Dual-bandpass 3-D FSS with close band spacing based on multiple square coaxial waveguides

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Abstract A three-dimensional (3-D) frequency selective surface (FSS) based on multiple square coaxial waveguides (SCWs) is proposed, which realizes a dual-bandpass response with close band spacing. In the unit cell of the proposed 3-D FSS, two concentric SCW propagation paths and one parallel-plate waveguide (PPW) propagation path are constructed by utilizing three square metallic tubes. Each SCW propagation path intrinsically generates one transmission pole by the square slot resonance, and the PPW propagation path produces another transmission pole by the half-wavelength resonance. In addition, two transmission zeros are introduced due to the counteraction of out-of-phase signals between different paths, improving the frequency selectivity. After properly adjusting the design parameters, two desired passbands are obtained around 4.4 and 4.845 GHz, and the band ratio is only 1.1. Through the electric-field distributions, equivalent circuit model and parameters study, the operating principle of the proposed 3-D FSS is investigated. A prototype of the proposed 3-D FSS is fabricated and measured. The measured results show that the proposed design can achieve a stable response under the incident angles up to 45° for both TE and TM polarizations.

Keywords: dual-bandpass, frequency selective surface (FSS), close band spacing, dual-polarized, square coaxial waveguides (SCWs)

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

1. Introduction

Frequency selective surfaces (FSSs) have been extensively investigated in the past decades [1]. They are used in the design of hybrid radomes, antenna sub-reflectors, absorbers, and polarizers [2, 3, 4, 5]. With the rapid development of multi-functional devices in the satellite communication system, the design of the FSSs with multi-band performances has attracted much attention [6, 7, 8, 9, 10, 11]. Furthermore, when two channels are closely located in the frequency domain, dual-bandpass FSSs with close band spacing are highly desirable for some applications.

In order to meet the above demands, some dual-bandpass 2-D FSSs were realized by using slots [12, 13], square loops [11, 14, 15], lumped elements [16, 17], convoluted designs [18, 19], or complementary structures [20, 21]. However, these FSSs had limitations of flatness of the passbands, because only one transmission pole was achieved in each passband. To address this issue, a 2-D FSS with two flat passbands was presented by using circular aperture-coupled patches [22], but no transmission zeros were introduced, leading to an unsatisfactory frequency selectivity. Then, multiple transmission zeros were designed to improve the selectivity by virtue of the electromagnetic coupling [23, 24]. Nevertheless, these cascaded 2-D FSSs usually had a relatively large electrical size, and their angular stability performances were poor. By using a via-based 2.5-D structure, the miniaturization of a closely-spaced dual-band FSS was realized [25]. Furthermore, an alternative approach was provided by 3-D structures with multi-mode resonant cavities [4, 9, 26, 27, 28, 29]. As described in [28], a new type of dual-bandpass 3-D FSS with arbitrary band ratios was investigated. Unfortunately, these two designs [25, 28] only operated under single polarization. Additionally, the band ratios of most of the above-mentioned FSSs are large, which cannot satisfy the requirements of the practical applications. Therefore, it still remains a problem for the existing designs to realize small band ratio, high selectivity, good angular stability, and dual polarizations simultaneously.

In this paper, a dual-bandpass 3-D FSS with close band spacing is proposed based on multiple square coaxial waveguides (SCWs). Being different from the unit cell design in [29], an additional SCW propagation path is added for constructing more resonant modes, and then the inner conductor is hollowed to reduce the weight of the original FSS. Furthermore, to facilitate printed circuit board processing, all metallic conductors are replaced by the thin metallic copper foil. Hence, the unit cell of the proposed 3-D FSS is achieved. This unit cell contains two concentric SCW propagation paths and one parallel-plate waveguide (PPW) propagation path. Definitively, each SCW path can provide a square slot resonance, and each PPW path can provide a half-wavelength resonance. Thus, three transmission poles are obtained by these three paths. Owing to the counteraction of out-of-phase signals between different paths, two transmission zeros are introduced for high selectivity. Ultimately, a dual-bandpass response with small band ratio is achieved. The measured results of an implemented FSS show that a stable response under the incident angles up to 45° for both TE and TM polarizations is realized.

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2. Structure and operating principle

2.1 Description of the proposed structure

Fig. 1(a) demonstrates the perspective view of the proposed 3-D FSS. The unit cell consists of two concentric SCW propagation paths and one PPW propagation path. These two SCW propagation paths, namely, path 1 and path 2, are constructed by three square metallic tubes combined with dielectric 1 and dielectric 2. The PPW propagation path is formed by the dielectric 3 layer sandwiched by the two adjacent outer tubes, which is defined as path 3. The relative dielectric constants of the dielectric 1, dielectric 2 and dielectric 3 are \( \varepsilon_1 \), \( \varepsilon_2 \) and \( \varepsilon_3 \), respectively. The thickness of the proposed FSS is denoted as \( t \). In addition, the detailed geometry of the unit cell is shown in Fig. 1(b) and (c). The period of the unit cell is represented by \( p \). The side lengths of these three tubes are \( a \), \( b \) and \( c \), respectively. The heights of these tubes are equal to the thickness of the proposed FSS. Note that the inner tube becomes a square waveguide filling with air, which will not influence the frequency response of the proposed 3-D FSS, because the operating band of the FSS is below the cut-off frequency \( (f \approx 16.13 \text{ GHz}) \) of this waveguide.

![Fig. 1. Structure of the proposed 3-D FSS: (a) Perspective view; (b) Top view of the unit cell; (c) Side view of the unit cell.](image)

Fig. 2 shows the simulated transmission and reflection coefficients of the proposed 3-D FSS by using of commercial ANSYS HFSS simulation software, with the design parameters \( (p, a, b, c, t) = (14.6, 13, 11, 9.3, 9.3) \text{ mm} \). The relative dielectric constant \( \varepsilon_1 = 2.9 \), \( \varepsilon_2 = 2.2 \) and \( \varepsilon_3 = 9.8 \). It is observed that a dual-bandpass response is obtained around 4.4 and 4.845 GHz, and the band ratio is only 1.1. In the lower band, one transmission pole is realized at \( f_{p1} \) (4.42 GHz). In the higher band, two transmission poles are achieved at \( f_{p2} \) (4.8 GHz) and \( f_{p3} \) (4.945 GHz). Besides that, two transmission zeros at \( f_{z1} \) (4.5 GHz) and \( f_{z2} \) (5.31 GHz), are located at the upper sides of each passband. The 3-dB bandwidths of the lower and higher band are 110 MHz (4.345–4.455 GHz) and 470 MHz (4.61–5.08 GHz), and the fractional bandwidths are 2.5 and 9.7%, respectively.

2.2 E-field analysis of the transmission zeros/poles

Fig. 3 illustrates the electric-field (E-field) distributions at three transmission-pole frequencies. As shown in Fig. 3(a), when the electromagnetic waves strike upon the FSS structure, the path 2 is mainly excited. It is seen that the E-field vectors at \( f_{p1} \) are uniformly distributed in path 2 with the same magnitude and direction. It is also observed from the top view that the E-field vectors at \( f_{p1} \) are basically concentrated in the square slot at interface of the path 2. Therefore, the pole \( f_{p1} \) is produced by the square slot resonance in path 2, and its resonant wavelength is approximately equal to the circumference of this slot. In Fig. 3(b), it is found that the E-field vectors at \( f_{p2} \) have a similar distribution to that at \( f_{p1} \), which reveals \( f_{p2} \) is generated by the square slot resonance in path 1. The circumference of the square slot in path 1 is less than that in path 2, therefore, the resonant frequency \( f_{p2} \) is higher than \( f_{p1} \). Fig. 3(c) shows the E-field vectors at \( f_{p3} \) mainly exist along the path 3 which supports the propagation of TEM waves. Meanwhile, the E-field vectors have the same magnitude but opposite directions in the upper half and the bottom half of the path 3. It indicates that the path 3 resonates at a half-wavelength, and corresponding resonant wavelength is equal to \( 2t \).

![Fig. 2. Simulated transmission and reflection coefficients of the proposed 3-D FSS.](image)

Fig. 4 shows the E-field distributions at two transmission-zero frequencies. As depicted in Fig. 4(a), the E-field vectors at \( f_{z1} \) are distributed in path 1 and path 2. Besides, it is also observed that the E-field vectors of both path 1 and path 2 have opposite directions at the output ports, where the E-field vectors are combined out of phase, resulting in one transmission zero at \( f_{z1} \). In Fig. 4(b), three propagation paths are excited at the same time. The E-field vectors at \( f_{z2} \) have the same direction in path 1 and path 2, while the E-field vectors in path 3 have opposite direction at the output ports, leading to the other transmission zero at \( f_{z2} \) due to the same operating principle of the transmission zero at \( f_{z1} \).

2.3 Equivalent circuit model

In order to provide direct physical insight into the operating principle of the proposed 3-D FSS, an ECM for normal
incidence is established based on the theory in [29], as demonstrated in Fig. 5.

The ECM shown in Fig. 5 can be divided into three series subnetworks: path 1, path 2, and path 3 subnetworks. Each subnetwork consists of the discontinuities between the propagation paths and free space, and the transmission line in the middle. As for the transmission lines, in path 1, it can be represented by the rectangular waveguide transmission line \((Z_1, \theta_1)\) under TE_{10} mode, with the short-side \(w_1 = 0.5(b - c)\) and the long-side \(l_1 = 2c + \sqrt{2}w_1\), as marked in Fig. 1(b). Similarly, in path 2, it can be equivalent to another rectangular waveguide transmission line \((Z_2, \theta_2)\) with \(w_2 = 0.5(a - b)\) and \(l_2 = 2b + \sqrt{2}w_2\). Meanwhile, in path 3, it can be considered as a PPW transmission line \((Z_3, \theta_3)\) under TEM mode with width of \(a\). The characteristic impedance \((Z_1, Z_2\) and \(Z_3\) \) and electrical length \((\theta_1, \theta_2 \) and \(\theta_3)\) of these three transmission lines can be obtained by the formulas in [29].

With regard to the discontinuities, they are modeled as the LC resonators consisting of \(L_1, L_2, \) and \(C_{l1}\) in path 1, and \(L_3, L_{l2}, \) and \(C_{l2}\) in path 2. The inductors \(L_1\) and \(L_3\) denote the self-inductance of the square loops on the terminals of the inner and outer tube, respectively. The inductors \(L_2\) and \(L_{l2}\) represent the equivalent inductances of the square loop on the terminals of the middle tube for the path 1 and path 2. The capacitors \(C_{l1}\) and \(C_{l2}\) are gap capacitances. The discontinuity of the path 3 is represented by the gap capacitance \(C_p\). The initial values of these electrical parameters are calculated by using the given formulas in [30], and the final parameter values of the ECM are obtained by curve-fitting.

Then, the transfer matrices of these three subnetworks can be expressed, respectively, as

\[
\begin{bmatrix}
A_1 & B_1 \\
C_1 & D_1
\end{bmatrix} = \begin{bmatrix}
\frac{1}{j\omega L_1} & -\frac{1}{j\omega C_1} \\
\frac{1}{j\omega L_1} & -\frac{1}{j\omega C_1}
\end{bmatrix}
\begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix}
\]

(1)

\[
\begin{bmatrix}
A_2 & B_2 \\
C_2 & D_2
\end{bmatrix} = \begin{bmatrix}
\frac{1}{j\omega L_2} & -\frac{1}{j\omega C_2} \\
\frac{1}{j\omega L_2} & -\frac{1}{j\omega C_2}
\end{bmatrix}
\begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix}
\]

(2)

\[
\begin{bmatrix}
A_3 & B_3 \\
C_3 & D_3
\end{bmatrix} = \begin{bmatrix}
\frac{1}{j\omega C_p} & 0 \\
\frac{j\omega C_p}{1}
\end{bmatrix}
\begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix}
\]

(3)

By connecting these three subnetworks in series, the impedance parameters of the ECM can be written as follows:

\[
Z_{11} = Z_{22} = \frac{A_1}{C_1} + \frac{A_2}{C_2} + \frac{A_3}{C_3}
\] (4)

\[
Z_{12} = Z_{21} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}
\] (5)

Consequently, the scattering matrix parameters can be obtained by using the following equations:

\[
S_{11} = S_{22} = \frac{(Z_{11} - Z_0) \cdot (Z_{22} + Z_0) - Z_{12}Z_{21}}{(Z_{11} + Z_0)^2 - Z_{12}^2}
\] (6)

\[
S_{21} = S_{12} = \frac{2Z_{12}Z_0}{(Z_{11} + Z_0)^2 - Z_{12}^2}
\] (7)

where \(Z_0\) is the characteristic impedance of transmission lines modeling the free space at both sides of the proposed 3-D FSS, and \(Z_0 = 377\ \Omega\).
Based on the ECM shown in Fig. 5, the effects on the resonant characteristics of the proposed 3-D FSS will be discussed for different design parameters. Fig. 7(a) shows the variation of the transmission-zero/pole frequencies with respect to the thickness $t$. It is seen that, $f_{p3}$ and $f_{z3}$ decrease significantly as the thickness $t$ increases. The reason is that an increasing $t$ can enlarge $\theta_t$. Because the thickness $t$ is irrelevant to the dimension of the square slots in path 1 and path 2, $f_{p1}$, $f_{p2}$ and $f_{z1}$ are unchanged. As seen in Fig. 7(b), basically, the variation of the period $p$ does not affect $f_{p1}$, $f_{p2}$ and $f_{z1}$, while $f_{p3}$ moves toward lower frequency and $f_{z2}$ moves toward higher frequency with an increasing $p$. This is because the gap capacitance $C_p$ decreases and characteristic impedance $Z_3$ increases as the period $p$ increases. In Fig. 7(c), an increasing $a$ leads to the enlargement of resonant wavelength of the square slot in path 2, resulting in a decrease of $f_{p1}$. Meanwhile, the decrease of $f_{z2}$ is on account of the decrease of $Z_2$ and $Z_3$. However, the variation of $a$ has no influence on the electrical length $\theta_l$ and the resonant wavelength of the square slot in path 1, so $f_{z1}$, $f_{p2}$ and $f_{p3}$ are unchanged. It can be observed from Fig. 7(d) that $f_{p1}$, $f_{p2}$, and $f_{z1}$ shift to the lower frequencies. It is because the resonant wavelengths of the square slots in path 1 and path 2 increase as $b$ becomes larger. $f_{p3}$ remains the same because $b$ does not affect $\theta_l$. Moreover, $f_{z2}$ changes slightly due to the change of $Z_l$ and $Z_3$. As depicted in Fig. 7(e), $f_{p2}$ changes greatly with the variation of side length $c$. It can be explained that the resonant wavelength of the square slot in path 1 increases when $c$ increases. At the meantime, $f_{z1}$ and $f_{z2}$ are changed because of the variation of $Z_1$. As $c$ increases, the $\theta_l$ and the resonant wavelength of square slot in path 2 have no change, thus, $f_{p1}$ and $f_{p3}$ are unchanged.

### 3. Implementation and measurement

A 3-D FSS prototype is fabricated to validate the proposed design. This implemented example is built up by five kinds of the building parts, as shown in Fig. 8(a). The building part A is one piece of long board made of TP-2 composite material ($\varepsilon_{r2} = 9.8$, $\tan\delta = 0.002$) with a thickness of 1.6 mm, in which 15 opening slots cut half way along the board are periodically created. The building part B1 and B2 are two types of short boards, which are made of F4B material ($\varepsilon_{r2} = 2.2$, $\tan\delta = 0.001$) with a thickness of 1.0 mm. The building part C1 and C2, similar to the part B1 and B2, are also short boards made of F4B material ($\varepsilon_{t1} = 2.9$, $\tan\delta = 0.001$) with a thickness of 0.85 mm. The assembly process illustrated in Fig. 8(b) basically contains three steps: 1) The building part A pieces are cross-jointed together through the opening slots to form a frame, which supports the path 3. 2) The building part B1 and B2 pieces are inserted into the frame one by one, for creating the path 2. 3) The same operation is performed for the building part C1 and C2 to construct the path 1. Besides, the junctions of the printed metal on each part are covered by conductive silver pulp for good contact. Ultimately, the implemented 3-D FSS is approximately 226 mm $\times$ 226 mm in size, and consists of 14 $\times$ 14 (196) unit cells, as shown in Fig. 8(c). The dimensions of the unit cell $p \times p \times t$ is $0.214\lambda_0 \times 0.214\lambda_0 \times 0.136\lambda_0$, where $\lambda_0$ is the free-space wavelength at the center frequency of the lower band.

Fig. 8(d) shows the photograph of the measurement setup of the free-space method. The implemented FSS is placed into the rectangular through-hole window in the center of the rotatable screen for the measurement of the angular stability. Two horn antennas operating from 1 to...
18 GHz are located about 120 cm apart from each side of the rotatable screen, so that a uniform plane wave strikes upon the FSS structure. In addition, the measurement system is surrounded by the absorbing screens. In terms of the transmission coefficients measurement, after obtaining the frequency response of the system with the FSS, the environment noise is eliminated by the measured results of an identically-sized metallic plate. Then, the propagation loss is eliminated by the normalization of the measured results without the FSS. In case of multipath effects, the time-domain gating function is also applied to calibrate the measured results.

As shown in Fig. 9, the measured and simulated transmission coefficients under oblique incidence for both TE and TM polarizations are compared, where a desired dual-bandpass response with close band spacing is obtained. It is also observed that the filtering performance of the proposed 3-D FSS is stable for the incident angles up to 45° for both TE and TM polarizations. The measured transmission losses at the center frequency of the lower and higher band are 1.36 and 1.03 dB under the normal incidence. The measured transmission losses are larger than the simulated ones due to the conductor losses which are not exactly considered in the simulated model, actual dielectric constant, measurement error, fabrication and assembly tolerances. However, the measured results have demonstrated the performances of the proposed 3-D FSS. Table I gives the comparison of the characteristics between the proposed 3-D FSS and several FSSs with similar responses, validating its advantages in small band ratio, good angular stability, dual polarizations, as well as relatively compact unit cell.

4. Conclusion

In this paper, a dual-polarized 3-D FSS based on multiple SCWs is presented to realize a dual-bandpass response with close band spacing. Two passbands are formed by the square slot resonances in SCW paths and the half-wavelength resonance in PPW path. Moreover, multiple transmission zeros are introduced due to the counteraction of out-of-phase signals between different paths for high selectivity. The operating principle of the proposed 3-D FSS is explained by the E-field analysis, ECM and parameters study. At last, the measured results of an implemented FSS verify the properties of the proposed design.

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Table I. Comparison of the FSS designs with similar responses

| Ref. | Band Ratio | Unit Cell Size and Thickness | Num. (TPz/TZs) | Polarization | Stability (TE/TM) |
|------|------------|------------------------------|----------------|--------------|------------------|
| [12] | 1.084      | 0.293l0 × 0.293l0 × 0.022l0 | 2/1            | dual         | 45°/45° (only sim.) |
| [14] | 1.53       | 0.133l0 × 0.133l0 × 0.077l0 | 2/1            | dual         | 30°/30° (only sim.) |
| [17] | 1.4        | 0.083l0 × 0.083l0 × 0.008l0 | 2/2            | dual         | 60°/60° (only sim.) |
| [22] | 1.37       | 0.495l0 × 0.495l0 × 0.05l0  | 4/0            | dual         | 30°/30° (only sim.) |
| [23] | 1.22       | 0.37l0 × 0.37l0 × 0.22l0   | 4/4            | dual         | 45°/45° (meas.)   |
| [24] | 1.45       | 0.38l0 × 0.38l0 × 0.135l0  | 4/4            | dual         | 30°/30° (meas.)   |
| [25] | 1.39       | 0.273l0 × 0.096l0 × 0.03l0  | 4/3            | single       | 60°/60° (only sim.) |
| [28] | 1.95       | 0.07l0 × 0.055l0 × 0.09l0  | 4/0            | single       | 40°/40° (meas.)   |

| This work | 1.1       | 0.214l0 × 0.214l0 × 0.136l0 | 3/2            | dual         | 45°/45° (meas.)   |

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