

Advanced Multi-Band Shared-Aperture Antenna Arrays

by
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under the supervision of

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Certificate of Original Authorship

I, Shangyi Sun, declare that this thesis is submitted in fulfilment of the requirements for the award of Doctor of Philosophy, in the School of Electrical and Data Engineering at the University of Technology Sydney.

This thesis is wholly my own work unless otherwise referenced or acknowledged. In addition, I certify that all information sources and literature used are indicated in the thesis.

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Abstract

The rapid evolution of communication systems, along with the continuous pursuit of platform miniaturization and cost efficiency, requires multiple antennas operating at different frequencies to share an extremely limited space to simultaneously support various standards. The coexistence of different antennas results in strong mutual interference, namely scattering and coupling, which leads to severe distortion of radiation patterns as well as degradation in isolation and impedance matching. Achieving simultaneous suppression of both scattering and coupling over wide bands remains a critical challenge in the design of high-performance multi-band shared-aperture antenna arrays.

The first contribution of this thesis is the realization of simultaneous suppression of cross-band scattering, cross-band coupling, and in-band coupling in a dual-band shared-aperture antenna array. Three multi-functional suppression structures, namely 2.5-dimensional cloak, defected ground structure, and resonance shifter, are proposed for the low-band (LB) and high-band (HB) antennas in the array. Moreover, the proposed suppression methods exhibit high robustness and universality.

The second contribution of this thesis is the development of methods for wideband simultaneous suppression of scattering and coupling in a dual-band antenna array. A segmented spiral LB radiator capable of scattering suppression across an ultra-wide band is designed under the guidance of characteristic mode analysis (CMA). With the help of serial resonators, the LB antenna achieves wideband impedance matching. The HB antennas loaded with parallel resonance shifters not only suppress the scattering generated in the LB but also achieve wideband matching.

The third contribution of this thesis is the development of a tri-band antenna array featuring wideband suppression of scattering and coupling. The LB antenna, employing the segmented

spiral and serial resonators, realizes scattering suppression across both the MB and HB, which cover the expected 5G-Advanced and 6G bands, while its operating band covers the entire 5G sub-6 GHz range. The middle-band (MB) and HB antennas adopt a planar magnetoelectric dipole structure, which prevents common-mode resonance and mitigates scattering. Furthermore, general suppression methods for tri-band arrays are developed.

In this thesis, the various scattering and coupling suppression methods developed for dual-band or tri-band shared-aperture antenna arrays effectively restore the radiation patterns while enhancing cross-band and in-band isolation over wide bandwidths. The measured results for all array prototypes show good agreement with simulation results. These techniques are applicable to current and future terrestrial and non-terrestrial wireless communication platforms, including but not limited to base stations, satellites, and unmanned aerial vehicles (UAVs).

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Chapter 1

Introduction

1.1 Background

The rapid evolution of mobile communication systems, coupled with the continuous pursuit of miniaturization and cost-efficiency of platforms, requires multiple antennas operating at different frequencies to share an extremely limited space in order to simultaneously support various standards. The coexistence of different antennas results in strong interference between antennas, namely scattering and coupling, which leads to severe distortion of radiation patterns and degradation of isolation and impedance matching.

1.1.1 Dual-Band Shared-Aperture Antenna Array

As shown in Figure 1.1, when the low-band (LB) antenna is positioned adjacent to the radiating high-band (HB) antennas in the commonly adopted interleaved configuration, HB currents are induced on the LB radiators under the excitation of the HB wave. These induced currents generate secondary radiation that interferes with the original HB radiation, resulting in distorted overall radiation patterns in the HB. This phenomenon is known as **cross-band scattering** in the HB [1].

The induced currents on the LB radiators can also flow into the LB antenna ports through the LB baluns, leading to **cross-band coupling** in the HB. Furthermore, the secondary radiation from the LB radiators may induce additional currents on other HB antennas, thereby intensifying **in-band coupling** among HB elements.

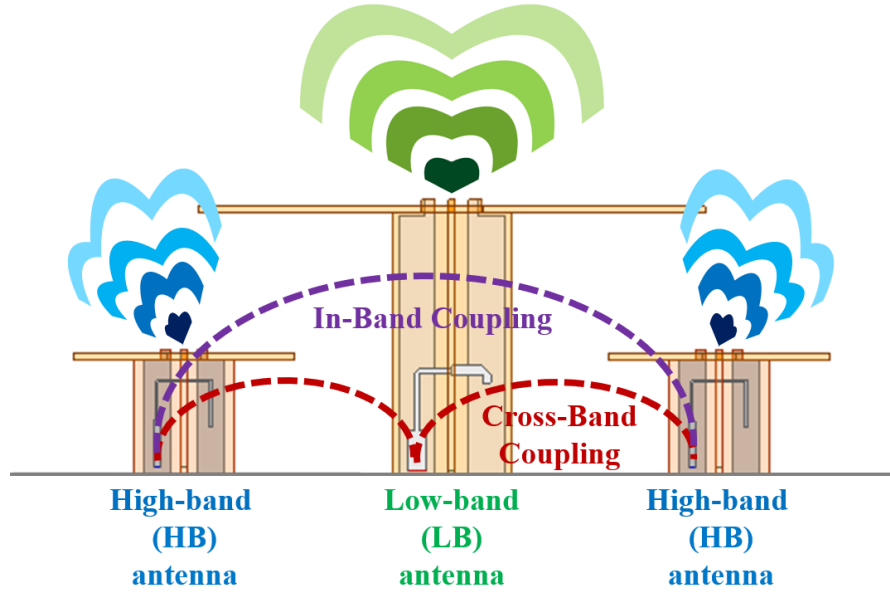


Figure 1.1: Cross-band scattering, cross-band coupling and in-band coupling in the dual-band shared-aperture antenna array.

Conversely, the HB antennas and the ground together form an integrated structure that can resonate at specific frequencies. When such a resonance point falls within the target LB, the radiation patterns of the LB antennas are significantly distorted. This interference is referred to as **cross-band scattering** in the LB.

Due to cross-band scattering, the main radiation beam of the LB or HB antennas may split or deviate from the boresight, causing the antenna to fail to provide the required coverage and resulting in signal loss in certain areas. Therefore, it is highly desirable to restore the radiation pattern as much as possible. Meanwhile, the mutual coupling between the LB and HB antennas, represented by the transmission coefficients between their ports, is also a non-negligible problem. Improving the isolation between antennas helps prevent inter-signal interference and maintains optimal impedance matching and radiation performance within each antenna's target frequency band. To enable the coexistence of multi-band antennas, all these scattering and coupling issues should be systematically but distinctly addressed.

To design a wideband dual-band shared-aperture antenna array, it is not sufficient to merely broaden the operating bandwidths of the LB and HB antennas. Instead, it is necessary to suppress scattering and coupling across the entire wideband while maintaining the broadband performance of both the LB and HB antennas. There is often a trade-off between achieving wideband suppression of scattering and coupling, and maintaining a wide operating bandwidth

of the antenna. When an LB radiator achieves scattering suppression over a wide HB through various methods or structural designs, its impedance matching bandwidth is often negatively affected. Similarly, for an HB antenna, structural loading can be applied to shift its resonance point out of the target LB; however, such structural modifications typically also deteriorate the impedance matching bandwidth of the HB antenna.

1.1.2 Tri-Band Shared-Aperture Antenna Array

In the tri-band shared-aperture antenna array, the scattering and coupling issues become more complex, since scattering and coupling occur between each pair of the three antenna elements, as shown in Figure 1.2. The LB antenna introduces scattering within the middle-band (MB) and HB, resulting in distortion of the radiation patterns of the MB and HB antennas. The MB antenna not only affects the radiation pattern of the HB antenna but may also generate resonance within the LB, causing distortion in the LB pattern. Similarly, the HB antenna may induce resonances in both the LB and MB, leading to radiation pattern distortion of the LB and MB antennas. In addition, coupling exists between any two antennas in the tri-band array, which affects the impedance matching and radiation performance of the antennas.

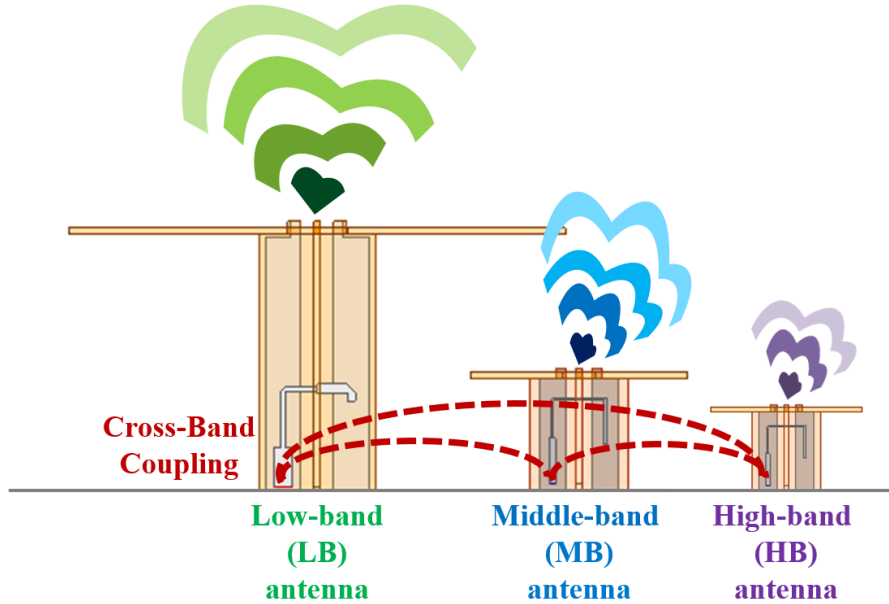


Figure 1.2: Cross-band scattering, cross-band coupling and in-band coupling in the tri-band shared-aperture antenna array.

In order to achieve holistic suppression within the tri-band shared-aperture antenna array, specific requirements must be met for each antenna. The LB antenna should effectively suppress

scattering within both the MB and HB. The MB antenna is required to suppress scattering in the HB while preventing any resonance in the LB. The HB antenna should maintain resonance-free operation in both the LB and MB. The isolation between any two antennas should be maintained at a relatively high level, typically above 20 dB. Furthermore, the implementation of the various suppression structures should not compromise the matching and radiation performance of the antennas.

The realization of a tri-band shared-aperture antenna array over wide bandwidths is extremely challenging, since the LB, MB, and HB antennas are required to suppress scattering and coupling in the other two wide bands while simultaneously achieving wideband impedance matching. As mentioned previously, there often exists a trade-off between wideband suppression and wideband matching. As the number of suppression bands increases and the suppression structures become more complex, the impact on impedance matching performance becomes more significant.

1.1.3 Applications

Shared-aperture antenna arrays have attracted increasing attention due to their capability to support multi-band and multi-function operation within a compact physical footprint. By integrating multiple frequency bands into a single aperture, these arrays significantly improve spectrum utilization efficiency while reducing system size, weight, and cost. At present, the most prominent applications of shared-aperture arrays are found in cellular base stations, where multi-band operation is essential for supporting legacy networks alongside emerging 5G-Advanced and 6G services. In addition, shared-aperture antenna systems are becoming highly attractive for non-terrestrial networks (NTNs), including unmanned aerial vehicles (UAVs), high-altitude platforms, and low-Earth-orbit (LEO) satellites, where stringent constraints on payload volume and platform integration exist.

1.2 Fundamental Theory

1.2.1 Characteristic Mode Theory

Since the following chapters will employ characteristic mode analysis (CMA) to guide the antenna design, this section provides a brief introduction to the characteristic mode theory and

its main parameters [2].

Characteristic mode analysis is essentially an eigenvalue problem. The eigenvalues and eigenvectors obtained from the eigenvalue equation represent certain intrinsic properties of a structure, such as resonant, radiative, and scattering characteristics.

In characteristic mode theory, the generalized eigenvalue equation is expressed as:

$$[X]\mathbf{J}_n = \lambda_n[R]\mathbf{J}_n \quad (1.1)$$

where $\mathbf{J}_n$ is the real eigenvector and λ_n is the real eigenvalue; and n is the index of the order of each mode.

Matrices $[R]$ and $[X]$ are the real and imaginary parts of the impedance matrix $[Z]$, respectively, satisfying:

$$[Z] = [R] + j[X] \quad (1.2)$$

$$[R] = \frac{[Z] + [Z]^*}{2} \quad (1.3)$$

$$[X] = \frac{[Z] - [Z]^*}{2j} \quad (1.4)$$

The eigenvalues reflect the resonant characteristics of the structure, specifically:

1. The magnitude of the eigenvalue is proportional to the total stored field energy in the radiation and scattering problems.
2. When $\lambda_n = 0$, the stored magnetic and electric energies are equal, and the mode is in a resonant state;
3. When $\lambda_n > 0$, the stored magnetic energy exceeds the electric energy, and the mode is in an inductive state;
4. When $\lambda_n < 0$, the stored electric energy exceeds the magnetic energy, and the mode is in a capacitive state.

The eigenvectors calculated from Equation (1.1) can be generally normalized to possess unit radiated power, which satisfies the following inner product criterion:

$$\langle [R]\mathbf{J}, \mathbf{J}^* \rangle = 1 \quad (1.5)$$

With this normalization, it can be observed that higher-order modes require larger electric currents to achieve the same unit radiated power compared to the lower-order modes. In this context, the magnitude of the normalized currents directly indicates the radiation efficiency of each mode.

The properties of the matrices $\mathbf{R}$ and $\mathbf{X}$ ensure that the real eigenvectors $\mathbf{J}_n$ are orthogonal with respect to both $\mathbf{R}$ and $\mathbf{X}$.

Accordingly, the orthogonality relations among the modal currents can be expressed as:

$$\langle \mathbf{J}_m, [R]\mathbf{J}_n \rangle = \langle \mathbf{J}_m^*, [R]\mathbf{J}_n \rangle = \delta_{mn} \quad (1.6)$$

$$\langle \mathbf{J}_m, [X]\mathbf{J}_n \rangle = \langle \mathbf{J}_m^*, [X]\mathbf{J}_n \rangle = \lambda_n \delta_{mn} \quad (1.7)$$

$$\langle \mathbf{J}_m, [Z]\mathbf{J}_n \rangle = \langle \mathbf{J}_m^*, [Z]\mathbf{J}_n \rangle = (1 + j\lambda_n) \delta_{mn} \quad (1.8)$$

where,

$$\delta_{mn} = \begin{cases} 1, & m = n \\ 0, & m \neq n \end{cases}$$

Because of this orthogonality, two orthogonal modes can be selectively excited to achieve circular polarization or suppress unwanted polarization in antenna design. Additionally, orthogonal modal properties are useful in the design of multiport or decoupled antenna structures.

Apart from the orthogonality of currents, the modal fields in the far-field region are also orthogonal. The far-field characteristic electric and magnetic fields satisfy:

$$\frac{1}{\eta} \oint_S \mathbf{E}_m \cdot \mathbf{E}_n^* dS = \delta_{mn} \quad (1.9)$$

$$\frac{1}{\eta} \oint_S \mathbf{H}_m \cdot \mathbf{H}_n^* dS = \delta_{mn} \quad (1.10)$$

The characteristic modes consist of a complete set of orthogonal modes. For a given external excitation, the induced current $\mathbf{J}$ on a perfect electric conductor (PEC) and the corresponding far-field radiation, represented by the characteristic electric field $\mathbf{E}$ and magnetic field $\mathbf{H}$, can be expressed as follows:

$$\mathbf{J} = \sum_n a_n \mathbf{J}_n \quad (1.11)$$

$$\mathbf{E} = \sum_n a_n \mathbf{E}_n \quad (1.12)$$

$$\mathbf{H} = \sum_n a_n \mathbf{H}_n \quad (1.13)$$

where a_n is the **Modal Weighting Coefficient (MWC)** of each mode. If the incident electric field is denoted as $\mathbf{E}^i$, then a_n can be expressed as:

$$a_n = \frac{\langle \mathbf{E}^i, \mathbf{J}_n \rangle}{1 + j\lambda_n} \quad (1.14)$$

The inner product $\langle \mathbf{E}^i, \mathbf{J}_n \rangle$ is referred to as the **Modal Excitation Coefficient (MEC)**, and the **Modal Significance (MS)** is defined as:

$$MS = \left| \frac{1}{1 + j\lambda_n} \right| \quad (1.15)$$

The modal weighting coefficient (MWC) and modal excitation coefficient (MEC) explain the ability of an external source to excite the characteristic currents, which depends not only on the structure itself but also on the position, amplitude, phase, and polarization of the external source. Proper utilization of these parameters can guide the design and optimization of antennas, for example, in determining the optimal feed location and suppressing undesired scattering.

The modal significance (MS) is an inherent property of the structure and is independent of external excitation. It represents the potential excitation capability between the external source and each mode.

By analyzing MWC and MS, the contribution of each mode under a given excitation source can be evaluated.

Compared with the eigenvalue, MS provides a more convenient way to characterize the resonant state of an antenna over a wide bandwidth. The potential relative operating bandwidth (BW) can be derived from the modal significance (MS):

$$BW = \frac{f_H - f_L}{f_{res}} \quad (1.16)$$

where f_{res} is the resonant frequency, f_H and f_L are the upper and lower frequencies, respectively. At these frequencies, the modal significance satisfies the following conditions:

$$MS(f_{res}) = 1 \quad (1.17)$$

$$MS(f_H) = MS(f_L) = \left| \frac{1}{1 + j\lambda_n} \right| = \frac{1}{\sqrt{2}} \quad (1.18)$$

The value of MS can be used to distinguish between significant and insignificant modes:

- When $MS \geq 1/\sqrt{2}$, the mode is considered a **significant mode**;
- When $MS < 1/\sqrt{2}$, the mode is considered a **non-significant mode**.

In the characteristic mode theory, a set of characteristic currents $\mathbf{J}_n$ is defined on the surface of a perfect electric conductor (PEC). These currents produce the corresponding characteristic electric fields $\mathbf{E}_n$ in free space. A phase delay exists between the tangential electric field and the characteristic current on the PEC surface, which is referred to as the **characteristic angle**, defined as follows:

$$\alpha_n = 180^\circ - \tan^{-1}(\lambda_n) \quad (1.19)$$

The phase information contained in the characteristic angle can be used to qualitatively analyze the radiation and scattering performance of each mode. For example, it can help determine whether two orthogonal modes can be simultaneously excited with a single feed in circularly polarized antenna design. In this case, two orthogonal modes with comparable modal significance (MS) and a 90° difference in characteristic angle are required to achieve an ideal axial ratio performance.

Usually, the characteristic angle α_n varies within the range of $[90^\circ, 270^\circ]$. The specific physical meanings of the characteristic angle in analyzing the radiation and scattering performance of the modes are as follows:

1. When $\alpha_n = 180^\circ$, the mode is at resonance.
2. When $\alpha_n = 90^\circ$ or 270° , the mode cannot radiate.
3. When $90^\circ < \alpha_n < 180^\circ$, the mode is inductive.
4. When $180^\circ < \alpha_n < 270^\circ$, the mode is capacitive.

The characteristic mode theory provides a comprehensive theoretical foundation for analyzing electromagnetic structures. The modal parameters, including the modal significance (MS), modal weighting coefficient (MWC), and characteristic angle (α_n), are important metrics for designing electromagnetic structures.

1.2.2 Antenna Scattering Theory

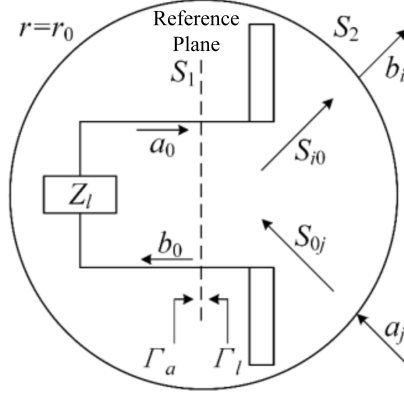


Figure 1.3: Scattering of antenna.

The scattering of antenna is illustrated in Figure 1.3. A cross section S_1 is arbitrarily selected on the feeding transmission line as the reference plane, while S_2 is a closed surface enclosing the antenna ($r = r_0$). The a_0 and b_0 denote the amplitudes of incident and reflected field at plane S_1 , respectively. The term a_j represents the amplitude of the incident field, while b_i denotes the amplitude of the radiated or scattered field [3].

The scattering characteristics of the antenna can be represented by the scattering matrix:

$$\begin{bmatrix} b_0 \\ b_i \end{bmatrix} = \begin{bmatrix} S_{00} & S_{0j} \\ S_{i0} & S_{ij} \end{bmatrix} \begin{bmatrix} a_0 \\ a_j \end{bmatrix} \quad (1.20)$$

From the scattering matrix:

$$b_0 = \sum_{j=0}^{\infty} S_{0j} a_j = S_{00} a_0 + \sum_{j=1}^{\infty} S_{0j} a_j \quad (1.21)$$

$$b_i = \sum_{j=0}^{\infty} S_{ij} a_j = S_{i0} a_0 + \sum_{j=1}^{\infty} S_{ij} a_j \quad (1.22)$$

When the antenna operates in radiation mode, the external incident field equals to zero, i.e., $a_j = 0$ ($j \neq 0$). Thus, Equations (1.21) and (1.22) reduce to

$$b_0 = S_{00} a_0 \quad (1.23)$$

$$b_i = S_{i0}a_0 \quad (1.24)$$

with

$$S_{00} = \Gamma_a = \frac{Z_{\text{in}} - Z_c}{Z_{\text{in}} + Z_c} \quad (1.25)$$

The matrix elements S_{00} and S_{i0} characterize the reflection coefficient and radiation property of the antenna.

In the receiving and scattering mode

$$a_0 = \Gamma_l b_0 \quad (1.26)$$

with

$$\Gamma_l = \frac{Z_l - Z_c}{Z_l + Z_c} \quad (1.27)$$

where Γ_l is the reflection coefficient of the load.

$$b_0 = \frac{1}{1 - \Gamma_l \Gamma_a} \sum_{j=1}^{\infty} S_{0j} a_j = \frac{1}{1 - \Gamma_l \Gamma_a} b_0^m \quad (1.28)$$

with

$$b_0^m = \sum_{j=1}^{\infty} S_{0j} a_j \quad (1.29)$$

where b_0^m denotes the receiving amplitude under matched condition.

Thus, the matrix element S_{0j} characterizes the receiving property of the antenna.

Similarly, the scattered wave amplitude can be written as

$$b_i = \sum_{j=0}^{\infty} S_{ij} a_j = S_{i0} \Gamma_l b_0 + \sum_{j=1}^{\infty} S_{ij} a_j \quad (1.30)$$

$$b_i^m = \sum_{j=1}^{\infty} S_{ij} a_j \quad (1.31)$$

where b_i^m denotes the scattering amplitude under matched condition.

Combining Equations (1.30) and (1.31) gives

$$b_i = b_i^m + \frac{\Gamma_l}{1 - \Gamma_l \Gamma_a} b_0^m S_{i0} \quad (\text{Scattering amplitude}) \quad (1.32)$$

Mapping the scattering amplitude in Equation (1.32) to the scattering field, the antenna scattering field can be expressed as

$$E^s(Z_l) = E^s(Z_c) + \frac{\Gamma_l}{1 - \Gamma_l \Gamma_a} b_0^m E_l^t \quad (1.33)$$

E_l^t denotes the radiated field generated under a unit-amplitude excitation source.

Defining

$$E^a(Z_l) = \frac{\Gamma_l}{1 - \Gamma_l \Gamma_a} b_0^m E_l^t \quad (1.34)$$

Thus

$$E^s(Z_l) = E^s(Z_c) + E^a(Z_l) \quad (1.35)$$

where $E^s(Z_c)$ is the structural-mode scattered field representing the scattering when the antenna is perfectly matched, and $E^a(Z_l)$ is the antenna-mode scattered field caused by load mismatch and reradiation.

Accordingly, the total radar cross section (RCS) can be decomposed as

$$\sigma_{\text{total}} = \left| \sqrt{\sigma_s} + \sqrt{\sigma_a} e^{j\phi} \right|^2 \quad (1.36)$$

where σ_s and σ_a are the structural-mode RCS and antenna-mode RCS, respectively, and ϕ is the phase difference between the two components.

In the far-field region, the scatterer illuminated by a plane wave can be regarded as a point target, and the distance r between the receiver and the scatterer is assumed to approach infinity. Accordingly, the RCS of antenna is defined as

$$\sigma = \lim_{r \rightarrow \infty} 4\pi r^2 \frac{|E_s|^2}{|E_i|^2} = \lim_{r \rightarrow \infty} 4\pi r^2 \frac{|H_s|^2}{|H_i|^2} \quad (1.37)$$

where E_i and H_i are the incident electric and magnetic fields illuminating the target, and E_s and H_s are the scattered fields observed at the receiver.

The unit of σ is m^2 , and it is commonly expressed in logarithmic scale with the unit of dBsm

$$\sigma_{\text{dBsm}} = 10 \log_{10} \sigma \quad (1.38)$$

Since the scattered field decays proportionally to $1/r$ in free space, the factor r^2 in Equation (1.37) cancels this dependence, making the RCS independent of the observation distance.

1.3 Research Objectives and Methodology

1.3.1 Research Objectives

Based on the scattering and coupling issues existing in dual-band and tri-band shared-aperture antenna arrays, as well as the difficulties in achieving effective suppression of scattering and coupling in wide bands, the specific research objectives of this thesis are:

1. To achieve simultaneous suppression of scattering and coupling in a dual-band shared-aperture antenna array; to develop methods with good universality and robustness that can be applied to dual-band arrays operating at different frequency bands or with various element configurations.
2. To achieve wideband suppression of scattering and coupling in dual-band shared-aperture antenna arrays, ensuring that the antenna performance in both the LB and HB is not degraded by the applied suppression method or structure.
3. To achieve wideband suppression of scattering and coupling in tri-band shared-aperture antenna arrays, effectively addressing various scattering and coupling problems within the antenna array while ensuring that antenna performance remains uncompromised.

1.3.2 Research Methodology

As described in Section 1.1.1, the cross-band scattering and coupling issues in shared-aperture antenna arrays originate from the HB induced currents excited on the LB antennas, which also further aggravate the in-band coupling between HB antennas. To mitigate the adverse effects caused by these induced currents, two methodologies are proposed in this thesis. The first approach introduces reverse HB induced currents through loading or slotting structures, where the scattered fields produced by equal-amplitude and out-of-phase currents cancel each other. The second approach suppresses the generation of HB induced currents; the segmented spiral structure designed in this thesis enables ultra-wideband suppression of HB induced currents. Furthermore, to extend the operational bandwidth of the segmented spiral when serving as LB dipole radiators, a serial resonator structure is proposed.

Conversely, HB antennas also severely distort the radiation patterns of LB antennas. This thesis mitigates such effects by introducing inductive/capacitive structures into the HB antennas or significantly reducing their profiles, thereby shifting the common-mode resonances responsible for pattern distortion away from the target LB.

1.4 Overview of Thesis

Chapter 2 provides a comprehensive review of the literature relevant to scattering and coupling suppression in multi-band shared-aperture antenna arrays. This includes studies on scattering suppression for LB and HB antennas, suppression of both cross-band and in-band coupling, simultaneous suppression of scattering and coupling, suppression achieved through modification of antenna array configurations, scattering and coupling suppression in tri-band and quad-band arrays, as well as Sub-6 GHz and millimeter-wave shared-aperture antenna arrays.

Chapter 3 presents a dual-band shared-aperture antenna array with simultaneous suppression of scattering and coupling. A 2.5-D cloak and a defected ground structure (DGS) are respectively implemented on the LB radiator and the LB balun, while a resonance shifter is introduced into the HB antenna. These structures can not only suppress cross-band scattering but also effectively mitigate both cross-band and in-band coupling. Moreover, the proposed method exhibits strong universality and robustness, making it applicable to different frequency bands and various array configurations. Under the two demonstrated configurations, the radiation

patterns of the LB and HB antennas are significantly restored, showing good consistency with those when they operate alone. Moreover, both the cross-band and in-band couplings in the dual-band array remain below -23 dB within the target bands.

Chapter 4 presents a dual-band shared-aperture antenna array that achieves simultaneous wide-band suppression of scattering and coupling. A segmented spiral radiator is developed for the LB antenna under the precise guidance of characteristic mode analysis (CMA), achieving scattering suppression over an ultra-wide band of 118%. To further enhance impedance matching, serial resonators are introduced to the LB antenna, extending the matching bandwidth to 0.69–1.00 GHz (36.7%). In addition, parallel resonance shifters are co-designed with the HB antenna to suppress LB scattering, and the HB antennas operate over a wide bandwidth of 1.63–2.83 GHz (53.8%). The radiation pattern distortions of both the LB and HB antennas are effectively mitigated, and both the cross-band and in-band isolations in the dual-band array exceed 20 dB.

Chapter 5 first presents a tri-band shared-aperture 5G/6G antenna array featuring wideband simultaneous scattering and coupling suppression. The LB antenna adopts segmented spirals as radiators, designed using the method proposed in Chapter 4. With the aid of serial resonators, an impedance bandwidth of 3.05–4.68 GHz (42.2%) is achieved, covering the 5G band of 3.3–4.2 GHz. The MB and HB antennas are both based on planar ME dipole structures, operating over 6.2–10.0 GHz (46.9%) and 10.0–15.6 GHz (43.8%), respectively, collectively spanning the expected 5G-Advanced and 6G spectrum of 6.425–15.35 GHz. The radiation patterns of all three antennas show no significant distortion, and the isolation between any two antennas in the proposed tri-band array is greater than 20 dB.

Chapter 5 then presents an LB antenna and an MB antenna designed for general tri-band shared-aperture antenna arrays. The LB antenna employs an uneven spiral radiator, which enables wideband scattering suppression covering both the MB and HB. With the aid of a serial resonator, the LB antenna achieves impedance matching within the band of 0.69–0.96 GHz. The MB antenna adopts a partial spiral radiator, providing effective scattering suppression within the HB (3.3–3.8 GHz). By introducing the proposed vortex line, a wide impedance bandwidth of 1.54–2.88 GHz (60.6%) is realized. In addition, a parallel LC structure is incorporated at the bottom of the LB balun to shift the resonance out of the target LB. The suppression techniques presented in this section exhibit better generality compared with those introduced in Section

5.2. Both the MB and HB antennas can operate at lower frequencies and employ conventional 3D dipole structures.

Finally, Chapter 6 summarises the results and implications of this thesis and briefly presents the ongoing and unfinished works related to multi-band antenna arrays for non-terrestrial networks (NTNs).

Chapter 2

Literature Review

2.1 Introduction

The coexistence of multi-band antennas within a shared-aperture array leads to interference, namely scattering and coupling, which causes severe distortion of radiation patterns and degradation of isolation and impedance matching. Distorted radiation patterns severely degrade antenna coverage and communication range, while low isolation induces mutual signal interference between ports. Numerous studies have focused on restoring radiation patterns and improving isolation by employing diverse scattering and coupling suppression methods. These efforts aim to satisfy common practical requirements, including minimal radiation pattern distortion and port-to-port isolation (both in-band and cross-band) exceeding 20 dB.

The cross-band scattering in the dual-band array occurs in both the low-band (LB) and high-band (HB) antennas. Section 2.2 discusses the suppression of scattering from the LB antenna that affects the HB radiation. The suppression structures include mantle cloaks, planar loaded structures, slots, L-shaped branches, U-shaped and folded structures, chokes and spirals, and frequency-selective surfaces (FSSs). They can generate out-of-phase scattered fields, induce reverse currents, or directly prevent the generation of HB-induced currents, thereby suppressing HB scattering. On the other hand, the suppression of scattering from the HB antenna that affects the LB radiation is summarized in Section 2.3. The LB scattering results from the common-mode resonance of the HB antennas within the target LB band. Therefore, connecting an inductor or a capacitor in series with the HB balun can shift the resonance out of the LB band. In addition, loading ferrite cores can convert the resonant electromagnetic energy into

thermal energy.

The coupling can be classified as cross-band coupling, which occurs between antennas operating at different bands, and in-band coupling, which occurs between antennas operating within the same band. The reduction of cross-band coupling is typically achieved through filtering techniques, which are outlined in Section 2.4. Section 2.5 presents the in-band coupling suppression methods. In addition to conventional approaches such as adding baffles, walls, and slots, other techniques include loading partially reflective surfaces (PRSs) and bending the arms of the dipoles. However, they are mainly designed to suppress the coupling without contributing to the scattering suppression.

Dual-functional structures capable of suppressing both scattering and coupling are reviewed in Section 2.6. Moreover, modifying the conventional interleaved configuration can also achieve simultaneous suppression of scattering and coupling. Section 2.7 introduces stacked, coplanar, and bowl-shaped configurations, which smartly avoid interference issues between cross-band antennas.

For the tri-band array, the suppression of scattering and coupling is much more complex than in dual-band arrays, as interference occurs between any two of the three types of antennas. Currently, existing works on tri-band arrays achieve suppression mainly by employing various methods proposed for dual-band arrays, which are discussed in Section 2.8. Furthermore, Sub-6 GHz and millimeter-wave shared-aperture antenna arrays have emerged as a recent research hotspot, and a brief overview is provided in Section 2.9.

2.2 Suppression of Scattering from the LB Antenna Affecting the HB Radiation

2.2.1 Mantle Cloak

In recent years, many scholars have conducted in-depth research on the scattering between cross-band antennas, especially the scattering from the LB antenna affecting the HB radiation. The mantle cloak is an early-studied loading structure for the scattering reduction in dipole antennas [4]–[12]. The mantle cloak is a kind of electromagnetic cover, which is selected and introduced into the scattering suppression in cross-band dipoles because it uses simpler loading

structures to achieve the scattering cancellation compared with other electromagnetic covers [4]. As shown in Figure 2.1 (a) and (b), the mantle cloaks consist of simple and periodic structures surrounding the LB dipoles [5], [6]. The scattering cancellation is achieved by introducing a reversed scattering field through the loading of the mantle cloak. When the scattering fields re-radiated by the cloak and the LB dipole are of reversed phase and equal magnitude, they can cancel each other out. The effect of the scattering cancellation can be intuitively reflected by the reduction of the scattering cross section (SCS) or radar cross section (RCS), and the scattering pattern of the LB dipole. After loading the optimized mantle cloak, the radiation patterns and reflection coefficients of the HB dipole are restored, while the radiation of the LB dipole is not affected by the cloak. Before and after loading the cloak on the LB dipole in Figure 2.1 (a), the radiation patterns of the HB dipole at 5 GHz are shown in Figure 2.1 (c), and the distortions in the patterns are reduced after the loading.

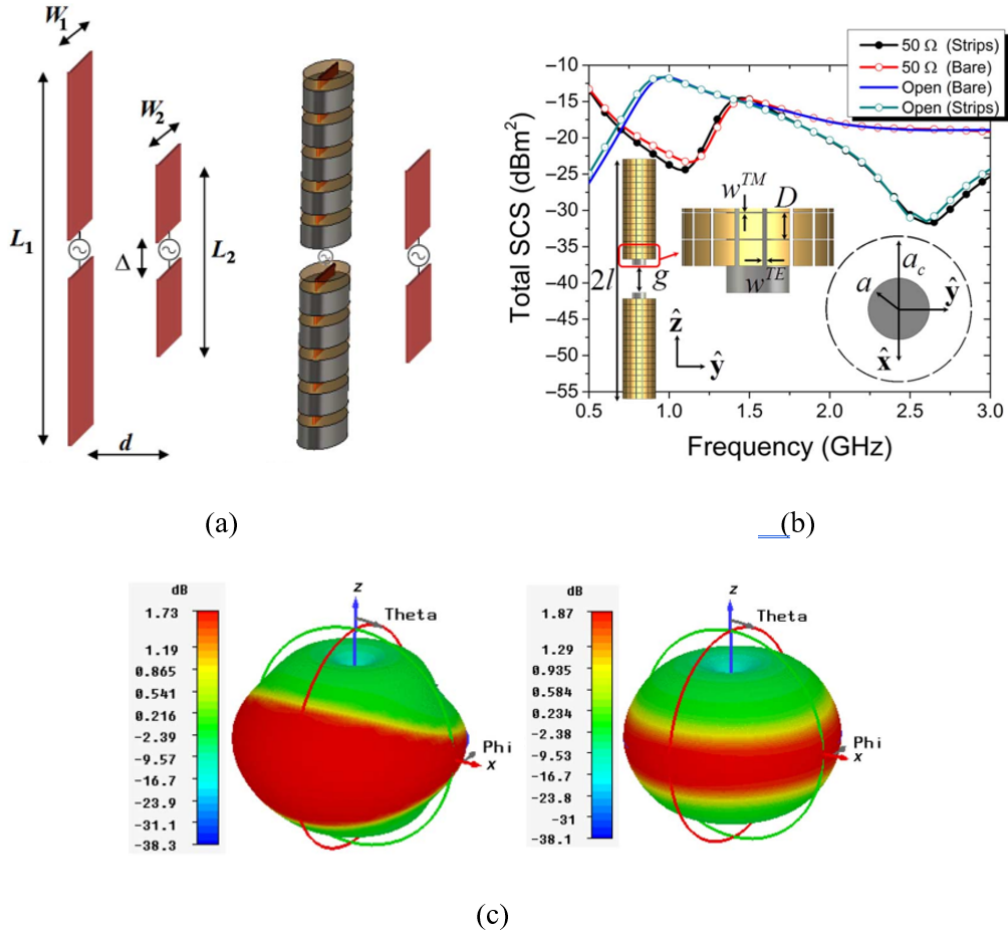


Figure 2.1: (a) LB dipole loaded with strip mantle cloak, (b) LB dipole loaded with patch mantle cloak, and (c) radiation patterns of HB dipoles before and after the loading of cloak.

2.2.2 Planar Loaded Structure

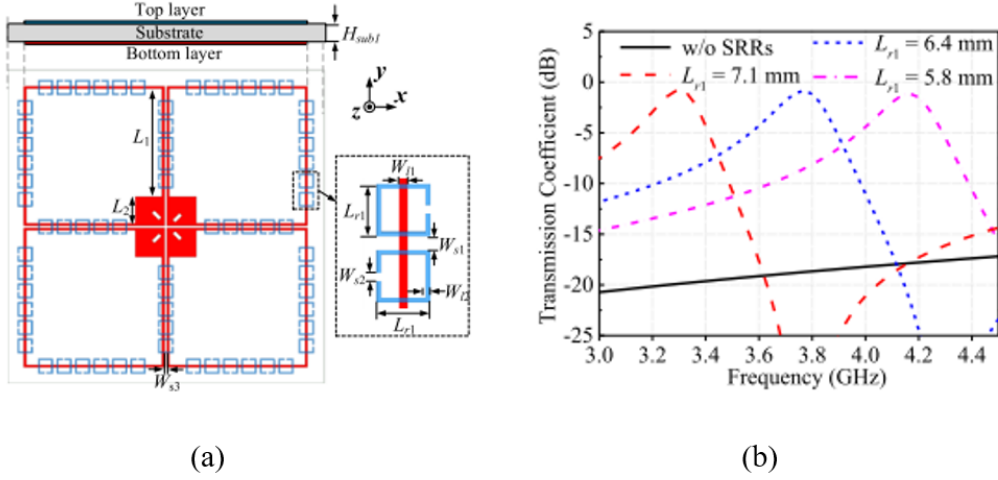


Figure 2.2: (a) LB dipole loaded with split-ring resonators (SRRs), and (b) transmission coefficients of the loaded LB dipole over the HB.

The above-mentioned traditional mantle cloaks are all three-dimensional structures that are loaded around the dipoles, causing great difficulties in fabrication and assembly. As depicted in Figure 2.2, the LB dipole working in 0.69–0.96 GHz is loaded with planar split-ring resonators (SRRs) [13] to suppress the scattering. After the SRRs are introduced, the modified LB dipole presents good band-pass characteristics in the HB (3.3–4.2 GHz). Therefore, the field radiated by the HB patch antenna is not affected, while the HB wave can pass through the loops of the LB dipole normally, making the patterns of the HB patch antenna return to normal [14].

Except for the SRRs, similar planar structures loaded on the LB antenna radiator include the mushroom structure [15], [16], patch structure [17], and the dielectric block structure [18]. The mushroom structure in [16] and the dielectric block structure in [18] are shown in Figure 2.3 (a) and (b), respectively. Their introduction can also suppress the reflection and scattering of the LB radiator and realize electromagnetic transmission within the HB.

2.2.3 Slot

The above works introduce additional cloaks or structures to suppress the scattering on the LB dipole, and the other works shown in Figure 2.4 etch various slots on the LB dipoles to generate reversed currents [19]. When the HB dipole is radiating, reversed currents are induced on both sides of the slots. When the amplitudes of these reversed currents are close, their secondary radiation fields (scattering fields) can cancel each other out, achieving the effect of scattering

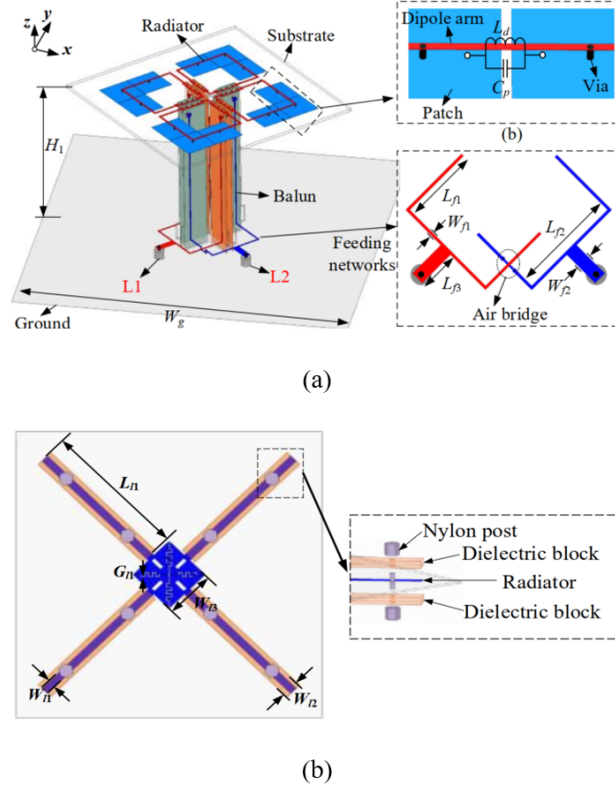


Figure 2.3: (a) Mushroom structure and (b) dielectric block structure loaded on the LB radiator for scattering suppression.

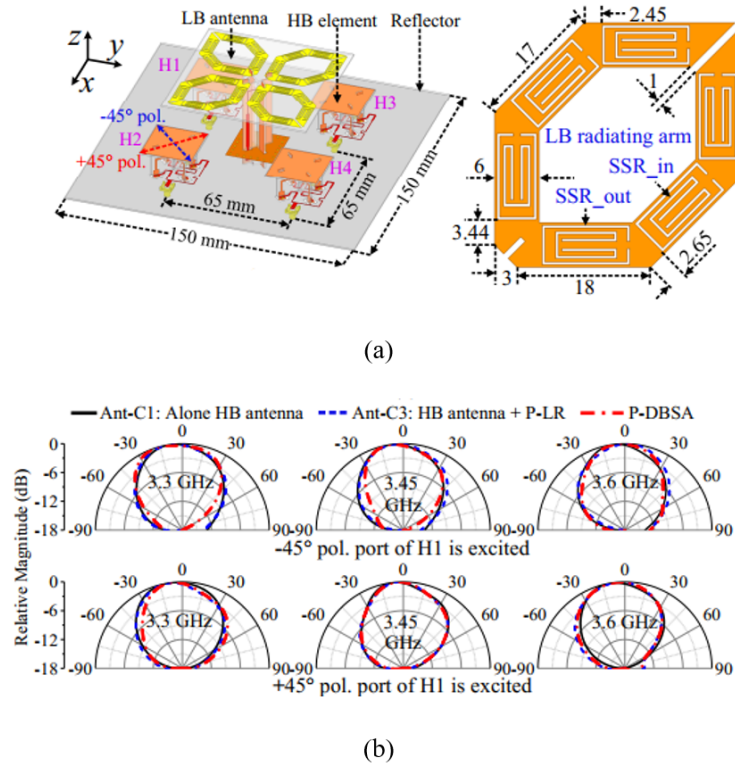


Figure 2.4: (a) LB dipole loaded with slots, and (b) patterns of the HB dipoles after the loading.

suppression. The suppression band is mainly determined by the length of the slots. The longer the length, the lower the suppression band, so the scattering suppression range can be quickly optimized and adjusted to other target bands. In Figure 2.4 (a), the LB dipoles loaded with slots working in 1.68–2.85 GHz achieve scattering suppression in the HB over 3.3–3.6 GHz. The patterns of the HB dipoles in Figure 2.4 (a) are well restored, as depicted in Figure 2.4 (b). In addition, the loadings of these slots do not affect the normal radiation or matching of the LB dipoles, so loading the slots is a good way to suppress the scattering on the LB dipole [20], [21].

The C-shaped or U-shaped slots in Figure 2.4 can only achieve scattering suppression in a narrow band, and the literature [22], [23], and [24] propose complex slots to achieve broadband scattering suppression. As given in Figure 2.5 (a) and (b), the slots on the LB radiator consist of U-shaped, cross-shaped, and square-loop-shaped slots, and the loading of each slot introduces reversed induced currents. The LB dipole after loading the slots can achieve scattering suppression within the HB (1.7–2.4 GHz), and the patterns of the HB dipole are well restored over this band. Moreover, the slots can improve the matching of the LB dipole, and its VSWR is less than 1.5 from 0.69 to 0.96 GHz. The scattering suppression effect of the combined slots is very good over a wide band, but the structures of the slots are complicated. The optimization process is very time-consuming, and it is very difficult to quickly adjust and apply to other suppression bands.

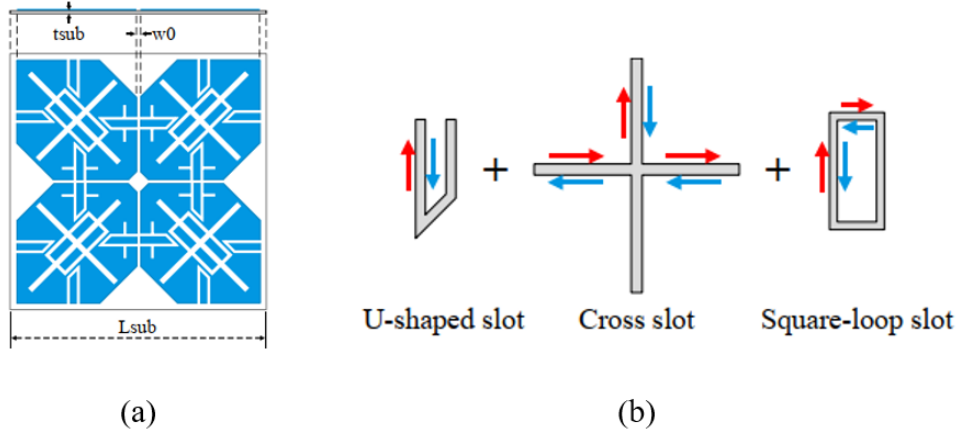


Figure 2.5: (a) LB dipole loaded with complex slots, (b) slots with different shapes.

2.2.4 L-Shaped Branch

In addition to etching slots on the LB radiator, adding L-shaped branches [25]–[29] to the LB radiator is also an effective method to suppress scattering in the HB as shown in Figure 2.6.

These L-shaped branches have a similar effect to an L-type filtering structure named open-ended stub, which can suppress the generation of HB-induced currents on the LB radiator and thereby suppress the HB scattering of the LB radiator. This method features a simple and low-cost structure and, through optimization, can achieve a relatively wide scattering-suppression bandwidth [26]. It has already been implemented in many commercial base station antenna products.

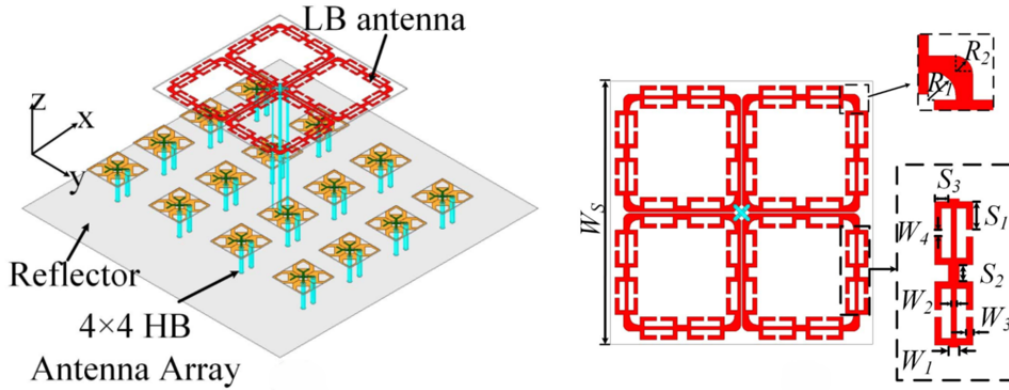


Figure 2.6: LB dipole loaded with the L-shaped branches in the dual-band antenna array.

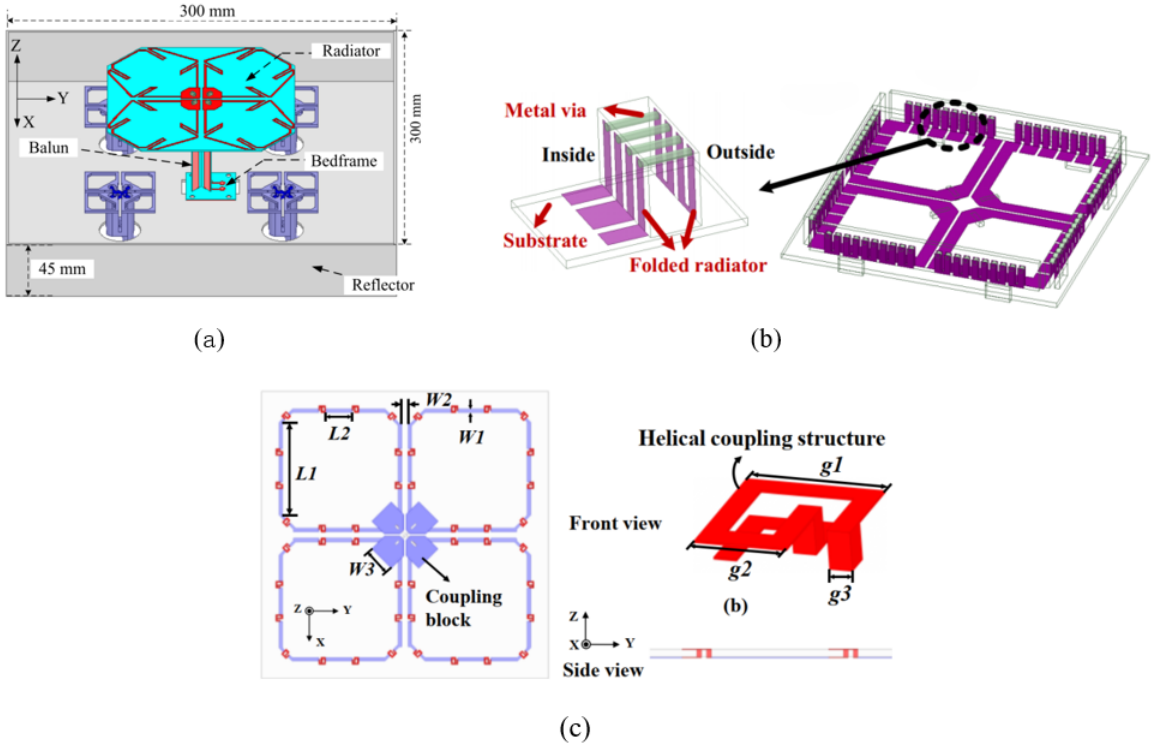


Figure 2.7: LB dipole loaded with (a) U-shaped structure, (b) meander line structure, and (c) helical line structure.

2.2.5 U-Shaped and Folded Structure

The scattering suppression methods introduced above all involve adding or removing structures on the LB radiator, while modifying the structure of the LB radiator itself can also achieve scattering suppression. As shown in Figure 2.7 (a), a part of the LB radiator is bent into a U-shaped structure [30]–[32]. Since the current distributions on the two arms of the U-shaped structure are almost of the same magnitude but opposite in direction and phase, most of the reradiation of the HB induced currents can be neutralized, resulting in scattering suppression and reduced HB pattern distortion. Moreover, bending the LB radiator into a meander line or helical line [33]–[35], as given in Figure 2.7 (b) and (c), can also achieve scattering suppression. It should be noted that a more complex structure or a greater number of bends does not necessarily lead to better performance.

2.2.6 Choke and Spiral

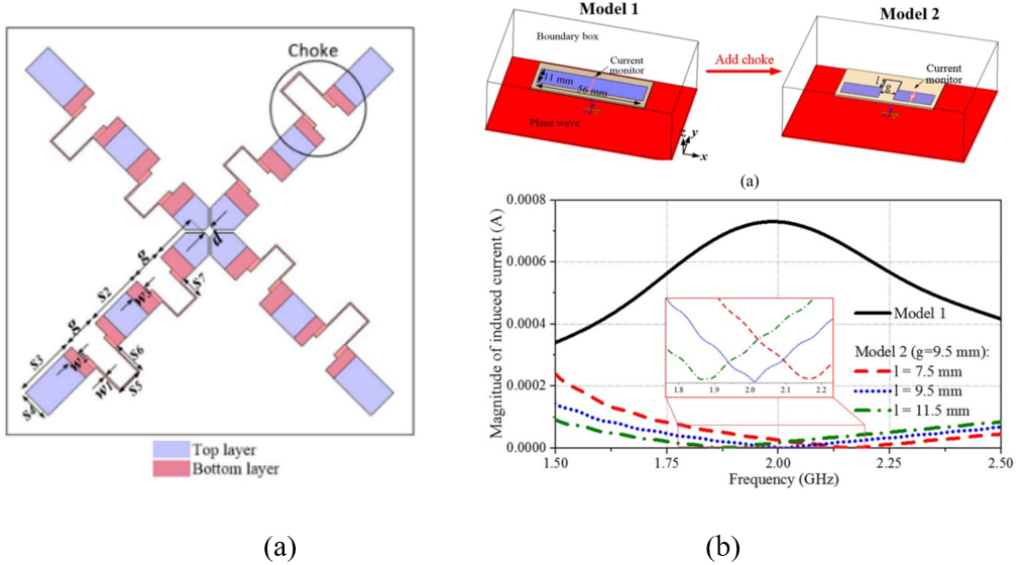


Figure 2.8: (a) LB dipole loaded with chokes, and (b) magnitudes of induced currents on the LB dipole before and after the loading.

As given in Figure 2.8 (a), loading chokes on LB dipoles is also an effective way for scattering suppression [36], [37]. Different from the introduction of reverse scattering fields or reversed induced currents in the above works, the chokes are introduced to suppress the HB-induced currents on the LB dipole. The radiator of the LB dipole is first divided into several sections, the length of which is shorter than the resonant length of the HB, so it is difficult for the HB wave to excite induced currents on the LB dipole when the HB dipole is working. However,

in order to ensure the normal flow of LB currents, the chokes composed of LC series-parallel circuits are loaded between these sections, and they present open to the HB and short to the LB. Thus, the HB scattering on the LB dipole is suppressed while the normal radiation of the LB dipole is maintained. After the loading, the magnitude of the HB-induced currents shown in Figure 2.8 (b) on the LB dipole is greatly reduced, so the HB patterns within 1.7–2.3 GHz are all well recovered. Although the introduction of chokes brings difficulties to the matching of the LB dipole, the LB dipole in this work can be matched and work normally from 0.82 to 1.0 GHz.

The spiral dipole [38] exhibited in Figure 2.9 (a) is actually a type of structure with its own chokes, in which the spiral structure itself presents open-circuit to the HB and short-circuit to the LB. The scattering suppression band can be adjusted by changing the pitch, radius, and slots of the spiral. The spiral in [38] has a good scattering suppression effect over the HB 3.3–3.7 GHz, and the HB patterns are well recovered. However, as shown in Figure 2.9 (b), the wider the scattering suppression bandwidth, the narrower the matching bandwidth of the LB dipole, and the working band of the LB spiral dipole is only 1.71 to 2.26 GHz after the loading. Therefore, a special impedance matching method is required for the spiral dipole to work in a wider band.

2.2.7 Frequency Selective Surface (FSS) and Metasurface

The frequency selective surface (FSS) exhibits transmission or reflection characteristics at specific frequency bands, making it a promising candidate for cross-band scattering suppression [39]–[43]. An FSS-based LB dipole operating in 1.8–2.7 GHz was first proposed in [39]. The square-ring-shaped radiator of the LB dipole is very similar to the unit cell of the FSS, so the LB radiator is designed and transformed into a band-pass FSS, as depicted in Figure 2.10 (a). The FSS has stop-band characteristics in the LB and pass-band characteristics in the HB. Such an FSS can make the LB dipole radiate normally and allow the HB wave to pass through with very weak reflection and scattering; therefore, the HB radiation patterns are not affected. Since the relative positions of the LB and HB dipoles are not fixed, the FSS should have good angular stability. The transmission coefficients of the proposed FSS remain stable from 0° to 60° over the HB 3.3–3.8 GHz, as given in Figure 2.10 (b). It is a clever design that integrates the design and simulation of the FSS and the LB radiator, resulting in a normally operating LB dipole antenna with suppressed HB scattering.

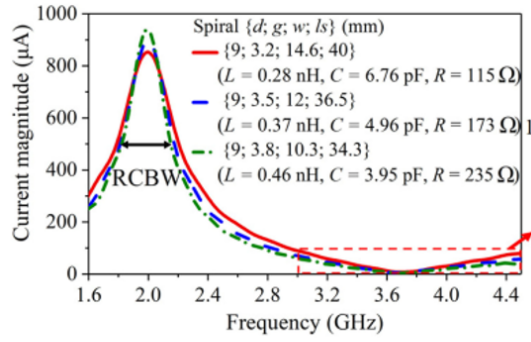
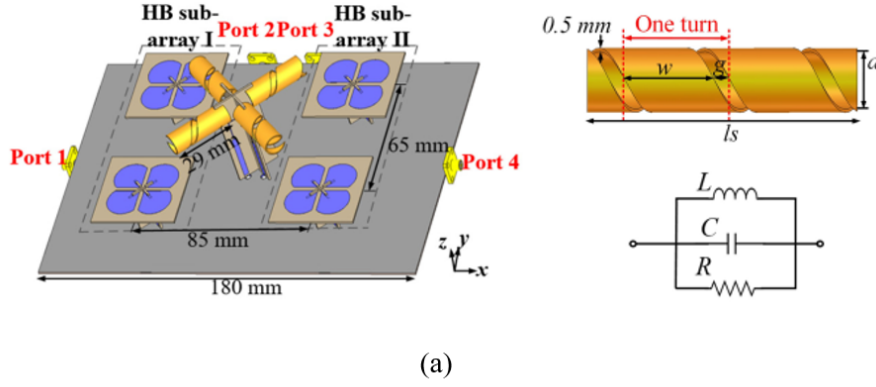


Figure 2.9: (a) Dipole array using the LB spiral dipole, and (b) current magnitude of the LB spiral dipole over the LB and HB.

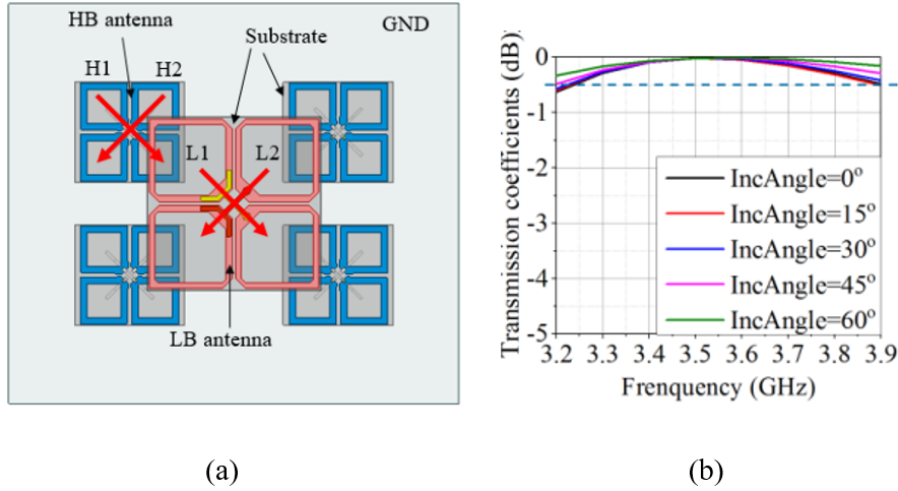


Figure 2.10: (a) Dual-band dipole array with the FSS-based LB dipole, (b) transmission coefficient of the FSS.

The FSS is only one typical representative of metasurfaces. In recent years, various types of metasurfaces have also been loaded onto the LB radiator or used directly as the LB radiator for HB scattering suppression. Examples include resistive metasurfaces [44], inhomogeneous

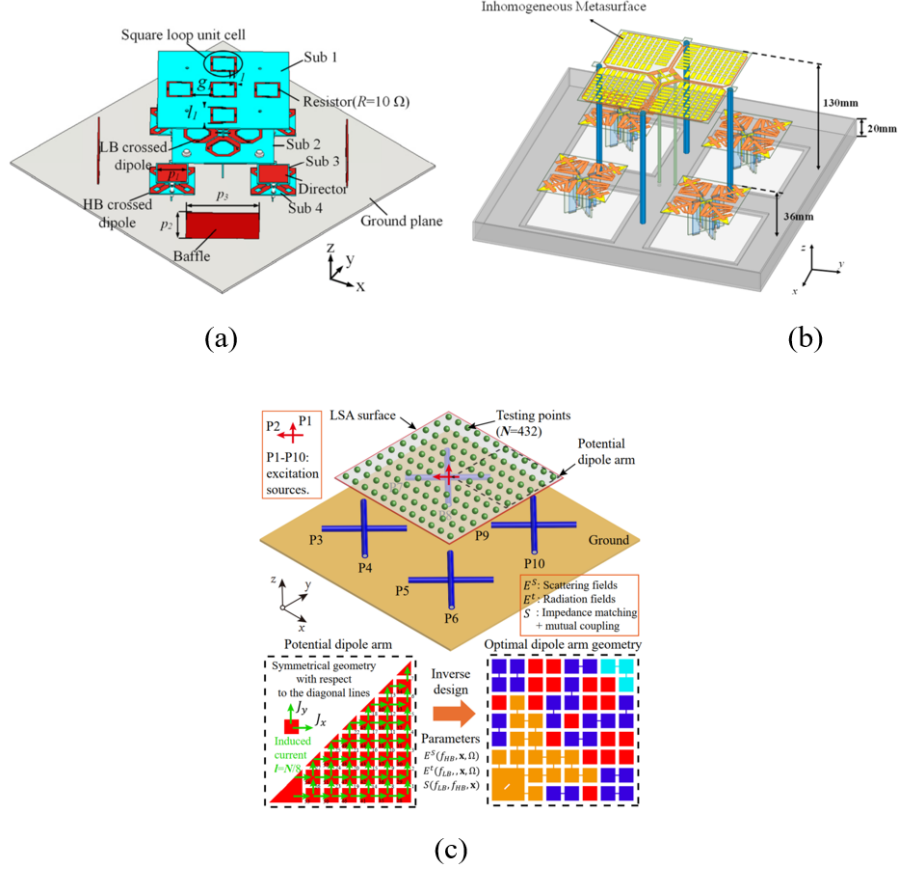


Figure 2.11: (a) Resistive metasurfaces, (b) inhomogeneous metasurfaces, and (c) pixelized metasurfaces co-designed with the LB antennas in the dual-band antenna arrays.

metasurfaces [45], and pixelized metasurfaces based on inverse design [46], as shown in Figure 2.11.

2.3 Suppression of Scattering from the HB Antenna Affecting the LB Radiation

The works mentioned above are all focused on suppressing the scattering on the LB dipoles, so that the HB dipoles are less affected and the HB patterns are recovered. Correspondingly, some scholars have noticed and analyzed the problem that the LB dipole is also affected by the HB dipole and proposed some solutions. At specific LB and HB frequencies, the HB dipole and the ground as a whole can resonate within the LB. The HB dipole can be equivalent to a top-loaded monopole placed perpendicular to the ground, with its overall size resonating within the LB. As a resonant structure, the HB dipole radiates LB waves outward, which makes the

patterns of the LB dipole show severe distortions.

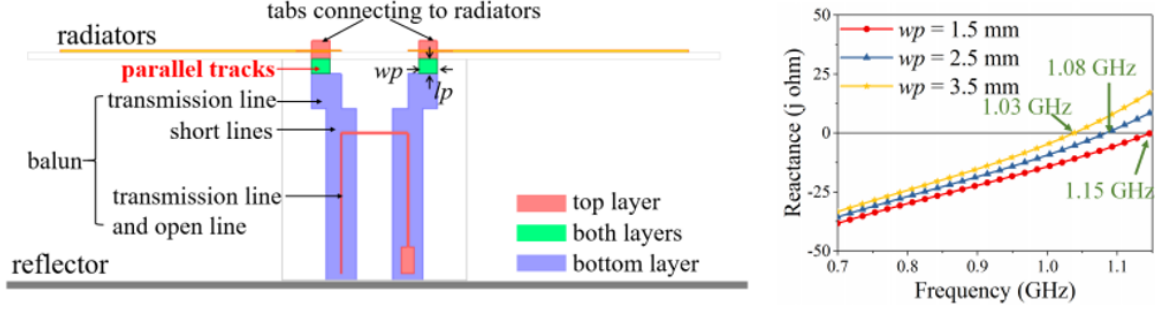


Figure 2.12: HB dipole loaded with capacitors, and reactance of the HB dipole over the LB.

In order to find and observe the resonance point, a port is added between the HB balun and the ground [47], and the impedance at the port is simulated as shown in Figure 2.12. When the reactance curve is equal to zero, it means that resonance occurs at this frequency point. To solve this problem, a capacitor is loaded between the HB radiator and the HB balun, as given in Figure 2.12. The loading of the capacitor moves the entire impedance curve to a higher frequency. When the resonance point is moved out of the LB, the HB dipole no longer resonates in the LB, and the LB patterns are restored. The loaded capacitor is composed of parallel patches etched on the substrate of the balun, but it destroys the original connected structure of the HB balun and brings great difficulties to the matching of the HB dipole. In addition, the resonance in the LB is actually periodic. When this resonance point is moved to a higher frequency and out of the LB, a new resonance point will enter the LB from the lower frequency. Therefore, the application of this method within a very wide LB may encounter difficulties. The working band of the LB dipole in this work is from 698 MHz to 960 MHz (31.6% relative bandwidth). After loading, the LB patterns are in very good agreement with the patterns when the LB dipole radiates alone. To move this resonance point out of the target LB, in addition to connecting a series capacitor to the HB antenna, methods such as connecting a series inductor [48], [49] or a parallel LC circuit [50] between the HB balun and the ground can also be adopted.

In [51], ferrite cores are loaded on the feeding cables of the HB dipole, as shown in Figure 2.13, to suppress the resonance and scattering from the HB dipole. When the LB-induced currents pass through the loaded ferrite cores, the ferrite cores are repeatedly magnetized under the external magnetic field generated by the induced currents. Through this repeated magnetization, part of the electromagnetic energy is converted into thermal energy by the ferrite cores, so that

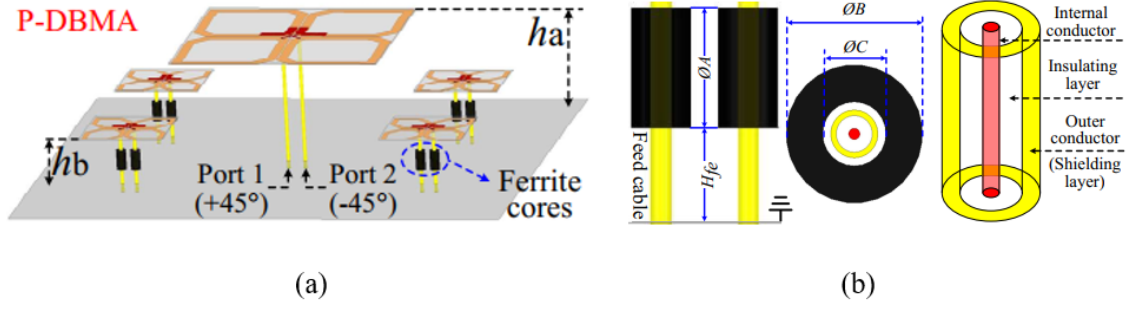


Figure 2.13: (a) Dual-band dipole array with the ferrite-loaded HB dipole, and (b) loaded ferrite cores around feed cables.

the energy re-radiated by the HB dipole is greatly reduced. Moreover, the ferrite cores only affect the induced currents on the outer surface of the coaxial cables, so they do not affect the normal feeding and signal transmission of the HB dipoles. In addition, the ferrite cores have the advantage of a wide scattering suppression band, but their assembly is relatively complex and the cost is high.

2.4 Suppression of Cross-Band Coupling Between the LB and HB Antennas

To suppress the cross-band coupling between patch antennas operating in different frequency bands, filtering structures [52]–[55] such as slots and open-loop resonators are employed, as shown in Figure 2.14. These filtering structures can prevent cross-band induced currents from entering the ports through the radiators of the patch antennas, which would otherwise degrade the isolation performance.

Similarly, cross-band coupling suppression between dipole antennas can also be achieved by employing various filtering structures [56]–[60]. As shown in Figure 2.15, parasitic loops are loaded above the radiators, while open stubs are integrated into the balun feeding network. The inclusion of these filtering structures does not adversely affect the radiation characteristics of the antennas. However, their individual filtering bandwidths are relatively narrow, so multiple filtering structures are required to work together to achieve a wide filtering band. For example, the HB antenna shown in Figure 2.16 is loaded with multiple parasitic loops to achieve coupling suppression over a wider band [16].

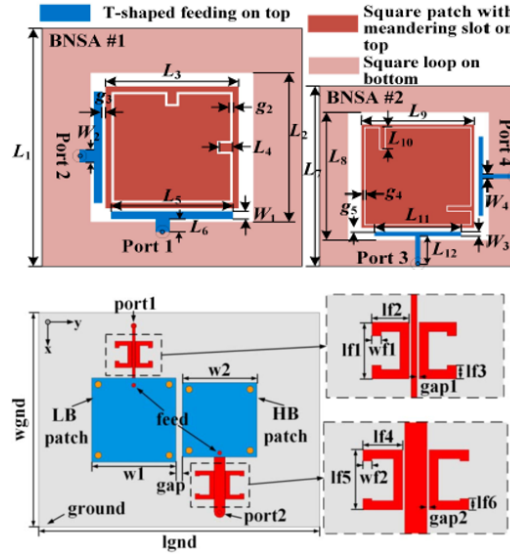


Figure 2.14: Slots and open-loop resonators are loaded on the patch antennas to improve the cross-band isolation.

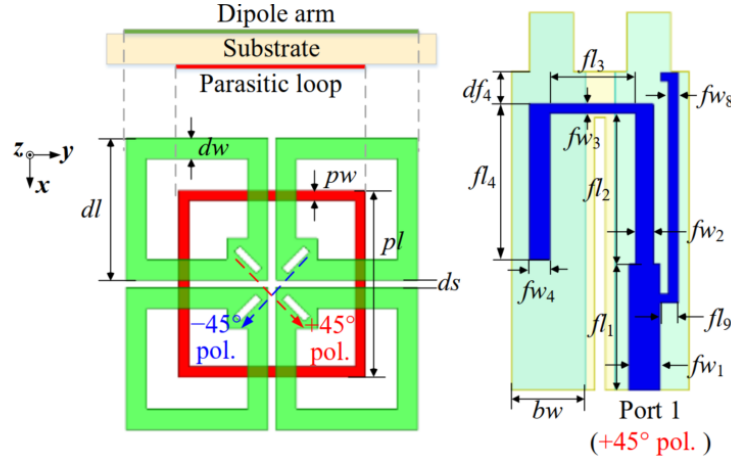


Figure 2.15: Parasitic loops and open stubs are loaded on the dipole antennas to improve the cross-band isolation.

2.5 Suppression of In-Band Coupling Between the HB Antennas

In dual-band shared-aperture antenna arrays, besides the necessity of suppressing cross-band coupling between the LB and HB elements, the in-band coupling among the HB antennas also needs to be suppressed. Conventional techniques for reducing mutual coupling typically involve introducing baffles, slots, or other isolating structures between antenna elements. In addition, several self-decoupling methods proposed for MIMO patch antenna arrays can be effectively

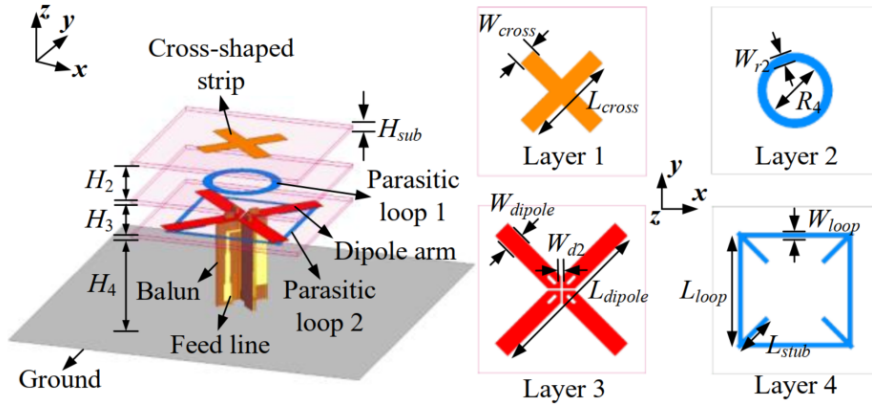


Figure 2.16: HB antenna loaded with parasitic loops for wideband cross-band coupling suppression.

adopted in the multi-band shared-aperture configurations [61], [62]. Furthermore, the use of differential feeding offers a promising solution for achieving extremely high isolation between closely spaced antenna elements [63]–[66].

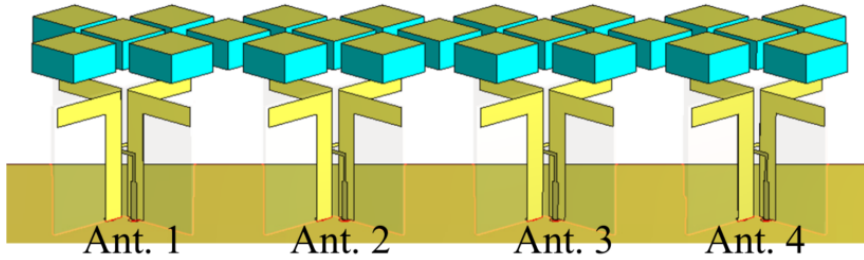


Figure 2.17: Partially reflective surface (PRS) is loaded above the antenna array to suppress in-band coupling.

For dual-polarized cross-dipole arrays, introducing a partially reflective surface (PRS) shown in Figure 2.17 above the array has proven to be an effective method for reducing mutual coupling [67]–[71]. The PRS partially reflects the radiated waves, and these reflected waves, being out of phase with the original coupling waves between adjacent elements, can cancel them out, thereby achieving effective in-band coupling suppression.

In addition, bending the arms of the dipole antenna is a simple and effective method for suppressing in-band coupling [67], [72], as depicted in Figure 2.18. Bending the two arms of a dipole antenna introduces additional current paths that effectively transform the original electric dipole into a combined electric and magnetic dipole structure. The bent or meandered sections generate magnetic dipole components that, together with the main electric dipole,

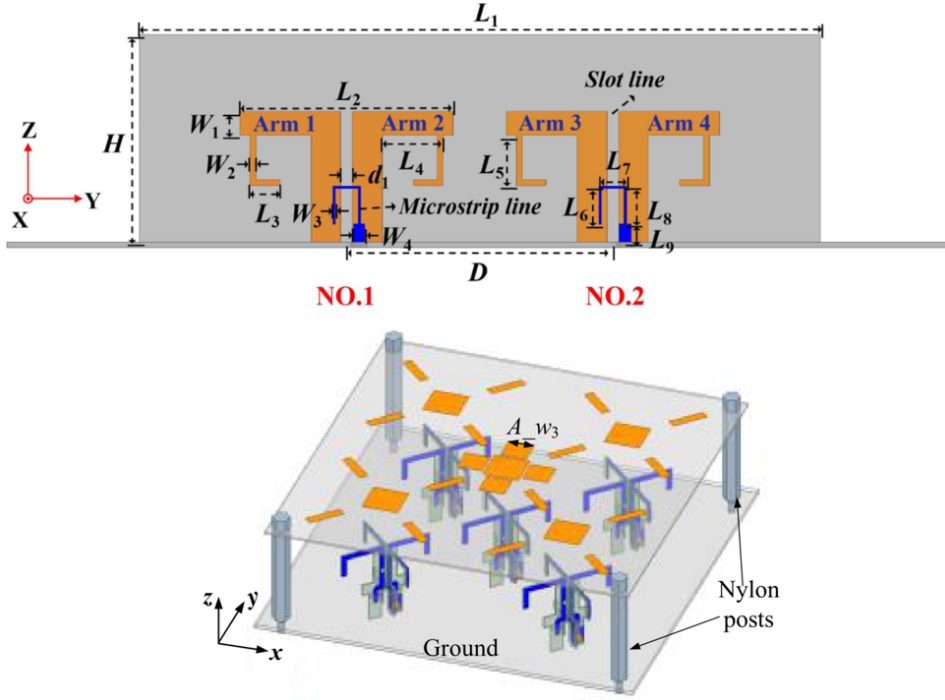


Figure 2.18: Dipole antennas with bent arms for in-band coupling suppression.

create multiple coupling paths between adjacent elements. When the amplitudes of these coupling currents are nearly equal but out of phase, they cancel each other, significantly reducing mutual coupling within the operating band. By carefully adjusting the bending geometry, such as the position and length of the meander lines, the phase and magnitude of these induced currents can be optimized to achieve in-band coupling suppression without requiring any external decoupling structure.

2.6 Simultaneous Suppression of Scattering and Coupling

The suppression structures introduced above for shared-aperture antenna arrays are mainly single-functional, as they can only be used to suppress either scattering or coupling. Several dual-functional suppression structures deserve to be highlighted, as they are capable of simultaneously suppressing both scattering and coupling [30], [32], [73]–[75].

The folded-dipole [32] and U-shaped [30] structures of the LB antennas, as shown in Figure 2.19, allow HB wave propagation with minimal interference and also prevent HB-induced currents from flowing into the LB ports, thus simultaneously suppressing cross-band scattering and

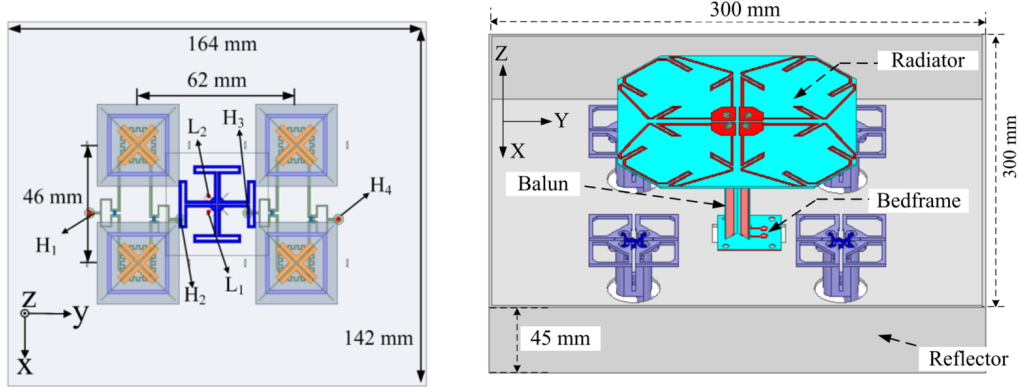


Figure 2.19: Folded-dipole and U-shaped structures of the LB antennas exhibiting dual-functional suppression characteristics.

coupling.

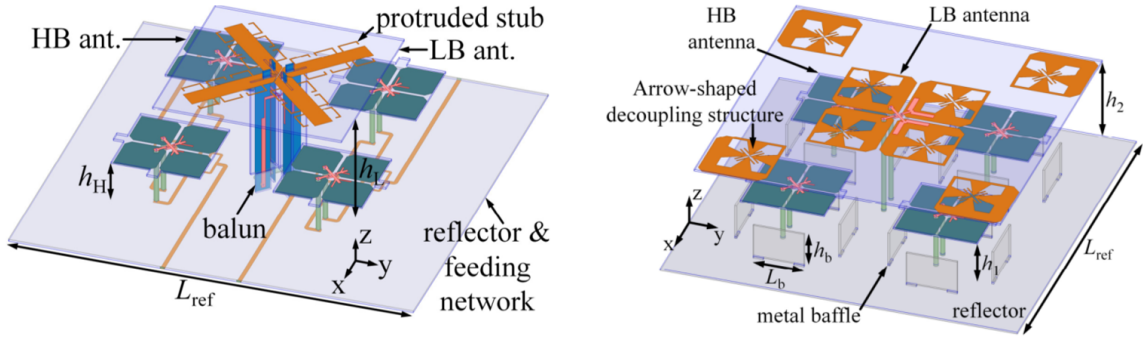


Figure 2.20: LB radiators co-designed with partially reflective surfaces exhibiting dual-functional suppression characteristics.

As shown in Figure 2.20, the LB radiators [74], [75], co-designed with partially reflective surfaces (PRSs), allow most of the HB waves to pass through. The HB waves partially reflected by the surface, being in opposite phase, cancel out the original HB coupling waves between HB antennas, thereby achieving both cross-band scattering and in-band coupling suppression at the same time.

2.7 Suppression of Scattering and Coupling by Modifying the Configuration of the Antenna Array

The antennas mentioned above all adopt a traditional configuration, with the dipoles positioned above a common ground plane. In recent years, several researchers have achieved suppression

of scattering and coupling by modifying the configuration of the antenna array.

2.7.1 Stacked Configuration

As exhibited in Figure 2.21 (a), an FSS layer is added above the LB dipole, while the HB dipoles are placed on the FSS layer [76]. The proposed FSS can reflect the HB wave and transmit the LB wave at the same time, as illustrated in Figure 2.21 (b), so the FSS can serve as the ground of the HB dipoles, meanwhile without affecting the radiation of the LB dipole below. In addition, the FSS is also a capacitive loading for the LB dipole, which can reduce the profile of the LB dipole while ensuring that the LB dipole operates in a wide band. Owing to the blocking of the HB wave by the FSS, the HB waves radiated by the HB dipoles cannot reach the LB dipole located below, so the patterns of the HB dipoles are not affected by the LB dipole. In this novel configuration [77], [78], good isolation and patterns can naturally be obtained. It is an effective approach to simultaneously suppress scattering and coupling; however, this configuration increases structural complexity, assembly difficulty, and cost, while reducing overall structural stability.

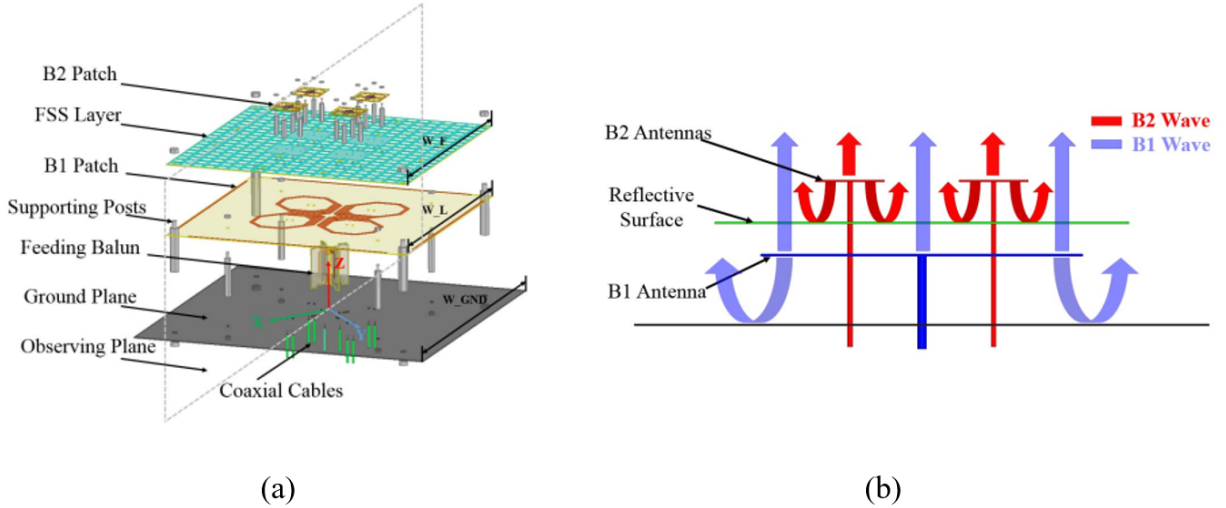


Figure 2.21: (a) The configuration of HB dipoles above the FSS and the LB dipole below the FSS. (b) The reflection and radiation of LB and HB waves.

This modified configuration, with the HB antennas placed above and the LB antennas below, has been widely adopted to realize shared-aperture antenna arrays. Different from inserting an FSS between the LB and HB antennas, an Indium Tin Oxide (ITO) surface [79], [80] or other structures [81], [82] are introduced to this stacked configuration to further enhance the scattering and coupling suppression effect.

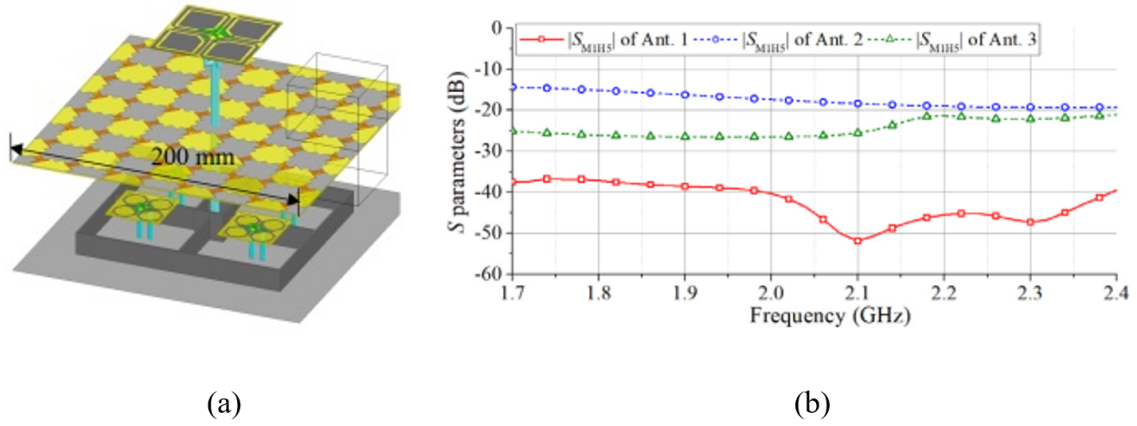


Figure 2.22: (a) The configuration of LB dipole above the FSS and HB dipoles below the FSS, and (b) isolations between the co-polarized LB and HB dipoles within the LB.

Contrary to the above articles in which the HB dipoles are placed above the LB dipole, the LB dipole is placed above the four HB dipoles in [83], [84], as shown in Figure 2.22 (a). An FSS is also inserted between the LB and HB dipoles, exhibiting high-pass characteristics for the HB band and stop-band characteristics for the LB band. Therefore, the FSS acts as a reflector for the LB dipole while simultaneously ensuring normal radiation of the HB dipoles. Compared with the traditional configuration where the LB and HB dipoles share the same ground plane, the isolation between the co-polarized LB and HB dipoles within the LB band is improved from 14 dB to 35 dB, as illustrated in Figure 2.22 (b). This improvement occurs because the LB waves radiated by the LB dipole cannot reach the port of the HB dipoles due to the blockage caused by the FSS. To prevent the radiation patterns of the HB dipoles from being affected by the LB dipole located above, the LB radiator is designed as an FSS with transmissive characteristics to the HB waves, which is the same method as that used in [39].

2.7.2 Coplanar Configuration

Apart from the stacked configuration, the coplanar configuration [85]–[88] has also been proposed to suppress the scattering and coupling within the antenna array as presented in Figure 2.23.

The HB dipoles working in 3.3-3.8 GHz are embedded in the loop arms of the LB dipole working in 0.69-0.96 GHz, and the radiators of the HB and LB dipoles are in the same plane [85]. In order to ensure the normal radiation of the HB dipoles and to achieve the isolation of the radiation space between the HB dipoles and the LB dipole, metal back cavities are loaded

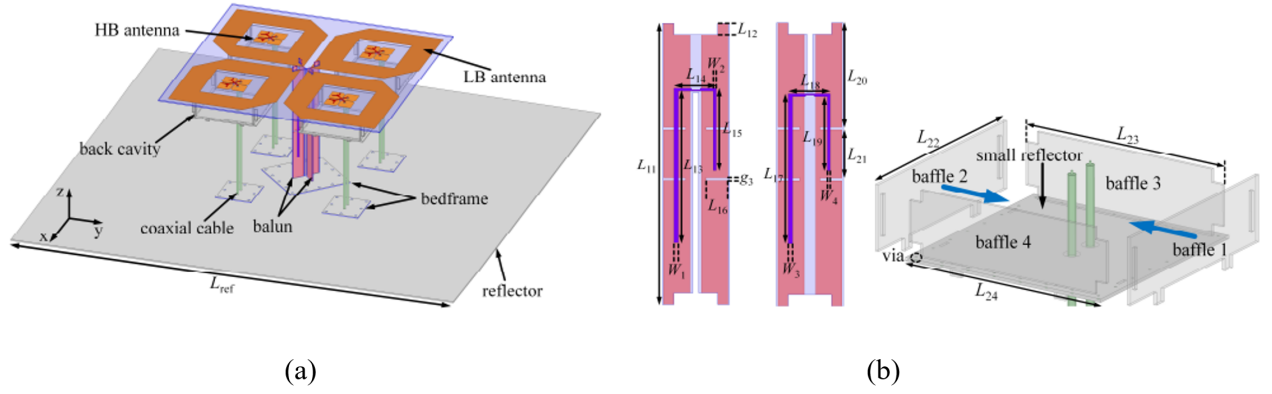


Figure 2.23: (a) Dual-band antenna array with coplanar configuration, (b) slotted LB balun and back cavity.

under the HB dipoles. The HB dipoles and the cavities are electrically small to the LB, so the radiation of the LB dipole is not affected. In addition, the open slots are loaded on the LB balun as a filtering structure, which can improve the isolation between the LB and HB dipoles within the HB. Through adopting the above configuration and loadings, the LB and HB dipoles obtain good isolation, and the radiation patterns of LB and HB dipoles remain stable. In other studies [86]–[88], a similar coplanar structure for the LB and HB antenna arrays is also adopted, effectively mitigating the cross-band scattering problem compared with the traditional side-by-side configuration.

2.7.3 Bowl-shaped Configuration

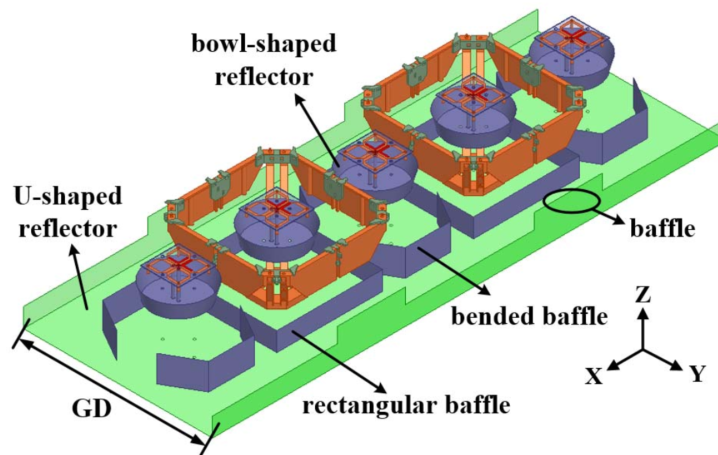


Figure 2.24: Dual-band shared-aperture array with a bowl-shaped configuration.

The bowl-shaped configuration [89]–[92], as shown in Figure 2.24, is also a common configura-

tion in dual-band antenna arrays, where four LB elements are symmetrically placed around one HB element. This symmetric configuration minimizes the physical blockage of the HB radiation by the LB elements, allowing the HB beams to radiate directly through the open bowl region. Moreover, the fourfold rotational symmetry leads to the cancellation of the scattered fields from different LB elements, further reducing the secondary radiation effects in the HB.

2.8 Suppression of Scattering and Coupling in the Tri-Band Antenna Array

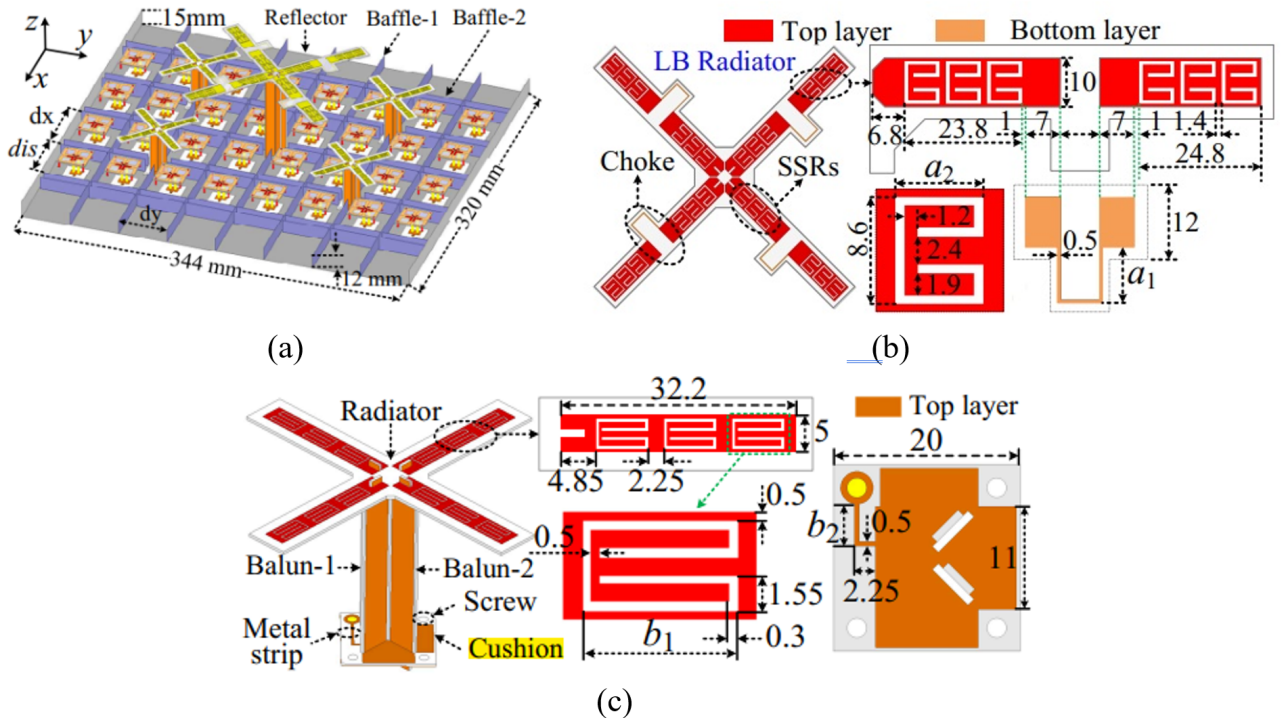


Figure 2.25: (a) Configuration of the tri-band antenna array, (b) LB dipole loaded with slots and chokes, and (c) MB dipole loaded with slots and a metal strip.

Compared to the dual-band dipole array, the scattering and coupling in the tri-band antenna array are more complex and the number of dipoles arranged in the limited space is also higher. Due to their electrical size, the HB dipoles are simultaneously affected by the scattering fields generated by the LB and MB (middle band) dipoles, while the MB dipoles are affected by the scattering fields generated by the LB dipoles. In addition, when the lower band dipoles are radiating, the higher band dipoles may resonate together with the ground in the lower bands. In summary, cross-band scattering and coupling may occur between any two of the

three antennas. At present, most tri-band shared-aperture arrays adopt the various scattering and coupling suppression techniques developed for dual-band arrays.

As shown in Figure 2.25, the tri-band dipole array [93] simultaneously employs the above-mentioned methods to suppress scattering. Chokes and slots are loaded onto the LB dipole operating in 0.79–0.96 GHz to suppress MB (1.71–2.17 GHz) and HB (3.4–3.6 GHz) scattering, while the slots loaded on the MB dipole are used to suppress HB scattering. In addition, a metal strip is inserted between the balun of the MB dipole and the ground, functioning as a loaded inductor. This modification shifts the resonant frequency out of the LB, preventing the MB dipole and the ground from resonating within the LB. After these loadings, the distorted radiation patterns of the LB, MB, and HB dipoles are all restored; however, the HB patterns still exhibit noticeable distortion compared with those of the standalone HB dipole. This is because the HB dipole is simultaneously affected by the scattering generated from one LB dipole and four MB dipoles, and these multiple scattering fields are not completely suppressed. Therefore, for HB scattering suppression in tri-band dipole arrays, it is necessary to adopt a more effective method that provides improved mitigation across a wider band.

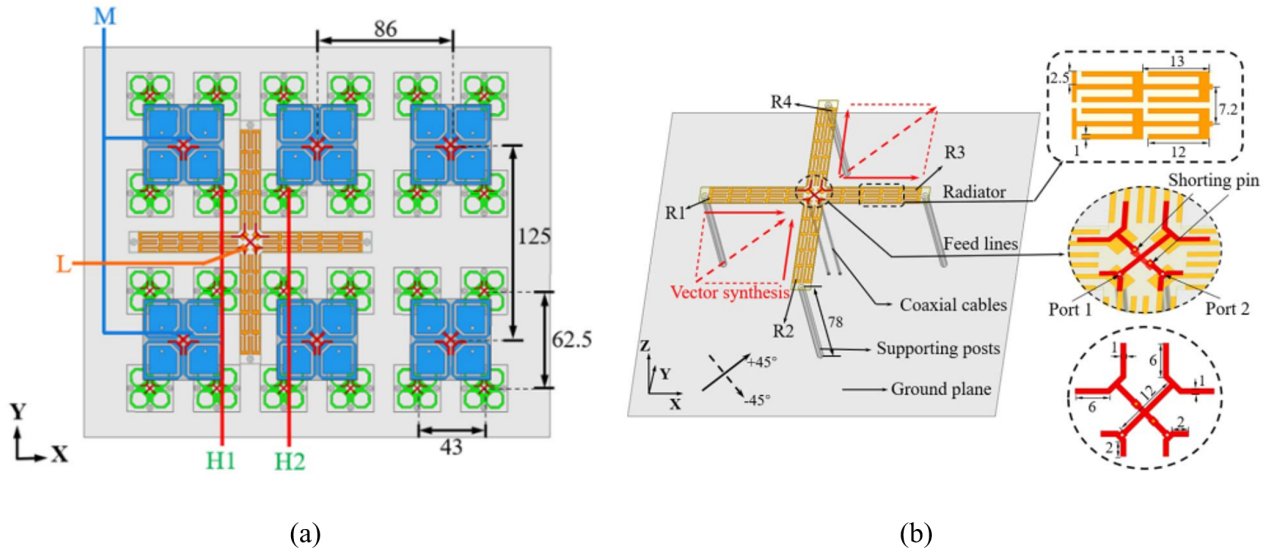


Figure 2.26: (a) Configuration of the tri-band antenna array and (b) rotated LB dipole with slots.

For the tri-band dipoles in [94], scattering suppression is achieved through the use of loaded slots, an FSS-based radiator, and a rotated radiator. As illustrated in Figure 2.26, comb-shaped slots are etched on the LB dipole operating at 0.76–0.88 GHz, and the LB dipole is rotated by 45° about its center. The $\pm 45^\circ$ polarized waves are radiated by the co-excitation of two adjacent

LB arms along the x - and y -directions. The MB radiator is designed as an FSS, which provides a passband for the HB waves. However, the scattering suppression bandwidth achieved by the loaded slots is narrow, making it difficult to simultaneously cover the MB (1.9–2.7 GHz) and HB (3.3–3.9 GHz) bands. Consequently, the rotation of the LB dipole plays a decisive role in achieving wideband scattering suppression. After rotation, the LB dipole radiator is no longer directly above the MB and HB dipoles, and the LB radiators are no longer aligned with the polarization directions of the MB and HB waves, thereby significantly reducing the scattering field reradiated from the LB dipole radiator [95].

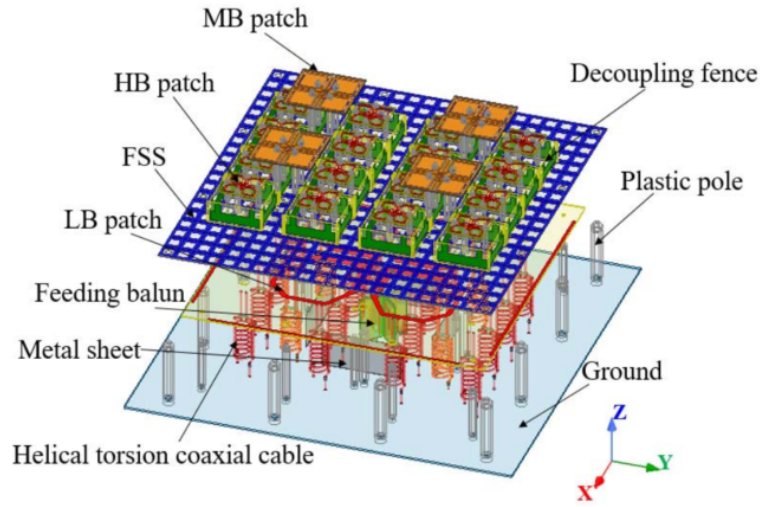


Figure 2.27: Configuration of the tri-band dipole array with FSS layer, helical torsion coaxial cables, and FSS-based radiators.

Tri-band dipole arrays based on a stacked configuration have been reported in [96], [97], and [98]. As illustrated in Figure 2.27 [96], an FSS is inserted between the LB (0.69–0.96 GHz) dipole and the MB/HB (1.8–2.7 GHz / 3.3–3.8 GHz) dipoles, while the MB radiator itself is designed as an FSS. In addition, helical torsion coaxial cables are co-designed with the feeding lines of the MB and HB dipoles. The above-mentioned methods and their corresponding suppression mechanisms have been described in detail earlier and are therefore not repeated here.

In tri-band antenna arrays, using an FSS with dual transmission bands (MB and HB) as the LB radiator is also an effective method [99]. The FSS-based LB radiator does not affect the normal radiation of the MB and HB antennas. As given in Figure 2.28, placing the MB and HB antennas on both sides of the FSS-based LB antenna [100], [101], or integrating the MB and HB antennas into a dual-band antenna [102], can effectively avoid other scattering and coupling problems within the tri-band array. Based on the same design concept of multi-

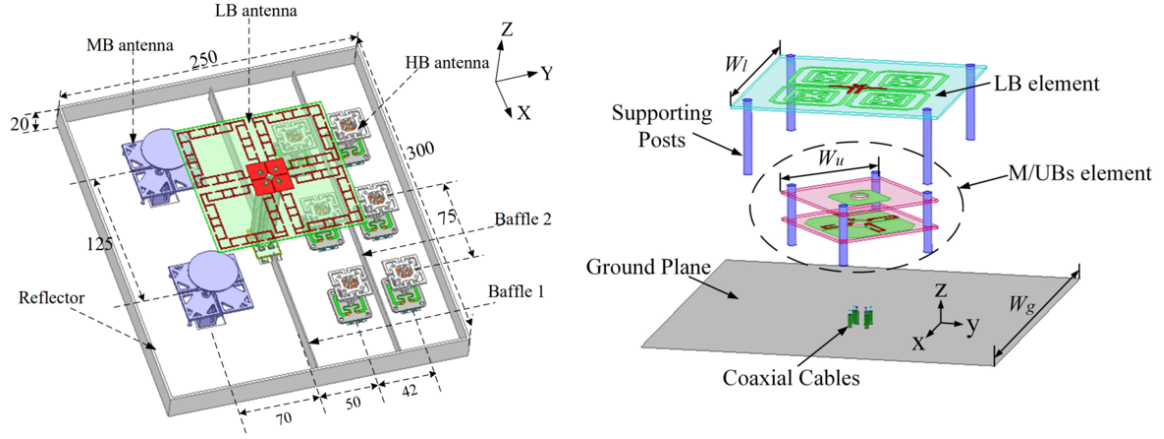


Figure 2.28: Configuration of the tri-band dipole array with FSS-based LB antenna.

band transmission FSSs, FSSs with triple transmission bands have also been employed as LB radiators in the design of quad-band antenna arrays [103], [104].

2.9 Sub-6 GHz and Millimeter-Wave Shared-Aperture Antenna Arrays

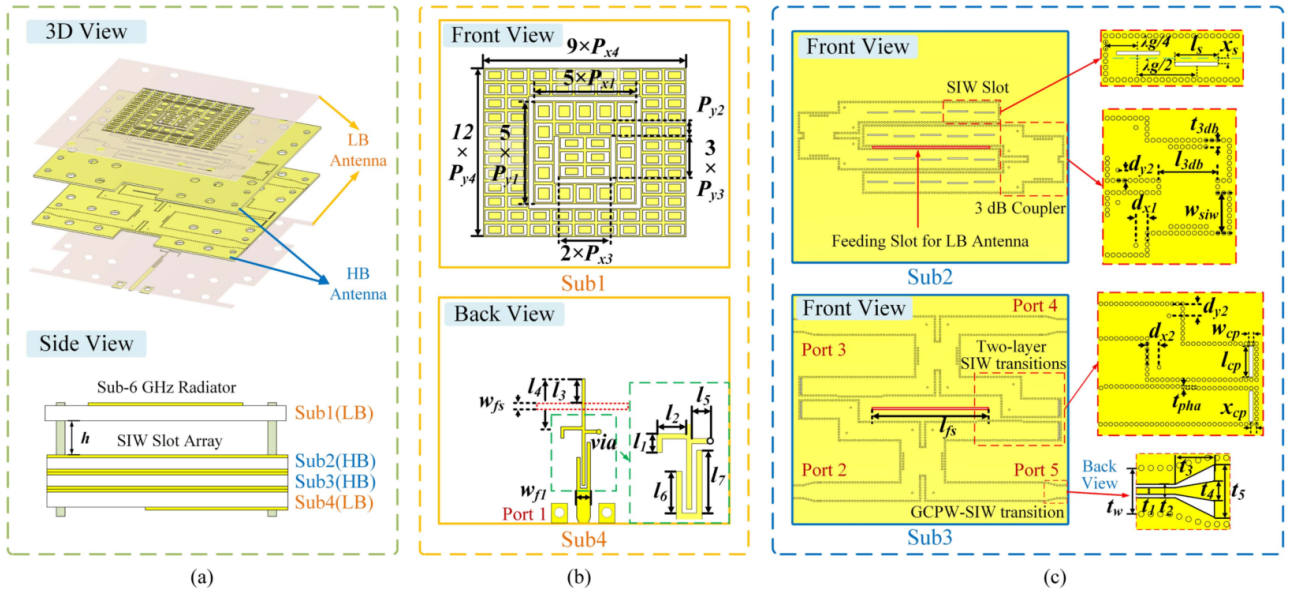


Figure 2.29: Geometry of the Sub-6 GHz and millimeter-wave shared-aperture antenna array.

(a) Total configuration. (b) LB antenna. (c) HB antenna.

The shared-aperture multi-band antenna arrays discussed above only focus on the Sub-6 GHz bands. In recent years, many studies have reported shared-aperture designs combining Sub-6 GHz and millimeter-wave (mainly 25–30 GHz) antennas [105], [106], [107], [108], as shown in

Figure 2.29. The millimeter-wave antenna provides wideband, high-speed data transmission, whereas the Sub-6 GHz antenna ensures wide-area coverage and reliable connectivity. Due to the large separation between the two target frequency bands, methods similar to those used for shared-aperture antennas within the Sub-6 GHz range can be employed to mitigate scattering and coupling issues, such as placing the millimeter-wave antennas above the Sub-6 GHz antennas. Owing to the significant difference in physical dimensions between the two frequency bands, some antenna structures can serve dual functions. For instance, the ground plane of a millimeter-wave antenna array can act as a radiator or a parasitic element for the Sub-6 GHz antenna, while the array elements of the millimeter-wave antenna can function as a metasurface operating in the Sub-6 GHz band.

2.10 Conclusion

This chapter reviews the scattering and coupling suppression techniques used in multi-band shared-aperture antenna arrays to restore radiation patterns and improve isolation. The reviewed works are mainly focused on dual-band and tri-band shared-aperture antenna arrays. Scattering suppression can be classified into that on the LB and HB antennas, while coupling suppression is divided into cross-band and in-band categories. Most of these suppression techniques are single-functional, meaning they can only mitigate either scattering or coupling. Modifying the antenna array configuration can simultaneously avoid both scattering and coupling issues, but this approach often introduces structural complexity, assembly challenges, and reduced mechanical stability.

The scattering and coupling issues in tri-band shared-aperture antenna arrays are more complex than those in dual-band arrays. Current tri-band works generally combine multiple suppression techniques proposed for dual-band antenna arrays; however, the use of various suppression structures leads to highly complex array configurations, while the achieved suppression bandwidth remains limited.

Therefore, the reported works still face challenges such as the inability to simultaneously achieve scattering and coupling suppression over a wide bandwidth, as well as the structural complexity and the trade-off between suppression and matching bandwidth. This thesis addresses these problems through several novel approaches and successfully achieves simultaneous wideband scattering and coupling suppression in both dual-band and tri-band shared-aperture antenna

arrays.

Chapter 3

Simultaneous Suppression of Scattering and Coupling in Dual-Band Antenna Array

3.1 Introduction

In this chapter, a holistic method is proposed to mitigate all three kinds of cross-band interactions, namely the scattering of LB antennas on HB radiation, the scattering of HB antennas on LB radiation, and the mutual coupling between HB and LB antennas. This is achieved by loading a 2.5-dimensional (2.5-D) cloak to the LB antenna, defected ground structures (DGSs) to the LB antenna balun, and resonance shifters to the HB balun. The efficacy of the proposed descattering and decoupling methods is demonstrated using a “1 LB + 1 HB” dual-band dual-polarized array to eliminate potential effects arising from array topology. The methods are also tested in a more practical “1 LB + 4 HB” array configuration. The results show that, in addition to the suppression of cross-band scattering and coupling, these methods also reduce in-band couplings between adjacent HB antennas, which is highly desirable in the collocation of multi-band antennas. In the proposed array, the radiation pattern distortions in both the LB and HB are substantially mitigated. The co-polarized and cross-polarized cross-band isolations achieve 23 dB and 28 dB, respectively, and the in-band isolation between the HB antennas exceeds 23 dB.

This “1+1” configuration consists of only one LB antenna and one HB antenna placed side-by-

side with an arbitrary relative position. One should note that other reported studies usually include two or more HB antennas. However, exciting more than one antenna intrinsically provides stronger anti-scattering ability than a single antenna. In contrast, when studying the scattering and coupling effects in this “1+1” configuration, the effects of the array topology can be excluded. Therefore, the proposed method can be applied to any complex antenna platforms to reduce the cross-band interactions.

The chapter is organised as follows. Following the introduction, Section 3.2 presents a method to load the LB (1.7-2.5 GHz) dipole radiators with a multi-functional 2.5-D cloak to suppress its cross-band scattering and coupling to HB radiation at 4.4-5.0 GHz. In Section 3.3, the influence of LB dipoles’ baluns on the HB radiation is identified for the first time. The resultant cross-band scattering and coupling are suppressed by etching DGSs on the baluns. In Section 3.4, to mitigate the negative effect on the LB radiation caused by the HB antennas, resonance shifters are introduced in the HB antennas. Section 3.5 evaluates the robustness of the suppression methods in two array configurations and various relative positions, while also presenting their suppression effects on in-band coupling between adjacent HB antennas. Experimental results are given in Section 3.6 and conclusions are drawn in Section 3.7.

3.2 Reducing the Cross-Band Scattering and Coupling to HB Radiation by Cloaking LB Radiator

3.2.1 Scattering Suppression

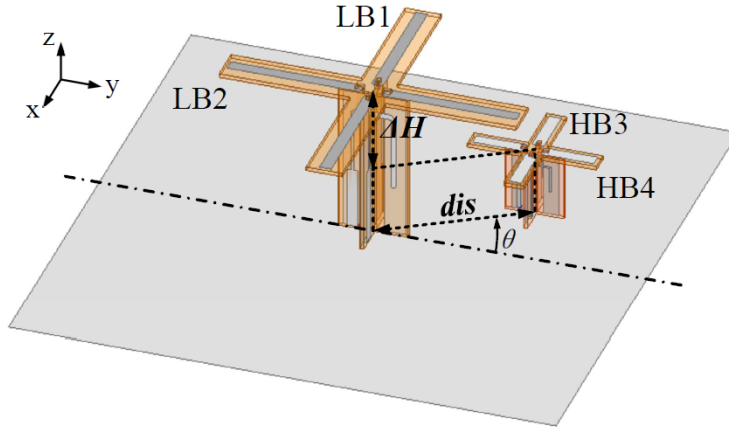


Figure 3.1: “1 LB + 1 HB” dual-band array configuration.

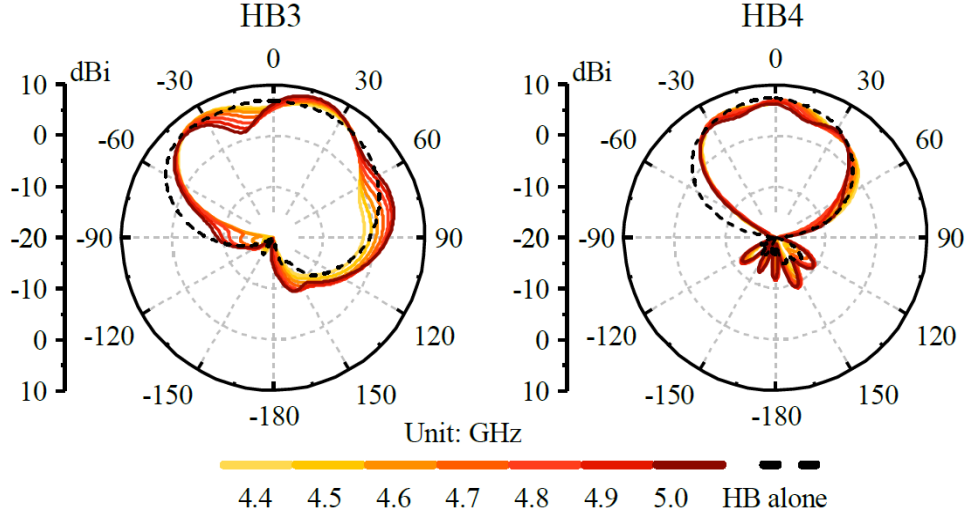


Figure 3.2: Radiation patterns in the yo z -plane of HB3 and HB4 when the LB dipole is present ($dis = 40$ mm, $\theta = 0^\circ$ and $\Delta H = 22$ mm).

When two antennas working in different frequency bands are closely placed together as shown in Figure 3.1, they interact with each other and degrade each other's performance. Due to their size and height differences, the LB antenna's scattering to the HB radiation is the dominant issue.

As shown in Figure 3.2, the HB radiation patterns exhibit severe distortions in the yo z -plane compared to the case when the HB antenna is radiating alone. To address this issue, a 2.5-D cloak made by patches and vias is loaded on the arms of the LB dipole, as shown in Figure 3.3. It has a similar working mechanism to the traditional 3-D mantle cloak that can achieve scattering cancellation [6], but with a much simpler and lighter structure. The 2.5-D cloak can achieve three objectives at the same time, i.e., suppressing cross-band scattering, cross-band coupling, and in-band coupling, which has not been reported in the published mantle cloaks or other loading structures.

Figure 3.3(a) gives the geometry of the original LB dipole that operates from 1.7 to 2.5 GHz. In Figure 3.3(b), the strip radiators and the loaded patches of the modified dipole are printed on the two sides of an FR4 substrate with a relative permittivity of 4.4, a loss tangent of 0.02, and a thickness of 1 mm. Metallic vias are employed to connect each loaded patch with the strip radiator to effectively choke the induced HB currents.

The working mechanism of the proposed 2.5-D cloak can be revealed by illuminating the modified LB dipole with x -polarized plane waves (parallel to the dipole) at the HB and observing

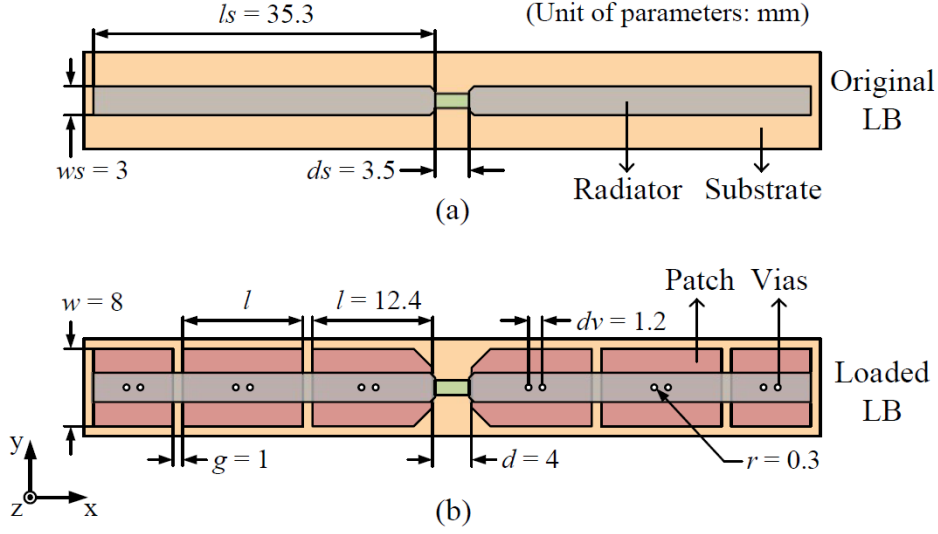


Figure 3.3: Geometry of (a) the original LB radiator and (b) the loaded LB radiator to suppress the scattering at 4.4-5.0 GHz.

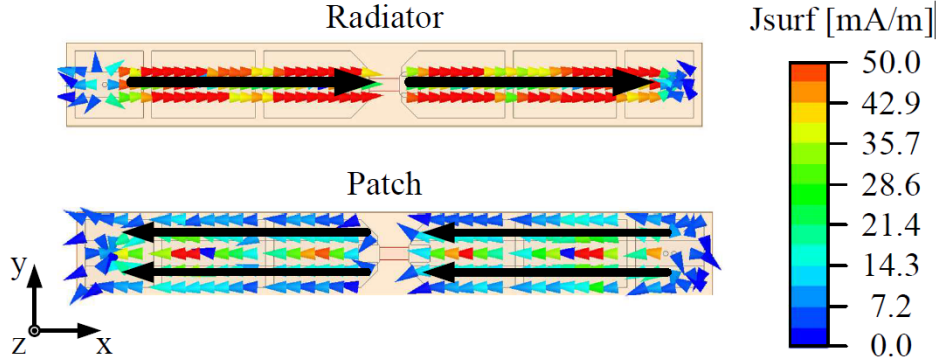


Figure 3.4: Current distributions on the strip radiators and loaded patches of the loaded LB dipole at 4.8 GHz.

the induced HB currents. Figure 3.4 depicts the surface currents on the strip radiator and the patches of the loaded LB dipole at 4.8 GHz. It is observed that reversed currents are induced on the strip radiators and the patches at the HB. By optimizing the geometry of the loaded dipole, the induced HB currents on the two sides can largely cancel each other, thus realizing scattering suppression.

When HB currents appear on the LB dipole, the LB dipole is excited with higher-order modes. To provide more insights, a characteristic mode analysis (CMA) [109], [110] is conducted. The modal significances (MSs) of the original and loaded LB dipoles are presented in Figure 3.5. As seen from Figure 3.5(a), there are three higher order modes with MSs that are higher than

0.4, indicating that they are very easily excited by an external source operating at the HB, e.g., a HB dipole nearby. As a consequence, the secondary-radiation (scattering) field of these excited modes will destroy the original current distribution on the HB. After loading the LB dipole with the 2.5-D cloak, however, all the higher order modes are effectively suppressed, as evidenced by the remarkably lower MSs for the modified LB dipole shown in Figure 3.5(b), thus avoiding re-radiation and pattern distortion.

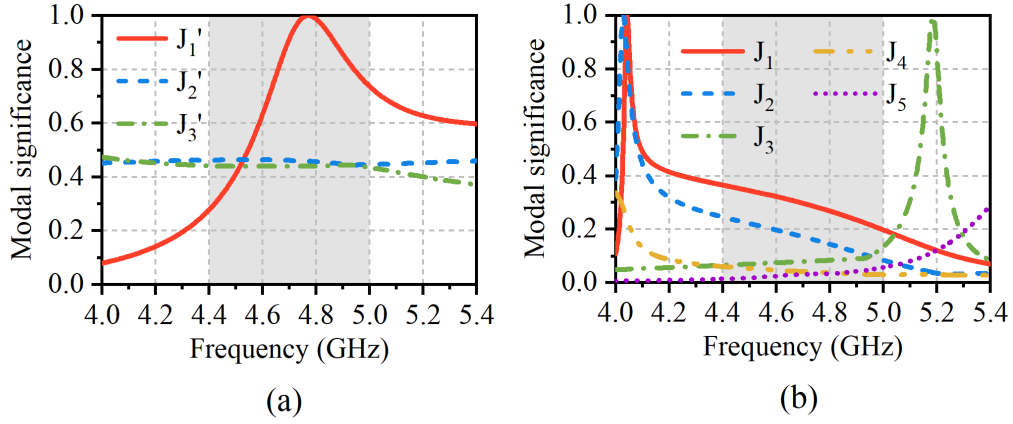


Figure 3.5: Modal significance of (a) the original LB dipole, (b) the loaded LB dipole.

To demonstrate how the descattering performance is improved by adding the 2.5-D cloak, Figure 3.6 compares the monostatic and bistatic radar cross sections (RCS) of the LB dipole with different loading conditions. It can be concluded from Figure 3.6(a) that adding both the patches and vias of the 2.5-D cloak is critical to the RCS reduction. As illustrated in Figure 3.6(b), the presence of the 2.5-D cloak significantly reduces the amount of energy scattered by the radiator in all directions.

The cross-dipole loaded with the 2.5-D cloaks is depicted in Figure 3.7. Additional L-shaped structures are utilized for impedance matching [111], [112]. Figure 3.8 plots the HB radiation patterns in the yo z -plane with the presence of the loaded LB antenna, when $dis = 40$ mm, $\theta = 0^\circ$, and $\Delta H = 22$ mm. Compared to the results shown in Figure 3.2, it can be found that the obvious pattern distortions are well eliminated after loading the 2.5-D cloak, and the restored patterns are almost the same as the patterns of the HB radiators when they are working alone. Please note that for the sake of conciseness, the xo z -plane radiation patterns of the HB antenna are not included here as they are less affected by the blockage of the LB antenna.

In addition, the designed 2.5-D cloaks exhibit good universality and flexibility. By simply

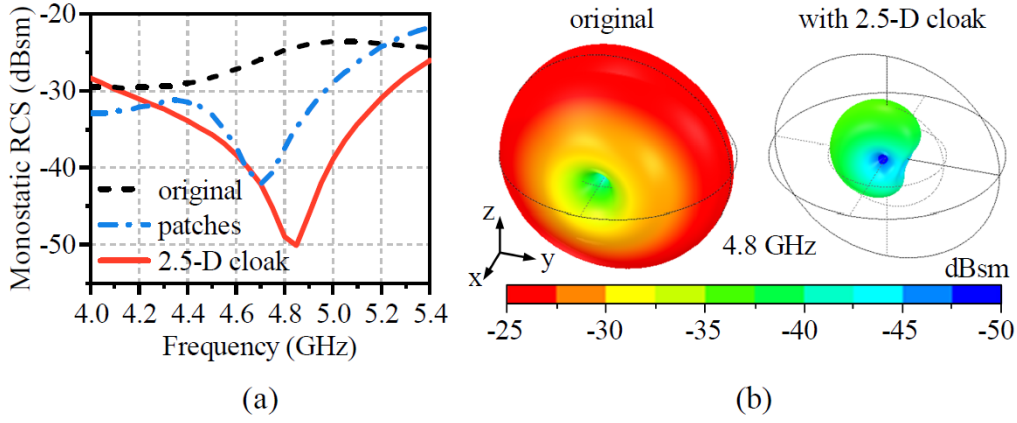


Figure 3.6: (a) Monostatic and (b) bistatic radar cross sections (RCS) of the LB dipole with different loading conditions.

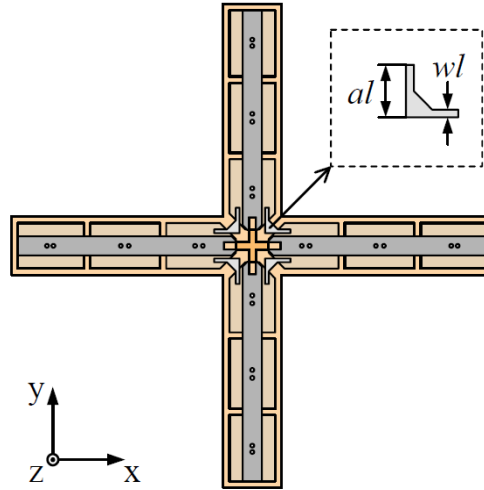


Figure 3.7: Geometry of the LB cross-dipole loaded with the 2.5-D cloak.

adjusting the length l , they can be applied to scattering suppression across different frequency bands. For example, as shown in Figure 3.9, when l is increased to 18.5 mm, the 2.5-D cloaks enable scattering suppression and radiation pattern restoration in the 5G band of 3.3–3.8 GHz. The radiation patterns before and after loading the 2.5-D cloaks over the 3.3–3.8 GHz band are presented in Figures 3.10 and 3.11, respectively. The following content will still focus on the HB operation in the 4.4–5.0 GHz range, where other scattering and coupling problems are also encountered.

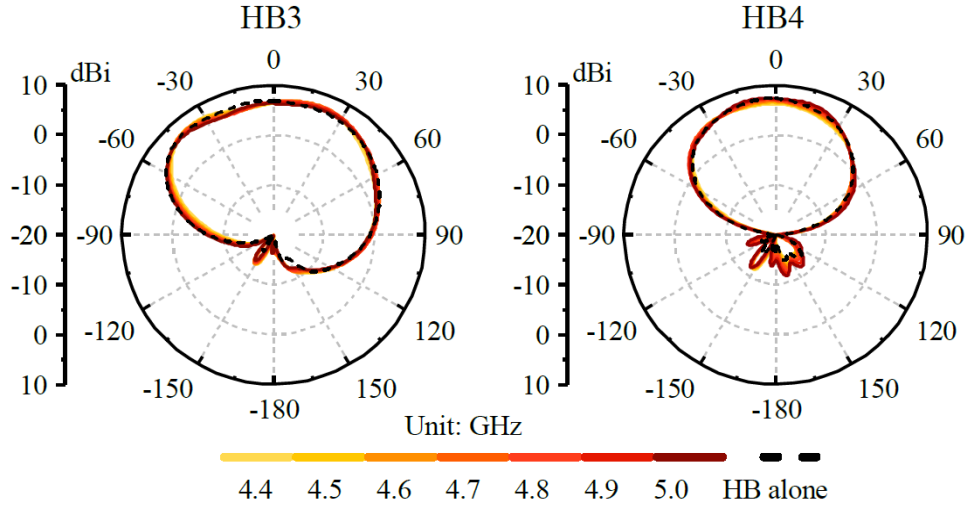


Figure 3.8: Restored radiation patterns in the yoz -plane of HB3 and HB4 when the loaded LB dipole is present.

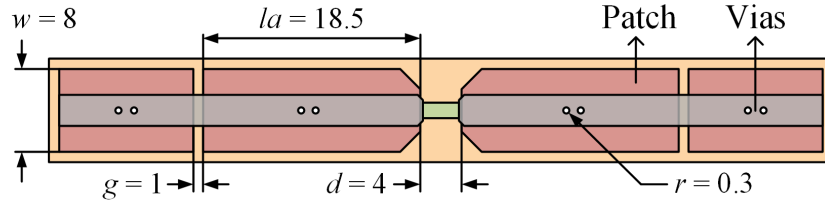


Figure 3.9: Geometry of the loaded LB radiator to suppress the scattering at 3.3-3.8 GHz.

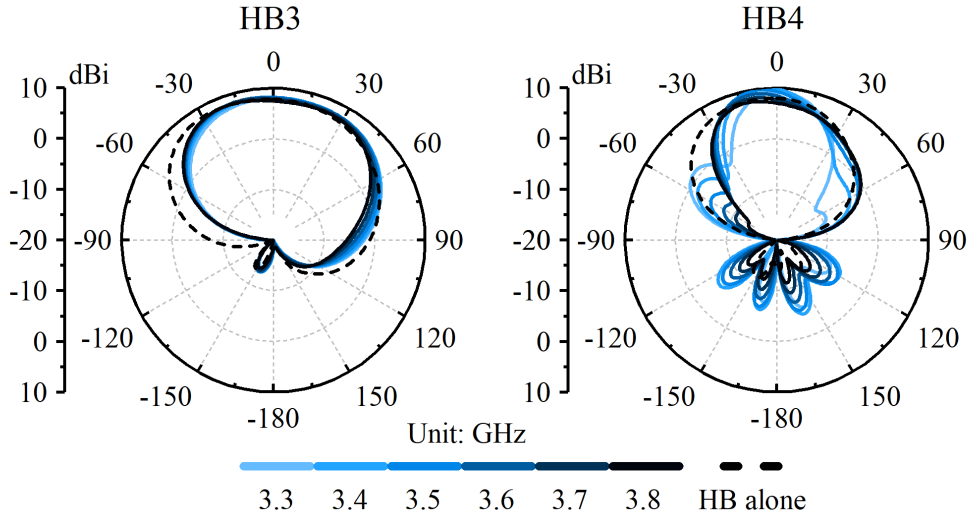


Figure 3.10: Radiation patterns in the yoz -plane of HB3 and HB4 when the LB dipole is unmodified.

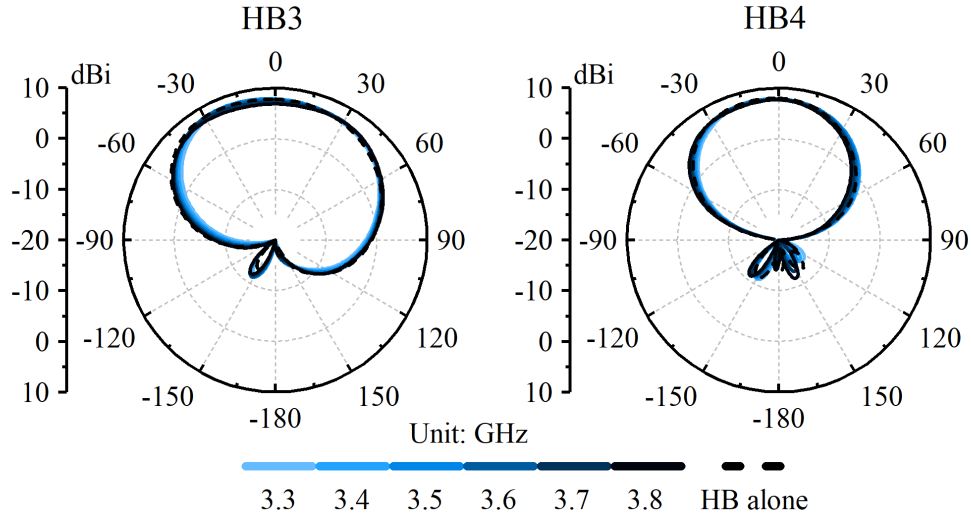


Figure 3.11: Radiation patterns in the yoz -plane of HB3 and HB4 when the LB dipole is loaded with the 2.5-D cloaks.

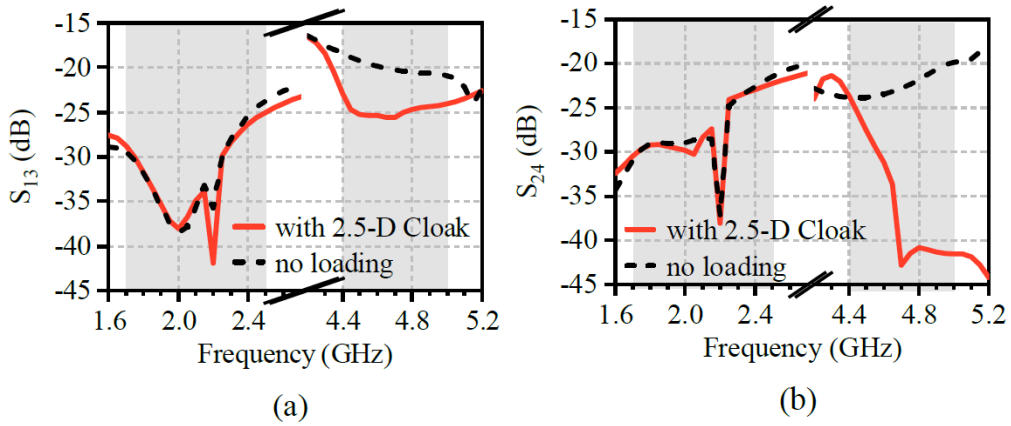


Figure 3.12: (a) S_{13} and (b) S_{24} before and after loading the 2.5-D cloak. Ports 1-4 are the feeding ports of LB1, LB2, HB3, and HB4, respectively.

3.2.2 Coupling Suppression

The proposed 2.5-D cloak can also reduce the cross-band coupling between the LB and HB antennas. The transmission coefficients between the co-polarized LB and HB dipoles (S_{13} and S_{24}) before and after loading the 2.5-D cloak are compared in Figure 3.12. As shown in the figure, the modification of the LB dipole significantly improves the port isolations in the HB, especially for S_{24} , while the isolations at the LB remain at the same level.

To explain this improvement, the current distribution on the LB2 when irradiated with HB4 is shown in Figure 3.13. After loading the 2.5-D cloak, reverse induced-currents are generated

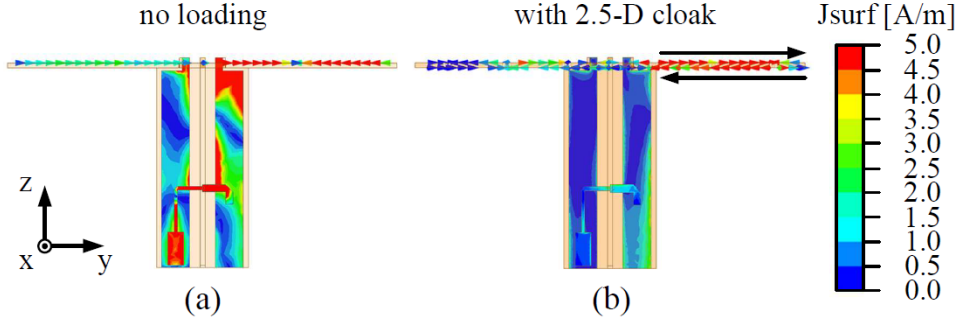


Figure 3.13: Current distributions at 4.7 GHz on (a) the original LB2 and (b) the LB2 with 2.5-D cloak when excited by HB4 radiation.

on the LB radiator and the cloak, resulting in a substantial reduction of the currents flowing into the LB port via the LB balun. This reduction leads to an improvement in the isolation between the ports of the LB and HB antenna.

3.3 Reducing the Cross-Band Scattering and Coupling to HB Radiation by Employing DGS Loaded LB Balun

3.3.1 Scattering Suppression

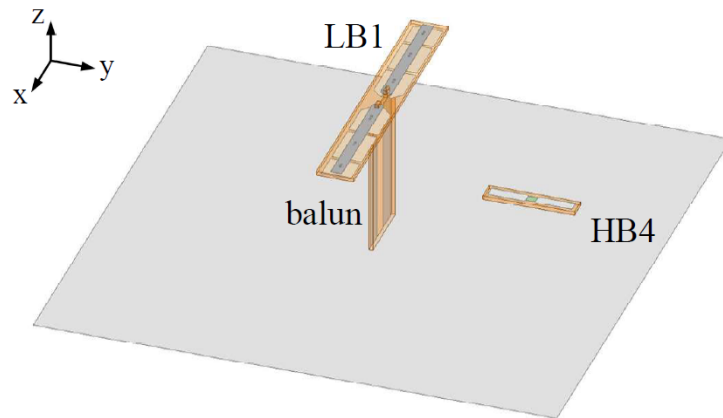


Figure 3.14: The “1 LB + 1 HB” dual-band array arrangement when the LB balun has the most significant effects on the HB radiation.

Although in Section 3.2 the LB dipole radiators’ scattering to the HB radiation has been largely eliminated, it is found that the balun of the LB dipole also blocks the HB radiation and has

a non-negligible effect on the HB performance in certain cases. Since the baluns are always perpendicular to the dipole arms, they usually have less effect on the HB radiation. However, when the LB baluns sit in the E-plane of the HB radiator, e.g., as shown in Figure 3.14, the balun of LB1 functions as a side reflector for the HB4 antenna, leading to a noticeable distortion. Figure 3.15 illustrates the change of HB radiation pattern before and after the LB balun is added. Although the deterioration is not as severe as the cases studied in 3.2, it introduces beam squint and narrows the half-power beamwidth.

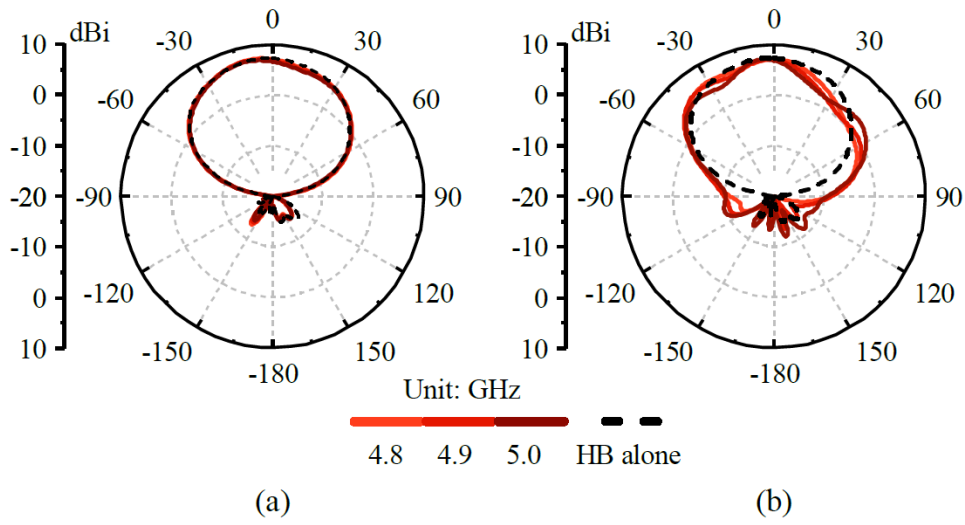


Figure 3.15: Radiation patterns of HB4 in the yoz -plane (a) without and (b) with the LB balun.

To suppress the LB baluns' scattering on the HB radiation, we propose to employ defected ground structures (DGSs) [113]–[115] and etch them on the baluns as depicted in Figure 3.16. The DGSs can choke the induced HB currents as demonstrated in Figure 3.17, thereby alleviating the pattern distortion at the HB. The suppression band can be adjusted by optimizing the heights of the DGSs ($hc1$ and $hc2$), and the slight height difference is introduced to extend the suppression bandwidth. The higher the height of the DGSs, the lower the suppression frequency.

The monostatic RCS of the LB balun with and without the DGSs is compared in Figure 3.18(a). It can be seen that the monostatic RCS is much lower when the DGSs are present. By replacing the original balun with the DGS-loaded one, the radiation pattern of the HB dipole is well restored as shown in Figure 3.18(b).

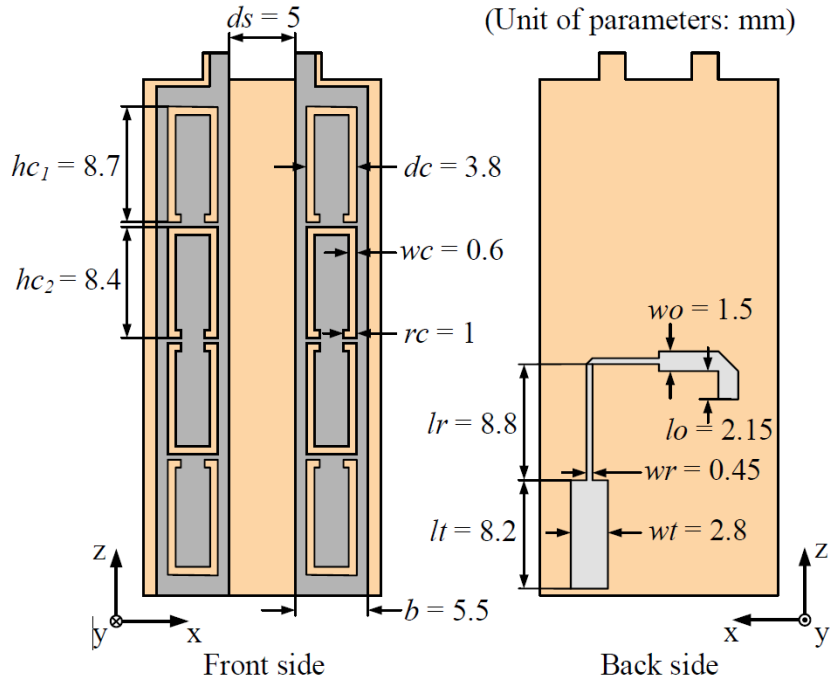


Figure 3.16: Geometry of the revised LB balun with DGSs.

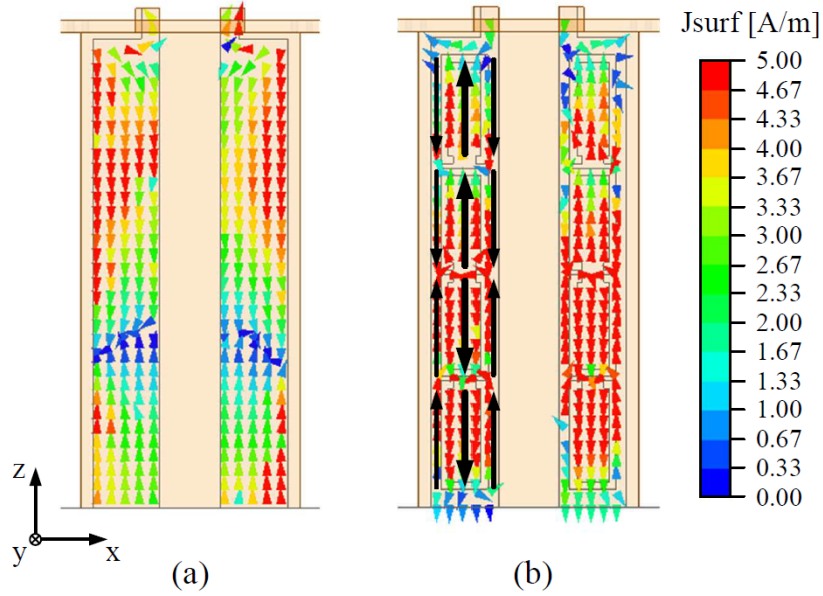


Figure 3.17: Current distributions at 4.9 GHz on (a) the original LB balun and (b) the DGS-loaded LB balun when excited by HB radiation.

3.3.2 Coupling Suppression

Similar to the 2.5-D cloak, the proposed DGSs can also suppress both the cross-band scattering and coupling at the same time. The transmission coefficients S_{13} and S_{24} with different loading conditions are compared in Figure 3.19. The introduction of the DGSs prompts a further decline

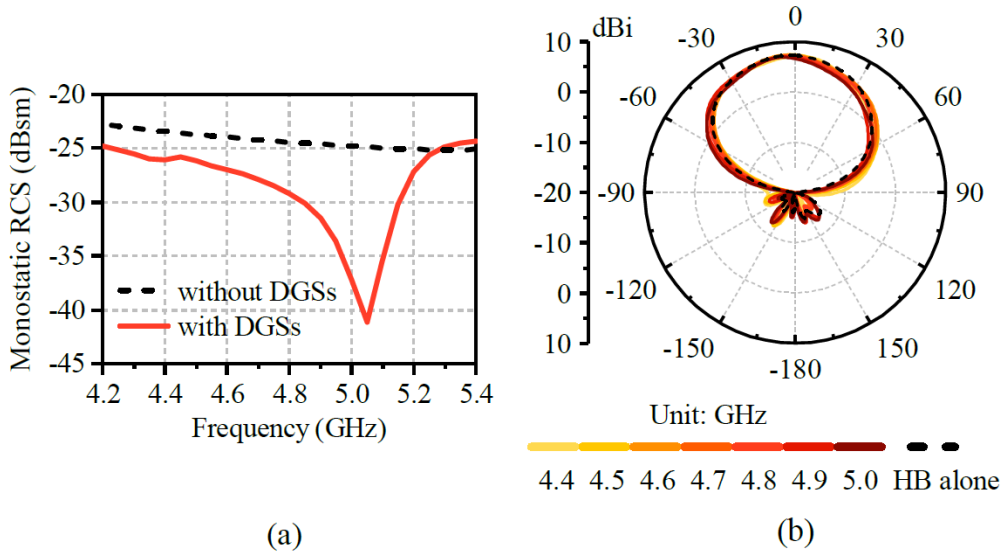


Figure 3.18: (a) Monostatic RCS of the LB balun, and (b) restored HB radiation patterns when the LB is loaded with DGSs.

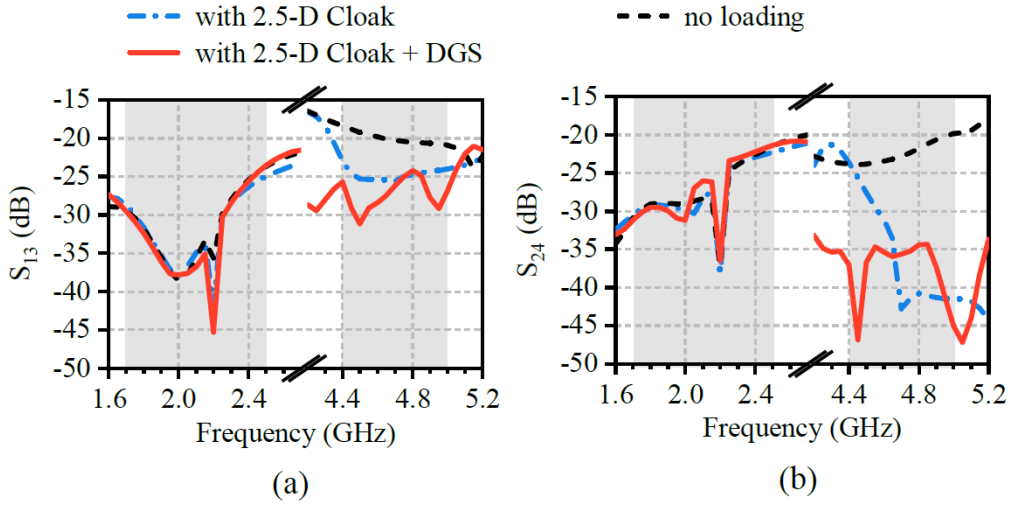


Figure 3.19: (a) S_{13} and (b) S_{24} with different loading conditions.

in S_{13} and S_{24} within the HB. It is noticed that the loadings on the LB dipole (2.5-D cloak and DGSs) only affect the isolation at the HB and the isolations in the LB remain unchanged. This is because both the 2.5-D cloak and DGSs are used to reduce the induced HB currents on the LB antennas, and they are designed to have minor effects on the LB operation.

To reveal the mechanism of the greatly improved isolations at the HB, Figure 3.20 depicts the current distributions on the LB baluns when the HB antennas are excited at 4.7 GHz. Before loading the DGSs, there are strong currents flowing into LB ports along the feeding

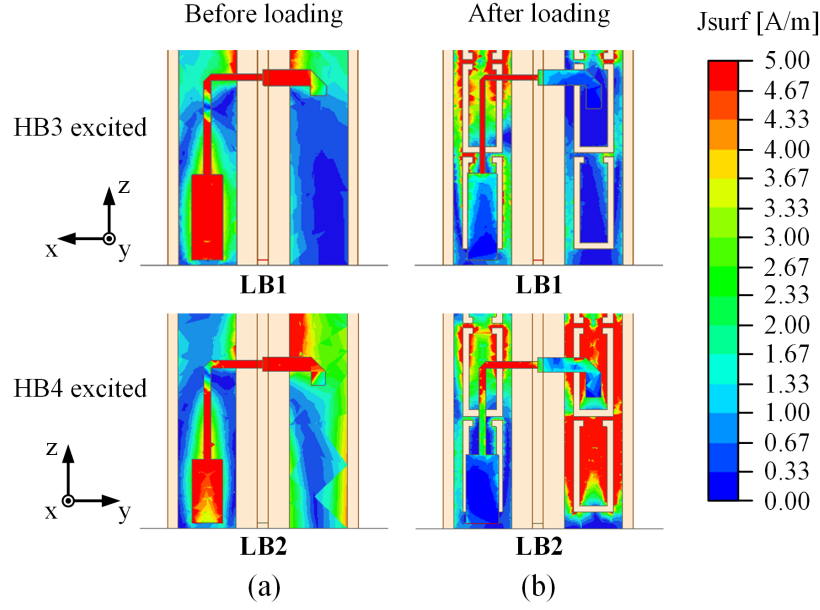


Figure 3.20: Current distributions at 4.7 GHz on the LB balun (a) before and (b) after loading DGSs.

lines of the LB dipole, resulting in a strong coupling between the LB and HB antennas. After the loading, the currents near the LB ports are significantly decreased, indicating substantial coupling reduction.

Although loading the slots on the balun ground has been used for cross-band decoupling, our study is the first to reveal an issue that has not been previously discussed: the impact of the LB balun on the HB patterns. Additionally, significant dual-functional effects are achieved using this method.

3.3.3 Finalized LB Cross-Dipole Antenna

To reduce the cross-band scattering of the LB dipoles to the HB radiation, the dipole radiators and baluns are loaded with the 2.5-D cloaks and the DGSs, respectively. The finalized LB cross-dipole is presented in Figure 3.21, and the S-parameters of the LB antenna are given in Figure 3.22. As shown in Figure 3.22, the reflection coefficients (S_{11} and S_{22}) of the two sub-dipoles are all < -10 dB from 1.67 to 2.59 GHz, and the isolation between the two LB ports is > 20 dB over this band. The radiation patterns in the yoz-plane of the two sub-dipoles are plotted in Figure 3.23. The simulated gain varies from 6.1 to 7.5 dBi over the impedance bandwidth.

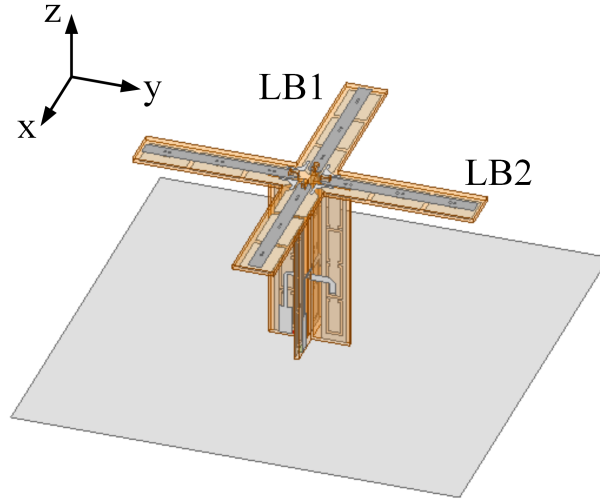


Figure 3.21: Geometry of the finalized LB cross-dipole with loadings on both the radiators and baluns.

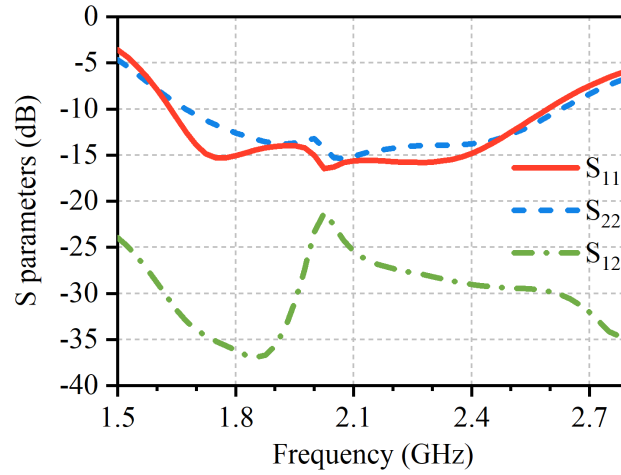


Figure 3.22: Simulated S-parameters of the finalized LB cross-dipole.

3.4 Reducing the Cross-Band Scattering and Coupling to the LB Radiation by Loading HB Balun With Resonance Shifters

3.4.1 Scattering Suppression

When two antennas working in different bands are closely placed together, the presence of the HB antenna could also degrade the performance of the LB antenna. As depicted in Figure

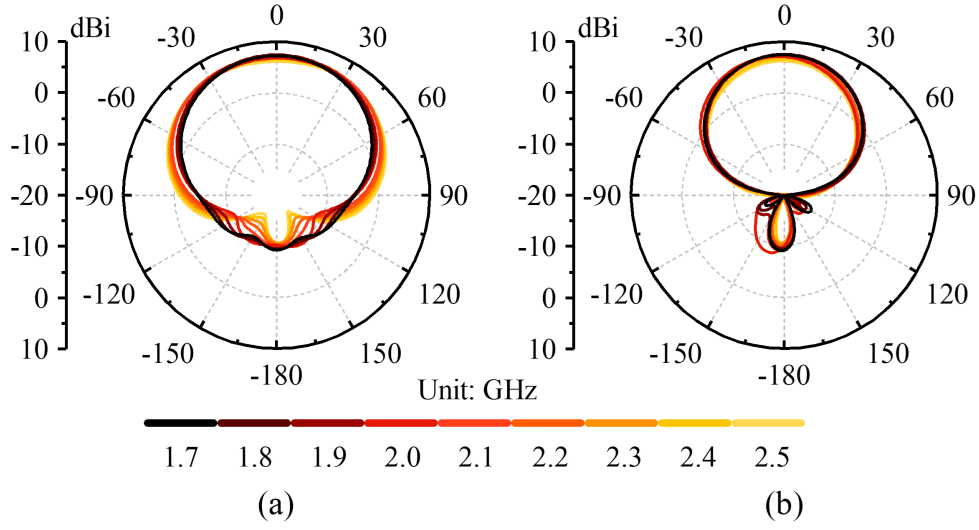


Figure 3.23: Radiation patterns in the yo z -plane of the finalized (a) LB1 and (b) LB2 antenna.

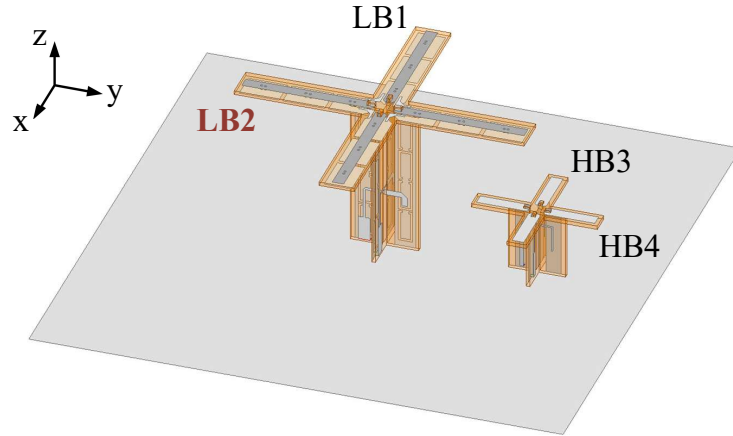


Figure 3.24: Configuration of the loaded LB cross-dipole antenna and a conventional HB cross-dipole antenna.

3.24, a traditional HB cross-dipole antenna working at 4.4-5.0 GHz is placed below the LB cross-dipole ($dis = 40$ mm, $\theta = 0^\circ$, $\Delta H = 22$ mm). With the HB antenna standing nearby, the patterns of the LB2 in the yo z -plane encounter severe distortions as shown in Figure 3.25(b) compared to the original patterns in Figure 3.25(a). Moreover, the realized gain of the LB dipole has been noticeably reduced with a maximum gain reduction of 2 dB at 1.75 GHz.

The pattern deterioration is due to the fact that the HB antenna works as a mushroom structure which couples to the LB antenna and introduces an unwanted resonance at LB frequencies. This can be demonstrated by evaluating the HB antenna's common-mode input impedance. As depicted in Figure 3.26(a), the HB antenna is slightly lifted up off the ground while the

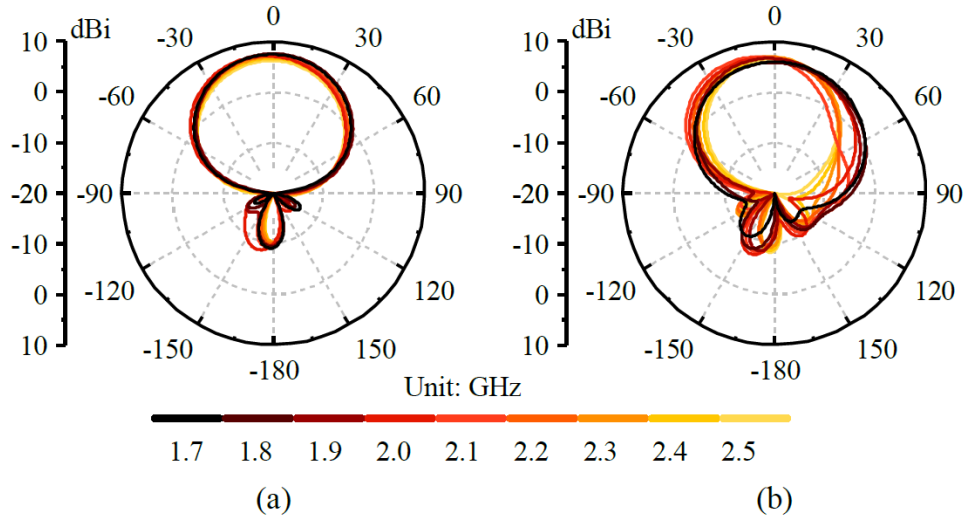


Figure 3.25: Radiation patterns of the LB2 in the yoz -plane (a) without and (b) with the HB antenna.

bottoms of the baluns are connected. By adding a discrete port between the bottom of the balun and the ground, one is able to evaluate the common-mode input impedance of the HB antenna. As revealed in Figure 3.26(b), there is a resonance at 2.1 GHz (the black dashed line) where the distortion to the LB pattern is the worst.

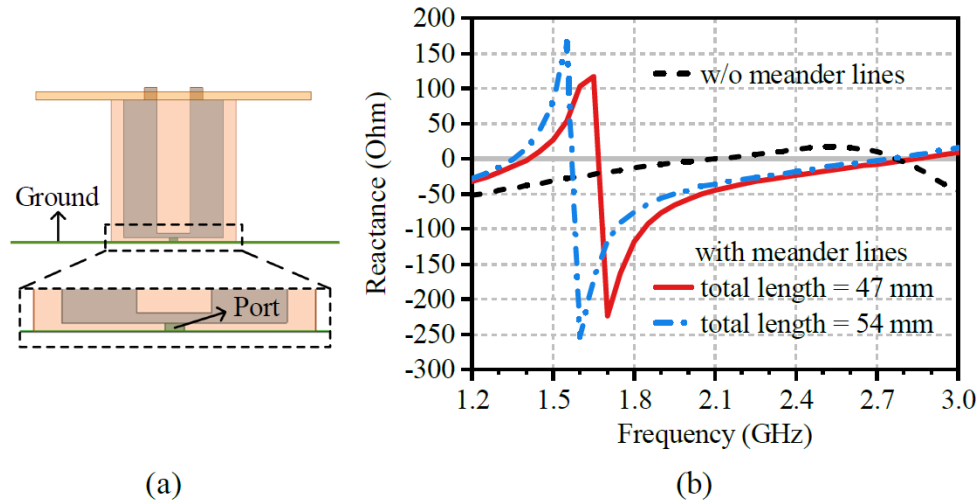


Figure 3.26: (a) Discrete port added to the HB antenna, and (b) the input reactance of the HB antenna at the discrete port.

The problem can be solved by introducing a resonance shifter in the form of meander lines to move this resonance out of the operating band (1.7-2.5 GHz) of the LB antenna. As shown in

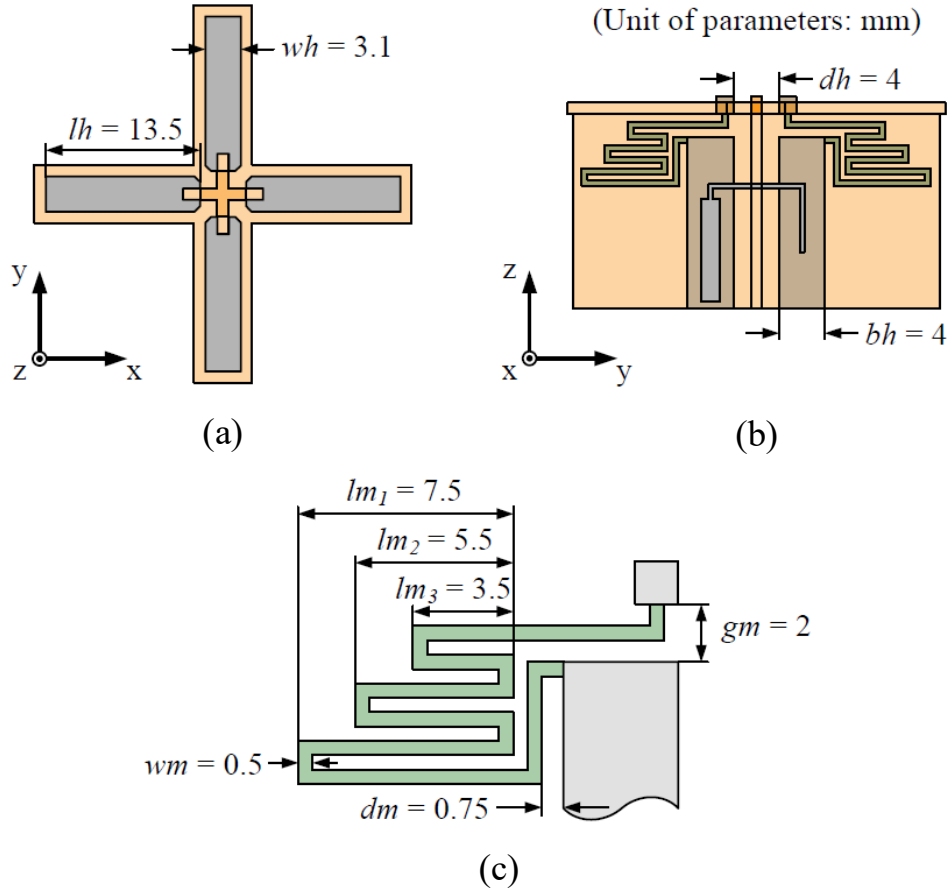


Figure 3.27: (a) Top view and (b) side view of the HB cross-dipole antenna with the resonance shifters, and (c) geometry of the resonance shifters.

Figure 3.27, the resonance shifters can be inserted between the radiators and the baluns of the HB antenna. As can be seen from Figure 3.26(b), the longer the meander lines, the lower the resonant frequency. On the other hand, the impedance matching of the HB antenna is more difficult with longer resonance shifters.

The resonance shifters (total length = 47 mm) added to the HB baluns have little effect on the HB radiation patterns, while the HB antennas are still well matched within the target frequency band. Figure 3.28 gives the simulated S-parameters of the altered HB antennas with the presence of the LB antennas when $dis = 40$ mm, $\theta = 0^\circ$. With the optimized resonance shifters, the LB2 radiation patterns are restored in the target band as evidenced in Figure 3.29.

3.4.2 Coupling Suppression

The isolations between the LB and HB antennas of the same polarizations, i.e., S_{13} and S_{24} , with and without the resonance shifters are compared in Figure 3.30. As expected, the resonance

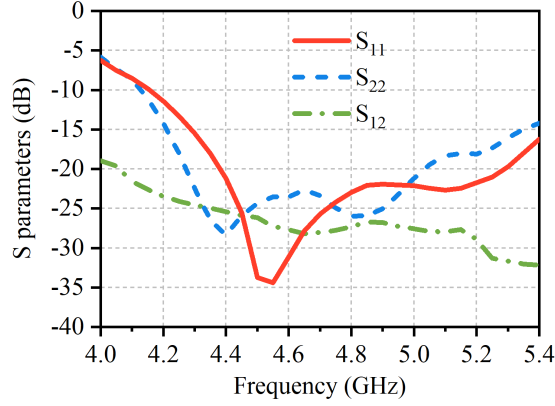


Figure 3.28: Simulated S-parameters of the loaded HB cross-dipole with the presence of the LB antennas.

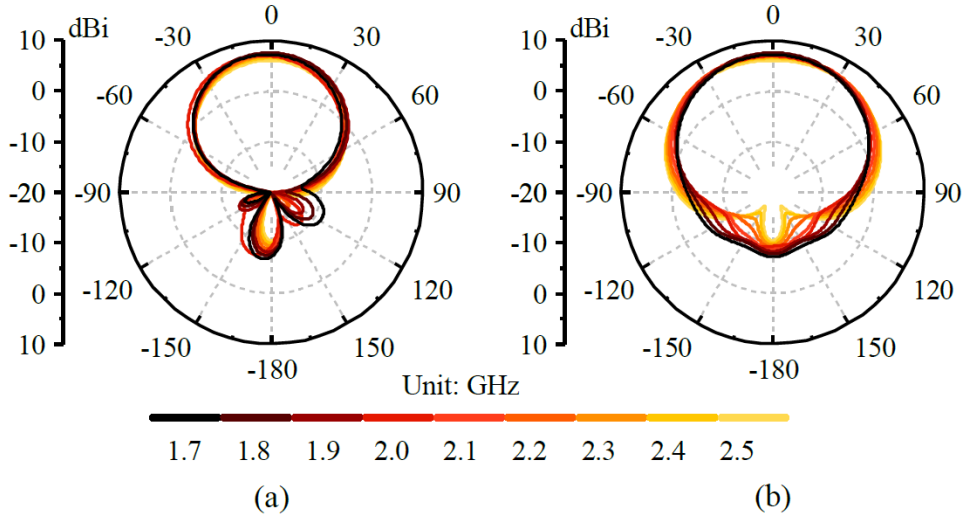


Figure 3.29: Restored radiation patterns of LB2 in the (a) yoz -plane and (b) xoz -plane.

shifters mainly affect the port isolations in the LB and have little effect on the HB. In the LB, despite there being a rise in the transmission coefficients near 2.1 GHz after the loading, the worst isolation across the band is better than that without the loadings. Actually, the low transmission coefficients without the resonance shifters near 2.1 GHz are misleading and there exists very strong coupling between the LB and HB antennas. In addition, Figure 3.31 compares the cross-band isolations for the cross-polarized ports (S_{14} and S_{23}), which exceed 28 dB throughout both the LB and HB.

To uncover the mechanism of the deceptive isolation degradation around 2.1 GHz, Figure 3.32 compares the current distributions on the HB dipoles before and after loading the resonance

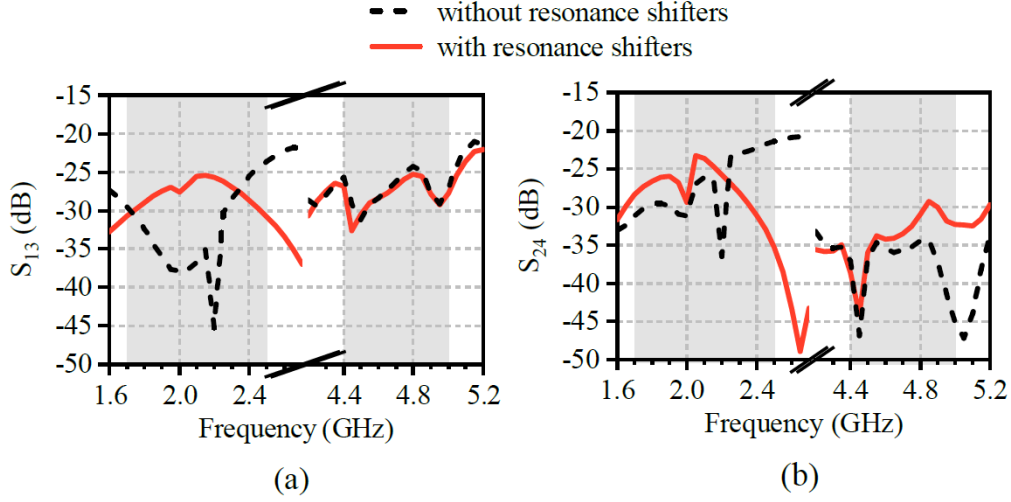


Figure 3.30: (a) S_{13} and (b) S_{24} with and without the resonance shifters.

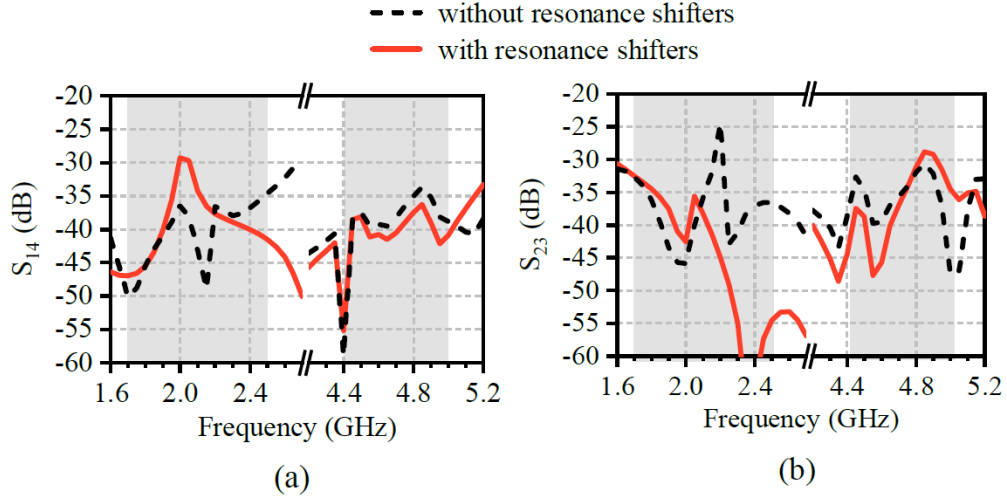


Figure 3.31: (a) S_{14} and (b) S_{23} with and without the resonance shifters.

shifters when the LB dipoles are excited at 2.1 GHz. As shown in Figure 3.32(a), for the unaltered HB antennas, although the transmission coefficients between the two dipoles are low, there are actually strong LB currents induced on the HB dipole. The induced currents are mainly distributed on the baluns and are directed to the ground and there is less current flowing into the HB ports, thereby the transmission coefficients between the LB and HB ports are low. However, there exist strong interactions between the LB and HB antennas and a noticeable portion of the LB radiation is redirected to the ground plane, leading to deteriorated LB patterns. As observed in Figure 3.32(b), after loading the resonance shifters, although there are more currents entering the HB ports (resulting in higher transmission coefficients), the resonance shifters prevent the currents from flowing to the ground plane and the overall

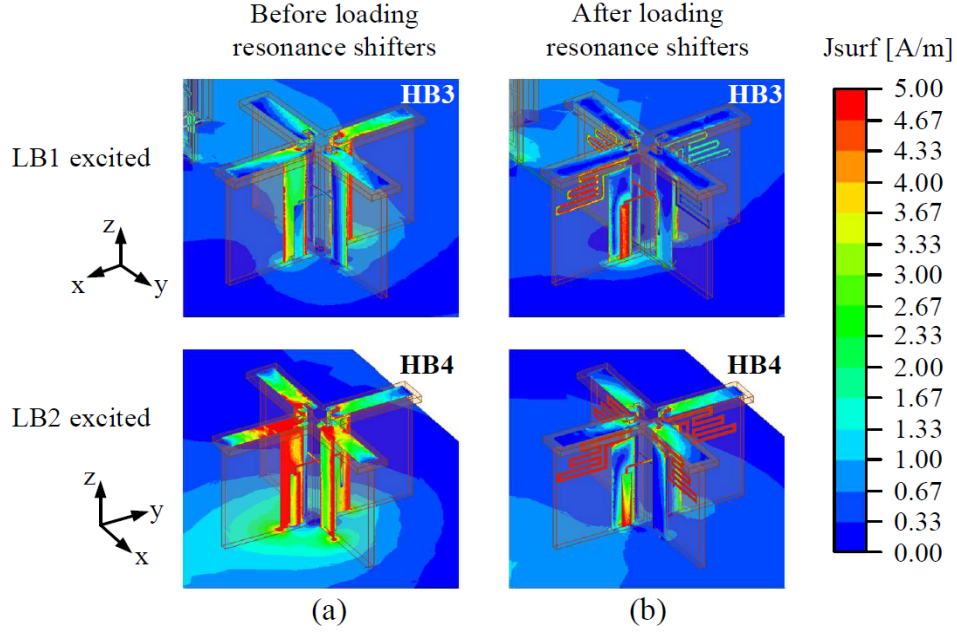


Figure 3.32: Current distributions at 2.1 GHz on the HB dipole (a) before and (b) after loading resonance shifters on HB dipole. (The HB dipoles are excited by the radiating LB dipoles.)

induced currents are less than the case without loadings.

3.5 Robustness Analysis of the Proposed Methods

3.5.1 “1 LB + 1 HB” Array

Figure 3.33 illustrates the radiation patterns of HB3 and HB4 in the xoz and $yo z$ planes when the HB antennas are placed near the LB dipole with $dis = 40$ mm, $\theta = 0^\circ$. The HB dipoles are also moved to another location with $dis = 40$ mm, $\theta = 45^\circ$ to test the robustness of the descattering techniques. Only the radiation patterns of HB3 are given in Figure 3.34 since those of HB4 are necessarily the same due to the symmetry. Figure 3.35 shows the radiation pattern of LB2 at 2.1 GHz under different dis and θ .

It can be concluded from the figures that the radiation patterns of both the LB and HB antennas are effectively restored with the loadings under various relative positions between the LB and HB antennas. This finding confirms the robustness of the proposed methods in restoring the antennas' radiation patterns to changes in antenna position.

The loadings' ability to reduce cross-band coupling has also been verified at different relative

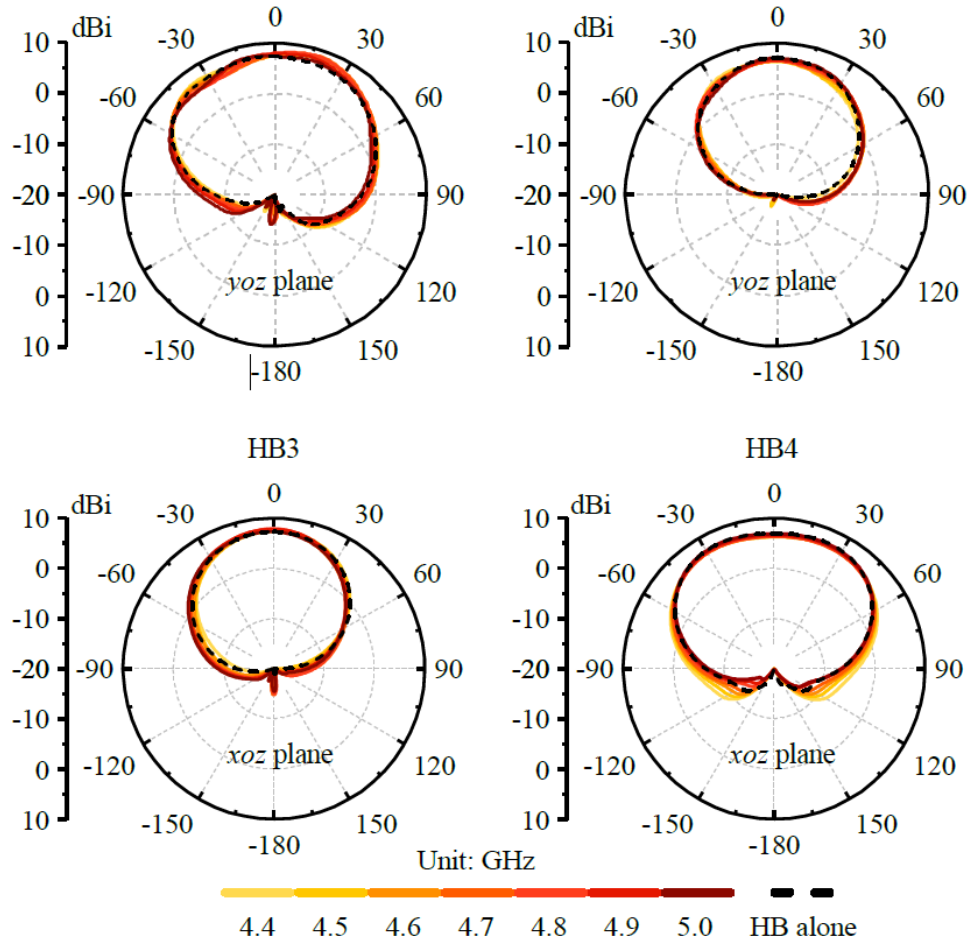


Figure 3.33: Radiation patterns of HB3 and HB4 in the xoz and yoz planes when $dis = 40$ mm, $\theta = 0^\circ$.

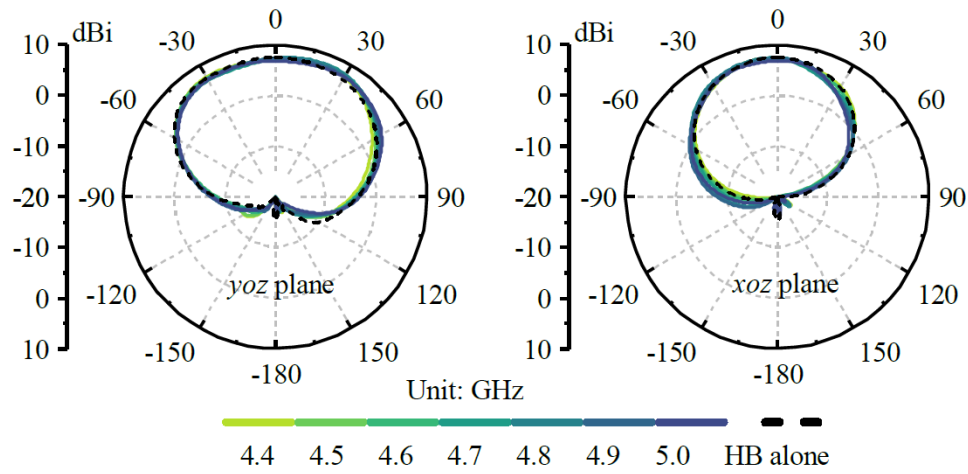


Figure 3.34: Radiation patterns of HB3 in the xoz and yoz planes when $dis = 40$ mm, $\theta = 45^\circ$.

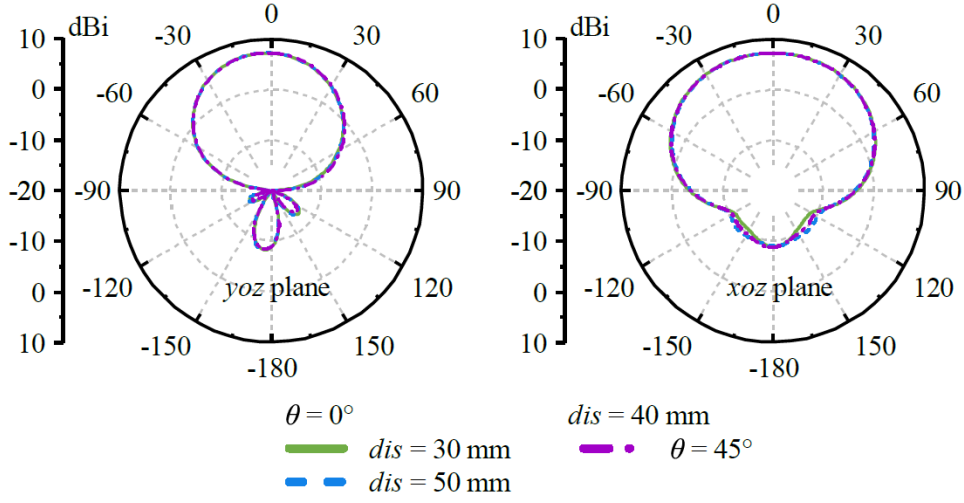


Figure 3.35: Radiation patterns of LB2 at 2.1 GHz with three different relative positions between the LB and HB dipoles.

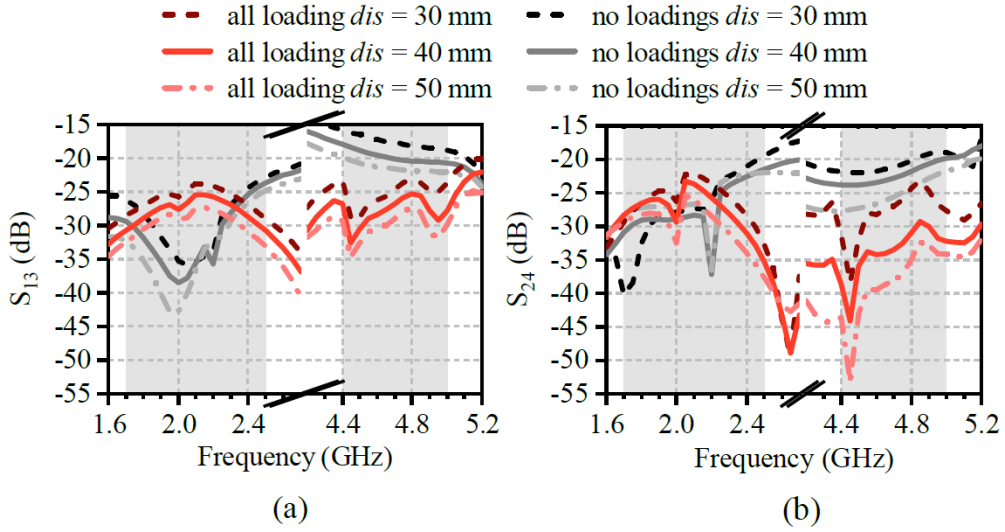


Figure 3.36: (a) S_{13} and (b) S_{24} with and without the loadings when $\theta = 0^\circ$ with different distances (dis).

positions between the LB and HB antennas, confirming their robustness. Figure 3.36 compares the transmission coefficients with and without the loadings when dis varies from 30 to 50 mm and $\theta = 0^\circ$. It is found that the proposed loadings consistently lead to decent isolation improvements at the HB. Then the isolations with and without loadings are further compared in Figure 3.37 by changing θ to 45° while keeping $dis = 40$ mm. The results are similar to those with $\theta = 0^\circ$ in Figure 3.36, i.e., a substantial enhancement at the HB and the overall isolation maintains > 25 dB throughout the LB.

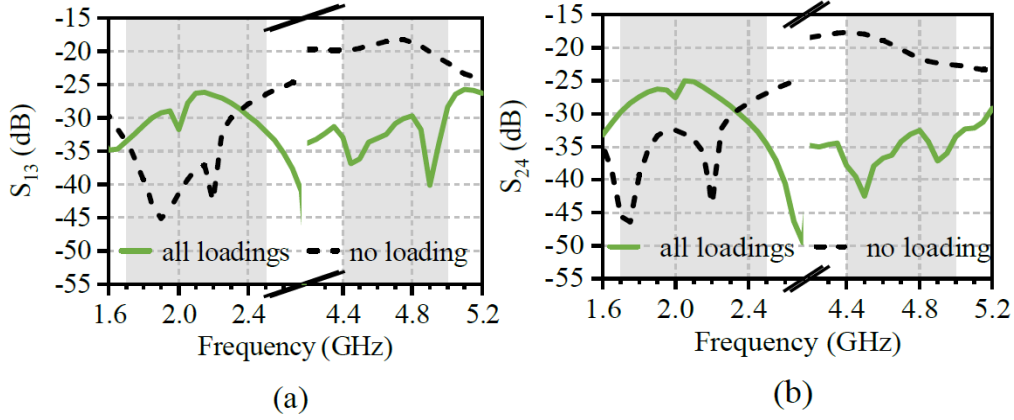


Figure 3.37: (a) S_{13} and (b) S_{24} with and without the loadings when $dis = 40$ mm, $\theta = 45^\circ$.

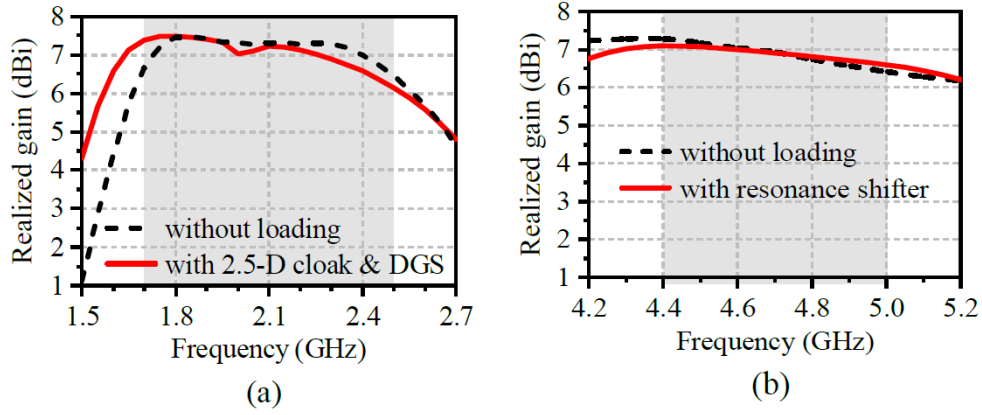


Figure 3.38: Realized gain of (a) LB antenna and (b) HB antenna under different loading cases.

Figure 3.38(a) and (b) compare the realized gains of the LB and HB antennas, respectively, before and after the loadings are applied to them when $dis = 40$ mm and $\theta = 0^\circ$. It can be seen that the realized gains are not affected much by the loadings and remain at the same level.

3.5.2 “1 LB + 4 HB” Array

To further verify the feasibility of the proposed suppression methods and their effect on in-band coupling between adjacent HB antennas, they are tested in a “1 LB + 4 HB” array configuration as shown in Figure 3.39. The radiation patterns of LB and HB antennas are shown in Figures 3.40 and 3.41, respectively. Similar to the results obtained from the “1 LB + 1 HB” array configuration, the radiation patterns of both the LB and HB elements in the “1 LB + 4 HB” array can still be restored, demonstrating good consistency with the patterns when

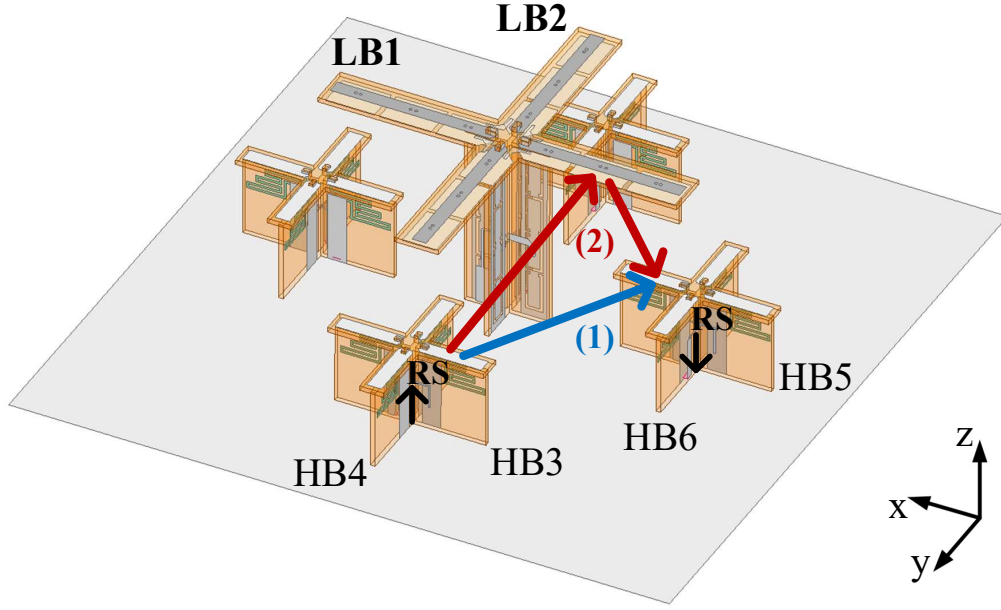


Figure 3.39: Configuration of the “1 LB + 4 HB” dual-band array and the coupling path between two adjacent HB elements.

the antennas are working alone. Given that the radiation patterns of each HB antenna are effectively restored, it can be concluded that the patterns of an HB array will exhibit minimal distortion, regardless of how the array is configured.

The cross-band couplings between the LB and HB antennas (S_{13} and S_{24}) in the “1 LB + 4 HB” configuration are almost identical to those in the “1 LB + 1 HB” configuration shown in Figure 3.36, which are compared in Figure 3.42. The similarity in the radiation pattern and cross-band isolation results between the “1 LB + 4 HB” and “1 LB + 1 HB” array configurations confirms the reliability of using the latter to validate cross-band decoupling and decoupling performance, while significantly reducing simulation time.

However, assessment of the impact of the proposed method on in-band coupling between adjacent co-polarized HB antennas (HB3 and HB5, and HB4 and HB6) can only be performed in the “1 LB + 4 HB” array configuration. Figure 3.43 compares the couplings between HB3 and HB5 and between HB4 and HB6 with different loading conditions. The figure reveals that all three methods have a positive impact on enhancing the in-band isolation, with particular effectiveness observed in the 2.5-D cloak and resonance shifter techniques. As shown in Figure 3.39, energy can be coupled from one HB element to its adjacent HB element (for example from HB3 to HB5) through two paths: (1) direct coupling between the HB antennas and (2)

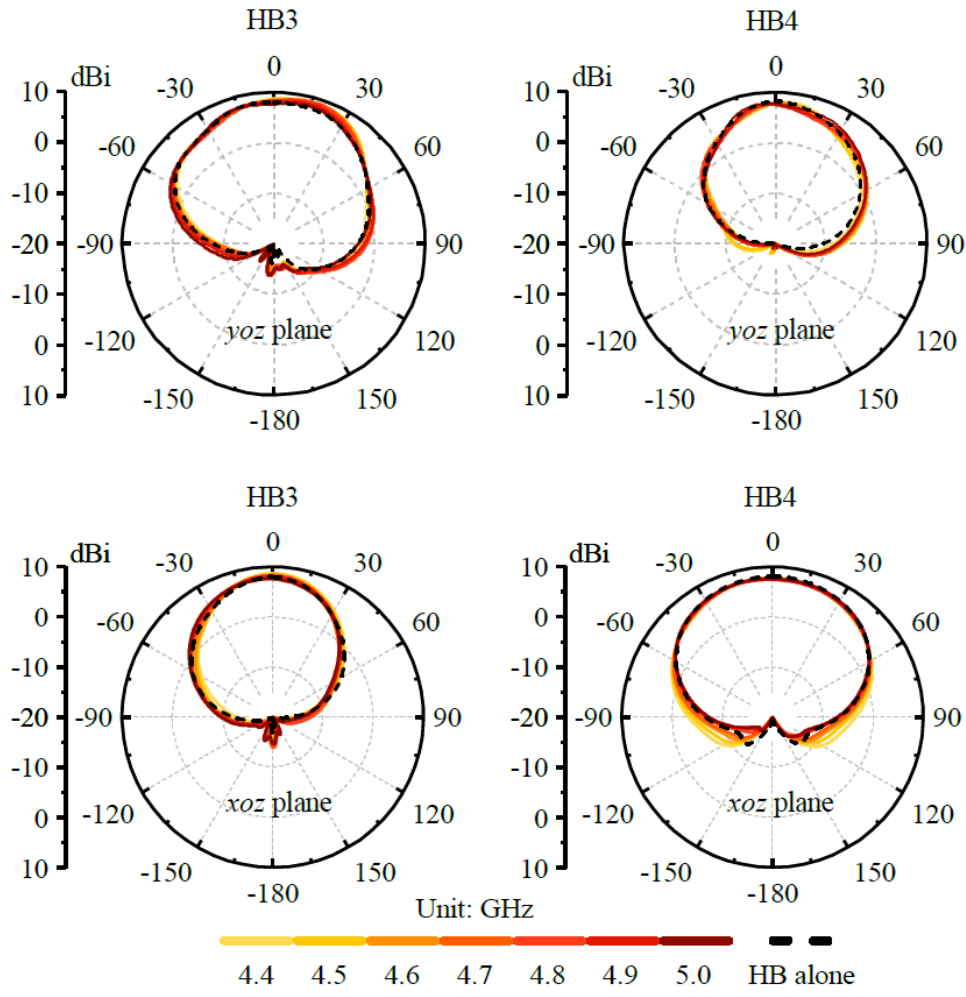


Figure 3.40: Radiation patterns of HB3 and HB4 in the "1 LB + 4 HB" array configuration. The patterns of other HB elements are similar due to symmetry.

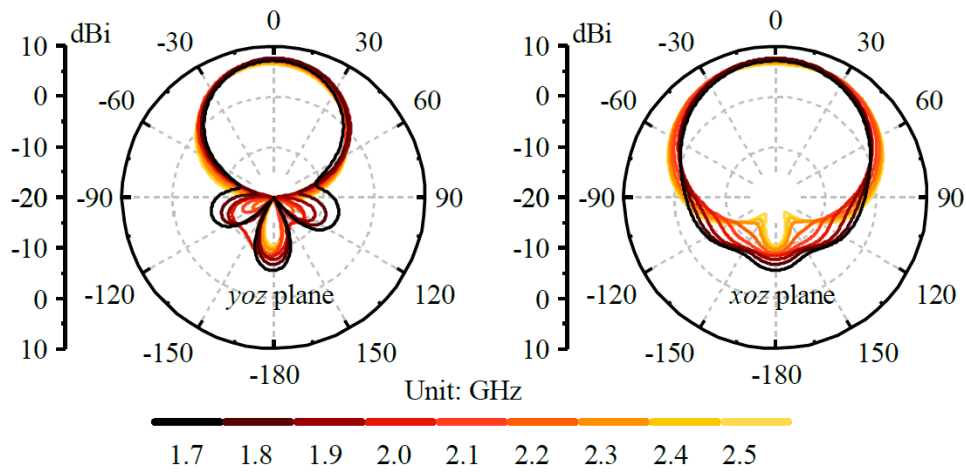


Figure 3.41: Radiation patterns of LB2 in the "1 LB + 4 HB" antenna array.

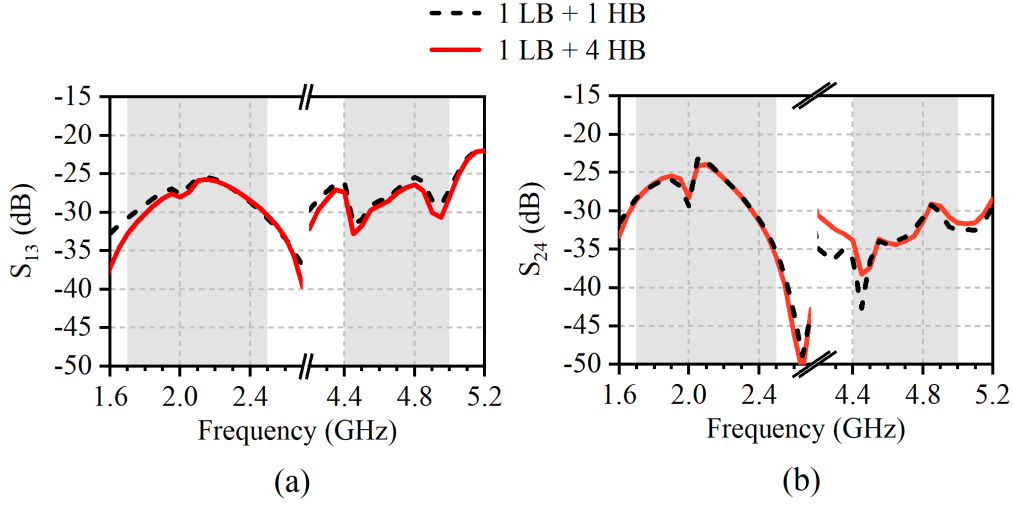


Figure 3.42: (a) S_{13} and (b) S_{24} in the “1 LB + 1 HB” and “1 LB + 4 HB” arrays.

indirect coupling via the re-radiation of the LB antenna. However, the introduction of the 2.5-D cloak significantly suppresses the scattering of the LB radiator in all directions in the HB, as depicted in Figure 3.6(b), thereby reducing the in-band coupling via path (2). Furthermore, as illustrated in Figure 3.39, introduction of the resonance shifters (RS) equivalently increases the length of the coupling path between the feed ports of two adjacent HB elements, thus improving the in-band isolation.

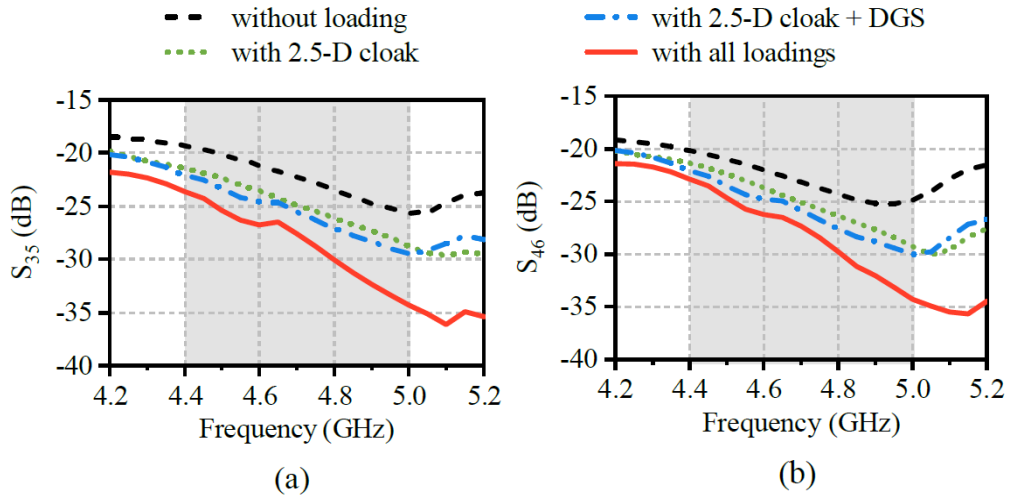


Figure 3.43: (a) S_{35} and (b) S_{46} in the “1 LB + 4 HB” array with different loading conditions.

To summarize, all three suppression methods proposed in this work demonstrate significant and effective suppression of cross-band scattering and coupling, irrespective of antenna positions. Furthermore, the 2.5-D cloak and resonance shifter also reduce in-band coupling between

adjacent HB elements, making them triple-functional.

3.6 Experimental Results

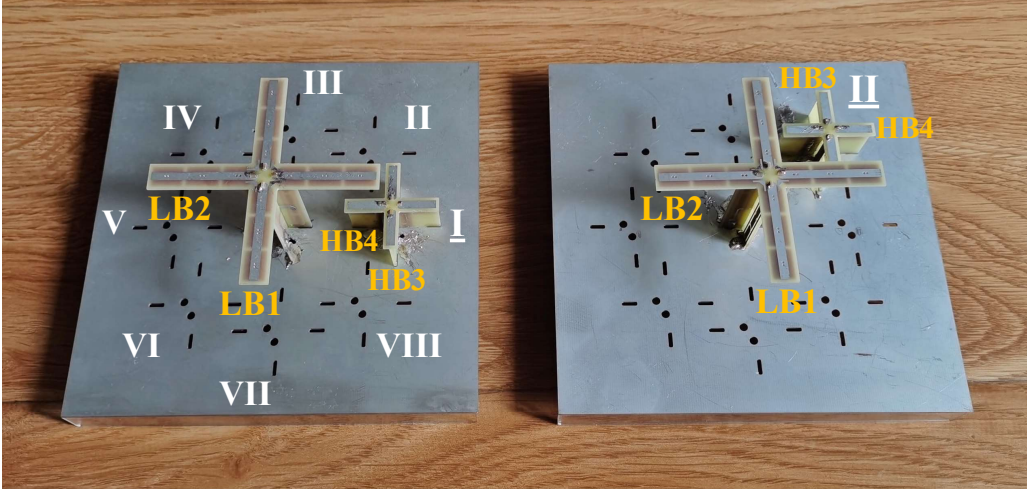


Figure 3.44: Prototypes of the proposed LB and HB cross-dipole antennas in Position I and II.

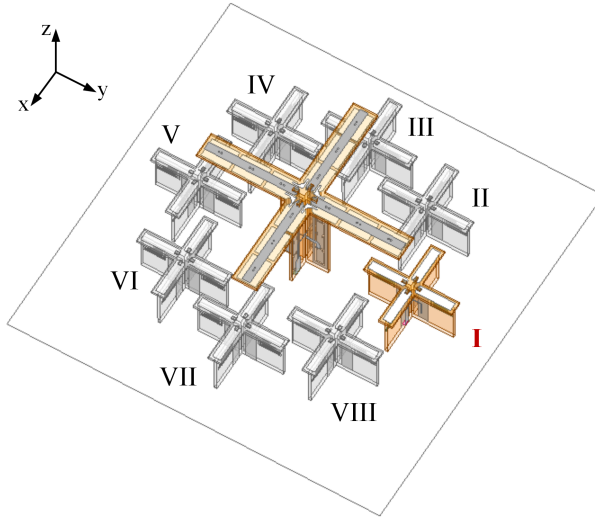


Figure 3.45: Different positions of the HB antennas in the dual-band array.

The proposed LB (1.7-2.5 GHz) and HB (4.4-5.0 GHz) antennas with loadings of a 2.5-D cloak to the LB antenna, a DGS to the LB antenna balun, and a resonance shifter to the HB balun were fabricated and assembled. They were then mounted on a ground plane for measurement as shown in Figure 3.44. The LB dipole is located at the centre of the ground plane and the HB dipole can be placed at eight different positions ($dis = 40$ mm and $\theta = 0^\circ + n \times 45^\circ$, where $n = 0, 1 \dots 7$) to demonstrate the robustness of the proposed descattering and decoupling methods

as given in Figure 3.45. Due to the symmetry of the structure, only the test results of positions I and II are given for brevity.

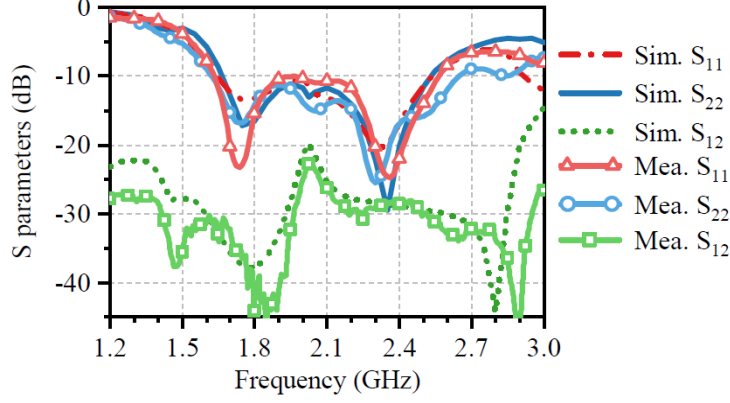


Figure 3.46: Simulated and measured reflection coefficients and polarization isolations of the LB cross-dipoles in Position I.

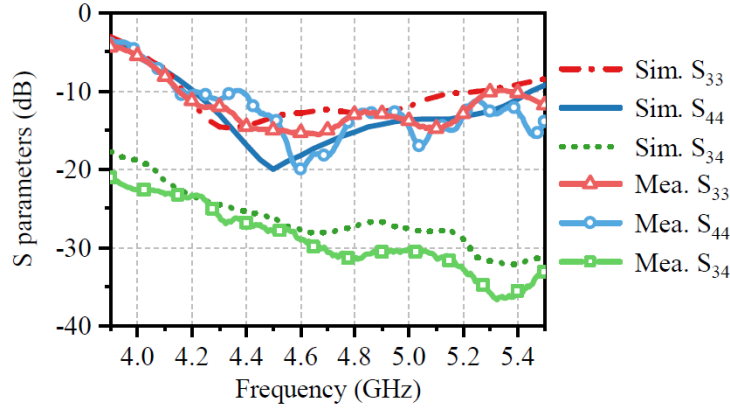


Figure 3.47: Simulated and measured reflection coefficients and polarization isolations of the HB cross-dipoles in Position I.

Figures 3.46 and 3.47 show the simulated and measured reflection and transmission coefficients of the LB and HB antennas, respectively, when the HB antenna stands at Position I. The measured results agree very well with the simulated ones. The measured reflection coefficients (S_{11} and S_{22}) of LB1 and LB2 are all < -10 dB from 1.64 to 2.56 GHz (43.8%) with the measured $S_{12} < -22$ dB over this band. The measured impedance bandwidth (S_{33} , $S_{44} < -10$ dB) of the HB antenna is 4.16-5.33 GHz (24.7%), and within the target HB (4.4-5.0 GHz) the isolation between the two HB ports is > 27 dB. Moving the HB dipole to Position II has little effect on the S parameters, so the results are not placed here for brevity.

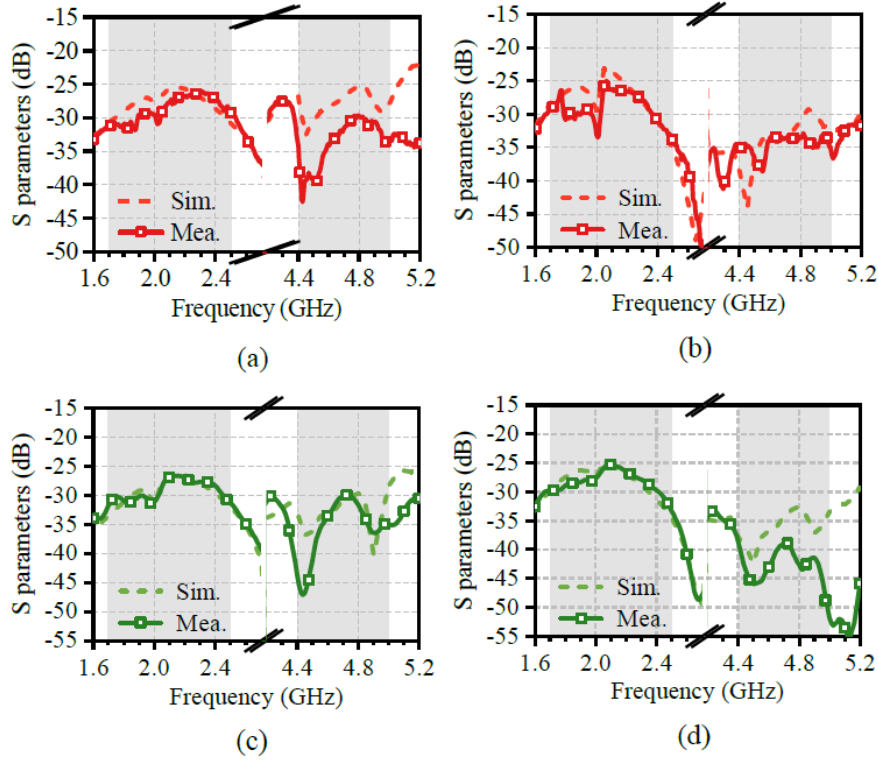


Figure 3.48: Simulated and measured (a) S_{13} and (b) S_{24} in Position I, and (c) S_{13} and (d) S_{24} in Position II.

As shown in Figure 3.48, the simulated and measured results of cross-band isolations at the LB and HB are generally in good agreement, except that the measured S_{13} in Position I and the measured S_{24} in Position II are slightly lower than the simulated values at the HB. This is attributed to the extra losses of the coaxial cables at the higher frequency. According to the measured results, the isolations between the LB and HB ports are all > 25 dB and > 29 dB at the LB and HB, respectively.

The measured radiation patterns of the HB antennas when they are in Positions I and II are given in Figures 3.49 and 3.50, respectively. For Position II, since HB3 and HB4 are symmetrical about a diagonal of the ground plane, only patterns of HB3 are given. According to the measured results, the HB radiation patterns are well restored despite the presence of the LB antennas, and the results are also consistent across the entire HB.

The measured radiation patterns of LB2 for Position I and II are all given in Figure 3.51, showing good consistency with the simulation. It can be concluded that the LB patterns are unaffected by the modified HB antenna under different relative positions between the LB and

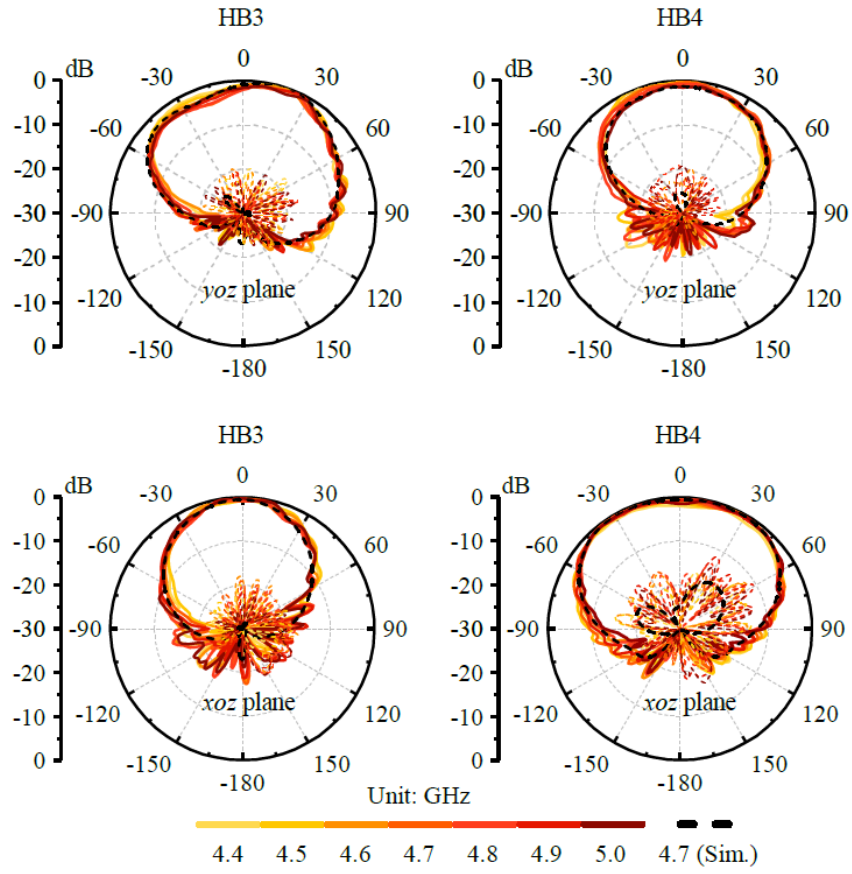


Figure 3.49: Measured radiation patterns of the two HB dipoles in the xoz and yoz planes when the HB antennas are standing at Position I.

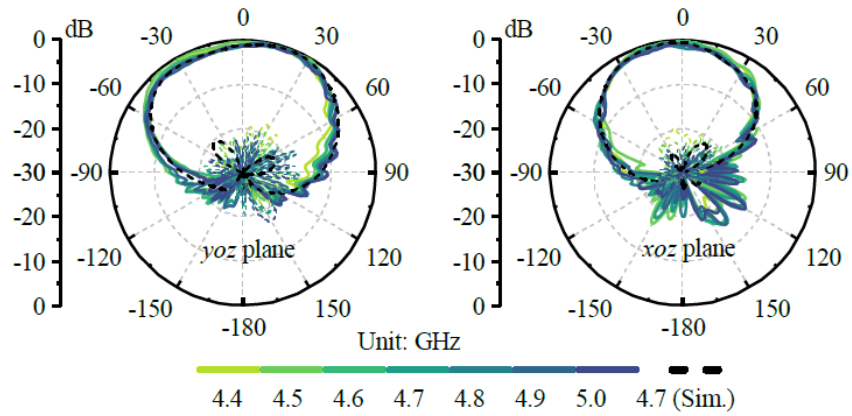


Figure 3.50: Measured radiation patterns of HB3 in the xoz and yoz planes when the HB antennas are standing at Position II.

HB antennas. In addition, the simulated and measured radiation patterns show very low cross-polarization radiation, further validating the effectiveness of the proposed methods in restoring

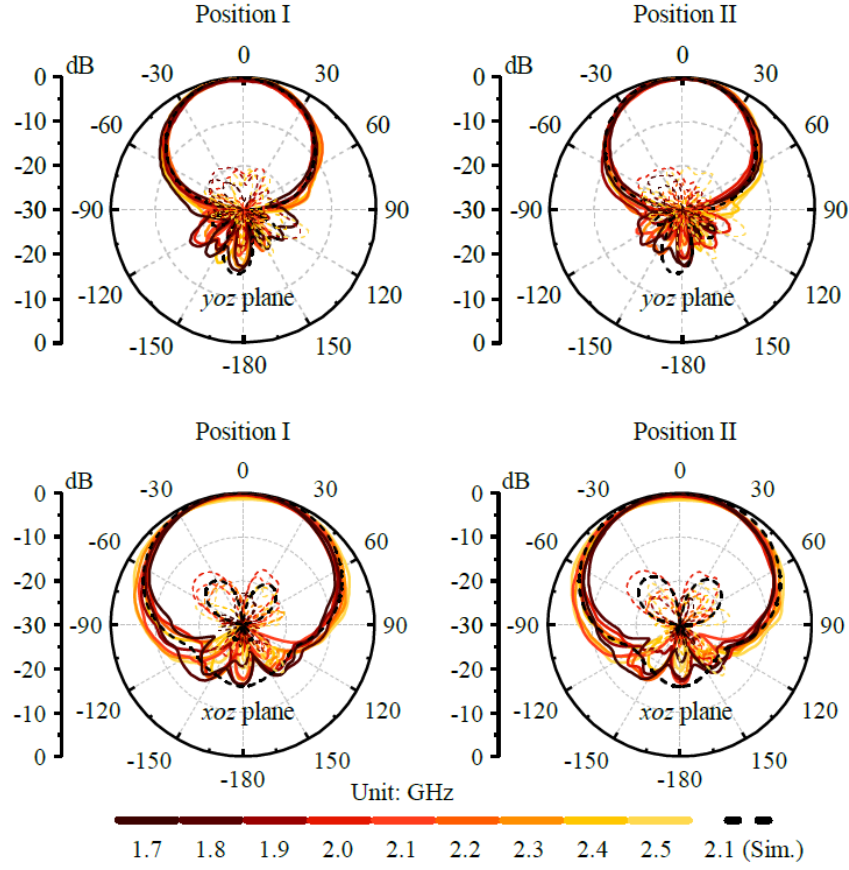


Figure 3.51: Measured radiation patterns of LB2 in the xoz and $yo z$ planes when the HB antennas are standing at Position I or II.

the radiation patterns.

To showcase the effectiveness of our proposed methods, Table 3.1 presents a comparison of cross-band interaction suppression techniques implemented in commonly used side-by-side configurations. Among the comparable works, only [16], [39] and our work are capable of dealing with all three types of interaction simultaneously. Our approach stands out for achieving the highest isolation compared to other methods owing to the three multi-functional structures and it is the only one that can address the cross-band scattering issue at the LB. Moreover, it is also the simplest and most cost-effective. In [39], extra modified metal fences (MMF) are used for each HB element to ensure good in-band isolation, while [16] employs two additional substrate layers between the HB and LB antenna and double-layered LB baluns. In contrast, our proposed interaction suppression structures are directly printed on the existing substrates, significantly reducing costs and simplifying fabrication. This makes our method ideal for 5G-and-beyond base station applications.

Table 3.1: Comparison of Cross-Band Interaction Suppression Techniques Implemented in Side-by-Side Antenna Configurations

Ref.	Proposed structure	Suppression function			Scattering-suppression /Operating band (GHz)	Cross-band isolation (dB)	In-band isolation (dB)
		Cross-band scattering	Cross-band coupling	In-band coupling			
[36]	Choke	at HB			0.82-1.0 (19.8%, LB) 1.71-2.28 (28.6%, HB)	N/A	N/A
[19]	Split-slot resonator	at HB			1.68-2.85 (51.7%, LB) 3.3-3.6 (8.7%, HB)	N/A	> 19 (HB)
[39]	FSS radiator	at HB			1.8-2.7 (40%, LB)	> 20 (LB)	> 20 (HB)
	Square ring		at LB		3.3-3.8 (14.1%, HB)	> 20 (HB)	
	Metal fence			at HB			
[16]	Mushroom patch	at HB			1.71-2.17 (23.7%, LB)	> 20 (LB)	> 20 (HB)
	Two-layer balun		at HB		3.3-3.8 (14.1%, HB)	> 20 (HB)	
	Parasitic loop		at LB				
	Cross-strip			at HB			
[13]	Split-ring resonator	at HB			0.69-0.96 (32.7%, LB) 3.3-4.2 (24%, HB)	N/A	N/A
This work	2.5-D cloak	at HB	at HB	at HB	1.64-2.56 (43.8%, LB)	> 23 (LB)	> 23 (HB)
	DGS	at HB	at HB		4.4-5.0 (12.8%, HB)	> 25 (HB)	
	Resonance shifter	at LB	at LB	at HB			

3.7 Conclusion

Mitigating cross-band scattering and coupling has become a major research focus in recent years. What sets this work apart from previous studies is that it addresses both of these issues simultaneously, without requiring any modifications to the array topology. Overcoming this challenge required a holistic approach that utilized the 2.5-D cloak, the DGS on the balun ground, and the resonance shifters. The DGS has dual functionalities, while the 2.5-D cloak and the resonance shifter are triple-functional, i.e., they can also reduce in-band coupling between adjacent HB antennas. These innovative methods work together seamlessly to achieve superior results in mitigating unwanted interactions between antennas in a collocated multi-band antenna array. The loading methods are tested under two polarizations, two array configurations, and various relative positions to demonstrate their robustness. An LB cross-dipole and an HB cross-dipole using the proposed techniques were fabricated and assembled. Measured

results agree very well with the simulated ones, validating the effectiveness and robustness of the proposed cross-band interaction suppression solution.

Chapter 4

Wide-Band Suppression of Scattering and Coupling in Dual-Band Antenna Array

4.1 Introduction

Achieving simultaneous suppression of cross-band scattering, cross-band coupling, and in-band coupling while maintaining the normal operation of antennas across wide bands presents a significant challenge. This is due to the requirement for both the LB and HB antennas to possess not only a wide impedance matching bandwidth but also a wide suppression bandwidth in which they do not interfere with the performance of other antennas. Extending all four bandwidths often conflicts with one another. Moreover, the simulation and optimization of numerous parameters in multi-functional structures are extremely challenging and time-consuming through a trial-and-error process.

This work successfully addresses these two challenges, and the novelty and contributions are outlined as follows:

(i) A dual-band antenna array is developed with operating bandwidths in the LB and HB ranges within 0.69-1.00 GHz (36.7%) and 1.63-2.83 GHz (53.8%), respectively. It is noteworthy that these operating bandwidths are the overlaps of impedance matching and suppression bandwidths of the antennas. The broadest HB suppression bandwidth is achieved with simultaneous

suppression of cross-band scattering, cross-band coupling, and in-band coupling.

(ii) A new segmented spiral structure is developed as the LB radiator with the precise guidance of characteristic mode analysis (CMA) rather than trial and error. It exhibits reduced radar cross-section (RCS) over an ultra-wideband of 1.10-4.29 GHz (118%), representing the broadest “invisible” bandwidth to our knowledge.

(iii) This work successfully addresses the conflicting tasks of maximizing the suppression bandwidth and impedance matching bandwidth for both the LB and HB antennas. For the LB antenna, serial resonators compensate for impedance matching degradation caused by maximizing scattering suppression with the segmented spiral. For the HB antennas, parallel resonance shifters connected to the HB balun mitigate negative impact of the HB antennas on LB radiation without significantly affecting HB antenna matching.

This chapter is organized as follows. Section 4.2 presents the design of the segmented spiral structure under the guidance of CMA, along with the design of the serial resonators for wide-band matching. Section 4.3 describes the configuration of the HB antennas with the parallel resonance shifters. The simulated scattering and coupling suppression results in the dual-band antenna array are provided in Section 4.4. Section 4.5 presents the experimental results of the fabricated antenna array. Finally, conclusions are drawn in Section 4.6.

4.2 Low-Band Antenna With Segmented Spirals and Serial Resonators

4.2.1 Characteristic-Mode-Guided Design of Segmented Spiral

When the LB antenna is positioned adjacent to the radiating HB antennas as in Figure 4.1, HB currents are induced on the LB radiator under the excitation of the HB wave. These induced currents give rise to secondary radiation, which interferes with the original HB radiation and results in distorted total radiation patterns in the HB. This phenomenon is known as **cross-band scattering** in the HB. The induced currents on the LB radiator also flow into the antenna ports through the LB balun, causing **cross-band coupling** in the HB [1]. Furthermore, the secondary radiation from the LB radiator excites secondary induced currents on the other HB antennas, intensifying **in-band coupling** between HB antennas. Consequently, the cross-band

scattering, cross-band coupling, and in-band coupling in the HB are all attributed to the HB induced currents on the LB radiators. Note that the target LB frequency range is from 0.69 to 0.96 GHz, while the HB frequencies can encompass 1.71-2.69 GHz and/or 3.3-3.8 GHz, as antennas working at these bands are usually collocated in 3G/4G/5G cellular base stations.

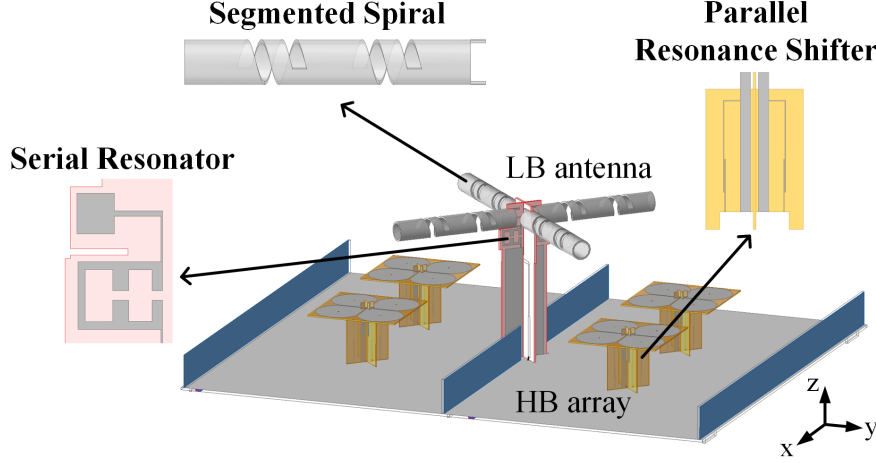


Figure 4.1: Configuration of dual-band array with one LB antenna interleaved with four HB antennas.

According to the characteristic mode theory, the total induced currents on the LB radiator under the given HB excitation can be decomposed into an infinite series of currents of orthogonal modes, and the contribution of each mode can be weighted by a complex modal weighting coefficient (MWC). In practice, the total induced currents can be approximated by combining currents of a few significant modes with larger $|MWC|$ values in the HB [116]. From the perspective of the characteristic mode theory, the desired interference suppression can be accomplished by reducing the $|MWC|$ values associated with the significant modes [109], [110].

The cylindrical tube shown in Figure 4.2(a) is the most traditional structure used as the radiating arm of a dipole antenna. This tube with a length of 108 mm and a diameter of 14 mm can be matched within 0.69-0.96 GHz as antenna radiators, and it will be compared as a reference case in the following parts. When the tube is illuminated by a plane wave polarized along its axis, the $|MWC|$ values of different modes are shown in Figure 4.2(b). The significant modes include the Modes t1 and t2, with the Mode t1 being the most dominant mode. Although the $|MWC|$ reaches its peak around 1.1 GHz, the resonance frequency is closer to the center frequency of the LB when the two tubes are combined into a dipole. Furthermore, its value remains high at higher frequencies, suggesting that the tube can be easily excited by waves at

higher frequencies as well. The modal current of Mode t1 at 2.6 GHz is shown in Figure 4.2(c) as a demonstration.

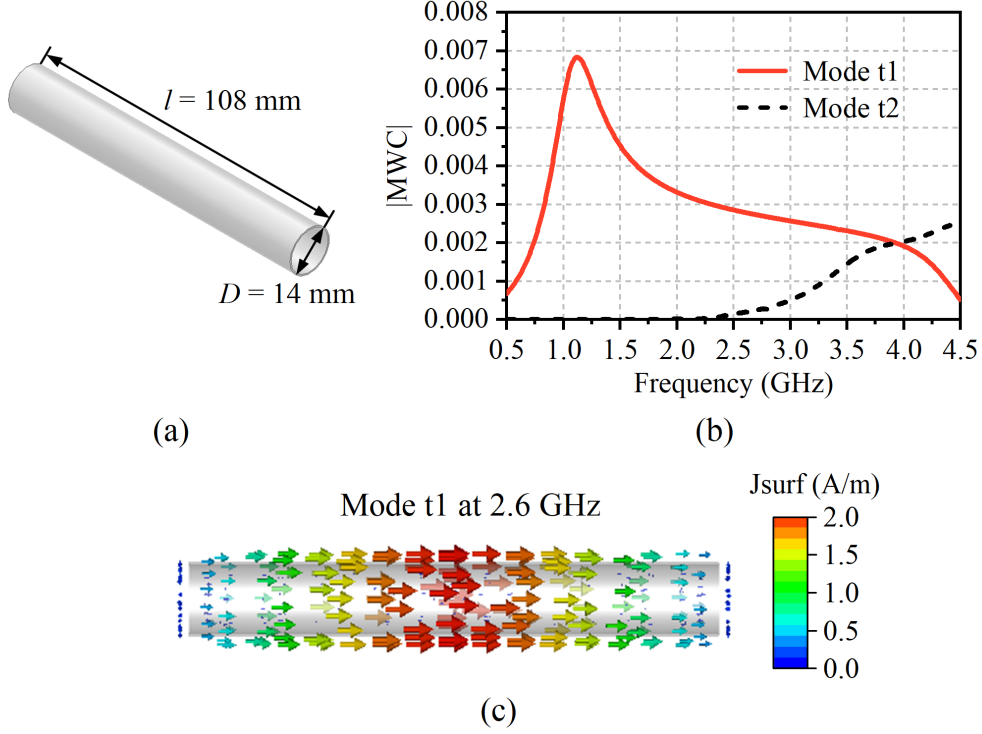


Figure 4.2: (a) Geometry of metal tube, (b) $|MWC|$ of the metal tube excited by plane wave, and (c) modal current distribution of Mode t1 at 2.6 GHz.

The target of this work is to modify the tube-based radiator to expand the scattering and coupling suppression band as much as possible while ensuring that the matching band covers at least 0.69-0.96 GHz. Therefore, it is imperative to significantly reduce the $|MWC|$ values of Mode t1 at higher frequencies while keeping the $|MWC|$ values at lower frequencies unchanged.

The spiral structure proposed in our previous work [38] appears to offer a promising solution for suppressing the strong induced current at higher frequencies on the cylindrical tube while allowing the flow of longitudinal currents at lower frequencies. The evenly distributed spiral is achieved by etching a continuous spiral slot of a constant pitch and width on the tube, as shown in Figure 4.3(a). Figure 4.3(b) shows the $|MWC|$ of the significant modes of the even spiral. The Mode e1 corresponds to the Mode t1 in Figure 4.2(b). It is evident that the spiral structure effectively suppresses Mode e1 over higher frequencies. However, it introduces several higher-order modes, which reduce the suppression bandwidth. To widen the suppression bandwidth, higher-order modes (Modes e2 and e3) need to be removed or shifted. The modal

E-field distributions of Mode e2 at 2.7 GHz and Mode e3 at 3.4 GHz are shown in Figure 4.3(c). For Mode e2, the E-field is much stronger in the middle part and the two regions near the ends of the slot. For Mode e3, the E-field is also stronger in the middle part. These two higher-order modes can be eliminated by short-circuiting the slots in the areas with strong E-fields.

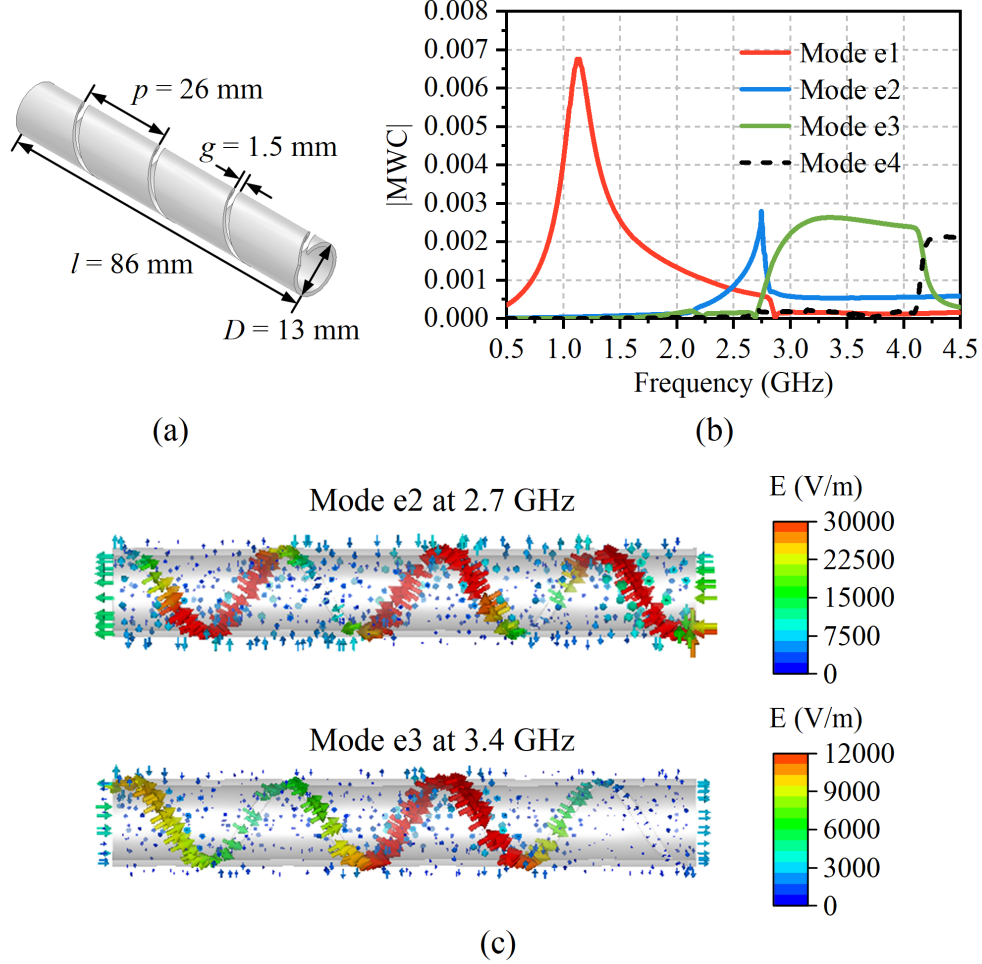


Figure 4.3: (a) Geometry of even spiral, (b) |MWC| of even spiral excited by plane wave, and (c) modal E-field distribution of Mode e2 at 2.7 GHz and Mode e3 at 3.4 GHz.

After short-circuiting the even spiral in the middle, it breaks into two segmented spirals, as shown in Figure 4.4(a). As given in Figure 4.4(b), the Modes e2 and e3 in Figure 4.3(b) are eliminated successfully, however, another higher-order mode named Mode 2 appears. Figure 4.4(c) presents its modal E-field distribution at 3.4 GHz. The areas with the strongest E-field are located in the two slots. However, the approach of short-circuiting the slots cannot be used in this case, as it would reshape the segmented spiral back into the original cylindrical tube. Therefore, instead of completely eliminating this mode, a more feasible approach is to suppress Mode 2 by modifying the slot widths of the two segmented spirals.

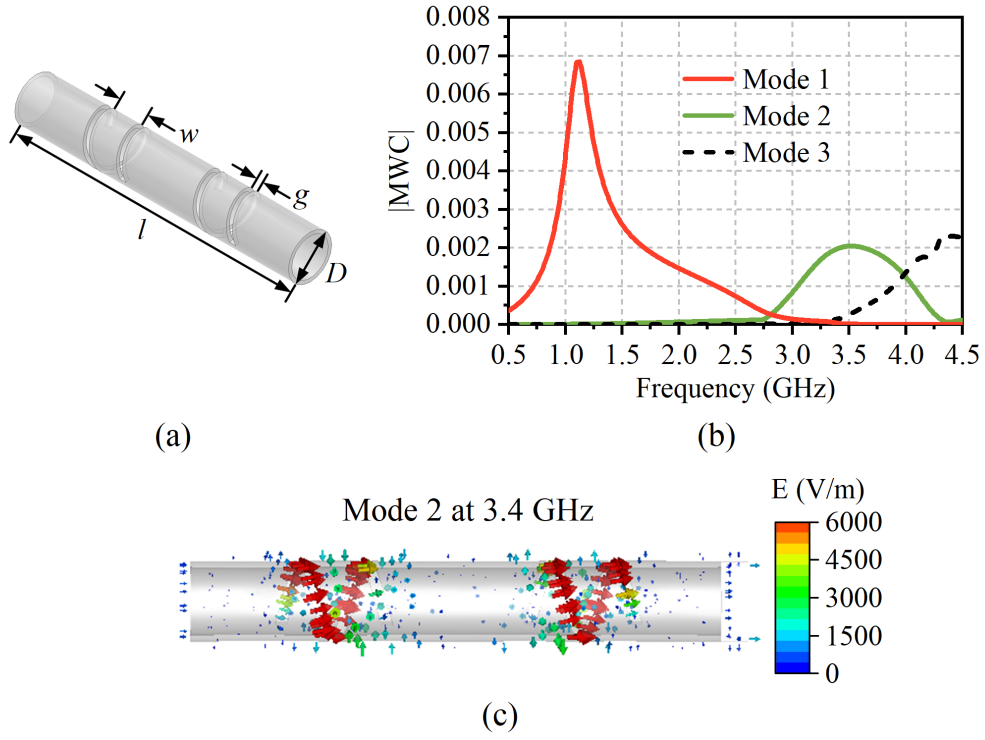


Figure 4.4: (a) Geometry of segmented spiral, (b) $|MWC|$ of segmented spiral excited by plane wave, and (c) modal E-field distribution of Mode 2 at 3.4 GHz.

The $|MWC|$ of segmented spiral with different slot widths are shown in Figure 4.5. As the slot width (g) increases from 2.1 mm to 6.1 mm, the $|MWC|$ of Mode 2 gradually decreases, thereby widening the suppression bandwidth. In addition, increasing g also slightly reduces the $|MWC|$ of Mode 1 at higher frequencies, further enhancing the effectiveness of interference suppression.

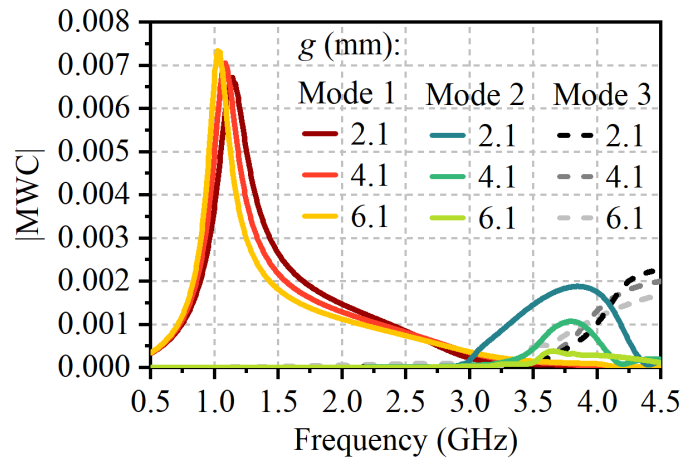


Figure 4.5: $|MWC|$ of segmented spiral with different width of the slots.

To demonstrate the advantage of the segmented spiral in scattering suppression, its monostatic

radar cross sections (RCSs) with different slot widths are compared to those of the reference cylindrical tube and the even spiral in Figure 4.6. The tube has RCS around -20 dBsm in the suppression band. The even spiral can reduce the RCS but is only effective in a limited frequency range. In contrast, the segmented spiral effectively reduces the RCS across a much wider band of 1.10-4.29 GHz (118%), with a maximum RCS reduction of 19 dB. Figure 4.7 illustrates the bistatic RCSs at 2.8 GHz of both the tube and the segmented spiral when $g = 4.1$ mm as an example. As observed from the figure, the developed segmented spiral realizes a remarkable reduction of scattering energy in all directions.

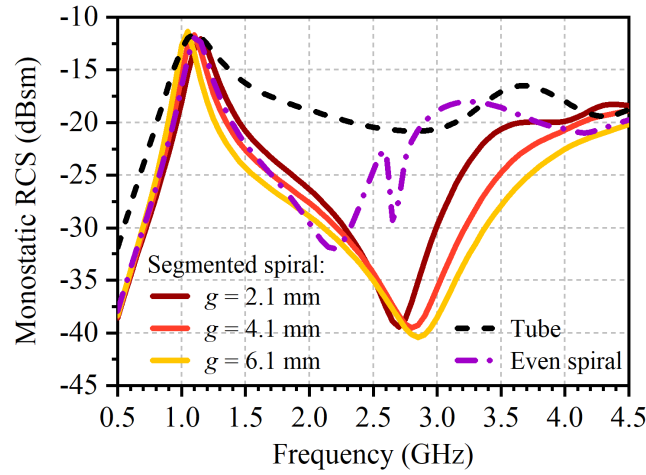


Figure 4.6: Monostatic RCSs of the tube, even spiral, and segmented spirals with different g .

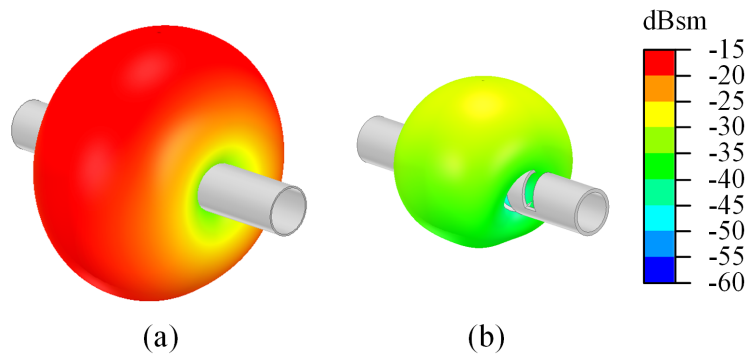


Figure 4.7: Bistatic RCSs of (a) the tube and (b) the segmented spiral at 2.8 GHz when $g = 4.1$ mm.

Moreover, it is noticed from Figure 4.6 that a larger g broadens the bandwidth with low RCS. Specifically, when $g = 6.1$ mm, the bandwidth over which the RCS is reduced when using the segmented spiral is almost twice as wide as when using the even spiral. The wideband

RCS reduction of the segmented spiral is attributed to the effective suppression of higher-order modes. The RCSs of different structures are consistent with the analysis with $|MWC|$ values.

Throughout the entire design process of the segmented spiral structure, CMA serves as an effective method to intuitively reflect the higher-order modes that contribute the most to interference in different frequency bands. It provides precise guidance for optimizing the structure to eliminate or suppress higher-order modes through mode currents or mode E-fields, thus achieving the final segmented spiral with outstanding interference suppression performance.

The segmented spiral design also exhibits a significant generalizability across different frequency bands, relative positions, and polarizations of the antennas. It can be used in various scenarios when multiple antennas working at different frequencies are placed in close proximity.

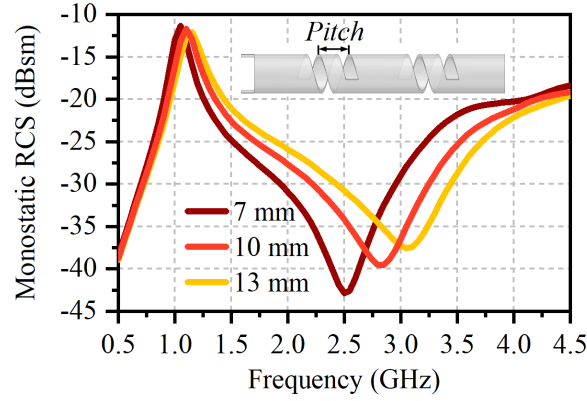


Figure 4.8: Monostatic RCSs of segmented spirals with different pitches.

Both LB and HB can be adjusted under the guidance of CMA by altering the structural parameters. The LB is primarily determined by the overall length of the spiral. The HB can be shifted by adjusting the *pitch* of spirals as shown in Figure 4.8. Moreover, as demonstrated in Figure 4.9, simply scaling all structural parameters of the segmented spiral by the same factor enables its application at other frequencies.

When the segmented spiral is employed as the LB antenna radiator, regardless of the polarizations of HB waves or the relative positions of the LB and HB antennas, the interference from the LB antenna on HB radiation will be significantly reduced. Figure 4.10 shows the monostatic RCSs of the segmented spiral under different incident angles, where all monostatic RCS curves exhibit good consistency.

The tolerance of the segmented spiral on the monostatic RCS is illustrated in the Figure 4.11.

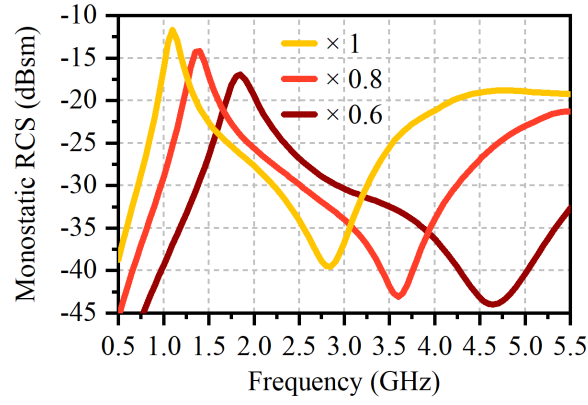


Figure 4.9: Monostatic RCSs of segmented spirals with different scaling factor.

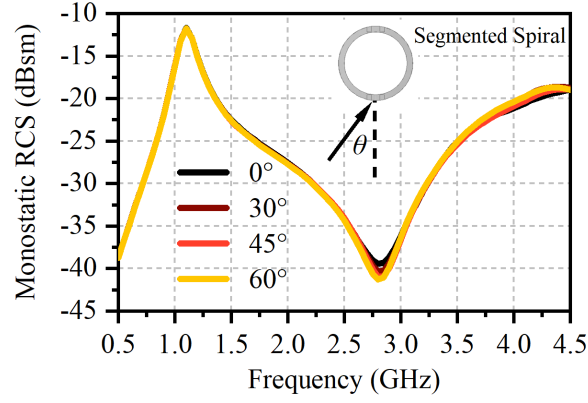


Figure 4.10: Monostatic RCSs of segmented spirals under different incident angles.

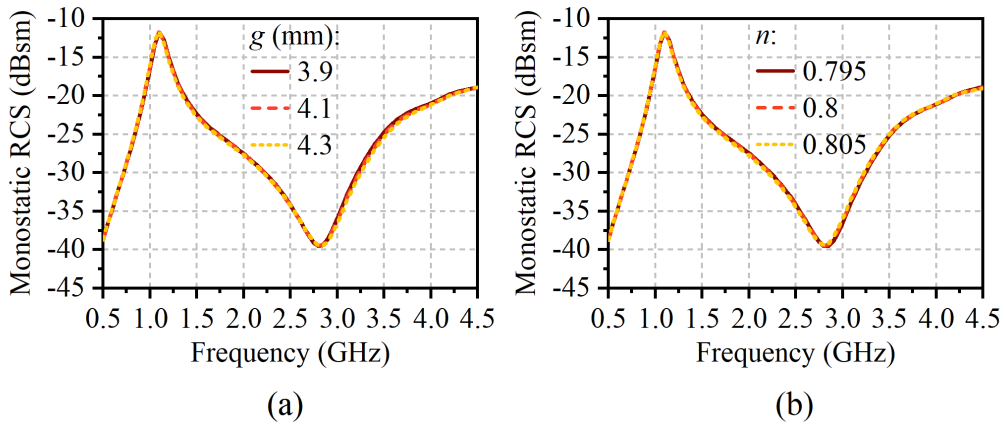


Figure 4.11: Monostatic RCSs of segmented spirals under different structure parameters.

The gap width (g) is varied within ± 0.2 mm, while the corresponding variation in the gap length caused by changes in the number of turns (n) is also approximately ± 0.2 mm. It can be observed that the monostatic RCS is not noticeably affected by these tolerances. In the

subsequent fabrication, the segmented spiral is realized using a laser cutting process, with a machining tolerance of ± 0.1 mm.

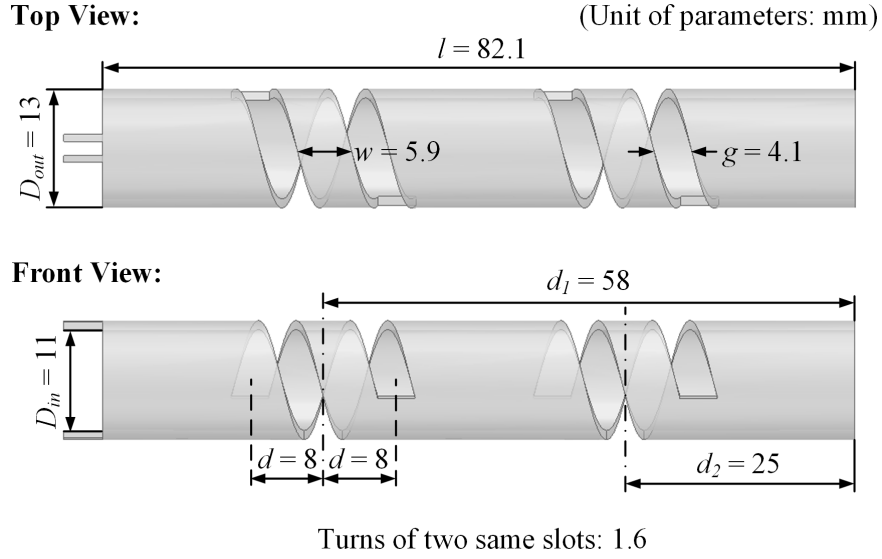


Figure 4.12: Geometry of the final segmented spiral.

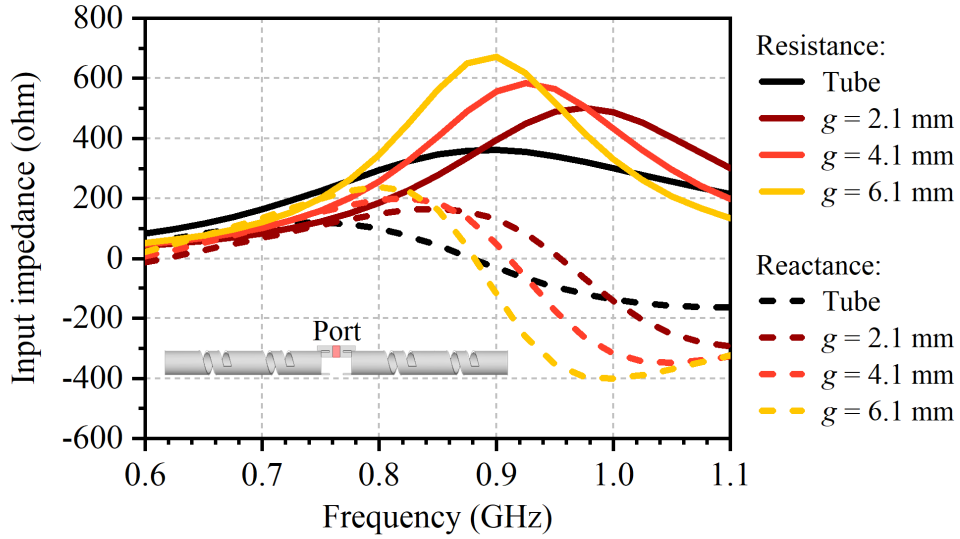


Figure 4.13: Input impedance of the LB dipole using the segmented spiral with different g or the reference tube.

The geometry of the final segmented spiral is shown in Figure 4.12, which is formed by etching two identical slots with turns of 1.6 on the tube. Although the suppression bandwidth is wider with a larger g , it is necessary to consider the interference suppression and matching characteristics at the same time since the segmented spiral is also used as the radiator of

antenna. A discrete port presented in Figure 4.13 is introduced between the two arms to measure the input impedance of the segmented spiral dipole. As shown in Figure 4.5, Figure 4.6, and Figure 4.13, with increasing g , the curves of Mode 1's $|MWC|$, RCS, and input impedance all become steeper at lower frequencies, which leads to greater challenges in LB matching. With an overall consideration of matching and suppression, $g = 4.1$ mm is chosen in our design. Compared with the reference tube, the variation in the resistance and reactance of the segmented spiral is significantly larger. An advanced impedance matching circuit needs to be designed to achieve satisfactory matching performance, and this will be presented in the next subsection.

4.2.2 Wide-Band Impedance Matching of Segmented Spiral Antenna by Serial Resonators

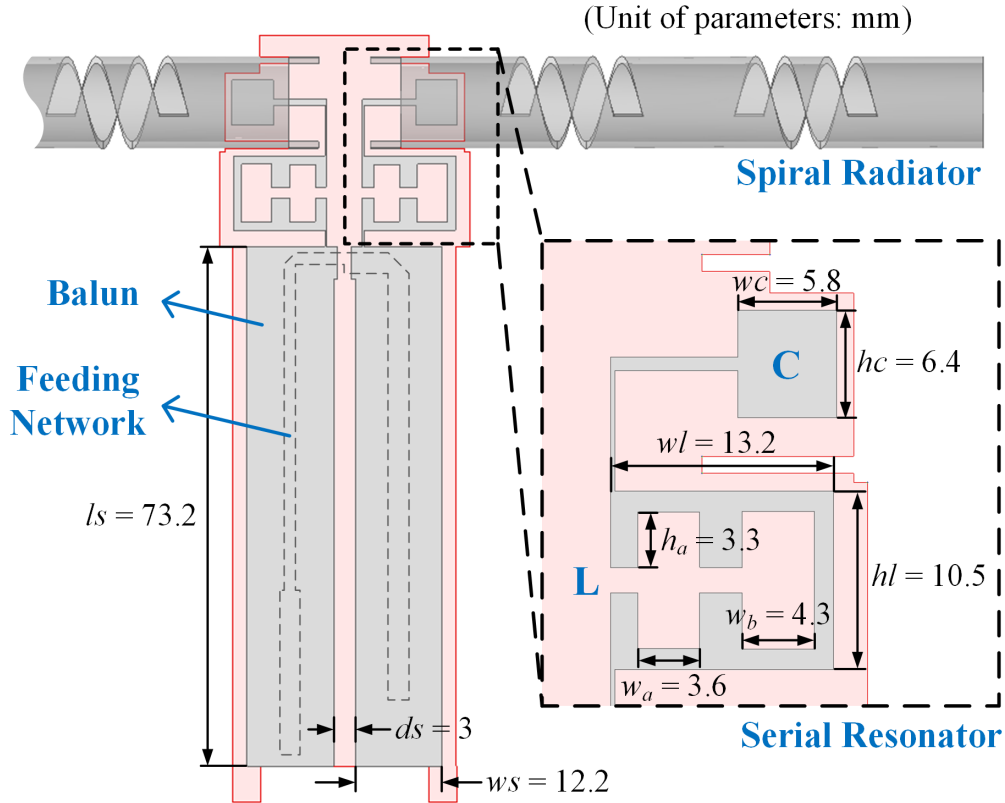


Figure 4.14: Geometry of the serial-resonator-integrated balun for segmented spiral antenna matching.

Serial resonators, as shown in Figure 4.14, are added between the radiators and the balun to achieve impedance matching for the segmented spirals. The balun along with the serial

resonators and feeding network are printed on two sides of an FR4 substrate with a thickness of 1.524 mm, a relative permittivity of 4.4, and a loss tangent of 0.02. A metal pad is inserted into the spiral radiator at the feeding position, forming a capacitor with the metal structure of the radiator. A double E-shaped strip is further added between the capacitor and the balun, serving as a quasi-inductor. The varied width of the metal strip of the double E-shaped strip also introduces some capacitance, providing more flexibility for parameter optimization.

Figure 4.15 illustrates the variation of the input impedance of the segmented spiral dipole with different loading conditions. The frequency band shown in the Smith chart is 0.69-1.00 GHz. For the segmented spiral dipole, its input impedance curve shown with “Spiral only” arc is far away from the matching point in the center of the Smith chart, and the corresponding radius of the arc is very large. Therefore, it is difficult for the spiral dipole to achieve impedance matching using the conventional balun structure. To reduce the radius of the “Spiral only” arc, the capacitor is first added, and the arc is rotated counterclockwise along the equal resistance circle in the Smith chart, obtaining a “Spiral + C” arc with a much smaller radius. However, it is further away from the matching point. The addition of the quasi-inductor makes the arc rotate clockwise. Compared with the arc of “Spiral only”, the arc of “Spiral + L + C” is closer to the matching point and the corresponding radius is smaller, bringing the possibility of impedance matching. Finally, the conventional balun with feeding network is introduced so that the “Spiral + L + C + balun” arc can wrap around the matching point of Smith chart with a small radius, indicating successful impedance matching.

The S parameters and radiation patterns in the yo -plane of the matched spiral antenna are presented in Figure 4.16(a) and (b), respectively. The reflection coefficients ($|S_{11}|$ and $|S_{22}|$) are less than -10 dB from 0.69 GHz to 1.00 GHz, and the port isolation $|S_{12}|$ exceeds 25 dB within this frequency range. The wideband matching performance validates the efficacy of the serial-resonator-integrated balun design. Stable radiation patterns across the band can be observed in Figure 4.16(b), demonstrating that the utilization of the segmented spiral radiator does not compromise the radiation performance of the antenna.

It is important to note that the coupled, non-contact connection between the serial resonator and the segmented spiral does not affect the RCS suppression characteristics of the segmented spiral. The monostatic RCSs of the segmented spirals with and without the serial resonator are compared in Figure 4.17. The RCS curves with and without the serial resonator overlap with

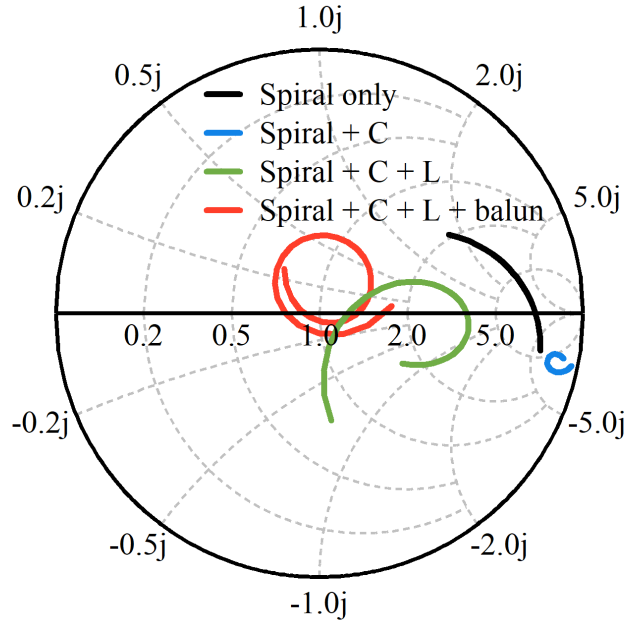


Figure 4.15: Input impedances of the segmented spiral dipole with different loading conditions.

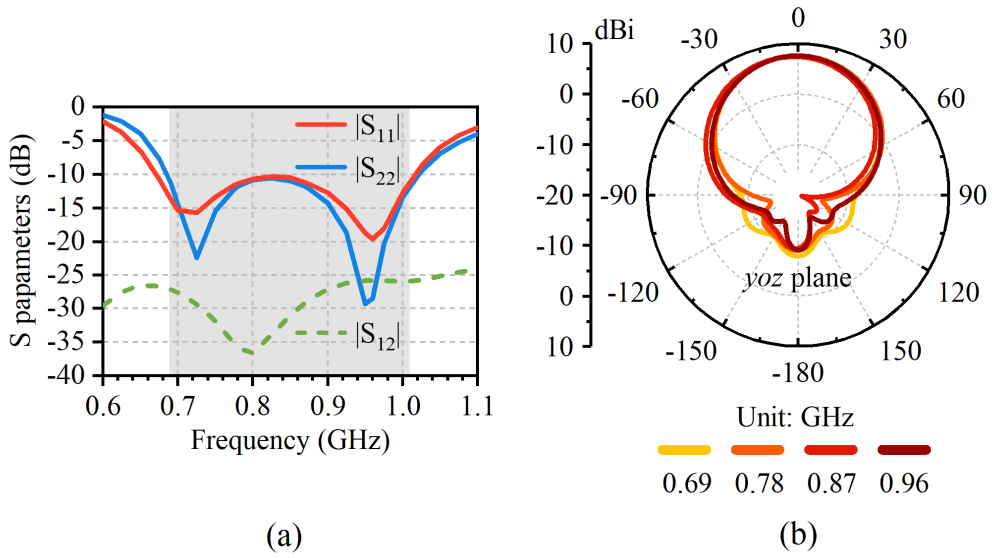


Figure 4.16: (a) S parameters and (b) radiation patterns in yoz -plane of the segmented spiral antenna.

each other except for a new suppression point at 3.95 GHz introduced by the serial resonator, but it is outside the band of interest. Therefore, the scattering-suppression characteristics of the segmented spiral are maintained after the integration of the serial resonator.

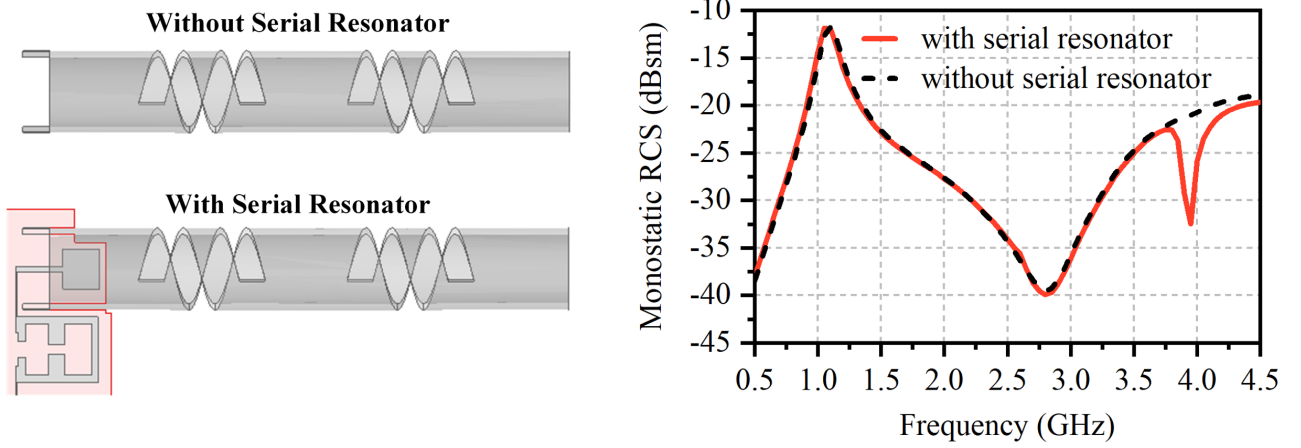


Figure 4.17: Monostatic RCSs of segmented spirals with and without serial resonator.

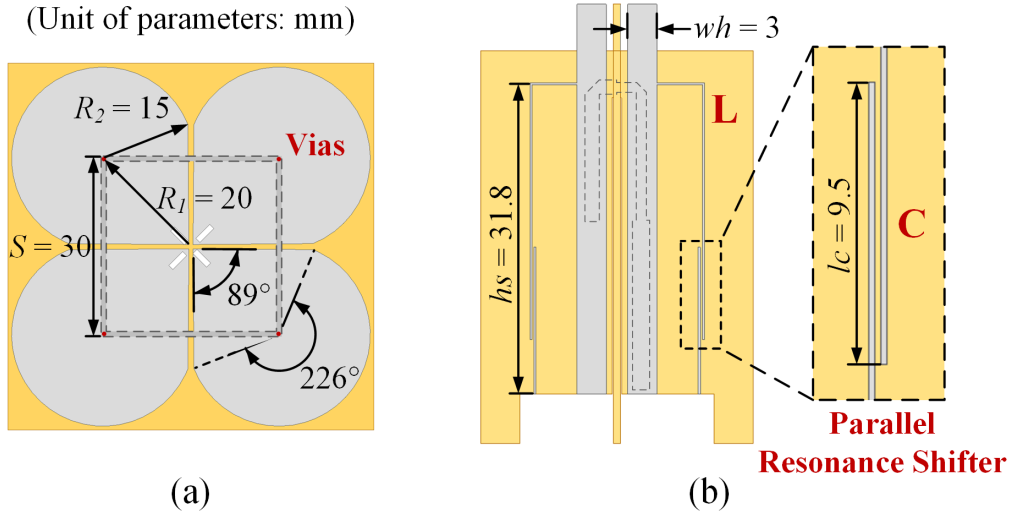


Figure 4.18: Geometry of (a) the HB radiator and (b) the HB balun.

4.3 High-Band Antenna With Parallel Resonance Shifters

The geometry of the HB antenna developed in this work is shown in Figure 4.18. A four-leaf-clover shaped radiator and a square loop are etched on the two sides of a substrate with a thickness of 0.762 mm, a relative permittivity of 3.48, and a loss tangent of 0.0037. The radiator and the loop are connected to each other through four metallized vias, as illustrated in Figure 4.18(a). Figure 4.18(b) depicts the balun and feeding network of the HB antenna. In addition to the conventional balun structure, a pair of parallel resonance shifters, each composed of two closely coupled thin strips, is connected in parallel with the balun. One end of each parallel resonance shifter is linked to the upper section of the HB balun, while the other end is grounded. These thin strips serve as inductors, and the coupling between them creates a capacitive effect.

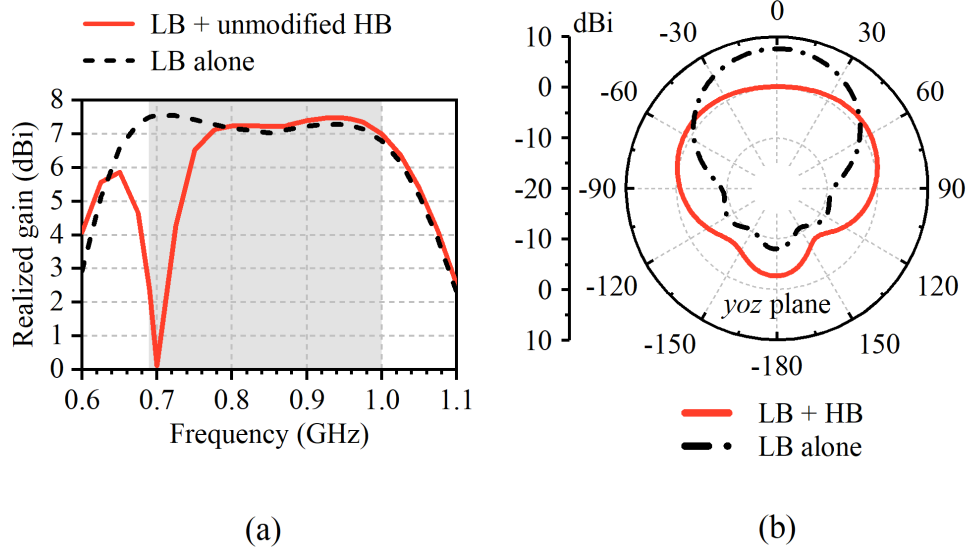


Figure 4.19: (a) Realized gain and (b) radiation patterns at 0.7 GHz of LB antenna.

The parallel resonance shifters are introduced to reduce the scattering of the HB antennas towards the LB radiation. When HB antennas are placed around the LB antenna in the traditional side-by-side configuration without the parallel resonance shifters, as depicted in Figure 4.1, the radiation performance of the LB antenna degrades substantially. Figure 4.19 illustrates the realized gain and radiation patterns of the LB spiral antenna at 0.7 GHz, both with and without the presence of HB antennas lacking parallel resonance shifters. It can be clearly observed that the boresight realized gain at 0.7 GHz sharply drops to 0 dBi, and the LB patterns at that frequency are severely distorted compared with the pattern when the LB antenna is working alone. These degradations result from the common-mode resonance of the HB antenna and the ground within the LB at around 0.7 GHz.

To understand this gain drop better, the E-field distribution when the LB antenna is excited at this frequency is plotted in Figure 4.20(a). Strong couplings among the LB and HB antennas are observed from the E-field distribution. The wave radiated from the LB antenna excites the HB antennas in common-mode, as evidenced by the current distribution on the HB baluns shown in Figure 4.20(b). The resonating HB antennas act as mushroom structures, creating an LC short circuit between the LB antenna and the ground. The coupling between the LB dipole arms and the HB aperture provides C, and the HB balun provides L. The wave radiated by the LB antenna is intercepted by the HB antenna, which results in strong LB induced currents. These induced currents are re-radiated and dissipated, resulting in distortion in the LB patterns

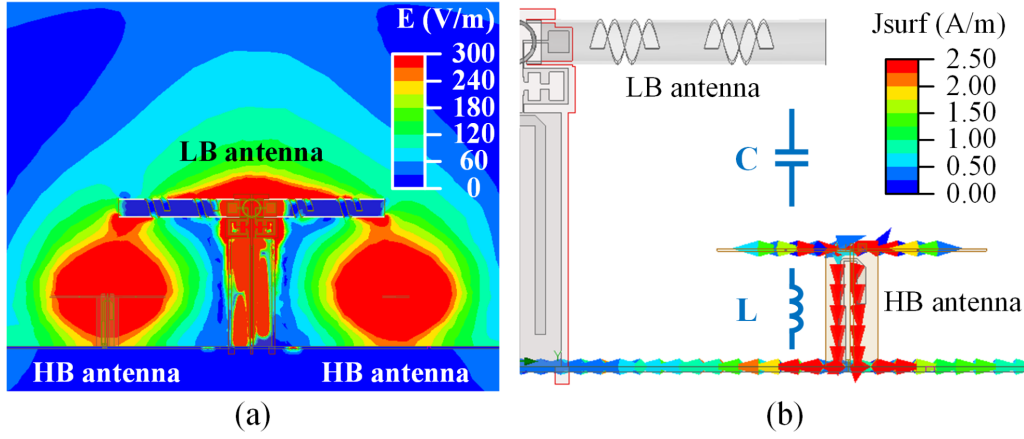


Figure 4.20: (a) E-field distribution near the dual-band array, and (b) induced LB currents on the HB antennas at 0.7 GHz when the LB antenna is excited.

and reduction in the LB boresight realized gain.

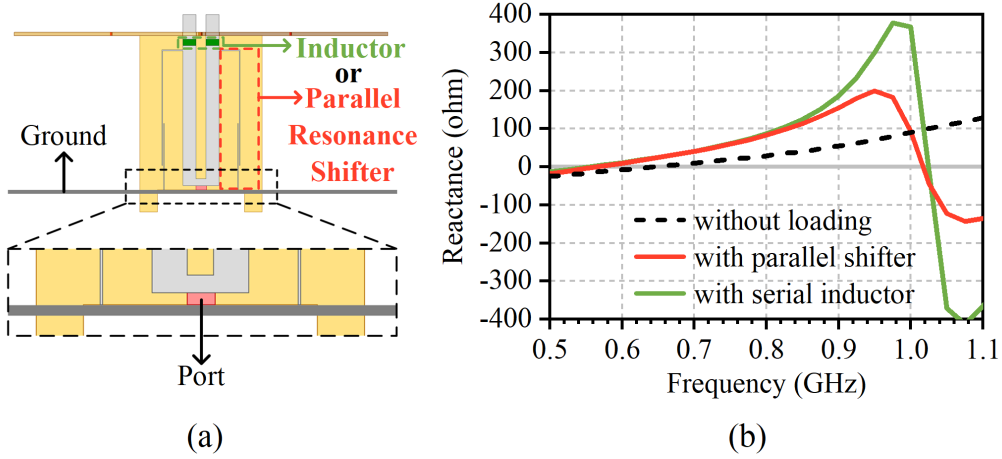


Figure 4.21: (a) Discrete port inserted to the HB antenna. (b) Input reactance of the HB antenna at the discrete port.

As shown in Figure 4.21(a), a discrete port is inserted between the HB balun and the ground, functioning as a probe to examine the common-mode resonance. The input reactance measured at the discrete port with different loading conditions is plotted in Figure 4.21(b). As indicated by the black dashed curve, the HB antenna fed by conventional balun exhibits a resonance near 0.7 GHz, causing the aforementioned distortion in the LB patterns.

With the inclusion of parallel resonance shifters, the resonance point shifts to a lower frequency at 0.56 GHz, as indicated by the red curve in Figure 4.21(b). This shift effectively places the resonance point well below the target LB range, which spans from 0.69 GHz to 1.00 GHz. It

is important to note that the desired resonance shift can also be achieved by inserting serial inductors (as illustrated by the green curve) or capacitors in the HB balun [47], [73]. However, the use of inductors or capacitors in series with the balun poses challenges for matching the HB antenna. For example, the operating bandwidths of the HB antennas reported in [47], [73] are only 12.8% and 25.6%, respectively.

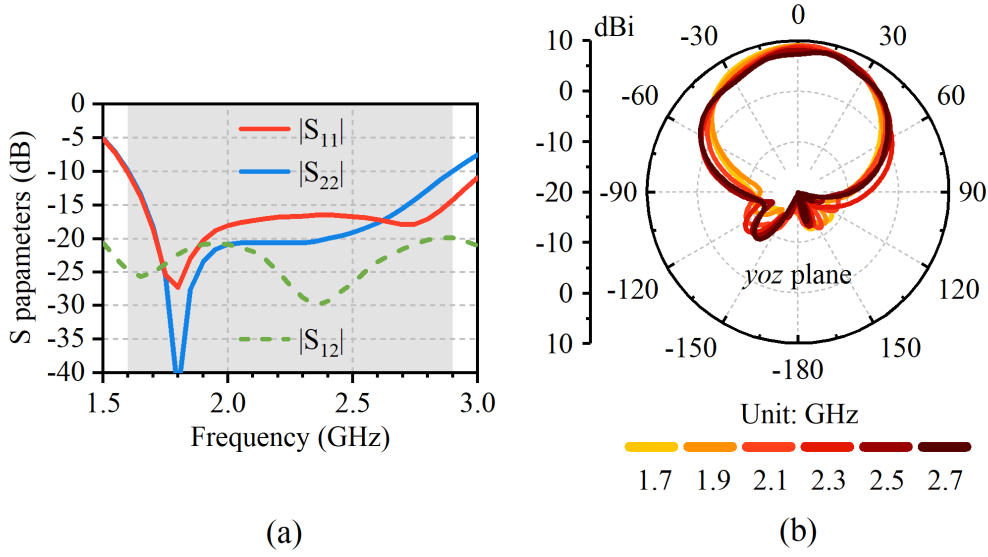


Figure 4.22: (a) S parameters and (b) radiation patterns of the HB antenna with parallel resonance shifters.

As demonstrated in Figure 4.22(a), the implementation of parallel resonance shifters with the HB antenna results in excellent matching across a wide band, spanning from 1.6 GHz to 2.9 GHz, achieving a bandwidth of 57.8%. Importantly, these parallel resonance shifters have no adverse effects on the radiation performance of the HB antennas, as evidenced by the radiation patterns presented in Figure 4.22(b).

The realized gain and radiation patterns of the LB antenna with the presence of the parallel-resonance-shifter-integrated HB antenna are shown in Figure 4.23(a) and (b). Stable gain and radiation patterns can be observed, verifying the effectiveness of the parallel resonance shifters in mitigating the influence from the HB antenna on the LB radiation.

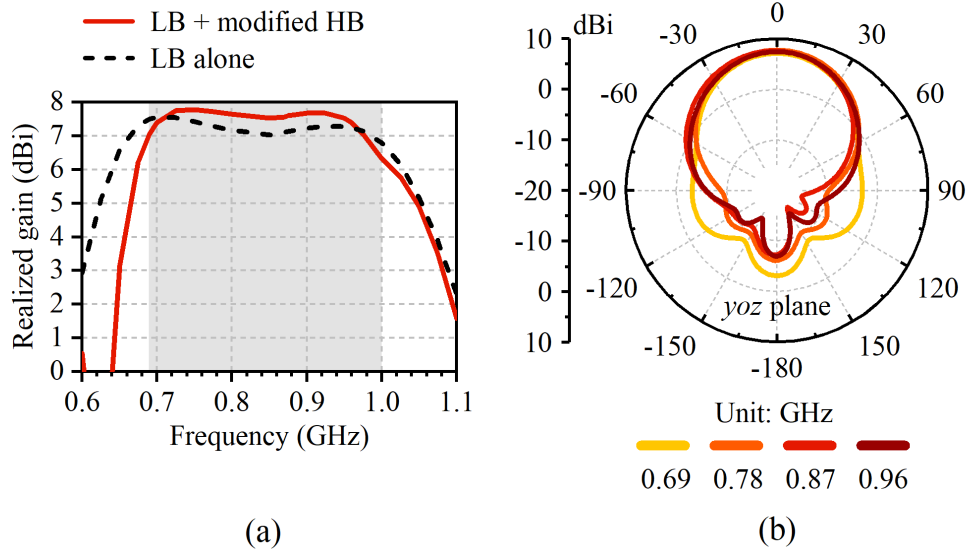


Figure 4.23: (a) Realized gain and (b) radiation patterns of the LB antenna with the presence of the parallel-resonance-shifter-integrated HB antenna.

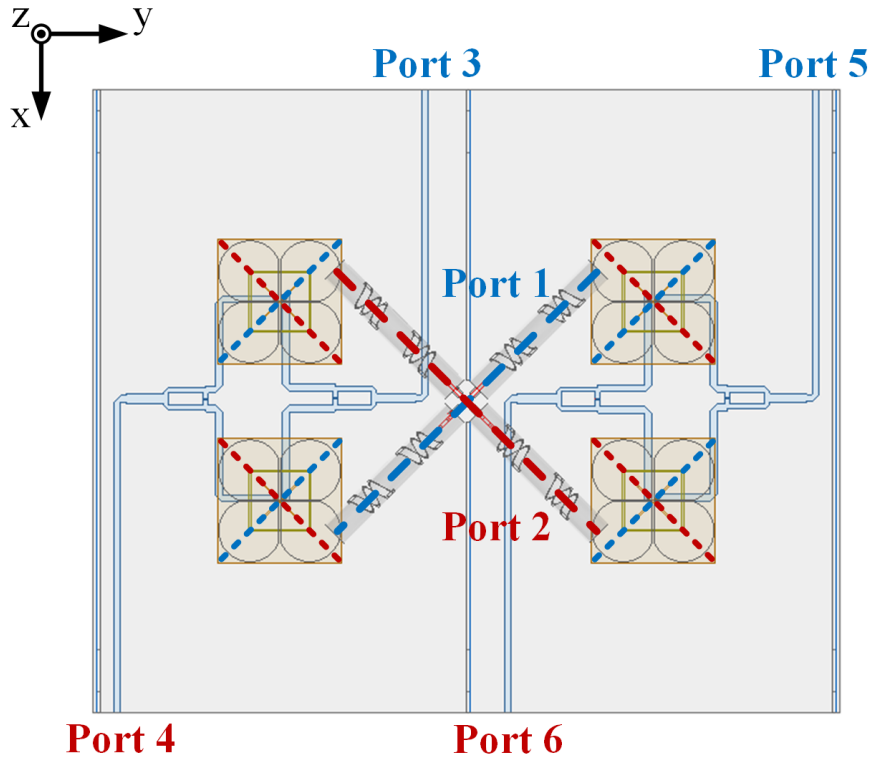


Figure 4.24: Configuration of interleaved dual-band antenna array.

4.4 Holistic Suppression of Scattering and Coupling in Collocated Dual-Band Array

The modified HB antennas and the segmented spiral LB antenna are arranged in the interleaved dual-band array to verify the suppression performance of the designed segmented spiral in the

HB. The array configuration is shown in Figure 4.24, with the marked ports and respective polarizations of the LB and HB antennas.

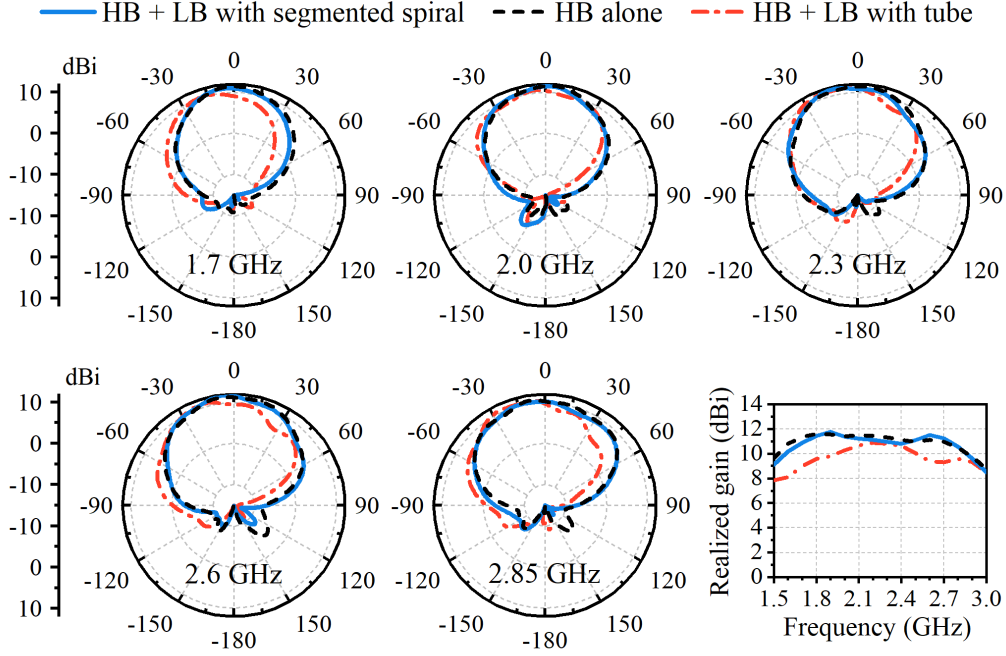


Figure 4.25: Simulated radiation patterns and realized gains of the HB antennas when Port 3 is excited, with or without the presence of different LB antennas.

When the HB Port 3 and Port 5 in the antenna array are individually excited, HB radiation patterns in the yo z plane under different cases are shown in Figures 4.25 and 4.26, respectively. In the reference case, the tube-based LB antenna significantly distorts the HB patterns and degrades the realized gain. In contrast, with the presence of the segmented spiral LB antenna, the HB patterns and realized gain closely resemble those when HB antennas operate alone across the entire HB.

The results show that the LB antenna with the segmented spiral arm is electromagnetically transparent to the HB, validating the effectiveness of the designed segmented spiral in suppressing cross-band scattering across the wide HB. The wideband scattering suppression capability of the designed LB antenna is attributed to the guidance provided by the CMA method in the structural design.

Besides its effectiveness in suppressing cross-band scattering, the segmented spiral as the LB radiator also has the capability to suppress cross-band coupling in the HB. Figure 4.27 presents the transmission coefficients between co-polarized LB and HB dipoles in the HB. When the

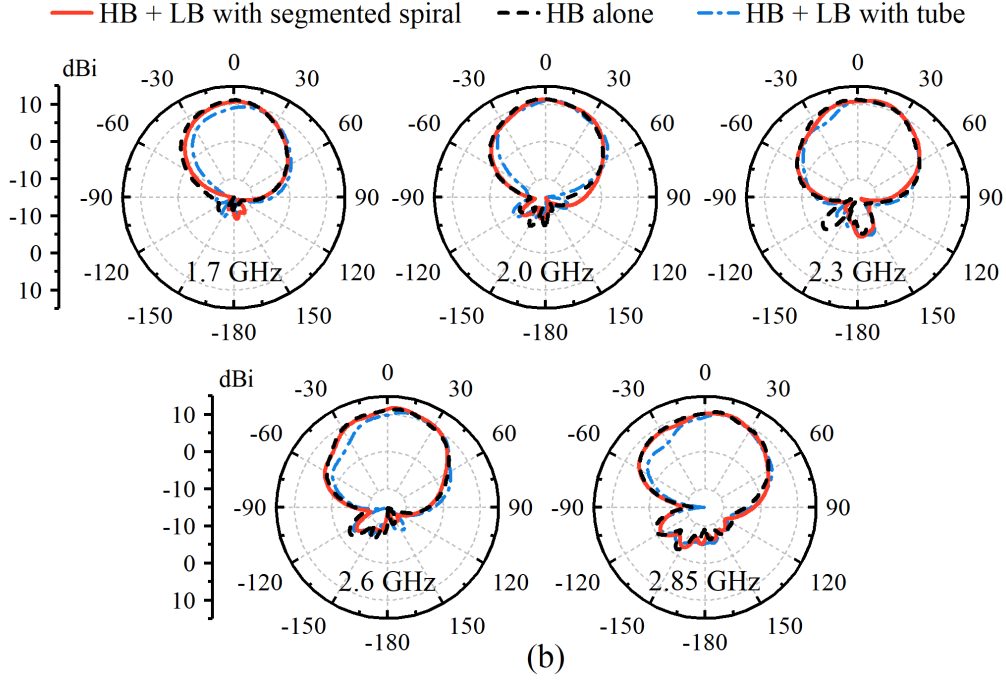


Figure 4.26: Simulated radiation patterns of the HB antennas when Port 5 is excited with and without the presence of different LB antennas.

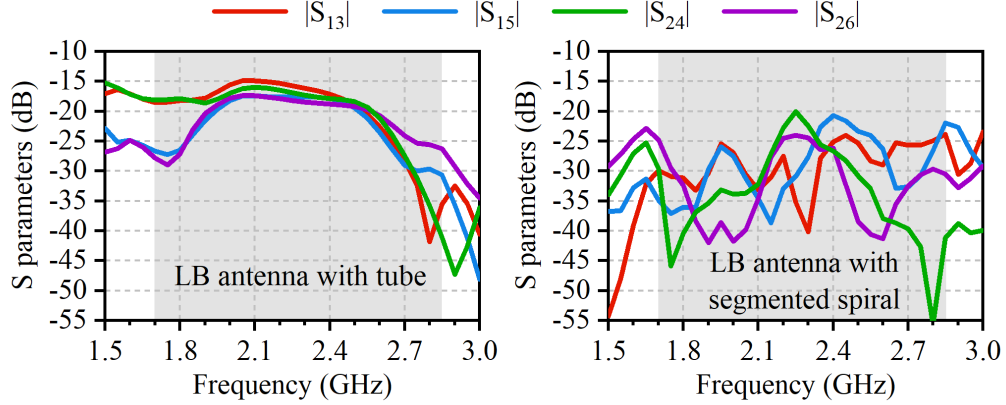


Figure 4.27: Transmission coefficients between co-polarized LB and HB dipoles, when the LB antenna adopts reference tube radiators or segmented spiral radiators.

LB antenna uses the reference tube shown in Figure 4.2 as radiators, the isolation is only slightly higher than 14 dB, as shown in Figure 4.27(a). By replacing the reference tube with the segmented spiral, all the isolation are improved to more than 20 dB across the band, as shown in Figure 4.27(b). This improvement is because the segmented spiral structure effectively suppresses the excitation of higher-order modes in the HB so that fewer HB currents are induced on the LB radiator. Therefore, fewer HB currents flow into the LB ports through the balun,

achieving better cross-band isolation in the HB. Similarly, the isolation between cross-polarized LB and HB dipoles is improved from 22 dB to 25 dB by replacing the LB radiator from the reference tube to the segmented spiral, as shown in Figure 4.28.

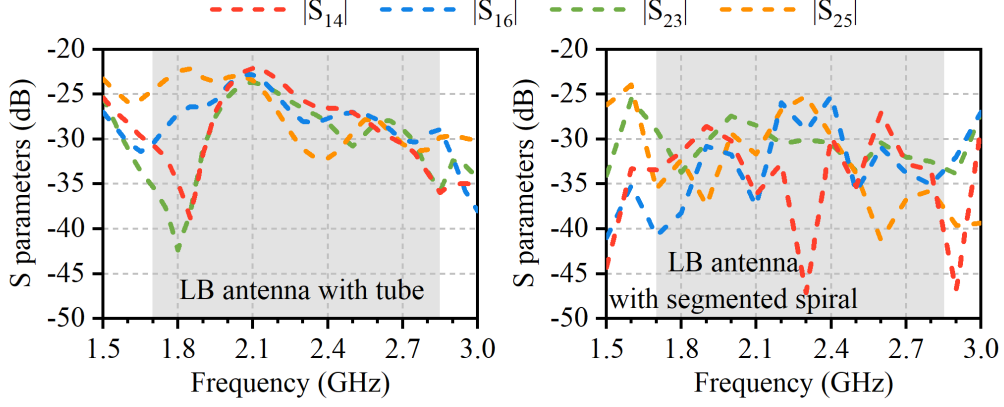


Figure 4.28: Transmission coefficients between cross-polarized LB and HB dipoles, when the LB antenna adopts reference tube radiators or segmented spiral radiators.

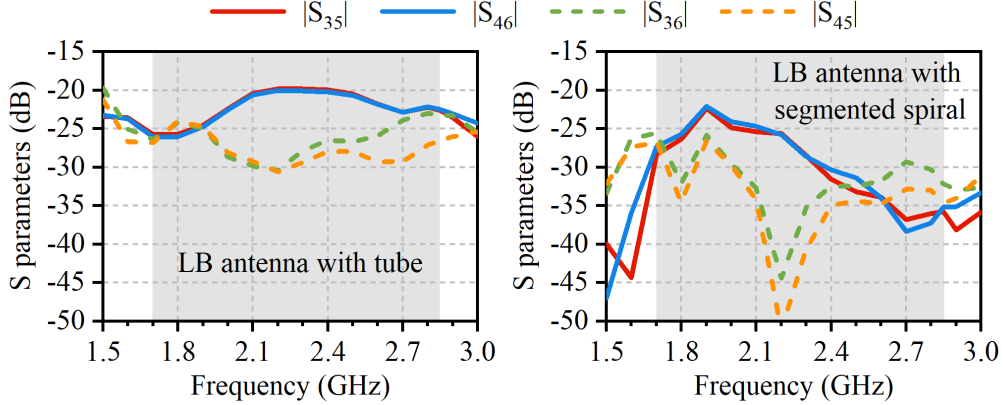


Figure 4.29: Transmission coefficients between co-/cross-polarized HB dipoles, when the LB antenna adopts reference tube radiators or segmented spiral radiators.

In addition to improving cross-band isolation in the HB, the use of the segmented spiral as the LB radiator also contributes to the enhancement of in-band isolation between HB antennas. As shown in Figure 4.29, replacing the reference tube with the segmented spiral as the LB radiator leads to a notable increase in both co-polarized and cross-polarized isolation in the HB. The lowest co-polarized and cross-polarized isolations are increased from 19 dB and 23 dB to 22 dB and 26 dB, respectively. The isolation improvement is more pronounced from 2.1 GHz to 2.85 GHz. The poor isolation when using the reference tube as the LB radiator is because when an HB antenna is excited, HB energy can be coupled to adjacent HB antennas through

the scattering from the tube, resulting in degradation of the in-band isolation. However, the segmented spiral effectively suppresses the HB scattering of the LB radiator in all directions as shown in Figure 4.7, thereby leading to improvement of the in-band isolation.

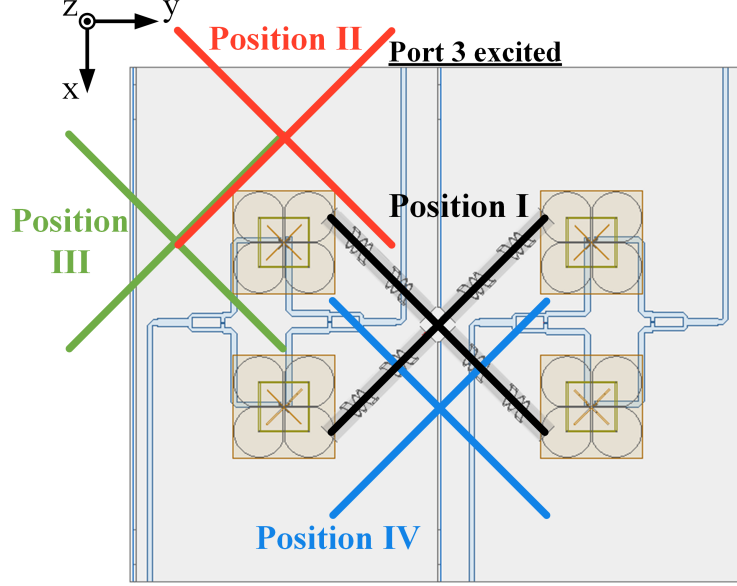


Figure 4.30: Different positions of the designed LB antenna.

The designed dual-band array also exhibits good robustness. Figure 4.31 presents the radiation patterns of the HB antennas when the segmented-spiral-loaded LB antenna is placed in different positions as shown in Figure 4.30. The results obtained remain consistent across different incident angles or positions, demonstrating the segmented spiral possesses excellent generalizability in different polarizations and different relative positions.

In conclusion, the designed segmented spiral, when used as an LB radiator, simultaneously achieves three functions: suppressing cross-band scattering, cross-band coupling, and in-band coupling across a broad HB (1.7-2.85 GHz, 50.5%).

4.5 Experimental Results

The prototype of the dual-band antenna array has been fabricated, as shown in Figure 4.32. Figure 4.33(a) shows the measured reflection coefficients of the LB antenna in the array. The measured reflection coefficients ($|S_{11}|$ and $|S_{22}|$) of dual-polarized LB antenna are all less than -10 dB from 0.69 GHz to 1.00 GHz (36.7%), which agree very well with the simulated ones shown in Figure 4.16(a). As shown in Figure 4.33(b), the measured impedance bandwidth

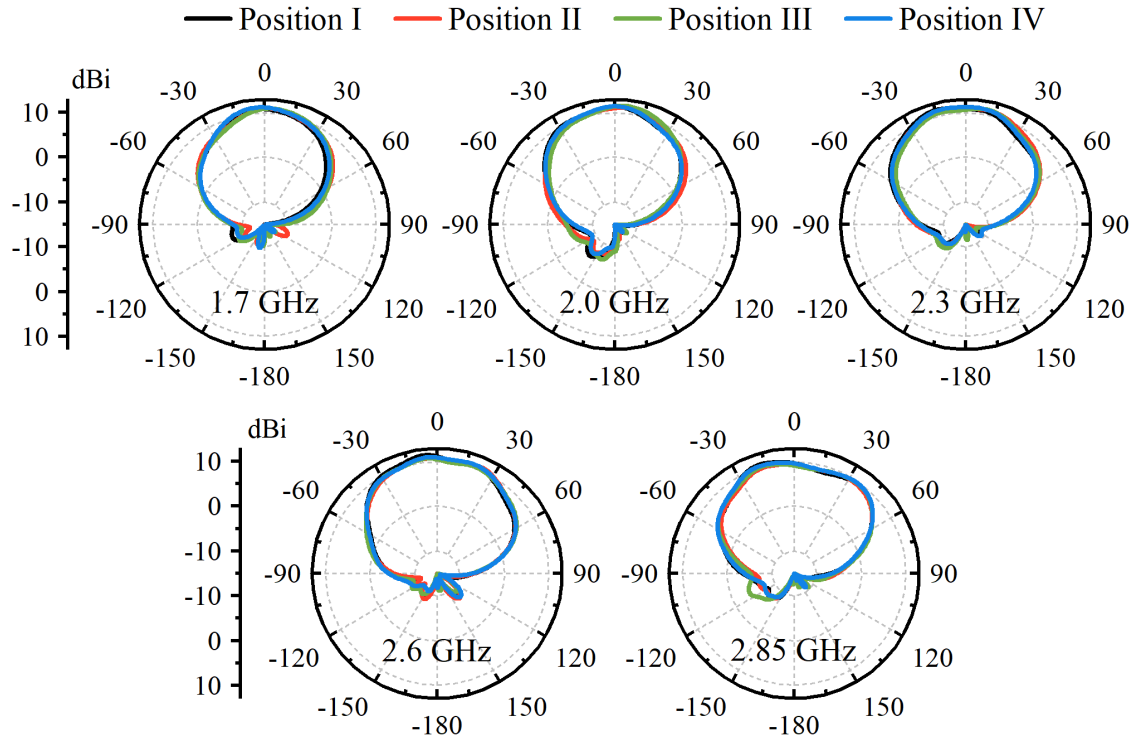


Figure 4.31: Radiation patterns of the HB antennas when the segmented-spiral-loaded LB antenna is placed in different positions (Port 3 in the array is excited).



Figure 4.32: Prototype of the dual-band antenna array.

($|S_{33}|$, $|S_{44}|$, $|S_{55}|$, and $|S_{66}| < -10$ dB) of the HB elements is 53.8% from 1.63 GHz to 2.83 GHz, showing good consistency with the simulated results in Figure 4.22(a).

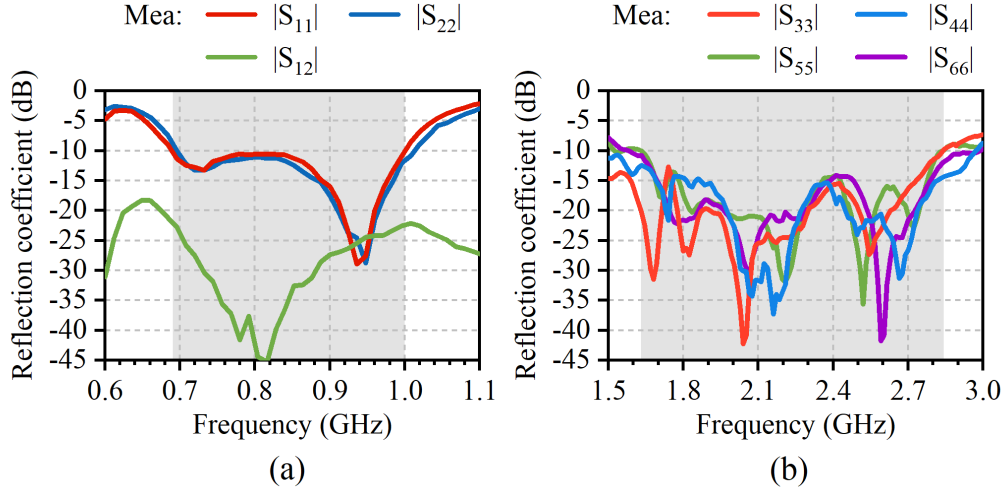


Figure 4.33: Measured reflection coefficients of (a) the LB antenna and (b) the HB antennas.

The measured radiation patterns of the HB and LB antennas in the yoz plane are presented in Figures 4.34 and 4.35, respectively. Due to the symmetry of the array, only the radiation patterns corresponding to the excitation of HB Port 3, HB Port 5, and LB Port 2 are displayed. The measured radiation patterns agree well with the simulated results, and both show undistorted radiation patterns despite the presence of the cross-band antenna elements, demonstrating the effectiveness of the proposed scattering suppression methods. The measured realized gains of the LB and HB antennas in boresight are in the ranges of 7.0-8.3 dBi and 10.7-12.1 dBi, respectively.

The performance of the designed segmented spiral LB antenna in reducing the cross-band coupling and in-band coupling over the HB is shown in Figures 4.36 and 4.37. The measured cross-band isolations between co-/cross-polarized LB and HB antennas are shown in Figure 4.36(a) and (b) across the HB. All the isolations for the co-polarized HB and LB antennas are higher than 20 dB, and all the isolations for the cross-polarized antennas exceed 24 dB. The measured in-band isolations between co-polarized and cross-polarized HB ports are shown in Figure 4.37. The isolations between co-polarization ports are higher than 22 dB, and between cross-polarization ports exceed 26 dB throughout the HB. All the measured results regarding isolations demonstrate the effectiveness of the segmented spiral structure in reducing cross-band coupling and in-band coupling in the HB. Along with its capability to suppress cross-band scattering, the segmented spiral structure offers a tri-functional performance that enhances the overall performance of the antenna array.

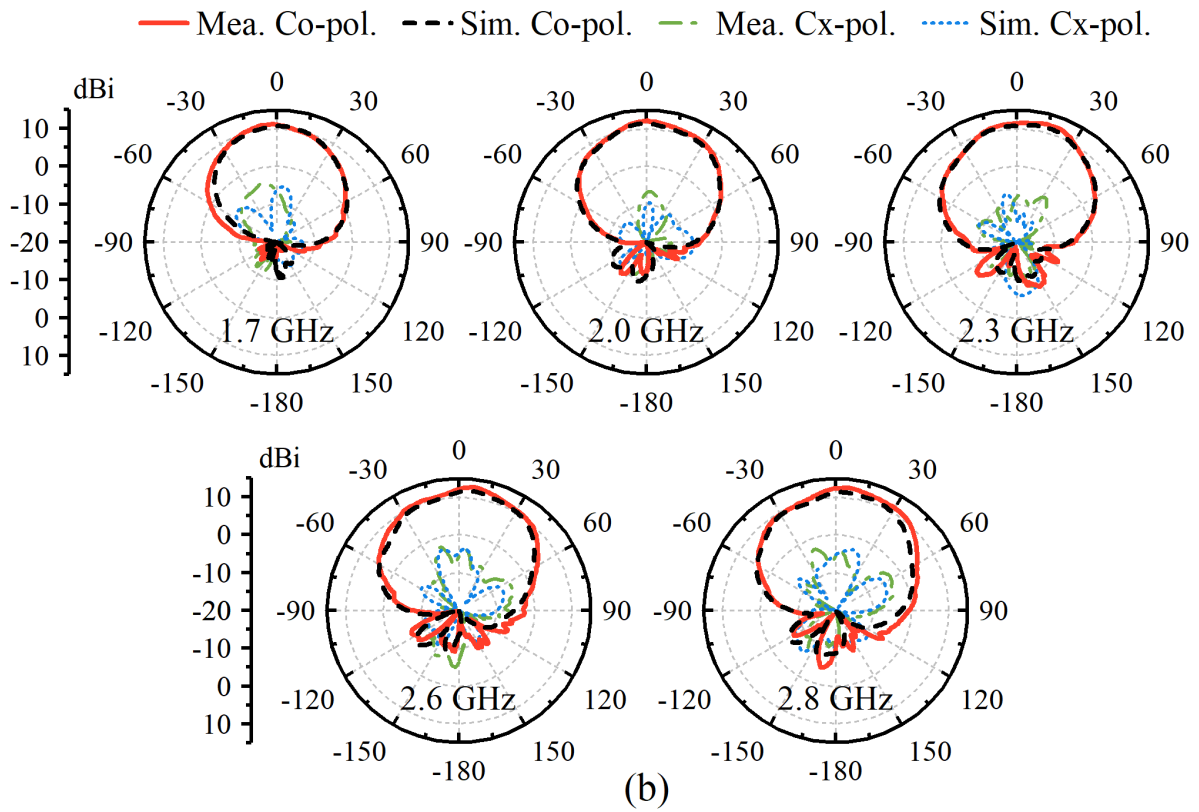
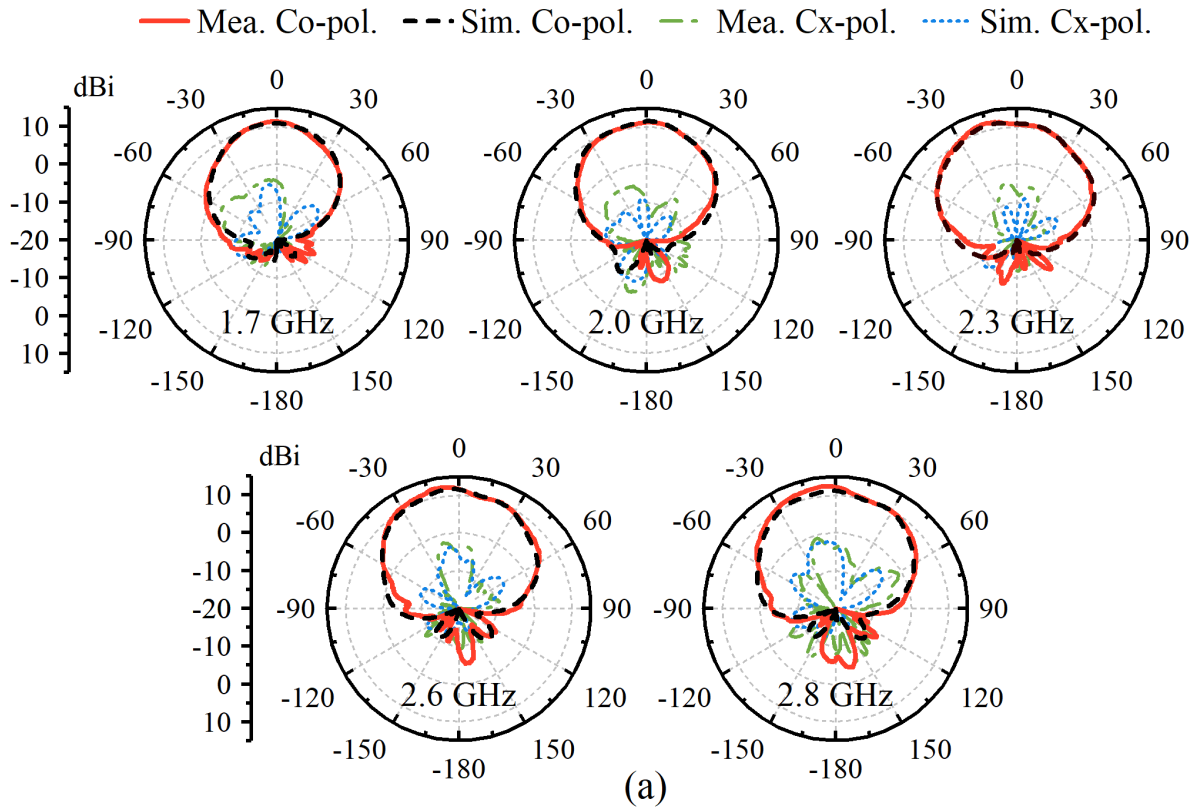


Figure 4.34: Measured and simulated radiation patterns of the HB antennas in the $yo z$ plane when (a) Port 3 and (b) Port 5 are excited.

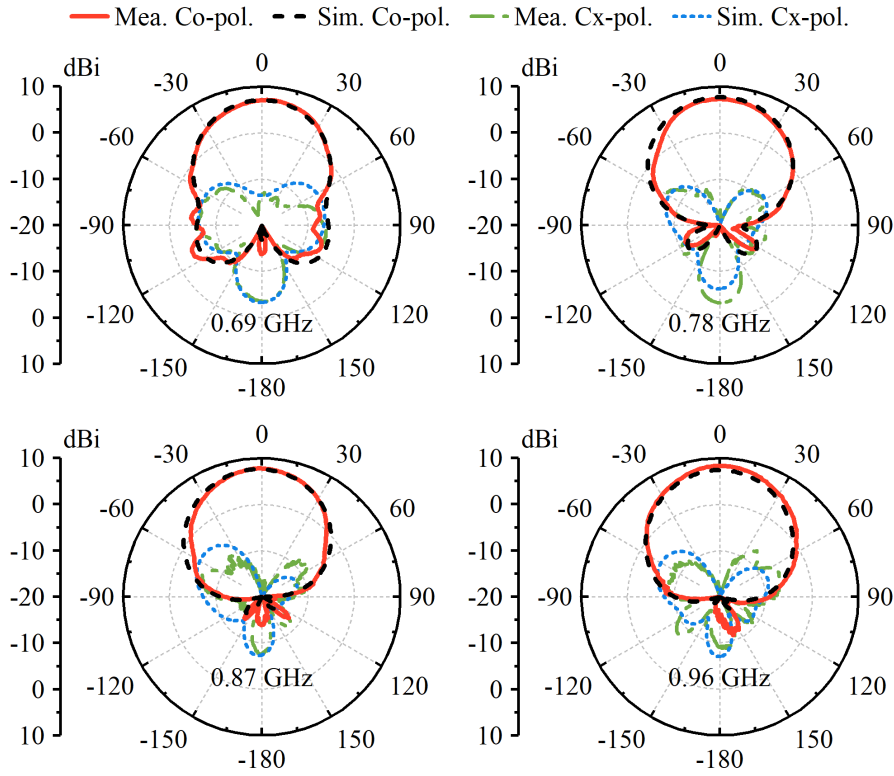


Figure 4.35: Measured and simulated radiation patterns of the LB antenna in the yoz plane, when Port 2 is excited.

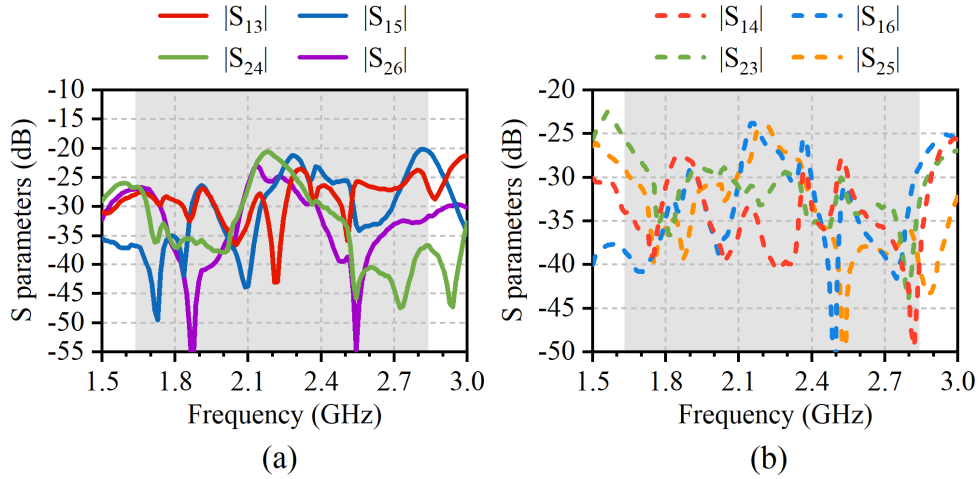


Figure 4.36: Measured transmission coefficients between (a) co-polarized LB and HB antennas, and (b) cross-polarized LB and HB antennas.

To demonstrate the advantages of techniques developed in this work, their performance is compared with that of other techniques for scattering and coupling suppression, as listed in Table 4.1. This work achieves the tri-functional suppression in HB, featuring the widest HB

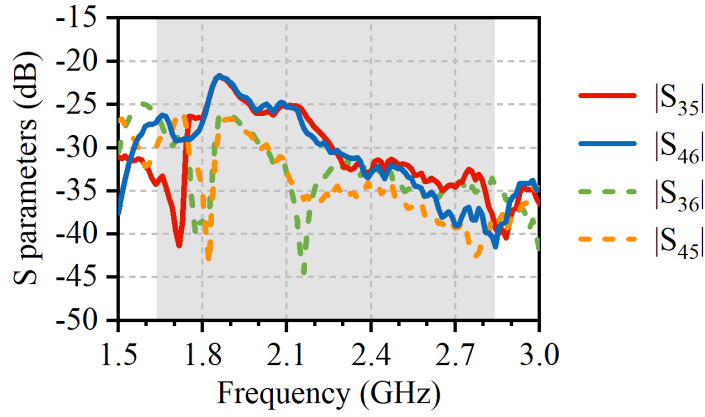


Figure 4.37: Measured transmission coefficients between co-polarized and cross-polarized HB antennas.

bandwidth of 53.8%, while concurrently maintaining LB matching across a wide band of 36.7%.

Table 4.1: Comparison of Multi-Functional Techniques for Scattering and Coupling Suppression in Interleaved Dual-Band Antenna Arrays

Ref.	Proposed structures	Suppression function			Scattering-suppression /Operating band (GHz)	Cross-band isolation (dB)	In-band isolation (dB)
		Cross-band scattering	Cross-band coupling	In-band coupling			
[73]	LB: 2.5-D cloak + DGS HB: Serial resonance shifter	✓	✓	✓	LB: 1.64-2.56 (43.8%) HB: 4.4-5.0 (12.8%)	> 25 (HB) > 23 (LB)	> 23 (HB)
[38]	LB: Spiral	✓			LB: 1.71-2.26 (27.7%) HB: 3.3-3.7 (11.4%)	N/A	> 25 (HB)
[47]	LB: Double-arm + Chokes HB: Capacitor	✓			LB: 0.698-0.96 (31.6%) HB: 1.7-2.2 (25.6%)	N/A	N/A
[32]	LB: Folded-dipole + Loops HB: Two loops + Two strips	✓	✓	✓	LB: 2.26-2.73 (19%) HB: 3.0-4.3 (36%)	> 25 (HB) > 25 (LB)	> 22 (HB)
[30]	LB: U-shape structure HB: Serial LC structure	✓	✓		LB: 0.69-0.96 (32.7%) HB: 1.7-2.7 (45.5%)	> 20 (HB)	N/A
[75]	LB: Partially reflective surface	✓		✓	LB: 1.7-2.2 (25.6%) HB: 3.3-3.6 (8.7%)	N/A	> 17.5 (HB)
[74]	LB: Radiator with stubs	✓		✓	LB: 1.7-2.2 (25.6%) HB: 3.3-3.6 (8.7%)	N/A	> 24.6 (HB)
This work	LB: Segmented spiral HB: Parallel resonance shifter	✓	✓	✓	LB: 0.69-1.00 (36.7%) HB: 1.63-2.83 (53.8%)	> 20 (HB) > 20 (LB)	> 22 (HB)

It is important to highlight that enhancing the operating bandwidth of dual-band array presents a particularly challenging task due to the need to simultaneously meet four constraints as

illustrated in Figure 4.38: 1) the LB antenna must exhibit good suppression characteristics within the selected HB; 2) The radiation performance and matching bandwidth of the developed LB antenna are not compromised; 3) the HB antenna should not adversely affect the radiation of the LB antenna; and 4) the HB antennas must be well matched over a wide bandwidth. The segmented spiral, coupled with the serial resonator and the parallel resonance shifter, addresses all the challenges and offers a unified pathway for the enhancement of the operating bandwidth.

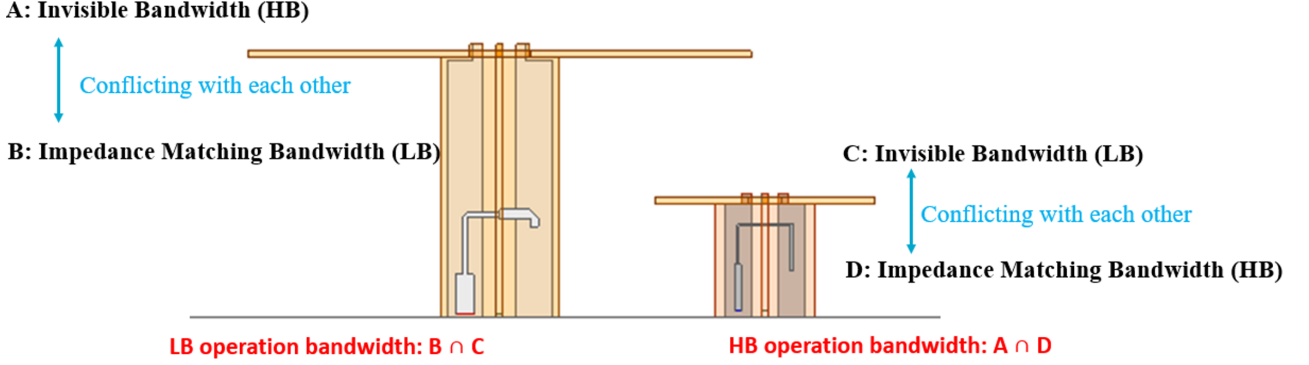


Figure 4.38: Illustration of the four conflicting bandwidths in the collocated dual-band antenna array.

Moreover, the segmented spiral structure in this work demonstrates remarkable RCS reduction over an ultra-wide bandwidth of 118% from 1.10 to 4.29 GHz. In this specific study, the HB antennas were chosen to operate from 1.63 to 2.83 GHz as a demonstration. However, it is important to note that the LB antenna is also “electromagnetically invisible” to antennas operating in another band, such as from 3.3 to 3.8 GHz for 5G applications.

4.6 Conclusion

To simultaneously suppress the cross-band scattering, cross-band coupling, and in-band coupling across wide bands in dual-band arrays, a tri-functional segmented spiral is developed as the radiator under the precise guidance of CMA. The segmented spiral effectively mitigates the higher-order modes responsible for the scattering and coupling, achieving a wide suppression bandwidth. The serial resonators are added between the radiator and balun to widen the matching bandwidth of the antenna. The antenna is successfully matched from 0.69 GHz to 1.00 GHz (36.7%). The designed antenna as an LB antenna is incorporated in a dual-band dual-polarized antenna array to verify its suppression performance in the HB. For the HB antennas in the array, the parallel resonance shifters are co-designed with the HB balun to

suppress LB scattering caused by the HB antennas. Both simulated and measured results of the array demonstrate that the radiation patterns of HB and LB antennas in the array are in good consistency with those obtained when they radiate alone. Moreover, the cross-band coupling and in-band coupling across the HB are all reduced compared to the LB antenna with a conventional tube radiator. The exceptional tri-functional performance of the segmented spiral is attributed to the effectiveness of the CMA-guided design approach. Compared to the dual-band arrays presented in other works, the array in this work shows the widest HB suppression bandwidth, across which various scattering and coupling are holistically suppressed.

Chapter 5

Wide-Band Suppression of Scattering and Coupling in Tri-Band Antenna Arrays

5.1 Introduction

The scattering and coupling issues in a tri-band shared-aperture antenna array are more complex than those in a dual-band antenna array, since scattering and coupling occur between any two of the three antennas. Moreover, enabling all three antennas to operate over wide bands is even more challenging.

In Section 5.2, based on the method proposed in Section 4.2, an LB radiator capable of achieving scattering suppression over a wide bandwidth is proposed. The suppression band covers the target MB and HB (6.425–15.35 GHz) of the tri-band antenna array. In addition, the MB and HB antennas employ a planar ME dipole structure, which mitigates some of the scattering and coupling issues inherent in the tri-band antenna arrays. However, such a planar structure is applicable only to the MB and HB antennas operating at higher frequency bands, and the operating bands of the MB and HB antennas are adjacent to each other.

In Section 5.3, more universal approaches for addressing scattering and coupling in tri-band shared-aperture antenna arrays are proposed. Both the MB and HB antennas employ a conventional three-dimensional dipole configuration, and their operating bands are separated by

a distinct frequency gap rather than being adjacent. Various scattering and coupling issues in the tri-band antenna array are effectively resolved using multiple techniques.

5.2 A Tri-Band Shared-Aperture Base Station Antenna Array Covering 5G Mid-Band and 6G Centimetric Wave Band

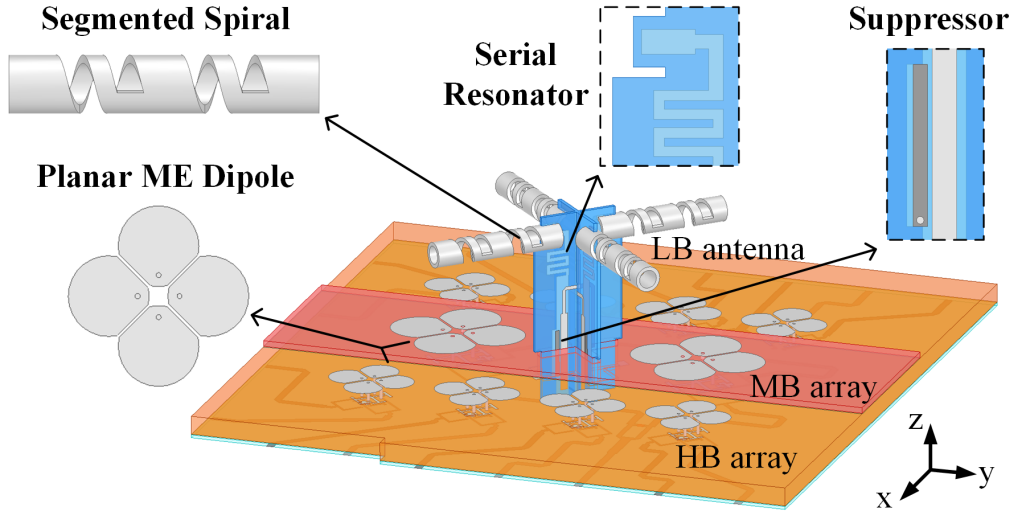


Figure 5.1: Configuration of the developed tri-band array.

In this section, as shown in Figure 5.1, a tri-band 5G/6G antenna array is developed with operating bandwidths in the LB, MB, and HB ranges of 3.05-4.68 GHz (42.2%), 6.2-10.0 GHz (46.9%), and 10.0-15.6 GHz (43.8%), respectively. The radiation pattern distortions caused by scattering within the LB, MB, and HB have been effectively mitigated or avoided, and all cross-band and in-band coupling has been suppressed to below -20 dB.

The novelty and contributions are outlined as follows:

- (i) **Ultra-Wide-Range in All Three Bands, Enabling Full Coverage for 5G and 6G:** The proposed tri-band shared-aperture antenna array features the widest bandwidths among existing designs. The LB antenna supports the 5G sub-6 GHz band 3.3-4.2 GHz, while the MB and HB antennas collectively cover the 5G-Advanced and 6G range of 6.425-15.35 GHz, ensuring seamless multi-generation connectivity.
- (ii) **Electromagnetic Transparency via Characteristic Mode Analysis (CMA):** To

prevent the LB radiator from distorting the radiation of the MB and HB antennas, the LB segmented spiral radiators are strategically designed using CMA to exhibit electromagnetic transparency across both the MB and HB. This design achieves a radar cross-section (RCS) reduction across 4.7–21.5 GHz (128.2%), demonstrating ultra-wideband scattering suppression.

(iii) **Impedance Compensation for Matching:** To address the degradation in LB impedance matching caused by the scattering suppression across an ultra-wide band, a pair of serial resonators within the LB balun is designed. These resonators effectively compensate for the impedance mismatch, achieving a matching bandwidth of 42.2%. Additionally, suppressors are placed near the LB ports to further reduce cross-band coupling between the LB and MB antennas.

(iv) **Low-Profile Planar ME Dipole Design:** Both the MB and HB antennas adopt a planar magnetoelectric (ME) dipole structure. This low-profile design, simplifying the assembly process, prevents common-mode resonances in both the LB and MB, and reduces the scattering from the MB antenna in the HB.

This section is organized as follows. Subsection 5.2.1 and Subsection 5.2.2 detail the design of the LB and MB/HB antennas, respectively. The scattering and coupling suppression results in the tri-band array are provided in Subsection 5.2.3. Subsection 5.2.4 presents the experimental results of the fabricated antenna array. Finally, the conclusions are summarized in Subsection 5.2.5.

5.2.1 LB Antenna Design

6G has now entered the research stage, with expectations of introducing several frequency bands within 7.125-15.35 GHz to achieve a balance between capacity and coverage [117], [118], [119]. Moreover, the band 6.425-7.125 GHz has been allocated mainly for the 5G-Advanced and may serve as a foundational band for 6G in certain countries [118]. The higher, wider, and more numerous operating bands for 5G and 6G pose new challenges for the design of antenna elements and multi-band arrays. For instance, if antennas continue to adopt the commonly used 3D dipole, the significantly reduced size will introduce difficulties in fabrication and assembly. More importantly, the use of additional and wider bands further complicates the interference problems within the antenna array.

The objective of this work is to design a multi-band shared-aperture antenna array for 5G and

6G base stations. The LB antenna is designed to cover the most commonly used 5G band of 3.3–4.2 GHz. As the frequency allocations for 6G have not yet been finalized, the MB and HB antennas in this work are designed to collectively cover the entire candidate 6G bands as well as the 5G-Advanced band of 6.425–15.35 GHz. This band exhibits a relative bandwidth of up to 82%, which makes it extremely challenging for an LB antenna to achieve effective scattering and coupling suppression across such a wide frequency range. Nevertheless, the segmented spiral, designed under the guidance of CMA and presented in Chapter 4 of this thesis, demonstrates great potential for realizing ultra-wideband suppression.

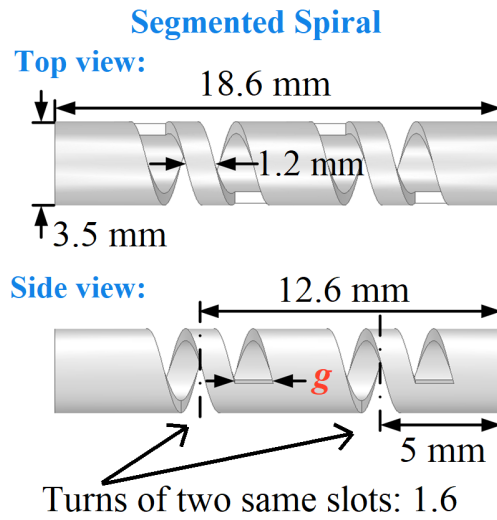


Figure 5.2: Geometry of the segmented spiral.

Following the CMA method described in Chapter 4, Section 4.2.1, a new segmented spiral is designed as shown in Figure 5.2, whose scattering-suppression bandwidth covers 6.425–15.35 GHz, while its operating band as an LB radiator also spans 3.3–4.2 GHz.

Monostatic radar cross sections (RCSs) of the segmented spiral and the reference tube are compared in Figure 5.3. The segmented spiral effectively reduces the RCS across a wide band of 4.7–21.5 GHz (128.2%), with a maximum RCS reduction of 20 dB. It covers the target suppression band of 6.425–15.35 GHz.

When the designed segmented spiral is employed as the radiator arm of the LB dipole, a port is inserted between the two arms to examine the input impedance of the dipole antenna, as shown in Figure 5.4. Within the LB target band of 3.3–4.2 GHz, the variation of the segmented spiral is significantly larger than that of the reference tube, imposing considerable challenges on the impedance matching of the LB antenna.

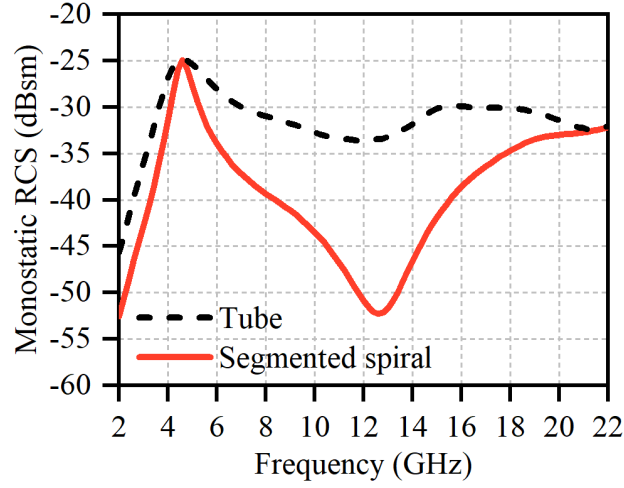


Figure 5.3: Monostatic RCSs of the segmented spiral and the tube.

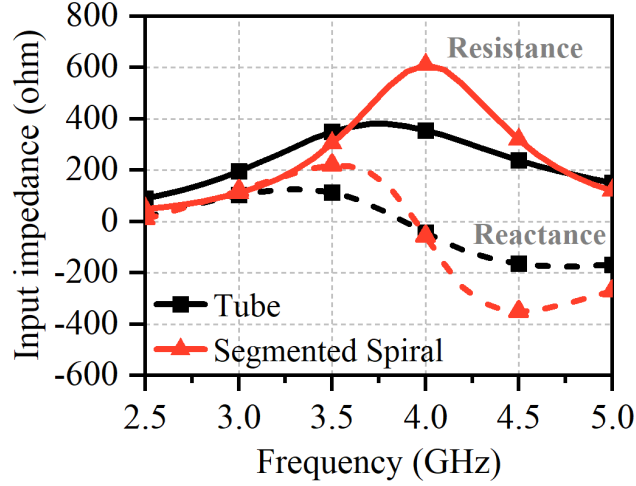


Figure 5.4: Input impedance of the LB dipole using the segmented spiral or the tube.

Similarly, the method described in Chapter 4, Section 4.2.2 is employed to reduce the fluctuation of the input impedance curves within the target band, thereby enabling the LB antenna to achieve impedance matching over a wider frequency band. As shown in Figure 5.5, the serial resonators are introduced between the LB segmented-spiral-based radiators and the conventional balun feeding structures. The serial resonators, balun, and feeding lines are printed on both sides of a substrate with a relative permittivity of 2.2 and a thickness of 0.508 mm. A metal pad is inserted into the spiral radiator at the feeding position, forming a capacitor with the metal structure of the radiator. Different from the complex inductive structure presented in Chapter 4, Section 4.2.2, the inductive element is simplified into a meander line. This configuration facilitates straightforward tuning of the required inductance, with the line length directly determining the inductance value.

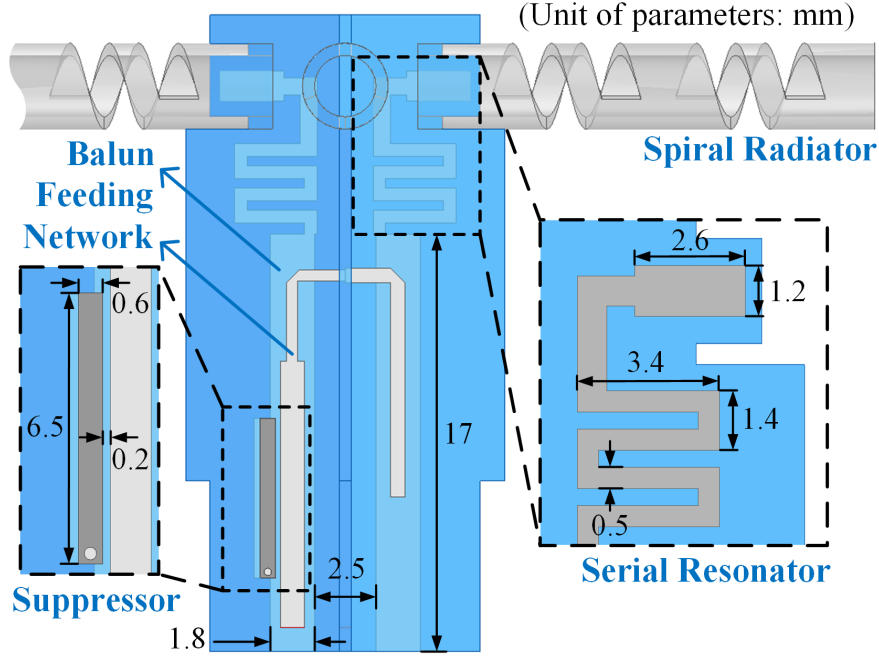


Figure 5.5: Geometry of the LB segmented spiral antenna with integrated serial resonators and suppressors in the balun.

In addition, the suppressor composed of a strip, vias, and ground is placed adjacent to the balun, near the LB port, and is used to enhance the cross-band isolation between the LB and MB antennas within the MB band. This structure essentially functions as a filtering element, which is commonly employed in the design of filtering antennas. Nevertheless, its operating bandwidth is relatively narrow. In this work, it serves as an auxiliary element to suppress cross-band coupling over a localized frequency band, keeping the cross-band isolation between the LB and MB antennas above 20 dB across the MB.

The S parameters of the matched LB spiral antenna are presented in Figure 5.6, and ports 1 and 2 represent the two ports of the LB cross-dipole antenna shown in Figure 5.5. The reflection coefficients ($|S_{11}|$ and $|S_{22}|$) are less than -10 dB in 3.05-4.51 GHz, and the polarization isolation $|S_{12}|$ exceeds 20 dB within this range. The wideband matching performance validates the efficacy of the serial-resonator-integrated balun design. Moreover, the introduction of the suppressors does not adversely affect the impedance matching of the proposed LB cross-dipole antenna.

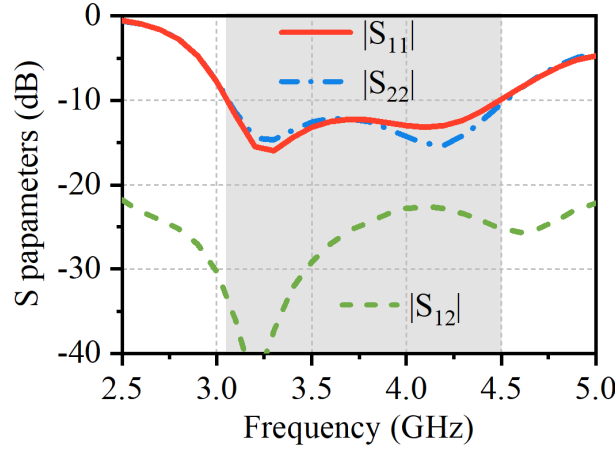


Figure 5.6: Reflection coefficients ($|S_{11}|$ and $|S_{22}|$), and transmission coefficient $|S_{12}|$ of the LB segmented spiral antenna.

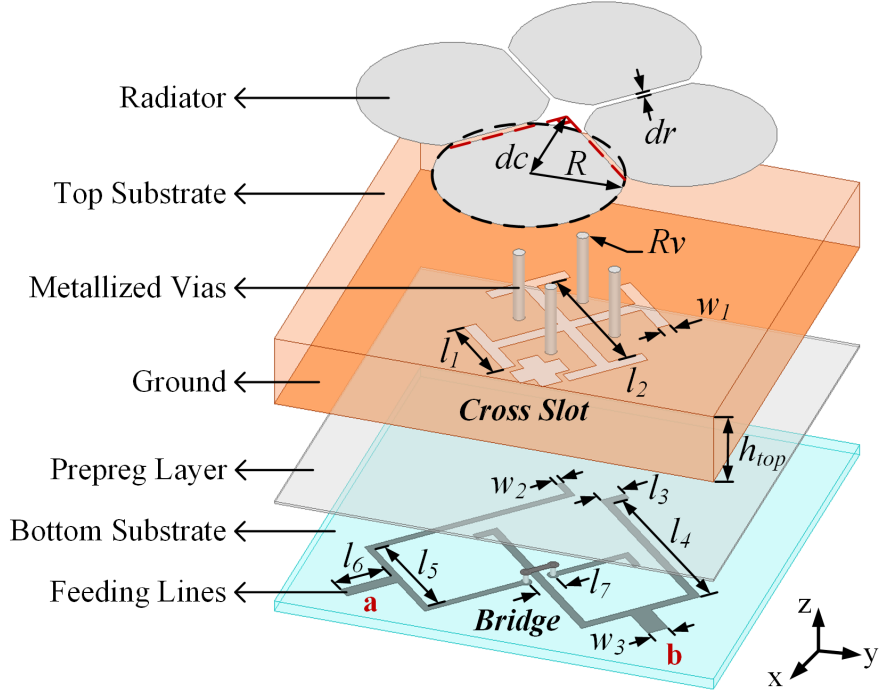


Figure 5.7: Geometry of the MB and HB planar ME dipole antennas.

5.2.2 MB and HB Antenna Design

The anticipated band for 5G-Advanced and 6G is 6.425–15.35 GHz. The required bandwidth is very wide, and the frequency is significantly higher than those of 3G, 4G and 5G (sub-6 GHz). If a conventional cross-dipole is still used, the designed antenna size will be extremely small, posing significant challenges for fabrication and assembly. To achieve the required wide operating bandwidth, the planar magnetoelectric (ME) dipole structure [120], [121] is a strong

Table 5.1: Geometrical Parameters of the MB and HB Antennas

	R	dc	dr	Rv	h_{top}	l_1
MB	4.2	5	0.35	0.25	3.43	3.8
HB	2.65	3	0.37	0.2	2.54	2.4
	l_2	l_3	l_{4a}	l_{4b}	l_{5a}	l_{6a}
MB	6.9	1.2	9.7	8.5	5	2.5
HB	4.4	0.75	6	6	3.2	1.6
	l_{6b}	l_7	w_1	w_2	w_{3a}	w_{3b}
MB	1.6	1.3	0.6	0.35	0.6	1
HB	1.6	1.2	0.4	0.3	0.6	0.6

candidate for this multi-band array design.

The designed planar ME antenna structure, serving as the MB/HB elements in the tri-band array, is shown in Figure 5.7. The four-leaf-clover-shaped radiator is connected to the ground through four vias, while the $\pm 45^\circ$ oriented fork-shaped microstrip lines excite the antenna through the Jerusalem-cross slots in the ground. To mitigate and avoid the coupling between the two polarizations of the antenna, the slot reserved for the bridge on the ground is designed in a cross shape, and the microstrip lines in the two directions have different lengths. The top and bottom substrate layers use materials with a thickness of h_{top} (relative permittivity of 2.2) and 0.406 mm (relative permittivity of 3.55), respectively. A prepreg layer with a thickness of 0.102 mm is inserted between them during the lamination process. The other parameter values of the MB and HB antennas are provided in Table 5.1. Note that they share a similar structure but have different dimensions.

In a tri-band antenna array, employing planar ME dipoles as the MB and HB antenna elements can avoid common-mode resonance [47] within the LB and MB, due to their significantly lower profile compared to traditional 3D dipole antennas. Moreover, the MB and HB radiators are positioned at almost identical heights, effectively mitigating the cross-band scattering [85]–[88] from the MB radiators in the HB.

According to the simulation results shown in Figure 5.8, the proposed MB and HB planar ME

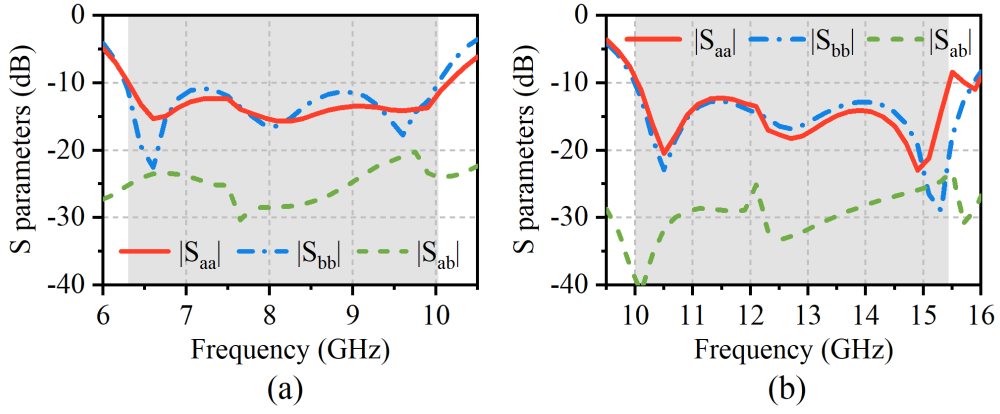


Figure 5.8: Reflection coefficients ($|S_{aa}|$ and $|S_{bb}|$), and transmission coefficient $|S_{ab}|$ of the (a) MB antenna and (b) HB antenna.

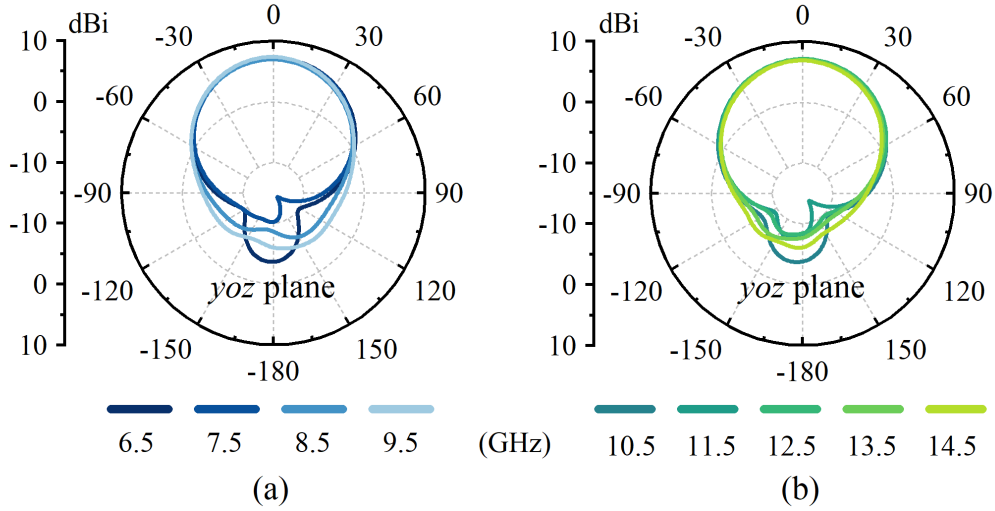


Figure 5.9: Radiation patterns of the (a) MB antenna and (b) HB antenna.

dipoles work in the bandwidths of 6.3-10.0 GHz and 10.0-15.45 GHz, respectively, covering the 5G-Advanced and 6G range of 6.425–15.35 GHz. Their polarization isolations $|S_{ab}|$ exceed 20 dB across the operating bands. Moreover, the radiation patterns of the MB and HB antennas are shown in Figure 5.9, which are stable and exhibit good consistency over their respective bands.

5.2.3 Holistic Suppression of Scattering and Coupling in the Tri-Band Array

As illustrated in Figure 5.10, the proposed LB, MB, and HB antennas are compactly arranged in an interleaved tri-band array to validate the scattering and coupling suppression performance

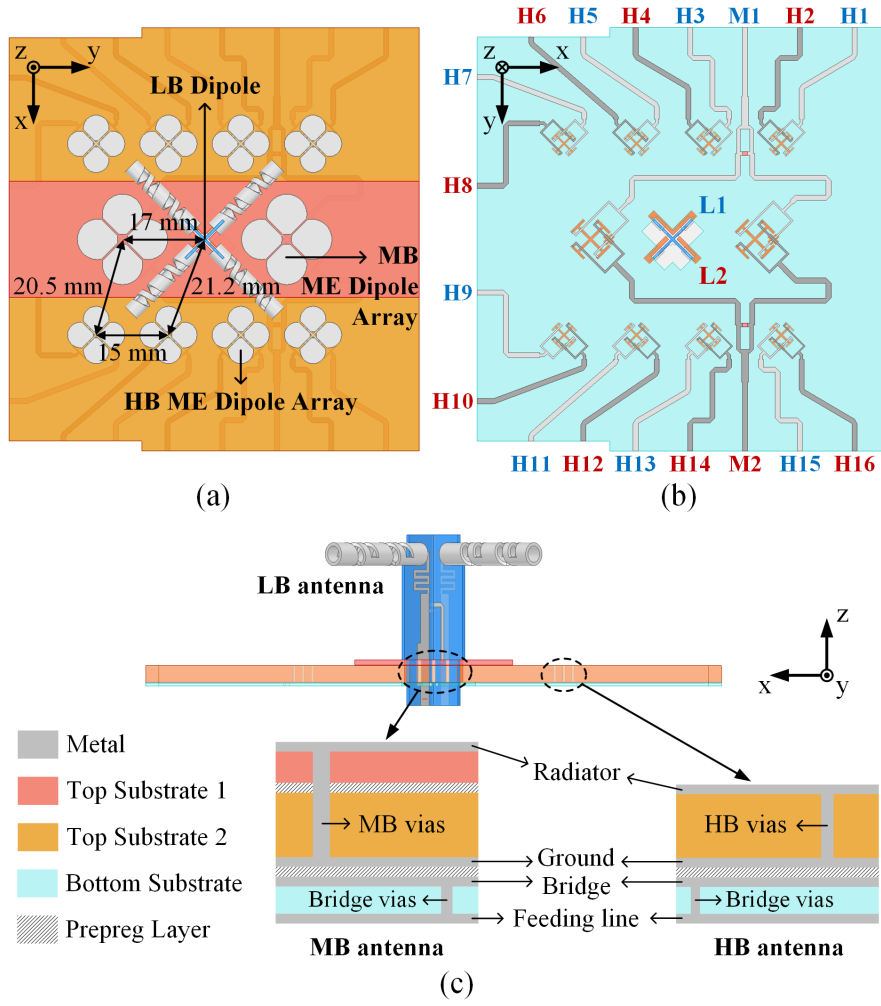


Figure 5.10: (a) Top view, (b) bottom view, and (c) side view of the tri-band antenna array.

of the proposed methods and structures. Although the MB and HB antennas have identical bottom substrate thicknesses, their top substrate thicknesses differ. Therefore, the MB and HB antenna arrays are designed as a planar structure composed of three laminated dielectric layers, with a slot etched in the central region to accommodate the proposed LB antenna. The minimum center-to-center spacings between the LB and MB elements, LB and HB elements, and MB and HB elements are 17 mm ($0.22 \lambda_{LB}$), 21.2 mm ($0.27 \lambda_{LB}$), and 20.5 mm ($0.55 \lambda_{MB}$), respectively.

The S-parameters, including the reflection and transmission coefficients of the LB, MB, and HB antennas in the tri-band array, are shown in Figure 5.11. The reflection coefficients ($|S_{11}|$ and $|S_{22}|$) of the LB, MB, and HB antennas remain below -10 dB within 3.05–4.53 GHz, 6.2–10.0 GHz, and 10.0–15.6 GHz, respectively. In addition, the transmission coefficients ($|S_{12}|$) are lower than -20 dB across the corresponding frequency bands. These results are consistent with the

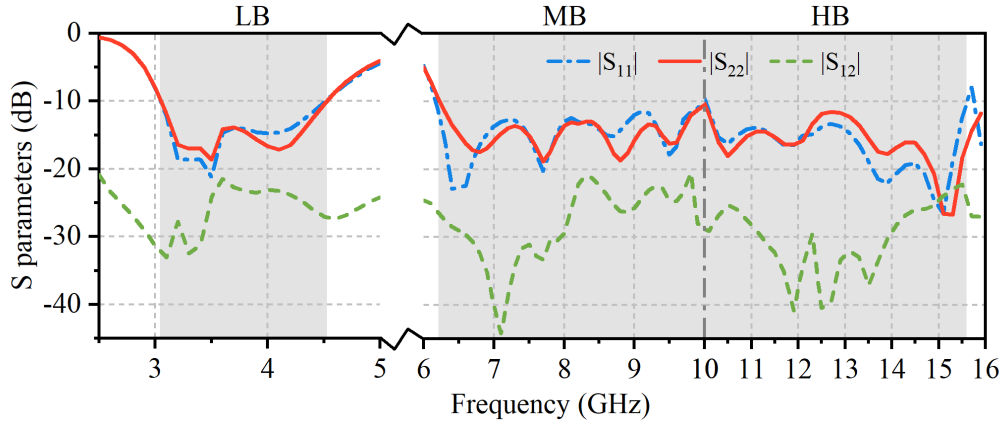


Figure 5.11: S parameters of the LB, MB, and HB antennas in the tri-band array.

S-parameters obtained when the antennas operate individually, indicating that the impedance matching and polarization isolation characteristics of the LB, MB, and HB antennas are not affected by the presence of other closely spaced antennas in the proposed shared-aperture tri-band array.

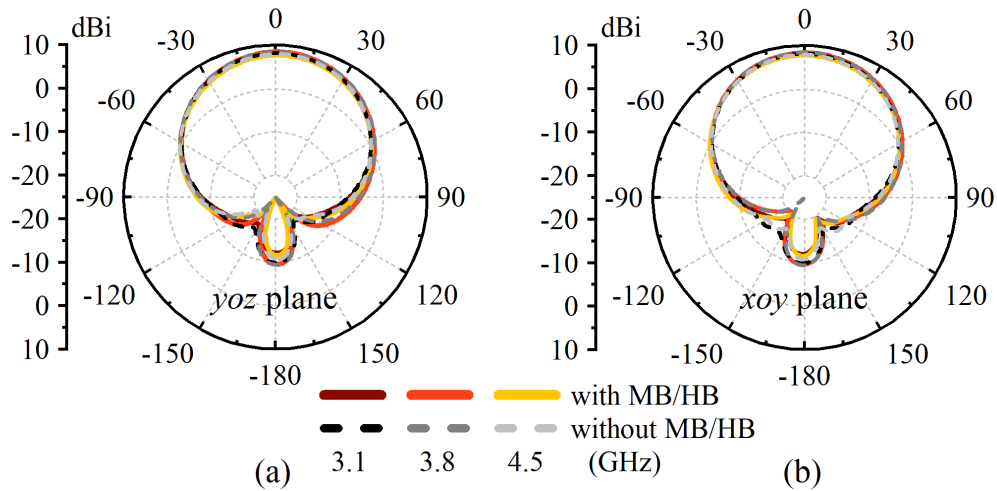


Figure 5.12: Simulated radiation patterns of the LB antenna in (a) yoz plane and (b) xoz plane, when port L1 is excited.

Figure 5.12 compares the simulated radiation patterns of the LB antenna in the presence and absence of the MB and HB antennas. Owing to the proposed planar ME dipole structure, the common-mode resonance that could be induced by the MB and HB antennas is effectively avoided within the LB. Consequently, the LB antenna, surrounded by the MB and HB antennas, maintains undistorted radiation patterns over the entire band, in good agreement with those of the standalone LB antenna.

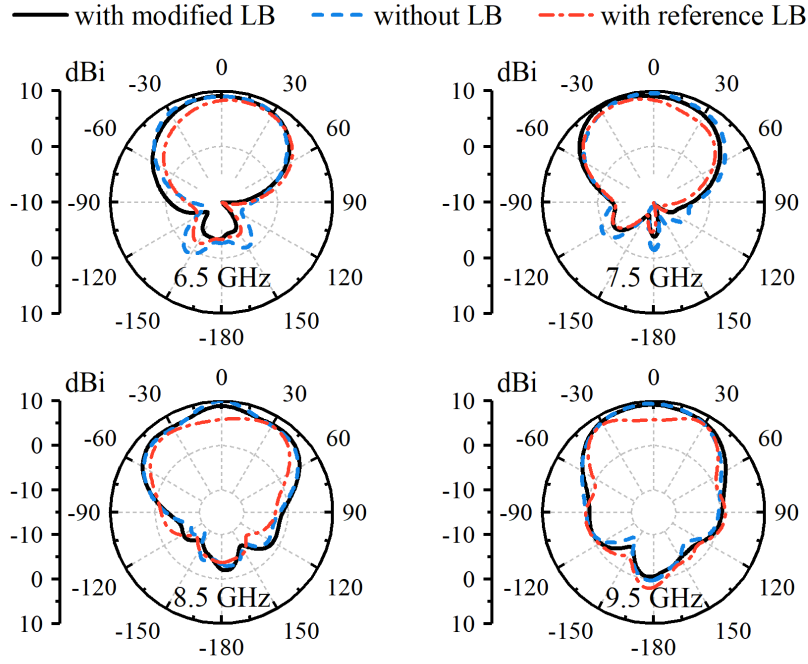


Figure 5.13: Simulated radiation patterns of the MB antennas with and without the presence of different LB antennas, when port M1 is excited.

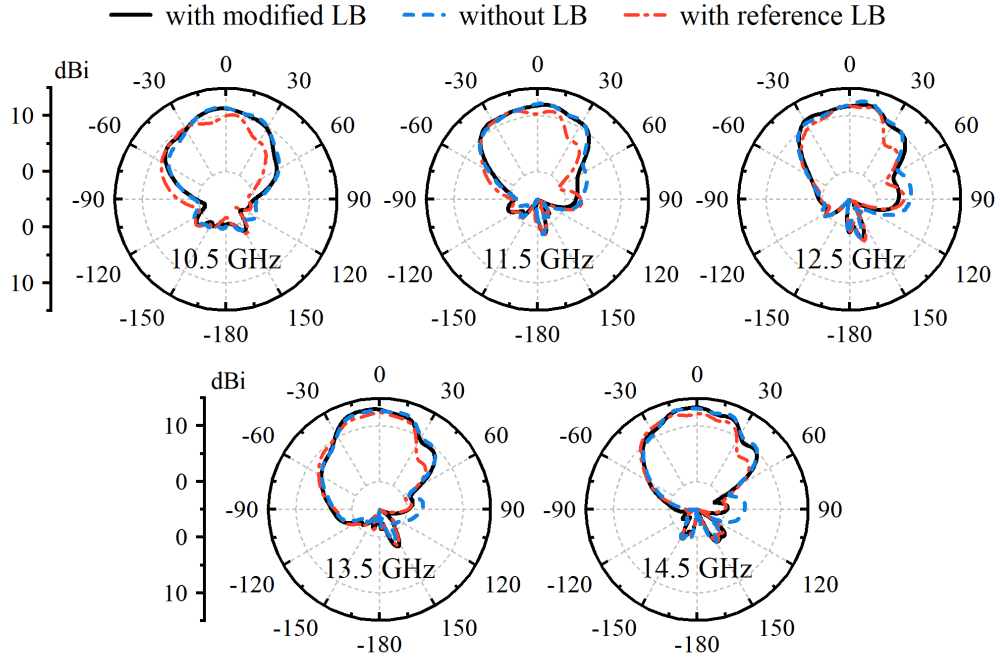


Figure 5.14: Simulated radiation patterns of the HB antennas with and without the presence of different LB antennas, when ports H1-4 are excited.

The MB and HB radiation patterns in the xoz plane under different cases are compared in Figure 5.13 and Figure 5.14, respectively. In the reference case, the conventional tube-based

LB antenna significantly distorts the MB/HB radiation patterns and degrades the realized gain. In contrast, employing the modified LB antenna with MB/HB-transparent segmented spirals results in minimal impact on the MB/HB radiation patterns and realized gain, which remain closely aligned with those of the MB/HB antennas operating independently, without the LB antenna.

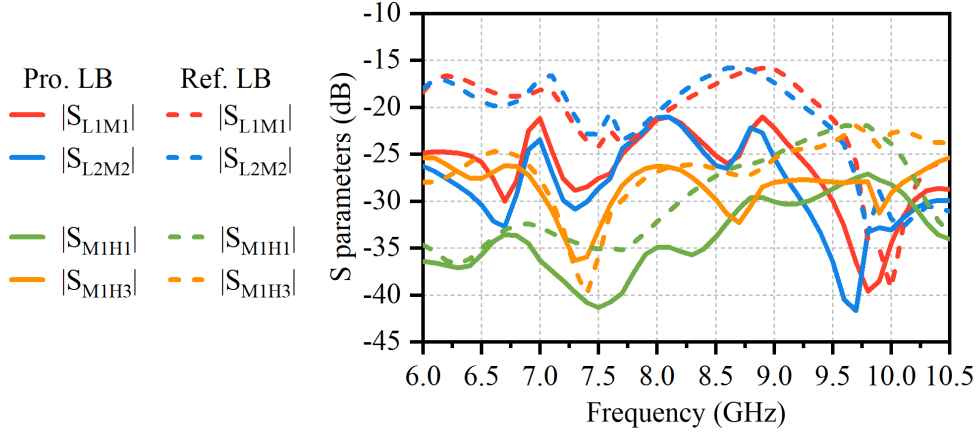


Figure 5.15: Transmission coefficients in the MB between the LB, MB, and HB antennas.

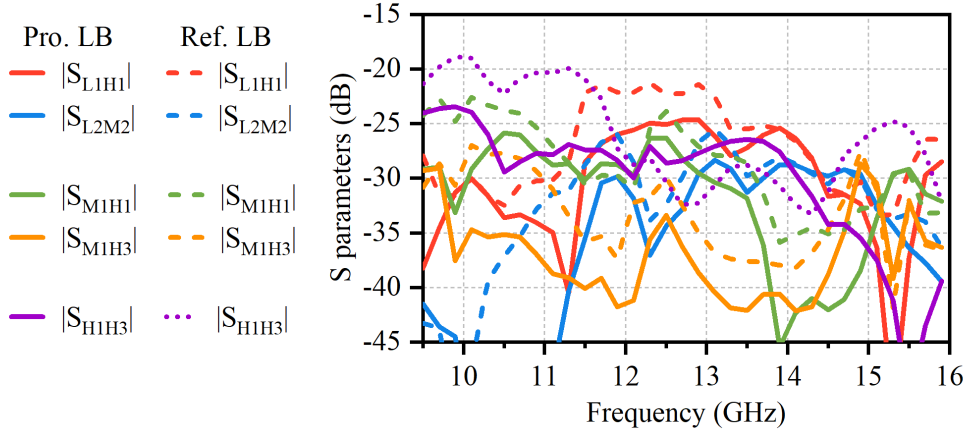


Figure 5.16: Transmission coefficients in the HB between the LB, MB, and HB antennas.

In addition to effectively suppressing cross-band scattering and restoring radiation patterns, the proposed suppression structures are also capable of mitigating coupling, thereby enhancing the isolation in the tri-band antenna array. Figure 5.15 and Figure 5.16 show representative isolations between the LB, MB, and HB antennas in the tri-band array employing either the proposed LB antenna or the reference tube-radiator LB antenna. The proposed LB antenna achieves higher isolations in both the MB and HB compared to the reference case. The MB/HB-transparent segmented-spiral-based LB radiator induces and re-radiates less MB/HB energy

than the reference tube radiator, resulting in reduced induced currents flowing to the LB ports and weaker coupling to the other MB/HB antennas. In addition, the suppressor in the LB balun further blocks the MB-induced currents from entering the LB ports. As a result, the cross-band coupling in the MB is reduced to below -20 dB, while both the cross-band and in-band coupling in the HB are suppressed to around -25 dB.

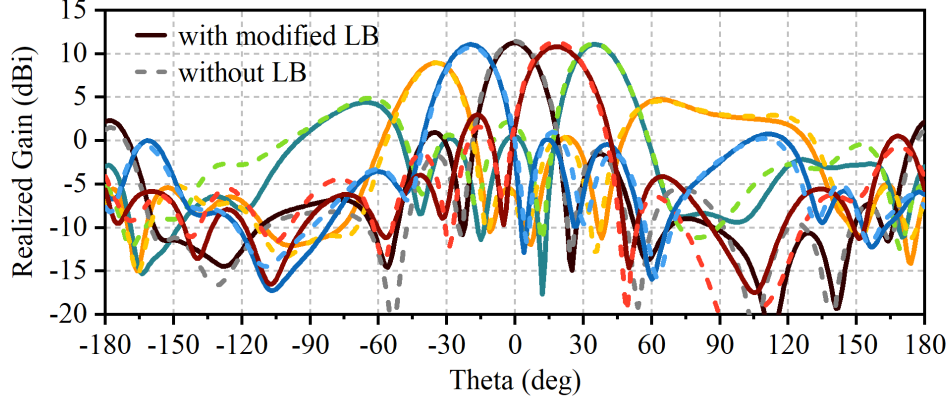


Figure 5.17: Radiation patterns of the HB antenna array scanned in the yo z -plane at 12.65 GHz.

To examine whether the radiation beams of the HB antennas are affected during angle scanning, H1, H3, H5, and H7 are excited with equal amplitude and identical phase differences are introduced between adjacent ports. The radiation patterns of the HB antenna array scanned in the yo z -plane under different phase differences are shown in Figure 5.17. It can be observed that the HB beams exhibit neither distortion nor angular deviation with the presence of the modified LB antenna and they remain consistent with those obtained in the absence of the LB antenna. This demonstrates that the proposed LB antenna exhibits good angular stability in its scattering suppression characteristics.

5.2.4 Experimental Results

The prototype of the proposed tri-band antenna array and the far-field test environment are shown in Figure 5.18(a) and (b), respectively. Figure 5.19 shows the simulated and measured S parameters of the antennas. The measured reflection coefficients ($|S_{11}|$ and $|S_{22}|$) of the LB, MB, and HB antennas are less than -10 dB in 3.05-4.68 GHz (42.2%), 6.2-10.0 GHz (46.9%), and 10.0-15.6 GHz (43.8%), respectively, which agree well with the simulated ones. Meanwhile, each of the three antenna types maintains over 20 dB polarization isolation across its respective

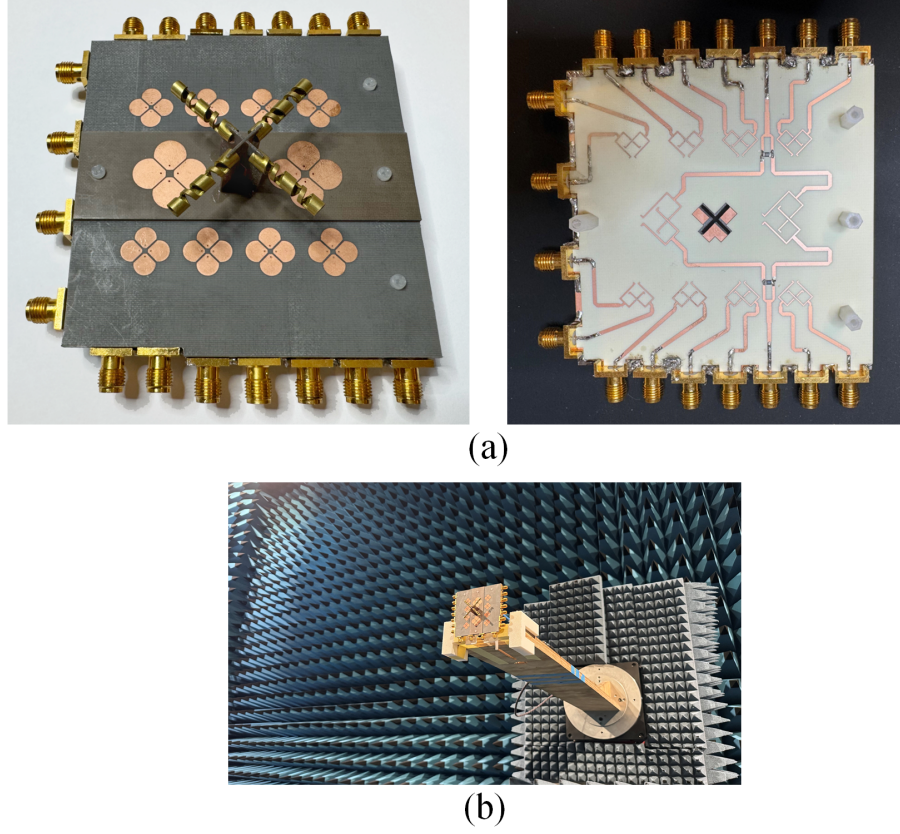


Figure 5.18: (a) Prototype of the tri-band antenna array and (b) far-field test environment.

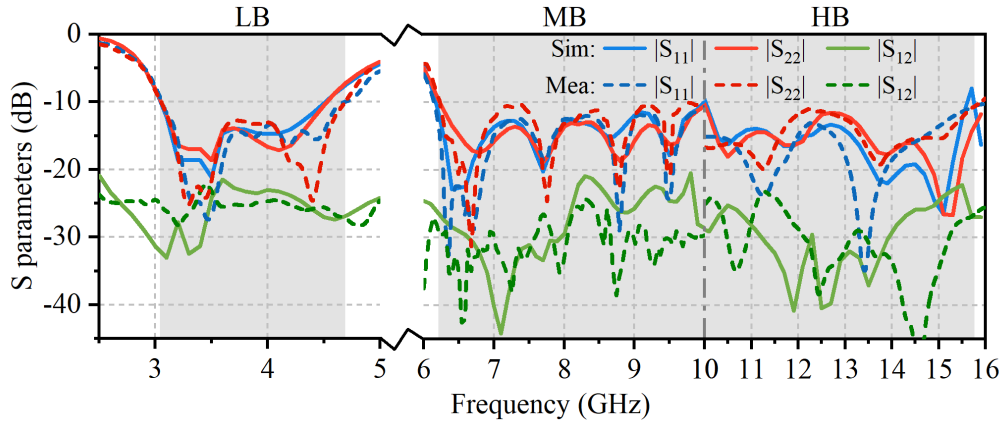


Figure 5.19: Simulated and measured S parameters of the LB, MB, and HB antennas in the tri-band array.

operating band.

The simulated and measured radiation patterns of the LB antenna are compared in Figure 5.20. The measured radiation patterns show good agreement with the simulated results, both exhibiting no significant distortion. These results confirm the effectiveness of the proposed pla-

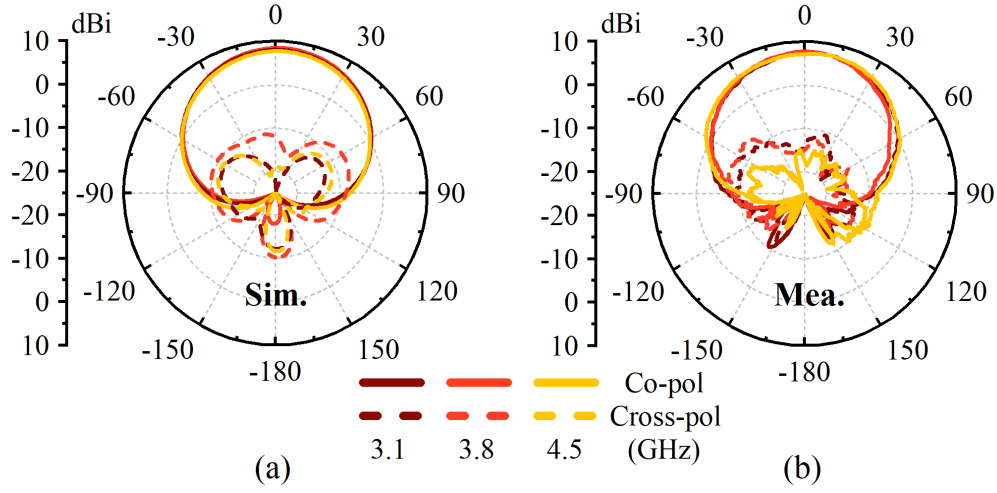


Figure 5.20: (a) Simulated and (b) measured radiation patterns of the LB antenna in yoz plane, when port L1 is excited.

nar ME dipole structure for the MB and HB antennas in avoiding the common-mode resonance commonly occurring in the LB.

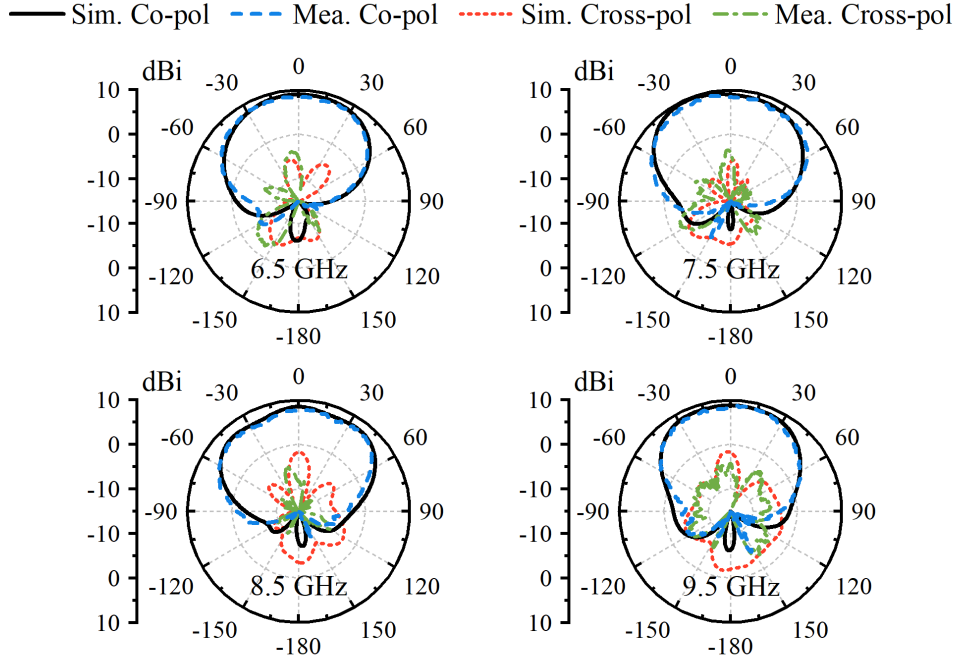


Figure 5.21: Simulated and measured radiation patterns of the MB antennas in the array.

The measured radiation patterns of the MB and HB antennas in the xoz plane are shown in Figures 5.21 and 5.22, respectively. The measured radiation patterns agree well with the simulated results. These undistorted MB/HB patterns validate the efficacy of the designed MB/HB-transparent segmented spiral in suppressing cross-band scattering across the ultra-

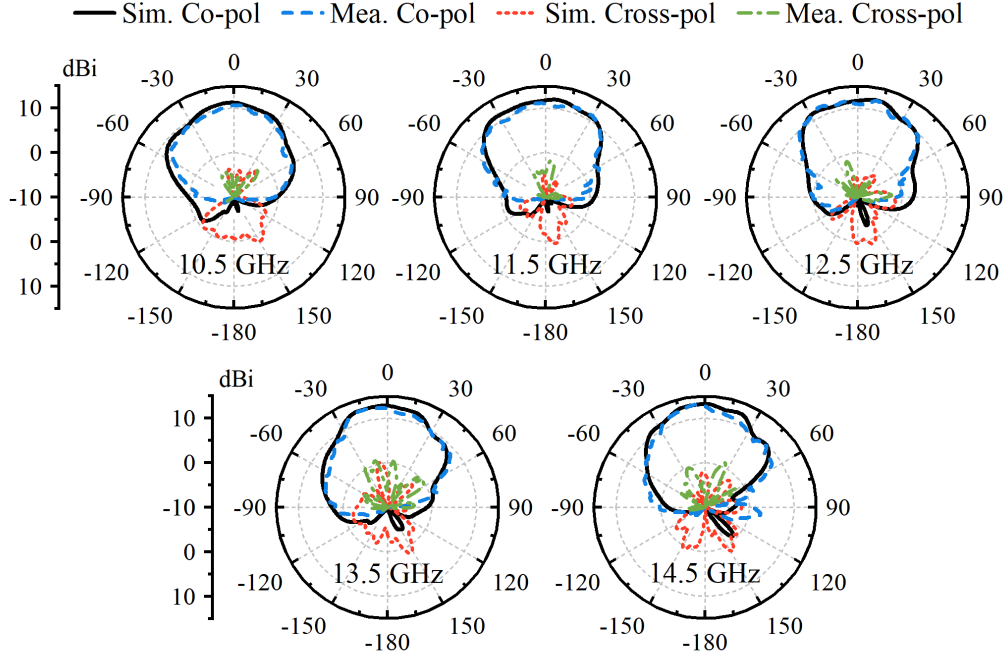


Figure 5.22: Simulated and measured radiation patterns of the HB antennas in the array.

wide frequency bands.

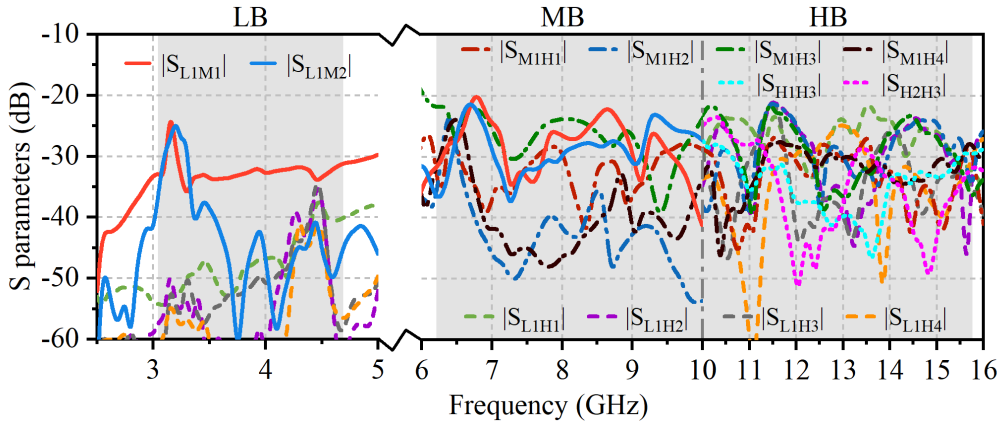


Figure 5.23: Measured transmission coefficients between the LB, MB, and HB antennas in the tri-band antenna array.

The measured isolation between any two ports in the proposed tri-band antenna array exceeds 20 dB across the three target bands, and representative results are shown in Figure 5.23. Therefore, the proposed suppression techniques not only achieve undistorted radiation patterns but also ensure good cross-band and in-band isolation over wide bands, holistically suppressing both scattering and coupling in the tri-band antenna array.

To demonstrate the advantages of the techniques developed in this work, their performance

is compared with that of other techniques for scattering and coupling suppression in tri-band arrays, as listed in Table 5.2. This work achieves tri-functional suppression of scattering and coupling without increasing the complexity of the array configuration, featuring the widest reported bandwidths across all three bands: LB, MB, and HB.

Table 5.2: Comparison of Techniques for Scattering and Coupling Suppression in Tri-Band Antenna Arrays

Ref.	Proposed structures	Suppression functions	Scattering-suppression /Operating band (GHz)	Cross-band isolation (dB)	In-band isolation (dB)
[99]	LB: Dual-band FSS radiator MB/HB: DRA	Cross-band scattering	LB: 1.85-2.15 (15%) MB: 3.4-3.6 (5.7%) HB: 5.4-5.6 (3.6%)	N/A	N/A
[93]	LB: Choke + Slot MB: Slot + Metal strip	Cross-band scattering	LB: 0.79-0.96 (19.4%) MB: 1.71-2.17 (23.7%) HB: 3.4-3.6 (5.7%)	N/A	N/A
[94]	LB: 45° rotation + slot MB: FSS radiator	Cross-band scattering	LB: 0.76-0.88 (14.6%) MB: 1.9-2.7 (34.8%) HB: 3.3-3.9 (16.7%)	N/A	N/A
[96]	Stacked arrangement with FSS MB: FSS radiator + Helical cable HB: Fence + Helical cable	Cross-band scattering Cross-band coupling In-band coupling	LB: 0.69-0.96 (32.7%) MB: 1.8-2.7 (40.0%) HB: 3.3-3.8 (14.1%)	> 20 (LB) > 19 (MB) > 19 (HB)	> 20 (HB)
[97]	LB/MB: A dual-band antenna with split rings HB: Stacked arrangement	Cross-band scattering Cross-band coupling	LB/MB: 1.427-1.518 (6.2%) 1.92-2.18 (12.7%) HB: 3.3-3.8 (14.1%)	> 20 (LB/MB) > 20 (HB)	N/A
This work	LB: Segmented spiral + Suppressor MB/HB: Planar ME dipole	Cross-band scattering Cross-band coupling In-band coupling	LB: 3.05-4.68 (42.2%) MB: 6.2-10.0 (46.9%) HB: 10.0-15.6 (43.8%)	> 20 (LB) > 20 (MB) > 20 (HB)	> 20 (HB)

5.2.5 Conclusion

This work proposes a tri-band shared-aperture 5G/6G antenna array featuring wideband scattering and coupling suppression, which is achieved by a modified LB antenna with segmented spirals, serial resonators, and suppressors, together with MB and HB antennas based on the planar ME dipole structures. The LB antenna (3.05–4.68 GHz, 42.2%) covers the 5G band of 3.3–4.2 GHz, while the MB (6.2–10.0 GHz, 46.9%) and HB (10.0–15.6 GHz, 43.8%) antennas collectively span the expected 5G-Advanced and 6G spectrum of 6.425-15.35 GHz. Across all three bands, the antennas maintain undistorted radiation patterns, and the isolation between any two ports in the array exceeds 20 dB. The measured results show good agreement with

the simulations, validating the effective suppression of scattering and coupling over the wide bands. Therefore, the proposed tri-band shared-aperture 5G/6G antenna array is a promising candidate for future 6G applications.

5.3 General Scattering and Coupling Suppression Methods for Tri-Band Shared-Aperture Arrays

In Section 5.2, the MB and HB antennas jointly covering 6.425–15.35 GHz employ a planar ME dipole structure. All of their components are etched on multilayer laminated dielectric substrates, and the overall antenna height corresponds to the thickness of the laminated substrates. Consequently, this type of planar structure can only be applied to the MB and HB antennas operating at relatively high frequency bands; otherwise, the substrate thickness would become excessively large. In addition, the operating bands of the MB and HB antennas are adjacent to each other. Therefore, the methods proposed in Section 5.2 are subject to certain limitations.

In this section, several more general scattering and coupling suppression techniques applicable to tri-band arrays are proposed. In the proposed array, both the MB and HB antennas employ a conventional three-dimensional dipole structure, and their operating bands are not adjacent but separated by a certain frequency gap.

An LB antenna employing an uneven spiral radiator is proposed in Section 5.3.1, which enables scattering suppression over a wide bandwidth covering both the MB (1.7–2.7 GHz) and HB (3.3–3.8 GHz). With the aid of a serial resonator, the LB antenna achieves good impedance matching within the target LB of 0.69–0.96 GHz.

An MB antenna employing a partial spiral radiator is proposed in Section 5.3.2, achieving effective scattering suppression within the HB (3.3–3.8 GHz). By introducing the proposed vortex line, a wide impedance bandwidth of 1.54–2.88 GHz (60.6%) is realized. In addition, a parallel LC structure is incorporated at the bottom of the LB balun to shift the resonance generated by the MB antenna out of the target LB band.

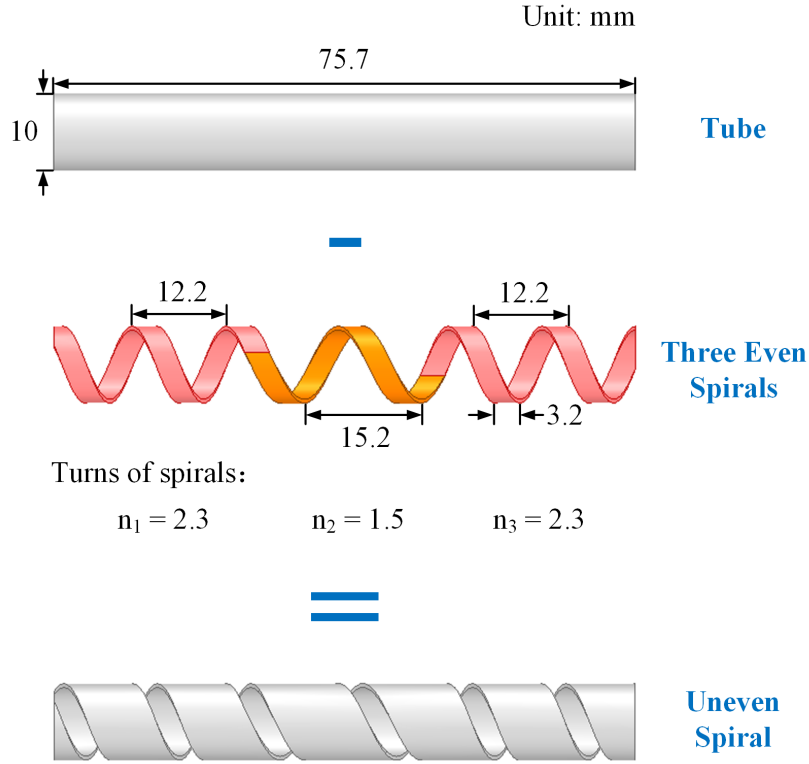


Figure 5.24: Development and geometry of the uneven spiral.

5.3.1 LB Design

To realize scattering and coupling suppression for the LB radiator within the target bands MB (1.7–2.7 GHz) and HB (3.3–3.8 GHz), an uneven spiral structure is proposed, as illustrated in Figure 5.24. The uneven spiral is realized by etching three even spirals on a metallic tube. All even spirals share the same width of 3.2 mm, while the middle even spiral has a larger pitch of 15.5 mm and the two side even spirals have smaller pitches of 12.2 mm. Their respective numbers of turns are 2.3, 1.5, and 2.3. Consequently, the uneven spiral exhibits a wider central section and narrower side sections, with the spiral edges forming a continuous curve.

The monostatic RCS of the designed uneven spiral is shown in Figure 5.25. Compared with the reference tube, it achieves significant RCS reduction in the target MB (1.7–2.7 GHz) and HB (3.3–3.8 GHz), with the reduction value exceeding 10 dB over both bands.

The RCS curve of the segmented spiral proposed in Chapter 4 is also presented in Figure 5.25. Similarly, the segmented spiral is employed as the radiator of the dipole antenna operating at 0.69–0.96 GHz. It can also achieve RCS reduction in the 1.7–2.7 GHz band; however, the reduction value drops sharply over the HB (3.3–3.8 GHz), making it unable to provide effective

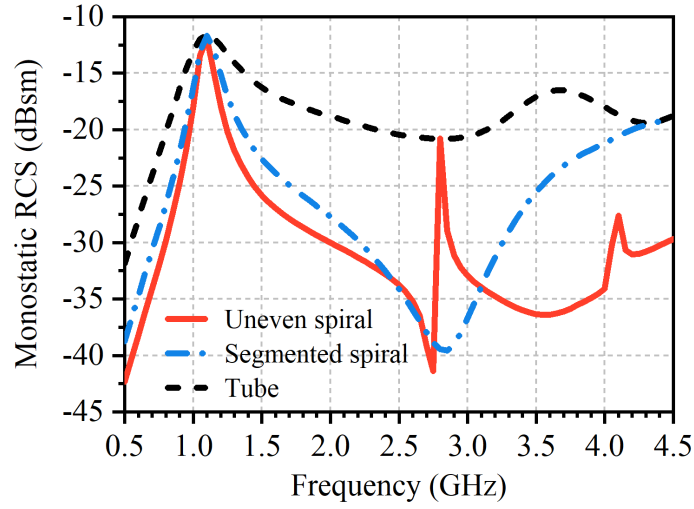


Figure 5.25: Monostatic RCSs of the uneven spiral, segmented spiral and tube.

scattering and coupling suppression in the HB. The uneven spiral proposed in this work obtains effective RCS reduction in both the MB and HB by adjusting the pitch and length of different sections. Moreover, the pronounced fluctuation on the RCS curve is intentionally positioned at 2.9 GHz, lying between the target MB and HB.

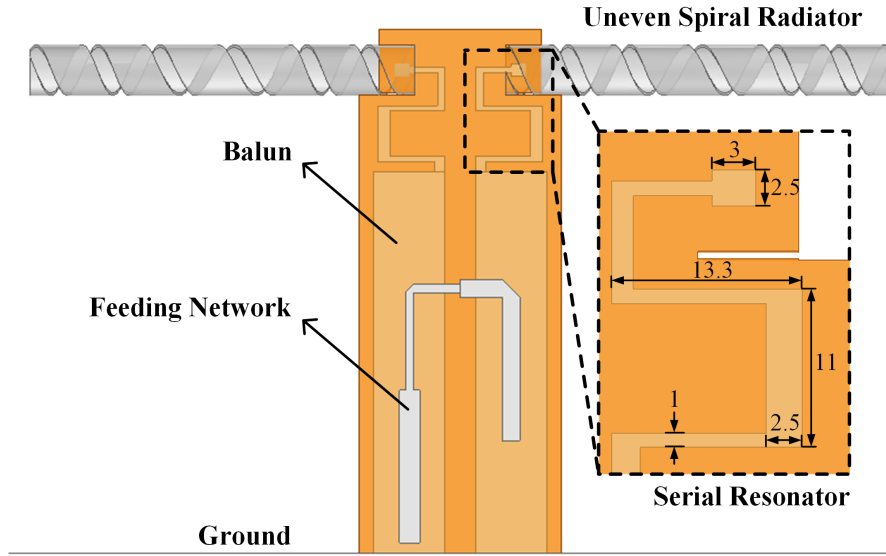


Figure 5.26: Geometry of the LB uneven spiral antenna with integrated serial resonator.

As illustrated in Figure 5.26, the designed uneven spirals are used as the LB dipole radiator. In order to realize impedance matching in the target LB (0.69–0.96 GHz), the serial resonator previously adopted in Chapter 4 and in Section 1 of Chapter 5 is likewise incorporated between the uneven spirals and the balun.

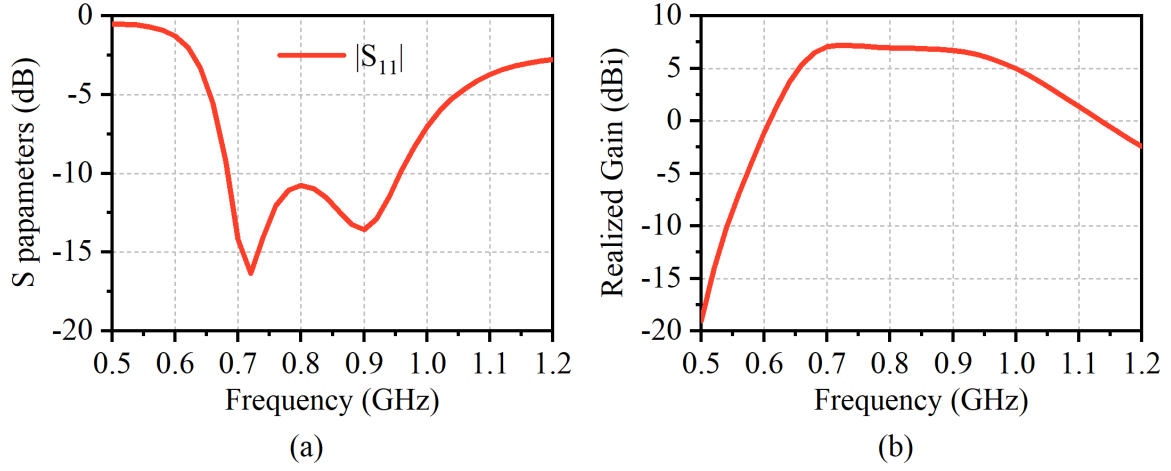


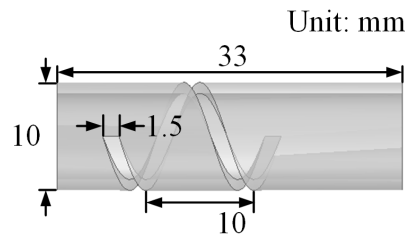
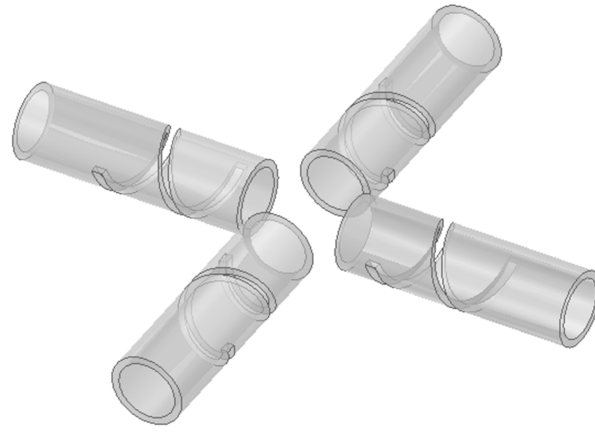
Figure 5.27: (a) Reflection coefficient and (b) realized gain of the LB uneven spiral antenna.

The reflection coefficient and the realized gain of the LB uneven spiral antenna are given in Figure 5.27 (a) and (b), respectively. In the target LB of 0.69–0.96 GHz, the $|S_{11}|$ of the LB uneven spiral antenna is below -10 dB and exhibits a stable realized gain. Owing to the uneven spiral and the serial resonator, the suppression bandwidth of the proposed LB antenna is significantly enhanced compared with the LB segmented-spiral-based antenna presented in Chapter 4, while its matching bandwidth and radiation performance are preserved without compromise.

5.3.2 MB Design

The radiator of the MB antenna is required to achieve impedance matching and radiation over the wide band of 1.7–2.7 GHz, while realizing effective scattering suppression in the HB (3.3–3.8 GHz). Since the MB and HB are spectrally close, it is highly challenging to switch rapidly from radiation to scattering suppression within such a narrow frequency interval.

A partial spiral structure is proposed, as shown in Figure 5.28, which exhibits excellent scattering-suppression characteristics in the target HB (3.3–3.8 GHz). As given in Figure 5.29, compared with the reference tube, the monostatic RCS of the partial spiral is significantly reduced within the 3.3–3.8 GHz, with the reduction exceeding 10 dB across the entire range. From 2.7 GHz (the highest frequency point of the 1.7–2.7 GHz band) to 3.3 GHz (the lowest frequency point of the 3.3–3.8 GHz band), the RCS value of the partial spiral decreases by 10 dB within a bandwidth of only 0.6 GHz.



Turns of slot: 1.5

Figure 5.28: Geometry of the partial spiral for MB radiator.

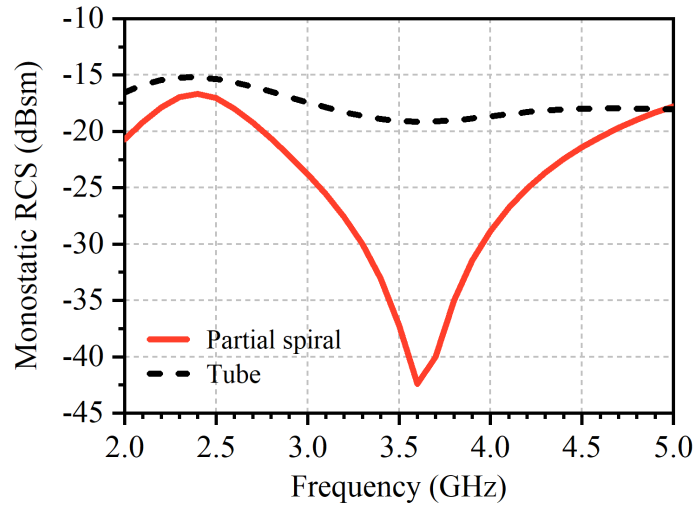


Figure 5.29: Monostatic RCSs of the partial spiral and tube.

By inserting the serial resonator, which has been repeatedly employed in the previous sections, between the partial spiral radiators and the balun, impedance matching in the 1.7–2.7 GHz (45.5%) can be achieved. To obtain a broader impedance bandwidth for MB antenna, the meander line structure shown in Figure 5.30 is proposed. Different from the serial resonator, which has separate capacitor and inductor structures, the designed meander line integrates the

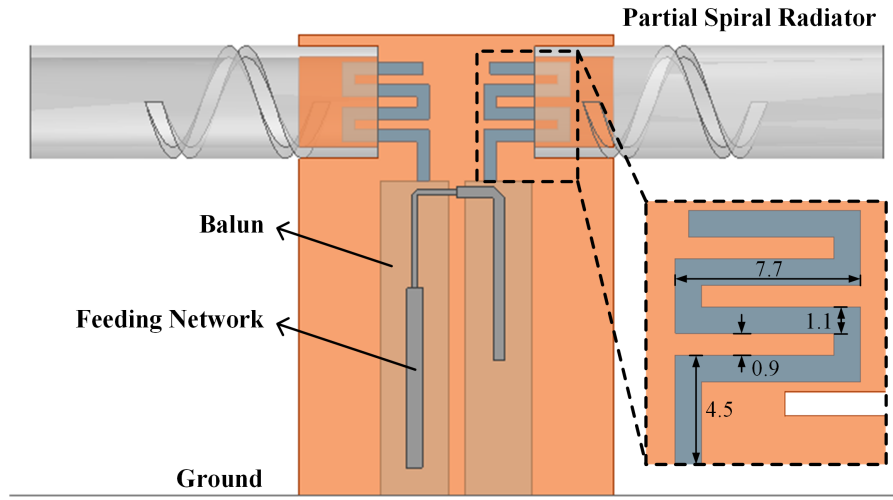


Figure 5.30: Geometry of the MB partial spiral antenna with the meander lines.

capacitor and inductor into one structure. The meander line inserted into the partial spiral radiators forms capacitance, while the bends of the line itself provide inductance. The insertion depth of the meander line into the radiator and the width of the meander line determine the capacitance value, while the length of the meander line determines the inductance value. This integrated LC structure significantly reduces the space occupied between the MB radiators and the balun, thereby providing additional room for the balun and feeding networks to realize wide impedance matching.

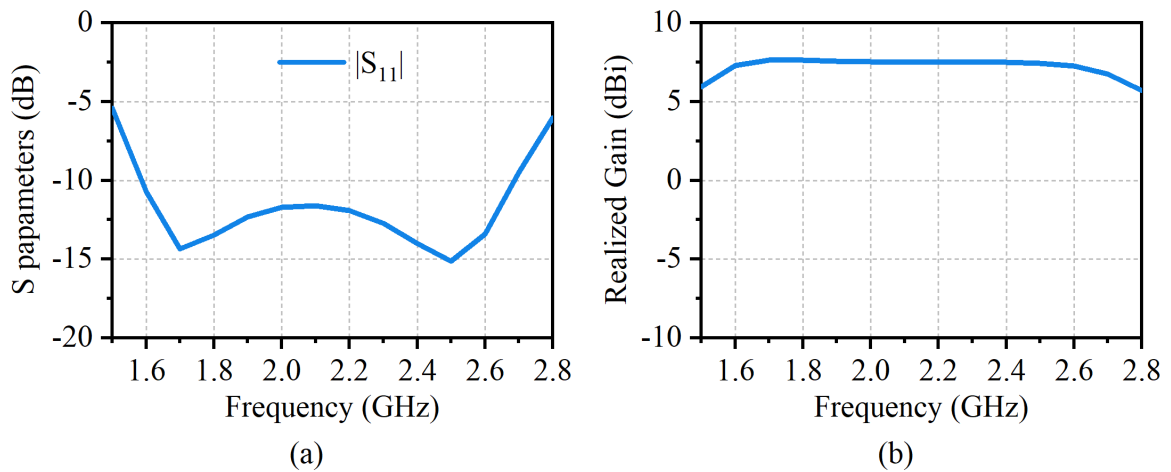


Figure 5.31: (a) Reflection coefficient and (b) realized gain of the MB partial spiral antenna with the meander line.

In addition to its space-saving advantage, the meander line contributes to a wider impedance

matching bandwidth for the MB antenna. As shown in Figure 5.31, the $|S_{11}|$ of the MB antenna is below -10 dB within the range of 1.6–2.7 GHz (51%), and the realized gain remains very stable across this band.

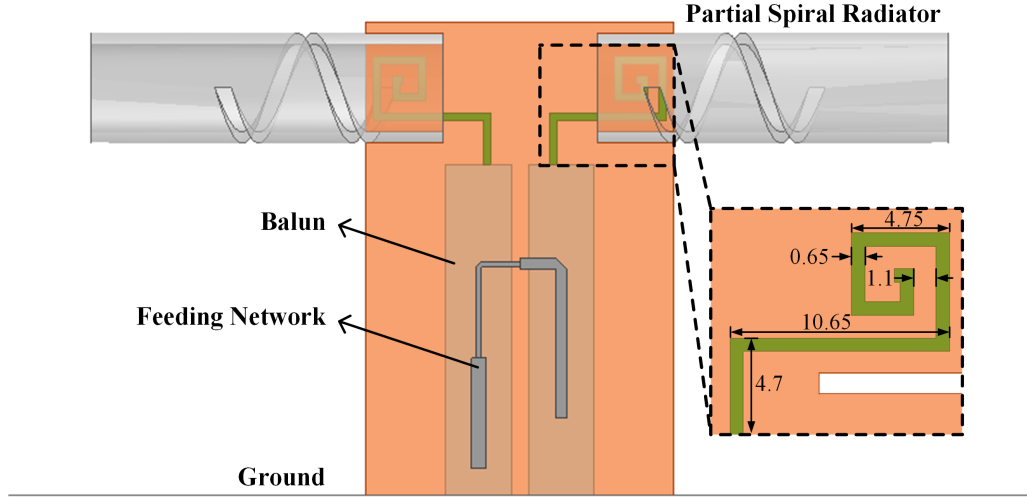


Figure 5.32: Geometry of the MB partial spiral antenna with the vortex lines.

Following the same design concept of the integrated LC structure with meander lines, a vortex line structure is proposed, as shown in Figure 5.32. Similarly, the vortex line forms a capacitor with the partial spiral radiators, while the vortex structure itself provides inductance. The width of the vortex line determines the capacitance value, while its length determines the inductance value.

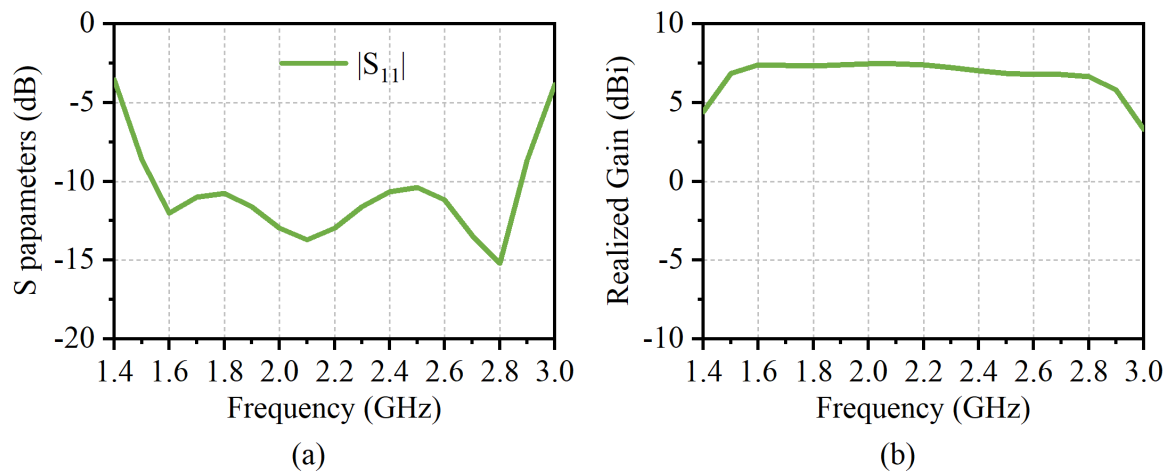


Figure 5.33: (a) Reflection coefficient and (b) realized gain of the MB partial spiral antenna with the vortex line.

As illustrated in Figure 5.33, by loading the vortex line, the MB partial spiral antenna attains an enhanced impedance matching bandwidth ranging from 1.54 to 2.88 GHz, corresponding to a relative bandwidth of 60.6%. A stable realized gain is also maintained throughout this wide band.

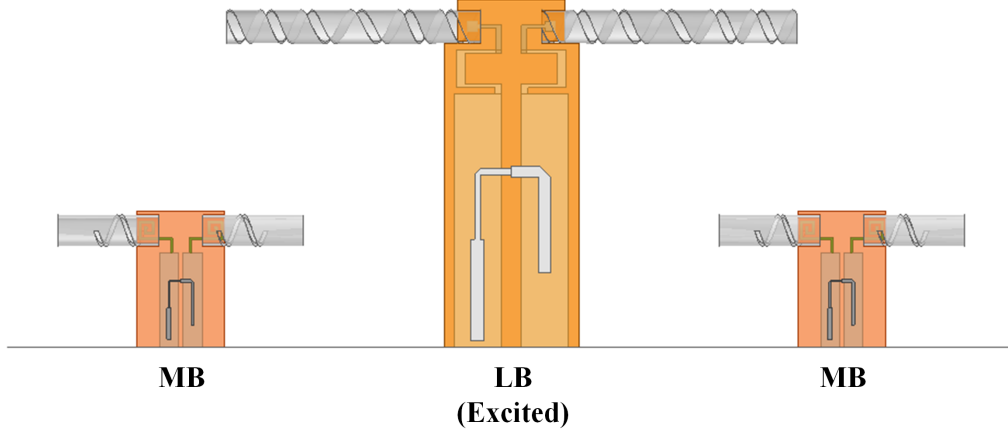


Figure 5.34: Two MB partial spiral antennas are placed beneath the LB uneven spiral antenna.

For the MB antenna in a tri-band shared-aperture antenna array, in addition to achieving scattering suppression in the HB, wideband impedance matching and radiation in the MB, it is important to investigate whether its presence influences the radiation characteristics of the LB antenna.

As shown in Figure 5.34, the two MB partial spiral antennas are placed beneath the LB uneven spiral antenna, and the three antennas are aligned collinearly. From Figure 5.35, it can be observed that the introduction of the MB antenna causes a noticeable reduction in the realized gain of the LB antenna in the higher frequency region, compared with the case when the LB antenna operates alone. This phenomenon can be attributed to the common-mode resonance introduced by the MB antenna structure in the higher frequency region of the LB. The LB antenna excites both MB antennas simultaneously, thereby causing a degradation in its realized gain at boresight.

To recover the realized gain of the LB antenna, the resonance point generated by the MB antenna must be moved outside the target LB of 0.69–0.96 GHz. Adding a series inductor or capacitor structure to the MB antenna is still an effective approach. However, since the vortex line has already been loaded between the radiator and the balun of the MB antenna to achieve wideband impedance matching, introducing an additional capacitor or inductor at this location

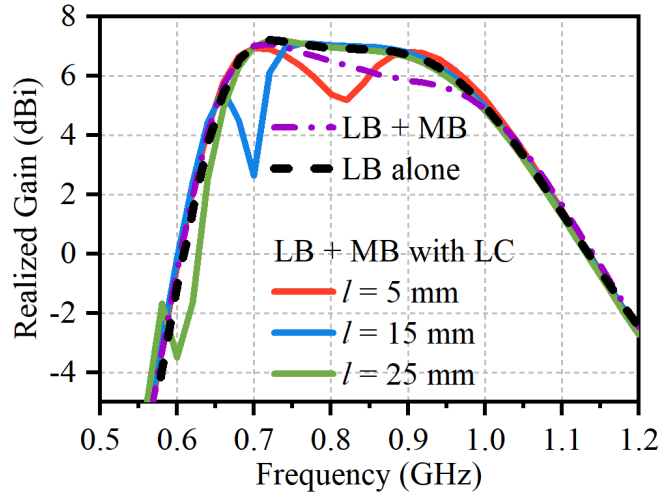


Figure 5.35: Realized gain of the LB uneven spiral antenna under different cases.

would inevitably affect the matching performance of the MB antenna. Therefore, the capacitor or inductor can only be loaded between the balun and the ground.

Adding a series capacitor at this position can shift the resonance point to a higher frequency. However, the realized gain of the LB antenna still exhibits a degradation, even though the resonance point has moved far away from the target LB. Conversely, introducing a series inductor at this position can shift the resonance point to a lower frequency. Moreover, once the resonance point is moved out of the target LB, the realized gain within the LB does not experience degradation. However, moving this resonance point outside the LB band necessitates an inductor with a large inductance value. When implemented with a metallic strip, such an inductor would require a very long strip length.

As shown in Figure 5.36, a parallel LC structure is loaded between the MB balun and the ground. By increasing the inductance value, namely the length of the strip that constitutes the inductor, the resonance point can be shifted toward lower frequencies and moved out of the target LB as presented in Figure 5.35. At $l = 25$ mm, the resonance introduced by the MB antenna is shifted to 0.6 GHz, and the realized gain within the LB (0.69-0.96 GHz) shows no degradation. It is worth noting that the required inductance in this parallel LC is significantly smaller than that of using a single series inductor.

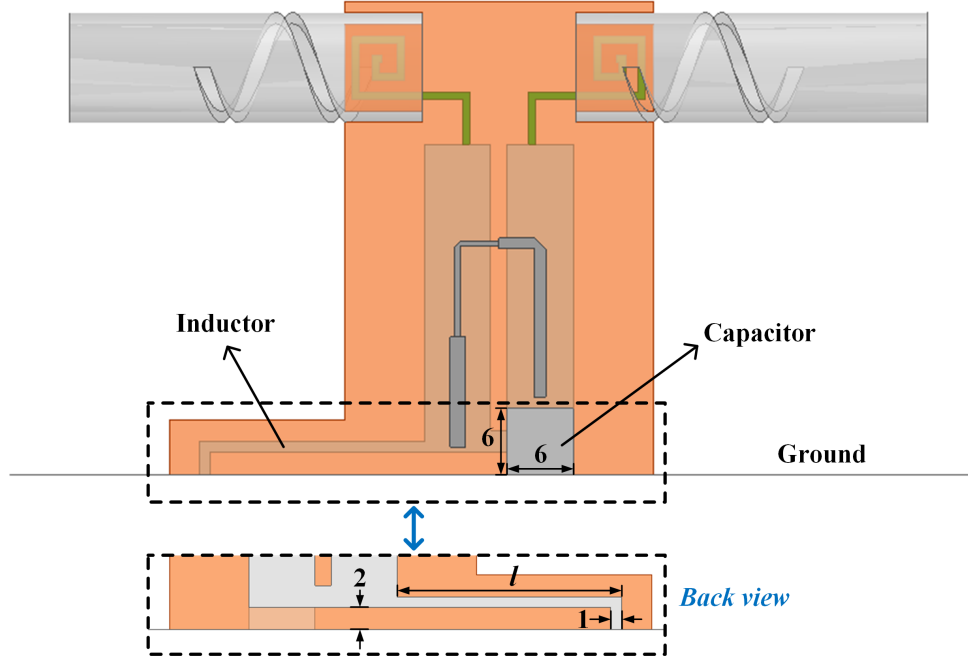


Figure 5.36: Geometry of the MB antenna with the parallel LC structure.

5.3.3 Conclusion

This section proposes an LB antenna and an MB antenna for a tri-band shared-aperture antenna array, using different scattering and coupling suppression methods. The LB antenna employs an uneven spiral as the radiator, which enables scattering suppression over a wide band, covering both the MB and HB. With the aid of a serial resonator, the LB antenna achieves impedance matching within the target band of 0.69–0.96 GHz. The MB antenna adopts a partial spiral as the radiator, achieving effective scattering suppression within the HB (3.3–3.8 GHz). With the proposed vortex line, a wide impedance bandwidth of 1.54–2.88 GHz (60.6%) is realized. In addition, a parallel LC structure is incorporated at the bottom of the LB balun to shift the resonance generated by the MB antenna out of the target LB.

The suppression methods, as well as the LB and MB antennas proposed in this section, exhibit greater generality compared with those in Section 5.2. Both the MB and HB antennas can operate at relatively low frequencies and employ conventional 3D dipole structures.

Chapter 6

Conclusions and Future Work

6.1 Summary of Outcomes

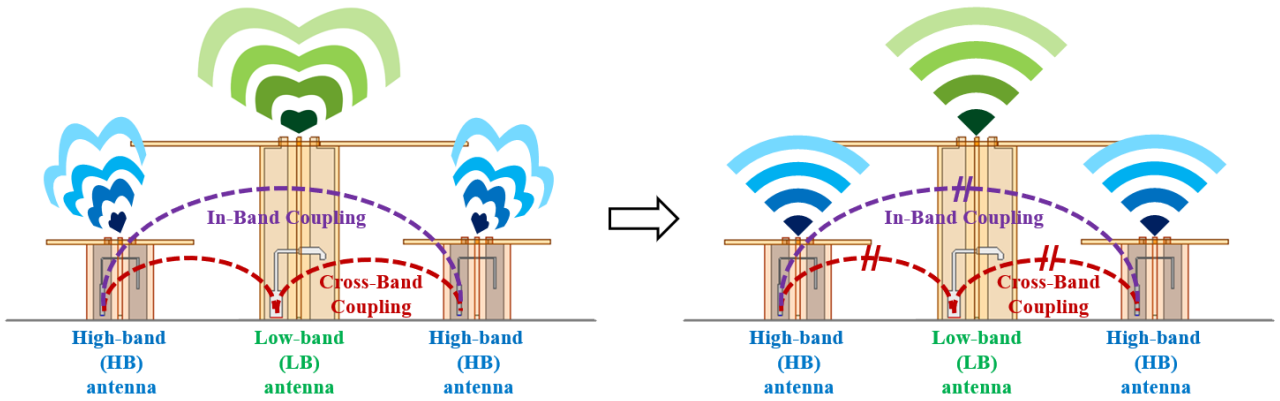


Figure 6.1: Schematic diagrams of the works in Chapter 3 and 4.

The outcomes of this thesis are summarized as follows:

A dual-band shared-aperture antenna array with simultaneous suppression of scattering and coupling is proposed in Chapter 3. A 2.5-D cloak and a defected ground structure (DGS) are respectively implemented on the LB radiator and the LB balun, while a resonance shifter is introduced into the HB antenna. These structures can not only suppress cross-band scattering but also effectively mitigate both cross-band and in-band coupling. Moreover, the proposed method exhibits strong universality and robustness, making it applicable to different frequency bands and various array configurations. Under the two demonstrated configurations, the radiation patterns of the LB and HB antennas are significantly restored, showing good consistency with those when they operate alone. Moreover, both the cross-band and in-band couplings in

the dual-band array remain below -23 dB within the target bands.

A dual-band shared-aperture antenna array that achieves simultaneous wideband suppression of scattering and coupling is proposed in Chapter 4. A segmented spiral radiator is developed for the LB antenna under the precise guidance of characteristic mode analysis (CMA), achieving scattering suppression over an ultra-wide band of 118%. To further enhance impedance matching, serial resonators are introduced to the LB antenna, extending the matching bandwidth to 0.69–1.00 GHz (36.7%). In addition, parallel resonance shifters are co-designed with the HB antenna to suppress LB scattering, and the HB antennas operate over a wide bandwidth of 1.63–2.83 GHz (53.8%). The radiation pattern distortions of both the LB and HB antennas are effectively mitigated, and both the cross-band and in-band isolations in the dual-band array exceed 20 dB.

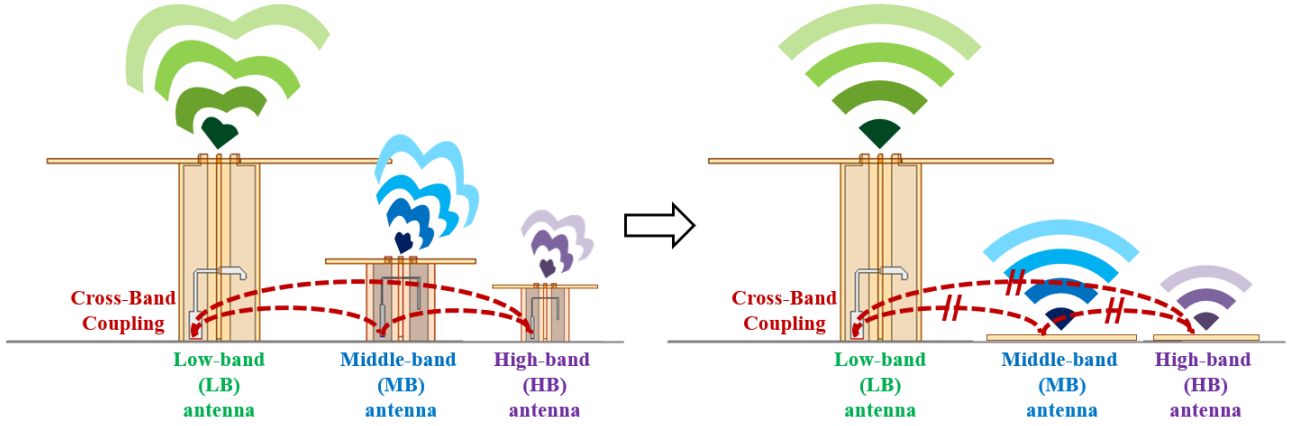


Figure 6.2: Schematic diagrams of the work in Chapter 5, Section 5.2.

A tri-band shared-aperture 5G/6G antenna array featuring wideband simultaneous scattering and coupling suppression is proposed in Chapter 5, Section 5.2. The LB antenna adopts segmented spirals as radiators, designed using the method proposed in Chapter 4. With the aid of serial resonators, an impedance bandwidth of 3.05–4.68 GHz (42.2%) is achieved, covering the 5G band of 3.3–4.2 GHz. The MB and HB antennas are both based on planar ME dipole structures, operating over 6.2–10.0 GHz (46.9%) and 10.0–15.6 GHz (43.8%), respectively, collectively spanning the expected 5G-Advanced and 6G spectra from 6.425 to 15.35 GHz. The radiation patterns of all three antennas show no significant distortion, and the isolation between any two antennas in the proposed tri-band array is more than 20 dB.

Improved scattering and coupling suppression methods and structures applicable to general tri-band shared-aperture antenna arrays are proposed in Chapter 5, Section 5.3. Both the MB and

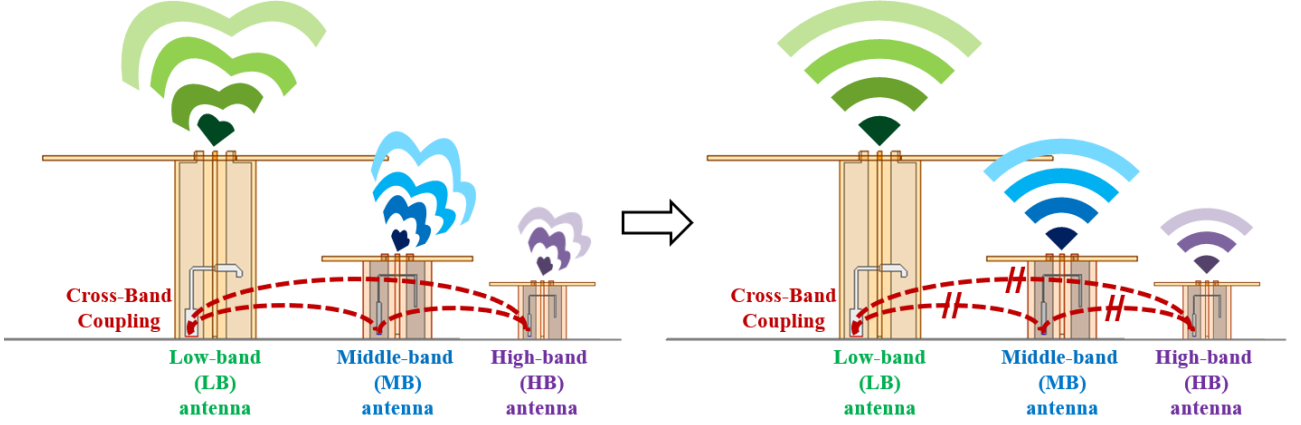


Figure 6.3: Schematic diagrams of the work in Chapter 5, Section 5.3.

HB antennas can operate at lower frequencies and employ conventional 3D dipole structures. The MB antenna adopts a partial spiral radiator, providing effective scattering suppression within the HB (3.3–3.8 GHz). By incorporating a parallel LC structure at the bottom of the LB balun, the unwanted resonance is shifted out of the target LB. In addition, a wide impedance bandwidth of 1.54–2.88 GHz (60.6%) is realized by introducing the proposed vortex line. The LB antenna employs an uneven spiral radiator, which enables wideband scattering suppression covering both the MB and HB. With the integration of serial resonators, the LB antenna achieves impedance matching within the band of 0.69–0.96 GHz.

6.2 Current and Future Work: Multi-Band Antenna Arrays for Non-Terrestrial Networks (NTNs)

The dual-band and tri-band shared-aperture antenna arrays proposed in Chapters 3–5 are mainly intended for base station applications in 3G, 4G, 5G, and future 6G mobile communication systems.

In future 6G networks, Non-Terrestrial Networks (NTNs) will become a key component of the communication infrastructure, enabling the deep integration of traditional terrestrial base stations such as macro and small cells with non-terrestrial nodes including satellites, unmanned aerial vehicles (UAVs), and high-altitude platform stations (HAPS). This integration aims to achieve seamless global coverage across oceans, deserts, polar regions, aerial routes, and other remote areas, thereby eliminating the coverage gaps in terrestrial networks. Meanwhile, antennas operating over multiple frequency bands are expected to be deployed on the same platform

to support multiple communication standards simultaneously. Such convergence of terrestrial and non-terrestrial systems will be fundamental for realizing ubiquitous, high-capacity, and resilient connectivity in the 6G era.

Compared with terrestrial base stations, the platforms in NTN, such as satellites and UAVs, provide much more limited space for the deployment of antennas. Therefore, the suppression of scattering and coupling in multi-band shared-aperture antenna arrays is expected to be more challenging in such compact NTN platforms, which constitutes the main topic of our ongoing and future work.

A CubeSat, the smallest form of satellite, is chosen as the platform for the dual-band shared-aperture antenna array. Since its structure consists of multiple cubic units with dimensions of $10\text{ cm} \times 10\text{ cm} \times 10\text{ cm}$, the ground plane of the antenna array is designed to be $10\text{ cm} \times 10\text{ cm}$ in size, as shown in Figure 6.4.

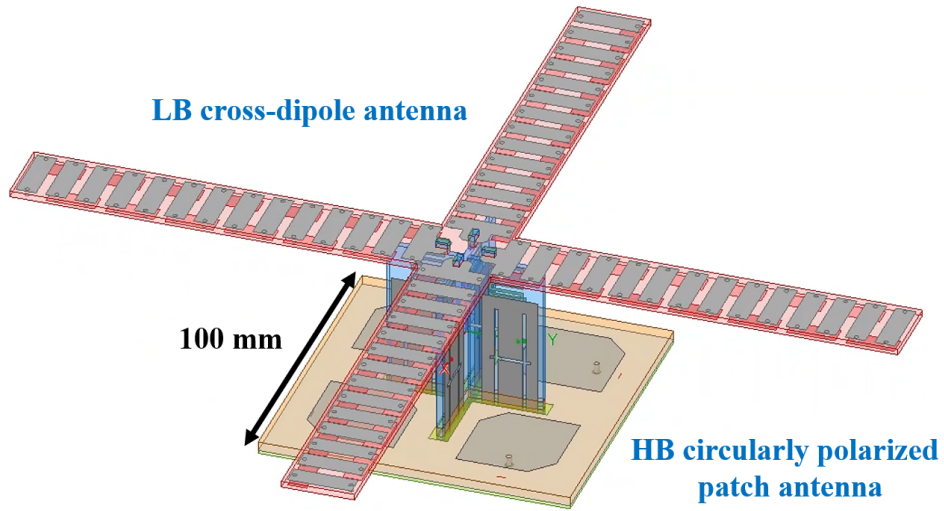


Figure 6.4: Dual-band shared-aperture antenna array for CubeSat.

In the dual-band antenna array, as shown in Figure 6.4, the LB dipole antenna operates at 400–440 MHz, while the HB antenna is a circularly polarized patch antenna composed of four elements, operating in the 2.2–2.5 GHz. The radiator of the LB antenna employs a 2.5-D structure, in which the metal patches on the upper surface of the dielectric layer are arranged in parallel, while patches on the lower surface are arranged in a staggered manner. The upper and lower metal patches are interconnected through vias, forming a folded LB radiator that effectively suppresses scattering in the HB.

In addition, an H-shaped slot is etched on the LB balun to suppress the scattering generated by the balun in the HB and to further improve the cross-band isolation between the LB and HB antennas within the HB. Owing to the incorporation of these suppression structures, the radiation pattern of the HB antenna remains unaffected by the LB antenna, despite the extremely close spacing. Moreover, the cross-band isolation between the LB and HB antennas is more than 20 dB within the target frequency bands.

Considering the limited space for the CubeSat platform and the deployment of the LB antenna, the LB radiator should be divided into two sections connected by a hinge with a spring. During the launch phase, the LB radiator is folded into an L-shaped configuration and mechanically restrained to minimize the occupied volume. After the CubeSat reaches orbit, the restraining structure is released, allowing the LB radiator to automatically deploy. The overall deployment process and mechanical scheme are similar to those commonly adopted for satellite solar panels.

Bibliography

- [1] Y. J. Guo and R. W. Ziolkowski, “Advanced antenna array engineering for 6G and beyond wireless communications,” in 2022.
- [2] Y. Chen and C.-F. Wang, *Characteristic modes: Theory and applications in antenna engineering*. John Wiley & Sons, 2015.
- [3] S. Gong, Y. Liu, G. Fu, *et al.*, *Microwave Technology and Antennas*. Beijing, China: Higher Education Press, 2014.
- [4] A. Monti, J. Soric, A. Alu, F. Bilotti, A. Toscano, and L. Vegni, “Overcoming mutual blockage between neighboring dipole antennas using a low-profile patterned metasurface,” *IEEE Antennas and Wireless Propagation Letters*, vol. 11, pp. 1414–1417, 2012.
- [5] H. M. Bernety and A. B. Yakovlev, “Reduction of mutual coupling between neighboring strip dipole antennas using confocal elliptical metasurface cloaks,” *IEEE Transactions on Antennas and Propagation*, vol. 63, no. 4, pp. 1554–1563, 2015.
- [6] J. C. Soric, A. Monti, A. Toscano, F. Bilotti, and A. Alù, “Dual-polarized reduction of dipole antenna blockage using mantle cloaks,” *IEEE Transactions on Antennas and Propagation*, vol. 63, no. 11, pp. 4827–4834, 2015.
- [7] Z. H. Jiang, P. E. Sieber, L. Kang, and D. H. Werner, “Restoring intrinsic properties of electromagnetic radiators using ultralightweight integrated metasurface cloaks,” *Advanced Functional Materials*, vol. 25, no. 29, pp. 4708–4716, 2015.
- [8] A. Monti, J. Soric, M. Barbuto, *et al.*, “Mantle cloaking for co-site radio-frequency antennas,” *Applied Physics Letters*, vol. 108, no. 11, 2016.
- [9] S. Vellucci, A. Monti, M. Barbuto, A. Toscano, and F. Bilotti, “Use of mantle cloaks to increase reliability of satellite-to-ground communication link,” *IEEE Journal on Multi-scale and Multiphysics Computational Techniques*, vol. 2, pp. 168–173, 2017.

- [10] A. Rajput and K. V. Srivastava, "Dual-band cloak using microstrip patch with embedded U-shaped slot," *IEEE Antennas and Wireless Propagation Letters*, vol. 16, pp. 2848–2851, 2017.
- [11] S. Vellucci, A. Monti, M. Barbuto, *et al.*, "On the use of nonlinear metasurfaces for circumventing fundamental limits of mantle cloaking for antennas," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 8, pp. 5048–5053, 2021.
- [12] G. Moreno, A. B. Yakovlev, H. M. Bernety, *et al.*, "Wideband elliptical metasurface cloaks in printed antenna technology," *IEEE Transactions on Antennas and Propagation*, vol. 66, no. 7, pp. 3512–3525, 2018.
- [13] S. J. Yang, Y. Yang, and X. Y. Zhang, "Low scattering element-based aperture-shared array for multiband base stations," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 12, pp. 8315–8324, 2021.
- [14] R. Wu, J.-H. Lin, F.-C. Chen, and L. H. Ye, "Dual-broadband aperture-shared base station antenna array using double decoupling techniques," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 10, pp. 3262–3266, 2024.
- [15] G. Su, W. Che, W. Yang, *et al.*, "Low-scattering dipole antenna using mushroom-shaped structure for applications in dual-band shared-aperture array," *IEEE Antennas and Wireless Propagation Letters*, vol. 22, no. 1, pp. 159–163, 2023.
- [16] S. J. Yang, R. Ma, and X. Y. Zhang, "Self-decoupled dual-band dual-polarized aperture-shared antenna array," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 6, pp. 4890–4895, 2022.
- [17] H. Huang, X. Li, and Y. Liu, "A method to suppress the coupling of dual-frequency antennas," *IEEE Antennas and Wireless Propagation Letters*, vol. 20, no. 8, pp. 1458–1462, 2021.
- [18] R. C. Dai, H. Su, S. J. Yang, J.-H. Ou, and X. Y. Zhang, "Broadband electromagnetic-transparent antenna and its application to aperture-shared dual-band base station array," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 1, pp. 180–189, 2023.
- [19] Y.-L. Chang and Q.-X. Chu, "Broadband dual-polarized electromagnetic transparent antenna for cross-band scattering suppression," *IEEE Antennas and Wireless Propagation Letters*, vol. 21, no. 7, pp. 1452–1456, 2022.

- [20] H. Yuan, F.-C. Chen, W.-F. Zeng, and Q.-X. Chu, "Dual-band base station antenna array with cross-band scattering and in-band coupling suppression," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 5, pp. 3983–3991, 2023.
- [21] Z. J. Xiao, Y. F. Cao, J. S. Lin, Y. Lan, and Q. Xue, "Dual-band shared-aperture base-station antenna array with dual polarization using filtering magnetoelectric dipole antenna," *IEEE Open Journal of Antennas and Propagation*, vol. 5, no. 1, pp. 82–89, 2024.
- [22] Y.-S. Wu, Q.-X. Chu, and H.-Y. Huang, "Electromagnetic transparent antenna with slot-loaded patch dipoles in dual-band array," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 9, pp. 7989–7998, 2022.
- [23] Q.-X. Chu, Y.-S. Wu, and Y.-L. Chang, "A novel electromagnetic transparent antenna in dual-band shared-aperture array," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 10, pp. 9894–9899, 2022.
- [24] J. Tao, P. Tang, H. Yang, *et al.*, "Design of electromagnetic transparent antennas based on drude metasurface for high-band compatibility," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 6, pp. 4911–4919, 2024.
- [25] W. Niu, B. Sun, G. Zhou, and Z. Lan, "Dual-band aperture shared antenna array with decreased radiation pattern distortion," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 7, pp. 6048–6053, 2022.
- [26] Y. He, W. Huang, Z. He, L. Zhang, X. Gao, and Z. Zeng, "A novel cross-band decoupled shared-aperture base station antenna array unit for 5G mobile communications," *IEEE Open Journal of Antennas and Propagation*, vol. 3, pp. 583–593, 2022.
- [27] Z. Chen, T. Xu, J.-F. Li, L. H. Ye, and D.-L. Wu, "Dual-broadband dual-polarized base station antenna array with stable radiation pattern," *IEEE Antennas and Wireless Propagation Letters*, vol. 22, no. 2, pp. 303–307, 2023.
- [28] X. W. Dai, D. L. Mi, H. Hong, S. Y. Lin, and G. Q. Luo, "Dual-polarized antenna with suppression of cross-band scattering in multiband array," *IEEE Antennas and Wireless Propagation Letters*, vol. 20, no. 8, pp. 1592–1595, 2021.
- [29] C. Liang, W. Yang, Q. Xue, W. Che, and H. Wong, "Wideband high-transparent dipole antenna based on frequency selective surface and its applications in dual-band shared-aperture array," *IEEE Antennas and Wireless Propagation Letters*, vol. 24, no. 3, pp. 572–576, 2025.

- [30] F. Jia, Z. Zheng, Q. Wang, *et al.*, “A new multi-band multi-array antenna configuration with scattering suppression for radiation pattern distortion mitigation of base station,” *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 7, pp. 6006–6011, 2022.
- [31] J. Jiang and Q.-X. Chu, “Dual-band shared-aperture base station antenna array based on 3-D chokes,” *IEEE Antennas and Wireless Propagation Letters*, vol. 22, no. 4, pp. 824–828, 2023.
- [32] X. Liu, S. Gao, B. Sanz-Izquierdo, *et al.*, “A mutual-coupling-suppressed dual-band dual-polarized base station antenna using multiple folded-dipole antenna,” *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 12, pp. 11 582–11 594, 2022.
- [33] W. Xie, F.-C. Chen, and H. Yuan, “A compact dual-band shared-aperture base station antenna array,” *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 5, pp. 1573–1577, 2024.
- [34] A. Pallaris and D. Sjöberg, “Design and measurement of a planar metasurface cloaked dipole antenna element for multifunction apertures,” *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 2, pp. 1182–1191, 2024.
- [35] C. Guo, B. Zhou, G. Su, and F. Xue, “A novel electromagnetic transparent antenna for 4G/5G dual-band share-aperture array,” *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 4, pp. 1171–1175, 2024.
- [36] H.-H. Sun, C. Ding, H. Zhu, B. Jones, and Y. J. Guo, “Suppression of cross-band scattering in multiband antenna arrays,” *IEEE Transactions on Antennas and Propagation*, vol. 67, no. 4, pp. 2379–2389, 2019.
- [37] R. Li, Y. Zhang, B. Xiao, and Y. Cui, “An aperture-shared dual-band antenna array based on upstanding dipoles,” *IEEE Antennas and Wireless Propagation Letters*, vol. 22, no. 11, pp. 2675–2679, 2023.
- [38] H.-H. Sun, H. Zhu, C. Ding, B. Jones, and Y. J. Guo, “Scattering suppression in a 4G and 5G base station antenna array using spiral chokes,” *IEEE Antennas and Wireless Propagation Letters*, vol. 19, no. 10, pp. 1818–1822, 2020.
- [39] D. He, Q. Yu, Y. Chen, and S. Yang, “Dual-band shared-aperture base station antenna array with electromagnetic transparent antenna elements,” *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 9, pp. 5596–5606, 2021.

- [40] Y. Wang, H. Su, R. C. Dai, L. T. Chen, L. H. Ye, and X. Y. Zhang, "A systematic approach to design spatial filtenna for aperture-shared base-station array," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 10, pp. 7792–7803, 2023.
- [41] L. Feng, H. Li, C. Wu, *et al.*, "Radiative deep-subwavelength frequency-selective surface-based cross-band decoupling method for base station antenna," *IEEE Transactions on Antennas and Propagation*, vol. 73, no. 2, pp. 771–781, 2025.
- [42] S. J. Yang and X. Y. Zhang, "Frequency selective surface-based dual-band dual-polarized high-gain antenna," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 3, pp. 1663–1671, 2022.
- [43] C. Chen, "A dual-band circularly polarized shared-aperture antenna for 1U cubesat applications," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 5, pp. 3818–3823, 2022.
- [44] J. H. Zhai, Y. F. Cao, Q. Xue, and W. Che, "Cross-band decoupling method for dual-band aperture-shared antenna array using metasurfaces," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 2, pp. 2001–2006, 2024.
- [45] Y. Fan, Y. Cheng, and Y. Dong, "A wideband shared-aperture dual-band base station antenna array based on inhomogeneous metasurface," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 2, pp. 463–467, 2024.
- [46] Y. Zhang and X. Chen, "Design of shared-aperture multi-band MIMO antenna array by scattering minimization," *IEEE Transactions on Antennas and Propagation*, pp. 1–1, 2025.
- [47] H.-H. Sun, B. Jones, Y. J. Guo, and Y. H. Lee, "Suppression of cross-band scattering in interleaved dual-band cellular base-station antenna arrays," *IEEE Access*, vol. 8, pp. 222 486–222 495, 2020.
- [48] S. Chainon, G. Coquille, A. Hilary, T. Julien, J. Plet, and J. Wang, "Multi-frequency antenna," *WO2015101138A1, Jul*, vol. 9, 2015.
- [49] Y. Zhu, Y. Chen, and S. Yang, "Helical torsion coaxial cable for dual-band shared-aperture antenna array decoupling," *IEEE Transactions on Antennas and Propagation*, vol. 68, no. 8, pp. 6128–6135, 2020.
- [50] J. Jiang and Q.-X. Chu, "Broadband suppression of common-mode current interference in dual-band base station antenna arrays by loading an LC circuit," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 1, pp. 224–228, 2024.

- [51] Y.-L. Chang and Q.-X. Chu, "Ferrite-loaded dual-polarized antenna for decoupling of multiband multiarray antennas," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 11, pp. 7419–7426, 2021.
- [52] M. Li, R. Wang, J. M. Yasir, and L. Jiang, "A miniaturized dual-band dual-polarized band-notched slot antenna array with high isolation for base station applications," *IEEE Transactions on Antennas and Propagation*, vol. 68, no. 2, pp. 795–804, 2020.
- [53] J.-F. Qian, S. Gao, B. Sanz-Izquierdo, H. Wang, H. Zhou, and H. Xu, "Mutual coupling suppression between two closely placed patch antennas using higher-order modes," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 6, pp. 4686–4694, 2023.
- [54] Y. Zhang, X. Y. Zhang, L.-H. Ye, and Y.-M. Pan, "Dual-band base station array using filtering antenna elements for mutual coupling suppression," *IEEE Transactions on Antennas and Propagation*, vol. 64, no. 8, pp. 3423–3430, 2016.
- [55] J. Guo, F. Liu, L. Zhao, G.-L. Huang, Y. Li, and Y. Yin, "Isolation improvement of two tightly coupled antennas operating in adjacent frequency bands using filtering structures," *IEEE Open Journal of Antennas and Propagation*, vol. 1, pp. 207–214, 2020.
- [56] X. Liu, K. Ning, S. Xue, L. Ge, K. W. Leung, and J.-F. Mao, "Printed filtering dipole antenna with compact size and high selectivity," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 3, pp. 2355–2367, 2024.
- [57] C. F. Ding, X. Y. Zhang, Y. Zhang, Y. M. Pan, and Q. Xue, "Compact broadband dual-polarized filtering dipole antenna with high selectivity for base-station applications," *IEEE Transactions on Antennas and Propagation*, vol. 66, no. 11, pp. 5747–5756, 2018.
- [58] W. Duan, Y. F. Cao, Y.-M. Pan, Z. X. Chen, and X. Y. Zhang, "Compact dual-band dual-polarized base-station antenna array with a small frequency ratio using filtering elements," *IEEE Access*, vol. 7, pp. 127 800–127 808, 2019.
- [59] S. J. Yang, W. Duan, Y. Y. Liu, H. Ye, H. Yang, and X. Y. Zhang, "Compact dual-band base-station antenna using filtering elements," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 8, pp. 7106–7111, 2022.
- [60] Y. Liu, S. Wang, N. Li, J.-B. Wang, and J. Zhao, "A compact dual-band dual-polarized antenna with filtering structures for Sub-6 GHz base station applications," *IEEE Antennas and Wireless Propagation Letters*, vol. 17, no. 10, pp. 1764–1768, 2018.

- [61] Q. X. Lai, Y. M. Pan, and S. Y. Zheng, "A self-decoupling method for mimo antenna array using characteristic mode of ground plane," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 3, pp. 2126–2135, 2023.
- [62] Q. Liu, L. Zhu, and W.-J. Lu, "Decoupling of closely spaced square patch antennas using short-circuited microstrip line," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 2, pp. 728–732, 2024.
- [63] Y.-N. Chen, C. Ding, H. Zhu, and Y. Liu, "A $\pm 45^\circ$ -polarized antenna system with four isolated channels for in-band full-duplex (IBFD)," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 4, pp. 3000–3010, 2023.
- [64] Y.-N. Chen, C. Ding, H. Zhu, *et al.*, "A four-port shared-aperture in-band full-duplex antenna array based on a novel common-mode and differential-mode combination method," *IEEE Transactions on Antennas and Propagation*, vol. 73, no. 9, pp. 6973–6978, 2025.
- [65] X. Wang, C. Ding, G. Zhao, S. Li, Y. Chen, and H. Sun, "Differential and symmetrical decoupling network for differentially fed antennas," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 3, pp. 1129–1133, 2024.
- [66] X. Wang, C. Ding, G. Zhao, C. Teng, S. Li, and H. Sun, "Compact dual-polarized self-decoupling shared-aperture antenna array for 6G systems," *IEEE Antennas and Wireless Propagation Letters*, pp. 1–5, 2025.
- [67] R. C. Dai, J. H. Ou, H. Su, and X. Y. Zhang, "A simple-structured dual-polarized staggered array with enhanced isolation at E- and H-planes," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 5, pp. 4664–4669, 2024.
- [68] B. Liu, X. Chen, J. Tang, A. Zhang, and A. A. Kishk, "Co- and cross-polarization decoupling structure with polarization rotation property between linearly polarized dipole antennas with application to decoupling of circularly polarized antennas," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 1, pp. 702–707, 2022.
- [69] Y. Da, X. Chen, and A. A. Kishk, "In-band mutual coupling suppression in dual-band shared-aperture base station arrays using dielectric block loading," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 10, pp. 9270–9281, 2022.
- [70] X. L. He and Y. Zhang, "A polarization convert reflective surface-based wideband antenna array decoupling structure," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 2, pp. 883–887, 2024.

- [71] Y. Xu, R. Li, Z.-X. Liu, Y. Hu, C. Wang, and Y. Liu, “Dual-polarized crossed-dipole base station antenna array with high-isolation and wide-angle scanning,” *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 11, pp. 8394–8403, 2024.
- [72] Z. Chen, D. Lu, M.-C. Tang, *et al.*, “Dipole antenna arrays with inherently high port isolation levels,” *IEEE Antennas and Wireless Propagation Letters*, vol. 24, no. 8, pp. 2512–2516, 2025.
- [73] S.-Y. Sun, C. Ding, W. Jiang, and Y. J. Guo, “Simultaneous suppression of cross-band scattering and coupling between closely spaced dual-band dual-polarized antennas,” *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 8, pp. 6423–6434, 2023.
- [74] Y. Li and Q.-X. Chu, “Dual-band base station antenna array using the low-band antenna as parasitic decoupler,” *IEEE Antennas and Wireless Propagation Letters*, vol. 21, no. 7, pp. 1308–1312, 2022.
- [75] Y. Li and Q.-X. Chu, “Self-decoupled dual-band shared-aperture base station antenna array,” *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 7, pp. 6024–6029, 2022.
- [76] Y. Zhu, Y. Chen, and S. Yang, “Decoupling and low-profile design of dual-band dual-polarized base station antennas using frequency-selective surface,” *IEEE Transactions on Antennas and Propagation*, vol. 67, no. 8, pp. 5272–5281, 2019.
- [77] Y. Chen, J. Zhao, and S. Yang, “A novel stacked antenna configuration and its applications in dual-band shared-aperture base station antenna array designs,” *IEEE Transactions on Antennas and Propagation*, vol. 67, no. 12, pp. 7234–7241, 2019.
- [78] Y. Zhu, Y. Chen, and S. Yang, “Integration of 5G rectangular MIMO antenna array and GSM antenna for dual-band base station applications,” *IEEE Access*, vol. 8, pp. 63 175–63 187, 2020.
- [79] F. Du, Y. Cao, X. Xiu, W. Che, and Q. Xue, “Compact dual-band shared-aperture base-station antenna array using transparent wideband absorbing metasurface and radiator-reused scheme,” *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 3, pp. 2502–2512, 2024.
- [80] P. Du, Y. Cao, W. Che, and Q. Xue, “Low-profile dual-band dual-polarized aperture-shared antenna array using bifunctional metasurface,” *IEEE Transactions on Antennas and Propagation*, vol. 73, no. 7, pp. 4274–4285, 2025.

- [81] D. Zhang and Q. Wu, "A shared-aperture antenna with high-aperture reuse efficiency and high isolation based on the quasi-yagi structure," *IEEE Transactions on Antennas and Propagation*, vol. 71, no. 11, pp. 8504–8513, 2023.
- [82] L. Ge, Y. Wang, M. Du, W. Liu, and Y. Zhao, "A dual-band dual-polarized base station antenna array with isolation enhancement," *IEEE Open Journal of Antennas and Propagation*, vol. 4, pp. 871–877, 2023.
- [83] Y. Zhu, Y. Chen, and S. Yang, "Cross-band mutual coupling reduction in dual-band base-station antennas with a novel grid frequency selective surface," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 12, pp. 8991–8996, 2021.
- [84] R. Chao Dai, S. Sun, H. Su, and X. Y. Zhang, "Analysis and design of dual-modal filtenna for dismountable aperture-shared array," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 6, pp. 4806–4817, 2024.
- [85] Y. Li and Q.-X. Chu, "Coplanar dual-band base station antenna array using concept of cavity-backed antennas," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 11, pp. 7343–7354, 2021.
- [86] Y. Li and Q.-X. Chu, "Shared-radiator design of dual-band coplanar base station antenna array using cavity-backed slots," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 12, pp. 8985–8990, 2021.
- [87] Y. He, S. Wang, G. Wei, C. Ding, and Y. Jay Guo, "A new buffering scheme for shared-aperture dual-band base station antenna array utilizing tight-coupling concept," *IEEE Transactions on Antennas and Propagation*, vol. 73, no. 6, pp. 3492–3502, 2025.
- [88] X. Wang, C. Ding, S. Li, *et al.*, "A low-profile dual-band self-decoupled shared-aperture patch antenna array for 6G beamforming," *IEEE Transactions on Antennas and Propagation*, 2025.
- [89] H. Huang, Y. Liu, and S. Gong, "A novel dual-broadband and dual-polarized antenna for 2G/3G/LTE base stations," *IEEE Transactions on Antennas and Propagation*, vol. 64, no. 9, pp. 4113–4118, 2016.
- [90] R. Wu and Q.-X. Chu, "A compact, dual-polarized multiband array for 2G/3G/4G base stations," *IEEE Transactions on Antennas and Propagation*, vol. 67, no. 4, pp. 2298–2304, 2019.

- [91] Y. He, C. Ding, C. Chang, G. Wei, and Y. J. Guo, "A bowl-shaped filtering antenna with wideband cross-band scattering mitigation for dual-band base stations," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 8, pp. 6723–6728, 2024.
- [92] X.-Y. Cheng, C. Ding, and R. W. Ziolkowski, "Dual-band shared-aperture dielectric resonator antenna (DRA) with suppressed cross-band interactions," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 7, pp. 5694–5704, 2024.
- [93] Y.-L. Chang and Q.-X. Chu, "Suppression of cross-band coupling interference in tri-band shared-aperture base station antenna," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 6, pp. 4200–4214, 2022.
- [94] X. Lu, Y. Chen, S. Guo, and S. Yang, "An electromagnetic-transparent cascade comb dipole antenna for multi-band shared-aperture base station antenna array," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 4, pp. 2750–2759, 2022.
- [95] W. Niu, B. Sun, and X. Huang, "A filtering and electromagnetic-transparent antenna for triple-band aperture-shared base station antenna array," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 1, pp. 244–248, 2024.
- [96] D. He, Y. Chen, and S. Yang, "A low-profile triple-band shared-aperture antenna array for 5G base station applications," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 4, pp. 2732–2739, 2022.
- [97] Y. Li, W. Yang, Q. Xue, and W. Che, "A miniaturized dual-polarized tri-band antenna based on multimode and stacked scheme for multiband aperture-shared base-station applications," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 5, pp. 4647–4652, 2024.
- [98] H. Huang, X. Li, and Y. Liu, "A way to improve mutual isolation and radiation pattern of triple-band antenna," *IEEE Antennas and Wireless Propagation Letters*, vol. 21, no. 7, pp. 1433–1436, 2022.
- [99] Y. Sun, J. Zhang, P. Mei, S. Luo, W. Fu, and S. Zhang, "Tri-band dual-polarized shared-aperture antenna arrays with wide-angle scanning and low profile for 5G base stations," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 3, pp. 2455–2467, 2024.
- [100] J.-X. Chen, X.-F. Wang, L.-L. Yang, X.-R. Wang, X.-F. Gu, and W.-W. Yang, "Asymmetrical low-RCS FSS-based antenna for tri-band shared-aperture array," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 5, pp. 1563–1567, 2024.

- [101] P. Liu, F. Jia, Y. Zhang, G. Su, Q. Wang, and X. Y. Zhang, "Dual-polarized dipole antenna with dual-band spatial filtering response for aperture-shared triband base station array application," *IEEE Antennas and Wireless Propagation Letters*, vol. 22, no. 12, pp. 3057–3061, 2023.
- [102] G.-N. Zhou, B.-H. Sun, Q.-Y. Liang, S.-T. Wu, Y.-H. Yang, and Y.-M. Cai, "Triband dual-polarized shared-aperture antenna for 2G/3G/4G/5G base station applications," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 1, pp. 97–108, 2021.
- [103] Y. Jia, H. Zhai, J. Yin, Y. Wang, and Y. Liu, "A quadruple-band shared-aperture antenna array with multiband radiation pattern restorations," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 10, pp. 7722–7735, 2024.
- [104] J. Yin, B. Wu, X.-Z. Song, Y.-T. Zhao, and Y. Liu, "Design of wide-band and multi-band shared aperture array with multi-order response," *IEEE Transactions on Antennas and Propagation*, pp. 1–1, 2025.
- [105] L. Wei, W. Yang, Q. Xue, and W. Che, "Aperture-shared design of millimeter-wave phased array and Sub-6 GHz antenna using diversified metasurface structures," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 8, pp. 6362–6373, 2024.
- [106] X.-H. Ding, W.-W. Yang, H. Tang, L. Guo, and J.-X. Chen, "A dual-band shared-aperture antenna for microwave and millimeter-wave applications in 5G wireless communication," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 12, pp. 12 299–12 304, 2022.
- [107] W.-W. Yang, Q.-H. Zhang, X. Fan, X. Geng, W. Qin, and J.-X. Chen, "A Sub-6 GHz/millimeter-wave dual-wideband shared-aperture antenna by reusing the mushroom structure," *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 10, pp. 8010–8015, 2024.
- [108] J. Yin, X. Li, S. Ni, J. Zhang, and Q. Wang, "Electromagnetic-transparent-structure-based quad-band shared-aperture antenna with multibeam in millimeter wave," *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 12, pp. 4124–4128, 2024.
- [109] H. Sheng and Z. N. Chen, "Radiation pattern improvement of cross-band dipoles using inductive-loading mode-suppression method," *IEEE Transactions on Antennas and Propagation*, vol. 67, no. 5, pp. 3467–3471, 2019.

- [110] Q. Wu, W. Su, Z. Li, and D. Su, "Reduction in out-of-band antenna coupling using characteristic mode analysis," *IEEE Transactions on Antennas and Propagation*, vol. 64, no. 7, pp. 2732–2742, 2016.
- [111] Q.-X. Chu, Y.-L. Chang, and J.-P. Li, "Crisscross-shaped $\pm 45^\circ$ dual-polarized antenna with enhanced bandwidth for base stations," *IEEE Transactions on Antennas and Propagation*, vol. 69, no. 4, pp. 2341–2346, 2021.
- [112] R. Wu and Q.-X. Chu, "Resonator-loaded broadband antenna for LTE700/GSM850/GSM900 base stations," *IEEE Antennas and Wireless Propagation Letters*, vol. 16, pp. 501–504, 2017.
- [113] A. Erentok, P. Luljak, and R. Ziolkowski, "Characterization of a volumetric metamaterial realization of an artificial magnetic conductor for antenna applications," *IEEE Transactions on Antennas and Propagation*, vol. 53, no. 1, pp. 160–172, 2005.
- [114] C.-C. Lin, P. Jin, and R. W. Ziolkowski, "Single, dual and tri-band-notched ultrawide-band (UWB) antennas using capacitively loaded loop (CLL) resonators," *IEEE Transactions on Antennas and Propagation*, vol. 60, no. 1, pp. 102–109, 2012.
- [115] L.-M. Si, Q.-L. Zhang, W.-D. Hu, *et al.*, "A uniplanar triple-band dipole antenna using complementary capacitively loaded loop," *IEEE Antennas and Wireless Propagation Letters*, vol. 14, pp. 743–746, 2015.
- [116] F. H. Lin and Z. N. Chen, "A method of suppressing higher order modes for improving radiation performance of metasurface multiport antennas using characteristic mode analysis," *IEEE Transactions on Antennas and Propagation*, vol. 66, no. 4, pp. 1894–1902, 2018.
- [117] *6G spectrum - enabling the future mobile life beyond 2030*, *Ericsson.com*, [Online]. Available: <https://www.ericsson.com/en/reports-and-papers/white-papers/6g-spectrum-enabling-the-future-mobile-life-beyond-2030>, May 2024.
- [118] *Spectrum for 6G explained*, *Nokia.com*, [Online]. Available: <https://www.nokia.com/6g/spectrum-for-6G-explained/>, Jun. 2023.
- [119] Y. J. Guo, C. A. Guo, M. Li, and M. Latva-aho, "Antenna technologies for 6G – advances and challenges," *IEEE Trans. Antennas Propag.*, 2025, early access.
- [120] M. Wang and C. H. Chan, "Dual-polarized, low-profile dipole-patch array for wide bandwidth applications," *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 9, pp. 8030–8039, 2022.

- [121] X. Lu, F. Wu, Z. H. Jiang, Z. Jiang, K.-F. Tong, and W. Hong, “A broadband dual-polarized me-dipole antenna array for millimeter-wave applications,” *IEEE Antennas and Wireless Propagation Letters*, vol. 23, no. 11, pp. 3456–3460, 2024.