Sensing Microwave-Terahertz Detectors Based on Metal-Semiconductor-Metal Structures With Symmetrical *I–V* Characteristic

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***Abstract*—We discuss the characteristics of new sensing elements based on a symmetrical metal-semiconductor-metal structure, which are designed for detection of microwave-terahertz signals. The schemes of connection of the sensing elements in the high-frequency path and the low-frequency measuring circuit are considered. Expressions for the volt-watt sensitivity and the noise-equivalent power are obtained. Comparison of the obtained characteristics with those of the detector zero-bias Schottky (Mott)-barrier diode is carried out. It is shown that the characteristics of the symmetrical sensing elements are comparable with or exceed in some cases similar characteristics of the detector zero-bias diodes.**

***Index Terms*—Metal-semiconductor-metal structure, microwave-terahertz detector, schottky (Mott)-barrier transition.**

I. Introduction

THE ASSIMILATION of the terahertz (THz) frequency range (0.1–10 THz) has been going on for over 30 years [1]. However, the problem of creating a simple, robust, and obtainable radiation detector remains topical to date. While until recently the detectors were mainly developed on the basis of bolometers and mixers, there is now an increase in interest in detector systems. Detector receivers are suitable for applications that do not require an ultrahigh spectral resolution. Their use simplifies the multi-element system of detection and the scheme of registering of signals at low frequencies compared with the systems based on heterodyne detectors and provides a gain in speed compared with the bolometers [2]. Structures with asymmetric nonlinear current-voltage (*I-V*) characteristic are commonly used as the sensing elements of the detectors. As an example, we mention the interband tunneling diodes [3], planar doped barrier diodes [4], semimetal-semiconductor transition diodes [5], and Schottky (Mott)-barrier diodes [6]–[9].

In this paper, we study the possibility of using as a sensing element a microwave or terahertz detector with a symmetrical structure containing metal-semiconductor-metal (MSM) Schottky (Mott)-barrier transitions connected in series. The barrier transitions connected in such a way have a symmetrical *I-V* characteristic, and the dc voltage component does not appear at the metal leads when a sinusoidal voltage is applied to them. However, each half-period of the voltage biases one of the metal-semiconductor (MS) periods in the forward direction, causing a flow of direct current. During the transition process, this current removes part of the charge carriers from the semiconductor. In the steady state, the semiconductor charge is different from the equilibrium one, and a dc reverse voltage is present in each transition. The principle of operation of the detector with the proposed sensing element consists in registering of this voltage in several possible ways.

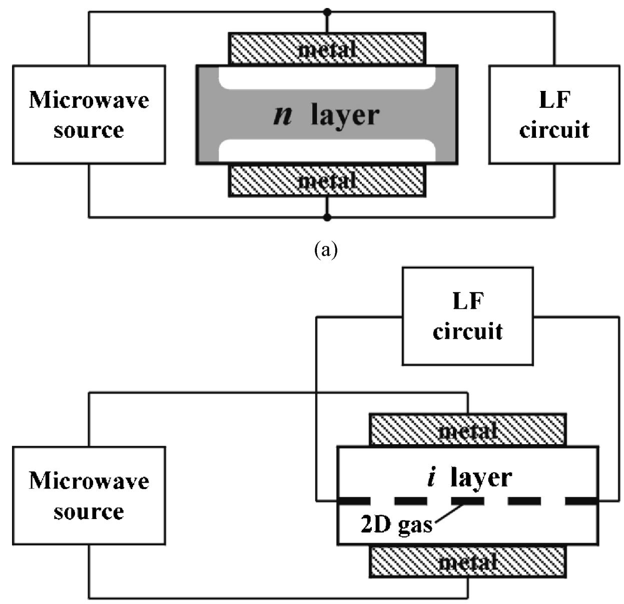

Fig. 1. Metal-semiconductor-metal sensing elements with vertical geometry. (a) Two-terminal network. (b) Four-terminal network.

We consider two types of structures used as the sensing elements. The structures of the first type are two-terminal networks. Fig. 1(a) shows one of their possible designs. The metal layers in this structure are applied on the different sides of a thin doped semiconductor layer. The planar geometry, in which the metal layers are deposited next to each other on one side of a thin doped semiconductor layer formed on a strongly doped semiconductor substrate, is also feasible. The reverse

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voltages occurring in the MS transitions when a microwave signal is fed to a two-terminal network change the differential resistance and the differential capacitance of each transition. As a result, the impedance of the two-terminal network also changes. This effect can be registered using a low-frequency measuring circuit or on the direct current.

The structures of the second type are four-terminal networks. In these structures, a conducting channel with the low-frequency circuit connected to it is formed under the MS transitions in the semiconductor. The reverse voltages occurring in the transitions when a microwave signal is fed to them change the carrier density in the channel due to the field effect. As a result, the conductivity of the channel also changes, and this effect is registered by a low-frequency measuring circuit or on the direct current. Fig. 1(b) shows one of the possible designs of the four-terminal network. In this structure, the conducting channel is a two-dimensional electron gas in the semiconductor i layer. The geometry can be planar when the barrier transitions are close to each other. The doping can be uniform in the semiconductor layer. In this case, the geometry can be more complex, with the multilateral girth of the channel by metal layers [10].

To match the sensing element of the detector with the antenna and the low-frequency circuit, it is needed to decrease its differential resistance. This is achieved by using MS transitions with low effective barrier height. Such a lowering of the effective barrier height can be provided by different methods, for example, by δ-doping of a semiconductor near its boundary with the metal [11], [12].

To demonstrate the described effect, the static *I-V* characteristics of the MSM structures with low-barrier Schottky transitions are shown in Fig. 2. The structure had a planar geometry on a strongly doped n^+ substrate (see the inset in Fig. 2). The area of each barrier transition was $50 \times 50\,\mu\text{m}^2$. The lowering of the effective Schottky barrier height was achieved by surface δ-doping of the semiconductor layer by silicon [12]. The effective barrier height of the transitions amounted to 0.29 eV. The semiconductor n layer was 400 nm thick and the doping level was $10^{16}\,\text{cm}^{-3}$. The *I-V* characteristics were measured under conditions where the structure is affected by a sinusoidal voltage $U = U_0 \cdot \sin(\omega t)$ with the frequency $f = \omega/2\pi = 1\,\text{MHz}$. The curve with the maximum slope corresponds to the zero amplitude of the sinusoidal voltage, the intermediate slope, to the amplitude $U_0 = 30\,\text{mV}$, and the minimum slope, to the amplitude $U_0 = 50\,\text{mV}$.

II. The Measurement Scheme with a Two-Terminal Network

The Consider the structure of a two-terminal network shown in Fig. 1(a). The structure represents two Schottky-barrier transitions connected in series to meet each other. Fig. 3 shows the equivalent scheme of a two-terminal network. Here, $C(V_1)$ is the capacitance of the forward-biased transition, which depends on the voltage V_1 in it, $C(-V_2)$ is the capacitance of the backward-biased transition, which depends on the voltage V_2 in it, the functions $I(V_1)$ and $-I(-V_2)$ describe the *I-V* characteristics of the corresponding transitions, and r is the

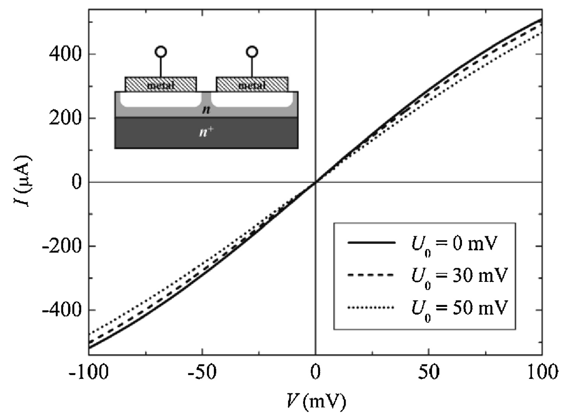

Fig. 2. *I–V* characteristics of the metal-semiconductor-metal structure with low-barrier Schottky transitions with the different-amplitude sinusoidal voltage affecting the structure. Inset: the structure of the two-terminal network with planar geometry.

resistance of the quasi-neutral region of the n layer, which is connected in series with the transitions. Let a sinusoidal voltage at a frequency ω with amplitude U_0 and a dc voltage V_0 be applied to the two-terminal network. The time dependences of the voltages and currents are described by the following system of equations

$$\begin{cases} \left[C\left(V_1\right)\dfrac{dV_1}{dt} + I\left(V_1\right)\right] r + V_1 + V_2 = U_0\left[V_0/U_0 + \sin\left(\omega t\right)\right] \\ \left[C\left(-V_2\right)\dfrac{dV_2}{dt} - I\left(-V_2\right)\right] r + V_1 + V_2 = U_0\left[V_0/U_0 + \sin\left(\omega t\right)\right] . \end{cases} \tag{1}$$

We assume that the voltages U_0 and V_0 are small compared with the characteristic voltages of the nonlinearity for the conduction $\left(\frac{dI}{dV}\middle/\frac{d^2I}{dV^2}\right)\Big|_{V=0}$ and the displacement $\left(C\middle/\frac{dC}{dV}\right)\Big|_{V=0}$ currents. For the solution of system (1) we represent the dependences of the transition capacitance and the conduction current in the form of the voltage series expansions

$$C\left(V\right) = C_0 + b\,V \tag{2}$$

$$I\left(V\right) = \frac{1}{R_0}\,V + \alpha\,V^2 + \gamma\,V^3 \tag{3}$$

where C_0 and R_0 are the differential capacitance and the differential resistance of the MS transition for a zero voltage, and

$$b = \left.\frac{dC}{dV}\right|_{V=0} \tag{4}$$

$$\alpha = \frac{1}{2}\left.\frac{d^2I}{dV^2}\right|_{V=0} \tag{5}$$

$$\gamma = \frac{1}{6}\left.\frac{d^3I}{dV^3}\right|_{V=0}. \tag{6}$$

Substituting (2) and (3) into (1), we seek a solution to the system of equations in the form of the voltage series expansions $V_1(t)$ and $V_2(t)$ over powers of the small parameter $2\alpha R_0 U_0$ up to the third power inclusive. The characteristic time of transition to the steady-state oscillation is $\tau = R_0 C_0$.

We calculate the total current through a two-terminal network and the power $P_\omega = I_\omega(t) \cdot U_0 \sin(\omega t)$ absorbed at the frequency ω. Here, $I_\omega(t)$ is the current at the frequency ω. Averaging the current and power over the microwave oscillation period, we find the small change ΔR_{TT} in the differential resistance of a two-terminal network on the direct current, which is divided by the average absorbed power $\langle P_\omega \rangle$.

Let us discuss the scheme of the low-frequency circuit. The two-terminal network is connected to the transmission line of a microwave signal or directly to the antenna, which receives microwave radiation, and is connected in series, through a low-pass filter, to the dc circuit with the load resistance R_l and the source of dc voltage E (Fig. 3). The measured signal is the on-load voltage. The volt-watt sensitivity of the detector with a two-terminal network is given by

$$S_{TT} = \frac{\Delta R_{TT}}{\langle P_\omega \rangle} \frac{R_l \mathrm{E} K}{(R_l + R_{TT})^2}$$
$$= \left\{ \frac{(\omega\tau)^2 \frac{\alpha b}{C_0} + \left[3 + (\omega\tau)^2\right] \alpha^2 R_0}{1 + (\omega\tau)^2} - \frac{3}{2}\gamma \right\}$$
$$\times \frac{4 R_0^2 R_{TT} R_l \mathrm{E} K}{\left(2 + r/R_0\right)^2 \left(1 + \omega^2 C_0^2 \frac{r R_0}{(2 + r/R_0)}\right) (R_l + R_{TT})^2} \quad (7)$$

where K is the ratio of the microwave signal power absorbed in the two-terminal network to the incident power. Intrinsic resistances of the filter and the voltage source are neglected. Assume that the *I-V* characteristic of the transitions has the exponential dependence [6]

$$I = S A_R T^2 \exp\left(-\frac{\Delta}{kT}\right) \left[\exp\left(\frac{qV}{n\,kT}\right) - 1\right] \quad (8)$$

where n is the non-ideality factor, k is the Boltzmann constant, T is the temperature, q is the elementary charge, A_R is the effective Richardson constant, Δ is the effective barrier height of the transition, and S is the transition area. Let us choose the bias voltage $\mathrm{E} = 2nkT/q$, which corresponds to the voltage drop less than nkT/q in each transition. When the matching conditions $R_l = R_{TT}$ are fulfilled in the dc circuit, from (7) and (8) we obtain the following expression for the volt-watt sensitivity in the high-frequency region $\omega\tau >> 1$

$$S_{TT} \approx \frac{b}{C_0} \frac{R_0}{\left(2 + r/R_0\right)^2 \left(1 + \omega^2 C_0^2 \frac{r R_0}{\left(2 + r/R_0\right)}\right)} K. \quad (9)$$

Here, b/C_0 is the nonlinearity parameter of the capacitive current. If the capacitance of the transitions is voltage independent ($b = 0$), then the effect of increasing differential resistance of the MSM structure is absent at high frequencies. In a similar way, for the low-frequency region $\omega\tau << 1$ we obtain

$$S_{TT} \approx \frac{q}{n\,kT} \frac{R_0}{\left(2 + r/R_0\right)^2 \left(1 + \omega^2 C_0^2 \frac{r R_0}{\left(2 + r/R_0\right)}\right)} K. \quad (10)$$

In this case, the effect is proportional to the nonlinearity parameter q/nkT of the conduction current.

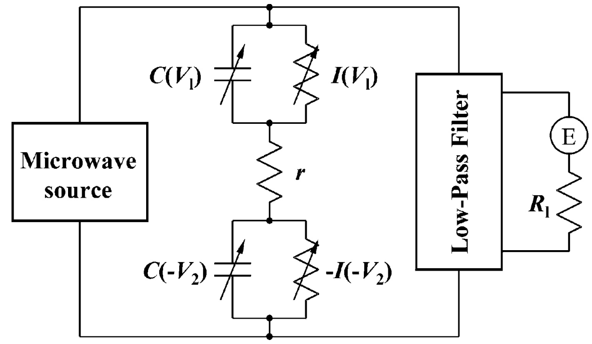

Fig. 3. Equivalent scheme of connection of the two-terminal network.

Assume that the thermal noise determined by the differential resistance of the transitions near the zero voltage and by the load resistance gives the main contribution to the detector noise at small voltages $|V_1|, |V_2| < nkT/q$. In this case, the noise-equivalent power of the detector with a two-terminal network is given by

$$\mathrm{NEP}_{TT} = \frac{\sqrt{4kT\,B \frac{R_{TT}}{2}}}{S_{TT}} = \frac{\sqrt{4kT\,B\left(R_0 + r/2\right)}}{S_{TT}} \quad (11)$$

where B is the frequency band.

We note that when a microwave signal is fed to the two-terminal network, simultaneously with the increase in its differential resistance, its differential capacitance changes (the sign of the change is signal-frequency dependent). Hence, the principle of operation of the detector can be based on measurement of the capacitance of the two-terminal network by using a low-frequency sounding signal.

III. The Measurement Scheme With a Four-Terminal Network

Consider the structure of a four-terminal network shown in Fig. 1(b). As in the previous case, a microwave signal is fed to the metal leads of the series-connected metal-semiconductor-metal transitions with Mott barriers. The equivalent scheme of the microwave circuit of a four-terminal network is completely similar to the equivalent scheme of a two-terminal network considered before. The charge $Q(V_1)$ in the channel under the forward-biased transition and the charge $Q(-V_2)$ in the channel under the backward-biased transition are functions of the voltage: $Q(V) = Q_0 + C_0 V + \frac{b}{2} V^2$. Using the solution to (1) obtained earlier for the particular case $V_0 = 0$, we find the time dependences of the charge values in the channel under the transitions. Averaging over the microwave oscillation period, we find the relative change in the charge in the channel under the transitions, which is correspond to the average power absorbed in the transitions $\frac{\langle \Delta Q \rangle}{Q_0 \langle P_\omega \rangle}$. The conducting channel of the four-terminal network is connected to the dc measuring circuit. As in the case of a two-terminal network, this circuit consists of the source of dc voltage E and the load R_l. The small relative change in the charge of the channel under the transitions is equal to the small relative change $\frac{\Delta R_c}{R_c}$ in the resistance of the channel under the transitions. For the volt-watt sensitivity of the detector with a four-terminal network we obtain the following expression

$$S_{\mathrm{FT}} = \frac{\langle \Delta Q\rangle}{Q_0 \langle P_\omega\rangle} \frac{R_{\mathrm{c}} R_{\mathrm{l}} \mathrm{E}\, K}{(R_{\mathrm{l}} + R_{\mathrm{c}} + r_{\mathrm{c}})^2} = \left(\alpha R_0 - \frac{1}{2}\frac{b}{C_0}\right)\frac{C_0}{Q_0} \times \frac{2R_0 R_{\mathrm{c}} R_{\mathrm{l}} \mathrm{E}\, K}{\left(2 + r/R_0\right)\left(1 + \omega^2 C_0^2 \frac{rR_0}{(2+r/R_0)}\right)(R_{\mathrm{l}} + R_{\mathrm{c}} + r_{\mathrm{c}})^2} \tag{12}$$

where r_c is the resistance of the channel outside the metal-semiconductor transitions. For the structure with a two-dimensional electron gas as the channel, we have $b = 0$, and when the *I-V* characteristic of the transitions is of the exponential form (8), from (12), for $R_{\mathrm{l}} = R_{\mathrm{c}} + r_{\mathrm{c}}$, we obtain

$$S_{\mathrm{FT}} = \frac{1}{2}\frac{q}{nkT}\frac{C_s}{qn_s}\frac{E}{4}\frac{R_c}{R_c + r_c} \times \frac{2R_0}{(2 + r/R_0)\left(1 + \omega^2 C_0^2 \frac{r/R_0}{(2+r/R_0)}\right)} \tag{13}$$

where C_s is the capacitance of the metal-semiconductor transition per unit area and n_s is the surface carrier density in the channel. The volt-watt sensitivity is proportional to the parameter $\frac{C_{\mathrm{s}}}{q n_{\mathrm{s}}}$, which for a two-dimensional electron gas in GaAs with a surface carrier density of $7 \times 10^{11}\,\mathrm{cm}^{-2}$ and a depth of 100 nm from the surface is $1\,\mathrm{V}^{-1}$. The quantity S_{FT} is also proportional to E; therefore, an increase in the volt-watt sensitivity can be achieved by applying a higher dc voltage. It should be taken into account that an increase in E leads to an increase in the voltage drop in the channel, which is neglected in the simple model we are using in this work.

Assume that the thermal noise determined by the differential resistance of the barrier transitions near the zero voltage, the channel and the load resistances give the main contribution to the noise of the detector with a four-terminal network. In this case, it is easy to obtain an expression for the noise-equivalent power of the detector

$$\mathrm{NEP}_{\mathrm{FT}} = \sqrt{8kTBR_0}\frac{(2 + r/R_0)\left(1 + \omega^2 C_0^2 \frac{rR_0}{(2+r/R_0)}\right)}{\frac{q}{nkT} R_0 K} \times \sqrt{1 + 4\left(\frac{qn_{\mathrm{s}}}{\mathrm{E}C_{\mathrm{s}}}\right)^2 \frac{R_c}{R_0}\left(\frac{R_c + r_c}{R_c}\right)^3}. \tag{14}$$

The second term under the square root in (14) becomes unity when the thermal noise of the differential resistances of the transitions and the thermal noise of the resistances in the low-frequency circuit become equal. This condition specifies a characteristic value of the dc voltage E, at which its further increase leads to a weak decrease in the noise-equivalent power.

IV. Discussion of the Results

We now compare the characteristics of the radiation detectors based on symmetrical sensing elements and the conventional zero-bias detector. The volt-watt and threshold sensitivities of the square-law detector based on a Schottky (Mott)-barrier diode in the mode without the dc bias are determined by the expressions [13]

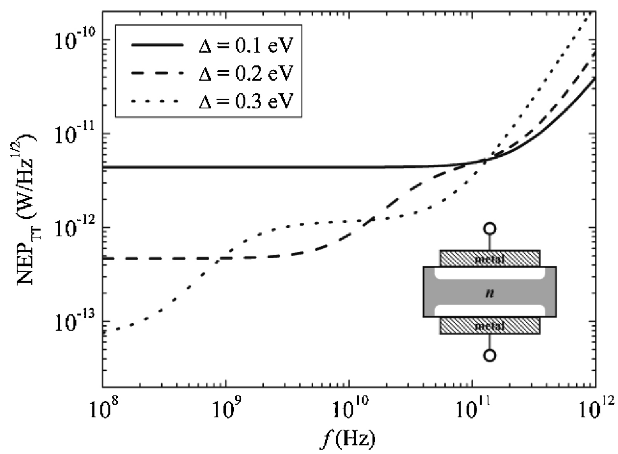

Fig. 4. Noise-equivalent power of the detector based on a two-terminal network as a function of the received-signal frequency for the structures with different effective Schottky barrier heights. Inset: the geometry of the two-terminal network.

$$S_{\mathrm{D}} = \alpha_{\mathrm{D}} R_{\mathrm{D}} \frac{R_{\mathrm{l}}}{R_{\mathrm{l}} + R_{\mathrm{D}} + r_{\mathrm{D}}} \times \frac{R_{\mathrm{D}}}{\left(1 + r_{\mathrm{D}}/R_{\mathrm{D}}\right)\left(1 + \omega^2 C_{\mathrm{D}}^2 \frac{r_{\mathrm{D}} R_{\mathrm{D}}}{1 + r_{\mathrm{D}}/R_{\mathrm{D}}}\right)} K \tag{15}$$

$$\mathrm{NEP}_{\mathrm{D}} = \frac{\sqrt{4kTB\,(R_{\mathrm{D}} + r_{\mathrm{D}})}}{S_{\mathrm{D}}} \tag{16}$$

where C_{D} and R_{D} are the differential capacitance and the differential resistance of the barrier transition of the diode at a zero voltage, r_{D} is the spurious resistance connected in series with the barrier transition, and α_{D} is the quantity determined by an expression similar to (5). One can introduce the critical frequency

$$\omega_{\mathrm{c}} = \sqrt{\frac{1 + r_{\mathrm{D}}/R_{\mathrm{D}}}{C_{\mathrm{D}}^2 r_{\mathrm{D}} R_{\mathrm{D}}}}. \tag{17}$$

The volt-watt sensitivity decreases with increasing frequency at frequencies higher than ω_c and has a weak frequency dependence at frequencies lower than ω_c. It follows from (7) and (12) that the critical frequency for the two- and four-terminal networks is given by

$$\omega_{\mathrm{c\ TT,\,FT}} = \sqrt{\frac{2 + r/R_0}{C_0^2\, r\, R_0}}. \tag{18}$$

With the identical barrier transitions, equal series resistances, and $r/R_0 << 1$, the critical frequency for the symmetrical sensing elements is a factor of $\sqrt{2}$ greater than that for the diode.

Fig. 4 shows the dependences of the noise-equivalent power of the detector based on a two-terminal network on the frequency of the received signal. The geometry of the two-terminal network is shown in the inset. The curves were obtained for the GaAs structures with different effective Schottky-barrier heights Δ, eV, namely, 0.1, 0.2, and 0.3.

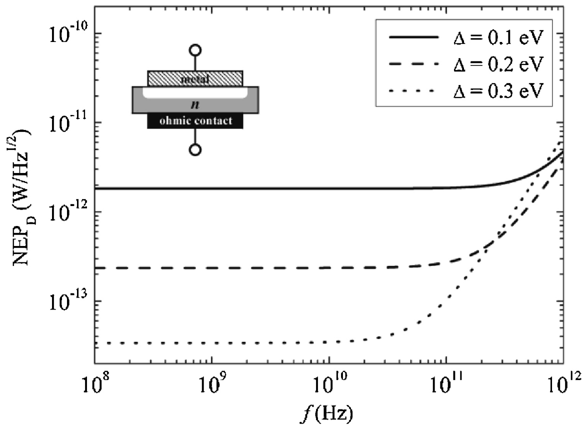

Fig. 5. Noise-equivalent power of the Schottky diode detector as a function of the received-signal frequency for the diodes with different effective Schottky barrier heights. Inset: the geometry of the diode.

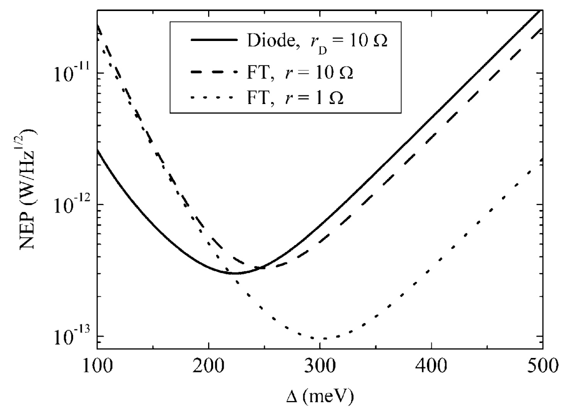

Fig. 6. Noise-equivalent powers of the Mott-barrier diode detector and the detector based on a four-terminal network with Mott barriers as functions of the effective barrier height for different values of the series resistance at the measured-signal frequency $f = 10^{11}$ Hz.

We used the following parameters for the calculation: $n = 1$, $T = 300$ K, and $S = 10^{-7}\,\text{cm}^2$. The semiconductor layer was 200 nm thick, the semiconductor doping level was $N = 5 \times 10^{16}\,\text{cm}^{-3}$, $R_1 = 2R_0 + r$; $\text{E} = 2kT/q$, and $K = 1$. For the dependence of the barrier-transition capacitance on the voltage, we used the relation adapted from [14]

$$C(V) = S\sqrt{\frac{Nq^2\varepsilon\varepsilon_0}{2kT}}\frac{1-\exp\left[q(V-V_F)/kT\right]}{\sqrt{\exp\left[q(V-V_F)/kT\right]-1-q(V-V_F)/kT}} \tag{19}$$

where V_F is the flat-band voltage, ϵ is the relative dielectric permittivity of the semiconductor, and ε_0 is the dielectric permittivity of vacuum. For comparison, Fig. 5 shows similar dependences for the Schottky-barrier diode detector. The geometry of the diode is shown in the inset in Fig. 5. The semiconductor layer was 100 nm thick, i.e., less than half that in the MSM structure. In accordance with this, $r_D = r/2$ and $\omega_c = \omega_{c\,TT}$. The remaining parameters of the diode structure and the structure of the two-terminal network coincided. The resistance of the ohmic contact to the semiconductor layer was assumed equal to zero. The load resistance R_l was assumed to be much larger than the diode resistance $R_D + r_D$. It is seen in Figs. 4 and 5 that the noise-equivalent powers for the diode and the two-terminal network are close at low frequencies. At high frequencies, NEP_{TT} has severalfold larger values than NEP_D. The limiting values of the noise-equivalent power can be compared for the chosen geometry of the structures. For the measured-signal frequencies satisfying the condition $\omega\tau << 1$, using (10), we obtain

$$\frac{\text{NEP}_{TT}}{\text{NEP}_D} \approx 2 + \frac{r}{R_0} \approx 2. \tag{20}$$

For $\omega\tau >> 1$, in view of (9), we find

$$\frac{\text{NEP}_{TT}}{\text{NEP}_D} \approx 2\frac{q}{nkT}\frac{C_0}{b} \xrightarrow[V_F \to 0]{} \frac{6}{n}. \tag{21}$$

Here, we used dependence (19), which yields an estimation for the nonlinearity parameter of the capacitive current in the flat-band limit: $b/C_0 = q/3kT$. The characteristics of the detector with a two-terminal network can be improved due, e.g., to the enhancement of the nonlinearity of the capacitive current by using a nonuniform (decaying away from the surface) doping profile of the semiconductor layer.

Fig. 6 shows the noise-equivalent power of the Mott-barrier diode detector (for $r_D = 10\,\Omega$) and the detector based on a four-terminal network with Mott barriers (for two values of r, namely, 10 and 1 Ω Hz) as functions of the effective barrier height at the measured-signal frequency $f = 10^{11}$. The geometries of the four-terminal network and the Mott diode are shown in the inset in Fig. 7. The calculation was performed for the GaAs structures at $T = 300$ K, $n = 1$, and $K = 1$. For the calculation, we used the following parameters of the four-terminal network: channel length $a = 200$ nm, channel width $w = 50\ \mu$m ($S = a \times w = 10^{-7}\,\text{cm}^2$), semiconductor i layer thickness 200 nm; $r_c = 0$, $n_S = 10^{11}\,\text{cm}^{-2}$, $R_l = R_c$, and $\text{E} = 0.1$ V. For the Mott diode detector, the i layer was 100 nm thick, $S = 10^{-7}\ \text{cm}^2$, and $R_l >> R_D + r_D$. It is seen from the dependences that the minimum NEPs of the diode and the four-terminal network with equal series resistances $r_D = r = 10\,\Omega$ virtually coincide. The absence of the contribution of the ohmic contacts in the microwave circuit is an important feature of these sensing elements. For this reason, the series resistance r of a four-terminal network can be very small. The dependence in Fig. 6, which is constructed for the four-terminal network with $r = 1\ \Omega$, demonstrates a decrease in the minimum NEP with decreasing series resistance. Fig. 7 shows the NEP dependences of the same structures as functions of the measured-signal frequency for the Δ values corresponding to the minima in Fig. 6.

It should also be mentioned that the absence of the ohmic contacts in considered structures (both in two- and four-terminal networks) simplifies the manufacturing technology of the real devices, eliminates unwanted impurity diffusion processes along the surface when ohmic contacts are annealed, and permits one to move toward reducing the lateral dimensions of the planar topology. This, in turn, reduces the capacitance of the barrier transitions, decreases the series resistance, and improves the detector sensitivity at high frequencies.

The higher admissible level of the signal amplitude compared with the diodes is another important feature of these

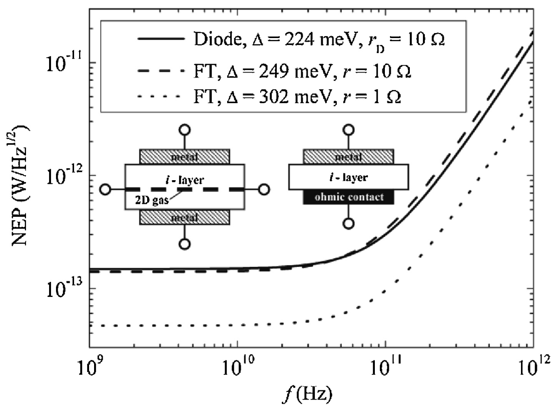

Fig. 7. Noise-equivalent powers of the Mott-barrier diode detector and the detector based on a four-terminal network with Mott barriers as functions of the measured-signal frequency for different values of the series resistance and optimal values of the effective barrier height.

sensing elements. Applying a high-amplitude periodic voltage to the diode results in that a high conduction current in the forward direction flows through it in each half-period of oscillation. The diode will break if the energy absorbed over the period exceeds the removed energy. After a large-amplitude ac voltage is applied to the MSM structure, the transition process occurs, during which a reverse voltage increases in each MS transition, that reduces the direct current through the transitions. The transition process stops when the direct and reverse currents transfer an equal amount of charge through the transitions over the oscillation period. The initial stage of the transition process can occur very quickly, for the picosecond time [14]. Due to this, the absorbed energy may not be enough to destroy the structure.

It is necessary to note that the description of the metal-semiconductor barrier transition using an equivalent scheme with the nonlinear conductance and nonlinear capacitance connected in parallel, which is valid for up to the subterahertz frequencies, probably becomes unsatisfactory with increasing frequency where the effects due to the inertia of electrons, finite times of transit and scattering, skin effect, etc. [15]–[17] become significant. These issues require further study.

V. Conclusion

We have proposed new sensing elements based on low-barrier metal-semiconductor-metal structures with a symmetrical I-V characteristic for the electromagnetic radiation detectors. The sensing elements represent two- or four-terminal networks. Feeding a high-frequency signal changes the impedance of the two-terminal network, or the channel conductivity in the case of the four-terminal design. The principle of operation of the detector consists in registering these changes on the direct current or at a low frequency. Schemes of connection of the sensing elements to the high-frequency path and the low-frequency measurement circuit are proposed. Expressions for the volt-watt sensitivity and NEP of the detector with various sensing elements are found. The obtained characteristics were compared with the characteristics of the Schottky and Mott detector diodes. It is shown that for the same barrier transitions and equal series resistances, the critical frequency of the proposed sensing elements is a factor $\sqrt{2}$ greater than in the diode.

We compared the NEPs of the two-terminal network and the Schottky diode with a half series resistance. It is shown that at the received-signal frequencies smaller than or of the order of $(2\pi\tau)^{-1}$, the NEPs of the two-terminal network and the diode have close values. At these frequencies, the NEP of the two-terminal network is inversely proportional to the nonlinearity parameter $\left(\frac{d^2I}{dV^2}\Big/\frac{dI}{dV}\right)\Big|_{V=0}$ of the conduction current (for the diode, the case is this over the entire frequency range). At frequencies much higher than $(2\pi\tau)^{-1}$, the NEP of the two-terminal network is inversely proportional to the nonlinearity parameter $\left(\frac{dC}{dV}\big/C\right)\big|_{V=0}$ of the capacitive current. At these frequencies, the ratio of the NEP of the two-terminal network and the diode is proportional to the ratio of the nonlinearity parameters of the capacitive current and the conduction current.

The NEP of the four-terminal network based on a vertical MSM structure with Mott barriers was compared with that of a Mott diode having half the thickness of the semiconductor i layer. It is shown that the four-terminal network and the diode with equal series resistances have about the same NEP over the entire frequency range if the optimal values of the barrier heights are selected. Because of the absence of the contribution of the ohmic contacts in the microwave circuit, the series resistance of a four-terminal network can be much smaller than in the diode. In this case, the characteristics of the four-terminal network exceed similar characteristics of the diode.

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