

Design of a Compact Wave-Absorbing Plate for Suppressing Ground Reflection in MWPT Field Experiments

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ABSTRACT: Microwave wireless power transfer (MWPT) offers significant advantages for charging unmanned vehicles over distances on the order of 100 m in atmospheric environments. To accurately measure the beam efficiency in field experiments, it is critical to suppress the impact of ground reflection on the field distribution generated by the beam. This paper presents the design of a compact wave-absorbing plate. The plate is composed of two dielectric waveguides arranged in an alternating side-by-side configuration. One waveguide is periodically loaded with metal patches along the propagation direction to absorb horizontally polarized incident waves, while the other is designed to absorb vertically polarized waves. Slots on the top surface are employed to couple the incident wave energy into the waveguides. Simulation results indicate that at 10 GHz, the reflection coefficients for both horizontal and vertical polarizations remain below -20 dB for incident angles ranging from 60° to 75° . In terms of volume, the proposed absorber achieves an 85% reduction in absorbing material consumption compared with conventional structures. It can be obliquely deployed on the ground as an array along the propagation path of the microwave beam to effectively attenuate ground reflection.

1. INTRODUCTION

In recent years, microwave wireless power transfer technology has been continually improved [1]. Large-distance experiments are performed, and advanced transmitting antennas are presented and applied [2–4]. The technology has also attracted attention in the application of charging unmanned vehicles [5–7]. In the X-band, microwave wireless power transfer utilizes non-diffracting propagation characteristics of microwave beams to achieve long-distance transmission on the order of tens of meters, which meets the distance requirement for charging unmanned vehicles [8]. Moreover, X-band microwaves have low atmospheric loss. In clear or rainy weather, the propagation loss is typically smaller than 1 dB/km. These advantages make it suitable for long-range wireless charging of unmanned devices such as drones [9].

Field experiments for measuring the power transmission efficiency can be conducted on the ground. The propagation distance is proportional to the square of the transmitting antenna aperture and inversely proportional to the microwave wavelength [10]. To obtain accurate measurement results, it is necessary to eliminate the reflection of microwave beams by the ground. Therefore, microwave absorbing materials need to be laid on the ground to reduce the power of reflected waves. For example, the planar reflect-array antenna can produce a microwave focusing beam in X-band and can easily achieve microwave power transfer over a distance of 20 meters with a 1.5-meter aperture. The microwave will arrive at the ground with an incident angle of $\sim 75^\circ$ at the half propagation distance of 10 m when the transmission antenna array is elevated up to a height of 2.6 m.

By now, a commonly available microwave absorber is a pyramid-shaped wave-absorbing foam, as shown in Fig. 1(a). When the incident microwave is perpendicular to the ground, the geometry is a slowly-tapered pyramid which gradually matches the wave impedances from free space to the lossy foam. This design can operate in a wide frequency range determined mainly by the effective attenuation frequency range of the lossy material. When the microwave incident at a grazing angle, e.g., when the incident angle exceeds 60° , the wave-absorbing capacity of the pyramidal wave-absorbing foam array is significantly weakened [11, 12].

The structure in Fig. 1(a) can be tilted to absorb the microwaves with large incident angles, as shown in Fig. 1(b). It can be expensive to cover the ground in the microwave power transmission (MPT) experiment at 10 GHz, considering that an area of 200 m^2 should be covered by the absorber shown in Fig. 1(b).

It is our purpose in this work to design a compact and low-cost absorbing structure that can effectively attenuate the incident microwave power with $\theta < 30^\circ$ as shown in Fig. 1(b) and operate in a narrow frequency range containing 10 GHz to suppress the ground reflection in 10-GHz MPT experiments. In previous studies, dielectric lossy bars have been applied for microwave attenuation in X-band [13, 14]. The dielectric bars work physically as an array of dielectric rectangular waveguides. The microwave power can be attenuated during the propagation along the waveguides.

The carbon-based materials, alloys, and nano-composites featured in recent studies offer microwave engineering enhanced design flexibility through lightweight structures and broadband impedance matching, facilitating the minia-

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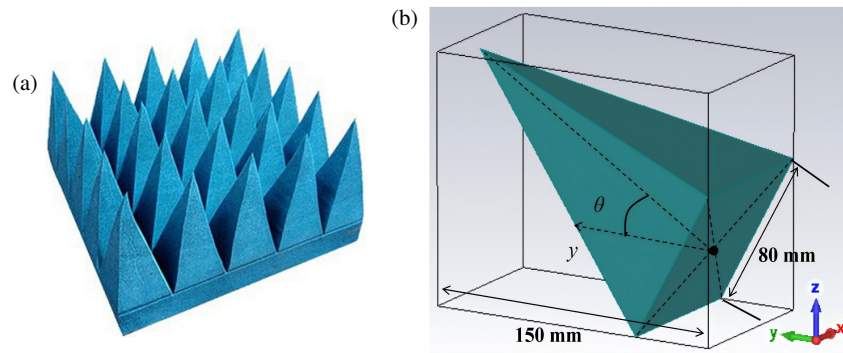


FIGURE 1. Pyramidal wave-absorbing foam array and microwave reflections between pyramids. (a) Geometry and (b) tilted geometry.

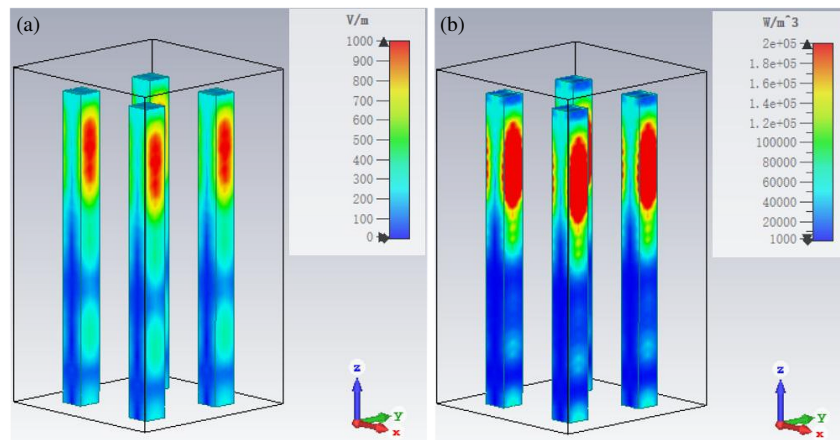


FIGURE 2. Simulated physical picture of beat-wave propagation along DRW. (a) Distribution of E -field on surface and (b) distribution of power loss density.

turization of antennas, filters, and microwave absorbers in microwave frequency bands [15–22].

In this paper, dielectric waveguides are specifically designed to attenuate the grazing-incident microwaves. The taper structure and the gradual matching of the wave impedance will not be adopted. However, the lossy material of the dielectric waveguides is the same as that of the pyramidal wave-absorber shown in Fig. 1(a). In Section 2, the attenuation properties in dielectric rectangular waveguides (DRWs) will be discussed. The lossy dielectric waveguides are also tilted as shown in Fig. 1(b), and they can hardly work for both polarizations of the electric field of the incident plane wave. Two waveguides are presented to solve the problem of polarizations in Section 3. Metal patches are periodically attached to the surface of the DRW along the propagation direction to absorb horizontally-polarized incident waves. Dielectric waveguides with and without metal patches are combined as a plate absorber in Section 4, and the optimized attenuation properties are compared using the pyramidal wave-absorbing array by full-electromagnetic simulation. At last, a conclusion is included in Section 5.

2. ATTENUATION IN DIELECTRIC RECTANGULAR WAVEGUIDES

It is expected that the DRWs can suppress power reflection in MPT experiments. Since the distance between the transmit-

ting antenna and the attenuator is much larger than the wavelength at 10 GHz, the incident microwave can be regarded as a TEM wave arriving on the top surface of the DRWs without loss of generality. In this section, the attenuation characteristics of DRWs are investigated for both perpendicular and inclined orientations relative to the ground, as illustrated in Fig. 1.

As reported by Bao et al. [13], two modes can be excited in a DRW under TEM incident wave illumination: one is the guided mode, and the other is the leaky mode. A beat-wave pattern arises from the difference in longitudinal phase constants between these two modes.

Figure 2 shows the simulated beat-wave propagation field pattern obtained from Computer Simulation Technology (CST) Microwave Studio, where the structure is excited by an x -polarized TEM plane wave incident from the upper z boundary. The transverse boundaries are set as unit-cell periodic boundaries in CST Microwave Studio. The lossy medium has a relative permittivity of approximately 6 and a loss tangent of 0.21. The foam material is selected as a pure dielectric loss wave-absorbing foam, which has advantages in loss capacity in the X-band compared with magnetic loss wave-absorbing foams [15–22]. This boundary configuration is adopted consistently for all simulations in this work.

The cross-section edge length is 8 mm, and the length of the DRW is 100 mm. A distinct beat-wave pattern can be observed in Fig. 2(a). The amplitude of the E -field is distributed in al-

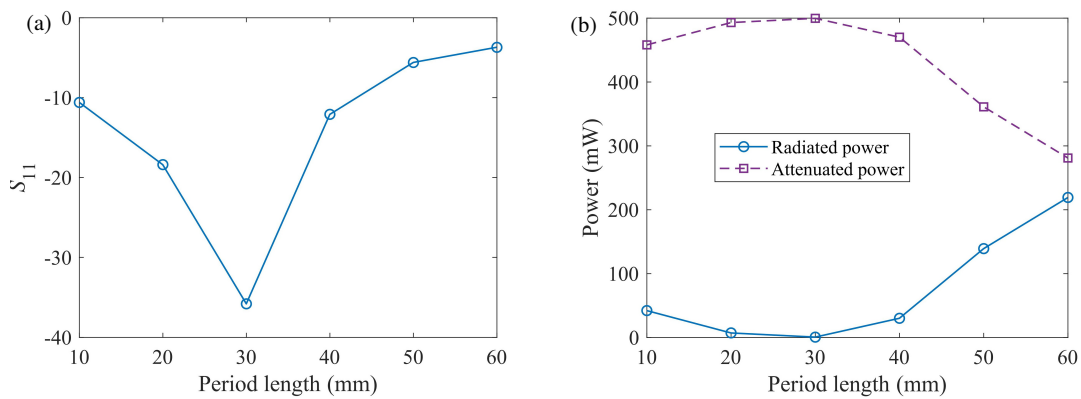


FIGURE 3. Simulated attenuation properties of DRW perpendicular to ground. (a) S_{11} and (b) radiated and attenuated powers.

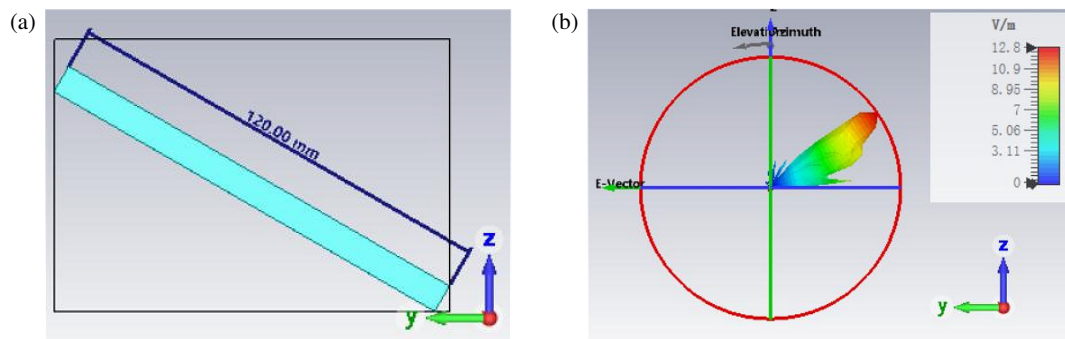


FIGURE 4. Inclined DRW with incident angle of 60° . (a) Geometry and (b) E -field radiation pattern.

ternating regions of high and low intensities along the propagation direction. In the low-intensity regions, the attenuation of the lossy foam will be weak. So, a reasonable longitudinal length should be chosen to contain the high-intensity regions for effective attenuation, as shown in Fig. 2(b).

As shown in Fig. 2(b), the power loss density at 10 GHz decreases rapidly along the propagation direction. After a propagation distance of approximately two wavelengths, the amplitude of the electric field on the DRW surface drops by one order of magnitude, leading to a low S_{11} at the upper z boundary. Most of the power loss occurs near the incident surface.

Consider the DRWs in Fig. 2(a) as an array of absorbing structures. The periodic lengths in both transverse directions (i.e., the x and y directions) are identical. The longitudinal length of the DRWs remains 100 mm. The perfectly electrically conducting (PEC) surface is set at the lower z boundary. Microwave power may be reflected and radiated outside the simulation domain, rather than being attenuated within the DRWs. The sum of the radiated power and dielectric power loss equals 0.5 W, which corresponds to the excitation power used in the simulations.

Figure 3 presents the S_{11} parameter and radiated power as functions of the period length. As the period length increases from 10 mm to 30 mm, the gaps between DRWs become larger, and the reflection at the top surface decreases. With sufficient attenuation, both the S_{11} and radiated power decrease. It should be noted that the incident and reflected waves belong to the same mode.

However, when the period length increases from 30 mm to 60 mm, the array becomes too sparse, and attenuation becomes insufficient, leading to increases in both S_{11} and radiated power. The PEC boundary on the lower z surface reflects any microwave power that is not fully attenuated.

Figure 4(a) shows an inclined DRW with a length of 120 mm. The angle between the DRW and the ground (modeled as a PEC boundary) is 30° , corresponding to an incident angle of 60° . The gap between adjacent DRWs in the x -direction is 30 mm, which is selected based on the lower reflection observed in Fig. 3(a).

The attenuation characteristics are investigated for both vertically and horizontally polarized incident waves. The horizontally polarized wave corresponds to the TE mode at the $+z$ port, where the electric field (E -field) is x -polarized. The vertically polarized wave corresponds to the TM mode at the $+z$ port, where the magnetic field (H -field) is x -polarized.

For convenience, TE and TM excitations are used to denote horizontally and vertically polarized incident waves, respectively, throughout this work.

When the incident angle is 60° , i.e., the wave vector of the incident wave is parallel to the inclined dielectric leaky-wave (DRW) structure in Fig. 4(a), the simulated attenuation characteristics as a function of the dielectric bar length are plotted in Fig. 5. The radiated power increases with the longitudinal length, since the DRW array forms a transmitting aperture whose size enlarges as the length increases. Consequently, for longer dielectric bars, an increasing amount of microwave

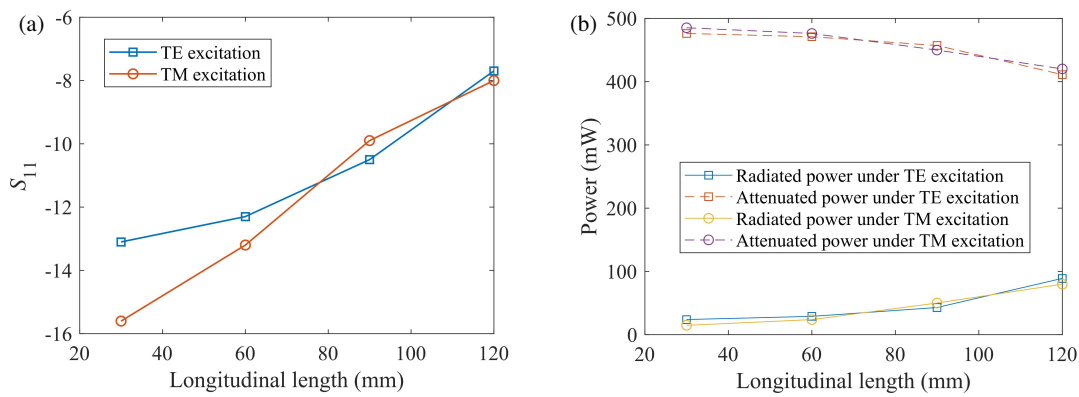


FIGURE 5. Simulated attenuation properties of inclined DRW array. (a) S_{11} and (b) radiated and attenuated powers.

power cannot be coupled into the DRWs and is instead radiated into free space as radiation modes, owing to the enlarged transmitting aperture. As illustrated in Fig. 5, the S_{11} coefficients for both TE and TM excitations increase with increasing radiated power and dielectric bar length. The reflection angle is approximately equal to the incident angle, as shown in Fig. 4(b).

As shown in Fig. 4, it is found that the S_{11} coefficients under TE and TM excitations are different. The PEC ground influences the TE and TM waves differently. For the TE excitation, the electric field is polarized along the x direction and the x -component of the electric field should be zero at the surface of the PEC ground. This causes a stronger reflection of the TE wave. It should be mentioned that the dielectric rectangular waveguide also has a cut-off frequency. The edge length of the waveguide, which is the core structural factor, should be large enough to allow the propagation of the excited modes.

By comparing the results in Fig. 5, it can be observed that the total reflections under TE and TM excitations differ when the DRWs are tilted. Before tilting the dielectric bars, identical periodic translational symmetry exists in both the x and y directions. However, after tilting the structure, the symmetry along the y direction changes, which leads to the discrepancy in the S_{11} coefficients under the TE and TM excitations.

A more sophisticated structure is needed to suppress reflections and enhance the coupling from free space into the lossy waveguides. The reflection can hardly be reduced by merely adjusting the period length in either the x or y direction. As discussed in Fig. 3, a denser absorbing array gives rise to a stronger reflection, while a sparser array results in poorer attenuation performance. To achieve efficient coupling from free space to the DRWs, the continuity of the tangential fields must be well satisfied. In particular, the longitudinal phase constants should be closely matched.

3. DESIGN OF ABSORBING DIELECTRIC WAVEGUIDES

To achieve a reflection level below -20 dB for both TE and TM excitations, two types of dielectric leaky waveguides (DRWs) are designed to attenuate the TE and TM modes, respectively. One is the metal-patch-attached dielectric rectangular waveguide

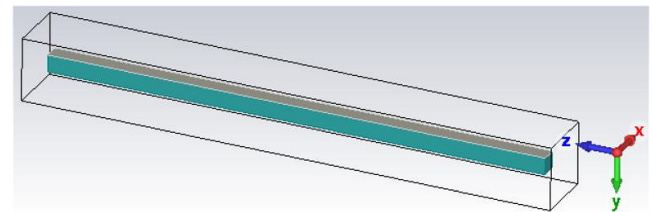


FIGURE 6. Geometry of metal-patch-attached dielectric rectangular waveguide.

uide (MPADRW); the other is the conventional DRW without metal patches.

The geometry of the MPADRW is illustrated in Fig. 6. The gray region represents the PEC metal patch loaded on the DRW, while the green bar denotes the lossy dielectric material. The dielectric has a relative permittivity of approximately 6 and a loss tangent of 0.21, identical to that of the aforementioned DRWs. The incident plane wave is launched from the $+z$ boundary and propagates along the $-z$ direction. The electric field of the TE mode at the $+z$ port is x -polarized, whereas that of the TM mode is y -polarized. An open boundary condition is applied at the $-z$ surface.

The MPADRW is suitable for absorbing microwave power under TE-mode excitation. Fig. 7 presents the electric field distributions in the xoy plane and the surface power loss densities under TE and TM mode excitations, respectively. The electric field strength inside the dielectric waveguide will be larger than that in free space, which is the requirement of the normal electric field boundary condition. Under this condition, the microwave power is coupled into the MPADRW through the sidewalls along the polarized direction of the E -field. For TM excitation, a sidewall along the y direction is blocked by the metal patch so that the electric field distributes mainly in free space, as shown in Fig. 7(b). However, the metal patch will not block the electric field under TE excitation.

Consequently, the attenuation under TE-mode excitation reaches approximately 0.5 W, corresponding to the total input power at the $+z$ port, whereas the attenuation under TM-mode excitation is insufficient.

As observed in Fig. 7(c), under TE-mode excitation, the microwave power can be effectively attenuated inside the

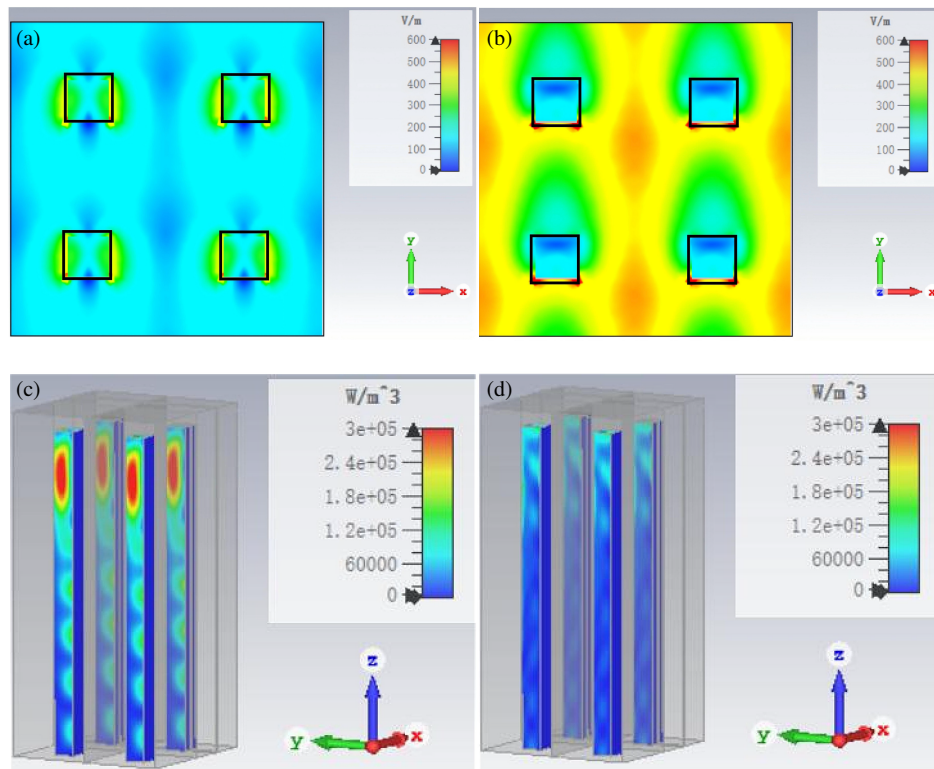


FIGURE 7. Attenuation properties of MPADRW under different excitations. (a) *E*-field pattern in the cross plane under TE mode excitation, (b) *E*-field pattern in cross plane under TM mode excitation, (c) power loss density on the surface under TE mode excitation, and (d) power loss density on the surface under TM mode excitation.

MPADRW, with power loss concentrated in the material regions adjacent to the metal patch. The power loss density decreases along the propagation direction, since the incident microwave power is substantially absorbed. It can be expected that the MPADRW can be more compact while still sufficiently attenuating the incident TE-mode wave. In contrast, as shown in Fig. 7(d), the power loss density is much lower under TM-mode excitation.

The thinner MPADRW fails to maintain its advantage in absorbing incident TE-mode waves. As illustrated in Fig. 8, the thickness along the direction normal to the metal patch is reduced from 8 mm to 2 mm. In comparison, the transverse gap between adjacent dielectric waveguides remains equal to a wavelength of 30 mm.

From the electric field distribution in the *xoy* plane (Fig. 8(a)), the electric field is concentrated in the spacing between adjacent waveguides. As shown in Fig. 8(b), the power loss inside the waveguide is weak and is localized near the corner region. The poor attenuation arises because the phase constant of the guided wave deviates significantly from that in free space.

Under TE excitation, the wave propagating along the thinner MPADRW has a phase constant smaller than that in free space, which is inversely proportional to the longitudinal propagation distance corresponding to a phase shift of 360° . As shown in Fig. 8(c), this distance is approximately 40 mm, resulting in the mismatch of the longitudinal phase constants. In contrast, for the square-shaped DRW, this characteristic distance returns to

28 mm (Fig. 8(d)), which is close to 30 mm for the TEM wave in free space. So, the microwave power is coupled into the square-shaped DRW well.

The phase velocity of the excited wave exceeds the speed of light due to the relatively large phase constant. Therefore, slowing down the phase velocity is necessary to improve the coupling and attenuation performance.

It is well known that periodic conducting structures can reduce the phase velocity in a waveguide, and such structures are referred to as slow-wave structures. In this work, the metal patch is divided into a series of small patches attached to one side of the MPADRW. Each patch has a width of 8 mm and a longitudinal length of 8 mm.

The simulated power loss of the modified MPADRW is presented in Fig. 9. As the period length of the metal patches increases, the dielectric loss first increases and then decreases for period lengths larger than 19 mm. The maximum attenuation is achieved when the period length of the metal patches is approximately 19 mm. The longitudinal distance corresponding to a 360° phase shift also first increases and then decreases, approaching 30 mm at a period length of around 19 mm.

The matching between the phase constants in free space and those in the dielectric waveguide is essential for achieving high attenuation efficiency. The core mechanism is to match the phase constant of free-space microwave to that of the fundamental guided mode for the MPADRW. Otherwise, under the continuity boundary condition of tangential fields, higher-order modes will be excited, and the coupling efficiency will drop.

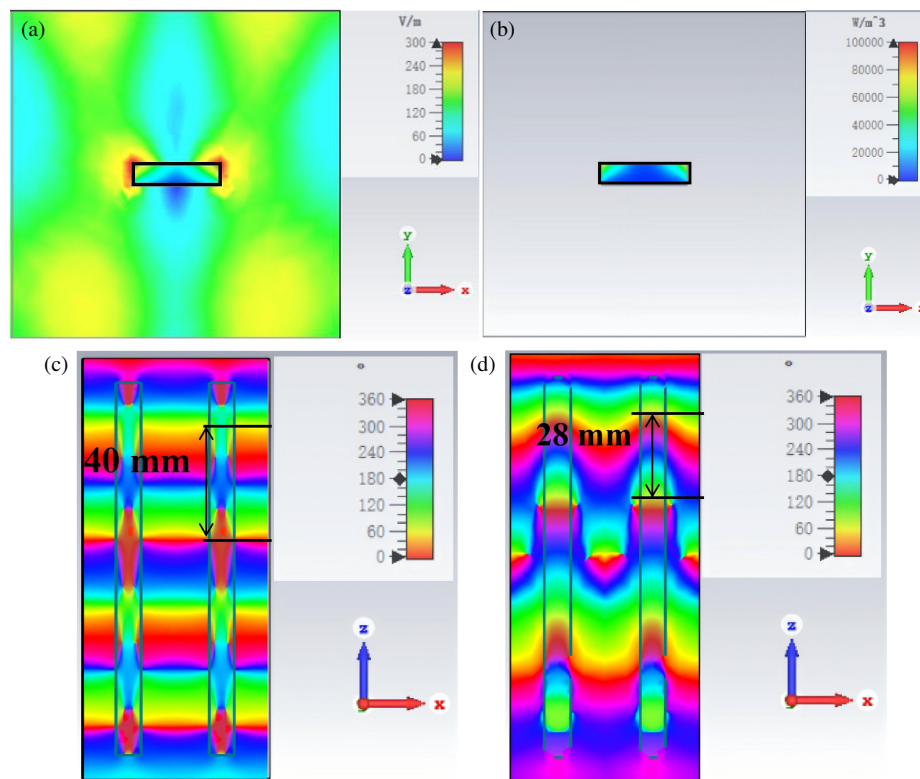


FIGURE 8. Attenuation properties of thinner MPADRW under TE mode excitations. (a) E -field pattern in cross-plane, (b) power loss density in cross-plane, (c) phase distribution in xoz plane for thinner MPADRW, and (d) phase distribution in xoz plane for square-shaped DRW.

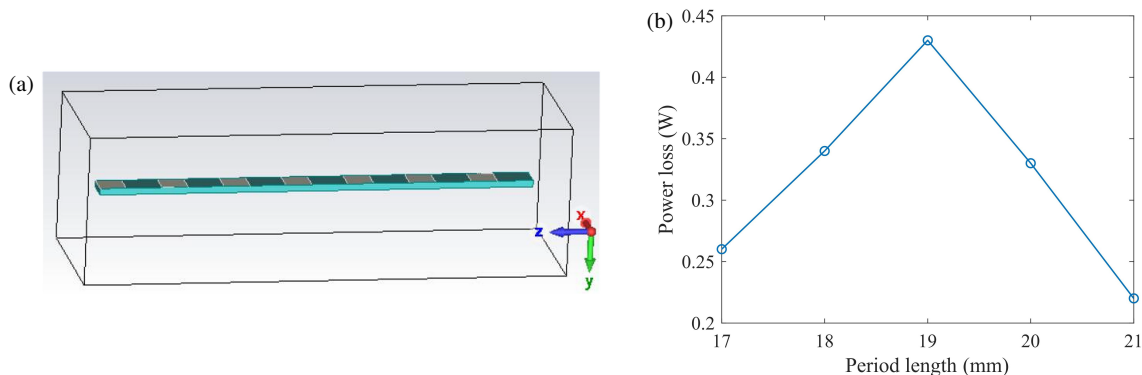


FIGURE 9. Reasonable choice of period length of metal patches. (a) Geometry of modified MPADRW and (b) power loss of TE wave as functions of period length of metal patches.

The DRW without metal patches is employed to absorb incident TM-mode waves. As illustrated in Fig. 10(a), the thickness of this DRW has been adjusted to match that of the MPADRW, facilitating their integration into a single unit. The incident TM wave is incident from the $+z$ port, with its wave vector parallel to the waveguide.

Figure 10(b) presents the attenuation characteristics of the thinner DRW as a function of the transverse period length. The dielectric loss decreases with an increase in the transverse period length; specifically, the sparser the array of thinner DRWs, the poorer the attenuation performance. Therefore, a more compact absorbing structure composed of the two types of DRWs described in Section 3 is required to effectively attenuate both the incident TE and TM waves.

4. ABSORBING ARRAY OF DIELECTRIC WAVEGUIDES

From the discussion in Section 3, it can be found that the thinner DRWs with and without metal patches can effectively attenuate incident TE and TM waves, respectively. By alternately arranging these two types of DRWs in a line, a compact absorbing plate without a taper structure can be modeled, as shown in Fig. 11. This designed absorbing array operates at 10 GHz, and the lossy dielectric material used has a relative permittivity of approximately 6 and a loss tangent of 0.21. In both the x and y directions, the absorbing plates are closely placed to form the absorbing array. The angle between each plate and the ground — modeled as a PEC boundary at the $-z$ surface — is $\theta = 30^\circ$. The S_{11} coefficients for TE and TM excitations can be affected by the conductivity of the ground, which in practice is signifi-

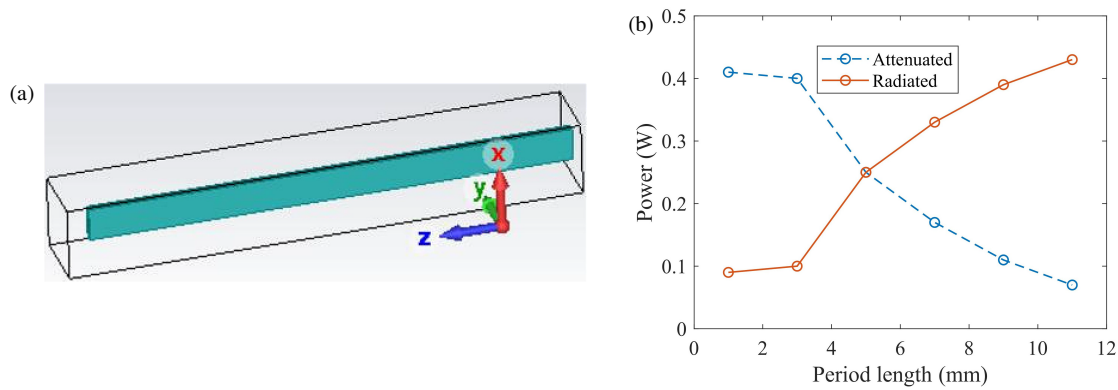


FIGURE 10. Attenuation properties of thinner DRW. (a) Geometry of thinner DRW and (b) radiated and attenuated power as functions of transverse period length.

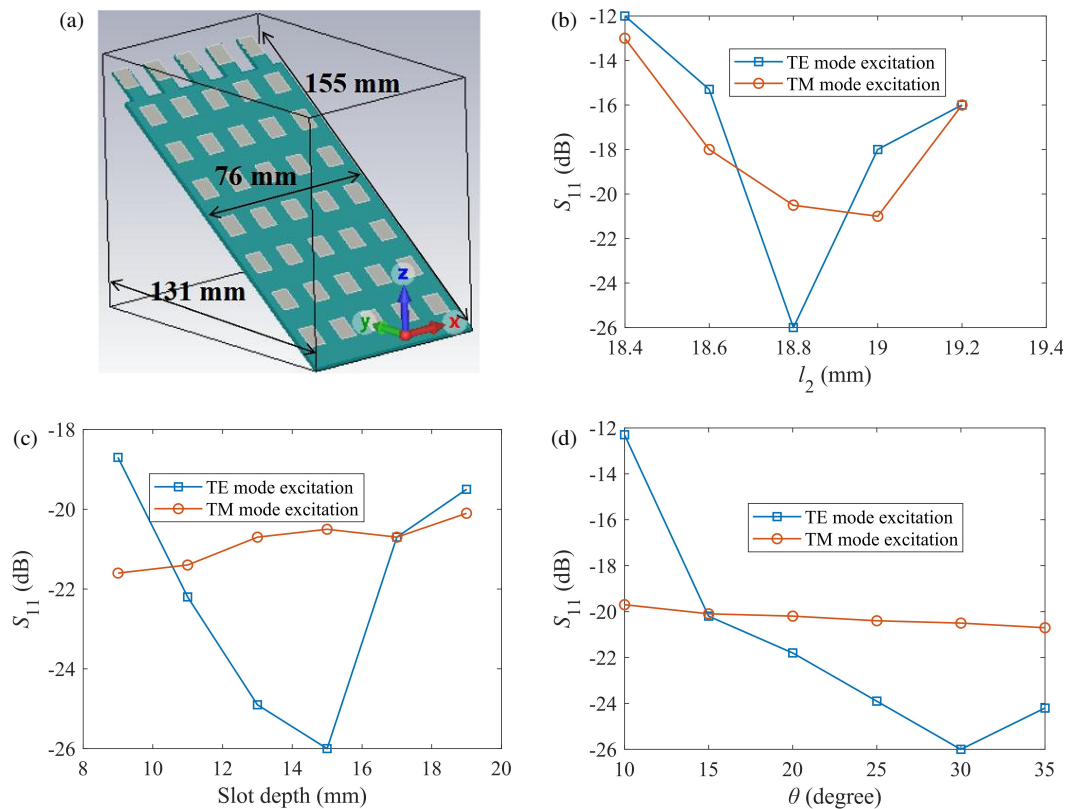


FIGURE 11. Optimization of attenuation properties of designed absorbing array. (a) Geometry, (b) S_{11} vs l_2 , (c) S_{11} vs slot depth, and (d) S_{11} vs θ .

cantly influenced by the soil moisture content. Therefore, it is suggested that a layer of aluminum foil should be placed over the ground to act as a PEC boundary, which is quite cheap and recyclable.

The incident wave always impinges normally on the absorbing plate. The widths of the thinner DRW (without metal patches) and the MPADRW (with metal patches) are denoted by w_1 and w_2 , respectively, and the period length in the x direction is the sum of w_1 and w_2 . The longitudinal length and period length of the metal patches are denoted by l_1 and l_2 , respectively, while the length of the waveguide is denoted by l , and the thickness of the plate is denoted by t . A slot is designed at the top of the thinner DRWs to enhance the coupling of the

incident wave, with a depth of 15 mm (i.e., half a wavelength). With $\theta = 30^\circ$, the above-mentioned parameters after optimization by CST MS are as follows: $w_1 = 8.1$ mm, $w_2 = 7.1$ mm, $t = 2$ mm, $l_1 = 9$ mm, $l_2 = 18.8$ mm, and $l = 150.4$ mm. The width of the slot in the x direction is equal to w_1 . The optimization results are also shown in Fig. 11.

It can be observed from Fig. 11(b) that parameter l_2 affects the coupling and S_{11} coefficients for both TE and TM mode excitations. Mutual coupling occurs between the two dielectric waveguides, so the optimal value of l_2 differs slightly from that shown in Fig. 9(b) for a simple MPADRW array.

The slot depth exhibits a significant influence on the reflection characteristic under TE mode excitation. Since a thinner

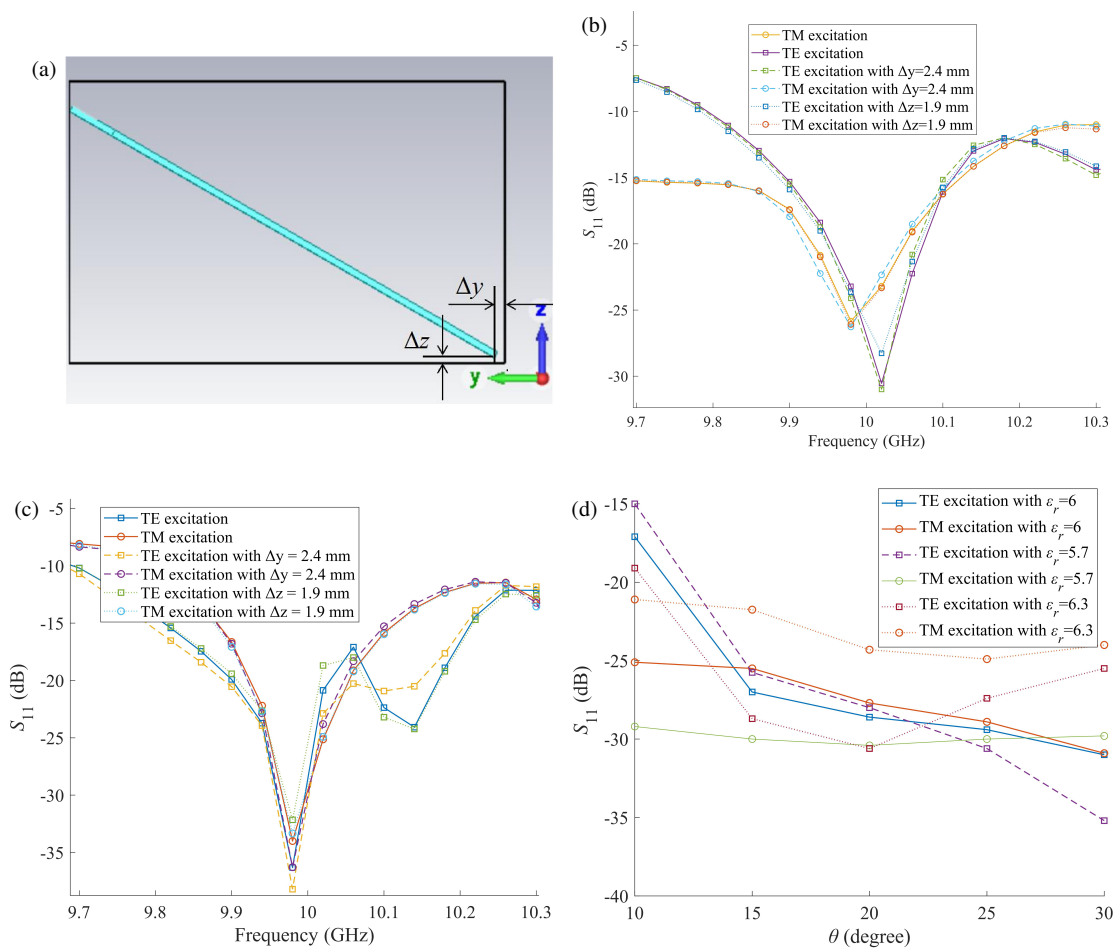


FIGURE 12. Attenuation properties of displaced absorbing structure. (a) Geometry, (b) S_{11} vs frequency for $\theta = 15^\circ$, (c) S_{11} vs frequency for $\theta = 30^\circ$, and (d) S_{11} vs θ at 10 GHz.

DRW cannot effectively couple the incident TE mode, the slot acts as a resonant cavity for the incoming TE wave. Consequently, the coupling of the TE wave is enhanced when resonance is established at a slot depth of half a wavelength.

As the angle θ between the absorbing plate and the ground plane varies, the incident TE wave is perturbed by the ground, since the x -polarized electric field must be zero at the ground surface. Thus, the S_{11} value for TE mode excitation becomes large as θ decreases, whereas the S_{11} value for TM mode excitation can remain low.

The absorber shown in Fig. 11(a) is optimized at 10 GHz and aims for a reflection coefficient below -20 dB. For $\theta = 15^\circ$, the S_{11} coefficients are close to -20 dB, and the corresponding operating bandwidth is reduced to a single frequency of 10 GHz. For a better operation bandwidth, the position of the absorber is optimized, while the structure of the absorbing plate remains unchanged. Fig. 12(a) shows the geometry of the placement. The absorber array is elevated from the ground by Δz , and the gap between adjacent plates along the y -axis is denoted as Δy . When $\Delta y = 2.3$ mm, $\Delta z = 1.8$ mm, the simulated S_{11} coefficients for $\theta = 15^\circ$ and $\theta = 30^\circ$ are plotted in Fig. 12(b) and Fig. 12(c), respectively. The S_{11} values are required to be higher than -20 dB for both TM and TE modes in the operation frequency band. For $\theta = 15^\circ$, a band-

width of $\sim 0.7\%$ is obtained, which is centered at ~ 10 GHz. For $\theta = 30^\circ$, the bandwidth varies little, and the frequency center moves to ~ 9.98 GHz. As a result, the narrow operation band ranges from 9.965 GHz to 10.015 GHz. Outside this range, the microwave power coupled into the lossy plates will decrease because of bad matching between the phase constants of the incident wave and the guided wave in the structure. At 10 GHz, the S_{11} coefficients stay below -20 dB at different values of θ for both excitations as shown in Fig. 12(d).

The sensitivity of the S_{11} coefficients to Δy , Δz , and the relative permittivity of the lossy foam, denoted by ϵ_r , is shown in Figs. 12(b)–(d). In Figs. 12(b) and (c), the errors in Δy and Δz are set to be 0.1 mm, which can be achieved by typical manufacturing techniques. It can be observed that the resonant frequency point, where the minimum reflection occurs, can be shifted by the displacements in the y and z directions. The resonant point varies ~ 0.01 GHz when Δy or Δz is increased by 0.1 mm. The influences of ϵ_r on the reflection coefficients are also investigated. An error of ± 0.3 in the ϵ_r is assumed. Obvious differences are found in Fig. 12(d). For TE excitation, the variation of S_{11} reaches ~ 2 dB for $\theta < 25^\circ$ and becomes larger for $\theta = 30^\circ$ due to the shift of the operation band. For TM excitation, the variation of S_{11} reaches ~ 4 dB when the error in ϵ_r is 5%.

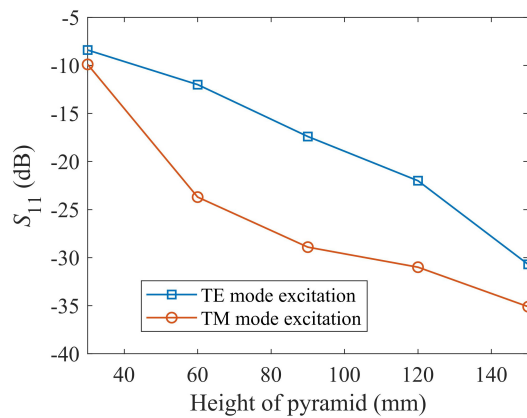


FIGURE 13. Attenuation properties of an absorbing array with pyramid-shaped lossy foam.

Further, a comparison between the proposed absorbing structure and conventional pyramid-shaped foam is conducted. The pyramid foam is tilted as illustrated in Fig. 1(b), with the angle between its symmetry axis and the ground plane also set to 30° . The simulated S_{11} results at 10 GHz are presented in Fig. 13.

Compared to the traditional inclined pyramid-shaped absorbing foam, the compact absorbing plate proposed in this work has a much narrower operation frequency band. The thinner the plate is, the narrower the bandwidth is. In Fig. 12, the optimized bandwidth of only 0.5% is obtained. Considering that the output signal of a microwave radiation source at 10 GHz also has a typical bandwidth of $\sim 0.4\%$. The thickness of the absorbing plate should not be reduced for effective absorption of the microwave power radiated from a practical source. This is the limitation of the material volume or the thickness of the plate. As a result, the optimized thickness of the plate is 2 mm. The volume ratio of the absorbing material per unit area required for the compact absorbing plate can be calculated when the thickness of the plate is 2 mm. Without loss of generality, it is calculated for $\theta = 30^\circ$, and it is required to achieve reflections below -20 dB for both TE and TM mode excitations. For the pyramid foam, a height of approximately 120 mm is required, and the volume of the lossy foam in a unit cell is 256000 mm^3 ($80 \times 80 \times 120/3 \text{ mm}^3$, as shown in Fig. 1(b)). The unit cell covers an area of 12000 mm^2 on the ground plane. So, the requirement of lossy foam for the pyramid absorber is 0.0213 m^3 per square meter. In contrast, the proposed absorbing plate has a volume of approximately 23560 mm^3 ($2 \times 76 \times 155 \text{ mm}^3$) in a unit cell, and the cell covers an area of $\sim 10108 \text{ mm}^2$ ($76 \times 133 \text{ mm}^2$, as shown in Fig. 11(a)). The requirement of the proposed absorber is $\sim 0.002 \text{ m}^3$ per square meter. Finally, the proportion of material consumption reduction achieved is 90%, calculated as $(0.0213 - 0.002)/0.0213$.

An acceptable trade-off is achieved between the operating frequency bandwidth and the overall volume. It is worth noting that the material cost is generally proportional to the volume of the lossy medium.

5. CONCLUSION

This paper focuses on suppressing ground reflection in outdoor beam efficiency measurements for microwave wireless power

transfer (MWPT) used in unmanned vehicle charging, designing a compact absorbing plate for 10 GHz grazing-incident microwaves. The plate combines two dielectric wave-guides arranged alternately: one with periodic metal patches absorbs horizontal polarization (TE mode), the other absorbs vertical polarization (TM mode), with top slots enhancing wave coupling. Simulations show its reflection coefficients for both polarizations are below -20 dB at 60° – 75° incidence, corresponding to angles with the ground ranging from 15° to 30° . Compared with the traditional pyramidal absorbing foam, the new structure uses 90% less absorbing material (only 10% volume required) and can be obliquely arrayed along the microwave beam path to effectively attenuate ground reflections. The high cost for large-area applications can be avoided.

The practical performance of the designed absorber is planned to be verified by an experiment, which is supported by an ongoing microwave wireless power transfer (MWPT) project. The incident microwaves can be radiated from a horn antenna. The wave is reflected and attenuated by the absorber array. The validity of the absorber will be shown by the comparison of the field strengths of the reflected wave with and without the existence of the absorber.

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