A Numerical Observation on the Self-Oscillating Tunnel-Diode Mixer

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***Abstract*—The anomalous behavior previously observed in a self-oscillating tunnel-diode mixer has been analyzed numerically. The theoretical results obtained explain the phenomena of the corresponding conversion loss becoming infinite when the oscillation magnitude of a self-oscillating tunnel-diode mixer is maximum.**

INTRODUCTION

In an experimental self-oscillating tunnel-diode mixer [1], it was reported that the conversion loss became infinite when the magnitude of the self-oscillations was maximum. This pronounced dependency of the gain on the oscillation magnitude was left unexplained. In this short paper the analytical gain expression of a self-oscillating tunnel-diode mixer is used to explain the effect. A simple polynomial representation of the nonlinear tunnel-diode conductance is employed and from this expression the magnitude of the oscillations is evaluated. High-frequency inductive and capacitive effects have been neglected. It is found that the maximum magnitude of the oscillations corresponds to a zero value for the first Fourier coefficient of the time-dependent negative conductance, and as a consequence, zero gain or infinite loss results.

ANALYSIS

A simple but accurate representation of the tunnel-diode conductance curve is given by [2]

$$\frac{dI}{dV} = G(V) = G_{\max} \frac{256(V - V_p)(V - V_v)^3}{27(V_v - V_p)^4} \tag{1}$$

where $G_{\max}$ is a positive quantity indicating the maximum value of the negative conductance, V is the voltage across the diode, and V_p and V_v are the voltages corresponding to the peak and valley currents of the tunnel-diode, respectively. By appropriate choice of V_v, (1) can be made to fit the actual conductance curve of the tunnel diode over the used region of the characteristics. Fig. 1 shows a comparison of the actual conductance curve of the tunnel diode MA 4605 B and that of (1).

When the tunnel diode oscillates, the voltage across it, assuming to be sinusoidal, is given by

$$V = V_B + \frac{V_0}{2}(e^{j\omega_0 t} + e^{-j\omega_0 t}) \tag{2}$$

where V_B is the bias voltage, V_0 is the peak voltage of the fundamental frequency of oscillation, and ω_0 is the angular oscillation frequency. Following Scanlan [3], V_0 is found as

$$V_0^2 = (V_v - V_p)^2 \left[3x(1 - 2x) + \left(28x^4 - 28x^3 + 9x^2 - \frac{27}{32}\bar{g} \right)^{1/2} \right] \tag{3}$$

where $x = (V_v - V_B)/(V_v - V_p)$ and $\bar{g} = G_g/G_{\max}$. G_g is the load conductance presented to the tunnel diode at the oscillation frequency and is approximately the same as that at signal frequency for the self-oscillating tunnel-diode mixer.

For small signals the time-dependent linearized tunnel-diode conductance can be expanded by Fourier series as

$$G(t) = G_0 + \sum_{n=1}^{\infty} G_n(e^{jn\omega_0 t} + e^{-jn\omega_0 t}). \tag{4}$$

After substituting (3) and (2) into (1), the coefficients of the resultant expression are equated to the coefficients of (4). This procedure gives the Fourier coefficients required by the conversion gain expression.

For a short-circuited image tunnel-diode mixer the resonant conversion gain is given by

$$A = \frac{4G_0^2\gamma_1^2 G_L G_g}{[G_0^2\gamma_1^2 - (G_L + G_0)(G_g + G_0)]^2} \tag{5}$$

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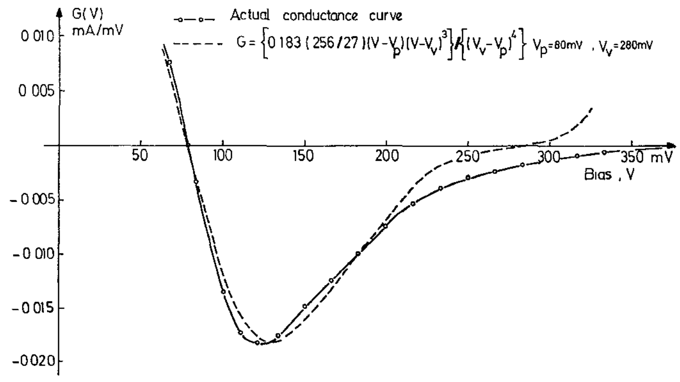

Fig. 1. Measured and computed conductances for the Microwave Associates germanium diode MA4605B.

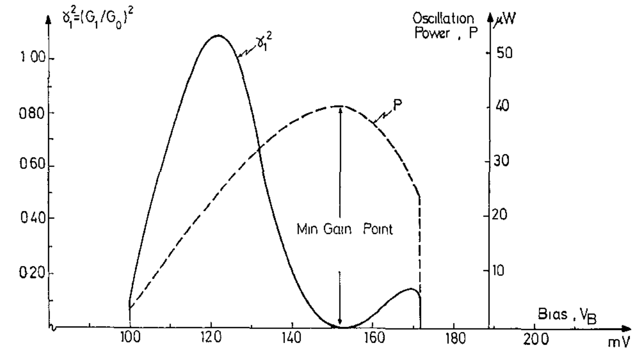

Fig. 2. Variation of the first Fourier coefficient of (4) and the oscillation power of (6), with the applied bias $V_B \cdot G_g = 0.012$ mho, diode: MA4605B.

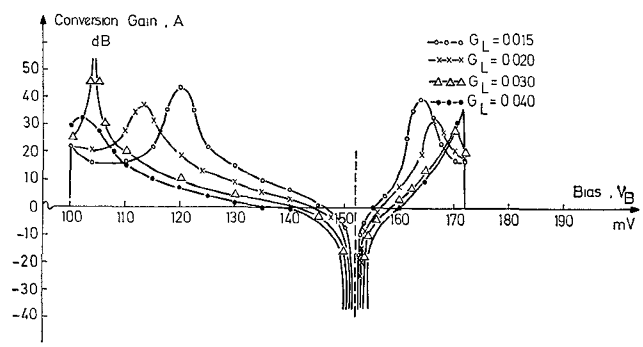

Fig. 3. Theoretical conversion gain curves for different values of $G_L \cdot G_g = 0.012$ mho, diode: MA4605B.

where $\gamma_1 = G_1/G_0$ and G_1 and G_0 are defined by (4). G_L is the load conductance at the intermediate frequency.

A numerical analysis is carried out to obtain γ_1^2 and G_0 and the variation of γ_1^2 is plotted as a function of applied bias voltage V_B, as shown in Fig. 2. In the same figure the oscillation power P dissipated in G_g and obtained from

$$P = \frac{V_0^2}{2} G_g \tag{6}$$

is also shown for comparison purposes. As seen from Fig. 2 the maximum value of P corresponds to the zero value of γ_1^2. At this bias voltage, the gain expression of (5) becomes zero which means infinite conversion loss.

Fig. 3 shows the theoretical conversion gain curves obtained from (5) for different values of G_L and for a fixed value of G_g. There is no need to vary both G_L and G_g since (5) is symmetric in G_L and G_g. Varying G_g would create extra complications in the analysis since it effects the oscillation magnitude through (6) and (3). The form of these theoretical curves is the same as those obtained from the experimental self-oscillating tunnel-diode mixer [1].

Conclusions

The following points are worthwhile considering:

1) The relatively small magnitudes of self-oscillations correspond to large gains. This is in agreement with the results obtained for externally applied local-oscillator tunnel-diode mixers [4].

2) In a tunnel-diode mixer with external local oscillator, both the bias voltage and the local oscillator magnitude can be varied independently. In the case of self-oscillating tunnel-diode mixer, the choice of bias voltage also fixes the oscillation magnitude and hence the infinite conversion loss condition becomes an inherent property of the mixer.

3) By suitable choice of G_L, the critical dependency of gain on the bias voltage can be minimized. However, in such an optimization procedure the noise figure of the mixer should also be considered.

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A New Method for Calculating the Capacitance of a Circular Disk for Microwave Integrated Circuits

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***Abstract*—A method for calculating the capacitance of a circular disk on a dielectric substrate backed by a ground plane is presented. Hankel transforms and Galerkin's method are used to derive the expression for the capacitance. Numerical results are compared with the experimental data and good agreement is reported.**

The increasing use of integrated circuits (IC's) at microwave frequencies has created a great deal of interest in the theoretical and experimental studies of microstrip lines and other similar structures. However, most of these studies are concerned with the properties of infinitely long transmission lines [1]. In actual microwave IC's, many finite-sized or lumped elements are employed to realize the desired functional devices. Hence, the analysis of these finite-sized elements is also important; however, to date, very little has been reported on the analysis of such elements.

Among the finite elements, a rectangular microstrip was recently analyzed by Farrar and Adams [2] and Itoh *et al.* [3]. Another typical finite element is the circular disk (see Fig. 1) for which reliable design data are lacking. In the present short paper a new method is presented for calculating the total capacitance of the circular disk under the quasi-static approximation. The method is an extension of the spectral domain technique developed in [3].

The first step is to write Poisson's equation for the potential ϕ in the cylindrical coordinate

$$\frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial\phi}{\partial r}\right)+\frac{\partial^2\phi}{\partial z^2}=-\frac{1}{\epsilon_0}\rho(r)\delta(z-d) \tag{1}$$

in which ρ is the charge distribution on the disk, and the θ terms vanished because of the circular symmetry. Let us now introduce the Hankel transform of the order zero:

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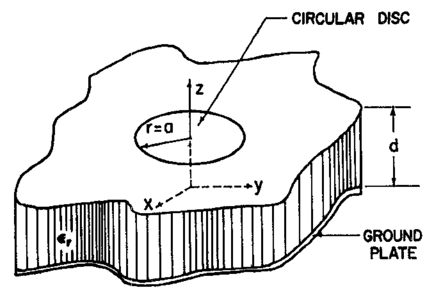

Fig. 1. Circular disk for the microwave integrated circuit.

$$\tilde{\phi}(\alpha,z)=\int_0^\infty \phi(r,z)J_0(\alpha r)r\,dr. \tag{2}$$

Upon Hankel transforming (1) we obtain

$$\left(\frac{d^2}{dz^2}-\alpha^2\right)\tilde{\phi}(\alpha,z)=-\frac{1}{\epsilon_0}\tilde{\rho}(\alpha)\delta(z-d) \tag{3}$$

where

$$\tilde{\rho}(\alpha)=\int_0^a \rho(r)J_0(\alpha r)r\,dr$$

is the Hankel transform of the charge distribution. The general solution of (3) which satisfies the boundary condition $\tilde{\phi}(\alpha,0)=0$ and the radiation condition $\tilde{\phi}(\alpha,+\infty)=0$ is

$$\tilde{\phi}(\alpha,z)=\begin{cases}A(\alpha)\sinh\alpha z, & 0<z<d\\ B(\alpha)\exp\left[-\alpha(z-d)\right], & z>d.\end{cases} \tag{4}$$

The unknown coefficients $A(\alpha)$ and $B(\alpha)$ are determined so that the interface conditions

$$\tilde{\phi}(\alpha,d+0)=\tilde{\phi}(\alpha,d-0)$$

$$\frac{\partial}{\partial z}\tilde{\phi}(\alpha,d+0)-\epsilon_r\frac{\partial}{\partial z}\tilde{\phi}(\alpha,d-0)=-\frac{1}{\epsilon_0}\tilde{\rho}(\alpha)$$

are satisfied. Upon eliminating A and B we obtain

$$\tilde{G}(\alpha)\tilde{\rho}(\alpha)=\tilde{\phi}_i(\alpha,d)+\tilde{\phi}_0(\alpha,d) \tag{5}$$

where

$$\tilde{G}(\alpha)=\frac{1}{\epsilon_0\alpha[1+\epsilon_r\coth\alpha d]}$$

$$\tilde{\phi}_i(\alpha,d)=\int_0^a J_0(\alpha r)r\,dr=\frac{a}{\alpha}J_1(\alpha a)$$

$$\tilde{\phi}_0(\alpha,d)=\int_a^\infty \phi(r,d)J_0(\alpha r)r\,dr.$$

Equation (5) corresponds to the integral equation in the conventional space-domain formulation where the convolution integral appears instead of the product $\tilde{G}\tilde{\rho}$ found in (5). Note that (5) contains two unknowns $\tilde{\rho}$ and $\tilde{\phi}_0$. However, as will be shown shortly, $\tilde{\phi}_0$ is eliminated in the process of solution.

Galerkin's method is now applied to (5). As the first step toward this, $\tilde{\rho}(\alpha)$ is expanded in terms of the known basis functions $\tilde{\rho}_n(\alpha)$:

$$\tilde{\rho}(\alpha)=\sum_{n=1}^N d_n\tilde{\rho}_n(\alpha)$$

$$\tilde{\rho}_n(\alpha)=\int_0^a \rho_n(r)J_0(\alpha r)r\,dr \tag{6}$$

where $\rho_n(r)$, the inverse transforms of $\tilde{\rho}_n(\alpha)$, are chosen so that they are zero for $r>a$. Substituting (6) in (5) and taking an inner product of one of the $\tilde{\rho}_n$ with (5), we have

$$\sum_{n=1}^N K_{mn}d_n=a_m,\qquad m=1,2,\cdots,N \tag{7}$$

$$K_{mn}=\int_0^\infty \tilde{\rho}_m(\alpha)\tilde{G}(\alpha)\tilde{\rho}_n(\alpha)\alpha\,d\alpha \tag{8}$$