New Modes of Operation for Avalanche Diodes: Frequency Multiplication and Upconversion

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Abstract—**The nonlinear properties of avalanche diodes can be exploited for use as frequency multipliers and upconverters. Among the various possible modes of avalanche device operations, the two mentioned above evoke great interest because of their high conversion efficiency. Basic principles, theoretical predictions, and experimental results are outlined in some detail and indicate both the present state of development as well as the performance achieved using avalanche devices operated in this manner.**

Introduction

FREQUENCY upconverters and multipliers, when used in microwave communication systems, ought to display a high conversion efficiency. A frequency upconverter may, for instance, constitute the last stage of the repeater in a circular waveguide millimeter-wave communication system. The distance between two repeaters depends on the output power which in turn is affected by the conversion losses of the said upconverter.

Such frequency multiplications and upconversions are normally achieved by using the nonlinear capacitance effect found in varactor diodes. The limiting factor in them which is due to parasitic series resistance, however, becomes rather pronounced in the millimeter-wave region. For such an upconverter, the minimum conversion loss is close to 5 dB. Besides, the limited variations of the nonlinear capacitance with voltage do not permit a direct frequency multiplication at a harmonic order greater than four.

It may be interesting to use the avalanche diode as a frequency multiplier or upconverter by making use of its behavior as a nonlinear inductor [1]–[3], [5]. The high degree of nonlinearity of the avalanche process renders itself consistent to frequency multiplication of very high harmonic orders. In addition, the parasitic series resistance is much smaller than that in the varactor diodes since the depletion layer width remains constant even under large-signal operation. Thus conversion losses may be reduced considerably. The principle of the avalanche diode frequency multiplication was reported by Constant [1] and some important results have been published [5]–[9]. Recently, Ntake and Conn [10] considered the case of a varactor P-I-N diode driven into avalanche breakdown. Their analysis, however, is quite different. Frequency conversion was first considered by Evans and Haddad [2]. In 1974, Vaterkowski [11] gave the first results concerning avalanche diode upconverters in millimeter-wave region.

In this paper, the principle of the nonlinear operating characteristics of an avalanche diode and the methods of the theoretical treatment are first outlined. The design considerations and experimental results for frequency multiplication and upconversion are then presented.

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Principle of Operation

The dynamic operating characteristics of an avalanche diode can be best understood by dividing the active zone of the device into two parts: an avalanche zone and a transit zone. From Read's equation, it can be easily deduced that the avalanche zone behaves as a nonlinear inductor [1]–[3], [5]. It contrasts with the nonlinear capacitance effect found in varactor diodes. The "Manley Rowe" power relations show that the avalanche zone can operate as a parametric amplifier, a harmonic generator, or a frequency up- and downconverter, with a conversion efficiency at the desired frequency approaching 100 percent. The nonlinear behavior of an avalanche diode is due only to the nonlinearity of the avalanche zone. The influence of the transit zone is quite different and its dimensions must be optimized for each mode of operation.

The complete "exact" calculation of avalanche diode performances is a complicated task: many different numerical cases must be considered in order to obtain the optimum operating conditions. Therefore, in addition to a general calculation method, we have developed several less expensive approximative ones.

The fundamental general device relations (electron and hole current continuity and Poisson's equations) are solved with a minimum of simplifying assumptions [8], [12], [13]; these equations are reformulated as finite differences, time and space being divided into equal intervals. Solution of these equations at time t and any point x is obtained by an iterative process, until a specified accuracy is reached (prediction correction method). The total voltage $V(t)$ and current $I(t)$ calculated in this manner are then decomposed in a Fourrier expansion and the various powers and impedances are subsequently obtained.

The simplified calculations [8], [9], [12] are based on the division of the active device into an avalanche and a drift zone. It is also assumed that the avalanche zone width is constant. This approximation is usually used by several authors [2]–[4].

The conduction current density J_c generated in the avalanche zone is derived from the modified Read's equation or Lee's equation [15]. They are

$$\frac{\tau_\delta}{M} \times \frac{dJ_c}{dt} = J_c \left[\int_0^\delta \alpha(E)\, dx - 1 \right]$$

(modified Read's equation)

$$\frac{2}{(v_n + v_p)M} \frac{dJ_c}{dt} = J_c \frac{\int_0^\delta \alpha_n(E)[\exp(\int_0^x - (\alpha_n(E) - \alpha_p(E))\, dx')]\, dx - 1}{\int_0^\delta [\exp(\int_0^x - (\alpha_n(E) - \alpha_p(E))\, dx')]\, dx}$$

(Lee's equation)

where $E = E(t + \tau_y)$ is the electric field which is stipulated at the beginning of the calculation; τ_δ is the avalanche zone transit time; v_n and v_p are electrons and holes velocities; and α_n and α_p are ionizations rates for electrons and holes. The transit time and mobile space charge effects are to a certain extent taken into account; the coefficients M and τ_y values are obtained from the results given by the general method (usually $M \simeq 2.8$ and $\tau_y \simeq \tau_\delta/9$). From this, the total current J_T can be easily obtained: $J_T = J_c + \varepsilon(\partial E/\partial t)$.

For the analysis of the drift zone two approaches are generally used as follows.

a) The first one is achieved by solving Poisson's equation (formulated as finite differences) with the assumption that the carriers drift at their saturated velocity until the electric field drops below the level E_s obtained in the undepleted zone. This value E_s is related to the total current density by

$$J_T = q \times N_D \times v(E_s) + \varepsilon \frac{\partial E_s}{\partial t}.$$

b) The second approach is an extension of Gilden and Hines' equations [20] into the case of a periodic structure [13]. Quite simple analytical formulas can then be derived. However, as opposed to the preceding method, it does not account for the parametric effects [14] and thus it is essentially useful for punchthrough diodes.

In most cases, the theoretical evaluation of diode performances is achieved by means of the simplified approximative methods in the following manner:

1) the solution of Lee's equation for the avalanche zone and Poisson's equation for the drift zone in the frequency multiplication study;
2) the solution of Read's equation and Gilden and Hines' equation in the frequency conversion study.

The approximate model can also be used to study a diode with an external circuit connected, which constitutes a quite suitable refinement for the theoretical design of frequency multiplier circuits. In this case the dissipated power at the input frequency and the load impedances at the other (idler and output) frequencies are stipulated. The steady-state operating conditions are then obtained by means of an iterative calculation method [18].

Frequency Multiplication

Harmonics are generated whenever a sinusoidal source drives a nonlinear impedance, which is the case in the avalanche diode. Hence an avalanche diode may be used as a convenient harmonic generator [1]–[3]. Preliminary works, undertaken by Allamando and Vanborren [5] have shown that the advantage of the avalanche multiplier is not significant when compared with varactor types under low harmonics low frequency operation. On the other hand, in an avalanche diode the nonlinear dependence of the electronic current upon the applied RF electric field is so strong that this high degree of nonlinearity permits a high order of harmonic generation in a single stage, with no special provisions made for idler current flow.

It is obvious that the avalanche current waveform is very rich in harmonics. Besides, the quick rise of the current peak (down to a few picoseconds) indicates high order high frequency harmonics up into the millimeter-wave range. Hence the avalanche diode appears to be a very attractive high harmonic millimeter-wave multiplier.

Diode Design

For high order harmonic operation, it is better to aim for short transit time diodes approaching the case of a sharp current pulse. In order to eliminate series resistance, it is desirable to have the p-n junction bounded by heavily doped p^+ and n^+ regions. So the diode is generally made as a p^+-n-n^+ structure. If a very short transit time is desired, the heavily doped region (n^+ and p^+) must be sufficiently close to the junction, so that the depletion layer extends to this heavily doped region for reverse voltages much less than the breakdown voltage. Such devices are punched through at breakdown. We shall use in this paper the punchthrough factor (PTF) as it has been introduced by Evans [16].

Computer calculations have shown that the most suitable devices should be highly punched through with electrical field profiles close to P-I-N types [8], [9]. This is illustrated in Fig. 1(a) where the theoretical optimum output power is plotted against the PTF for an X 10 multiplier. This can be explained by the fact that the multiplication effect is located in the avalanche zone and that high voltage levels can be the more easily applied across it as the PTF is higher. This means that these devices provide negligibly small output power in IMPATT mode.

The optimum value of the avalanche zone width and also the layer doping concentration is determined by taking into account two opposite factors:

1) power levels at the different frequencies increase with the avalanche zone width δ, and
2) the nonlinearity of the avalanche zone is a decreasing function of δ.

Numerical calculations [8], [9] have shown that the product $\omega\tau_\delta$ must remain close to unity for the output frequency ($N_D \simeq 2 \times 10^{16}$ At/cm^3 at 35 GHz).

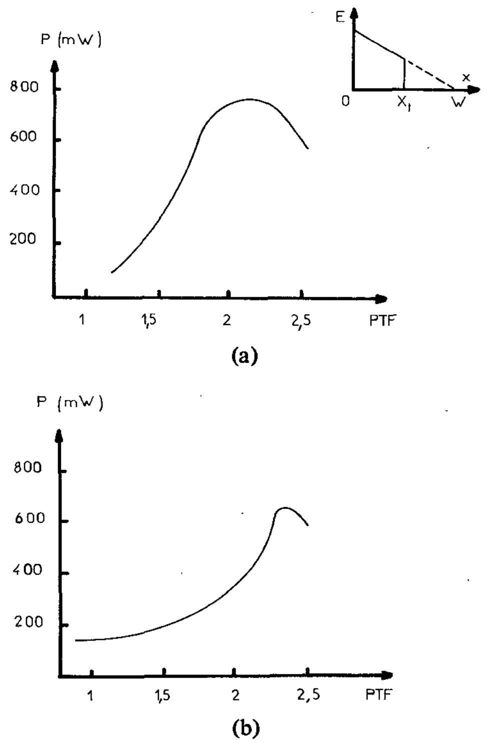

Fig. 1. Variation of output power as function of punchthrough factor for X 10 multiplier. (a) Theoretical curve: $N_D = 2 \times 10^{16} at \times \mathrm{cm}^{-3}$, $J_0 = 10\ \mathrm{kA} \times \mathrm{cm}^{-2}$, $F_{\mathrm{out}} = 35$ GHz. (b) Experimental curve: $N_0 = 2 \times 10^{16} at \times \mathrm{cm}^{-3}$, $J_0 = 10\ \mathrm{kA} \times \mathrm{cm}^{-2}$, $F_{\mathrm{out}} \cong 35$ GHz.

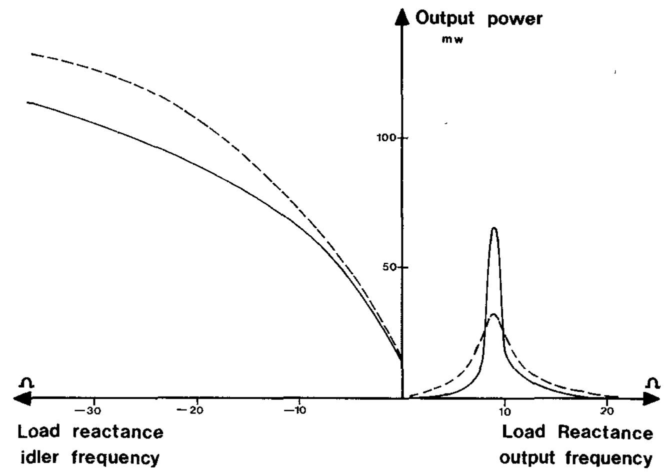

Fig. 2. Calculated effect of external circuit on output power of X 10 multiplier. Right side: effect of output frequency load impedance ($Z_{10} = R_{10} + jX_{10}$) when no idler circuit is introduced (circuit impedance is zero at all undesired harmonic frequencies); continuous line: $R_{10} = 1\ \Omega$; dotted line: $R_{10} = 2.5\ \Omega$. Left side: effect of capacitive tuning on second and third idler frequencies: $Z_{10} = 2.5 + j8\ \Omega$; continuous line: tuning on second harmonic; dotted line: tuning on third harmonic; conditions: $F_{\mathrm{out}} = 35$ GHz, Si diode, $S = 0.4\ 10^{-8}\ \mathrm{cm}^2$, $W = 0.6\ \mu\mathrm{m}$, $R_S = 0.5\ \Omega$, $V_B = 20$ V, $I_0 = 150$ mA, $P_1 = 200$ mW.

Though it is the particle current that determines the RF properties of the device, there is an important point concerning the capacitance. It dominates the diode impedance at high frequencies and since there is a minimum matchable impedance, the capacitance limits the area of a device operating at a specific output frequency. However, this limitation is much less critical for frequency multipliers than for IMPATT oscillators because the performances in this mode of operation depend not only on the output characteristics but also on the input ones. The optimum junction area was found to range from 0.2 to 0.4 $\times\ 10^{-4}\ \mathrm{cm}^2$ for Q band operation and from 0.1 to 0.15 $\times\ 10^{-4}\ \mathrm{cm}^2$ for E band operations, for highly punched-through Si diodes (PTF $\simeq$ 2.2).

Circuit Design

The fact that a nonlinear inductance enables us to have a highly efficient harmonic generator does not guarantee that maximum power is being converted from the input frequency to the desired output frequency. Maximum power conversion, at maximum efficiency, can be achieved only by careful circuit design. When maximum output power is not required, the circuit design may be simple and usually does not employ any deliberately introduced idler circuits. However these idlers are always present, and so it is interesting to study their influences on the multiplier performances. This is summarized in Fig. 2.

If the voltage waveform is of the type

$$V = V_0 + V_1 \sin \omega t + V_n \sin (n\omega t + \phi_n)$$

comprising only of the dc, input and output components, the load impedance must be zero at all the other harmonic frequencies. In this case (right side of Fig. 2), it can be seen that maximum output power is obtained for a very small load resistance. The load reactance tuning becomes critical as the load resistance is reduced. Maximum conversion efficiency of about 30 percent can be obtained in this case.

Let us now examine what happens if a reactive load is introduced at one of the idler frequencies. The first important point is that one must be careful not to develop too high a voltage at this idler frequency which would result in a premature saturation of the diode output power and a decrease of the conversion efficiency. This means that the diode capacitance must not be tuned to resonance at this frequency and that it is better to avoid an inductive load. On the left side of Fig. 2, it can be seen that the capacitive tuning of the idler circuit is not critical at all and provides a great increase in the conversion efficiency which can reach a value of 60 percent. This is a very important result because it shows that in contrast to the varactor diodes, the idler circuit is not selective. This can be explained by the fact that the high degree of nonlinearity of the avalanche process causes the conduction current to present all the harmonic components up to very high frequencies, even under voltage sinusoidal waveform conditions. The introduction of a voltage at the considered idler frequency via the idler circuit causes the nth output of the particle current to increase. This increase of the output current component will in turn increase the output voltage for a given output circuit. This cumulative effect results in an increase of the output power.

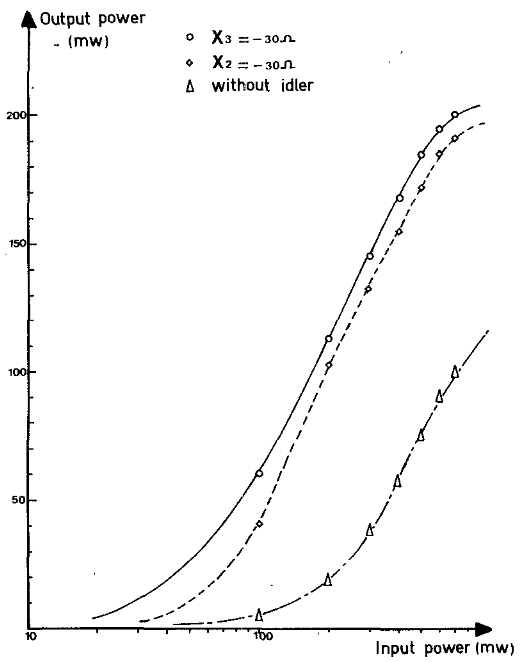

Fig. 3. Calculated output power variations of X 10 multiplier as function of drive level with and without idler tuning. Same conditions as in Fig. 2: $Z_{10} = 2.5 + j\,8\ \Omega$.

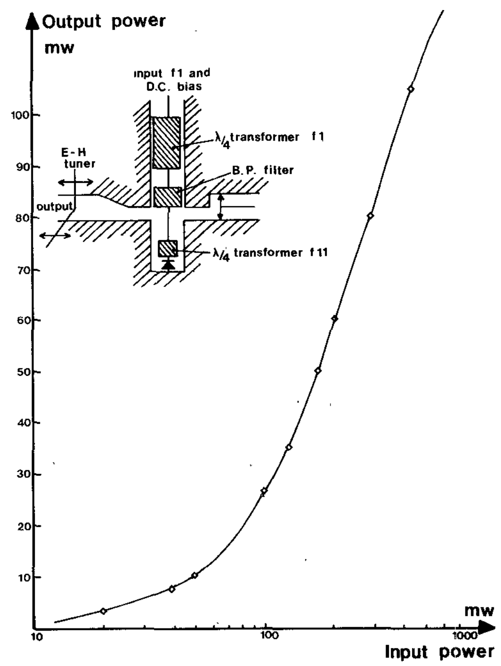

Fig. 4. Experimental output power variations of X 11 multiplier as function of drive level. $F_{out} = 38.8$ GHz, Si diode, $V_B = 20$ V, $W = 0.6\ \mu$m, $S = 3.10^{-9}$ cm^2.

The multiplier can be tuned up more completely by using capacitive idler tuned on both second and third harmonic frequency (for $X_2 = -20\ \Omega$ and $X_3 = -30\ \Omega$, the conversion efficiency reaches 90 percent).

Theoretical Calculation

A typical theoretical curve of output powers versus input power is presented in Fig. 3. For Si diodes in Q band, the output circuit tuning is kept constant, while the input is matched at each point. A capacitive loading at the second or third harmonic frequency results not only in a greatly improved output power as it has been already explained but also in an earlier efficiency saturation, due to a higher value of the instantaneous voltage.

The output power is also an increasing function of dc bias current [6], [8] until saturation occurs; for each value of input power, there is a value of dc bias current which gives a maximum output power. The variations of the output negative resistance, with respect to input power and dc bias current, are similar to those of the output power.

Another interesting result is that the output power and conversion loss are slowly decreasing functions of the order of frequency multiplication harmonic. With a given output frequency, the variation is close to 3 over n for harmonic orders ranging from 10 to 40 [7].

Experimental Results

Diffused Si p$^+$-n junctions with various doping profiles have been used for the experimental studies. These diodes were made by the LEP[1] [8]. The multiplier tests were made with a circuit whose basic structure is a coaxial-to-waveguide transition [6], [7]. The diode terminates a matched low-impedance coaxial line which passes through the waveguide at the center of the broad wall. This coaxial line is coupled to the waveguide by a coaxial-to-waveguide transition matched by a tapered ridge section. This also constitutes a high pass filter allowing tuning on the idler frequencies. Fine tuning of the output is achieved by a movable short at one side and an *E-H* tuner at the other.

The output power and conversion efficiency obtained from various semiconductor wafers show that multiplication diodes should be designed with a high punchthrough factor [8] and with junction areas in quite good agreement with those predicted from theory. For instance, in Fig. 1(b), the experimental variations of optimum output power are plotted as a function of the punchthrough factor of different diodes. These results are close to the theoretical predictions [Fig. 1(a)]. Typical data of Si diodes for an output frequency in the Q band are: V_B = 20-V PTF $\simeq 2$; $W \simeq 0.6\ \mu$m; $S = 0.3 \times 10^{-4}$ cm^2.

Experimental output power variations for an X 11 multiplier versus input power are presented in Fig. 4. The diode bias current was optimized at each point; but circuit tuning was held fixed at the values corresponding to the highest efficiency for a 20-dBm input power. This figure shows that, in spite of its relatively simple circuitry, this avalanche multiplier performs well since a conversion efficiency of 30 percent was achieved for a 20-dBm input power. The deviation from the theoretical curves shows that this harmonic generator was not yet completely tuned up.

Fig. 5 illustrates the high order multiplication capabilities of an avalanche multiplier. The output frequency was kept

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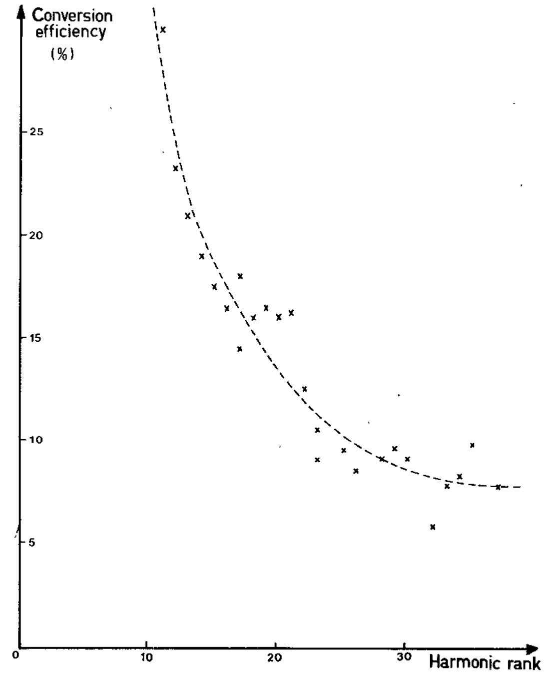

Fig. 5. Harmonic order dependance of multiplier conversion efficiency using same diode as in Fig. 4. $F_{\text{out}} \# 35$ GHz.

fixed around 35 GHz and the harmonic order was increased by decreasing the input frequency. The bias current as well as the circuit tuning were optimized at each point. At each point, at least a 15-dB ratio is achieved between the power on the selected frequency and on all the other parasitic harmonic frequencies. This curve (Fig. 5) brings out the fact that the avalanche multipliers are well adapted to obtain high order multiplications. Some experiments have been successfully performed using similar multipliers with the output frequency in the E band. No particular deterioration in the performances were observed. Harmonic frequencies of up to 140 GHz were made available.

All the results presented here were obtained for medium output power levels usually required in the actual millimeter-wave systems. The avalanche frequency multiplier is capable of much greater performances in this frequency range [6], [8]. A 600-mW output power was obtained from an X 10 multiplier at an output frequency of 35 GHz with a 10-dB conversion loss, while an X 35 multiplier provided 300-mW output power and 13-dB conversion loss. An instantaneous bandwidth of 3.3 GHz was obtained from an X 9 multiplier of 420-mW maximum output power provided with fixed tuning adjustments. The noise level for Si avalanche multipliers was substantially lower than for Si IMPATT oscillators [8]. The improvement is of the order of 26 dB, at 10 kHz from the carrier frequency. Experiments have shown that, when the avalanche multiplier is driven by a highly stabilized source (10^{-10}), it provides an output signal of the same stability [18].

Thus we see that the avalanche multipliers exhibit a high degree of short-term frequency stability, of conversion efficiency, and of harmonic order. It should be noted, however, that these multipliers need specific highly punched-through diodes quite different from IMPATT devices.

Frequency Conversion

There are two basic types of upconverters: the upper sideband upconverter (USUC) has $f_3 = f_1 + f_2$, while for the lower sideband upconverter (LSUC), $f_3 = f_2 - f_1$. Usually a varactor diode is used as a nonlinear component. As described above, the nonlinear characteristic of the avalanche process can be utilized to effect frequency conversion [2], [3], [15]. In this section we consider an avalanche diode of the USUC type only.

In a large-signal USUC, all the frequencies f_1 (intermediate frequency: IF), f_2 (local oscillator frequency: LOF), and f_3 (upper sideband frequency: USF) correspond to large amplitudes of the avalanche electric fields, and only numerical solutions are applicable. In the upconverter case, we assume that the avalanche diode electric field waveform contains only the frequencies of interest (IF, LOF, USF), i.e., the avalanche diode is short circuited by the cavity at all the spurious frequencies.

All the calculations are made with IF = 1.5 GHz, LOF = 33.25 GHz, USF = 36.75 GHz. These frequencies have been selected in order to eliminate in the calculation the multiplication effect described above, and to minimize the integration period of Read's equation, which must be the least common multiple of LO and IF period. The computer program is used to determine the optimum diode structure and load impedances and to predict the USUC performances.

Fig. 6 shows the variations of the input and output powers versus the drift zone width ($W - \delta$) of the Si avalanche diode used. Calculations with different phase shift values were made for each of the various electric fields in the avalanche zone. If we consider an avalanche diode without transit time effect ($W - \delta = 0$), the output and input powers are given approximately by Manley–Rowe's equations, as in varactor diodes. But when the drift zone width increases, US power, IF power, and LO power increase. However, the LO power attains a saturation level whereas US power continues to increase. The conversion efficiency can then become greater than zero decibel. This interesting feature is provided by the transit time effect, which gives under suitable conditions an amplification of the US signal. Such a result is unattainable with varactor diodes. This is the greatest advantage of avalanche diodes used as USUC over varactor diodes.

Another important aspect of USUC is the diode impedance values at LOF and USF. These variations are represented in Fig. 7, where the corresponding output powers are given. The input power is kept constant at 400 mW and the LO power with matched circuit at 200 mW. These curves show that the diode impedance at the LOF and the load impedance at the USF must be about the same in order to reap a maximum of output power and conversion efficiency (point 6).

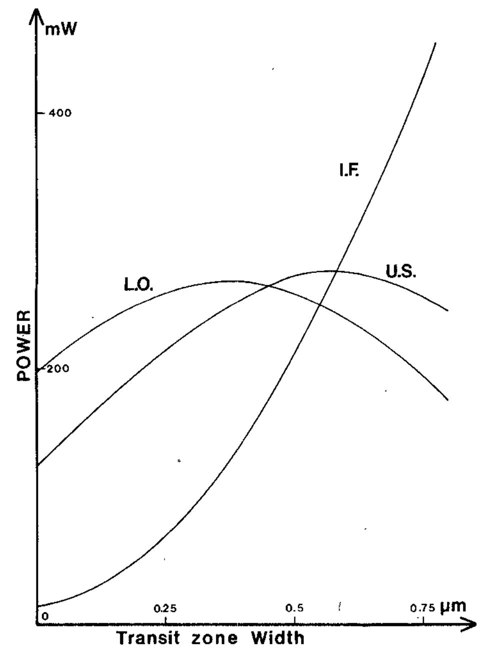

Fig. 6. Calculated variations of input and output power versus drift zone width. Diode diameter 50 μm; avalanche zone width 0.5 μm; dc bias: 100 mA. Electric fields: $E_{IF} = 2.8\ 10^3$ V/cm, $E_{LO} = 6.6\ 10^4$ V/cm, $E_{US} = 6.9\ 10^4$ V/cm. Phases: $\phi_{IF} = 0$, $\phi_{LO} = 0$, $\phi_{US} = 0.22\ \Pi$.

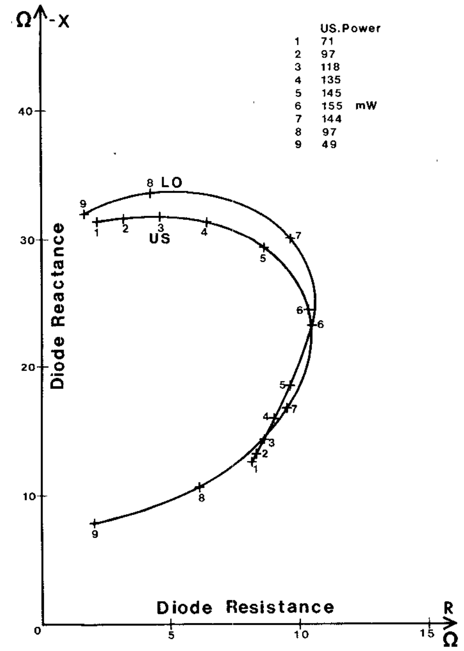

Fig. 7. Theoretical variations of diode reactance versus absolute value of diode resistance, according to different output US power. Diode characteristics: avalanche zone width: 0.5 μm, drift zone width: 0.6 μm, diameter: 50 μm, dc bias: 100 mA, IF power: 400 mW, LO power: 200 mW.

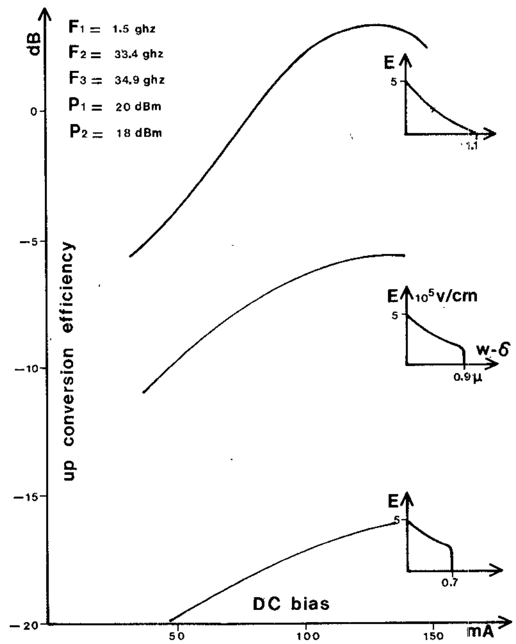

Fig. 8. Experimental upconversion efficiency versus dc bias for diodes with different electric field profiles. Diode characteristics: diameter: 50 μm, avalanche zone width: 0.5 μm.

Single-drift diffused silicon junctions, made at the LEP[1] and double-drift junction made at the DMH of Thomson CSF[2] were used. For the single-drift type, the junction diameter is usually close to 50 μm, and the breakdown voltage is about 25 V, while the corresponding values for the double-drift type are 70 μm and 30 V.

Silicon diodes with different doping profiles have been tested (Fig. 8). In this figure, variations of conversion efficiency versus the dc bias are represented for diodes with different transit zone widths. In these cases, the local oscillator at 33.5 GHz gave 18-dBm input power. The intermediate frequency oscillator at 1.5 GHz gave 20 dBm. In these experiments it appears that the US power increases with the depletion layer width according to theoretical predictions. We note, however, that a conversion efficiency of +3 dB is obtained with the diode having a depletion layer width of 1.1 μm, and such a diode is similar to that used for IMPATT oscillators. Whereas it should be noted that all output signal disappears when the LO signal alone is removed [19]. Conversion efficiencies are similar for single- and double-drift diode in the actual state of development.

Two different basic mountings have been tested, the first one used is a reflection system and the second is a transmission system. Fig. 9 shows the reflection diode mount. Impedance matching is achieved by means of a coaxial waveguide transformer and a movable slug in the coaxial. Fine tuning is obtained with four dielectric screws. The input IF signal is fed into the diode from a coaxial line through a millimeter-wave filter while the local oscillator

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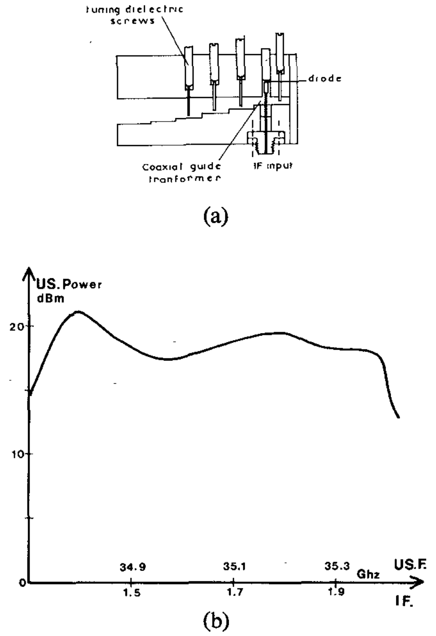

Fig. 9. (a) Diode mount of reflection USUC. (b) bandwidth for reflection USUC experiment. Diode characteristics: diameter: 50 μm, avalanche zone width: 0.5 μm, drift zone width: 0.6 μm, dc bias: 150 mA, LF power: 24 dBm, LO power: 20 dBm.

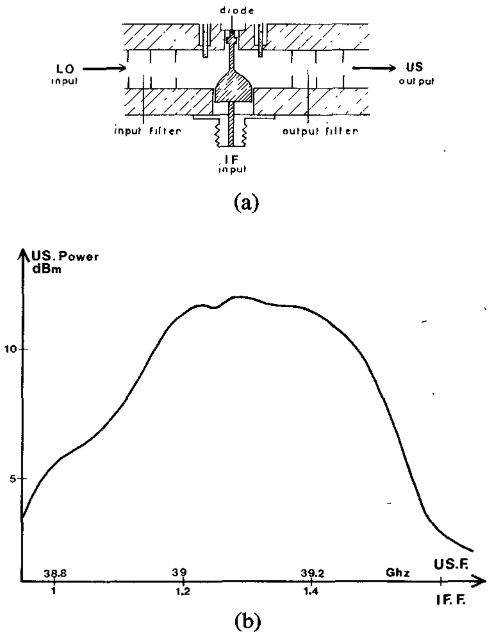

Fig. 10. (a) Diode mount of transmission USUC. (b) Bandwidth for transmission USUC experiment. Diode characteristics: double-drift junction, diameter: 65 μm, dc bias: 100 mA, IF power: 23 dBm, LO power: 20 dBm.

signal is fed from the waveguide. The US signal is separated from the LO signal by means of a circulator. Fig. 9(b) gives the bandwidth obtained under these conditions. The 3-dB bandwidth is 680 MHz, but the ripple is about 3 dB in the whole bandwidth.

Fig. 10(a) shows the diode mount used in the transmission system. LO and US signals are separated by means of two bandpass filters integrated in the diode mount, in order to obtain a cavity between these two filters. The input filter has a narrow bandwidth while the output bandpass filter has a bandwidth of 350 MHz. This structure gives, under identical conditions of input powers and dc bias, the same output power as in the reflection system. However, it is easier to tune circuit impedances for maximum output power and for minimum ripple in the entire bandwidth [Fig. 10(b)]. Moreover, in the transmission mount, the use of integrated filters enables us to suppress all the spurious frequencies. Group delay variations and phase shift variations have also been measured. The typical results show that the phase shift variations versus the IF input power are always less than 3° per dBm and the group delay variations in the entire bandwidth of the USUC system are less than 2 ns [19]. These experimental data are quite consistent with transmission system requirements.

Conclusion

In this paper applications of avalanche diodes as frequency multipliers and upconverters have been described. The high degree of nonlinearity of the avalanche process enables the realization of the high order high efficiency frequency multipliers. When this same nonlinearity is combined with the negative resistance of IMPATT diodes, amplifying frequency upconverters in the millimeter-wave range can be achieved. It should be noted that these applications require circuits less complicated than those used in conventional varactor devices.

The avalanche multipliers provide signal sources when direct signal generation becomes difficult to realize. Their high harmonic order capabilities in a single stage eliminate some of the main difficulties of varactors chains such as bandwidth and interstage impedance matching and filtering. Thus avalanche multipliers can find extensive use in the millimeter-wave range as local oscillators for receivers, parametric amplifier pumps, and carrier power sources for transmitters.

The avalanche frequency upconverters appear very attractive because of their high conversion efficiency. They can also serve as output circuits in millimeter-wave transmitters or repeaters. Their high output power and efficiency are suitable for power amplification in the millimeter-wave range.

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InP Gunn-Effect Devices for Millimeter-Wave Amplifiers and Oscillators

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Abstract—**CW InP Gunn oscillator performance has been extended up in frequency to the 26.5–40 and 50–75-GHz ranges. CW power outputs of 78 mW at 56 GHz have been attained to date. Amplifier evaluation in *Ka* band yielded useful gain from 26.5 to 40 GHz in two half-band circuits with noise figures ranging from 12.4 to 16.5 dB on flat profile devices. In a narrow-band amplifier circuit at 23 GHz, a device noise figure of 10.1 dB was obtained at 9-dB gain. A description of material growth, evaluation techniques, and device designs is also presented.**

I. Introduction

Interest in InP for transferred-electron oscillators and amplifiers has been increasing recently owing to its potential for high efficiency and low noise. Experimental results have been reported which confirm these attractive characteristics. Peak output powers of 8.5 W at 18-percent efficiency and 5 W at 24-percent efficiency have been reported under low-duty-cycle pulsed operation in *Ku* band [1]. Although the thermal conductivity of InP is higher than GaAs, because of InP's high threshold field (10 kV/cm) power densities are high and efficient CW operation is limited to 18 GHz and above (except for very low *nl* product amplifier devices) by thermal restraints. Under CW operation, efficiencies up to 10 percent at 22 GHz and 6 percent at 27 GHz have also been achieved [2], [24]. CW narrow-band reflection amplifiers have provided noise figures as low as 7.5 dB at 33 GHz [3] and 10.7 dB at 14 GHz [4], while theoretical predictions of 4 dB for optimum cathode notch structures have been published [5].

The investigations reported in this paper are being directed toward development of high-efficiency InP sources and wide-band low-noise amplifiers above 26 GHz. In order to accomplish this, vapor-phase InP epitaxial growth reactors have been optimized for control of the thin layers required for high-frequency operation. Material characterization methods have been designed for increased measurement accuracy on InP so that layers can be evaluated for carrier concentrations and impurity levels over a wide range of doping. Device fabrication methods, originally utilized on GaAs devices, have been adapted for InP requirements.

II. Device Technology

The application of InP for CW transferred-electron oscillators and amplifiers in the millimeter-wave range provides significant performance improvements over the more widely utilized GaAs devices. In particular, InP is a

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