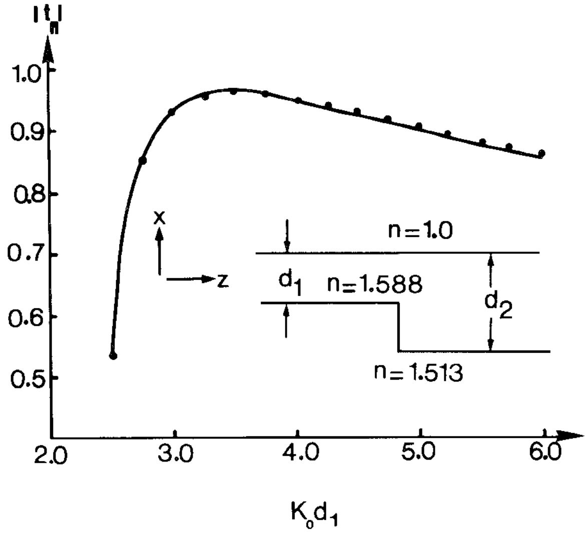

Fig. 2. Asymmetric step discontinuity. —— rigorous mode matching technique; ●●●● BPM.

the same case using the residue calculus, and recently Nishimura *et al.* [10] used an approximate integral equation to calculate the relative radiated power P_r (the ratio of the power lost by radiation to the incident power). Fig. 1 shows the results of the BPM (solid dots) presented in this letter and the three previous techniques. The variation of P_r is shown as a function of $k_0 d_1$ when the ratio d_2/d_1 is kept constant at 0.5; the field incident on the step from the right is the fundamental TE mode. As a second example we consider the asymmetric step studied rigorously by Boyd *et al.* [11] using a discretized representation of the radiation field. Fig. 2 shows the results of this method and those obtained by the BPM (solid dots) in evaluating the magnitude of the transmission coefficient $|t_n|$ as a function of $k_0 d_1$ ($d_1/d_2 = 0.5$ and the wavelength is 0.6328 μm). The comparison is fairly favorable because the relative forward radiated power is much higher than the backscattered radiated power for a step ratio $d_1/d_2 = 0.5$, as pointed out by Marcuse [9], so that the reflected field can be neglected; hence the losses are mainly due to forward scattering.

III. Conclusions

To the author's knowledge, the BPM applied to step discontinuities is presented for the first time in this paper. The applicability of this method to symmetric and asymmetric steps is checked and the results are compared with those of four other methods.

We think that the BPM is efficient for analyzing a wide class of step discontinuities, and perhaps it will be the most convenient method for dealing with step discontinuities between guides of arbitrary refractive index distribution, for example graded or buried waveguides.

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Low-Phase-Noise Gunn Diode Oscillator Design

ROBERT A. STRANGEWAY, MEMBER, IEEE, T. KORYU ISHII, SENIOR MEMBER, IEEE, AND JAMES S. HYDE

Abstract —**Low-phase-noise Gunn diode oscillators with an operating frequency of 35 GHz and an output power of 100 mW are designed, fabricated, and tested. The phase noise is −132 dBc/Hz to −125 dBc/Hz at 100 kHz offset from the center frequency. This low phase noise is obtained by closely coupling the stabilizing transmission cavity resonator and the Gunn diode oscillator coaxial line while loosely coupling the transmission cavity to the output waveguide following van der Heyden's approach.**

I. Introduction

A stable low-phase-noise microwave oscillator is always useful as a signal source for synchronized communications and scientific precision measurements. In the past, various approaches have been tried to reduce the phase noise of various types of microwave oscillators [1]–[10]. According to published references, the phase noise of a center frequency of 35 GHz ranges from −115 to −70 dBc/Hz at an offset frequency of 100 kHz from

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R. A. Strangeway is with the Department of Electrical Engineering and Computer Science, Marquette University, Milwaukee, WI. He is also with the Milwaukee School of Engineering, Milwaukee, WI

T. K. Ishii is with the Department of Electrical Engineering and Computer Science, Marquette University, Milwaukee, WI 53233.

J. S. Hyde is with the Department of Electrical Engineering and Computer Science, Marquette University, Milwaukee, WI. He is also with the Medical College of Wisconsin, Milwaukee, WI.

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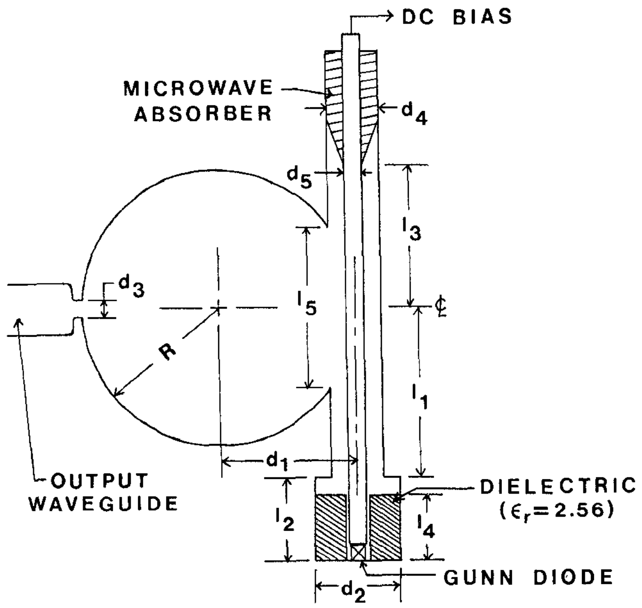

Fig. 1. Schematic diagram of a transmission-cavity-stabilized Gunn diode oscillator:

$d_1 = 6.4$ mm	$l_1 = 11.8$ mm	$l_2 = 1.9$ mm
$d_2 = 3.6$ mm	$l_3 = 12$ mm	$l_4 = 0.8$ mm
	$l_5 = 8.4$ mm	
$d_3 = 2.1$ mm	$d_4 = 3.2$ mm	$d_5 = 1.6$ mm.

Quiescent cavity length = 7.6 mm. $R = 6.35$ mm.

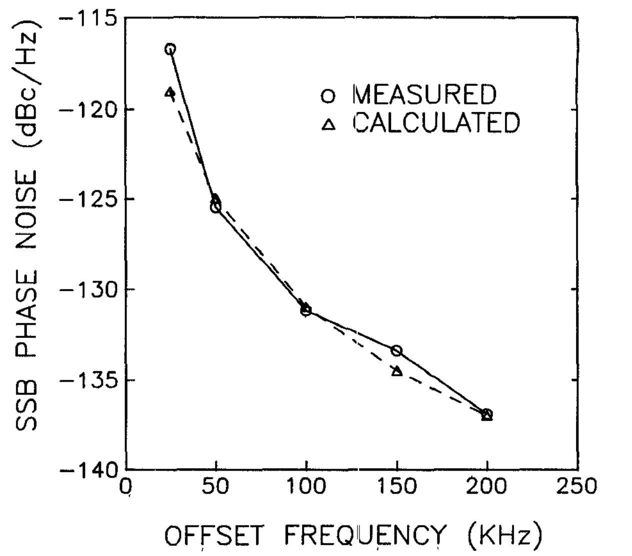

Fig. 2. Single sideband phase noise as a function of offset frequency from the carrier of the oscillator.

the center frequency with an output power of the order of 20 dBm [1], [4]–[6], [10].

In this paper a design of a low-phase-noise Gunn diode oscillator operating at 35 GHz with an output power of 20 dBm and lower phase noise than previously published is presented. The oscillator is stabilized by a transmission cavity in van der Heyden form [11].

II. Structure of the Oscillator

A schematic diagram of the transmission cavity stabilized Gunn diode oscillator circuit in the van der Heyden configuration [11] is shown in Fig. 1. The circuit is designed to operate at 35 GHz. As seen from Fig. 1, a Gunn diode MA49837 is mounted at the lower end of a short-circuited coaxial line. The upper end of the coaxial line is terminated by a tapered microwave absorbing material, Eccosorb. A dc bias voltage for the Gunn diode is fed through the center conductor of the coaxial line from the top end of the coaxial line. To this coaxial line, a TE_{011} mode cylindrical cavity resonator is coupled by an elliptical opening of 8.4 mm × 3.2 mm, as seen in Fig. 1. The axis of the cylindrical cavity resonator is perpendicular to the axis of the coaxial line. As seen from this figure, a large coupling window exists along the intersection of the cavity wall and the outer conductor wall of the coaxial line. This structure ensures a strong coupling between the Gunn diode and the stabilizing cavity [11].

At a location diametrically opposite to the coupling window, a tiny coupling hole 2.06 mm in diameter is drilled. This hole size is experimentally determined to obtain just enough desirable output power to the WR-28 rectangular waveguide across the mechanical tuning range, which is 20 dBm. The cavity resonator size is determined by the resonance frequency of 35 GHz in the TE_{011} mode.

For precise tuning of the stabilizing cavity resonator, one of the end plates of the cavity resonator is controlled both manually and piezoelectrically by Piezoelectric Disc 6080-5. The other end plate of the cavity resonator is controlled by Unimorph Disc 61065.

III. Oscillator Performance

The oscillator performance factors to be inspected include the power output, the mechanical and electrical frequency tuning ranges, the temperature dependency of center frequency, and the phase noise results.

The power output over the mechanical frequency tuning range from 34.5 GHz to 35.5 GHz is more than 20 dBm, with a maximum power of 21.5 dBm at 35 GHz.

The electromechanical tuning range using the piezoelectric Unimorph is approximately 120 MHz for 0 to 250 V_{dc} applied, but it varies from Unimorph to Unimorph. The temperature coefficient of the oscillation frequency of this oscillator is approximately −600 kHz/°C. The electronic tuning of this oscillator by the dc bias voltage applied to the Gunn diode is found to be approximately 1.1 kHz/mV.

Two experimental units were fabricated for phase noise tests. The phase noise performance was −125.6 dBc/Hz for unit 1 and −131.0 dBc/Hz for unit 2, both at a 100 kHz offset from the carrier. Unit 2's peak power output was 96 mW at the time of this test while unit 1's was 135 mW. Also, passive coupling tests show that unit 2 is significantly undercoupled compared to unit 1. Hence, unit 2 has a higher Q_{ext} and a lower phase noise is expected. Also, the units contain different Gunn diodes which may have significantly different noise figures.

The frequency offset from the carrier was changed and the phase noise of unit 2 was remeasured to verify the consistency of the phase noise results. Then the phase noise was calculated using the expected dependence of $1/\Delta f^2$ [12] under the assumption that the phase noise at a 100 kHz offset is the value that was experimentally obtained.

The calculated results are shown in Fig. 2 with the measured results. The phase noise was measured using an HP8566A spectrum analyzer. The phase noise measurement at this offset has been made numerous times with different video amplifiers as a further check and the results were the same. For comparison [1]–[10], the measured value of the loaded Q of this resonator was 1750 and the noise figure of the Gunn diode used is considered to be in the range of 25 to 35 dB [13].

IV. Remarks and Discussion

The repeatability of the observed phase noise is, in the worst case, ± 1.5 dB, based on the numerous measurements taken to date. At 25 kHz offset, the larger measured phase noise may be due to the $1/f$ noise effects. It was too difficult to measure any closer to the carrier because of the narrow frequency spans required to measure the calibration sidebands and the simultaneous locking of the microwave carrier at the spectrum analyzer.

The measured phase noise of the present 35 GHz klystron was -103.0 dBc/Hz at 100 kHz offset. The phase noise test of unit 2 was repeated under the same conditions of microwave power and environmental conditions as those that the klystron was tested under, and its phase noise was consistent with previous measurements. Thus, unit 1 is 22 dB better than the klystron and unit 2 is 27 dB better.

The amount of phase noise shown in Fig. 2 should be compared to the previously published values of the phase noise of various oscillators of 35 GHz with comparable output power in the past [1]–[10]. The phase noise of a MIC YIG tuned GaAs FET was -70 dBc/Hz [1]; a cavity-stabilized phase-locked IMPATT oscillator produced -107 dBc/Hz [2], a cavity-stabilized phase-locked Gunn oscillator -98 dBc/Hz [3]; a phase-locked oscillator -110 dBc/Hz [4]; a phase-locked IMPATT sweep oscillator -106 dBc/Hz [5]; a cavity-stabilized Gunn oscillator -115 dBc/Hz [6]; a free-running IMPATT oscillator -92 dBc/Hz [7]; a cavity-stabilized phase-locked IMPATT oscillator -81 dBc/Hz [8]; a phase-locked klystron -89 dBc/Hz [9]; and a klystron -101 dBc/Hz [10]. Thus the oscillator designed for this research has smaller phase noise than previously published oscillators.

V. Conclusions

Both calculations and measurements show that the transmission-cavity-stabilized 35 GHz 100 mW Gunn diode oscillator designed and fabricated as described in this paper has a low phase noise of $-125 \sim -132$ dBc/Hz. This low phase noise was accomplished by making the coupling between the terminated coaxial line Gunn diode oscillator and the stabilizing cavity resonator as large as possible and the coupling to the output as small as practical for sufficient power output.

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