A compact coplanar waveguide dual-band bandpass filters based on defected ground structures

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Abstract A dual-band bandpass filter using coplanar waveguide (CPW) has been proposed in the paper. To generate the first passband, a gap is adopted in the central strip of the CPW main transmission line, and two pairs of spiral-shaped defected ground structures (DGSs) are added to the ground planes. Then, the loaded four L-shaped slots in the central strip of the CPW main transmission line can achieve the second passband. The measured central frequencies of the dual-band bandpass filter are 3.59 GHz and 5.87 GHz. The proposed filter has good dual-passband performance and good stopband suppression.

key words: dual-band bandpass filter, coplanar waveguide, defected ground structures

Classification: Microwave and millimeter-wave devices, circuits, and modules

1. Introduction

With the rapid development of wireless communication technology, more and more communication systems work in two frequency bands, so the demand for dual-band microwave systems has been increased dramatically, such as WiMAX and WLAN communication systems. Filter is one of the key components, so, not only the performance requirements of the filter are greatly improved, but also the demand for their dual-band frequency response characteristics is increased. Therefore, dual-band bandpass filters with good performances and compact sizes have great development prospects [1, 2, 3, 4, 5, 6, 7]. There are plenty of methods that can be adopted to realize the dual-band performance. For example, adopting the stepped-impedance resonator (SIR) to achieve the dual-band bandpass filters with the great performance [8, 9, 10, 11, 12, 13]. In [12], using SIRs and loaded tuneable capacitors to realize each band independently controllable to design a differential dual-band bandpass filter. Stub-loaded resonators (SLR) are introduced to reduce the size of the dual-band bandpass filters [14, 15, 16, 17]. In [16], SLRs are adopted to realize the dual-band differential-mode bandpass filter with a controllable resonant frequency. But the stopband performance of the filter needs to be improved.

In recent years, CPW technology has been widely studied due to the circuit design requirements of integration and miniaturization [18, 19, 20, 21, 22, 23, 24, 25, 26, 27]. In [20], a composite coupling structure with the CPW and microstrip circuits was used to achieve a dual-band bandpass filter, which adopts two Y-shaped resonators in the CPW transmission line, and two rectangular ring resonators with microstrip circuits are constructed on the other side of the dielectric substrate. In [21], using the circuit of conductor-backed coplanar waveguide (CBCPW), high selective dual-band bandpass filters are achieved. However, these two structures are designed on two surfaces of the substrate.

In this paper, a dual-band bandpass filter with controllable central frequencies using the CPW circuit is proposed. Each operating central frequency of two passbands can be controlled by adjusting the sizes of the DGSs and the lengths of L-shaped slots, respectively. The proposed dual-band method is different from the traditional method of cascading two different single-band filters with different operating frequency points, which would increase the circuit size. The proposed filter makes full use of the central strip and ground planes of the CPW main transmission line to achieve dual-band performance, and the sizes of the proposed filters are not increased, which meets the requirement of a compact design.

2. CPW bandpass filter with spiral-shaped DGSs

The structure of the CPW single-band bandpass filter is given in Fig. 1(a), which is composed of a central strip of the CPW main transmission line with a gap, and two ground planes with four spiral-shaped DGSs. The spiral-shaped DGSs and the gap could generate a passband and have stopband rejection characteristics.

The whole equivalent circuit of the CPW single-band bandpass filter is given in Fig. 1(b), where series inductor \( L_1 \) and shunt capacitors \( C_1 \) are the equivalent parameters of the CPW transmission line. The gap works as the series capacitor \( C_0 \). Each spiral-shaped DGS introduces additional inductance and capacitance for the CPW main transmission line, which can be represented by a capacitor \( C_s \) and an inductor \( L_s \). The equivalent series capacitor \( C_s \) and the series
inductor \( L_s \) of the spiral-shaped DGS can be expressed

\[
C_s = \frac{\omega_s}{Z_0 g_1 \left( \omega^2_{01} - \omega^2_0 \right)} = \frac{f_s}{4\pi Z_0 \left( f_{01}^2 - f_s^2 \right)}
\]

(1)

\[
L_s = \frac{1}{\omega^2_0 C_s} = \frac{1}{4\pi^2 f_{01}^2 C_s}
\]

(2)

A detailed analysis of the spiral-shaped DGS has been given in [28]. The dimensions of the single-band bandpass filter as denoted in Fig. 1(a) are \( w_0=2.2 \) mm, \( w_1=9.0 \) mm, \( d_0=0.4 \) mm, \( g=0.2 \) mm, \( a=3.1 \) mm, \( b_1=0.3 \) mm, \( b_2=0.3 \) mm, \( d=1.8 \) mm. The frequency response of the single-band filter is plotted in Fig. 1(c), where the filter centers at 3.5 GHz, and the 10-dB fractional bandwidth is 11.4%, from 3.23 GHz to 3.63 GHz, and the insertion is less than 1.30 dB.

![Diagram](image1)

**Fig. 1.** The proposed CPW single-band filter (a) Layout (b) Equivalent circuit (c) S-parameters

### 3. CPW dual-band bandpass filter

#### 3.1 The resonator with L-shaped slots

Fig. 2(a) shows the L-shaped slots [29], which is composed of a CPW main transmission line and two L-shaped slots in the central strip, where the metal part and the etched part are represented by gray and white colour, respectively. The equivalent transmission line model and its simplified model of the L-shaped slots that ignores the inner couplings are given in Fig. 2(b), and Fig. 2(c).

![Diagram](image2)

**Fig. 2.** L-shaped slots (a) Layout (b) Equivalent transmission line model (c) Simplified equivalent transmission line model

For each transmission line of the L-shaped slots, \( Z_a, Z_b \) are the characteristic impedance, and \( \theta_a, \theta_b \) are the phase shift. According to Fig. 2(c), the transition matrix is given by

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = M_1 \times M_2
\]

(3)

\[
M_1 = \begin{bmatrix}
1 & 0 \\
-f 2 Z_a \cot \theta_a & 1
\end{bmatrix}
\]

(4)

\[
M_2 = \begin{bmatrix}
\cos \theta_b & jZ_b \sin \theta_b \\
jY_b \sin \theta_b & \cos \theta_b
\end{bmatrix}
\]

(5)

Thus, the parameters of the transition matrix can be calculated as

\[
A = \cos \theta_b
\]

(6)

\[
B = jZ_b \sin \theta_b
\]

(7)

\[
C = jZ_b \cos \theta_b + j2Z_b \sin \theta_b \cot \theta_a \]

(8)
D = \frac{2Z_u \cos \theta_i \cot \theta_a - Z_\theta \sin \theta_b}{2Z_u \cot \theta_a} \quad (9)

And \( \theta_i (i = a, b) \) can be obtained by

\[ \theta_i = \frac{\omega_0 l_i \sqrt{\varepsilon_{re}}}{c} (i = a, b) \quad (10) \]

where \( l_i \) (i=a, b) are the physical length of each line, \( \omega_0 \) is the resonant angular frequency of the L-shaped slots resonator, and \( c \) is the velocity of light in free space. \( \varepsilon_{re} \) is the effective permittivity, and \( \varepsilon_{re} \) is satisfied with [30]

\[ \varepsilon_{re} = \frac{\varepsilon_r + 1}{2} \quad (11) \]

By using the transition matrix parameter, the transmission coefficient \( S_{21} \) of the L-shaped resonator can be expressed as [31]

\[ S_{21} = \frac{2}{A + B + CZ_a + D} \quad (12) \]

When the value of \(|S_{21}|\) equals zero, the frequency point at transmission zero of the L-shaped slots resonator can be given as

\[ f_e = \frac{c(2n + 1)}{4l_i \sqrt{\varepsilon_{re}}}, n = 0,1,2,3,... \quad (13) \]

When the L-shaped resonator works at the resonant frequency point, in the ideal condition the input signal of the resonator should be completely transmitted to the output port, so \(|S_{21}|\) should be equals to 1. Then, from the equation (12), the characteristic impedance and phase shift of each transmission line of the L-shaped resonator should be satisfied

\[ \left( \frac{P}{Q} \right)^2 + \left( \frac{M}{N} \right)^2 = 4 \quad (14) \]

\[ P = 4Z_u \cos \theta b \cot \theta_a - Z_\theta \sin \theta_b \quad (15) \]

\[ Q = 2Z_u \cot \theta_a \quad (16) \]

\[ M = 2Z_u Z_\theta^2 \sin \theta b \cot \theta_a + Z_\theta^2 Z_\theta \cos \theta b + 2Z_\theta^2 Z_u \sin \theta b \cot \theta_a \quad (17) \]

\[ N = 2Z_u Z_\theta \cot \theta_a \quad (18) \]

The central frequency for the second passband is 5.8 GHz, and the substrate with 1 mm thickness and relative permittivity \( \varepsilon_r = 2.65 \) has been adopted. For impedance matching, two ports of the filter are designated to 50 \( \Omega \), so \( w_0 = 2.2 \text{ mm, } g = 0.2 \text{ mm, as shown in Fig. 2(a). The L-shaped slots are designed to embed into the central strip of the CPW main transmission line to conveniently integrate with the first single-band filter with spiral-shaped DGSs to achieve dual-band performance. It can be seen from Fig. 2(a) that, } 2w_a + w_b + 2w_1 = w_a, \text{ and } w_2 = 2.2 \text{ mm. Considering the limitation of fabrication accuracy, } w_a, w_b, w_f \geq 0.2 \text{ mm, so, } 0.2 \text{ mm} \leq w_b \leq 1.4 \text{ mm, then we can obtain that } Z_\theta \leq 104 \Omega, 62 \Omega \leq Z_u \leq 104 \Omega. \text{ To minimize the circuit size, the values of } \theta a \text{ and } \theta b \text{ should be less than 0.5\pi. According to these analyses, } Z_\theta = 100 \text{ ohm, } Z_u = 67 \text{ ohm, } \theta a = 0.40\pi, \theta b = 0.43\pi \text{ are a set of specific solutions satisfying the equations (14)-(18). From } Z_\theta = 100 \text{ ohm, } Z_u = 67 \text{ ohm, we can obtain } w_a = 0.3 \text{ mm, } w_b = 1.0 \text{ mm. According to equation (10), it can be calculated out that } l_i = 7.7 \text{ mm, } l_2 = 8.0 \text{ mm. According to the equation (13), the transmission zero is calculated to 7.21 GHz.}

The simulated s-parameters of the CPW L-shaped slots resonator is shown in Fig. 3. It can be seen from Fig. 3 that adopted L-slots in the central strip of the CPW main transmission line can generate resonance. The transmission zero frequency is located at 7.23 GHz, and the resonant frequency is located at 5.77 GHz. The simulations agree with the above analysis. Moreover, the CPW L-shaped slots resonator can achieve another passband at an upper frequency for designing the dual-band filter.

![Fig. 3. Simulated s-parameters of CPW L-shaped slots resonator](image-url)
Considering the passband performance and stopband suppression, the central frequencies and the transmission-zero frequencies of the proposed filter can be affected by several key dimensions including the length and width $a$ of spiral-shaped DGS, the length of L-shaped slots $l_a$, the width $d_0$ of the gap, and the parameters are analyzed as follows. The central frequency of the first passband can be mainly controlled by the length and width $a$ of spiral-shaped DGS. Fig. 5 plots the s-parameters of the proposed filter with different spiral-shaped DGS sizes. The central frequency decreases with the increase of length and width $a$. Moreover, the central frequency of the second passband remains unchanged essentially. Therefore, when $a$ is chosen to 3.1 mm, the first passband centers at 3.5 GHz, which is satisfied with WiMAX communication frequency band requirements.

The performance of the second passband can be mainly controlled by the L-shaped slots length $l_a$. Fig. 6 gives the simulations of the proposed filter with different lengths $l_a$ of the L-shaped slots, where the central frequency of the second passband is decreased when the lengths $l_a$ is increased, while the central frequency of the first passband remains 3.5 GHz. Meanwhile, with the increased length $l_a$, both of the first and second transmission zero frequencies remain 2.79 GHz and 4.57 GHz, respectively. So, when $l_a=7.7$ mm, the central frequency of the second passband is close to 5.8 GHz, which is satisfied with WLAN communication frequency band requirements.

The width of the gap $d_0$ can also control the frequencies of the transmission zeros and the central frequencies of the proposed filter. Fig. 7 illustrates the s-parameters of the CPW dual-band bandpass filter. As the gap width $d_0$ is increased, the central frequency of the first passband remains the same. However, when the gap width $d_0$ is increased from 0.4 mm to 0.8 mm, both central frequencies of the second passband and the third transmission zero frequency is increased to a higher frequency. When the gap width $d_0$ is increased to 1.0 mm, the central frequency for the second passband is increased to a higher frequency. To make the proposed filter have a good passband performance, it can be chosen that $d_0$ is 0.6 mm. After the above analysis, the final optimized sizes of the bandpass filter are recorded in Table. I as the notations marked in Fig. 4.

### Table I. Optimized dimensions of the proposed filter (unit: mm).

| Parameters | Value | Parameters | Value |
|------------|-------|------------|-------|
| $a$        | 3.1   | $l_a$      | 8.0   |
| $b_1$      | 0.3   | $l_b$      | 1.1   |
| $b_2$      | 0.3   | $w_h$      | 0.3   |
| $w_1$      | 9.0   | $w_b$      | 1.0   |
| $g$        | 0.2   | $w_d$      | 0.3   |
| $d$        | 0.8   | $l_d$      | 2.2   |
| $l_a$      | 7.7   | $l_1$      | 3.6   |

### 3.3 Fabrication and measured results

All of the simulations are operated by an electromagnetic simulation software of HFSS. The CPW dual-band bandpass filter is fabricated on an F4B substrate with a thickness of 1mm and relative permittivity $\varepsilon_r=2.65$, loss tangent $\delta=0.005$, and the size is 26.0 mm $\times$ 20.6 mm. The photograph of the proposed filter is shown in Fig. 8. The filter is measured by Agilent 8510C network analyzer. Fig. 9 shows the simulated and the measured s-parameter for the dual-band bandpass filter. The simulated central frequencies are 3.50 GHz and 5.85 GHz, while measured central frequencies are 3.59 GHz and 5.87 GHz, with the 10-dB fractional bandwidths of 5.8% and 5.4%. For the measurement results, in the lower passband of 3.52 GHz-3.73 GHz, the return loss is better than 10 dB, and the insertion loss is 1.62 dB at 3.59 GHz.
the upper passband of 5.75 GHz-6.07 GHz, the return loss is better than 10 dB, and the insertion loss is 1.69 dB at 5.87 GHz. There are three transmission zeros outside the passbands, and $|S_{21}|$ are -28.9 dB, -42.1 dB, and -37.0 dB at frequencies of 2.99 GHz, 4.50 GHz, and 6.85 GHz, respectively. Moreover, the stopband rejection between the two passbands is more than 20 dB from 4.34 GHz to 4.77 GHz. The stopband rejections are more than 20 dB in the ranges from 1.0 GHz to 3.17 GHz and from 6.31 GHz to 8.0 GHz. The measurements agree with the simulation results for a slight frequency shift, which demonstrates the validity of designing the proposed dual-band bandpass filter.

![Fig. 8. Photograph of the fabricated dual-band bandpass filter](image)

The performance comparisons between the proposed bandpass filter and some other CPW bandpass filters listed in Table II, where RL and IL represent the return loss and insertions loss, respectively. FBW$_{10\,\text{dB}}$ represents the 10-dB fractional bandwidth at each central frequency, $f_{t2}$ is the transmission-zero frequency, and $\lambda_g$ represents the guided wavelength at the first central frequency. The comparisons in Table II shows that the proposed CPW bandpass filter has a good performance with a compact size.

![Fig. 9. Simulated and measured s-parameters of the proposed bandpass filter](image)

### Table II. Performance comparisons of bandpass filters

| Ref | $f_1$/$f_2$ (GHz) | at $f_1$/$f_2$ | FBW$_{10\,\text{dB}}$ | $f_{t2}$ (GHz) | Size |
|-----|-------------------|---------------|-----------------|---------------|------|
|     | RL (dB) | IL (dB) |                 |               |      |
| [20] | 2.45/3.55 | 13/35 | 1.58/1.22 | 11.8% | 4.2%  | 0.61$\lambda_g$ | 0.38$\lambda_g$ |
| [21] | 2.4/3.8 | 33/12 | 2.0/1.7 | 5.3% | 11.6 | 1.92/2.9/3.04/4.4 | 0.49$\lambda_g$ | 0.30$\lambda_g$ |
| [22] | 1.8/3.2 | 21/20 | 1.2/1.5 | 5.4% | 5.3% | 2.58/3.72 | 0.75$\lambda_g$ | 0.59$\lambda_g$ |

### 5. Conclusions

In this paper, a bandpass filter using spiral-shaped DGSs and L-shaped slots with the CPW circuit are presented. Two pairs of spiral-shaped DGSs and a gap can generate the first passband, and four L-shaped slots are introduced to generate the second passband, then a dual-band bandpass filter is achieved, which is applied in WiMAX and WLAN system. The two passbands can be individually controlled by adjusting the sizes of DGSs and the lengths of L-shaped slots, respectively. There are several transmission zeros for the dual-band bandpass filter that can improve the stopband rejection. The proposed bandpass filter has a compact structure, good stopband suppression, easy to design, and are simple to be manufactured, and convenient to connect with other components.

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