Multifunctional Dual-Band Energy Selective Surface Based on Two-Layer Diode-Loaded FSS

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October 31, 2023

Abstract

This article presents a multifunctional dual-band energy selective surface (ESS) based on a two-layer resonant circuit model with high-power microwave (HPM) protection ability for two bands. An equivalent circuit model (ECM) engineering multiple stopbands and passbands with combinations of parallel and serial resonators is proposed for dual-band design. By loading diodes in the resonator accordingly, the operating band can be switched from low insertion loss (IL) state to high shielding effectiveness (SE) state power-dependently. Additionally, the independence of two bands is greatly improved by utilizing two separate layers to mount the diodes, attributing to the enhanced difference of induced voltage between diodes in different layers resulting from shielding effect of top layer. Various functions in two-frequency co-incidence scenarios are analyzed, demonstrating the advantage of independent modulation. Based on the ECM, a two-layer ESS is designed. Two wide bands with IL<1dB under low-power incidence and SE>20dB under HPM are achieved, due to second-order resonance produced by double-layer configuration. The multiple nonlinear functions of the design are verified by field-circuit co-simulations. Several prototypes are measured, exhibiting 0.6&25.2 dB and 0.7&27.3 dB of IL&SE at two central operating frequencies for low-power and high-power incidences, respectively, with good agreement with the simulations.
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Index Terms—energy selective surface (ESS), adaptive shielding, frequency selective surface (FSS), dual-band.

I. INTRODUCTION

Facing the complex electromagnetic (EM) environment, radar and communication systems require special EM protection cover to prevent damage caused by high-power microwave (HPM) [1], [2]. Some conventional EM protection methods, for instance, limiters [3], isolators [4], absorbers [5], [6], have been widely used to shield HPM. However, these techniques have limitations because they usually block all the EM waves indistinguishably whether it is harmful HPM or useful communication signal in the other band. Therefore, some functional devices, such as frequency-selective surfaces (FSSs) [7]–[9], frequency-selective rasorbers (FSRs) [10]–[15], have been proposed to shield the HMP in the stopband and transmit the communication signal in the passband simultaneously. In recent years, to enable the FSS to switch from transmission-state to shielding-state in the same band power-adaptively, a new concept of energy selective surface (ESS) with selfactuated protection ability was proposed and attracted intense interest [16]–[34]. The power-dependent designs are usually realized by integrating ESS with nonlinear elements or special materials, such as diodes [16]–[32], phase-change materials [33], and film materials [34]. For low-power signal, the nonlinear components are OFF, ensuring high transmittance. In contrast, under HPM incidence, the nonlinear devices are triggered on due to strong fields, shielding the wave adaptively.

The insertion loss (IL) and shielding effectiveness (SE) represent transmissions under low-power and high-power incidences, respectively. Low IL is highly demanded for communication, and high SE is crucial for HPM protection. Essentially, previous ESS designs can be classified as several categories based on the frequency responses. The first is low-pass/high-pass type [16], [17]. The most of diode-loaded grid ESSs are operating in this mode, which generate low-pass and high-pass responses according to power-dependent switching of diodes. However, it is difficult to obtain extremely low IL and high SE simultaneously due to lack of impedance zero and pole. Later, by introducing bandpass and high-pass responses, another type of ESS is implemented by mounting diodes into square slot bandpass FSS [18]–[20]. This configuration can achieve very low IL at the resonance frequency of bandpass FSS under low-power incidence, but the SE is still limited when the diode is on under HPM. Recently, Zhou et al. proposed a double-resonance method to realize power-dependent switching between bandpass parallel LC bandpass circuit and serial LC band-stop circuit, obtaining low IL (0.02 dB) and high SE (35 dB) at the same resonance frequency [21]. However, all the ESSs mentioned above are operating in only single band. In complicated scenarios, ESSs with multiple operating bands are desired, especially when the frequency band of HPM from aggressor is different from the communication band.

Two methods for designing multiband ESS is proposed in [22], including multiple parallel LC circuits and cascading serial LC circuits. By engineering the multiple resonance bands and loading nonlinear diodes in the LC resonators, the transmission zeros and poles are controlled power-dependently. A dual-band design based on two parallel LC circuits has been designed and measured, showing 0.5 dB IL and 30 dB SE in both bands. Nevertheless, in the experiments, the two operating bands were not modulated independently as expected. It is mainly because the diodes in different resonators are activated simultaneously, since the difference of induced voltage in two resonators is much lower than the required power variation (20 dB) for the diodes switching from OFF to ON. Additionally, the SEs at two operating frequencies deteriorate as 12 dB for high-power incidence, originating from only one impedance zero when all the diodes are on. Therefore, ESS with independent modulated multibands and high SE in both bands is still challenging.

Manuscript received xxx, xx, 2023; revised xxx, xx, 2023. This work was supported by Natural Science Foundation of Zhejiang Province under Grant LY22F010001, National Natural Science Foundation of China under Grant 61971250, Natural Science Foundation of Ningbo (2022J098, 202003N4109), and the Fundamental Research Funds for the Provincial Universities of Zhejiang. (Corresponding author: Gaoming Xu, Yi Chen)
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II. DUAL-BAND ESS ANALYSIS AND DESIGN

For a one-layer ESS, the two important qualities IL and SE can be calculated as [29]

\[ IL = SE = \left| S_1 \right| = \left| 20 \log \left( \frac{Z_{\text{ESS}}}{Z_0} \right) \right| = \left| 20 \log \left( 1 + \frac{Z_0}{2Z_{\text{ESS}}} \right) \right| (\text{dB}) \]  

(1)

where \( Z_0 \) is the wave impedance in free space and \( Z_{\text{ESS}} \) is the surface impedance of ESS. ESS demands distinct transmission characteristics for low-power and high-power signals, i.e. low IL for low-power signals and high SE for high-power signals within the same operating band. Therefore, \( Z_{\text{ESS}} \) should be infinite value at the working frequency to obtain \( 0 \) dB IL for low-power signal. Whereas, for HPM incidence, the surface impedance \( Z_{\text{ESS}} \) should be switched into near-zero value to obtain large SE, providing adaptive protection. Here we introduce diodes into FSS to realize the power-dependent responses. The diodes are in turned-OFF state at low incident power level, which can be modeled as a capacitor. For HPM, the diodes will be triggered on, and can be regarded as a small resistance.

A. Circuit Model of One-layer Dual-Band ESS

As demonstrated in [35], parallel and series LC (PLC and SLC) resonators are two basic circuits to create transmission poles (passband) and zeros (stopband) at the resonance frequency of \( 1/2\pi\sqrt{LC} \), since the impedance of PLC and SLC are infinite and zero at resonance frequency, respectively. In addition, the impedance of PLC outside the resonance becomes much smaller, and the impedance of SLC outside the resonance becomes much larger. To implement ESS with two passbands, two impedance poles and three impedance zeros are required.

To address these issues, we propose a multifunctional dual-band ESS based on two-layer diode-loaded FSS. In Section II, an equivalent circuit model (ECM) approach to engineer the transmission poles and zeros for dual-band ESS design is presented. By connecting a serial LC resonator, a parallel LC resonator and a diode-loaded parallel LC resonator in series on one layer, the ESS achieve two transmission poles with three adjacent transmission zeros at low-power scenario and convert into circuit with one transmission pole and two adjacent transmission zeros at high-power scenario, ensuring high SE for the switchable band. In addition, by cascading a similar combination of resonators with power-dependent switchability for the other band on separate layer, the independence of the two operating bands is improved remarkably due to giant enhancement of the induced voltage difference between two diodes owing to the shielding effect of top layer. Then four additional operating modes in two-frequency co-incidence scenarios are analyzed attributing to the independent operating bands. Based on the ECM, a dual-band ESS is realized with IL less than 1 dB in two wide bands (17.5\% and 22.4\% bandwidth), and SE greater than 45 dB at the central frequencies of both bands. Section III gives the nonlinear simulations results for both single-frequency incidence and two-frequency co-incidence obtained from field-circuit co-simulations, verifying the versatility of the ESS. Three prototypes of the ESS are fabricated and measured using parallel plate waveguide (PPW) and rectangular waveguide setups in Section IV. Conclusions are given in Section V.
incidence since the diode is equivalent to a resistor $R_d$. The impedance of the circuit can be calculated as
\[
Z = j\omega L_1 + \frac{1}{j\omega C_1} + \frac{1}{R_d} + \frac{1}{j\omega C_{a1}} + \frac{1}{j\omega L_3}.
\] (3)

Since the resistance $R_d$ is usually smaller than the impedance of $L_{a1}$ by three orders in the band of interest, for simplicity of theoretical analysis, the resistance is negligible [21], [22] and the inductance $L_{a1}$ can be regarded as being shorted. Therefore, the total reactance can be calculated as summation of the reactance of serial $L_{11}C_{1}$ resonator and parallel $L_{b1}C_{b1}$ resonator. It can also be observed from Fig. 2 (b) that there is still an impedance pole at the resonance frequency $f_o$, corresponding to a transmission peak, and two impedance zeros $f_1$ and $f_2$ corresponding to two stopbands are synthetized. In this way, the passband $f_0$ is closed because the parallel resonators $L_{a1}C_{1}$ is eliminated when the diode is turned on by strong field intensity, whereas the other passband $f_0$ remains unaffected.

It is important to note that although the two transmission poles are not affected by the values of $L_1$ and $C_1$ in series, the impedance zero $f_2$ is affected by serial $L_{11}C_{1}$ resonator. Therefore, the values of $L_1$ and $C_1$ need to be designed carefully, otherwise, transmission zero $f_2$ may fall far away from the frequency $f_3$, resulting in poor SE for HPM at the frequency $f_0$.

Analogously, by employing nonlinear diode in the second parallel resonator with resonance frequency of $f_0$, the passband $f_0$ can also be controlled by the incident field intensity independently. However, in real experiment, if the two diodes to control the passbands $f_0$ and $f_0$ are positioned in the same layer, the diodes will turn on simultaneously when they are exposed to high power fields [22]. Thus, the independent modulation of the two working bands will be challenging in one-layer configuration. Moreover, the elimination of both two PLC resonators will lead to only one impedance/transmission zero at $f_c$, resulting in decrease of SE at the two switchable passbands $f_0$ and $f_0$, as demonstrated in [22].

**B. Mechanism of Two-layer ESS with Better Independence of Two Operating Bands**

To address the issues above, we propose a double-layer dual-band ESS with improved independence of the two bands, as illustrated in Fig. 3(a). The circuit model includes two diodes, $D_1$ and $D_2$, located in the first and second layers, controlling the switch of the passbands $f_0$ and $f_0$, respectively. Under low power incidence, the parallel $L_{a1}D_1$ and $L_{a2}D_2$ resonators have the same resonant frequency $f_0$, while the $L_{b1}C_{b1}$ and $L_{b2}C_{b2}$ resonators have the same resonant frequency $f_0$. Thus, the two cascaded surfaces exhibit two transmission bands at $f_0$ and $f_0$ for low power signals. For high-power incident scenarios, the two-layer ESS exhibits four different operating modes dependent on the frequency and direction of incident HPM due to asymmetry of the nonlinear-surfaces combination, as shown in Fig. 3(b). When HPM at $f_0$ illuminates the first layer, the diode $D_1$ is turned on, reflecting the high power signal at $f_0$ from the first layer. This prevents the simultaneous turn-on of the diode $D_2$ since the high-intensity fields are shielded by the first layer and cannot reach the diode $D_2$ on the second layer. Consequently, the low-power signal at $f_0$ can pass through the whole two layers due to the resonators $L_{b1}C_{b1}$ and $L_{b2}D_2$, as depicted in case 1 of Fig. 3(b). Similarly, as shown in case 4 of Fig. 3(b), independent control of operating band $f_0$ should also achieve when a high-power signal at frequency $f_0$ illuminates the ESS from the 2nd layer, with the transmission band $f_0$ maintaining open. This functionality with independent control of two bands are of great value in radar radome, which can protect our devices from HPM at foe band and maintain the transmission of friend communication signals at the other operating band simultaneously.

When the frequency of HPM illuminating the ESS from right side is located in the transmission band $f_0$ of the 2nd layer, as shown in case 2 of Fig. 3(b), the high-intensity wave can pass through the layer and activate the diode $D_1$ in the 1st layer. In this way, the diode $D_2$ will be activated simultaneously since strong voltages will be induced in the 2nd layer when the high-power wave passes the 2nd layer. Then all the signals at both two operating bands will be shielded. Similar operating mode occurs in case 3 of Fig. 3(b). In these scenarios, the ESS loses independence of two operating bands, since the ratio between the induced voltages of two diodes is not large enough to ensure
different switching states of each diode. This phenomenon also occurs in the experiment of [22], where the diode-loaded resonators for two operating bands are designed in the same layer. Therein, the difference of induced voltage across two diodes is 6 dB (power difference 12 dB), much lower than the required 20-dB power variation for diode to switch from OFF to ON. Therefore, the other diode will be triggered when one diode is turned on. In our design, a feasible solution for better independence is proposed by enhancing the difference of induced voltage via a two-layer configuration, which works in case 1 and case 4.

It is also worth noting that the values of $C_1 \cdot L_1$ and $C_2 \cdot L_2$ in the two layers need to be optimized to compensate the inductive and capacitive reductance of the resonators $L_{b1} \cdot C_{b1}$ and $L_{a2} \cdot C_{a2}$ at the corresponding operating band, respectively. In this way, when the respective diode is activated, the shielding peak or transmission zero will not fall away from the switchable operating frequency band.

C. Structure Design of the Two-layer Dual-Band ESS

Based on the theoretical analysis, a two-layer dual-band ESS with operating frequencies of 2.8 GHz and 4.75 GHz are designed. The unit cell configuration is depicted in Fig. 4. The ESS consists of two PCB layers separated by an air spacer with thickness of $h=12.5$ mm. The PCB substrate is F4B with permittivity of 2.2, loss tangent of 0.001 and a thickness of 0.5 mm. The patterns on the front surfaces of both two layers are metal strips with different width connected by lumped capacitors and diodes. The width and length of strips with vias and the gap between them are identical in the same layer. Split rings are designed on the bottom surfaces, which are connected with the front surfaces by four metal vias with radius of 0.15 mm. By placing the split rings on the bottom surface, the impacts on other metal structures are reduced, and the geometric parameters of the split rings can be optimized to exhibit the desired inductance values. Two lumped capacitors of 0.3 pF are loaded in the first layer and two capacitors of 0.47 pF are used in the second layer. The diodes used in the ESS are NSR201MX, with an equivalent capacitance $C_d$ of 0.15 pF at low power and an equivalent resistance $R_d$ of 1.5 Ω at high power [36].

The equivalent circuit model (ECM) of the dual-band ESS for TE wave incidence is derived as shown in Fig. 5. $L_{a1}$, $L_{b1}$, and $L_{a2}$, $L_{b2}$ indicate the equivalent inductors of the split rings on the bottom surface of each layer (Fig. 4(d) and Fig. 4(e)), respectively. The metal strips connecting adjacent unit cells are modelled as inductors $L_1$ and $L_2$. The capacitors $C_1$ and $C_2$ are constructed by the strip gap in Fig. 4(b) and the gap loaded with lumped capacitor in Fig. 4(c), respectively. The thin F4B substrates and the air spacer are represented by transmission lines with characteristic admittance $Y_1$ and $Y_0$, respectively. The circuit parameters are initially derived based on the empirical formulas in [37], and then tuned using the particle swarm optimization algorithm to fit the simulation results.

The ILs and SEs of the ESS under normal incidence are simulated and calculated according to the ECM, respectively. As shown in Fig. 6, two operating bands with central frequency of 2.85 GHz and 4.91 GHz are observed. By altering the states of diodes $D_1$ and $D_2$ independently, the lower and higher operating bands can be switched from transmission state to shielding state, respectively. For low-power scenario where all diodes are turned-OFF, the two IL < 1 dB bands cover from 2.60 to 3.10 GHz (bandwidth of 17.5 %) and 4.36 to 5.46 GHz (bandwidth of 22.4 %). The flat and wide passbands are attributed to the second-order resonance in double-layer FSS [21]. For high-power scenario where all diodes are turned-ON, the SE in each band is larger than 14.5 dB and 18 dB, respectively, and the SE reaches 45 dB and 46 dB at 2.8 GHz and 4.75 GHz, respectively. Fig. 6(c) and 6(d) illustrate the independent control of each operating band by setting one diode...
turned-ON. It can be seen that low IL in the transmission band and high SE in the shielding band maintain in these scenarios, revealing good independence of two operating bands. The IL and SE results calculated according to the ECM in Fig. 5 are in excellent agreement with full-wave simulations, verifying the design method.

Since the thickness of the whole ESS is also an important parameter in the design, we conduct a parametric study of the thickness of air spacer $h$ between the two layers. As shown in Fig. 7(a), when the thickness $h$ decreases, the second band-pass resonance in the lower operating band moves towards higher frequency, making the ripple more obvious in the transmission band. Moreover, the SE in the edges of the operating bands becomes worse for a thinner thickness $h$, producing undesired transmission peaks, as shown in Fig. 7(b). Therefore, a distance of $h=12.5$ mm is chosen for obtaining a flat transmission band in the low-power scenario and high SE with large bandwidth in the high-power scenario.

The simulation results of IL and SE for TE polarization under oblique incidences for four different operating modes are presented in Fig. 8. Apparently, the ESS has stable transmission zeros and poles for incident angles up to 60°. However, the ripple in the lower transmission band becomes more obvious for 60° of incident angle. Moreover, the bandwidth of the higher transmission band decreases slightly when the angle increases.

III. NONLINEAR SIMULATIONS FOR SINGLE-FREQUENCY AND TWO-FREQUENCY INCIDENCES IN TWO BANDS

To evaluate the nonlinear performance of ESS under varying incident power, we conduct CST field-circuit co-simulations in this section. As shown in Fig. 9, the co-simulation joints full-wave simulation and circuit simulation automatically. A SPICE model for the diode provided by the manufacturer onsemi [40] is applied. The package of diode contains two parasitic components, namely the inductor of 0.45 nH and the capacitor of 50 fF. The 0.3 pF capacitor is the lumped device in parallel with the diode in the first layer. Both single-frequency incidence and two-frequency co-incidence scenarios are simulated.

A. Single-Frequency Incidence

First, a single-frequency incidence simulation is carried out. The selected incident signal is the sinusoidal signal illuminating the ESS from left side, whose frequency and amplitude can be adjusted. The transmission coefficient of the ESS at different incident frequency versus various incident field intensities can be obtained. The simulation results for five incident frequencies 2.6, 2.80, 3.00, 4.75, and 5.00 GHz in the two operating bands with varying field intensities are shown in Fig. 9(b). For the designed ESS, IL is less than 1 dB when the electric field intensity is less than 20 V/m for all five frequencies. Then the SE increases rapidly with increasing incident power, reaching up to 18 dB, 29.5 dB, 30.1 dB, and 27 dB, with field intensities of 3500 V/m at 2.6, 2.8, 4.75 and 5 GHz. For 3 GHz incidence, the SE reaches a maximum of 20.8 dB with field intensity of 1200 V/m. Generally, when the electric field intensity is greater than 2000 V/m, the SE of the five frequencies is greater than 18 dB. It should be mentioned that due to the asymmetry of the two-layer ESS, the transmissions under opposite illumination.
directions may differ when the diodes are activated with different degrees. However, the difference is tiny (less than 0.14 dB at 2.8 GHz, and less than 0.06 dB at 4.75 GHz according to co-simulations) between the transmissions under HPM from opposite sides. This is because the activation degrees of diodes are approximately the same since the right layer is transparent for the incident wave at the nonlinear operating frequency of the left layer, and the transmission should be identical due to reciprocity.

**B. Two-Frequency Co-incidence**

When two sinusoidal signals illuminate the ESS simultaneously, the ESS should show different responses for each signal, depending on the power spectral density and incident direction. In two-frequency co-incidence simulations, we generate a superposition of two sinusoidal functions with different frequencies in each operating band using Matlab and import it into CST as the incident wave. The field intensity of each frequency component can be set at different levels.

First, when the incident field intensities of both two frequencies are high, the ESS will shield both two high-power signals, as predicted in Fig. 6(b). When the two-frequency sets are (2.60 GHz, 4.36 GHz), (2.8 GHz, 4.75 GHz) and (3.00 GHz, 5.45 GHz) with equal intensity of 2 kV/m, the corresponding SEs obtained from the co-simulations are (18.90 dB, 19.71 dB), (26.56 dB, 27.07 dB) and (20.16 dB, 26.58 dB), respectively, in good agreement with the results in full-wave simulations.

Then, the four incident scenarios in Fig. 3(b) are simulated. The field intensity of low-power sinusoidal signal for communication is maintained at 15 V/m, and the field intensity of the other sinusoidal signal from the aggressor varies from 20 V/m to 3500 V/m, i.e. from low-power to high-power one. As shown in Fig. 10, the frequency set (2.8 GHz, 4.75 GHz) is chosen as an example, and the IL/SE performances for both two frequencies versus different intensity of incident field from the aggressor are simulated. In case 1 [Fig. 10(a)] and case 2 [Fig. 10(b)], the low-power signal with fixed intensity is at 4.75 GHz (higher operating band), and the other incident signal with varied intensity is at 2.8 GHz (lower operating band). In case 3 [Fig. 10(c)] and case 4 [Fig. 10(d)], the fixed low-power signal is at 2.8 GHz, and the varied signal is at 4.75 GHz.

It can be seen from Fig. 10(a) that when the field intensity of aggressor signal (2.8GHz) is lower than 200 V/m, the IL of low-power communication signal is less than 1 dB, while the SE of aggressor signal reaches 13 dB for intensity of 200 V/m. Notably, in case 1, when the intensity of high-power signal increases above 3500 V/m, the IL for the low-power signal at 4.75 GHz stabilizes around 5.2 dB, and the SE for the high-power signal at 2.8 GHz increases up to 29.5 dB. The large 24.3 dB difference of transmission shows good independence of the two operating bands. Nevertheless, when the high-power signal illuminates from the opposite side, as shown in case 2 of Fig. 10(b), the IL of communication signal at 4.75 GHz increases to 16 dB when the intensity of high-power signal increases to 2000 V/m. Only 11 dB difference of transmission for two operating bands means deteriorated independence. Similarly, when the aggressor signal is set at 4.75 GHz, as shown in case 3 [Fig. 10(c)] and case 4 [Fig. 10(d)], the ILs for low-power communication signal at 2.8 GHz are also distinct when the high-power signal illuminates the ESS from different sides. In case 4, with intensity lower than 1200 V/m, IL of the communication signal at 2.8 GHz is less than 5 dB. It will increases up to 10.5 dB when the control signal intensity increases to 3500 V/m. In spite of this, 20 dB difference of transmission for two operating frequencies still exhibits. When the high-power incidence illuminates as case 3, with intensity of 1200 V/m, the IL of the communication signal at 2.8 GHz will be about 15 dB, meaning a worse independence of two operating bands in this scenario.
Fig. 11. The ratio between the induced voltage of diode $D_1$ and $D_2$ at different frequencies in (a) lower operating band and (b) higher operating band. The diodes are set in different states for comparison.

To reveal the mechanism of improved independence of two operating bands by two-layer configuration, the ratios of induced voltage of diodes in different layers at different frequencies are illustrated in Fig. 11. Giant enhancements for the difference of induced voltage between the two diodes are achieved by the strong shielding effect of the activated layer. When the diode $D_1$ is ON, the first layer can shield the incident wave in the lower band, resulting in much weaker EM response at diode $D_2$. As shown in Fig. 11(a), the ratio peak ($V_{D1}/V_{D2}$) at 2.8 GHz is about 32 (power difference 30 dB) when the diode $D_1$ is OFF, which is much larger than the ratio of 4 (power difference 12 dB) when the diode $D_1$ is ON. For the incidence in higher operating band, the peak ratio ($V_{D2}/V_{D1}$) reaches 34 (power difference 30.6 dB) when the diode $D_2$ is OFF. For the incidence in higher operating band, the peak ratio ($V_{D2}/V_{D1}$) reaches 34 (power difference 30.6 dB) when the diode $D_2$ is ON, and is 6 (power difference 15.6 dB) when the diode $D_2$ is OFF. From these results, it is concluded that the prominent shielding effect of separated layer enhances the difference of power level in different diodes dramatically, improving the independence of control of two operating bands.

IV. MEASUREMENT AND VERIFICATIONS

To verify the performance of the multifunctional dual-band ESS design, several prototypes with different size adaptive to each test platform are fabricated. It is worth mentioning that our prototype and waveguide setup only support single frequency incidence experiment. The main reason is that the fundamental $TE_{10}$ mode for the two operating bands are not supported in one type of waveguide simultaneously. Therefore, the two-frequency co-incidence verification is only feasible in free-space test, where ultra-high-power microwave radiating and protection systems are required. Due to the lack of these conditions, the high-power experiments are conducted in the closed waveguide for health and safety.

For low power incidences test, a conventional parallel plate waveguide (PPW) setup is used. As shown in Fig. 12(a), a prototype with 10 unit cells is placed in middle of the PPW. Some foam absorbers with the same height as the PPW are placed around to eliminate reflections at the edges [38], [39]. Two cone antennas are used to provide a smooth transition from the coaxial cables to the PPW for broadband impedance matching. Time gating is applied to filter the noise caused by multiple reflections [40]. As shown in Fig. 12(b), the measured transmission at a low incident power level exhibits two transmission peaks with IL of 0.7 dB at 2.80 GHz and IL of 0.6 dB at 5 GHz and three adjacent transmission dips. The results are in good agreement with the simulation results, in spite of small frequency shifts of two passbands and three transmission dips, and small discrepancy of the bandwidth and minimum insertion loss. These differences may be attributed to fabrication errors of the prototype and test errors from the PPW setup. Nevertheless, the dual-band bandpass characteristic of our design for low-power incidence scenario is verified.

Fig. 12. (a) The PPW setup for low power incidence measurements. (b) The measured and simulated transmissions for low-power incidence.

Fig. 13. Rectangular waveguide setup for the ESS measurement under incidences with different power levels.

Fig. 14. ESS samples with different size. (a) The first layer and (b) the second layer in the waveguide WR-284. (c) The first layer and (d) the second layer for the waveguide WR-187.
As shown in Fig. 13, the performance of ESS under incidences with increasing power levels is measured using a rectangular waveguide setup. An incident signal with varying power is generated by a signal generator connected with a power amplifier. The circulator is used to protect the power amplifier. The signal passing through the waveguide is attenuated by an attenuator before measured by the spectrum analyzer for protection. Two standard rectangular waveguides, WR-284 (Inner dimension: 34.04 mm× 72.14 mm) and WR-187 (Inner dimension: 22.149 mm × 47.549 mm), are used for the tests of two different frequency bands, respectively. Fig. 14 shows the prototypes with corresponding sizes, which is a 3×7 array (30 mm×70 mm) and a 2×4 array (20 mm×40 mm), respectively. The metallic vias equivalent to metallic walls are designed on the edges of the ESS, minimizing the influence of the gap between the ESS and waveguide walls. Additionally, a metal flange with the same thickness as the air layer and the same size of hole as the waveguide is placed between the two layers of ESS. Initially, the transmitted power of the whole route without the ESS is measured to obtain the loss of route. Then the ESS is loaded and the transmitted powers at different incidence power levels are recorded by tuning the amplifier. The SEs of the ESS are calculated by normalizing the power with the ESS and that without the ESS. The SEs for different incidence frequencies are measured by varying the frequency of signal generator.

As shown in Fig. 15(a), the IL/SE results at six frequencies in the two operating bands are measured, with the incident power increasing from -30 dBm to 30 dBm. The IL/SEs increase from 0.6 dB to more than 20 dB (27.3 dB at 5.3 GHz). When the incident power is less than -15 dBm, the ILs are less than 1 dB. However, as the power increases further from -15 dBm to 30 dBm, the shielding effect enhances gradually due to the trigger of diodes. The observed nonlinear transmission performance has good agreement with the co-simulation results. Fig. 15(b) illustrates the SE results at various frequencies under high incident power levels of 30 dBm. In this high-power scenario, high SEs larger than 15 dB are observed in the frequency range of 2.6–5.9 GHz, indicating that both two operating bands are shielded when the ESS faces HPM illumination. The measured results agree well with the co-simulation results.

The designed dual-band ESS offers both low IL and high SE under low and high power incidences in two wide operating bands. A comparison between the proposed ESS and other designs reported in the literatures is provided in Table I. It is observed that most previous designs operate in single band. Moreover, compared to the dual-band ESS in [22], our design also exhibits superior performance in bandwidth of IL<1dB and SE>20dB owing to the second-order resonance. Additionally, the proposed ESS also offers better IL and higher SE in both bands in the experiments, benefiting from better independence of diodes in such a dual-layer configuration.

### V. CONCLUSION

A new design method of dual-band ESS is presented to get two switchable operating bands dependent on incident power. Two-layer configuration with two diode-loaded resonators designed on separate layers offers great benefit for the independence of two operating bands. The ESS exhibits low IL and high SE at low and high power levels in two bands. A variety of functions for each band are achieved according to the single-frequency incidence and two-frequency co-incidence scenarios with different incidence directions and powers. The multi-functionality is analyzed theoretically and studied via both full-wave and field-circuit simulations. Notably, the ESS is capable of shielding HPM in one band and transmitting low-power signals in another band simultaneously with an enormous transmission difference of 24.3 dB. Measurements for low-power and varying-power incidences are carried out in PPW and rectangular waveguides, respectively, showing good agreement with the simulations. The ESS can be used in radome of dual-band communication systems, providing two transmission windows for low-power communication signals and closing them adaptively when illuminated by high-power interference signals in both bands. Moreover, the independent control ability for two bands has great prospects in radome applications where HPM protection for one band and efficient communication in another band are required simultaneously.

### REFERENCES


