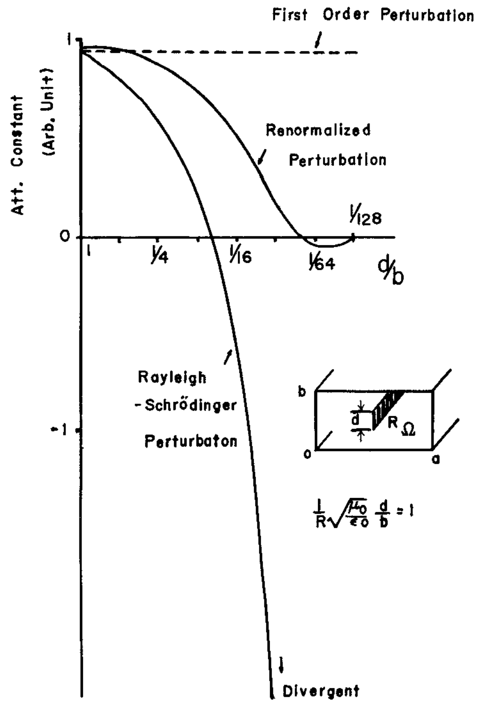

Fig. 1. Example of the renormalized perturbation method for the inhomogeneously loaded waveguide. The load $(1/R)\cdot\sqrt{(\mu_0\cdot\epsilon_0)}\cdot d\cdot b$ is 1.0, frequency $2a/\lambda_0$ is 1.4, and a/b is 2.0. The load is a thin-film resistor in the center of the waveguide.

divergence of the second-order perturbation in the quantum electrodynamics [7].

The "Renormalization Theory" is introduced by analogy with quantum mechanics to exclude the divergence. The perturbation operator $L' = L + \delta L_\alpha + \delta L_\beta$ is substituted for L in our renormalized perturbation method. δL_α and δL_β are the new operators used to cancel the real and imaginary part of the second-order term in (1)

$$\delta L_\alpha = -\operatorname{Re}\lim_{d\to 0}\sum_{j\neq i}^{\infty}\frac{\Gamma_{ij}\cdot\Gamma_{ji}}{\gamma_i^0-\gamma_j^0}$$
$$\delta L_\beta = -\operatorname{Im}\lim_{d\to 0}\sum_{j\neq i}^{\infty}\frac{\Gamma_{ij}\cdot\Gamma_{ji}}{\gamma_i^0-\gamma_j^0} \tag{3}$$

which are divergent with $d/b\to 0$ [7]. δL_α and δL_β are proportional to $1/(d/b)$ and $1/(d/b)^2$, respectively, in the region of small d/b:

$$\delta L_\alpha = A_\alpha\cdot 1/(d/b)$$
$$\delta L_\beta = A_\beta\cdot 1/(d/b)^2. \tag{4}$$

In computer-aided calculation, we cannot execute Γ_{ij} for the infinitely small d/b (at the same time infinitely small R) region and the infinite summation of the second-order term in (1). To be finite, we have to "cut off" the d/b value and the concerned mode number. For actual computer calculation, first, we assume the d_0/b value small enough, and determine the factors A_α and A_β in (4) to cancel the second-order term in (1) at this stage. For the region $d>d_0$, we postulate that these factors A_α and A_β are constant. If $d\neq d_0$, the second-order term in (1) and δL are not cancelled with each other. The residual second-order term subtracted by δL is the observed value at $d>d_0$ [7].

An example is shown in Fig. 1. The concerned mode number is 24 051, which cannot be treated by the variational method.

The load $1/R\cdot\sqrt{(\mu_0/\epsilon_0)}\cdot(d/b)$ is 1.0. The cutoff value d_0/b for computer-aided calculation is 1/128 ($R\approx 3\Omega^\square$). The Rayleigh–Schrödinger method shows the strong divergence, whereas our renormalized perturbation method gives physically reasonable values. The calculation time for our new method is proportional to N, which is smaller than for the variational or finite difference methods.

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Gunn-Effect Power Limiter

C. S. AITCHISON AND R. DAVIES

***Abstract*—The possibility of using the nonlinearity of the Gunn device current–voltage characteristic to provide microwave power limiting is discussed. Initial pulsed and CW results are presented which demonstrate limiting action.**

I. Introduction

The use of the Gunn effect in an oscillator is well known. Little attempt has been made to exploit the nonlinear nature of the Gunn effect mechanism for other applications such as mixers, harmonic generators, parametric amplifiers, and limiters.

Sterzer [1] has used the Gunn effect to construct an amplitude modulator. Aitchison [2] and others have observed parametric amplification.

This correspondence suggests that the Gunn-effect device could be used as a microwave power limiter.

II. Discussion

It is known that the Gunn-effect dc current–voltage characteristic is nonlinear and, while varying from sample to sample, is often of the form shown in Fig. 1; only some of the apparent nonlinearity is due to heating.

Work at microwave frequencies at Mullard Research Laboratories has demonstrated that a similar characteristic is available up to 3 GHz and it will be assumed in this discussion that this characteristic exists beyond 3 GHz.

The application of a microwave signal to a Gunn sample mounted in shunt with a transmission line of suitable impedance will produce a voltage across the Gunn sample; the current which flows will be a function of the applied bias and the magnitude of the voltage. A Fourier analysis of the current will show a mean (dc) component plus a component at the incident frequency (plus other components at higher frequencies which are neglected). Both the mean component and the incident frequency component will be functions of power level and the detailed behavior will be a function of the low signal bias.

In particular, the small signal conductance at the turnover point will be zero and using an established simple equivalent circuit of the Gunn sample in the form of a shunt capacitance and a shunt conductance we may expect a Gunn sample placed across a transmission line to have a small insertion loss when the appropriate bias is applied.

Increase in applied power at the incident frequency will have two effects—both of which will introduce a finite insertion loss thereby giving a limiting action. The mean value of conductance will change,

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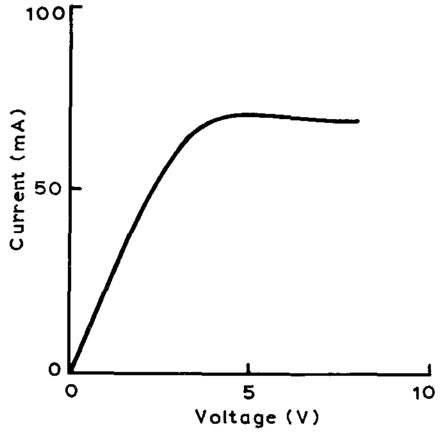

Fig. 1. A dc current-voltage characteristic of a typical Gunn device (CXY 11).

TABLE I

CW Transmission Loss Characteristic of Gunn Device as a Function of Incident RF Power Level

Input Power Level (dB)	Insertion Loss (dB)
−14	0
−12	0.4
−10	1.0
−8	2.0
−6	2.6
−4	3.4
−2	4.0
−1	4.2
0 (850 mW)	4.4

TABLE II

Pulsed Transmission Loss Characteristic of Gunn Device as a Function of Incident Peak RF Power Level

Input Power Level (Peak) (dB)	Insertion Loss (dB)
−20	0
−18	0
−16	+0.1
−14	−0.1
−12	+0.4
−10	−0.2
−9	−0.2
−8	−0.1
−7	+0.6
−6	+1.6
−5	+2.8
−4	+3.8
−3	+4.7
−2	+5.3
−1	+5.8
0	+6.0

due in part to a significant voltage swing over the characteristic. Similarly, the bias situation will change due to the mean (time average) value of dc resistance changing. Both of these effects will result in an absorption of power since the shunt conductance will increase.

III. Experimental Results

In an endeavor to test the principal of this hypothesis the following simple experiment has been performed. A Gunn sample mounted in a pill encapsulation has been placed across a 50-Ω coaxial line and biased to 4.5 V and the transmission characteristic examined with CW power at 6 GHz as the power level was increased. The insertion loss of the system was measured at each power level. Table I shows the results obtained using a CXY 11.

The VSWR at all levels was not large (less than 2.4). It should be noted that the bias value was adjusted to give limiting in the range of power levels shown above. Limiting at higher levels should be obtainable at lower bias levels.

A similar pulse power experiment has been undertaken at 9375 MHz with a pulse length of 10 μs and a PRF of 5 kHz. A bias voltage of 4.6 V was applied.

Table II shows the variation of insertion loss as a function of frequency 0 dB is $1\frac{1}{2}$-W peak and 75-mW mean.

It will be noted that the limiting action occurs at a peak power similar to that of the CW case, suggesting that the limiting action is essentially a peak effect.

IV. Comments

It is clear that there is evidence of limiting action. Since neither the impedance of the line nor the Gunn-effect characteristics have been selected to optimize any limiting effect, it is reasonable to assume that a design system would produce more than $4\frac{1}{2}$ dB of compression.

It is interesting that compression is observed with a Gunn device since unlike other semiconductor limiters it does not rely on the behavior of a junction. It is therefore possible that the behavior under pulse conditions will be attractive, particularly as typical Gunn samples can dissipate substantially more than 1 W of mean power.

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Top-Wall and Branch-Waveguide Hybrids for Millimeter Wavelengths

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***Abstract*—Top-wall and multiple-branch waveguide couplers were developed as hybrid junctions for millimeter wavelengths, and their electrical characteristics were measured. For construction of the 55-GHz top-wall coupler, electroforming techniques were used; for the 94-GHz branch-guide coupler, the branch lines were cut directly into the wall of the main guide. Copper losses were less than 0.2 dB for the two types of couplers.**

I. Introduction

The hybrids discussed herein were developed for radiometer RF circuits operating at 55 and 94 GHz. The physical features, the design and construction details, and the measured performance characteristics are described.

II. Description and Construction of Hybrids

Two forms of hybrids were designed for two frequency ranges in the millimeter wavelength region: 1) a top-wall coupler [1] for the 53.5- to 56.2-GHz frequency range, and 2) a multiple-branch waveguide 3-dB coupler [2] for the 88- to 96-GHz range. The exploded views of the two hybrids are illustrated in Figs. 1 and 2, respectively.

A. Top-Wall Hybrid, 53.5 to 56.2 GHz

The top-wall hybrid is a junction between two parallel lengths of waveguide joined along a common broad-wall surface. Coupling slots introduced in this common broad wall of the waveguide facilitate the

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