Read-Type Varactors for Parametric Amplifier Applications

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Abstract—The use of Read-type HI–LO doping profiles in varactors for parametric amplifier applications is shown to result in improved performance over conventional structures. Optimal diode doping levels, layer thicknesses, and pump drive levels are derived which give specified frequency performance while minimizing pump power requirements, minimizing noise, maximizing dynamic range, and reducing amplifier sensitivity to pump power fluctuations. The optimum device design is based on environmental limitations such as pump power, circuit losses, and impedance levels, and the unavoidable diode series resistance level. Design examples are given for 10- and 100-GHz parametric amplifiers.

I. Introduction

THE PARAMETRIC amplifier remains an important source of low-noise amplification in many applications, particularly at millimeter-wave frequencies and frequencies beyond the capability of microwave transistors (MESFET's and BJT's). As frequencies of operation increase, the confines within which the amplifier must operate become increasingly restrictive and performance correspondingly degrades. Inadequate or low pump power and increasing circuit and device losses are among the more important limitations encountered at millimeter-wave frequencies. Within these and other restrictions it is readily apparent that the varactor itself should be designed to maximize the quality of amplification at the signal frequency. With the ever improving technology for device fabrication and the availability of complex, multilayer doping profiles, such as are used in Read-type IMPATT diodes, device designs for "optimal" parametric amplifier operation become a serious proposition.

In carrying out such a device design, it is important that it be based on and specified in terms of the environmental limitations such as pump power, circuit losses, and matching impedance levels at the signal frequency. Although there have been many detailed and extensive analyses of the varactor-diode parametric amplifier [1], [2], of which the more general references have been cited, there is little information regarding optimal device design in terms of the aforementioned limitations.

In this paper, parametric amplifiers using Read-type HI–LO doping profiles are analyzed, and it is shown that substantial improvements in performance over "conventional" uniformly doped structures are possible with these devices. In particular, there is an optimum way to choose the various doping levels and layer widths and pump the diode to minimize pump power requirements, minimize noise, or maximize dynamic range and insensitivity to pumping level fluctuations. The results are expressed in terms of pump power limitations, unavoidable series resistance levels and negative-resistance levels at the signal frequency, and device area restrictions. An appropriate profile is easily determined once these limitations are specified.

In the following sections a qualitative comparison of conventional and Read-type HI–LO varactors shows the basic differences in operation and identifies the expected advantages of the Read structure. The detailed analysis gives optimal solutions and quantitative device design information.

II. Basic Properties and Advantages of the Read-Type Varactor for Parametric Amplifiers

Many of the advantages Read-type or HI–LO diodes have over conventional diodes for use in parametric amplifiers can be inferred qualitatively by considering their incremental elastances S (elastance $=1/$capacitance) as a function of diode removed charge q_R. The elastance waveform $S(t)$ when the varactor is periodically pumped is important in determining many amplifier properties. The elastances as a function of removed charge for both a one-sided uniform profile and a Read HI–LO profile are shown in Fig. 1 along with $S(t)$ waveforms resulting from *sinusoidal* charge pumping. It is appropriate to assume sinusoidal charge pumping in a parametric amplifier because 1) current or charge pumping has been shown for conventional diodes to result in generally larger modulation ratios[1] than voltage pumping [1], 2) the nature of the varactor model and pump circuitry results in approximately sinusoidal charge pumping at microwave frequencies, and 3) this assumption allows straightforward solutions for various amplifier properties to be obtained such that device and pumping level optimizations are possible.

For both of the cases shown in Fig. 1, the diode is assumed to punch through before breakdown with uniform doping in each layer, and the minimum elastance

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[1]The ratio of the fundamental component of $S(t)$ to the maximum elastance variation possible (S_{max}).

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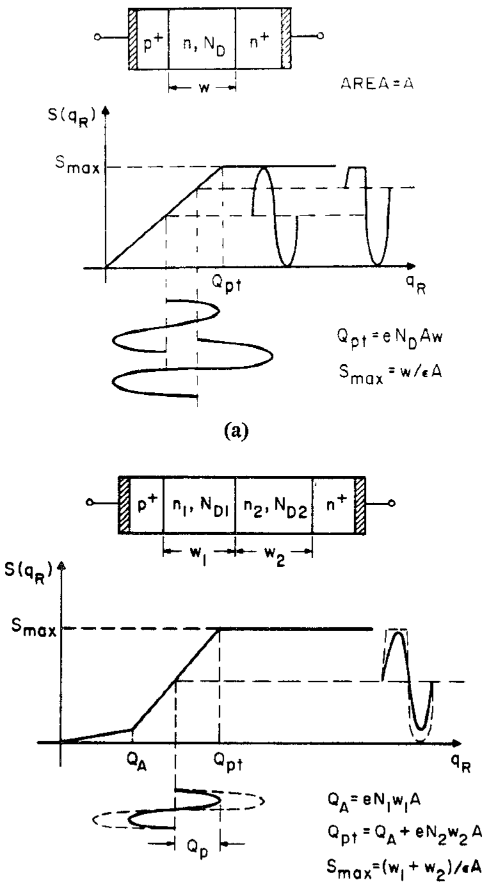

Fig. 1. Elastance waveforms obtainable by sinusoidal pumping of (a) a conventional varactor and (b) a Read-type varactor.

near forward bias ($q_R \triangleq 0$) is taken as zero. The elastance waveforms achievable for the conventional varactor are indicated in Fig. 1(a), where forward bias ($q_R<0$) is avoided primarily for noise minimization. Below punchthrough, the largest modulation ratio achievable is 0.25 and occurs when $q_R(t)=Q_0+Q_1\cos\omega_p t$ has $Q_0=Q_1=Q_{pt}/2$, where ω_p is the pump frequency. Only slightly higher modulation ratios than this occur for pumping above punchthrough; the largest value being 0.268 where $Q_0=Q_1=0.65\ Q_{pt}$.

In the case of the Read profile, depending on the value of Q_A, sinusoidal charge pumping can produce an $S(t)$ approaching a square wave to give a modulation ratio approaching the maximum value [1] of $1/\pi$ or 0.318. Furthermore, the offset Q_A reduces the possibility of forward bias and the associated noise performance degradation. Another apparent advantage with this profile is that of reduced sensitivity of the elastance waveform, and hence amplifier gain, to fluctuations (AM noise) when the $S(t)$ is substantially clipped. In addition, since the parametric amplifier dynamic range limitation is usually the result of signal frequency power levels large enough to affect the pumping and $S(t)$ waveform [1], [2], a strongly driven Read diode should have higher signal saturation levels than coventional structures. This will be demonstrated in the next section.

It may not be apparent that the benefits of the Read diode can be obtained without large charge variations and correspondingly large pump power; however, as the doping level of the low region is reduced, less charge pumping is required to overdrive the diode, but the series resistance goes up correspondingly. Since the series resistance of the diode greatly affects the amplifier properties, an optimum doping profile and pumping level can be derived. For the Read profile, the parameters of interest will be:

1) the profile design for lowest pump power to achieve a given level of negative resistance at the signal frequency;

2) the design for lowest noise at a given value of pump power;

3) the sensitivity of the amplifier to pumping fluctuations and expected improvement in dynamic range. These topics are addressed quantitatively in the next section.

III. Optimization of High–Low Profiles

The equations which govern the varactor diode parametric amplifier for constant series resistance are well known [1], [2], particularly for the simplest case when only the signal, idler, and pump frequencies are considered. Varactor pumping does, in fact, result in periodic modulation of the series resistance and this will additionally contribute to the frequency conversion effects in a parametric amplifier. While in some cases the effect of such resistance modulation may be substantial, for the profiles studied here they modify the optimal solutions and doping parameters by only a few percent while greatly complicating the analysis. Since the fabrication tolerance on any given profile is probably several percent, resistance modulation effects (including thermal noise cross correlations) are neglected. A detailed analysis indicates that the average series resistance as pumped is important but higher frequency components can be neglected. The nonlinear varactor model is shown in Fig. 2(a), with the large- and small-signal models under pumped conditions shown in Fig. 2(b) and (c).

A. Descriptive Equations for the Simplest Case

For small-signal operation of a pumped varactor, the diode voltage v_d and current i_d are related by [1]

$$V_d(t)=S(t)\int i_d(t)dt+R_0 i_d(t)+e_n(t) \tag{1}$$

where R_0 is the average series resistance, $e_n(t)$ is the associated thermal noise, and $S(t)$ is the varactor incremental elastance as pumped which can be expressed as

$$S(t)=\sum_{-\infty}^{\infty} S_K e^{jk\omega_p t}, \qquad S_{-K}=S_K^* \tag{2}$$

where ω_p is the pump frequency. In the case of the simplest amplifier, the diode current $i_d(t)$ is specified as

$$i_d(t)=I_s e^{j\omega_s t}+I_i^* e^{-j\omega_i t} \tag{3}$$

where ω_s is the signal frequency and $\omega_i=\omega_p-\omega_s$ is the idler frequency. Substituting (3) into (1) gives the Fourier volt-

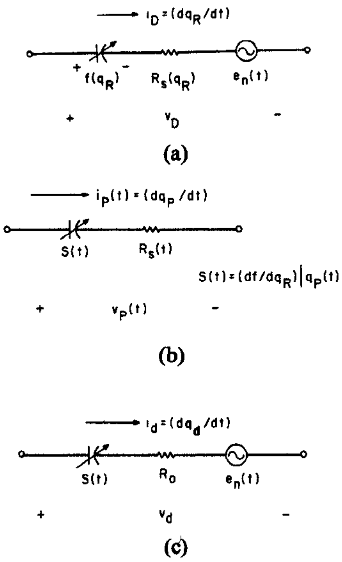

Fig. 2. Various varactor models for analyzing large-signal periodic and small-signal operation. (a) Varactor nonlinear model $[v_D=f(q_R)+R_s(q_R)(dq_R/dt)+e_n(t)]$. (b) Large-signal model of pumped varactor for $q_R=q_P(t)+q_d(t)[v_P(t)=\int S(t)i_P(t)dt+R_s(t)i_P(t)]$. (c) Small-signal model of pumped varactor $[v_d=S(t)\int i_d(t)dt+R_0 i_d(t),\ R_0=\langle R_s(t)\rangle]$.

age amplitudes at the signal and idler frequencies as

$$V_s=\left[R_0+\frac{S_0}{j\omega_s}\right]I_s+\frac{S_1}{-j\omega_i}I_i^*+E_{ns} \tag{4a}$$

and

$$V_i^*=\frac{S_{-1}}{j\omega_s}I_s+\left[R_0+\frac{S_0}{-j\omega_i}\right]I_i^*+E_{ni}^* \tag{4b}$$

where E_{ns} and E_{ni}^* represent the spot noise voltages around ω_s and ω_i, respectively, from the thermal noise $e_n(t)$ and have mean square values in a Δf bandwidth of

$$\overline{E_{ns}^2}=\overline{E_{ni}^2}=4kT_dR_0\Delta f \tag{5}$$

where T_d is the diode temperature. When resistance modulation is neglected, $\overline{E_{ns}E_{ni}^*}=0$.

When the case of a lossless, tuned idler circuit is assumed, i.e.,

$$Z_i^*=-\frac{V_i^*}{I_i^*}=\frac{S_0}{j\omega_i} \tag{6}$$

then, at the signal frequency

$$V_s=Z_{in}I_s+E_n \tag{7}$$

where $Z_{in}=R_{in}+jX_{in}$ with

$$R_{in}=R_0-\frac{|S_1|^2}{\omega_s\omega_iR_0} \tag{8}$$

$$X_{in}=-\frac{S_0}{\omega_s} \tag{9}$$

and

$$E_n=E_{ns}-\frac{S_1}{-j\omega_iR_0}E_{ni}^*. \tag{10}$$

In addition to the input negative resistance level, two other important quantities regarding amplifier performance are the noise measure M defined by

$$M=\frac{T}{T_0}\triangleq\frac{\overline{|E_n^2|}}{4kT_0(-R_{in})\Delta f} \tag{11}$$

where $T_0=290$ K and T is the effective noise temperature [1], and the average pump power P_p given by

$$P_p=\langle R_s(t)i_p^2(t)\rangle \tag{12}$$

where $\langle\ \rangle$ signifies time average, $i_p(t)$ is the pump current, and $R_s(t)$ is the time-varying series resistance. Using (8) and (10) in (11) gives the noise measure as

$$M=\frac{T_d}{T_0}\left(1+\frac{|S_1^2|}{\omega_i^2R_0^2}\right)\Big/\left(-1+\frac{|S_1^2|}{\omega_s\omega_iR_0^2}\right). \tag{13}$$

Equations (8), (12), and (13) will now be evaluated for a varactor diode with a high–low doping profile sinusoidally charge pumped to establish the criteria for optimum performance.

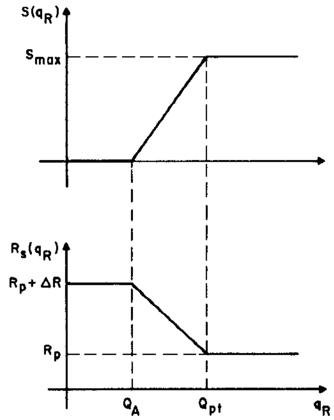

Fig. 3. The incremental elastance $S(q_R)$ and series resistance $R_s(q_R)$ for an idealized Read HI–LO profile.

B. Application to an Idealized High–Low Read Profile

The elastance as a function of removed charge for a HI–LO diode was shown in Fig. 1(b). For analysis, it will be assumed that the diode is a p^+-n_1-n_2-n^+ structure with n_1 and n_2 uniformly doped layers. Furthermore, the n_1-layer is taken as highly doped compared to the n_2-layer such that Q_A is finite but $S(Q_A)\ll S_{max}$. Such profiles are readily achieved in practice. With these restrictions, the elastance-charge $S(q_R)$ and resistance-charge $R(q_R)$ relations are as indicated in Fig. 3. Here, ΔR is the total resistance of the n_2-layer and R_p is the external and contact resistance. For operation like a conventional diode, pumping would be restricted to $Q_A\leqslant q_R\leqslant Q_{pt}$. In terms of the profile parameters

$$S_{max}=\frac{w}{\epsilon A} \tag{14a}$$

$$\Delta R=\frac{w}{q\mu N_DA} \tag{14b}$$

and

$$Q_N=Q_{pt}-Q_A=qN_DwA \tag{14c}$$

where w and N_D are the width and doping of the n_2-layer, respectively, A is the diode area, Q_N is the total charge in the n_2-layer, μ is the mobility, and q is the electronic charge. Now for a given charge pumping, $q_R(t)=q_p(t)>0$, $S(t)$, $R_s(t)$, and R_0 can be found for use in (8), (12), and (13). Since $R_s(q_R)$ can be written as

$$R_s(q_R)=R_p+\Delta R\left[1-\frac{S(q_R)}{S_{\max}}\right] \tag{15}$$

$S(t)$ and $R_s(t)$ have basically the same time dependence for any $q_p(t)$.

Based on previous arguments, the pumping current is assumed sinusoidal, and hence $q_p(t)$ is written as

$$q_p(t)=Q_0+Q_p\cos\omega_p t, \qquad Q_p<Q_0. \tag{16}$$

For maximum modulation ratio $|S_1|/S_{\max}$, it is easily shown that the pumping of $S(q)$ should be symmetrical, i.e., Q_0 is chosen as

$$Q_0=\frac{Q_A+Q_{pt}}{2} \tag{17}$$

and hence

$$\frac{S_0}{S_{\max}}=\frac{1}{2} \tag{18}$$

$$R_0=R_p+\frac{\Delta R}{2} \tag{19}$$

and

$$\frac{S_1}{S_{\max}}=F(\beta)=\begin{cases}\dfrac{\beta}{4}, & \beta\leqslant 1\\[2ex] \dfrac{1}{2\pi}\left[\sqrt{1-\dfrac{1}{\beta^2}}+\beta\sin^{-1}\dfrac{1}{\beta}\right], & \beta>1\end{cases} \tag{20}$$

where

$$\beta\triangleq\frac{2Q_p}{Q_N} \tag{21}$$

is defined as the pump drive factor. Hence $\beta\leqslant 1$ corresponds to typical operation of a conventional varactor.

Using (20) now gives the input resistance as

$$R_{in}=R_0-\frac{F^2(\beta)S_{\max}^2}{\omega_s\omega_i R_0} \tag{22}$$

the noise measure as

$$M=\frac{T_d}{T_0}\left(1+\frac{F^2(\beta)S_{\max}^2}{\omega_i^2R_0^2}\right)\Big/\left(-1+\frac{F^2(\beta)S_{\max}^2}{\omega_s\omega_i R_0^2}\right) \tag{23}$$

and the average pump power as

$$P_p=\frac{\omega_p^2Q_p^2R_0}{2}=\frac{\omega_p^2\epsilon^4A^4}{8\mu^2}\frac{R_0}{\Delta R^2}\beta^2S_{\max}^4. \tag{24}$$

The three parameters which relate to the profile (doping and width) and the diode pumping are $S_{\max}$, ΔR, and β. Thus for given pump and signal frequencies and any area A with associated R_p, these three variables can be adjusted to give optimum performance in some prescribed sense.

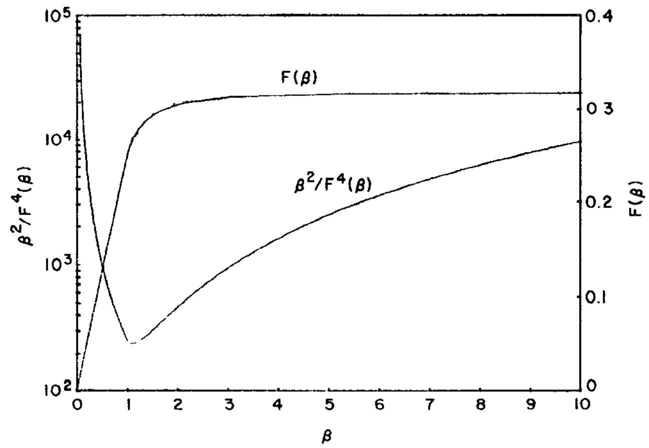

Fig. 4. The functions $F(\beta)$ and $\beta^2/F^4(\beta)$ associated with pumping level.

Of perhaps greatest interest are the two cases given below:

1) maximizing the amount of negative resistance at ω_s for a given amount of pump power or, equivalently, minimizing the pump power needed to achieve a particular amount of negative resistance;

2) minimizing the noise measure for a given amount of pump power.

To carry out the optimizations, the following normalized variables are introduced:

$$\alpha=\frac{\Delta R}{R_p} \tag{25a}$$

$$r=-\frac{R_{in}}{R_p} \tag{25b}$$

and

$$m=M\frac{T_0}{T_d}=\frac{T}{T_d}. \tag{25c}$$

In Case 1, (22) is used to eliminate $S_{\max}$ in (24) to give

$$P_p=P_0G(\alpha,\beta;r) \tag{26}$$

where

$$P_0=\frac{\omega_s^2\omega_i^2\omega_p^2\epsilon^4A^4R_p^3}{\mu^2} \tag{27}$$

and

$$G(\alpha,\beta;r)=\frac{[1+(\alpha/2)]^3[1+(\alpha/2)+r]^2}{\alpha^2}\frac{\beta^2}{F^4(\beta)}. \tag{28}$$

For any value of r, values of α and β can be found to minimize $G(\alpha,\beta;r)$ in the usual way. Shown in Fig. 4 is a plot of $F(\beta)$ along with $\beta^2/F^4(\beta)$, indicating an optimum value β_r very close to 1.1 independent of r, and dictating that the pump power should be used to overdrive the n_2-layer and partially clip the $S(t)$ waveform. The minimum value of $\beta^2/F^4(\beta)$ is 241.4. It is apparent that the benefit of overdrive in the HI–LO profile is inconsequential to that available from a fully pumped conventional diode.[2] However, it will be shown later that overdrive can greatly improve dynamic range and gain sensitivity to pump power.

The optimum value α_r which minimizes G is found by setting $\partial G/\partial\alpha = 0$ and is given by

$$\alpha_r = \sqrt{\left[\frac{1+r}{3}\right]^2 + 8\left[\frac{1+r}{3}\right]} - \frac{1+r}{3}. \quad (29)$$

For small $r, \alpha_r \to 4/3$, while for large $r, \alpha_r \to 4$. Hence the value of the n-layer resistance ΔR is optimally a few times R_p. With respect to α and β, $G(\alpha, \beta; r)$ has rather broad minimums so that these parameters are not overly critical.

From the optimum values α_r and β_r, S_{max}, w and the doping N_D can be determined using (14). The width and doping are given by

$$w = w_0 \frac{\{[1+(\alpha_r/2)][1+(\alpha_r/2)+r]\}^{1/2}}{F(\beta_r)} \quad (30a)$$

where

$$w_0 = \epsilon A R_p \sqrt{\omega_s \omega_i} \quad (30b)$$

and

$$N_D = N_0 \frac{\{[1+(\alpha_r/2)][1+(\alpha_r/2)+r]\}^{1/2}}{\alpha_r F(\beta_r)} \quad (31a)$$

with

$$N_0 = \frac{\epsilon\sqrt{\omega_s \omega_i}}{q\mu}. \quad (31b)$$

Of interest is that N_0 is independent of A and R_p and increases with frequency.[3]

In the minumum noise case, (24) is used to eliminate S_{max} in (23) to give

$$m = \frac{1 + KR\dfrac{F^2(\beta)\alpha}{\beta[1+(\alpha/2)]^{5/2}}}{-1 + K\dfrac{F^2(\beta)\alpha}{\beta[1+(\alpha/2)]^{5/2}}} \quad (32)$$

where

$$K = \sqrt{\frac{8P_p}{P_0}} \quad (33)$$

and

$$R = \omega_s/\omega_i. \quad (34)$$

To optimize $M > 0$ with respect to α and β for any value of K and R, the value of $F^2(\beta)\alpha/\beta[1+(\alpha/2)]^{5/2}$ must be maximized. The maximum value of $F^2(\beta)/\beta$ occurs at β_m which is the same as that found for the minumum value of $\beta^2/F^4(\beta)$ in Case 1, i.e., $\beta = 1.10$. The value α_m which maximizes $\alpha/[1+(\alpha/2)]^{5/2}$ is easily determined as $\alpha_m = 4/3$. Since α_m and β_m are constants, the values of doping and width can be found for any value of r by using (30) and (31) with α_m and β_m replacing α_r and β_r. Also, for a given value of r, α_m and β_m can be used in $G(\alpha, \beta; r)$ to obtain the normalized pump power level P_p/P_0. Shown in Figs. 5 and 6 are plots P_p/P_0, m at maximum negative resistance, w/w_0, and N_D/N_0 as functions of r for both Case 1 and Case 2. As observed in Fig. 5 at a given value of pump power, a minimum noise solution results in only a slightly lower value of negative resistance than the maximum possible; in fact, the actual noise measure is fortunately only slightly different for the two cases. The ratio for the same T_d of the lowest noise temperature T_m to the noise temperature T_r at maximum negative resistance is shown as a function of pump power for several ω_s/ω_i values in Fig. 7. Less than 1-dB improvement is possible. Also shown in the same figure is the degradation in negative resistance occurring when a minimum noise solution is used.

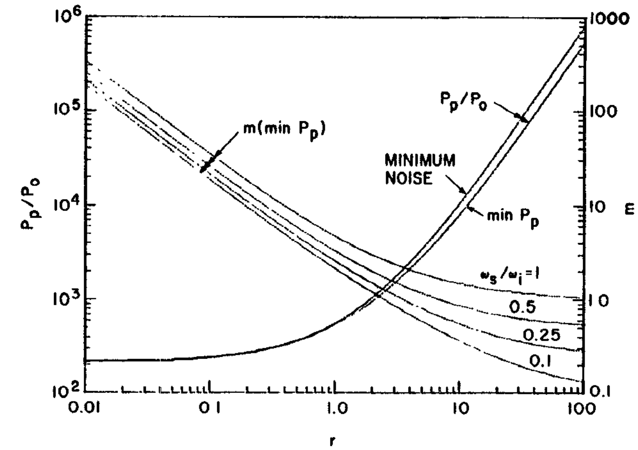

Fig. 5. Minimum pump power requirements and noise measure as a function of normalized resistance level r.

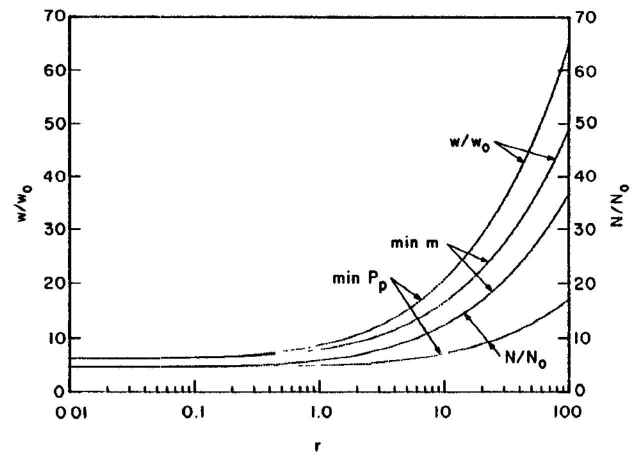

Fig. 6. Optimum profile doping and width of the LO region as a function of normalized resistance r.

Optimum doping levels and widths of the LO region, the minimum pump power, and the resulting noise temperature m for 10- and 100-GHz GaAs diode parametric

[2] $\beta = 1.0$ is equivalent to a fully pumped conventional diode.

[3] Note that $Q_s = |X_{in}/R_{in}|$ at ω_s is obtained from $2\sqrt{(\omega_s/\omega_i)}\, Q_s = (w/w_0)/r$ for any r.

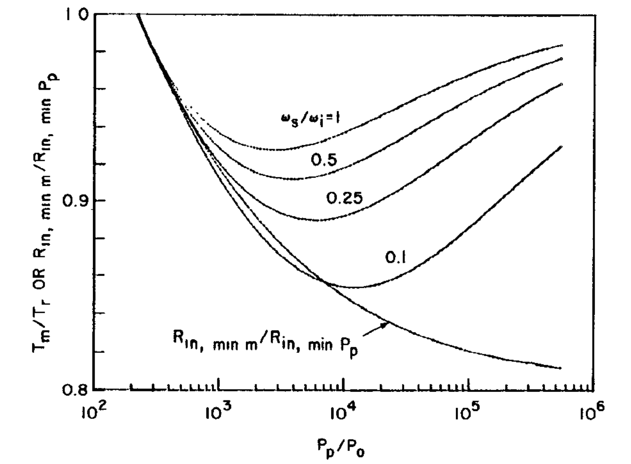

Fig. 7. Ratio of noise measures and negative resistance levels for the minimum pump power case and minimum noise measure case.

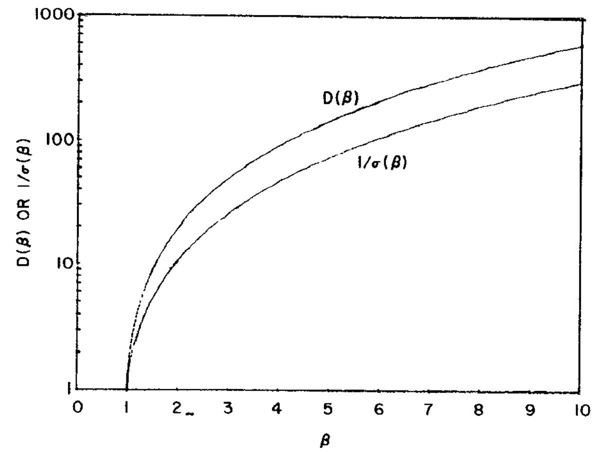

Fig. 10. Dynamic range improvement or reduction in resistance sensitivity as a function of drive factor β.

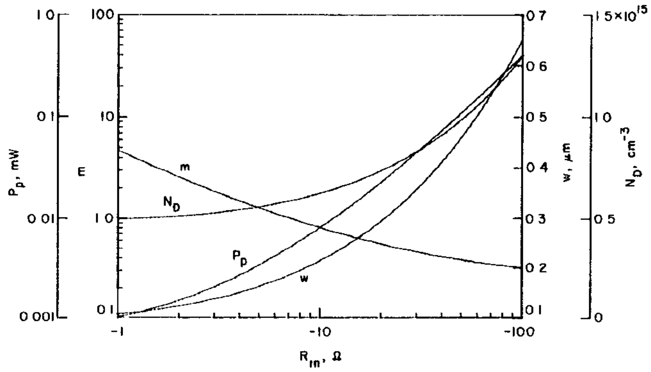

Fig. 8. Profile parameters, minimum pump power requirements, and noise temperature for a 10-GHz parametric amplifier pumped at 50 GHz. $f_p=50$ GHz; $f_s=10$ GHz; diameter$=25$ μm; $R_p=2$ Ω.

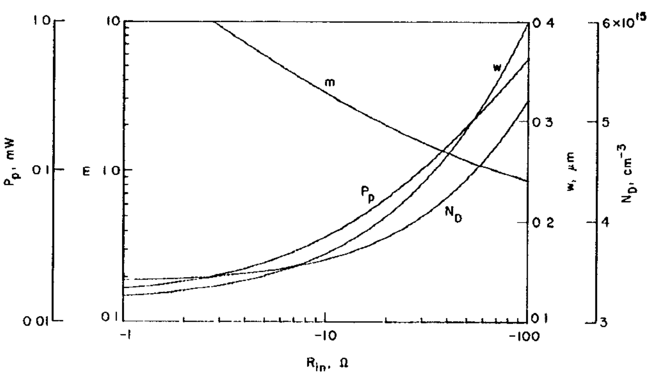

Fig. 9. Profile parameters, minimum pump power requirements, and noise temperature for a 100-GHz parametric amplifier pumped at 300 GHz. $f_p=300$ GHz; $f_s=100$ GHz; diameter$=5$ μm; $R_p=10$ Ω.

amplifiers are shown in Figs. 8 and 9, respectively, as functions of the negative resistance at the signal frequency. A diode area and corresponding external series resistance must be specified for these calculations, and the doping of the HI region is assumed large compared to that of the LO region. This assumption is justified because of the doping levels obtained. If the diode area can be halved at the same value of R_p, the pump power requirements are reduced by a factor of 16. It should be noted that, from the values of dopings and widths obtained for the LO region, the relatively high doping and low width required of the HI region to satisfy the assumption of Fig. 3 should be easily obtainable with present fabrication technology. Hence the original assumption regarding the profile is reasonably self-consistent.

As noted earlier, the $\beta^2/F^4(\beta)$ curve of Fig. 4 indicates a negligible reduction in pump power from clipping of the $S(t)$ waveform, implying little benefit in this regard from the Read profile. However, clipping of the $S(t)$ waveform can greatly improve the *dynamic range* of the parametric amplifier as well as substantially reduce the sensitivity of the amplifier gain to fluctuations in the pump power. The dynamic range will be defined as the amount of signal input power P_s required to cause a change in gain ΔG in the amplifier. Following the approach of Penfield and Rafuse [1] yields

$$P_{s,\,\max}=P_p\left(\frac{\Delta G}{G}\right)\frac{\sqrt{G}}{(G-1)^2}\frac{\omega_s}{\omega_p}D(\beta) \tag{35}$$

where

$$D(\beta)=\begin{cases}1, & \beta\leqslant 1\\ \dfrac{[2\pi F(\beta)]^2}{\left[\beta\sin^{-1}\dfrac{1}{\beta}\right]\left[\beta\sin^{-1}\dfrac{1}{\beta}-\sqrt{1-(1/\beta^2)}\right]}, & \beta>1.\end{cases} \tag{36}$$

This result agrees with that of [1] for $\beta\leqslant 1$. The function $D(\beta)$ increases rapidly as a function of β, as shown in Fig. 10. An appropriate interpretation of (35) is that for a given gain G and specified ΔG, the value of $P_{s,\,\max}$ with respect to P_p increases with the drive factor β. Hence for a given amount of P_p and realized gain G, clipping of the $S(t)$ waveform can afford a substantial improvement in dynamic range. At the optimum β of 1.1, $D(\beta)$ is approximately 5 dB. Of course, for fixed P_p, as β deviates from 1.1, the negative-resistance level will decrease slightly and make the matching problem to achieve a gain G at ω_s somewhat more difficult.

Another benefit of overdrive is in the fractional sensitivity of R_{in} to fractional variations in the pump power. Calculating this sensitivity gives

$$\sigma(\beta)=\frac{dR_{in}/R_{in}}{dP_p/P_p}\simeq\frac{\beta\frac{dF(\beta)}{d\beta}}{F(\beta)}$$

$$=\begin{cases}\dfrac{\beta\sin^{-1}\frac{1}{\beta}-\sqrt{1-(1/\beta^2)}}{\beta\sin^{-1}\frac{1}{\beta}+\sqrt{1-(1/\beta^2)}}, & \beta\leqslant 1.\\ 1, & \beta\leqslant 1.\end{cases} \quad (37)$$

and a plot of $1/\sigma(\beta)$ is also indicated in Fig. 10 as a function of β. From the curves of Fig. 10 it is apparent that substantial benefits in parametric amplifier performance are available by using Read-type or HI–LO varactors.

IV. Conclusions

It has been shown that the performance of parametric amplifiers can be improved in general and specifically by use of Read-type HI–LO varactors. Optimum values for doping levels and widths of the LO region and pump drive levels have been specified which optimize certain performance criteria when external limitations can be specified. As these diodes become more available, their use in parametric amplifiers will increase, particularly at millimeter-wave frequencies where pump power limitations and circuit losses are most severe.

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A 40-GHz Digital Distribution Radio with a Single Oscillator

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Abstract—**New 40-GHz band digital radio equipment is described. In the equipment we adopted a new circuit configuration consisting of a single IMPATT diode oscillator which functions as both transmitter frequency converter and receiver local oscillator simultaneously. The principal system design factors, a unique IMPATT diode oscillator mount configuration, and test results are described. The compact radio equipment is designed so that it ensures excellent cost performance for communication systems in local trunk service, even in short hop applications resulting from rainfall attenuation in the new band.**

I. Introduction

IN INTRACITY and suburban areas, voice and data business communications are expanding rapidly. In such areas it is difficult to use buried cables, while there also exists the problem that the radio spectrum is very congested. Under these circumstances, in late 1974 the FCC made the 40-GHz band available for such applications as local trunk service.

In this paper, newly developed 40-GHz digital radio equipment is described. For this radio equipment to be used in local trunk service, it must firstly be of low cost, since the channel loadings are small, i.e., up to 24 channels, and the radio span is only up to ten miles. Secondly, it must interface with the digital signal format, i.e., PCM, which is economically appropriate for a variety of signals such as voice, data, FAX, etc. Thirdly, it must be easy to install and maintain

Considering the above requirements, we have successfully developed a unique transmitter–receiver configuration consisting of a single IMPATT diode oscillator which functions as transmitter oscillator, up converter, and receiver local oscillator simultaneously. As a result of this, we achieved considerable cost reduction while obtaining smaller size, lower power consumption, better reliability, and easier maintenance. The waveguide circuits, which

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