Progress In Electromagnetics Research B, Vol. 16, 277–290, 2009

COMPACT NARROWBAND BANDPASS FILTER USING DUAL-MODE OCTAGONAL MEANDERED LOOP RESONATOR FOR WIMAX APPLICATION

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Abstract—In this paper, a new design of a compact narrowband bandpass filter is proposed. This new narrowband bandpass filter is designed using an octagonal form of dual-mode closed-loop microstrip ring resonator based on a meander structure in order to achieve compactness. The designed filter has a 3 dB fractional bandwidth (FBW) of 5% at 2.3 GHz. The filter has been fabricated on Taconic CER-10 substrate having 0.64 mm thickness and a relative dielectric constant of 10. Experimental results show good agreement with simulated values. Apart from WiMax, this new model of filter is also useful for WLAN and mobile communication applications, since it is compact in size, low loss, and low cost with good performance of elliptic response with sharp rejection and adequate fractional bandwidth.

1. INTRODUCTION

Microwave filters play important role in many wireless and communication systems such as satellite and cellular mobile organizations. In such kind of systems some factors in designing microwave filters are of primary importance. Compact size, low cost, low weight, high performance, and low loss are some of these factors that are essentially required for enhancing the system performance and reducing fabrication cost [1, 2]. Parallel coupled microstrip filters have been used for many years since 1958, first proposed

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by Cohn [2]. Although they have many advantages such as wide range of filter fractional bandwidth (FBW) of 5% to 50% and simple design procedure, there are some issues on this type of filter. Major disadvantages of this type of filter include the length of parallel coupled filter that is too long, and it further increases with the order of filter. Some techniques have been developed to solve this problem. One of them was using hairpin-line filters including folded $\lambda/2$ resonator structures [2]. Afterwards, microstrip ring resonators in any shape were considered as the building blocks of microstrip bandpass filters in which they are widely used, because they satisfy the mentioned demands that are essential for microwave filters.

On the other hand, microstrip ring resonators have many interesting characteristics including small size, low fabrication cost, and narrowband [3]. In 1969, one technique for measuring wavelengths and dispersion characteristics of the relative permittivity of a microstrip line was developed by Troughton with the help of ring resonators [4]. Microstrip ring resonators are also used for the measurement of phase velocity, and effective dielectric constant. Since no open-end effects need to be considered in microstrip ring resonators, using them instead of the linear resonator for dispersion measurements is more desirable [4, 5].

The curvature of the ring affects the resonance frequencies, and this effect becomes larger by using lines with small impedances and substrate materials with small relative permittivity. Therefore, wider resonators are more affected in the higher-order resonances by the curvature, and as the width increases, this influence becomes stronger. For resonators with impedances higher than $20\,\Omega$ (for example 50, 90, and $110\,\Omega$) the effective relative permittivity is almost independent of the length of the resonator [4]. One another factor that affects the resonant frequency is the coupling gap. To calculate the coupling gap effects on the resonant frequency, the equivalent circuit of coupling gap that is modeled by a π-network is included in the equivalent circuit of the ring resonator [5]. By using this model it can be found that as the coupling gap decreases, the resonant frequency becomes lower, but for most ranges of the coupling gap size, the influences on resonant frequency are small and negligible [5]. The main discussing concept in this paper is on dual-mode ring resonators. During recent years, they have been considered more and greatly used for microwave bandpass filter in wireless local area network (WLAN) applications and mobile communication systems [6].

Microstrip dual-mode filters have many interesting characteristics, such as narrowband, high Q, easy-to-design, and compact in size. The main advantage of these types of filters is that in dual-mode, each resonator operates as a double tuned resonant circuit and therefore, an n-degree filter can be achieved in more compact configurations due to the halved number of resonators [6, 7]. In addition to using dual-mode filters, there are some other useful methods to achieve a compact size in filter designing. One of these techniques is to have different parts of filter bent. This could be the best solution to get more compact sizes especially for filters with stubs and long straight transmission lines. Ultra-wideband filter reported in [8], and wideband filter reported in [9] are among the structures making use of method of bending the lines.

Although miniaturizing of microwave filters can be done by using substrates with high dielectric constant, or lumped element or different transmission lines such as coplanar waveguide (CPW), reduction in size by changing the geometry of the filters using microstrip transmission lines is more desirable. This is because high dielectric permittivity will often introduce more surface waves and losses, while filters using coplanar waveguide (CPW) can be found in quite compact sizes. However, they also introduce more insertion losses. Hence, microwave filters using coplanar waveguide are seldom used in the millimeter-wave range, despite their wide applicability in monolithic microwave integrated circuits (MMICs) due to their ability in easily integrating series and shunt elements and simplicity of fabrication [10–12].

2. DUAL-MODE MICROSTRIP OCTAGONAL LOOP RESONATOR

As the first step in the design of this new proposed filter, a dual-mode microstip resonator (basic part of our filter) has been designed which resonates at 2.3 GHz on a Taconic CER-10 substrate having $h = 0.64\,\text{mm}$ thickness and a relative dielectric constant of $\varepsilon_r = 10$. The layout of the resonator is depicted in Fig. 1. It is shown in Fig. 1 that the resonator is excited by using gap coupling method where input port, Port 1 and output port, Port 2 are spaced with a gap, G symmetrically on each side. By applying the feedlines gap, $G = 0.1\,\text{mm}$, and coupling stubs, $W_1 = 0.2\,\text{mm}$, optimum insertion loss is obtained. Equations expressed in [7] are used to synthesize W/h (the conductor width W and substrate thickness h of microstrip). The circumference ℓ_c of the ring resonator is expressed as [1, 13]:

$$\ell_c = n\lambda_g \tag{1}$$

where n is the number of mode and λ_g is the guided wavelength. Consequently, the resonant frequency f_0 (the center frequency of the bandpass filter in which this resonator will be used) can be expressed as [13]:

$$f_0 = \frac{nc}{\ell_c \sqrt{\varepsilon_{eff}}} \tag{2}$$

where $c = 3 \times 10^8$ m/s is the speed of light in free space; ℓ_c is the circumference of the ring resonator; ε_{eff} is the effective dielectric constant. As can be seen in Equation (2), the center frequency of the filter is a function of the filter structure. In this design for the first mode ($n = 1$), $\ell_c = \lambda_g = 50.26$ mm. Therefore each side of octagonal ring resonator, $a = 6.28$ mm ($a \approx \lambda_g/8$). The trace width (W) of 0.6 mm will produce a 50 Ω characteristic impedance line. The dimensions of the resonator structure are shown in Table 1.

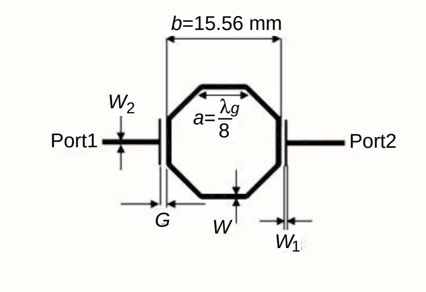

Figure 1. Single mode octagonal ring resonator.

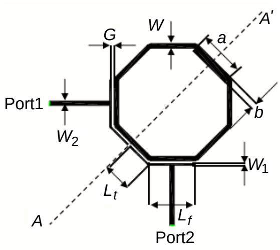

Figure 2. Conventional structure of dual-mode bandpass filter [6].

Table 1. Dimensions of the single mode resonator, with reference to the layout depicted in Fig. 1.

Parameter	Dimension (mm)
a	6.28
W	0.6
W_1	0.2
W_2	0.6
G	0.1

3. DESIGN OF NARROWBAND ELLIPTIC BANDPASS FILTER USING DUAL-MODE MICRISTRIP OCTAGONAL LOOP RESONATOR

A microstrip dual-mode resonator in any shape and usually symmetrical in two dimensions (2-D) can be described by Wheeler's cavity model [7]. According to this model, different modes corresponding to the electromagnetic (EM) fields inside the cavity are described in terms of TM^z_{mn0} modes (where Z is perpendicular to the ground plane). Different modes introduce unlimited resonant frequencies. Normally, without any perturbation to the symmetry of the cavity, degenerate modes that have the same resonant frequency with orthogonal field distributions have no coupling and effect to each other. When some perturbations are added to the symmetry of the structure, the field distributions of degenerate modes will be no longer orthogonal and coupled to each other. In this condition, two coupled degenerate modes act like two coupled resonators, and a two-pole dual-mode microstrip filter can be achieved. This is the simplest dual-mode bandpass filter using a single dual-mode resonator [7].

Typically, according to basic microwave concept, in dual-mode microstrip resonators, with symmetrical structure, two orthogonal feedlines produce the two degenerate modes, and these degenerate modes can be coupled to each other by adding a perturbation at $\varphi = 45°$. The perturbation dimension ($a \times b$) controls the coupling between two degenerate modes and bandwidth of the filter [6]. The conventional model of dual-mode octagonal-loop bandpass filter is illustrated in Fig. 2, and its dimensions are shown in Table 2. This bandpass filter layout has a total dimension of 15.86 mm × 15.86 mm operating at 2.3 GHz, with fractional bandwidth of 5%. By using the feedlines structure and dimensions shown in Fig. 2 and Table 2, the optimum frequency response can be achieved.

After designing the resonator as the first step, the design procedure for the filter will be continued by designing the loop resonator in compressed form, feedlines structure, determining the tab position of the feedlines, and perturbation size (according to the desired bandwidth), respectively.

The proposed new filter in this paper is shown in Fig. 3, where it has three main differences compared to the previous design in Fig. 2 from [6]. First, the most important difference is that it includes one meander side which results in solely vertical symmetry. This structure leads to more compact size, because as mentioned before the resonance frequency is determined by mean circumference of the closed-loop resonator. Therefore, by decreasing the length of other

Table 2. Dimensions of "dual-mode filter" with reference to the layout depicted in Fig. 2.

Parameter	Dimension (mm)
a	5.95
b	0.48
W	0.6
G	0.1
W_1	0.2
W_2	0.6
L_f	6.23
L_t	3.22

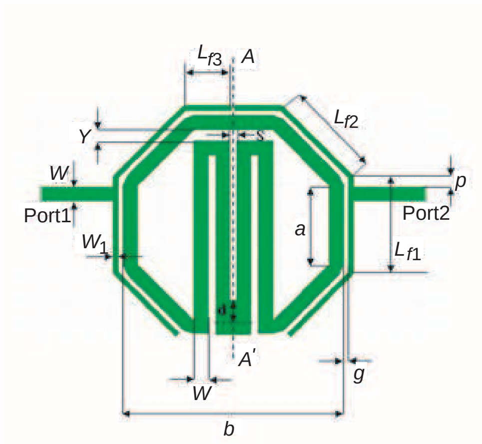

Figure 3. Compressed dual-mode microstrip bandpass filter.

sides and appending to one side length in meander form, the total mean circumference and consequently the resonant frequency will be static.

Secondly, the lack of horizontal symmetry itself resembles perturbation and causes the degenerate modes to be coupled to each other. But only one rectangular perturbation is needed to control the coupling between two degenerate modes by its dimension. As the filter layout is symmetrical only across the line $A - A'$, so the rectangular patch perturbation, $(s) \times (d)$ can be placed across this line as shown in Fig. 3. This perturbation affects the mean circumference of the dual-mode filter and consequently the resonance frequency will be shifted [14]. Therefore, by optimizing the length of the loop sides and using an EM simulator software [15], the exact center frequency can be achieved.

Finally, the third difference appears in the feedlines position. This filter design despite the conventional form is fed by a pair of feedlines along a straight line separated symmetrically. In some microwave circuits and networks, this kind of non orthogonal feedlines are more desirable and more useful. The tab position p of the feedlines, determines the position of the two transmission zeros and also the symmetry of the frequency response of the filter, although the passband responses are unchanged. Fig. 4 illustrates the effect of the tab position p on the frequency response of the filter. The filter in this design has almost symmetric frequency response for tab position of $p = 0.36$ mm.

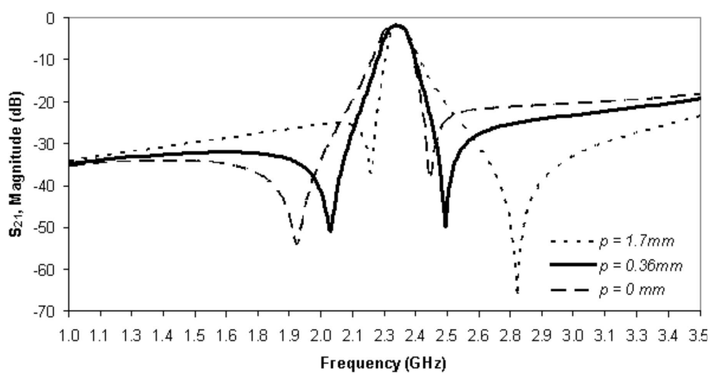

Figure 4. Variation of the filter frequency response for different tab position of feedlines.

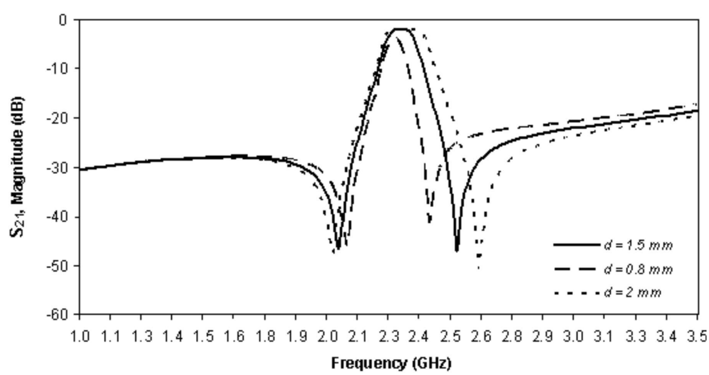

Figure 5. Bandwidth variation versus different perturbation size.

Figure 5 shows the variation of filter bandwidth against the perturbation size, d. According to the acceptable amounts for insertion loss and return loss, the valid range for perturbation size is $0.8\,\text{mm} < d < 2\,\text{mm}$.

The acceptable range for 3 dB bandwidth is $70\,\text{MHz} < \text{BW}_{3\,\text{dB}} < 120\,\text{MHz}$ corresponding to the valid range of perturbation size $0.8\,\text{mm} < d < 2\,\text{mm}$. By choosing the perturbation size $d = 1.5\,\text{mm}$ the fractional bandwidth (FBW) of 5% is achieved. Because the perturbation size affects the mean circumference of the resonator, the center frequency is shifted, so for different desired bandwidths, the resonator size should be optimized to get the exact center frequency.

Since a dual-mode resonator operates as two resonators, coupling coefficient exists between them. Coupling coefficient between resonators can be extracted by using the EM simulator [15] and by using the general formula described by [7]

$$k_{12} = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2} \tag{3}$$

where f_1 and f_2 are the resonant frequencies of mode 1 and mode 2, respectively. Since the mode splitting of the dual-mode changes according to the perturbation size, therefore the coupling coefficient is a function of perturbation size d. The variation of the coupling coefficient against the size of the perturbation element is illustrated in Fig. 6.

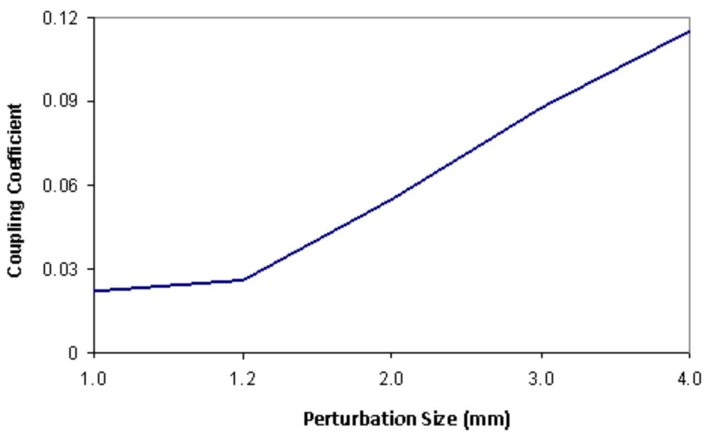

Figure 6. Simulated coupling coefficient versus the size of perturbation.

Table 3. Dimensions of proposed "dual-mode filter" with reference to the layout depicted in Fig. 3.

Parameter	Dimension (mm)
a	3.3
g	0.1
L_{f1}	3.6
L_{f2}	3.885
L_{f3}	1.82
W	0.6
W_1	0.2
s	0.3
d	1.5
p	0.365
Y	0.46
b	9.11

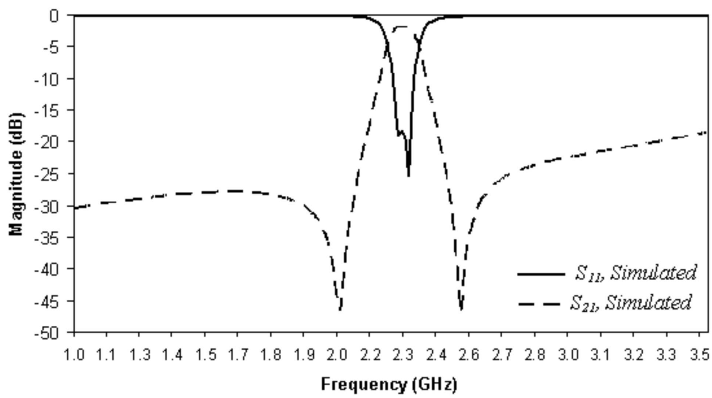

Figure 7. The simulation results for proposed dual-mode filter.

As can be seen from the Fig. 6 the coupling coefficient increases by the increment of the perturbation size. The optimum perturbation size in this design is $d = 1.5$ mm that results in the coupling coefficient of 0.043. The feedlines structure and dimensions used in this design and the selected sizes of Y, S, and G lead to have more compact size with optimum frequency response.

The dimensions of the filter illustrated in Fig. 3 are shown in Table 3, and Fig. 7 shows the simulation results of the proposed filter [15].

The simulation result shown in Fig. 7 exhibit that the specification for a narrowband filter at 2.3 GHz is obtained. The minimum insertion loss is 1.68 dB, and the return loss is better than 20 dB. The fractional bandwidth is about 5% (120 MHz). The two transmission zeros are located at 2.05 GHz, and 2.51 GHz, having sharp rejection of more than 45 dB. The total size of this compact bandpass filter is approximately 9.1 mm × 9.1 mm.

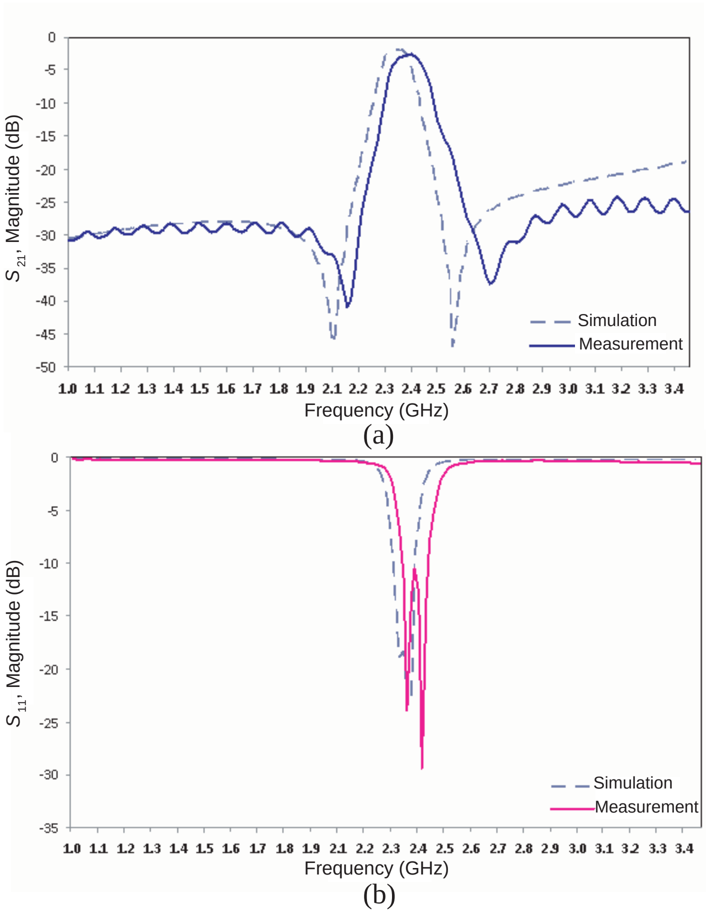

Figure 8. (a) The simulated and measured S_{21} frequency responses of the Filter. (b) The simulated and measured S_{11} frequency responses of the filter.

4. EXPERIMENTAL RESULTS

The filter is fabricated using standard photolithography process on Taconic CER-10 having relative permittivity constant of 10, 0.64 mm of substrate thickness and loss tangent, $\tan\delta = 0.0035$. The simulated and measurement results are presented in Fig. 8.

Figure 8 shows that the simulated and measured results are in good agreement. The measured fractional bandwidth is about 5%, the return loss is better than 25 dB and the minimum insertion loss of the filter is 2.35 dB. Conductor loss and connector mismatches are the main factors to contribute to the total loss. Of course by using a substrate material with lower loss tangent the insertion loss in the passband will be decreased. By using superconductors and designing with very narrow gap feedlines using micromachining technique can reduce the insertion loss to its minimum, but it should be mentioned that the size of the coupling gap between the feedlines and ring resonator affects not only the strength of coupling but also the resonant frequency. Furthermore, micromachining is costly and superconductors need the use of cryostats that makes the application limited and not robust. This is a reason why conventional photolithography fabrication is maintained in this paper.

Figure 9 shows the photograph of the BPF and its dimensions with respect to Fig. 3 and Table 3. As shown in Fig. 9, two 50 Ω SMA connectors are joined to 50 Ω feedlines to connect to the Vector Network Analyzer (VNA) for measurement in the best matching condition. The measurement was performed using a N5230A vector network analyzer.

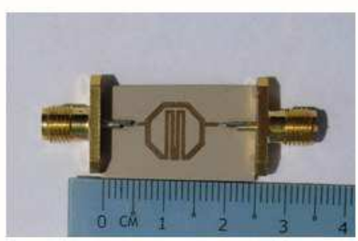

Figure 9. The top view of the fabricated dual-mode bandpass filter.

5. CONCLUSION

In this paper, a new type of compact narrow bandpass filter using dual-mode microstrip octagonal loop resonator is designed and tested. By adding the perturbation to the filter layout the coupling between the two degenerate modes can be controlled. By choosing the optimized perturbation size, $d = 1.5\,\text{mm}$, the desired fractional bandwidth and frequency response can be achieved. This new type of bandpass filter is fed by a pair of feedlines that are located along a straight line. The tab position of the feedlines controls the symmetry of the frequency response and the transmission zeros position. In this design the minimum measured insertion loss is 2.35 dB, and the return loss is better than 25 dB operating at 2.3 GHz with a fractional bandwidth of 5% with two transmission zeros on both sides of the passband. The total size of this layout is $9.1\,\text{mm} \times 9.1\,\text{mm}$. This new dual-mode filter leads to size reduction of 41%, 27.5%, 18%, with respect to the conventional octagonal design depicted in Fig. 2, conventional square ring resonator design [3] and meander square ring resonator [6], respectively, at the same center frequency. This configuration also in addition to its compact size has lower insertion loss and better return loss compared to the rectangular form reported in [14] at the same center frequency. This is due to its enhanced feedlines and lower curvature effects in its structure form. Finally, it should be noted that there can be found techniques which give more compact size as well as less insertion loss; among them is micromachining technique in building MEMS and SAW filters. However, this technique is considerably costly and complicated compared to standard photolithography technique introduced here. The MEMS filter designed for WiMax systems reported in [16], for instance, has a total size of $3.5\,\text{mm} \times 2\,\text{mm}$ and insertion loss less than 1 dB in the same frequency band. However, our new filter is designed to support WiMax application, and it offers compactness in size with good performance and sharp rejection. Not only that, it can be fabricated with simple but practical standard conventional photolithography technique which is cheaper compared to other fabrication techniques.

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