Powder Core Dielectric Channel Waveguide

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***Abstract*— A powder-filled rectangular groove in the surface of a plastic substrate has been demonstrated as a dielectric waveguide at 94 GHz. Propagation losses as low as 0.09 dB/cm were measured by direct transmission with nickel-aluminum titanate powder in a polypropylene substrate and with barium tetratitanate powder in a polytetrafluoroethylene (PTFE) substrate; values as low as 0.06 dB/cm were deduced from ring resonator measurements. Guide wavelengths measured for various combinations of guide dimensions, powders, and substrates agree within 10% with values predicted by the approximate theory of Marcatili for the E^y_{11} mode. Effective loss tangents for the powders at 94 GHz were calculated from waveguide attenuation measurements, using Marcatili's field solutions. Ring resonators fabricated by filling a groove in a polypropylene substrate with nickel-aluminum titanate powder exhibit Q's as high as 2400 at 94 GHz in an 8 cm diameter ring. Coupling to the resonators was achieved with adjacent straight powder core channel guides as directional couplers.**

I. Introduction

Dielectric waveguides have been widely studied for use in millimeter wave integrated circuits [1] since they generally exhibit much lower millimeter wave attenuation than structures which rely on metal surfaces for guiding, such as microstrip, slotline, and coplanar waveguide. (Hollow metal guide is low-loss, but not suitable for integrated circuits.) In addition, dielectric waveguides are particularly well-suited for coupled waveguide structures, because the evanescent fields of the guided mode extend outside the core.

We have demonstrated a powder-filled groove in the surface of a plastic substrate as a low-loss dielectric channel waveguide at 94 GHz. Directional couplers and ring resonators with Q's as high as 2400 at 94 GHz have been implemented with this waveguide. Such powder core channel guides may be attractive for low cost passive integrated circuits such as feed structures for antennas.

In the first half of this paper, we briefly survey theories applicable to rectangular dielectric channel waveguide, describe our powder core channel waveguides, and discuss the measurements used to characterize the performance of straight segments. The second half of this paper treats the design and testing of ring resonators, and the design of the straight guides used to couple to the resonators. In the appendix, we derive an equation relating the attenuation of the E^y_{pq} modes of rectangular channel waveguide to the loss tangents of the core and its surroundings. This equation is then used to calculate 94 GHz effective loss tangents for powders used in our experiments.

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II. Powder Core Channel Waveguide

A. Motivation

The cost of conventional millimeter waveguide and waveguide components is substantially higher than that of their longer wavelength counterparts because of their proportionally smaller dimensional tolerances. As a consequence, millimeter wave waveguide components are not mass-produced; rather, they are machined or assembled individually (or electroforming mandrels are machined individually) so that, in a sense, the finished part represents "stored precision machinist's labor." The situation is no different for the usual dielectric waveguide or image guide component machined from teflon or silicon—in fact, the machining task may be even more difficult due to the ductile or brittle nature of dielectric materials. Metallic integrated circuits produced photolithographically on a dielectric substrate is one answer to this dilemma; however, MMIC is not an exact replacement for conventional waveguide components, but is more a complementary technology. One motivation for the present study was to see if millimeter waveguide and waveguide components could be realized (ultimately) by injection molding or embossing plastic substrates, so that the "stored machinist's labor cost" resides in a precision form usable for thousands of components rather than being delivered with every piece. While, for convenience, we have used precision machining in our laboratory demonstrations, we have also demonstrated that waveguide sections at 94 GHz can be formed by thermoplastic means (for example) with negligible change in characteristics [2].

The choice of placing the precision fabrication in the substrate rather than the core is somewhat arbitrary. We could have chosen to mold the waveguide cores from low-loss polymers and then surround them with foam or powdered material of lower dielectric constant (possibly powders of the core material, which automatically have lower dielectric constants). Our choice of a solid, precision-formed substrate filled with a powder exhibiting a higher dielectric constant was actually chosen as a continuation of our studies on flexible dielectric millimeter waveguide using powdered cores [3].

B. Survey of Theoretical Approaches

The rectangular dielectric channel waveguide is shown in cross section in Fig. 1. No analytical solution exists for the propagating modes of this guide. However, there are several

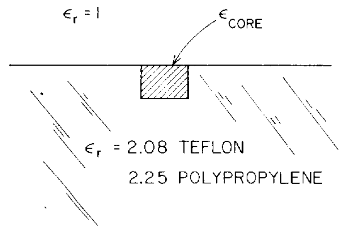

Fig. 1. Cross section of rectangular dielectric channel waveguide.

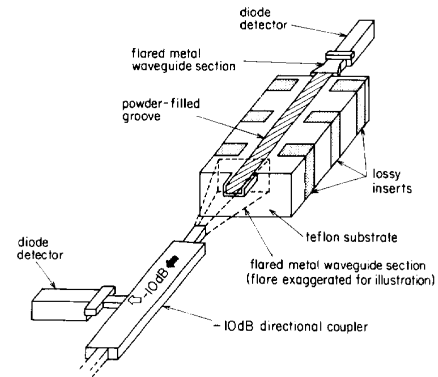

Fig. 2. Set-up for measuring attenuation and guide wavelength.

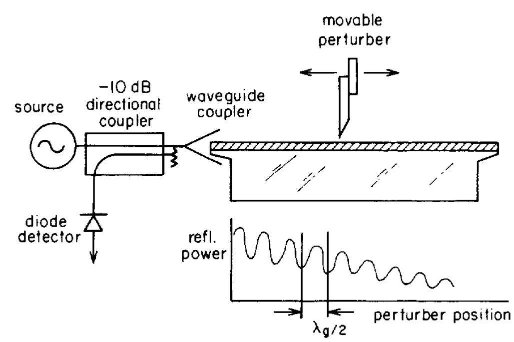

Fig. 3. Guide wavelength measurement.

numerical methods which yield approximate values of the propagation constants. Most of these techniques, such as those proposed by Yeh et al. (finite elements) in [4], [5] and (finite differences) [6], require the use of large computers, Marcatili's approximate-mode method [7], on the other hand, is fairly easy to use since the computations involved are much simpler. Marcatili's method makes an approximation based on the criterion that the refractive index of the core is similar to that of the cladding. That is, $(1-n_{\text{clad}}/n_{\text{core}} \ll 1)$. However, selected cases analyzed with other methods agree reasonably well with guide wavelengths calculated from Marcatili's theory for the E_{11}^y mode [6], [7] even when the criterion $(1-n_{\text{clad}}/n_{\text{core}} \ll 1)$ is invalid.[1] Indeed, good agreement is obtained even for a rod with relative dielectric constant equal to 13 surrounded by air [6]. For these reasons, Marcatili's method was used here to predict the propagation constant of the E_{11}^y mode of powder core channel waveguide.

C. Guide Wavelength and Attenuation Measurements

A rectangular groove was milled into the surface of a low-loss (PTFE or polypropylene) substrate and was filled with a powder having a high-dielectric constant to form the core of a dielectric waveguide (Fig. 1). With this configuration, the powder could be packed from the top to assure a sufficiently uniform density along the length of the groove. Rectangular grooves with cross-sectional dimensions varying less than ±0.001 inches or ±0.025 mm from the specified values (typically 1 mm × 1 mm) could be milled with relative ease. This degree of dimensional accuracy was found to be sufficient at 94 GHz to produce a guide wavelength uniform within our measurement accuracy.

The guide wavelength and loss per unit length were measured for the fundamental vertically polarized (E_{11}^y) mode of various powder core channel waveguides using the set-up shown in Fig. 2. On each end of the substrate the dielectric-filled groove was extended with a thin-walled trough of substrate material. This trough fitted snugly into the end of a slightly flared section of WR-10 metal waveguide to couple to the dielectric guide. Lossy inserts made from Emerson and Cumming MF-110 absorber were placed at non-periodic intervals in the substrate 3 mm from the groove to attenuate any substrate modes (that is, propagation of energy through the substrate other than via the desired guided mode) that might have been excited at the coupling point and resulted in end-to-end coupling via the substrate.

To measure the guide wavelength, a knife-edged metal shorting plane was held mechanically just above the surface of the powder, as shown schematically in Fig. 3. The perturber reflects a small fraction of the power traveling along the waveguide toward the feed, where it interferes with the reflection from the input coupler. The amplitude of this interference changes as the relative phase between these two signals changes. Thus, as the perturber was moved along the length of the groove, a sequence of maxima and minima in reflected power was sensed with a −10 dB directional coupler and a Schottky diode, as shown in Fig. 3. The guide wavelength is twice the distance the perturber is moved between successive minima.

For various combinations of guide dimensions, dielectric powders, and substrate materials, the guide wavelengths were compared to the values predicted by Marcatili's approximate theory [7] for the fundamental vertically polarized mode. (No beats were observed in the measured patterns of reflected power versus perturber position, indicating that the waveguides were single-mode, as intended. Also, the polarization of the guided mode matched the vertical polarization of the excitation, as expected.)

In order to use Marcatili's theory, the dielectric constants of the powders were needed. The density of the powder in

[1] We use Marcatili's notation here, despite the difficulty that the magnetic field associated with the "E_{pq}^y" mode is mainly x-directed and therefore cannot be designated "H_{pq}^y" without confusion. Marcatili solves this problem by rarely referring to the magnetic field. See also [6] for an alternative way of naming modes.

TABLE I
COMPARISON OF MEASURED GUIDE WAVELENGTH WITH PREDICTION OF MARACATILI

Powder Type	Substrate Type	Width of Groove (mm)	Depth of Groove (mm)	Powder Density (g/cm^3)	Relative Dielectric constant	λg (Meas.) (mm) $\pm .02$	λg (Marcatili) (mm)
D-30**	T	0.94	0.94	$1.95 \pm .07$	$5.78 \pm .35$	1.86	$1.96 \pm .08$***
D-38**	T	1.12	1.12	$1.77 \pm .04$	$5.0 \pm .4$	2.06	$1.9 \pm .1$
MCT40*	T	1.83	1.04	$1.33 \pm .02$	$3.77 \pm .07$	2.07	$2.11 \pm .02$
D-8512*	T	1.27	1.10	$1.55 \pm .03$	$3.85 \pm .07$	2.16	$2.18 \pm .02$
D-8512*	T	1.27	1.10	$1.47 \pm .02$	$3.55 \pm .22$	2.22	$2.26 \pm .06$
D-8512*	P	1.47	1.10	$1.55 \pm .02$	$3.85 \pm .08$	2.04	$2.09 \pm .02$

Substrate T is PTFE
Substrate P is polypropylene

*** The uncertainty in the guide wavelength predicted by Marcatili's theory is estimated from the uncertainty in the dielectric constant of the powder.
** 70% of particles between 100μm and 43μm, 30% less than 43μm
* 100% of particles less than 43μm

the groove was determined by precision weight measurement, and previously-measured curves of dielectric constant versus density were used to find the effective dielectric constant of the powder packed into the groove. The dielectric constants of the powders had been measured at 10 GHz using the shorted-waveguide technique. These measurements were made at 10 GHz because of the difficulty of controlling the length of a powder sample in a shorted WR-10 metallic waveguide sufficiently accurately to measure its dielectric constant at 94 GHz. The effective dielectric constant of a powder composed of low-loss dielectric material should not vary much between 10 GHz and 94 GHz if the powder grains are small relative to wavelength at 94 GHz. (For further discussion, see [2].)

The powders used were the same as for the flexible guide work described in [3]: Trans-Tech D-30 nickel-aluminum titanate, Trans-Tech D-38 barium tetratitanate, Trans-Tech MCT-40 magnesium calcium titanate, and Trans-Tech D-8512, an "improved" barium tetratitanate. For D-8512, MCT-40, and for one batch of D-30, all particles were less than 43 μm in size. For D-38 and for a second batch of D-30, 70% of the particles were between 100 μm and 43 μm and 30% were less than 43 μm. Trans-Tech gives $\epsilon' = 31$ and $\tan\delta < .0002$ for solid D-30 at 10 GHz, $\epsilon' = 37$ and $\tan\delta < .0005$ for solid D-38 at 6 GHz, $\epsilon' = 40$ and $\tan\delta < .002$ for solid MCT-40 at 6 GHz, and $\epsilon' = 38.6$ and $\tan\delta < .0005$ for solid D-8512 at 6 GHz; no data at 94 GHz is known, other than that in this report.

To determine the loss-per-unit length of a channel waveguide, the power transmitted from end-to-end was measured by a detector connected to the flared section of metal waveguide surrounding the trough on the far end of the substrate (Fig. 2). Shorted stub ("E/H") tuners were added to match the coupling sections. The power detected at the far end could not be significantly increased by adding the E/H tuners, so we assume that the couplers are reasonably well matched by themselves. In addition, removing the lossy substrate inserts did not affect the power received at the far end, indicating that little power is lost to substrate modes. A third detector connected to a small horn antenna was used as a movable probe to determine that the power radiated from the couplers and guide was small. Taken together, these observations indicate that almost all of the incident power was coupled into the dielectric waveguide, so that the difference between the incident power and the power detected at the far end mostly represents dielectric waveguide loss. The loss per unit length, including a small error due to coupling loss, is then this loss divided by the length of the dielectric waveguide.

A comparison between the measured values of the guide wavelength with those predicted for the E_{11}^y mode by Marcatili's approximate theory is given in Table I for various powders in plastic substrates at 94 GHz. Although Marcatili's theory assumes that the refractive index of the core is similar to those of the surrounding media and is thus only an approximation, the wavelengths it predicts for the E_{11}^y mode (Table I) are in reasonable agreement with those measured. Hence, we conclude that the dielectric constants of the powders used were not too high for Marcatili's theory to be useful in predicting guide wavelength.

A selection of typical measured values of loss per unit length for straight powder-filled grooves in a plastic substrate are given in Table II. The measurements were made at 94 GHz. For given substrate material and channel dimensions, the

TABLE II
ATTENUATION OF POWDER CORE CHANNEL WAVEGUIDES

Powder Type	Substrate Type	Width of Groove (mm)	Depth of Groove (mm)	Density of Powder (g/cm³)	Loss (dB/cm)
D-30**	P	1.17	1.13	$1.75 \pm .03$	$0.34 \pm .01$
MCT 40*	T	1.50	1.05	$1.26 \pm .03$	$0.14 \pm .01$
D-8512*	T	1.27	1.10	$1.47 \pm .02$	$0.09 \pm .01$
D-30*	P	1.17	1.13	$1.80 \pm .03$	$0.17 \pm .01$
D-30*	P	1.17	1.13	$1.68 \pm .03$	$0.09 \pm .01$

** 70% of particles between 100μm and 43μm, 30% less than 43μm
* 100% of particles less than 43μm

Substrate T is PTFE, $\tan\delta$ = .0002 @ 94 GHz (Ref. 15)
Substrate P is polypropylene, $\tan\delta$ = .0002 @ 94 GHz (Ref. 16)

transmission loss increased with powder density. This variation is consistent with a closer confinement of the mode energy to the core region, which has higher material loss, and with increasing powder loss tangent.

III. Bending Loss and Ring Resonators

Having demonstrated the realizability of straight waveguide sections with useful low loss and the applicability of an existing simple theory to describe the guide, a logical next step was to determine the bending loss of such guides; straight sections alone would not be useful in making complex circuits. Fortunately, Marcatili had also published a theory for curved waveguides [8], and we wished to see how well it applied to the values to be employed here. For other theories of propagation on curved dielectric waveguides, see [9]–[13].

For guides whose curvature radius is large compared to both wavelength and channel width, Marcatili treats curvature as a perturbation. The field distributions of his modes of curved guide are thus only slightly different from those of straight guide. Each mode of the curved guide is given the name of the associated mode of straight guide. Like Marcatili's theory for straight channel guides, the theory of curved guides assumes that the differences in refractive index between the core and the surrounding media are small.

For our powders and substrate materials, Marcatili's theory predicts that, at 94 GHz, bending losses would be insignificant compared to our measured absorptive losses in straight guides for radii of curvature greater than about 1 cm. We checked this prediction by measuring the transmission losses of guides machined in 180° circular arcs, coupling in and out exactly as we had with the straight sections. We found that a 1 cm radius of curvature gave a measured loss much higher than that of a straight guide. On the other hand, the measured losses of 180° arc guides with 4 cm and 5 cm radius of curvature were

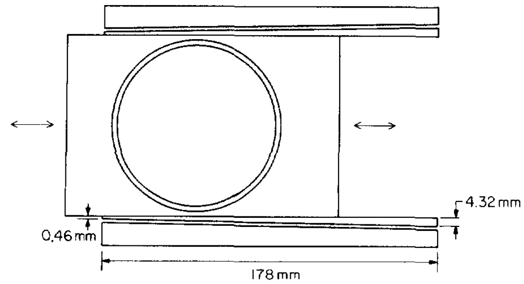

Fig. 4. Coupling scheme used to achieve adjustable distances between a ring resonator and two straight waveguides.

comparable to straight sections. Something seemed to be amiss between experiment and theory.

Coupling to 180° arcs proved more difficult than coupling to the straight sections, particularly for the small radius arcs (1 cm). We were concerned that the transition from the straight "trough" section inserted in the flared metal waveguide to the curved section might produce a radiative loss.

It occurred to us that a neat technique to eliminate this possible source of error was to use 360° arcs, or ring resonators. The guide loss could then be determined from the measured Q of the resonator under very loosely-coupled conditions. We decided to build one ring resonator with a 4 cm radius and another with a 5 cm radius. Coupling to the ring resonators was achieved by placing straight channel guides in proximity to the ring guide. We know of no theory to analyze such a coupling structure; Marcatili treats only the coupling between parallel straight guides [7], predicting a coupling that is approximately an exponentially decreasing function of the spacing. We reason that the coupling for the ring-to-straight guide spacing will also vary exponentially with the guide spacing. Thus, we developed a technique that allowed this spacing to be adjusted easily. Straight channel guides were positioned on opposite sides of the ring to couple power in and out of the ring. The resonator and the guides used for coupling were each built on individual substrates. Material was cut away from two opposing edges of the resonator substrate until each edge was only 0.38 mm from the channel. The substrate of each straight guide had an edge that was cut at an angle to the channel. For both of these substrates, the distance between the channel and the edge was 0.46 mm at one end and 4.32 mm at the other. When the three substrates were placed together as shown in Fig. 4, the separations between the resonator and the coupling guides could be adjusted from about 0.9 mm to 4.7 mm by sliding the substrates with respect to one another.

The substrate material was chosen to be polypropylene, since it is easier to machine than PTFE. The powder was chosen to be D-30 nickel-aluminum titanate, with all particles less than 43 μm in size. This powder gave the least lossy straight guides at 94 GHz, by measurement on straight powder core channel guides and on cylindrical dielectric guides made by filling hollow PTFE tubes with powder [2], [3].

Marcatili's theory of straight channel waveguide was used to choose the channel dimensions of the guides used for coupling to the resonator so that only the E^y_{11} mode would propagate,

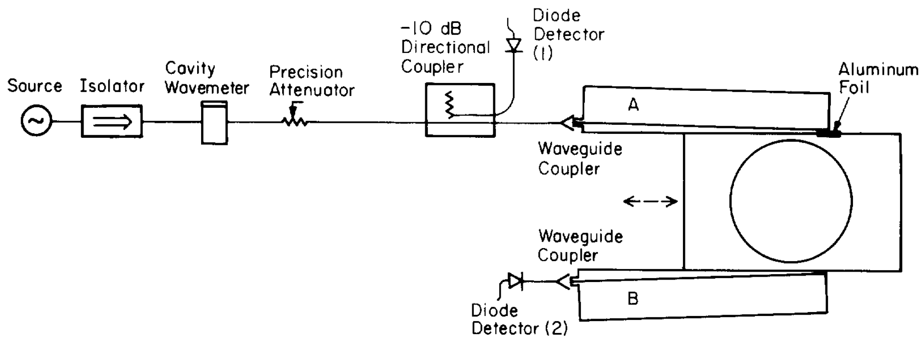

Fig. 5. Set-up for measuring Q's of ring resonators.

and the penetration into the substrate of the evanescent fields would be substantial. Since the horizontal penetration increases as the channel width decreases, a narrow width channel was desired. In Marcatili's analysis, the *depth* of the channel does not affect the horizontal penetration, but the channel depth was limited to less than about 1.1 mm by the inner dimensions of the flared WR-10 metal waveguide sections used for coupling to metal guide (Fig. 5). We picked 1.35 mm for the width and 1.05 mm for the depth because these values satisfied the design criteria for low powder densities, where waveguide propagation loss is lowest.

The goal of the ring resonator design was to choose the optimum channel dimensions given the properties of the materials to be used. Marcatili's theory predicts that bending loss is independent of the channel depth when the plane of curvature is horizontal, so the depth was chosen equal to that of the straight coupling guides in order to enhance the coupling between these guides and the resonator. Marcatili's theory also predicts that the bending loss of the E_{11}^y mode decreases as the channel width increases. Consequently, we chose the width as large as possible (1.83 mm) without allowing propagation of higher-order vertically polarized modes for the range of expected powder dielectric constants.

A. Experiment

The experimental set-up is shown in Fig. 5. For the measurement of Q, the resonator was operated as a transmission filter. Since the loaded Q value of a resonator approaches the unloaded Q in the limit of zero coupling [14], we wanted the coupling between the resonator and the straight guides to be sufficiently weak that the difference between the loaded and unloaded Q's would be less than the other errors in the measurement. This was accomplished by sliding the substrates to decrease the coupling until further decreases yielded no measurable increase in Q. Data were taken with the filter operating with a 40 dB insertion loss, a value more than sufficient to make the loaded Q equal to the unloaded Q to within the experimental uncertainty. The 40 dB insertion loss was obtained by setting the distance between the adjacent walls of the two waveguide channels to about 1.4 mm.

The Q was measured by varying the frequency of the source (mechanically tuning a Varian VRB-2113B23 klystron[2]) and observing the response of the resonator with the detector at (2).[2] The amount of power incident on waveguide A was monitored with the detector at (1) and the precision attenuator was used to keep this power level constant as the frequency was changed. The frequencies of the resonator's peak response and half-power response points were measured with the cavity wavemeter. The Q is the frequency of the peak response divided by the difference in frequency of the half-power points. Since the cavity wavemeter used to measure frequency had a Q comparable to that of the ring resonators, the uncertainties in our measured values of Q are relatively large.

We used several checks to determine that the measured Q was actually that of the ring and not the result of a spurious resonance elsewhere in the measurement set-up. First, with the frequency tuned to a resonance, the placement of a small piece of lossy ferrite over any portion of the powder ring caused a 10 dB drop in received power at (2). Secondly, placing lossy ferritte inserts at various positions elsewhere in or on the substrate of the resonator had no effect on the performance. These observations show that power was propagating through the ring's powder channel and not through the substrate. (The purpose of the aluminum foil tab shown in Fig. 5 was to prevent radiation from the end of waveguide A from entering the resonator's substrate.) Finally, the resonant frequency could be tuned by adjusting the height of a piece of polypropylene positioned above the ring.

B. Conclusions

Using the results of our previous measurements on straight channel waveguide and applying the formula $Q = k_z/2\alpha$ [8], where k_z and α are the propagation and attenuation constants of the straight guide, the predicted ring resonator Q's would be no larger than 1700 if bending losses were neglected. ($.09 \pm .01$ dB/cm loss and $\lambda_g = 2.04 \pm .02$ mm for the E_{11}^y mode give $Q = 1500 \pm 200$.) Thus, we are surprised that some of the actual measured Q values (Table III and Fig. 6) exceed 1700. These results suggest that measurement of the Q of a ring resonator may be a better method for determining waveguide dissipative and scattering losses than end-to-end transmission on a straight guide. The discrepancy in propagation loss values

[2] A 94 GHz sweeper would obviously have been a better choice, but the klystron was the only source available.

TABLE III
MEASURED Q VALUES OF RING RESONATORS

Density of Powder (g/cm^3)	Dielectric Constant	Radius of Curvature (cm)	Frequency (GHz)	Measured Q
1.76 ± .02	4.28 ± .07	4.0	94.39	1100 ± 200
1.86 ± .02	4.65 ± .08	4.0	94.61	1300 ± 200
1.88 ± .02	4.73 ± .09	4.0	94.24	2400 ± 400
1.95 ± .02	5.06 ± .10	4.0	94.61	1600 ± 200
2.10 ● .02	5.90 ± .13	4.0	94.31	1200 ± 200
1.67 ± .02	4.02 ± .06	5.0	94.86	810 ± 100
1.70 ± .02	4.10 ± .06	5.0	93.20	930 ± 150
1.78 ± .02	4.35 ± .07	5.0	94.32	1300 ± 200
1.83 ± .02	4.53 ± .08	5.0	94.28	1600 ± 200
1.88 ± .02	4.73 ± .09	5.0	94.47	1900 ± 200
1.89 ± .02	4.78 ± .09	5.0	94.44	1000 ± 200

Powder: Nickel-aluminum titanate (Trans-Tech D-30).
All particles less than 43 μm.
Channel width: 1.83 mm
Channel depth: 1.09 mm for R = 5.0 cm
1.05 mm for R = 4.0 cm

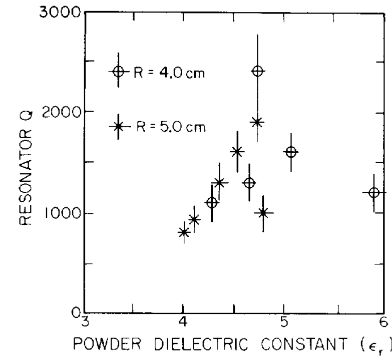

Fig. 6. Measured Q values of ring resonators versus the dielectric constant of the powder (Trans-Tech D-30 nickel aluminum titanate).

between straight guides and ring resonators is likely due to losses incurred in coupling to the straight guides.

We tentatively attribute the initial increase in Q with increasing powder density to reduced bending loss concomitant with increased propagation constant. However, as powder density increases, dielectric loss increases as the fields become more confined to the (relatively) lossy channel and the effective loss tangent of the powder increases. Eventually this effect becomes dominant and the Q begins to decrease with increased powder dielectric constant. Propagation of a higher-order mode (or modes) may also have contributed to the eventual decrease in Q, particularly for the 5 cm radius ring for which the decrease in Q was abrupt.

IV. SUMMARY

Waveguides consisting of a low-loss, high-dielectric constant powder packed in a rectangular channel in the surface of a plastic substrate exhibit losses as low as 0.09 dB/cm at 94 GHz. These waveguides appear to be attractive for passive millimeter wave integrated circuits. Guide wavelengths calculated with Marcatili's approximate theory are in reasonable agreement with experimental measurements at 94 GHz. Powder loss tangents at 94 GHz were calculated from attenuation measurements. Ring resonators implemented with powder core channel waveguide exhibit Q's as high as 2400 at 94 GHz implying waveguide loss of approximately 0.06 dB/cm or 6 dB/m. Coupling to the resonators was achieved via adjacent straight powder core channel waveguides.

V. APPENDIX

COMPUTING WAVEGUIDE LOSS USING MARCATILI'S THEORY

Marcatili's theory [7] applies to lossless rectangular dielectric waveguide. The difficulty in obtaining a closed-form

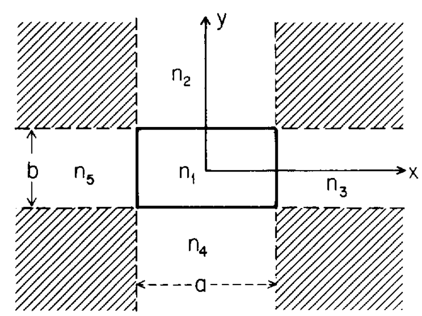

Fig. 7. Cross section of dielectric waveguide analyzed by Marcatili.

solution for this waveguide is contained in the mixed boundary conditions in the "corner regions" shown shaded in Fig. 7. Marcatili observes that, for a guided mode, a very small fraction of the power propagates in the corner regions. Hence it is reasonable to ignore the fields in these regions. Fields are assumed to exist only in regions 1–5 and these are matched only at the boundaries of region 1. Marcatili simplifies the problem further by assuming that the refractive index difference between region 1 and any other region is small. Stated precisely, Marcatili assumes that $1 - n_i/n_1 \ll 1$, $i = 2, 3, 4, 5$. (For our channel guides, the left-hand side of this inequality typically equaled 0.25 for $i = 3, 4, 5$ and 0.5 for $i = 2$; thus, strictly speaking, we do not satisfy Marcatili's approximation.)

As a result of these assumptions, the modes found by Marcatili have almost purely transverse fields (TEM) and can be grouped into two sets, E^y_{pq} and E^x_{pq}. For both families of modes, the subscripts p and q indicate the number of extrema of the transverse field components in the x and y directions, respectively.

The propagation constant of a mode, k_z, is found by matching tangential field components at the edges of region 1 in two steps, equivalent to superposing two slab waveguides: $n_2 - n_1 - n_4$ and $n_3 - n_1 - n_5$. For E^y_{pq} modes,

$$k_z^2 = k_1^2 - k_x^2 - k_y^2,$$

where $k_1 = n_1 k$, $k = 2\pi/\lambda$, and λ is the free space wavelength. The constant k_x is found by solving the transcendental characteristic equation for the $n_3 - n_1 - n_5$ slab,

$$k_x a = p\pi - \tan^{-1}(k_x \xi_3) - \tan^{-1}(k_x \xi_5)$$

in which

$$\xi_3 = ((\pi/A_3)^2 - k_x^2)^{-1/2}$$
$$\xi_5 = ((\pi/A_5)^2 - k_x^2)^{-1/2}$$

and

$$A_3 = \lambda/(2(n_1^2 - n_3^2)^{1/2}) \qquad A_5 = \lambda/(2(n_1^2 - n_5^2)^{1/2}).$$

The constant k_y is found by solving the transcendental characteristic equation for the $n_2 - n_1 - n_4$ slab,

$$k_y b = q\pi - \tan^{-1}((n_2/n_1)^2 k_y \eta_2) - \tan^{-1}((n_4/n_1)^2 k_y \eta_4),$$

in which,

$$\eta_2 = ((\pi/A_2)^2 - k_y^2)^{-1/2} \qquad \eta_4 = ((\pi/A_4)^2 - k_y^2)^{-1/2}$$

and

$$A_2 = \lambda/(2(n_1^2 - n_2^2)^{1/2}) \qquad A_4 = \lambda/(2(n_1^2 - n_4^2)^{1/2}).$$

A similar procedure is used to find the propagation constants of the E^x_{pq} modes.

Marcatili's theory can be used to calculate waveguide loss as a function of material losses for low-loss rectangular dielectric channel waveguides in the same way the calculation is made for low-loss metal waveguide. Namely, we assume that the dielectric losses are so small in all regions that the fields are the same as for the lossless case [17]. Losses due to dimensional imperfections (e.g., waveguide wall roughness) and to scattering from material inhomogeneities are not included in this analysis. We also note that for our rectangular dielectric channel guide, regions 3–5 in Marcatili's analysis (Fig. 7) have identical material properties.

Since the material losses are assumed to be small, they can be taken into account by multiplying each of Marcatili's field components by the factor $\exp(-\alpha z)$. Here 2α is given by the ratio of the average power dissipated per unit length, A, to the average power flowing along the guide, F. These quantities, in turn, are found from the field components given by Marcatili, assuming that the loss per unit volume is everywhere proportional to the square of the electric field. In the following, we take Marcatili's arbitrary field constant, M_1, equal to unity.

For the E^y_{pq} modes, α can be expressed as[3]

$$\alpha = A/2F, \tag{1}$$

[3] This expression for α could be written as a single large equation, but the form used below proved convenient for calculation.

where

$$A = (\omega\epsilon_1'/\tan\delta_1 P_1 + \omega\epsilon_4' \tan\delta_4(P_3 + P_4 + P_5) + \text{and } \omega\epsilon_2' \tan\delta_2 P_2)/(\pi/k_z)$$

$$P_1 = (c_1^2 I_1 I_2 + c_2^2 I_1 I_3)(\pi/2k_z)$$

$$P_2 = (c_3^2 + c_4^2) d_2^2 I_1 I_4(\pi/2k_z)$$

$$P_3 = (c_5^2 I_2 d_1^2 + c_2^2 I_3 d_1^2) I_5(\pi/2k_z)$$

$$P_4 = (c_7^2 + c_8^2)(d_4)^2 I_1 I_6(\pi/2k_z)$$

$$P_5 = P_3$$

$$d_2 = \cos((k_y b/2) + \gamma)\exp(b/(2\eta_2))$$

$$d_1 = \cos(k_x a/2)\exp(a/(2\xi_3))$$

$$d_4 = \cos((-k_y b/2) + \gamma)\exp(b/(2\eta_4))$$

$$I_1 = (a/2) + \sin(k_x a)/(2k_x)$$

$$I_2 = (b/2) + \sin(k_y b)\cos(2\gamma)/(2k_y)$$

$$I_3 = (b/2) - \sin(k_y b)\cos(2\gamma)/(2k_y)$$

$$I_4 = (\eta_2/2)\exp(-b/\eta_2)$$

$$I_5 = (\xi_3/2)\exp(-a/\xi_3)$$

$$I_6 = (\eta_4/2)\exp(-b/\eta_4)$$

$$c_1 = (k_1^2 - k_y^2)/(\omega\epsilon_0 n_1^2 k_z)$$

$$c_2 = k_y/(\omega\epsilon_0 n_1^2)$$

$$c_3 = ((1/\eta_2)^2 + k_2^2)/(\omega\epsilon_0 n_2^2 k_z)$$

$$c_4 = 1/(\eta_2 \omega\epsilon_0 n_2^2)$$

$$c_5 = (k_3^2 - k_y^2)/(\omega\epsilon_0 n_3^2 k_z)$$

$$c_7 = ((1/\eta_4)^2 + k_4^2)/(\omega\epsilon_0 n_4^2 k_z)$$

$$c_8 = 1/(\eta_4 \omega\epsilon_0 n_4^2)$$

$$\gamma = \tan^{-1}(n_1^2/n_2^2 \eta_2 k_y) - k_y b/2$$

$$k_2 = n_2 k$$

$$k_3 = n_3 k = k_4 = n_4 k.$$

Note that $\omega\epsilon_i' \tan\delta_i P_i/(\pi/k_z)$ is the average power dissipated per unit length in region i. The variable F can be expressed as

$$F = W_1 + W_2 + W_3 + W_4 + W_5,$$

where

$$W_1 = c_1 I_1 I_2/2$$

$$W_2 = c_3 I_1 I_4 d_2^2/2$$

$$W_3 = c_5 I_2 I_5 d_1^2/2$$

$$W_4 = c_7 I_1 I_6 d_4^2/2$$

$$W_5 = W_3.$$

Here, W_i represents the average power flowing in region i.

In the equations above, $\epsilon' = \mathrm{Re}\,(\epsilon)$ and ϵ_0 is the permittivity of free space. Also, the subscripts 1, 2, 3, and 4 after the variables k, ϵ, n, and $\tan\delta$ associate them with the corresponding regions shown in Fig. 7. In addition, we note that α depends implicitly on the mode indexes p and q through k_x and k_y.

If the effective loss tangents of our powders had been known at 94 GHz, (1) could have been used to predict the transmission losses of our channel guides. However, since the 94 GHz effective powder loss tangents were unknown, (1) was used to calculate them from the measured transmission losses of the E_{11}^y mode, using literature values [15], [16] of the loss tangents of the substrate materials. (The dielectric properties of heterogeneous media, such as powders, are often called "effective" to distinguish them from the bulk properties of the constituent materials. In this paper, the term "effective" has been omitted when the properties being discussed are clearly those of powders.)

Some of the results are given in Table IV. As shown there, the effective loss tangents increased with powder density, as expected [2]. This effect causes the waveguide loss to increase. However, the rise in waveguide loss accompanying increased powder density is not solely due to the increase in effective loss tangent for the powder. The increased effective dielectric constant of the powder also caused a greater fraction of the power of the guided mode to travel in the core, rather than in the lower loss substrate.

As shown in Tables II and IV, waveguides made with 43 μm D-30 powder exhibited much lower losses than those using the larger grain D-30. As a result, the computed values of effective loss tangent for the 43 μm D-30 were smaller. The fact that higher losses were measured for the D-30 powder with larger particles suggests that the extra loss is due to scattering. However, we were not able to detect significant scattered radiation with the small horn-plus-Schottky-detector probe for *any* powder waveguide. In addition, theories of scattering by powders [18], [19] predict negligible scattering loss for any of these powder sizes. We do not know why the coarse D-30 powder is more absorptive.

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