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an antenna volume of $40 \times 12 \times 4 \text{ mm}^3$, which is much smaller than previous internal antenna designs. A prototype of the proposed antenna is being designed and tested. Good results of efficiency and gain illustrate the potential use of the proposed antenna for tablet computer applications.

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Theoretical Limitations for TM Surface Wave Attenuation by Lossy Coatings on Conducting Surfaces

Sanghoon Kim and Daniel F. Sievenpiper

Abstract—In this work, we theoretically analyze the limitations for TM surface wave attenuation on lossy coated conducting surfaces containing electric and/or magnetic loss. We use both an analytical approach as well as numerical simulations, and find excellent agreement between them. We also find that the loss can be described by a simple approximate expression for a wide range of material properties. Furthermore, we analyze lossy slabs with simple equivalent circuit boundaries on top, such as may be provided by frequency selective surfaces or other patterned structures. We find that such composite lossy coating can exceed the attenuation of a simple lossy slab, but with limited bandwidth. We also find that only by increasing permeability, and not permittivity, can the peak absorption frequency be lowered for a given thickness without reducing the relative absorption bandwidth.

Index Terms—Attenuation measurement, conducting materials, dielectric materials, microwave propagation, surface waves.

I. INTRODUCTION

It is well-known that a grounded dielectric slab can support bound surface waves [1], [2]. The bound waves are related to the ordinary surface currents that occur in any metal surface in the limit where the dielectric thickness approaches zero. The surface currents contribute significantly to interference between nearby antennas or electronics [3], [4], and suppressing the currents such as by using a lossy coating can be an important tool for interference reduction [5], [6]. An analytical solution to the attenuation by a dielectric slab has been provided by Attwood [7], where he calculated loss terms for both the conductor loss in the metal surface and dielectric loss due to the slab. However, there has been no attempt to analyze general trends or theoretical limits of the attenuation by lossy coatings, particularly for surface waves. The closest general analysis is for plane waves at normal incidence [8]. Furthermore, Attwood's original analysis cannot be applied to many modern lossy coatings which may contain magnetic materials, as well as patterned composite materials such as frequency selective surfaces [9] or metamaterials [10]. In this work, we extend Attwood's analysis to include both of these effects and we verify our analytical solution with numerical simulations. We refer to lossy slabs which may contain impedance sheets, FSS, or other such structures collectively as "coatings". Our analytical solution allows us to sweep a wide range of material properties to derive general trends and establish theoretical limits. In particular, we find that the attenuation by a lossy slab can be approximated by a simple formula. Adding an impedance sheet on top of the slab that can be described by an equivalent circuit can provide greater attenuation than the slab alone, but over limited bandwidth. Thus, lossy coatings combined with frequency selective surfaces, metamaterials or other such patterned surfaces can be effective narrowband absorbers.

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Fig. 1. Geometry for TM surface wave propagating along the z direction with the electric field decaying in the y direction. Schematic of the structure and properties of materials; region 1 is assumed to be a perfect conductor ($\sigma = \infty$), region 2 is a slab of thickness ' a ' with loss properties, given by the imaginary components of relative permittivity (ϵ_2'') and permeability (μ_2''), region 3 is air. There is a possible impedance sheet between the region 2 and region 3.

In this communication, a transverse magnetic (TM) type wave is assumed as a uniform surface wave propagating parallel to the surface of a dielectric slab on a conducting surface. We focus on TM waves because most absorbing coatings are electrically thin, however because the purpose of this communication is to establish trends, we will study a wide range of thickness.

Here, we follow Attwood's analysis and label the conductor as region 1, the slab as region 2 and the surrounding air as region 3. In order to consider only the effects of the slab, we assume that region 1 is a perfect electric conductor as shown in Fig. 1. Region 2 is a dielectric material of thickness ' a ' with relative permittivity (ϵ_2) and permeability (μ_2). The loss properties are represented by the imaginary permittivity (ϵ_2'') and permeability (μ_2''). In region 3, we assume the material outside of the dielectric slab is air with relative permittivity and permeability of unity. We also consider the case where an impedance sheet is located between the slab and the air region. The sheet is considered to have infinitesimal thickness such as a thin frequency selective surface [9].

II. ANALYTICAL SOLUTION

To determine the theoretical limitations of attenuation by a lossy slab, first we consider an analytical solution. The analytical solution is developed from Maxwell's equations, assuming a traveling TM surface wave. Here, we follow Attwood's analysis [7], and also add the effects of magnetic loss, and lossy or reactive impedance sheets.

A. Without Impedance Sheet

For the transverse fields in the slab and air regions, we first start with Maxwell's equations under source-free, linear and homogeneous space conditions. A propagating TM surface wave along the z direction is considered as a monochromatic wave. The wave components are functions of y , and independent of x . For an electrically conductive boundary at $y = 0$ as shown in Fig. 1, the tangential electric field is 0 at the boundary. From this condition, the field components are represented as below. For the dielectric slab in region 2,

$$E_{z2}^{slab} = jK_z \frac{k_2}{\omega\epsilon_2} \sin(k_2 y) \quad (1)$$

$$E_{y2}^{slab} = K_z \frac{\beta}{\omega\epsilon_2} \cos(k_2 y) \quad (2)$$

$$H_{x2}^{slab} = -K_z \cos(k_2 y). \quad (3)$$

For the air in region 3,

$$E_{z3}^{air} = jK_z \frac{k_2}{\omega\epsilon_2} e^{+k_3 a} \sin(k_2 a) e^{-k_3 y} \quad (4)$$

$$E_{y3}^{air} = K_z \frac{k_2}{k_3} \frac{\beta}{\omega\epsilon_2} e^{+k_3 a} \sin(k_2 a) e^{-k_3 y} \quad (5)$$

$$H_{x3}^{air} = -K_z \frac{k_2}{k_3} \frac{\epsilon_3}{\epsilon_2} e^{+k_3 a} \sin(k_2 a) e^{-k_3 y}. \quad (6)$$

Here, K_z is an amplitude of tangential component in (3) of the slab, $K_z \propto -(\omega\epsilon_2/k_2)$. The relations between the propagation factor, permittivity, permeability and frequency in both regions are developed from the wave equation. The relations are represented below.

$$k_2^2 = \omega^2 \mu_2 \epsilon_2 - \beta^2 \quad (7)$$

$$(jk_3)^2 = -k_3^2 = \omega^2 \mu_3 \epsilon_3 - \beta^2 \quad (8)$$

$$k_2^2 + k_3^2 = \omega^2 (\mu_2 \epsilon_2 - \mu_3 \epsilon_3). \quad (9)$$

At the boundary, $y = a$, the impedance is continuous, because the tangential components of electric field E and magnetic field H are continuous. From this condition, the propagation constant in the air region is represented as below.

$$k_3 = k_2 \frac{\epsilon_3}{\epsilon_2} \tan(k_2 a). \quad (10)$$

Substituting (10) into (9), we obtain an equation for the propagation constant in the slab.

$$k_2^2 \left[1 + \frac{\epsilon_3^2}{\epsilon_2^2} \tan^2(k_2 a) \right] = \omega^2 (\mu_2 \epsilon_2 - \mu_3 \epsilon_3). \quad (11)$$

Using the above relations, we can solve for β , k_2 and k_3 from (8), (11) and (10). In formulating an equation for the attenuation by the loss properties of the slab, we had to consider the fields in both dielectric and air regions. The attenuation is proportional to the time average loss in the slab and to the time average power of the wave propagating in both regions. To obtain the time average power propagating along the z direction, we determine the time average Poynting vector in both regions.

Hence, in each region, the time average propagating power is derived as below. For the slab,

$$\overline{P_{z2}^{slab}} = +K_z^2 \frac{1}{4k_2} \frac{\beta}{\omega\epsilon_2} \left[k_2 a + \frac{1}{2} \sin(2k_2 a) \right]. \quad (12)$$

For the air region,

$$\overline{P_{z3}^{air}} = +K_z^2 \frac{1}{4} \frac{\epsilon_3}{\epsilon_2} \frac{k_2^2}{k_3^3} \frac{\beta}{\omega\epsilon_2} \sin^2(k_2 a). \quad (13)$$

To obtain the time average loss in the slab including both dielectric and magnetic loss, we calculate real power dissipated in the slab. Hence, the loss of time average propagating power in the slab with dielectric and magnetic loss properties is given below.

$$\overline{P_{loss}^{slab}} = +K_z^2 \frac{\omega\epsilon_2''}{4(\omega\epsilon_2)^2} \left\{ (\beta^2 + k_2^2) a + \frac{\beta^2 - k_2^2}{2k_2} \sin(2k_2 a) \right\} + K_z^2 \frac{\omega\mu_2''}{4} \left\{ \frac{1}{2k_2} \sin(2k_2 a) + a \right\}. \quad (14)$$

Finally, the attenuation of TM surface waves on a lossy slab is given by

$$\alpha = \frac{\overline{P_{loss}^{slab}}}{2 \left(\overline{P_{z2}^{slab}} + \overline{P_{z3}^{air}} \right)}. \quad (15)$$

B. With Impedance Sheet

In the case of applying the impedance sheet to the top of the slab, we assume it is represented by an equivalent transmission line system in which the slab, the impedance sheet and the air region are connected in parallel. To develop an analytical solution for the attenuation with the impedance sheet, we need to calculate modified propagation constants for both the slab and air regions. The transverse resonance method (TRM) [1], [11] has been used to calculate the modified propagation constants with the impedance sheet. The propagation constants that satisfy the TRM are determined by equating the admittance looking down into the slab, and up into the air region, including the susceptance component of the impedance sheet [12]. We can construct an equation at the boundary that is expressed in terms of the field components in both slab and air regions and the susceptance B of Y^{sheet} as shown in Fig. 1.

$$\frac{H_{x3}}{E_{z3}} \Big|_{y=a}^{y=\infty} + B = \frac{H_{x2}}{E_{z2}} \Big|_{y=a}^{y=0}. \quad (16)$$

After applying the previous field components to (16), we obtain the propagation constants, k_2 and k_3 that include properties of the impedance sheet.

$$\frac{\omega \varepsilon_3}{k_3} = \frac{\omega \varepsilon_2}{k_2} \cot(k_2 a) - B. \quad (17)$$

Therefore, the modified propagation constant in the air region is derived from (17). In addition, adding the sheet generates an additional loss term due to the resistance of the sheet at the boundary $y = a$. For the sheet loss, we consider the electric fields in the z direction, parallel to the sheet.

$$\frac{P_{sheet}}{P_{loss}} = \frac{\frac{1}{2} \frac{1}{R_{sheet}} \left[K_z^2 \frac{k_2^2}{(\omega \varepsilon_2)^2} \sin^2(k_2 a) \right]}{2 (\bar{P}_{z2} + \bar{P}_{z3})}. \quad (18)$$

Therefore, the sheet loss is linearly added to the attenuation, α , as given below.

$$\alpha = \frac{\frac{P_{slab}}{P_{loss}} + \frac{P_{sheet}}{P_{loss}}}{2 \left(\frac{P_{z2}}{P_{slab}} + \frac{P_{z3}}{P_{air}} \right)}. \quad (19)$$

By applying the TRM to calculate the modified propagation constants, we now have an expression for the attenuation which includes the properties of the impedance sheet on the top of the slab. We assume that the dielectric slab and ground plane provide inductance, so our analysis focuses on sheets having resistance and capacitance, and we consider both series and parallel RC circuits for the impedance sheet.

III. NUMERICAL SOLUTION

As discussed above, analytical solutions have been developed for TM surface wave attenuation by a lossy slab and an impedance sheet. To validate the analytical solutions, we compare the results to numerical solutions. In obtaining the numerical solutions, we simulated the same structure as shown in Fig. 1 using Ansoft HFSS. To determine the attenuation of TM surface waves, we calculated the complex dispersion curves for TM waves using an eigenmode solver in HFSS. We also tried various approaches using a driven modal solution, but the eigenmode solution provided results that were the cleanest and easiest to interpret. We simulated a small piece (2 mm length by 10 mm width) of the slab and impedance sheet as shown in Fig. 2. In the $x - y$ plane, we applied periodic boundary conditions (master and slave boundaries). We then enter a phase difference across the structure corresponding to a wave vector, and the eigenmode solver provides the corresponding frequency of that TM surface wave mode. In the $y - z$ plane, we applied perfect H boundaries because we are simulating TM surface waves. In $x - z$ plane, the bottom and top planes are perfect E boundaries with 50 mm height between the planes including 10 mm height of the slab. At the surface of the dielectric slab, we applied an impedance sheet or lumped RLC boundary. In the slab, we used a material with relative

Fig. 2. Schematic for the numerical solutions and assigned boundaries. In the $x - y$ plane, master and slave boundaries are applied, in the $y - z$ plane, perfect magnetic conductor (PMC) boundaries are applied, in the $x - z$ plane, perfect electric conductor (PEC) boundaries are applied.

complex permittivity and permeability. The relative complex values, $\varepsilon_r = 10 + j0.001$ and $\mu_r = 1 + j0.001$ are used in most simulations. Also, we examined a wide variety of material properties in this study. At the impedance sheet, using the lumped RLC boundary, a range of resistance and capacitance were studied. As part of the eigenmode solution, we also obtain the quality factor Q of each mode which is a function of the real and imaginary parts of frequency, ω and ω'' .

$$Q = \left| \frac{\omega'}{2\omega''} \right|. \quad (20)$$

This definition of Q indicates that it is a measure of how quickly the mode would decay in time if it were excited uniformly across the surface at some starting time, and then allowed to decay without additional excitation. However, what we need is the rate of decay along the surface for a wave starting at one end with continuous excitation. These two quantities are related through the group velocity of the wave, which itself is a function of frequency.

$$\alpha = \frac{\omega''}{v_g} = \omega'' \frac{dk}{d\omega}. \quad (21)$$

Here, k is the wave number, and ω and ω'' represent real and imaginary angular frequencies, respectively. The magnitude of electric field decreases along the length of the surface by $E \propto e^{-jkz}$. Therefore, the modified wave number can be expressed by $k = \phi/z$. Here z is the length of the structure in our simulation, and ϕ is the phase difference applied to the periodic boundary conditions in the z -direction. Substituting (20) and the modified wave number into (21) gives the equation for attenuation using calculated values in the HFSS eigenmode solver, and then we numerically differentiate to obtain

$$\alpha = \omega'' \frac{dk}{d\omega} = \frac{\omega'}{2Q} \frac{1}{z} \frac{d\phi}{d\omega} = \frac{\omega_{n+1}}{2Q} \frac{1}{z} \frac{\phi_{n+1} - \phi_n}{\omega_{n+1} - \omega_n}. \quad (22)$$

Where n and $n + 1$ represent two successive simulations.

IV. RESULTS

A. Validation of Analytical Solution

In previous sections, we discussed the analytical and numerical solutions, separately. To verify the analytical solutions, we have performed a detailed comparison with the numerical solutions from HFSS simulations. The matching plot shown in Fig. 3 is representative of a range of different conditions. In the plot, the impedance sheet is modeled as a series RC circuit, but we have also checked the parallel case for consistency. Multiple peaks in the plot correspond to different TM surface waves and are labeled TM_0 , TM_1 , etc. In the plot, the solid lines

Fig. 3. Matching plot between the analytical and numerical solutions with an impedance sheet on the slab of dielectric loss 0.001 and magnetic loss 0.001 with resistance of $377 \Omega/\text{square}$ and reactance of $1 \Omega/\text{square}$, respectively.

Fig. 4. Envelope limit line for the dielectric slab without any impedance sheet. Properties of the dielectric slab are the relative permittivity 10, permeability 2, dielectric loss tangent 0.001 and magnetic loss tangent 0.001 with thickness 1 cm. The unit factor, U , of the enveloped limit line is 1.2.

represent the analytical solutions and the dots represent the numerical solutions. Because of the close agreement between the analytical and numerical solutions for a wide range of properties for the slab and the impedance sheet, we are convinced that the analytical solution for the attenuation is formulated correctly.

B. Theoretical Limitations

After having validated our analytical model, we now explore a wide range of material and impedance sheet properties to identify trends and limits. For the following discussion, we initially consider the case where there is no impedance sheet on the slab. In every case we find that the attenuation shows a series of peaks corresponding to the surface wave modes, and that these peaks form an envelope that can be approximated by a linear function that is consistent over a wide range of material properties. An example is shown in Fig. 4. We designated the line as the 'envelope limit line' which is given by

$$\alpha \cong U \cdot f \cdot c \cdot \varepsilon_2 \cdot \mu_2 \cdot \tan \delta. \quad (23)$$

$$\tan \delta = \frac{\mu_2 \varepsilon_2'' + \mu_2' \varepsilon_2}{\mu_2 \varepsilon_2 - \mu_2'' \varepsilon_2'}. \quad (24)$$

Here, f and c represent frequency and speed of light in free space, and loss tangent, $\tan \delta$ expressed in (24) [13], represents the combined dielectric and magnetic loss properties. In (23), there is a factor, U , which has a value that is approximately unity for a wide range of material properties, and can be considered to be unity as a simple guideline for estimating attenuation.

Fig. 5. Comparison of attenuation values when adding an impedance sheet to the dielectric slab: the dashed line shows the envelope limit line with $U = 1.3$. the solid and the dot dashed lines show parallel and series impedance sheets, respectively, with resistance of $1000 \Omega/\text{square}$ and capacitance of $0.1 \text{ pF}/\text{square}$. Note that we have only plotted the first of a series of attenuation peaks.

Fig. 6. Peak frequency as a function of the capacitance of an impedance sheet in parallel connection of resistance and capacitance values. The dot dashed line, dashed line, and solid line represent different sheet resistances of 100, 377 and $1000 \Omega/\text{square}$, respectively.

Even under a broad range of loss tangents from 0.001 to 1.0, U is still within this range. When the slab is electrically thick, the attenuation deviates from (23), but only by a small amount. Thus, (23) provides a valuable guideline for estimating attenuation. It is noteworthy that the thickness does not affect the slope of the envelope limit line, although the thickness does determine the frequency of the first mode, and the frequency spacing of the modes.

If an impedance sheet is added to the slab, the attenuation can be substantially greater than the envelope limit line as shown in Fig. 5, but over limited bandwidth. The peak frequency is determined by the capacitance of the sheet and thickness of the slab, as shown in Figs. 6 and 7, while the maximum attenuation value is determined by the resistance of the sheet, and the thickness and loss parameters of the slab. Increasing the slab thickness reduces the peak frequency, but also reduces the absolute bandwidth of the peak, as shown in Fig. 7, as measured by the full width of half maximum (FWHM).

Thus, the relative bandwidth, as determined by the absolute bandwidth at the FWHM divided by the peak frequency, is essentially independent of the slab thickness and the properties of the sheet, as shown in Fig. 8. The permeability and permittivity of the slab affect the relative bandwidth in different ways, as shown in Fig. 9.

Increasing either of these parameters will reduce the peak absorption frequency. However, increasing permeability does so without affecting the relative bandwidth, which increasing permittivity substantially reduces relative bandwidth. This is consistent with previous studies of

Fig. 7. Attenuation value for different thickness of the slab with an impedance sheet having resistance of $377 \Omega/\text{square}$ and capacitance of $0.1 \text{ pF}/\text{square}$. The thicknesses of the slab are 0.5 and 2.0 cm, respectively.

Fig. 8. Relative bandwidth for difference thicknesses of the slab as a function of capacitance. This is for an impedance sheet represented by a parallel RC circuit.

Fig. 9. Relative bandwidth for different slab properties as a function of relative permittivity or permeability up to 100. The solid, dot dashed and dashed lines are for swept relative permeability with the fixed permittivity of 2, 5 and 10. The thick solid line represents a swept relative permittivity with a permeability of 10.

absorption of normally incident waves [8], where permeability has a beneficial effect on bandwidth, but permittivity does not.

Although the resistance of the sheet does not significantly affect the peak frequency or bandwidth, it does affect the maximum attenuation value. The attenuation values of the peak of the first curve are related with the resistance of the impedance sheet as shown in Fig. 10. In the plot, the value of the attenuation decreases with increasing resistance. Additionally, the curves for the attenuation values with the impedance

Fig. 10. Peak attenuation values as a function of resistance of the impedance sheet, which is represented by a parallel RC circuit. The solid line and dot dashed line represent different capacitances, $0.1 \text{ pF}/\text{square}$ and $1.0 \text{ pF}/\text{square}$, respectively. The dashed line represents the envelope limit line for the slab alone.

sheet are compared to a dashed straight line corresponding to the envelope limit line of the slab only. Thus, the impedance sheet can enhance the attenuation compared to the slab alone, but at the expense of bandwidth.

V. CONCLUSION

We have developed an analytical model for TM surface wave attenuation by grounded lossy slabs including both dielectric and magnetic losses. Using the transverse resonance method we have extended the solution to include an impedance sheet which may represent a frequency selective surface, metasurface or other periodic structure. We find that for the case of the bare slab, the attenuation can be approximated by a simple formula for a wide range of material properties. By making practical assumptions about available materials, this formula can be used to estimate the limits on absorption for homogeneous linear lossy slabs. We have also found that the use of an impedance sheet can provide substantially greater attenuation than what is possible with a simple lossy slab, but with limited bandwidth. For an impedance sheet that can be described in terms of effective sheet capacitance and sheet resistance, the capacitance and the slab thickness primarily determine the frequency of the attenuation peak, and the resistance determines the maximum value of attenuation. Although the absolute bandwidth is determined by both the capacitance of the sheet and the thickness of the slab, the relative bandwidth is nearly independent of all of these parameters. Furthermore, increasing permittivity to reduce the peak absorption frequency also reduces relative bandwidth, whereas increasing permeability does not. These results can be used as a guideline for the development of broadband or narrowband coatings for attenuation of TM surface waves.

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A Dual Circularly Polarized Waveguide Antenna With Bidirectional Radiations of the Same Sense

Yang Zhao, Zhijun Zhang, Kunpeng Wei, and Zhenghe Feng

Abstract—A dual circularly polarized waveguide antenna with bidirectional radiations of the same sense is proposed in this communication. Bidirectional circular polarization (Bi-CP) of the same sense was obtained by two identical rectangular metal slices installed on one lateral side of the waveguide with an intersection angle of 45° . These two metal slices were horizontally perpendicular to each other and vertically spaced by a quarter guided wavelength. A rat-race coupler was employed to excite the two metal slices with the same amplitude, but with a 0° or 180° phase difference depending on the selection of two inputs. One sense of Bi-CP was realized when the two metal slices were fed in phase and the opposite sense of Bi-CP could be obtained when they were fed out of phase. A prototype for 2.4-GHz WLAN application was tested to verify our design. The measured common bandwidth for 10-dB return loss and 3-dB axial ratio at the two feed ports was 230 MHz (9.6%, 2.29–2.52 GHz) and 210 MHz (8.6%, 2.35–2.56 GHz), respectively. The measured isolation between the two feed ports was better than 30 dB over the whole operating band.

Index Terms—Bidirectional, circular polarization, dual sense, same sense.

I. INTRODUCTION

Circularly polarized antennas are usually required in many wireless communication systems because they can allow the avoidance of polarization alignment between the transmitting (Tx) and receiving (Rx) terminals and thus enhance link consistency, which is advantageous

in comparison to linearly polarized ones. In principle, circularly polarized waves are usually excited by two orthogonal linearly polarized components with the same amplitude, but with a 90° phase difference. The practical designs of antenna with circular polarization (CP) include single-fed method with structural perturbation, dual-fed technique, and structural rotation of several linearly polarized elements [1]–[3], etc. Meanwhile, a single antenna operating with both senses of CP [4]–[8] is much preferred for the reason that it can integrate the traditionally separated Tx and Rx antennas into one and thus reduce system complexity. Besides, the adoption of two orthogonal signals with dual senses of CP, in this case left-hand CP (LHCP) and right-hand CP (RHCP), also allows a more efficient utilization of the electromagnetic spectrum and increases the link capacity. A patch antenna coupled by two crossed slots with four arms serially fed by a single microstrip line was proposed in [4] to realize dual senses of CP through the excitation at either end of the microstrip line. A planar dual circularly polarized array was presented in [5] whose element comprises an arc-shaped patch and an extended stub, and it can radiate LHCP or RHCP depending on the feed port that was selected. An open-end waveguide integrated with a septum polarizer was proposed in [6]–[8] to achieve dual CP senses, and two individual coaxial ports were located at the two sides of the septum for both senses of CP. But all these designs focus on unidirectional radiation with both senses of CP.

On the other hand, bidirectional CP (Bi-CP) is needed in various modern wireless applications, such as a microcellular base station, a radio frequency identification system, an indoor high-speed WLAN, a coal mine channel, a bridge or tunnel communication, etc. Slot radiators are good candidates for Bi-CP and have been studied by many researchers. However, the existing design will produce CP of opposite senses in the two different radiating directions [9]–[11]. This means that if an LHCP is produced at one end, RHCP would unavoidably be generated at the opposite end. No information can be exchanged if the polarizations are mismatched in that opposite direction. For the realization of CP of the same sense in the two opposite directions, a back-to-back arrangement by two circularly polarized patch elements was proposed in [12]–[14]. In [12], a slot-coupling feeding approach was utilized to excite two back-to-back corner-cutting circularly polarized patches. A coplanar waveguide feeding structure was employed in [13], [14] to excite two back-to-back partially overlapped rectangular patches for Bi-CP of the same sense. However, these designs are realized just by replicating and superposing unidirectional circularly polarized elements located at the front and back sides of a common feeding scheme, and they have insufficient axial ratio bandwidth. A waveguide antenna with Bi-CP of the same sense was designed in [15], but only one sense of CP was available. In addition, the realization of dual senses of CP with bidirectional radiations of the same sense has not been published in literature.

In this communication, we propose a novel design of dual circularly polarized waveguide antenna with bidirectional radiations of the same sense for either CP. Both senses of CP are obtained for a single antenna by selecting two different input ports. The proposed structure incorporated two rectangular metal slices installed inside a square-aperture waveguide, as that proposed in [15]. These two metal slices were perpendicular to each other in the horizontal plane, and separated by a quarter guided wavelength in the vertical direction. For the realization of dual senses of CP, a rat-race coupler was used in this study to feed the two metal slices along their diagonal line with the same amplitude, but with a 0° or 180° phase difference. When the two metal slices were fed with the same phase, one sense of Bi-CP was realized, but Bi-CP of the opposite sense could be obtained when they were fed with anti-phase. The design details are presented in the following sections.

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