Application of Inverted Strip Dielectric Waveguides for Measurement of the Dielectric Constant of Low-Loss Materials at Millimeter-Wave Frequencies

TATSUO ITOH, SENIOR MEMBER, IEEE, AND FWU-JIH HSU, MEMBER, IEEE

***Abstract*—An entirely new method for measuring dielectric properties of slab-type materials is developed by using a novel dielectric waveguide structure originally designed for millimeter-wave integrated circuits. The method entails the measurement of the stopband of the grating structure created in the dielectric waveguide. Several examples of measurement results are reported.**

I. Introduction

With increasing use of dielectric resonators [1], dielectric waveguides [2], and printed transmission lines at higher frequencies, simple methods for accurately predicting the properties of dielectric materials become very important. There are currently a number of methods available for measuring dielectric properties of material media at microwave and millimeter-wave frequencies. One typical method is the use of waveguides or waveguide cavities, which are either partially or completely filled with dielectric materials to be measured [3]. Cavities made of striplines or microstrip lines have also been used [4], [5]. In free-space environment, scattering of electromagnetic energy from a dielectric sphere can be used for evaluating the dielectric property of the sphere [6].

The paper proposes a novel approach to measurement of dielectric constants of low-loss slab materials at microwave and millimeter-wave frequencies. The method makes use of stopband phenomena observed in periodic structures created in an inverted strip dielectric (IS) waveguide developed recently [7]. The principle of the measurement process will be explained in the following paper.

II. Principle of Measurement

The cross section of the IS waveguide is shown in Fig. 1(a). The major portion of the wave energy propagates in that portion of the guiding layer immediately above the dielectric strip [2]. When periodic grooves are created in the strip as shown in Fig. 1(b), the propagation characteristic of the IS waveguide is periodically modulated and a grating structure is created. Depending on the electrical length of the grating period d, the structure works either as a leaky-wave antenna or a band-reject filter [7].

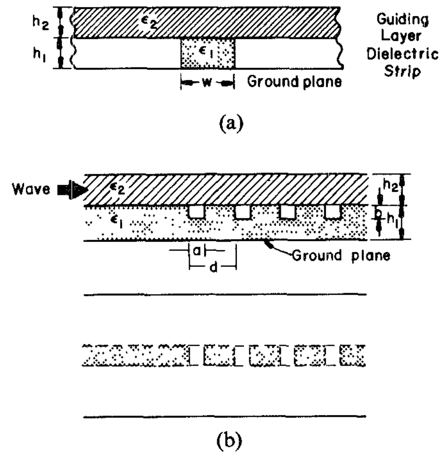

Fig. 1. Experimental setup. (a) Cross section. (b) Side and top views.

Although the antenna scheme may be used it is not as practical as the filter structure as the field measurement is required. Hence, we will use the filter method in this paper. The frequency response of the filter is a function of all the dimensions of the IS waveguide, the grating period, profile of the grating, and the dielectric constants of the materials involved. Hence, if one measures the frequency response of the filter, the dielectric constant of one of the materials may be determined provided all other parameters are known.

If we use a bandstop filter scheme, it is possible to find out the dielectric constant of one of the materials in the waveguide by simply measuring the frequency related to the stopband, such as the center frequency, the lower band edge, or the upper band edge.

In the proposed measurement setup, we use a slab material with unknown dielectric constant as the guiding layer of the IS waveguide. The periodic grooves are

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

F.-J. Hsu is with the Anaconda Telephone Company, Anaheim, CA 92801.

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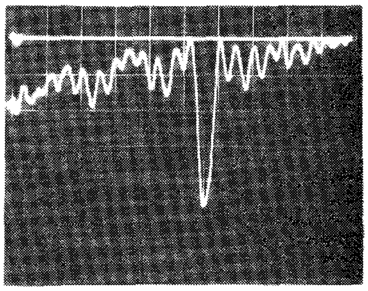

Fig. 2. Typical sweep response from the grating structure: 20 mV/div, 12.4–18 GHz.

created as shown in Fig. 1 in the dielectric strip which has a known dielectric constant. When we excite this structure with a sweep source a strong reflection is observed at the stopband, provided the frequency range of the sweeper covers the stopband. A typical response from an IS waveguide with periodic grooves in its dielectric strip is shown in Fig. 2.

Let us now discuss the algorithm to obtain the dielectric constant from the measured data. First, if we assume that the dielectric constant ϵ_2 of the guiding layer is known, the center frequency f_c of the stopband is given by

$$f_c = F(h_1, h_2, \epsilon_1, w, d, a, b, \epsilon_2) \tag{1}$$

where h_1, w, and ϵ_1 are the thickness, width, and dielectric constant of the dielectric strip, h_2 is the thickness of the guiding layer, and d is the grating period. Also, a and b are the length and the depth of the grooves. F indicates that f_c is a function of all the parameters in the parentheses. No closed form of F is available. However, values of f_c are given numerically, as explained in the Appendix.

Our problem at hand is to solve (1) for ϵ_2, that is, to obtain

$$\epsilon_2 = F^{-1}(h_1, h_2, \epsilon_1, w, d, a, b, f_c). \tag{2}$$

Since there is no closed-form expression for (1), its inversion (2) does not have an analytical form either. However, we may use either a graphical method or a computer-optimization technique to accomplish the solution for ϵ_2. In the graphical method, a number of curves for f_c versus ϵ_2 are generated numerically from (1) for a number of structural parameters. Examples of curves are presented in Fig. 3. Once these curves are available, it is possible to obtain ϵ_2 graphically from the measured value of f_c for a corresponding set of structural parameters. For structures not provided with curves some interpolation procedures may be used.

Computer-optimization techniques, though more time consuming, generally provide more accurate answers than the graphical method. In such a method, a subroutine to perform numerical solutions of (1) is used repeatedly for an assumed value of ϵ_2 until the computed value of f_c becomes virtually identical to the measured f_c. This iteration scheme is carried out in a systematic manner, and the final assumed value of ϵ_2 is considered as that of the unknown material.

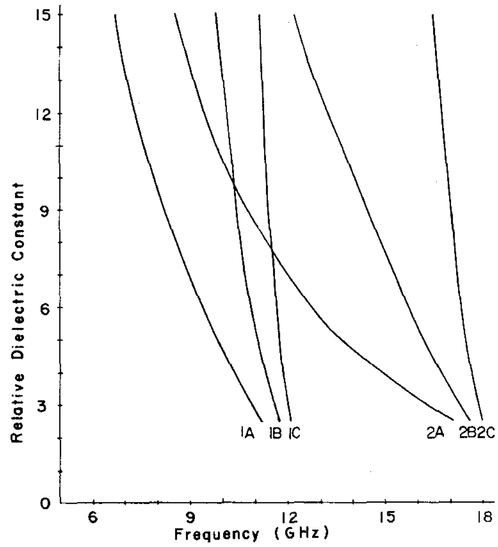

Fig. 3. Calculated results for stopband frequency f_c versus relative dielectric constant; h_2 is the thickness of the dielectric slab, d is the period of the grating, $h_1 = 6.35$ mm, $w = 16$ mm, $a = 1.59$ mm, $b = 3.16$ mm, $\epsilon_1 = 2.1$ (Teflon).

(1A) $h_2 = 6.35$ mm $d = 10$ mm;
(1B) $h_2 = 3.175$ mm $d = 10$ mm;
(1C) $h_2 = 1.5875$ mm $d = 10$ mm;
(2A) $h_2 = 6.35$ mm $d = 6.25$ mm;
(2B) $h_2 = 3.175$ mm $d = 6.25$ mm;
(2C) $h_2 = 1.5875$ mm $d = 6.25$ mm.

III. Measurement Procedures

The IS waveguide with grating sections can be assembled from a large ground plane, a rectangular dielectric strip (ϵ_1) with periodic grooves, and a slab material to be measured which is used as the guiding layer. It is helpful in practice if the period of grooves is determined in such a way that the stopband appears in the frequency range, where the field is well guided in the IS waveguide with the guiding layer made of a commonly used material such as fused quartz. Each component is securely positioned by mechanical pressure between the guiding layer and the ground plane by means of clamps located away from the waveguiding portion. Since any air gap between dielectric materials themselves and the ground plane may be a potential source of error in the final results, smooth dielectric surfaces and relatively strong mechanical pressure are essential. In order to excite the IS waveguide assembly efficiently and to resolve the stopband clearly, a good transition from the conventional rectangular metal waveguide to the assembly is important. We have accomplished this requirement by the use of a tapered rectangular dielectric waveguide. The pointed end of the tapered structure is inserted into the rectangular metal waveguide which is connected to a microwave sweep source by way of a reflectometer setup. The other end of the tapered section is flat and butt-jointed to the end surface of the slab material immediately above the dielectric strip (ϵ_1).

The measurement procedure is as follows. 1) First, obtain a sweep response which is the ratio of the input

TABLE I
EXAMPLES OF MEASURED DIELECTRIC CONSTANTS

Material	ϵ_2 (supplied values)	ϵ_2 (measured)
Stycast HiK	3.75	3.75
Plexiglass	2.59	2.58

power to and the reflected power from the IS waveguide assembly. The range of sweep is adjusted until the stopband appears. Read the center frequency f_c of the stopband. Measure all the pertinent physical dimensions involved. 2) By using either a graphical method or a computer-optimization algorithm explained earlier, obtain the dielectric constant ϵ_2 of the unknown slab material.

IV. Results and Discussions

To demonstrate the validity of the method, several measurements have been performed in the frequency range up to 18 GHz. Although there is no inherent limit in frequency, we could not use any higher frequency due to unavailability of sweeper sources higher than 18 GHz. Examples of measured results are listed in Table I. It is seen that the results obtained in the present method are quite accurate. Note that for these examples we used a computer-optimization algorithm for inversion of (1).

We will now summarize some of the features of the present method.

1) Since the quantity to be measured is the frequency, the error involved in the measurement process itself is small.

2) The method is simple to use and the material to be measured can be easily replaced in the setup since the IS waveguide assembly, including the grating section, is held in position by applying the mechanical pressure between the guiding layer (unknown material) and the ground plane by means of clamps at the location away from the waveguiding portion.

3) The dissipation factor cannot be measured in the present setup. If the attenuation due to the material loss can be isolated, such a factor could be measured. However, in most open structures, it is extremely difficult to separate the radiation loss and material loss.

4) The accuracy of the methods depends on the inversion scheme of the measured data. For instance, both, the method to obtain β_G and β_{NG} and the one for β by using (A1) are approximate, although they are usually quite accurate for the types of structures used here. Another problem is the convergence of the optimization routine. Sometimes, the convergence could be slow and one may be advised that the use of graphical method to obtain initial value in the optimization routine could be helpful.

The method proposed here does not replace other existing methods, in many of which both real and imaginary parts can be measured. Rather, the present method provides a simple alternative when only the real parts of dielectric constants in low-loss materials are needed.

V. Conclusions

We proposed a simple method for measuring the dielectric constant of slab-type materials for microwave and millimeter-wave applications. A number of features of the method are listed as well as results of some example measurements. The latter agreed well with data supplied by manufacturers.

Appendix
Derivation of Equation (1)

Although a more exact analysis may be done, we present here a simple method to obtain the dispersion relation in the grating section in the IS waveguide operated in the surface-wave region. We will model the grating section as periodically cascaded transmission line. The complex phase constant β can be approximated in a manner similar to the one found in [8]. Each unit cell of the grating consists of two transmission lines; the grooved section of length a and the nongrooved section of length $(d-a)$. In this analysis, any junction susceptance between these two transmission lines will be neglected.

We will write the $ABCD$ matrix for each transmission-line section, and subsequently obtain the $ABCD$ matrix for the unit cell. Since the input and output signals of the unit cell differs only by a certain complex phase β, we obtain an eigenvalue equation for β, which is

$$\cos\beta d=\cos(\beta_G a)\cos[\beta_{NG}(d-a)]-\tfrac{1}{2}(\beta_{NG}/\beta_G+\beta_G/G_{NG})\cdot\sin(\beta_G a)\sin[\beta_{NG}(d-a)] \quad \text{(A1)}$$

where β_G and β_{NG} are the phase constants in the grooved and nongrooved sections of the IS waveguide. In the stopband $\beta d=\pi-j\alpha d$, where α is the attenuation constant caused by the mode coupling between space harmonics. The center frequency f_c is where α becomes a maximum.

The phase constants β_{NG} and β_G can be obtained by solving the eigenvalue problems associated with the IS waveguides with or without a gap between the guiding layer (ϵ_2) and the dielectric strip (ϵ_1). The details of the derivation of β_{NG} are given in [2]. To obtain β_G, we need to modify the method in [2] so that the air region between the guiding layer and the dielectric strip is taken appropriately into account. We need to modify eq. (3) in [2] and write an expression for the air-gap region. Obviously, the effective dielectric constant in such a region is different from the no-air-gap portion and, hence, eqs. (4) and (5) in [2] should be modified. However, all the remaining procedures in [2] remain unchanged for derivation of β_G.

The center frequency f_c can be computed in the following manner. First for a given set of structural and material parameters, we compute β_{NG} and β_G versus frequencies. Then (A1) is solved for β, and we obtain f_c at which α becomes maximum.

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Low Cost *X*-Band MIC BARITT Doppler Sensor

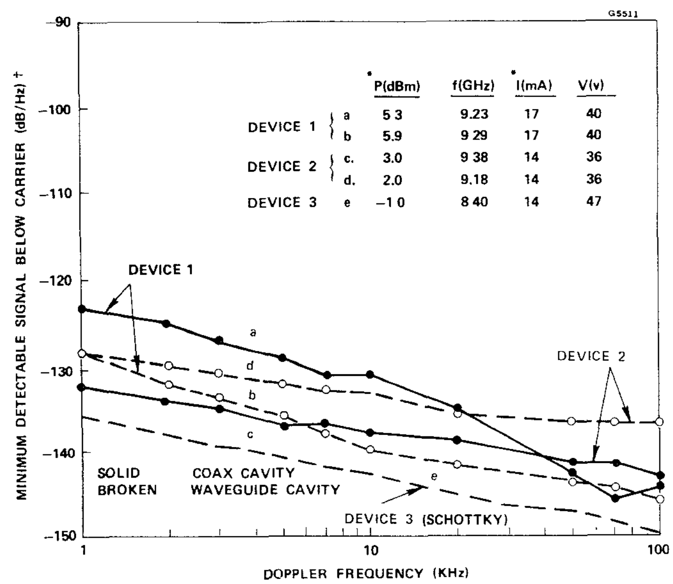
SIANG PING KWOK, MEMBER, IEEE, AND KENNETH P. WELLER, MEMBER, IEEE

***Abstract*—An MIC *X*-band BARITT self-mixing oscillator has been developed. A minimum detectable signal below carrier of −139 dB/Hz at 100 kHz away from carrier was achieved at 1-mW signal carrier. A low cost, compact and sensitive hybrid MIC Doppler sensor module was constructed, incorporating the BARITT MIC circuit and a microstrip antenna.**

I. Introduction

The superior self-mixing property of the BARITT device offers potential advantages for Doppler sensor applications over IMPATT and Gunn devices [1]–[3]. However, a major factor inhibiting the widespread use of microwave Doppler sensors is the cost of reliable microwave components. Microwave integrated circuit (MIC) technology using microstrip offers advantages for miniaturization, reliability, and low cost.

This paper describes the design of the BARITT device, the MIC oscillator circuit, the sensitivities of the self-oscillating mixer, and a hybrid MIC *X*-band Doppler sensor module, incorporating the BARITT MIC circuit and a microstrip antenna.



*P, I for maximum MDSBC.
† for S+N/N = 3 dB.

Fig. 1. BARITT self-mixing sensitivity.

II. Device Design and Self-Mixing Sensitivity

There are two figures of merit that can be applied to self-oscillating detectors. One is the minimum detectable signal (MDS) level which is related directly to the noise figure of the detector. For a self-mixing oscillator, the output power may be relatively low for the best MDS. The second figure of merit is the minimum detectable signal below carrier (MDSBC). It is directly related to the maximum range capability of the system, since it is the ratio of transmitted power to minimum detectable received signal. Experimentally, it was found that an excessively large oscillator output power may actually reduce the MDSBC ratio, as shown in Fig. 1. It is suggested that the oscillator noise and conversion loss are the dominant factors. Thus the BARITT device design for this application was aimed primarily at maximizing the oscillator MDSBC and not its power output.

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S. P. Kwok is with the Torrance Research Center, Hughes Aircraft Company, Torrance, CA 90509.

K. P. Weller was with the Torrance Research Center, Hughes Aircraft Company, Torrance, CA. He is now with TRW Inc., Redondo Beach, CA 90278.

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