Progress In Electromagnetics Research B, Vol. 3, 131–141, 2008

NOVEL DESIGN OF DUAL-MODE BANDPASS FILTER USING RECTANGLE STRUCTURE

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Abstract—A compact dual-mode filter is proposed by using rectangle structure. The filter has the characteristics of compact structure, low insertion loss and so on. Several attenuation poles in the stopband are realized to improve the selectivity of the proposed bandpass filter. The experimented results were in good agreement with simulated results.

1. INTRODUCTION

DUAL-MODE microstrip bandpass filters have been investigated by many researchers for applications in both wired and wireless communication. Recently, the synthesis theory of microwave filters presenting two passbands mostly use frequency-variable transformations [1]. However, the strong attenuation is required for practical applications. Many new structures, such as stepped impedance resonators (SIRs) or parallel coupling [2, 3] or equal-length coupled-serial-shunted lines [4], have been proposed for a dual-mode bandpass filter. For dual-mode operation, a perturbation is introduced in the resonator in order to couple its two degenerate modes. Depending on the position and size of the perturbation, different filter responses can be obtained. Transmission zeros at finite frequencies can also be generated and controlled by the same mechanism. In [5], a dual-mode dual-band bandpass filter was initially reported. Unfortunately, this solution suffers from high insertion loss and none transmission zeros in the stopband. And an extra matching network is needed to combine them. Recently, the dual-mode resonator using patch [6] or square structure has attracted many attentions for its low insertion loss and compact structure in design of single band filter [11–29]. A dual-mode filter with stacked loop structure is proposed in [7]. However, the stacked loop structure may introduce higher cost and difficulties in fabrication.

In this paper, we introduce a new microstrip rectangle loop dual-mode filter. The filter, with lower insertion loss, provides the better transmission band.

2. RECTANGLE DUAL-MODE RESONATOR

As a kind of special filters based on a variety of symmetric dual-mode resonating structuresdual-mode filter can be equivalent to dual tunable resonator circuit in practical application [8, 9]. Therefore, the number of the patch can be decreased 50 percent and the size of circuit can also be reduced in the current application. Based on the characteristics above, the dual-mode filter can be used sharply in miniaturized communication system.

Some people such as Wolf have advanced many designs of microstrip dual-mode bandpass filters, owning the same characteristics, which introduce asymmetry feed-lines, slots or pins and so on for a perturbation in the resonator in order to couple its two degenerate modes, with tuning the correlative parameters of circuit so as to obtain the work condition of dual-mode resonator. In the conventional design of filters, circular and square patches have used widely. The degenerate modes of a square ring are coupled by a perturbation at one or more corners of the square. However, rectangle microstrip structure was used in the design of filter in this paper for its smaller size.

Different filter responses can be obtained with different positions and sizes of the perturbation, which is analyzed in detail in [6]. According to the analysis of resonant mode theory and slow-wave effect, the function of rectangle patch is equal to cut a part of the structure [6].

The fundamental resonance occurs when λ_g is the perimeter of the rectangle, where λ_g is the guided wavelength.

$$\lambda_g = \frac{c}{f\sqrt{\varepsilon_{eff}}} \tag{1}$$

where c is the velocity of light in free space, and ε_{eff} is the effective dielectric constant of the substrate. According to (1), while a resonant frequency is fixed, λ_g is decreased to realize size reduction as ε_{eff} increased. Similarly, for a fixed ε_{eff}, the resonant frequency f is decreased as the perimeter increased.

3. RECTANGLE DUAL-MODE BANDPASS FILTER

Figure 1 shows the proposed rectangle structure. So far, the study of microstrip rectangle structure filters has received little attention. Based on the former concerns on filtering characteristics, a novel dual-mode bandpass filter is presented. Given the lengths of side of the structure, the waveguide wavelength corresponding to the passband is

$$\lambda_{g1} = 2(a1 + b1) \tag{2}$$

where $a1$, $b1$ are the lengths of side of the rectangle.

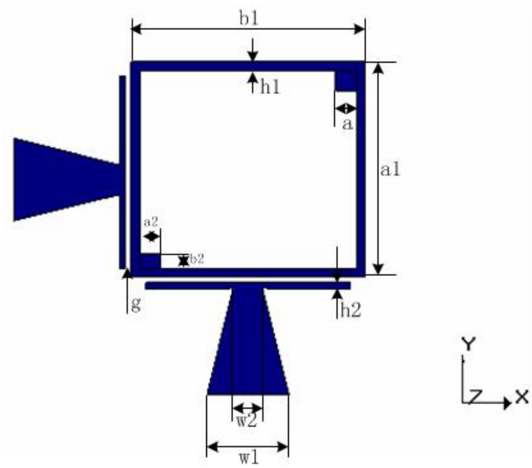

Figure 1. Configuration of the dual-mode bandpass filter.

The two different perturbations (the top one: a is the length of the side of the square; the bottom one: a_2, b_2 are the lengths of the side of the rectangle.) can change the field distribution and inspire the degenerate modes, meanwhile, the cross couplings between the degenerate modes including electric and magnetic couplings can not only generate attenuation poles but also cause the resonant frequency of higher harmonic wave to shift. In this case, it can be used for the miniaturization of the filter design.

By inserting an electric wall and a magnetic wall into the symmetry plane of the equivalent circuit, respectively, we can obtain

$$f_e = \frac{1}{2\pi\sqrt{(L - L'_m)(C - C'_m)}} \tag{3}$$

$$f_m = \frac{1}{2\pi\sqrt{(L + L'_m)(C + C'_m)}} \tag{4}$$

As can be known that both the magnetic and electric couplings have the same effect on the resonant frequency shifting. In other words, they reduce or enhance the stored flux or charge of the single resonant circuit at the same time when the electric wall or the magnetic wall is inserted.

From the Equations (3) and (4), we can obtain the mixed coupling coefficient k_B:

$$k_B = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} = \frac{CL'_m + LC'_m}{LC + L'_m C'_m} \tag{5}$$

When

$$L'_m C'_m \ll LC$$

$$k_B \approx \frac{L'_m}{L} + \frac{C'_m}{C} = k'_M + k'_E \tag{6}$$

Which clearly indicates that the mixed coupling is resulted from the superposition of the magnetic and electric couplings.

In this part we work with a fixed coupling topology as shown in Fig. 2.

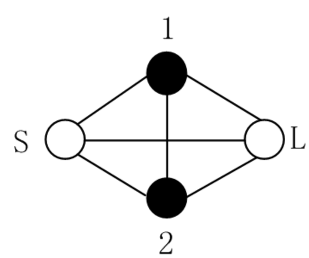

Figure 2. Topology of the proposed filter.

For any two-port lossless filter network composed of a series of N intercoupled resonators, the transfer and reflection functions may be expressed as a ratio of two Nth degree polynomials

$$S_{11}(\omega) = \frac{F_N(\omega)}{E_N(\omega)} \qquad S_{21}(\omega) = \frac{P_N(\omega)}{\varepsilon E_N(\omega)} \tag{7}$$

where ω is the real frequency variable related to the more familiar complex frequency variable s by $s = j\omega$. For a Chebyshev filtering function, ε is a constant normalizing S_{21} to the equiripple level at $\omega = \pm 1$ as follows:

$$\varepsilon = \frac{1}{\sqrt{10^{RL/10} - 1}} \cdot \frac{P_N(\omega)}{F_N(\omega)}, \quad \omega = 1 \tag{8}$$

when RL is the prescribed return loss level in decibels and it is assumed that all the polynomials have been normalized such that their highest degree coefficients are unity. $S_{11}(\omega)$ and $S_{21}(\omega)$ share a common denominator $E_N(\omega)$, and the polynomial $P_N(\omega)$ contains the transfer function transmission zeros.

Using the conservation of energy formula for a lossless network $S_{11}^2 + S_{21}^2 = 1$ and (3)

$$S_{21}^2(\omega) = \frac{1}{1 + \varepsilon^2 C_N^2(\omega)} = \frac{1}{(1 + j\varepsilon C_N(\omega))(1 - j\varepsilon C_N(\omega))} \tag{9}$$

where

$$C_N(\omega) = \frac{F_N(\omega)}{P_N(\omega)} \tag{10}$$

$C_N(\omega)$ is known as the filtering function of degree N and has a form for the general Chebyshev characteristic [6]

$$C_N(\omega) = \cosh\left[\sum_{n=1}^{N} \cosh^{-1}(x_n)\right] \tag{11}$$

where

$$x_n = \frac{\omega - 1/\omega_n}{1 - \omega/\omega_n} \tag{12}$$

and $j\omega_n = s_n$ is the position of the nth transmission zero in the complex s-plane.

The first step in the polynomial synthesis procedure is to replace the $\cosh^{-1}$ term in (11) with its identity

$$C_N(\omega) = \cosh\left[\sum_{n=1}^{N} \ln(a_n + b_n)\right] \tag{13}$$

where

$$a_n = x_n, \quad b_n = \left(x_n^2 - 1\right)^{1/2} \tag{14}$$

Then

$$C_N(\omega) = \frac{1}{2}\left[\exp\left(\sum \ln(a_n+b_n)\right) + \exp\left(-\sum \ln(a_n+b_n)\right)\right]$$
$$= \frac{1}{2}\left[\prod_{n=1}^{N}(a_n+b_n) + \frac{1}{\prod\limits_{n=1}^{N}(a_n+b_n)}\right] \tag{15}$$

Multiplying the second term in (11) (top and bottom) by $\prod_{n=1}^{N}(a_n-b_n)$ yields

$$C_N(\omega) = \frac{1}{2}\left[\prod_{n=1}^{N}(a_n+b_n) + \prod_{n=1}^{N}(a_n-b_n)\right] \tag{16}$$

Because

$$\prod_{n=1}^{N}(a_n+b_n).\prod_{n=1}^{N}(a_n-b_n) = \prod_{n=1}^{N}\left(a_n^2-b_n^2\right) \tag{17}$$

In the bottom line of the second term will always be unity. This is easily verified by substituting for a_n and b_n using (13).

Equation (9) may now be written in its final form by substituting for a_n, b_n and x_n using (11) and (13) as follows:

$$C_N(\omega) = \frac{1}{2}\left[\frac{\prod\limits_{n=1}^{N}(c_n+d_n) + \prod\limits_{n=1}^{N}(c_n-d_n)}{\prod\limits_{n=1}^{N}\left(1-\frac{\omega}{\omega_n}\right)}\right] \tag{18}$$

where

$$c_n = \omega - \frac{1}{\omega_n}, \quad d_n = \omega'\left(1-\frac{1}{\omega_n^2}\right)^{1/2}, \quad \omega' = \left(\omega^2-1\right)^{1/2} \tag{19}$$

a transformed frequency variable.

According to the analysis above, the normalized coupling matrix can be synthesized as:

$$[M] = \begin{bmatrix} 0.0000 & 0.0187 & -0.0103 & 0.0002 \\ 0.0187 & -0.0246 & -0.0160 & 0.0096 \\ -0.0103 & -0.0160 & 0.0246 & 0.0187 \\ 0.0002 & 0.0096 & 0.0187 & 0.0000 \end{bmatrix}$$

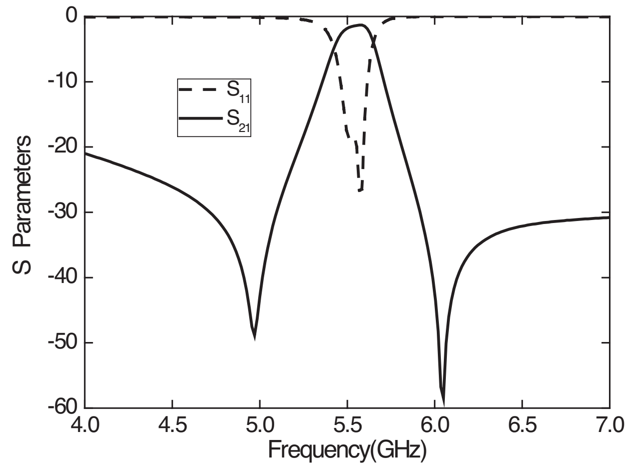

Figure 3. Simulation results of proposed filter.

As shown in Fig. 3, the simulated results of the proposed dual-mode filter are obtained using simulator IE3D V10 based on MOM. IE3D from Zeland Software Inc is used for this purpose as well as for the analysis and design of the filter. From which, low insertion loss, high selectivity, and good isolation characteristics can be observed clearly.

4. FABRICATE FILTER AND MEASURED RESULT

The proposed dual-mode bandpass filter is fabricated with design parameters as follows: $a = 0.7\,\mathrm{mm}$, $a1 = 7.3\,\mathrm{mm}$, $b1 = 8\,\mathrm{mm}$, $a2 = 0.7\,\mathrm{mm}$, $b2 = 0.5\,\mathrm{mm}$, $h1 = 0.3\,\mathrm{mm}$, $h2 = 0.2\,\mathrm{mm}$, $w2 = 1\,\mathrm{mm}$, $g = 0.2\,\mathrm{mm}$, and $w1 = 2.8\,\mathrm{mm}$ is the width of $50\,\Omega$ microstrip feed-line on a 1 mm thick dielectric substrate with a relation dielectric constant of 2.65. The photograph of the fabricated filter is shown in Fig. 4.

The filter is measured with Agilent 8719ES network analyzer. The results are shown in Fig. 5. Two attenuation poles in the stopband as follows: 4.94 GHz, 6.10 GHz. Return loss larger than 20 dB is achieved in the passband.

Compared with the simulation results, slightly shift to the higher frequency occurred between the simulation results and the measurement results for the machining tolerance. The insertion loss in passband is mainly due to the conductor loss of the rectangle resonators. Good agreement between the simulation and measurement is achieved.

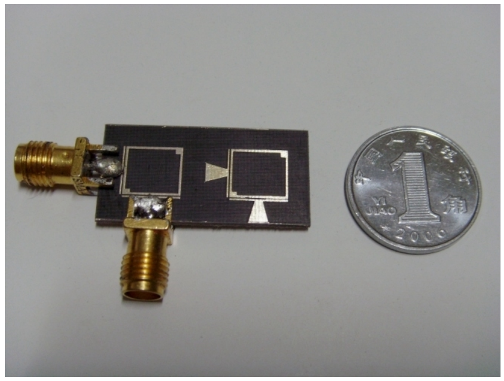

Figure 4. Photograph of the fabricated filter.

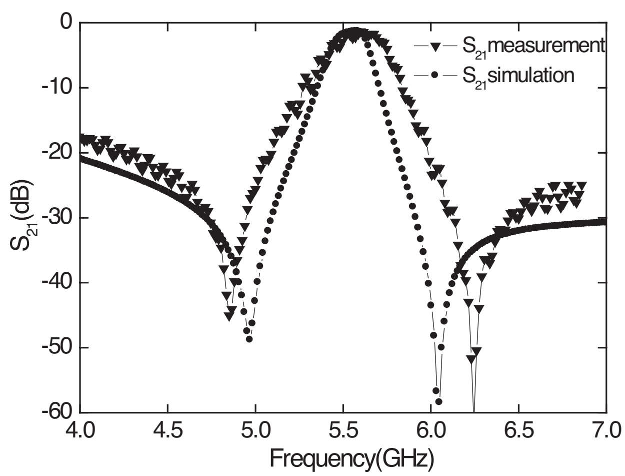

Figure 5. Comparison of the simulated and measured results.

5. CONCLUSION

In this paper, a dual-mode bandpass filter using rectangle structure is presented. Several attenuation poles in the stopband are realized. Numerical simulations using IE3D show the feasibility of the dual-mode bandpass filter. It has been shown that the proposed microstrip filter can provide good selection and better insertion loss. The measured results show a good agreement with the simulation.

ACKNOWLEDGMENT

This work is supported by the National Natural Science Foundation of China (NSFC) under Contract No. 60501023.

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