Nonradiative Dielectric Waveguide T-Junctions for Millimeter-Wave Applications

T. YONEYAMA AND S. NISHIDA

Abstract —**A high-performance T-junction has been fabricated based on the nonradiative dielectric waveguide. Widths of the main arm and subarm were determined experimentally. The T-junction was provided with a dielectric stub and thin-metal patches which proved to be effective matching elements in improving the performance. Well-balanced outputs in excess of −4 dB were obtained over a frequency range of 2 GHz at a center frequency of 35 GHz.**

I. Introduction

Since the nonradiative dielectric waveguide (NRD-guide) is a low-loss dielectric waveguide which can completely suppress undesired radiation at curved sections and discontinuities of circuits as described in a previous paper [1], it can be used to great advantage for millimeter-wave integrated circuits. Various circuit components, such as couplers [2] and filters [3], have already been constructed by taking advantage of the nonradiating nature of the NRD-guide. The T-junction is another circuit component which is of practical importance.

In this paper, an NRD-guide T-junction will be described with emphasis on techniques of empirically optimizing the performance. Although dielectric waveguide junctions have been constructed before by employing a rectangular rod waveguide [4] and an image guide [5], they are not necessarily satisfactory in their power dividing capability. The NRD-guide T-junction is expected to overcome the imperfect power division difficulties inherent in other dielectric waveguide T- and Y-junctions, because there is no radiation at all.

II. Fabrication of NRD-Guide T-Junction

Fig. 1 shows a fabricated NRD-guide T-junction. It is a dielectric T-shaped structure sandwiched between parallel metal plates separated by a distance of less than half a wavelength, according to the principle of NRD-guide operation. Teflon, whose relative dielectric constant is 2.04 at millimeter wavelengths, was used as strip material because of its low-loss nature. As expected, this structure functioned as a practical T-junction, unaffected by radiation, if it was excited at port 1 with electromagnetic waves whose electric field was parallel to the metal plates.

The junction performance can be evaluated by measuring the output power at port 2 and port 3, when port 1 is excited. The feeding guide is called the main arm, while the remaining parts are called the stubarms in the following discussion. Since measurements were made at 35 GHz, the plate separation of the NRD-guide was chosen to be 4.0 mm so as to be less than half a wavelength.

Since, as is well known, any three-port network cannot be simultaneously matched at all ports, matching was attempted only at port 1 to extract half the amount of input power from port 2 and half from port 3. For this purpose, the fabricated T-junction was provided with a dielectric stub in the form of an extension of the main arm, and thin rectangular metal patches were attached on both the free surfaces of the main arm. The sub

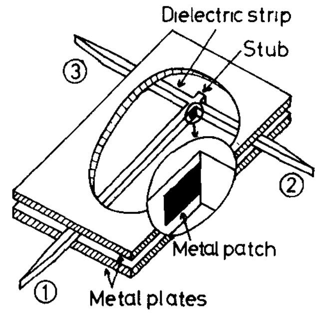

Fig. 1. Sketch of a fabricated NRD-guide T-junction

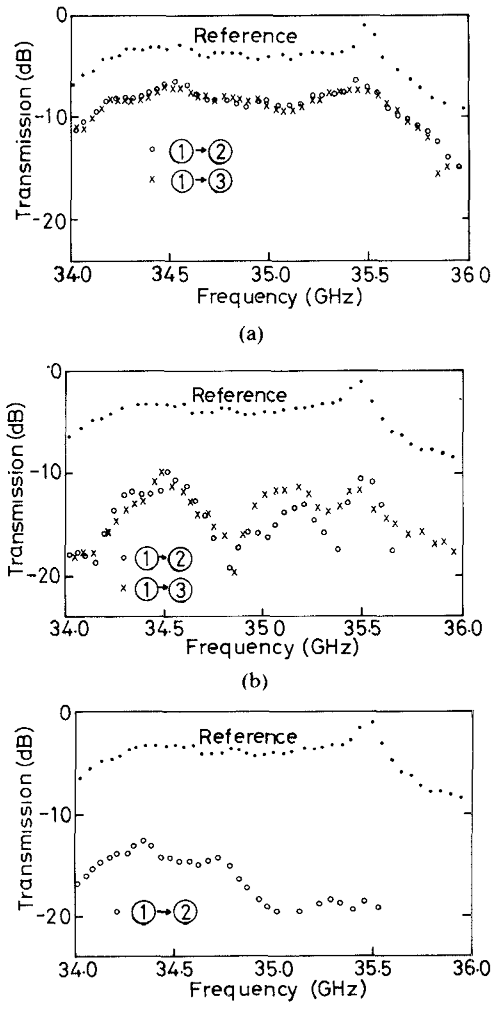

Fig 2 Measured performances of NRD-guide T-junctions of various types. (a) T-junction with a dielectric stub of 3.5 mm in length. The main arm and subarm are 3.5 mm and 3.0 mm in width, respectively. (b) T-junction without a dielectric stub. The widths of the main arm and subarm are the same as those in (a). (c) T-junction with a dielectric stub of appropriate length. The main arm is the same in width as that in (a), but the subarm is 5.0 mm in width.

Manuscript received January 4, 1985; revised May 15, 1985.

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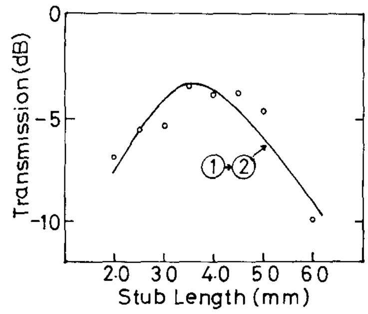

Fig. 3. Output power level measured at 34.5 GHz as a function of stub length for a specific NRD-guide T-junction whose main arm and subarm widths are the same as those in Figs 2(a).

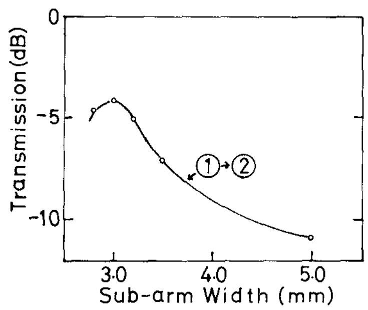

Fig. 4. Output power level measured at 34.5 GHz as a function of width of the subarm for a fixed width of the main arm, 3.5 mm.

and the metal patches serve as effective matching elements to cancel out undesired reflection at the junction.

The effect of the dielectric stub on T-junction performance will be discussed first. Fig. 2(a) shows the T-junction performance measured over a frequency range of 34 to 36 GHz. The main arm and subarm were 3.5 mm and 3.0 mm in width, respectively. The optimum length of the stub was found to be 3.5 mm in this case. The reference curve at the top corresponds to the transmission of a straight NRD-guide equal in length to the T-junction path length. The other two plots, which almost coincide, show power levels detected at either end of the subarm when the port 1 was excited. The power levels are well-balanced, but are 4 dB or more below the reference due to residual reflections at the junction.

The characteristics of another T-junction, which has no dielectric stub, are shown in Fig. 2(b). The power levels are greatly reduced and fluctuate with frequency. By comparing this figure with Fig. 2(a), it may be seen that the stub is essential for improving the T-junction performance. Fig. 3 shows the output power level measured at 34.5 GHz as a function of the stub length for a specific T-junction whose main arm and subarm are 3.5 mm and 3.0 mm in width, respectively. The optimum stub length can be seen to be about 3.5 mm.

The next consideration is concerned with the optimum widths of the main arm and subarm. Fig. 2(c) shows the performance of another T-junction whose subarm is 5.0 mm in width. Though the length of the stub was adjusted carefully, no increase in the power level was detected. Comparison of Figs. 2(a) and (c) indicates that the widths of the dielectric strips are also important parameters in designing NRD-guide T-junctions. Fig. 4 shows the output power level measured at 34.5 GHz as a function of the width of the subarm for a fixed width of the main arm, 3.5 mm. The optimum width of the subarm is seen to be 3.0 mm for the 3.5-mm-wide main arm.

Even if the widths of the dielectric strips are designed properly and the length of the stub is optimized, there is still another problem, namely, the suppression of residual reflection at the junction. During the experiments, it was noticed that thin metal patches attached to the free surfaces of a dielectric strip serve as very effective matching elements with broad-band characteristics. In order to facilitate fabrication of these patches, Teflon sheeting, 0.14 mm in thickness and deposited with copper, was chosen. Various sizes of rectangular copper patches were etched and then attached at different positions on the main arm. Measurements were repeated until a satisfactory performance could be attained.

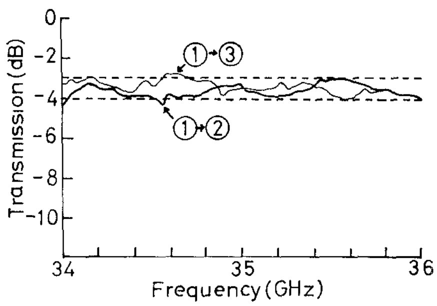

Fig. 5. Measured performance of NRD-guide T-junction of the final form as illustrated in Fig. 1.

The T-junction having the characteristics plotted in Fig. 2(a) was adopted as the prototype. As previously mentioned, its main arm and subarm are 3.5 mm and 3.0 mm in width, respectively, and it is provided with a dielectric stub 3.5 mm in length. The width, length, and position of the metal patches affect T-junction performance in different ways. After optimizing these parameters with respect to the bandwidth of operation and output power level, the characteristics of the NRD-guide T-junction as shown in Fig. 5 could be obtained. The metal patches used were 2.0 mm in width and 5.0 mm in length, and were attached at the base of the junction as illustrated in Fig. 1. They are not only useful for achieving excellent electrical performance of T-junctions but, being small in size and perfectly flat and low in profile, they are also quite practical.

Output power levels are within the range -3 dB to -4 dB over a bandwidth of at least 2 GHz around 35 GHz. Small irregularities superimposed on a large-scale variation in power-level curves may be attributed to inaccuracies in machining. Therefore, if the technique of machining is improved, such irregularities should be eliminated.

III. Conclusion

A practical T-junction has been fabricated at 35 GHz by taking advantage of the nonradiating nature of the NRD-guide. The fabricated T-junction includes a dielectric stub and thin metal patches attached to the free surfaces of the main arm as matching elements. The widths of the main arm and subarm were optimized with respect to the output power level. Well-balanced outputs in excess of -4 dB were obtained over a bandwidth of 2 GHz at a center frequency of 35 GHz. The technique of matching described here may also be useful for other NRD-guide circuit components.

Acknowledgment

The authors wish to thank N. Hayashi and Y. Sasaki for their assistance in computation and measurements.

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A Novel Method for the Analysis of Microwave Two-Port Active Mixers

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***Abstract*—An analysis method for a microwave mixer, based on an active nonlinear device, is developed. The method is a two-port extension of Kerr's work on a one-port (diode)-type mixer and utilizes the harmonic balance approach for the large-signal analysis portion. It results in a relatively fast and efficient program, the use of which is demonstrated by simulating the operation of a mixer based on a NEC MESFET for a range of local oscillator available power and frequency conditions.**

I. Introduction

Microwave mixers have normally been based on Schottky-barrier diodes. Lately, major advances in GaAs technology have led to a new type of an active microwave mixer based on a two-port device, the MESFET, which is likely to find wide use in the future [1]–[5]. The advantages of such a mixer are a) conversion gain, b) self-oscillation option, c) natural isolation between device terminals (makes filters unnecessary), and d) compatibility with GaAs monolithic technology.

While experimental data on MESFET mixers is available, little has been published so far on theoretical methods for two-port mixer analysis. Preliminary investigation of this topic was performed by Pucel *et al.* [1], Ntake [2], Harrop [3], and Kurita and Morita [4]. A more systematic approach by Mass [5] uses Kerr's single-port method [6], [7], originally applied to diode mixers, for each of the nonlinear elements in the FET equivalent circuit.

This paper proposes a new approach to the general analysis of two-port mixers which is a true two-port extension of Kerr's method. The active device is represented by a mathematical two-port model and not by a specific equivalent circuit. This simplifies the adaptation of the analysis program to other devices: one needs only change the subroutine that calculates the two-port parameters. The method is quite general and any type of mixer can be analyzed (ordinary, subharmonic, high IF, etc.).

Moreover, as signal analysis is performed by the harmonic balance method, which is very fast and accurate, any number of the LO harmonics and small signals can be considered to improve accuracy.

Manuscript received February 7, 1985; revised June 6, 1985.

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II. Outline of the Mixer Analysis Method

As in Kerr's work [6], [7], the analysis method is composed of large- and small-signal analyses parts. The first is performed on the nonlinear circuit excited by the large LO signal alone. As the other signals present are very small, it is assumed that they do not affect the device parameters and all the pumping effects are due to the LO signal only. From the voltage and current waveforms obtained, conductance-like and capacitance-like waveforms are derived which are then used in the small-signal analysis to characterize the device as a linear time varying network. The relations between the small-signal voltages and currents of the various frequencies are obtained from the network in the form of an admittance matrix whose elements are 2×2 submatrices. Combining this matrix with the source and load constraints at each frequency makes possible the calculation of the mixer performance parameters, such as conversion gain and input and output impedances. The two parts of the analysis are detailed below.

III. Large-Signal Analysis

The large-signal analysis step uses the standard two-port representation for the device. Six coefficients which are nonlinear functions of the voltages $V_1(t), V_2(t)$ are used to relate the currents and voltages

$$I_1(t) = I_{\text{con}_1} + (CV)_{11}\frac{dV_1(t)}{dt} + (CV)_{12}\frac{dV_2(t)}{dt} \qquad (1)$$

$$I_2(t) = I_{\text{con}_2} + (CV)_{21}\frac{dV_1(t)}{dt} + (CV)_{22}\frac{dV_2(t)}{dt} \qquad (2)$$

where

$I_1(t)$ current waveform at port 1

$I_2(t)$ current waveform at port 2

$V_1(t)$ voltage waveform at port 1

$V_2(t)$ voltage waveform at port 2.

A computer model that calculates the six coefficients for a given V_1, V_2 in the case of a MESFET was presented by Madjar and Rosenbaum [8], [9] and later modified by Green *et al.* [10]. This model is used here.

The complete circuit of the MESFET is shown in Fig. 1 in which the box at the center represents the intrinsic FET.

In order to obtain the current and voltage waveforms in the steady state, the harmonic balance method [11] is used because of its efficiency (short computing times). The method divides the network into linear and nonlinear parts and then matches the currents between the two parts at their interface. The voltages at the interface (which are of course the same for the two parts), are used to derive these currents. Treating the interface voltages as variables and the difference in the respective currents at the interface as the function for an optimization routine leads to the correct solution for voltages and currents at the interface nodes. As these quantities are presented by their Fourier series, it is the Fourier coefficients that are the optimization variables. The ZSPOW optimization routine from IMSL (International