Fig. 6(a) and 6(b), respectively [20]. Figure 7 demonstrates the simulated frequency responses (S_{21}) to which it can be seen that the asymmetrical parallel-coupled line has notch responses. The odd- and even- mode notches appear at the frequencies of 6.36 GHz and 6.0 GHz, respectively, closed to the first spurious response frequency of the proposed resonator. These odd- and even-mode responses will certainly affect the resonator, resulting in a bandstop characteristic; therefore, superior suppression of the first spurious response may be obtained.

IV. Filter Design and Measured Results

1. Coupling Structures

The IE3D simulator, which is based on the method of moments which proves to be quite accurate in its prediction, was used to simulate the frequency responses of the three basic coupling structures. The coupling in each structure can be specified by the two dominant resonance frequencies, which are split off from the resonance condition due to the electromagnetic coupling. The coupling coefficients M_{ij} for resonators i and j would then be extracted from the simulated frequency responses by using [15].

$$M_{ij} = \pm \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2}, \tag{2}$$

where f_1 or f_2 corresponds to either even- or odd-mode resonance frequency. The external quality factor may be characterized by

$$Q_e = \frac{f_0}{f_{3dB}} \quad , \tag{3}$$

where f_0 and f_{3dB} are the resonance frequency and 3 dB bandwidth of the input or output resonator when it alone is externally excited. The layouts of the two-resonator and four-resonator bandpass filters using the proposed microstrip slow-wave open-loop resonators and 0° feed structures, of which the signals at the input and output are in phase, are shown in Fig. 8 and Fig. 11, respectively.

2. Two-Resonator Bandpass Filter

The first filter is an example of a two-resonator bandpass filter, which consists of the proposed microstrip slow–wave open-loop resonators. The filter was then fabricated on an RT/Duroid 3003 substrate with a relative dielectric constant of 3 and a thickness of 1.524 mm. In order to have a maximally flat passband bandwidth of 60 MHz (or the fractional bandwidth of FBW=3%) at f_0 equal to 2 GHz, the external quality factor Q_e and the coefficient M_{21} were calculated as

$$Q_e = \frac{c_0 c_1}{FBW} = 47.1, \tag{4}$$

$$M_{21} = -\frac{FBW}{\sqrt{c_1 c_2}} = -0.021, \tag{5}$$

where c_0, c_1, and c_2 are element values. The spacing S_{12} between the resonators in Fig. 8 was adjusted to obtain the appropriate external quality factor and coupling coefficient with the aid of an IE3D EM simulator [16]. The filter parameters were obtained in the following: $L_1 = 7$ mm, $L_2 = 0.5$ mm, $L_3 = 4.25$ mm, $L_4 = 9$ mm, $L_5 = 9$ mm, G = 0.5 mm, and W = 0.5 mm, with spacing S_{12} equal to 1.12 mm.

Figure 9 shows a photograph of the fabricated filter. The size of this two-resonator filter is about $0.095\lambda_{go}$ by $0.202\lambda_{go}$, where λ_{go} is the guided wavelength of a 50-Ω line on this substrate at the center frequency. The fabricated filter was then measured on an Agilent 8719ES network analyzer, with the measured and theoretical performance as shown in Fig.10. Figure 10(a) gives the details of the passband response. Due to the 0° feed structure, we can notice the appearance of two transmission zeros near the passband [14]. Figure 10(b) shows the wideband response. We can notice that the measured result shows a slight deviation in the center frequency and bandwidth. Also from the measured data, we found that the passband insertion loss is approximately 2.5 dB at the center frequency of 2.03 GHz, which is mainly due to the conductor loss of copper. The passband insertion loss curve of the proposed bandpass filter is sharper than that of the conventional filter. The passband return loss is greater than 20 dB and the out-of-band rejection is better than 20 dB at the lower stopband and 20 dB at the upper stopband. The two transmission zeros are at 1.91 GHz with 42 dB rejection and 2.13 GHz with 31 dB rejection, respectively. The filter exhibits a wide upper stopband with a rejection better than 20 dB up to about 10 GHz, and the first spurious response

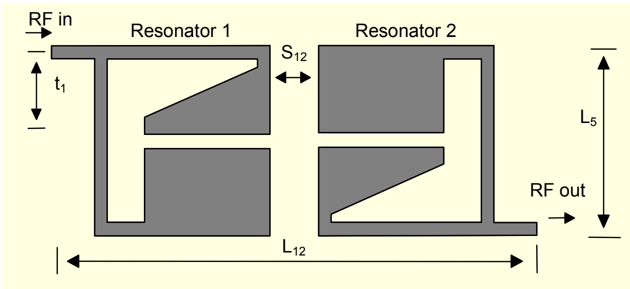

Fig. 8. Layout of the two-resonator designed filters with L_1 = 7 mm, L_2 = 0.5 mm, L_3 = 4.25 mm, L_4 = 9 mm, L_5 = 9 mm, W = 0.5 mm, G = 0.5mm, L_{12} = 27.62 mm, and t_1 = 4 mm.

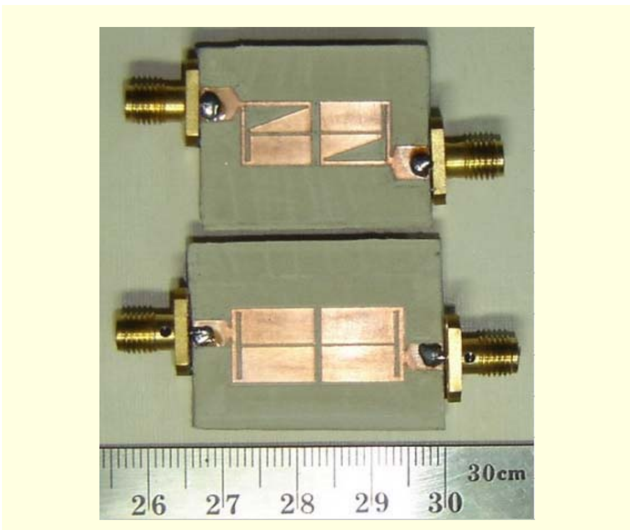

Fig. 9. A photograph of the proposed filter (top) and conventional filter (bottom).

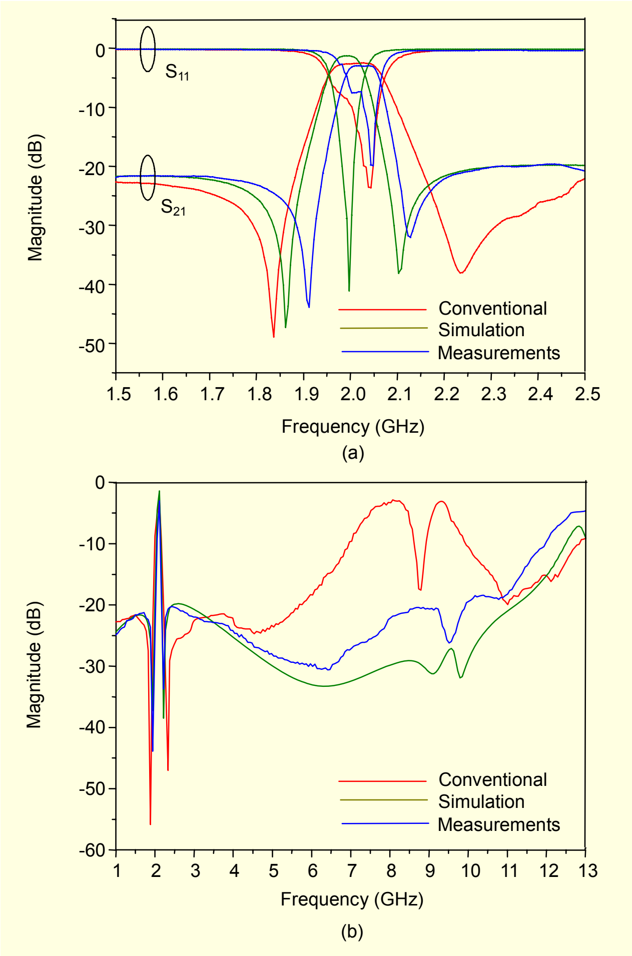

Fig. 10. Comparison of measured and simulated responses of the two-resonator slow-wave open-loop resonator filters' (a) passband response and (b) wideband response.

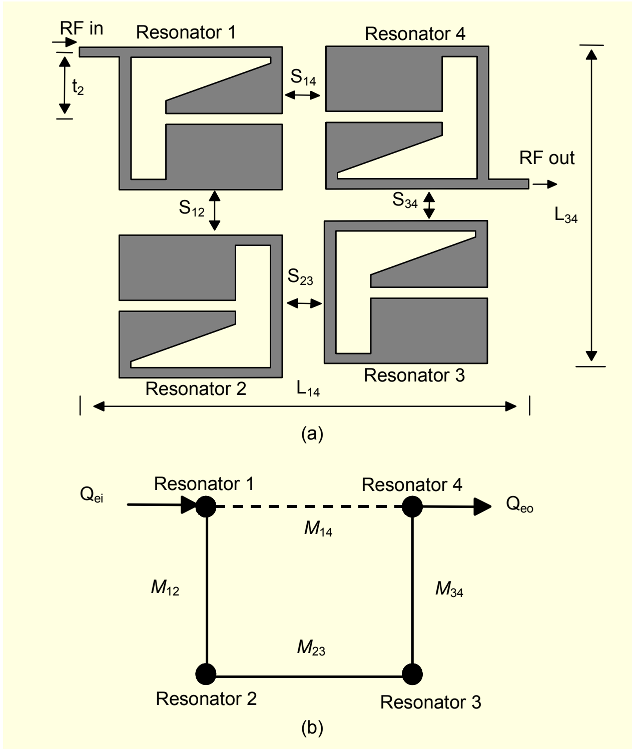

Fig. 11. (a) Layout of the four-resonator designed filters with L_1 = 7 mm, L_2 = 0.5 mm, L_3 =4.25 mm, L_4 = 9 mm, L_5 = 9 mm, W = 0.5 mm, G = 0.5 mm, S_{14} = 1.25 mm, S_{23} = 1.5 mm, S_{34} = 1.62 mm, S_{12} = 2.75 mm, L_{34} = 20.75 mm, L_{14} = 28 mm, and t_2 = 4 mm. (b) Typical coupling structure.

is about 6 dB at 12.7 GHz. The first spurious frequency is higher compared with the conventional structure because of the higher dispersion effect of the loading capacitance [15].

3. Four-Resonator Cross-Coupled Filter

The second filter is an example of a four-resonator cross-coupled microstrip slow-wave open-loop resonator filter, designed and fabricated on the same microwave substrate and same resonator size of the designed two-resonator filter. The configuration of a four-resonator cross-coupled microstrip slow-wave open-loop resonator filter is shown in Fig. 11(a). Figure 11(b) shows the typical coupling structure of the four-resonator cross-coupled bandpass filter, where each node represents a resonator. The solid line and dotted line represent the direct coupling route and the cross-coupling route, respectively. The filter specifications are the center frequency of 2 GHz, a passband ripple of 0.01 dB, and the passband bandwidth of 60 MHz.

The filter could be synthesized by using a method described in [22], from which the lumped-element values of a low pass prototype were determined as c_0 = 1, c_1 = 0.95947, c_2 = 1.42292, J_1=–0.21083, and J_2 = 1.11769. The design parameters of the bandpass filter (namely, the elements of the coupling matrix and the input/output single-loaded external Q_e) could then be calculated as

$$M_{12} = M_{34} = \frac{FBW}{\sqrt{c_1 c_2}} = 0.025, \tag{6}$$

$$M_{23} = \frac{FBW \cdot J_2}{c_2} = 0.023, \tag{7}$$

$$M_{14} = \frac{FBW \cdot J_1}{c_1} = -0.006, \tag{8}$$

$$Q_e = \frac{c_0 c_1}{FBW} = 31.99. \tag{9}$$

The filter design is based on knowledge of the coupling coefficients of the three basic coupling structures, which are electric, magnetic, and mixed coupling, as described in [15]; therefore, the offset between resonators 2 and 3 is 1.25 mm. The

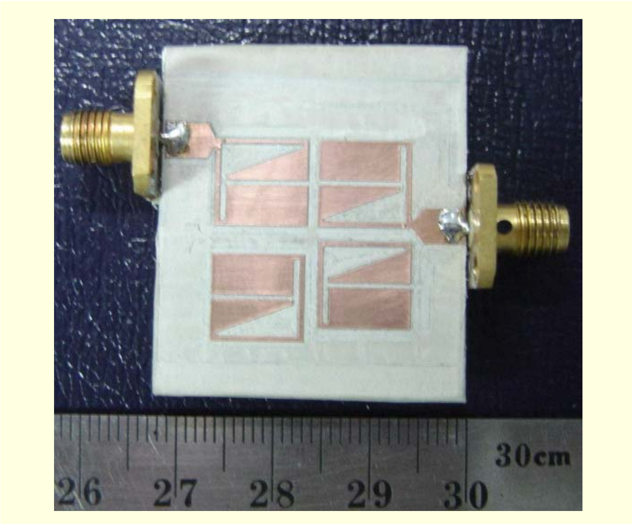

Fig. 12. A photograph of the fabricated four-resonator bandpass filter using the proposed microstrip slow-wave open-loop.

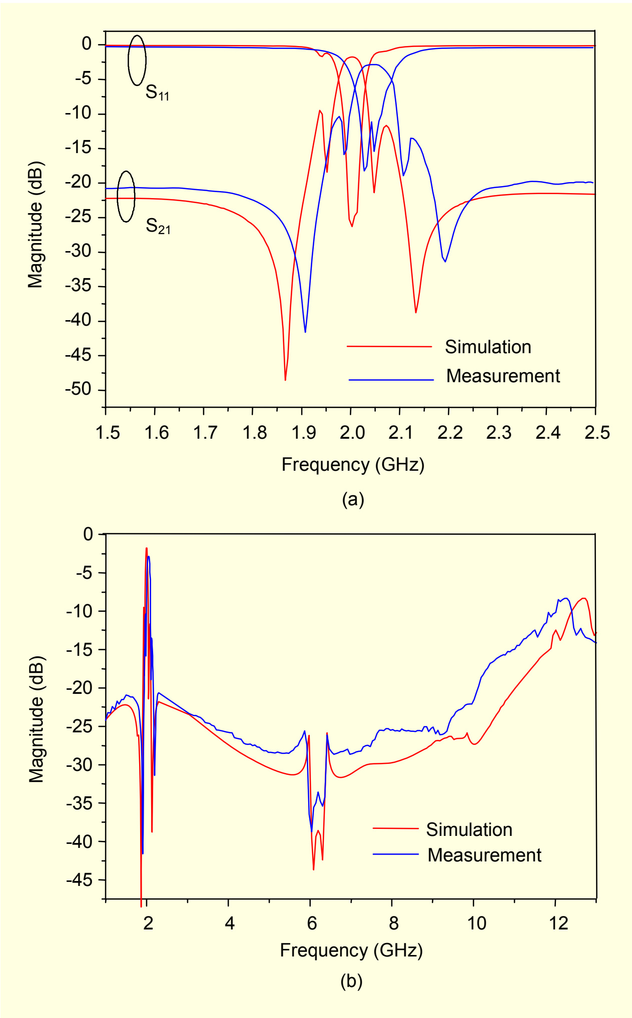

Fig. 13. Comparison of measured and simulated responses of the four-resonator slow-wave open-loop resonator filter (a) passband response and (b) wideband response.

spacing between the resonators in Fig. 11(a) has been determined as the following: S_{14} = 1.25 mm, S_{23} = 1.5 mm, S_{34} = 1.62 mm, and S_{12}= 2.75 mm. Figure 12 shows a photograph of the fabricated four-resonator bandpass filter using new microstrip slow-wave open-loop resonators. In this case, the size of the filter is only $0.219\lambda_{go}$ by $0.206\lambda_{go}$. Figure 13(a) shows the measured and theoretical data. It can be expected that there is a single pair of transmission zeros near the passband due to the cross-coupling effect. The effect of the two transmission zeros at 1.98 GHz and 2.09 GHz is observed. It can be also clearly observed that there are two extra transmission zeros on opposite sides of the passband due to the 0^{o} feed structure. One of the extra transmission zeros is at 1.93 GHz and the other is at 2.21 GHz. The measured passband insertion loss is approximately 2.78 dB at the center frequency of 2.05 GHz, which again is attributed to the conductor loss of copper. The passband return loss is greater than 18 dB. The out-of-band rejection is better than 22 dB at the lower stopband and 20 dB at the upper stopband. In Fig. 13(b), the filter exhibits a wide upper stopband with a rejection better than 25 dB up to about 10.5 GHz. It can also be clearly seen that there is an unnatural dip in the response at f_1 around 6 GHz. The reason for this dip is that the bandstop characteristic of the asymmetrical parallel coupled-line of the proposed resonator causes superior suppression of the spurious response as previously discussed.

V. Conclusions

We have presented two bandpass filters designed using the proposed microstrip slow-wave open-loop resonators. The size of the resonators has been optimized from the standpoint of the resonators' unload *Q*-factor. These bandpass filters have been designed at an operating frequency of about 2 GHz and a bandwidth of 60 MHz. The filters are not only compact in the size, but also have a wider upper stopband as a result of the asymmetrical parallel-coupled line section of the proposed resonator. The measured responses have very good agreement with our simulation expectations.

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