Single-Conductor Strip Leaky-Wave Antenna

Wanchu Hong and Yu-De Lin

***Abstract*—This paper presents a leaky-wave antenna with only a single-conductor strip on a substrate without a practical ground plane. The full-wave integral equation method is used to investigate the EH_{01} leaky mode of this single-conductor strip structure. When this single-conductor strip is on a thin substrate with a low dielectric constant, a broad-band radiation regime can exist for the EH_{01} leaky mode. The balanced microstrip lines and the inverted balanced microstrip lines are used to excite this EH_{01} leaky mode. This work presents both the numerically simulated and measured data. The measured bandwidth of a voltage standing-wave ratio $\leqq 2$ is from 6.55–13.75 GHz (2.34:1). In this case, the normalized phase constant is very close to 1 and this results in a fixed main-beam radiation pattern in the end-fire direction. This leaky-wave antenna with only a single rectangle stripline of length 110 mm can achieve the same broad-band performance as a tapered microstrip leaky wave antenna with a length 310 mm, as previously reported by the authors.**

***Index Terms*—Broad-band antenna, leaky-wave, microstrip line balun, single-conductor strip.**

I. INTRODUCTION

THE EXISTENCE of the leaky modes on various open planar transmission lines such as the microstrip line, the slot line, the strip line, the coplanar waveguide and the coplanar strips, is well known and has received considerable attention [1]–[12]. This is owing to that inadvertent excitation of these leaky modes may incur undesirable radiation and crosstalk in microwave circuits. Most open transmission lines can be used as leaky-wave antennas based on the space-wave leakage of the higher order leaky modes [2]–[9]. For such planar open transmission lines, the space-wave radiation bandwidth of leaky modes is around 20% and is limited by the bandwidth of the feeding network and the dielectric constant of the substrate. For example, the center-fed aperture-coupled leaky wave microstrip antenna with a uniform strip width has a bandwidth of only 23% [6]. For wireless broad-band communication applications, a leaky-wave antenna with a wider radiation bandwidth could be useful. To enhance the bandwidth, a tapered microstrip leaky-wave antenna is developed in [7]–[9]. Because this broad bandwidth is a composition of several individual leaky-wave regimes radiated from different strip widths, the antenna length is much longer than a single uniform microstrip antenna [9].

This paper presents a novel leaky-wave broad-band antenna that has only a single conductor strip on a substrate but no ground plane. That absence of a ground plane allows both TE_0 and TM_0 surface-wave modes to exist in this single-conductor strip structure, causing some interesting characteristics to be exhibited. The rest of this paper is organized as follows. Section II applies the integral equation method in the spectral

Manuscript received October 24, 2002; revised June 9, 2003. This work was supported in part by the MOE Program for Promoting Academic Excellence of Universities under Grant 89-E-FA06-2-4 and in part by the National Science Council under Grant NSC-91-2213-E-009-130.

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Digital Object Identifier 10.1109/TAP.2004.829854

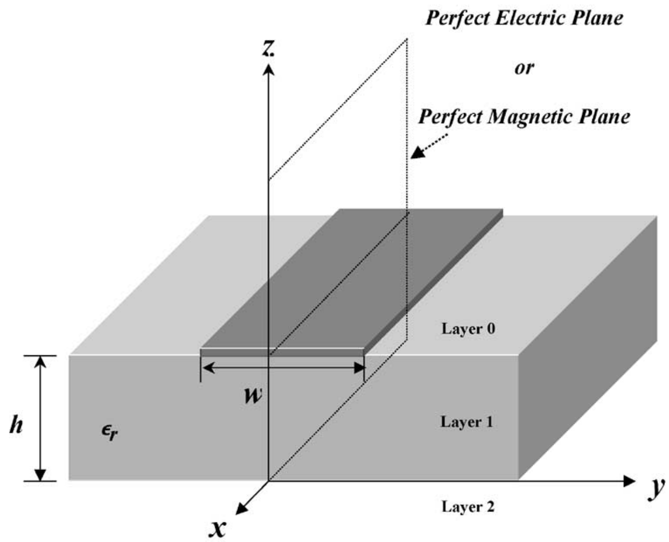

Fig. 1. Single-conductor strip structure. The substrate is denoted as Layer 1, while the air layers above and below the substrate are denoted as Layer 0 and Layer 2, respectively. The vertical plane denoted by dark dotted line indicates the virtual perfect electrical wall or perfect magnetic wall according to the odd or even modes analysis.

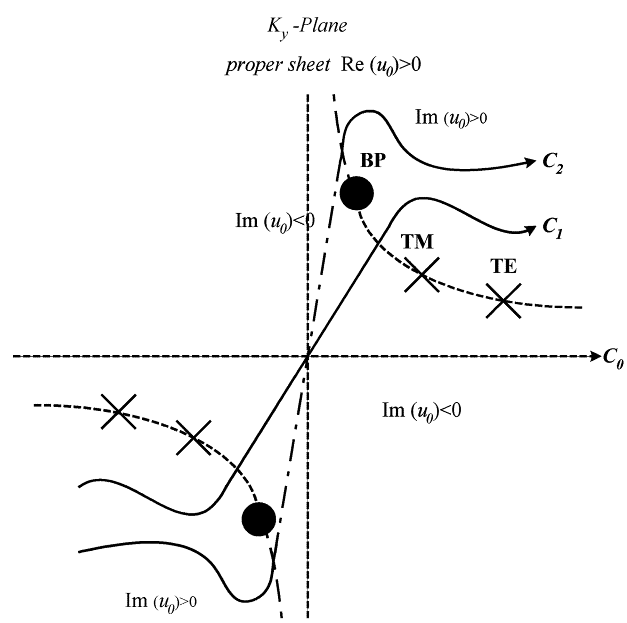

Fig. 2. Integral paths of the inverse Fourier transform on the ky plane. •: branch points; ×: proper surface poles (TE_0, TM_0); dashed lines: branch cuts; dashed-dotted line: path on the improper sheet.

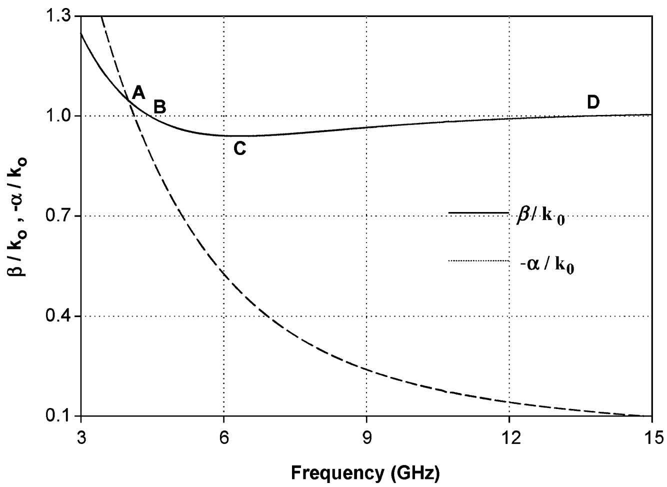

Fig. 3. Normalized phase constant β/k_0 and the normalized attenuation constant $-\alpha/k_0$ of the EH_{01} leaky-wave mode of the single-conductor strip leaky-wave antenna ($w = 20$ mm, $h = 0.508$ mm, $\varepsilon_r = 2.2$). Four points are indicated; the intercept point A: $(\beta/k_0, \mathrm{freq}) = (1.047, 4.0)$; crossing point B: $(\beta/k_0, \mathrm{freq}) = (1.000, 4.5)$; minimum point C: $(\beta/k_0, \mathrm{freq}) = (0.9397, 6.3)$; end point D: $(\beta/k_0, \mathrm{freq}) = (1.000, 13.7)$.

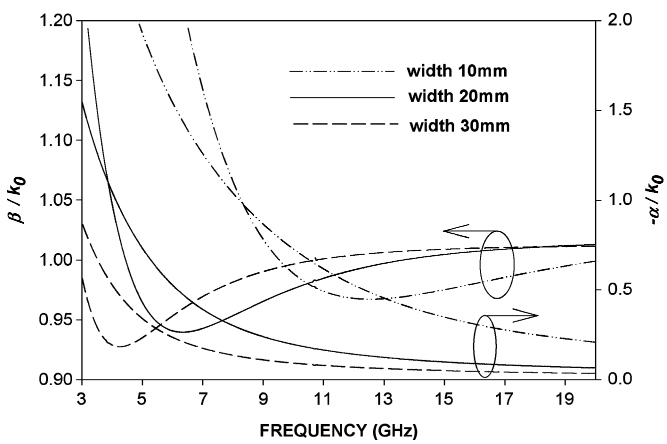

Fig. 4. Normalized phase constant β/k_0 and the normalized attenuation constant $-\alpha/k_0$ of the first higher order leaky mode for different strip widths ($h = 0.508$ mm, $\varepsilon_r = 2.2$).

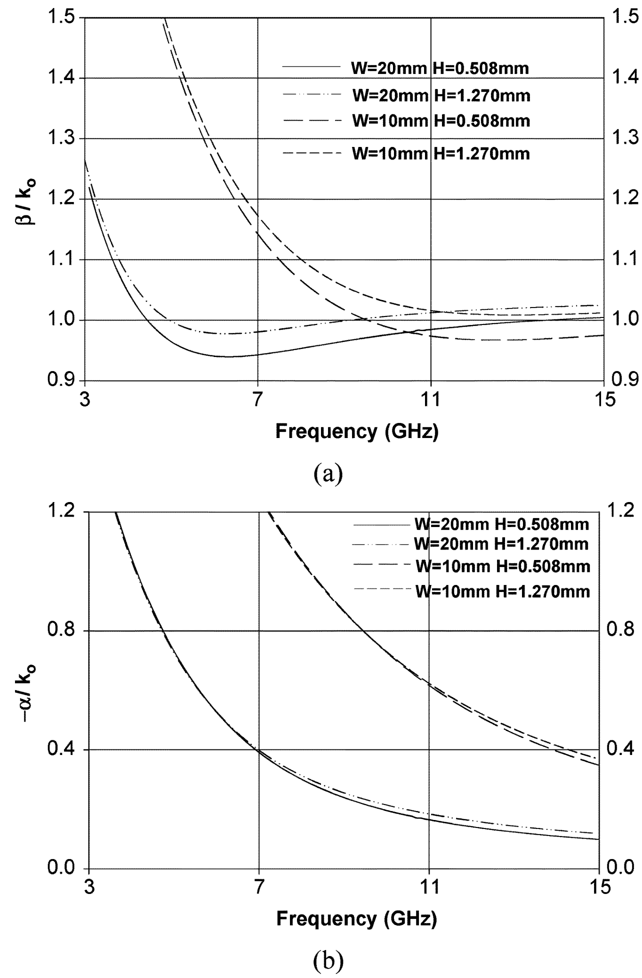

Fig. 5. Normalized propagation constant of the structure with $\varepsilon_r = 2.2$ for different strip widths and substrate thicknesses. (a) The normalized phase constant (b) the normalized attenuation constant.

domain and the method of moments to analyze this structure. The numerical results indicate that the EH_{01} leaky mode of this single-conductor strip structure on a thin substrate with a low dielectric constant has a broad-band radiation regime. Section III proposes a broad-band planar feeding structure to excite properly this EH_{01} leaky mode. This broad-band planar feeding structure consists of the balanced microstrip lines and the inverted balanced microstrip lines. Section IV presents the measurement results of this single-conductor strip leaky-wave antenna to verify the broad-band nature of the EH_{01}leaky mode.

II. Analysis and Numerical Results

Fig. 1 shows the structure of the single-conductor strip leaky-wave antenna. Only one perfect conducting strip of width w is located at the interface between a semi-infinite air layer and an isotropic lossless substrate, with a dielectric constant of ε_r and a thickness of h. In this structure, the strip is uniform along the longitudinal x directoin and no practical ground metal plane is placed on the substrate. This single-conductor strip structure is analyzed based on the integral equation formulation and the dyadic Green's function in the spectral domain [4], [5].

A. Integral Equation Formulation

To simplify the analysis, the structure is assumed to be infinite in both the x and y directions, and the conducting strip is assumed to be infinitesimally thin, such that the surface current density J_s on the strip has components in the x and y directions only. The structural symmetry is such that a wave's propagation along such a structure can be expressed as even or odd modes corresponding to even or odd symmetry about the plane that can be replaced by a perfect magnetic or electric wall in the center of the conducting strip for the purpose of analysis. Hence, the longitudinal currents on this strip are forced to be evenly or oddly symmetric according to which mode is under consideration.

For fields that propagate in the positive x direction, the surface current J_s can be expressed as

$$J_s = [\hat{x}J_x(y) + \hat{y}J_y(y)]\, e^{-jk_x x} \tag{1}$$

where $J_x(y)$ and $J_y(y)$ represent the x-directed and y-directed currents, respectively. The complex propagation constant is

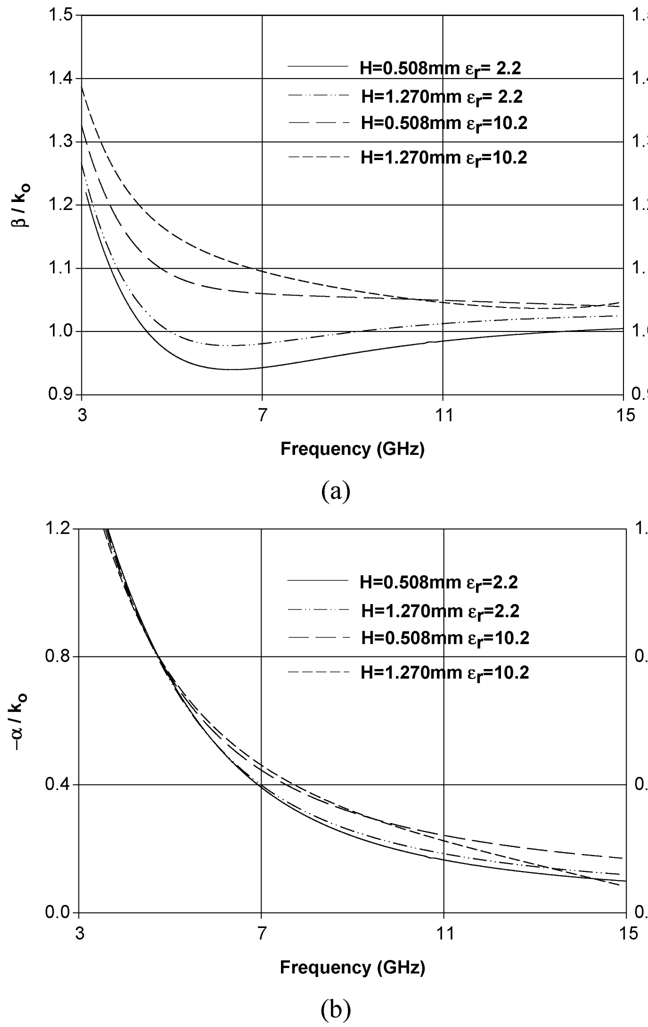

Fig. 6. Normalized propagation constant of the structure with $w = 20$ mm for different dielectric thicknesses and dielectric constants. (a) The normalized phase constant (b) the normalized attenuation constant.

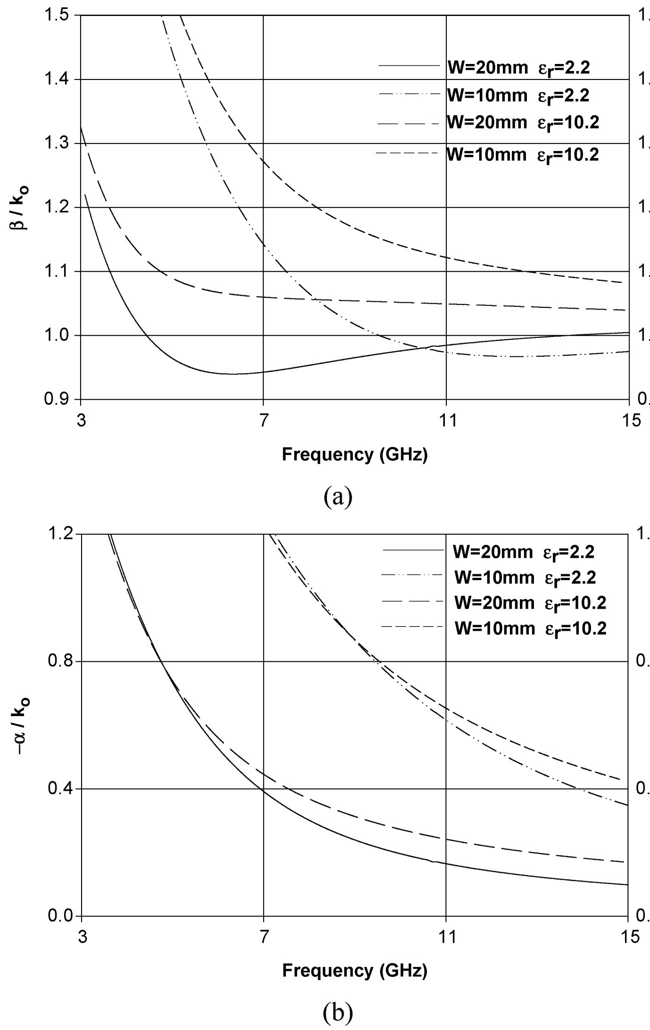

Fig. 7. Normalized propagation constants of the structure with the dielectric thickness $h = 0.508$ mm for different strip widths and dielectric constants (a) the normalized phase constant (b) the normalized attenuation constant.

$k_x = \beta - j\alpha$, where β is the phase constant and α is the attenuation constant. The terms $J_x(y)$ and $J_y(y)$ can be expanded approximately as linear combinations of known basis functions

$$J_x(x) \simeq \sum_{n=1}^{N} a_n J_{xn}(x) \quad J_y(x) \simeq \sum_{m=1}^{M} b_m J_{ym}(x) \tag{2}$$

where a_n and b_m are unknown current coefficients. The choice of the expansion functions should correspond to the odd or even symmetry of the currents for the mode of interest. The electric field can be represented by a dyadic Green's function and the surface current density $\boldsymbol{J}_s$ on the strip as [4], [5]

$$E(x,y,z) = \frac{1}{2\pi} \int_{\frac{-w}{2}}^{\frac{w}{2}} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \widetilde{G}_{EJ}(k_x,k_y,z) e^{jk_x(x-x')} \times e^{jk_y(y-y')} dk_x dk_y J_s(x',y') dx' dy'. \tag{3}$$

For this single-conductor strip structure, the dyadic Green's function can be expressed as

$$\tilde{G}_{EJ}^{ij} = \left[\frac{f_e^{ij}(k_x,k_y)}{D_e(k_{yp})} + \frac{f_m^{ij}(k_x,k_y)}{D_m(k_{yp})} \right] e^{-u_0(z-h)} \tag{4}$$

$$u_0 = \sqrt{k_y^2 + k_x^2 - k_0^2} \tag{5}$$

$$k_{yp} = \sqrt{k_s^2 - k_x^2} \tag{6}$$

where k_0 is the wavenumber in free-space and k_s is the wavenumber associated with the surface wave mode of the dielectric slab structure and i and j indicate the x and y directions, respectively. The Appendix lists the dyadic Green functions for $z \geqq h$. The other the dyadic Green functions are presented in . The complex propagation constant k_x along the uniform strip can be determined by solving the determinant equation derived from the Galerkin moment method, under the boundary condition of a zero tangential electric field on the strip.

The choice of the integration path used to determine the complex propagation constant, k_x, depends on the physical interpretation of the integration path since the path of integration is an

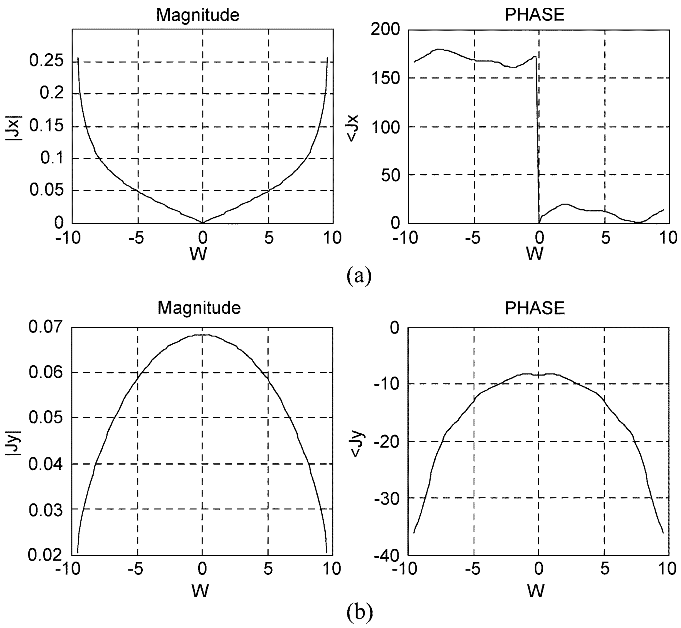

Fig. 8. Current distributions of the first higher order mode at 10 GHz: (a) the longitudinal currents (b) the transverse currents ($w = 20$ mm, $h = 0.508$ mm, $\varepsilon_r = 2.2$).

open contour in the k_y plane from $-\infty$ to ∞ [4], [5]. Both the TE_0 and the TM_0 surface wave modes exist for a dielectric slab structure, so two proper surface-wave poles of k_{yp} are present in the k_y plane. As present in Fig. 2, the path C_2 is used to determine the solution for the leaky-wave mode that energy leaks into both the space wave and the surface wave [4], [5]. This path lies partly on the improper sheet of the k_y plane in the region between the branch cuts and includes both proper surface-wave poles for the TE_0 and the TM_0 modes.

B. Numerical Results

Consider the single-conductor strip structure in Fig. 1, with a strip width of $w = 20$ mm, a substrate of dielectric constant of $\varepsilon_r = 2.2$ and a thickness of $h = 0.508$ mm. Fig. 3 plots the normalized phase constant β/k_0 and the normalized attenuation constant $-\alpha/k_0$ against the frequency of the EH_{01} leaky mode. As shown in the figure, the normalized phase constant β/k_0 declines from 1.22 at 3.0 GHz to a minimum of 0.9397 at 6.3 GHz, and then increases very slowly to 1 at 13.7 GHz. In the microstrip line structure, the leaky wave radiation begins at the frequency at which the real propagated power starts to exceed the imaginary propagated power, and ends when $\beta/k_0 \geqq 1$ [5]. In this single-conductor strip structure, the intercept point of $\beta/k_0 = -\alpha/k_0 = 1.047$ is at 4.0 GHz, but at this frequency, the normalized phase constant β/k_0 (and the attenuation constant $-\alpha/k_0$) exceeds not only unity one but also those of both the TE_0 surface mode (1.0003) and the TM_0 surface mode (1.0001). The space-wave radiation regime of this EH_{01} leaky-wave mode under the condition of $\beta/k_0 \leqq 1$ is from 4.5 GHz to 13.8 GHz. Fig. 4 plots the normalized propagation constants of the EH_{01} leaky mode against frequency for strips of various widths. Because of the resonance of the transverse currents, the leaky regime obviously shifts to a higher frequency for a narrower strip. Also, a narrower strip increases of the attenuation constant. When the thickness of the substrate increases, as in Figs. 5 and 6, the normalized phase constant (β/k_0) increases, since the effective strip width is reduced by the fringing

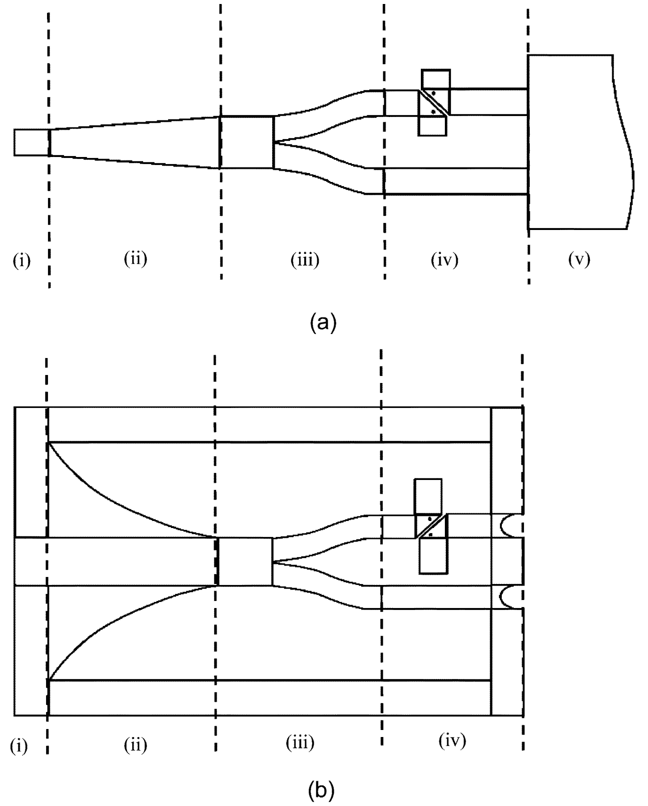

Fig. 9. Diagram of the broad-band balun structure used to excite the first higher order leaky mode of the single-conductor strip antenna. This balun structure consists of a closed loop that connects back to the original ground plane ($h = 0.508$ mm, $\varepsilon_r = 2.2$). This feeding structure consists of (i) a conventional microstrip line; (ii) a microstrip-to-balanced-microstrip-line transition; (iii) a balanced-microstrip-line T-junction power divider; (iv) one set of the balanced microstrip lines is changed to form the inverted balanced microstrip lines and returned back to the original ground plane of the microstrip line; (v) single-strip antenna. (a) On the upper substrate; (b) on the lower substrate side.

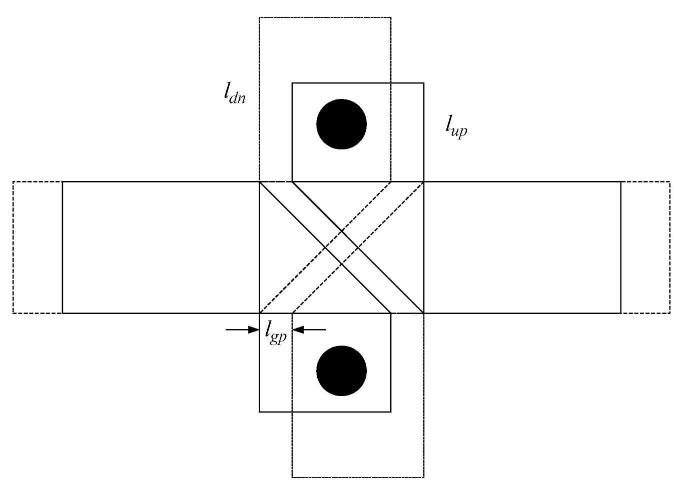

Fig. 10. Inverted balanced microstrip line. The length of the bent stubs: $l_{up} = 1.60$ mm and $l_{dn} = 2.75$ mm; the gap width: $l_{gp} = 0.15$ mm; the radius of vias is 0.3 mm; the slanted angle is 45 degrees. The black circles: the vias; the solid lines: the strips on the upper substrate side; dashed lines: the strips on the lower substrate side.

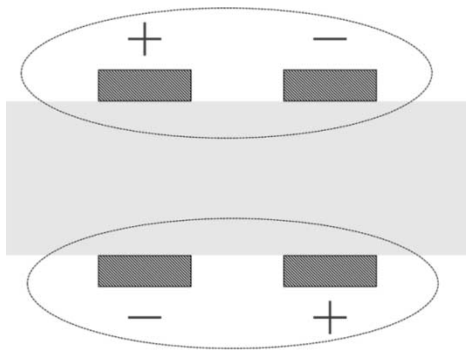

Fig. 11. Two pairs of the broad-band planar baluns on the upper and lower substrate sides, respectively.

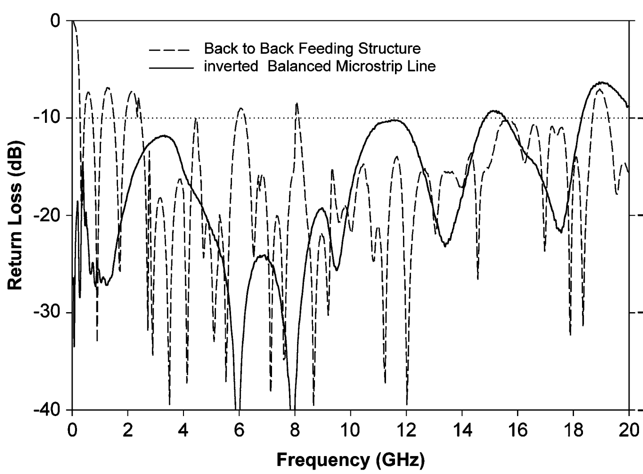

Fig. 12. Measured return loss. Solid line: the back-to back feeding structure; dashed line: the inverted balanced microstrip line.

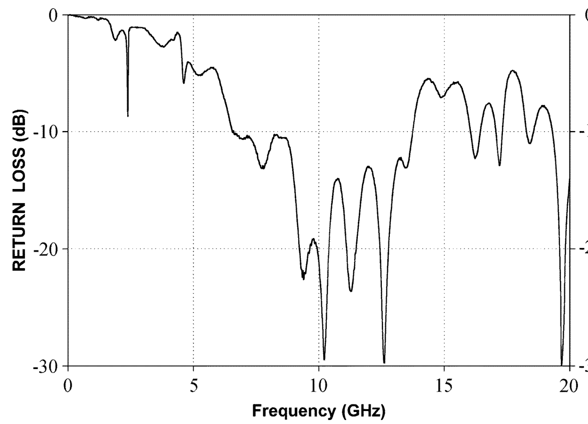

Fig. 13. Measured return loss of the single-conductor strip antenna.

effect. As shown in Figs. 6 and 7, a substrate with a high dielectric constant will rapidly increase the normalized phase constant. If the normalized phase constant always exceeds unity one, the leaky wave radiation may not occur. Most of the energy is focused under the strip when the dielectric constant is high or the substrate is thick. The energy leaks out to the air much more easily if the thickness or the dielectric constant of the substrate is lower or smaller. Increasing the dielectric constant of the substrate also increase the attenuation constant. The attenuation constant of a thin substrate is less than that of a thick substrate but the difference is small. Fig. 8 plots the longitudinal and transverse current distributions of the EH_{01} leaky mode. On the conductor strip, the distributions of the longitudinal currents are in the opposite direction and those of transverse currents are in the same directions. These current distributions confirm that the mode is the EH_{01} mode, and help in designing an appropriate feeding structure for this EH_{01} mode.

III. Broad-Band Feeding Structure

A broad-band planar feeding structure based on the balanced microstrip lines and the inverted balanced microstrip lines is developed to feed this single-conductor strip and thus excite the EH_{01} leaky mode. As shown in Fig. 9, gradually tapering the ground plane to a width equal to the strip width w makes conventional microstrip line a balanced microstrip line, with a strip of positive voltage on the upper side of the substrate and a strip of negative voltage on the lower side of the substrate [8], [11]. The position of the positive strip on the upper side of the substrate of the inverted balanced microstrip line structure is exchanged with that of the negative strip on the lower side of the substrate after a microstrip phase inverter illustrated in Fig. 10. Each of these positive and negative strips is terminated with a chamfered right-angled bend in an opposite direction. The subsequent strip is also headed with a chamfered right-angled bend but in the other direction. A slanted gap separates two strips on the same side of the substrate. The positive strip on the upper substrate side is connected vertically through a cylindrical via to the subsequent strip on the lower side of the substrate. The negative strip on the lower side is similarly connected to the subsequent strip on the upper side. Hence, the positions of the positive and negative strips alternate. The bent stubs l_{up}, and l_{dn} on the upper and lower sides of the substrate, respectively, can be used to compensate for the reactance induced by the via holes and the slanted gaps, and may have different lengths. The details of the design method for such an microstrip phase inverter structure are presented in [8]. Using a T-junction balanced-microstrip-line power divider with inverted balanced microstrip lines substituted for balanced microstrip lines in one of the two output ports, two pairs of broad-band planar baluns can be formed as shown in Fig. 11. No background metal plane is located beneath the single-conductor strip, so the balun on the lower substrate side may be floating, resulting in a disturbed radiation pattern. Fig. 9 depicts a method for preventing such disturbed radiation. The strips of the balun on the lower side of the substrate are connected to each other and returned back to the original ground plane of the microstrip line. Hence, only the balun on the upper side of the substrate feeds the single-conductor strip and a closed metal loop is formed on the lower side of the substrate. On this closed metal loop, two semicircles beneath the two feeding strips are etched out with diameters equal to the widths of the feeding strips, respectively, to enhance the electrical field transition from feeding points to the single-conductor strip. Fig. 12 indicates that the back-to-back measured bandwidth of this feeding structure exceeds 6:1 from 2.5–15.2 GHz for a voltage standing-wave ratio (VSWR) $\leqq 2$.

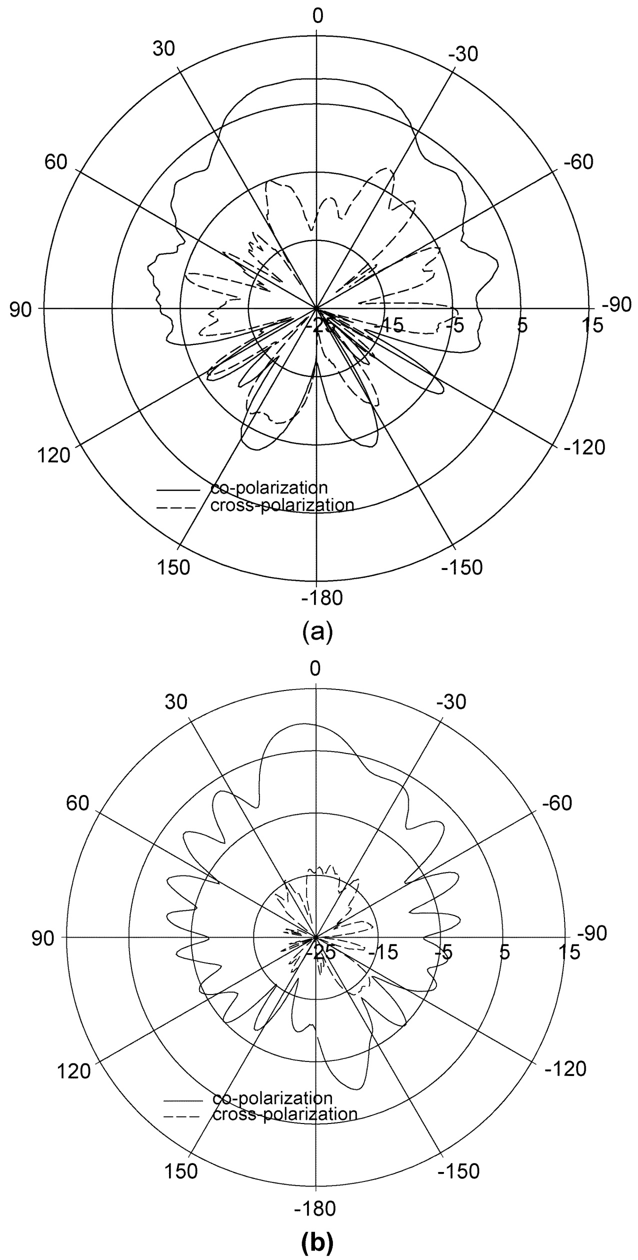

Fig. 14. Measured copolarization and cross-polarization radiation patterns of (a) H plane ($x - z$ plane) and (b) E plane ($x - y$ plane) E_ϕ fields at 11 GHz.

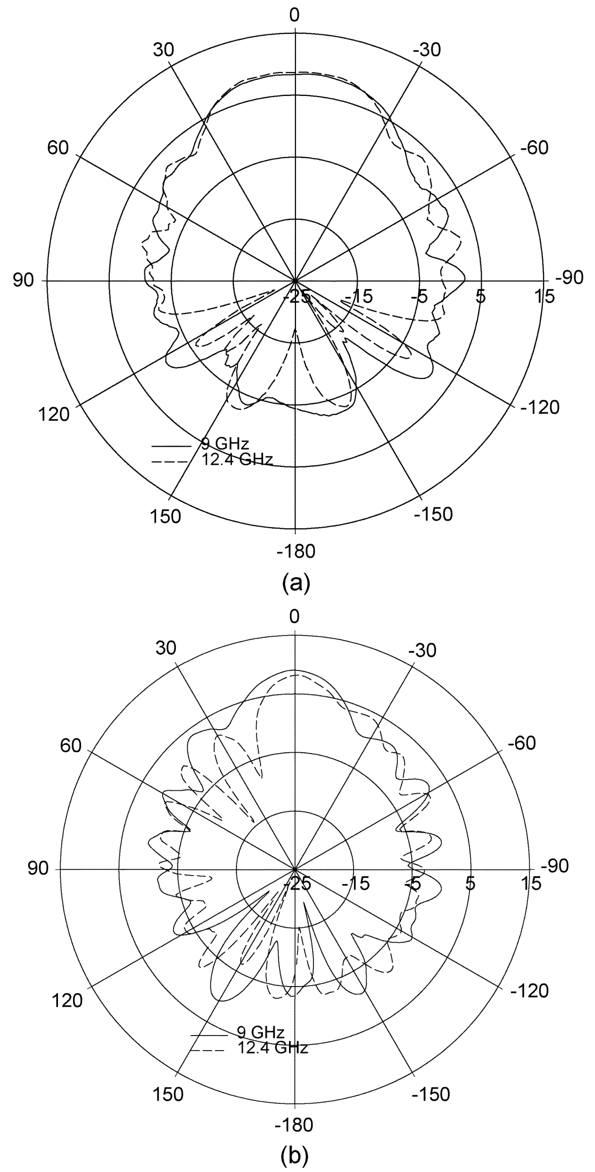

Fig. 15. Measured copolarization patterns of (a) H plane ($x - z$ plane) and (b) E plane ($x - y$ plane) E_ϕ fields at 9 and 12.4 GHz.

IV. Performance of Antenna

Careful attention is paid to the feeding points on the edge of strip to exploit the leaky-wave bandwidth as much as possible. As shown in Fig. 9, the feeding points are offset from the edges by 3.485 mm. Fig. 13 presents the return loss of a single-conductor strip leaky-wave antenna with a strip width of $w = 20$ mm and a length of 110 mm, on a substrate with a dielectric constant of $\varepsilon_r = 2.2$ and a thickness of $h = 0.508$ mm. This antenna has a $\mathrm{VSWR} \leqq 2$ from 6.55–13.75 GHz, yielding a relative bandwidth of 2.1:1, more than an octave. This measured data justifies the classification of this antenna as a broad-band antenna [12]. Fig. 14 presents the measured copolarization and cross-polarization radiation patterns of the H plane ($x - z$ plane) and the E plane ($x - y$ plane) fields at 11 GHz. The cross-polarization level of the E plane E_ϕ radiation pattern is relatively large compared to that of the H plane E_ϕ radiation pattern. The copolarization patterns of the H plane ($x - z$ plane) and the E plane ($x - y$ plane) E_ϕ fields at 9 and 12.4 GHz, as shown in Fig. 15, have peak gains of 9.03 and 8.22 dBi, respectively. According to Fig. 3, the value of the normalized phase constant ranges from 0.94–1.0 in the space-wave leaky regime; the radiation pattern is almost in the endfire direction and has a fixed mainbeam in the far-field region. For the same radiation bandwidth, this single-conductor strip leaky-wave antenna, with a length of 110 mm, is much shorter than the broad-band tapered microstrip leaky-wave antenna, which is 310 mm long [9].

V. Conclusion

This paper describes the basic characteristics of the EH_{01} leaky-wave mode of the single-conductor strip structure. Both surface wave modes TE_0 and TM_0 exist in the single-conductor strip structure, so the behavior of the space-wave and surface-wave modes differs from that of open transmission lines covered with ground planes, and so needs further investigation. A broad-band feeding structure is developed and a feeding method is also proposed to excite this EH_{01} leaky mode. The measured results of return loss and radiation patterns reveal the broad-band characteristics of the structure.

APPENDIX

The dyadic Green functions of the single-conductor strip structure for $z \geqq h$ are listed as follows:

$$f_e^{xx} = \frac{jk_0}{2\pi\omega\varepsilon_0}\frac{1}{n_\rho^2}\left[n_x^2 n_0 n_1(sn_1 + \varepsilon_r cn_0)\right] \quad \text{(A-1)}$$

$$f_m^{xx} = \frac{jk_0}{2\pi\omega\varepsilon_0}\frac{1}{n_\rho^2}\left[-n_y^2(cn_1 + sn_0)\right] \quad \text{(A-2)}$$

$$f_e^{xy} = \frac{jk_0}{2\pi\omega\varepsilon_0}\frac{n_x n_y}{n_\rho^2}\left[n_0 n_1(sn_1 + \varepsilon_r cn_0)\right] \quad \text{(A-3)}$$

$$f_m^{xy} = \frac{jk_0}{2\pi\omega\varepsilon_0}\frac{n_x n_y}{n_\rho^2}[cn_1 + sn_0] \quad \text{(A-4)}$$

$$f_e^{yx} = f_e^{xy} \quad \text{(A-5)}$$

$$f_m^{yx} = f_m^{xy} \quad \text{(A-6)}$$

$$f_e^{yy} = \frac{jk_0}{2\pi\omega\varepsilon_0}\frac{1}{n_\rho^2}\left[n_y^2 n_0 n_1(sn_1 + \varepsilon_r cn_0)\right] \quad \text{(A-7)}$$

$$f_m^{yy} = \frac{jk_0}{2\pi\omega\varepsilon_0}\frac{1}{n_\rho^2}\left[-n_x^2(cn_1 + sn_0)\right] \quad \text{(A-8)}$$

$$D_e = n_1(sn_1 + \varepsilon_r cn_0) + \varepsilon_r n_0(cn_1 + \varepsilon_r sn_0) \quad \text{(A-9)}$$

$$D_m = n_1(sn_1 + cn_0) + n_0(cn_1 + sn_0) \quad \text{(A-10)}$$

$$n_i^2 = \left(\frac{k_\rho}{k_0}\right)^2 - \left(\frac{k_i}{k_0}\right)^2 \quad i = 0, 1\text{Layer} \quad \text{(A-11)}$$

$$k_\rho^2 = k_x^2 + k_y^2 \quad \text{(A-12)}$$

$$n_x = \frac{k_x}{k_0} \quad \text{(A-13)}$$

$$n_y = \frac{k_y}{k_0} \quad \text{(A-14)}$$

$$s = \sinh(u_1 h) \quad \text{(A-15)}$$

$$c = \cosh(u_1 h). \quad \text{(A-16)}$$

ACKNOWLEDGMENT

The authors would like to thank Dr. T.-L. Chen and Dr J.-W. Sheen for their valuable discussions.

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