EXPERIMENT ON AN APERTURE-COUPLED LEAKY-WAVE ANTENNA WITH A PENCIL BEAM

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ABSTRACT: *A frequency-scanning leaky-wave antenna with an aperture-coupled feeding structure is presented. The aperture-fed structure is used to excite the microstrip first higher mode. In order to suppress the dominant mode, we add a sequence of covered wire in the center of this antenna. The measured H-plane main beam is a pencil beam, and is continuously scanned 12° as the frequency varies from 9.85 to 10.27 GHz. This feeding structure is very suitable for active phase antenna array applications.* © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 21: 5–8, 1999.

Key words: *leaky-wave antenna; aperture-fed antenna; microstrip antenna*

I. INTRODUCTION

Recently microstrip leaky-wave antennas (LWAs) have been the subject of extensive research [1]. The feeding structure is one of the important factors which determine the efficiency of the LAW. Here, we demonstrate an X-band LWA on one substrate, which is coupled to an open-ended microstrip line feed on another parallel substrate, through an aperture in the ground plane which separates the two substrates (see Fig. 1). The aperture-fed configuration has been proposed by Pozar [2], and it provides an excellent shielding between the feed network and the radiation element. The radiation from the feed network cannot interfere with the main radiation pattern because of this separating ground plane. Because of this virtue, the feeding signal can be placed below the antenna, so that the whole circuit size can be reduced. In addition, using the LWA as the radiation element, the main beam can be a pencil beam, and can scan in the elevation plane with variable frequencies [3]. The microstrip LWA is not only a simple structure, but also has the advantages of easy fabrication, narrow beam, frequency scanning, and is very suitable for integrated antenna application. In this letter, the LWA is fed from the aperture to excite the first higher mode [4], leaks in the form of a space wave, and covered wires are added in the center of the microstrip LWA in order to suppress the dominant mode [5]. By controlling the frequencies of the feeding signal, the main beam can be scanned in the elevation plane.

II. DESIGN OF THE APERTURE-COUPLED LEAKY-WAVE ANTENNA

Figure 1 shows the structure of the aperture-coupled leaky-wave antenna. The antenna has to be fed in the center of one side to excite the first higher mode. A sequence of covered wire was inserted in the center of the antenna. The frequency of the leaky region is decided by the width of the antenna. The size of the aperture and the open-ended stub can be used to determine the antenna impedance. Here, a commercially available CAD tool IE3D is utilized to help us design the matching network [6]. The width of the open-end stub is determined to be the equivalent impedance of a 50 Ω microstrip line at 10 GHz, and the aperture width is about $0.1\lambda_g$ when the aperture length is decided to be close to $1.25\lambda_g$. In addition, the open-ended stub length is adjusted to match the input resistance of the antenna.

A prototype of the aperture-fed leaky-wave antenna is designed and fabricated on RT/Duroid substrate with $\varepsilon_r = 2.2$ and a thickness of 20 mils. To excite the first higher mode,

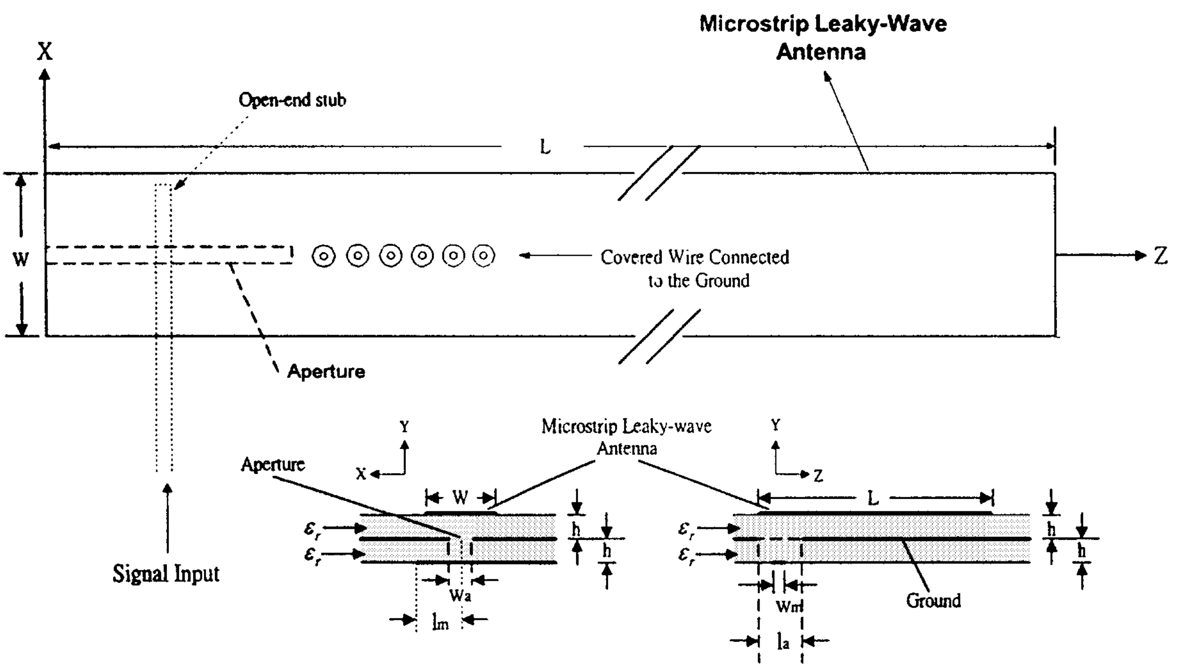

Figure 1 Configuration of the aperture-coupled leaky-wave antenna. L = 125 mm, W = 11.2 mm, ε_r = 2.2, h = 0.508 mm, l_a = 25 mm, w_a = 2.02 mm, w_m = 1.54 mm, l_m = 17.06 mm

the dimensions of the LWA are determined to be 11.2 mm long and 125 mm wide. The aperture is of width w_a = 2.02 mm and length l_a = 25 mm (see Fig. 1). A 1.54 mm wide microstrip line extends 17.06 mm past the aperture, and is termined as an open end.

The radiation characteristics of a microstrip LWA can be realized by obtaining its complex propagation constant $\beta + j\alpha$ in the leaky (radiation) region, where β is the phase constant and α is the attenuation constant. We employed the rigorous (Wiener–Hopf) solutions mentioned in [7] to obtain the complex propagation constant. The variations of phase constant β and attenuation constant α as a function of the frequency are shown in Figure 2. Figure 2 also shows the measured scattering parameter S_{11} of our antenna, and we have found that the leaky region is narrow for this structure, which is about 600 MHz, at ~ 9.85–10.45 GHz. The geometry and coordinate system for the microstrip leaky-wave antenna are shown in Figure 3. The guided wave will leak as a leaky mode

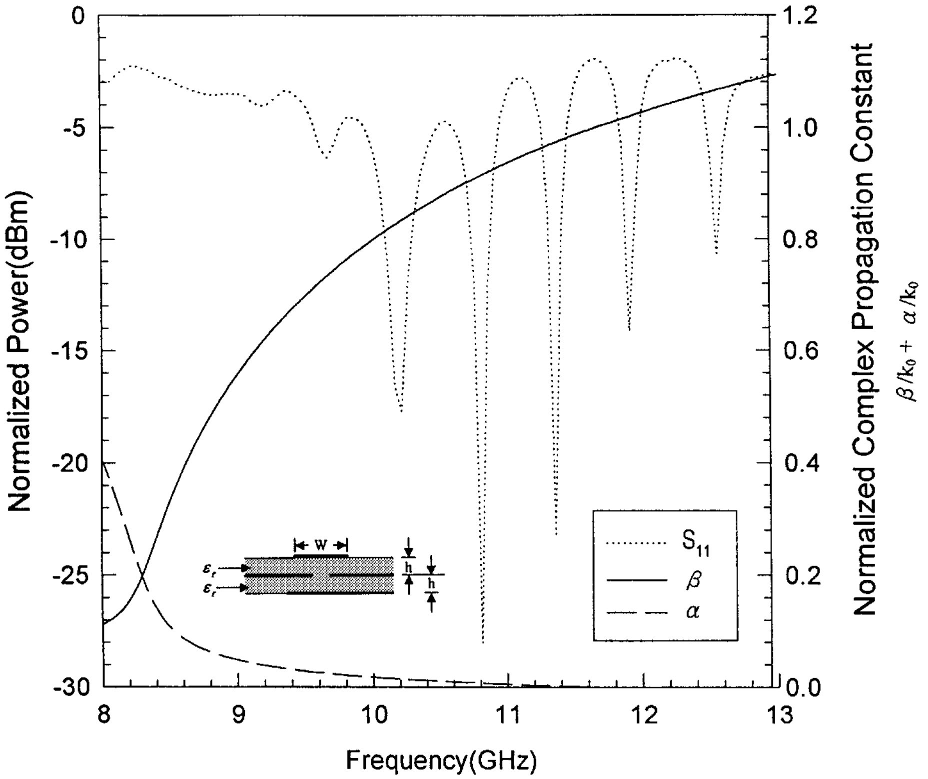

Figure 2 Normalized complex propagation constant and the S-parameter of the first higher mode for the microstrip leaky-wave antenna (h = 0.508 mm, w = 11.2 mm, ε_r = 2.2, and k_0 is the free-space wavenumber)

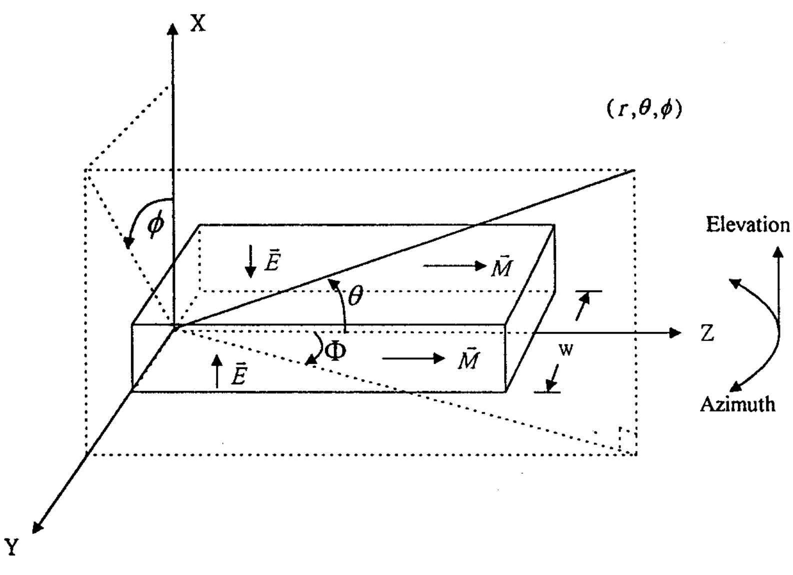

Figure 3 Geometry and coordinate system for the microstrip leaky-wave antenna. The width and the length of this antenna are 11.2 and 125 mm

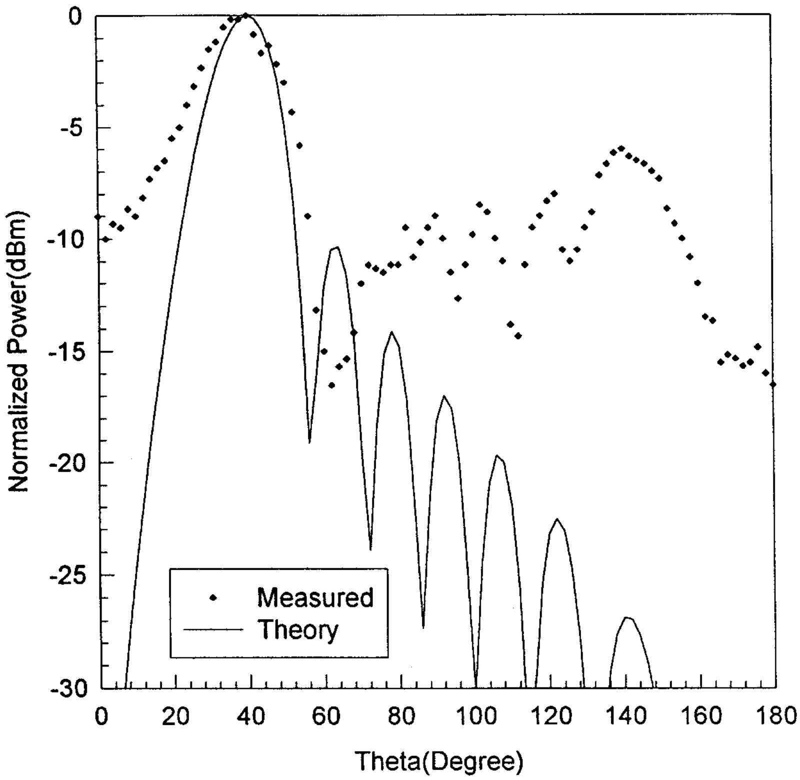

Figure 4 Theoretical and measured far-field radiation patterns ($\varphi = 0$) of the aperture-coupled leaky-wave antenna at 9.85 GHz

into the space wave and surface wave when $\beta < k_0$. θ is the elevation angle between the main-beam direction and the end-fire direction, and it can be calculated by using an approximation relationship of $\theta = \cos^{-1}(\beta/k_0)$. According to this relationship, it can be predicted that the main beam position will change with the frequency.

III. THEORETICAL AND EXPERIMENTAL RESULTS

Figure 4 shows the theoretical and measured radiation patterns of our aperture-coupled LWA at 9.85 GHz. The theoretical prediction of the radiation pattern is calculated by applying Huygens' principle at the far-zone field [8, 9].

Here, the aperture-fed leaky-wave antenna exhibits a tuning bandwidth from 9.85 to 10.47 GHz, and the variation of scanning angle is 12°. Figure 5 displays the measurement results of the main beam swings up from the end-fire direction (Z-axis) as the operating frequency decreases from 10.47 to 9.85 GHz. The observed narrow beamwidth in Figure 5 is due to the attenuation constant α decreasing when the operating frequency becomes higher (see Fig. 2). The maximum effective radiated power (ERP) of this antenna is approximately 21.05 ± 0.7 dBm. The power level variation is due to the variation of the impedance of the aperture-fed LWA as the frequency varied.

IV. CONCLUSION

A technique utilizing the aperture-coupled structure for a leaky-wave antenna has been demonstrated in this letter. The structure provides a simple fabrication to achieve the tunable and narrow main beam, and the most important characteristic of eliminating the interference of the feed and the radiation elements. We found a measured beam-scanning angle of 12°. This antenna-feeding structure is very suitable for the applications of an active phase antenna array.

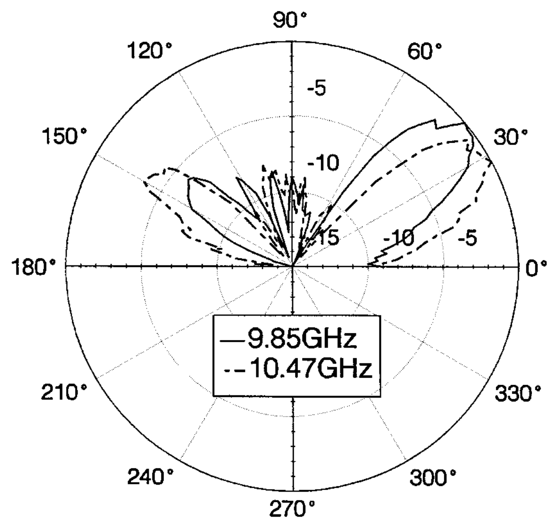

Figure 5 H-plane (y–z plane) frequency-scanned radiation patterns

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MODAL CIRCUIT DECOMPOSITION OF LOSSY HOMOGENEOUS UNIFORM QUASI-TEM COUPLED TRANSMISSION LINES

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ABSTRACT: *This paper presents a detailed analysis and discussion of the conditions on which the realizability of modal circuit decomposition for lossy homogeneous quasi-TEM coupled transmission lines stands. It is shown that the requirement of a real nonsingular modal transformation matrix cannot always be met. With the exception of some limit situations (lossless conductors or very high-frequency operation), equivalent modal circuits can only be obtained, in general, for transmission-line configurations exhibiting very peculiar symmetry properties.* © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 21: 8–11, 1999

Key words: *multiconductor transmission lines; quasi-TEM analysis; modal circuit decomposition*

1. INTRODUCTION

Driven by the ever-expanding applications of VLSI, MIC, MMIC, and MHMIC structures, many papers devoted to the quasi-TEM analysis of multiconductor transmission lines have been published in the past decades. Among the many issues addressed, a considerable concern has been expressed about the conditions for the existence of realizable equivalent modal circuits [1–4] whose implementation via SPICE or other CAD tools [5–6] permits easy simulation models for transient analysis of n-coupled line circuits.

This paper presents a new contribution to the subject of modal circuits. Its purpose is to rigorously clarify the circumstances under which lossy multiconductor lines may really be simulated by equivalent modal circuits.

2. MODAL DECOMPOSITION

The equivalent modal circuit for a uniform n-coupled multiconductor transmission line, depicted in Figure 1, is composed of two n-port ideal transformers with real transformation ratios and n uncoupled transmission lines [1].

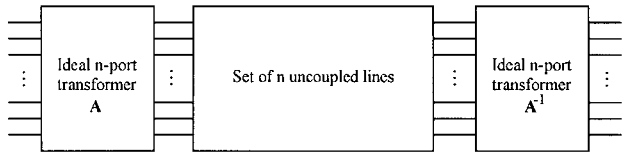

Figure 1 The modal circuit is composed of: 1) an ideal n-port transformer with winding ratios chosen so that natural voltages and natural currents are transformed into modal voltages and modal currents according to $\mathbf{V} = \mathbf{T}\mathbf{V}_{\text{modal}}$ and $\mathbf{I} = \mathbf{T}^{-1t}\mathbf{I}_{\text{modal}}$; 2) a set of n uncoupled transmission lines, each line being described by a given propagation constant γ and by a given characteristic impedance Z_w; and 3) a second ideal n-port transformer, identical to the first one, but with windings reversed, assuring the transformation $V_{\text{modal}} = \mathbf{T}^{-1}\mathbf{V}$ and $\mathbf{I}_{\text{modal}} = \mathbf{T}^t\mathbf{I}$

For determining the modal circuit constitutive parameters, we start with the well-known frequency-domain generalized Telegrapher's equations [7–8]

$$\begin{cases} \dfrac{d}{dx}\mathbf{V} = -\mathbf{Z}\mathbf{I} \\ \dfrac{d}{dx}\mathbf{I} = -\mathbf{Y}\mathbf{V} \end{cases} \rightarrow \begin{cases} \dfrac{d^2}{dx^2}\mathbf{V} - \mathbf{Z}\mathbf{Y}\mathbf{V} = 0 \\ \dfrac{d^2}{dx^2}\mathbf{I} - \mathbf{Y}\mathbf{Z}\mathbf{I} = 0 \end{cases} \tag{1}$$

where $\mathbf{V}(x)$ and $\mathbf{I}(x)$ are the column vectors of conductor voltages and conductor currents, respectively, and $\mathbf{Z}$ and $\mathbf{Y}$ are the impedance and admittance matrices per unit length, respectively. As usual, both $\mathbf{Z}$ and $\mathbf{Y}$ matrices can be broken down into their real and imaginary parts, that is, $\mathbf{Z} = \mathbf{R} + j\omega\mathbf{L}$ and $\mathbf{Y} = \mathbf{G} + j\omega\mathbf{C}$, where $\mathbf{R}$, $\mathbf{L}$, $\mathbf{G}$, and $\mathbf{C}$ are real symmetric positive-definite matrices.

The diagonalization of the $\mathbf{ZY}$ product, through a similarity transformation $\mathbf{T}$, yields the diagonal matrix of eigenvalues λ:

$$\mathbf{T}^{-1}\mathbf{Z}\mathbf{Y}\mathbf{T} = \lambda. \tag{2a}$$

The diagonalization of the $\mathbf{YZ}$ product, through a similarity transformation $\mathbf{T}_i$ (with $\mathbf{T}_i = \mathbf{T}^{-1t}$) yields the same diagonal matrix of eigenvalues:

$$\mathbf{T}_i^{-1}\mathbf{Y}\mathbf{Z}\mathbf{T}_i = \lambda. \tag{2b}$$

Note that the matrix eigenvalue equation in (2b) is not independent of (2a) as, in fact, one equation is just the transpose of the other. It should be added, further, that $\mathbf{ZY}$ and $\mathbf{YZ}$ are similar matrices since $\mathbf{ZY} = \mathbf{Y}^{-1}(\mathbf{YZ})\mathbf{Y}$.

As for the modal circuit shown in Figure 1 we have, on the one hand, that the evaluation of the transmission matrix $\mathbf{A}$ for the n-port ideal transformers is based on the knowledge of $\mathbf{T}$ and $\mathbf{T}_i$ [1] via

$$\mathbf{A} = \begin{bmatrix} \mathbf{T} & 0 \\ 0 & \mathbf{T}^{-1t} \end{bmatrix}$$

and, on the other hand, the set of n uncoupled lines is characterized by a diagonal matrix γ of modal propagation constants, given by $\gamma = \lambda^{1/2}$ [7], and by a diagonal matrix $\mathbf{Z}_w$ of modal characteristic impedances given by $\mathbf{Z}_w = (\mathbf{T}\gamma)^{-1}\mathbf{Z}\mathbf{T}_i$ [7].