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Detection of nanosecond-scale, high power THz pulses with a field effect transistor

S. Preu,^{1,2,3} H. Lu,² M. S. Sherwin,¹ and A. C. Gossard²

¹*Physics Department and Institute for Terahertz Science and Technology, University of California, Santa Barbara, California 93106, USA*

²*Materials Department, University of California, Santa Barbara, California 93106, USA*

³*Chair for Applied Physics, University of Erlangen-Nuremberg, 91058 Erlangen, Germany*

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We demonstrate detection and resolution of high power, 34 ns free electron laser pulses using a rectifying field effect transistor. The detector remains linear up to an input power of 11 ± 0.5 W at a pulse energy of 20 ± 1 μ J at 240 GHz. We compare its performance to a protected Schottky diode, finding a shorter intrinsic time constant. The damage threshold is estimated to be a few 100 W. The detector is, therefore, well-suited for characterizing high power THz pulses. We further demonstrate that the same detector can be used to detect low power continuous-wave THz signals with a post detection limited noise floor of $3.1 \mu\text{W}/\sqrt{\text{Hz}}$. Such ultrafast, high power detectors are important tools for high power and high energy THz facilities such as free electron lasers. © 2012 American Institute of Physics. [<http://dx.doi.org/10.1063/1.4705986>]

I. INTRODUCTION

In recent years, a variety of detector concepts for THz radiation have been developed, fitted to specific needs. Most detector types were optimized for detecting small amounts of THz power of typical tabletop THz sources. In terms of fast, coherent detectors, Schottky diode mixers are capable to resolve signals with hundreds of MHz modulation frequency. Schottky diodes can also be used for direct detection with few $\text{pW}/\sqrt{\text{Hz}}$ noise equivalent power (NEP).¹

These detectors are extremely sensitive but cannot measure high power pulses easily. Radiation from free electron lasers (FELs) or gyrotrons, with peak powers of hundreds of W to kW, have to be heavily attenuated to prevent destruction of these sensitive detectors.

Many materials or semiconductor structures respond in some way to such high power levels. However, the responses are typically highly non-linear² and therefore ineligible for detection and pulse characterization.

The work horses for measuring the pulse energy are thermal detectors, such as pyrometers,³ pyroelectric detectors,⁴ or photoacoustic power meters. Their damage threshold is high enough to withstand these extraordinarily high power levels and their response is linear. However, the detector time constants are typically in the range of tens to hundreds of ms. Such detectors are thus much too slow to resolve the pulse shape with typical rise and fall times in the nanosecond range.

There are only a few detection concepts that allow for both linear detection of high power signals and short time constants. Non-linear media have been explored for up-conversion of an FEL pulse on an optical carrier and subsequent optical spectroscopy was demonstrated.⁵⁻⁷ Although this method is very fast and even allows for measuring the frequency of the FEL pulse,^{6,7} the optical setup is relatively bulky and cannot easily be transferred between experiments. In terms of fast response, photon drag detectors are frequently used.⁸ They typically offer time constants in the range or be-

low 1 ns.⁹ However, their responsivity is in the range of only a few 10^{-5} V/W and their noise floor is very high, generally impeding measurements of powers around or below watt level. Superlattice structures consisting of a large number of coupled quantum wells have also been employed for detection of high power, short pulsed THz radiation.^{10,11} Due to dynamic localization, an incident THz field reduces the current through the minibands that are formed by the coupled wells. The detectors are extremely fast, with intrinsic relaxation times in the range of 100 fs. A maximum linearity range up to 5 W and a dynamic range of 50 dB (Ref. 10) could be demonstrated in continuous-wave (CW) measurements. However, accurate quantum well superlattice structures are difficult to grow.

Recently, a rectification effect in field effect transistors (FETs) became an attractive tool for THz detection.¹²⁻¹⁶ Noise equivalent powers below $100 \text{ pW}/\sqrt{\text{Hz}}$ (Refs. 17-19) of such detectors were demonstrated. In contrast to amplifying FETs, the rectification effect also persists and remains efficient above frequencies where the transistor has gain,²⁰ simplifying the processing requirements as larger structures can be used. Kachorovskii and Shur²¹ also pointed out by a theoretical study that such detectors should be capable of resolving modulations in the high GHz range.

In this paper, we investigate rectifying FET THz detectors with potentially handheld operation that remain linear up to 11 W (20 ± 1 μ J) total input power (pulse energy). We estimate a damage threshold in the few 100 W range for frequencies around 300 GHz, minimizing the risk of destruction even under improper handling. We show that the same detector is very sensitive at room temperature operation and can detect $\mu\text{W}/\sqrt{\text{Hz}}$ pulsed and CW THz signals. It is, therefore, by orders of magnitude more sensitive than a photon drag detector and about one order of magnitude more sensitive than the superlattice detector.¹⁰ Furthermore, its responsivity is also at least 2-3 orders of magnitude higher than that of the photon drag detector. We show that the detector is fast enough to

resolve FEL pulses with a full-width half-maximum (FWHM) of 34 ns with detector rise and fall times below 11 ns. The detectors were processed implementing inexpensive UV contact lithography.

II. SAMPLE STRUCTURE

The investigated n-FETs consist of a two dimensional electron gas (2DEG) in an $\text{Al}_{0.3}\text{Ga}_{0.7}\text{As}$ -GaAs-structure. The carrier concentration was determined to be $n^{(2D)} = 5.49 \times 10^{11}/\text{cm}^2$, the room-temperature mobility is $\mu = 6900 \text{ cm}^2/\text{Vs}$. The 80 nm wide 2DEG well is 65 nm below the surface. The channel is remotely doped with a silicon (Si) delta doping of $6 \times 10^{12}/\text{cm}^2$, 10 nm away from the channel. The geometric layout of the devices is illustrated in Figs. 1(a) and 1(b). The detectors consist of N individual mesas of length L_M , each with a gate (G), source (S), and drain (D) electrode. These N mesas are arrayed to cover an area of $S \times W$ in the range of the THz spot size. They are connected in parallel. An incident THz beam with electric field polarized along the source-drain (SD) axis (S-direction in Fig. 1(a)) induces a SD THz bias. This bias modulates the carrier velocity in the channel. At the same time, the incident THz field also modulates the gate-to-channel bias which alters the carrier concentration in the channel. The common modulation of carrier concentration and carrier velocity results in a net rectified component that is proportional to the incident THz power. This rectification effect still works much above frequencies where the transistor has gain. A detailed and more accurate theoretical description of the detection principle can be found elsewhere.^{20,22,23} Three different layouts have been investigated as summarized in Fig. 1(d). The nomenclature is the number of mesas, followed by “M” and the approximate device length in microns. The first sample (4M145) has relatively loosely spaced electrodes. For the second sample, the number of mesas was doubled but the total size maintained, resulting in a much denser electrode layout. The third sam-

ple consists of the same mesa structure as sample 4M145 but its size was doubled, resulting in 8 mesas of width $W = 292 \mu\text{m}$. This sample is used to investigate the effect of the radiation resistance of the large area detector, which depends on the device size.²⁴

The samples were processed with contact UV lithography. Ohmic Ge/Au/Ni/Au contacts were deposited and thermally annealed. Subsequently, a wet etch with $\text{H}_2\text{O}_2:\text{H}_2\text{SO}_4:\text{H}_2\text{O} = 8:1:80$ was used to structure the mesas. Then, Al:Au gates were deposited. In the same deposition, the source and drain contacts were wired to large contact pads. To be able to address the gates, a dielectric spacer was deposited on the source wires and a subsequent metallization layer, on top of the source contact layer, connected all gates to another large contact pad.

III. EXPERIMENTAL RESULTS

A. Setup

The measurements were performed at the FEL of the University of California, Santa Barbara. The FEL operated at 240 GHz in two modes: Firstly, the quasi-CW mode, with pulse lengths in the range of $1.8 \mu\text{s}$ at almost constant pulse power. Secondly, a cavity dumped mode, where the FEL cavity was dumped by an optically gated Si-mirror.²⁵ High power pulses with durations slightly shorter than the cavity round trip time of 38 ns are obtained. The FEL beam is directed on a 45° oriented wire grid. The transmitted beam is used for pulse energy calibration with a photoacoustic detector. The reflected beam travels through a series of wire grid attenuators and is focused by a parabolic mirror on a FET that is mounted on a hyperhemispherical Si-lens. The Si-lens improves the coupling efficiency of the THz beam to the (relatively small) FET. A packaged version of a FET detector with integrated Si-lens is illustrated in Fig. 1(c). From the geometrical size, only about $15\% \pm 5\%$ of the power is incident on

FIG. 1. (a) Schematic sample layout of sample 4M145. The source, gate, and drain contacts are labelled by the letters S, G, and D, respectively. The individual N mesas are connected in parallel. (b) Cross section through a mesa, showing the ohmic source and drain contacts and the Schottky gate contact. The silicon lens is not to scale. (c) Packaging example of the FET detectors with integrated 1 cm silicon lens on a 0.5" mounting post. (d) Layout dimensions of the investigated devices.

the devices 4M145 and 8M145, assuming a diffraction limited THz spot at 240 GHz. Device 8M290 receives about $48\% \pm 12\%$. A fraction of the beam is directed onto an electrostatic discharge (ESD) protected Schottky diode detector (Virginia diodes WR4.3ZBD, ESD limited rise time of 1.8 ns) for pulse width characterization and for reference. The detection signals of the photoacoustic detector, the FET and the Schottky diode are fed into a LeCroy WavePro 950 oscilloscope with an integration time of 0.25 ns. The input impedance can be switched between $50\ \Omega$ and $1\ \text{M}\Omega$.

B. Detector characterization with high power THz pulses

1. Linearity and responsivity

First, we investigated the linearity of the detectors by attenuating the incident beam using the wire grid attenuator setup. A rotatable wire grid is mounted between two parallel wire grids. The transmitted power can be set by the angle of the center wire grid. In addition, we added a polyvinylchloride (PVC) attenuator with a transmittance of 12% for low power measurements. The oscilloscope was set at an input bias of $1\ \text{M}\Omega$. The detected voltage vs. incident quasi-CW FEL power is depicted in Fig. 2 for 4M145 for three biases. When operated closer to the threshold bias of $-1.48\ \text{V} \pm 0.05\ \text{V}$ for all three devices, the responsivity increases as expected for a rectifying FET.^{20,23}

The error bars indicate the FEL fluctuations in the range of 7% and the oscilloscope noise of 0.5 mV. At 0V gate bias, the detector remains linear up to an incident power of 9 W ($\sim 1\ \text{dB}$ down power) with a responsivity of 1.3 mV/W and then saturates. Above 9 W the shape of the detected signal also starts deviating from that measured with the fast Schottky diode. The responsivity of sample 8M145 was about 65% larger, but it started saturating already at about 4 W at 0 V gate bias. When operated closer to the threshold, the responsivity of the detectors increases but, as we will see later, its time con-

FIG. 2. Generated output voltage of device 4M145 for quasi-CW FEL pulses at zero gate bias. Inset: Zoom into linear range. The detector response is shown for $U_{GS} = 0\ \text{V}$, $-0.8\ \text{V}$ and $-1.3\ \text{V}$ with increasing responsivity for operation closer to the threshold.

stant increases, too, attributed to a larger device impedance. At $-0.8\ \text{V}$ gate bias, the responsivity of device 4M145 increased by 75% to 2.3 mV/W. Its linearity range remained the same. The responsivity of device 8M145 increased even to 96 mV/W but it became nonlinear already at about 0.2 W.

The responsivity of sample 8M290 was almost identical to that of device 4M145 with a 1 dB down power of 11 W at 0V gate bias. Due to the larger device area, the total incident power on device 8M290 is about three times of that of 4M145. However, the larger size also results in a smaller radiation resistance. In Ref. 24, Eq. 90, we calculated the radiation resistance of a large area device of radius ρ that consists of small dipoles of length $l_D = g_{sa}$, providing

$$R_{rad} = \frac{20\Omega/n_{eff} \times (l_D k_{THz})^2}{1 + (0.82\rho k_{THz})^2}, \quad (1)$$

where n_{eff} is the effective index of refraction on a detector on a semiconductor-air interface and $k_{THz} = 2\pi/\lambda_{THz}$. The numerator of Eq. (1) is the (Hertzian) radiation resistance of a small area element of the large area with an average dipole moment of length l_D . The denominator of Eq. (1) takes interference of the Hertzian dipoles within the detector area into account. Due to non-resonant operation at THz frequencies, the effective dipole length l_D is small compared to the mesa dimension. As the source-drain distance and the losses (due to the sheet access resistance ρ_{acc}) of both devices are identical, the numerator of Eq. (1) is identical for both devices. We can therefore use this equation to compare the responsivity of device 8M290 and 4M145. The responsivity, $\mathfrak{R}$, of the detector scales with the square of the THz bias at the structure which is given by the radiation resistance as $\mathfrak{R} \sim (U_{DS}^{THz})^2 = 2R_{rad}P_{THz}$. The radiation resistance of device 8M290 is a factor of 3.2 smaller than that of device 4M145 due to the interference-related term in the denominator of Eq. (1). Thus, the effect of collecting more power approximately cancels the effect of a smaller radiation resistance. Both devices should approximately show the same responsivity, in excellent agreement with the experiment. That is, a larger device size does not necessarily result in a higher detection efficiency. As the peak intensity limits the linearity range, the larger size will also not improve the linearity range if the THz beam is not expanded on the larger device. This also agrees with the experiment as the linearity range of the larger device was only about 18% larger than that of the smaller device. This small improvement is most likely due to a slightly different focal alignment of 8M290.

For device 8M145, the denominator of Eq. (1) is identical to that of device 4M145 as the same array diameter was used. Since the number of mesas was doubled for device 8M145, its dipole density is a factor of two higher. This results in a factor of 4 higher bias generated by device 8M145. Due to the shorter source-drain distance, the access resistance of device 8M145 is $\sim 70\%$ smaller than that of 4M145. We have shown in Ref. 20 that the access resistance of the FET is responsible for a linear roll-off of the detector efficiency with frequency because the access resistance (R) and the gate capacitance (C) form an RC element. Using Eq. 8 from that reference, we estimate that the losses in device 8M145 are about a factor of 12 smaller than those in device 4M145 at 240 GHz. The

responsivity of device 8M145 should thus be $4 \times 12 = 48$ times better than that of 4M145. At -0.8 V gate bias, the experimental responsivity of device 8M145 is 42 times higher than that of device 4M145, close to the theoretical expectation.

2. Detector time constant vs. gate bias

To determine the optimum gate bias for detection of quasi-CW pulses, we measured the FET signal for various gate biases as shown in Fig. 3. A low gate bias results in a small device resistance and therefore short time constants. Therefore, the FEL pulse is well resolved (cf. Fig. 3, -0.8 V $< U_{GS} < 0$ V). However, a low gate bias also results in lower responsivity. When approaching the threshold, both the responsivity and the time constant increase. After excitation with the THz pulse, the signal decays with an exponential tail when the pulse is switched off. From that tail we can fit the time constant of the detector which is plotted in the inset of Fig. 3. Due to a pulse rise and fall time in the range of 200 ns of the quasi-CW pulses, a gate bias around -0.8 V seems to be optimal for high responsivity measurements while maintaining a time constant short enough to resolve the quasi-CW FEL pulse.

To refine the data on the time constant of the detector, the FEL was operated in the cavity dumped mode. The laser cavity was emptied with an optically gated Si-wafer mirror, resulting in pulse widths in the range of the round trip time of the pulse of about 38 ns. The resulting peak power is about 5 times higher. Fig. 4(a) shows the comparison of the FET signal and the signal from the Schottky diode for an oscilloscope input impedances of $R_O = 50$ Ω and 1 M Ω for device 4M145.

At $R_O = 50$ Ω , the pulse width measured with the Schottky diode is 34.3 ± 0.5 ns (Gaussian FWHM). For $R_O = 1$ M Ω input impedance, it is drastically broadened compared to the signal at $R_O = 50$ Ω . This broadening originates from

FIG. 3. Quasi-CW signals measured with different gate biases for device 4M145 at $R_O = 1$ M Ω . The data are smoothed by a 5 point average. Inset: Detection time constant. The fall time of the pulse, obtained from the Schottky diode measurement, limits the minimum measurable detector time constant. This limit is indicated by the solid line.

the RC time caused by R_O and the ESD protection. The time constant at $R_O = 1$ M Ω is determined from the exponential decay to $\tau_S = 111 \pm 3$ ns. The signal measured with the FET, however, shows very similar pulse widths for both oscilloscope settings. At $R_O = 50$ Ω , the pulse width obtained with the FET detector is 33.8 ± 1 ns, slightly shorter than that from the Schottky diode measurement. We, therefore, conclude that the response time of the FET operated at $R_O = 50$ Ω is comparable to or shorter than the pulse rise time. The slightly larger width obtained from the Schottky diode is attributed to its ESD limited rise time. In contrast to the Schottky diode, the FET signal at $R_O = 1$ M Ω is only slightly broadened by 30% to a value of 43.6 ± 1.1 ns, demonstrating an extremely short intrinsic time constant of the FET, shorter than that of the ESD protected Schottky diode. The fit of the exponential decay of the FET signal at the falling edge with $R_O = 1$ M Ω provides a $1/e$ decay time of $\tau_{tot} = 28 \pm 1$ ns which is a convolution of the detector time constant and the pulse $1/e$ fall time of $\tau_{pls} = 11$ ns, obtained from the measurement with the faster device 8M145 in Fig. 4(b). From that, we can extrapolate a FET time constant in the range of $\tau \approx [1/\tau_{pls} - 1/\tau_{tot}]^{-1} \approx 18 \pm 1$ ns at $R_O = 1$ M Ω .

Device 8M145 performed even better due to its smaller access resistance. Even at $R_O = 1$ M Ω and $U_{GS} = -0.8$ V, the detected signal is practically identical with that obtained from the Schottky diode as shown in Fig. 4(b). At smaller gate biases, the FET data even show a shorter $1/e$ fall time (11 ± 0.04 ns) than the Schottky diode (15 ns ± 1 ns), indicating a shorter intrinsic time constant. We believe that the measurement is limited by the actual pulse fall time as the time constant obtained for $U_{GS} = -0.2$ V and -0.4 V are identical and agree with the values obtained at $R_O = 50$ Ω . Therefore, we conclude that the FET does not noticeably broaden the pulse and its time constant is much shorter than the pulse $1/e$ fall time of 11 ns.

For both the Schottky diode and the FETs, the signal at the oscilloscope is much smaller at $R_O = 50$ Ω compared to $R_O = 1$ M Ω as most of the power drops at the access resistance of the detectors and along the biasing lines. As the FETs can resolve the pulse even at $R_O = 1$ M Ω , most data shown in this paper were taken at this input resistance. The Schottky diode, however, would be too slow and was therefore operated at $R_O = 50$ Ω .

3. Damage threshold

We also investigated the damage threshold of FET mixers. The large area devices could not be destroyed by a maximum input power of 70 ± 7 W (corresponding to a pulse energy of 125 ± 12 μ J). An antenna-coupled device with about 28 times smaller active area dimensions was also investigated. Although highly non-linear, the device withstood more than 7 W FEL power. It was destroyed when switching abruptly to 70 W. The large area devices could just be investigated to a power level of 70 W. None of the devices failed under quasi-CW operation, nor under short pulse operation. From the failure power of the antenna-coupled device, we estimate that

FIG. 4. (a) Normalized FEL signal obtained from the Schottky diode at $R_O = 50 \Omega$ (top) and $R_O = 1 \text{ M}\Omega$ (2nd from top) and the device 4M145 at 50Ω (3rd) and $1 \text{ M}\Omega$ (bottom). The data are smoothed by a 5 point average. For clarity, the graphs are vertically offset. (b) Normalized voltage detected with 8M145 at $R_O = 1 \text{ M}\Omega$ for various gate biases. The signal obtained with the Schottky diode ($R_O = 50 \Omega$, red) is shown together with the data obtained at $U_{GS} = -0.8 \text{ V}$. They are almost identical. The graphs are vertically offset for clarity. Inset: Time constants obtained from 8M145 vs. gate bias at $R_O = 50 \Omega$ (red) and $R_O = 1 \text{ M}\Omega$ (black).

the large area devices should at least tolerate $7 \text{ W} \times 28 \approx 200 \text{ W}$, but more likely several 100 W of input power, minimizing the risk of destruction by improper handling.

4. Dynamic range

In terms of dynamic range, device 4M145 and 8M290 performed best. From the (dc) resistance of 4M145 of $R_{DS}(U_{GS} = -0.8 \text{ V}) = 270 \Omega$ (for device 8M145, $R_{DS}(-0.8 \text{ V}) = 90 \Omega$), we calculate the intrinsic thermal noise limited NEP of $870 \text{ nW}/\sqrt{\text{Hz}}$ ($12 \text{ nW}/\sqrt{\text{Hz}}$). In the measurement, however, the noise of other components, such as the oscilloscope, limits the NEP. For the measurement bandwidth of $BW = 1 \text{ GHz}$, a noise at the oscilloscope of $U_N = 0.5 \text{ mV}$ was obtained. For device 4M145, this noise level corresponds to a NEP of $NEP = U_N/(BW\mathfrak{R}) = 6.5 \mu\text{W}/\sqrt{\text{Hz}}$, where $\mathfrak{R}$ is the device responsivity. This results in a bandwidth-normalized dynamic range of $61 \text{ dB}/\sqrt{\text{Hz}}$. In order to investigate the origin of the noise, we evaluated the measured voltage noise at the absence of a THz signal for different gate biases of the FET. When approaching the threshold of the transistor, its resistance, R , strongly increases, so does its thermal noise since $U_N = \sqrt{4k_B T R \times BW}$. However, the measured voltage noise only increased when operating the device very close to the threshold ($U \sim -1.5 \text{ V}$), clearly indicating that the noise originated from the input impedance of the oscilloscope at bias levels far away from threshold ($U > -1.2 \text{ V}$). The theoretical NEPs may be reached by implementing a low noise amplifier between FET and oscilloscope. The detection voltage is then amplified to values above the noise floor of the oscilloscope. The measured signal is then limited by the FET (and amplifier) noise, not by the post detection unit at a sufficient amplification factor.

Device 8M145 shows an NEP of $0.2 \mu\text{W}/\sqrt{\text{Hz}}$ resulting in a similar dynamic range of $60 \text{ dB}/\sqrt{\text{Hz}}$. The theoretical noise floor of this device is calculated to $12 \text{ nW}/\sqrt{\text{Hz}}$. Val-

ues for both devices were obtained at an oscilloscope input impedance of $R_O = 1 \text{ M}\Omega$ and a gate bias of -0.8 V .

C. Continuous-wave characterization

In order to confirm these low NEP values, we also characterized the devices in a CW setup at 300 GHz implementing the lock-in technique (SR 830) with a preamplifier (SR 560) at a modulation frequency around 3.7 kHz . The peak to peak responsivity for all investigated devices was about a factor of 1.2–2 lower than in the pulsed experiments. This may be attributed to alignment issues or the higher frequency which may result in slightly different responsivities. An amplifier with an input noise of $4 \text{ nV}/\sqrt{\text{Hz}}$ (corresponding to an input resistance of $1 \text{ k}\Omega$ at 300 K) was used prior lock-in detection. For device 4M145 the noise floor was $3.1 \mu\text{W}/\sqrt{\text{Hz}}$ and $0.4 \mu\text{W}/\sqrt{\text{Hz}}$ for device 8M145, close to the values obtained from the pulsed measurements. Similar to the pulsed measurements, the noise is dominated by the amplifier noise of $4 \text{ nV}/\sqrt{\text{Hz}}$ for gate biases $U_{GS} > -1.2 \text{ V}$, where the device resistance is smaller than $1 \text{ k}\Omega$. Only at operation close to the threshold the noise is limited by the thermal noise of the FET. As in the pulsed measurements, an appropriately designed post-detection measurement unit with an input noise lower than that of the detectors should allow for reaching the thermal noise limited values.

The operation frequency of these detectors is not limited to 240 GHz . In fact, with a CW setup we successfully tested the detectors also between 500 and 700 GHz . At higher frequencies, the responsivity is lower and the diffraction-limited spot size is smaller, reducing the maximum linearity range. A silicon lens may not be necessary any more as the diffraction limited spot becomes smaller as the frequency increases. By removing the silicon lens, the high linearity range may be recovered.

IV. CONCLUSIONS

We developed and characterized field effect transistors for detection of high power, few ns THz pulses with a linearity range up to 11 W and a destruction threshold above 200 W at 240 GHz. Three different device layouts without antennas were tested and the devices were individually packaged for simple handling. A bandwidth-normalized dynamic range of $61 \text{ dB}/\sqrt{\text{Hz}}$ was achieved. The NEP was $0.2\text{--}6.5 \mu\text{W}/\sqrt{\text{Hz}}$, depending on the device layout and limited by the post-detection setup. These values were confirmed by CW measurements at 300 GHz. Extremely short detector time constants (much) below 11 ns, limited by the time resolution of the pulse, could be demonstrated. Such detectors can fill the gap in fast THz detection between ultra-sensitive but low damage threshold detectors such as Schottky diodes, and high power but insensitive detectors such as the photon drag detector. The responsivity and maximum tolerable input power can be tailored to user-specific needs by the density of electrodes.

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⁹See a list of commercial THz detectors for high power pulses at <http://www.physik.uni-regensburg.de/forschung/ganichev/english/fields/advertisment/pdd99.pdf>.

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