A Quasioptically Stabilized Resonant-Tunneling-Diode Oscillator for the Millimeter- and Submillimeter-Wave Regions

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Abstract—**A semiconfocal open-cavity resonator has been used to stabilize a resonant-tunneling-diode waveguide oscillator at frequencies near 100 GHz. The high quality factor of the open cavity resulted in a linewidth of approximately 10 kHz at 10 dB below the peak, which is about 100 times narrower than the linewidth of an unstabilized waveguide oscillator. This technique is well suited for resonant-tunneling-diode oscillators in the submillimeter-wave region.**

I. Introduction

THE OSCILLATION frequency of the double-barrier resonant-tunneling diode (RTD) has recently been extended up to 712 GHz [1], which makes it the fastest solid-state electronic oscillator demonstrated to date at room temperature. A major challenge in operating solid-state oscillators at frequencies above 100 GHz is the design of the resonator. Conventional resonators, such as those based on closed cavities or radial transmission lines, exhibit an unloaded quality factor Q_u that decreases with increasing frequency because of increases in the ohmic losses of metallic surfaces. Open resonators, such as those used in lasers, provide a much higher Q_u but are difficult to integrate with lumped-element, solid-state oscillators. We have combined a waveguide RTD oscillator and a high-Q_u semiconfocal cavity to form a quasioptical oscillator operating at frequencies near 100 GHz.

At present, the primary application of the RTD oscillator is a low-noise local oscillator (LO) for high-sensitivity radiometers operating in the submillimeter-wave region ($f \gtrsim 300$ GHz). In this application the linewidth must be less than about 100 kHz and the oscillator should be frequency tunable by at least $\pm 1\%$ of the nominal center frequency. The quasioptical oscillator demonstrated here provides the required narrow linewidth, and can be easily scaled down in size for operation at higher frequencies.

II. Quasioptical Oscillator Design

The schematic diagram of our quasioptical oscillator designed for the 100 GHz region is shown in Fig. 1. The RTD is mounted in a standard-height WR-6 (0.065 × 0.0325 in) rectangular waveguide that opens abruptly to a round 0.075-in-diameter coupling hole within the middle of a flat metallic plate. This plate forms one reflector of a semiconfocal open resonator. The TEM_{00N} modes of this resonator have a Gaussian transverse intensity profile with $1/e$-point loci as shown in Fig. 1. The spot diameter of these modes at the flat reflector is designed to be larger than the diameter of the coupling hole. This makes the coupling between the waveguide and the open cavity fairly weak, which is necessary to realize a large loaded quality factor Q_l for the open resonator.

In our first implementation of the open resonator, the flat mirror was an aluminum plate, and the spherical mirror was made from stainless steel and had a radius of curvature of 3.0 cm. The length of the cavity, D, was mechanically adjustable about the semiconfocal value of 1.5 cm. The upper limit of Q_l for this cavity is the unloaded quality factor Q_u. This can be estimated by assuming that the only power loss suffered by the Gaussian beam is the ohmic loss in the stainless steel mirror, which is much greater than the loss in the aluminum mirror. For a resonator in which the loss of one mirror dominates, a useful expression is $Q_u = D/\delta$ [2], where δ is the skin depth given by $\delta = (\rho/\mu\pi f)^{1/2}$ (MKSA units) ρ is the resistivity, μ is the permeability, and f is the frequency. Taking $D = 1.5$ cm, $f = 103$ GHz, and $\rho = 72 \times 10^{-6}$ Ω-cm (the dc resistivity of stainless steel [3]), we find $\delta = 1.3$ μm and $Q_u = 1.2 \times 10^4$.

The waveguide portion of the oscillator, shown in cross section in Fig. 2(a), is similar to the circuit used in all of our previous RTD oscillators operating above 100 GHz. The diode is dc biased by a coaxial circuit that suppresses spurious oscillations by means of a very lossy section of transmission line placed in close proximity to the diode chip. The lossy material is an iron-loaded epoxy. The equivalent circuit is given in Fig. 2(b). The active region of the RTD is represented by a large-signal conductance G in series with an inductance L_{QW}, both in parallel with a capacitance C_D. The capacitance is attributed primarily to the depletion region of the RTD. The inductance is

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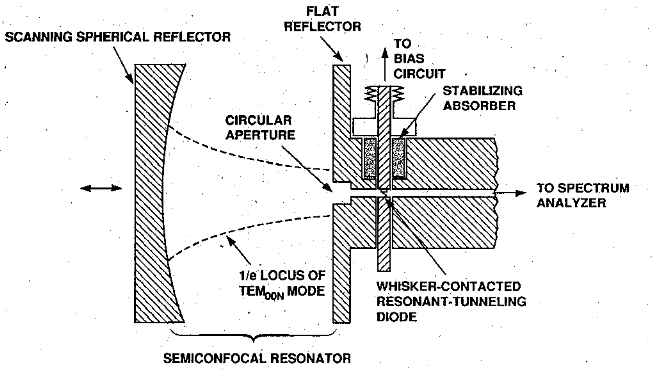


Fig. 1. Schematic cross-sectional diagram of quasioptical resonant-tunneling-diode oscillator designed to operate in the 100 GHz region.

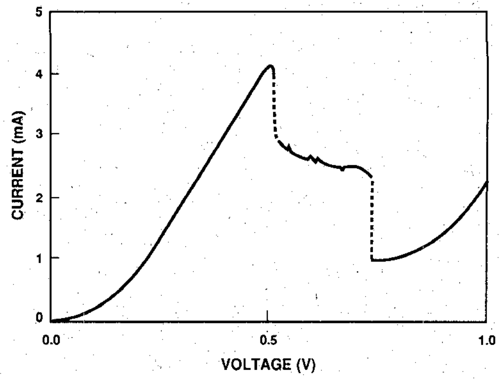


Fig. 3. Current-voltage curve of an $In_{0.53}Ga_{0.47}As/AlAs$ RTD mounted in the WR-6 waveguide at room temperature.

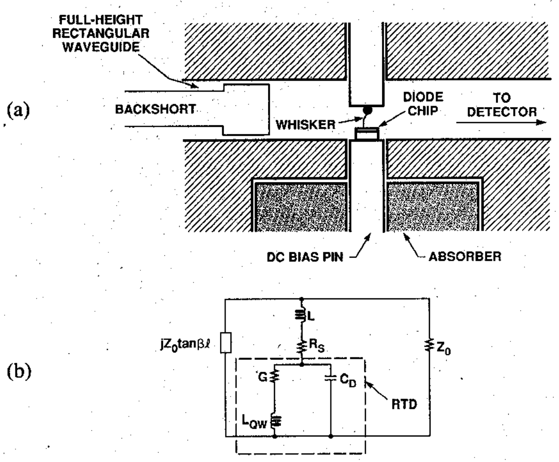


Fig. 2. (a) Cross-sectional diagram of the waveguide part of the quasioptical oscillator with a tunable backshort replacing the open cavity. (b) Equivalent circuit of the waveguide oscillator.

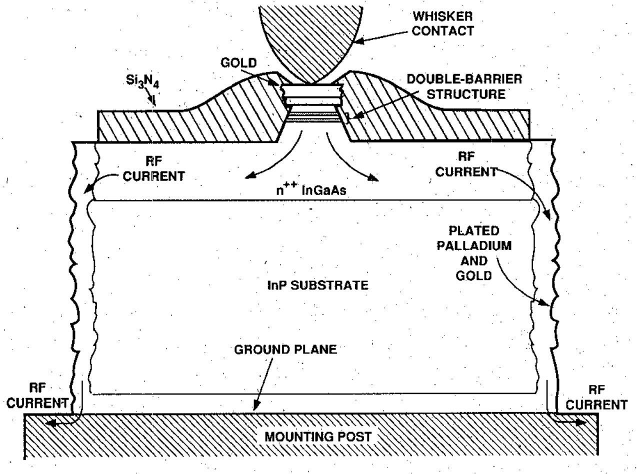


Fig. 4. Cross section of RTD chip showing Si_3N_4 hole for whisker contact. The figure also shows the path followed by the RF current between the top contact and the ground plane.

attributed to the time delay of resonant tunneling [4], but is not important in the present experiment because this time delay is much less than the period of an oscillator operating near 100 GHz. The elements representing the active region are in series with a parasitic resistance R_S that is composed of a number of ohmic-loss mechanisms in the RTD chip. Each chip contains several mesa-geometry RTD's, one of which is contacted by a whisker. The whisker is mounted on a post that penetrates somewhat into the waveguide, and the electrical effect of the whisker is represented by the inductance L. The effect of the backshort is represented by the reactive element $jZ_0 \tan \beta l$, where Z_0 is the characteristic impedance, β is the propagation constant for the fundamental TE_{10} mode, and l is the separation between the backshort and the whisker. The parallel combination of Z_0 and $jZ_0 \tan \beta_l$ yields a series impedance $Z_L = Z_0 \sin^2 \beta l + j(Z_0/2) \sin 2\beta l$. The elements Z_L, C_D, and L form a low-Q series resonance that supports oscillations when G is sufficiently negative.

III. High-Speed Resonant-Tunneling Diode

The RTD used in the present experiment was made from the InGaAs/AlAs materials system. It has a theoretical maximum oscillation frequency of 900 GHz, and has previously demonstrated the most powerful oscillations we have observed to date above 100 GHz [5]. It consists of two 1.4-nm-thick undoped AlAs barriers separated by a 5.5-nm-thick undoped $In_{0.53}Ga_{0.47}As$ quantum well. It was grown by molecular beam epitaxy on an n^+ InP substrate. The room-temperature current-voltage (I–V) curve of a diode having an active area of 4 μm^2 is shown in Fig. 3. It has a peak-to-valley current ratio of approximately 4.2, and a peak current density of 1.0×10^5 A cm^{-2}. The plateau in the I–V curve between the peak and valley voltages is the region of oscillation. The discontinuities (shown in Fig. 3 as dashed lines) connecting the plateau to the peak and valley points are a result of self-rectification of the oscillation by the RTD.

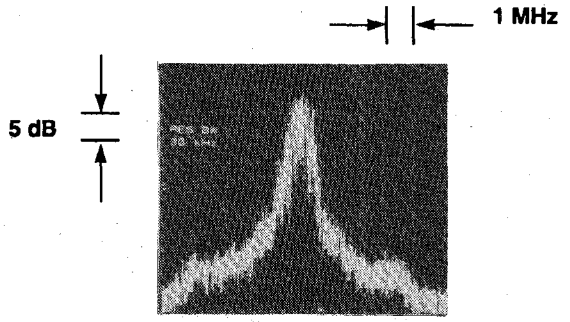
In RTD's intended for stable high frequency oscillators, it is important to achieve a robust whisker contact and a low-resistance current path from the active region of the device to the ground plane. The robust contact is obtained by entrapping the whisker in a hole in a Si_3N_4 layer which covers the RTD, as shown in Fig. 4. The Si_3N_4 layer is deposited after the RTD mesas are fabricated on the wafer. The holes are defined by photolithography and reactive-ion etching. An additional benefit of the Si_3N_4 layer is that it acts to passivate the $In_{0.53}Ga_{0.47}As$ sidewalls of the RTD mesa [6]. The low-resistance current path is achieved in the manner shown in Fig. 4. Between the RTD mesa and the edges of the chip, the RF current flows in a heavily doped n^{++} epitaxial layer grown on top of the InP substrate. The ac conductivity of this layer is roughly 10 times that of the substrate. From the edge of the chip down to the post (ground plane), the current flows through a palladium/gold layer plated on the sidewalls of the chip. The overall series resistance from the double-barrier structure to the bottom of the chip is approximately 0.5 Ω at dc, increasing to about 1.0 Ω at 600 GHz.

IV. Experimental Results

The output radiation of the quasioptical oscillator propagates down the waveguide to a Schottky-diode mixer where it is down converted to the frequency range of a microwave spectrum analyzer. Experimental power spectra are shown in Fig. 5 for an RTD oscillator operating near 103 GHz. The broad spectrum in Fig. 5(a) results from inserting a backshort into the waveguide at the open end. In this case, the oscillator operates with the waveguide resonator shown in Fig. 2. The width of the spectrum 10 dB below the peak is approximately 1 MHz, which is unsuitable for most applications.

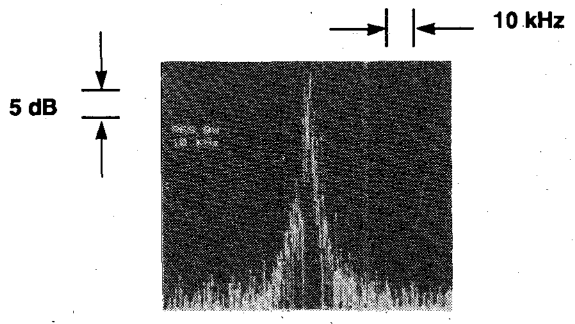
Upon removing the short and exposing the open cavity, the spectrum shifts and becomes much narrower. The expansion of this spectrum, shown in Fig. 5(b), yields a linewidth of approximately 10 kHz at 10 dB below the peak. This is 100 times narrower than the waveguide-oscillator linewidth. The center frequency of the cavity resonance is determined, as in all open resonators, by the spatial separation of the reflectors. By varying this separation, we were able to tune the stabilized power spectrum over a range of about 0.3 GHz at a fixed RTD bias voltage. A greater tuning range of approximately 5 GHz was obtained by varying the bias voltage. With each change of bias, the open cavity had to be adjusted to establish a new oscillation frequency.

The average power of the waveguide and quasioptical oscillators corresponding to the spectra in Fig. 5(a) and (b) was found to be approximately −17 and −19 dBm, respectively. The lower power of the quasioptical resonator may be a result of some diffraction loss to free space by the semiconfocal cavity. Because the RTD is a negative resistance oscillator, its theoretical maximum power P_{max} can be estimated directly from the I-V curve. A useful estimate, first derived for p-n Esaki tunnel diodes [7], is $P_{max} = (3/16)\,\Delta I \cdot \Delta V$, where ΔI and ΔV are the differences between peak and valley currents and between valley and peak voltages, respectively. For the present diode, we find $\Delta I = 3.1$ mA and $\Delta V = 0.22$ V, so that $P_{max} = -9$ dBm. The discrepancy between this and our measured powers is typical for the best RTD oscillators operating near 100 GHz.



WAVEGUIDE OSCILLATOR

(a)



QUASIOPTICAL OSCILLATOR

(b)

Fig. 5. (a) Power spectrum of the waveguide oscillator without the semiconfocal open cavity. (b) Power spectrum of the quasioptical oscillator with the semiconfocal cavity.

V. Analysis

The nonzero linewidth of an RTD oscillator is attributed to phase fluctuations caused by noise processes in the RTD. In this case, one expects that the linewidth should depend directly on the RTD noise power and reciprocally on the loaded quality factor. The dc bias conditions were the same for both spectra in Fig. 5, and hence the intrinsic RTD noise characteristics should have been the same for both cases. The factor of 100 difference in linewidth is therefore attributed to a difference in the Q_l of the two oscillators.

The Q_l of the quasioptical oscillator is estimated by assuming that it is approximately equal to the Q_l of the semiconfocal cavity coupled to the waveguide. Thus we can apply the general expression $Q_l^{-1} = Q_u^{-1} + Q_e^{-1}$, where Q_e is the (external) quality factor for the reactive part of the open cavity plus the load circuit. An expression applicable to the semiconfocal resonator is $Q_e = 4\pi f_0 D/Tv_g$ [8], where f_0 is the resonant frequency, v_g is the group velocity of the radiation in the cavity, and T is the net coupling transmissivity. We estimate T by assuming that the power coupled from the open cavity to the waveguide is given by the ratio of the power contained within the area (πR^2) of the coupling hole to the total power in the TEM_{00N} Gaussian mode. This leads to $T = 1.0 - \exp(-2R^2/\omega_0^2)$, where ω_0 is the characteristic width of the

Gaussian beam at the flat mirror.[†] For our semiconfocal cavity, the expression $\omega_0 = (\lambda D/\pi)^{1/2}$ applies [9], which yields $\omega_0 = 0.4$ cm. From this we find $T \cong 0.1$, which results in $Q_e = 1.3 \times 10^3$ for our open cavity. Combining this with the value of Q_u derived in Section II yields $Q_l \cong 1.2 \times 10^3$.

The Q_l of the waveguide oscillator is estimated from the equivalent circuit of Fig. 2. Experience has shown that this circuit oscillates with greatest power when Im $[Z_L]$ is inductive and Re $[Z_L]$ is fairly small, i.e., when $n\pi \lesssim \beta l \lesssim (n + 1/4)\pi$, where n is an integer. In this case the quality factor (for a series resonance) is $Q_l \cong X_c/(\text{Re }[Z_L] + R_s)$, where $X_c = (2\pi f_0 C)^{-1}$ is the capacitive reactance at resonance. In the experiment reported here, we used a 4 μm^2 RTD having $R_s \cong 12\ \Omega$ and $C \cong 5$ fF, and the backshort was adjusted so that Re $[Z_L] \sim 60\ \Omega$. Thus we estimate $Q_1 \sim 5$ for the waveguide oscillator. The ratio of the theoretical loaded quality factors of the two oscillators is 240, which is of the same order as the inverse ratio of the linewidths.

VI. INCREASED FREQUENCY AND POWER

The quasioptical oscillator is ideally suited to operate far into the submillimeter-wave region where the RTD is currently the only solid-state source that operates at room temperature. The unloaded quality factor of the resonator will certainly increase with frequency as the skin depth in the mirrors decreases. Values of Q_u over 10^6 are routinely available from semiconfocal resonators in the infrared region of the spectrum.

One could obtain significantly more power and maintain the advantage of quasioptical stabilization by properly implementing an array of RTD oscillators rather than the single-element oscillator demonstrated here. In principle, one could configure many waveguide-mounted RTD's with a single semiconfocal resonator. However, a more practical approach for very high frequencies is a planar RTD array based on microstrip-circuit techniques. The key point in either approach is that the oscillators all lie in an equiphase plane of the open-cavity mode and are thus synchronized by the high-Q_l resonance. This method of power combining has been used to obtain CW power levels up to 0.32 W from both planar FET oscillator arrays and Gunn diode oscillator arrays operating near 10 GHz [10]. It should be a useful technique for obtaining milliwatt levels of power from RTD oscillators in the submillimeter-wave region. A prototype planar RTD quasioptical oscillator having a single element has recently been demonstrated near 10 GHz [11].

VII. SUMMARY

A semiconfocal resonator has been used to stabilize a waveguide-based RTD oscillator at a frequency of 103 GHz. The stabilized oscillator linewidth was approximately 10 kHz, which is about 100 times narrower than the linewidth of the waveguide oscillator alone. The quasioptical RTD oscillator should be suitable as a local oscillator for superconducting tunnel-junction mixers up to frequencies of at least 400 GHz.

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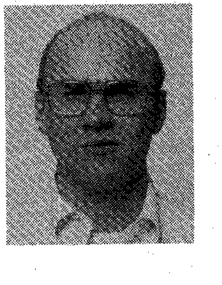

Elliott R. Brown obtained the Bachelor of Science degree (Summa Cum Laude) in physics from U.C.L.A. in 1979. In 1985 he was granted the Ph.D. in applied physics from the California Institute of Technology. His thesis concerned the application of cyclotron resonance in InSb for heterodyne conversion in the submillimeter-wavelength region.

In 1985 Dr. Brown joined the research staff at MIT Lincoln Laboratory to work in the area of millimeter and submillimeter-wave quantum elec-

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[†] The quantity ω_0 is the distance from the center of the Gaussian beam to the point where the electric field is down by $1/e$.

tronics. His research has focused on resonant-tunneling devices and quantum-well intersubband optical devices.

Christopher D. Parker was born in Boston, MA, on September 17, 1931. He attended the University of Maine, Orono, majoring in engineering physics.

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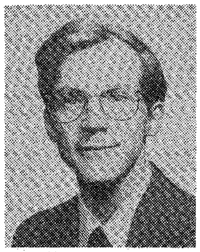

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In 1977, he joined Motorola, Inc. From 1979 to 1981, he was with Scientific-Atlanta, where he engaged in research and development pertaining to cable television systems. In September 1983, he joined the faculty of the University of Massachusetts at Amherst, where he is presently Associate Professor of Electrical Engineering. His current research interests include the application of quasi-optical techniques to millimeter-wave circuits and subsystems.