TABLE I
PEAK POWER-HANDLING CAPABILITY OF WR-90 WAVEGUIDE HOUSING FINLINES WITH DIFFERENT GAP WIDTHS AT 9.6 GHz (*X*-BAND)

d (mm)	Z_0(ohms)	1st trial (kW)	2nd trial (kW)	Average (kW)
0.5	145 6	5.60	5 43	5 51
1.0	188.4	9.50	8.38	8 94
1 5	213.7	10 75	10 75	10 75
2 0	238 3	12 00	11 50	11 75
2 5	261 2	19 06	18 02	18.54

Repetition of pulse is 1000 pulses/s with pulse duration of 0.8 μs. Conductor thickness $g = 34$ μm and substrate thickness $s = 762$ μm.

TABLE II
PEAK POWER-HANDLING CAPABILITY OF WR-28 WAVEGUIDE HOUSING FINLINES WITH DIFFERENT GAP WIDTHS AT 35.5 GHz (*Ka*-BAND)

d (mm)	Z_0(ohms)	1st trial (kW)	2nd trial (kW)	Average (kW)
0 2	168.5	0.94	0 96	0 95
0 5	217 5	3.80	3.15	3 48

Repetition of pulse is 60 pulses/s with pulse duration of 16.6 μs. Conductor thickness $g = 9.9$ μm and substrate thickness $s = 254$ μm.

TABLE III
COMPARISON BETWEEN THEORETICAL AND EXPERIMENTAL RESULTS FOR UNILATERAL FINLINE STRUCTURES AT 9.6 GHz

d (mm)	Peak power (kW) lower limit	Expt (kW)	% error	Peak power (kW) upper limit
0 5	2 125	5.510	61 4	49 95
1 0	5 349	8 938	40 2	125 7
1 5	8.830	10 75	17 9	207 6
2 0	11 76	11.75	00 0	276.4
2 5	14.25	18 54	23 1	335 0

WR-90 waveguide is used with conductor thickness $g = 34$ μm and substrate thickness $s = 762$ μm.

TABLE IV
COMPARISON BETWEEN THEORETICAL AND EXPERIMENTAL RESULTS FOR UNILATERAL FINLINE STRUCTURES AT 35.5 GHz

d (mm)	Peak power (kW)	Expt (kW)	% error
0 2	1 002	0 950	5 47
0.5	3.240	3.475	6.76
1 0	6 109	—	—
2.0	9 663	—	—

WR-28 waveguide is used with conductor thickness $g = 9.9$ μm and substrate thickness $s = 254$ μm.

to determine accurately, especially when the electric field becomes highly inhomogeneous, which is the case for narrow gaps. The calculated maximum power values correspond to the model that assumes the breakdown to occur only over the short distance observed on the samples after experiments. This value is about 12 μm and depends on the size of the irregularities of the profiles. The theoretical upper limit for which the electric breakdown would occur along the entire structure is about one order of magnitude above the experimental values, but does not correspond to a realistic situation.

VI. CONCLUSION

The theoretical model to predict the peak power-handling capacity of finline structures was verified by an experimental procedure carried out at 9.6 GHz (*X*-band) and 35.5 GHz (*Ka*-band). Theoretical and experimental values compare fairly well, especially for structures with gap widths larger than 1.5 mm. The best prediction is made when it is assumed that the breakdown occurs only over a small distance along the line, which corresponds to the size of the irregularities of the profiles (about 12 μm). This value can be used to determine the ionization rate for extrapolation to higher frequencies.

The analysis of the samples after breakdown showed that, for pulsed operations, no significant heat is produced due to the relatively low average power involved in this case. It is found that at 35.5 GHz (*Ka*-band), the maximum peak power that can be handled by a unilateral finline of 220 Ω characteristic impedance is about 3.5 kW under narrow pulse signal operation.

In conclusion, the theoretical breakdown model developed and presented previously by the authors [1] is appropriate for predicting the peak power-handling capability of finlines for the purpose of establishing limits of safe operation under laboratory conditions.

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An Evanescent-Mode Tester for Ceramic Dielectric Substrates

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***Abstract*—The TE_{01} mode in a cylindrical waveguide at a frequency below cutoff is used to probe a ceramic dielectric substrate located on the central plane between input and output coupling loops. Maximum transmission occurs at a frequency determined by the waveguide radius, the substrate thickness, and the dielectric constant. The dielectric constant and**

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loss tangent are obtained from the resonant frequency and the absorption bandwidth. The measurement is insensitive to the position of the substrate in the gap between waveguide sections, and no intimate contact is required.

I. Introduction

The dielectric constant and loss tangent of ceramic substrates are data essential to the design of microstrip and integrated microwave circuits. Accordingly, accurate measurement methods are required by the circuit designer and the substrate manufacturer as well. For both, the simplicity and convenience of this method are important, but the user and producer may judge these merits by different criteria.

The methods proposed in the literature require intimate contact between the dielectric and a circuit. The contact may be achieved either by metallization or by pressing a circuit on a soft substrate against the ceramic [1]. With metallization of two or all six sides of the substrate, a cavity resonator is formed [2]–[4], and the required data can be calculated from frequency and Q measurements. Other procedures, dependent on metallization, derive their results from measurements on specially fabricated circuits or from circuits such as microstrip lines [5] that may be provided by the manufacturer.

The merits of these methods are distributed between convenience and accuracy. When metallization is required, convenience for the user depends largely on the metallization pattern supplied by the manufacturer. For the producer of unmetallized substrates, metallization is a costly and inconvenient requirement. Since it is an irreversible process that wastes substrates, it is suited only for sampling a production run. The pressure contact with the circuit [1] is more suited for production testing, although it requires calibration with a standard substrate.

The need to establish intimate contact between the dielectric and a conductor can be eliminated by using as a probe an electric field that has no component normal to the dielectric surface. If the sample is cylindrical, this may be accomplished by placing it in a cylindrical waveguide of equal diameter or in a radial waveguide. In both cases, the resonance of a mode with azimuthal electric fields is measured [6]. The method proposed here combines the cylindrical and radial waveguides so that the sample need not be cylindrical. The dielectric is placed on the midplane between two cylindrical waveguides, as shown in Fig. 1. Transmission through the sample of the TE_{01} mode is a maximum at a frequency below cutoff in the waveguides but above cutoff in the cylindrical portion of the sample. In the radial guide ($r > a$), the fields are rapidly evanescent, and the extension of the sample beyond the waveguide radius has little effect on the frequency. Any fields normal to the dielectric surface occur in the evanescent region and are of no consequence. The resonance frequency is close to that of the $TE_{01\delta}$ mode of a dielectric resonator of diameter $2a$ and height $2d$.

II. Analysis of Operation

Modes that propagate or are evanescent both in the empty portions of the waveguide and in the dielectric need not be considered; the terminations at the ends eliminate any narrowband resonance phenomena. It is assumed as well that a mode that is evanescent in the waveguide but propagates in the dielectric is evanescent in the radial waveguide in the gap beyond the radius a. This justifies the use of the conducting wall boundary condition at $r = a$ in the dielectric. The error that follows from this approximation can be estimated by a perturbation calculation.

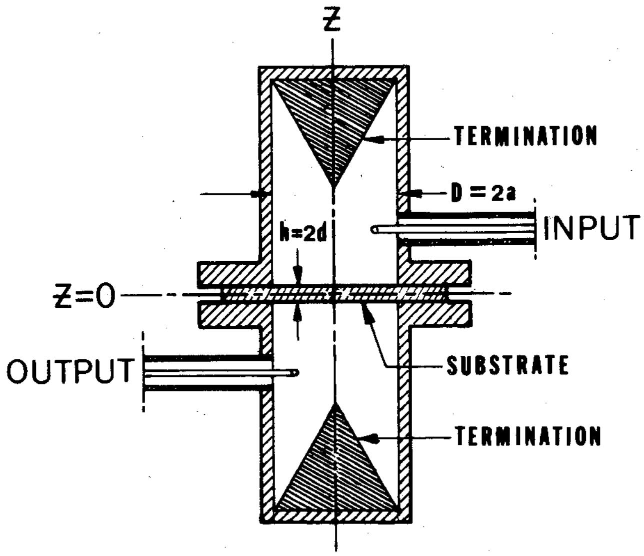

Fig. 1. Schematic cross section of substrate tester.

The boundary conditions at the dielectric surfaces lead to the eigenvalue equations

$$(\theta \sin\theta - \gamma\cos\theta)(\theta\cos\theta + \gamma\sin\theta) = 0 \tag{1}$$

for TE modes, and

$$(\theta \sin\theta - K\gamma\cos\theta)(\theta\cos\theta + K\gamma\sin\theta) = 0 \tag{2}$$

for TM modes. The notation is defined as follows:

$$\theta^2 = K\theta_0^2 - \theta_c^2, \qquad K = (\epsilon'/\epsilon_0) \tag{3}$$

$$\gamma^2 = \theta_c^2 - \theta_0^2 \tag{4}$$

$$\theta_c = (x_{lm} d/a) \tag{5}$$

$$\theta_0 = (\omega d/c). \tag{6}$$

For a TE_{lm} mode, x_{lm} is the mth zero of the derivative of J_l, and for a TM_{lm} mode it is the mth zero of J_l. If the assumptions are correct, a resonance exists such that (4) is positive, and (1) or (2) has a solution for θ in terms of γ. Then, the dielectric constant K can be calculated from (3).

In both (1) and (2), the first term is zero for even modes, and the second vanishes for odd modes. For the reasonable constraints of $K < 100$ and $(d/a) < 0.05$, there are no odd-mode solutions. The effect of the substrate on TM modes is relatively small, and the even-mode TM resonant frequencies are close to their cutoff frequencies. For the TE_{01} and TM_{11} modes, which have the same cutoff frequency, the substrate produces a wide separation between the two even resonances.

The modes one expects to observe in order of increasing frequency are $TE_{11}, TE_{21}, TE_{01}, TE_{31}, \cdots$. With small coupling loops in the transverse plane, coupling to the TM modes is very weak. The resonances of the first four TE modes are well separated and easily identified. All these frequencies can be used to calculate K, but only the TE_{01} mode gives reliable results. The relation between K and this frequency is shown by the curves of Fig. 2.

At resonance of the TE_{01} mode, the reciprocal of the unloaded Q is

$$(1/Q_0) \equiv D = \left[(\epsilon''/\epsilon') + (\delta/a)(\theta_c/\theta_0)^2 U\right] / [1 + U] \tag{7}$$

where δ is the skin depth in the waveguide, $(\epsilon' - j\epsilon'')$ is the

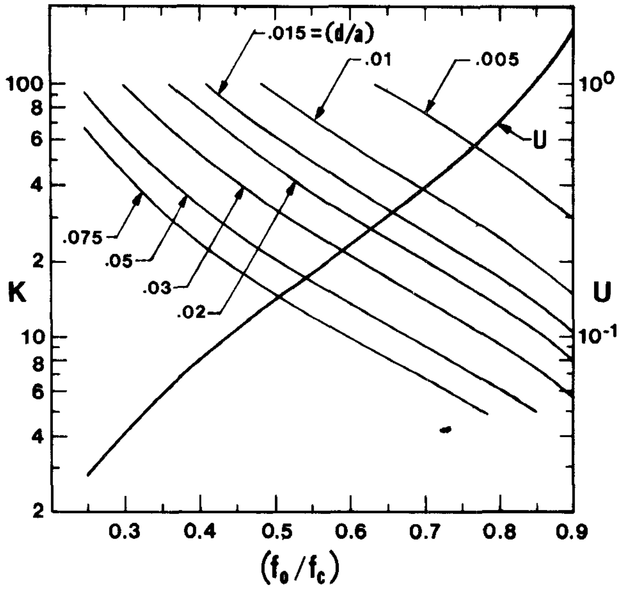

Fig. 2. Curves of K versus (f_0/f_c) of the TE_{01} mode with (d/a) as parameter. Also shown is the energy ratio U for $(d/a) = 0.03$.

complex permittivity of the substrate, and

$$KU = [\cos^3\theta]/[(\theta + \sin\theta\cos\theta)\sin\theta]. \tag{8}$$

The quantity U is the ratio of the electric energy in the waveguide to that in the substrate. As shown in Fig. 2, it increases rapidly as the cutoff frequency is approached and K decreases. It is insensitive to (d/a) over its practical range. The contribution to D from conductor losses, represented by the second term in the numerator of (7), cannot be made negligibly small, and it cannot be measured at the correct frequency in any simple way. Calculations of that term (Fig. 3), using the bulk conductivity of brass, show that it increases more or less as the root of the cutoff frequency.

Three sources of error in the determination of K are uncertainties in the measurement of a, d, and f. These are approximated

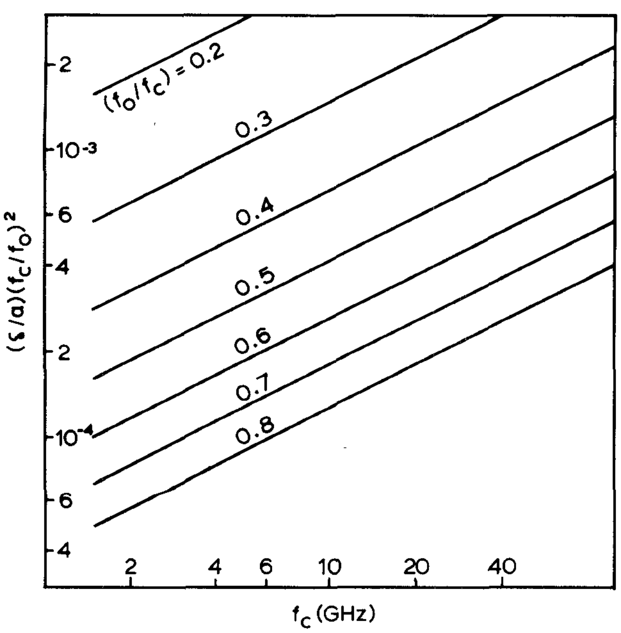

Fig. 3. Plot of $(\delta/a)(f_c/f_0)^2$ versus f_c for brass waveguide.

TABLE I
GAP EFFECT

Dimensions	Calculated K	$\Delta K/K$
$h = 0.0980$ cm	38.1 with gap	0.005
$D = 3.801$ cm	37.9 without gap	0.003 by eq. (12)

by the following expressisons:

$$(\Delta K/K)_a = -2(\theta_c/\theta_0)^2(U + 1/K)(\Delta a/a) \tag{9}$$

$$(\Delta K/K)_d = -\frac{2}{K}\left(\frac{\theta}{\theta_0}\right)^2 \frac{1}{1+\sin 2\theta/2\theta}\left(\frac{\Delta d}{d}\right) \tag{10}$$

$$(\Delta K/K)_{\theta_0} = -2(1+U)(\Delta\theta_0/\theta_0). \tag{11}$$

The coefficients of $(\Delta a/a)$, $(\Delta d/d)$, and $(\Delta\theta_0/\theta_0)$ are of the order (1–5). The frequency error coefficient is likely to be the largest, but the frequency can be most accurately measured.

A fourth source of error is the assumption that the electric field in the substrate is zero at $r = a$. This can be estimated as a frequency perturbation [7] produced by removing the stored energy of the evanescent radial modes in the space $r > a$. The approximate result is

$$(\Delta K/K) = (4/\pi K)(\theta_c/\theta_0)^2\left[1 - K(2\theta_0/\pi)^2\right]^{-1/2}(d/a). \tag{12}$$

The approximation is made by truncating the Fourier expansion in evanescent modes and equating the remaining coefficient to its asymptotic form. The validity depends on how small (d/a) is. For practical parameter values, $(Kd/a) \sim 0.25$, and an overestimate of (12) is

$$(\Delta K/K) \sim 10(d/a)^2. \tag{13}$$

Errors from this source are usually below 1 percent, and (12) can be used to correct the data.

With several testers of differing radii the TE_{01} mode can be measured over some band of frequencies. The maximum radius must be somewhat less than the width of the substrate. The minimum radius is determined by available instrumentation and the maximum permissible value of (d/a). Fig. 2 suggest that it should be possible to span up to one octave.

III. EXPERIMENTAL RESULTS

Many resonances can be observed as maxima in the transmission through the test substrate. The lower modes can be identified by assuming a mode index, and assessing that assumption by the corresponding calculated value of the dielectric constant. If the assumption is false, the calculation yields a dielectric constant far removed from the expected value. The modes thus identified are TE_{11}, TE_{21}, TE_{01}, TE_{31}, TE_{41}, TE_{12}, TE_{51}, and TE_{02}. All modes are even with respect to z; no TM modes have been identified.

The calculations of K from these resonant frequencies give values that are low for all but the TE_{01} and TE_{02} modes. These are also distinguished by the insensitivity of the frequency to movement of the substrate. The substrate can fit loosely in the holder, but movement in neither the z direction nor the transverse direction has a significant effect. Although the TE_{02} mode gives good data, it is too close in frequency to other modes to allow certain identification. The TE_{01} mode is always well separated and identifiable as the third in the sequence.

TABLE II
MEASURED VALUES OF K FOR VARIOUS MATERIALS AND THICKNESSES

Material	K10 P125			K38 NPO						K70 NPO		
Thickness	0.0632(cm)			0.02336		0.0648		0.1282		0.1004		
Diameter	f(GHz)	K*	K	f	K	f	K	f	K	f	K*	K
3.82(cm)	8.59	9.83	9.80	7.83	38.0	5.63	38.2	4.29	38.1	3.65	68.1	67.7
3.21	9.91	9.86	9.82	8.88	38.4	6.26	38.3	4.75	38.1	4.03	68.1	67.7
2.60	11.7	9.82	9.78	10.29	38.5	7.10	38.3	5.38	37.9	4.54	68.2	67.8
1.99	14.3	9.84	9.77	12.31	38.5	8.32	38.2	6.30	37.8	5.29	68.3	67.5
Bridge	1(Mhz)		10.00					(1MHz)	38.1	1(MHz)		67.3

In order to assess the error that comes from the extension of the substrate beyond the radius of the waveguide (gap effect), two measurements were made on a substrate with a nominal K value of 38 and a thickness of 0.0385 in. For the first, the substrate was square (2 in.$\times$2 in.) extending outside the 1.5 in. waveguide diameter. For the second, a circle was cut from the center to fit inside the tester. The results of these measurements are shown in Table I. The gap produces an apparent increase in the value of K, as is to be expected from the inductive property of the gap fields. The fractional error in K is as close to that predicted by (12) as can be expected from truncated calculations. Moreover, the somewhat loose fit of the circular substrate into the waveguide tends to increase the frequency and lower the calculated dielectric constant.

The results of measurements on a variety of substrate materials and sizes are shown in Table II. The bottom line shows results obtained by a bridge measurement of capacitance at 1 MHz. For this, the substrate is metallized and a circular electrode with guard is formed by etching a circular line. The agreement between the microwave data and the bottom line shows, as is generally accepted, that the dielectric constants are independent of frequency.

To increase the resonance frequency for a given sample, a must be reduced; the increase of (d/a) tends to increase the gap error with frequency. The gap error is illustrated in the first and last columns by the uncorrected values of K (designated K^*). The corrections, made with (12), are less than 1 percent except for the entry in the last row, last column, where (d/a) is greatest.

Apart from dielectric dispersion, discrepancies between the microwave results and those derived from capacitance data can be attributed to errors in the measurement of d and to surface damage of the sample. A positive error in the thickness measurement produces a positive error in the K calculated from the capacitance data, but the error from the frequency calculation is negative. Since both error coefficients are of order one, the discrepancy between the calculated K values is approximately twice the error in the thickness measurement. The thickness error can be reduced by taking the geometric mean of the tester and bridge values. This procedure applied to alumina (column one) gives $\langle K \rangle = 9.895 \pm 0.005$ as the corrected average.

If the ceramic is relatively soft, lapping and polishing leave some debris embedded in a thin surface layer, and its dielectric constant is less than the bulk value. The effect on the TE_{01} frequency is equivalent to a small positive error in the measurement of d, and the calculated K is low. The effect on the capacitance calculation is also an underestimate but amplified by a factor of $(K-1)$. If, for example, K increases from one to its bulk value of 26 in a layer of thickness 2×10^{-4} cm on a 1 mm sample, the capacitance calculation is 10 percent low; the TE_{01} calculation is about 0.4 percent low.

To test the accuracy of the procedure, the transverse dielectric constant of sapphire was measured. The result, $K = 9.405$ at 11.6 GHz, is close to the published value of 9.34 [8].

The error in the loss tangent that results from uncertainty in the effective conductivity of the waveguide can be minimized by using low-loss material and choosing the frequency range so that the energy ratio $U \ll 1$. Measurement and calculations from (7) for the K38, NPO material at 4.29 GHz give

$$(\epsilon''/\epsilon') = (2.65)[1 \pm (0.11)(\Delta\sigma/\sigma)]10^{-4}. \tag{14}$$

A 10 percent uncertainty in conductivity has less effect on (ϵ''/ϵ') than the uncertainty in the bandwidth measurement.

IV. CONCLUDING REMARKS

With suitable software and instrument automation, this method of testing is fast, repeatable, and accurate. Although a loose fit of the substrate in the gap has negligible influence on the resonance frequency, the coupling is sensitive to its position. Greater care is required to obtain repeatable data on the dissipation factor.

The test method is limited to materials that are isotropic in the plane of the sample. The range of dielectric constants is limited on the low end by the proximity of the resonance to the cutoff frequency; at the high end the limitation is related to the coupling problem at a frequency far below cutoff. An estimate of the range is 5–1000. Nonuniformity in the transverse plane can be explored qualitatively if the waveguide diameter is considerably smaller than the substrate.

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