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Theoretical Investigation of Large-Signal Noise in Nanometric Schottky-Barrier Diodes Operating in External Resonant Circuits

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We report Monte Carlo simulations of electronic noise in heavily doped nanometric GaAs Schottky-barrier diodes operating in series with a parallel resonant circuit when a high-frequency large-signal voltage is applied to the whole system. Significant modifications of the noise spectrum with respect to the unloaded diode are found to occur in the THz-region.

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1. Introduction

One of the most promising directions to develop solid-state sources of THz radiation working at room temperature is a frequency multiplication in GaAs Schottky-barrier diodes (SBDs) [1]. To extract the high-frequency signal, the SBD is usually loaded by a given external resonant circuit, which, due to feedback, can modify considerably the intrinsic noise spectrum of the unloaded SBD [2]. The aim of this article is to study such an influence by both an analytical approach and Monte Carlo particle (MCP) simulations of the circuit performance of a nanometric GaAs n^+n-metal SBD of Ref. [2] operating under cyclostationary conditions.

2. Results and discussion

We consider a third harmonic generation in a SBD connected in series with a parallel resonant circuit consisting of a load resistance R, an inductance L and a capacitance C_l as shown schematically in Fig. 1a when a microwave voltage $U(t) = U_0 + U_1 \sin(2\pi\nu_0 t)$ at frequency $\nu_0 = 200$ GHz is applied to the whole sys-

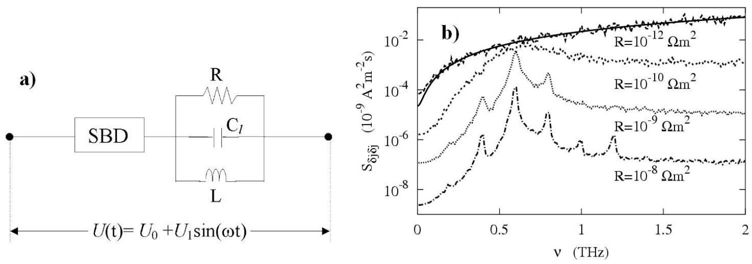

Fig. 1. (a) Scheme of the series connection of the SBD with a parallel resonant circuit, and (b) spectral density of current fluctuations through various load resistances R calculated by MCP approach.

tem and output power is extracted from R. For the scheme presented in Fig. 1a, the circuit equations can be written as:

$$C_l \frac{\mathrm{d}U_\mathrm{c}}{\mathrm{d}t} + j_L + \frac{U_\mathrm{c}}{R} = j_\mathrm{d}, \qquad L\frac{\mathrm{d}j_L}{\mathrm{d}t} - U_\mathrm{c} = 0, \qquad U_\mathrm{d} + U_\mathrm{c} = U(t), \tag{1}$$

where U_d and U_c are the voltage drops at the terminals of the SBD and the external circuit, respectively, j_L is the current density flowing through the inductance L,

$$j_\mathrm{d}(U_\mathrm{d}, t) = j_\mathrm{d}^\mathrm{RS}(t) + C_\mathrm{g}\frac{\mathrm{d}U_\mathrm{d}}{\mathrm{d}t} \tag{2}$$

is the total current density flowing through the SBD, $j_\mathrm{d}^\mathrm{RS}(t)$ — the drift component of the current directly calculated by the MCP procedure in accordance with the Ramo–Shockley theorem, C_g — the geometrical capacitance of the SBD. In calculations, the capacitance of the external circuit is $C_l = 0.5C_\mathrm{g}$ and the tuning of the resonator is performed by the external inductance which, for the resonance at the frequency of the third harmonic, $\nu_\mathrm{res} = 3\nu_0 = 600$ GHz, is equal to $L = 1.02 \times 10^{-23}$ H m^2.

Figure 1b reports the results of MCP simulations for the spectral density of current fluctuations $S^R_{\delta j \delta j}(\nu)$ through different load resistors R. For comparison, the solid line shows the spectral density of current fluctuations for the unloaded SBD, $S_{\delta j \delta j}(\nu)$, under the same biasing conditions $U_0 = 0.6$ V, $U_1 = 0.2$ V, $\nu_0 =$ 200 GHz. When the load resistance is very small, the short circuit operation holds and the spectrum of $S^R_{\delta j \delta j}(\nu)$ practically coincides with $S_{\delta j \delta j}(\nu)$ of the unloaded SBD. With the increase in R, we observe: (i) a suppression of the low- and high-frequency wings of the spectrum centered on the frequency of the third harmonic, (ii) the formation of a peak centered around the frequency of the third harmonic as well as the onset of a series of additional peaks centered around other harmonics.

For the analytical consideration the current flowing through the SBD is approximated as:

$$j_\mathrm{d}(U_\mathrm{d}) = j_\mathrm{d}^\mathrm{s}(U_d) + \frac{d}{dt}Q^\mathrm{s}(U_\mathrm{d}), \tag{3}$$

where the drift component, $j_{\mathrm{d}}^{\mathrm{s}}(U_{\mathrm{d}})$, is taken from the static current–voltage ($I-V$) relation, and the SBD varactor capacitance, $C_v(U_{\mathrm{d}}) = \mathrm{d}Q^{\mathrm{s}}(U_{\mathrm{d}})/\mathrm{d}U_{\mathrm{d}}$, is included as the rate of time-variations of the SBD static total charge $Q^{\mathrm{s}}(U_{\mathrm{d}})$ induced by variations of U_{d}. By combining Eqs. (1) to (3), linearizing them with respect to fluctuations, and going to the spectral representation, one obtains finally fluctuations of voltage drop as:

$$\delta U_{\mathrm{d}}(\omega) = -Z(\omega)[\delta j_{\mathrm{d}}(\omega) + i\omega \sum_{n\neq 0}^{\pm\infty} C_n^v \delta U_{\mathrm{d}}(\omega - \omega_n)], \tag{4}$$

where

$$Z(\omega) = \frac{\mathrm{i}\omega}{\overline{C}(\omega_{\mathrm{res}}^2 - \omega^2 + \mathrm{i}\omega\nu_{RC})} \tag{5}$$

is the net small-signal impedance of the considered system (SBD + the resonant circuit) which describes the fluctuations of the voltage drop at the SBD. Here $\overline{C} = C_l + C_0^v$, $\omega_{\mathrm{res}}^2 = 1/\overline{C}L$, $\overline{R} = RR_{\mathrm{d}}{}^0/(R + R_{\mathrm{d}}^0)$, $\nu_{RC} = 1/(\overline{CR})$, R_{d}^0 is the SBD average resistance, C_n^v the Fourier coefficients of the varactor capacitance C_v.

In the square brackets of the r.h.s. of Eq. (4), the first term corresponds to the noise source related to fluctuations of the drift current flowing through the unloaded SBD. The second term describes the influence of the $C-V$ nonlinearity on voltage-drop fluctuations, $\delta U_d(\omega)$, when the SBD operates in an external resonant circuit, and it can be considered as an additional noise source induced by the harmonic modulation of the varactor capacitance. However, the dependence of $\delta U_{\mathrm{d}}(\omega)$ upon $\delta j_{\mathrm{d}}(\omega)$ becomes implicit at a given frequency ω since the second term in Eq. (4) involves frequency mixing. Such an implicit dependence can be resolved by using sequential iterations with respect to C_n^v. In the zero-order approximation, by omitting the second term in square brackets of Eq. (4), the spectral density of voltage-drop fluctuations takes the form:

$$S_{U_{\mathrm{d}}U_{\mathrm{d}}}^0(\omega) = |Z(\omega)|^2 S_{jj}^{\mathrm{d}}(\omega), \tag{6}$$

where $S_{jj}^{\mathrm{d}}(\omega)$ is the spectral density of the current noise source of the unloaded SBD. Figure 2a compares the result of the zero-order approximation with MCP results at low and high values of the load resistance. As follows from Fig. 2a, the zero-order approximation provides a good description of the $S_{U_{\mathrm{d}}U_{\mathrm{d}}}(\omega)$ spectrum in practically the whole frequency range of interest, including the resonant frequency $\nu_{\mathrm{res}} = 3\nu_0$ at which the external circuit is tuned. However, it does not describe extra resonance-like noise contributions which appeared around frequencies $\nu_{\mathrm{res}} \pm n\nu_0$. To understand their origin the influence of the $C-V$ nonlinearity on the spectrum of fluctuations must be taken into account. In the first-order approximation one obtains:

$$S^1_{U_\mathrm{d}U_\mathrm{d}}(\omega) = |Z(\omega)|^2[S^\mathrm{d}_{jj}(\omega) + (\omega)^2 \sum_{n\neq 0}^{\pm\infty} |C^v_n|^2 |Z(\omega-\omega_n)|^2 S^\mathrm{d}_{jj}(\omega-\omega_n)], \qquad (7)$$

where the second term can be considered as an additional source of fluctuations of the current flowing through the SBD induced by the harmonic modulation of the varactor capacitance. Accordingly, the modified total spectrum is shown in Fig. 2b, which reports the first-order iteration for $S_{U_\mathrm{d}U_\mathrm{d}}(\omega)$ in comparison with direct MCP simulations (respectively, solid and dotted lines).

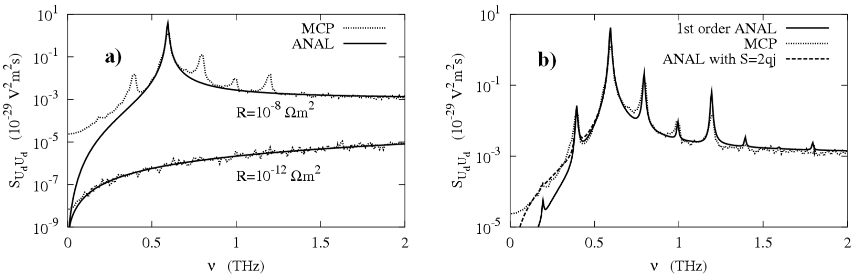

Fig. 2. Spectral densities of the total voltage drop fluctuations on the load resistor R calculated by the MCP technique (dotted curves) and by the analytical approach in: (a) zero-order and (b) first-order approximations (solid lines).

As follows from Fig. 2b, the first-order iteration well describes qualitatively and quantitatively the origin of the extra noise at the harmonics of the applied signal in the whole frequency region of interest excluding the low-frequency region where the noise remains underestimated with respect to the MCP results. The origin of this discrepancy is related with the increase in the mean current $\overline{j}_0$ associated with the increase in R which is not accounted for by the noise source of the unloaded SBD. By using the corrected shot-noise value one can recalculate the circuit noise. The result is presented in Fig. 2b by the dashed line which practically coincides with MCP results.

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