provides a useful approximation for the purposes of this example. The embedding impedance Z_e seen by the diode at the LO frequency and its harmonics is tabulated in Fig. 3.

Fig. 4 shows the convergence parameter $|Z_d(jn\omega_{LO})|/|Z_e(jn\omega_{LO})|$ as a function of harmonic number n, when 16 harmonics are considered. For 300 iterations, convergence is reasonably complete up to the 11th harmonic. For 500 iterations $|Z_d|$ is within 0.5 percent of $|Z_e|$ up to the 15th harmonic.

The diode current and voltage waveforms are shown in Fig. 5. The computation time per cycle of iteration, when 16 harmonics are considered, is 3 ms on an IBM 360/95, and 2 s on an IBM 360/50.

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Wide-Band Characteristics of a Coaxial-Cavity Solid-State Device Mount

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Abstract—**An experimental comparison is made between two theoretical approaches to evaluation of the driving-point impedance at the terminals of a gap in the center conductor of a coaxial cavity. The study shows that for wide-band characterization, radial-wave modal-field analysis provides greater accuracy than the conventional transmission-line approach.**

I. INTRODUCTION

This short paper reports a comparison between two methods for the analysis of a coaxial cavity over a wide frequency range. The cavity, shown in Fig. 1, is assumed to be formed of perfectly conducting material, and contains a gap in the center conductor, within which a solid-state device may be located. The purpose of the study was to determine the driving-point impedance (i.e., the reactance of the lossless structure) viewed from the gap terminals, for a wide range of frequencies and cavity dimensions.

Attention has previously been confined to the lowest order resonance of the cavity, using Green's function [1] and radial line analyses [2], [3], in addition to conventional TEM-mode transmission-line theory. Our interest in wide-band characterization arises from the need for knowledge of harmonic-frequency impedance values in the design of solid-state oscillators [4], [5].

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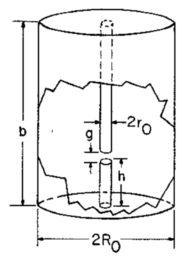

Fig. 1. Lossless coaxial cavity for use as a solid-state device mount.

II. THEORETICAL ANALYSIS

Two approaches to determination of the impedance viewed from the center-conductor gap are described briefly here.

A. Transmission-Line Analysis

This analysis yields the circuit shown in Fig. 2. The inductance L accounts for magnetic energy storage in the annular region of length g, external to the gap. The capacitances C_1' and C_2 represent the gap discontinuity, determined from the approach of Green [6] and Dawirs [7]. C_1' represents the series gap capacitance of these authors, modified by subtraction of a parallel-plate capacitance C_0, since we are interested in the impedance which loads a packaged device mounted in the gap; this distinction has also been made by Getsinger [8]. Using Fig. 2, the susceptance $B_R = -1/X_R$ is given by

$$B_R = \omega C_1' - [\omega L + Z_0(T_1 \tan \beta l_1 + T_2 \tan \beta l_2)]^{-1} \quad (1)$$

where

$$T_i = (1 - \omega C_2 Z_0 \tan \beta l_i)^{-1}, \qquad \text{for } i = 1,2.$$

B. Radial-Wave Analysis

The input reactance is calculated by establishing two sets of radial waves, outward bound and inward traveling, and imposing perfect-conductor boundary conditions at $r = R_0$. Using the concept of complex Poynting vector power, the radiation impedance at the gap terminals is determined by an approach similar to that of Eisenhart and Khan [9]; this impedance reduces to a reactance jX_R for the lossless structure considered here. Summing over all possible modes, we obtain

$$B_R = \sum_{n=0}^{\infty} \frac{(K_{gn})^2}{Z_n}$$

where

$$Z_n = j\left[\frac{(1+\delta_{n0})b\eta}{4\pi r_0}\frac{\beta_n}{k}\right]\left[\frac{J_0(\beta_n r_0)Y_0(\beta_n R_0) - J_0(\beta_n R_0)Y_0(\beta_n r_0)}{J_1(\beta_n r_0)Y_0(\beta_n R_0) - J_0(\beta_n R_0)Y_1(\beta_n r_0)}\right]$$

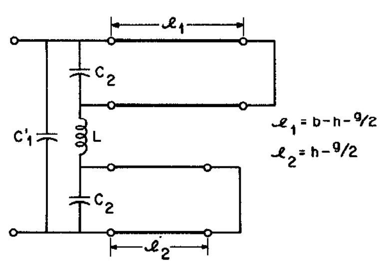

Fig. 2. Equivalent circuit of the driving-point reactance, using the transmission-line approach.

for propagating modes, and

$$Z_n = -j\left[\frac{(1+\delta_{n0})b\eta}{4\pi r_0}\frac{\gamma_n}{k}\right]\left[\frac{K_0(\gamma_n r_0)I_0(\gamma_n R_0) - I_0(\gamma_n r_0)K_0(\gamma_n R_0)}{K_1(\gamma_n r_0)I_0(\gamma_n R_0) + I_1(\gamma_n r_0)K_0(\gamma_n R_0)}\right]$$

for evanescent modes, and

$$K_{gn} = \cos\left(\frac{n\pi h}{b}\right)\frac{\sin\left(\frac{n\pi g}{2b}\right)}{\left(\frac{n\pi g}{2b}\right)}$$

is the gap coupling factor [9].

$$\beta_n = [k^2 - (n\pi/b)^2]^{1/2}$$

$$\gamma_n = [(n\pi/b)^2 - k^2]^{1/2}$$

and

$$\eta = \left[\frac{\mu}{\epsilon}\right]^{1/2}$$

$$k = \omega(\mu\epsilon)^{1/2}.$$

The $J_i(x)$ and $Y_i(x)$ are the Bessel functions of the first and second kind, respectively, of order i and argument x; $K_i(x)$ and $I_i(x)$ are the corresponding modified functions.

III. EXPERIMENTAL RESULTS

The reactance was measured from the gap terminals looking into the cavity by use of a subminiature coaxial probe of the form described by Eisenhart and Khan [9]. The modeling and measurement procedures are similar to those previously reported [9].

Fig. 3 presents a comparison between the measured values of reactance, over a 2–22-GHz range, and the values determined by the two theoretical approaches described in the preceding. The comparison was carried out for two values of $b/2R_0$. Fig. 3(a), for the cavity with $b/2R_0 = 2.15$, indicates that both analyses give similar results, and their theoretical values are almost identical. Fig. 3(b), for the cavity with $b/2R_0 = 1.04$, shows that the radial-wave analysis successfully predicts several resonant frequencies which are not determined by the transmission-line approach. Thus use of the modified transmission-line theory gives erroneous results for higher resonant frequencies of the latter cavity.

IV. DISCUSSION

Using the radial-wave theory, we can examine the effect of variation in position of the solid-state device along the center-conductor length of a coaxial cavity. Fig. 4 shows the driving-point reactance, from the gap terminals, for the cases where the gap is at the bottom of the cavity and where it is in the center. It is evident that the pole-zero spacing is markedly aperiodic, and that the spacing varies with gap position. Thus Fig. 4 shows that this spacing may be varied, and some tailoring of the reactance variation accomplished, through suitable choice of the gap position. The zeros around 10.1 and 16.5 GHz in Fig. 4 were eliminated through movement of the gap to the center of the cavity.

V. CONCLUSION

A comparison between the modified transmission-line approach and the radial-wave approach to the analysis of coaxial cavities having a center-conductor gap shows the superiority of the radial-wave analysis when the cavity has a diameter of the same magnitude as the length. Both analyses are accurate for long cavities. The radial-wave analysis shows that the driving-point reactance characteristic can be modified, to some extent, by variation in the position of the gap along the cavity center conductor.

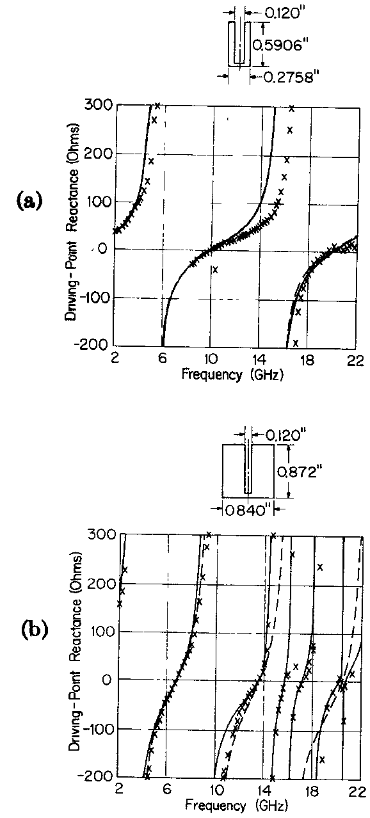

Fig. 3. Comparison between theoretical and experimental values of the driving-point reactance. The solid line is for the radial-wave analysis; the broken line is for the transmission-line approach; the discrete points are measured data. (a) A cavity with $b/2R_0 = 2.15$ and $R_0/r_0 = 2.30$. (b) A cavity with $b/2R_0 = 1.04$ and $R_0/r_0 = 7.00$.

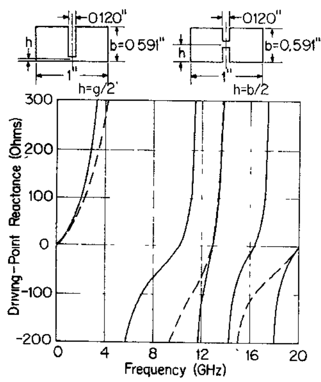

Fig. 4. Driving-point reactance of the cavity, using the radial-wave analysis, for two gap positions. The solid line is for $h = g/2$; the broken line is for $h = b/2$. For both cavities $R_0 = 8.34r_0$; $g = 0.12R_0$.

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Wide-Bandwidth Millimeter-Wave Gunn Amplifier in Reduced-Height Waveguide

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***Abstract*—A *Ka*-band millimeter-wave Gunn-amplifier structure with 15-GHz gain$^{1/2}$-bandwidth product has been fabricated in reduced-height waveguide. Measurements were taken of the terminal admittance of the diode and its mount and used in the computer optimization of impedance matching transformer sections.**

SUMMARY

A reduced-height waveguide structure housing a Gunn diode has produced *Ka*-band gain$^{1/2}$-bandwidth products of approximately 15 GHz. Previously, large-bandwidth millimeter-wave Gunn amplifiers have been constructed using coaxially mounted diodes iris coupled to waveguide [1], although amplification at *X* band with moderate bandwidths has been reported [2] using reduced-height guide. It is expected that millimeter-wave InP diodes will shortly become more available and it is hoped that low-noise-figure amplifiers (such as the often quoted 7.5 dB [3]) can be constructed with useful gain over large bandwidths.

The purpose of this work was initially to use available supercritically doped GaAs diodes[1] to do the following.

1) Find a waveguide structure which would be both stable and capable of large bandwidth operation with reasonable gain.

2) Develop a method which can be used for experimentally determining the circuit effects of the diode and its mount (including parasitics). The latter, which is called "terminal" admittance in this short paper, is of particular importance in circuit design.

3) Computer optimize a matching structure given the terminal admittance found above.

In the experiments to be described, the Gunn diodes were center mounted in reduced-height double-cosine tapered cavities. A compressible gold-plated bellows allowed vertical movement of the diode as shown in Fig. 1. The 0.118-in-diam heat sink was allowed to penetrate into the 0.010-in reduced-height cavity in order to create a waveguide discontinuity found essential for large-bandwidth operation. With no penetration of the heat sink, useful gain could be found (moving an adjustable short) for only one or two hundred MHz. Maximum bandwidth (7.5 GHz) was obtained for the heat sink inserted about halfway into the cavity.

In the original structure a second double-cosine taper behind the diode was terminated by a spring short formed by doubling over a thin strip of 0.280-in brass. Two extremes of gain and bandwidth found by adjusting the diode height and short position are shown in Fig. 2.

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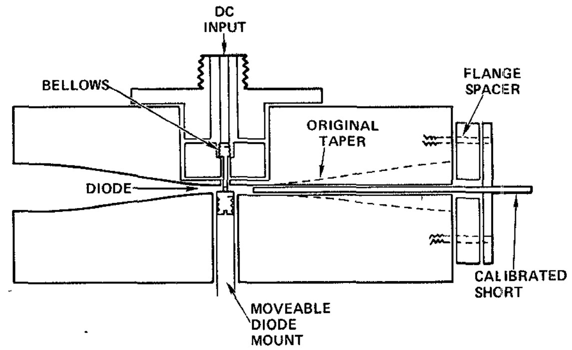

Fig. 1. Original Gunn-diode amplifier mount. Modification allowed measurements of diode terminal admittance.

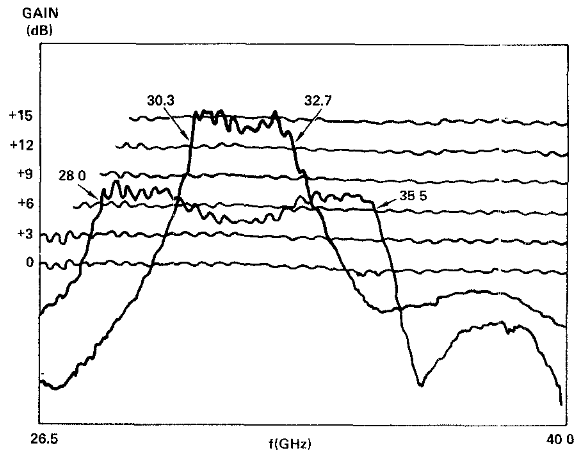

Fig. 2. Measured gain with two different diode and short positions.

In order to make meaningful measurements on reflection coefficients it was necessary to change the original structure as shown in Fig. 1 so that one side of the cavity was merely reduced-height (0.010 in) guide. A metal strip can be positioned accurately within the reduced-height guide by using waveguide flanges of various widths as spacers.

Two-port *S* parameters are found for the double-cosine cavity (between the input flange and the position of the diode) by removing the diode and filling the holes in the waveguide walls with machined slugs. For each frequency of interest, reflection coefficient data are taken for a minimum of three known positions of the sliding short behind the terminal (diode) position. After all the data are taken for the cavity *S*-parameter determination, additional reflection coefficient data are taken with the biased diode in place and the short set at some position where the circuit is stable. From the new data, using the previously calculated cavity parameters, the terminal admittance can be calculated.

Once the diode terminal admittance and matching circuit *S* parameters are found, a check on the accuracy of the calculated values can easily be made. With the diode vertical position unchanged, the calibrated short is set to other positions that also give stable gain and measured reflected gain is compared to calculated gain.

Results are shown in Fig. 3 for the diode terminal admittance data taken with the short 0.142 in from the diode center. Using the *S* parameter and terminal admittance data, amplifier gain was calculated for the short 0.411 in from center and compared to measured reflectometer gain at the latter position (Fig. 4). Other comparisons between calculated and measured results, not so closely matched, can be shown to be due, in part, to the effect of higher order TE_{LO} modes which are created at the diode post discontinuity and reflected back from the short.

The diode terminal admittance data of Fig. 3 were used in a computer optimization program which determines a matching *N*-section transformer. For example, (Fig. 5), a two-section transformer (six