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Three-Dimensional Dual-Polarized Frequency Selective Structure with Wide Out-of-Band Rejection

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Abstract—A new three-dimensional bandpass frequency-selective structure (FSS) is proposed for dual-polarized applications. Each square unit cell of the proposed FSS consists of vertical and horizontal double-sided parallel-strip lines (DSPSLs) and a thin metallic plate. Furthermore, the parallel-strip lines of every DSPSL are connected together through via holes. Using this arrangement, multimode resonators are established in the proposed structure. At lower frequencies, the substrate mode propagates and two resonators along the DSPSL are constructed, leading to a bandpass response with two transmission poles. At higher frequencies, the air mode is reflected by and resonates along the inserted plate, resulting in a stopband with two transmission zeros. The operating principle for this FSS is explained with the aid of an equivalent circuit model. Moreover, a parametric study of the proposed FSS is also carried out and design guidelines are formulated. As an example, a prototype of the proposed FSS is designed, fabricated, and tested. Measured results demonstrate that the FSS exhibits dual polarizations, high out-of-band rejection, and stable frequency response under both transverse electric (TE) and transverse magnetic (TM) polarizations of an obliquely incident wave.

Index Terms—Double-sided parallel-strip line, dual-polarization, frequency-selective structure.

I. INTRODUCTION

Traditional frequency-selective surfaces, which are normally made of a two-dimensional (2-D) periodic arrangement of identical unit cells (such as patches, strips, apertures, slots, and the like), have been widely investigated [1], [2]. Unfortunately, most of these single-layer 2-D frequency-selective surfaces suffer from poor filtering characteristics, such as poor selectivity and unstable angular response. Subsequently, several novel multilayer and three-dimensional (3-D) frequency-selective surfaces were proposed to alleviate the above-mentioned shortcomings of traditional 2-D surfaces [3]-[7]. A bandpass frequency-selective surface by integrating two patch antenna arrays and one array of non-radiating resonant structures in between was described in [3]. Multiple transmission poles in the passband were obtained and one transmission zero was produced in the lower rejection band. The flatness of the passband was thus improved and a fast roll off in the rejection band was obtained. Another type of frequency-selective surface with high selectivity was presented by using multilayer topology in [4], [5]. An additional transmission zero near the passband was created by introducing an extra series LC resonator in one of the layers. In [6] and [7], several frequency-selective surfaces based on substrate integrated waveguides (SIWs) were proposed and quasi-elliptic filtering responses were achieved by introducing cross couplings between SIW cavities and slots etched on the top and bottom surfaces of these SIWs. A 3-D frequency selective surface made of an array of circular tubes was described in [8], where the frequency response was mainly determined by the tube thickness.

Recently, a novel concept of 3-D frequency-selective structure (FSS) was reported in [9], where multiple transmission zeros/poles can be produced by a 3-D FSS with multimode cavities/resonators. Compared to 2-D surfaces, the filtering responses of 3-D FSSs can be greatly improved by introducing transmission zeros/poles at desired frequencies. Based on this concept, a new type of 3-D FSS consisting of a 2-D periodic array of vertical microstrip line resonators was proposed in [10]-[12] to achieve quasi-elliptic bandstop filtering response. Transmission poles and zeros were realized by using resonances and couplings of two quasi-transverse electromagnetic (TEM) modes excited by the shielded microstrip line. On the other hand, by combing this microstrip-line array with an array of rectangular waveguide resonators, a quasi-elliptic bandpass FSS with two transmission poles and two transmission zeros located near the passband was realized in [13]. Although quasi-elliptic bandpass responses were observed in [3]-[7], [13], these designs cannot achieve wideband out-of-band rejection together with high selectivity, which is actually desired in many practical applications. Furthermore, the FSSs presented in [9]-[13] only operate under a single polarization, which further limits their applications.

In this paper, a new 3-D bandpass FSS with wide stopband characteristic is presented for dual-polarized applications. In each unit cell, multimode resonators for both transverse electric (TE) and transverse magnetic (TM) polarizations are constructed by using two printed double-sided parallel-strip lines (DSPSLs): one pair in the horizontal direction and another pair in the vertical direction, as well as an inserted metallic plate. At lower frequencies, signals can pass through the structure and a bandpass response with a transmission pole is realized around the resonant frequency of the DSPSL. Besides, the parallel strip lines of each DSPSL are connected together, which can introduce one more resonator in the passband leading to a wider bandwidth. The inserted metallic plate in the air region is employed to provide another two resonators that produce two transmission zeros at higher frequencies. A practical realization of the proposed FSS operating at the center frequency of 7.5 GHz is designed. Under the normal incidence,
the designed FSS has a relative 3-dB passband bandwidth of 18.4% and a bandwidth of about 78.4% for the out-of-band rejection better than 20 dB.

II. DESCRIPTION OF THE FSS AND DSPSL
A. Description of the FSS
The proposed 3-D FSS, illustrated in Fig. 1(a), consists of two arrays of vertically and horizontally placed DSPSLs and a number of inserted metallic plates. Figs. 1(b) and (c) show the structural details of a square unit cell, where the dash line represents the periodic boundary. It is seen that the two parallel strip lines of each DSPSL are connected together by a centered via hole. In addition, a square and thin metallic plate is inserted in the air region, which is in contact with the inner strip lines. Periods along the $x$- and $y$-axes are denoted by $b$. The thicknesses of the proposed FSS and the inserted metallic plate along the $z$-axis are represented by $l$ and $l_m$, respectively. The distances between the metallic plate and two ends of the DSPSL are $l_1$ and $l_2$. The line width of each strip line and the diameter of the via hole are $w$ and $D$, respectively. It is noted that the horizontal and vertical DSPSLs have the same topology and dimension, and they are responsible for TE and TM polarizations of the incident wave, respectively. It is therefore seen that the proposed FSS is insensitive to the polarization of an incoming wave under the normal incidence. In the following sections, we will only discuss the operating principle for the horizontally placed DSPSL under the normal incidence.

![Fig. 1. Geometry of the proposed FSS consisting of vertically and horizontally placed DSPSLs and inserted metallic plates. (a) Perspective view of the proposed FSS and one unit cell; (b) Front view of a unit cell in the xy plane; (c) Side view of a unit cell in the yz plane.](image)

B. Double-Sided Parallel-Strip Line
Fig. 2 shows the cross-section and electric (E-) field distribution of a DSPSL, which comprises two identical strip lines, one on the top layer of the substrate and the other on the bottom. According to the descriptions in [14] and [15], a DSPSL can be considered as a printed finite conductor parallel plate waveguide, which supports quasi-TEM waves. Furthermore, the quasi-TEM field distribution remains unchanged if a virtual ground plane with infinite extent is inserted at the center of the substrate and parallel to the strip conductors, as shown in Fig. 2. Therefore, the DSPSL can be considered as a combination of two identical back-to-back microstrip lines and the DSPSL has similar characteristic to that of microstrip lines [15]. Compared with a microstrip line, a DSPSL has several advantages: (i) it has wider line width than a microstrip line with the same characteristic impedance; (ii) it gives shorter wavelength than a microstrip line with the same line width. In this design, the DSPSL is employed not only based on the above two advantages, but also its propagating quasi-TEM modes in the entire frequency range. If DSPSLs are replaced with traditional microstrip lines, the ground plane of horizontal and vertical microstrip lines will connect together and they may form a dielectric-filled square waveguide, which is more dispersive. The proposed FSS is somehow similar to the structure described in [13]. However, waveguide modes are excited and propagate in the structure described in [13], thus resulting in narrow bandwidth and also narrow out-of-band rejection.

![Fig. 2. E-field lines on the cross section of a DSPSL.](image)

III. THE OPERATING PRINCIPLE

It may be useful to recall the structure presented in [10] and its operating principle. When the E-field of an incident plane wave is perpendicular to the strip lines, two propagating paths, the air and substrate paths, are formed along the shielded microstrip lines. Correspondingly, air and substrate quasi-TEM modes are excited and propagate in the two paths, respectively. At lower frequencies, most signals go from one port to the other through the air path, thus leading to a lowpass response. At higher frequencies, signals can pass through both paths. Owing to the fact that the guided wavelength of substrate mode is smaller than that of air mode, signals coupled through two paths will have a phase difference. Transmission zeros are then obtained at frequencies where the two propagating modes are combined out of phase. Transmission poles are produced when the individual length of the air or substrate mode reaches approximately half a guided wavelength.

In each unit cell of the present design, two horizontal and vertical via holes are introduced at the center of the corresponding DSPSLs and a metallic plate is inserted in the air region, which are different from the structure in [10]. The substrate propagating path still exists through DSPSLs, which...
means that the substrate quasi-TEM mode can also be excited and propagate though it may be perturbed by the introduced via holes. However, the air propagating path is blocked by the inserted metallic plate because signals coupled to the air path will be reflected by the inserted metallic plate. Therefore, the air quasi-TEM mode may not link the input and output ports though they may still be excited. Transmission zeros are then achieved at frequencies where the air mode meets their resonant conditions. Compared with the structure in [10], the present FSS can provide more resonators in the operating frequency range and has very different filtering response. Fig. 3 shows the simulated S-parameter results of the proposed FSS, where a good bandpass filtering response with a wide stopband at higher frequencies can be observed under the normal incidence.

Fig. 3. Simulated S-parameter results of the proposed bandpass FSS using full-wave EM simulator CST-MWS \((b = 8.0 \text{ mm}, w = 4.0 \text{ mm}, l = 12 \text{ mm}, l_s = 4.65 \text{ mm}, l_1 = 3.05 \text{ mm}, D = 1.7 \text{ mm}, e_r = 3.0, d = 1.524 \text{ mm}, \text{TE incidence, } \theta = 0^\circ\).

A. Resonant Characteristics of the Substrate Mode

As discussed above, only the substrate propagating path along DSPSL exists linking the input and output ports of the proposed FSS. At a very low frequency, although most signals are coupled to the substrate path due to the existence of the metallic plate, the proposed FSS still exhibit strong reflection because the two parallel strip lines are connected by a via hole. However, the DSPSL section will resonate when the frequency increases to a certain value. A passband with two transmission poles can then be obtained around the resonant frequencies of the DSPSL with the via hole, as shown in Fig. 3. The propagating and resonant characteristics of the FSS can be explained by an equivalent circuit model shown in Fig. 4. It is seen that the equivalent circuit model contains two series sub-networks representing the air and substrate paths. In the substrate path, the DSPSL is represented by a transmission line with equivalent characteristic impedance \(Z_d\) and electrical length \(\theta\). The coupling between the input/output ports and the DSPSL is denoted by capacitor \(C_d\). The inductor \(L\) denotes the effect of the via hole connecting the two strip lines of a DSPSL.

The E-field lines of a DSPSL (in the yz plane) at the two resonant frequencies \(f_{p1}\) and \(f_{p2}\) are shown in Figs. 5(a) and (b). It is seen that, at resonant frequency \(f_{p1}\), the E-field vectors of the DSPSL have the same magnitude and direction at both sides of the via hole, which means the transmission pole at \(f_{p1}\) is produced by the resonance between half of the DSPSL and the via hole. In the equivalent circuit model, this resonator can be represented by \(R_1\), which can also be seen as a short-circuited transmission line resonator [16]. At resonant frequency \(f_{p2}\), the E-field vectors at both sides of the via hole along the DSPSL still have the same magnitude but opposite directions, which means the transmission pole at \(f_{p2}\) is provided by the resonance of the entire DSPSL. In the equivalent circuit model, this resonance is denoted by resonator \(R_2\), which can be equivalent to an open-circuited transmission line resonator [16].

![Fig. 4. Equivalent circuit model of the proposed FSS.](image)

![Fig. 5. (a) and (b) E-field distributions for the substrate mode at frequencies \(f_{p1}\) and \(f_{p2}\); (c) and (d) E-field distributions for the substrate mode at frequencies \(f_{p1}\) and \(f_{p2}\).](image)
modes at frequencies $f_1$ and $f_2$. It is seen that the $E$-field vectors
coupled from the input ports and then decreases to zero when these
signals reach the metallic wall. The resonant
characteristics of this case can also be explained by the
equivalent circuit model shown in Fig. 4. In the air path of this
equivalent circuit, two transmission-line sections $(Z_a, \theta_3)$ and
$(Z_a, \theta_4)$ denote the left and right sides of the separated air path,
respectively. $Z_a$ is the equivalent characteristic impedance of
the air mode; $\theta_3$ and $\theta_4$ are the electrical lengths of the left
and right sides of the air path. Due to the coupling between vertical
and horizontal DPSLs, the equivalent characteristic impedance $Z_a$
is lower than the standalone impedance. The inserted metallic plate is represented by a short circuit between
$(Z_a, \theta_3)$ and $(Z_a, \theta_4)$. The discontinuity between the
input/output ports and the air region of the presented FSS is
denoted by capacitor $C_w$. As discussed earlier, both transmission-line sections $(Z_a, \theta_3)$ and $(Z_a,$
$\theta_4)$ will resonate and produce transmission zeros at frequencies above the
passband. In order to facilitate the description, resonators $R_3$
and $R_4$ shown in Fig. 4 are used to represent the resonant
characteristics of transmission-line sections $(Z_a, \theta_3)$ and $(Z_a,$
$\theta_4)$. It should be clear that $R_3$ and $R_4$ can be seen as two
short-circuited transmission line resonators. More detailed
simulation results can be found in the parametric analysis in
Section IV.

IV. PARAMETRIC STUDY AND DESIGN GUIDELINES

A. Parametric Study

In order to establish a generally applicable design procedure
for the proposed FSS, a parametric study is conducted. The
analysis is carried out by simulating different variations of the
proposed structure with the help of the full-wave simulator CST
MWS. As shown in Fig. 3, there are six specifications when
designing this FSS: two transmission-pole frequencies $f_{p1}$ and
$f_{p2}$, two transmission-zero frequencies $f_{z1}$ and $f_{z2}$, $RBW_{3dB}$
and $RBW_{20dB}$. $RBW_{3dB}$ and $RBW_{20dB}$ are defined as follows:

$$RBW_{3dB} = \frac{\text{Bandwidth of } |S_{21}| \geq -3\text{dB}}{(f_{p1} + f_{p2})/2} \times 100\% \quad (1)$$

$$RBW_{20dB} = \frac{\text{Bandwidth of } |S_{21}| \geq -20\text{dB}}{(f_{p1} + f_{p2})/2} \times 100\% \quad (2)$$

where $RBW_{3dB}$ represents relative 3-dB bandwidth of the
passband and $RBW_{20dB}$ denotes 20-dB relative bandwidth of
out-of-band rejection in higher frequencies.

Figs. 6(a) and (b) show the variation of resonant frequencies and
bandwidths with respect to the diameter $D$ of the via hole
and the width $w$ of DSPSL, respectively. It is seen from Fig. 6(a)
that $D$ only affects the resonant frequency $f_{p1}$ and a smaller $D$
leads to a lower $f_{p1}$. This can be explained by the equivalent
circuit model in Fig. 4. The inductance $L$ is mainly determined
by $D$, which is only included in resonator $R_1$. Therefore, the
resonant frequency $f_{p1}$ varies with $D$, while other resonant
frequencies remain more or less the same. Furthermore, the
bandwidth of passband ($RBW_{3dB}$) increases first and then
decreases with a growing $D$. When $D$ is small, $f_{p1}$ is much lower
than $f_{p2}$ and the structure exhibits a dual-band bandpass
response, thus leading to a narrow bandwidth. When $D$ is very
large, $f_{p1}$ and $f_{p2}$ may merge together, which obviously results in
a narrow bandwidth. As observed in Fig. 6(b), all of the resonant
frequencies will change when $w$ varies. This is because the strip line is present in all of the resonators $R_1$, $R_3$, $R_3$
and $R_4$. The frequencies $f_{p1}$ and $f_{p2}$ of transmission poles
decrease with an increasing $w$, while the frequencies $f_{z1}$ and $f_{z2}$
of transmission zeros increase. This is because increasing $w$
will greatly enlarge the coupling capacitor $C_d$ in resonators $R_1$
and $R_4$, thus resulting in lower resonant frequencies. It is also
noted that increasing $w$ leads to smaller impedance for the strip
line, which means the values of $Z_q$ and $Z_w$ become smaller and
so do the electrical lengths of $\theta_1$, $\theta_3$ and $\theta_4$. Therefore, the resonant frequencies $f_{z1}$ and $f_{z2}$ become higher accordingly.
Furthermore, both $RBW_{3dB}$ and $RBW_{20dB}$ increase when
enlarging $w$, which suggests that $w$ should be as larger as possible
to obtain wide bandwidths.

Fig. 7 shows the resonant frequencies and relevant
bandwidths under different values of $l_3$ or $l_4$ ($l_3/4\lambda$).
It is seen that when adjusting $l_3/4\lambda$, the frequencies $f_{p1}$ and $f_{p2}$ remain almost
unchanged, while those of transmission zeros vary accordingly.
This is because $l_1$ and $l_4$ are only related to resonators $R_1$ and $R_4$
respectively, which have no influence on $R_3$. As seen in
Fig. 7, when increasing $l_3$ while fixing $l_4 = 4.5$ mm, $f_{p1}$ is
lowered significantly, while $f_{z2}$ is unchanged at around 11.85
GHz. The same situation can be observed for \( f_2 \) when adjusting \( l_4 \) and fixing \( l_3 = 4.5 \text{ mm} \). This means that the two transmission zeros provided by resonators \( R_3 \) and \( R_4 \) can be adjusted independently while retaining the passband performance, which provides more freedom for the FSS design. It should be noted that when \( l_3 = l_4 = 4.75 \text{ mm} \), the metallic plate is placed at the middle of the DSPSL and the two transmission zeros are merged into one, as shown in Fig. 7. This is understandable because \( R_3 \) and \( R_4 \) have the same length, resulting in the same resonant frequency. Furthermore, it should be mentioned that \( f_1 \) and \( f_2 \) may not be adjusted too far away from each other for a large RBW\(_{20\text{dB}}\) because the ripple in the stopband will become noticeable. Therefore, \( l_3 \) and \( l_4 \) should be properly adjusted to achieve a wider bandwidth for the out-of-band rejection.

Fig. 8 shows resonant frequencies and bandwidths as functions of period \( b \) and thickness \( l \) of the proposed FSS. It is observed that \( f_1 \) and \( f_2 \) decrease with increasing \( b \) and \( l \), respectively. This is attributed to the fact that the resonant lengths of \( R_3 \) and \( R_4 \) increase when enlarging \( b \) and \( l \). The frequency \( f_1 \) decreases with an increasing value of \( b \), while \( f_2 \) increases. This can be explained by the inverse relationship between the strip line’s impedance and the coupling between these closely spaced strip lines. The difference between \( f_{p1} \) and \( f_{p2} \) can then be widened by using a large \( b \). Meanwhile, RBW\(_{20\text{dB}}\) decreases with a large \( b \), while it increases with an increasing \( l \). Moreover, RBW\(_{3\text{dB}}\) increases initially and then decreases when \( b \) increases. This is expected because when \( b \) is too small, \( f_{p1} \) and \( f_{p2} \) are too close to each other; when \( b \) is too large, the difference between \( f_{p1} \) and \( f_{p2} \) becomes too large and the structure exhibits a dual-band bandpass response, which is similar to the situation when \( D \) is small.

- Period \( b \) should be selected much smaller than the operating wavelength for a larger out-of-band rejection bandwidth and stable frequency performance even under a large variation of the angle of incidence. However, \( b \) may not be selected too small since it may be difficult and costly to fabricate and assemble the structure.
- The thickness \( l \) is approximately equal to \( \lambda_g/2 \), where \( \lambda_g \) is the wavelength of the quasi-TEM mode at the desired center frequency. \( \lambda_g \) is approximately \( c/f_0\sqrt{\varepsilon_{\text{eff}}} \), where \( c \) is the speed of light in free space, \( f_0 \) denotes the desired center frequency, and \( \varepsilon_{\text{eff}} \) represents the effective dielectric constant of the structure.
- Since a large \( w \) can lead to a wide bandwidth, the strip width \( w \) is chosen to be close to \( (b - d) \). Since the diameter \( D \) only influences the first transmission pole, it can thus be used as a parameter for obtaining the desired bandwidth for the passband.
- The location and thickness of the metallic plate is determined by \( l_3 \) and \( l_4 \), which only influence the out-of-band rejection performance in higher frequencies. Decreasing \( l_3 \) or \( l_4 \) leads to a higher \( f_1 \) or \( f_2 \), and vice versa. \( l_3 \) and \( l_4 \) can be adjusted individually to obtain a wider bandwidth for out-of-band rejection.

![Fig. 7. Variation of resonant frequencies and bandwidths with respect to \( l_3 \) or \( l_4 \) (\( b = 8 \text{ mm}, l = 12 \text{ mm}, w = 4 \text{ mm}, D = 1.6 \text{ mm}, \varepsilon_r = 3.0, d = 1.524 \text{ mm}, \) TE incidence, \( \theta = 0^\circ \)).](image)

B. Design Guidelines

Based on the understanding gained through parametric studies, the following guidelines are suggested for a fast design of the proposed FSS.

- A substrate material with low dielectric constant \( \varepsilon_r \) may be preferred. Although this may lead to a relative larger and thicker FSS, it is easy to fabricate and assemble the proposed FSS with a less stringent tolerance requirement.

![Fig. 8. Variation of resonant frequencies and bandwidths with respect to \( b \) and \( l \) (\( l_{\text{m}} = 3.5 \text{ mm}, w = 4 \text{ mm}, D = 1.6 \text{ mm}, \varepsilon_r = 3.0, d = 1.524 \text{ mm}, \) TE incidence, \( \theta = 0^\circ \)).](image)
C. Design Example

A prototype of the proposed bandpass FSS with a center frequency of 7.5 GHz is designed. The transmission-pole and transmission-zero frequencies are designed to be \( f_{p1} = 7.1, f_{p2} = 7.9, f_{z1} = 10.5 \), and \( f_{z2} = 15.3 \) GHz, respectively. In this design, all the DSPSLS are printed on a substrate of Rogers 4230 \( (\varepsilon_r = 3.0, d = 1.524 \, \text{mm}, \tan\delta = 0.0023) \). The thickness of the copper on both sides of the substrate is 1 oz. The dimensions of the FSS unit cell are as follows: \( b = 8.0 \, \text{mm}, w = 4.0 \, \text{mm}, l = 12 \, \text{mm}, l_3 = 4.65 \, \text{mm}, l_4 = 3.05 \, \text{mm}, D = 1.7 \, \text{mm} \).

![Graph](image)

Fig. 9. Simulated results of the designed FSS under oblique incidence angles for (a) TE polarization and (b) TM polarization.

Fig. 9 illustrates the simulated S-parameter results of the designed FSS for both TE and TM polarizations under various incident angles. It is observed that under TE polarization, the ripple in the passband increases with the growing incident angle, which is mainly attributed to the decrease of the wave impedance of the incidence wave seen at the striking interface for a large incident angle [1]. RBW_{3dB} decreases from 22.9% to 19.7% when the incident angle increases from 0° to 60°. Meanwhile, the transmission zero at \( f_{z1} \) is very stable when the incident angle varies. However, the frequency \( f_{z2} \) reduces with the increasing incident angle, thus leading to a reduced bandwidth for the out-of-band rejection. When the incident angle increases from 0 to 60°, RBW_{20dB} varies from 86.3% to 51.1%. Under TM polarization, the passband ripple decreases slightly as the incident angle increases. This is mainly because the wave impedance of the incidence wave increases a little in this case [1]. RBW_{3dB} varies from 22.9% to 21.4% when the incident angle increases from 0° to 60°. Although the transmission zeros and the bandwidth of the stopband are stable, a parasitic peak (around 12 GHz) in the stopband occurs under oblique incidence, as shown in Fig. 9(b). This peak can be attributed to the coupling of two closely placed vertical and horizontal strips because the E-field vectors have longitudinal components under TM polarization and oblique incidence.

![Graph](image)

Fig. 10. Simulated results of the proposed FSS with loading stubs under oblique incidence angles for (a) TE polarization and (b) TM polarization.

In this design, two small stubs with the same dimensions are introduced to connect the vertical and horizontal strips to remove these peaks under TM polarization and oblique incidence. Fig. 10 shows the simulated S-parameter results of the modified FSS with two loading stubs. It is seen that the filtering responses under TE polarization remain unchanged with and without loading stubs. Under TM polarization and oblique incidence, the filtering responses change slightly except that these peaks in the stopband disappear. The location and line width of this stub can be optimized and its effect may be carefully accounted for. In this design, the length and width of each stub are optimized to be 1.2 mm and 0.2 mm respectively, as shown in Fig. 11.

V. IMPLEMENTATION AND MEASURED RESULTS

A. Implementation

Fig. 11 shows the implementation and assembly details of a
unit cell in our design. The DSPSLs are printed on the Rogers 4230 substrate and are cut into 2-D pieces, as shown in Fig. 12(b). These horizontal and vertical 2-D pieces are cross-joined together to form the required structure through the slots cut half way along the circuit board. Furthermore, in every unit cell, the subs for cancelling coupling between DSPSLs under TM polarization are in good contact after vertical and horizontal circuit boards cross-inserted. Additionally, the metallic plate for providing transmission zeros is realized using aluminum due to its light weight. As discussed above, the location of the aluminum plate is a very important parameter for resonant frequencies in the stopband and this plate should be fixed tightly. In this realization, this aluminum plate is secured in the unit cell by using four short metallic rods, which go through the via holes in the circuit board and extend into the aluminum plate, as shown in Fig. 11.

**Fig. 11.** Assembly details of pieces forming one unit cell.

**Fig. 12.** Photographs of the fabricated FSS. (a) Perspective view; (b) Top view of the fabricated DSPSLs.

**B. Measured Results**

Fig. 12 shows the photo of our fabricated prototype, which is approximately 185 mm × 193 mm in size and consists of 552 unit cells. Both transmission and reflection coefficients are measured by the free-space method using two horn antennas in an anechoic chamber, similar to the measurement set-up described in [13]. Fig. 13 shows measured and simulated S-parameter results of the fabricated FSS under different (TE and TM) polarizations and incident angles (0°, 20° and 40°). It is observed that the frequency performance of the proposed FSS is very stable under different incident angles. Furthermore, it is seen that the fabricated FSS can successfully provide a passband with two transmission poles at 7.4 and 7.8 GHz and a stopband with two transmission zeros close to 10 and 14.6 GHz under normal incidence.

![Fig. 13. Measured results of the fabricated FSS under oblique incidence for (a) TE polarization and (b) TM polarization.](image)

It is observed that measured transmission zero and pole frequencies have small shifts compared to their simulated ones. Such discrepancies between measured and simulated results are attributed to assembly errors. Due to these errors, the measured RBW_{3dB} is 18.4%, which is slightly narrower than the simulated one (22.9%). The measured RBW_{3dB} is 78.4% under normal incidence. In addition, the measured insertion loss at the center frequency of the passband is 0.72 dB under the normal incidence, which is greater than the simulated one (0.13 dB). A larger measured insertion loss may be due to the surface
The wideband rejection has been explained through equivalent circuit models and the designed FSS has been successfully verified by measured results.

VI. CONCLUSION

This paper has presented a new 3-D dual-polarized frequency-selective structure with wide stopband response. The good bandpass response has been realized through substrate modes, which propagate through a 2-D periodic array of double-sided parallel-strip lines. The wideband rejection has been provided by the resonances of air modes due to the inserted metallic plate. The operating principle of the proposed FSS has been explained through equivalent circuit models and the designed FSS has been successfully verified by measured results.

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Bo Li was born in Hunan, China, in 1984. He received the B.S. and Ph.D. degrees both in communication engineering from Nanjing University of Science and Technology, China, in 2006 and 2011, respectively.

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