

Figure 10 Width and temperature errors on profile retrieval in simulations by data fusion of 3, 35, 60, and 94 GHz radiometers

with a bandwidth of 2 GHz. The length of the lossy line to be considered is 1 mm, and its lineic resistance is 100 Ω /cm. In order to simplify our study, we assume losses to be only due to the dc resistance of the line; we then neglect phenomena such as skin effect.

We create a steep hot spot of 200 μ m with $\Delta T = 5^{\circ}$ C. In Figure 9, we can see the results of the inversion process applied to the radiometric data. We show that it is not possible to conclude about the true thermal gradient from one radiometer data. However, data fusion of the radiometric data at different frequencies in the adaptive filter allows a very good estimation of the temperature ($\varepsilon_T = 0.5\%$) and of the width of the hot spot.

For the case of less severity on retrieval conditions, we give in Figure 10 the ε_w and ε_T errors as functions of the hot spot width. For example, a resolution of 100 μ m for 5°C is retrieved with a width error about 20% and a temperature error less than 10%.

7. CONCLUSION

The use of an inversion process based on a Kalman filtering in a nonclassical application brings a good retrieval of the shape and temperature of a thermal gradient. We obtain quantitative information about the temperature, with an *a priori* knowledge of the weighting functions involved in the radiometric measurement.

We have shown, by means of a simulation, the possibility of simplifying a correlation radiometer by using small bandwidth systems. But another consequence concerns the improvement of resolutions (size and temperature) if we use radiometers with very different bandwidths. It is then possible to define the pattern of a thermal microsensor. This type of system can be dedicated to the measurement of temperature inside a three-dimensional electronic circuit, where infrared measurement is not possible. The measurement of temperature in microelectronic components becomes possible if we work with several millimetric radiometers.

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X-BAND ACTIVE TWO-TERMINAL LEAKY-WAVE ANTENNA

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ABSTRACT: An active-integrated two-terminal leaky-wave antenna design is demonstrated in this paper. An X-band two-terminal-fed microstrip leaky-wave antenna and an active source, a varactor-tuned oscillator HEMT VCO, are integrated on the same plane. The space wave, due to the first higher order mode, dominates the leakage of the radiated energy in this experiment. This approach, as compared to a traditional single-terminal feeding leaky-wave antenna, creates a dualbeam radiation pattern, and has the advantages of multidirection and no reflected wave caused by the open end of the radiating element. Measured results on the experimental antenna show that the total scanning angle of two beams of this antenna configuration is approximately 44°, the maximum effective isotropic radiated power (EIRP) is close to 17.5 dBm for the right beam and 16.7 dBm for the left beam at 10.4 GHz, and the power difference between the two beams is less than 1 dB. Between 9 and 11.5 GHz, the return loss is less than -10 dB, and the transmission coefficient shows that the power radiates into space. © 1998 John Wiley & Sons, Inc. Microwave Opt Technol Lett 19: 100–103, 1998.

Key words: two-terminal feeding; dual-beam scanning; balanced matched load

INTRODUCTION

Active integrated antennas have become an important topic in the microwave and millimeter-wave area over recent years. They have been applied to wireless communication, the mobile collision avoidance system, and satellite transmission. Because of the integration of active devices and passive radiated elements, it is usually not essential for active antennas to have external injected power, and they can reduce the size, cost, fabrication, and structure of the overall system. Active leaky-wave antennas have the above advantages; meanwhile, they have a wide bandwidth, pencil beam, and excellent performance of frequency scanning ability. By adjusting the bias voltage of an active device, e.g., VCO, the main beam can be controlled electronically to scan in wide angles. Active leaky-wave antennas have been previously investigated in [1, 2].

Here, an active phase-shifterless dual-beam scanning antenna (see Fig. 1) by integrating a varactor-tuned HEMT VCO with the microstrip two-terminal feeding leaky-wave antenna is developed. Using this configuration, a dual-beam pattern is created. We adjust the varactor dc bias voltage and tune the operating frequencies to control the dual-beam position. Therefore, an active dual-beam scanning leaky-wave antenna can be used as a frequency-scanning antenna, and a wider covered scanning region due to its dual-beam scanning performance will be obtained. Another advantage of this design is that there is no reflected wave due to the unmatched load of the open end of the leaky wave in a finitelength situation. Although the reflected wave can be suppressed by longer length [3] or the array topology [2], this balanced matched load of the two-port design can be thought of as a matched load on both sides of the open end in this circuit; thus, this two-port design can suppress the reflected wave.

DESIGN

The configuration of the active dual-beam scanning leakywave antennas (see Fig. 1) is fabricated in this paper. The



Figure 1 Configuration of the active dual-beam scanning leaky-wave antenna

circuit uses a 0.508 mm thick RT/Duroid substrate with relative permittivity $\epsilon_r = 2.2$, and an NE42484A low-noise GaAs HEMT is used as the oscillator device. The HEMT VCO, the matching circuit, and a two-terminal feeding leaky-wave antenna are integrated on the same plane. The VCO is designed using the negative-resistance method utilizing the commercially available CAD tool HP-EEsof Libra. An ALPHA CVG 7864 GaAs package varactor is used as a tuning varactor in this experiment, and the capacitance for a tuning voltage of 1–11 V ranges approximately from 11 to 0.3 pF in the data sheet.

To excite the first higher order mode, this microstrip leaky-wave antenna is fed asymmetrically. The width W and the length L of the leaky-wave antenna are 11 mm and 15 cm, respectively. The dimension of this two-terminal leakywave antenna is chosen empirically so that the space wave dominates the radiating power, and the first higher order mode can be excited within the operating frequency. Because the operating frequency is below the cutoff frequency, the normalized phase constant β/k_0 is less than 1, and the space wave dominates the leakage [4]. The simple T-type power divider provides an equal power split from the active source to the two feeding terminals of the leaky-wave antenna. The matching circuit is designed to obtain the maximum radiating power within the operating frequency range. By this design procedure, the radiator can have maximum power input, and the portion of the reflected wave also can be reduced. In addition, the loading effect will have a minimal effect on the oscillator's operation.

THEORETICAL AND EXPERIMENTAL RESULTS

We employed a rigorous (Wiener–Hopf) solution mentioned by [5] to obtain the normalized complex propagation constant $\beta - j\alpha$ of the first higher order mode in its leakage range for the leaky-wave antenna, where β is the phase constant and α is the attenuation constant. The variation of β and α as a function of frequency is shown in Figure 2. According to the value of Figure 2, we calculate and plot the theoretical beam-scanning radiation patterns (see Fig. 3) of this active



Figure 2 Normalized complex propagation constant $\beta - j\alpha$ of the first higher mode for the particular microstrip leaky-wave antenna. h = 0.508 mm, w = 11 mm, and $\varepsilon_r = 2.2$; k_0 is the free-space wavenumber



Figure 3 Theoretical radiation patterns (*X*–*Z* plane) of the active dual-beam scanning leaky-wave antenna at 9.7, 10.4, and 11.3 GHz

dual-beam scanning leaky-wave antenna. By adjusting the bias voltage of the HEMT voltage-controlled oscillator VCO, the free-running frequency of the HEMT VCO can be varied from 9.7 to 11.3 GHz. Under the far-field condition, a measured effective isotropic radiated power (EIRP) of 17.5 dBm for the right beam of this antenna with the HEMT VCO biased is approximately observed at 10.4 GHz, and the measured power (EIRP) of the left beam is 16.7 dBm at the same frequency. Figure 4 shows a comparison of the theoretical and measured radiation patterns of this microstrip active dual-beam leaky-wave antenna at 10.4 GHz, in the situation where there is a long length (L = 15 cm). Figure 5 presents the measured dual-beam scanning radiation pattern of this prototype for three frequencies, 9.7, 10.4, and 11.3 GHz.



Figure 4 Comparison of the theoretical and measured radiation patterns of the microstrip dual-beam leaky-wave antenna at 10.4 GHz, in the situation where there is a long length (L = 15 cm)



Figure 5 Measured polar radiation patterns (X-Z plane) of the active dual-beam scanning leaky-wave antenna at 9.7, 10.4, and 11.3 GHz

Measured results show that the total scanning angle is 44°, from 24° to 46° for the right beam, and from 128° to 150° for the left beam. In Figure 6, a comparison of the theoretical and measured scanning angles of the right beam is presented from 9.7 to 11.3 GHz. In Figure 7, the measured return loss S_{11} and the transmission coefficient S_{21} of the passive twoterminal leaky-wave antenna are shown. Here, S_{11} is approximately less than -10 dB from 9.5 to 11.5 GHz, and S_{21} shows that the injected power leaks into space from the leaky-wave antenna.

CONCLUSION

An X-band active dual-beam scanning leaky-wave antenna has been demonstrated. We successfully make use of this



Figure 6 Comparison of the theoretical and measured scanning angles of the right beam from 9.7 to 11.3 GHz



Figure 7 Measured S-parameters of the microstrip passive dualbeam scanning leaky-wave antenna

two-feed topology for the active leaky-wave antenna in order to create a dual-beam radiation pattern, and vary the frequency of the HEMT VCO to control a two-directional scanning beam. The total measured scanning angle of this active antenna can be steered over a range of angle 44°. The measured S-parameters, the return loss S_{11} and the transmission coefficient S_{21} , show that this dual-beam scanning leaky-wave antenna has the properties of wide bandwidth and high efficiency. This antenna design can be very useful in an automotive radar system, mobile communication, and a personal communication system (PCS); meanwhile, it can also be easily implemented into a monolithic leaky-wave array module.

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A NEW JUNCTION MODEL WITH TWO WIRES OF DIFFERENT RADIUS FOR THE TLM METHOD

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ABSTRACT: A new wire-junction node for modeling a junction with two wires of different radius using the TLM method is presented. The scattering matrix of the node is provided, and the numerical results are compared with the moment method solutions traditionally used for modeling wires and wirelike structures. © 1998 John Wiley & Sons, Inc. Microwave Opt Technol Lett 19: 103–107, 1998.

Key words: transmission-line matrix method; time-domain algorithm; electromagnetic propagation; wave scattering

I. INTRODUCTION

The transmission line matrix (TLM) method is a time-domain algorithm that has been used in the resolution of several electromagnetic propagation and scattering problems. EMC is one of the important problems that can be dealt with using the TLM method to model the coupling between the incident EM wave with the radiating systems or receptors and the associated electrical structures frequently consisting of wires and junctions of conducting wires.

Modeling structures of wires, and more specifically junctions of wires, by means of the TLM method is a difficult task [7], and their treatment in the literature is relatively scarce in spite of their importance [1, 8]. Generally, several approaches can be used to model wires by the TLM method. These are the separated solutions [1], which decouple the field and wire simulations and perform them separately; the integrated method [1], which considers the wire as a thin conducting structure upon which the incident fields reflects; the diacoptic methods [2], which model the wire and the medium in which it is placed with irregular coarse mesh dimensions; and finally, the wire nodes. In the latter technique, the structure of wire is modeled by a special node which is different from the commonly used symmetrical condensed node (SCN), and so requires the introduction of special modeling elements.

However, in order to model wire junctions by the TLM method, it is only possible to use the integrated solution method which requires very fine discretization of the medium, and thus a large amount of memory space and computation time, to provide satisfactory results. The second technique is to use a special node with wire junctions inside [7]. However, up to now, there has been only one special node that allows us to model a wire junction with the same radius [7].

In this paper, we propose a symmetrical condensed node for modeling a junction with two wires of different radius. This node, as another special node of single wires, contains the additional capacitances and inductances introduced by the wires, and makes use of the "mixed two-step" model which provides the most satisfactory results in the simulation of single wires [6].

We have tested the proposed model by calculating the currents on different arms of wires junction, and the numerical results are compared with the moment method solutions.