

# Reconfigurable Metasurface for Near-Field Efficient Wireless Power Transfer System



Wang Yunfeng<sup>1</sup>, Xu Zhiyu<sup>2</sup>, Zhang Hongtao<sup>1</sup>

(1. Key Lab of Universal Wireless Communications, Ministry of Education of China, Beijing University of Posts and Telecommunications, Beijing 100876, China;

2. ZTE Corporation, Shenzhen 518057, China)

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**Abstract:** Wireless power transfer (WPT) offers a promising solution for powering distributed devices in smart cities, significantly simplifying system complexity and accelerating urban automation. In traditional WPT systems, far-field schemes are costly, energy-intensive, and inefficient, while near-field transmission offers higher efficiency but suffers from limited distance. To address the limited transmission distance of near-field WPT, this paper derives an electromagnetic model for WPT using reflective metasurfaces and presents the design of a high-performance 3-bit reconfigurable intelligent surface (RIS). Based on theoretical models, we investigate the effects of phase quantization and array size on focusing efficiency. Simulation results guide the subsequent fabrication of the RIS prototype. Experimental measurements on the fabricated prototype validate the beam-focusing effect and evaluate the overall WPT efficiency.

**Keywords:** reconfigurable intelligent surface; near-field beam focusing; wireless power transfer

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## 1 Introduction

Wireless power transfer (WPT) enables the transfer of electrical energy from the power source to a load through electromagnetic waves, without physical wire connections. It offers features such as on-demand charging, non-contact operation, safety, reliability, and simultaneous charging of multiple devices, eliminating cables and greatly enhancing convenience and flexibility.

WPT technology has the potential to revolutionize traditional charging methods, significantly improving the flexibility, convenience, and reliability of equipment such as electric vehicles and smart power systems. It can drive innovation across sectors including electric vehicles, smart grids, medical devices, home appliances, and aerospace. At the same time, it ensures secure and stable grid operation and enhances energy efficiency, offering vast market potential.

Unlike electromagnetic induction charging, which only supports short-range power transfer, WPT based on radio frequency (RF) energy harvesting can operate over longer distances, providing greater flexibility for power-driven devices<sup>[1-2]</sup>. A complete WPT system requires determining the

receiver's location and adjusting the transmission system accordingly, followed by signal emission, transmission, and reception. Current research on transmission systems mainly focuses on three categories: far-field RF transmitters<sup>[3-4]</sup>, fixed antennas<sup>[5-6]</sup>, and energy transfer assisted by metasurfaces<sup>[7-9]</sup>. However, far-field RF power transmission suffers from low efficiency and limited application scenarios. It is heavily affected by path loss, and the electric field distribution is diffuse, resulting in low efficiency. It can only power low-energy devices through specialized collection and processing at the receiver. In contrast, near-field radiation offers higher transmission efficiency, better meeting the functional needs of indoor devices<sup>[10-12]</sup>. The focusing efficiency of metasurfaces depends primarily on phase control accuracy, the reflection loss coefficient, and phase compensation precision. Phase quantization accuracy is limited by the unit resonator component design. Current adjustable metasurface prototypes commonly use PIN diodes<sup>[13-15]</sup> and varactor diodes<sup>[16]</sup> for phase tuning, but PIN-diode-based designs are typically limited to discrete phase states, while varactor-diode-based designs may face a tradeoff between the achievable phase-tuning range and tuning loss, requiring further improvement in transmission efficiency.

Over the past two decades, research on metamaterials and

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metasurfaces has experienced explosive growth<sup>[17]</sup>. Metasurfaces are artificial materials designed to achieve electromagnetic phenomena that are either unavailable or difficult to obtain in nature. With their wave manipulation characteristics, they are implemented as an array of specially arranged scatterers in a two-dimensional periodic layout<sup>[17–18]</sup>. Subsequently, the theory and applications of metasurfaces have been developed in tandem within and beyond the antenna and propagation communities<sup>[19–20]</sup>. Groundbreaking work in the field of gradient metasurfaces<sup>[21]</sup>, along with the manipulation of amplitude and phase of incident wavefronts using nanostructures<sup>[22]</sup>, has sparked significant interest in the control of light wavefronts with metasurface devices. The engineering concepts of phased arrays, reflective beams, and emission arrays have further highlighted the scientific and commercial value of these novel devices, increasing interest in metamaterials and metasurfaces<sup>[23–25]</sup>.

The transmission of microwave energy through antenna arrays<sup>[26]</sup> offers high efficiency, but the supporting equipment is complex and costly due to the use of traditional antennas. Energy focusing using a fixed metasurface<sup>[10]</sup> shows good performance at a preset focusing position, but the fixed unit design makes it difficult to achieve consistent and efficient energy transfer when the focusing position changes. By using PIN diodes, 1-bit and 2-bit reflective WPT reconfigurable intelligent surface (RIS) can be achieved<sup>[11, 27]</sup>. However, PIN diodes introduce design complexity and limit control accuracy, which reduces energy transfer efficiency. This paper proposes a 3-bit RIS unit based on low-loss varactor diodes, which significantly reduces the antenna cost and control complexity of the approach in Ref. [26] and improves WPT efficiency compared to lower-bit phase quantization in Refs. [11] and [27]. Compared to Ref. [10], it can achieve consistent energy transfer efficiency at different focusing positions. Our research on quantization accuracy and array size has guided the design of an efficient RIS for energy transmission.

In this paper, we propose a feasible and efficient WPT strategy that utilizes reflective RIS to achieve near-field energy-focused transmission. The rest of this paper is organized as follows. We discuss the efficient near-field WPT strategy using RIS and establish a transmission efficiency model in Section 2. Section 3 presents the design of the 3-bit RIS and simulations of transmission efficiency based on it. We also fabricate prototypes and present actual measurement results in Section 4. Finally, Section 5 concludes the paper.

## 2 RIS-Aided Near-Field WPT Framework

In WPT systems, a key metric is energy transmission efficiency, which is significantly influenced by the power transmission beam. This paper proposes an efficient power beam-forming strategy that integrates the location information of the device to be charged and the RIS. The strategy automatically calculates and allocates high-gain near-field energy beams to

meet the energy transmission requirements across different scenarios. Fig. 1 illustrates the application scenario of an RIS in an indoor WPT system.

### 2.1 Reflective RIS and Numerical Model

In recent years, there has been growing interest in developing dynamic metasurfaces for constructing advanced reconfigurable devices. The concept of RIS has been introduced, enabling real-time control over electromagnetic wave propagation. Similar to digital circuit technology, the electromagnetic responses of metasurface elements are represented by discrete digital codes. By modifying these code sequences, functional control of both far-field and near-field electromagnetic behavior can be achieved.

A metasurface can be defined as an array of polarizable elements that induce discontinuities in electric and magnetic fields on both sides of the surface. The macroscopic description of metasurfaces aims to express these field discontinuities as functions of the surface-averaged electric and magnetic polarization densities. This macroscopic approach replaces the detailed spatial distribution of individual elements with analytical relations between the electric and magnetic fields on either side of the surface.

The homogenization of the surface treats it as having continuous electric and magnetic polarization densities. This homogenization incorporates the unit cell's size, shape, and material properties directly into these analytical relations. This provides essential design guidance for metasurface elements.

The analytical relations linking the fields on both sides of the surface to the homogeneous polarization densities are known as the Generalized Sheet Transition Conditions (GSTCs)<sup>[28]</sup>. This macroscopic description allows the polarization densities to be represented using magnetizability tensors instead of polarizability, relying only on the averaged fields. These magnetizability tensors, much like dielectric permittivity and magnetic permeability for volumetric metamaterials,

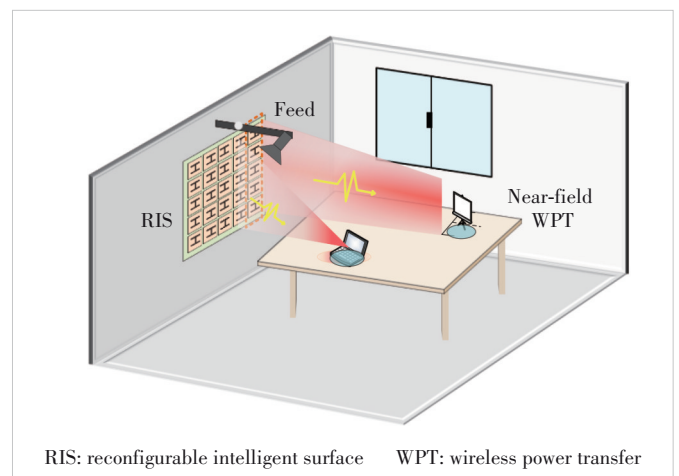


Figure 1. Application scenarios of RIS in a WPT system

act as effective parameters for the metasurface. They depend on the unit cell's topology, geometry, spacing, material properties, and wavelength. The electric ( $\mathbf{P}$ ) and magnetic ( $\mathbf{M}$ ) surface polarization densities can be formulated as follows:

$$\mathbf{M}(x, y) = \sqrt{\varepsilon/\mu} \bar{\chi}_{me}(x, y) \mathbf{E}_{av}(x, y) + \bar{\chi}_{mm}(x, y) \mathbf{H}_{av}(x, y) \quad (1),$$

$$\mathbf{P}(x, y) = \varepsilon \bar{\chi}_{ee}(x, y) \mathbf{E}_{av}(x, y) + \sqrt{\mu\varepsilon} \bar{\chi}_{em}(x, y) \mathbf{H}_{av}(x, y) \quad (2),$$

where  $\varepsilon$  and  $\mu$  are vacuum permittivity and permeability, respectively;  $\bar{\chi}_{ab}(x, y)$ , for  $a = \{e, m\}$  and  $b = \{e, m\}$ , are the electric-electric, electric-magnetic, magnetic-electric, and magnetic-magnetic surface-averaged (or simply surface) susceptibility dyadics, which are the macroscopic quantities of interest for a simple but accurate synthesis and analysis of metasurfaces;  $\mathbf{E}_{av}$  and  $\mathbf{H}_{av}$  are the surface-averaged electric and magnetic fields, where the average is computed over a small surface area of the order of the wavelength (including one or more unit cells depending on the setup).

Once the average reflection coefficient of the metasurface is determined, we incorporate analytical theories of incident, reflected, and surface electromagnetic waves to meet various performance requirements. This approach facilitates the derivation of a general susceptibility function, which helps optimize the design parameters of the unit cells<sup>[29]</sup>. These analytical frameworks enable fine-tuning of the metasurface characteristics to achieve desired functionalities, ensuring that it performs effectively in different scenarios. The surface electromagnetic field results from the electromagnetic discontinuity introduced by the metasurface structure. Its analytical expression can be derived from the Generalized GSTCs. These conditions relate the fields on either side of the metasurface to the surface polarization densities, thereby providing a framework for understanding how the metasurface interacts with incident electromagnetic waves and influences the fields in its vicinity.

$$\begin{aligned} E_{x,y}^{rt}(x, y) &\propto \mathbf{s}_{E,x,y}^{rt}(x, y) E_x^i(x, y) + \mathbf{s}_{E,x,y}^{rt}(x, y) E_y^i(x, y) + \\ &\mathbf{s}_{H,x,y}^{rt}(x, y) H_x^i(x, y) + \mathbf{s}_{H,x,y}^{rt}(x, y) H_y^i(x, y) \end{aligned} \quad (3),$$

$$\begin{aligned} H_{x,y}^{rt}(x, y) &\propto \xi_{E,x,y}^{rt}(x, y) E_x^i(x, y) + \xi_{E,x,y}^{rt}(x, y) E_y^i(x, y) + \\ &\xi_{H,x,y}^{rt}(x, y) H_x^i(x, y) + \xi_{H,x,y}^{rt}(x, y) H_y^i(x, y) \end{aligned} \quad (4),$$

where the superscripts i, r and t denote the incident, reflected, and transmitted components of the fields, respectively;  $E_{x,y}(x, y)$  and  $H_{x,y}(x, y)$  represent the tangential components of the electric and magnetic fields, respectively;  $\mathbf{s}_{\{E,H\},\{x,y\}}^{rt}(x, y)$  and  $\xi_{\{E,H\},\{x,y\}}^{rt}(x, y)$ , derived from a matrix that depends solely on the metasurface structure (specifically the surface magnetic susceptibility function), express each tangential component of

the surface electromagnetic fields on both sides of the metasurface as a linear combination of the tangential components of the incident field. In the case of perfect anomalous reflection, analyzing the reflective surface electromagnetic field provides valuable insights for the design of broadband metasurface structures, enabling precise control over the reflected waves and enhancing metasurface performance.

Under the assumption of perfect anomalous reflection, conditions  $E_x^t(x, y) = E_y^t(x, y) = 0$  and  $H_x^t(x, y) = H_y^t(x, y) = 0$  exist. The calculations for  $E_y^r(x, y)$  and  $H_x^r(x, y)$  are based on the conventional scenario. Here,  $\chi_{ee}^{yy}(x, y)$  and  $\chi_{mm}^{xx}(x, y)$  represent the surface-averaged electric and magnetic susceptibility functions. Due to  $\chi_{ee}^{yy}(x, y) \chi_{mm}^{xx}(x, y) = -4/k^2$ , the corresponding two constitutive relations in the reflected electromagnetic field are not linearly independent. This implies that multiple electromagnetic field transformations satisfy the constitutive relations. Considering a solution of the form  $H_x^r(x, y) = e(x, y) E_y^r(x, y)$ , where  $e(x, y)$  is a given function such that  $E_y^r(x, y)$  and  $H_x^r(x, y)$  satisfy Maxwell's equations, we obtain an explicit expression for the reflected electromagnetic field as a function of the incident electromagnetic field under these assumptions:

$$\begin{aligned} E_y^r(x, y) &= \frac{-\frac{j\omega\varepsilon}{2} \chi_{ee}^{yy}(x, y)}{e(x, y) + \frac{j\omega\varepsilon}{2} \chi_{ee}^{yy}(x, y)} E_y^i(x, y) - \\ &\frac{1}{e(x, y) + \frac{j\omega\varepsilon}{2} \chi_{ee}^{yy}(x, y)} H_x^i(x, y) \end{aligned} \quad (5),$$

$$\begin{aligned} H_x^r(x, y) &= \frac{-\frac{j\omega\varepsilon}{2} e(x, y) \chi_{ee}^{yy}(x, y)}{e(x, y) + \frac{j\omega\varepsilon}{2} \chi_{ee}^{yy}(x, y)} E_y^i(x, y) - \\ &\frac{e(x, y)}{e(x, y) + \frac{j\omega\varepsilon}{2} \chi_{ee}^{yy}(x, y)} H_x^i(x, y) \end{aligned} \quad (6).$$

From the above expressions, the reflected electric and magnetic fields are represented as linear combinations of the incident electric and magnetic fields. Generally, there is cross-coupling between the electromagnetic fields. For any given matrix, its pseudoinverse always exists and is unique. This means that for a given metasurface structure (represented by the surface magnetic susceptibility function and the incident field), one can always find the corresponding reflected and transmitted fields. Therefore, the metasurface structure can be reverse-engineered from the reflected field, providing guidance for unit cell design.

## 2.2 Near-Field WPT Strategy Using RIS

To achieve near-field energy focusing using a reflective metasurface, a numerical model is first established to compute the electric field  $E$  in the Fresnel region, assuming the metasurface consists of  $I \times J$  reflective units, each of size  $px \times py$ . The emitting source is located at point  $(x_f, y_f, z_f)$ , and the target focal point is at point  $(x_d, y_d, z_d)$ . Fig. 2 shows a schematic diagram of phase compensation for reflective focusing metasurfaces.

Assuming polarization along the  $y$ -axis in the model, the scattering  $E$ -field in the Fresnel region can be represented using the plane wave angular spectrum (PWAS) method<sup>[11]</sup>:

$$E_y(x, y, z) = \frac{j}{2} \exp(-jk_0 z) \cdot \sum_{i=0}^{I-1} \sum_{j=0}^{J-1} \left\{ E_t(x_i, y_j) \cdot \left[ C(t_2) - jS(t_2) \right] - \left[ C(t_1) - jS(t_1) \right] \right\} \cdot \left\{ \left[ C(t_2') - jS(t_2') \right] - \left[ C(t_1') - jS(t_1') \right] \right\} \quad (7)$$

where  $k_0 = 2\pi/\lambda$  is the wave number in free space,  $E_t(x_i, y_j)$  is the tangential  $E$ -field at the center of each unit, and  $C(u)$  and  $S(u)$  are the Fresnel integrals, which can be written as:

$$C(u) = \int_0^u \cos\left(\frac{\pi}{2} t^2\right) dt, \quad S(u) = \int_0^u \sin\left(\frac{\pi}{2} t^2\right) dt$$

$$t_1 = \sqrt{\frac{k_0}{\pi z}} \left( \frac{p_x}{2} + x - mp_x + \frac{(M-1)p_x}{2} \right)$$

$$t_2 = \sqrt{\frac{k_0}{\pi z}} \left( -\frac{p_x}{2} + x - mp_x + \frac{(M-1)p_x}{2} \right)$$

$$t_1' = \sqrt{\frac{k_0}{\pi z}} \left( \frac{p_y}{2} + y - np_y + \frac{(N-1)p_y}{2} \right)$$

$$t_2' = \sqrt{\frac{k_0}{\pi z}} \left( -\frac{p_y}{2} + y - np_y + \frac{(N-1)p_y}{2} \right) \quad (8)$$

The representation of each unit's  $E_t(x_i, y_j)$  is given as follows:

$$E_t(x_i, y_j) = A_{ij} \exp(j\varphi_{ij}) \cdot E_0(x_i, y_j) \times \exp\left(jk_0 \sqrt{(x_f - x_i)^2 + (y_f - y_j)^2 + z_f^2}\right) \quad (9)$$

where  $A_{ij}$  and  $\varphi_{ij}$  represent the reflection amplitude and phase shift of the  $ij$ -th unit cell, respectively;  $E_0(x_i, y_j)$  denotes the amplitude of the incident tangential electric field from the spatial source at the location of the  $n$ -th unit cell. To generate a focused spot at the desired position, the phase shifts of all units must align coherently at the target location.

According to electromagnetic wave propagation theory and

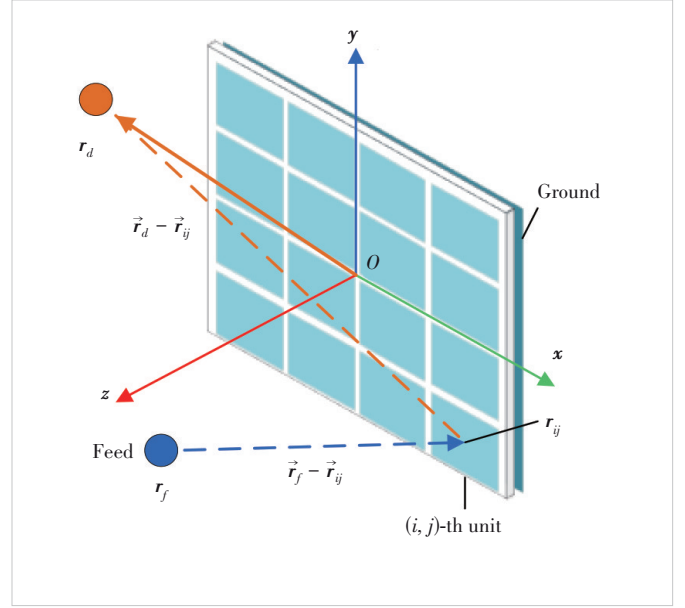


Figure 2. Schematic diagram of phase compensation for reflective focusing metasurfaces

the principle of coherent beam superposition, the phase equals the product of the wavenumber and the optical path. The optical path difference between each unit on the reflective focusing surface and the target focal point can be expressed as  $|\vec{r}_d - \vec{r}_{ij}|$ , as shown in Fig. 2. Assuming the reflective focusing surface serves as the reference phase plane, the phase distribution required to generate the focus is located behind the reference plane wave. Therefore, the phase distribution necessary for focusing in space can be expressed as:

$$\varphi_d(x_i, y_j) = \exp\left(-jk_0 |\vec{r}_d - \vec{r}_{ij}|\right) \quad (10)$$

Using conventional horn-fed excitation, the amplitude distribution on the reflective focusing surface is predetermined, as is the phase excitation distribution. When designing the reflective focusing surface, it is essential to account for the initial phase and amplitude distributions. Assuming the reflective focusing surface serves as the phase reference plane, the phase distribution of the incident feed on the metasurface can be expressed as:

$$\varphi_f(x_i, y_j) = \exp\left(jk_0 |\vec{r}_f - \vec{r}_{ij}|\right) \quad (11)$$

The required phase shift distribution for the reflective focusing surface is the difference between the focusing phase distribution necessary to direct the energy towards the focal point and the initial phase distribution of the incident feed on the metasurface. Thus, it can be expressed as:

$$\Delta\varphi_{ij}(x_i, y_j) = \varphi_d(x_i, y_j) - \varphi_f(x_i, y_j) \quad (12)$$

By setting the feed source position, the focal point, and the arrangement of all units on the metasurface, the phase compensation can be calculated through optical path differences and assigned to the metasurface elements. This enables energy synthesis at the desired focal point.

Using the above formulas and incorporating position information, a phase compensation matrix can be computed. The electric field amplitude and phase are then superimposed on the focusing surface to simulate the energy focusing effect of the reflective metasurface. Additionally, the impact of quantization bit depth and array aperture on focusing performance is preliminarily explored, providing insights for subsequent element and array design.

### 2.3 Quantization and Aperture Analysis

Focusing efficiency is a crucial metric that directly reflects the effectiveness of focusing-based WPT systems. In this section, the efficiency  $\eta$  is defined as the ratio of the total energy  $P_d$  within the focused area at the reference plane to the incident power  $P_f$  from the feed source on the metasurface:

$$\eta = \frac{P_d}{P_f} \quad (13).$$

The focal area is defined as the circular region on the focal plane that encompasses the point of maximum electric field intensity and the surrounding area with an intensity drop of 3 dB, as illustrated in Fig. 3. The total energy within the focal area  $P_d$  can be obtained by integrating the Poynting vector over this region. Similarly, the incident power is calculated by integrating the Poynting vector over the plane of the receiving horn antenna.

Existing research indicates that in near-field beam steering

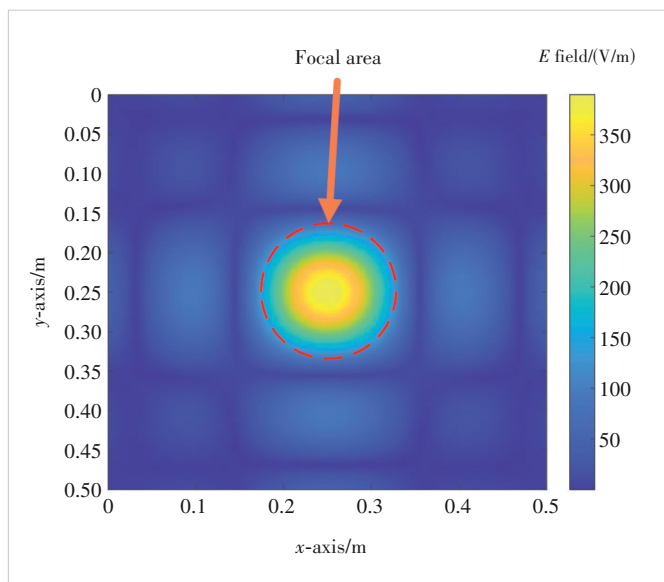


Figure 3. Electric field distribution of the metasurface focal area

scenarios, a 1-bit quantization scheme results in a gain reduction of approximately 3 dB<sup>[11]</sup>. Increasing the quantization precision with more bits can mitigate this gain loss. Thus, a higher phase quantization resolution improves the focusing efficiency of the reflective focusing metasurface. However, the control overhead introduced by multi-bit quantization is significant and will substantially increase system costs. Therefore, it is crucial to balance the number of quantization bits against the focusing efficiency. This section illustrates this issue via simulation.

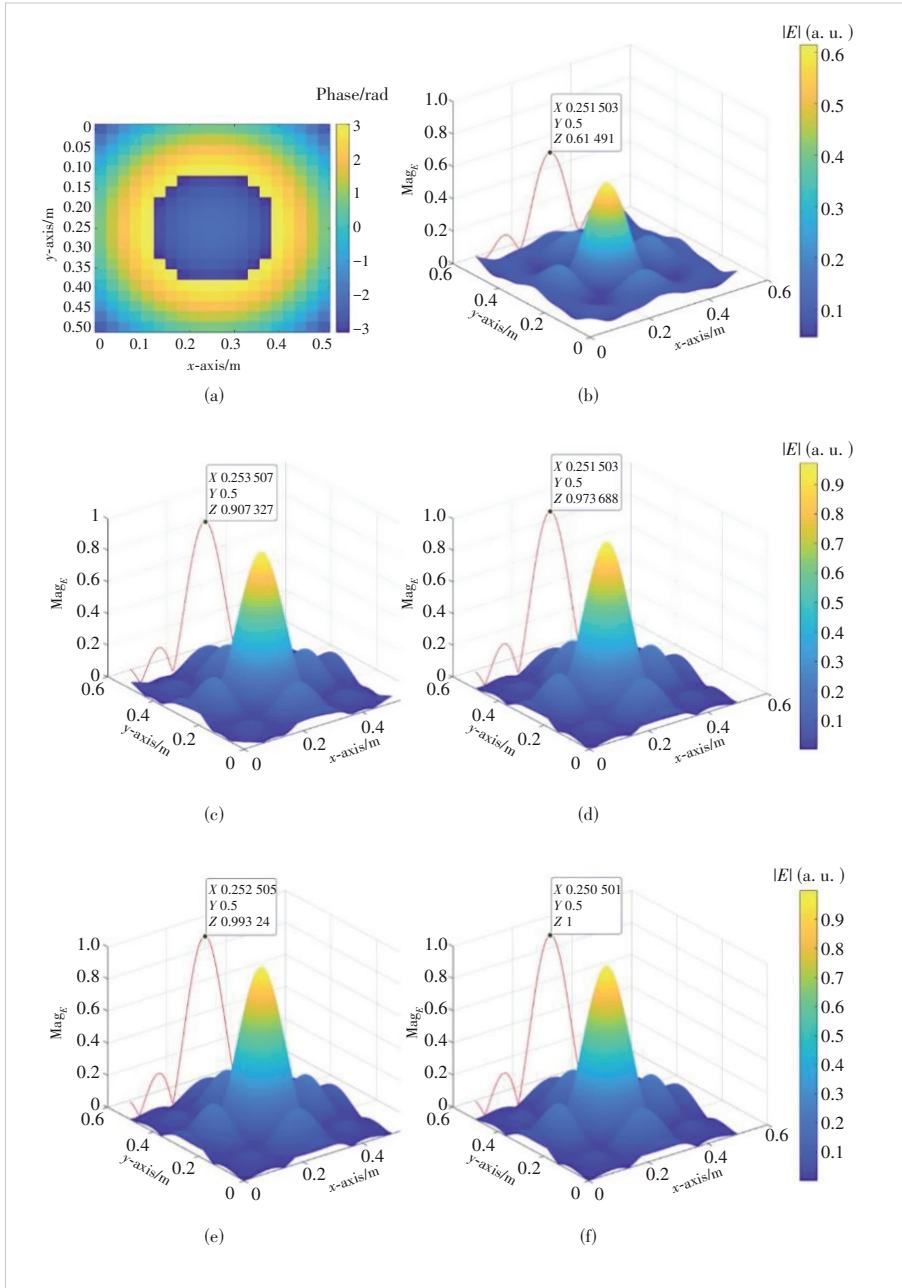
We set the unit size to 25 mm × 25 mm with an incident wave frequency of 5.8 GHz. A 20 × 20 array forms a 0.5 m × 0.5 m aperture, with the focal point fixed at the center, 1 m away from the metasurface. By calculating the compensation phases for each unit, a non-quantized compensation phase matrix is obtained. The matrix is then quantized to 1-bit, 2-bit, 3-bit, and 4-bit precision. The strongest electric field value at the focal point is compared for each quantization level (Fig. 4).

From Fig. 4, as the quantization precision increases from 1 bit to 4 bit, the electric field intensity at the focal center gradually rises (all normalized to the maximum under non-quantized control), but the rate of increase diminishes. This indicates a trend towards saturation. Specifically, increasing from 1 bit to 2 bit improves the focal intensity by 47.6%; from 2 bit to 3 bit, by 7.4%; and the 4-bit improvement is smaller. However, increasing the number of control bits significantly increases control link bandwidth requirements and system costs. After weighing performance against cost, this study adopts a 3-bit quantization scheme.

To determine the influence of array size on focusing performance, we simulate arrays of different sizes. Fig. 5 shows the simulated focusing effect. In the simulations, the array sizes are set to 10 × 10 units, 20 × 20 units, and 40 × 40 units, with dimensions of 0.25 m × 0.25 m, 0.5 m × 0.5 m, and 1 m × 1 m, respectively, focusing at distances of 1 m and 5 m from the metasurface center. Since the theoretical simulations do not account for path loss, the focused power can only be compared at the same distance. Non-quantized phase is used as the compensation phase, and the influence of the feeding path is ignored, assuming the reflected power of each unit is uniform across all array sizes.

From Fig. 5 and the focal area definition, we determine the focal spot sizes. At a focusing distance of 1 m, the focal spot sizes for the 10 × 10, 20 × 20, and 40 × 40 arrays are 19 cm × 19 cm, 9.5 cm × 9.5 cm, and 5.2 cm × 5.2 cm, respectively. At a focusing distance of 5 m, the focal spot size of the 10 × 10 array exceeds the size of the transmitting array, while those for the 20 × 20 and 40 × 40 arrays are 48 cm × 48 cm and 23.2 cm × 23.2 cm, respectively.

In practical WPT applications, the energy receiving area is typically fixed. Reducing the focal spot size increases the received power. However, increasing the array aperture multi-



**Figure 4. Phase compensation and normalized electric-field distributions: (a) non-quantized compensation phase distribution; (b) 1-bit quantization; (c) 2-bit quantization; (d) 3-bit quantization; (e) 4-bit quantization; (f) non-quantized electric-field distribution (reference)**

plies the number of metasurface units, leading to a significant rise in costs. Thus, it is essential to balance the final array size against the transmission distance, aperture size, and focusing efficiency.

From Fig. 5, doubling the array size significantly reduces the focal size. At closer distances, the large-aperture gain of larger arrays is evident. For example, at 1 m, the  $40 \times 40$  array has 300% more elements than the  $20 \times 20$  array. The focal size decreased from  $9.5 \text{ cm} \times 9.5 \text{ cm}$  to  $5.2 \text{ cm} \times 5.2 \text{ cm}$ ,

and the area decreased to 30% of its original size. At 5 m, the  $10 \times 10$  array focal spot exceeds the aperture size, while the  $20 \times 20$  and  $40 \times 40$  arrays produce focal spots of  $48 \text{ cm} \times 48 \text{ cm}$  and  $23.2 \text{ cm} \times 23.2 \text{ cm}$ , respectively. The focal spot size of the  $40 \times 40$  array decreases to 23% of that of the  $20 \times 20$  array. We consider the near-field and far-field boundaries for antennas. The boundary distance is given by:

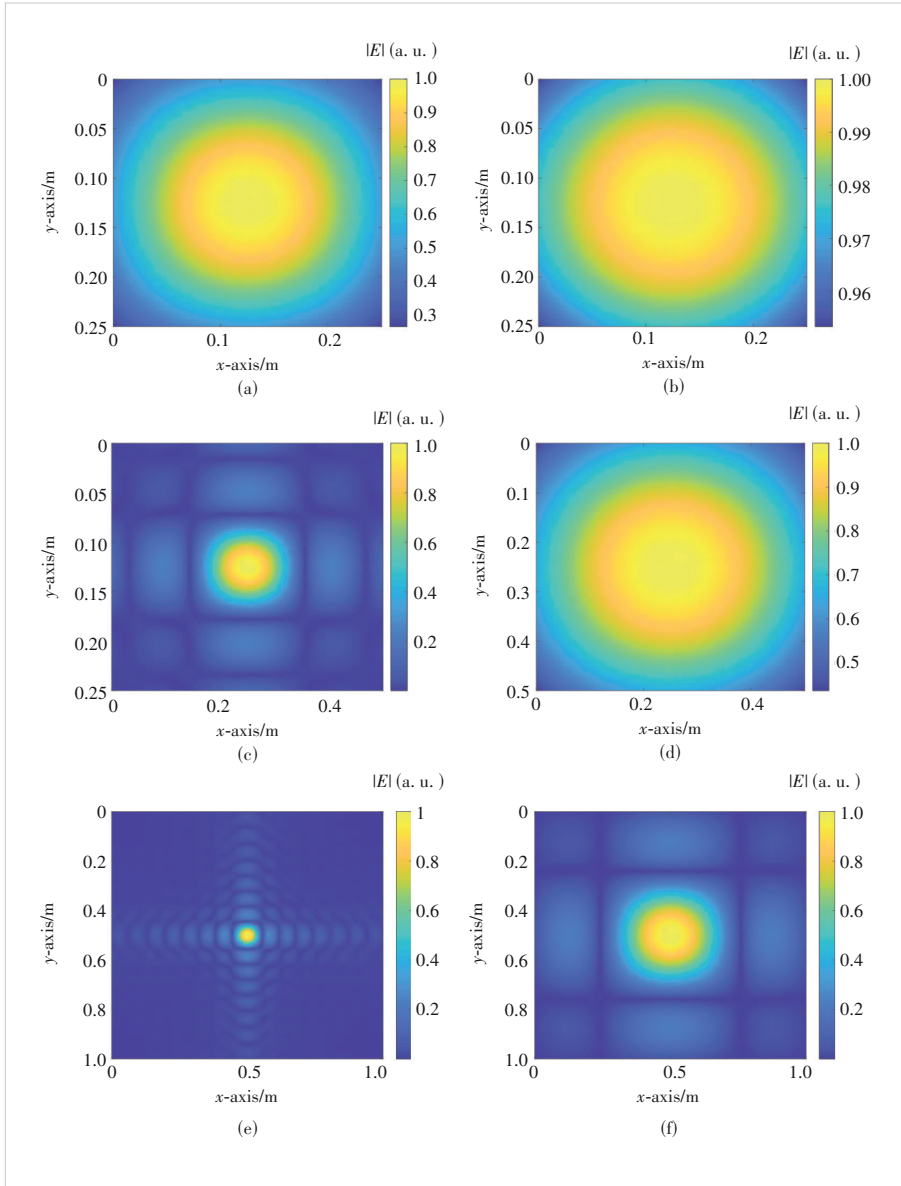
$$r_{\text{far}} \approx \frac{2D^2}{\lambda} \quad (15),$$

where  $D$  is the antenna aperture (in meters),  $\lambda$  is the wavelength (in meters), and  $r_{\text{far}}$  is the boundary distance between the far field and near field (in meters). Distances greater than this calculated value are far-field; distances less than this value are near-field. In the 5.8 GHz frequency band, the near-field range for the  $20 \times 20$  array is less than 9.7 m, while that for the  $40 \times 40$  array is less than 38.7 m. Based on the above data, within the 5 – 10 m near-field focusing range, the performance improvement of the  $40 \times 40$  array over the  $20 \times 20$  array is less than the increase in the number of units. Therefore, the  $20 \times 20$  array is a more practical choice.

### 3 Design of 3-Bit Reflective RIS and Simulation

#### 3.1 3-Bit Reflective RIS Design

Based on the theoretical analysis in Section 2.1 and the simulation results in Section 2.3, we designed a high-reflectivity, reconfigurable, focusing metasurface unit with a dog-bone-shaped patch for 5.8 GHz operation. Each unit consists of two dielectric substrate layers of different thicknesses and three copper metal patch layers of different shapes, which are bonded using prepreg materials. The patch is disconnected in the middle, and a varactor diode, specifically the MAVR-000120-1411 model produced by MACOM, is used to connect the two separated patches. The bottom layer consists of the DC bias line for the varactor diode and, through via connections, links to the top patches. The control terminal is connected via a feeding line, enabling adjustments to the input voltage at the control terminal, which



**Figure 5.** Normalized electric field distribution of different array sizes and focus distances: (a)  $10 \times 10$  array at (0, 0, 1 m); (b)  $10 \times 10$  array at (0, 0, 5 m); (c)  $20 \times 20$  array at (0, 0, 1 m); (d)  $20 \times 20$  array at (0, 0, 5 m); (e)  $40 \times 40$  array at (0, 0, 1 m); (f)  $40 \times 40$  array at (0, 0, 5 m)

ultimately changes the phase response of the unit. Fig. 6 shows the unit design. Table 1 lists the geometric parameters and material properties.

We used the full-wave electromagnetic simulation software ANSYS High-Frequency Structure Simulator (HFSS) to simulate the electromagnetic properties of the designed metasurface unit. For the varactor diode, three series lumped ports were used to model its resistor-inductor-capacitor (RLC) series equivalent circuit, as shown in Fig. 7, with boundary conditions assigned to its capacitance, resistance, and inductance. To determine the varactor diode parameters corresponding to 3-bit phase control, a Floquet port was employed for 2D

periodic S-parameter analysis. A parameter sweep was set up for the capacitance  $C_1$  in the RLC circuit, scanning from 0.15 pF to 0.95 pF.

The 3-bit control corresponds to eight states, with the phase response of each unit differing by  $45^\circ$  between adjacent states. The total phase shift across the eight states must be at least  $315^\circ$  to cover a  $360^\circ$  phase range. The reflection phase and amplitude curves for the incident polarization are obtained from the  $S_{11}$  data. Figs. 8 and 9 show the parameter curves for the eight capacitance values. Table 2 lists the capacitance, encoding, bias voltage, and phase for each state.

From the  $S_{11}$  phase curves, the phase difference between adjacent states is approximately  $45^\circ$ , with the maximum difference exceeding  $315^\circ$  at 5.8 GHz. Due to phase deviations in practical systems, achieving perfect phase intervals is challenging. Additionally, the limited control capability of the diode makes it difficult to meet large phase differences across a wide frequency band. Therefore, a phase deviation criterion is introduced as the evaluation standard for multi-bit unit design.

To quantify the phase error versus frequency, the equivalent bit number has been employed in previous works for programmable metasurface reflectors with discrete phase steps. Pereira et al. proposed a quality indicator for the phase response of discretely tunable reflect arrays as the equivalent bit number<sup>[30]</sup>. The equivalent bit number is defined as:

$$N_{\text{bit}} = \log_2 \left( \frac{360}{\sqrt{12} \sigma} \right) \quad (16),$$

where  $\sigma$  is the phase standard deviation (in degrees), calculated from the set of phase differences,  $\Delta\varphi$ , between different discrete states:

$$\sigma = \sqrt{\frac{\sum_{m=1}^M \Delta\varphi^3}{12 \times 360}} \quad (17).$$

The upper error limit for 3-bit control is  $16.25^\circ$ . After calculation, the standard phase deviation of the designed unit in

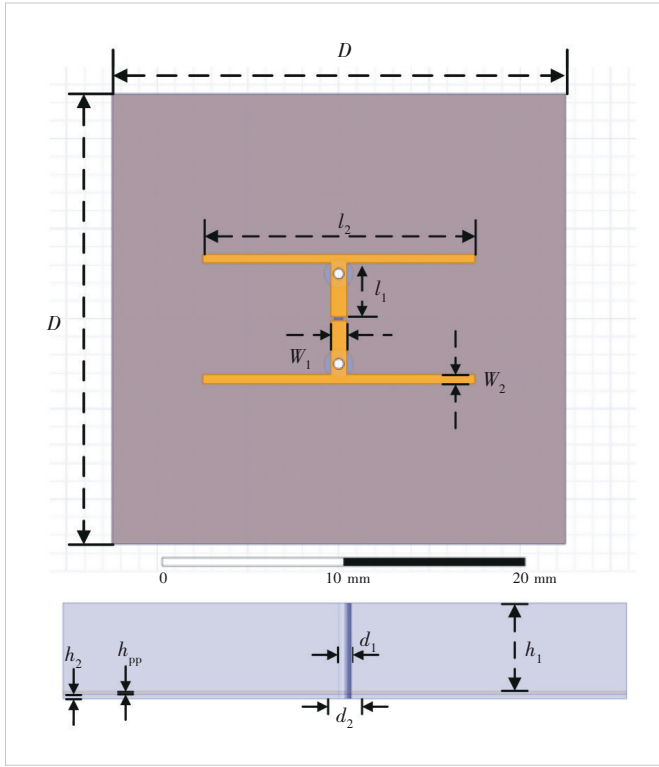


Figure 6. Design diagram of the unit

this study is  $13.3^\circ$ , which is below the upper limit, indicating that the design meets the 3-bit control requirements.

From the  $S_{11}$  amplitude curves, the reflection amplitudes for the eight modulation states are all greater than  $-3.3$  dB within the operating frequency band, indicating low energy loss. The unit thus achieves phase control with nearly constant reflectivity. This is attributed to the dielectric substrate materials, whose appropriate dielectric constant enhances the reflection efficiency of the metasurface.

### 3.2 Simulation and Discussion

By periodically extending the unit model from Section 3.1 in two dimensions, we es-

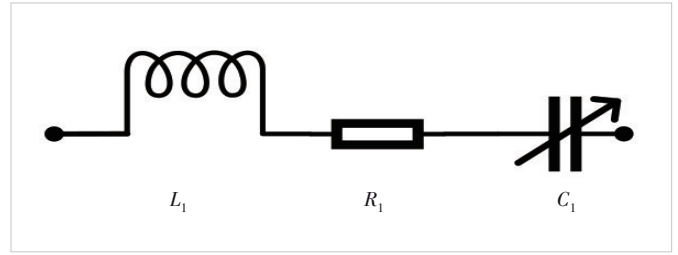


Figure 7. Equivalent circuit of the varactor diode, modeled as a series RLC circuit

tablished an array model in HFSS. Incident antenna feeds were used as inputs to investigate the focusing effect and efficiency of the array.

Microstrip antennas were used as feeds owing to their smaller volume, which can significantly reduce the obstruction of reflected focused beams. Because of their poor directivity, the microstrip antenna and the metasurface were placed at a location that ensures high feed efficiency, with the feed source at  $(0, 0, 0.08)$  m. The incident power is 1 W. We used a  $10 \times 10$  metasurface array and set the focusing position at the center of the reference plane, 1 m away from the center of the metasurface, to evaluate the theoretical focusing performance. The compensation phase of each element is calculated using the phase calculation formula in Section 2.2.

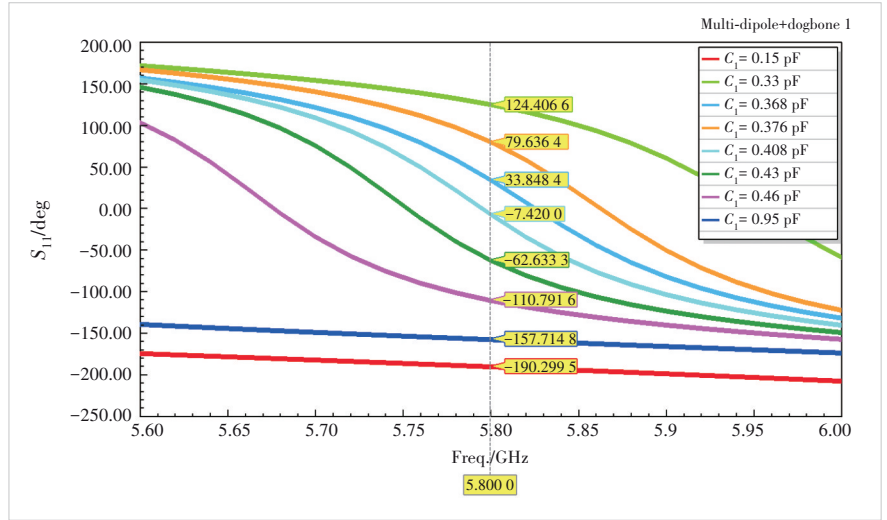
Figure 8.  $S_{11}$  phase curves for different capacitance values

Table 1. Geometric parameters and material properties of the unit design

|                      |                                  |                                                                                    |                   |
|----------------------|----------------------------------|------------------------------------------------------------------------------------|-------------------|
| Geometric Parameters | Element dimensions and thickness | $D = 25$ mm                                                                        | $h_1 = 4$ mm      |
|                      |                                  | $h_2 = 0.2$ mm                                                                     | $h_{pp} = 0.1$ mm |
|                      | Patch parameters                 | $w_1 = 0.9$ mm                                                                     | $l_1 = 2.985$ mm  |
|                      |                                  | $w_2 = 0.5$ mm                                                                     | $l_2 = 15$ mm     |
|                      | Via dimensions                   | $d_1 = 0.6$ mm                                                                     | $d_2 = 1.6$ mm    |
| Material Properties  | Dielectric substrate             | F4B material (with a relative permittivity of 2.65 and a loss tangent of 0.001 5)  |                   |
|                      | Adhesive material                | Prepreg material (with a relative permittivity of 2.5 and a loss tangent of 0.004) |                   |

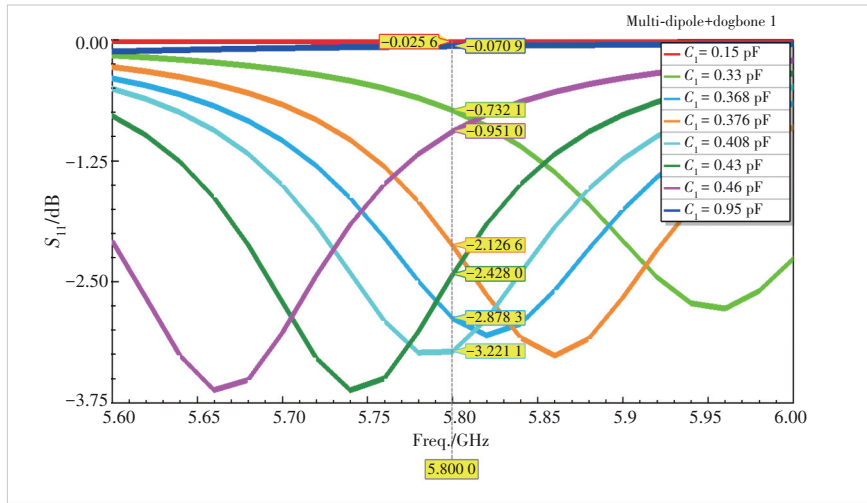
Figure 9.  $S_{11}$  amplitude curves for different capacitance values

Table 2. 3-bit encoding correspondence table

| Encoded Value | Bias Voltage/V | Capacitance/pF | Resistance/ $\Omega$ | Inductance/nH | Phase/deg |
|---------------|----------------|----------------|----------------------|---------------|-----------|
| 000           | 9.60           | 0.150          | 0.88                 | 0.02          | 124.41    |
| 001           | 4.56           | 0.330          | 0.88                 | 0.02          | 79.64     |
| 010           | 4.08           | 0.368          | 0.88                 | 0.02          | 33.85     |
| 011           | 3.70           | 0.376          | 0.88                 | 0.02          | -7.42     |
| 100           | 3.32           | 0.408          | 0.88                 | 0.02          | -62.63    |
| 101           | 2.94           | 0.430          | 0.88                 | 0.02          | -110.79   |
| 110           | 2.12           | 0.460          | 0.88                 | 0.02          | -157.71   |
| 111           | 0              | 0.950          | 0.88                 | 0.02          | -190.30   |

The phase compensation and focusing effect diagrams are shown in Fig. 10.

When the  $10 \times 10$  array is focused at the center of the reference plane 1 m away, the size of the focused area is approximately  $19 \text{ cm} \times 19 \text{ cm}$ , which aligns with the simulation results in Section 2.3, verifying the accuracy of the WPT array focusing simulation. To determine the focusing efficiency of the array, the field calculator in HFSS software was used. We selected the focal area on the reference surface that receives energy from the metasurface as the integration area. The real part of the Poynting vector was integrated over this area. The total energy in the focal region is 0.11 W, while the transmission power of the microstrip antenna is 1 W. According to the efficiency formula, the focusing efficiency is 11%. We note that we did not use a horn antenna with better directivity to reduce antenna obstruction in the simulation.

Using a horn antenna as the feed would further improve the focusing efficiency.

For the RIS-based WPT system simulated in Fig. 10, we set reference surfaces at different distances and calculated the focusing power using the Poynting vector integral from the simulation. The beam focusing efficiency mentioned in Section 2.3 was calculated at each distance. For comparison, we also included a high-gain beam as a reference. As shown in Fig. 11, at the set focusing distance, the focusing efficiency exceeds 11%, but it decreases significantly as the distance deviates. This indicates that deviation in target device positioning can lead to substantial changes in focusing efficiency, which can be resolved with more accurate location information.

Compared with the high-gain beam, the focusing beam achieves significantly higher efficiency at the focus distance, while the high-gain beam maintains relatively uniform efficiency across distances. For long-distance transmission, high-gain beams have advantages. However, for the short-range energy transmission scenario discussed in this paper, near-field beam focusing is more suitable, demonstrating the feasibility of the proposed scheme.

#### 4 WPT Experiment

Based on the metasurface elements discussed in Section 3.1 and the array size discussed in Section 2.3, we fabricated a  $20 \times 20$  WPT RIS prototype. The RIS panel measures  $500 \text{ mm} \times 574 \text{ mm}$ , as shown in Fig. 12. The DC bias line for controlling the reverse bias voltage of the diode is guided to the back of the RIS via a through-hole, and the input voltage is adjusted by a controller with 20 sets of 21 pins per group. The  $21 \times 20$

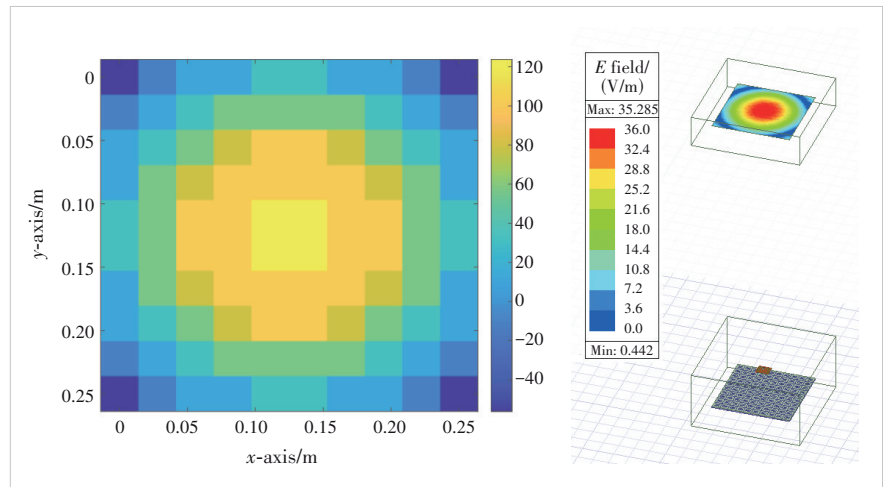


Figure 10. Phase compensation (left) and focusing effect (right) of the array

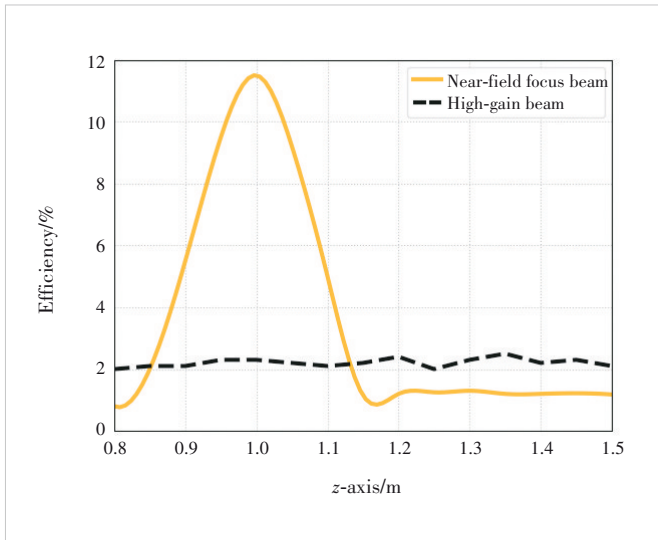


Figure 11. WPT efficiency versus distance for near-field focus beam and high-gain beam

(total of 420) bias lines are used because the  $20 \times 20$  array has 400 units, with each column equipped with a grounding bias line to enhance grounding and system stability. The RIS is regulated by a high-performance FPGA control board, which can output 480 DC channels. The RIS is connected to the FPGA control board via a ribbon cable.

We measured the focusing effect of near-field RIS in a microwave anechoic chamber. The configuration of the measuring device is shown in Fig. 13. The RIS is fed by a horn antenna, which is set to oblique incidence at positions  $(0, -0.8, 0.5)$  m and operates at a frequency of 1 – 18 GHz. Because near-field feed blockage is more severe than the far-field<sup>[11]</sup>, the focusing effect decreases significantly when the focusing position is behind the horn antenna. Therefore, oblique incidence is used for energy feeding. However, the oblique incidence results in nonuniform illumination of the RIS aperture, which reduces the focusing efficiency<sup>[31]</sup>.

The test used planar near-field scanning with a low-gain rectangular waveguide probe to scan a  $50 \text{ cm} \times 50 \text{ cm}$  plane.

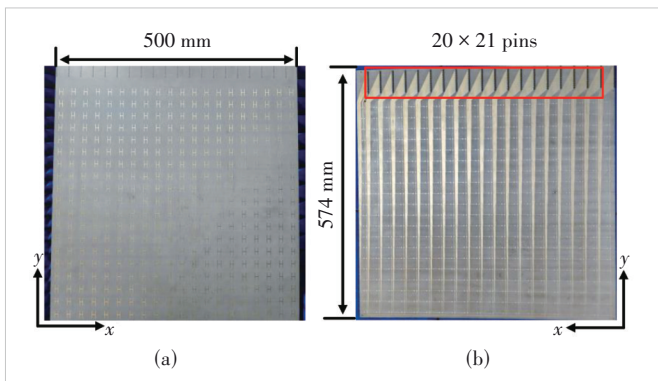


Figure 12. Fabricated 3-bit RIS: (a) front view; (b) back view

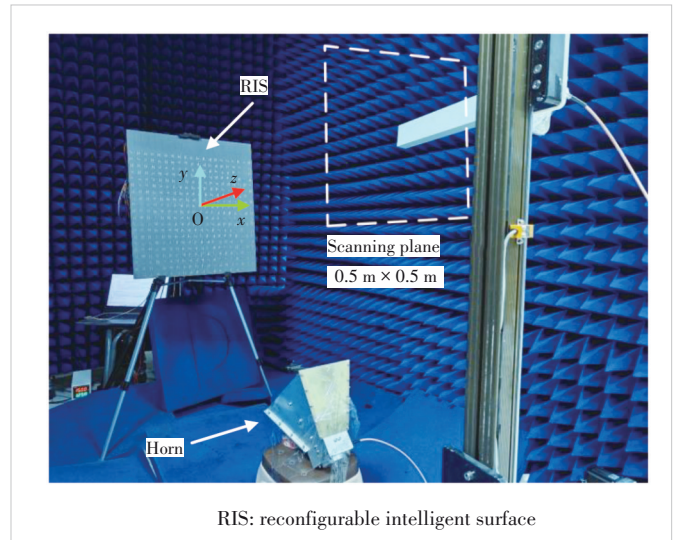
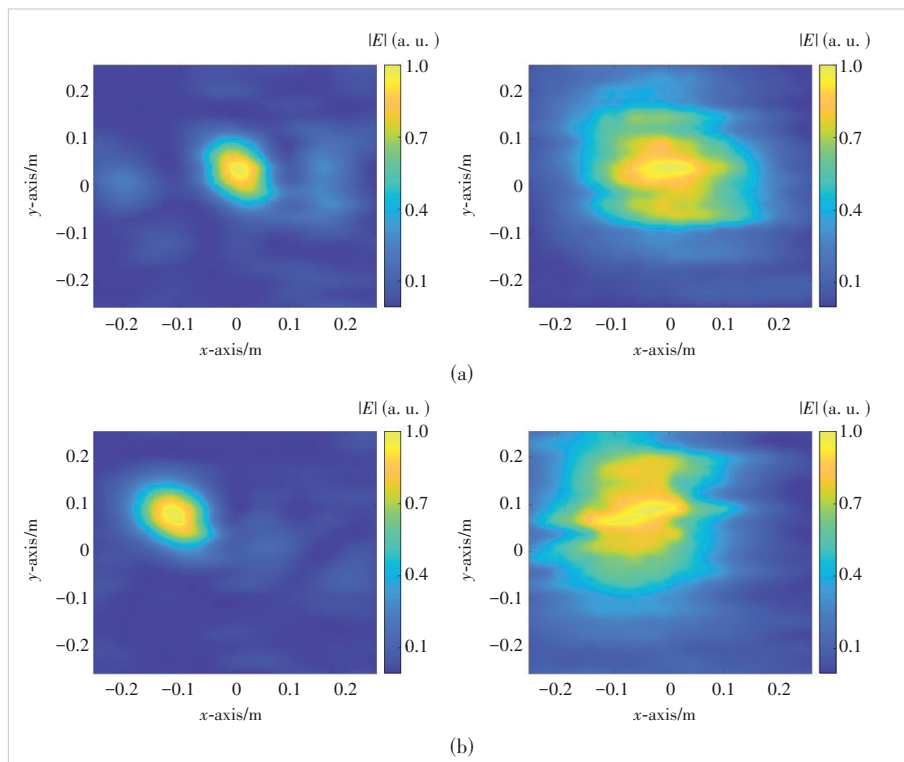


Figure 13. Near-field measurement setup for focal spots in the anechoic chamber

During the test, the polarization mode of the RIS and probe remained consistent. Due to site limitations, the focusing distances were set to 1 m and 3 m, with focusing positions at  $(0, 0, z)$  m on the focal plane (where  $z$  is 1 m and 3 m) and at  $(-0.1, -0.1, z)$  m, respectively. The results are shown in Fig. 14. For the convenience of observation, all electric field distribution results are normalized.

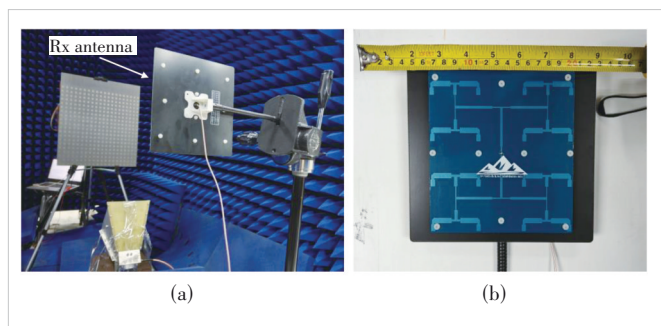
It can be observed that there are obvious focal points at the designed focal positions, which verifies the rationality of phase compensation for near-field focusing. The focal point contours in Fig. 14 appear elliptical, likely due to oblique incidence and the non-uniform dielectric constant of the F4B substrate used. As the distance increases, the focal point size grows significantly: at 1 m, it is  $9.5 \text{ cm} \times 9.5 \text{ cm}$ ; at 3 m, it is  $27 \text{ cm} \times 27 \text{ cm}$ . The test results at a focusing distance of 1 m are consistent with the simulation results. For the 3 m case, the simulated focus size (with the configuration modified accordingly in Section 2.3) is  $27.6 \text{ cm} \times 27.6 \text{ cm}$ , and the measured results also match the simulation results. These test results indicate that the designed near-field focusing scheme and 3-bit RIS can focus wireless power at the desired position and remain effective at longer distances.

To verify the effectiveness of energy focusing, we used a directional microstrip antenna array as the RX and connected it to a VNA to measure the received power. The transmitting antenna configuration was the same as above. The microstrip antenna is a commercial antenna array measuring  $14 \text{ cm} \times 14 \text{ cm}$ , with a nominal gain of 20.5 dBi at 5.8 GHz. Placed at  $(0, 0, z)$  (where  $z$  is 1 m and 3 m), the measurement setup and RX antenna are shown in Fig. 15. The energy transmission efficiency is determined by the ratio of the received power to the transmitted power, with the transmitted power fed into the horn antenna set at  $-5 \text{ dBm}$ . Measurements show that the energy transmission efficiency is about 10% ( $-15 \text{ dBm}$ ) at a dis-



**Figure 14.** Measured normalized electric field distributions with focusing distance (represented as  $z$ ) set to 1 m (left column) and 3 m (right column): (a) focusing position at  $(0, 0, z)$  m; (b) focusing position at  $(-0.1, -0.1, z)$  m

able part of daily life. In this study, we designed a near-field reflective focusing metasurface with a 3-bit quantization precision and high reflectivity. Simulations were conducted to investigate its specific energy focusing efficiency and the advantages provided by its reconfigurability. We explored the phase control mechanism of the metasurface, achieving beam focusing through phase compensation based on near-field beam coherent superposition theory. Analysis identified phase quantization precision and array aperture as the two primary factors affecting focusing performance and provided optimal parameter values. A metasurface unit based on a dog-bone-shaped patch structure was designed and simulated, using varactor diodes to achieve  $360^\circ$  non-quantized phase control with high reflectivity at 5.8 GHz. Actual measurements further validated the feasibility of the focusing strategy and unit design for near-field WPT.



**Figure 15.** Measurement setup for WPT efficiency: (a) anechoic chamber setup; (b) Rx antenna

tance of 1 m and about 2.5% ( $-21$  dBm) at a distance of 3 m. It is noted that the focusing efficiency significantly decreases at 3 m because the receiving area is fixed, while the 3 dB focal spot is significantly larger than the receiving end area, so most energy is not received. Compared with the 1-bit scheme in Ref. [11], the 3-bit scheme achieves more efficient near-field WPT at longer distances.

## 5 Conclusions

The development of wireless power transfer technology has spanned a century, and it is foreseeable that it will be widely adopted in the near future, becoming an indispens-

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### Biographies

**Wang Yunfeng** received his bachelor's degree in communication engineering from Beijing University of Posts and Telecommunications (BUPT), China in 2024. He is currently working toward a master's degree in communication and information engineering at the School of Artificial Intelligence, BUPT. His research interests include the emerging technologies of 5G wireless communication networks.

**Xu Zhiyu** received his PhD degree in electrical and electronic engineering from the Hong Kong University of Science and Technology (HKUST), China in 2002. He is currently a senior architect at ZTE Corporation. He has authored or co-authored more than 10 articles in international journals and conferences, and has filed more than 10 patents. He is the author of two technical books. His research interests include 6G wireless communication and signal processing.

**Zhang Hongtao** (htzhang@bupt.edu.cn) received his PhD degree in communication and information systems from Beijing University of Posts and Telecommunications (BUPT), China in 2008. He is currently a full professor at BUPT. He has authored or coauthored more than 100 articles in international journals and conferences, and has filed more than 50 patents. He is the author of ten technical books. He is a senior member of the IEEE. His research interests include 5G wireless communication and signal processing.