Current Saturation in Submillimeter Wave Varactors

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Abstract—**In semiconductor devices the speed of electrons cannot exceed certain limits. This phenomenon will affect varactor multipliers as well as other high frequency devices where the RF current through the active part of the device is primarily displacement current. Hence, we expect at some point "saturation" of the varactor output power. We will, in this paper, discuss this phenomenon in some detail and show that it severely deteriorates the multiplier performance at higher frequencies. Single barrier varactors (SBV) should have an advantage over GaAs Schottky diode varactors because they can be fabricated on InAs and stacked in a series array, allowing for lower current densities and higher power handling.**

I. Introduction

COMPUTER analysis can be used to accurately predict the performance of millimeter and submillimeter wave Schottky barrier diode mixers and frequency multipliers over a wide range of operating conditions [1], [2]. However, these analyses depend on an accurate equivalent circuit for the nonlinear element over the range of RF drive levels and frequencies expected in actual operation. The commonly used varactor diode equivalent circuit consists of a parallel combination of voltage dependent capacitance and conductance in series with a diode resistance (Fig. 1(a)). The nonlinear capacitance is due to the voltage variable width of the depletion layer and the nonlinear conductance is due to the I-V curve of the diode. Both can be found from an analytic approximation or from measured data. The resistance term is more difficult to obtain from experimental measurements. It can be approximated from information on the device undepleted layer width, the spreading resistance, and the contact resistance (see e.g. [3], [4]). The resistance may be constant or voltage variable. The best multiplier efficiency theoretically occurs when the multiplication is purely reactive. If the diode is driven into forward conduction, multiplication becomes a hybrid of resistive and reactive multiplication which degrades the efficiency. Theoretical performance derived using this equivalent circuit agrees well with experimental performance of harmonic multipliers at lower frequencies or power levels. However, as the frequency and power level increase disagreement between theory and experiment increase. Erickson [5] has reported experimental efficiencies in harmonic multipliers that match the theoretical results at low pump power levels, but not at higher power levels. The experimental degradation of efficiency occurs well before the diode is driven into forward conduction. He suggested that the disagreement might be caused by device heating at the higher power levels which was not included in the device model. The purpose of this paper is to analyze limitations on varactor performance and then to compare the modified models with the experimental results.

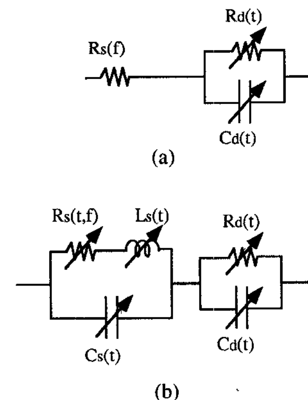

Fig. 1. Schottky diode varactor models with: (a) constant series resistance; (b) series resistance, a parallel displacement current capacitance both changing with the width of the undepleted epitaxial layer, and added inductance L_s related to the inertia of the electrons.

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II. Basic Limitations

In the most common model for the Schottky diode varactor, the *displacement current* (i_d) through the depleted region is

$$i_d(t) = C_d(t) \cdot \frac{dV_d(t)}{dt} \tag{1}$$

where $C_d(t)$ is the depletion capacitance and $V_d(t)$ is the voltage over the depletion region. Hence we expect i_d to increase with the pump power and the pump frequency. In the varactor diode, the displacement current must be matched by the *electron conduction current* (i_e) *through the undepleted semiconductor*, where i_e is

$$i_e = A_d \cdot n_e \cdot v_e(t) \cdot e. \tag{2}$$

A_d is the diode area, n_e the electron density (in our case

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$n_e = N_d$, the doping density), v_e the electron velocity and e the charge of the electron. Most millimeter wave and submillimeter wave varactor multipliers reported to date use GaAs Schottky diodes. Since the electron velocity in GaAs reaches a maximum of about $2.2 \cdot 10^5$ m/s at about 3.2 kV/cm [6] applied dc field, a saturation phenomenon is expected when $i_d > i_{sat}$ where

$$i_{sat} = i_{e,max} = N_d \cdot v_{e,max} \cdot e \cdot A_d. \tag{3}$$

This current limiting phenomenon can be modeled as an *effective series resistance* which increases rapidly with higher power levels, causing considerable deterioration in the multiplier performance.

Experimental evidence can be found by studying the 2 × 80 GHz multiplier reported by Erickson [5]. This high efficiency doubler uses two diodes (6P4 from the University of Virginia) in a balanced configuration. Erickson found that the shape of the measured efficiency versus power curve did not follow the theoretical predictions. We confirmed this in an independent calculation using the Siegel and Kerr program [1] as shown in Fig. 2. The maximum electron current, i_{sat}, as calculated from (3) for this particular diode (6P4; see Table I) is 44 mA. The displacement current, i_d, equals this electron current, i_e, for a pump power of 11 mW (per diode). Furthermore, with this current through the 11 Ω series resistance the field over the 1 μm epitaxial layer will be greater than 3.2 kV/cm.

The maximum pump power used experimentally by Erickson was 120 mW or 60 mW per diode; hence the maximum displacement current, i_d, exceeds i_{sat} by a large margin. We consequently decided to investigate how the efficiency of the 2 × 80 GHz multiplier varies with increasing series resistance using a large signal analysis. The graphed calculation in Fig. 2 shows the calculated efficiency versus input power parameterized by increasing series resistance for this multiplier. In addition, it shows the experimental results obtained by Erickson. We have also scaled these results by 1.3 dB, the estimated loss in the output waveguide of the multiplier mount (losses of the input circuit have been neglected). Optimum values of the input and output embedding impedances and bias voltages were determined by the computer program since the experimental values were not known. Poor fit between theory and experiments at low input powers may result from this. We conclude that a series resistance that increases from about 10 Ω at low pump powers to about 30 Ω at high pump powers can explain the measured efficiency degradation for input powers above about 20 mW (10 mW per diode). Above input powers of about 40 mW the diode is driven with voltages beyond the break-down voltage. This will cause further losses not included in our theoretical analysis (i.e., there will be an even larger series resistance) and some increase in noise [5].

Several other phenomena may play a role in explaining the observed fall off in efficiency at higher pump powers. One is that *the edge of the depletion region cannot move faster than allowed by the maximum electron velocity*,

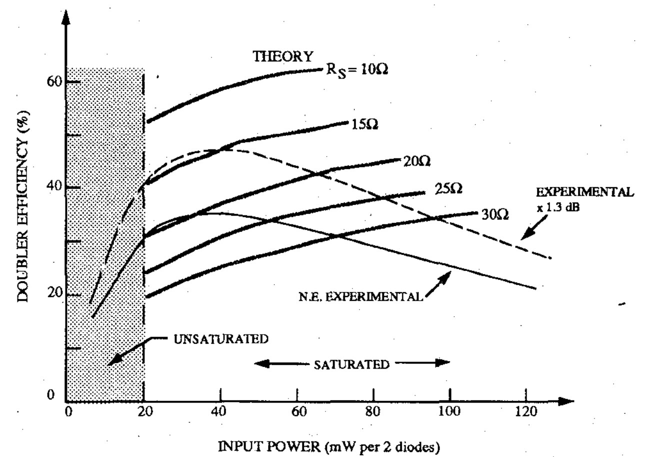

Fig. 2. Measure efficiency (full line: measured; dashed line: ohmic losses of 1.3 dB subtracted) for the 2 × 80 GHz multiplier [5] compared to theoretical efficiencies obtained for different series resistances. The shadowed area indicates where theory suggests the multiplier is unaffected from current saturation.

i.e., $dW(t)/dt < v_{e,max}$ where W is the width of the depletion region. Hence the capacitance, $C(t) \sim 1/W(t)$, may not vary as rapidly as calculated by the multiplier large signal analysis program. Investigating $dW(t)/dt$ for the Erickson doubler shows that $dW(t)/dt > v_{e,max}$ for input powers >40 mW (20 mW per diode). Hence, for this multiplier we will assume that this phenomenon is less important than the saturation of the current. However, for higher frequency multipliers the limited speed of the capacitance variation may be a more serious concern. *The acceleration of the electrons is finite* and introduces an inductance, L_s, as shown in Fig. 1(b) (see e.g. [7], [8]). An approximate calculation shows that $\omega L_s << R_s$ for pump frequencies below about 100 GHz:

$$\omega L_{epi}(t) = R_{epi}(t) \cdot \frac{\omega}{\omega_{es}} \approx R_{epi}(t) \cdot \frac{f_{GHz}}{f_{es}}. \tag{4}$$

The epilayer electron scattering frequency, $f_{es} = 1/(2\pi m^* \mu)$ (where μ is the low field mobility) is of the order 800 GHz at the low fields, high doping levels and room temperatures found in the experimental multipliers. The scattering frequency increases with increasing field.

Another factor may be *the displacement current through the undepleted epitaxial layer* (represented by Fig. 1(b) by $C_s(t)$). We discuss this briefly later. Below pump frequencies of about 100 GHz it has little impact. However, for higher pump frequencies it must be taken into account.

Heating of the diode seems not to be important for the diodes investigated. The maximum heating of the diode was theoretically calculated assuming that the thermal resistance was $1/4\pi\kappa T$, where κ is the thermal conductivity of GaAs (see [6]). An upper limit on the effect of heating for the Erikson multiplier can be derived as follows. We assume a maximum of 48 mW (80% of 60 mW for one diode despite mount losses and power delivered to the external circuit at the 3rd, 4th, etc. harmonic) pump power is absorbed in the epilayer of the 6P4 diode. This leads to the conclusion that the heating of the epilayer is less than

27 K and that the series resistance increases (mobility decreases [6]) by less than 5%. In the case of another diode discussed below, 2T2 (see Table I), the maximum power absorbed by the diode is less than 20 W, yielding a temperature increase less than for the 6P4 diode. The saturation velocity will drop by about 10% [6] for a 30 K temperature increase. Hence, the temperature effect is much smaller than the effect due to the electron velocity upper limit.

The *spreading resistance* in the heavily doped substrate can be ignored. For a typical varactor, the epitaxial layer region is 5000 to 10 000 Å. Since the epitaxial layer resistivity is much larger than the bulk resistivity, and since a large fraction of the epitaxial layer is undepleted during most of the pump cycle, R_{epi} will be the dominating contribution to the series resistance.

The time variation of the width of the undepleted epitaxial layer that causes $R_s(t)$, $L_s(t)$ and $C_s(t)$ to be time varying, does not change the general picture. Raisanen *et al.* [9] have investigated the influence of the time variation of $R_s(t)$, $L_s(t)$ and $C_s(t)$. They found a considerably increased theoretical efficiency at low drive levels, while for a large pump power the theoretical efficiency is only slightly higher than assuming a constant series resistance ($R_s = R_{epimax}$) and neglecting the influence from C_s and L_s. This can be understood, since the mean of the time varying series resistance is lower than the experimentally determined series resistance, which is measured under heavily forward bias conditions. For high pump powers our computer simulations shows that the maximum current through the diode is obtained for near zero and moderately small negative voltages when most of the epitaxial layer is still undepleted, indicating that the effective series resistance then becomes nearly equal to the experimentally determined one.

III. Analysis Using a Drift Diffusion Model Approach

To gain more insight into the current saturation problem, a large signal time dependent version of a drift diffusion-based device model simulation has been modified to assess the large signal equivalent circuit of varactors. The simulation solves Poisson's equation and the current continuity equation self-consistently in the time domain to find the total current (the sum of the electron current and the displacement current) as a function of time through the structure. The simulation is driven with an RF voltage and the resulting current waveform is found. The in-phase Fourier component of the current at the drive frequency is associated with the resistance, and the out-of-phase component is associated with the reactance. The device large signal capacitance can be found from the reactance and the frequency. The RF mobility of electrons in GaAs as a function of frequency and drive level used in this simulation were determined using a large signal Monte Carlo calculation of the epitaxial region following the technique described in [10] modified to include impurity

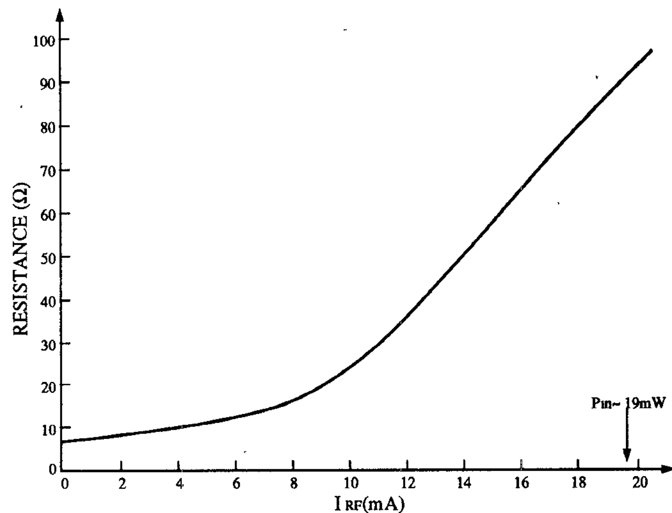

Fig. 3. The large signal series resistance at 200 GHz for the 2T2 diode as obtained from the drift-diffusion model.

scattering. The inductance, L_s, related to the inertia of the electrons is neglected. This simulation has been used to study a variety of varactor operating conditions. The varactor structure used in the simulation is that of the submillimeter wave varactor, 2T2, from the University of Virginia. It has a 0.5 micron long epitaxial layer with a doping of $1 \cdot 10^{17}$ cm^{-3}.

Fig. 3 shows the large signal resistance of the complete device pumped with a pure sinusoidal voltage at 200 GHz as a function of RF current. For the device under consideration, the time-dependent voltage across the device is $V(t) = V_{DC} + V_{RF} \sin(\omega_p t)$. The resulting current is also a function of time. The frequency domain representation of the current can be found by a Fourier transform and will have both sin and cos components at the pump frequency and its harmonics. The large signal resistance is defined to be $R_{RF} = V_{RF}/I_{RF}$, where I_{RF} is the magnitude of the sin-component (in phase with the voltage) of the current at the pump frequency. The cos magnitude of the current is associated with the reactive part of the impedance at the pump frequency. We have chosen the bias voltage $V_{DC} = V_{RF}$. These conditions correspond to pumping the varactor with the higher harmonic frequencies shorted. A more complete analysis could include the effect of voltages at higher frequencies or a combination of the varactor model and a nonlinear circuit analysis program. Similar effects have been studied in IMPATT diodes [7], [8] and references at end of [10].

For combinations of RF drive level and frequencies that produce a small current in the epitaxial region, the mobility is approximately equal to the low field mobility, and the device resistance is small. As the RF drive level increases, the current through the device increases, the voltages and fields in the undepleted region increase, the mobility goes down and the resistance increases. As can be seen in Fig. 3, the resistance has increased many times near the saturation current. The major contribution to the increase is due to the decrease of the mobility at high

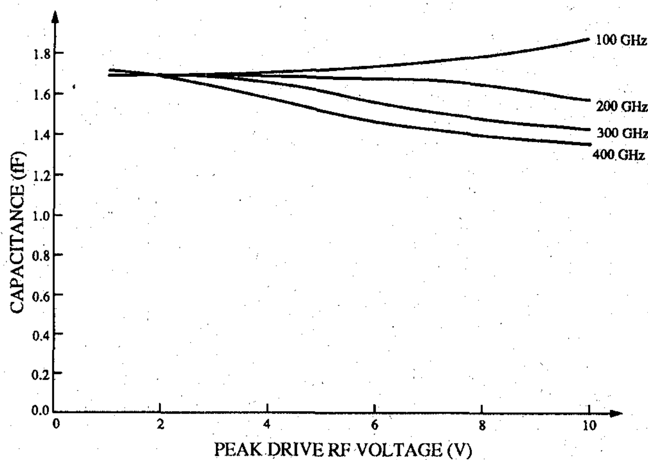

Fig. 4. Device capacitance versus drive voltage as obtained from the drift diffusion model for different frequencies.

fields, while a negligibly small part of this device resistance is related to power loss at harmonic frequencies within the diode, in particular in the series resistance (the sinusoidal drive voltage ensures that no harmonic power is lost in the external circuit). The result in Fig. 3 shows a major limitation on the combination of RF pump power and frequency in varactor multipliers.

The varactor equivalent circuit must be further modified at higher frequencies. Consider the varactor equivalent circuit shown in Fig. 1(b) where the capacitance C_s shunting the epitaxial resistance accounts for the displacement current through the undepleted part of the epitaxial layer. At "low" frequencies and small drive powers, the impedance of this extra capacitance is usually much larger than the resistor impedance. Most of the RF current will flow through the resistance and the effect of this capacitance will be small. However, conditions will change as the drive level or frequency increase. The parallel combination of the resistor and capacitor will act like a current divider. Increasing the RF drive at a constant frequency will increase the impedance of the resistor as shown in Fig. 3. This will shunt a larger fraction of the current through the capacitance. If the undepleted epitaxial layer capacitance were to completely shunt the resistance, the device would appear to be a series combination of two capacitors, the net capacitance would go down and become constant. This phenomenon is illustrated in Fig. 4. Notice that this saturation effect would also degrade the harmonic multiplier performance since the multiplier performance depends on the nonlinear nature of the device capacitance.

The results of the diffusion model and Monte Carlo simulation of the multiplier diode show several limitations not included in the simple equivalent circuit of Fig. 1(a). The problems include (1) large signal current limits in the undepleted epitaxial region which can occur at all frequencies, (2) large signal shunting of the undepleted region which can reduce the device nonlinearity and occurs at frequencies of several hundred GHz, and (3) inductive electron effects which can be ignored for most varactor applications but will have a major impact on very high frequency multipliers.

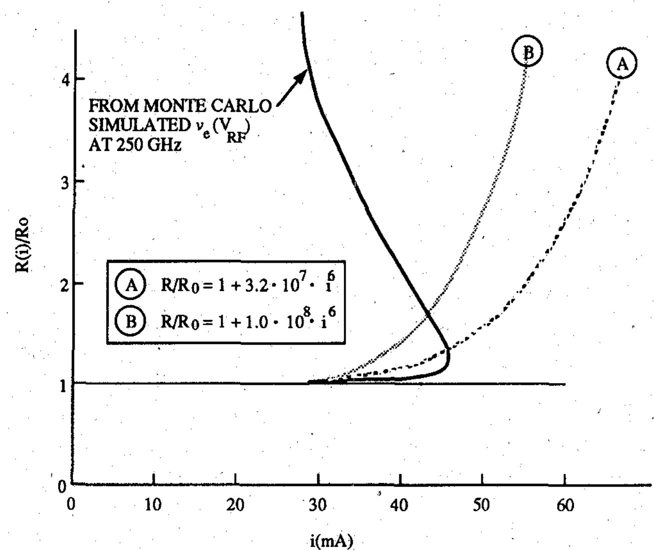

Fig. 5. Resistance versus drive current for the 6P4 varactor obtained from Monte Carlo simulations and Eq. (4), and resistance versus drive current according to Eq. (5). (A) and (B) are used for calculation of efficiencies as described in Fig. 6.

IV. Approximate Approach Using a Current and Time Dependent Series Resistance

From the discussion above, it is obvious that the series resistance increases at high drive levels. Hence, we suggest that multiplier performance can be analyzed approximately using a current-dependent series resistance. In order to get a reasonable closed form current dependence, we have studied results published on electron velocities derived using Monte Carlo technique for bulk GaAs driven by a sinusoidal electric field $E = E_0 \cos(\omega t)$ [10]. Since it is possible to translate the electron velocity (v) into a current density ($j = vN_D e$), and since the field may be translated into a voltage ($V = L \cdot E$, L is the device length), it is possible to derive a high resistance viz:

$$R_s(E) = \frac{i_0(E)}{i_0^2(E) + i_i^2(E)} E \cdot L \tag{5}$$

where i_0 is the current component in-phase (proportional to the in-phase velocity component) and i_1 the current component out-of-phase with the field E. The device length, L and the field, E, is eliminated taking the ratio $R_s(E)/R_s(E = 0)$. In Fig. 5 is plotted $R_s(E)/R_s(E = 0)$ versus the current $(i_0^2 + i_1^2)^{1/2}$ derived using the area and doping concentration of the 6P4 diode. This plot shows a two-valued curve for high drive levels. However, the importance of this curve is *not* that it can be used for predicting an exact current dependence of the series resistance, *but* can be used for a reasonable estimation. Notice that the series resistance derived using the full diode model (Fig. 3) is single valued everywhere. Arguing that a major aim of this paper is to demonstrate the influence of the velocity saturation on the performance of multipliers and

TABLE I
DIODE DATA

	2 × 80 GHz f_p = 80 GHz 6P4	3 × 160 GHz f_p = 160 GHz 2T2	3 × 330 GHz f_p = 330 GHz 2T2	3 × 330 GHz f_p = 330 GHz SBV
N_d nominal (cm^{-3})	$3.0 \cdot 10^{16}$	$1.0 \cdot 10^{17}$	$1.0 \cdot 10^{17}$	$1.7 \cdot 10^{17}$
N_d used by us (cm^{-3})	$3.5 \cdot 10^{16}$	$1.0 \cdot 10^{17}$	$1.0 \cdot 10^{17}$	$1.7 \cdot 10^{17}$
t_{epi} (μm)	1.0	0.5	0.5	
R_{so} (Ω) measured	10	14[1], 12[2]	14[1], 12[2]	12
C_0 (fF) measured	21	5.5[1], 6.5[2]	5.5[1], 6.5[2]	20
$V_{\text{break-down}}$ measured (V)	20	11[1], 8.5[2]	11[1], 8.5[2]	
Area (μm^2) Nominal/Adjusted[3]	33/33	5/6[3]	5/6[3]	3.5
Assumed v_{max} (m/s)	$2.4 \cdot 10^5$	$2.4 \cdot 10^5$	$2.4 \cdot 10^5$	$2.4 \cdot 10^5$
$i_{sat} = AN_d e v_{max}$ (mA)	44	23	23	23

[1]Univ. Va.;
[2]From Ref. [2];
[3]Adjusted according to measured C_0. See the text.

since it was indeed only possible to modify the Siegel and Kerr multiplier program including a single valid current dependence, we decided to use a current-dependent series resistance defined accordingly:

$$R_s(i) = R_{so}(1 + a \cdot i_{mA}^6) \tag{6}$$

With proper choice of the parameter "a" and the exponent (6 in this case), the desired dependencies are obtained. The choice of the functional form of (6) is somewhat arbitrary, but it does give two desired effects:

i) The apparent series resistance at the pump frequency and the harmonics increases with the pump power or current.
ii) The current waveform is modified (clipped) as required.

In Fig. 5 the resistance is depicted for two values of the parameter "a" to be used for multiplier performance simulation in the next paragraph.

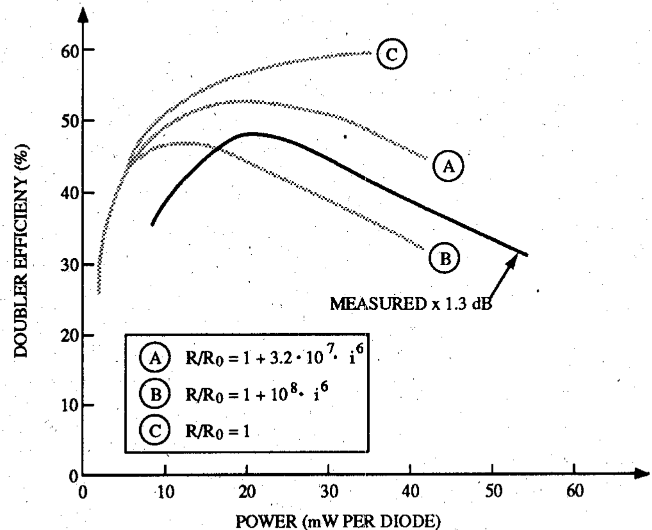

Fig. 6. Intrinsic efficiency versus pump power (see Fig. 2) for the Erickson 2 × 80 GHz multiplier compared to calculated efficiency using the current dependent series resistance (A) and (B) as suggested in Fig. 5.

V. RESULTS

The efficiency of several multipliers were investigated theoretically using a large signal analysis computer program with and without the current-dependent series resistance. These included the 2 × 80 GHz multiplier and a 3 × 160 GHz multiplier that could be compared to the experimental results reported by Erickson. In addition, two 3 × 330 GHz multipliers were studied, one using a high frequency GaAs Schottky barrier varactor and the other using a SBV. Table I summarizes the diode parameters used in the large signal analysis. The 6P4 millimeter wave varactor diode and the 2T2 submillimeter wave varactor diode, both from the University of Virginia, were used in the 2 × 80 GHz and 3 × 160 GHz multipliers developed by Erickson. Also given are the characteristics for a GaAs/AlGaAs SBV with a 200 Å effective barrier width.

2 × 80 GHz Multiplier

The maximum measured efficiency of the 2 × 80 GHz multiplier was 34%. Since the losses in the mount were about 1.3 dB, the experimental intrinsic conversion efficiency is about 46%. For the 2 × 80 GHz multiplier, R_{so} was chosen equal to the measured series resistance of 10 Ω. We selected two values for "a" in (6), as indicated in Figs. 5 and 6. In estimating i_{sat} we have assumed the doping concentration to be $3.5 \cdot 10^{16}$ cm^{-3} rather than the nominal doping concentration of $3.0 \cdot 10^{16}$ cm^{-3}. A best fit to the experimental curve is obtained by using a value between the two values for "a" used for the current-dependent resistance curves shown in Fig. 5. The results are summarized in Table II.

The relation, $i_{RF} < i_{sat}$, is violated for input powers as low as of 11 mW. The depletion edge velocity exceeds the maximum electron velocity at larger powers. For 40

TABLE II
CALCULATED PERFORMANCE OF THE ERICKSON 2 × 80 GHz AND 3 × 160 GHz MULTIPLIERS

	P_{in} (mW)	i_{RF} (mA)	dW/dt 10^5 (m/s)	η	R_s (Ω)	a
	11	42.0	2.0	52	10	—
2 × 80 GHz	44	69.0	2.8	54	10	—
$i_{sat} \approx 44$ mA	11	41.5	1.7	46	$R(i)$	$3.2 \cdot 10^7$
	40	62.7	2.6	41	$R(i)$	$3.2 \cdot 10^7$
	2	23.7	≈ 1.7	22	12	—
3 × 160 GHz	20	47.0	3.5	>40	12	—
$i_{sat} = 23$ mA	2	23.0	1.6	20	$R(i)$	$1.53 \cdot 10^9$
	20	33.0	2.8	9	$R(i)$	$1.53 \cdot 10^9$

P_{in} is the power absorbed by the varactor, and η the intrinsic efficiency.

TABLE III
CALCULATED PERFORMANCE OF 990 GHz SCHOTTKY BARRIER DIODE VARACTOR (2*T*2) MULTIPLIER

	P_{in} (mW)	i_{RF} (mA)	dW/dt 10^5 (m/s)	η	R_s (Ω)	a
	2	24	2.3	4.1	12	—
3 × 330 GHz	10	53	4.6	11.6	12	—
$i_{sat} \approx 23$ mA	2	23	2.0	2.3	$R(i)$	$1.53 \cdot 10^9$
	10	31	2.9	0.6	$R(i)$	$1.53 \cdot 10^9$

mW input power per diode (60 mW/diode was used as a maximum in Erickson's experiment) the $i_{RF} >> i_{sat}$ and $dW/dt > v_{max}$. Due to these effects, the calculated efficiency with the current-dependent series resistance is about 25% lower than when it is neglected, consistent with the measured results. The comparison between the theoretical calculations and the experimental results over the full range of pump power is shown in Fig. 6. The agreement shown here gives us some confidence in evaluating the 3 × 160 GHz multiplier using the same technique.

3 × 160 GHz Multiplier

The 3 × 160 GHz multiplier was measured to have a maximum output power of 0.7 mW for an input power of about 25 mW. We assume losses of about 5 dB for the 3 × 160 GHz multiplier by scaling the 1.3 dB waveguide loss for the 2 × 80 GHz multiplier. If the input losses are 2 dB and the output losses 3 dB, for 25 mW input power, about 15 mW reaches the diode, and the intrinsic output power is 1.4 mW. Hence, the experimental intrinsic multiplier efficiency is 9.3%. From Table II it is seen that i_{RF} is greater than i_{sat} for input powers as low as 2 mW. The impact on the performance of this higher frequency multiplier due to the current saturation is much more severe than for the 2 × 80 GHz multiplier. For the analyses, a lower value of the parameter "a" was used, scaled by the area and the doping concentration of the 2T2 diode. The area is actually more uncertain for this small area diode than for the 6P4 diode. We determined an "effective area" by comparing the capacitances as measured by Erickson [5] of the 6P4 and 2T2 diodes, and scaling from the area of the 6P4 diode; $A(2T2) = A(6P4) \cdot \{C(2T2)/C(6P4)\} \cdot \{N_d(6P4)/N_d(2T2)\}^{1/2}$. The efficiency calculated including the saturation effect is a factor of almost five lower than the efficiency calculated for the constant series resistance case ($R_s = 12$ Ω). The predicted efficiency of 9% for large pump power coincides with the measured one. Interestingly, the reverse bias voltage for optimum performance of this tripler was near the breakdown voltage ($V_{bias} = 5$–6 V, $V_{break\text{-}down} = 8.5$ V), corresponding to a negative voltage swing considerably larger than the break-down voltage.

3 × 330 GHz Multiplier

A theoretical case of a 3 × 330 GHz multiplier was also analyzed. Results are shown in Table III. Again, the saturation severely deteriorates the efficiency: at large input powers, the efficiency drops from almost 12% to 0.6%, a factor of 20! In fact, assuming these calculations are valid, it may not be possible to obtain more output power than 60 μW at 1 THz using the 2T2 diode. If no deterioration due to current saturation were present, the maximum predicted output power would be 1.2 mW.

The Single Barrier Varactor (SBV)

The single barrier varactor is an MBE-grown mesa diode, with ohmic contacts. It has been shown experimentally that this diode has a considerable potential as a varactor diode for multiplier applications [10]. In GaAs, an AlGaAs barrier in the middle of the mesa blocks all conduction current. The diode exhibits a symmetrical C-V characteristic, which causes only odd harmonics to be created when the varactor is pumped (zero bias assumed), similar to a back-to-back configuration of two Schottky varactors. Since this diode relies on a voltage variable depletion region to generate the nonlinear capacitance, it will have exactly the same problem with current saturation in a multiplier application as the Schottky varactor diode.

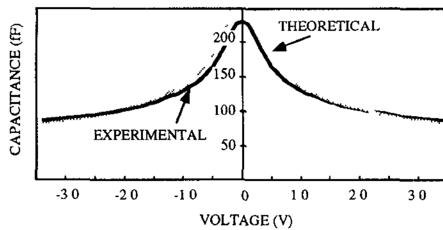

Fig. 7. Capacitance versus bias voltage for a single barrier varactor.

TABLE IV
PREDICTED PERFORMANCE OF A SINGLE BARRIER VARACTOR 3 × 330 GHz MULTIPLIER

P_{in} (mW)	i_{RF} (mA)	dW/dt 10^5 (m/s)	η	R_s (Ω)	a
2	18	3.6	7.2	12	—
10	39	6.6	17.9	12	—
2	17	3.6	6.9	$R(i)$	$1.53 \cdot 10^9$
10	31	5.1	6.8	$R(i)$	$1.53 \cdot 10^9$

We have measured the C-V characteristic of such a diode (see Fig. 7). We found that the zero bias capacitance deviated from the capacitance expected from the simple calculation, $C_{max} = (A_d \epsilon)/W_{barrier}$, by about a factor of three. An interpretation of this phenomenon is that the effective barrier thickness is larger than $W_{barrier}$. The lower zero bias capacitance indicates an effective barrier thickness of 600 Å rather than 200 Å. An obvious possible contribution to this excess barrier thickness stems from the fact that there are 50 Å undoped regions on both sides of the barrier. In addition, the deviation arises from the discontinuity of the Fermi level at the doped/undoped barrier interface.

A few large signal analysis simulations of a SBV diode multiplier have been carried out, as summarized in Table IV. For the particular case shown in Table IV, it is assumed that $N_d = 1.7 \cdot 10^{17} \text{cm}^{-3}$, $W_{barrier} = 200$ Å (this is now an *effective* width), $C_0 = 20$ fF, $R_s = 12\ \Omega$. This hypothetical SBV was designed to have the same $i_{sat} = 23$ mA as the 2P2 diode, yet it suffers less from saturation effects. The efficiencies without and with the current-dependent series resistance are 17.9% and 6.8%, respectively. This is only a factor of 2.6 reduction as compared to a factor of 20 reduction for the 2T2 varactor. The SBV has a 10–20% larger dynamic cutoff frequency, but this is not large enough to explain the difference. One possible explanation may arise from the device symmetry. Since the SBV exhibits a symmetrical C-V curve, it generates no currents at the idler frequency (2nd harmonic) and hence suffers no resistive losses at the 2nd harmonic.

A limitation of these devices as they are currently fabricated is that they start conducting at a lower voltage than designed based on the barrier height and thickness. Multiplier large signal analyses show that when the diode is pumped to the extent that the magnitude of the resistive current (as determined from the I-V curve) becomes approximately equal to the magnitude of the capacitive current, the efficiency of the multiplier is deteriorated. This limitation along with that arising from the current saturation may be overcome by using a number of diodes in series. It should be possible to make an epitaxial layer with a stack of e.g. four diodes in series. If each is 7500 Å thick, a stack of four results in an epitaxial layer of 3 μm. Such a diode offers several advantages:

i) The area can be made n times larger for a given impedance, where n is the number of stacked barriers.
ii) The influence of the ohmic contact will become n times less important due to the larger area.
iii) The impact of the current saturation and the depletion edge speed will be considerably relaxed since for constant power the current density goes down with the number of series diodes (area), leading to higher efficiency.
iv) The lower current density or drive voltage per series diode will allow using low bandgap materials such as InAs, yielding still lower series resistance and higher cut-off frequencies.

VI. MIXERS

Most likely GaAs Schottky barrier diode mixers will also experience current saturation. An analysis of a 600 GHz mixer using a 1 μm diameter diode with a driving power of 1 mW shows diode currents that exceed $i_{e,max}$ by a factor of 6 (private communication with Dr. I. Mehdi).

VII. SUGGESTIONS FOR FUTURE RESEARCH

We have identified a serious limitation of semiconductor multipliers using a voltage variable depletion region to generate reactive multiplication when operating at high frequencies and/or high input power levels. This arises from the saturation of electron velocity in the semiconductor. Whether a more exact theory using a more precise device model will predict lower or higher efficiency is not known. There are several approximations involved in our calculations, as mentioned above. There are also inconsistencies such as that the velocity of the depletion edge calculated in two different ways do not agree, viz.:

$$v = \frac{A\epsilon}{\Delta t} \cdot \left(\frac{1}{C(t)} - \frac{1}{C(t + \Delta t)} \right) \neq \frac{i_d(t)}{N_d e A} \tag{7}$$

where $i_d(t)$ is the current through the varying capacitance. Typically the maximum velocity determined from the depletion current $i_d(t)$ is the larger one and the maxima do not occur simultaneously. It is, however, quite clear that the current dependence of the series resistance is very important and will deteriorate the performance of the multiplier. For 1 THz, diodes with higher doping than 2T2 and correspondingly thinner epitaxial layer (the breakdown voltage will be lower, and the corresponding epitaxial layer thickness smaller) will suffer less from saturation, and are expected to be more efficient.

To date most varactors have been fabricated from GaAs which exhibits a saturation velocity of about $2\text{–}3 \cdot 10^5$ m/s. Other semiconductors, such as InAs have higher saturation velocities and may provide a more optimum material for the varactors. However, other parameters affect varactor performance as well. In particular, the breakdown voltage of the device is critical. More studies are needed to determine the relative tradeoff of these effects in the GaAs and the InAs materials systems. In addition, different varactor architectures, which reduce the impact of the current saturation may be available, such as forming a series stack of single barrier varactors.

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