

# REACTIVE LOADED MICROSTRIP LEAKY WAVE ANTENNA FOR LOW COST BEAM STEERING APPLICATIONS

Gijo Augustin, S. V. Shynu, P. Mohanan, C. K. Aanandan, and K. Vasudevan

Centre for Research in Electromagnetics and Antennas, Department of Electronics, Cochin University of Science and Technology, Kochi 682 022, India

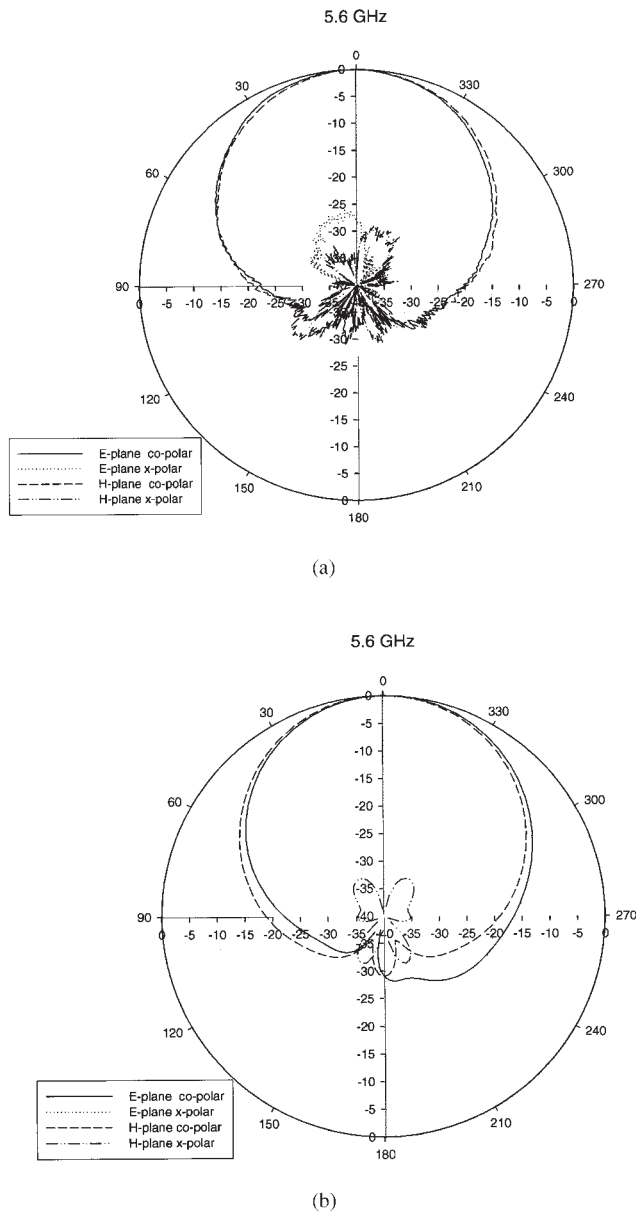
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**ABSTRACT:** A novel fixed frequency beam scanning microstrip leaky wave antenna is reported. The beam scanning at fixed frequency is achieved by reactive loading. Simulation and measured results shows frequency scanability of  $80^\circ$  as well as fixed frequency beam steering of  $68^\circ$  over the  $-10$  dB impedance band of 4.56–5.06 GHz © 2006 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 48: 2299–2301, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21912

**Key words:** leaky wave antenna; beam steering

## 1. INTRODUCTION

The microstrip leaky wave antenna (MLWA) has been applied in many fields, such as cruise control, because of its simple feeding configuration, low profile, and frequency scanning [1, 2]. When the operating frequency changes in the leaky region, the main lobe scans in the H-plane. However, this frequency dependence of the scanning angle, limits the use of these types of antennas if narrow frequency bandwidth is available and wide angle of coverage is required. Fixed frequency beam scanning is the right alternative,



**Figure 4** Radiation pattern of proposed antenna at 5.6 GHz: (a) measured and (b) simulated

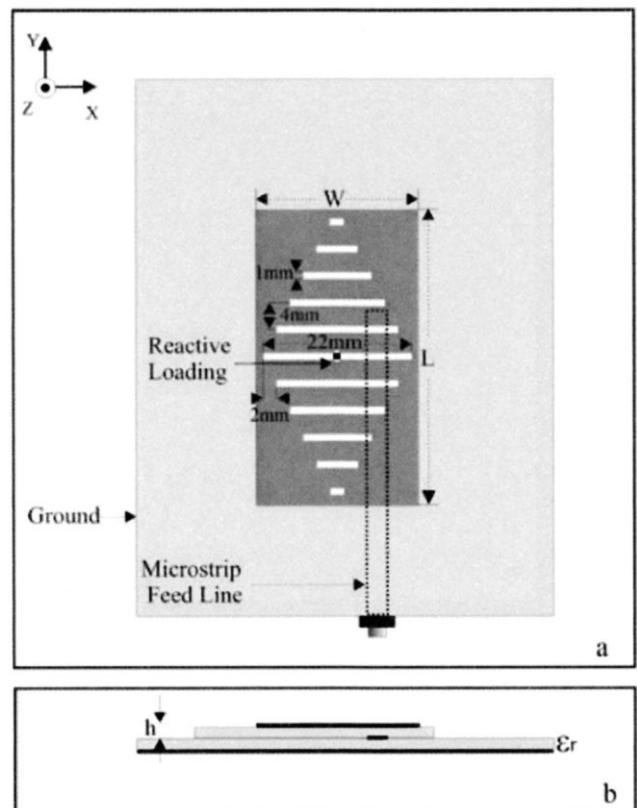
## ACKNOWLEDGMENT

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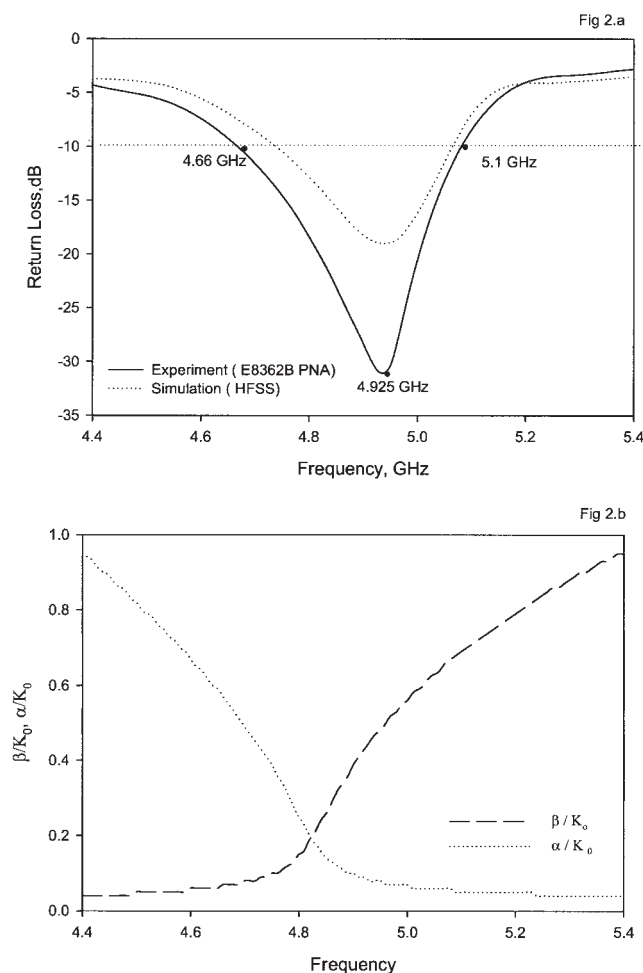


**Figure 1** Geometry of proposed MLWA: (a) Top view and (b) side view.  $L = 44$  mm,  $W = 24$  mm,  $h = 1.6$  mm

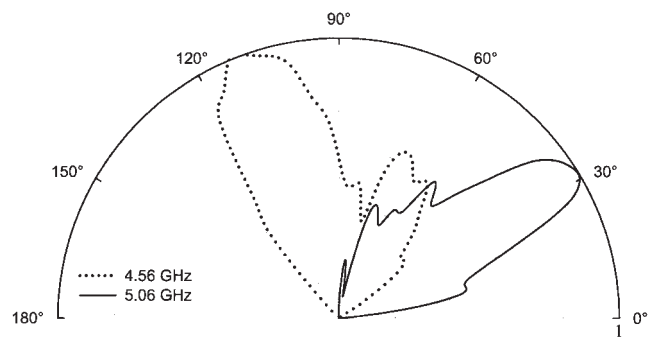
but new feeding configurations and phase shifter circuits integrated to the antenna are required [3]. As the phase velocity along the microstrip and the cut off frequency of its first higher order mode are functions of the reactance across the slots, capacitive loading along the leaky slot line provides a simple method for fixed frequency beam steering [4, 5]. In this letter we propose a MLWA, loaded with MIM capacitors, capable of both frequency scanability and fixed frequency beam steering. The proposed antenna design and experimental results are presented.

## 2. ANTENNA DESIGN

The layout of the MLWA is shown in Figure 1. It consists of a rectangular microstrip patch of  $L = 44$  mm,  $W = 26$  mm, fabricated on a substrate of dielectric constant,  $\epsilon_r = 4.7$  and  $h = 1.6$  mm. A central slot of length 22 mm with five horizontal slots of linearly decreasing length are loaded on either side, in the patch. Ansoft High Frequency Structure Simulator (HFSS™ v10) is used to optimize the dimensions and positions of the horizontal slots. The optimization is done to suppress the dominant resonant mode of the original patch and to excite the first higher mode with a single  $50 \Omega$  microstrip feed line. The suppression of the dominant mode is evident from the computed surface current density plot shown in Figure 5 of the patch at 4.9 GHz. The leaky region of the antenna is characterized by its normalized phase constant ( $\beta/k_0 < 1$ ), the value of which determines the direction of the radiated



**Figure 2** (a) Return loss characteristics of the LWA. (b) Behavior of normalized phase constant  $\beta/k_0$  and attenuation constant  $\alpha/k_0$  of first higher order leaky mode of LWA over the leaky band



**Figure 3** Measured H-plane radiation patterns at 4.56–5.06 GHz

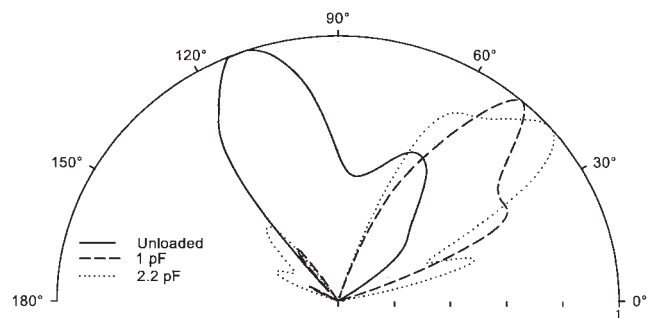
beam. To understand the radiation characteristics of the MLWA, we obtained its complex propagation constant ( $\alpha + j\beta$ ) as a function of frequency. In Figure 2, the space leaky region is fixed as 4.56–5.06 GHz, corresponding to an antenna configuration of  $L = 44$  mm,  $W = 26$  mm,  $h = 1.6$  mm, and  $\epsilon_r = 4.7$ . Since  $\beta$  is a function of the surface reactance of the radiating patch, fixed frequency beam steering is accomplished by reactively loading the slot line embedded in the patch. This is achieved by integrating MIM capacitors (Q-Max, Capax Technologies) of different values across the center horizontal slot.

## 3. RESULTS AND DISCUSSION

The experimental return loss characteristics, measured using Agilent E8362B PNA Series Network Analyzer, is shown in Figure 2. A good matching below  $-10$  dB over the predicted leaky band of 4.56–5.06 GHz is observed. The frequency scanning feature of the antenna is shown in Figure 3. When the frequency is varied from 4.56 to 5.06 GHz, the direction of the main radiated beam in the elevation plane is tilted from  $31^\circ$  to  $112^\circ$  resulting a frequency scanability of  $81^\circ$ . Effect of reactive loading over the slot line is studied from the radiation pattern measured at 4.9 GHz for different values of loaded reactance (Figure 4). The unloaded LWA has a radiation beam directed towards  $113^\circ$ . Loaded tilt angle is  $45^\circ$  and  $41^\circ$  for capacitance values 1 and 2.2 pF, respectively. We observe that a variation of capacitance from 1 to 2.2 pF gives a tilt of the radiated beam by  $4^\circ$ . Surface current distribution is shown in Figure 5.

## 4. CONCLUSION

A novel reactively loaded slot line LWA which finds application in side looking radars or any type of low cost tracking systems is presented. The antenna is capable of both frequency scanning of  $80^\circ$  ( $160^\circ/\text{GHz}$ ) and fixed frequency beam steering of  $68^\circ$ .



**Figure 4** Measured H-plane radiation patterns of capacitively loaded LWA ( $C = 2.2$  and  $5.6$  pF) and unloaded LWA at 4.9 GHz

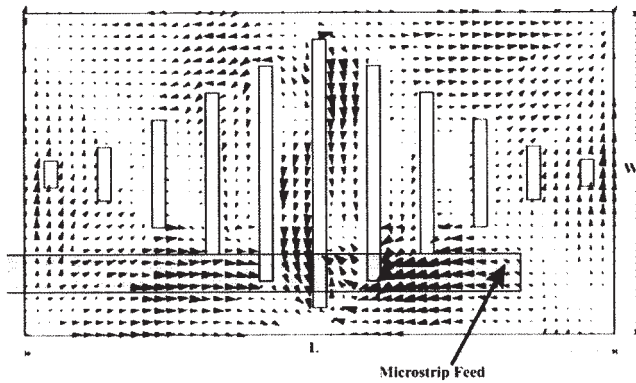


Figure 5 Surface-current distribution of the LWA at 4.9 GHz

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## ELECTRONICALLY CONTROLLED RECIPROCAL PHASE SHIFTER

G. Collin,<sup>1</sup> T. Ditchi,<sup>2</sup> E. Géron,<sup>1</sup> and S. Holé<sup>2</sup>

<sup>1</sup> LEG

Ecole Supérieure de Physique et de Chimie Industrielle  
Paris, France

<sup>2</sup> LISIF, Université Paris 6, 10 rue Vauquelin, 75005 Paris, France

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**ABSTRACT:** A reciprocal phase shifter principle is proposed, of very simple design with only one control voltage, which enables the continuous variation of the phase at a frequency of 1.89 GHz. An implementation on Teflon substrate using Wilkinson dividers and a second-order filter using one variable capacitor is presented. © 2006 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 48: 2301–2303, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21911

**Key words:** phase shifter; microstrip line; transmission line; microwave

### 1. INTRODUCTION

In many applications, for instance in the field of telecommunications, it is necessary to implement phase shifters having a large phase range and a flat amplitude response. As a typical example, multielement arrays used to create adaptive antennas require the use of one phase shifter per element to allow the control of their radiation pattern. Many technologies can be used for that purpose [1–5]. In this paper we present a phase shifter of very simple design with only one control voltage, which enables the continuous variation of the phase at a frequency of 1.89 GHz. It is composed

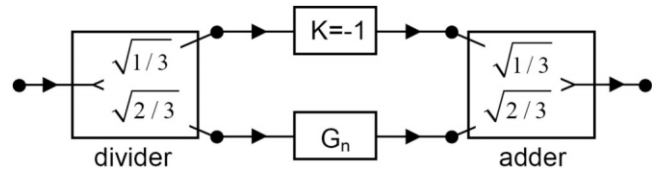


Figure 1 Block diagram of an implementation of the transfer function

of two power dividers, an half-wave transmission line, and a filter, the nature and order of which play an important role on the phase shifter features. An implementation on Teflon substrate using Wilkinson dividers and a second-order filter using one variable capacitor is presented.

### 2. PRINCIPLE

The transfer function  $H_n$  of a perfect phase shifter of order  $n$  can be expressed as the ratio of two  $n$  order polynomials as:

$$H_n(s) = \frac{1 - a_1 s + a_2 s^2 - a_3 s^3 + \dots + (-1)^n a_n s^n}{1 + a_1 s + a_2 s^2 + a_3 s^3 + \dots + a_n s^n}, \quad (1)$$

where  $s = j2\pi f$ ,  $f$  is the frequency,  $a_i$  are positive real numbers and  $j = \sqrt{-1}$ .

$H_n$  has an amplitude equal to 1 and its argument is a function of  $a_i$  and  $f$ . The phase of  $H_n$  can be adjusted by varying the parameters  $a_i$ . The range of variation of the phase increases with the order  $n$ . It reaches  $\pi$  when  $H_n$  is a first-order transfer function,  $2\pi$  for a second order, and  $n\pi$  for an  $n$ -order function.

An easy physical implementation of  $H_n$  can be made by expressing Eq. (1) as a sum of two terms:

$$H_n(s) = 1 - 2 \frac{a_1 s + a_3 s^3 + \dots}{1 + a_1 s + a_2 s^2 + a_3 s^3 + \dots} = -3 \left[ -\frac{1}{3} + \frac{2}{3} G_n(a_i, s) \right]. \quad (2)$$

The first term is a constant whereas the second depends on  $s$  and on the  $a_i$  coefficients. Figure 1 shows the block diagram of an implementation of such a transfer function characterized by an attenuation of 1/3, a 180° phase shift, but a phase range identical to that expressed by Eq. (2).

$G_n$  is the transfer function of a filter which depends on the order  $n$  of  $H_n$ .  $G_1$  is associated with a first-order high pass filter and  $G_2$  with a second-order pass band filter:

$$G_1(s) = \frac{a_1 s}{1 + a_1 s}, \quad G_2(s) = \frac{a_1 s}{1 + a_1 s + a_2 s^2}, \dots \quad (3)$$

Higher-order transfer functions can be implemented for obtaining larger phase range. For instance,  $G_3$  is a high pass filter followed by a cut band filter.

This block diagram is symmetrical and reciprocal as soon as all blocks are.

### 3. IMPLEMENTATION

Our target was to make a phase shifter in a frequency domain around 2 GHz with a narrow frequency band (<1%). The phase shifter behavior has been studied around the central frequency. All transfer functions presented below are calculated assuming a 50 Ω generator and a 50 Ω load.