ORIGINAL RESEARCH PAPER

Novel 4-bit second-order multifunction frequency selective surface

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Abstract
An innovative four-bit second order multifunction active frequency selective structure controlled via four PIN diodes with nine different combinations is introduced. The proposed design is equipped with electromagnetic switching, frequency switching and polarization selection at two passbands. The coding states 0000 and 1111 offer transmission to the passbands centred at 1.96 and 4.98 GHz frequencies and shield to the bands centred at 4.98 and 1.96 GHz for transverse electric (TE)/transverse magnetic (TM) waves, respectively. The coding state 1010 offers transmission to the TE and TM waves at 1.96 GHz and 4.98 GHz, respectively and shields the TM and TE waves at 1.96 GHz and 4.98 GHz, respectively. Contrarily, the coding state 0101 offers transmission to the TE and TM waves at 4.98 GHz and 1.96 GHz, respectively and shields the TM and TE waves at 1.96 GHz and 4.98 GHz, respectively. The coding states 1000/0100 and 1101/1110 allow transmission to the TE wave at 1.96 and 4.98 GHz frequencies, respectively and shields to both the passbands at TM wave. Conversely, the coding states 0001/0100 and 1011/0111 allow transmission to the TM wave at 1.96 and 4.98 GHz frequencies, respectively and shields to both the passbands at TE wave. The coding state 1100/0011 provides shield to all the frequencies below 7.7 GHz. The designed structure is validated through the measured response of the fabricated prototype.

1 | INTRODUCTION

Frequency selective surfaces (FSSs), with vast applications in the field of microwave engineering, known as spatial filters are constructed by an assembly of different resonant and non-resonant elements organized on a lossy substrate [1]. These spatial filters, as a shield, are applied to secure sensitive equipment, biological safety from excessive RF/microwave radiation exposures, radar cross section (RCS) enhancement, RCS reduction techniques, absorbing surfaces for stealth augmentation, satellites equipped with multiple antennas and multifunction systems in modern military systems [2–15]. There are numerous microwave radiating sources which can be harmful to humans and important industrial equipment [16, 17]. FSS can filter out the unwanted dangerous radiation for safety of biological beings and valuable equipment while becomes transparent to the desired frequency of use in the closed environments [18, 19]. Frequency response with spectral changes can be transformed from one filter characteristic to other by selecting active frequency selective spatial filters. Mechanical, liquid/fluid based and electronically manipulable FSS replace the passive FSS for multifunction and complex communication systems as active filters. Also, an electronically reconfigurable meta-material FSS, with comparative advantages of low price, high switching speed and easily installable due to its low weight, can be tuned using variable capacitance diodes or switched between functions of transmission, absorption and shielding using PIN diodes [20–36].

The frequency, angle and polarization dependence make the spatial filter more complex than guided wave filters. The challenge for an active filter is to be equipped with electromagnetic switching between the transmission, reflection and absorption with good selectivity, higher roll off, shielding with better rejection, polarization selection, angle insensitivity with polarization independent behavior for single and multiple frequencies of operation. A cascade of single order structures separated by dielectric spacers leads to maximal flat

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744 | IET Microw. Antennas Propag. 2021;15:744–753. Wileyonlinelibrary.com/journal/mia2
Butterworth or rippled Chebyshev frequency response. The dielectric spacers act as impedance inverters when thickness of the dielectric is selected close to quarter wavelengths [37]. Second-order, single passband active FSS with tuning characteristics using three metallic layers arranged on two substrates has been introduced in [35]. First-order multifunction (polarization selection and electromagnetic switching) FSSs were contributed into literature by Li et al. in [20, 21]. A multifunction FSS switchable between reflection and absorption with wide band and polarization independent property has been proposed in [22]. A multifunction reconfigurable FSS equipped with switching between transmission, reflection and absorption for a single polarized plane wave was added into research in [24]. Ghosh et al. introduced fluid guided multifunction FSSs [26, 27] which cannot switch quickly from one state to other state. A multifunction FSS loaded with switching between transmission and absorption characteristics holding additional tuning quality was proposed in [25], which used four PIN diodes and four varactor diodes. The designs proposed in [20–27] are first-order, two bit guided and have single band filtering property.

The design introduced by the authors possesses the originality of 4-bit selection with 9 independently controllable different coding states (CS). This proposed research comprises multifunction response with second order filtering performance at two passbands that has not been proposed in the literature till date. This design is also capable of shielding for single and dual band frequencies with independent control of CS. The angle insensitivity can be observed from 0° to 45° with no undesired resonances in the whole range of passband frequencies. This article is presented under different sections with brief descriptions and is as follows: The design structure is explained in Section 2. The Section 2 also throws light on the simulation results. The details of measurement equipment and measured results along with the simulation results are described in Section 3. The conclusion is made in the final Section 4.

### 2 | THE DESIGN STRUCTURE AND DISCUSSION ON SIMULATION RESULTS

Top, bottom and side view of the unit structure is shown in Figure 1. Four copper layers are arranged on front and rear sides of the two substrates separated by an air gap $S_p$. Both the lossy substrates (F4BMK) have same thickness equal to $S_T$ and relative dielectric constant $\varepsilon_r = 2.2$. Identical top metal layers on both substrates are organized in orthogonal pattern to identical bottom metal layers. This arrangement of the structure leads to the second-order response of the filter. Optimized geometric dimensional lengths of the filter are summarized in the Table 1. Four coding bits, assigned to biasing of the PIN diodes in different layers of the second order active frequency selective surface (SAFSS), guide the multi behavior response. The $DB_{S1}$ and $DT_{S1}$ are two bits assigned to the bottom and top diodes of upper substrate, respectively. Similarly, $DB_{S2}$ and $DT_{S2}$ are the bits assigned to the bottom and top diodes of upper substrate.

![Figure 1](image)

**FIGURE 1** Unit view of the proposed SAFSS: (a) top view; (b) bottom view and (c) side view

| Parameter | $P$ | $C_F$ | $W_a$ | $W_b$ | $W_p$ | $S_F$ | $S_T$ |
|-----------|-----|-------|-------|-------|-------|-------|-------|
| Value (mm)| 16  | 1.0   | 0.1   | 0.1   | 8     | 12    | 0.5   |
lower substrate separated by air thickness \( s_p \), respectively. The binary bit 1 means 1.8 V DC power supplied to whole single layer and binary bit 0 means that no voltage is supplied. The CS designation to biasing of the PIN diodes embedded in the four metallic layers is summarized in Table 2. CST Microwave Studio, which is a commercially available simulation solver, is opted to simulate the SAFSS. The proposed filtering structure is integrated by sub wavelength unit cells. Equivalent circuit composed of lumped components is used as a tool to understand the role of passive and active elements used in the spatial filters under normal incidence of the plane wave. The inductor model represents microstrip dipole elements when aligned parallel to the electric field and maximum currents are distributed in the same direction at the resonant frequency. Capacitor is modeled at gap between the microstrip dipoles oriented vertical to the electric field. Series LC resonator contributes a transmission zero and a transmission pole is oriented vertical to the electric field. Parallel combinations of the inductor and capacitor when electric field of the plane wave is in the \( y \)-direction. Bottom microstrip dipole elements show inductive behavior to the same plane wave with electric field aligned along the \( y \)-direction. A small resistance \( R_{\text{on}} = 2 \, \Omega \) is identical to the ON state of the PIN diode. The PIN diode in the OFF state is modeled by the parallel combination of high resistance \( R_{\text{OFF}} = 5000 \, \Omega \) and a capacitance \( C_d = 0.17 \, \text{pF} \) [38, 39]. The top layer is identical to a series LC resonator which contributes a transmission zero. The bottom layer furnishes inductive behavior. A passband is achieved by combined effect of top series LC resonator and bottom inductive patch elements.

In the off state of the PIN diodes, the total capacitance of the structure is equal to the parallel combination of the diode capacitance and the gap capacitance between microstrip dipole elements. In this state, a passband is achieved at lower frequency of 1.96 GHz. The pole \( f_{p1} \) and zero \( f_{01} \) frequencies for OFF state of all the PIN diodes can be approximated by:

\[
f_{01} = \frac{1}{2\pi\sqrt{L_T(C_g + C_d)}}
\]

where \( L_T \) is the inductances of top microstrip dipole elements. \( L_B \) denotes the inductance of the bottom inductive bias lines. \( C_g \) is the gap capacitance between top layer microstrip dipoles and \( C_d \) is the capacitance of the PIN diode in the OFF state. The equivalent capacitance and inductance values of the microstrip dipole elements can be estimated from the equations in [40, 41].

Second-order Butterworth or Chebyshev filtering response of the equivalent circuit can be followed by the methods proposed in [41–44]. The capacitive effect in the ON state of the PIN diode, contributed only by the capacitive gap between the microstrip dipole elements, is reduced. The passband shifts to 4.98 GHz frequency due to the reduced capacitance between the inductive microstrip elements. The pole \( f_{p2} \) and zero \( f_{02} \) frequencies of ON state of all the PIN diodes can be approximated by:

\[
f_{02} = \frac{1}{2\pi\sqrt{L_T(C_g)}}
\]

\[
f_{p2} = \frac{1}{2\pi\sqrt{(L_T + L_B)(C_g)}}
\]

No bias to the top layer PIN diodes, in a single structure, is responsible for the TE wave selection and no bias to the bottom PIN diodes in the selection of the TM wave to pass through the structure at 1.96 GHz frequency. Conversely, it is true for the ON state of the PIN diodes at 4.98 GHz frequency. The simulated frequency response for 9 CS at TE and TM modes of obliquely incident plane wave are shown in Figures 2–4. Figure 2 shows the transmission coefficient at four different CS 0000, 1010, 0101 and 1111. CS 0000 is transparent to dual polarized plane wave at the center frequency of 1.96 GHz and −10 dB fractional bandwidth (FBW) of this band is 44%. CS 1111 is transparent to the dual polarized plane wave at centre frequency of 4.98 GHz with −10 dB FBW of 15%. These two CS validate electromagnetic response inversion and frequency switching between two bands with the advantage of shielding to other than the passband frequencies. The CS 1010 allows the transmission to the TE wave at 1.96 GHz, whereas the CS 0101 allows transmission to the TM wave at 4.98 GHz frequency. Alternatively, the CS 0101 allows transmission to the TM wave at 1.96 GHz, whereas the CS 1010 allows transmission at 4.98 GHz to the TM wave. These two CS 1010 and 0101 verify the polarization selection at two bands with alternate CS. It is observed that transparency of one mode at a resonant frequency is opaque to the other mode at same frequency. In CS 1010, both the top layers are with no bias to PIN diodes and select the TE wave to pass through the structure at 1.96 GHz, and the forward bias to both the bottom layer PIN diodes permit the TM wave to pass thorough the structure at 4.98 GHz frequency.

| Coding state | DBg2 | DTSb | DBs | DTs |
|--------------|------|------|-----|-----|
| 0000         | OFF  | OFF  | OFF | OFF |
| 0001         | OFF  | OFF  | OFF | ON  |
| 0011         | OFF  | OFF  | ON  | OFF |
| 1000         | ON   | OFF  | OFF | OFF |
| 1010         | ON   | OFF  | ON  | OFF |
| 1011         | ON   | OFF  | ON  | OFF |
| 1100         | ON   | ON   | OFF | ON  |
| 1101         | ON   | ON   | OFF | ON  |
| 1111         | ON   | ON   | ON  | ON  |
Figure 2 depicts the simulated response for next four CS 1000, 0001, 1101 and 1011. The CS 1000/0010 and 1101/1110 allow transmission to the TE wave at 1.96 GHz and 4.98 GHz frequencies, respectively and act as a shield to TM mode at all the frequencies. On the contrary, the CS 0100/0001 and 1011/0111 allow transmission to the TM wave at 1.96 and 4.98 GHz frequencies, respectively and act as a shield to all the frequencies at TE wave incidence. These four CS are also responsible for polarization selection at two passband frequencies of 1.96 and 4.98 GHz with difference of shielding from CS 1010 and 0101 for entire frequency range of 1–7.7 GHz.

In terms of filtering response, the four identical CS 1100, 0011, 1001, and 0110 provide shield to all the frequencies below 7.7 GHz as depicted in Figure 3. At the CS 1100, the top layer structure is transparent to dual polarized plane wave at frequency of 1.96 GHz and opaque to frequency of 4.98 GHz, whereas the bottom layer structure is opaque to the dual polarized plane wave at 1.96 GHz and transparent to 4.98 GHz. As a result, the wave transmitted through the one-layer structure at a frequency is reflected by the opaque behavior of the other layer structure. The two weak resonances at 1.96 and 4.98 GHz occur simultaneously, which can be observed in Figure 4. Converse is true for CS 0011 with same resultant response as that of CS 1100. It is noted that the shielding to all frequencies is less than –20 dB for CS 1100 and 0011 at

**FIGURE 2** Simulated transmission coefficient for different CS at oblique incidence of dual polarized plane wave: (a) 0000; (b) 1010; (c) 0101 and (d) 1111

**FIGURE 3** Simulated transmission coefficient for different CS at oblique incidence of dual polarized plane wave: (a) 1000; (b) 0001; (c) 1101 and (d) 1011

**FIGURE 4** Simulated transmission coefficient for CS 1100 at oblique incidence of dual polarized plane wave
normal incidence of the plane wave. The second-order filtering response has also higher roll off, better selection at passband frequencies and very good shielding to other than the passband frequencies. The oblique angle incidence in the simulation results is observed to be angle insensitive with no spurious and unwanted reflections in the entire frequency range of 1 ~ 7.7 GHz. However, the bandwidth variation for two passbands is noted differently at two modes on an obliquely incident plane wave. This dependency of bandwidth variation is related to the loaded quality factor of the resonators which changes distinctively with wave impedance at increasing incident angles of plane wave [26, 45].

Table 3 presents functionality of all the differently working CS. Distinguished feature of the second order multifunction structure presented in this research opens the new domain of higher bit coding for higher order multifunctional filtering response.

The design is further modified by adding inductive lines of width \( W_s = 0.1 \) mm as shown in Figure 5. This modified basic structure was proposed in [46]. These inductive lines form a series LC resonator together with the bias lines of width \( W_b \). This LC resonator adds a transmission zero at a higher frequency of around 8 GHz. However, insertion loss reduces with the addition of these inductive lines. The simulation results for the final design structure of Figure 5, at air separation distance \( SP = 12 \) mm and inductive line width \( Wa = 0.1 \) mm, are shown in Figures 6-8. The effect of variation in the width of these inductive lines for three CS is shown in Figure 9. It is observed that the resonance introduced by inductive lines shifts to lower frequencies due to increased capacitance with increase in the width \( W_s \). The resonant frequency of 4.98 GHz also shifts toward lower frequencies with increase in \( W_s \). However, at \( W_s = 3 \) mm, the two resonances split in different direction due to increased mutual coupling between the microstrip dipole elements of width \( W_p \) and inductive lines of width \( W_s \). With increase in \( W_s \) from 0.1 to 3 mm, the higher frequency band shifts from 4.98 to 4.1 GHz and reduction in insertion loss is also observed.

TABLE 3 The summarized responses at different working CS in terms of functionality

| Binary coding state | Frequency (GHz) | TE   | TM   |
|---------------------|-----------------|------|------|
| 0000                | 2               | Pass | Pass |
| 0000                | 5               | Shield | Shield |
| 1010                | 2               | Pass | Shield |
| 1010                | 5               | Shield | Pass |
| 0101                | 2               | Shield | Pass |
| 0101                | 5               | Pass | Shield |
| 1111                | 5               | Shield | Shield |
| 1111                | 2               | pass | pass |
| 0001                | 2               | All shield | Pass |
| 1000                | 2               | Pass | All shield |
| 1011                | 5               | All shield | Pass |
| 1101                | 5               | Pass | All shield |
| 1100                | –               | All shield | All shield |
Two resonances around 1.96 GHz correspond to two resonating structures separated by an impedance inverter. Two resonances can be governed by air separation thickness $S_p$ to achieve maximally flat or rippled response [37]. Effect of change in air gap between the two layers is shown in Figure 10. It is observed that the passband ripple at 1.96 GHz reduces with increase in air gap and becomes approximately flat at $S_p = 20$ mm. However, the insertion loss increases at the 4.98 GHz band with an increase in the air gap. Optimum air separation distance $S_p = 12$ mm and inductive line width $W_a = 0.1$ mm, for the design structure of Figure 5, are selected in simulation results depicted in Figures 6–8. The insertion loss in the ripple at the 1.96 GHz band varies from 2.7 to 1.1 dB, whereas the insertion loss of 3.5 dB is observed at the 4.98 GHz frequency band.

3 | PROTOTYPE FABRICATION AND MEASUREMENT VALIDATION OF THE DESIGN

The two identical fabricated panels of the two-layer SAFSS incorporating 17 × 17 unit cells has total dimensions of 272 mm × 272 mm.

Top view of one full structure is shown in Figure 11(a). Parallel biasing topology used in this design for 4 × 4 cells is shown in Figure 11(b). The positive and negative bias lines are added on each metallic layer to feed DC supply to the PIN diodes for switching purpose. The nylon dielectric screws, with 4 mm diameter, are used to hold the two structures together. Expanded polystyrene (EPS) foam of thickness 8.0 mm as spacer is used at borders between the two layers. The EPS foam also provides support to minimize curvature in the structures. The dielectric spacers have almost no effect on the frequency response of the structure. The PIN diodes BAR64-02V manufactured by Infineon technology are soldered manually between the metallic microstrip dipoles. The pyramidal absorber frame with aperture (for SAFSS embedding) in
the centre is used for absorbing the interfering reflections from multiple sides and maximum transmission through the prototype. Two dual power supplies are used to independently bias the PIN diodes used in the four metallic layers. The effect of biasing wires is minimized by routing them on back side of the absorber frame. Two horn antennas operating from 0.8 GHz to 6 GHz are connected to Vector Network Analyzer N5245A. The SAFSS is fixed in the centre between the two horn antennas with sufficient distance to justify the plane wave criteria. Figure 11(c) is presented to show the front and side view of the measurement setup. AC and DC isolation is achieved using inductors of 47 nH in the track of biasing lines. Firstly, the response is measured at each angle from $0^\circ$ to $45^\circ$ of the dual polarized plane wave with SAFSS embedded in the frame. Secondly, the response is measured without the SAFSS at each step of the measurement. Then actual calibrated response of the fabricated prototype is achieved by subtracting the results of second step from the first ones. The frequency response measured for seven CS is shown in Figures 12 and 13. The oblique angle insensitivity can also be observed in the measured frequency response. The measured and simulated results resemble to prove the validity of design structure with some inconsistencies. The simulated insertion loss at $S_p = 8$ mm around 1.96 GHz resonance varies from 4.6 to 1.7 dB and measured insertion loss at this resonance peak is 2.3 dB. The simulated insertion loss at the 4.98 GHz band is 3.5 dB, whereas the measured insertion loss of 2.1 dB is observed at this resonance peak.

A frequency shift in the measure results is observed from Figures 12 and 13. Drift in dielectric parameters such as permittivity and thickness of the fabricated prototype may be the reasons for shift in the frequency. The parasitic effects of the PIN diodes and soldering errors add another reason to differences between the measured and simulated response. The measured bandwidth is observed to be wider at higher frequency band of 4.98 GHz due to reduced quality factor of the fabricated prototype [47]. The diffraction phenomena from edges of the

**Figure 10** Simulated transmission coefficient at variation of the air gap $S_p$ for different CS. (a) 0000; (b) 1111; and (c) 0011

**Figure 11** Full structure top view with biasing scheme for 4x4 unit cells of SAFSS structure and measurement testing setup: (a) full structure; (b) biasing methodology; and (c) test setup
finite fabricated structure cause mismatch as the design structure is simulated with infinite size. The curvature effects, due to flexibility of the two-layer structure, have been tried to minimize using the EPS foam, however, these effects still exist and cause misalignment between the two layers which cannot be ignored in the measured response. In spite of these tolerances in parameters of the measurement process, the proposed innovative idea is corroborated by the two-layer fabricated panel with the second-order multifunction filtering characteristic.

4 | CONCLUSION

Here, a 4-bit second-order structure with intelligently controlled multi-functional frequency response at two bands has been proposed and verified by the measured prototype. TE/TM wave selection, frequency switching and inversion between transmissions/shieldings have been discussed for various CS. The SAFSS is resonant at two passbands of 1.96 and 4.98 GHz frequencies with fractional bandwidths equal to 45% and 15% calculated at −10 dB, respectively. For the lower resonant frequency of 1.96 GHz, the structure is miniaturized to λ/10, and λ/4 for the higher resonant frequency of 4.98 GHz. This miniaturization leads stability to the plane wave oblique angles from 0° to 45°. The parallel biasing concept used in this design requires only 1 ~ 2 V DC supply to the whole structure. This innovative design can open a gateway for complex communication systems used in military and satellites with multifunction requirements. Sensitive and important devices can also be safeguarded from high power microwave radiation using the shielding CS of the design [48]. Distinctive property of transmission at the desired frequency band with better shielding to other frequencies can be used in the industrial, scientific and medical fields.

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