A Planar Quasi-Optical Subharmonically Pumped Mixer Characterized by Isotropic Conversion Loss

KARL D. STEPHAN, MEMBER, IEEE, AND TATSUO ITOH, FELLOW, IEEE

Abstract —By using a subharmonically pumped circuit in a quasi-optical planar mixer, we have found it possible to use an LO frequency of one-half the normal value with little added circuit complexity. This circuit shows conversion loss as low as 8.6 dB $\pm$2 dB at 14 GHz. Through the means of a newly defined quasi-optical mixer parameter called isotropic conversion loss (L_{iso}), we find that performance of the mixer system degrades less than 10 dB from an RF input of 14 GHz to 35 GHz, which is more than twice the designed RF frequency.

I. Introduction

A QUASI-OPTICAL mixer combines the functions of a receiving antenna and a mixer in one device. As several groups have recently shown [1], [2] such mixers can be fabricated on a planar dielectric substrate by photolithographic techniques. Such simple methods promise to make imaging arrays of quasi-optical mixers a reality in the near future. One significant problem remaining, however, is the question of providing local oscillator (LO) power to a large array of such mixers.

The LO power requirement for a quasi-optical mixer circuit is essentially the same as that of a conventional mixer using the same number and kind of diodes. For typical microwave diodes, the power needed is about 5 to 50 mW per device, depending on diode barrier height and material. Many quasi-optical mixer designs couple the LO power through free space. While by far the simplest method of distributing power to a number of mixers, this free-space technique involves substantial loss between the LO feed horn and the mixers facing it. For arrays of usable size, the total LO requirement will approach the multi-watt range. Above 30 GHz, where the most interesting imaging applications lie, such LO power levels are hard to obtain economically.

II. Subharmonic Mixing

By using a properly designed subharmonically pumped mixer [3], one can maintain good mixing performance at millimeter wavelengths while using an LO frequency of one-half the value required for a conventional mixer. Since a given power level is much easier to achiever at 60 GHz than 120 GHz, for example, a quasi-optical mixer employing such a subharmonic pump has a great practical advantage.

One particularly efficient type of subharmonically pumped mixer uses a pair of antiparallel (head-to-tail) diodes. A simple ideal switch model shows that the pair of diodes turns on and off twice per LO cycle, effectively doubling the LO frequency. Later in this paper we will describe a quasi-optical mixer using this approach, but first we shall explore the methods currently used to characterize quasi-optical mixers in general.

Manuscript received May 17, 1983; revised August 23, 1983. This work was supported in part by the U.S. Army Research Office under Contact DAAG29-81-K-0053.

K. D. Stephan was with the University of Texas at Austin. He is now with the Department of Electrical and Computer Engineering, University of Massachusetts, Amherst, MA 01002.

T. Itoh is with the Department of Electrical Engineering, University of Texas at Austin, Austin, TX 78712.

III. Conversion Loss Versus L_{iso}

The development of a new class of devices brings with it the obligation to define appropriate and meaningful measures of performance for them. These measures should allow direct comparison of different devices in different laboratories, and should ideally be established through laboratory procedures that are easy and simple to perform with high precision. In this section, we propose to show that conversion loss L as defined for conventional mixers may not completely describe the performance of quasi-optical mixers. This is because the antenna and the mixing element are inseparable in a true quasi-optical mixer, preventing independent measurements on either half alone. Instead of conversion loss L, we define a new parameter L_{iso} for quasi-optical mixers. The quantity L_{iso} is not only more comprehensive but is also much easier to measure accurately than conventional conversion loss.

The need for the new parameter L_{iso} arises from a fundamental difference between conventional and quasi-optical mixers. The input signal of a conventional mixer is measured in watts available at the well-defined input port of the network element called the mixer. On the other hand, a network approach is unsuitable when one considers an array of quasi-optical mixers forming a kind of microwave photographic film at the focus of an imaging system. While a clearly defined single-mode port cannot be found for such a system, one can easily measure the power *density* at the focal plane in W/m^2. This essential difference is elucidated in the following examples.

In Fig. 1 is illustrated a conventional receiving system consisting of an antenna feeding a mixer whose IF output is proportional to the incoming RF wave intensity. The

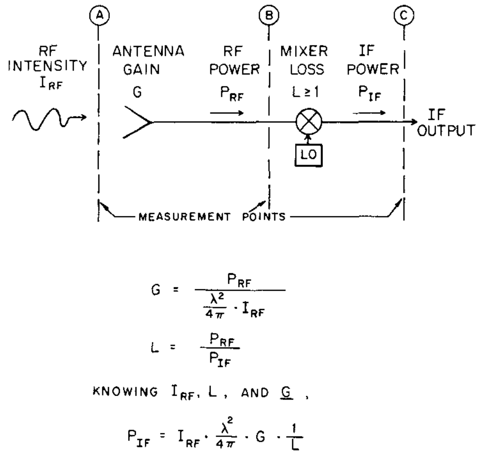

Fig. 1. System calculations for separate antenna and mixer.

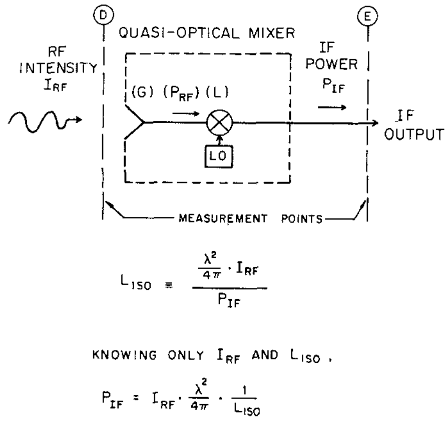

Fig. 2. $L_{\rm iso}$ used to characterize performance of quasi-optical mixer.

dimensions of this intensity are power per unit area (W/m^2). For simplicity, we shall make the following assumptions. 1) The incoming single-frequency plane wave arrives from the direction in which the antenna portion of the mixer has maximum gain. 2) No power is reflected at the mixer input or the IF output. 3) The image conversion loss L_I is much larger than the signal conversion loss L_S. The third condition insures that the overall conversion loss L is approximately equal to L_S. For mixers with substantial response at the image frequency, the following discussion applies directly if L is modified according to [4]:

$$\frac{1}{L} = \frac{1}{L_S} + \frac{1}{L_I}. \tag{1}$$

Three points in this system are conveniently accessible for measurement. We can measure the incoming wave intensity I_{RF} at point A directly by replacing the receiving antenna with a standard-gain antenna feeding a power meter. Since the antenna and mixer are separate components, we can remove the mixer input at point B and measure RF power P_{RF}. And finally the IF output power P_{IF} can be measured without disturbing the system at all.

With these three measurements, we can calculate the performance of the antenna and mixer separately, and then use the antenna gain and mixer loss to find the IF output for a given intensity at the input. This last ratio of IF output to RF intensity at the mixer is all the system designer needs, although for diagnostic purposes it may be convenient to know the separate contributions of antenna gain G and mixer loss L to the overall system performance.

Contrast the conventional system just described with the quasi-optical mixer system illustrated in Fig. 2. By examining this figure, one can see clearly why a statement of conversion loss (L) *only* is inadequate to describe the system performance. Without knowing the gain (G) of the antenna portion of the quasi-optical mixer, it is impossible to predict the IF output arising from a known incoming wave intensity. A mixer connected to a high-gain antenna will deliver more IF power with a given RF illumination than the same mixer fed from a low-gain antenna. Under certain conditions, it may be possible to separate the antenna and mixer to measure (G). With (G) and (L) known, both conversion loss and system performance can be predicted from measurements at points D and E. Where the antenna and mixer cannot be separated, indirect means must be used to find antenna gain. One indirect method used in the results to follow involves integrating extensive antenna radiation patterns to obtain a directivity for the antenna portion of the quasi-optical mixer. Assuming no losses, this directivity equals the antenna gain (G), allowing the calculation of P_{RF} and conversion loss. But for reasons we will state, such indirect methods are awkward at best and prone to numerous errors.

Especially at millimeter wavelengths, quasi-optical mixers tend to become highly integrated structures. If the structure is disturbed with cables or other devices intended to measure available RF power and antenna gain, the conditions of measurement are no longer identical to the conditions of use as a mixer, and the results thus obtained are cast into doubt. Even the indirect gain measurement mentioned above is difficult to perform with the local oscillator feed system in place, which it should be for realistic antenna patterns. In Fig. 2, the dashed line around the interior of the quasi-optical mixer indicates the relative inaccessibility of the antenna–mixer path. Given this problem, the only quantities available for direct measurement are the RF intensity at point D and the IF output power at point E.

Using these two quantities, we can define a new parameter which not only characterizes the quasi-optical mixer as well as separate statements of L and G, but is much easier to measure precisely. We call this new measure $L_{\rm iso}$ since it is equivalent to the mixer conversion loss one would deduce if the antenna were assumed to be isotropic, i.e., with $G = 1$. The definition of $L_{\rm iso}$ is simply the ratio of conventional conversion loss L to antenna gain G at the input signal wavelength λ:

$$L_{\rm iso} \equiv \frac{L}{G} = \frac{\frac{\lambda^2}{4\pi} \cdot I_{RF}}{P_{IF}}. \tag{2}$$

Neither L nor G is independently determinable without internally disturbing the mixer, but the ratio $L/G = L_{\rm iso}$ is. Only the incoming RF intensity I_{RF} at D and the output power P_{IF} at E are needed to find $L_{\rm iso}$. Once $L_{\rm iso}$ is measured, the system designer uses the following equation to find P_{IF} given a certain intensity I_{RF}:

$$P_{IF} = I_{RF} \cdot \frac{\lambda^2}{4\pi} \cdot \frac{1}{L_{\text{iso}}}. \tag{3}$$

Strictly speaking, L_{iso} is a function of the incoming wave's direction, since it includes the angle-dependent antenna gain term G. Nevertheless, if a plane wave incident normal to the antenna surface is established as a standard for quasi-optical mixers, a system designer can use L_{iso} by merely finding the equivalent plane wave intensity at the focal plane of his quasi-optical system. For waves arriving off-axis, corrections for the angular dependence of L_{iso} must be made, but the basic utility of the measure remains.

It should be emphasized that the parameter L_{iso} is most useful for quasi-optical mixers in isolation, unaided by lenses or other gain-enhancing devices. When a system is tested which incorporates both a focussing element and a mixer, L_{iso} may become numerically small because of the high effective gain of the total system. But for mixers, especially planar types, whose antenna dimensions are limited to a few wavelengths at most, antenna gain will be modest and L_{iso} will remain a convenient and easily measured quantity. One such mixer we will now describe.

IV. The Bowtie Antenna

A basic requirement for the antenna portion of a subharmonically pumped quasi-optical mixer is that it must efficiently receive the incoming RF signal as well as the LO power at about one-half the frequency of the RF signal. While separate narrow-band antennas could be used for the RF and LO waves, a simpler approach is to choose an antenna structure with a bandwidth exceeding one octave. Based on this criterion, the bowtie antenna was selected for use. Other workers have had good experience with similar antennas in quasi-optical applications [5].

The metal-only bowtie antenna was the subject of an intensive experimental study by Brown and Woodward at RCA [6]. They tabulated extensive impedance and radiation pattern data which showed that a properly proportioned bowtie antenna is capable of matching a 300-Ω line with less than 2:1 VSWR over a frequency range of nearly one octave.

An equivalent-circuit model was found to account reasonably well for both the radiation pattern and the input impedance of the metal-only bowtie. If the antenna has dimensions given in Fig. 3 (fin length = r_0 and fin half-angle = ψ), the input impedance is well modeled over the useful frequency range by the equivalent circuit shown in Fig. 4. The derivation of this model is given elsewhere [7], but is based on the fact that for wavelengths longer than about $r_0/4$, the bowtie excites primarily the TM_{11} mode of free space, which has a dipole-like radiation pattern. Plots of impedance calculated from the model in Fig. 4 are compared with measurements of Brown and Woodward in Figs. 5 and 6, and agreement is seen to be reasonably good for such a simple model.

To make a usable structure incorporating delicate beam-lead diodes, we decided to support the entire bowtie an-

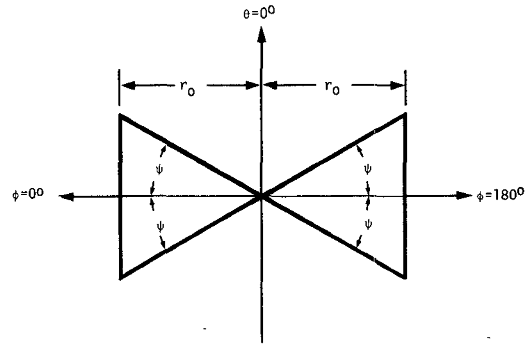

Fig. 3. The metal-only bowtie antenna.

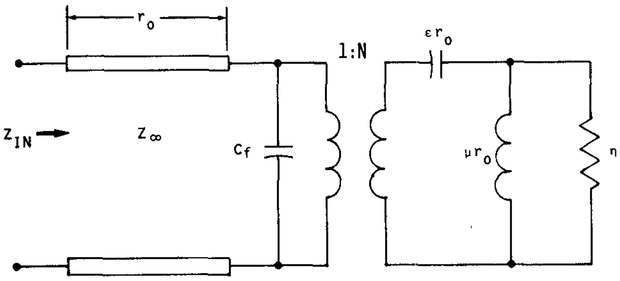

Fig. 4. Simplified equivalent circuit of bowtie antenna using only TM_{11} mode.

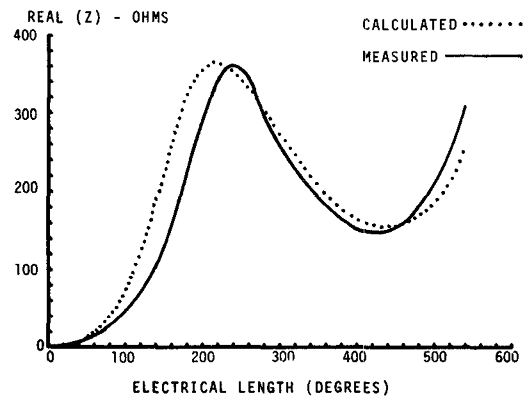

Fig. 5. Calculated and measured real part of metal-only bowtie antenna impedance versus electrical length.

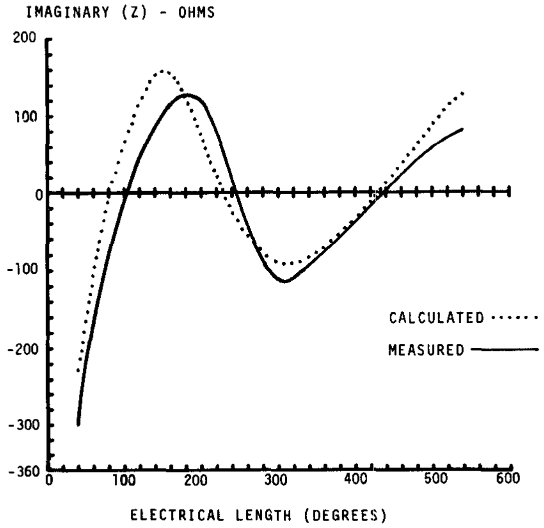

Fig. 6. Calculated and measured imaginary part of metal-only bowtie antenna impedance versus electrical length.

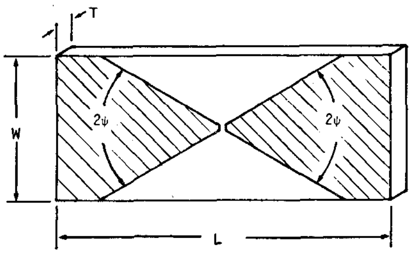

Fig. 7. Bowtie antenna on a dielectric substrate.

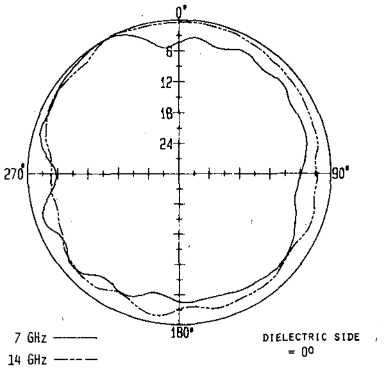

Fig. 8. E-plane patterns, 7 and 14 GHz, bowtie antenna.

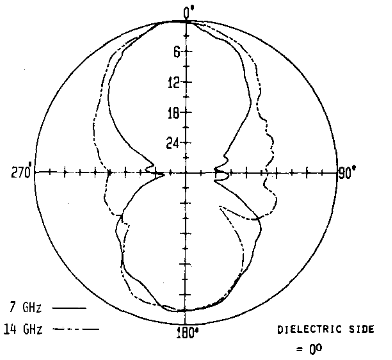

Fig. 9. H-plane patterns, 7 and 14 GHz, bowtie antenna.

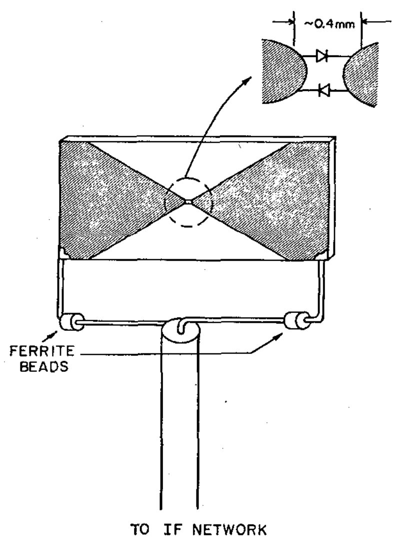

Fig. 10. Subharmonic mixer using antiparallel diodes and bowtie antenna.

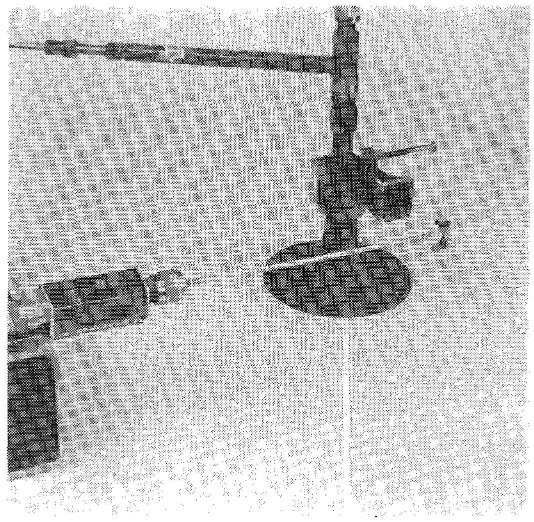

Fig. 11. Photograph of bowtie mixer in front of LO horn.

tenna on a dielectric substrate as shown in Fig. 7. If the dielectric is relatively thin and has fairly low permittivity, the behavior of the antenna will not deviate excessively from that of the metal-only structure. The substrate also permits forming the antenna by photolithography and etching a conductive cladding from the surface.

The dimensions of the antenna used are: length $L = 22.6$ mm, width $W = 9.45$ mm, and thickness $T = 1.57$ mm. The dielectric constant of the substrate was about 2.2, and the fin half-angle $\psi = 30°$. Radiation patterns using this antenna were measured at 7 and 14 GHz, and are shown in Figs. 8 and 9. By adding two antiparallel diodes and IF takeoff leads as shown in Fig. 10, the subharmonic mixer was completed. The ferrite beads allow the VHF IF signal to pass but tend to block RF currents that would otherwise distort the radiation pattern. A photograph of the mixer in front of the LO feed horn is shown in Fig. 11.

V. Mixing Experiments at 14 and 35 GHz

After extensive pattern measurements were made at 14 GHz, numerical spherical integration was used to find the maximum directivity of the dielectric-supported bowtie. This was equated to maximum theoretical gain of the antenna, and was found to be about 4.5 dB on the dielectric side. This figure was used to calculate mixer conversion loss L in Table I. No pattern measurements were made at 35 GHz, but isotropic conversion loss L_{iso} could still be measured at this millimeter wavelength, since no independent determination of antenna gain is needed for L_{iso}.

The experimental setup is shown in Fig. 12. The RF source on the left was monitored with a dual directional coupler both to measure power delivered to horn #1 and to insure that all the forward power was in fact radiated. The horn–mixer distance R_1 was chosen so that the mixer was in the far field of the horn, allowing the intensity I_{RF} to be calculated from the power at the horn and its gain. To avoid excessive LO power loss, the distance R_2 from the LO feed horn #2 to the mixer was made as small as possible, varying from 1 to 5 cm. The presence of this horn in the near field of the antenna perturbs the radiation

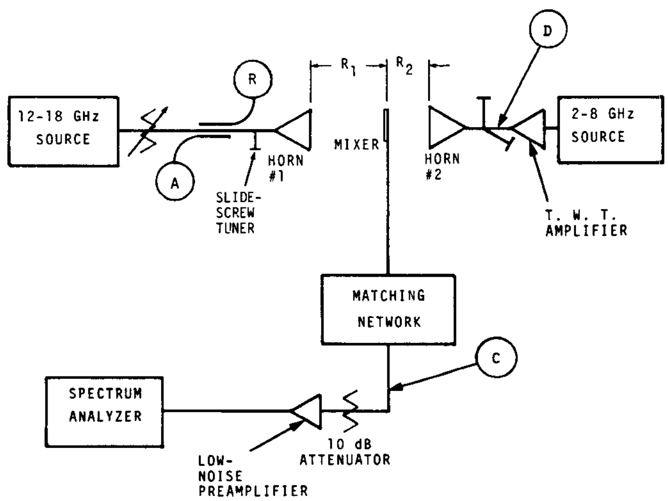

Fig. 12. Apparatus for conversion loss measurements.

TABLE I
BOWTIE SUBHARMONIC MIXER PERFORMANCE

Max. C_T per diode (pF)	f_{RF} (GHz)	f_{LO} (GHz)	f_{IF} (MHz)	L (dB)	L_{iso} (dB)
0.15	13.68	7.02	367	16.0	11.5
0.1	14.29	7.00	290	8.6	4.1
0.1	13.68	7.00	313	9.2	4.7
0.15	34.52	17.09	348	--	25.1
0.1	34.52	17.11	300	--	13.6
			Estimated Errors:	±2 dB	±1.5 dB

pattern and can lead to disagreement between the antenna gain in use and the gain measured with an isolated antenna. These problems affect only conversion loss L, since the definition of L_{iso} automatically includes any such system effects. Local oscillator power delivered to horn #2 was measured at point D.

The IF system was conventional, using a VHF matching network feeding a 10-dB attenuator followed by a low-noise amplifier. The presence of the attenuator, unfortunate from a noise-figure standpoint, was made necessary by the poor return loss of the amplifier input. The attenuator provided a known load impedance at point C where IF output power was measured by a substitution method.

As Table I indicates, two types of diodes were used in the mixer circuit. The diodes having $C_T = 0.15$ pF were Hewlett-Packard 5082-2299 silicon beam-lead devices, while the 0.1-pF units were H-P 5082-2264 diodes. The 0.1-pF devices gave much better performance at both *Ku*-band (14 GHz) and *Ka*-band (35 GHz). The best conversion loss obtained was 8.6 dB$\pm$2 dB, for an IF of 290 MHz. Conversion loss using the same LO frequency but the lower sideband was slightly higher (9.2 dB), but this change is probably due to slight shifts in the radiation pattern from 14.29 GHz to 13.68 GHz.

It should be stressed that the same antenna dimensions were used at both *Ku*-band and *Ka*-band. At 34.52 GHz the bowtie is being used at more than twice its designed upper frequency limit, yet the isotropic conversion loss L_{iso} increases only 9.5 dB (from 4.1 dB to 13.6 dB). A smaller bowtie designed for 35 GHz would undoubtedly show better performance. Note that the error associated with L_{iso} is smaller than the conversion loss error (± 1.5 dB versus ± 2 dB). This is due to the added uncertainties of the separate antenna gain measurements used to calculate L.

VI. SUMMARY

We have shown how the newly defined parameter L_{iso} characterizes quasi-optical mixers more completely than conventional conversion loss L. This new parameter was used in measurements of a subharmonically pumped mixer using a bowtie antenna, although it can be applied to any form of quasi-optical mixer or receiver. A simple equivalent-circuit model predicted reasonably stable antenna impedance over the 7–14-GHz frequency range used for the LO and RF inputs. Antiparallel beam-lead diodes were mounted at the terminals of a dielectric-supported bowtie to form a mixer whose measured conversion loss was 8.6 dB at 14.29 GHz. The corresponding L_{iso} at that frequency was 4.1 dB. Mixer performance as measured by L_{iso} degraded less than 10 dB when the RF input was raised to 34.52 GHz, indicating that a properly scaled version of the mixer should be very useful at millimeter wavelengths.

VII. CONCLUSIONS AND FURTHER WORK

A single quasi-optical subharmonically pumped mixer has been shown to work well. The most promising areas of application involve arrays of such devices at the focal plane of an imaging system. Before such arrays are attempted, questions of mutual coupling and IF feed techniques must be addressed. Also, the fundamental criterion of receiver performance is noise temperature, which was not measured for the device described. The most meaningful measurement of noise temperature will be one characterizing the mixer, antenna, and image-forming element (lens or reflector) as an integrated system rather than measurements of individual components. Once these problems are surmounted, we can expect the fabrication of a truly integrated array on a GaAs or other semiconductor dielectric, in which the mixer diodes are formed directly on the same substrate used for the array.

ACKNOWLEDGMENT

The authors would like to thank those who reviewed this manuscript for suggesting a number of helpful clarifications and useful references.

REFERENCES

[1] L. Yuan, J. Paul, and P. Yen, "140 GHz quasi-optical planar mixers," in *1982 MTT Int. Microwave Symp. Dig.*, pp. 374–375.
[2] K. Stephan, N. Camilleri, and T. Itoh, "A quasi-optical polarization-duplexed balanced mixer for millimeter-wave applications," *IEEE*

Trans. Microwave Theory Tech., vol. MTT-31, pp. 164–170, Feb. 1983.

[3] M. V. Schneider and W. W. Snell Jr., "Harmonically pumped stripline down-converter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, pp. 271–275, Mar. 1975.

[4] M. V. Schneider, "Metal-semiconductor junctions as frequency converters," in *Infrared and Millimeter Waves*, K. J. Button, Ed. New York: Academic Press, 1982, ch. 4, pp. 248–249.

[5] D. B. Rutledge and M. S. Muha, "Imaging antenna arrays," *IEEE Trans. Antennas Propagat.*, vol. AP-30, pp. 535–540, July 1982.

[6] G. H. Brown and O. M. Woodward, "Experimentally determined radiation characteristics of conical and triangle antennas," *RCA Review*, vol. 13, pp. 425–452, Dec. 1952.

[7] K. Stephan, "A study of microwave and millimeter-wave quasi-optical planar mixers," Ph.D. dissertation, University of Texas at Austin, May 1983.

Karl D. Stephan (S'77–M'77) was born in Fort Worth, TX, on December 18, 1953. He received the B.S. degree in engineering from the California Institute of Technology, Pasadena, CA, in 1976, and the M. Eng. degree from Cornell University, Ithaca, NY, in 1977.

In 1977, he joined Motorola, Inc. in Fort Worth, TX. From 1979 to 1981, he was with Scientific-Atlanta, Atlanta, GA, where he engaged in research and development pertaining to cable television systems. After working at Hughes Aircraft in the summer of 1982 as a Student Engineer, he received the Ph.D. degree in electrical engineering from the University of Texas at Austin in 1983. In September 1983, he joined the faculty of the University of Massachusetts at Amherst, where he is presently Assistant Professor of Electrical Engineering.

Dr. Stephan is a member of Tau Beta Pi.

Tatsuo Itoh (S'69–M'69–SM'74–F'82) received the Ph.D. degree in electrical engineering from the University of Illinois, Urbana, in 1969.

From September 1966 to April 1976, he was with the Electrical Engineering Department, University of Illinois. From April 1976 to August 1977, he was a Senior Research Engineer in the Radio Physics Laboratory, SRI International, Menlo Park, CA. From August 1977 to June 1978, he was an Associate Professor at the University of Kentucky, Lexington. In July 1978, he joined the faculty at the University of Texas at Austin, where he is now a Professor of Electrical Engineering. During the summer 1979, he was a Guest Researcher at AEG-Telefunken, Ulm, West Germany. Since September 1983, he has held the Hayden Head Centennial Professorship of Engineering at the University of Texas.

Prof. Itoh is a member of the Institute of Electronics and Communication Engineers of Japan, Sigma Xi, and Commission B of USNC/URSI. He serves on the Administrative Committee of IEEE Microwave Theory and Techniques Society and is the Editor of IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES. He is a Professional Engineer registered in the State of Texas.

The Microstrip Open-Ring Resonator

INGO WOLFF, MEMBER, IEEE, AND VIJAI K. TRIPATHI, MEMBER, IEEE

***Abstract*—The open-ring microstrip resonator is analyzed by utilizing the two-dimensional magnetic wall model. The solution and the numerical results for the eigenvalues and the electromagnetic fields for various resonant modes are presented. It is shown that the experimental results are in good agreement with the theoretical predictions based on this model.**

I. INTRODUCTION

OPEN-RING microstrip resonators have been proposed for applications in microwave filters [1] and as planar antenna elements [2]. The structure is analyzed in this paper by utilizing the two-dimensional magnetic wall model, and the results computed for eigenvalues (resonant frequencies) and electromagnetic field distribution are presented. The problem is similar to that of the disc and the closed-ring resonators, which have been studied extensively in recent years for applications as resonators and planar antenna elements [2]–[10]. The magnetic wall model, though an approximate one, has been successfully used in the past for many microstrip patch geometries, including discs and annular rings.

Manuscript received May 18, 1983; revised August 3, 1983.

I. Wolff is with the Department of Electrical Engineering, Duisburg University, Bismarckstr. 81, 4100 Duisburg, Germany.

V. Tripathi was on sabbatical leave at the Department of Electrical Engineering at Duisburg University, Duisburg, Germany. He is with the Department of Electrical and Computer Engineering, Oregon State University, Corvalis, OR.

II. THE MODEL AND THE EIGENVALUE PROBLEM FOR THE MICROSTRIP OPEN-RING RESONATOR

The microstrip open-ring resonator, as shown in Fig. 1(a), consists of a planar ring segment having an inner radius r_i, an outer radius r_a, and an angle α of the open segment on a substrate of height h. The corresponding magnetic wall model for the resonator is defined in Fig. 1(b). This consists of a ring with effective inner and outer radii $r_{i,\text{eff}}$ and $r_{a,\text{eff}}$, respectively, vertical magnetic walls

0018-9480/84/0100-0102$01.00 ©1984 IEEE