Towards multipixel THz Schottky diode detector with a single RF output line

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Abstract. We propose the design of a dual-pixel array of Schottky diodes. Each diode is fixed between the bow-tie antenna arms on top of a SI-GaAs membrane acting as a waveguide backshort for efficient coupling of the antenna to the feedline of high-directivity horn. The detector utilizes a single RF output line: microwave reflectometer is used for the readout. The pixels are equipped with dual-mode resonator filters to eliminate the cross-talk. We evaluate the design proposed via numerical simulation and performance tests of the array subunits: *NEP* of 300 pW/Hz$^{0.5}$ and dynamic range of 24 dB are demonstrated at 137.5 GHz.

1. Introduction

Terahertz coherent and direct detectors are in demand in numerous applications nowadays. The advantages of GaAs planar Schottky diode (PSD) direct detector are mainly related to a) its implementation-dependent time constant, which is determined only by the diode intrinsic RC-circuitry and output lines RF losses; b) decent sensitivity at room temperature; c) the ease of integration with complex RF circuits ensured by the strict control and reproducibility of the diode parameters. PSD detector also possesses a wide dynamic range beneficial in both the terahertz data transfer and medical imaging applications. However, it is difficult to combine high barrier diodes in array due to the need of DC biasing. This issue is traditionally solved via the use of Mott diodes providing low barriers resulting in high sensitivities if operated with zero bias at room temperature. Alternatively, it is possible to use a microwave frequency comb to readout an array of Schottky diodes [1].

In this paper we report on the evaluated performance of a multipixel terahertz Schottky diode detector with a single RF output line.

2. Detector design and performance tests

2.1. Dual pixel detector design

In accordance with the design proposed, a dual-pixel array of Schottky diodes is arranged as shown in figure 1. Each diode is fixed between the bow-tie antenna arms on top of a back metalized SI-GaAs membrane acting as a waveguide backshort for efficient coupling of the antenna to the feedline of high-directivity horn. There is only one RF output line intended for the injection of a probing microwave dual-frequency comb to readout the pixels [1] electrically connected by the Wilkinson power combiner (WPC). The pixels are equipped with miniaturized dual-mode resonator filters [2] to

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eliminate the cross-talk. We evaluate the design proposed via numerical simulation and performance tests of the array subunits.

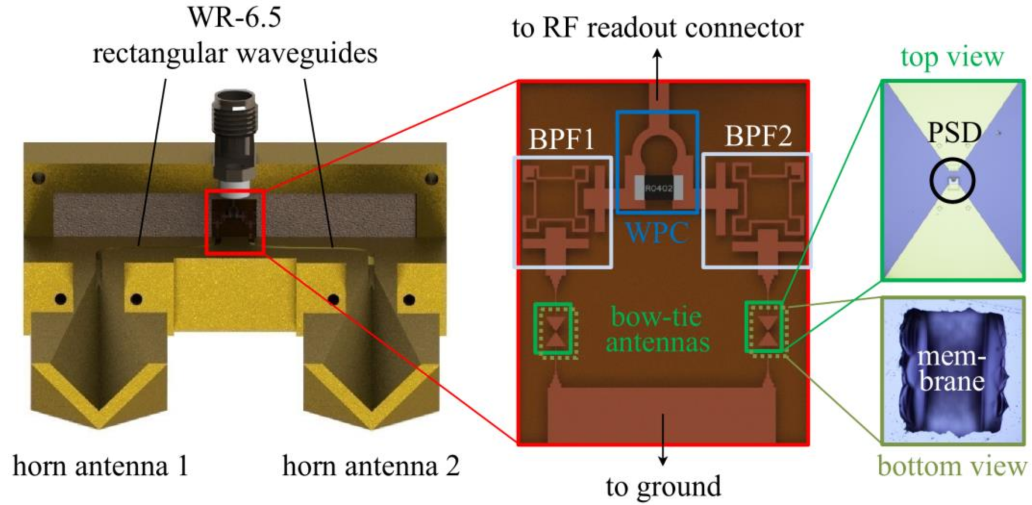

Figure 1. Dual pixel detector design.

The footprint of the miniaturized band-pass filters (BPF) developed for GaAs substrate by the means of EM modeling is within 1.36×1.36 mm^2, which corresponds to a nearly quarter-wave linear dimension given their central frequencies of 18.5 and 20.3 GHz; the fractional bandwidths of ~6 % ensure the pixels isolation of 19.6 dB. The insertion losses introduced by WPC are within 2.2-2.3 dB for each frequency channel, and the return losses are in the range of 15.7-19.8 dB. Please be advised that the inset with a microstrip circuitry of the detector chip is provided to scale, and a 100 Ω 0402 SMD resistor can be used to estimate actual linear dimensions of the microstrips.

Referring to figure 1, the bow-tie antennas are centered with respect to WR-6.5 rectangular waveguides in a way that their arms are oriented along the waveguides' E-planes. For such transitions, EM modeling suggests input return loss of less than -15 dB in the frequency range of 140-160 GHz and insertion loss as low as -3 dB. The diagonal horn antennas [3] possess gain of 23.5 dB and the half power beam width of 12º. Rigidness of the detector chip is due to thick SI-GaAs frame surrounding the membrane.

2.2. Readout method and its evaluation

The detector design adopts planar diodes with a Schottky contact diameter of 3 μm and series resistance of <30 Ω (our fabrication process is described elsewhere [4]) as sensors. It is known that Schottky diode features non-quadratic rectification characteristics at certain operating conditions [5]. Since the probing microwave power can be set for each pixel individually to tune the performance, we theoretically and experimentally investigated the impact of various bias conditions (BC) of the diode on input dynamic range (*IDR*) and the optical noise equivalent power (*NEP*) of the detector.

According to the thermionic emission model for an ideal Schottky diode, the forward bias current-voltage characteristic (IVC) can be described by

$$I_{ideal} = I_s\left(\exp\left(qV_b\left(\eta k_B T\right)^{-1}\right) - 1\right). \tag{1}$$

In case of a real Schottky diode, one has to account for the impact of series resistance (R_s) on the shape of IVC at high values of forward bias voltage (V_b) as

$$I_{real} = I_s\left(\exp\left(q\left(V_b - I_{real}R_s\right)\left(\eta k_B T\right)^{-1}\right) - 1\right). \quad (2)$$

The saturation current (I_s) is defined as

$$I_s = AA_R T^2 \exp\left(-q\Phi_{b0}\left(k_B T\right)^{-1}\right). \quad (3)$$

Here η is the ideality factor, q is the elementary electron charge, k_B is the Boltzmann constant, T is the physical temperature of the diode measured in Kelvin, A is the Schottky contact area, A_R is the Richardson constant, and Φ_{b0} is the zero bias Schottky barrier height. Introducing the Lambert W function (W_0), one can explicitly represent equation (2) as [6,7]

$$I_{real} = \eta k_B T\left(qR_s\right)^{-1} W_0\left(qR_s\left(\eta k_B T\right)^{-1} I_s \exp\left(q\left(V_b + I_s R_s\right)\left(\eta k_B T\right)^{-1}\right)\right) - I_s. \quad (4)$$

The values of η, Φ_{b0}, R_s are extracted upon approximation of the experimental IVC by equation (4) with the aid of the least squares method. In our studies, we rely on symbolic calculations and make use of a cross-platform computer algebra system "Maxima". Flat band barrier height (Φ_{bf}) is further calculated as [8]

$$\Phi_{bf} = \eta\Phi_{b0} - \left(\eta - 1\right)k_B T q^{-1} \ln\left(N_c N_d^{-1}\right),$$

where N_d is the semiconductor impurity concentration, N_c is the effective density of states in the conduction band.

In our experiments, Schottky diode with η = 1.56, Φ_{b0} = 0.577 eV, R_s = 14.1 Ω, Φ_{bf} = 0.891 eV is simultaneously biased by a source-meter and microwave frequency synthesizer. The latter also acts as a source of probing signal to readout the diode response to a sub-THz incident power. The output/input power ratio (P_{out}/P_{in}) is measured with the aid of a spectrum analyzer and a sub-THz CW source equipped with precision waveguide attenuator. Referring to figure 2, the slope of a "linear-response" region gradually grows from 1.34 to 1.88 along with a BC set's serial number. This corresponds to the gradual increase of the percentage of AC component (δI_{mw}) induced by the probing signal of 1.4 GHz in the diode bias current (I_b) from 1.25 % for BC#1 to 217.5 % for BC#9. We fix the DC current component (I_{DC}) to 200 μA in all the measurements and observed that $\Delta P_{out}/\Delta P_{in}$ reaches its maximum of 1.91 for BC#10, when the diode is biased by direct current solely and conventional readout scheme is employed. IDR > 30 dB can be achieved in both of the above readout options. The current components are related as $I_b = I_{DC}(1 + \delta I_{mw})$.

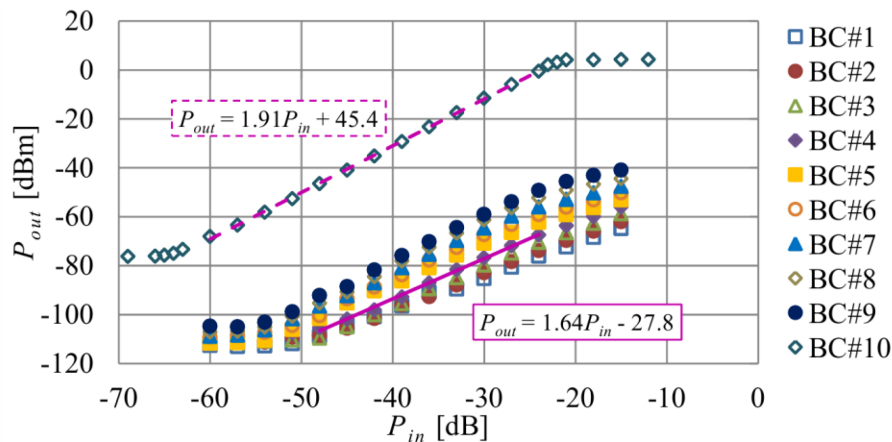

Figure 2. Linearity curves measured for a Schottky diode at different bias conditions.

To analyze the trends in the changes of linearity of a Schottky diode upon increase of the percentage of AC component in its bias current, we calculated a number of time series [t, I_{real}] with the aid of Monte-Carlo method. Analytic approach was inconvenient for the calculation due to the need to deal with modified Bessel functions of the first kind naturally appearing because of $\exp(\xi_0 \sin(\xi(t))$ under integral operator. Indeed, when the diode is under microwaves, its bias voltage is defined as

$$V_b = V_{DC} + V_{AC,0}\left(\sin\omega t\right), \tag{5}$$

where V_{DC} is the DC voltage from a source-meter, $V_{AC,0}$ is the amplitude of AC voltage generated in the diode by frequency synthesizer, and ω is the angular frequency of AC power provided by the synthesizer. Combining equations (4) and (5), one obtains time-dependent current through the diode as

$$I_{real}(t) = \eta k_B T\left(qR_s\right)^{-1} W_0\left(qR_s\left(\eta k_B T\right)^{-1} I_s \exp\left(q\left(\left(V_{DC} + V_{AC,0}\left(\sin\omega t\right)\right) + I_s R_s\right)\left(\eta k_B T\right)^{-1}\right)\right) - I_s. \tag{6}$$

For further analysis, we divide the $I_{real}(t)$ dependence obtained in 32 sections of small-signal response. Sufficiency of such a linear interpolation is justified by the analysis of linearity curves provided in figure 3. Average current value ($\langle I_{real}\rangle$) over the period of microwave probing signal is further calculated. Equating this value to I_b measured for a given BC set's serial number, one acquires the corresponding $V_{AC,0}$ value.

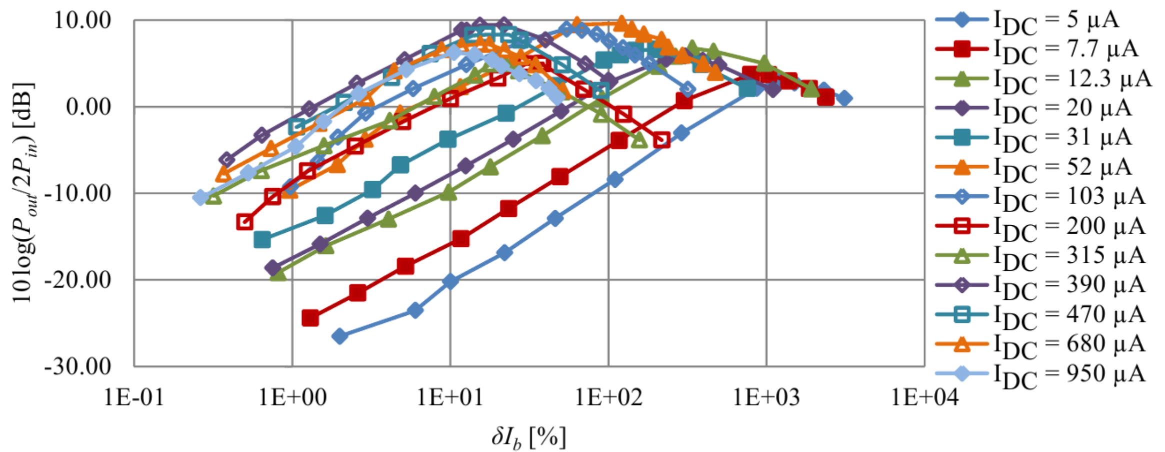

Figure 3. Linearity of the terahertz current generation as a function of DC bias.

Reflected microwave power at a 50 Ω load is further calculated as [9]

$$P_{50}(t) = \left|V_{AC}(t)\right|^2 50\left(\left(50 + \mathrm{Re}\left(Z_{PSD}(t)\right)\right)^2 + \mathrm{Im}\left(Z_{PSD}(t)\right)^2\right)^{-1} \tag{7}$$

with

$$Z_{PSD}(t) = R_s + \left(\left(R_{diff}(t)\right)^{-1} + i\omega C_{tot}\right)^{-1}. \tag{8}$$

Here $R_{diff}(t)$ is the diode time-dependent differential resistance and $Z_{PSD}(t)$ is the diode time-dependent impedance. For its calculation, we assume that total capacitance of the diode element (C_{tot}) is solely determined by the parasitic capacitance value, which is estimated as 4 fF for our diodes [1]. The junction capacitance (C_j) is neglected, since the diode operates as a varistor (figure 4), and

instantaneous value of its bias voltage stays within the range from 0.55 to 0.81 V during all the measurements. Such voltages significantly exceed built-in potential (V_{bi}) for the diodes studied.

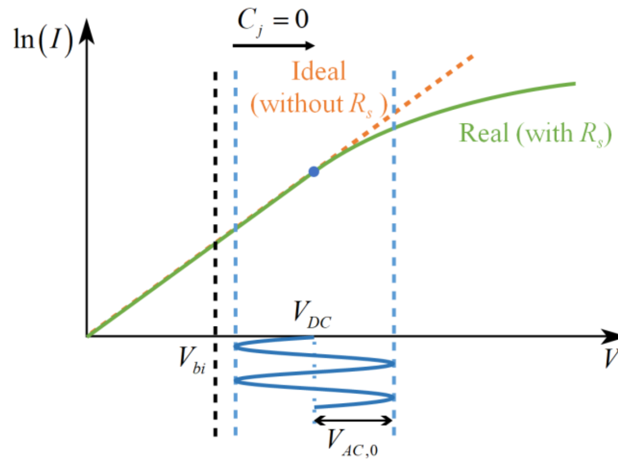

Figure 4. Real and ideal IVCs of a varistor diode.

Referring to figure 5, linearity of the diode response to microwaves gradually grows along with a BC set's serial number. The trend observed is qualitatively consistent with the experimental results.

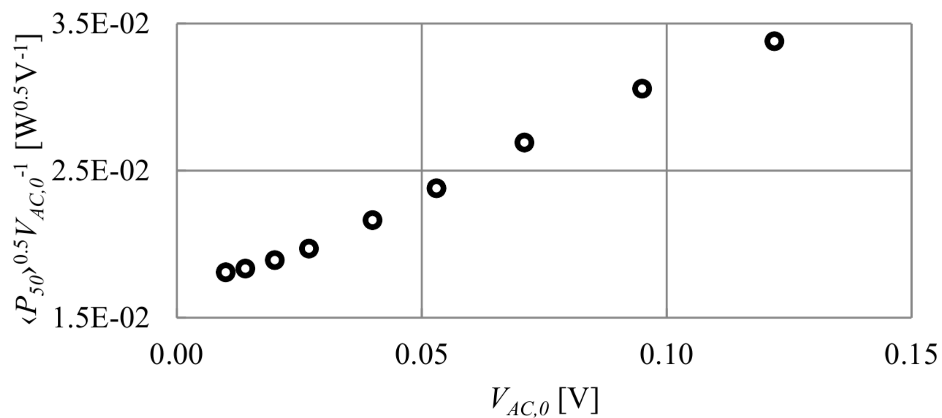

Figure 5. Normalized reflected microwave voltage at a 50 Ω load averaged over the period of microwave probing signal as a function of the amplitude of AC voltage generated by it in a Schottky diode (i.e. as a function of a BC set's serial number).

As a sweet spot of the detector's linearity and sensitivity, we measured $\Delta P_{out}/\Delta P_{in} = 1.64$, $IDR = 24$ dB and $NEP = 300$ pW/Hz$^{0.5}$ at signal frequency of 137.5 GHz for the diode biased by $I_{DC} = 200$ µA with $\delta I_{mw} = 10$ % (figure 2: BC#4).

3. Conclusion

We propose the design of a dual-pixel array of Schottky diodes. The detector utilizes a single RF output line: microwave reflectometer is used for the readout. We evaluate the design proposed via numerical simulation and performance tests of the array subunits. Given the capabilities of the input terahertz optics and the microwave distribution network developed, it seems feasible to expand the

design of a dual-pixel array of Schottky diodes to that of the linear array with a noticeably bigger number of elements.

Acknowledgments

We acknowledge support of the Russian Science Foundation grant No. 19-72-10156.

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