Single- and Dual-Band High-Order Bandpass Frequency Selective Surfaces Based on Aperture-Coupled Dual-Mode Patch Resonators

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Abstract—In this paper, a class of bandpass frequency selective surfaces (FSSs) based on aperture-coupled dual-mode patch resonators (AC-DMPRs) are proposed to achieve single- and dual-band high-order filtering responses with low profiles. Initially, a basic resonator of a square patch with diagonal corner truncations is theoretically investigated so as to demonstrate that two orthogonal modes can be simultaneously excited in a single patch resonator. Then, to eliminate the cross-polarized reflection caused by the orthogonal modes, a composite resonator, including four of such corner-truncated patches with 90° rotation between each two adjacent patches, are constructed and analyzed. Furthermore, by arranging two of such composite resonators in a back-to-back manner through coupling apertures on the middle metallic layer, an FSS element of three-layer AC-DMPR is formed. Compared with traditional AC-PRs, the resonant modes of AC-DMPR are increased twice, thus giving more design flexibility to achieve high-order performance. To validate the design concept, a single-band fourth-order bandpass FSS is designed by introducing magnetic couplings in the AC-DMPRs. Moreover, dual-band second-order bandpass FSSs with magnetic and/or electric couplings in the AC-DMPRs is also designed. Based on the even- and odd-mode analysis method, equivalent circuit models are established to explain the operating principles of the proposed structures. Finally, the designed FSSs are fabricated and measured. Good agreement between the measured and simulated results well validates the conceptual designs.

Index Terms—Aperture-coupled patch resonators, dual-mode, frequency selective surface, high-order filtering response.

I. INTRODUCTION

Frequency selective surfaces (FSSs), also known as spatial filters for electromagnetic waves from microwaves to optical signals, have been extensively studied over the past decades. They are usually composed of periodically arranged metallic patch or aperture elements, and are designed to manipulate amplitudes, phases, and polarizations of the incident electromagnetic (EM) waves [1], [2]. Among them, a traditional bandpass FSS possesses single-layer topology and exhibits a first-order filtering response, which fails to meet the performance requirement of single-/multi-band high-order filtering response in advanced radar and communication systems [3]–[6].

To achieve a high-order bandpass filtering response with sharpened rejection skirts, it is straightforward to be realized by cascading multiple first-order bandpass FSSs with quarter-wavelength spacers. Unfortunately, the overall thickness of these cascaded structures should be too large, thus leading to bulky volume [7], [8]. In [9], [10], a kind of multilayer bandpass FSS with non-resonant elements has been proposed to obtain low profile and single-band high-order filtering response. Moreover, dual-band high-order performance can also be realized by introducing resonant elements between non-resonant elements in different metallic layers [11], [12]. However, the filtering order of this kind of FSS is strictly dependent on the number of metallic layers. Another alternative approach to realize high-order bandpass FSS is using three-dimensional (3-D) cavity elements [13]–[16]. These 3-D FSSs can provide more degrees of freedom in design to realize superior features, such as single-/multi-band high-order filtering response [17]–[19]. However, the main disadvantages of these 3-D FSSs are their complicated topologies and large assembly tolerance, which extremely restrict their applications.

On the other hand, aperture-coupled patch resonators (AC-PRs), consisting of back-to-back patch resonators with coupling apertures etched on the middle common ground, have been developed in planar filters to provide another dimension in flexible design and achievement of high-order filtering functionality [20]. Based on this concept, AC-PRs with different configurations [21]–[26] have been developed to realize single-/multi-band high-order bandpass FSSs. In [21], an FSS based on AC-PR elements has been proposed to realize second-order bandpass filtering response. It has been mentioned that the coupling apertures are non-resonant and
arranged at the position where the magnetic field intensities of the patch resonators are strongest, thus providing magnetic coupling between the back-to-back patch resonators. In [22], an FSS element on AC-PRs with dominant electric couplings has been presented, where two transmission zeros are produced near the skirts of the passband and thereby enhancing the frequency selectivity. In order to realize high-order filtering response and improve the frequency selectivity, FSS elements based on AC-PRs with resonant coupling apertures, also named as antenna-filter-antenna (AFA) modules, have been proposed [23]. It has been found that FSSs based on AC-PRs also have advantages of high-frequency selectivity, low profile, and stable filtering performance under oblique incident angles.

In this paper, a novel periodic element of aperture-coupled dual-mode patch resonators (AC-DMPR) is proposed to realize high-order bandpass FSSs, where the resonant modes are increased twice than traditional AC-PRs. First, in order to generate two orthogonal modes with different resonant frequencies in a square patch, two diagonal corners of the square patch are truncated. Then, four of such corner-truncated patch resonators with 90° rotation between each two adjacent ones are arranged together to eliminate the cross-polarized reflection caused by the two orthogonal modes. Furthermore, to achieve high-order bandpass filtering response, the composite corner-truncated patch resonators are placed in a back-to-back manner through coupling apertures on a middle metallic layer, where an FSS element of three-layer AC-DMPR is constructed. By reasonably designing the apertures to provide magnetic and electric couplings, single- and dual-band high-order bandpass filtering responses can be obtained respectively. To investigate the operating principles of the proposed FSSs, equivalent circuit models are established and analyzed. The remainder of this paper is organized as follows: The operation principles of single and four square patches with diagonal corner truncations are investigated in section II. In section III and IV, a single-band fourth-order and two dual-band second-order bandpass FSSs are designed and analyzed, respectively. In section V, two prototypes of the designed single- and dual-band FSSs are fabricated and measured to verify our concept. Finally, concluding remarks are given in section VI.

II. SQUARE PATCH WITH DIAGONAL CORNER TRUNCATIONS

In the design of planar dual-mode microwave bandpass filters and circularly-polarized antennas [27]–[30], the diagonal corner truncation of a single square patch resonator is a well-known technique to control resonant properties of its orthogonal modes along diagonal lines. Based on this concept, this technique is extended to design dual-mode FSS elements.

A. Single Square Patch with Diagonal Corner Truncations

Fig. 1(a) shows the geometry of the single square patch resonator with diagonal corner truncations. It can be seen that the whole patch resonator possesses a double-layer topology, consisting of a corner-truncated square patch on the top layer and a square shape of metallic conductor on the bottom layer. Assuming that the incidence of a linearly polarized EM wave and periodic boundary conditions (PBCs) are constructed in this element, as illustrated in Fig. 1(a), the excited electric and magnetic field intensities are concentrated between the top-layer patch and bottom-layer conductor, thus leading to the generation of cavity modes [27]. Moreover, since the bottom layer is fully printed by metallic conductor, the incident wave is all reflected and only reflection coefficients can be obtained.

In order to understand the operation mechanism, the incident electric field ($E_i$) is decomposed into two orthogonal and equal components $E_{i\parallel}$ and $E_{i\perp}$ along the diagonal lines $AA'$ and $BB'$ respectively, as shown in Fig. 1(a). On the other hand, as indicated in [27], a square patch supports two orthogonal modes along two diagonal lines respectively. These two orthogonal modes, denoted as mode A and mode B in Fig. 1(b), are independent, which obviously bring different

![Fig. 1. Geometry and resonant E-field intensities of the basic resonator of single square patch with diagonal corner truncations. (a) Geometry and simulation set-up. (b) Mode A and Mode B of a square patch resonator with diagonal truncations.](image)

![Fig. 2. Simulated and calculated S-parameter results of the square patch with diagonal corner truncations under y-polarized incidence. (Physical dimensions: $P = 8 \text{ mm}, D = 5.3 \text{ mm}, c = 1.6 \text{ mm}, h = 0.813 \text{ mm}, \varepsilon_r = 3.55.$)](image)
resonant frequencies due to the truncation of diagonal corners. Therefore, based on the mode orthogonality and structural symmetry, the incident \( \mathbf{E} \)-field component \( \mathbf{E}_A \) (or \( \mathbf{E}_B \)) can only excite mode A (or mode B), resulting in a reflected component \( \mathbf{E}_r^A \) (or \( \mathbf{E}_r^B \)). With coordinate transformations, the reflected \( \mathbf{E}_r^A \) and \( \mathbf{E}_r^B \) can be expressed as

\[
\mathbf{E}_r^A = \frac{1}{2} \mathbf{e}_x \mathbf{E}_r \begin{pmatrix}
(S_{11})_x & +
(S_{11})_y \end{pmatrix} + \frac{1}{2} \mathbf{e}_y \mathbf{E}_r \begin{pmatrix}
(S_{11})_y & -
(S_{11})_x \end{pmatrix}
\]

\[
\mathbf{E}_r^B = \frac{1}{2} \mathbf{e}_x \mathbf{E}_r \begin{pmatrix}
(S_{11})_y & +
(S_{11})_x \end{pmatrix} + \frac{1}{2} \mathbf{e}_y \mathbf{E}_r \begin{pmatrix}
(S_{11})_x & -
(S_{11})_y \end{pmatrix},
\]

where \( (S_{11})_x = |S_{11}|_x e^{i\phi} \) and \( (S_{11})_y = |S_{11}|_y e^{i\phi} \) are the reflection coefficients of the whole resonator under \( \mathbf{E}_A \)- and \( \mathbf{E}_B \)-polarized incidences respectively. In this way, the total reflected \( \mathbf{E} \)-field vector \( \mathbf{E}_r \) can be obtained by

\[
\mathbf{E}_r = \mathbf{E}_r^A + \mathbf{E}_r^B
\]

\[
= \frac{1}{2} \mathbf{e}_x \mathbf{E}_r \begin{pmatrix}
((S_{11})_y + (S_{11})_x) & +
((S_{11})_y - (S_{11})_x) \end{pmatrix} + \frac{1}{2} \mathbf{e}_y \mathbf{E}_r \begin{pmatrix}
((S_{11})_y + (S_{11})_x) & -
((S_{11})_y - (S_{11})_x) \end{pmatrix}
\]

where \( (S_{11})_x = \frac{1}{2}((S_{11})_y + (S_{11})_x) = \frac{1}{2}(|S_{11}|_y e^{i\phi} + |S_{11}|_x e^{i\phi}) \) and \( (S_{11})_y = \frac{1}{2}((S_{11})_y - (S_{11})_x) = \frac{1}{2}(|S_{11}|_y e^{i\phi} - |S_{11}|_x e^{i\phi}) \).

Herein, \( (S_{11})_x \) and \( (S_{11})_y \) are defined as x-polarized (cross-polarized) and y-polarized (co-polarized) reflection coefficients of the whole patch resonator under y-polarized incidence, respectively. In order to obtain \( (S_{11})_x \) and \( (S_{11})_y \), the whole patch resonator under the incidences of \( \mathbf{E}_A \)- and \( \mathbf{E}_B \)-polarized EM waves respectively is analyzed by using the full-wave simulator CST Microwave Studio (CST-MWS), as illustrated in Fig. 1(b). Then, the simulated magnitudes and phases of the reflection coefficients \( (S_{11})_x \) and \( (S_{11})_y \) are shown in Fig. 2. It can be observed that, both the magnitude responses \( (S_{11})_x \) and \( (S_{11})_y \) are equal to unity (0 dB), while the phase responses \( \phi_{11} \) and \( \phi_{11} \) are different due to the different resonant lengths for mode A and mode B. Therefore, \( (S_{11})_x \) and \( (S_{11})_y \) can be simplified as

\[
(S_{11})_x = \frac{1}{2} e^{i\phi} \left(1 + e^{i\Delta\phi}\right)
\]

\[
(S_{11})_y = \frac{1}{2} e^{i\phi} \left(1 - e^{i\Delta\phi}\right),
\]

where the \( \Delta\phi = (\phi_{11})_x - (\phi_{11})_y \) is the phase difference of the reflection coefficients between mode A and mode B under \( \mathbf{E}_A \) and \( \mathbf{E}_B \)-polarized incidences respectively. Based on the above understandings, it can be concluded that, under y-polarized incidence, the corner truncation of a square patch resonator can produce phase differences between mode A and mode B, thus leading to the existence of cross- and co-polarized reflections. Fig. 2 also presents the comparison of the magnitudes of \( (S_{11})_x \) and \( (S_{11})_y \), calculated from (4)-(5) and those obtained from CST-MWS directly (under y-polarized EM wave incidence). Very good agreement between them can be observed over the entire frequency band, which further verifies the validation of our analysis procedure. Moreover, a flat reflection band is achieved for cross-polarized reflection, and two reflection zeros at \( f_1 \) and \( f_2 \) are obtained for co-polarized reflection, which are produced at the positions of \( \Delta\phi = 180° \). As indicated in (4)-(5), when \( \Delta\phi = 180° \), \( |(S_{11})_x| = 0 \) and \( |(S_{11})_y| = 1 \). Then, the y-polarized incidence wave will be all converted to x-polarized reflection, thus leading to co-polarized reflection zeros (or cross-polarization reflection poles). This mechanism is similar to that commonly used in reflectarray to twist the linearly-polarized incident wave by 90° [28].

To investigate the relationship between the existence of cross- and co-polarized reflections and the degree of corner truncation, the variation of the phase difference \( \Delta\phi \) with respect to different dimensions of the truncations is presented in Fig. 3. Without corner truncation \( (c = 0 \text{ mm}) \), the square patch operates as a single-mode resonator and resonates at TM_{000} (TM0_{xxy}) mode under y-polarized incidence [27]. In this case, \( \Delta\phi = 0 \), \( |(S_{11})_x| = 1 \) and \( |(S_{11})_y| = 0 \), only co-polarized reflection can be obtained. When \( c \) increases from 0 to 1.4 mm, \( \Delta\phi \) increases accordingly, and the y-polarized incidence wave is gradually converted to x-polarized reflection, as shown in Fig. 3. As \( c \) reaches to 1.4 mm, there is one specific frequency that satisfies \( \Delta\phi = 180° \) and a co-polarized reflection zero (or cross-polarized reflection pole) is obtained. When \( c > 1.4 \text{ mm} \), there are two specific frequencies that satisfy \( \Delta\phi = 180° \) and two co-polarized reflection zeros (or cross-polarized reflection poles) are achieved, as shown in Fig. 3. Therefore, the number of co-polarized (or cross-polarized) reflection zeros (or poles) entirely depends on the degree of corner truncation. Finally, it can be concluded that, with corner truncations, a square patch can be transformed from a single-mode resonator to a dual-mode one, thus providing more resonances without increasing the number and size of the resonator.

**B. Four Square Patches with Diagonal Corner Truncations**

Although the single square patch with diagonal corner truncations can operate as a dual-mode resonator, the undesired cross-polarized reflection of such resonator is produced and maintained at a very large level because of the excitation of two
orthogonal modes (mode A and mode B). In order to address this problem, four of such corner-truncated square patches, denoted as resonators I and II in Fig. 4, are arranged together to construct a new composite resonator, where $90^\circ$ rotation of corner truncations between each two adjacent sub-resonators can be observed. Fig. 5 illustrates the excited $E$-field intensities of the composite resonator under $y$-polarized incidence. It can be observed that mode A and mode B are both generated in each corner-truncated square patch, which indicates that the dual-mode characteristic is still maintained. Since the composite resonator is symmetrical with respect to the $x=P$ plane, as well as the diagonal lines $AA'$ and $BB'$ respectively, the reflected $E_y$ and $E_a$ can only have the same components along the $y$-direction, which leads to $(S_{11})_A = (S_{11})_B$. Therefore, the total reflected $E$-field vector $\vec{E}_r = \vec{E}_y^* + \vec{E}_a^*$ and the cross-polarized reflection is zero $(S_{11})_x = 0$ as indicated in (3). Then, the magnitude of the co-polarized reflection coefficients is calculated as

$$
|S_{11}| = \frac{1}{2} \left| 1 + e^{i\Delta \varphi} \right| = 1, \quad \Delta \varphi = (\varphi_1)_a - (\varphi_1)_b = 0.
$$

To verify this point, the simulated reflection coefficient of the composite resonator under $y$-polarized incidence is also shown in Fig. 5, where only co-polarized reflection with the value of 0 dB is observed. So far, the DMPR with four sub-elements is successfully constructed to be a basic FSS element. Then, this basic element is applied to design high-order bandpass FSSs in the following.

III. SINGLE-BAND FOURTH-ORDER BANDPASS FSS

A. Geometry and Performance

In order to realize a bandpass FSS, two identical composite resonators shown in Fig. 4 are arranged in a back-to-back manner and then are periodically duplicated in the $x$-$y$ plane, as shown in Fig. 6(a). The whole FSS element consists of three metallic layers separated by two identical dielectric substrates, where the top and bottom layers have the corner-truncated square patches, and the middle metallic layer is etched with coupling apertures. Moreover, the whole FSS element contains four sub-elements, each of which is an AC-DMPR, as discussed above. To show the detailed configuration, the front and cross-section views of a sub-element, including two cross-orthogonal apertures sandwiched between two back-to-back patch resonators, are displayed in Fig. 6(b)–(d). It is seen that these two cross-orthogonal apertures, possessing...
that, four prospective transmission poles are achieved at incidence as obtained by using CST-MWS. It can be observed the diagonal lines the top- and bottom-layer patches, and are also along the different widths and lengths, are etched beneath the center of equivalent circuit model. (b) Even-mode circuit. (c) Odd-mode circuit.

Fig. 7. Simulated S-parameter results and resonant E-field intensities of the single-band fourth-order bandpass FSS under ψ-polarized incidence. (Physical dimensions: P = 8 mm, D = 5.3 mm, c = 1.2 mm, l1 = 2.8 mm, l2 = 2 mm, w1 = w2 = 0.3 mm, h = 0.813 mm. Circuit parameters: 1/J01 = 280, L1 = 0.184 nH, C1 = 0.765 pF, K01 = 0.095, 1/J02 = 255, L2 = 0.096 nH, C2 = 1.2 pF, K02 = 0.0464). Fig. 8. Equivalent circuit model of the AC-DMPRs based single-band FSS. (a) Equivalent circuit model. (b) Even-mode circuit. (c) Odd-mode circuit.

different widths and lengths, are etched beneath the center of the top- and bottom-layer patches, and are also along the diagonal lines AA’ and BB’, respectively. Based on the coupled-resonator theory in microwave filters [30], two AC-DMPRs can produce a four-pole bandpass filtering response.

Fig. 7 shows the simulated S-parameter results of the proposed single-band fourth-order FSS under ψ-polarized incidence as obtained by using CST-MWS. It can be observed that, four prospective transmission poles are achieved at f_{even}, f_{odd}, f_{even}, and f_{odd}, respectively. The -3dB absolute bandwidth (ABW) of the passband is 2.68 GHz (12.12 to 14.80 GHz), resulting in a fractional bandwidth (FBW) of 20%.

B. Analysis and Design

In order to gain an insight into the operation principle, the resonant E-field intensities of the proposed single-band bandpass FSS are also illustrated in Fig. 7. It can be obtained that, the transmission poles at f_{even} and f_{odd} are produced by mode A, and the transmission poles at f_{even} and f_{odd} are provided by mode B. Therefore, it can be obtained that AC-DMPRs are constructed for the passband. Moreover, at these resonant frequencies, the E-field intensity is always at its minimal value at the positions of the coupling apertures. In other words, the apertures are etched at the position where the maximum magnetic field intensity is achieved at these resonant frequencies. Therefore, it can be well figured out that, the back-to-back patch resonators are magnetically coupled by the cross-orthogonal apertures [31].

Then, the equivalent circuit model of the proposed FSS can be established accordingly, as shown in Fig. 8(a). It can be observed that mode A and mode B are denoted by LC circuits, whose resonant frequencies are determined by the dimensions of the DMPR [27]. The input and output (I/O) ports with characteristic impedance of Z0 = 120 π represent the wave impedance in free space. The couplings between the I/O ports and the resonant modes (mode A and mode B) are represented as admittance inverters J01 and J02, respectively [32]. Besides, the magnetic couplings provided by the middle-layer apertures are denoted by mutual inductances L_{01} and L_{02}, where L_{01} = L_{02} and L_{02} = L_{2}K_{01} (K_{01} and K_{02} are the magnetic coupling

Fig. 9. Variation of K_{01} and K_{02} with respect to different aperture dimensions. (a) K_{01} and K_{02} vs. l1 and w1; (b) K_{01} and K_{02} vs. l2 and w2. (Physical dimensions: P = 8 mm, D = 5.3 mm, c = 1.2 mm, h = 0.813 mm.)

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coefficients provided by the apertures for mode A and mode B respectively) [31]. In order to precisely extract these coupling coefficients, the even- and odd-mode analysis method is then adopted, as shown in Fig. 8(b) and (c) [30]. As a result, the even- and odd-mode resonant frequencies, $f_{\text{even}}$, $f_{\text{odd}}$, $f_{2\text{even}}$, and $f_{2\text{odd}}$ are given by

$$f_{\text{even}} = \frac{1}{2\pi \sqrt{(L_1 + 2L_{\text{m}})C_1}}, \quad f_{\text{odd}} = \frac{1}{2\pi \sqrt{L_1C_1}}$$

$$f_{2\text{even}} = \frac{1}{2\pi \sqrt{(L_2 + 2L_{\text{m}})C_2}}, \quad f_{2\text{odd}} = \frac{1}{2\pi \sqrt{L_2C_2}}.$$  

(7)

It can be obtained from (7) that the odd-mode resonant frequencies $f_{\text{odd}}$ and $f_{2\text{odd}}$ are only attributed to resonances of the DMPPRs, while the even-mode resonant frequencies $f_{\text{even}}$ and $f_{2\text{even}}$ are determined by the DMPPRs and the coupling apertures [31]. Then, the coupling coefficients $K_{\text{me}}$ and $K_{\text{ne}}$ can be calculated by

$$K_{\text{me}} = \frac{x_1}{1-x_1} + x_1 = \frac{(f_{\text{odd}})^2 - (f_{\text{even}})^2}{(f_{\text{odd}})^2 + (f_{\text{even}})^2}. \quad (8)$$

$$K_{\text{ne}} = \frac{x_1}{1-x_1} + x_2 = \frac{(f_{2\text{odd}})^2 - (f_{2\text{even}})^2}{(f_{2\text{odd}})^2 + (f_{2\text{even}})^2}. \quad (9)$$

Since the $E$-field energy can only be coupled through the coupling aperture that is transverse to the direction of $E$-field vectors, the cross-orthogonal apertures can independently control the coupling coefficients $K_{\text{me}}$ and $K_{\text{ne}}$. Fig. 9 shows the variation of $K_{\text{me}}$ and $K_{\text{ne}}$ with respect to different aperture dimensions, where $K_{\text{me}}$ and $K_{\text{ne}}$ can be independently enlarged by increasing the dimensions of two apertures respectively. It should be noted that the even- and odd-mode resonant frequencies in (8) and (9) are obtained from CST-MWS by setting the middle metallic layer of the FSS element with a perfect magnetic and electric wall respectively [30], [31]. In addition, the admittance inverters $J_{01}$ and $J_{02}$ of the equivalent circuit model are extracted by ADS Schematic using the curve fitting method. After obtaining all the circuit parameters, we can gain the S-parameter results of the proposed equivalent circuit model by using ADS Schematic, as shown in Fig. 7. Very good agreement can be observed between the results from EM simulation and those from the equivalent circuit model.

Finally, to obtain a desirable fourth-order bandpass filtering response, the general design guidelines can be concluded as: (i) First, design the patch size and diagonal corner truncations to achieve the resonant frequencies $f_{\text{odd}}$ and $f_{2\text{odd}}$. (ii) Second, carefully choose the dimensions of each coupling aperture to obtain the desired coupling coefficients $K_{\text{me}}$ and $K_{\text{ne}}$ so as to determine $f_{\text{even}}$ and $f_{2\text{even}}$ respectively; (iii) Finally, slightly adjust the dimensions of the FSSs to make sure that all the resonant frequencies are located at the desired locations.

IV. DUAL-BAND SECOND-ORDER BANDPASS FSS

A. Geometry and Performance

With the proposed AC-DMPRs, dual-band bandpass FSS can also be realized, as shown in Fig. 10. Similar to that of the single-band fourth-order bandpass FSS design, each sub-element in the dual-band design is also constructed by two back-to-back identical truncated-corners square patches with apertures etched in the middle conductive layer. The difference is that one of the original cross-orthogonal rectangular apertures are replaced by a pair of triangular apertures, which are allocated near the un-truncated corners of the patch, as illustrated in Fig. 10(b)-(d).

Fig. 11 shows the simulated S-parameter results of the proposed dual-band bandpass FSS under $\gamma$-polarized incidence as obtained by using CST-MWS. It can be seen that four transmission poles ($f_{\text{odd}}, f_{\text{even}}, f_{2\text{odd}}, f_{2\text{even}}$) are split into two groups, and two passbands are achieved around $f_1$ (12.12 GHz) and $f_2$ (15.05 GHz), respectively. The frequency band ratio $f_2/f_1$ is as small as 1.24. Moreover, two transmission zeros ($f_3$ and $f_4$) are located at both sides of the first passband, which enhance the isolation between these two operation passbands.

B. Analysis and Design

To understand the operation mechanism, Fig. 11 illustrates the front view of the resonant $E$-field intensities of the proposed dual-band bandpass FSS. It can be seen that mode A and mode B are responsible for the first and second passband respectively.
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Besides, it is observed that the triangular apertures are placed at the position where the maximal $E$-field intensity is achieved, and thus the back-to-back patch resonators are electrically coupled by the triangular apertures in the first passband. Besides, since the direction of each rectangular aperture (etched beneath the center of each patch) is parallel to the direction of the resonant $E$-field vectors, the first passband is not affected by the rectangular apertures. Moreover, both the rectangular and triangular apertures are allocated at the position with minimal electric (or maximal magnetic) field intensity in the second passband. Therefore, it can be concluded that, in the first passband, the back-to-back DMPRs are electrically coupled by the triangular apertures only, while they are magnetically coupled by both the triangular and rectangular apertures in the second passband [31]. Furthermore, the transmission zeros at both sides of the first passband are produced by cross-couplings between the path directly through the triangular apertures [22].

Based on the previous discussions, the equivalent circuit model of the dual-band FSS is established, as shown in Fig. 12. Since mode A of the back-to-back DMPRs are electrically coupled, the coupling between them can be represented by a mutual capacitance $C_{em} = C_{e}K_{e}$ ($K_{e}$ is the electric coupling coefficient provided by the triangular apertures for mode A). The magnetic coupling provided by the rectangular aperture for mode B is denoted by a mutual inductance $L_{m} = L_{2}K_{m}$, which is similar to the situation shown in Fig. 8. Moreover, a direct coupling path is introduced through the triangular apertures, where the coupling is represented by an inductance $L_{3}$ [22], [31]. Then, the even- and odd-mode resonant frequencies $f_{odd}$, $f_{even}$, $f_{even}$, and $f_{odd}$ are given by

$$f_{odd} = \frac{1}{2\pi\sqrt{L_{1}C_{1}}} \hspace{1cm} f_{even} = \frac{1}{2\pi\sqrt{L_{2}(C_{1} - 2C_{m})}}$$

$$f_{even} = \frac{1}{2\pi\sqrt{(L_{2} + 2L_{m})C_{2}}} \hspace{1cm} f_{odd} = \frac{1}{2\pi\sqrt{L_{c}C_{2}}}$$

(10)

It can be obtained from (10) that the odd-mode resonant frequencies $f_{odd}$ and $f_{even}$ are only attributed to resonances of the DMPRs, while the even-mode frequencies $f_{even}$ and $f_{odd}$ are determined by the DMPRs and the coupling apertures [31]. The coupling coefficients $K_{e}$ and $K_{m}$ can be calculated by

$$K_{e} = \frac{x_{1}}{1 - x_{1}}, \hspace{0.5cm} x_{1} = \left(\frac{f_{even}}{f_{odd}}\right)^{2} + \left(\frac{f_{even}}{f_{odd}}\right)^{2}$$

(11)
Similarly, the variation of coupling coefficients $K_e$ and $K_m$ with respect to different aperture positions and dimensions are then extracted and shown in Fig. 13. It is illustrated that, $K_e$ and $K_m$ can be simultaneously enlarged by increasing the value of $l_1$ and $w_1$. In addition, the inductance $L_3$ and the admittance inverters $J_01$, $J_02$, and $J_03$ representing the couplings between the I/O ports and the FSS element can be obtained by using ADS Schematic with a curve fitting method. The S-parameter results of the proposed equivalent circuit model are then simulated by ADS Schematic, as shown in Fig. 11. Reasonable agreement can be attained between the results from the EM simulation and those from the equivalent circuit model.

Finally, the general design guidelines of the second-order dual-band bandpass FSS can be concluded as: (i) First, design the patch size and diagonal corner truncations to achieve $f_{odd}$ and $f_{odd}$ according to the given operation frequencies and band ratio $f_s/f_1$; (ii) Second, choose the dimension of $s$ to make sure that the triangular apertures are at the position where the resonant $E$-field of the AC-DMPRs is strongest. (iii) Third, carefully choose the dimensions of each coupling aperture to obtain the desired coupling coefficients $K_e$ and $K_m$ so as to determine $f_{low}$ and $f_{low}$ respectively; (iv) Finally, slightly adjust the dimensions of the FSS to make sure that all the resonant frequencies are at the desired locations.

It should be noted that the roll-off of the first passband is sharp as two transmission zeros are introduced at its two skirts, while that of the second passband is not very fast. This is because the back-to-back DMPRs are electrically coupled by the triangular apertures in the first passband, while they are magnetically coupled by the rectangular apertures in the second passband. It is indeed true that if electrical couplings are introduced for the second passband, two transmission zeros can also be provided at two skirts of the second passband, thus leading to a fast roll-off characteristic for both passbands. Based on this concept, a modified dual-band bandpass FSS is designed, as shown in Fig. 14(a). It can be seen that the rectangular apertures are allocated beneath the truncated corners of the square patches, thus leading to electrical couplings between the back-to-back DMPRs for the second passband. The corresponding equivalent circuit model is shown in Fig. 14(b), where the electrical couplings between the DMPRs are denoted by $C_{m1}$ and $C_{m2}$ respectively. Fig. 15 shows the simulated S-parameter results obtained by CST-MWS and ADS Schematic. It can be seen that two pairs of transmission zeros ($f_1$, $f_2, f_3$, and $f_4$) at two skirts of both passbands are provided for the modified design, which can significantly improve the roll-off characteristic.

V. EXPERIMENTAL VERIFICATION

To validate our proposed design concept, two prototypes of the single- and dual-band FSS with one pair of transmission zeros designs are fabricated and measured. The dielectric substrates are chosen as the RO4003C materials with $\varepsilon_r = 3.55$, $\tan(\delta) = 0.0029$. The substrate thickness is chosen as 0.813 mm and 0.406 mm for the single- and dual-band FSSs, respectively. Besides, their overall sizes of the fabricated prototypes are the same, which are 160 mm $\times$ 160 mm (including 10 $\times$ 10 elements). Fig. 16 shows the photographs of the measurement
set-up for the fabricated FSS prototypes. The overall measurement system is established in an anechoic chamber, which mainly contains transmitting and receiving antennas, a rotatable absorbing screen with a central test window, a Vector Network Analyzer (VNA), and cables. As shown in Fig. 16, the screen is placed at the center of the transmission/reflection path between the two antennas. In the measurement, the device under test (DUT) is placed at the center of the test window of

Fig. 16. The free space measurement system. (a) Measurement of transmission coefficients. (b) Measurement of reflection coefficients.

![Diagram](image1)

![Diagram](image2)

(a) (b)

Fig. 17. Measured and simulated results of the single-band fourth-order bandpass FSS under different incident polarizations and angles. (a) y-polarized incidence. (b) x-polarized incidence.

![Graph](image3)

(a) (b)

Fig. 18. Measured and simulated results of the dual-band second-order bandpass FSS with one pair of transmission zeros under different incident polarizations and angles. (a) y-polarized incidence. (b) x-polarized incidence.
the rotatable screen. It should be mentioned that the distance \(d\) between the antennas and the DUT should satisfy the far-field distance requirement of the antennas. With the measurement setup, we record the original data of transmission responses between the antennas. For oblique incidence measurement, we appropriately rotate the rotatable screen or the antennas to the desired angle \(\theta\) and record the transmission responses between the antennas accordingly. Then, to eliminate the effects of environment noise, multiple reflections, and propagation loss of the EM waves, three situations, including the measured FSS, an identically-sized empty window, and a same-sized metallic plate, are measured, respectively. A complete calibration and normalization process, as described in [9] and [16], has been carefully considered to obtain accurate measurement results.

Figs. 17 and 18 show the measured and simulated S-parameter results of the designed single- and dual-band bandpass FSSs under different incident polarizations and angles. It can be seen that the measured results agree well with the simulated ones, which also indicate that the frequency performances of both designed FSSs are stable under different incident polarizations and angles (up to 40°). For the dual-band FSS design, the measured results of the second passband have little shifts compared with the simulated ones, which may be attributed to fabrication tolerances. Under normal incidence, the measured in-band minimum and maximum insertion losses of the single-band FSS are 1.1 dB and 2.9 dB respectively. For the dual-band FSS, the ranges of the measured insertion losses within the first and second passbands are from 0.9 to 2.4 dB and from 1.73 to 2.95 dB, respectively. The extra measured insertion loss in the second passband may be due to the surface roughness of the metallic layer and the increased substrate loss of the RO4003C used in this design.

In order to better exhibit the attractive advantages of the proposed bandpass FSSs, Tables I and II tabulate comparisons between our proposed designs and other existing designs in the literature. In addition to the properties of the ultra-thin total thickness, our proposed structures have their merit of higher-order bandpass filtering response with less metallic layers and simple topologies. For single-band performance, the technique proposed in [23] can also achieve a fourth-order bandpass filtering response with three metallic layers. Nevertheless, those structures are generally polarization-sensitive and suffer from a narrow angular range of operation (\(\theta < 20^\circ\)). For dual-band performance, a small band ratio of our structure can be designed by only tuning the dimensions of the truncated corners and coupling apertures. In this way, an extremely small band ratio of 1.23 can be achieved with our proposed technique.

In addition, the roll-off of our proposed single-band fourth-order design is relatively slow. However, compared with the traditional AC-PR FSS design [21], our proposed design features higher-order and sharper roll-off. In order to further improve the roll-off, transmission zeros should be introduced at two skirts of the passband, as done in the proposed dual-band bandpass FSS.

### VI. CONCLUSION

In this paper, a novel periodic element of AC-DMPR has been presented and studied to design single- and dual-band high-order bandpass FSSs. The operation principle of the corner-truncated square patch resonator has been analyzed to demonstrate that two orthogonal modes can be excited in a...
single square patch. Then, a composite resonator, including four of such corner-truncated patches with 90° rotation between each two adjacent patches, has been developed to eliminate the cross-polarized reflection caused by the orthogonal modes. Based on this, a bandpass FSS element based on AC-DMPRs, formed by two of such composite resonators arranged in a back-to-back manner with coupling apertures on the middle metallic layer, has been constructed. With the aid of equivalent circuit models, a fourth-order single-band and a second-order dual-band bandpass FSSs have been designed, fabricated, and measured, respectively. The measured results are in good accordance with the simulated ones, showing that stable performance can be satisfactorily maintained for different incident polarizations and angles. Benefiting from the low-profile property and good angular stability, the proposed FSSs are believed to be useful for conformal FSS applications.

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