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 to ordinary 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.

1. 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. 1a). 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. ref. [3,4]). The resistance can be constant or voltage variable, depending on the nature of the solution. The best multiplier efficiency theoretically occurs when the multiplication is purely reactive. If the diode is

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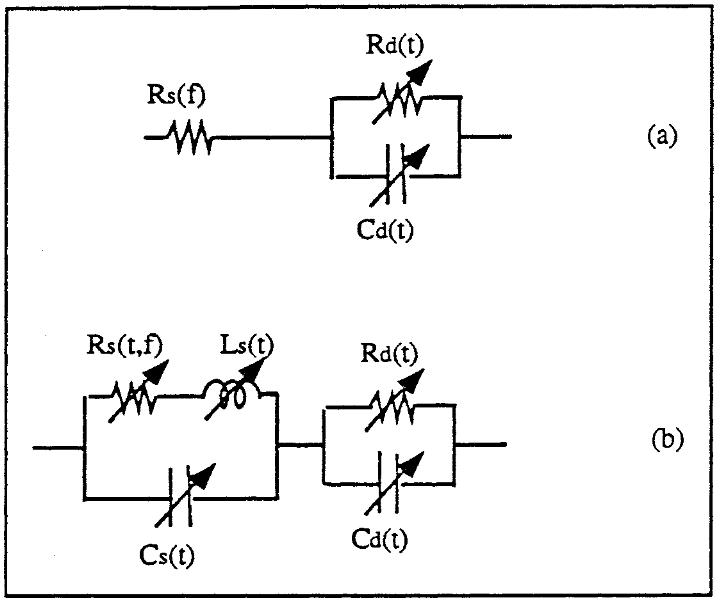

Figure 1: Schottky diode varactor models with: (a) constant series resistance; (b) series resistance, a parallel displacement current capacitance both changing with the width of the undepleted epitaxial layer, and added inductance L_s related to the inertia of the electrons.

driven into forward conduction, multiplication becomes a hybrid of resistive and reactive multiplication which degrades the efficiency. This equivalent circuit approach has worked well in the design 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 efficiencies in harmonic multipliers that match the computer simulation 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 problem might be caused by device heating at the higher power levels. The purpose of this paper is to analyze limitation on varactor performance and then to compare the modified models with the experimental results.

2. Basic Limitations.

In the most common model for the Schottky diode varactor (see Fig. 1a), the *displacement current* (i_d) through the depleted region is

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

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 model of Fig. 1a 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 n_e=N_d, the doping density), v_e the electron velocity and e the charge of the electron. Basically all millimeter wave and submillimeter wave varactor multipliers use GaAs Schottky diodes. Since it is well known that in a DC electric field, the electron velocity in GaAs reaches a maximum of about $2.2 \cdot 10^5$ m/s at about 3.2 kV/cm [6], 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 modelled as an *effective series resistance* which increases rapidly with higher power levels, causing a considerable deterioration in the multiplier performance.

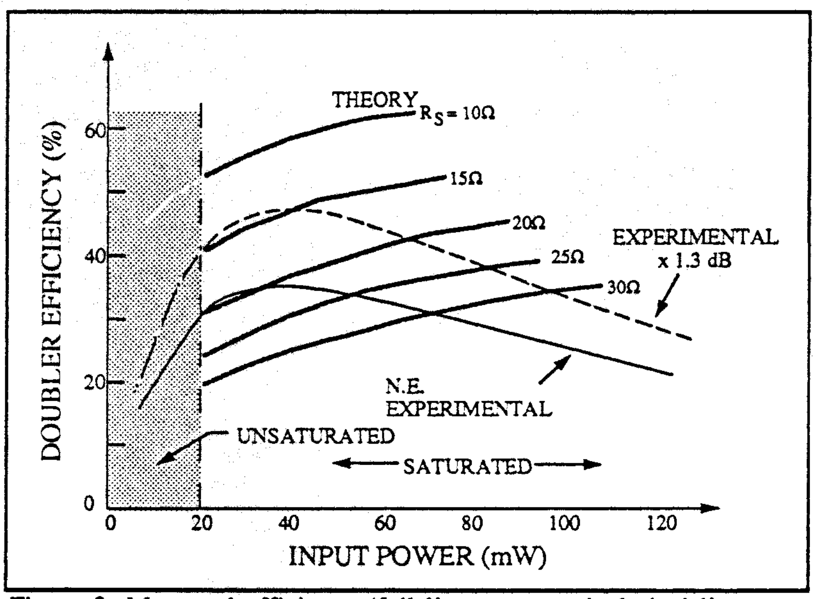

Figure 2: Measured efficiency (full line: measured; dashed line: ohmic losses of 1.3 dB subtracted) for the 2x80 GHz multiplier [5] compared to theoretical efficiencies obtained for different series resistances. The shadowed area indicates where theory suggests the multiplier is unaffected from current saturation.

Experimental evidence can be found by studying the 2x80 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 fall off of the measured efficiency versus power 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 Eq. (3) for this particular diode (6P4; see Table 1) 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} a large margin (compare Table 2). We consequently decided to investigate how the efficiency of the 2x80 GHz multiplier varies with increasing series resistance. The graph shown in Fig. 2 shows calculation of the efficiency versus input power with an increasing series resistance as a parameter. In addition it shows the experimentally obtained results as obtained by Erickson. We have also scaled these results by 1.3 dB, the estimated loss in the waveguide multiplier mount. We conclude that a series resistance that increases from 11 Ω a 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).

In addition, several other phenomena may play a role in explaining the 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*, i. e. $dW(t)/dt < v_{s,max}$ where W is the width of the depletion region. Hence the capacitance variation $C(t) \sim 1/W(t)$ may not vary as rapidly as calculated by the multiplier simulation computer program. Investigating dW(t)/dt for the Erickson doubler, shows that $dW(t)/dt > 3 \cdot 10^5$ m/s 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. 1b (see e.g. ref. [7,8]). An approximate calculation shows that $\omega L_s \ll R_s$ for pump frequencies below about 100 GHz (f_{es} is about 800 GHz).

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

Further *the displacement current through the undepleted epitaxial layer* (represented in Fig. 1b by $C_{Rs}(t)$) is not of any major concern at pump frequencies of about 100 GHz and below. However, for higher frequencies it must be taken into account.

Heating of the diode seems not to be of any particular importance for the diodes investigated. Estimates of the heating suggests that when each 6P4 diode absorbs 48 mW, the heating will not exceed 30 °C, increasing the series resistance (decreasing the mobility [6]) by less than 5%. For another diode discussed below, 2T2 (see Table 1), the maximum power absorbed by the diode is of the order 20 mW yielding a temperature increase less than that for the 6P4 diode. The saturation velocity in the 2T2 diode will drop with about 10 % [6] for a 30 °C temperature increase.

Basically, the *spreading resistance* in the heavily doped substrate can be ignored. For a typical varactor, the epitaxial layer region is 5000 to 10000 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 most of the series resistance. In fact, when the varactor is driven at about the maximum allowable power, R_{epi} is time-varying between a maximum value (R_{epimax}) valid when the $W_{depl}\approx 0$, and a minimum value approximately equal to zero.

The time variation of the width of the undepleted epitaxial layer causing $R_s(t)$, $L_s(t)$ and $C_{Rs}(t)$ to be time varying, also 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 $C_{Rs}(t)$. 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 ($R_s \approx R_{epimax}$) and neglecting the influence from C_{Rs} 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 during 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.

3. Analysis Using a Drift Diffusion Model Approach.

To gain more insight into the current saturation problem, a large signal time dependent version of a device simulation has been modified to assess the large signal equivalent circuit of varactors. The simulation solves Poison'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 90 degree phase component is associated with the reactance. The device large signal capacitance can be found from the reactance and the frequency. A simple approximation for the velocity that includes a constant low field mobility region and a saturated high field velocity is used. 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.

A wide variety of dynamic or velocity overshoot effects are possible in GaAs. These effects can modify the simple current transport model discussed above. A large signal Monte Carlo simulation of the epitaxial region using a technique described in [10] was modified to include impurity scattering. This simulation was then used to find the RF mobility of electrons in GaAs as a function of frequency and drive level. The results of this simulation were used to set the mobility and saturated velocity in the diffusion model. The varactor structure used in the Monte Carlo 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. Similar effects has been studied in IMPATT diodes [7, 8 and references at end of 10]. For such combinations of RF drive level and frequencies that produce a small current where the fields and voltages in the epitaxial region are small, 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. In fact, since the device is being driven by a sinusoidal wave, the current through the device is a non-sinusoidal function of time. The resistance shown in Fig. 3 is an average resistance over the 200 GHz RF pump cycle. Although part of this device resistance is related to power loss at harmonic frequencies in the series resistance, the major contribution is due to the decreased mobility. The result in Fig. 3 shows a major limitation on the combination of RF pump power and frequency in varactor multipliers. This resistance information is used to study efficiency reduction in multipliers. Based on this simulation, a current dependence for the diode series resistance is derived.

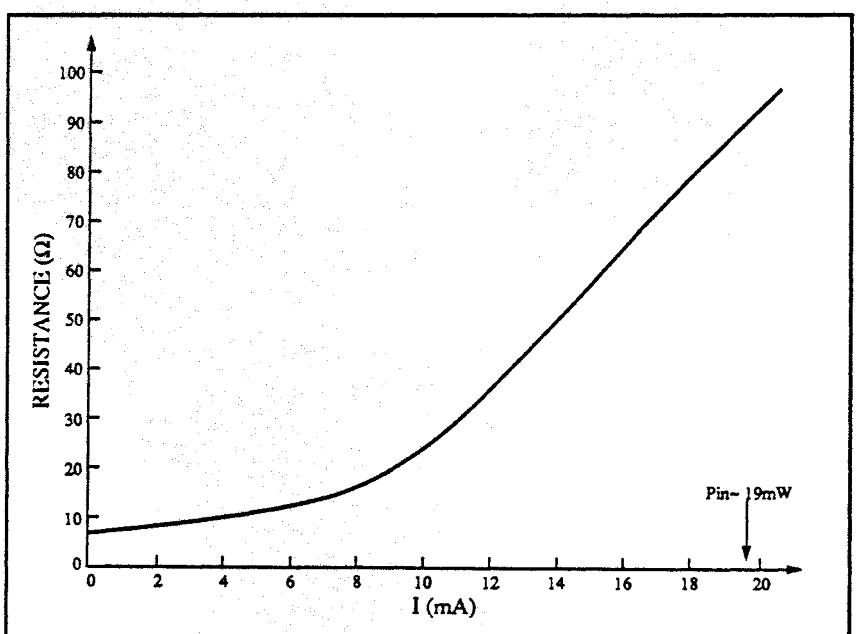

Figure 3: The large signal series resistance at 200 GHz for the 2T2 diode as obtained from the drift-diffusion model.

The varactor equivalent circuit must be further modified at higher frequencies. Consider the varactor equivalent circuit shown in Fig. 1b where the capacitance C_{R_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.

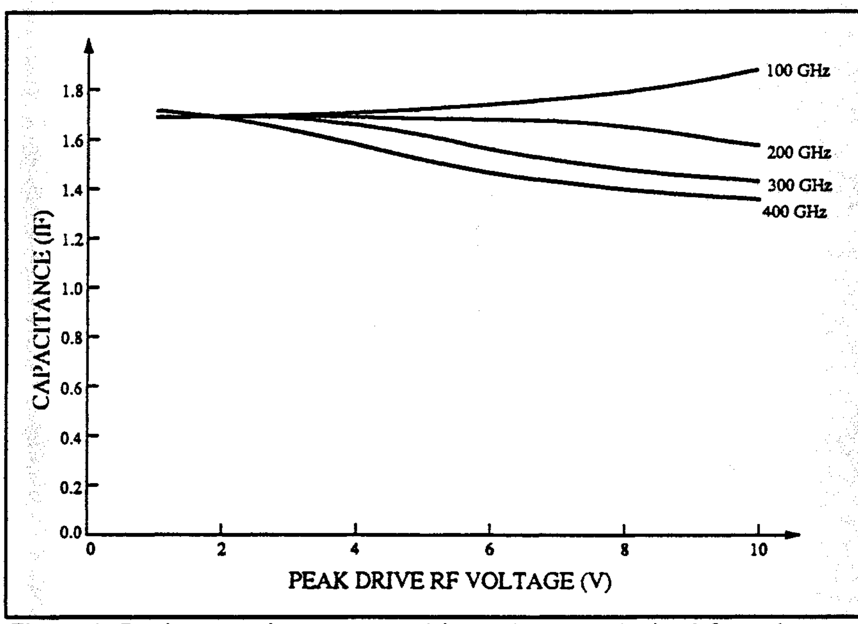

Figure 4: Device capacitance versus drive voltage as obtained from the drift diffusion model for different frequencies.

This brief description of varactor models shows several limitations with the simple equivalent approach typically used with large signal device circuit programs. 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 mixers. The results will be used in the next section to see the effect on varactor frequency multipliers.

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

As mentioned above, in GaAs, the electron velocity in a DC electric field reaches a maximum of about $2.2 \cdot 10^5$ m/s [6]. Electron velocities in bulk GaAs have been calculated for different RF electric fields at 50, 250 and 1000 GHz using Monte Carlo simulations [10]. At 250 and 1000

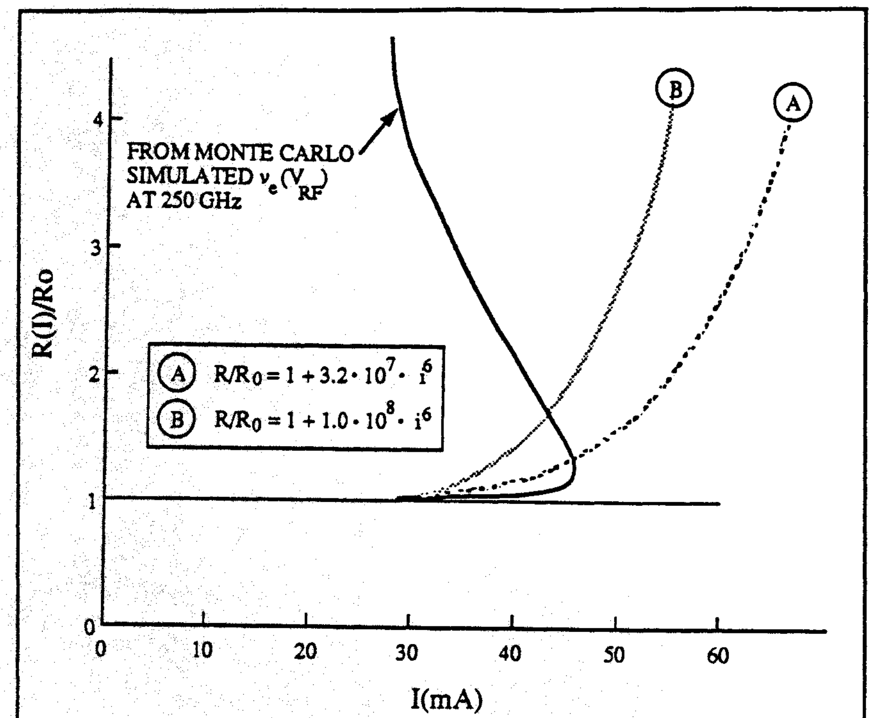

Figure 5: Resistance versus drive current obtained from Monte Carlo simulations and Eq. (4), and resistance versus drive current according to Eq. (5) ((A) and (B)) and used for calculation of efficiencies as described in Fig. 6.

GHz in addition to the in phase component v_o, an inductive, i. e. delayed velocity component v_1, is present . Hence in a high frequency electric field, the net electron velocity $(v_o^2 + v_1^2)^{1/2}$ is slightly larger due to overshoot phenomena. The maximum velocity at 50, 250 and 1000 GHz is about $2.85 \cdot 10^5$, $2.9 \cdot 10^5$ and $3.4 \cdot 10^5$ m/s at approximately 4.5, 4.5 and 10 kV/cm, respectively. Extrapolating this data to DC, the velocity is $2.8 \cdot 10^5$ m/s at about 4 kV/cm, which can be compared with $2.2 \cdot 10^5$ at 3.2 kV/cm [6]. It is evident from the discussion above that the velocity versus field dependence at high frequencies is considerably modified as compared to the DC curve [6].

Although the current (velocity) can be divided up into one in phase ($i_0 \propto v_o$) and one 90 degrees out of phase ($i_1 \propto v_1$) component, we can also model this as a series configuration with a total current $(i_o^2 + i_1^2)^{1/2} = (v_o^2 + v_1^2)^{1/2} \cdot N_d \cdot e \cdot A_d$ with one voltage component in phase and one inductive voltage component out of phase with the current. The "inductive voltage" can be tuned out by a proper choice of embedding circuit. The effective series resistance can then be identified as

$$R_s(E) = \frac{i_0(E)}{i_0^2(E) + i_1^2(E)} E \cdot L \tag{5}$$

In Fig. 5, $R(E)/R(E{\approx}0)$ is shown as a function of the total current $(i_o^2 + i_1^2)^{1/2}$ for the field E for the 6P4 diode. (E is the electric field and L is an arbitrary length and will disappear upon normalization.)

Since the current is limited at high frequencies, an approach with a current dependent series resistance was chosen. In order to have a single valued R(i) curve, we define the current dependent series resistance accordingly:

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

By a proper choice of the parameter "a" and the power for the current (here chosen equal to 6 although this is not critical), $R_s(i(t))$ will increase dramatically for currents near and above the maximum current defined by Eq. (3) or the curve in Fig. 5 for $R(E)/R(E{\approx}0)$. This choice of the current dependent series resistance will

- i. cause the apparent series resistance at the pump frequency and the harmonics to increase with the pump power (compare Fig. 2), and
- ii. modify (clip) the current waveform approximately as required.

In Fig. 5 $R_s(i)$ is depicted for two values of the parameter "a" to be used for multiplier performance simulation in the next paragraph.

5. Results.

In Table 1 information is presented for two diodes, the 6P4 millimeter wave varactor diode, and the 2T2 submillimeter wave varactor diode. Both are built at University of Virginia. Erickson used the 6P4 diode in the 2x80 GHz multiplier, and the 2T2 diode in a 3x160 GHz multiplier. The maximum measured efficiency of the 2x80 GHz multiplier was 34%. Since the losses in the mount were about 1.3 dB, the intrinsic conversion efficiency is about 46%. The 3x160 GHz multiplier had 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 3x160 GHz multiplier by scaling the 1.3 dB waveguide loss for the 2x80

GHz multiplier. This results in an intrinsic efficiency is about 9%. The efficiency of these multipliers were investigated theoretically using ta large signal analysis computer program. The relevant data for the diodes are summarized in Table 1.

Table 1: ***Diode Data for the U. Va varactor diodes 6P4 and 2T2.***

	2x80 GHz *f_p=80 GHz* *6P4*	3x160 GHz *f_p=160 GHz* *2T2*	*3x330 GHz* *f_p=330 GHz* *2T2*
N_d nominal (cm^{-3})	$3.0 \cdot 10^{16}$	$1.0 \cdot 10^{17}$	$1.0 \cdot 10^{17}$
N_d used by us (cm^{-3})	$3.5 \cdot 10^{16}$	$1.0 \cdot 10^{17}$	$1.0 \cdot 10^{17}$
t_{epi} (μm)	1.0	0.5	0.5
R_{so} (Ω) measured	10	14[1], 12[2]	14[1], 12[2]
C_0 (fF) measured	21	5.5[1], 6.5[2]	5.5[1], 6.5[2]
$V_{break\text{-}down}$ measured (Volts)	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]
assumed v_{max} (m/s)	$2.4 \cdot 10^5$	$2.4 \cdot 10^5$	$2.4 \cdot 10^5$
$i_{sat} = AN_d e v_{max}$ (mA)	44	23	23

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

<u>*2x80 GHz Multiplier.*</u>

For the 2x80 GHz multiplier, R_{so} was chosen equal to the measured series resistance of 10 Ω. We selected two values for "a" in Eq. (6), as indicated in Fig. 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 $R_s(i)$ curves shown in Fig. 5. The results are summarized in Table 2.

Table 2: ***Calculated performance of the Erickson 2x80 GHz and 3x160 GHz multipliers.***

	P_{in} (mW)	*i_{RF} (mA)*	*dW/dt* *10^5(m/s)*	*η*	*R_s (Ω)*	*a*
2x80 GHz	11	42.0	2.0	52	10	–
i_{sat}=	40	69.0	2.8	54	10	–
44 mA	11	41.5	1.7	46	*R(i)*	$3.2 \cdot 10^7$
	40	62.7	2.6	41	*R(i)*	$3.2 \cdot 10^7$
3x160 GHz	2	23.7	≈1.7	22	12	–
i_{sat}=	12	47.0	3.5	>40	12	–
23 mA	2	23.0	1.6	20	*R(i)*	$1.53 \cdot 10^9$
	20	33.0	2.8	9	*R(i)*	$1.53 \cdot 10^9$

It is quite obvious from Table 2 that 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 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 is decreased with about 25% from its unsaturated value.

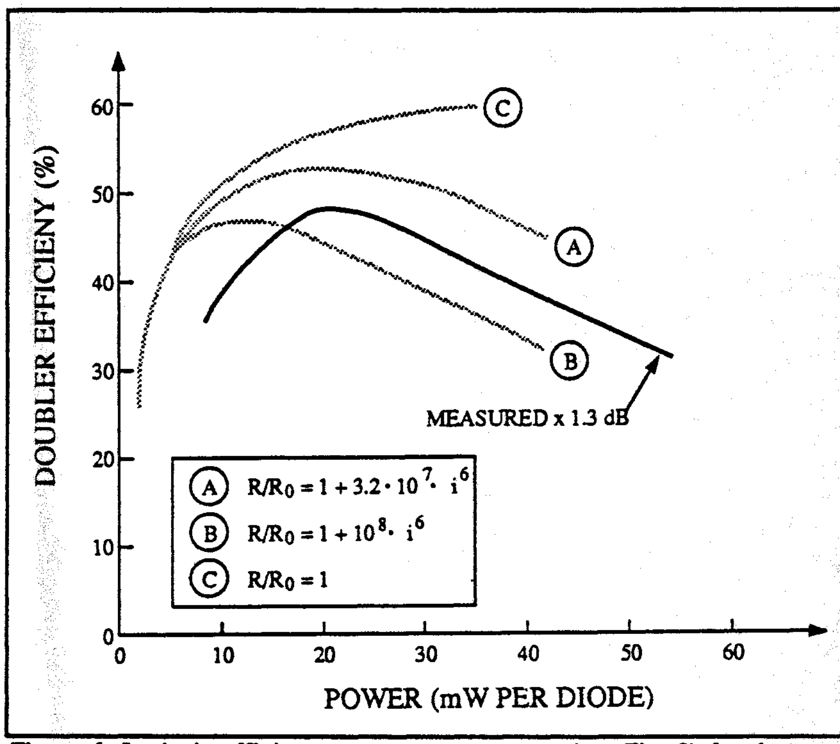

Figure 6: Intrinsic efficiency versus pump power (see Fig. 2) for the Erickson 2x80 GHz multiplier compared to calculated efficiency using the current dependent series resistance (A) and (B) as suggested in Fig. 5.

The agreement between theoretical calculations and the experiments are shown in Fig. 6. These results give some confidence in evaluating the 3x160 GHz multiplier using the same technique.

<u>*3x160 GHz Multiplier.*</u>

From Table 2 it is seen that i_{RF} is greater than i_{sat} for input powers as low as 2 mW. The situation seems now much worse than for the 2x80 GHz multiplier. For the analyses, the lower value of the parameter "a" was used, scaled taking into account 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; $A(2T2)=A(6P4)\cdot\{C(2T2)/C(6P4)\}\cdot\{N_d(6P4)/N_d(2T2)\}^{1/2}$. Computed results are shown in Table 2. The efficiency calculated including the saturation effect is now deteriorated by a factor of almost five as compared to the efficiency expected for the constant series resistance case (R_s=12 Ω). The predicted efficiency of 9% for large pump power coincides with the measured one assuming the ohmic losses of the mount is 5 dB. Interestingly Erickson pointed out that the reverse bias voltage for optimum performance of this tripler was near the break-down voltage (V_{bias} = 5-6 volts, $V_{break\text{-}down}$= 8.5 volts), and the negative voltage swing was considerably larger than 8.5 volts.

3x330 GHz Multiplier.

A theoretical case of a 3x330 GHz multiplier was also analyzed. Results are shown in Table 3. 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 calculation 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.

Table 3: ***Calculated performance of 990 GHz Schottky Barrier Diode Varactor (2T2) Multiplier.***

	P_{in} (mW)	i_{RF} (mA)	dW/dt 10^5 (m/s)	η	R_s (Ω)	a
3x330	2	24	2.3	4.1	12	-
GHz	10	53	4.6	11.6	12	-
i_{sat}=	2	23	2.0	2.3	R(i)	$1.53 \cdot 10^9$
23 mA	10	31	2.9	0.6	R(i)	$1.53 \cdot 10^9$

Discussion

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 inconsistences such as that the velocity of the depletion edge calculated in two different ways as demonstrated in Eq. (6) do not agree, viz.

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

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 break-down voltage will be lower, and the corresponding epitaxial layer thickness smaller) will suffer less from saturation, and are expected to be more efficient.

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 [11]. 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). The SBV exhibits a symmetric C-V characteristic 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 common Schottky varactor diode.

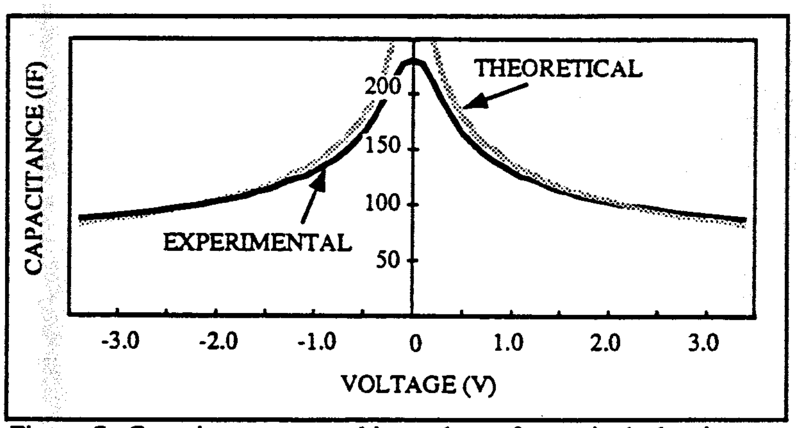

Figure 7: Capacitance versus bias voltage for a single barrier varactor.

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, $C_{max} = (A_d\varepsilon)/W_{barrier}$, 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 200Å. An obvious possible contribution to this excess barrier thickness stems from the fact that there are 50Å undoped regions on both sides of the barrier. In addition, the deviation may arise from the discontinuity of the Fermi level at the doped/undoped barrier interface.

A few simulations of a SBV diode multiplier have been carried out, showing similar results as for the Schottky varactor multiplier as summarized in Table 4. For the particular case shown in Table 4, it is assumed that $N_d = 2\cdot10^{17}$ cm^{-3}, $W_{barrier}$=200 Å (this is now an *effective* width), C_0=20 fF, R_s=12 Ω. It is seen that this device suffers less from saturation while the velocity of the edge of the depletion region seems to be a more serious problem. InAs based devices have fewer problems in this respect since the electron velocity is two to three times higher than for GaAs.

Table 4. ***Predicted performance of a Single Barrier Varactor 3x330 GHz Multiplier.***
($N_d = 2\cdot10^{17}$ cm^{-3}, $W_{barrier}$=200 Å, C_0=20 fF, R_s=12 Ω;
GaAs: $v_{max}\approx2.4\cdot10^5$ m/s, $i_{sat}\approx88$ mA; InAs: $v_{max}\approx5\cdot10^5$ m/s, i_{sat} >100 mA)

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)	$1.53\cdot10^9$
10	31	5.1	6.8	R(i)	$1.53\cdot10^9$

A limitation of these devices as the are currently made is that they start conducting at a finite voltage. Multiplier simulation shows 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 can be overcome by using a number of diode 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 worries about 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.

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 $i_{e,max}$ with a factor of 6 (private communication with Dr. I. Mehdi).

6. 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. To date most varactors have been fabricated from GaAs which exhibits a saturation velocity of about 2-$3 \cdot 10^5$ 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 break-down 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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