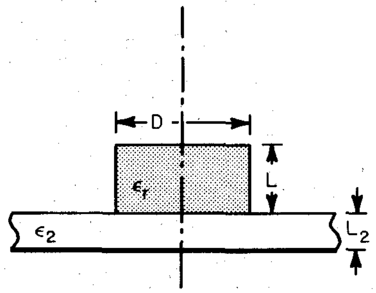

Fig. 2. Dielectric resonator mounted on a substrate.

(15), Q_c is then found from

$$Q_c = \frac{f_0}{\frac{\Delta f_0}{\Delta L_2} \cdot \delta}. \qquad (16)$$

In the example to be computed, the resonator is specified by $\epsilon_r = 38$, $D = 10.5$ mm, $L = 4.6$ mm, and the dielectric substrate is specified by $\epsilon_2 = 2.5$ and $L_2 = 0.762$ mm. Suppose that the unloaded Q, given by the manufacturer, is $Q_0 = 5000$. This Q-factor takes into account the dielectric losses of the resonator material only.

As a first step in the computation, the resonant frequency is evaluated from (13) by iteration, the result being $f_0 = 5.483$ GHz. Actually, the accuracy of method [8] is only about 2 percent, but we have to keep a sufficient number of digits in order to evaluate the derivatives from finite differences. If frequency is now increased by 0.1 percent, the increment in L is found to be $\Delta L = 0.01254$ mm. If L_2 is next increased for 0.1 percent, the corresponding increment is $\Delta L = .5925$ μm. Thus, the required derivative computed by (15) is

$$\frac{\Delta f_0}{\Delta L_2} = -0.3429 \text{ GHz/mm}. \qquad (17)$$

The sign is negative, because frequency decreases when L_2 is increased. To apply (2), the length L_2 has to be shortened by δ, i.e., increased by $-\delta$ so that the computed Q-factor comes out to be positive. From (1), the skin depth for a copper conductor is found to be $\delta = 0.8913$ μm, which then gives $Q_c = 17938$. Therefore, due to the contribution of conductor losses in the ground plane, the overall unloaded Q of the resonator in Fig. 2 will drop from 5000 to $(17938^{-1} + 5000^{-1})^{-1} = 3910$.

V. Corollaries

i) The derivation of (2) is valid only if the stored electric energy is stationary when the walls are receded for distance δ. Therefore, the field distribution must be of such a nature that the electric field normal to the shielding walls is zero. None of the HEM modes possess this property.

ii) The metal enclosure must possess the rotational symmetry. A box shaped as a parallelepiped would create normal components of the electric field on the walls, thus violating the assumption $\Delta W_e = 0$.

iii) Equation (2) is not an approximation, but it represents an exact relationship, as long as the assumptions utilized in the proof remain valid.

iv) The presence of the dielectric core inside the resonator is not required for the application of the incremental frequency rule. This may be seen by deriving an analytical expression for the Q-factor of a TE_{omp} hollow cylindrical resonator by evaluating the differential Δf in (2), instead of performing the conventional integration of the dissipated power along the walls. The result is identical with the published Q-factor value for hollow cylindrical resonator [9].

v) Since the quantities appearing in (2) can be observed by an experiment, the incremental frequency rule may be used in measurements.

vi) A partial Q-factor due to only one of the metal walls may be studied by receding only that wall and leaving the other walls intact. As an example, it is possible to analyze the effect of the metal tuning plunger on the Q-factor of the cavity, if the plunger is inserted into a cavity in such a way that the rotational symmetry is preserved.

Acknowledgment

I wish to thank Dr. A. Glisson and P. Wheless for their critical remarks.

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Loss Measurements of Nonradiative Dielectric Waveguide

TSUKASA YONEYAMA, MEMBER, IEEE, NORIO TOZAWA, AND SHIGEO NISHIDA, SENIOR MEMBER, IEEE

***Abstract*—A technique has been developed for precisely measuring the attenuation constant of the nonradiative dielectric waveguide (NRD-guide) at 50 GHz. The novelty of the present technique lies in incorporating the NRD-guide directional coupler into the measurement system and taking advantage of the total reflection of waves at the truncated end of the dielectric strip to facilitate the construction of the setup and to attain a high degree of accuracy in measurements. Measured attenuation constants were found to be about 13 dB/m for a polystyrene NRD-guide and 4 dB/m for a Teflon NRD-guide. These values indicate that the NRD-guide can be of practical use as a waveguide for millimeter-wave integrated circuits because of its low-loss nature as well as its radiation suppression capability. Calculation is also carried out in order to support measurements.**

Manuscript received December 5, 1983; revised March 12, 1984.
T. Yoneyama and S. Nishida are with the Research Institute of Electrical Communication, Tohoku University, Katahiri 2-1-1, Sendai, 980 Japan.
N. Tozawa is with Fijitsu, Ltd., Kawasaki, Japan.

0018-9480/84/0800-0943$01.00 ©1984 IEEE

I. Introduction

Although it resembles the H-guide both in structure and in field configuration, the nonradiative dielectric waveguide (NRD-guide) can be used to great advantage for millimeter-wave integrated circuits, since it almost completely suppresses radiation at curved sections and discontinuities due to the cutoff nature of the metal plates separated by a distance less than half a wavelength [1]. The versatility of the NRD-guide has been demonstrated by excellent performance of bends [2], directional couplers [3], and some other circuit components, such as ring resonators and filters [4]. Transmission loss of the NRD-guide has also been calculated and shown to be reasonably small, provided that a low-loss dielectric material is used [1]. This is to be expected because the conductor loss is not so large in the NRD-guide, in spite of the narrow plate separation due to the electric field being predominantly parallel to the metal plates as in the low-loss circular H_{01} waveguide. Though theoretical examination is promising, it has to be verified by measurements.

In this paper, the attenuation constant of the NRD-guide is measured at 50 GHz by means of a newly developed technique. The novelty of the present technique lies in incorporating the NRD-guide directional coupler into the measurement system and taking advantage of total reflection of waves at the truncated end of a dielectric strip. The NRD-guide coupler is especially useful for minimizing error in measurements, and the total reflection at the end of the strip serves to make the setup simple and compact by allowing effective path length of transmission to be twice as long as the actual strip length.

Measured attenuation constants were found to be about 13 dB/m for a polystyrene NRD-guide and about 4 dB/m for a Teflon NRD-guide. The latter is particularly small and is a direct consequence of the small loss factor of Teflon, which has $\tan\delta = 1.5\times10^{-4}$ at 50 GHz. Thus, the NRD-guide is proved to be superior, even in loss characteristics.

II. Measurement System

Fig. 1 is a sketch of the setup built for the present measurements. The top and bottom plates of the NRD-guide, 45 cm in length and 10 cm in width, were silver-plated with a thickness of 2 μm to reduce conductor loss. Dielectric strips can be held in place securely by the mechanical pressure between the two conducting plates without the use of any adhesive material. Polystyrene ($\epsilon_r = 2.56$) and Teflon ($\epsilon_r = 2.04$) were chosen as strip materials because of their own practical merits. Polystyrene is easier to work while Teflon is less lossy. Their loss factors at 50 GHz were found to be $\tan\delta = 9\times10^{-4}$ for polystyrene and to be $\log\delta = 1.5\times10^{-4}$ for Teflon by preliminary Q-measurements of disk resonators made of the respective dielectric materials. The polystyrene strip was 2.7 mm in height and 2.4 mm in width, and the Teflon strip was 2.85 mm in height and 3.2 mm in width. The large dimensions of the Teflon strip are favorable for reduction of transmission loss. The dielectric strips were cut down gradually to change the length from 40 to 3 cm. Electromagnetic waves come from the metal waveguide, enter the NRD-guide by way of the transition horn, and return by being reflected at the truncated end of the strip. In order to confirm this statement, the standing-wave pattern along the Teflon strip was measured by using a 1.5-mm-long unipole antenna consisting of an inner conductor of a very thin semirigid cable whose outer surface was coated with lossy paint to reject interferences. One of the results is shown in Fig. 2 in which the strip occupies the negative region of the z-axis and is truncated at $z = 0$. Standing-wave ratio (SWR) is more than 35 dB, which demonstrates that total reflection of waves actually takes place at the truncated end of the strip. The field decays in an exponential manner beyond $z = 0$ as expected.

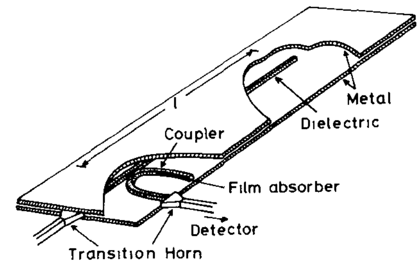

Fig 1 Setup built for loss measurements of NRD-guide. Metal plates are 45 cm in length and 10 cm in width.

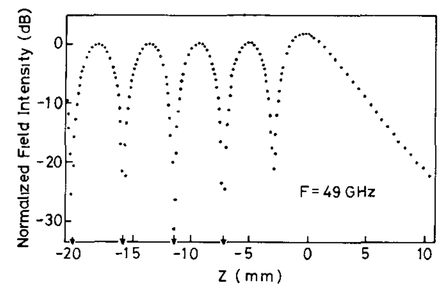

Fig. 2. Standing-wave pattern measured along a Teflon strip in NRD-guide at 49 GHz.

A 180° dielectric bend, 20 mm in curvature radius, was located in close proximity to the main strip so as to work as a directional coupler for monitoring field intensities of incident and reflected waves. One of its ends was terminated with a film absorber [4], and the other was tapered and led to a metal waveguide detector device by way of the transition horn. Attenuation was measured by turning the coupler forwards and backwards, and taking differences of the respective detector readings. The relative position of the dielectric guide and the coupler was maintained by using a spacer which was removed after the top plate was added. Such a process was repeated for different strip lengths and for different frequencies.

Actually, attenuation data thus obtained include the coupling effect as expressed by

$$2\alpha l - 10\log(1-K^2) \qquad (\text{dB}) \tag{1}$$

where α is the attenuation constant of the NRD-guide under measurement, l is length of the dielectric strip, and K is the amplitude coupling factor between the main strip and the coupler. Therefore, if measured values are plotted against round trip path length $2l$, they are located nearly in a straight line and the slope of the line determines the real attenuation constant of the NRD-guide, provided that the coupling factor is kept constant during the measurement.

To achieve high accuracy in measurements, the coupling factor has to be carefully optimized. If the coupling factor is too small, most power remains in the main guide, making a round trip between the truncated end of the strip and the input transition horn. This gives rise to frequency-dependent fluctuations in measurements and causes considerable deterioration in the accuracy of measurement. On the other hand, if the coupling factor is close to 0 dB, most power goes into the coupled arm and is subjected to multiple reflection between the input transition horn and the detector apparatus, including the transition horn and the detector itself. This also causes serious error in measurements. The directivity of the coupler may be another source of error, though it has

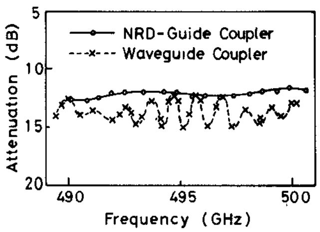

Fig. 3. Attenuation of polystyrene NRD-guide measured by using an NRD-guide coupler (solid curve) and a metal waveguide coupler (dotted curve) as a function of frequency.

been found to be better than 25 dB, even at a coupling of 0 dB [3]. Taking these facts into account, a coupling factor of about 5 dB was chosen.

A metal waveguide coupler might be considered to be more useful than the NRD-guide coupler. Indeed, at the very early stage of the research, a metal waveguide coupler was located preceding the transition horn. Such a setup, however, was found to lack accuracy. This is because a portion of the incident wave is reflected back toward the coupler by the transition horn and is mixed with the real signal returning from the far end of the dielectric strip. Error is serious, since the signal suffers a considerable amount of attenuation in a round trip propagation and is weak in intensity when it meets the spurious signal. An advantage of employing the NRD-coupler is that the spurious signal caused by multiple reflection, if any, is mixed with the incident wave which is always at a higher level of intensity; hence, the spurious signal is negligible unless the coupling factor of the coupler is too small or too large, as mentioned above.

An additional improvement can be attained if NiCr resistive sheets [4] are attached to both free surfaces of the dielectric strip at the midpoint between the input transition horn and the coupler. Because the real incident wave passes through the resistive sheets only once, while the multiply reflected signal passes through them three times at least, the latter is much more decayed than the former. Therefore, these resistive sheets work as a buffer attenuator for further suppressing undesirable interferences. The buffer of about 8-dB attenuation was found to be suitable for the present purpose.

For reference, Fig. 3 shows two typical attenuation curves as a function of frequency, one measured by using the NRD-guide coupler and the other measured by using the metal waveguide coupler. The dielectric strip was polystyrene and 42 cm in length. Strictly speaking, the length of the strip should be considered to be slightly shorter than this value for the NRD-guide system, depending on the position of the coupler. However, length is not a problem in this case. What is important is that the use of the NRD-guide coupler is capable of smoothing ripples of the attenuation curve to an acceptable level. Thus, the superiority of the NRD-guide system is definitely demonstrated.

III. Results and Discussions

The NRD-guide system was used to measure loss characteristics at 50 GHz. Measurements were repeated many times for different frequencies and for different lengths of the dielectric strip. The length of the strip should be considered to be the distance between the top of the coupler and the truncated end of the strip. Data for 50 GHz are presented in Fig. 4. The method of least squares is applied to determine the straight line which best fits the data. The slope of each line gives an attenuation constant for the corresponding dielectric material. Attenuation constants thus obtained are plotted as a function of frequency in Fig. 5, together with theoretical curves which will be explained later.

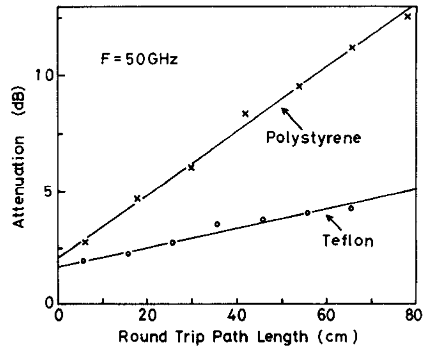

Fig. 4. Attenuation of polystyrene and Teflon NRD-guides measured at 50 GHz as a function of round-trip path length.

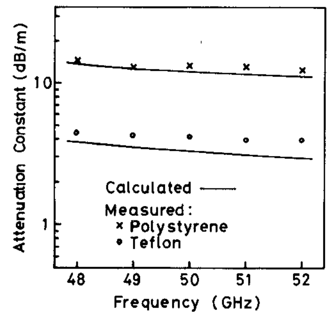

Fig. 5. Calculated and measured attenuation constants of polystyrene and Teflon NRD-guides as a function of frequency.

Attenuation constants of the polystyrene and Teflon NRD-guides were found to be about 13 and 4 dB/m, respectively. The major contribution to the low-loss nature of the Teflon NRD-guide is the small loss factor of Teflon, whose $\tan\delta = 1.5\times10^{-4}$ at 50 GHz. On the contrary, the loss factor of polystyrene is 9×10^{-4} at the same frequency. Therefore, it is clear that using a low-loss dielectric is a key requirement for reducing transmission loss of the NRD-guide. In this respect, it is interesting to note that the attenuation constant of a microstrip line has been reported to be 57 dB/m at 50 GHz [5]. The attenuation constants slightly decrease as frequency increases, as can be seen in Fig. 5. This is because the dielectric loss is almost constant, while the conductor loss decreases with the increasing frequency in the NRD-guide due to the fact that the electric field is predominantly parallel to the metal plates, just as in the low-loss circular H_{01} waveguide.

It may be valuable to compare measured data with theory. By applying the same perturbation technique as that for the H-guide [6], the attenuation constant α of the NRD-guide can be derived in the form

$$\alpha = \alpha_c + \alpha_d \tag{2}$$

where α_c and α_d are conductor and dielectric losses, respectively, and are given by

$$\alpha_c = \frac{2\pi^2 r_s}{\left(k_0^2+p^2\right)a^2}\,\frac{\epsilon_r k_0}{\beta a}\,\frac{pb\left(q^2+\epsilon_r^2p^2\right)+2\left(q^2+\epsilon_r p^2\right)}{pb\left(q^2+\epsilon_r^2p^2\right)+2\epsilon_r(\epsilon_r-1)k_0^2} \tag{3a}$$

$$\alpha_d = \epsilon_r \tan\delta\,\frac{p}{2\beta}\,\frac{k_0^2 b\left(q^2+\epsilon_r^2p^2\right)+2p\left(\epsilon_r k_0^2-2q^2\right)}{pb\left(q^2+\epsilon_r^2p^2\right)+2\epsilon_r(\epsilon_r-1)k_0^2} \tag{3b}$$

where

$$\beta = \sqrt{\epsilon_r k_0^2 - q^2 - (\pi/a)^2} \quad (4a)$$

$$r_s = \sqrt{\frac{\omega\mu_0}{2\sigma}} \Bigg/ \sqrt{\frac{\mu_0}{\epsilon_0}} = \sqrt{\frac{\omega\epsilon_0}{2\sigma}}. \quad (4b)$$

In the above equations, ϵ_r and $\tan\delta$ are the relative dielectric constant and loss factor of the dielectric strip, ϵ_0 and k_0 are the free space permittivity and wavenumber, β is the propagation constant of the dominant mode in the NRD-guide, σ and r_s are the conductivity and normalized surface resistance of the metal plates, a and b are the height and width of the dielectric strip, and the parameters p and q are the lowest eigenvalues of the following characteristic equations for TM surface waves supported by a dielectric slab with a thickness of b:

$$\epsilon_r p = q \tan\left(\frac{qb}{2}\right) \quad (5a)$$

$$p^2 + q^2 = (\epsilon_r - 1) k_0^2. \quad (5b)$$

Assuming the top and bottom plates to be silver-plated ($\sigma = 6.17 \times 10^7$ S/m), the attenuation constant is calculated for polystyrene ($\epsilon_r = 2.56$, $\tan\delta = 9 \times 10^{-4}$) and Teflon ($\epsilon_r = 2.04$, $\tan\delta = 1.5 \times 10^{-4}$) and compared with measured data in Fig. 5. Agreement between theory and measurements is quite satisfactory.

IV. Conclusions

A technique for measuring the attenuation constant of the NRD-guide was developed and applied to polystyrene and Teflon NRD-guides at 50 GHz. Measured attenuation constants are about 13 dB/m for a polystyrene NRD-guide and 4 dB/m for a Teflon NRD-guide. These results clearly show that the NRD-guide can be of practical use as a transmission line for millimeter-wave integrated circuits because of its low-loss nature, as well as its radiation suppression capability. The calculated attenuation constant is also found to be in a good agreement with measurements.

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Comments on "The Microstrip Open-Ring Resonator"

KARL B. NICLAS, SENIOR MEMBER, IEEE

Abstract—**Attention is drawn to the analogy existing between the electromagnetic fields of the open-ring resonator and the ring-sector waveguide [1], [2]. Except for the eigenvalues of γ and k, the spatial variations of their fields in the cross-sectional plane are solutions of the same differential equation [3].**

Manuscript received February 17, 1984.

The author is with Watkins-Johnson Company, 3333 Hillview Ave., Palo Alto, CA 94304.

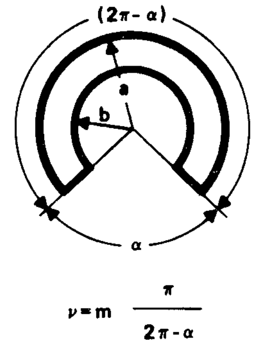

Fig. 1. Cross-sectional geometry of ring-sector waveguide and open-ring resonator.

In a recent paper[1], the open-ring microstrip resonator was analyzed by means of the two-dimensional magnetic wall model. When taking a look at the expressions, the authors derived for the components of the electromagnetic fields, one detects a striking analogy to those of the ring sector or coaxial sector waveguide which were published in 1961 [2]. The differences impacting the solutions for the field components of the coaxial sector waveguide and the open-ring resonator are the conducting walls versus the magnetic walls and the three-dimensional versus the two-dimensional geometry. As a result of the first (conducting versus magnetic walls), the magnetic-field components of the ring-sector waveguide as functions of the transverse coordinates r and ϕ are similar to the electric-field components fo the open-ring resonator. The same similarity exists between the electric-field components of the ring-sector waveguide and the magnetic-field components of the open-ring resonator. However, the second deviation (three-dimensional versus two-dimensional geometry) causes the transverse electric fields of the open-ring resonator not to be excited. The analogy between the fields of the two structures having the cross-sectional geometry of Fig. 1 is easily discernible from the proportionalities of the corresponding fields summarized in Table I. Here, the wavenumber k is related to the eigenvalue γ in accordance with $\gamma^2 = k^2 - \beta^2$, where $j\beta$ is the propagation constant of the waveguide.

As early as 1955, P. R. Clement and W. C. Johnson pointed out the analogy that exists between the uniform waveguide and its two-dimensional model [3]. As to the solutions of the differential equations which are found to be common to both structures, one could interpret the two-dimensional problem as a special case of the uniform waveguide problem, distinguished from each other only by a different set of boundary conditions. Since field patterns of uniform waveguides have been reported for many cross-sectional shapes, a familiarization with the work reported in [3] may be highly beneficial for the understanding of microstrip structures with corresponding geometries. A rigorous analysis of the coaxial sector waveguide, including field distributions and cutoff frequencies of the TE_{mn}- and TM_{mn}-modes, as well as an approximation for the characteristic impedance of the TE_{11} mode, is contained in [2]. As open-ring resonators have been proposed for filters and planar antennas, coaxial waveguides have been employed as coupling elements for traveling-wave amplifiers [4],

[1] I. Wolff and V K. Tripathi, *IEEE Trans Microwave Theory Tech.*, vol. MTT-32, pp. 102–107, Jan 1984