

Article

Design and Analysis of Dual-Frequency Energy–Frequency Composite Selective Surface with Dual-Period Nested Cross Fractals

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Abstract

This paper presents the design of a dual-frequency energy–frequency composite selective surface based on a double-period nested cross-fractal structure. The unit cell consists of a composite metallic layer loaded with diodes, an F4B dielectric substrate, and an intermediate layer with cross-shaped feeding line. The proposed model is structurally optimized and characterized using the periodic method of moments theory and the equivalent circuit method. In addition, its performance was verified through a comparative study. The results demonstrate that under low-power conditions, the surface achieves stable frequency-selective transmission at 2.4 GHz (S-band) and 4.2 GHz (C-band), enabling highly efficient signal transmission with an insertion loss of less than 0.6 dB. Under a high field strength, it automatically switches to an energy-selective state, providing full-band shielding effectiveness of ≥ 18 dB across a 2.0–5.0 GHz broadband, thereby achieving stealth functionality. The designed composite selective surface exhibits excellent angular stability and features a simple biasing network that does not require additional feeding lines. Thus, this study presents a new approach for designing such surfaces for operation in the microwave regime.

Keywords: energy–frequency composite selective surface; method of moments; dual-frequency; angular stability

1. Introduction

The stealth capabilities of weapons and equipment are of paramount importance in modern military warfare. However, achieving efficient electromagnetic stealth remains a critical challenge [1–6]. Research on integrated energy–frequency-selective surfaces (FSSs) [7] has overcome the limitation of single functionality in traditional devices [8–12] by enabling synergistic control of both the frequency and energy of electromagnetic waves within a single structure. Under low-power conditions, it allows for the normal transmission of the desired signals, fulfilling the frequency selection function. When confronted with high-power microwave threats, it automatically switches to an energy-shielding state, satisfying the requirements for out-of-band filtering and in-band limitation in spatial-field-intensity electromagnetic protection. This approach not only blocks strong electromagnetic pulses from entering the system but also effectively suppresses out-of-band electromagnetic interference.



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In recent years, significant progress has been made in circuit structures, operating frequency bands, and integration processes of energy–frequency composite selective surfaces. Keywon et al. [13] innovatively applied FSSs to radio-frequency (RF) energy harvesting. Their designed grid square-loop FSS operated without a matching network and could harvest energy from incident waves of arbitrary polarization, demonstrating the feasibility of integrating frequency selection with energy harvesting and thereby expanding the application prospects of integrated surfaces in the energy domain. In 2020, Qin et al. [14] proposed an ultra-wideband high-field protective device composed of a bandpass metamaterial cascaded with an active metamaterial. It achieves an insertion loss of <1.5 dB in the 3.38–4.33 GHz band under a low power, and a shielding effectiveness of >13 dB across the 0–20 GHz band under high-power microwave conditions. In 2021, Gao et al. [15] proposed a method to achieve broadband protection by cascading a metasurface with a switching effect and bandpass FSS. The surface designed based on this method transmits operational signals in the 4.02–4.44 GHz range with a transmission loss of <2 dB and provides protection against high-power microwaves across 0–8.85 GHz with a shielding effectiveness greater than 20 dB. In 2023, Zhong et al. [16] designed a dual-band energy-selective surfaces (ESS) based on a double-layer resonant circuit model. By loading diodes in separate layers to enhance the independence of the dual bands, they achieved an insertion loss of less than 1 dB under low-power conditions and a shielding effectiveness greater than 20 dB under high-power conditions, enabling independent control of energy and frequency across the two bands and refining the integrated design approach. Furthermore, Liu et al. [17] from Southeast University designed a metasurface focused on the Ku band. When irradiated by a high-power signal, the diodes conduct, switching the metasurface to the reflection mode with an isolation exceeding 8 dB below 13.7 GHz. In low-power scenarios, the diodes are OFF, and the metasurface operates in transmission mode with an insertion loss of less than 1 dB, achieving dual-frequency energy–frequency regulation. In the same year, a research team from the National University of Defense Technology designed an ultra-wideband high-intensity electromagnetic protection energy-selective surface with transceiving compatibility that can adaptively switch between the reflection and transmission states. Under low-power scenarios in the L, S, and C bands, the insertion loss is less than 1 dB, and the high-intensity electromagnetic protection effectiveness reaches 22 dB, providing an effective solution for the protection of ultra-wideband frequency-utilizing equipment [18]. In 2025, a review by Lv et al. [19] focused on the development of ESSs, covering four core areas: frequency bandwidth expansion, material innovation, functional enhancement, and application diversification, with particular attention to the structural design and related research advancements in integrated composite selective surfaces combining ESS and FSS. Furthermore, with the ongoing evolution of communication requirements, similar reconfigurable technologies have been widely implemented to control in more complex electromagnetic environments. For instance, movable time-modulated arrays can provide additional spatiotemporal degrees of freedom for covert communications [20]. Meanwhile, self-powered absorptive reconfigurable intelligent surfaces (RISs) have demonstrated significant potential for securing satellite–terrestrial integrated networks [21]. These frontier applications further underscore the necessity of developing composite selective surfaces with high-efficiency frequency control and high-power protection capabilities. In summary, the integrated composite design of ESSs and FSSs has become a core research topic.

Based on a periodically nested cross-shaped FSS, this study proposes a dual-frequency energy–frequency composite selective surface (EFSS) through structural innovation. A PIN diode with $R_s = 2\ \Omega$ and $C_T = 0.2\ \text{pF}$ under specific bias conditions is selected after in-depth analysis of its nonlinear characteristics. The design also incorporates a detailed bias network with an RF choke (RFC) and a DC block capacitor to effectively reduce the

nonlinear distortion and switching delay during state transitions. Compared to existing studies, the proposed design demonstrates significant performance advantages: under low-power conditions, it achieves precise frequency selection at 2.4 GHz and 4.2 GHz, with transmission loss below 0.6 dB—outperforming the approximately 1.5 dB loss reported in References [14,15] and thus significantly enhancing signal transmission efficiency. Under high-power conditions, the surface achieves full-band blocking across 2.0–5.0 GHz, with shielding effectiveness exceeding 18 dB. In contrast to Reference [17], this design not only covers a higher frequency range but also ensures fast response and high power tolerance during switching through accurate modeling and bias control of the diode’s nonlinear behavior, demonstrating substantial advantages and research value, and advancing the practical engineering application of integrated composite selective surfaces.

2. Unit Cell Model and Analysis Method

2.1. Fundamental Structure Model

For the selection of the unit cell structure, a cross-shaped geometry was adopted as the first-order fundamental element in this study because of its design simplicity, stable polarization characteristics, and ample symmetric redundant space, which facilitated the implementation of N-order structural loading. The FSS screen was arranged in a square lattice and tightly attached to a polytetrafluoroethylene substrate with a relative permittivity of $\epsilon_r = 2.65$ and a dielectric loss tangent of $\sigma = 0.001$; the specific structural parameters are listed in Table 1. Figure 1 shows the frequency response of the designed first-order unit array under normal incidence. Evidently, transmission occurred at $f = 5.8$ GHz, with the first-order operating band located in the C-band. Moreover, the structure exhibited desirable out-of-band rejection on both sides of the passband.

Table 1. Structural parameters of the FSS.

Parameter	D	L	W	R	h	ϵ_r	$\tan\delta$
Value/mm	19.6	1.2	19	3	1.1	2.65	0.001

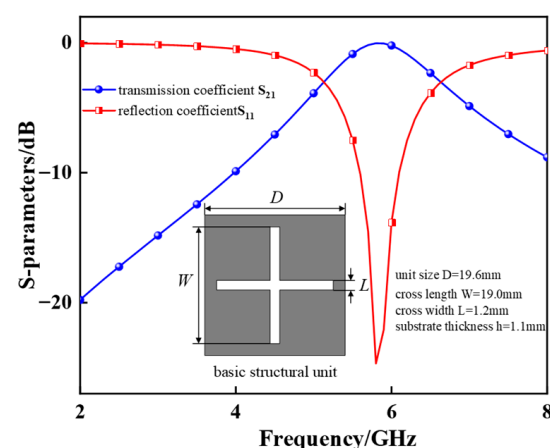


Figure 1. Frequency response of the cross-shaped FSS.

2.2. Numerical Analysis Method for Fractal FSS

This study primarily employed the periodic method of moments (MoM) for numerically solving the electromagnetic parameters of a fractal FSS [22]. Consider an FSS structure in free space as an example, as shown in Figure 2, the cross-shaped patches extend infinitely along the x - and y -directions. Here, θ denotes the angle between the incident electromagnetic wave and the normal vector of the FSS plane; ϕ represents the angle between the

projection of the wave vector on the xoy -plane and the x -axis; D_x and D_y are the periodic lengths in the x - and y -directions, respectively; and E_x^{inc} and E_y^{inc} correspond to the components of the incident electric field along the x - and y -directions.

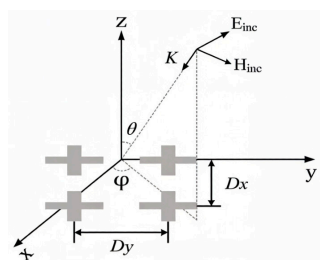


Figure 2. Schematic diagram of the cross-shaped periodic array.

According to Floquet's theorem for periodic structures, when a plane wave illuminates a periodic array, the amplitude of the induced current in any unit cell is identical to that in all the other unit cells. However, this idealized scenario assumes that patch elements possess infinite conductivity. For patches with finite conductivity, the boundary condition incorporating the surface resistance can be expressed as follows [23]:

$$\begin{bmatrix} E_x^s(x, y) \\ E_y^s(x, y) \end{bmatrix} + \begin{bmatrix} E_x^{inc}(x, y) \\ E_y^{inc}(x, y) \end{bmatrix} = Z_s \begin{bmatrix} J_x \\ J_y \end{bmatrix} \quad (1)$$

Here, Z_s denotes the surface resistance of the patch element, and E_x and E_y represent the components of the electric field along the x and y -directions, respectively. The superscripts s and i denote the scattered and incident fields, respectively, and J_x and J_y represent the induced surface current densities. By relating the electric field over the periodic unit cell to its surface current density using Equation (1), the Electric-Field Integral Equation (EFIE) describing the current distribution on the conductive patch can be derived as follows [24]:

$$-\begin{bmatrix} E_x^{inc}(x, y) \\ E_y^{inc}(x, y) \end{bmatrix} = \frac{2\pi}{j\omega\epsilon_0 D_x D_y} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \begin{bmatrix} k_0^2 - \alpha_m^2 & -\alpha_m \beta_n \\ -\alpha_m \beta_n & k_0^2 - \beta_n^2 \end{bmatrix} \times \tilde{G}(\alpha_m \beta_n) \begin{bmatrix} \tilde{J}_x(\alpha_m, \beta_n) \\ \tilde{J}_y(\alpha_m, \beta_n) \end{bmatrix} \exp(j\alpha_m x) \exp(j\beta_n y) - Z_s \begin{bmatrix} J_x \\ J_y \end{bmatrix} \quad (2)$$

In which

$$\alpha_m = \left(\frac{2m\pi}{D_x} \right) + k_x^{inc} \quad (3)$$

$$\beta_n = \left(\frac{2n\pi}{D_y} \right) + k_y^{inc} \quad (4)$$

In the above equations, K_x^{inc} and K_y^{inc} denote the x - and y -components of the incident wave vector, respectively; $\tilde{G}(\alpha_m \beta_n)$ represents the dyadic Green's function for free space; and m and n are the harmonic indices of the Floquet modes.

For FSSs, the roof-top basis function is the most commonly employed subdomain basis function. By expanding the induced surface current densities, J_x and J_y , in terms of roof-top basis functions and discretizing the unit cell into $N \times M$ segments, the expression for the induced current on the patch is obtained via the Fourier transform as follows:

$$J_x = \sum_{m=-M/2}^{M/2-1} I_x(n, m) B_x(n, m) \quad (5)$$

$$J_y = \sum_{m=-M/2}^{M/2-1} I_y(n, m) B_y(n, m) \quad (6)$$

In the above expressions, $I_x(n, m)$ and $I_y(n, m)$ represent the current distribution functions, and $B_x(n, m)$ and $B_y(n, m)$ denote the subdomain current basis functions along the x - and y -directions, respectively.

$$B_x(n, m) = \Lambda\left(n + \frac{1}{2}\right)\Pi(m) \quad (7)$$

$$B_y(n, m) = \Pi(n)\Lambda\left(m + \frac{1}{2}\right) \quad (8)$$

The roof-top basis function employs a triangular function as its spatial form:

$$A(n) = \begin{cases} 1 - \frac{|x - n\Delta x|}{\Delta x}, & |x - n\Delta x| \leq \Delta x \\ 0, & \text{other} \end{cases} \quad (9)$$

The pulse function is obtained by the following:

$$\Pi(m) = \begin{cases} 1, & |y - m\Delta y| \leq \frac{\Delta y}{2} \\ 0, & \text{other} \end{cases} \quad (10)$$

Applying the method of moments to Equations (5)–(10) determines the unknown induced current distribution on the unit cell surface. The scattered electric fields at various frequencies were then computed to yield the transmission and reflection coefficients of the structure.

2.3. Equivalent Circuit Method

An FSS is a two-dimensional periodic structure that filters incident electromagnetic waves based on frequency. It exhibits bandpass characteristics for aperture-type elements and band-rejection (or band-stop) characteristics for patch-type elements. The equivalent circuit method (ECM) [25] is an efficient, intuitive, and classical approach for analyzing FSS. This method represents the periodic structure of the FSSs at specific frequencies and incidence angles as an equivalent network composed of lumped elements, inductance (L) and capacitance (C), thereby transforming the complex problem of electromagnetic wave scattering into a simpler circuit analysis problem. When an electromagnetic wave is incident, the metallic strips parallel to the direction of the electric field can be modeled as equivalent inductors, whereas the apertures perpendicular to the direction of the electric field can be represented as equivalent capacitors. The cross-aperture FSS is a typical bandpass FSS whose unit cell is formed by etching a cross-shaped slot onto a continuous metallic plane. Based on the fundamental principles of the ECM, the response of such an aperture-type structure can be effectively modeled using a series LC resonant circuit. The corresponding equivalent circuit is shown in Figure 3 [26].

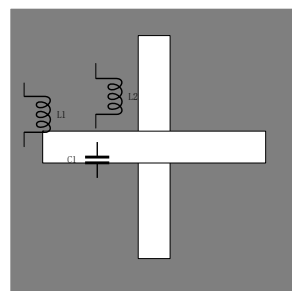


Figure 3. Equivalent circuit model of the center-cross structure.

For the series LC circuit model depicted in Figure 3, the equivalent impedance, Z , serves as the starting point for analyzing the frequency response. This impedance corresponds to that of the series of LC branches.

$$Z_{FSS} = j\omega L + \frac{1}{j\omega C} = j\left(\omega L - \frac{1}{\omega C}\right) \quad (11)$$

where Z_{FSS} represents the equivalent impedance of the FSS unit cell, and $\omega = 2\pi f$ denotes the angular frequency. At lower frequencies, the impedance, Z_{FSS} , is capacitive, which causes most of the incident signals to be reflected. At higher frequencies, Z_{FSS} becomes inductive, which results in the reflection of most signals. At the resonant frequency f , $\omega L = \frac{1}{\omega C}$, and this condition is satisfied, leading to a Z_{FSS} of 0. This indicates that the series LC branch is in resonance and effectively behaves as a short circuit.

To achieve dual-bandpass characteristics, a second-order LC-coupled bandpass filter circuit consisting of two parallel-series LC resonant tanks was inherently formed. Each series LC tank corresponds to a specific resonance point that determines the passband central frequency of the FSS. These two LC resonant tanks were coupled via electromagnetic fields, and their interaction created passbands within specific frequency ranges, collectively enabling the selective transmission of electromagnetic waves at two distinct frequencies. As illustrated in Figure 4, the free space on both sides of the FSS is equivalent to a transmission line with a characteristic impedance of $Z_0 = 377 \Omega$, where Z_d denotes the impedance of the dielectric substrate, which can be represented by a transmission line model of an equivalent circuit comprising a series inductor, $L_h = \mu_0\mu_r h$, and a shunt capacitor, $C_h = \epsilon_0\epsilon_r h/2$.

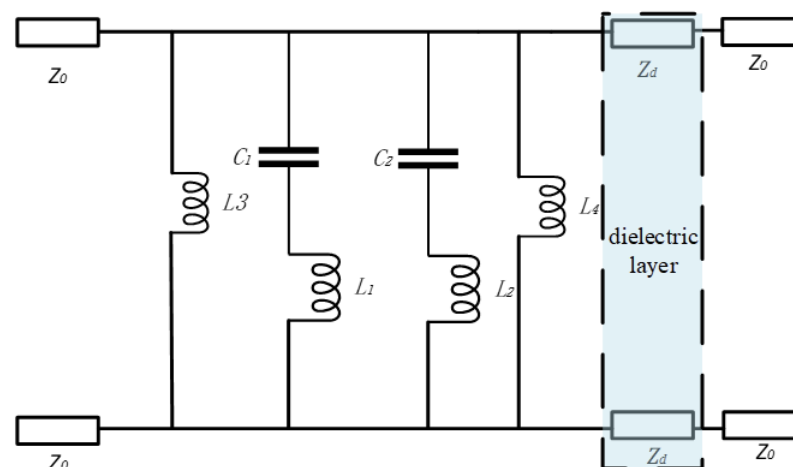


Figure 4. Equivalent circuit diagram of the dual-frequency energy-frequency composite selective surface.

3. Analysis of Numerical Results

3.1. Periodically Nested FSS Within a Unit Cell

Based on the equivalent circuit model topology shown in Figure 4, a dual-period nested fractal structure is constructed by extending a conventional simple cross-aperture model [27–29]. This nested configuration can be conceptualized as the integration of different types of (or the same type of) multiple periodic structures, with different electrical sizes, following specific nesting or combination rules. Fractal elements that expand outward from the original FSS unit to form self-similar patterns represent a particular form of periodic nesting. Such structures leverage the inherent properties of self-similarity and iterations to achieve multiband performance within a single layer. Assuming negligible mutual coupling among different nested elements, the overall response of the structure

can be approximated by a simple superposition of the resonant couplings. In this design, four second-order unit cells are loaded into the redundant space of the first-order array unit. The size ratio between the first- and second-order elements is 2.5. The resulting frequency-response characteristics of the periodically nested FSS array are illustrated in Figure 5.

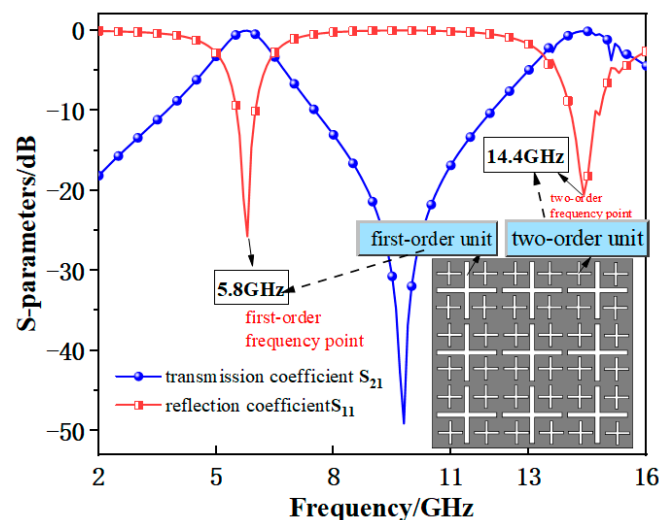


Figure 5. Frequency response of the intra-unit periodically nested FSS.

As observed in Figure 5, the transmission frequencies appear at 5.8 and 14.4 GHz, and the ratio between these resonant frequencies approximately equals the unit scaling factor. Because both the periodic unit arrays in this FSS structure are cross-shaped, the design exhibits low sensitivity to the incident angle and polarization of the incoming wave, resulting in high stability. For finite arrays, the number of N -th-order unit cells is four times that of the $(N - 1)$ -th-order array. This increase in the number of unit cells effectively mitigates edge effects, offering a novel solution for practical engineering applications.

Figure 6 shows the electric-field distributions excited at the first-order (5.8 GHz) and second-order (14.4 GHz) resonances for the first- and second-order unit arrays, respectively, under normal incidence of a TM-polarized plane wave. The results demonstrate that the dual-periodic nested structure excites two distinct modes of field resonance. The resonant coupling at the first-order passband is primarily excited by the first-order unit cells, whereas the coupling at the second-order passband is dominated by the four smaller second-order unit cells with a reduced electrical size. Moreover, the coupling strength is well maintained in both cases.

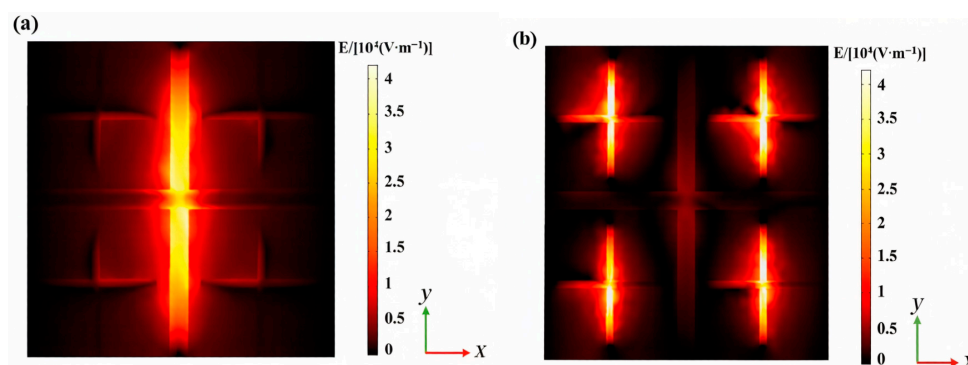


Figure 6. Electric-field distribution at the nested resonant frequencies of the unit cell: (a) 5.8 GHz (first order) and (b) 14.4 GHz (second order).

Based on the equivalent circuit model illustrated in Figure 4, a corresponding circuit is constructed using the in ADS 2020 software, and its S-parameters are analyzed. Figure 7 compares the S_{21} transmission coefficients obtained from the ADS 2020 equivalent circuit simulation and full-wave electromagnetic simulation (performed using CST 2018) for the intra-unit periodically nested FSS structure. This comparison is primarily focused on the S_{21} parameter to underscore the consistency between the equivalent circuit model predictions and full-wave simulation. The corresponding values of the equivalent circuit components, inductance (L) and capacitance (C), are summarized in Table 2.

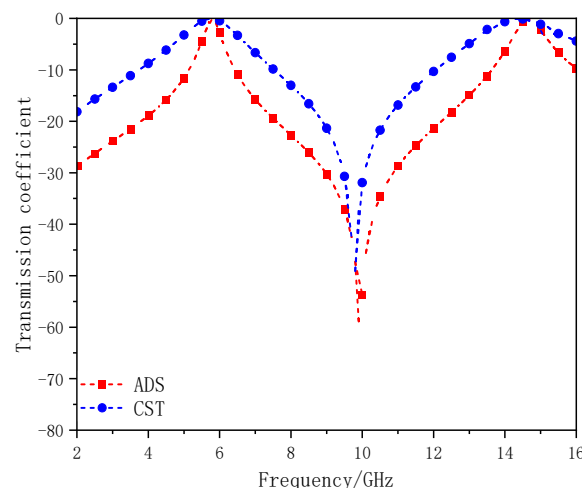


Figure 7. Comparison of S-parameters obtained from ADS 2020 and CST 2018 software for the intra-unit periodically nested structure.

Table 2. Equivalent circuit parameters of the nested unit cell structure.

Parameter	C_1	C_2	L_1	L_2	L_3	L_4
Value	0.56 pF	0.66 pF	0.46 nH	0.01 nH	0.96 nH	1.02 nH

For the intra-unit periodically nested structure, the S_{21} parameters (including both resonant frequency and strength) obtained from the ADS 2020 equivalent circuit simulation align well with those derived from the CST 2018 full-wave simulation (Figure 7). This excellent agreement between the results fully validates the rationality and correctness of the equivalent circuit model and supports the effectiveness of utilizing the ADS 2020 software to guide the FSS circuit design process.

3.2. Quantitative Analysis of Edge Effects and Numerical Errors in Finite Arrays

The scattered field, E_{total} , of a finite array can be interpreted as a convolution of an infinite array field, E_{infty} , with a truncation function. The electric-field distribution, $E_{finite}(x, y)$, of a finite array can be expressed as the product of the infinite array distribution, $E_{infinite}(x, y)$, and a window function, $W(x, y)$:

$$E_{finite}(x, y) = E_{infinite}(x, y) \cdot W(x, y) \quad (12)$$

where $W(x, y)$ defines the physical boundary of the array. According to Fourier-transform theory, the array factor, AF , of the finite array broadens in the far field:

$$AF_{finite}(k_x, k_y) = AF_{infinite}(k_x, k_y) * \text{sinc}(k_x L_x) \text{sinc}(k_y L_y) \quad (13)$$

Here, L_x and L_y are the total dimensions of the array.

When the total array size (L) exceeds the operating wavelength, λ ($L \gg \lambda$), the $\sin c$ function approximates the Dirac, δ , function. In this case, the response of the finite array mathematically converges to that of the infinite periodic model. The nested fractal design increases the total number of elements per unit area, N_{total} (the number of elements in N -th order is four times that in the elements of the $(N - 1)$ -th order). The resulting higher filling factor reduces the electromagnetic field discontinuity at the edges, further diminishing the influence of the correction terms. From the perspective of electromagnetic scattering theory, the relationship between the scattered field of the finite array, E_f , and that of the infinite periodic array, E_∞ , can be expressed as follows:

$$E_f \approx E_\infty \cdot \text{rect}(x/L) \quad (14)$$

In the spectral domain, this corresponds to the convolution of an ideal impulse function with the $\sin c$ function. Because the nested fractal structure proposed in this study significantly enhances the equivalent radiating magnetic current density per unit area, the beamwidth narrows, and the energy loss caused by edge diffraction is effectively suppressed. Therefore, the infinite periodic moment method (PMM) for parametric prediction is theoretically robust.

To quantify edge-truncation effects, full-wave simulations were conducted for arrays of different scales. Evidently, the overall frequency response trend of the 10×10 finite array aligns well with that of the infinite periodic assumption model, as evidenced by the significant resonant dip observed in the 9–10 GHz band. This result validates the predictive accuracy of the PMM in practical engineering (Figure 8). However, compared to the ideal infinite array, the finite array exhibits a notably worse matching depth at this resonance point, characterized by a shallower dip and narrower bandwidth. This characteristic feature confirms the existence of a performance trade-off, which is attributed edge effects in finite-sized arrays.

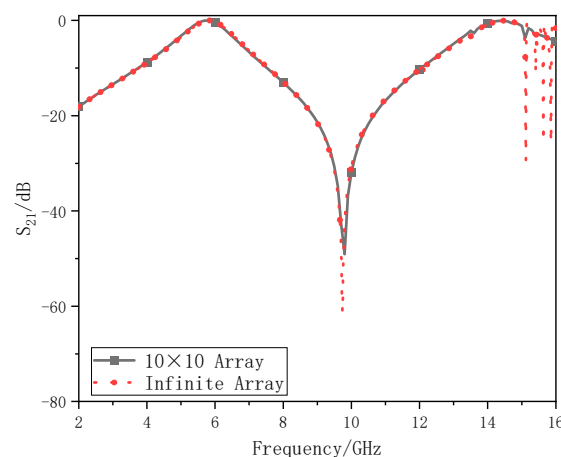


Figure 8. Comparison of simulated results for finite and infinite arrays.

3.3. Analysis of the Composite Selective Surface Structure and Its Parametric Effects

3.3.1. Diode Selection and Bias Network Analysis

Diode Equivalent Model and Nonlinear Characteristics Analysis

The performance of PIN diodes at high frequencies is significantly more complex than that of ordinary switches, and their equivalent circuits must account for package parasitic effects and internal physical structures. As shown in Figure 9a, under zero-bias conditions, the diode is equivalent to a network consisting of a parasitic inductance, L_s ; junction capacitance, C_T ; and parallel high resistance, R_p . L_s originates from the package leads, and C_T is determined by the dielectric properties of the intrinsic layer (I-layer).

When the diode is in the forward-bias state (Figure 9b), the I-layer is heavily injected with carriers, leading to a sharp drop in its impedance, which consequently behaves as a series resistance, R_s . This transition from a high-impedance capacitive state to a low-impedance resistive state forms the physical basis for the switching of RF switches, and the operating-environment-dependent parameter variations primarily reflect the nonlinear characteristics of diodes. The junction capacitance, C_T , is not a fixed value; it decreases with increasing reverse-bias voltage (V_R).

$$C_T(V_R) = \frac{C_{T0}}{(1 + V_R/\phi)^m} \quad (15)$$

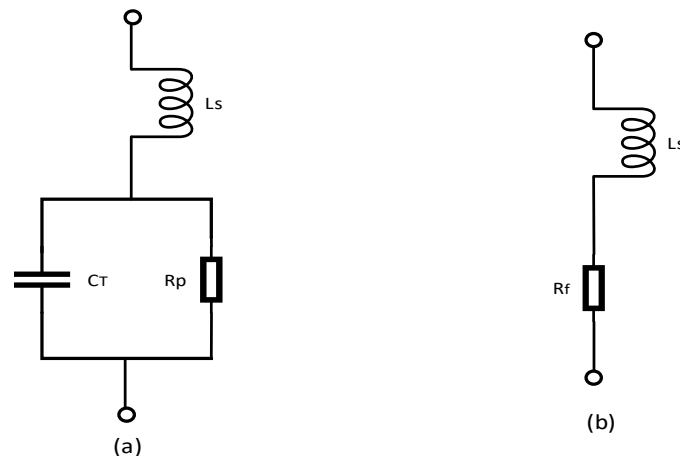


Figure 9. Equivalent circuits of PIN diode: (a) zero-bias equivalent circuit and (b) forward-bias equivalent circuit.

The forward resistance, R_s , is directly affected by the bias current (I_F) and typically satisfies $R_s \propto I_F^{-k}$, meaning that the driving capability of the bias network directly determines the switching loss. During high-speed switching, the diode exhibits a charge-storage effect, and its switching time is limited by the carrier lifetime, τ . However, within the frequency band investigated in this study, the signal period is considerably shorter than the carrier recombination time, and thus, this nonlinear effect is simplified to a quasi-static impedance.

To improve the simulation convergence speed and focus on system-level performance analysis, reasonable engineering simplifications were applied to the aforementioned model in this study. At high frequencies, R_p is extremely large (usually exceeding 10 k Ω), and the influence of L_s in the target frequency band is negligible; thus, when cut off, the diode is simplified to a pure capacitance, $C_{\text{off}} = C_T = 0.2$ pF. By contrast, in the conducting state, the inductive reactance generated by L_s is considerably smaller than the resistance at the current frequency; thus, the model is simplified to a pure resistance, $R_{\text{on}} = R_f = 2$ Ω . The optimal parameter combination, selected by commercial PIN diodes, achieves an optimal balance between isolation, $I_{so} \approx -20 \log(2\pi f C_T Z_0)$, and insertion loss, $IL \approx 10 \log(1 + \frac{R_s}{Z_0})$, thereby fulfilling the stringent requirements of this design for high-frequency switching.

In practical physical environments, the parasitic parameters of PIN diodes exhibit significant nonlinear characteristics, as shown in Figure 10a (junction capacitance curve) and Figure 10b (forward resistance curve). When considering the nonlinearity of the forward ON-resistance, as illustrated in Figure 10a, the forward dynamic resistance (r_f) of the diode is negatively correlated with the bias current, I_F . When the bias current increases from 0.1 mA to > 10 mA, the resistance drops rapidly from several hundred ohms to single-digit values. To balance low loss and power consumption, the bias point selected in this study corresponds to $I_F = 25$ mA, at which the equivalent resistance is $R_s = 2$ Ω . This condition explains the physical basis of the resistance in the simplified model. The voltage

and frequency dependence of the junction capacitance is shown in Figure 10b. Evidently, in the cutoff state, the junction capacitance (C_T) decreases nonlinearly with increasing reverse-bias voltage (V_R). When the reverse-bias voltage exceeds 4 V, and the operating frequency is in the RF band, the capacitance tends to stabilize. Therefore, adopting $C_T = 0.2$ pF as the equivalent value in the cutoff state accurately captures the isolation characteristics of the system at high frequencies.

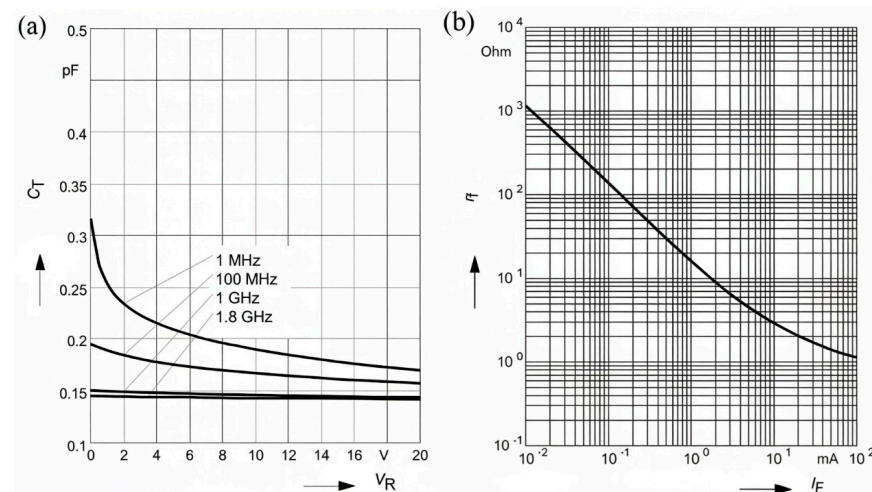


Figure 10. Datasheet diagram of PIN Diode BAR50-02L: (a) C_T - V_R and (b) r_f - I_F .

Diode Selection Process

Diode selection cannot rely solely on isolated parameters; rather, it must be comprehensively considered in conjunction with the frequency characteristics. The cutoff frequency (f_c) is the core indicator for evaluating the high-frequency switching performance of devices and is also a key basis for selection:

$$f_c = \frac{1}{2\pi R_s C_T} \quad (16)$$

where R_s is the forward-bias ON-resistance, and C_T is the reverse-bias junction capacitance. In this design, the intrinsic cutoff frequency, corresponding to the selected parameters ($R_s = 2 \Omega$; $C_T = 0.2$ pF) is approximately 398 GHz, which is significantly higher than the system operating frequency, and thus ensures the basic performance of the switch at high frequencies.

During the selection process, the following two key parameters must be constrained according to the system's requirements for signal transmission quality. In the ON state (Figure 9a), the loss generated by the switch is mainly attributed to R_s , and its simplified calculation formula is as follows:

$$IL(\text{dB}) \approx 10\lg\left(1 + \frac{R_s}{Z_0}\right) \quad (17)$$

where Z_0 denotes the characteristic impedance. To ensure overall system efficiency, the lowest possible forward-bias ON-resistance (R_s) should be that the insertion loss (IL) is controlled within 0.2 dB. In the cutoff state, the signal isolation is determined by the reverse-bias junction capacitance, C_T .

$$I_{so}(\text{dB}) \approx 10\lg\left(1 + \frac{1}{(4\pi f C_T Z_0)^2}\right) \quad (18)$$

The above equation shows that at high operating frequencies (f), a diode with an extremely low junction capacitance (C_T) should be selected to maintain sufficient isolation.

Combined with the nonlinear characteristic curves in the datasheet (see Figure 10), the selection process further takes into account the influence of the bias point on the parameters. The r_f - I_F curve shows that the slope of the resistance change tends to flatten after the bias current (I_F) reaches a certain threshold. The $2\ \Omega$ resistance selected in this study corresponds to the current saturation region, which can effectively suppress the nonlinear resistance fluctuations caused by signal swing. The C_T - V_R curve demonstrates that because the system's bias voltage is set in the flat section of the curve ($V_R > 5\text{ V}$), the nonlinear voltage dependence of the junction capacitance (C_T) is minimized, which ensures linearity in the cutoff state. The final parameter combination, selected based on the aforementioned quantitative analysis, achieves an optimal balance among loss, isolation, and nonlinear distortion.

Design and Operational Mechanism of the Diode Bias Network

To ensure the normal operation of the PIN diode in an RF system, a dedicated bias network must be designed to achieve effective isolation and coupling between the DC and RF signals. As shown in Figure 11, the network mainly injects a DC control signal (V_{Bias}) into the diode through a high-impedance RFC. Because the RFC exhibits an extremely high inductive reactance within the operating frequency band, it can effectively prevent the RF signal from leaking into the DC power supply terminal, thereby ensuring the integrity of the RF path.

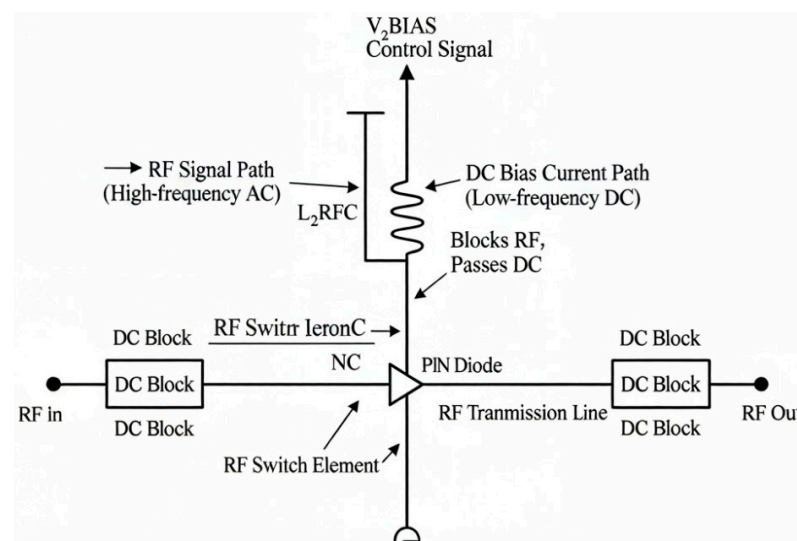


Figure 11. Diode bias network diagram.

DC blocking capacitors (DC blocks) are configured at the RF input (RF In) and output (RF Out) terminals. These capacitors allow for the passage of high-frequency AC signals while completely blocking the propagation of the DC signal source or load, thereby ensuring that the bias current flows only through the diode path. Upon varying V_{Bias} , the diode switches between forward- and reverse-bias states. In the forward-bias state, the bias network supplies a stable driving current, making the diode equivalent to an ON-resistance of $R_s = 2\ \Omega$; by contrast, in the reverse-bias state, it provides a cutoff voltage, making the diode equivalent to a cutoff capacitance of $C_T = 0.2\text{ pF}$.

The power consumption of the diode, which is determined by the bias current (I_F) and forward voltage drop (V_F), occurs mainly in the forward-conduction state ($P = I_F \cdot V_F$). As shown in Figure 10, the r_f - I_F curve demonstrates that an mA-level current is required to

maintain the $2\ \Omega$ low impedance, resulting in low total power consumption. Negligible power loss occurs under reverse bias owing to the extremely small leakage current in the circuit. The turn-off speed of a diode is primarily limited by its internal charge-storage effect. The selected device features an ultrasmall RC time constant (minimal $R_s C_T$ product) and optimized carrier lifetime, resulting in a nanosecond-level switching time. The bias network enables rapid current extraction (Figure 11), overcomes delays caused by charging and discharging of the nonlinear junction capacitance, and ensures precise waveform switching of the switch under high-frequency pulse environments.

3.3.2. Design of the Unit Cell Structure

To enhance the application diversity, an innovative design of a dual-frequency energy–frequency composite selective surface was developed in this study by structurally modifying the dual-period nested cross-fractal FSS and incorporating nonlinear elements. A cross-shaped FSS was introduced in the intermediate layer (Figure 12), which was cascaded with a 2.0 mm thick air layer. Furthermore, via structures (diameter: 0.2 mm) were incorporated at the corresponding positions throughout the three-layer configuration (Figure 13). These vias guaranteed stable interlayer interconnections and also functioned as feeding channels, enabling the efficient delivery of signals and power from external circuits to the entire energy-selective surface. Consequently, the proposed designed system achieved effective high-level electromagnetic shielding.

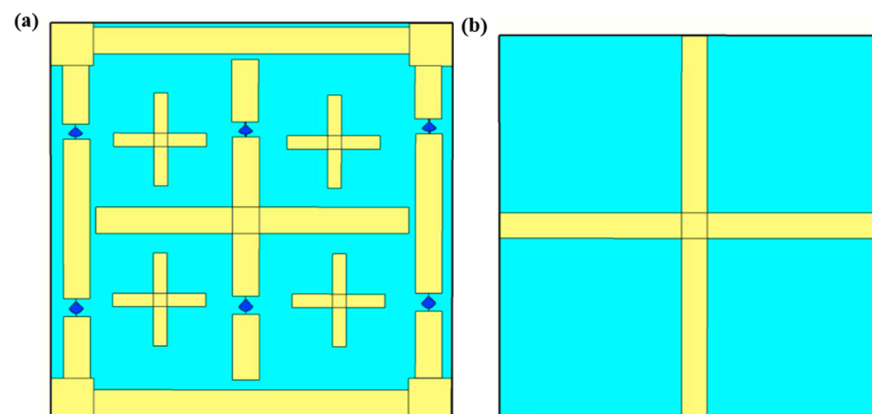


Figure 12. Unit cell structure of the top/bottom layer and intermediate layer. (a) Unit cell structure of the top and bottom layers. (b) Unit cell structure of the intermediate layer.

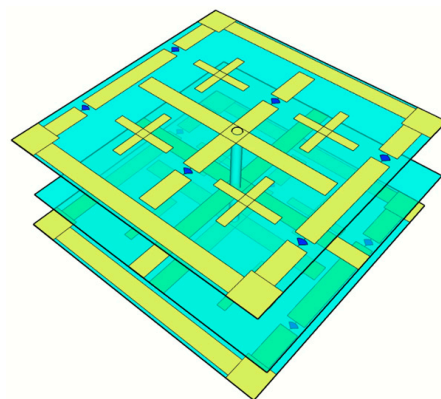


Figure 13. Overall configuration of the dual-frequency energy–frequency composite selective surface.

As shown in Figure 14, the proposed energy selective surface exhibits significant energy-shielding performance under different incident field strengths. Under low-field conditions, the diodes are in the OFF state, presenting a capacitance of 0.2 pF. Electromagnetic

simulations reveal distinct bandpass resonances at 2.4 and 4.2 GHz, indicating that the proposed design allows for the passage of signals at these frequencies, thus maintaining stable filtering performance. When the external field intensity increases beyond the triggering threshold, the PIN diodes switch ON, behaving as a resistance of $2\ \Omega$. This feature induces a fundamental transformation in the electromagnetic response of the surface, resulting in complete signal blocking across the 2.0–5.0 GHz frequency range. This full-band shielding effectively protects internal equipment from interference or damage caused by high-power electromagnetic signals.

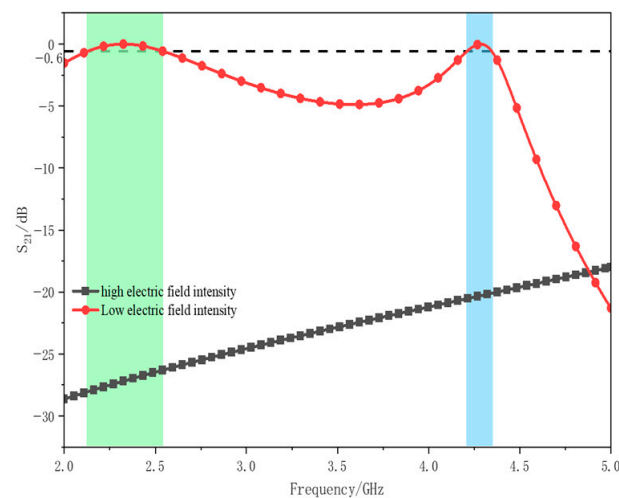


Figure 14. Frequency response of the dual-frequency energy–frequency composite selective surface.

According to the BAR50-02V datasheet, the threshold voltage (V_{th}) for the onset of conduction ranges from 0.6 V to 0.7 V. Once the voltage induced across the fractal gap reaches this threshold, the diode transitions from the OFF state to the ON state, triggering the redistribution of the surface electric field and achieving electromagnetic protection.

Under low-power signal incidence, the PIN diodes are in the OFF state, and their high impedance causes a mismatch, hindering the smooth transmission of electromagnetic waves. A portion of the energy fails to propagate through the structure and is instead reflected, superimposing around the diode regions. Consequently, these areas exhibit a high electric-field intensity, which is primarily concentrated near the left and right edge diodes, as shown in Figures 15 and 16.

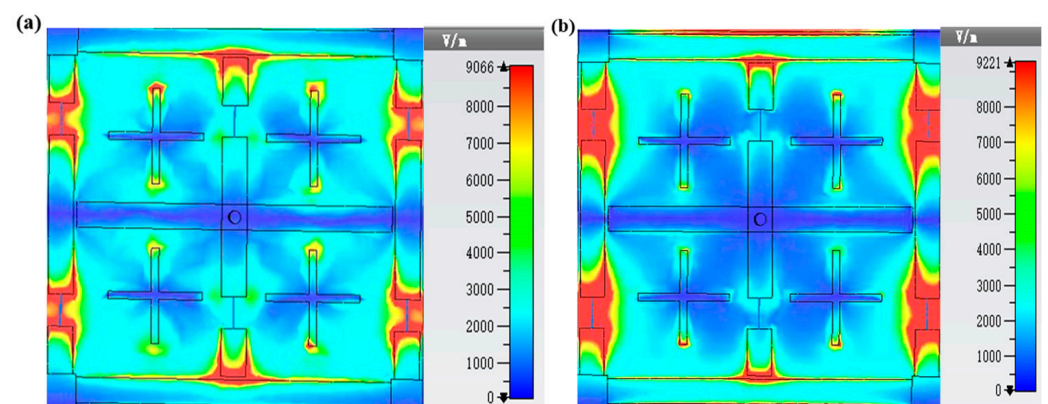


Figure 15. Electric-field distribution at resonance frequencies with diodes in the OFF state: (a) $f = 2.4$ GHz and (b) $f = 4.2$ GHz.

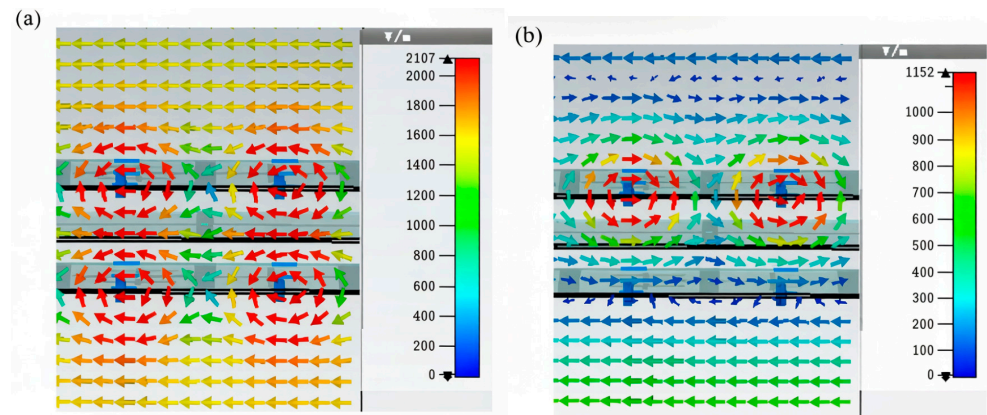


Figure 16. Electric-field-vector distribution at the resonance point when the diode is in the OFF state: (a) $f = 2.4$ GHz and (b) $f = 4.2$ GHz.

As shown in Figures 17 and 18, when the diodes are switched ON, the impedance of the energy-selective surface decreases rapidly, becoming smaller than the field strength at the resonant points under the diode-OFF state, and the impedance of the energy selective surface decreases rapidly when the diodes are switched ON. This phenomenon causes considerable mismatch between the wave impedance of the surface and that of free space, and thus, a large portion of the incident electromagnetic waves is reflected. Consequently, the energy transmitted through the surface is substantially reduced, which weakens the field strength at the original resonance points. The remaining energy is predominantly concentrated around the upper and lower edge patches in the center of the unit cell.

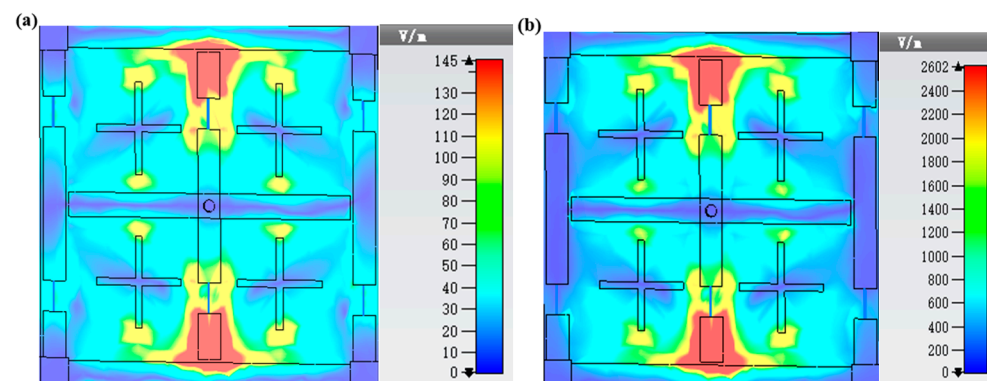


Figure 17. Electric-field distribution at resonance frequencies with diodes in the ON state: (a) $f = 2.4$ GHz and (b) $f = 4.2$ GHz.

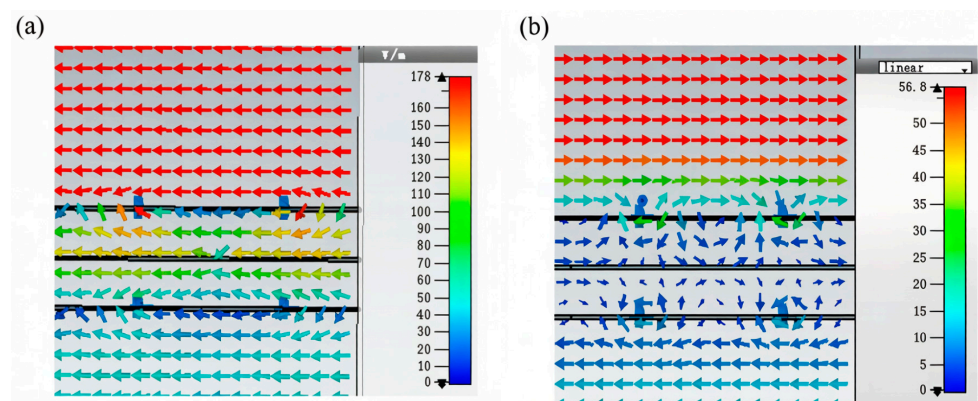


Figure 18. Electric-field-vector distribution at the resonance point when the diode is in the ON state: (a) $f = 2.4$ GHz and (b) $f = 4.2$ GHz.

The operating principle of the dual-frequency energy–frequency composite selective surface was analyzed using the ECM (Figure 19). The top and bottom layers are modeled as resonant circuits comprising a series of LC combinations in parallel with an inductor L , whereas the central cross patch and metal vias are represented by an inductor L_m , which adjusts the overall inductive reactance of the circuit. Here, Z_d is the impedance of the dielectric substrate. When the external incident field intensity is high, the PIN diode operates in the ON state, and its electrical behavior can be approximated by the resistor, R_i . This resistor absorbs a significant fraction of the high-field energy, thereby suppressing strong electromagnetic pulses. Under low-incident-field conditions, the PIN diode switches to the OFF state, and its characteristics are represented by the junction capacitance, C_j . This capacitance ensures the smooth transmission of low-field signals, ultimately realizing the core functionality of the energy-selective surface transmission under low fields and suppression under high fields.

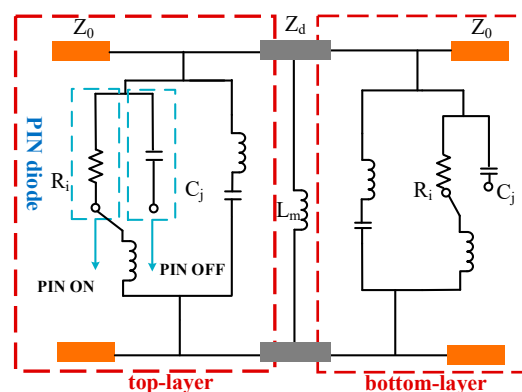


Figure 19. Equivalent circuit model of the dual-frequency energy–frequency composite selective surface.

To validate the proposed model, a cross-comparison verification was conducted using the equivalent circuit model (analyzed by the ADS 2020 software and a full-wave electromagnetic simulation (conducted using the CST 2018 software). Under both forward-bias (ON) and reverse-bias (OFF) states of the diode, the corresponding equivalent circuit models were established in the ADS 2020 software (Figures 20 and 21, respectively), and the specific values of each equivalent component were determined through parameter optimization (Tables 3 and 4). To verify the accuracy and consistency of the equivalent circuit modeling across different operating states, the S-parameter results obtained from the ADS 2020 simulations were compared quantitatively with the full-wave simulation results derived from the CST 2018 software under the same boundary conditions (Figures 22 and 23).

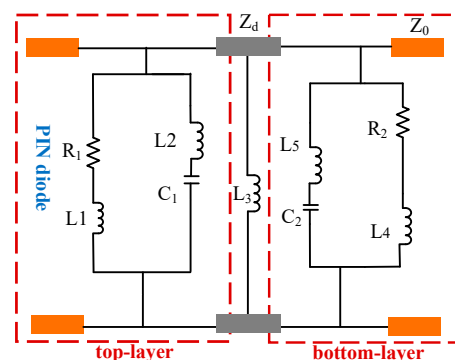


Figure 20. Circuit simulation diagram of the diode in conduction state in ADS 2020 software. (ON state).

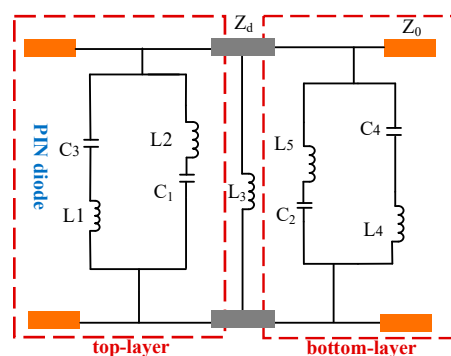


Figure 21. Circuit simulation diagram in ADS 2020 software when the diode is reverse-biased (OFF state).

Table 3. Component parameter values of the equivalent circuit when the diode is forward-biased (ON state).

L_1	L_2	C_1	L_3	L_4	L_5	C_2
1.84 nH	0.05 nH	0.57 pF	3.2 nH	1.32 nH	0.01 nH	0.01 pF

Table 4. Component parameter values of the equivalent circuit when the diode is reverse-biased (OFF state).

L_1	L_2	C_1	L_3	L_4	L_5	C_2
1.8 nH	0.01 nH	1.42 pF	0.98 nH	3.8 nH	2.91 nH	1.12 pF

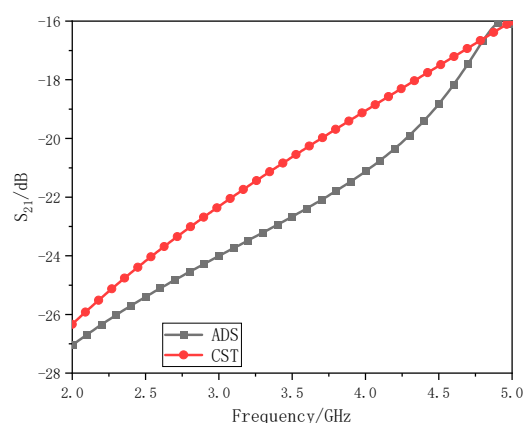


Figure 22. Comparative analysis of S_{21} parameters between ADS 2020 and CST 2018 software when the diode is forward-biased (ON state).

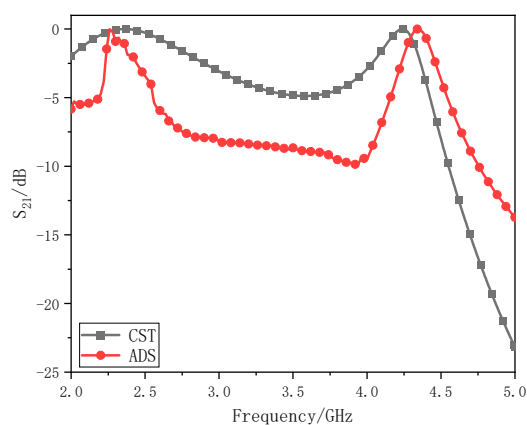


Figure 23. Comparative analysis of S_{21} parameters between ADS 2020 and CST 2018 software when the diode is forward-biased (OFF state).

3.3.3. Influence of Various Parameters on EFSS Performance

In this study, we focused on the unit cell structure of the proposed dual-frequency energy–frequency composite selective surface, specifically the key structural parameter of the dielectric substrate thickness (h). A single-variable parametric sweep analysis was conducted (Figure 24), and the influence of this parameter variation on the transmission coefficient, S_{21} , of the EFSS was systematically investigated through electromagnetic simulations. The parametric sweep results reveal that upon gradually increasing the dielectric substrate thickness (h), the shielding effectiveness of the ESS against high-intensity radiated fields degrades continuously. Moreover, the variation in loss exhibits a significant quantitative correlation for every 0.02 mm increase in thickness, and the insertion loss of the ESS increases by approximately 0.1 dB.

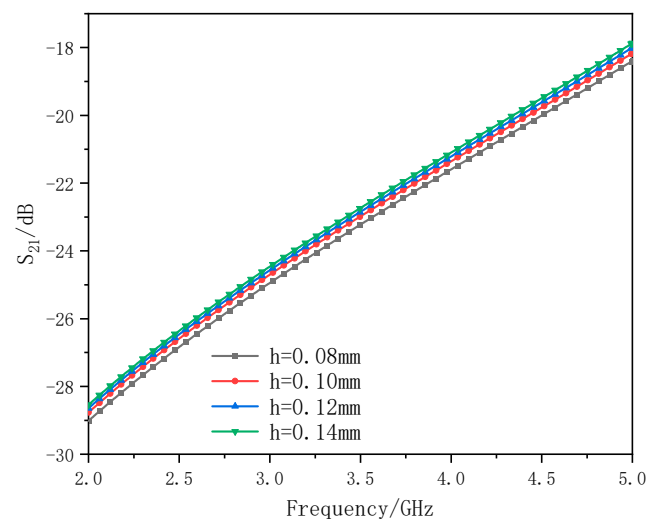


Figure 24. Influence of dielectric substrate thickness on the EFSS transmission coefficient.

To verify the environmental adaptability of the proposed structure in practical scenarios, a simulation-based analysis of its angular stability was conducted (Figure 25). Using the electromagnetic simulation software CST 2018, the reflection coefficients were measured across a wide range of incident angles, from 0° (normal incidence) to 75° (oblique incidence), within the target operating frequency band. The experimental results demonstrate that the design exhibits excellent angular stability from 0° to 60° . Specifically, the low-frequency resonance point (approximately 2.35 GHz) shows superior stability, with the response curves nearly overlapping and exhibiting negligible frequency drift. At the high-frequency resonance point (approximately 4.25 GHz), only a slight blueshift of approximately 0.1 GHz is observed as the incident angle increases, and this small shift is well within the acceptable tolerance for engineering applications.

Figure 26 illustrates the influence of key dimensional variations in the cross-patch structure on the transmission performance of EFSS. Figure 26a shows the variation in the transmission coefficient (S_{21}) across the frequency range of 2.0–5.0 GHz as the cross-patch width, L_1 , increases progressively from 0.6 mm to 0.9 mm. With increasing L_1 , the central frequency of the stopband shifts toward the low-frequency regime, while the stopband width changes slightly, and the suppression depth varies significantly within the 3.0–4.0 GHz range. Figure 26b demonstrates the impact of changes in the cross-patch length, D_1 , on the transmission performance. Similar to the variations in L_1 , upon changing D_1 , the stopband position shifts, and the suppression characteristics change. Overall, these results indicate that the electromagnetic response of the EFSS is sensitive to structural dimensions, and the operating band can be flexibly tuned by adjusting L_1 and D_1 .

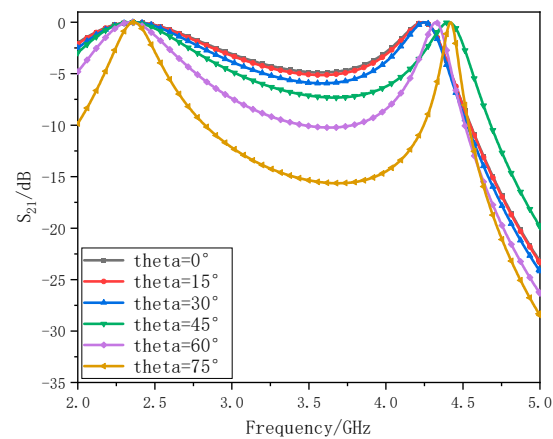


Figure 25. Frequency response of the dual-frequency energy–frequency composite selective surface under oblique incidence from 0° to 75° .

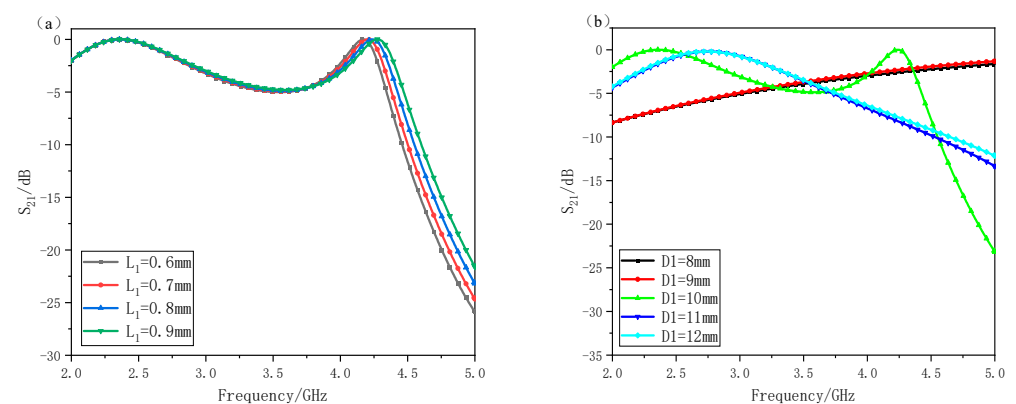


Figure 26. Impact of structural dimension variation on EFSS transmission performance. (a) Cross-patch width, L_1 . (b) Cross-patch length, D_1 .

In the frequency range of 2.0–5.0 GHz, the insertion loss of the structure with vias is generally lower than that of the structure without vias, with more pronounced improvement observed particularly in the 2.0–3.0 GHz band (Figure 27). This result indicates that the introduced vias significantly optimize the electromagnetic coupling path, reduce transmission losses, and enhance the energy transmission efficiency.

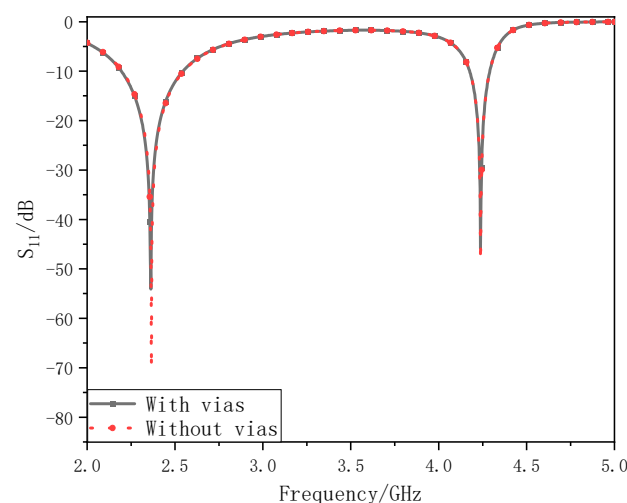


Figure 27. Impact of via presence on EFSS transmission performance.

3.3.4. Structural Sensitivity and Robustness Analysis

To further validate the reliability of the designed nested fractal structure for practical engineering applications and to evaluate its performance under non-ideal manufacturing conditions, a detailed sensitivity and robustness analysis was conducted, addressing uncertainties in manufacturing tolerances, non-ideal characteristics of active devices, and material properties.

Given the complexity of the ABA-type nested structures, manufacturing and assembly errors are inevitable during practical fabrication. In this study, we investigated the effects of the machining tolerance (± 0.1 mm) in the fractal patch position and the assembly tolerance (± 0.2 mm) in the air layer thickness on the device performance (Figure 28). The simulation results indicate that at both 2.4 and 4.2 GHz, the resonant frequency shifts by only $<2\%$. This result demonstrates that the proposed design exhibits strong robustness against errors introduced by standard printed circuit board (PCB) etching processes and manual lamination assembly.

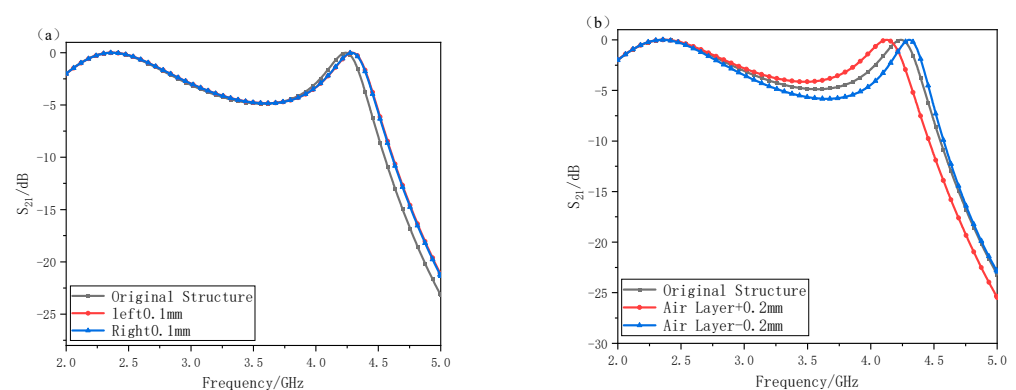


Figure 28. Impact of fabrication tolerances on transmission performance. (a) Effect of manufacturing tolerances in fractal patch position on transmission performance. (b) Effect of air-layer thickness on transmission performance.

Considering the occurrence of alignment errors during manual soldering or Surface Mount Technology (SMT), we performed a sensitivity analysis by shifting the diode position by ± 0.1 mm. As shown in Figure 29, the transmission performance remains stable, confirming the high fabrication tolerance of the proposed structure.

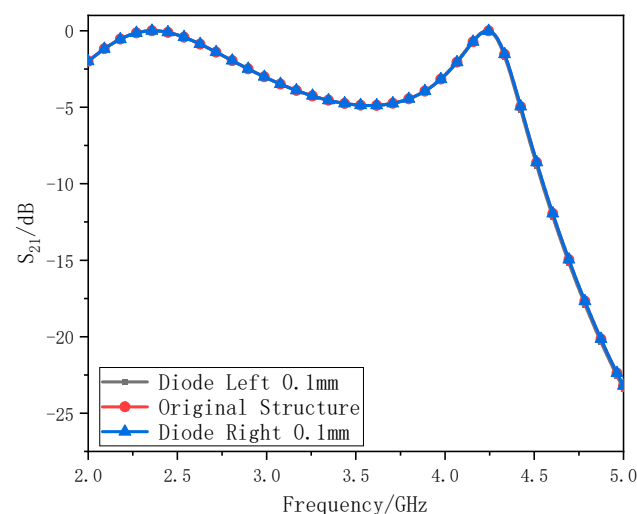


Figure 29. Variations in simulated transmission coefficients, S_{21} , with ± 0.1 mm shifts in diode position.

To ensure the reliability of the performance prediction, we incorporated the non-ideal modeling feature of Infineon BAR50-02V using its actual RF parameters (ON-state resistance, $R_s = 2\ \Omega$; OFF-state capacitance, $C_T = 0.2\ \text{pF}$). Considering the thermal drift effects under high-power conditions, a sensitivity analysis was conducted on R_s within the range of $1 - 10\ \Omega$ (Figure 30), which validated the reliability of the design in high-stress environments. Although an increase in R_s led to a slight rise in insertion loss, the core high-power shielding functionality remained fully intact.

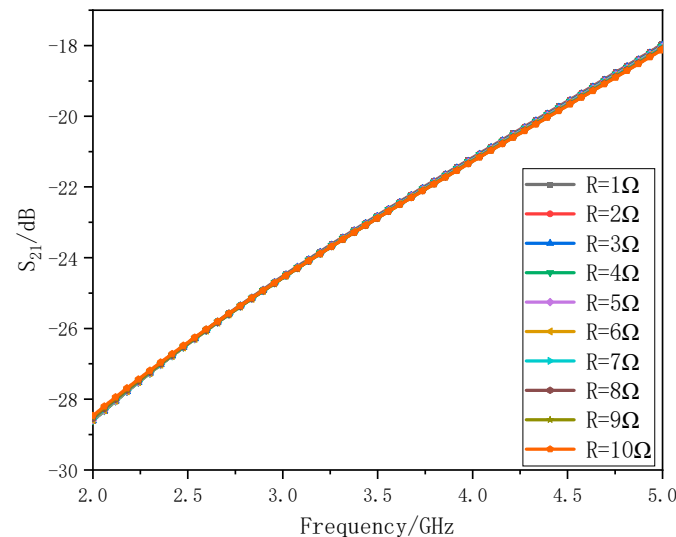


Figure 30. Influence of R_s variation on shielding effectiveness.

To address the parameter variations in commercial substrate materials across different batches, we investigated the sensitivity of the proposed structure to the dielectric constant, ϵ_r , of the substrate (Figure 31). A parametric scan around the standard value reveals that the transmission response does not show significant changes in terms of bandwidth and selectivity. This result indicates that the proposed design can be adapted to different batches of substrate materials without complex recalibration, thereby demonstrating its applicability in engineering practice.

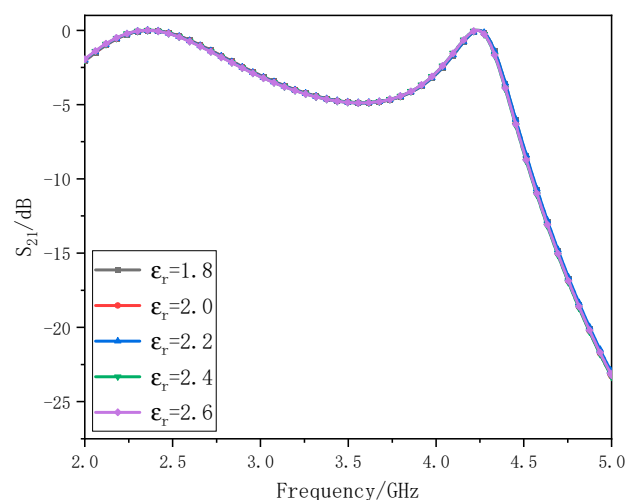


Figure 31. Influence of ϵ_r variation on shielding effectiveness.

3.3.5. Performance Trade-Off Analysis and Comprehensive Evaluation

The ABA-type triple-layer nested fractal structure designed in this study achieves an ultralow insertion loss of 0.6 dB while achieving a reasonable engineering trade-off among

bandwidth, angular stability, and fabrication cost. A low insertion loss serves as a critical metric to ensure communication transparency, and its relationship with the quality factor satisfies the following proportionality:

$$IL(dB) \approx 4.34 \cdot \frac{Q_u}{n \cdot Q_{ext}} \quad (19)$$

Although exceptional transmission performance is attained through triple-layer impedance matching, the highly selective resonance inherently narrows the fractional bandwidth (FBW) owing to the inverse relationship between the FBW and total quality factor. In Addition, the multilayer structures exhibit intrinsic phase shifts under oblique incidence. Using an effective refractive index model, the frequency drift behavior can be analyzed as follows:

$$f_{res}(\theta) = \frac{f_{res}(0)}{\sqrt{1 - (\frac{\sin \theta}{n_{eff}})^2}} \quad (20)$$

Despite maintaining high stability within 30°, frequency shifts occur above 45°. The reduction in bandwidth and angular dependence constitutes the essential physical trade-off required to achieve extremely low-loss performance at normal incidence and ensure nearly lossless signal transmission.

Although a design involving triple-layer PCB lamination and nested vias substantially increases the manufacturing complexity and cost, it forms the fundamental physical basis for enabling active self-triggered protection. According to the datasheet for the selected diode (BAR50-02V), the maximum allowable power dissipation per device is limited to 250 mW. The nested fractal structure significantly enhances the diode deployment density and evenly distributes the induced currents across the fractal layers. The corresponding thermal load-sharing mechanism is as follows:

$$P_{diss} = \frac{1}{2} I_{peak}^2 R_s \quad (21)$$

Leveraging this mechanism, the overall structure achieves superior a power-handling capability, outperforming conventional sparsely distributed designs. Thus, it effectively mitigates the risk of local thermal failure caused by high electric fields at the fractal gaps.

To ensure the reliability of the abovementioned model predictions and to evaluate their practical engineering value, non-ideal factors were incorporated into the simulations. The diode model not only includes realistic parasitic parameters ($L_s = 0.6$ nH and $C_T = 0.2$ pF) but also incorporates a sensitivity analysis of the ON-resistance over a range of 1–10 Ω :

$$Z_{diode} = R_s + j\omega L_s + \frac{1}{j\omega C_T} \quad (22)$$

These steps account for common practical variations, such as temperature drift, turn-ON voltage dispersion, and drive-current fluctuations. The experimental results (Figure 30) demonstrate the high transmission efficiency of the design within the communication band and its remarkable robustness under high-power electromagnetic stress.

4. Conclusions

In this study, we designed a dual-frequency energy–frequency composite selective surface using PIN diodes. The proposed structure comprised three layers: the top and bottom layers employed an innovative periodically nested cross-fractal patch structure, with an F4B material as the dielectric substrate, whereas the intermediate layer utilized a cross patch to facilitate feeding. A 0.2 mm air gap was inserted between adjacent layers, and

metallized vias were vertically incorporated to form the biasing network. The frequency response and energy regulation characteristics of the selective surface were investigated by analyzing the variation in the transmission coefficients under high- and low-field-strength conditions. The numerical results demonstrated that under low-power conditions, the structure exhibited precise dual-band bandpass characteristics, enabling efficient transmission at 2.4 and 4.2 GHz with extremely low insertion losses of 0.6 dB. These features ensured effective frequency selection and maintained the communication quality. When exposed to high-power microwave threats, the nonlinear components in the structure were rapidly activated, leading to a fundamental transformation of its electromagnetic response—automatically switching from a “transmission” mode to a “shielding” mode. In this state, the structure achieved full-band energy shielding, with a shielding effectiveness exceeding 18 dB across the 2.0–5.0 GHz frequency range, thereby realizing band stealth functionality. Furthermore, the structure maintained stable electromagnetic responses at incident angles ranging from 0° to 60°, demonstrating excellent angular stability. The unit cell model was analyzed using the ECM, and the results obtained from FDTD simulations and ECM were compared, confirming the validity of the numerical simulations. Although the feasibility of the model was thoroughly verified through comprehensive numerical simulations, the present study was conducted entirely within a simulated environment. Owing to the constraints of fabrication costs and experimental conditions, physical measurements were not performed. Consequently, the performance results presented herein can be considered theoretical predictions. In the future, we plan to develop a hardware prototype and conduct field experiments to further validate the robustness of the proposed model in complex real-world scenarios. The proposed composite selective surface overcomes the limitations of single-functional designs by enabling low-loss transmission in low-power scenarios, as well as by providing high-energy shielding in high-field-strength environments. These features effectively address the challenge of simultaneous precise control of frequency and energy. These characteristics significantly enhance the survivability of electronic systems in complex electromagnetic environments and broaden their adaptability, offering new insights into the integrated design of frequency and energy control in the microwave band.

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Conflicts of Interest: The authors declare no conflicts of interest.

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